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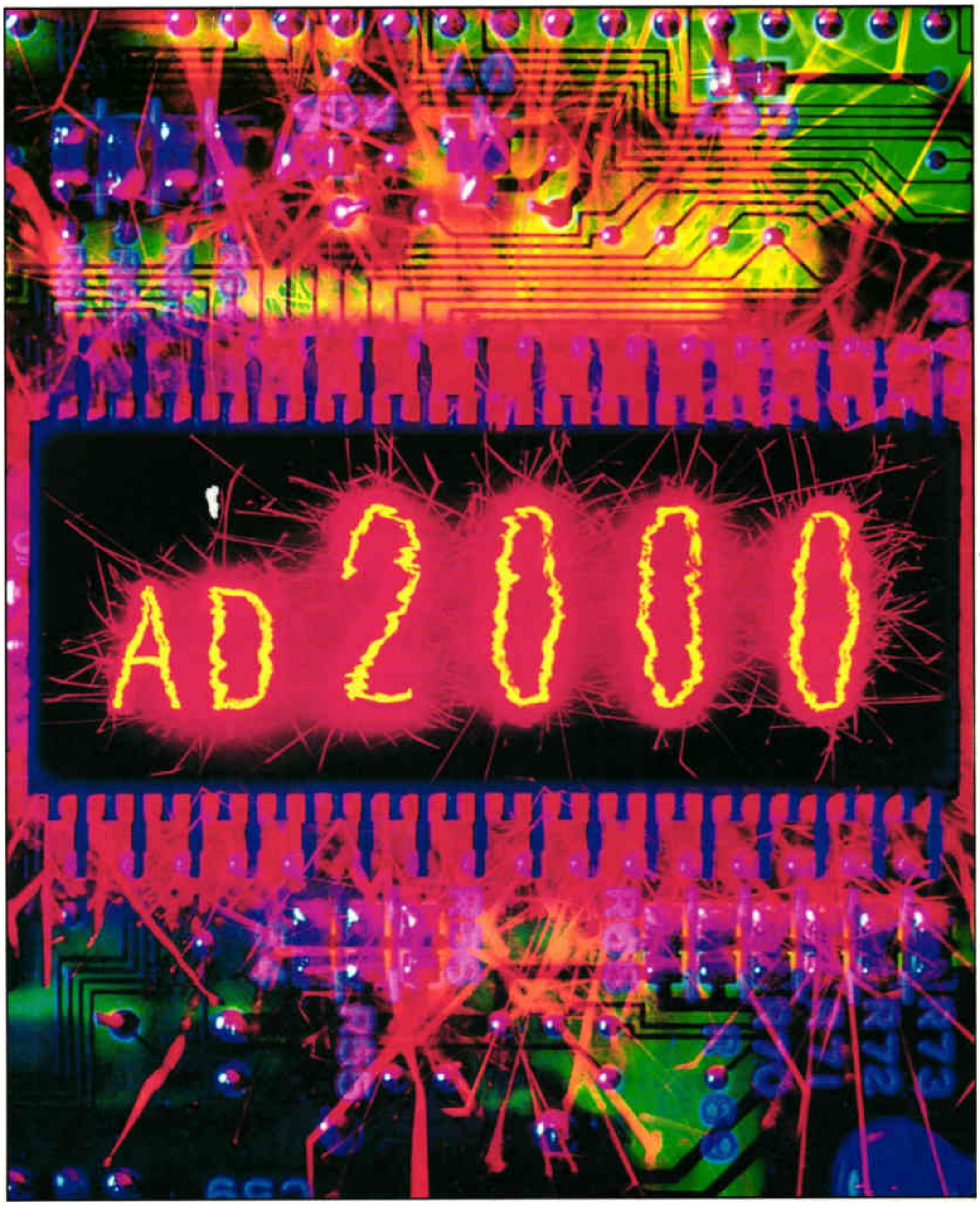
Year 2000 – is your PC safe?



High-level rf mixers

The perfect If oscillator?

Neural nets, fuzzy logic

Rfi sources and solutions

Aluminium and tantalum caps

400Hz in 3-phases

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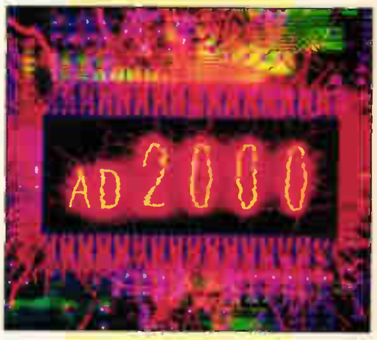
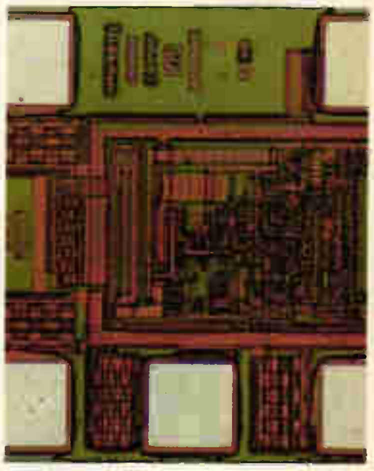
Contents

There's a number of active high-level rf mixer types but passive alternatives like this one still have applications in modern receiver designs. Find out why on page 464.



Photography Nicholas Rigg

What can neural networks and fuzzy logic tell us about the future of digital electronics? See page 512.



Are LM300 series op-amps dead? Find out on page 490.

JULY ISSUE ON SALE 29 MAY

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REED BUSINESS INFORMATION

Our life is frittered away by detail...

When Windows entered my consciousness years ago, I was terrified of it, because I thought I might disable the beast and not be able to recover. But there was no trouble; Windows 3.1 was robust, there was an enormous and well written manual and the help line worked as promised. All went well.

But that was then; this is now and I have had to "upgrade" – a marketing expression meaning "to shell out for a new computer" – a 50Mbyte hard disk, 33MHz 386 and 4Mbyte of ram being a bit limiting. This new marvel was already loaded with Windows 95 and accompanied by a great tribe of cd-roms, many containing advertisements and games that anyone aged over nine has so far found totally incomprehensible. Telephone support – a major selling point – was promised but there were no manuals, since the "help" text was said to be enough. Ha!

It has been my subsequent experience that Windows 95 is the flakiest system yet to be foisted on a receptive public. It is highly susceptible to wobbly mouse operation. If your aim is at all uncertain, or the simple truth is that you don't really know what you are doing, the most entertaining things can and do happen. "Shortcuts" appear on the desktop unbidden – they do on mine, at least – files turn up from nowhere, settings change without any intervention from me and the machine has an infuriating habit of locking up.

Then there is its apparent delight in performing "illegal" operations and implicitly blaming me for them; following the screen instructions after such a hiccup doesn't work, so you have to reset the computer, which means another run of the disk-scanner before it will consent to do any more work, having lost any you haven't saved. And the start-up menu: whoever decided to put Suspend next to Shut Down ought to be suspended and put down, preferably painfully. You can, of course, hide it, but you have to find that out after the first stress-inducing episode of blankness.

Casually ringing the ambitiously named help-line did little towards postponing an incipient seizure. I tried to speak to someone for three days, but was faced with a recording of a silver-tongued young lady every time, explaining that they were busy and would I please hold on.

I tried a computer "expert", but found myself listening to Mozart, who is wonderful, but as a Windows counsellor not particularly helpful. So I wrote to my supplier. That is many weeks ago and all I have yet received is a note to say I will shortly hear from them. I am becoming used to this *prima donna* of an operating system and the cynical way in which it and computers are now marketed, but cannot help feeling a little resentful.

Windows 95 gives me the impression of being hopelessly over-engineered, providing endless different methods of doing just about anything. It appears that programmers write facilities into it because they can be done, rather than because users need them. Maybe the programmers should be taught that simplicity is usually the winning approach in any branch of engineering.

On-screen help facilities are good, but they cannot

replace well produced manuals, particularly if there is nothing on the screen. An introductory booklet that came with the computer – a very slim volume indeed – is all right as far as it goes, but it doesn't go anywhere near far enough. If the simpler Windows 3.1 needed a tome, why doesn't this one?

Most of the cd-roms had little to say about themselves apart from a bit on the cover and some fairly off-hand help on screen.

Cost is, I assume, the reason for a lack of manuals, but I do think an effort should be made to provide something a little more helpful than the booklet on Windows 95 that is currently sent out, particularly some help on how to recover from something disastrous like a blank screen after the boot-up (the F8 procedure at "Starting Windows 95" gets you out of that one).

Clearly, the computer sellers' army of helpers is hard put to deal with the clamour for assistance and there is no reason why they should have to; it is the responsibility of people like Microsoft to ensure that there is enough printed instruction to enable customers trying to earn a living using Windows to do so without getting themselves in a fix.

Maybe the next version of Windows will be a bit simpler and a lot less vulnerable, but it seems unlikely.■

Philip Darrington

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Internet access via the three-pinned plug?

Nortel's technology for delivering Internet access over the mains power cables will be offered by several power utility companies by the end of this year.

"We have a considerable number of customers who are committed to feasibility trials," said Dr John Laycock, head of the project at Nortel in Maidenhead. Commercial systems could be in place by September, he claimed. "We'll have announcements next week for content providers and partners," Laycock added.

Companies that install the system will be able to offer a wider range of services than just Internet access. "We intend to do voice in the future," Laycock said. This would replace the BT land line, and could offer several phone lines at once.

Another application, aimed at small businesses, will back up the computer's hard disk every night. And since the user is permanently connected to the network, a digital camera can be added for surveillance of the home or office.

A production version of the Digital PowerLine technology was announced by Nortel at CeBIT.

"Last year we announced a first generation product for verification and trials," said Mark Main, the project's marketing manager. This was based on a card in a pc, but restricted the technology to consumers privileged to own a computer, said Main.

The new unit is separate from other equipment and can be used with Web browsers and network computers as well as standard pcs.

Base-station units will be located at transformer substations, modulating data onto the mains.

Each base station handles a 1Mbit/s data stream. Houses or businesses that have a link onto the system share the bandwidth.

• *See late news on page 456.*

New light on charging NiCd cells

Careful circuit design around NiCd batteries can be difficult, but not as troublesome as problems caused by early cell death.

Simple precautions can extend the working life of rechargeable batteries dramatically. "You mustn't forget that most of the public don't have a clue about battery technology. If they are going to operate your product, it needs protecting from unintended abuse," said John Hutchinson, engineering director at BayGen, the wind-up radio company.

The company has just released a product with an internal NiCd battery pack and Hutchinson has made it as idiot-proof as possible.

"There are essentially two ways to damage a NiCd battery, overcharging and over-discharging them. Most commercial products with NiCds, like hand vacuum cleaners and torches, do both," said Hutchinson.

Overcharging , according to Hutchinson, results in electrolyte evaporation and dendrite growth – dendrites are metal crystals that short out the cell internally.

"The majority of simple products charge their batteries continuously at C/5. This abuses them once they are fully charged and they don't last long," said Hutchinson.

Conventional wisdom is that C/10 is a safe rate for continuous charge, but Hutchinson has come to differ: "C/10 may be all right for some hours, but charging at this rate for hundreds of hours damages cells." He has found C/30 completely safe as a continuous charge current. "I arrived at this figure initially by talking to a number of cell suppliers. But it is hard to get a consistent story. The most that the makers agree on is that C/10 is worse than C/20."

C/30 takes around 40 hours to fully charge a flat cell. This is an unacceptably long time for BayGen's application, so some form of controlled fast charge was called for.

It chose C/5 as a fast charge rate, terminated by a six hour timer. "AC/5 current completely charges the battery in six hours, then the charger switches to a C/30 trickle charge."

Using a timer is not ideal, plugging the equipment in with an already full battery results in it being somewhat overcharged for six hours.

"Sensing battery parameters like temperature and pressure to determine when charge is complete is one way, but the sensors add a lot of cost. Other schemes use voltage sensing which isn't too expensive in hardware, even though you need a microcontroller."

Problems also occur during discharge when more than one cell in a series-connected battery runs out of charge before the others. "We detect the battery voltage and disconnect the load with a transistor when it drops below 1V per cell,"said Hutchinson, "this means that no cell ever gets reversed."

By limiting overcharge and overdischarge Hutchinson has made a product that is unlikely to be returned because its battery expired within the warranty period.

Steve Bush, *Electronics Weekly*

Discharge and charge rates of a battery are usually expressed as a ratio of its capacity. Discharging at C/10, drawing 400mA from a 4Ah battery for instance, empties a full battery in 10 hours. As charge efficiency of NiCd technology is around 1.4, charging an empty cell at C/10 completely fills it in around 14 hours.

Clearly a torch, but what's the handle for?

Offender tracking has a £100m tag

Electronics firms are being asked to tender for four major contracts, worth collectively over £100m, to extend electronic tagging of offenders throughout England and Wales.

Home Office Ministers have decided that system tests in areas such as Greater Manchester, West Yorkshire, Berkshire, Middlesex and Cambridgeshire have proved a success.

Despite teething troubles, Home Secretary Jack Straw has decided that the system – which has shown an 80 per cent success rate for 1000 persistent petty offenders and people on bail – is ready to go nationwide.

Now his Minister of State, Joyce Quin, is to let four regional contracts from 1999 for systems of a wrist/ankle tag and a receiver linked to the telephone system.

She said: "Companies from a wide range of relevant sectors including the security, telecommunications and electronics industries, are being encouraged to consider the business opportunities created by the increased use of tagging."

The tender details that the four regional contracts – which will let for five years with a two year extension option – will cover a minimum of 4500 offenders and be worth at least £100m.

The two firms involved in the existing trials – Securicor and Geographics – are favourites for the contracts but a Whitehall source stressed that it was an open competition for any company with the necessary technical expertise.

Firms can bid for one, more or all of the regional contracts by the closing date of April 21.

US skills crisis a myth?

US labour leaders have attacked high-tech companies' claims that there is a serious IT skills shortage in the US.

The Department for Professional Employees (DPE) has sent a letter of protest to the US Congress complaining that US technology firms' real motivation is not looking for workers but paying lower wages.

"There is no proven crisis regarding the demand for IT workers that justifies the drastic action of filling these jobs with foreign workers," said DPE chairman Morton Bahr. "We should be very careful about IT employers crying wolf just to enlarge the labour pool, depress salaries and benefits, and undermine working conditions."

Are 2000 solutions enough?

The DTI is stepping up its awareness campaign for the Millennium date change problem and calling on all sectors of the electronics industry to support its initiative.

"The electronics industry has a pivotal role to play," said Ian Edison, acting director of the DTI's Action 2000 programme. The focus for this activity, which will involve trade associations and institutions like the IEE, is the problem of finding and providing a fix for all embedded clocks.

"It is getting pretty close to the wire for many large companies,"said Peter Wilkinson of Action 2000.

Virtual Recruitment Fair – free entry

Electronics World's sister publication *Electronics Weekly* has produced the first ever Recruitment Fair on CD Rom.

The 'Virtual Recruitment Fair' uses 3D graphics and sophisticated animation techniques to bring a whole new dimension to job seeking for electronics employees.

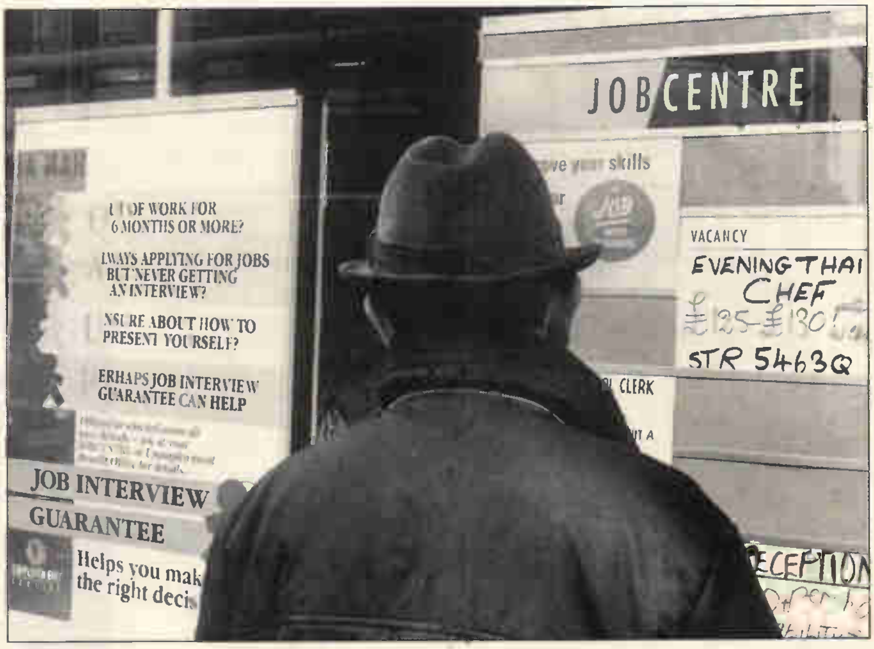
Once inside the prestigious life-like venue, you can complete your CV at registration, pick up a virtual carrier bag then visit the impressive array of interactive company stands in the Main Hall.

Take a seat in the big-screen cinema or enjoy a moment of tranquillity in the reading room while you read back issues of *Electronics Weekly* from the archives.

From the exhibition stands themselves, you can: view – and respond to with CV – well over a thousand live vacancies; connect to advertiser web sites; view and print company brochures; launch slide presentations and watch video footage to find out exactly what it's like to work for a prospective employer.

Recent research illustrates that some 84 per cent of UK electronics professionals have access to CD-Rom facilities – over 30 per cent more than have access to the Internet.

The Virtual Recruitment Fair CD-rom is distributed free to UK *Electronics World* subscribers with this issue. To obtain your free copy – whether you are in the UK or overseas – write to Grant Allaway, Recruitment sales, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS or fax your request on 0181 652 3400. Alternatively, e-mail grant.allaway@rbi.co.uk. Be sure to mention that you are an *Electronics World* reader.



Job hunting? Got a CD player? Then you can take a break from browsing the notice board. Check out hundreds of electronics-related vacancies at the Virtual Recruitment Fair – for free, whenever you want, and in the privacy of your own home.

Ten sign up for mains Internet

Ten public utility companies in Europe and Asia have signed up to use Digital PowerLine (DPL), the Internet-over-the-mains technology.

Nortel and Norweb Communications, which developed the technology, have formed a joint venture company called NOR.WEB to further market DPL.

Norweb is itself one of the ten utilities committed to installing DPL. Others include Vattenfall and Sydkraft in Sweden, RWE and EnBW of Germany, and Singapore Power. These companies provide electricity services to 35 million people, all of whom could potentially use DPL.

"We had an unprecedented response to the technology," said Steve Pusey, chief executive of the joint venture. The company received some 1,500 enquiries when DPL was unveiled last October. "We narrowed these down to 40 serious leads," he said.

The remaining 30 companies, located around the world, are currently in discussion with NOR.WEB with a view to introducing the fast Internet technology.

Digital PowerLine uses the existing mains cables to link homes, small offices and schools to the Internet at 1Mbit/s. "This is 20 times faster than the fastest dial-up modem available today," Pusey said.

NOR.WEB expects the first services to roll out by September, with up to 2000 homes connected by the year end.

Big brother back-off

The US government appears to be relaxing its position on the use of powerful encryption products.

Representatives of the FBI and the Department of Justice testified before the Senate subcommittee saying they would not seek key escrow legislation which would allow the government to read encrypted messages.

The subcommittee also heard from senior high-tech executives. "Strong encryption is the key to privacy on the Internet," said MCI Communication's chief lawyer Tim Casey. "Such privacy, in turn, is the key to realising the enormous potential of the Internet and global electronic commerce."

US high-tech companies have proposed a voluntary, industry-led approach to encryption policy rather then government controls or laws.

Intel outside?

Chief pretenders to the Intel throne, AMD and Cyrix, previewed microprocessors with a 100MHz system bus at CeBIT.

AMD's offering is the K6-3D, to be launched officially on May 27. Along with the system bus speed increase, it has improved floating point performance to speed up 3D graphics. "The 3D technology is 24 new instructions," said AMD's Claes Berglund. They will do for floating point what MMX did for integer data, he said.

Cyrix has yet to adopt the instructions. The two companies have first to agree whether to use the same set. "The plan is for Cyrix and IDT to use the same instructions as us," said Berglund. By early next year, AMD plans to release the K6-3D+.

"This integrates the level two cache on chip which will operate at the processor clock speed," said Berglund. He claimed the device will match the performance of Intel's Katmai processor.

Cyrix meanwhile seems more concerned with pushing for cheaper devices, with its 6x86MX and MediaGX processors. "In 1998 we will see a $599 price for a pc," said Forrest Norrod, a Cyrix senior director. "The same phenomenon which swept the consumer market in the US last year will hit Europe and Asia this year."

As with AMD, getting devices manufactured is Cyrix's biggest headache. "Volumes are critical - on our own Cyrix was limited to five or six million units a year," said Steve Tobak, v-p of marketing at Cyrix. Having been acquired by National, Cyrix's fortunes may change.

The 6x86MX is currently running through a 0.25μm pilot fab at National.

Hull to get fast ADSL phone link

Kingston Communications plans to offer its customers high-speed interactive services this year using asymmetric digital subscriber line (ADSL) technology.

The scheme, involving up to 170 000 customers in the Hull area, is an extension of a successful pilot scheme.

Huw Saunders, Kingston Communications' technology development director, estimates that the cost of implementation will exceed £10m, although this figure is dependent on customer uptake. "It is a significant investment for us," he said.

The ADSL network will offer a 2Mbit/s transfer rate with some services provided at 6Mbit/s.

A modem fitted in the home will provide a standard interface to a pc or set-top box. Initially the modem will be part of the network and be supplied to customers by Kingston Communications in order to retain control over the network. As standards are agreed within ADSL, other connection equipment will be used. Charges for services are still under review but it is likely that payment will be on a subscription basis.

Meanwhile, BT are pressing ahead with plans for a trial of ADSL this summer involving 2000 homes in west London.

Government reacts on low r&d spend

The government seems to have hit the panic button on the UK's high technology r&d performance. With r&d investment in the civil sector having fallen as a proportion of gross domestic product for three consecutive years, the Treasury has produced a consultative document to demonstrate its commitment to using tax incentives to promote r&d investment in UK firms.

According to the latest official figures, the UK's investment in research and development fell by two per cent in real terms in 1996, the third year in a row that r&d investment has fallen.

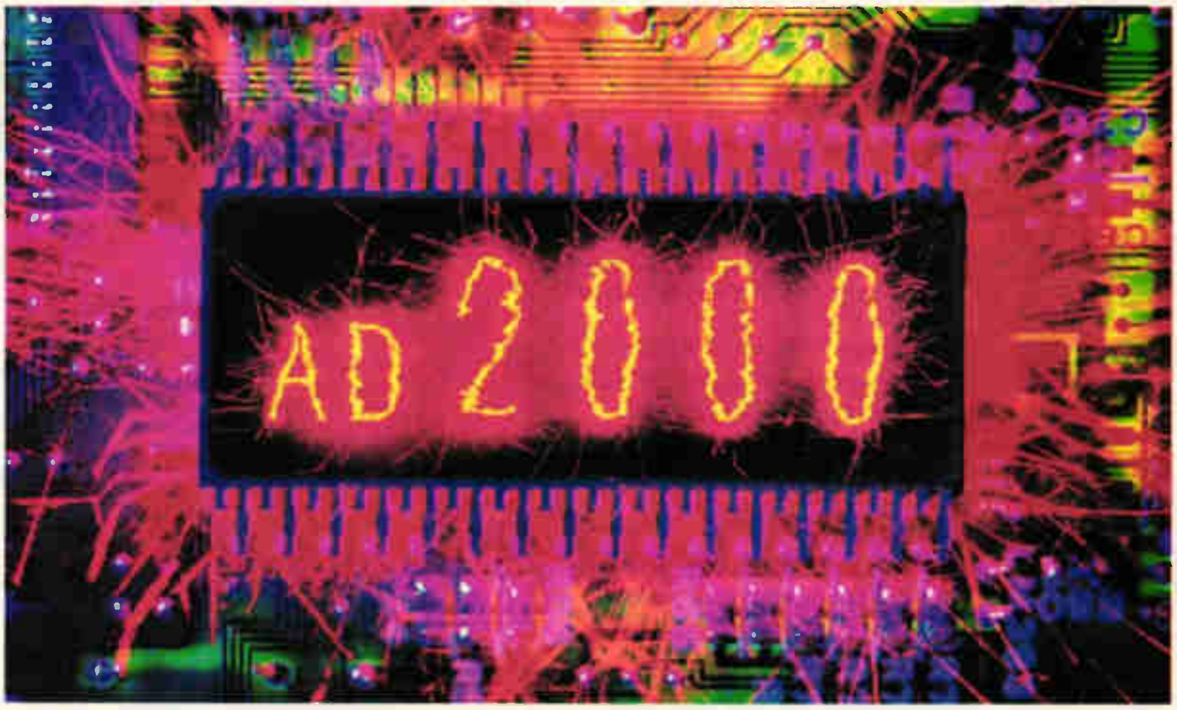
DSP like a microcontroller?

Texas Instruments has announced a single-processor dsp chip with microcontroller-like capabilities.

"This is a dsp designed to perform well in areas where microcontrollers are strong, like code density, support for the C language, and i/o and interrupt handling," said Jean-Marc Darchy, TI's European marketing and applications manager. "Our new core has ten per cent better code density running C than the *ARM7TDMI*, which is the industry benchmark."

Called the *TMS320C2000* architecture, the dsp core is said to be all-new. It will offer 100Mips performance at 3V using a 0.25μm process. TI is reluctant to talk about power consumption in cores but Darchy said: "The [older] C20x family consumes 1.1mA/Mips. The 2000 will be less, perhaps 0.8mA/Mips."

The first product will be a dedicated hard disk drive processor. "We are waiting to see how the industry receives this before we decide what other applications to aim the core at," said Darchy. ■



Year 2000 debugged

Robert Harcourt looks at how the Year 2000 changeover will affect your pc – if at all.

First the good news. New Year's Eve in 1999 falls on a Friday. This means that companies can shut down over the weekend to convert their pcs to cope with the year 2000. Another piece of good news is that pc real-time clocks usually implement the fact that the year 2000 is a leap year – although 1900 was not.

Apple Macintosh computers have not shown a problem with the year 2000 since at least Mac OS version 7. IBM pc compatibles on the other hand contain four or five system levels within which problems can occur with the date. At the lowest level is the real-time clock, which runs on a stand-by battery so that it does not stop running – in theory at least.

The real-time clock passes its date on to the basic input/output system, or BIOS, which only starts up at the moment the computer is switched on or re-booted. The real-time clock, and hence the BIOS, in many cases rolls over from midnight 1999 to midnight 1900, represented by 00:00:01.

At the third level is the operating system. Windows 3 Windows 95 and DOS, all interpret their version of the date from the date in the BIOS. However, they implement an algorithm so that no date before 1980 is possible. The RTC, BIOS and DOS clocks are shown in **Table 1** at a normal time and date before the year 2000.

Table 2 shows these same clocks after being allowed to roll-over – while the computer is left switched on – on new years eve to 2000. You can see that the DOS clock shows correctly the year 2000 but the real-time and BIOS clocks show 1900.

When the computer is switched on, BIOS starts up and loads DOS. At this point it is possible to invoke the DATE command. The DOS date command forces the BIOS and RTC year, **Table 3** into place if they are awry.

This is a complete solution for DOS, but it does not mean that the next level up – i.e. the application program – will function correctly. If your computer runs under DOS, and automatically runs an application program at start up, you will not have the opportunity to change the date. If this is the case, place the simple statement DATE in the file AUTOEXEC.BAT, using a text editor such as EDIT. In this way, each time the computer is switched on, the current date will be displayed and you will have the opportunity to correct it.

The disk operating system asks:

```
Enter new date (mm-dd-yy):
```

Although the variable *yy* implies only two characters for the year, in fact the year input by the user will be accepted as four characters. If the date displays correctly, just press the return key.

This measure is thought to be necessary only once, as roll-over to 2001 a year later is usually correctly performed.

The fourth and fifth levels

The fourth level involving clock information is application software or Windows running application software. Windows 3 and, to an extent, '95 relies on the disk operating system, and uses the DOS date. The fifth level is application software which may or may not make use of the date supplied by DOS. For individual users of pcs, all that is required is to invoke the DATE command at the C: prompt of DOS and correct any problems with the operating system DOS date.

With Windows '95 it is necessary to go to the Control Panel and enter a four-digit date. This is because, whether the machine is left switched on or off overnight at the Millennium, the hardware real-time clock will fail to roll-over to the year 2000 and will set the date to the 1st of January 1900, while the operating system clock goes to the 4 January 1980 for at least some older and not so old pcs, **Table 4**.

Interestingly, if the pc is left switched on, the software in the DOS clock will in many cases roll-over correctly to 2000, Table 3. But at the first boot operation, it will revert to 1980, Table 4.

This is a potential trap for anyone trying to test for the problem, because although DOS initially has the correct date, the real-time clock and BIOS have 1900 as the date, and 'steal' the 2000 date from DOS at boot-up. This implies problems for servers or other machines which are left switched on all the time. They may appear to be working without bugs until their next re-boot.

Planning ahead

For those of you who are planning ahead because you foresee problems with the Millennium changeover, a free diagnostic test is available at the web site http://y2000fix.com.uk. The firm operating this site is Year 2000 Consultants Limited. The company has a second site at http://Y2KPCPro.com.

Part of the free software is a program called *ViewCMOS*. This package allows you to observe the real-time, BIOS and DOS clocks at the same time, and in real time. The same company sells a fix program for those machines failing their test, called *Y2KPCPro*.

Table 1. How the three clock systems in a pc view time.

Clock system	Mnemonic	Date	Clock system time
Real-time hardware clock	RTC	20-03-1998	12:15:49
Basic input/output system clock	BIOS	20-03-1998	12:15:49
Operating system clock	DOS	20-03-1998	12:15:48.45

Table 2. How the three clock systems read after rolling over to 2000 while the computer is switched on.

Clock system	Mnemonic	Date	Clock system time
Real-time hardware clock	RTC	01-01-1900	00:00:01
Basic input/output system clock	BIOS	01-01-1900	00:00:01
Operating system clock	DOS	01-01-2000	00:00:00.98

Table 3. Entering a value after the DATE command in DOS forces all three clocks to the required time.

Clock system	Mnemonic	Date	Clock system time
Real-time hardware clock	RTC	01-01-2000	00:34:12
Basic input/output system clock	BIOS	01-01-2000	00:34:12
Operating system clock	DOS	01-01-2000	00:34:12.18

Table 4. All may be ok until you boot – at which point, the clock reverts to 1980.

Clock system	Mnemonic	Date	Clock system time
Real-time hardware clock	RTC	01-01-1900	00:33:02
Basic input/output system clock	BIOS	01-01-1900	00:33:02
Operating system clock	DOS	04-01-1980	00:33:01.98

Other firms selling year 2000 software for pcs are Greenwich Mean Time Inc., Euromax Electronics Ltd, Deeside Technology Ltd, Ninety Nine 2000, and Computer Experts Ltd. *Which* magazine has carried out a review of the products from these firms in its November 1997 and January 1998 issues.[1]

Are you Year-2000 compliant?

So far I have considered only the hardware and operating system aspects of the Year 2000 changeover. Problems with the software are a different story – especially where networks are concerned.

Some firms are spending thousands throwing away non-compliant pcs and networks in favour of Year-2000 compliant alternatives. But just what is meant by Year-20000 compliant?

The British Standards Institution has published 'Definition of Year 2000 Conformity Requirements.'[2] It states that, "Year 2000 conformity shall mean that neither performance nor functionality is affected by dates prior to, during and after the year 2000." There's more on this in the panel.

Other requirements cover the prohibition of the use of the date as an end-of-file marker. The sequence 9/9/99 is said to have been used by programmers as an end-of-file marker, or EOF, clearly limiting the life of the system.

The inferencing rule mentioned in the panel will not fit all situations: for example, if birth-dates are maintained

Year 2000 defined

This is an extract from the BSI publication, 'Definition of Year 2000 Conformity Requirements,' document reference DISC PD2000-1, explaining what Year 2000 compliance means:

Rule 1 No value for current date will cause any interruption in operation.

Rule 2 Date-based functionality must behave consistently for dates prior to, during and after year 2000.

Rule 3 In all interfaces and data storage, the century in any date must be specified either explicitly or by unambiguous algorithms or inferencing rules. An example of an inferencing rule is "e.g. two digit years with a value greater than 50 imply 19xx, while those with a value equal to or less than 50 imply 20xx.

Rule 4 Year 2000 must be recognised as a leap year.

in a database, and a particular person was born in 1948, the inferencing rule would change their birth-date to 2048, which could mean they died before they were born.

Data files with dates in them must specify the century when there is any sorting applied to chronologically sequence the data, otherwise 2000 – appearing as 00 – would be sorted before 1999, which appears as 99.

This could mean that a lot of re-writing of data-base or accounts software is required. DOS 6 and 6.22, Windows 3.1 and Windows 95 all carry the date of creation of files in the directory on pc's hard drive. These dates appear as only two digits, so the backup and archiving of files could become confused.

In summary

The date mechanism in pcs is not simple, and manual checks have to be meticulously carried out. In particular, it is necessary to check the date at switching on and on re-booting while switched on. It is necessary to do this at least twice to make sure the 21st century is not lost.

If you are using a pc that is not connected to a server, you will find it easy to correct any date problems with the DATE command. ■

References

1. *Which* , PO Box 44, Hertford SG14 1SH, November 1997 and January 1998 issues.
2. BSI, 389 Chiswick High Road, London W4 4AL: 'A Definition of Year 2000 Conformity Requirements.' Document reference DISC PD2000-1. Codes of practice are detailed in DISC PD2000-2 and a general outline of the year 2000 problem is presented in DISC PD2000-3. See www.bsi.org.uk/disc/year2000.html.

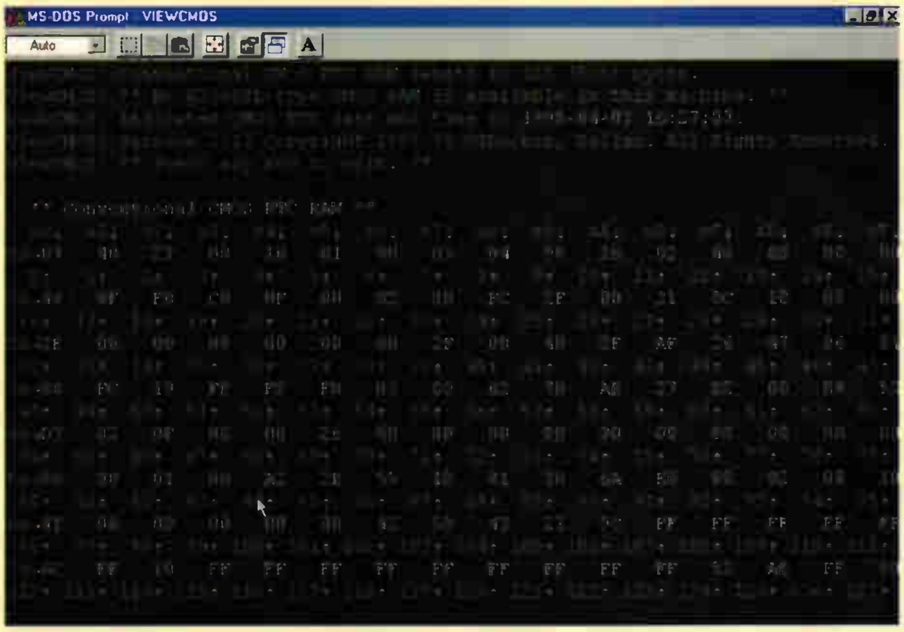
Why *Test2000* thinks its best

Other testers of this sort test the CMOS real-time clock, which essentially always fails to increment from 1999 to 2000. What is important is not the CMOS rtc date itself, but what the BIOS does with the CMOS RTC date. This is because the operating system uses the BIOS date and some applications use the BIOS date.

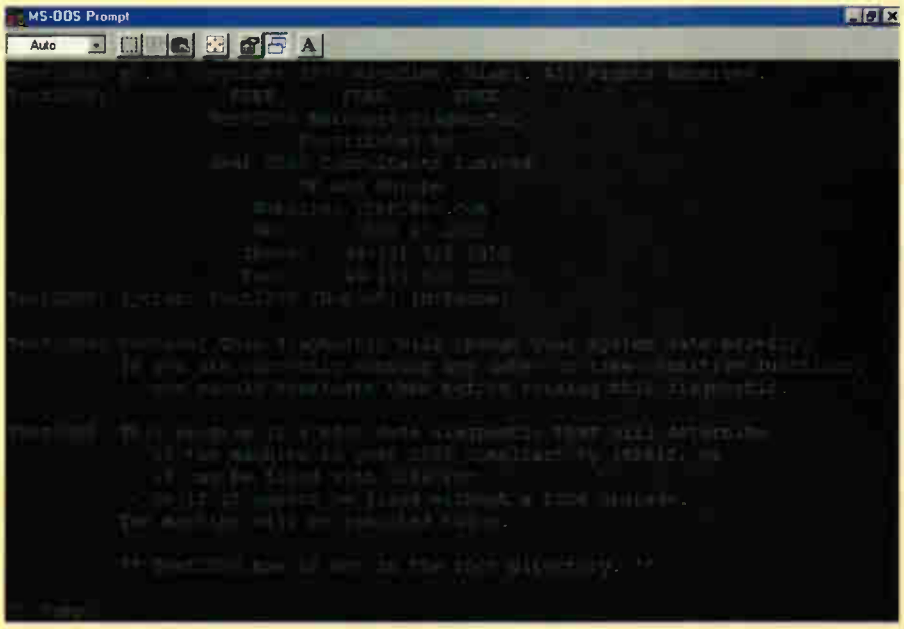
Although a fundamental design error that causes the year 2000 transition problem exists in the CMOS rtc, the CMOS rtc date itself is unimportant because no common application software uses it directly, so testing the CMOS rtc is immaterial, unnecessary and misleading. The BIOS date is what matters and despite its pervasive century error, the CMOS rtc date doesn't matter.

Other testers also test the operating system 1999 to 2000 increment, which will always pass. There is no reason to suspect an operating system date year 2000 failure. Windows 3.1 File Manager does odd things to the year in its displays, but those are only display artifacts; the underlying date is handled correctly.

Other testers also look at a variety of leap year concerns, none of which need any concern. No pc hardware leap year failures have ever been demonstrated, and there is no reason to suspect that pc hardware leap year errors exist. If another tester fails a leap year test, it is probably the testing program that has failed – not the hardware.



ViewCMOS allows you to see what's in your pc's CMOS to help you determine what your hardware real-time clock is going to do at the Year 2000 changeover. It shows the clock values update in real time. It is only free "when used for Year 2000 testing by computer users" so don't look at the non rtc locations!



Test 2000 is a free Year 2000 diagnostic tool. It checks whether you are going to have a problem, and whether or not the problem can be avoided by using the company's rectification tool Y2KPCpro – which isn't free. If that won't work, you are in for a BIOS upgrade.

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Audio signal generator

AG2601 is a portable mains-powered instrument covering 10Hz to 1MHz in five overlapping decades. Sinewave distortion between 500Hz and 50kHz is just 0.05%.

The AG2601 audio signal generator spans 10Hz to 1MHz in five overlapping ranges and features floating output and low distortion. This stable sine and square-wave oscillator is being made available to *Electronics World* readers at the fully-inclusive special price of £129. Its normal selling price is £129 *excluding* VAT and delivery.

Please use the coupon to order your signal generator, and address all correspondence relating to this order to Vann Draper Electronics at Unit 5, Premier Works, Canal Street, South Wigston, Leicester LE18 2PL, fax 0116 2773945 or tel. 0116 2771400.

AG2601 audio generator – specifications

General	
Frequency range	10Hz to 1MHz
Frequency stability	within ±2Hz
Output waveforms	sine, square
Output impedance	600Ω
Accuracy	±5%+2Hz, 10Hz-1MHz
	±3%+2Hz, 100Hz-100kHz
O/P floating voltage	within ±1.5dB
Sinewave characteristics	
Distortion	<0.05%, 500Hz to 50kHz
	<0.5%, 50Hz to 500kHz
Output voltage	8V rms, max
Output flatness	±1.5dB (1kHz)
Output impedance	600Ω
Squarewave characteristics	
Output voltage	15V pk-pk, min
Rise time	0.5μs
Synchronization input	
Input impedance	10kΩ
Maximum input	10V rms
Supply	
115/230V, 50/60Hz	
Physical data	
Dimensions	150 by 250 by 130mm
Weight	2.5kg

*Test leads supplied as standard

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*Overseas readers can also obtain this discount but details vary according to country.

Darren Conway* helps you decide which is the right mixer option for your receiver design.

High level rf mixers

The vast majority of radio receivers are based on the super heterodyne principle, where incoming rf signals are mixed with a local oscillator to produce an intermediate frequency. In most cases a receiver uses two mixers to convert from rf to the desired base band signal.

One of the key design choices for a receiver is the selection of mixers – and in particular the first stage mixer. The development of rf mixers over the past forty years has been a process of evolutionary development and refinement. Circuits have improved but the basic principles and techniques have remained largely unchanged.

The aim of this article is to review a representative sample of the current range of high level mixers and discuss the characteristics of each.

Discussion is limited to mixers used up to vhf and those that are readily available. You are encouraged to view the relevant data sheets and references in conjunction with this article as they contain a wealth of information that is not included here.

Although this discussion is limited to vhf, the basic principles and mixer characteristics are applicable at all frequencies. The purpose is to provide you with the information required to select the right mixer for the right application.

Why have a mixer?

The primary application of a mixer is to add an input frequency and a local oscillator frequency to obtain a single intermediate frequency, or IF, output.

Ideally, a mixer output would only contain the desired output frequencies, $F_{RF}+F_{LO}$ and $F_{RF}-F_{LO}$, where F_{RF} is the received rf signal and F_{LO} is the local oscillator input. But no such device exists.

In reality, mixers produce a whole range of frequency products

$$\pm mF_{rf} \pm nF_{lo} (n=0...\infty, n=0...\infty)$$

at the IF output. The unwanted harmonics are filtered out.

The balance of mixers

There are effectively three types of mixers which may all be implemented as passive or active. The basic passive circuits are shown in Figs 1, 2 and 3. For active mixers, the diodes are replaced with fets, mosfets or transistors.

An unbalanced mixer shown in Fig. 1 has no isolation between the two input ports and the output port. They are rarely used because lack of isolation between the local oscillator and the rf input results in unwanted transmissions from the antenna at the local oscillator frequency. These transmissions may be strong enough to contravene emc regulations, and even if they are within legal limits, they are still undesirable.

Detection of local oscillator transmissions has been successfully used in the past to locate clandestine or illegal receivers. The main advantage with unbalanced mixers is that they can operate over a very wide frequency range spanning greater than five decades. In certain applications, this may justify their use.

Single-balanced mixers go some way to solving the problem of unwanted transmissions by providing isolation between the local oscillator and the rf input as shown in Fig. 2. In addition, the local oscillator signal is suppressed at the intermediate-frequency output.

However, like the unbalanced mixer, the rf input appears

*Darren Conway, Lieutenant Commander, ME(Elect), MIPENZ, MIEE, C.Eng, psc, RNZN

at the IF output port. If it is decided to fabricate a mixer from discrete components, then the single balanced mixer offers a reasonable compromise between the superior performance of the double balanced mixer, and the simplicity of the unbalanced mixer.

Double-balanced mixers ideally offer infinite suppression of signals between the three ports. The only output should be the intermediate-frequency signals ($\pm mF_{RF} \pm nF_{LO}$) with the rf and local oscillator inputs fully suppressed.

In practice the local oscillator and the rf input are typically suppressed by about 50dB at the intermediate-frequency output. The ready availability of single package double balanced mixers often means that they are simpler to apply than either single balanced or unbalanced mixers. In most applications, the double balanced mixer shown in Fig. 3 will provide the best performance.

What makes a good mixer?

Practical mixers are complex to analyse and their performance is defined by a number of characteristics. The following define the major specifications of a mixer in order of importance.

Frequency range. Mixers are usually required for all receivers operating from very low frequency to tens of gigahertz. Typical mass-produced mixers operate to maximum frequencies in the range of 100MHz to 2.5GHz. The operating frequency range is a fundamental design characteristic that will in part determine the final selection of mixer type.

Dynamic range. This is one of the most important specifications for a mixer. The massive proliferation of rf transmitters and other interference sources means that the modern radio receiver is usually operating in the presence of significant rfi.

Even if the desired signal is always very weak – for example satellite transmissions – it is still important that a receiver has the ability to operate in the presence of strong signals so that the desired weak signal is not lost.

The lower limit of dynamic range is defined by the noise figure, while the upper limit is defined by the compression point, intermodulation products and burnout level.

Noise. Typically, mixers have noise figures ranging from 6dB to 20dB. The noise figure for passive mixers is about equal to the insertion loss. The noise figure for active mixers depends on the selected devices and circuit topology. It is usual but not essential in receiver design to include a low-noise rf amplifier ahead of the first mixer in order to improve the noise figure of the receiver system.

Gain. The ready availability of amplifiers that cover the radio-frequency spectrum means that the mixer is not generally required to have any gain. Excess mixer gain can reduce the dynamic range of the receiver.

In most cases, insertion loss through a mixer is also undesirable particularly in the case of passive mixers. Active mixers provide gain in the range of –1dB to +17dB while passive mixers have a typical insertion loss of between 5.5dB and 8.5dB.

Local-oscillator drive. The ideal mixer would be insensitive to both local oscillator level and harmonic content but in reality, the local oscillator specifications need to match the requirements of the mixer.

Passive double-balanced diode mixers require local oscillator levels from +7dBm up to +23dBm. Active mixers require local oscillator output levels ranging from –20dBm to +30dBm depending on the selected type. The design of the local oscillator is intimately related to the selected mixer type.

Isolation. Isolation is a measure of the mixer's ability to prevent a signal applied to a port, appearing at either of the other two ports. The only output should be the mixer products at the intermediate-frequency port. The degree of isolation depends on whether the mixer is unbalanced, single balanced or double balanced. Unbalanced mixers exhibit no isolation between ports. Double balanced mixers provide the best isolation between the three ports.

Impedance matching. The three ports of a mixer should be matched at each port. Mismatch in active mixers usually results in reduced mixer gain.

Passive mixers are particularly sensitive to mismatch at the intermediate-frequency output port which causes greater insertion loss and unwanted mixer products. Regardless of whether an active or passive mixer is used, due care should always be taken to implement proper matching to maximise mixer performance.

Simplicity. An important characteristic of any circuit is the ease of design and implementation. Complex designs are often complex to build and difficult to set up. A lower parts

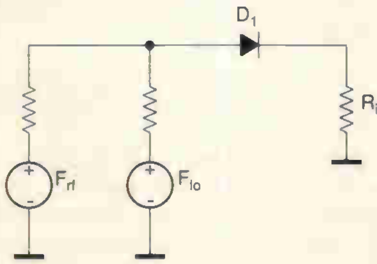
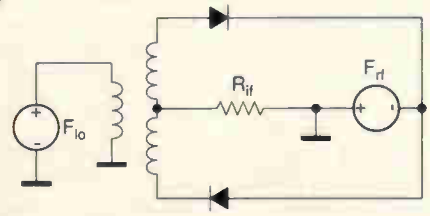


Fig. 1. With the unbalanced passive mixer, there is no isolation between the two input ports and the output port. This results in unwanted local-oscillator elements at the antenna.

Fig. 2. Below: single-balanced passive mixer solves the problem of local-oscillator feed through, but allows rf to pass to the output.



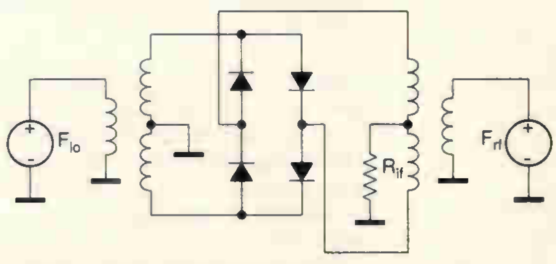


Fig. 3. The double-balanced passive mixer allows no local oscillator or rf components through to the output – in theory.

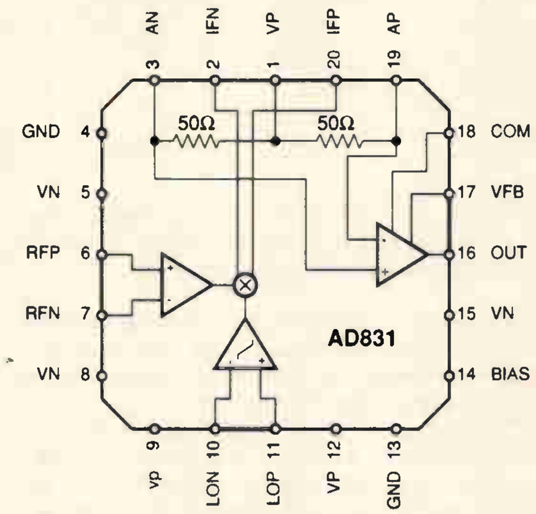


Fig. 4. Example of a high-level active double-balanced mixer – the bipolar AD831.

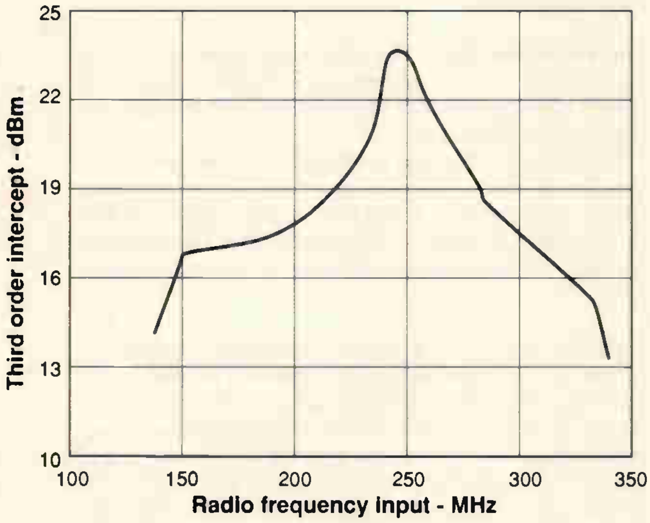


Fig. 5. Intermodulation versus frequency of the AD831 taken from its data sheet.

count has the added benefit of greater reliability, easier maintenance and a smaller spare parts inventory.

Overly complex design is a form of inefficiency that can have a significant effect on the through life cost of equipment. Good design concentrates on gaining the maximum performance from the minimum of components.

Integrated mixers

Given the superior performance of high level double balanced mixers and the availability of both passive and active devices in single packages, there is little incentive to use either an unbalanced or single balanced mixer. This assumes that the requirement for high level performance is one of the main design criteria. Applications demanding low power consumption or specialist functions may be more suited to mixers other than the high level double balanced mixers reviewed below.

AD831. The *AD831* is a modern example of a bipolar high-level active double balanced mixer. Housed in a PLCC package it includes a buffer amplifier with a 50Ω input and output, **Fig. 4**. The gain of the buffer amplifier can be set by appropriate feed back resistors. The buffer is configured as a balanced to unbalanced amplifier without the need for transformers.

The inclusion of the buffer amplifier with the mixer package makes this device particularly appealing as a first stage mixer. In most receiver designs, the first mixer is followed by a low-pass filter and then a buffer amplifier before the signal is applied to intermediate-frequency band-pass filters.

With the *AD831*, an additional two capacitors will implement a one-pole low-pass filter between the mixer outputs and the buffer inputs. The gain of the buffer amplifier can be set by two resistors to compensate for insertion loss in the following band-pass filters.

If a buffer gain of one is specified, then the feedback resistors are not required. Excessive gain reduces the gain bandwidth of the buffer amplifier and reflects noise back into the mixer.

When properly applied this single device and associated passive components can combine the mixer/filter/buffer functions into a very small space with no bulky tuneable components.

The *AD831* is specified for operation to 400MHz at the rf and local-oscillator inputs, but this needs qualifying. All performance curves are basically flat up to 100MHz. Above this frequency, some curves improve while others degrade.

Performance at about 200MHz is particularly ambiguous. If the application requires very good isolation between ports, then frequencies near 200MHz should be avoided. Local oscillator to rf isolation falls from 72dB at 100MHz, to a minimum of 45dB at 200MHz – which is still good by the way. However the high level performance peaks at 250MHz where the third order intercept reaches 23dBm and the –1dB compression point is 11dBm as shown in **Fig. 5**. It is an example how variable the important performance characteristics can be. This is not a particular fault of the *AD831* as all types of mixers both active and passive exhibit widely varying characteristics.

The *AD831* may be used with single or dual supplies. A programmable bias allows the user to define the power consumption. For the best third-order intercept performance, the *AD831* will typically draw 100mA quiescent current. The current may be reduced to 45mA for minimum power consumption.

The noise figure is specified at a high 20dB for a single ended voltage output. For receiver applications, this means that an rf amplifier ahead of the mixer is mandatory if a reasonable receiver noise figure is to be achieved.

In summary, the *AD831* is suitable for most mixer applications to at least 300MHz. Performance is very good up to 100MHz but performance varies at higher frequencies, particularly at 200MHz. The 20dB noise figure makes this device unsuitable for low noise applications.

The ability to combine a high performance mixer, low pass filter and buffer amplifier into a very small space with no tuneable components means that the AD831 would be a good choice for many applications.

SL6440. The *SL6440* is a bipolar double balanced mixer first produced by Plessey about 1980. In spite of its age, it is still a useful device with good performance.

The specified –3dB frequency range extends to 150MHz but I consider this to be a little optimistic for practical applications. The intermediate-frequency output level remains constant up to 50MHz after which the roll off is dependant on the supply voltages.

For a main dc supply voltage of 6V, the 1dB roll off occurs at 80MHz while for a supply of 12V dc, it occurs at just over 100MHz. I would conservatively rate this device as being suitable for applications up to 90MHz.

Mixer selection guide

For applications that require good performance, no tuned components, and a minimum of pcb real-estate, the *AD831* is an excellent choice. The implications of the high noise figure need to be considered.

The *SL6440* is an old device but for applications that require an average noise figure with a low level local oscillator it remains a good choice.

If very high level performance and gain are the key criteria, then the jfet mixer should be selected. Most applications will require the matching transformers to be hand crafted.

For those after the ultimate in high-level mixers, the mosfet mixer is supreme. But as with the jfet mixer, the effort required to implement the mixer and related circuitry makes this choice difficult to justify except for the most demanding applications.

When properly applied, the diode double balanced passive mixer combines the lowest noise figure with good high level performance. If low noise is important, and there is reasonable space and local oscillator power available, then a diode double balanced mixer is the right choice.

Almost all of the values shown in the **Table** are frequency dependent and can vary greatly depending on the exact operating conditions. Mixers should be chosen based on performance at the intended operating conditions. Prototypes should then be constructed and tested because it is highly unlikely that the test circuits in the data sheets will match exactly the applied circuit.

Table. Comparison between high-level mixers.

Characteristic	Units	Active jfet SBL1*	Diode RAY11*	Diode	Mosfet AD831	Bipolar SL6440	Bipolar
Frequency range	MHz	50-250	0.5-500	0.01-2500	0.2-500	0-400	0-90
Isolation rf local oscillator	dB	35	45	32	30	65	—
Isolation local oscillator – IF	dB	60	40	20	25	28	29
Overall noise figure (ssb)	dB	8	6	7	9	20	11
Local-oscillator drive level	dBm	+15	+7	+23	+16	–10	–5
Two-tone intercept point	dBm	+34	+16	+25	+44	+24	+30
Conversion gain	dB	+4	–5.6	–6.2	–7 to +18	0	–1
1dB compression	dBm	+13	+1	+15	+16	+10	+10

*MiniCircuits

The *6440* is contained in a 16-pin DIP package, **Fig. 6**. It includes two input and two output pins for balanced signals but may also be configured for unbalanced signals. The *6440* also includes an input pin to program the supply current which can be useful for battery operation.

It is possible to implement power load management by programming the supply current to a low value in the absence of a desired signal. When a signal is received, the supply current can be increased to improve the mixer performance. Such a feature would be relatively simple to include with the receiver squelch control.

Note that the supply current for most active mixers can be programmed by various means and this is not a unique feature of the *6440*. As a general guide for any mixer, either passive or active, intermodulation performance improves with increased supply current and voltage.

Like most active mixers, this device draws a significant current to achieve the best performance. The device requires two positive supply voltages. Rail V_{cc1} is for the mixer while the on chip oscillator buffer is supplied by V_{cc2}.

Typically the *SL6440* draws 30mA to 60mA, which may be increased with the use of a heat sink, or decreased to conserve power. The quiescent current drawn by the mixer is determined by the value of V_{cc2}, and the programming current I_p into pin 3. The absolute maximum power dissipation is specified at 1200mW.

For maximum suppression of intermodulation products, the programming current Ip should be between 10mA and 12.5mA giving a total supply current of about 40mA. The –1dB compression point is then about +7dBm but can be increased to +15dBm by increasing I_p to 30mA giving a total supply current of about 77mA. Reducing the programming current I_p below 5mA attenuates the wanted intermediate-frequency output signal.

Conversion gain or loss is determined by the input and output configuration and programmed current. For resistive, unbalanced inputs and outputs a conversion loss of about 1dB would be expected. The use of tuned, balanced input and output transformers will provide a measured conversion gain of about 4dB.

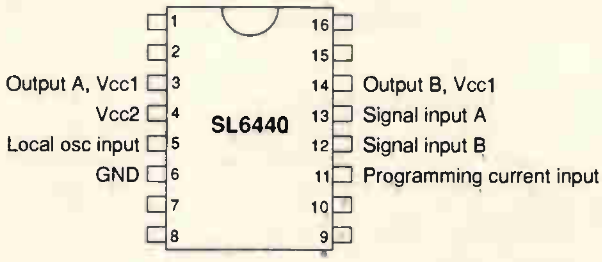


Fig. 6. Around since 1980, the SL6440 still has its uses. A drawback though is that it needs quite a few external components.

Using tuned transformers on the input and outputs offers the advantage of some gain and filtering without degrading performance. This is because balanced inputs and outputs evenly distribute current flow within the *SL6440*. In either case conversion gain is dependant on both the programming current I_p and the load impedance. Excessive gain increases the risk of saturating the mixer output transistors and should be avoided.

While adequate, the information provided on the data sheet could not be described as comprehensive. There is a small selection of tables, text and graphs detailing overall performance and design procedure.

No information is given about the internal workings of the device nor are there any details of frequency dependent impedance characteristics. The design of matching circuits must therefore be accomplished by trial and measurement.

In summary, the *SL6440* may be old but its performance is still good up to about 90MHz. If you need optimum performance then tuned input and output circuits would typically be used. Even in a basic application, about a dozen passive components are required around the *SL6440* which means that this device occupies a reasonably large area of pcb real-estate. The 11dB noise figure is about as good as can be expected from this type of mixer.

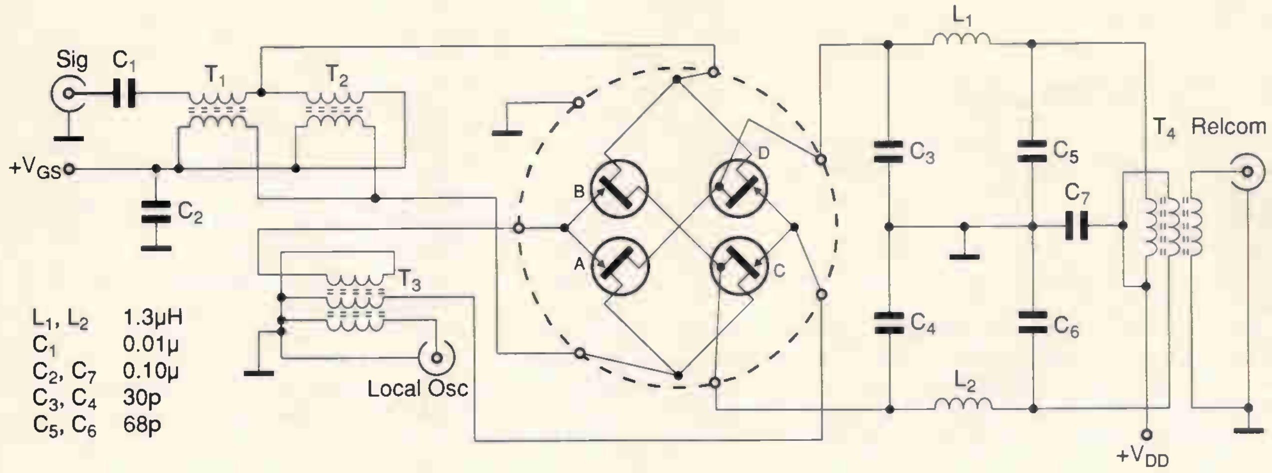


Fig. 7. Circuit for a double-balanced fet mixer by Ed Oxner.

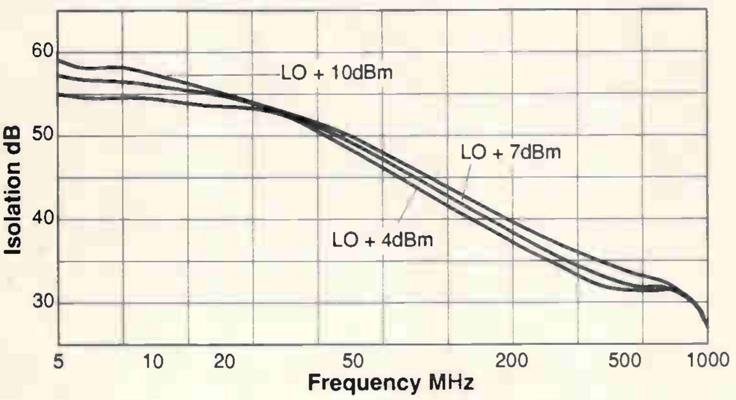


Fig. 8. Local oscillator to rf isolation performance for a diode mixer based on the Mini Circuits SBL-1.

Using junction and mos fets

Jfet mixer. The junction-fet mixer appears to offer the potential for very good performance. Conversion gain and intermodulation distortion characteristics are superior to typical passive mixers. Junction fets have an inherent square law response which reduces third order intermodulation distortion. And, like passive mixers, jfet mixers have a high burn-out level.

The main disadvantage is that there is no known source of ready-made units incorporating the jfets and transformers. Also a high local-oscillator drive is required.

Unlike double-balanced diode mixers, the jfet mixer has to be built up from individual components. Even commercially produced receivers such as the *ICOM R9000* uses a single balanced jfet mixer made from individual components.

The inability to obtain a single-package double-balanced mixer is a major disadvantage of the jfet mixer. Suitable transformers are difficult to obtain and for prototype work, hand made coils are usually necessary.

If optimum performance is to be achieved then the transformers need to be physically matched to high tolerances. It is possible to obtain good results with a combination of good construction techniques and a suitable balance adjustment.

Optimum power gain and noise do not occur at the operating point. The bias currents, local oscillator drive level and matching transformers must be properly selected to ensure that the fets operate in the square law region and that distortion is minimised. The lowest achievable noise figure is 8dB.

For optimum performance of a double balanced fet mixer, the fets should be perfectly matched which is very difficult to achieve using discrete fets due to the difficulties in precisely controlling characteristics between batches.

In practice matched characteristics can only be achieved if the fets are mounted within the same package and are made from the same silicon wafer. Such devices are available from Siliconix which manufactures dual (*U430*) and quad fets (*U350*) in single packages specifically for mixer applications.

For single-balanced mixers, the requirement for matched components is less stringent allowing the use of fets matched to within 10%.

Despite the problems of construction, the jfet mixer offers good performance. Their specifications are similar to that of a high-level diode mixer with the advantage of positive gain.

The major disadvantage is the high local-oscillator power requirement and the significant construction effort. For those of you wanting to experiment with jfet mixers, references 3 and 4 provide excellent details on the performance and practical construction of the mixer shown in **Fig. 7**.

Mosfet mixers. If there is a need for very high-level performance, then the jfets can be substituted for either monolithic dmos fet quads or a combination of rf power mosfets. Suitable monolithic devices include Calogic *SD8901* and the Siliconix *SD5000* from which very good results have been obtained. Double balanced mixers may operate the mosfets as switches with no drain voltage applied resulting in an insertion loss of about –7dB. Alternatively, drain voltage supplied to the mosfets may be used to give a mixer gain of up to +17dB.

One of the features of mosfets is that the gate drive voltages are roughly the same for for all mosfet types. This means that the local oscillator power needed for a very high level mosfet mixer is less than the power required for an equivalent diode mixer. A mosfet mixer giving very high level performance does not require an excessively high local oscillator drive.

While mosfet mixers provide excellent very high level performance, they suffer from the same problem as jfet mixers in that the transformers and surrounding circuitry result in a high parts count and the use of a significant area of circuit board. Such a mixer would only be required where the receiver was physically located near a high powered transmitter. Typically, mosfet mixers would be found in very high performance hf and mf receivers.

Double-balanced-diode mixer

The double balanced diode mixer is a design that is positively archaic but still has application in modern receiver design. Its ability to withstand high level input signals, low cost and the ready availability of small mass produced mixers has ensured its continued use.

If the insertion loss is tolerable, the characteristics compare favourably with other high level mixers. Perhaps the most widely used mixer type is the *SBL-1* from Mini-

SBL-1 diode mixer from Mini-Circuits. Although not an active device, this mixer produces good results in a wide range of receiver applications.

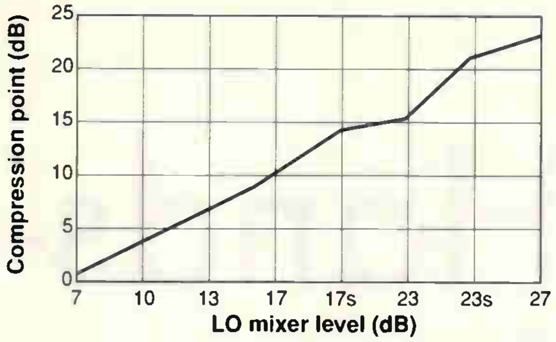


Fig. 9. Diode mixer 1dB compression point versus local oscillator level for various Mini-Circuits parts.

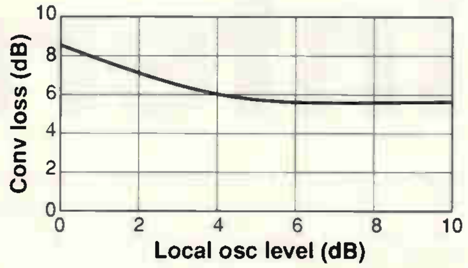


Fig. 10. Typical diode-mixer conversion loss versus local-oscillator level.

Circuits. It is cheap, compact and available. Mini-Circuits and other companies produce a wide range of passive mixers suitable for most mixer applications.

Like the other mixers described, the performance of the diode double balanced mixer is dependant on a variety of factors including but not limited to frequency, local-oscillator power, matching and temperature. The graph of local-oscillator to rf isolation shown in **Fig. 8** is typical of the variations seen in performance.

Proper matching of the intermediate-frequency output is essential to gain optimum performance from a diode double balanced mixer. The intermediate-frequency filter should feature a 50Ω resistive impedance across the rf spectrum. Failure to ensure proper matching will result in higher conversion losses and the generation of unwanted harmonics. In addition to impedance matching, the intermediate-frequency filter should attenuate all mixer products except the desired intermediate-frequency. Detail of the design of an intermediate-frequency filter for a passive diode mixer is the subject of a future article.

Diode mixers require a high-level local oscillator with at least +7dBm output. Some very high level passive mixers require up to +23dBm local oscillator drive. The relationship between local oscillator power and the 1dB compression point is shown in **Fig. 9** for a selection of diode mixers from Mini-Circuits.

The local-oscillator level also effects conversion gain as shown in **Fig. 10**. It is important that harmonics from the local oscillator are adequately suppressed. Harmonics from the local oscillator mix with the rf input and produce unwanted mixer products.

It can be difficult to design and build oscillators at the required output level combined with the desired spectral purity. This is particularly true for battery operated equipment where available power and voltage are relatively low.

Some mixers are available that include a built-in amplifier on either the local oscillator input, or at the intermediate-frequency output. These may ease the design problem in some situations. Local-oscillator power and purity has a major effect on the design and performance of the mixer stage and therefore the whole receiver system.

One of the main advantages with the diode double-balanced mixer is that being a passive device, it draws no quiescent current. This is somewhat offset by the high local-oscillator drive requirements and the need to compensate for the conversion loss.

Although diode mixers are available in very compact packages, the high level local oscillator and intermediate-frequency matching filter will normally require a significant circuit-board area. If high level and low noise are the prime requirements then the double balanced diode mixer remains the best choice.

In summary

The selection of a mixer for any receiver application depends on many variables. The information in the Table has been obtained from a number of sources. The values are representative of actual devices operating under typical conditions and in most instances these values will be different to those specified in data sheets.

As with most design choices, the selection of the right mixer for the right application is based on compromise. There is no single mixer that is suitable for all tasks and it is up to the design engineer to determine which characteristics should dominate the selection process. Selecting the right mixer is a matter of determining the most important characteristics and matching these to the appropriate mixer type. ■

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Hands-on Internet

Cyril's findings on the net include news of Microsoft's continuing legal battle, an announcement that Intel is to sell 'cpu kits' to pc oems, and a dedicated modelling tool for optical links.

To consolidate its processor program discussed last month, Intel is developing a series of cpu kits[1] for pc makers, designed to speed the production of 'Intel Inside' pcs. These all embracing kits will include the processor, chip set, motherboard, memory types, i/o interfaces and graphics buses, **Fig. 1**. Some pc vendors are reported to feel these kits will reduce their design options and may reduce their already small profit margins.

Other makers of compatible chip-sets however fear that 'Intel Inside' might soon become 'Only Intel Inside'.[2] As reported by *ZDNN*, the P6, or sixth generation, high speed interface bus for Pentium II is proprietary to Intel so cannot be used without permission or licensing.

Acer Laboratories, Silicon Integrated Systems and Via Technologies all have compatible chip-sets in customer evaluation, but these cannot be released for sale until legal problems have been resolved, leaving Intel as sole supplier.

Microsoft still under pressure

The Senate Judiciary Committee hearing on 'competition in the computer industry' debated the extent of Microsoft's power to dominate desktop operating systems and Internet browsers. As reported by *PCWeek*,[3] much discussion centred on Windows 95 and Explorer licensing.

The Senate Judiciary Committee oversees the DOJ. While the hearing produced no substantive result, it was apparently seen as a green light for the DOJ to continue its actions against Microsoft.

Fig. 1. Intel's new kits of CPU, motherboard, chip-set may make choices simpler, but will component costs increase for computer makers?

With the hearing over, the DOJ continues to probe all aspects of Microsoft, to determine whether to proceed with a full-scale anti-trust assault. According to *PC Week*,[4] the DOJ is conducting yet another round of depositions with computer oems to find ammunition for its next actions. One particular question regards performance degradation should a different browser be installed on top of Windows 98 with Internet Explorer integrated into it.

Windows 98

Retail shipment of Windows 98 is now planned for June 25.[5] Domestic users of Windows 95 in the USA and Canada can now purchase a preview release of Windows 98 on cd-rom for $29.95 as part of a beta preview programme for technically competent users.

This offer is available in other countries, although the price differs. For UK and Ireland the price is £24.99, which includes an upgrade voucher[6] to Internet Explorer Plus. You must already use Windows 95, have 125Mbyte available on your hard disk and have at least 16Mbyte ram with a 486DX/66 or better processor.

It is claimed that Windows 98 speeds up system performance and its more efficient filing system releases some 28% of hard disc space without compressing files. The operating system includes support for DVD-rom and the universal serial bus, or USB, port. This serial port allows a large number of peripherals to be connected and disconnected without turning the machine off.

Browsers

Last month I mentioned the 'Opera' browser. At the moment, Windows versions are available, but I access Internet using only OS/2 Warp4 or Merlin. I installed the 16 bit version of Opera 3.1 under Windows in OS/2 with no problems. But I had difficulty linking the browser with my native OS/2 dialler since it required a Winsock dialler. A 'proper' OS/2 version of Opera is apparently due shortly, so I have decided to await its arrival.

Many other companies are busy writing Net-based software. Browser.Com,[7] part of ClNet, gives a full listing. In its recent review of 'rebel' browsers, Opera was 'editors choice.' This page is also worth visiting if you use Internet Explorer 4 and are troubled with the "No such interface supported" error. Apparently caused by installing over an older beta version, this problem can be cured by uninstalling and reinstalling completely.[8] A simpler alternative correction is also possible, **Fig. 2**.

Junk mail

I recently mentioned that I receive far too much junk e-mail. If anything the problem has worsened lately. Not only has

Fig. 2. Visit Browsers.com to find the latest versions and performance test results of major or rebel browsers.

Fig. 3. Interested in modelling optical systems?

the quantity increased, but I frequently now get more than one copy of a mailing. Since I use an American based service provider and my e-mail address is on my home Web page, it is possible I receive more than the average UK user.

Junk e-mail or 'spam' as it is called, is a problem for both service providers and search engines. A service called Search Engine Watch recently ran some tests on search engines to check their spam content.

At one time Infoseek had a particular problem with junk e-mail and as a result took preventative action. In these latest search engine tests, it was considered equal best with Lycos.[9]

Apart from being annoying, junk e-mail can cause severe problems to Internet providers. In California, a bill giving legal recourse against 'spammers,' is now in hand. One vocal supporter of this bill – Pacific Bell Internet – was recently deluged with spam e-mail, over a four day period. Having 170 000 customers and a 50 000 spare capacity, the system ran out of processing space. The service provider responded by installing four more gateways, doubling its capacity to 440 000 users, at a cost of $500 000.

Simulation

If you are interested in optical-fibre circuit design, one page worth visiting is Gold.[10] An acronym for Gigabit Optical Link Designer, this new Australian simulator is specifically for optical systems.

Some of the core software in Gold is licensed from the

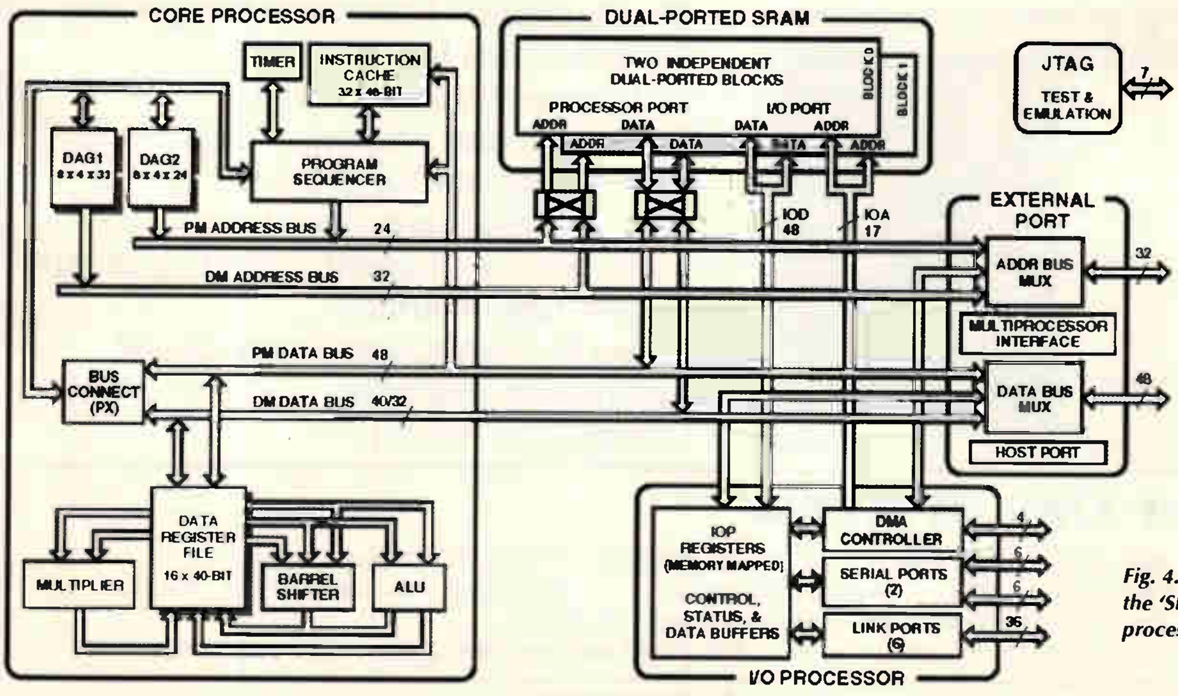

Fig. 4. Block diagram of the 'Sharc' 240 pin signal processor.

Where to surf

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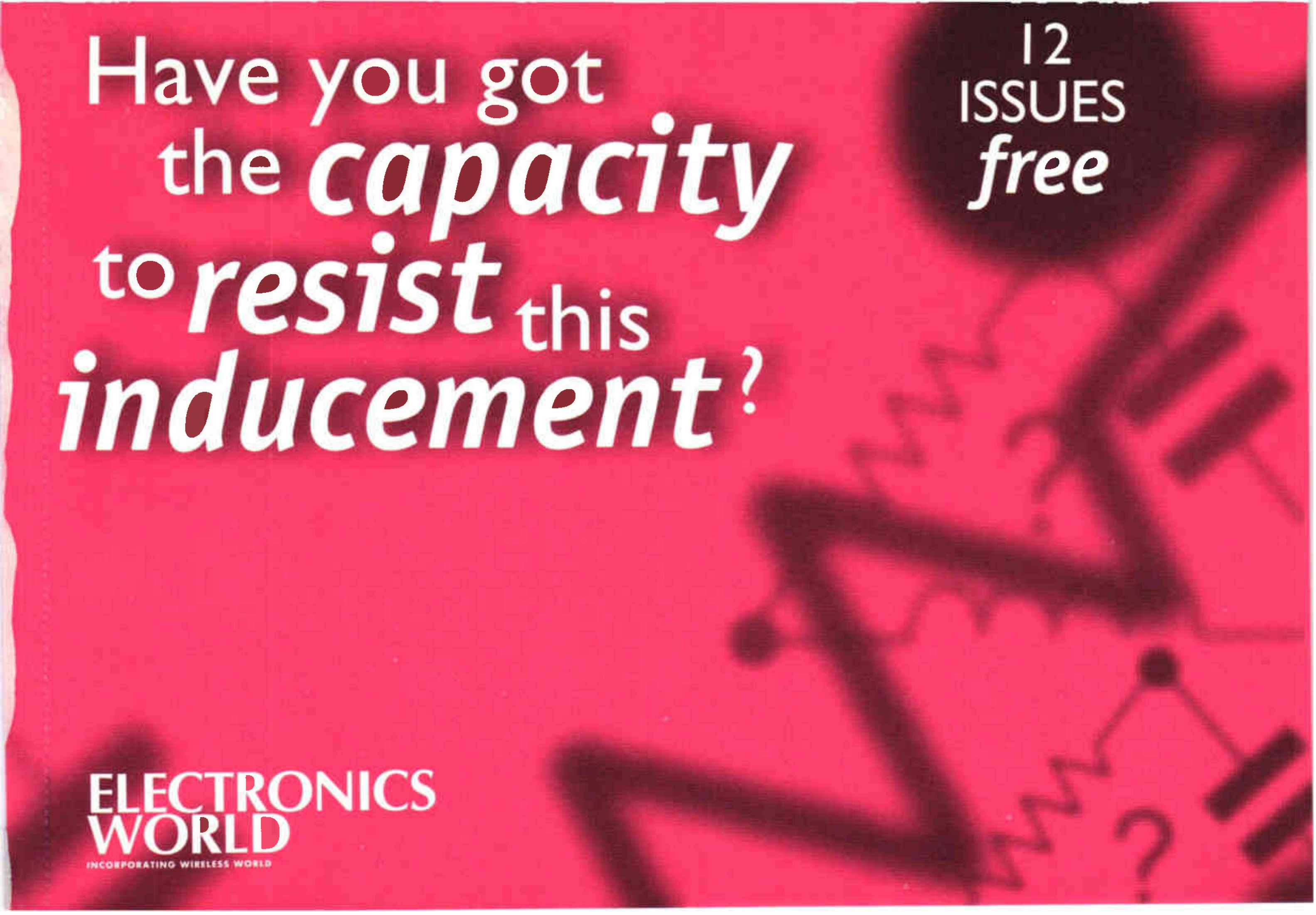
9. Search Engine Spam Survey. http://www.searchenginewatch.com/reports/spam.html
10. GOLD Virtual Photonics. http://www.vp.com.au/GOLD/GOLDGI.htm
11. Computer Design Online. http://www.computer-design.com
12. Attack your design with a SHARC. http://www.analog.com/SHARC/attack

Australian Photonics Co-operative Research Centre. The package runs on pcs under *LabVIEW* for Windows, but versions for other platforms will be developed. A Gold demonstration disk is available, but you need to have either the base or development versions of *LabVIEW* installed, Fig. 3.

Computer Design Online is an on-line magazine and designers' meeting place. It is dedicated to all aspects of serious computer aided design.[11] The magazine includes a searchable database of its past publications and provides support for all computer aided design decisions.

This magazine is sponsored in part by Analog Devices. When I visited it, the home page featured a flyer for Analog's new digital signal processor – a low-cost 240 pin 'Sharc' – and its evaluation kit. **Fig. 4.**[12] ■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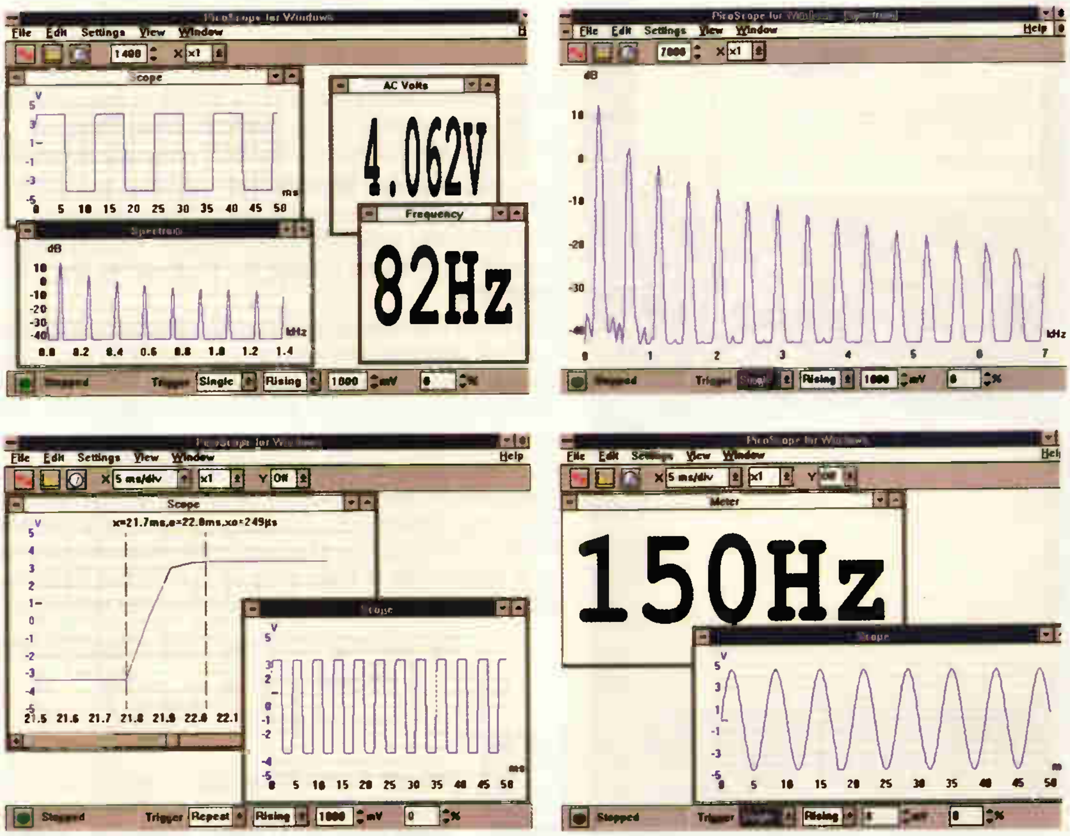
How to submit your ideas

The best ideas are the ones that save readers time or money, or that solve a problem in a better or more elegant way than existing circuits. We will also consider the odd solution looking for a problem – if it has a degree of ingenuity.

Your submission will be judged on its originality. This means that the idea should certainly not have been published before. Useful modifications to existing circuits will be considered though – provided that they are original.

Don't forget to say why you think your idea is worthy. We can accept anything from clear hand writing and hand-drawn circuits on the back of an envelope. Type written text is better. But it helps us if the idea is on disk in a popular pc or Mac format. Include an ascii file and hard-copy drawing as a safety net and please label the disk with as much information as you can.



Turn your PC into a high-performance virtual instrument in return for a circuit idea.

The ADC200-50 is a dual-channel 50MHz digital storage oscilloscope, a 25MHz spectrum analyser and a multimeter. Interfacing to a pc via its parallel port, ADC200-50 also offers non-volatile storage and hard-copy facilities. Windows and DOS virtual instrument software is included.

ADC42 is a low-cost, high-resolution a-to-d converter sampling to 12 bits at 20ksample/s. This single-channel converter benefits from all the instrumentation features of the ADC200-50.

Separating p-waves in cardiac waveforms

An important part of cardiac analysis is the measurement of rate variability, the interval between appearances of the p-wave recently being of most interest. Since the p-wave is a relatively small part of the waveform, it is more difficult to separate it from the rest of the electrocardiogram than the r-wave; this circuit performs that function and measures the p-p time interval.

In the block diagram of **Fig. 1**, the input is the ecg, which is amplified and compared with the threshold level Th_1 to produce a reference pulse in time with the r-wave. The comparator triggers two monostables, the outputs of which are shown in **Fig. 2** as (c) and (d), Anded to produce the waveform at (e), which falls within the p-wave region. This pulse, which occurs, due to the timing of the monostables, during the p-region, programs the amplifier with a gain of 10; for the rest of the cycle, it is zero.

Only the p-wave, therefore, is amplified, the output (f) being compared with Th_2 and the resulting pulse (h) triggering a further monostable to generate (k) and (l) by means of the two circuits shown in **Figs 3** and **4**. The pulse output (k) resets a 16-bit counter (four 7493s) and the one at (l) triggers a latch composed of four 7474s to hold the time count.

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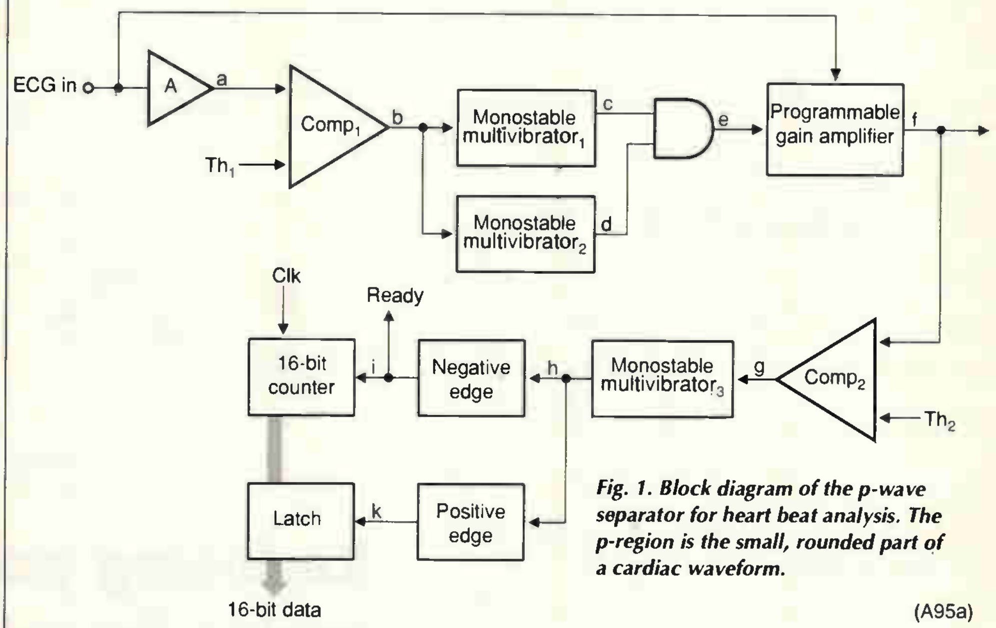


Fig. 1. Block diagram of the p-wave separator for heart beat analysis. The p-region is the small, rounded part of a cardiac waveform.

(A95a)

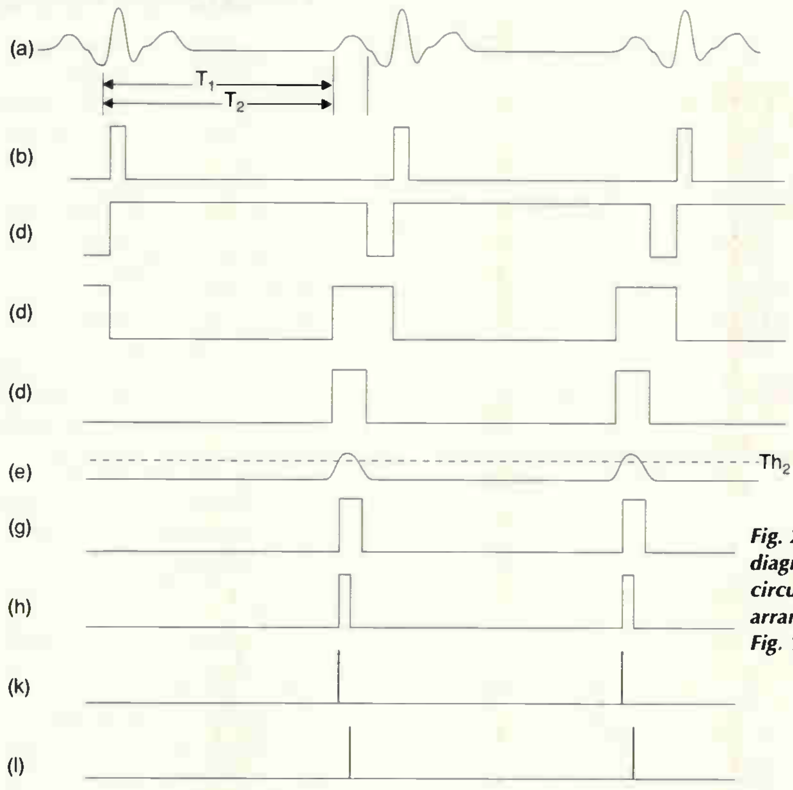
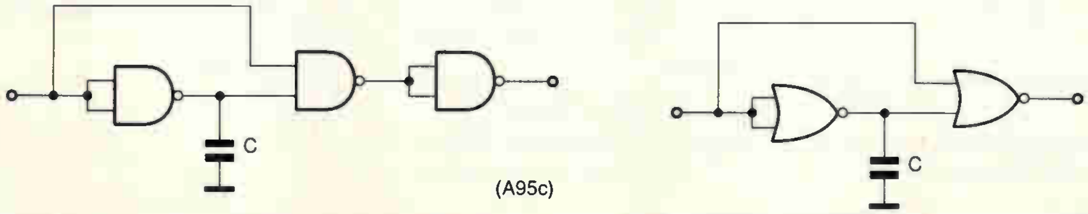


Fig. 2. Timing diagram for circuit arrangement of Fig. 1.

(A95b)



(A95c)

Fig. 3. Edge-deriving circuits for positive (a) and negative (b) edges of a positive-going square pulse.

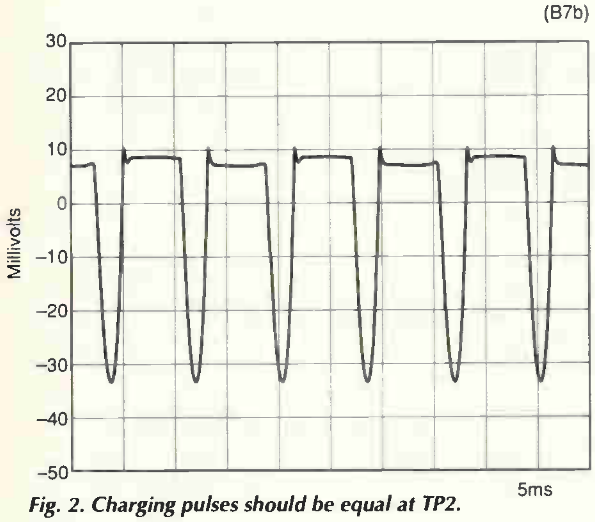


Fig. 2. Charging pulses should be equal at TP2.

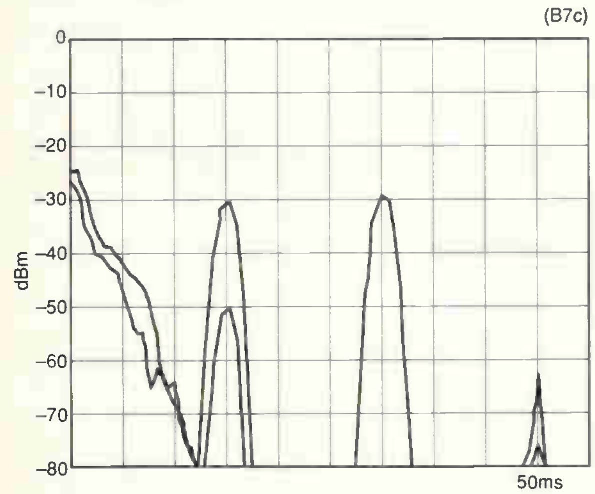


Fig. 3. 20dB improvement in power supply ripple after adjustment of R_2*.*

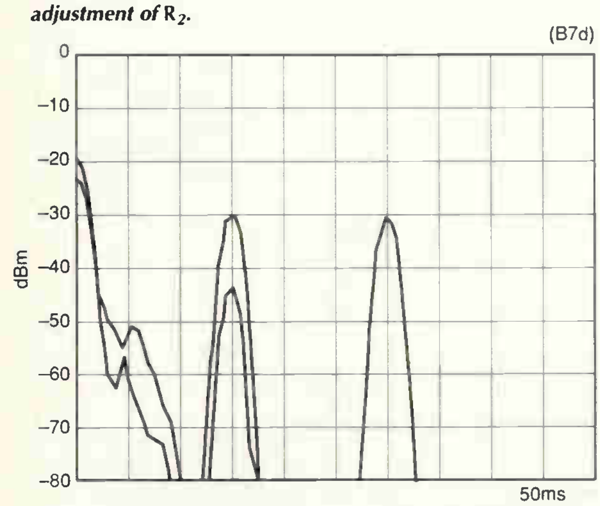


Fig. 4. Change of power frequency ripple when transformer heater leads to a directly-heated rectifier are reversed.

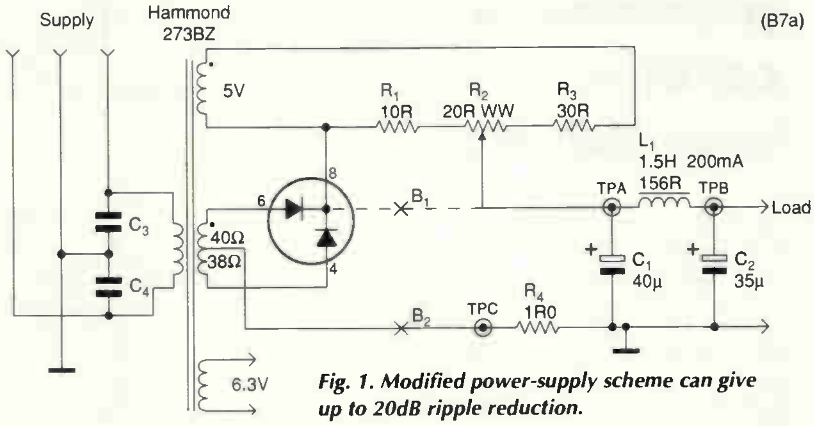


Fig. 1. Modified power-supply scheme can give up to 20dB ripple reduction.

Reducing power-supply mains-frequency ripple

If you have a centre-tapped mains transformer driving a full-wave rectifier, it is possible that 50/60Hz ripple is greater than it theoretically should be. This modification resolves the problem, producing a ripple reduction of about 20dB at mains frequency and odd harmonics.

With this modification, high voltages are present.

Because power transformers are normally wound serially, one side of the high-voltage winding is of higher resistance than the other. Although one expects the ripple to be primarily at even harmonics of the mains frequency, this effect inserts significant ripple at mains frequency and at odd harmonics, since the output filter network, *RC* or *LC*, is much less effective at the lower frequency.

The ripple does not occur in full bridge circuits and push-pull amplifiers are relatively invulnerable, but single-ended amplifiers, particularly the triode type, are affected. Depending on loading, between 2/3 to 4/5 of the ripple is applied to the load in a triode power stage and, since speakers have low-frequency resonances, the conditions are right for hum.

My own case is shown in **Fig. 1**, a power supply for a 1960s audio amplifier, the transformer being a 375V-0-375V model, the windings of which measure 38 and 40Ω. The 5V winding drives resistors $R_{1,2,3}$, which apply ac in series with the dc from the rectifier in place of the direct connection at B_1. Varying the potentiometer, which is a wirewound type, alters the polarity and aids or bucks each half of the ht winding.

The only other modification is R_4, which is included for use during adjustment, which is done by observing current pulses across this resistor at test point C; **Fig. 2** shows properly adjusted pulses. If the 6V ac winding is used, the break should be at B_2. Capacitors $C_{3,4}$ must be ceramic types, safety rated for connection the mains.

Both semiconductor and valve rectifiers have been tried, the valve type with varying results, and the same process can be applied to a choke-input filter. The improvement at 60Hz at the output was, in my case, from 75mV to 8.5mV and at 180Hz from 2.5mV to 1mV.

If your power supply has a directly-heated rectifier, there may be no need for all that. Simply reverse the heater leads to get about 15dB improvement. There is a 50:50 chance that they were the wrong way round.

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Low-noise, high input-impedance amplifier

Although this amplifier was developed for use with a guitar pickup, it may be used as a general-purpose circuit where small voltages must be amplified with very low noise and extremely high input impedance.

Input stage consists of a *U430* matched jfet pair in a differential arrangement, the output op-amp having feedback to set the gain. Balance requires that $R_{3,4}$ and $R_{5,6}$ are equal, voltage gain being given by $G=g_m R_5$, where g_m is the fet transconductance.

Input impedance is equal to R_1 which, with input bias current at 2pA, causes an input voltage of 20μV, negligible compared with the input offset of the jfet pair at 5mV. Bandwidth is 1MHz and is limited by the op-amp. Noise is about 2.5nV/√Hz. The input capacitor may be left out if dc coupling is needed.

Peter Goodson
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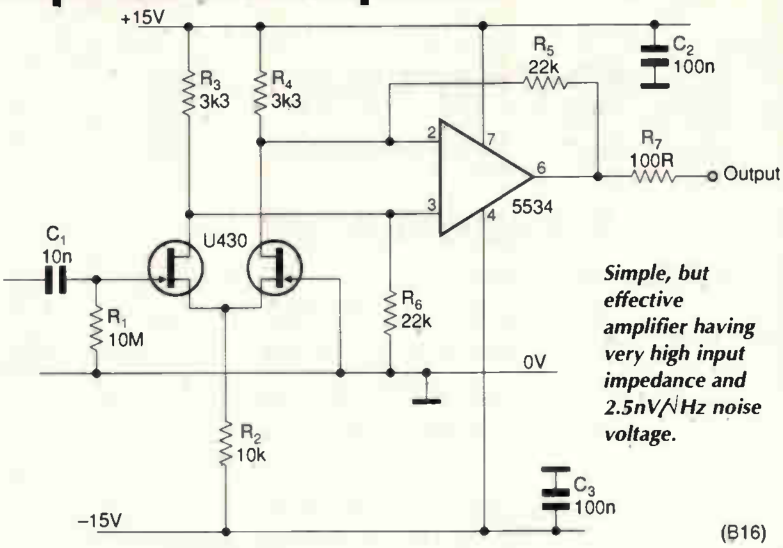

Simple, but effective amplifier having very high input impedance and 2.5nV/√Hz noise voltage.

Economical testing for digital recorders

Output from DAT, cd and full solid-state recorders, when fed with a single frequency, should be a pure sine wave without distortion, sidebands, harmonics or beats; a single spectral line. If this is not the case, the system is at fault and observing the extraneous spectral components helps identify the faulty stage.

Commonly, two sampling frequencies are in use: 44.1kHz for cds and 48kHz for DAT, the test signal being 11.111kHz and 12.345kHz respectively. These frequencies derive from a 1MHz crystal oscillator by way of a switchable ÷90 and ÷81 divider using two *4017*s.

Output from the circuit contains a signal within the bandwidth of the recording system and several harmonics outside it, including one near the Nyquist frequency. After being recorded and replayed, the signal should, theoretically, be sinusoidal, but the anti-aliasing filter, the dsp or the reconstruction filter will not be ideal and will produce additional frequencies. These extra lines in the spectrum can be used to identify the source of the error. Do not overload the input of the recorder, since digital recorders have absolutely no headroom.

Having selected the output frequency, make the level as high as possible, but within the dynamic range, which is, typically, –1dB to –0.5dB. Record this for a few minutes and replay, observing the waveform and spectrum. Additional frequencies shown depend on which input frequency has been selected and which system fault causes them, as in the **Table**.

Well adjusted systems with 16-bit accuracy should have a noise floor below –98dB, with no spectral component other than the signal. In practice,tests on various systems, some of them professional equipment costing up to £40000, indicate that all systems suffer from one kind of trouble or another.

Complex music generates a large number of spurious frequencies absent from the signal; it seems likely that this is part of an explanation for 'digital sound'.

Gerd Schmidt
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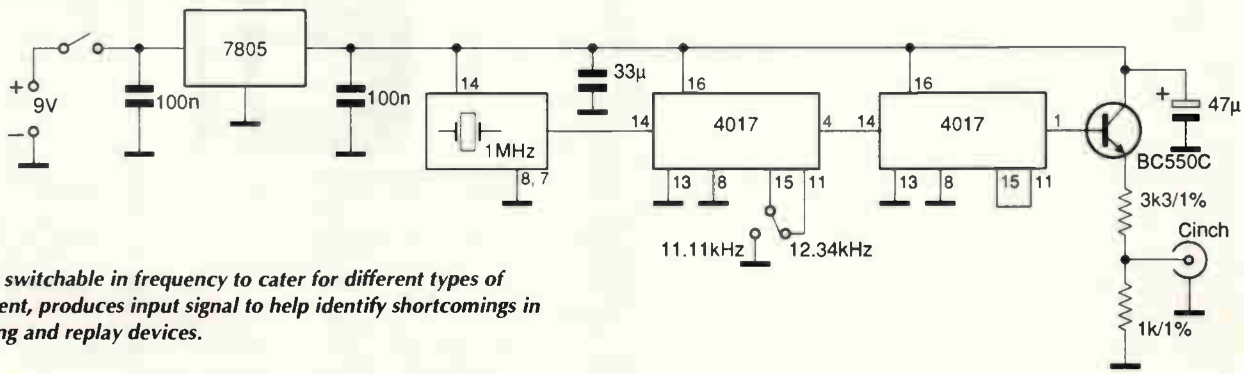

Simple circuit, switchable in frequency to cater for different types of digital equipment, produces input signal to help identify shortcomings in digital recording and replay devices.

Table. Fault examples.

Taking 44.1kHz, with the 48kHz signal in brackets:

A sampling stage fault, probably anti-aliasing filter, will give a line at,
344Hz (1.383kHz), 10.767kHz (10.963kHz), 11.455kHz (13.728kHz) or 21.878kHz(23.308kHz).

Errors in digital signal processing show at,
32.9989kHz (35.654kHz).

And a faulty reconstruction filter shows up at,
22.222kHz (24.69kHz) or 33.333kHz (37.03kHz).

Quietening switched-mode power supply rails

Ripple and spikes from a switched-mode power supply can exceed the noise margin of logic gate inputs, make monostable flip-flops oscillate or misfire and possibly cause embarrassment in phase-locked loops. Low-impedance capacitors across the rails may effect a cure, but some form of inductive filter is often needed, although it must be damped to avoid making matters worse. This circuit, shown in **Fig. 1a)**, provides an economical solution.

Inductance takes the form of a MMG-Neosid ferrite window core with a single wire through the two holes; capacitors are one good 220µF electrolytic and a number of ceramic types, which may be distributed around the board. This arrangement avoids excessive inductance and large capacitors, which would also be inductive.

Used with a run-of-the-mill 5V, 2A supply, the filter reduced ripple and spikes from 100mVpk-pk to 7mVpk-pk. **Figure 1b)** illustrates the method testing for the effect of other sources of trouble such as clock circuits; a square-wave generator producing a 5V output from 50Ω couples ±50mA into the 5V rail. The worst effect was 20mVpk-pk at 800kHz, with another of about the same amplitude at 20MHz,although the power-supply's feedback loop bandwidth is 5kHz, which makes for sags and overshoots.

If the circuit being supplied has its own ground connection, the arrangement of **Fig. 1c)** may be used, this having the advantage of some common-mode rejection. With a multi-output power supply, where all rails share a common 0V, it may be necessary to bring the 0V directly and use heavy cable for the 5V, common-mode connecting to the others. There is no easy method of overcoming dc errors without using properly isolated outputs.

Spikes have been found to cause more trouble than ripple and it is essential to attenuate anything that looks 'sharp'; and this also means that spikes coming back from the load must be decoupled by low-impedance capacitors. Aluminium types, such as the Rubycon *TWSS* and, rather better, the same company's *YXF* are good, the technique having improved in recent times, while the Sanyo *Oscon* is even better; AVX *MicroChip* tantalums are very good.

CJD Catto
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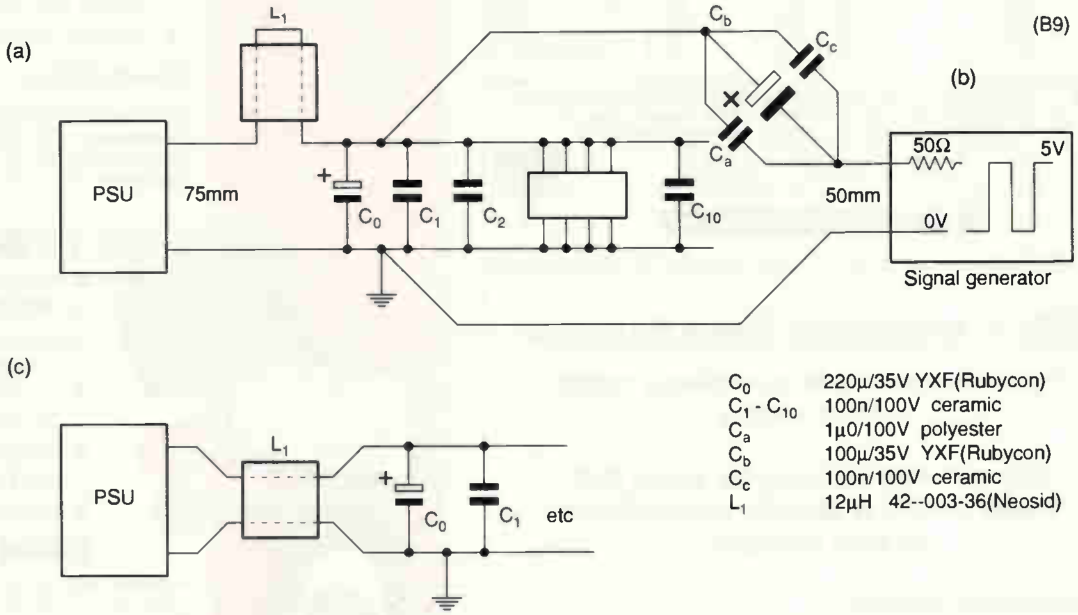

Fig. 1. Single turn on a ferrite window core, together with the right kind of capacitor, reduces ripple and noise from switchers by over 90%. At a) is the filter and b) shows how to test the power source for noise coming from the load. For circuit boards having their own ground connection, use the arrangement at c).

Wide-band, dc-isolated amplifier

In applications such as grounded-cathode photomultiplication, or scanning electron miscroscopy of dc-biased samples, it is necessary to use an isolation amplifier; wide bandwidth is also sometimes needed and is not often found in such circuits. This design is useful when ac common-mode potentials are not involved and it provides a rise time of less than 100ns.

Input splits into two paths: an optically isolated amplifier, $A_{1,2}$ and the optocoupler, and an ac coupling via C_2, both driving A_3. By virtue of the ac coupling, provided that the time constant R_6C_3 equals R_7C_2 at 20µs here, frequency response is flat and limited by the bandwidth of A_3, not by that of the optically coupled section.

Negative feedback using one diode in the *IL300* eliminates non-linearity in the optoisolator, and $R_{3,5}$ allow bipolar input. Gain of the low-frequency path into the inverting input of the output amplifier is unity and that of the ac-coupled path is 2. Since $R_6=2R_2$, overall gain of the lf path is 2. Capacitors $C_{1,4}$ preserve stability.

This is not a true isolation amplifier and the floating input and grounded output common rails must be well decoupled, in this case by C_5. If an unstabilised power supply is used – power connections not shown here – $R_{3,5}$ should be replaced by current sources. Overall frequency response may be extended to over 50MHz by the use of a faster op-amp for A_3; voltage op-amps should be used unless $R_{1,7}$ are

Interface between a 4-by-4 keypad and one microcomputer pin

ADC42 WINNER

Used with a microcomputer having an i/o pin that may be shorted to ground while in the high state, such as one of the 8051 family, this circuit enables a 4-by-4 keypad to use only that pin, instead of the four or five normally needed.

Neglecting, for the moment, the effect of R_9, current through R_{10} is small. Shorting Col_1 and Row_1 via the keypad, Q_1 being low and Q_5 high, causes current to flow from +5V through R_1, D_5 and D_1 through R_{10}, the consequent voltage drop being detected by comparator A_4.

Most states of the counter give either no path for the current to R_{10} or several but, for every switch position, there is one counter state giving an unambiguous indication of the switch state. For example, if Col_1 and Row_1 are shorted by the keypad, the counter code is 00011110 and for Col_2 and Row_1, 00101110 etc. The process of testing a switch therefore requires the computer to advance the counter to the desired count and to monitor A_4 output. *LM339* op-amps are open-circuit comparators; when on, the output is a short to ground and when off, open-circuit.

Initial conditions are such that point X_1 is held low by the computer for 100μs or more, A_3 has switched off and C_2 is charging to reset the counter. Current through R_9 drops a small voltage across R_{10} and the output of A_2 and therefore the negative input of A_4 are grounded and A_4 is off.

When the computer takes X_1 high, A_1 discharges C_1, A_2 goes high and A_4 pulls X_1 low again. Switching delays allow C_1 to discharge completely so that when A_4 switches A_1 off, it takes about 10μs for the voltage across C_1 to switch A_2 high again. For a count state and switch position causing the extra current to flow in R_{10}, A_4 does not switch on and X_1 stays high.

After driving X_1 high, the computer monitors it to detect the presense or absence of current through the keypad and then forces X_1 low again, notwithstanding the state of A_4. After 14μs, the trailing edge of the pulse at A_2 has incremented the counter, A_4 has switched off and the computer drives X_1 high to start the whole thing again until the counter holds the desired count for the switch under test.

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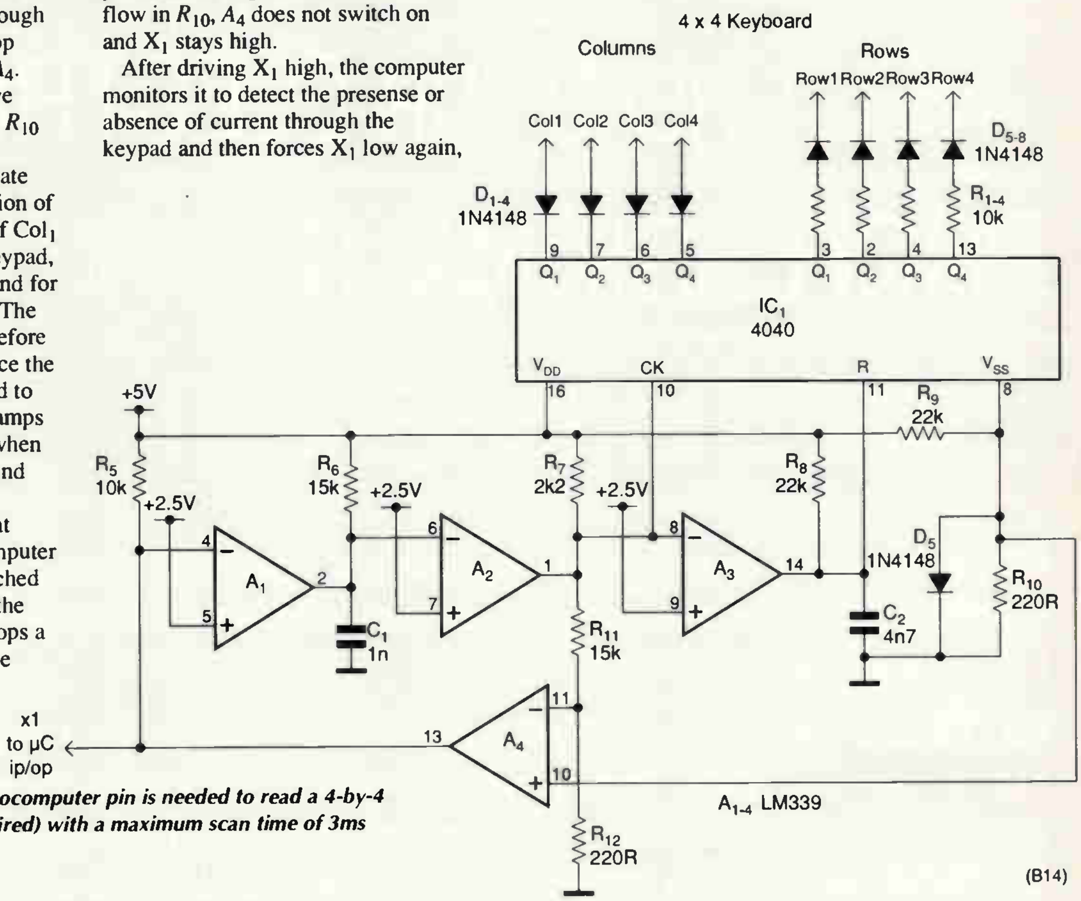

Only a single input/output microcomputer pin is needed to read a 4-by-4 keypad (or up to 6-by-6, if required) with a maximum scan time of 3ms (42ms for 6-by-6).

under 100Ω and both the values and voltage ratings of $C_{2,5}$ are increased. Minor adjustments in crossover frequency may be effected by replacing C_3 with 100pF across a 20pF trimmer; crossover time constant of 20μs should not be reduced to become comparable with the optocoupler time constant of about 1.5μs.

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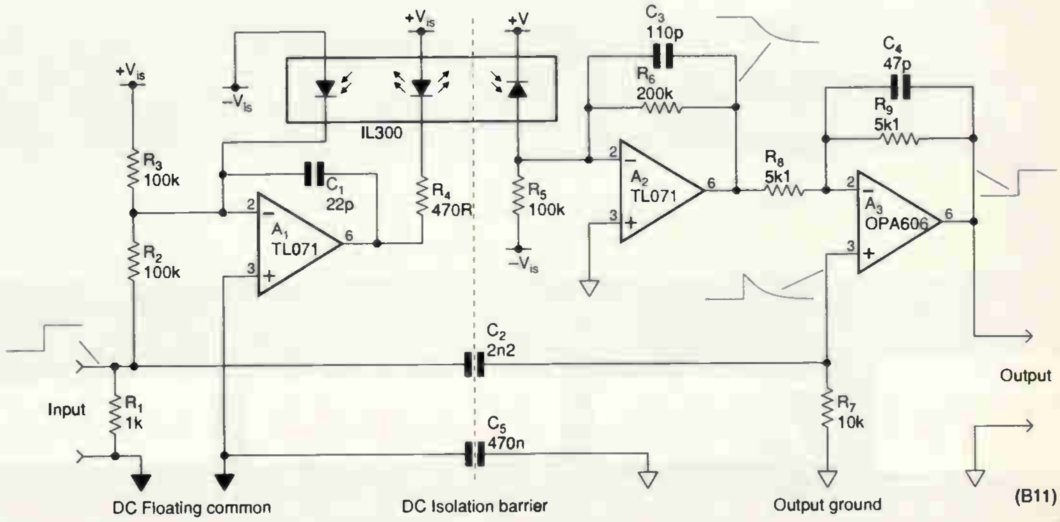

Improved-accuracy current mirror

Current mirrors of the form shown in **Fig. 1** achieves increased accuracy by virtue of the emitter follower to reduce input base current. Assuming that all three βs are equal,

$$A=\frac{1}{1+\dfrac{2}{\beta(\beta+1)}}$$

To improve bandwidth, Tr_1 may be biased by an emitter resistor which, unfortunately, reduces mirror accuracy by the increased base current.

To regain accuracy, the circuit of **Fig. 2**

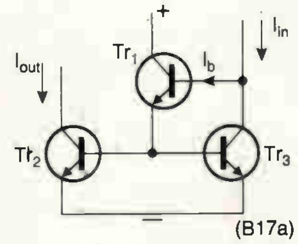

Fig. 1. Emitter follower in this current mirror improves bandwidth, but increased base current does nothing for accuracy.

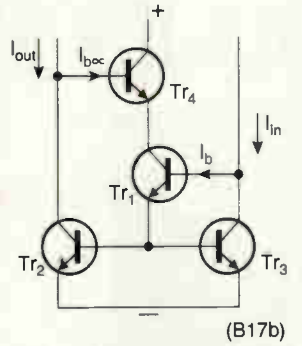

Fig. 2. Further base current added to output restores accuracy while retaining improved bandwidth.

offers an alternative approach by deriving an extra base current and adding it to the output. Again, all βs are equal and gain is now,

$$A=\frac{1+\dfrac{2}{(\beta+1)^2}}{1+\dfrac{2}{\beta(\beta+1)}}$$

Adding Tr_4 does have the effect of slightly lowering output impedance, but the improved accuracy and bandwidth normally outweigh the effect.

Figure 3 shows the final circuit, in which R_1 helps to reduce errors caused by offset voltage and R_2 sets the quiescent current in $Tr_{1,4}$. Reducing the value of R_2 does not affect the accuracy significantly, due to the effect of Tr_4.

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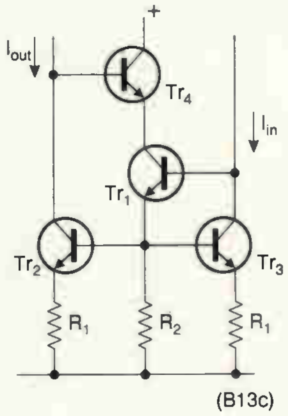

Fig. 3. Practical circuit of the wider bandwidth current mirror with uncompromised accuracy.

Longer-life led flasher

This 1.5V high-efficiency led flasher is an astable multivibrator with voltage doubler. The two transistor pairs are wired in Darlington mode to increase gain and reduce current consumption.

Values of C_1 and C_2 can be changed to adjust the flashing rate. The ultra-bright led is obtainable from RS Components and Farnell Electronics. Ordinary leds can be used, but the flashing intensity is very much reduced since the circuit is trimmed for minimum current consumption.

An ordinary 1N4148 type diode can be used instead of the 1N5819 schottky-barrier diode but led intensity will be reduced due to the higher voltage drop.

Idle current of the circuit is 0.2mA and with the led on, the current goes up to 0.3mA. With such a low current consumption, a 1.5V cell lasts a long time.

The popular *LM3909* led flasher from National Semiconductor idles at typical 0.55mA and with the led active it drains 0.75mA. According to National's handbook the *3909* will run for around 3 months continuously on an ordinary AA sized cell or for around 6 months on an alkaline type cell.

The circuit shown here should give you twice as much battery life.

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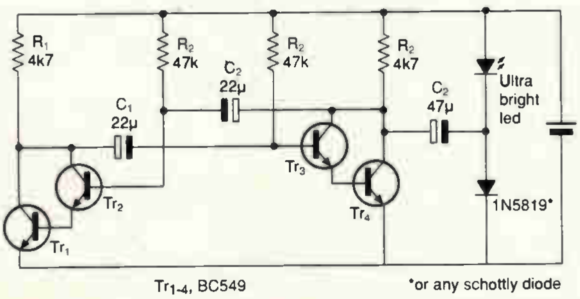
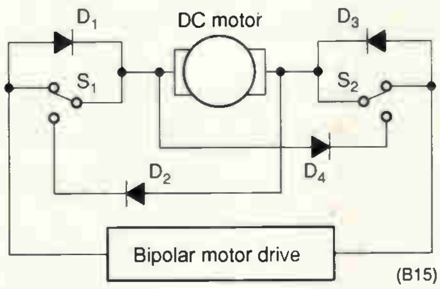


Comprising a multivibrator followed by a voltage doubler, this led flasher should run for up to a year on one alkaline AA cell.



A reliable (and cheap) method of stopping a dc servo motor very quickly without recourse to electronic sensing.

Rapid stop for dc motors

When used in servos, dc motors usually need travel limit switches. This circuit arrangement stops a motor quickly when the switch is tripped.

When the servo reaches a limit, the bipolar drive circuit should be able to reverse the motor; $D_{1,3}$ across the limit switches perform this function.

Dc motors stop much more quickly if a short circuit is placed across the terminals, since the motor then behaves as a generator. Diodes $D_{2,4}$ provide a virtual short circuit in the direction in which the motor is is moving when the limit is reached. This process is particularly suitable when moving high-inertia loads and does not rely on any exotic sensing devices; it is, therefore, more reliable when driving critical loads. It appears simple and somewhat obvious, but we have not seen it published before.

Limit switches and diodes must be suitably rated. ■

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Think of a sinewave, distort it to stabilise the amplitude, take away the harmonics you first thought of, and the answer's a pure sinewave. Not quite that straightforward, but that is roughly what Ian Hickman's latest invention amounts to.

A perfect variable oscillator?

This design more or less materialised by itself – perhaps harmonic serendipity describes it. Low distortion audio frequency oscillators have always held a fascination for me. Several designs have appeared in these pages over the years. Some, like the one described in reference 1, offer a modest performance – around 0.07% THD – while the configuration described in reference 2 improves on this by a factor of 20 or more.

Many low-distortion audio oscillator designs, like the one in reference 3, use a thermistor such as the famous R53, or the R54, for amplitude stabilisation. Though effective, this has certain disadvantages, one of which is the high cost of the device.

The main disadvantage though is a rise in third-harmonic distortion at frequencies below 100Hz, becoming quite marked at 20Hz. This is due to the period of the waveform beginning to approach the thermal time-constant of the thermistor, so that the element tends to cool between each half cycle of the waveform, as it passes through zero voltage.

With this in mind, some years ago I designed a 20Hz-20kHz audio oscillator[4] using diodes to stabilise the amplitude in place of the expensive R53. The advantage here is that the limiting action is totally independent of frequency.

At 0.04% total harmonic distortion plus noise over the whole audio frequency range, its performance was good, bearing in mind the low component cost. But in another application[5] even the diodes were dispensed with. Here, the oscillator consisted of a frequency-selective amplifier whose gain was allowed to reach infinity. The relationship between oscillators and selective amplifiers was discussed in reference 6 which, though predating op-amps, nevertheless contains some interesting ideas.

Another look at the problem

Being now equipped with an *HP3580A* audio frequency spectrum analyser, I decided it was time to take another look at just how good a low a distortion oscillator could be designed, without recourse to a thermistor for amplitude stabilisation.

As an exercise in making a silk purse out of a sow's ear, that standard low-cost standby of FET input op-amps, the *TL084* was selected. I started with the circuit of **Fig. 1**, which, apart from the frequency, is the same basic circuit as in reference 5. This circuit produces an output at approximately 1kHz, the peak to peak amplitude just reaching to the supply rails.

Output spectrum is shown in the upper trace of **Fig. 2**. This shows the measured at the output of the second integrator, deliberately plotted with rather a wide IF bandwidth. The largest harmonic is the third, being 63dB down on the fundamental or 0.07%. Other harmonics are about 70dB or more down.

Amplitude stabilisation took the form of reducing loop gain at the fundamental component, due to the output hitting the supply rails. In theory, the loop phase shift is exactly 180°, corresponding to a circuit Q of infinity; in practice, the phase shift will be a smidgeon more, due to the inevitable residual phase lag in the integrators, even though this is very small at 1kHz.

The output waveforms of all three op-amps in the circuit were substantially similar. All were limiting gently, to slightly different degrees. The obvious next step, therefore, was to confine the limiting to the first op-amp, so that the harmonic distortion products caused by the limiting were attenuated by the the 6dB/octave roll-off of each of the two following integrators.

This was achieved by raising the gain of the first op-amp, so that the amplitude at its output – the high-pass node of the

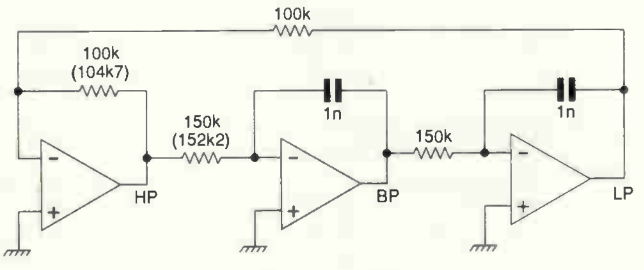

Fig. 1. Simple low distortion sinewave oscillator

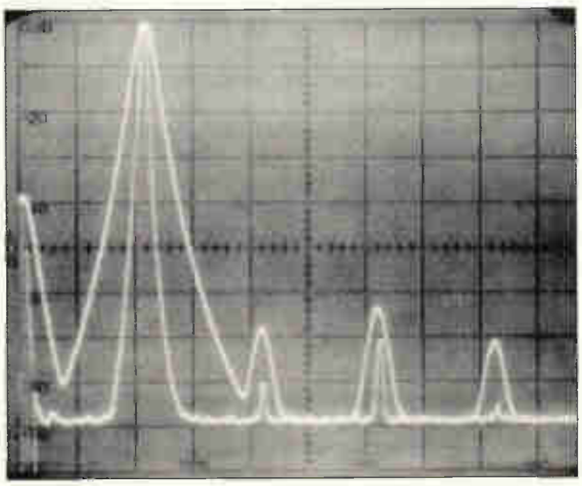

Fig. 2. Upper trace, output of the circuit of Fig. 1. Span 0-5kHz, 100Hz resolution (IF) bandwidth, video (post detection) filter off.
Lower trace, as above, but with component values in brackets, resolution bandwidth 30Hz.

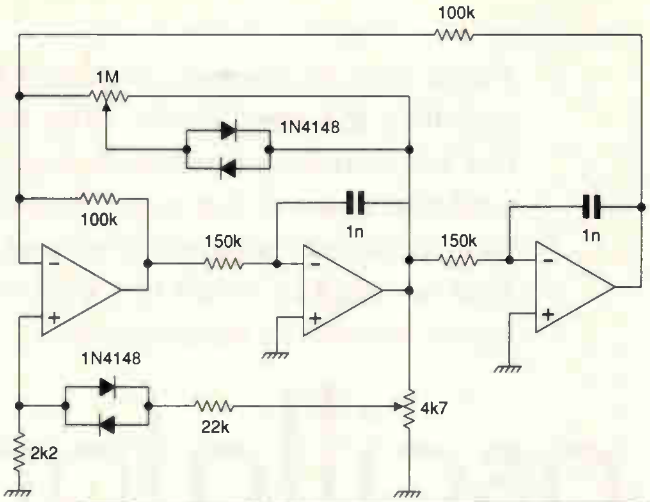

Fig. 3. Modified circuit designed to produce little or no third harmonic component at the lowpass output.

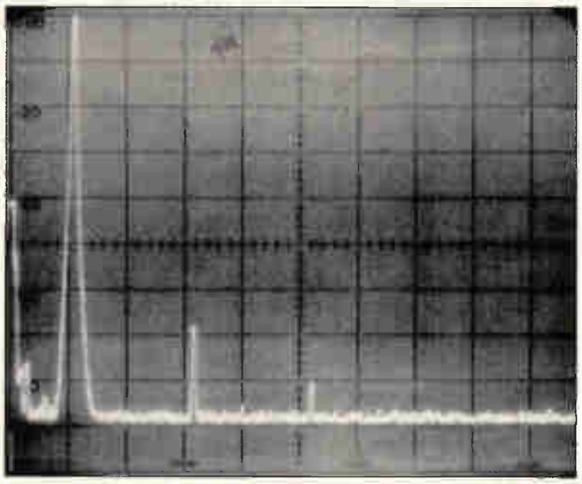

Fig. 4. Output of the circuit of Fig. 3. Span 0-10kHz, 100Hz resolution (IF) bandwidth, video filter off.

filter – was larger than at the output of the second integrator – the low-pass output. Its feedback resistor was therefore raised by 4.7kΩ, the integrators between them now providing a gain of 4.7% less than unity.

To do this, the frequency automatically adjusts itself up a little. If the integrator components remained the same, the first integrator would still run at a slightly higher amplitude than the second, so its input resistor was raised by 2.2kΩ, giving roughly equal amplitude non-limiting outputs at the bandpass and lowpass output nodes.

The spectrum was now as shown in Fig. 2, lower trace. For clarity, this has been recorded at a narrower IF bandwidth, and the slight increase in frequency is clearly visible, more so at the higher harmonics.

The third harmonic is now 70dB down on the fundamental, and others 80dB or more down. Incidentally, the reduction in the second harmonic was not all due to the component changes; the opportunity was taken to tweak one of the supply rails to more nearly equalize the clipping of the positive and negative peaks.

For such a simple circuit, a thd plus noise figure of 0.03% is not at all bad, and it is a very useful and economical arrangement for a fixed frequency oscillator. For a wide-range audio frequency oscillator it is less suitable, since the excess phase shift in the integrators will increase with frequency, leading to a much worse distortion figure at 10kHz and above. So my thoughts turned to the type of oscillator described in reference 4.

Diodes for stabilisation

This circuit incorporated negative feedback of such a value as to set the filter to a Q of 11. Positive feedback was applied from the bandpass output, BP, of a value exceeding the negative feedback, effectively providing the filter with an input derived from itself.

However, as the amplitude of the oscillation builds up, diodes start to limit the positive feedback, so that the fundamental component of it ceases to increase, or does so only marginally.

As the negative feedback continues to increase proportionally with the amplitude of the oscillation, an equilibrium point is reached, where the net loop gain at the fundamental just equals infinity. The circuit was arranged so that this occured at a level well below the maximum possible swing set by the supply rails, with the result that the second harmonic distortion was negligible.

An advantage of an oscillator such as this, based upon the state variable filter, is that the circuit can be used both as an oscillator, and simultaneously as a second-order elliptic low-pass filter. Any third harmonic, due to the limiting action of the diodes, present at the high-pass output will be attenuated by a factor of three in each of the integrators.

Furthermore, as each integrator provides a 90° phase shift regardless of frequency, the third harmonic present at the high-pass output is in antiphase to that present at the lowpass output. So by using an additional summing amplifer to combine one ninth of the highpass output with the full lowpass output, the third harmonic sits in a notch.

This results in a signal entirely free of any third harmonic, an arrangement used in reference 4. Meanwhile, the fundamental output is of course also reduced, but only by a ninth.

Handy though this arrangement is, there is a down side to it. While the third harmonic is outphased completely, the fifth harmonic is actually made 5dB worse. But there is still a net reduction in thd, since, due to the fairly soft limiting action of the diodes, the amplitudes of the harmonics at the high-pass output drop off much more rapidly than in an ideal squarewave.

Outphasing the fifth harmonic

It occurred to me that if I could arrange that the amplitude limiting method inherently did not produce third harmonic, then the fifth harmonic could be outphased instead – simply by adding a twentyfifth of the high-pass output to the low-pass output.

Reduction of selected harmonics by waveform shaping is a technique long familiar in power engineering. There, it can be used to minimise the filtering required at the output of an inverter, as in for example reference 7. So I experimented with a circuit designed to avoid producing third harmonic in the stabilisation department, permitting outphasing of the fifth.

This is shown in **Fig. 3**. Oscillation commences since there is positive feedback from the bandpass output via 1MΩ to the inverting input of the first op-amp. Initially, there is no negative feedback via the 22kΩ resistor to the noninverting input, as it is blocked by the diodes.

As the amplitude increases, their forward voltage is overcome, and thereafter the stronger negative feedback increases more rapidly than the positive feedback. This leads to equilibrium at a level which can be adjusted by means of the 4.7kΩ potentiometer. The 1MΩ potentiometer can then be adjusted to feed narrow additional positive feedback pulses at the tips of the waveform.

Following careful setting up, the output spectrum was as shown in **Fig. 4**. The third harmonic at the low-pass output is at –69dB, about as low as the diode-free circuit of Fig. 1. It is also lower – without the benefit of outphasing – than achieved by the circuit of reference 4 with diodes.

But setting up was fairly critical, and performance with tuning and with change of ambient temperature would clearly have been unsatisfactory. Nevertheless, the low level of fifth harmonic was encouraging, so the circuit was rearranged to provide low third harmonic output by means of outphasing.

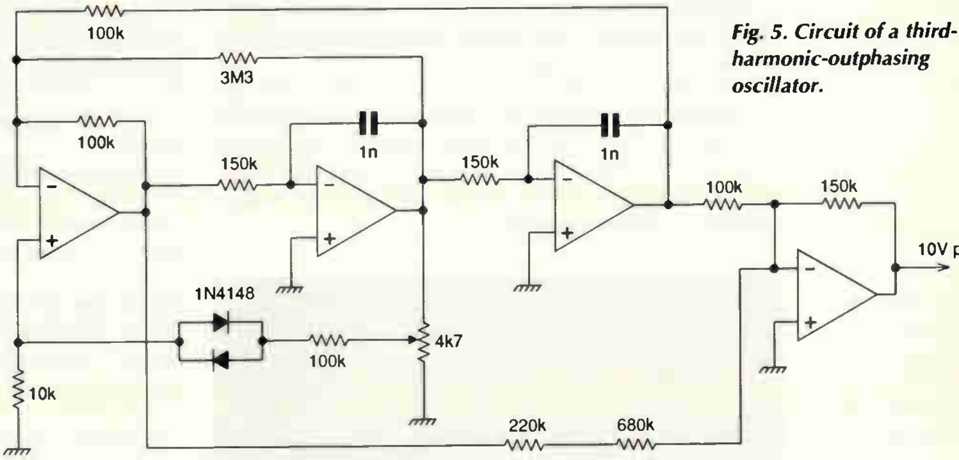

Fig. 5. Circuit of a third-harmonic-outphasing oscillator.

Why not outphase all the harmonics?

This circuit is shown in **Fig. 5**, and its performance in **Fig. 6**. The third harmonic is outphased completely, while compared with Fig. 4, the fifth harmonic has increased by the expected 5dB.

It would be fine though, I thought, if both the third *and* the fifth harmonics could be outphased, and why not the seventh and the others to boot? This led to the experimental arrangement of **Fig. 7**.

Here, the amount of positive feedback is fixed by the 3.3MΩ resistor, the desired amplitude of oscillation being set by the amount of negative feedback via the diodes. Having constructed the circuit as shown, with the 100kΩ potentiometer at mid-travel, on switching on, the spectrum which met my unbelieving gaze is shown in **Fig. 8**.

Where had all the harmonics gone? The following day, they were back again, but only just above the –89dB noise floor of the spectrum analyser. Successive adjustments of the two potentiometers nearly reproduced the previous day's astounding results, but the settings were critical, and the amplitude of the oscillation very unstable. So it was time to don the thinking cap.

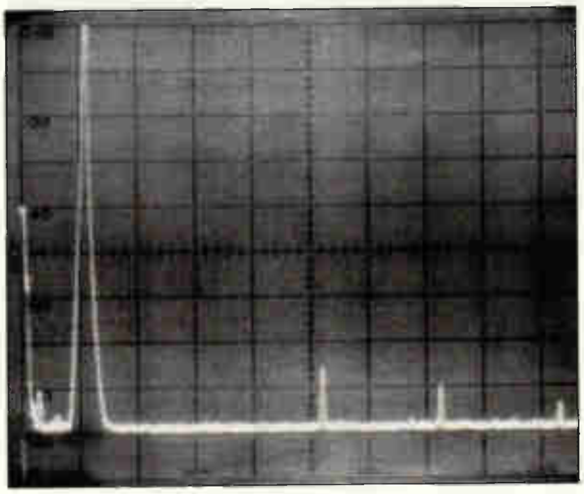

Fig. 6. Spectrum of the 10V peak to peak output of the circuit of Fig. 5. Spectrum analyser settings as for Fig. 4, but with video filter set to 'medium'.

The intention had been to use op-amp IC_4 as a difference amplifier to pick off the harmonics generated by the diodes. These were then fed into the inverting input of IC_1. With the identical harmonics fed also into the non-inverting input as part of the negative feedback signal, the harmonic distortion should be rejected by IC_1 as a common mode signal.

The idea is sound, but it is essential that the harmonics, and only the harmonics, be fed back to the inverting input of IC_1. For if the fundamental component of the near-squarewave appearing across the diodes be fed back as well, the amplitude stabilisation mechanism is defeated. It is the 'dead-space' created in the negative feedback signal by the diodes that ensures oscillator start-up.

Thereafter, when the signal becomes large enough for the diodes to start to conduct, the negative feedback increases with the amplitude of oscillation, faster than the positive feedback increases.

If the fundamental component of the diode drop is now fed back along with the harmonics, via the difference amplifier, to the inverting terminal of IC_1, then regardless of level, the ratio of positive to negative feedback becomes independent of level, and amplitude stabilisation is lost.

So the circuit was modified to make IC_4 a true difference amplifier, as shown in **Fig. 9**. Note that the ratio of the two resistors at the inverting input of IC_4 is equal to the ratio of the resistors at the other input – although these in fact also do double duty as the input divider chain for the negative feedback to IC_1.

With the wiper of R_B at the top of its travel – point B in Fig. 9 – IC_4 is a classic difference amplifier, with a near squarewave output representing the voltdrop waveform across the diodes. This, however, is not quite what is wanted; it contains the missing fundamental component of the negative feedback as well as the harmonics, as shown in the upper trace in **Fig. 10**.

The waveform input to the diodes is shown in the upper trace of **Fig. 11**, and note that the upper trace in Fig. 10 is in antiphase to this. Thus the voltage applied via the diodes and

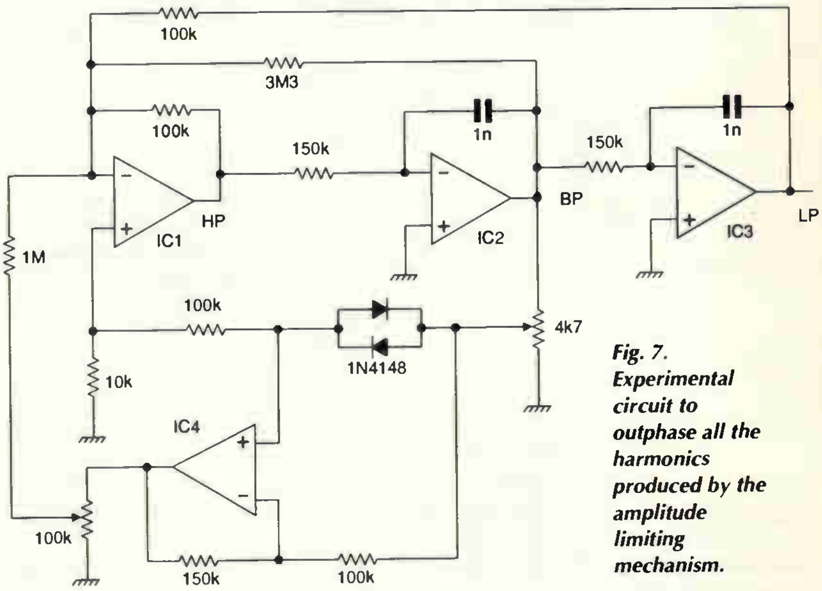

Fig. 7. Experimental circuit to outphase all the harmonics produced by the amplitude limiting mechanism.

resistors, as negative feedback to IC_1, is as shown in Fig. 11, lower trace. Note the dead-space, rather like an underbiased class B output stage.

This illustrates an important point. When any mechanism introduces harmonic distortion onto a pure sinewave, the amplitude of the fundamental component of the distorted waveform will differ from that of the original.

To ensure outphasing of the harmonics by common mode cancellation in IC_1 without upsetting the amplitude stabilisation, the waveform fed to that op-amp's inverting input should consist of the harmonics only.

These only exist at the downstream side of the diodes. The fundamental component of the near-square waveform across the diodes represents the amount by which the fundamental input to the diodes exceeds that at their output. The difference amplifier IC_4 can be arranged to see a slightly smaller sinewave at its inverting input.

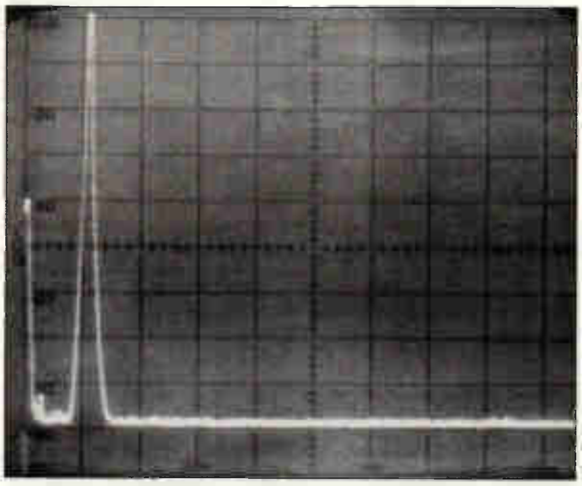

Fig. 8 . Output of the circuit of Fig. 3. Span 0-10kHz, 100Hz resolution (IF) bandwidth, video filter off.

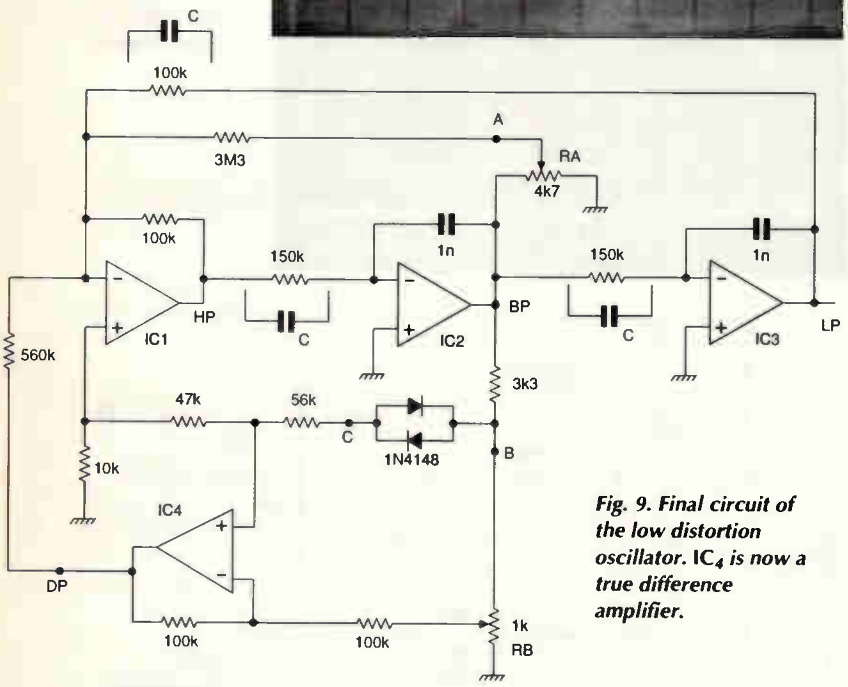

Fig. 9. Final circuit of the low distortion oscillator. IC_4 is now a true difference amplifier.

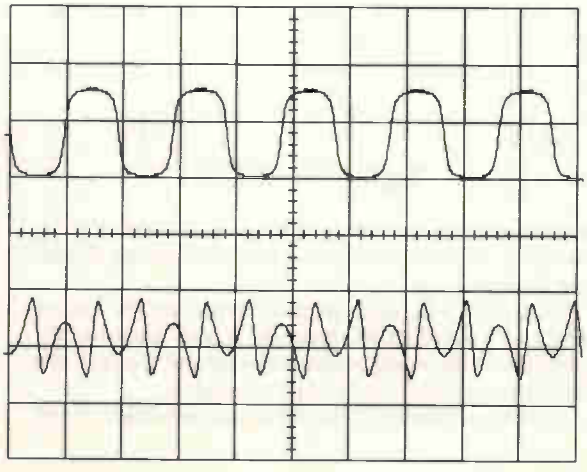

Fig. 10. Upper trace (1V/div), the waveform appearing across the diodes contains a component at the fundamental frequency. Lower trace (0.2V/div), as the upper trace, except with the fundamental component removed. Horizontal: 500µs/division.

Resistor R_B in Fig. 9 allows this to be adjusted to equal the amplitude of the fundamental component of the negative feedback signal shown in the lower trace of Fig. 11. The result is that the output of the difference amplifier IC_4 then contains the harmonic components *only*, as shown in Fig. 10, lower trace.

This adjustment is very easily carried out, by displaying the spectrum of the waveform at the distortion products measuring point DP in Fig. 9, on a spectrum analyser, and adjusting R_B to minimise the fundamental component displayed.

Measuring the results

Output of the circuit of Fig. 9, measured at the point LP, displayed a third harmonic component just visible above the spectrum analyser's noise floor, with higher harmonics not visible at all.

To increase the measurement range, a twin-tee network was made up, using four 15kΩ, 1% metal film resistors and four 10nF, 2.5% polystyrene capacitors. Figure 9 used 2.5% 1n polystyrene capacitors, and 1% metal film resistors throughout by the way.

Response of the twin-tee was plotted, using the spectrum analyser's internal tracking generator, and is shown as the upper trace in **Fig. 12**. With no adjustment, the twin-tee notch reduced the fundamental of the oscillator by 45dB, permitting the input level to the spectrum analyser to be increased by 40dB without overloading.

The measurement noise floor was now about 120dB below the oscillator's output, as can be seen in Fig. 12, lower trace. Allowing for the residual attenuation of the notch, the third harmonic is seen to be at –88dBc, and the fifth at –104dBc, while the second and fourth are lost in the oscillator's phase noise sidebands.

Some practical extensions

While the circuit has not to date been developed into a complete 20Hz to 20kHz audio oscillator, some initial tests were carried out to see if this would be practicable.

To test the performance with change in ambient temperature, an electronic thermometer was used in conjunction with a hair-dryer. The thermometer probe was placed adjacent to the diodes, and the hairdryer at a fair distance, so that the whole circuit was bathed in the same air stream, and left to reach equilibrium.

The temperature was thus raised from 20° C to 30°C. The oscillator's output level rose by roughly 2dB, but the distortion figure remained unchanged.

Next, the circuit was modified to operate at 10kHz by changing the two integrator 150kΩ resistors to 15kΩ. As expected, amplitude control was lost, due to the increased residual phase shift in the integrators. This was largely corrected by fitting three 1.8pF capacitors shown as *C* in Fig. 9.

Some re-adjustment of R_A was still necessary, but thereafter, the distortion was, as far as could be determined without another twin-tee filter, similar to that measured previously.

The exact value required for the compensating capacitors *C* will depend upon layout, and may well be 1.8pF or even less with a good pcb design – the experiments were carried out using, for speed, *Experimentor* prototyping plug-board. This exhibits a fairly high stray capacitance between adjacent points, rendering it completely unsuitable for rf work.

Even at audio, the strays can, as here, pose a problem. Bandswitching and fine frequency control, to cover 20Hz to 20kHz, can be implemented as in reference 4.

Figure 13 shows some simple additions which could be accommodated with an additional quad op-amp. In theory, IC_5 should outphase even the residual third harmonic distortion visible in Fig. 12 – but don't expect perfection. After all, the typical thd of the *TL084* used in all the circuits illus-

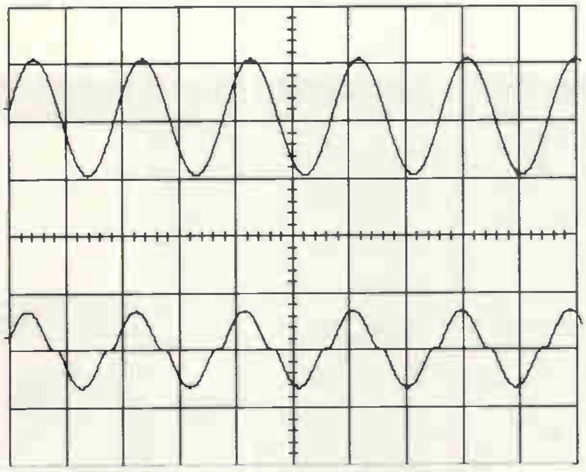

Fig. 11. Upper trace, the waveform at point B in Fig. 9. Lower trace, the waveform at point C in Fig. 9. Both traces, 2V/division vertical, 500μs/division horizontal.

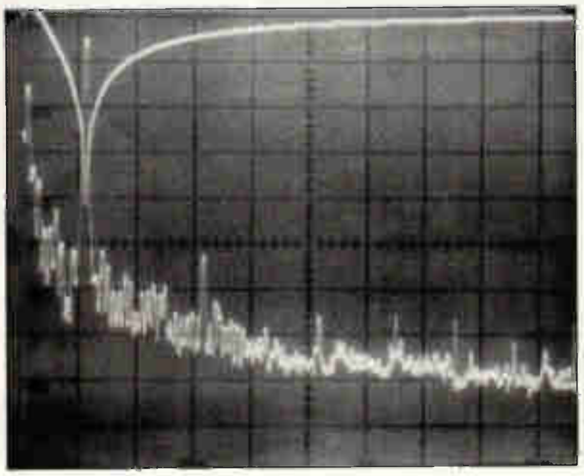

Fig. 12. Upper trace, the response of a 1kHz (nominal) notch filter. Lower trace, the LP output of the circuit of Fig. 9 with the fundamental in the notch, and the sensitivity of the spectrum analyser increased by 40dB. Span 0-10kHz, 30Hz bandwidth, video filter at 'medium', 20s/div sweep speed.

trated, is listed as 0.003% or –90dBc.

Together with IC_5, IC_6 provides a choice of a 600Ω output, or a variable voltage source for driving high impedance loads. Amplitude monitoring is provided by IC_7, enabling any output level variations with temperature to be trimmed out, by adjustment of R_A.

The value of R_m is determined by the sensitivity of the meter used, while another resistor R_p, of suitable value, may be included if desired, for meter protection in the event of overload.

Finally, IC_8 provides a means of adjusting R_B so that the output of difference amplifier IC_4 contains zero contribution at the fundamental. As R_B is adjusted, the sudden complete disappearance of the fundamental is very obvious, no need for the aid of a spectrum analyser

A switch to turn the stage off once the adjustment is complete, is clearly a good idea. Resistors R_A and R_B should preferably be multi-turn types, and for convenience can be made adjustable via the oscillator's front panel.

Setting up

Setting up the circuit for optimum performance is very straightforward, and no special equipment is needed. The sequence is as follows.

Disconnect the 560kΩ resistor from the point DP and connect it to ground instead. This isolates the effect of R_B, preventing it interacting with the next step, while ensuring the effective positive feedback ratio is unchanged.

Set R_A to give the desired output level. Assuming 15V supply rails, 10V peak to peak is a suitable value.

Using the audio monitor of Fig. 13, adjust R_B for zero fundamental component in the output of IC_4; this is conveniently carried out at 1kHz.

Disconnect the 560kΩ resistor from ground and reconnect it to IC_4's output. Check that this does not alter the level at the LP output. ■

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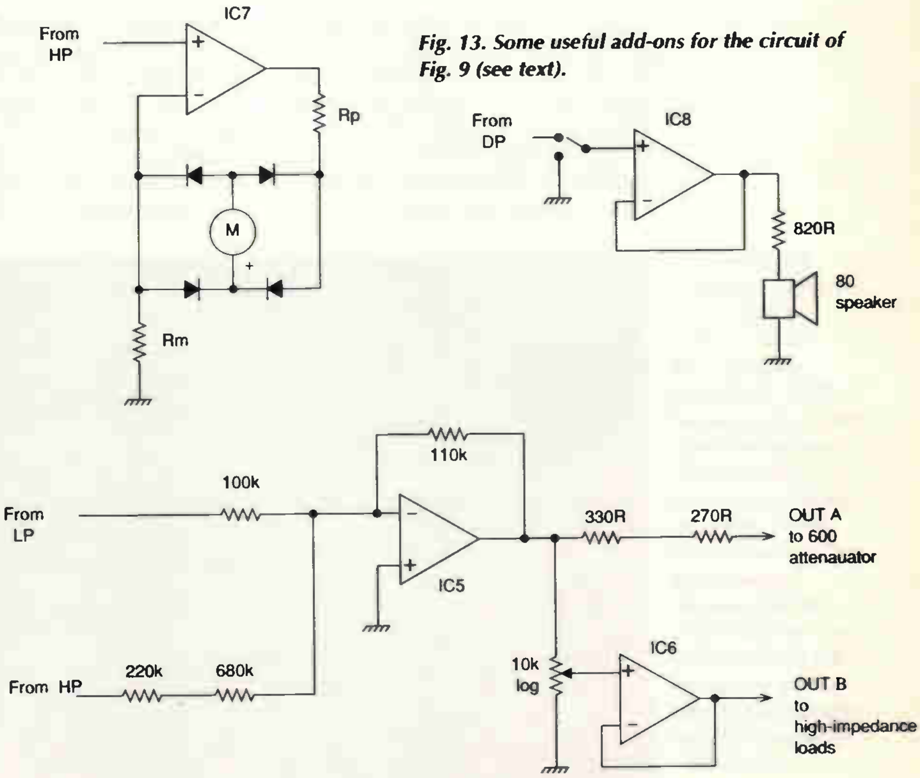
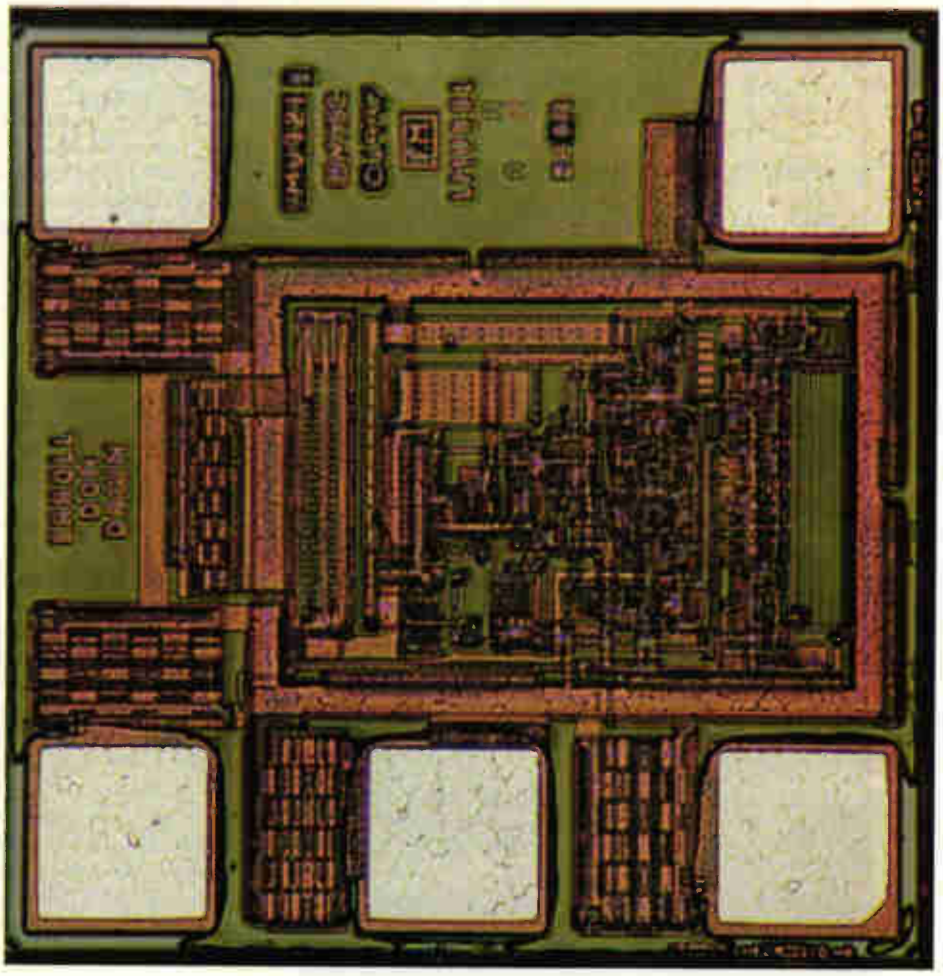


Fig. 13. Some useful add-ons for the circuit of Fig. 9 (see text).

Born again



Richard Ball reports on how *LM300* family op-amps – which are now a quarter of a century old – have been relaunched with a familiar feel but a new set of specs.

How many op-amps can you get from a six-inch wafer? In National Semiconductor's case, with its *LMV300* family, then the answer is an incredible 80000.

National claims it has re-invented the op-amp. The truth is less dramatic, but is certainly significant.

A quarter of a century after its introduction, the company has redesigned the *LM300* family of op-amps using a relatively advanced 0.8μm BiCMOS manufacturing process. For designers of modern equipment, the most important outcome is that supply voltage now ranges from 2.7 to 5.5V. Other parameters have been improved, and because of the huge numbers of die per wafer, cost is actually reduced.

In order to get 80000 die from a wafer, National reduced the size of the pads from four to three mils (75μm). The scribe street – the gap between each die removed when the wafer is cut with a diamond saw – was also reduced to 50μm. National got its supplier to redesign the saw to get this narrow cut.

The LMV321, a single op-amp, is available in the world's smallest package – the SC70-5. The five-pin device measures just 2mm on each side. By summer, the company expects to have a chip scale package, with solder bumps placed directly over the pads on the device. But placing such a device may prove difficult with the die being only 0.43mm along a side.

Cyrille Claustres, National's European marketing manager, explained the company's reasons for the redesign. "Many op-amps were introduced up to 25 years ago. They cannot meet today's voltage and packaging requirements," he said.

Devices that can do the job are available, but as Claustres said: "Specialised low power, small products cannot meet the requirement of low cost."

The redesigned family includes the *LM324*, arguably the most popular device used today.

Continued over...

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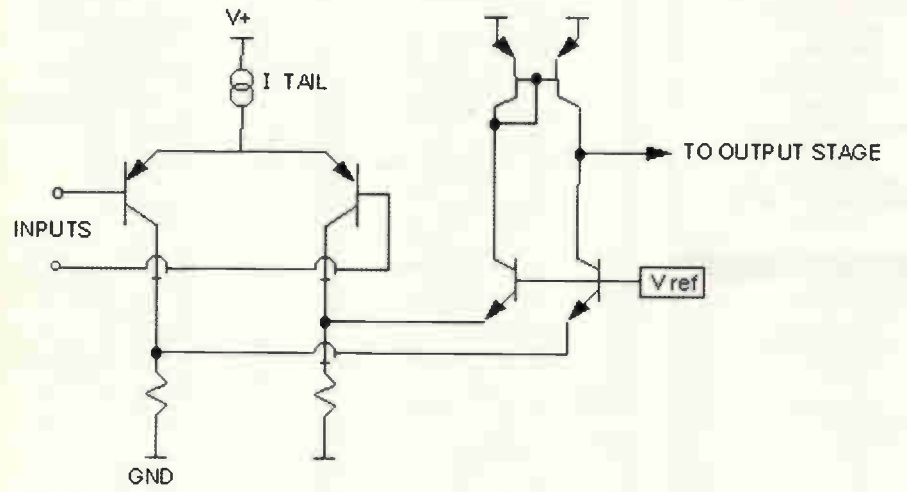

The input section of the op-amp uses a differential pair with a folded cascode stage.

Both op-amps and comparators have been announced, in single, dual and quad form. Part numbers are *LMV321, 358, 324* for op-amps and *LMV331, 393* and *339* for comparators. These devices are drop in replacements for their *LM* ancestors. "Very often when we introduce a new product, designers have to relearn how to use it – not with the *LMV*," said Claustres.

Prices range from 26¢ for single and dual devices in SOIC packages to 30¢ for a quad *LMV324* or *339* device in a TSSOP.

In order to meet, and in some cases exceed, the specification of the *LM300* devices, some careful design was required. The major change is of course the use of a BiCMOS process. This has both advantages and disadvantages. Because the process is less than a micron, thousands of die are available from a wafer, reducing cost even though there are four times as many transistors than the original design's 13.

If CMOS devices are available, then power consumption can be reduced. Bipolar transistor have better matching characteristics and drive capabilities. The main downside of BiCMOS is that it is essentially a digital process, modified for analogue devices. "We had to come up with a completely new architecture," admitted Claustres.

Just to make the redesign even more difficult, the designers set themselves the task of eliminating as far as possible the crossover distortion in the *LM300* op-amps. "*LM324*s are not stable in some applications due to crossover distortion," said Claustres.

The new design, with a class AB output amplifier, has rail to rail outputs with little distortion. Tested in a typical audio configuration, the device had a signal-to-noise ratio of 71dB and total-harmonic-distortion plus noise figure of 0.02%.

On the output, CMOS transistors made rail to rail output easier to implement and maintained a 1MHz gain bandwidth product with up to 200pF of output capacitance. At low values of output capacitance, the gain bandwidth product increases to 3MHz.

National is being deliberately secretive about the actual design of the output stage. Chief designer Erroll Dietz would only say: "All I would like to release for now is that the output stage has an extra gain stage sourcing current and uses an n-p-n for sinking and a PMOS for sourcing."

Another problem with the original design is the input common mode voltage range. The inputs cannot go closer than 1.5V to the supply. With a 15 or 30V supply this is fine, but at 2.7V, this does not leave much room for manoeuvre.

National solved this by changing the front end to a bipolar differential pair with a folded cascode configuration. The original design also used a differential pair, but needed a darlington pair on the inputs.

This new arrangement allows the input common mode voltage range to extend to ground and to within 0.8V of the positive supply. Using bipolar transistors rather than MOS types on the input improves noise and matching characteristics.

The majority of the remaining circuit uses CMOS which brings the benefit of reduced power consumption.

More modern and stringent electromagnetic copmpatibility tests have also been taken into consideration. Electrostatic discharge protection of 250V on the old devices has been increased to 900V on the *LMV321* and 2000V on the dual and quad devices.

National is taking the opportunity to design a range of devices on its new BiCMOS process. "The problem with this process is it's pad limited," said National's Seamus Coyle. So the company is looking at further integration. "We will put a three op-amp instrumentation amplifier onto one die," he said. This combines four op amps and eight resistors on a single die. The die could easily be squeezed into an SOT23-5 package.

There are many other building blocks that could be integrated. "We also have an a-to-d converter in this process," Coyle said. ■

Specification table

Supply current	100µA/channel @ 2.7V 130µA/channel @ 5.0V
Input offset voltage	7mV (max) @ 25°C
Input bias current	11nA (typ) @2.7V
Common mode rejection ratio	63dB (typ)
Power supply rejection ratio	60dB (typ)
Input common mode voltage range	V_-–0.2V to V_+–0.8V (typ)
Output swing	V_-+120mV to V_+–40mV into 2k (typ)
Slew rate	1V/µs
Gain-bandwidth product	1MHz with *C*=200pF
Phase margin	60°

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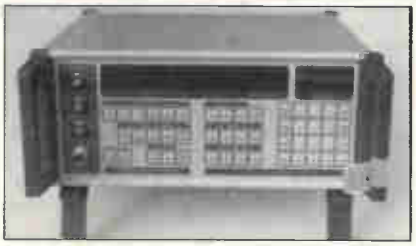

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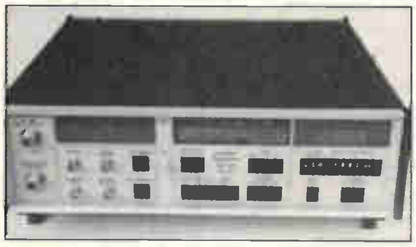

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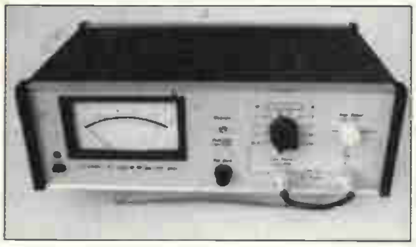

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Is it necessary to re-form electrolytic capacitors that have been stored for years? Why do tantalum capacitors fail short circuit? Do electrolytic capacitors have an extremely high dielectric constant? Cyril Bateman knows.

Understanding capacitors

Aluminium and tantalum options

Aluminium and tantalum electrolytic capacitors uniquely balance *CV* product against physical size. While film or ceramic capacitors are readily available up to 10μF, higher capacitance values are physically large and very expensive. Very large value electrolytic capacitors are available at low cost, and in physically small packages.

The possibility of forming an insulating film on the surface of aluminium was first observed by Wheatstone in 1854. In 1908 a rolled type aluminium capacitor was produced in USA and a viscous electrolyte of ammonium tetraborate and glycerine was developed in Germany.[1] These two essential discoveries resulting ultimately in the modern electrolytic capacitor.

Regardless of construction or dielectric used, every capacitor is composed essentially of two conducting surfaces, called electrodes, separated by an insulator.[2] Insulation materials produce differing capacitance values for the same area and thickness, according to their dielectric constant, or 'K' value.

The permittivity of free space is used as the base unit since all other materials – including air – exhibit increased capacitance. These K values, which range from 1.00059 for air to greater than 12000 for high-K ceramics, significantly influence the capacitance achieved.

Electrolytic capacitors are manufactured using so-called valve metals, the most common being aluminium and tantalum. Valve metals have the ability to form an insulating, semi-conducting and protective surface oxide film. Aluminium oxide has a K value of 8, while that for tantalum pentoxide is 27.6, approximately.[3]

Aluminium oxide and tantalum pentoxide films grown on the highest purity metals provide very high-quality, low-loss insulators. Aluminium oxide has a dielectric strength approaching the theoretical strength[4] as predicted by the ionic theory of crystals.

Capacitance value depends on the product of the electrode area and the K value of the insulating dielectric. It is inversely proportional to the distance separating the electrodes. For any chosen dielectric, the capacitance attained depends totally on the insulators thickness and surface area.

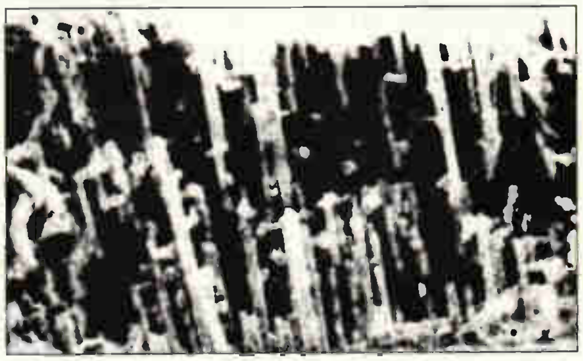

Fig. 1. Edge on view, much enlarged, of a 100 micron thick etched and formed foil showing large tunnels connecting both foil faces. Taken using an electron microscope. This foil has become porous to liquids.

When a flexible electrode system that conforms precisely with the insulator's surface is used. The effective or apparent area can then be increased by roughening or abrading, increasing capacitance without increase of physical size. This effect is called surface gain.

Frequently, electrolytic capacitors have been described as having an extremely high dielectric constant. This is wrong, the K values of 8 and 27.6 are correct,[3] so how can such high capacitance values be attained?

Compared to metallised-plastic film capacitors, which use dielectric film thickness of a micron and above, electrolytic capacitor dielectric films are some fifty times thinner.

This extremely thin dielectric results in electrolytics being the most highly voltage stressed of all capacitors. However this combination of a very thin dielectric having a K value of 8, or even 27.6, still does not explain how such large capacitance values can be attained.

Surface gain – aluminium

I mentioned roughening or abrading the insulators surface to increase its apparent area. You can see an example of this by closely examining a piece of aluminium kitchen foil. Usually, one side is smooth and shiny, the other matt. Under a times-ten magnifying glass, this matt surface is visibly embossed, providing some surface gain.

The earliest electrolytic capacitors attained usable surface gains by spraying molten aluminium onto a carrier. Known

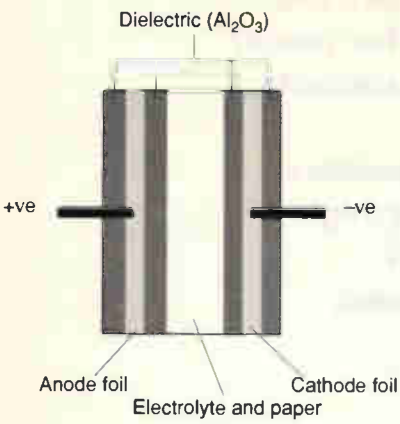

Fig. 2. Simplified sketch of an aluminium electrolytic capacitor, showing anode and cathode aluminium oxide dielectric films in contact with highly conductive electrolyte.

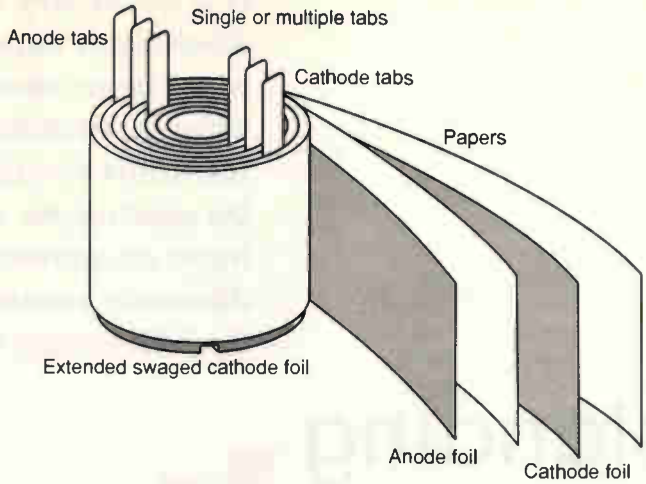

Fig. 3. This view of a partially wound element, shows a large multiple tabbed radial leaded capacitor. An alternative arrangement using a single anode and cathode tab is more common. Also shown is an extended cathode foil, which improves heat dissipation. With its turns swaged to connect each other, this extended cathode foil can substantially reduce capacitor equivalent series resistance.

as 'fabricated plate', this technique was popular for high voltage capacitors. It was quickly superseded for low voltage capacitors though by chemical etching methods.

Today, the surface gain of aluminium electrolytic capacitor foils is attained by electro-chemical etching in a bath of electrolyte. Reduced to basics, an electrolyte is simply a liquid that conducts electricity, having a low resistance or high conductivity, **Fig. 1**.

Common examples of electrolytes are car battery acid and sea water. The aluminium to be etched is placed in a bath of electrolyte and current is passed. There's more about this in the panel entitled 'Aluminium etching.'

Surface gain – bead tantalum

The modern bead tantalum capacitor is fabricated from fine particles of tantalum powder surrounding a small tantalum wire. These particles are compressed into pellets or slugs. Compression pressure is controlled to restrict the pellet's density.[5] The object is to produce a porous body having a very large internal surface area.

To increase the cohesion of these tantalum particles and ensure electrical conductivity, the compressed slugs are sintered at very high temperatures in vacuum furnaces. This provides a mechanically rugged, porous structure.

The high sintering temperature performed in vacuum has a purifying effect. Tantalum is a refractory metal and needs a sintering temperature approaching 2000°C. At such temperature, most impurities evaporate, and are removed by the vacuum pumping system.

Having attained a large surface area, both aluminium and tantalum oxide dielectrics are grown *in situ* onto all-metal surfaces. This involves an electro-chemical process called 'forming'. The 'formed' oxide fully covers and insulates all visible metal surfaces and all invisible cavities within the metal.

Forming aluminium oxide

Placed in a bath of suitable electrolyte and with a positive voltage applied to the foil, aluminium oxide, Al_2O_3, grows on the foil's surfaces.

Depending on the electrolyte used, two main forms of this oxide can be grown. A hard non-porous oxide layer is deliberately grown when the aluminium foil is to be used as a capacitor dielectric or other insulator. When the aluminium is to be 'anodised' to provide a coloured decorative finish, a porous oxide is grown.

An aggressive electrolyte simultaneously re-dissolves some existing oxide, resulting in a porous oxide layer. This process is commonly called anodising. In other applications, the pores in the oxide are be filled with coloured dyes, resulting in the popular decorative finish.

Using a non-aggressive electrolyte, a slowly growing, impervious and continuous oxide film is obtained. This non-porous, hard oxide film grown on super-pure aluminium is an excellent insulator that can be formed to withstand at least 600 volts.

I mentioned earlier that the dielectric oxide formed is very thin. Its thickness is self limiting. Oxide growth at any one voltage slows and almost ceases with time, ultimately attaining some 14Å thickness[3] for each volt applied.

Foil used in commercial capacitors is formed at a voltage at least 120% of the capacitor's rated voltage. This ensures low leakage current and provides the overvoltage capability needed for the 'surge voltage' claimed.A commercial 10V aluminium electrolytic capacitor will have a dielectric thickness of some 0.02 microns. There's more on this topic in the panel entitled 'Aluminium forming'.

Forming oxide for tantalum devices

Being less reactive than aluminium oxide, tantalum pentoxide is generally grown in an acidic bath. Since the current paths into the centre of a slug can be quite tenuous, a strong, highly conductive electrolyte is used.

Tantalum pentoxide is an excellent insulator, growing to a thickness of 17Å per formation volt applied to the slug.[3] This oxide makes a highly amorphous film, but if subjected to a high field strength at a fault site, it can become crystalline, resulting in a short circuited capacitor.[5] To reduce the risk of this happening, tantalum capacitors are usually formed at much higher voltages than are aluminium capacitors, for the same rated voltage.

Putting it together

Mainstream aluminium electrolytic capacitors are wound then impregnated with liquid electrolyte, while most tantalum capacitors are porous slugs impregnated with a dry electrolyte. However many permutations on these techniques are possible.

Dry or solid aluminium electrolytic capacitors offer an intermediate approach between these two extremes, and wound foil tantalum also wet electrolyte slug tantalum capacitors are also produced.

Wet tantalum capacitors are usually fully hermetically sealed using glass metal seals, for good reason. While protecting the capacitor from moisture, the glass seals also protect the outside world from the electrolyte. For example, sulphuric acid or lithium chloride based electrolytes have been used.[6]

Many wet tantalum capacitors use similar strong electrolytes, so casual dismantling is not advised.

For brevity I only detail the mainstream products here, i.e. those that are commonly available from distributors.

Aluminium – winding methods

Aluminium is non-rusting because, on exposure to air, it forms a protective visually transparent oxide film on its surface. If this naturally occurring oxide film is mechanically removed, aluminium has a characteristic smell, which disappears as the oxide film regrows.

As a result, all aluminium electrolytic capacitor foils and connecting tabs, whether deliberately formed or not, are covered in oxide. Obviously thickness of the oxide coating differs. Naturally occurring oxide is extremely thin, equivalent perhaps to some 1.5 electrical volts of formation. The lowest voltage foil is formed to 8-10 volts, and is thus 5 or 6 times thicker.

Similar in construction to the foil and paper capacitors described in a previous article,[7] aluminium electrolytic capacitors comprise two aluminium electrode foils, interwound with separating papers and using tab connections. The whole winding is then vacuum impregnated.

Here all similarity ends. Being an extremely good insulator, the impregnant used in paper capacitors is the true dielectric. Aluminium capacitor electrolyte on the other hand is an extremely good conductor, the aluminium oxide layers form the dielectric, **Fig. 2**.

When the lowest possible equivalent series resistance is needed, multiple connecting tabs, dispersed along the foils, are common – especially for large windings. Smaller low esr windings can be wound using single tabs, connected to the centre of each foil.

The cathode foil only of low voltage radial capacitors can also be wound extended beyond the papers along one edge. Each cathode turn can then be short circuited to all other cathode turns by passing over the rim of a high speed rotating aluminium 'swaging' wheel, **Fig. 3**.

I used this intermediate approach many years ago to design some very low profile 4000µF, 63V capacitors. These were custom designed for the first mass-produced 100W per channel, hi-fi stereo amplifier. They probably represented the first production capacitors to combine a central tabbed anode with an extended and swaged cathode foil. The construction ensured extremely low esr and inductance, in a case 45mm diameter and only 50mm tall.

Most small aluminium electrolytic capacitors however are wound using a buried-foil winding. External connections are made using two formed aluminium connecting tabs, frequently placed towards the outer end of the winding.

One exception to this is some very small diameter axial capacitor elements that are wound onto an aluminium spindle or riser. The anode foil is attached to this riser, which remains as part of the finished capacitor, **Fig. 4**.

The cathode connection for this style of axial capacitor can be made using short tabs connected to the outer end of the cathode foil. Pressure contact with the capacitor case is made by trapping the tabs between the case and its rubber end-seal. Alternatively an extended cathode tab is welded to the bottom of the case, before inserting the wound element. These constructions exhibit from 20 to 85nH of self inductance, according to case size.

Inductance myth

Since these capacitors are wound, many writers erroneously ascribe very high self-inductance as being inevitable in aluminium electrolytic capacitors. The largest inductance value found browsing several catalogues, was less than 100nH. This is equivalent to some 10-12cm of connecting wire or printed circuit track.

Radial-mounting electrolytic capacitors use multiple distributed tab connections, or single tab connections made close to the mid-turn of each winding. These exhibit an extremely small winding inductance – less than 20nH and 35nH respectively.[8] The major contribution to this is from the connecting tabs and the capacitors leadout wires, **Fig. 5**.

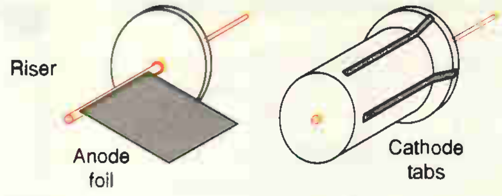

Fig. 4. One common construction used for the smallest axial wound capacitors. Disadvantages are its pressure contact between cathode tabs and case and increased self inductance due to foil connections being at extreme opposite winding ends. But with small windings, self inductance remains acceptable.

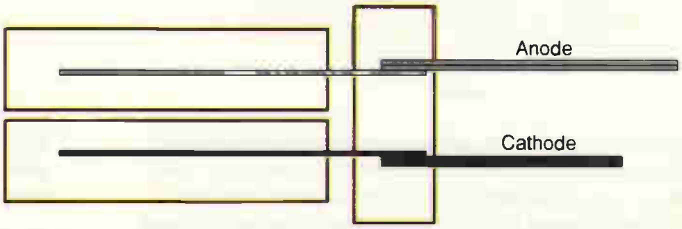

Fig. 5. Sectional view of typical small radial capacitor showing centre foil tab connections, minimising self inductance, which is typically much less than 20nH.

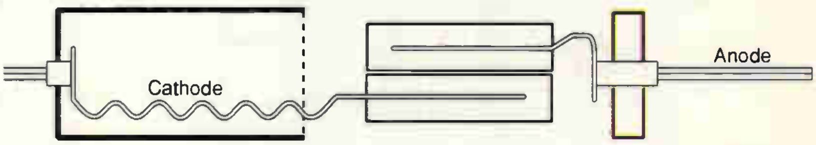

Fig. 6. Sectional view of a typical axial capacitor construction used for medium and larger sizes. While having centre foil tab connections, self inductance is increased due to the excess cathode tab length needed to permit assembly.

Axial-leaded capacitors exhibit more inductance. While larger sizes may be centre tabbed, connecting the cathode foil to the case, requires a longer tab. Surplus tab is folded when inserting the capacitor winding into the case.

Inevitably, an axial capacitor exhibits higher self inductance compared with the same sized radial leaded alternative, as illustrated in **Fig. 6**.

To demonstrate the effects constructional differences of small aluminium electrolytic capacitors have in practice. I used the application test circuit and phase meter,[9] described in 'Fazed by Phase' to make comparison measurements.

I measured four capacitor from 10µF to 100µF. These values are frequently used in audio amplifier coupling circuits. When possible, I selected 50V rated bi-polar radial, polar radial and polar axial aluminium electrolytic capacitor constructions. Since 10µF 35V tantalum bead capacitors were available, they too were measured.

The un-sanitised measured plots, **Figs 7-10**, show departure from ideal or theoretical behaviour starting at well below 1kHz. The esr curves are almost constant with frequency, resulting in capacitor phase angles approaching 45° at the critical mid-audio frequencies. The bi-polar capacitor clearly out-performs the other aluminium types.

Regardless of these differences in construction, to ensure a long service life, the electrolyte used must provide the oxygen needed to regrow any damaged oxide film. It must also be chemically inert to both aluminium and its oxides. Due to electro-chemical potential effects, no other metal can be permitted to contact the electrolyte. Being in contact with electrolyte, the capacitor case must be made of aluminium or a non-metal.

Any two pieces of aluminium in contact with electrolyte and subjected to a voltage differential will grow aluminium oxide. So pressure contacts, as used in foil and paper capac-

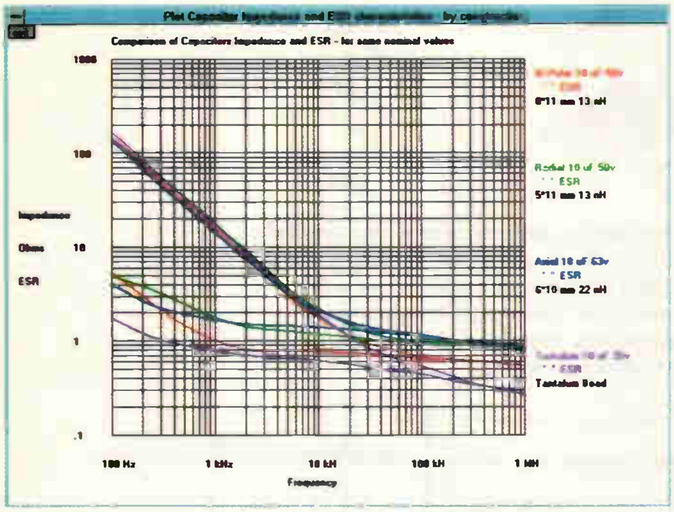

Fig. 7. As measured, plots of impedance and esr for bi-polar radial, polar radial, polar axial and bead tantalum 10µF capacitors. Shows the Panasonic bi-polar performs best of aluminium variants, but all types have near 45° phase by 10kHz and deteriorating performance from as low as 1kHz.

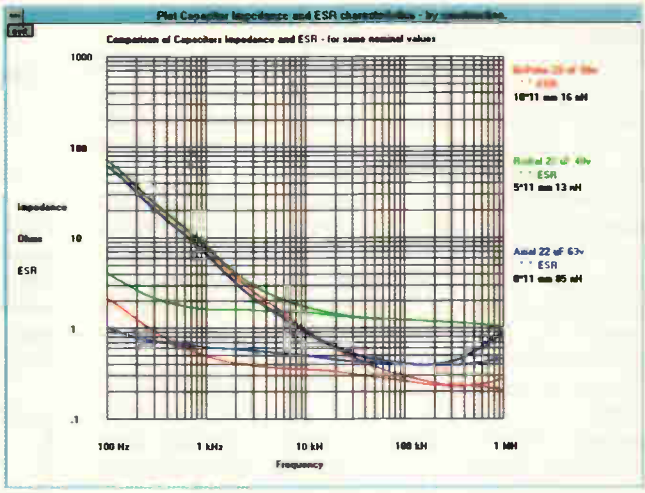

Fig. 8. As for Fig. 7 aluminium types but 22µF values. Where possible, 50V ratings or nearest possible were used for these comparisons. At this value the bi-polar capacitor is the best choice.

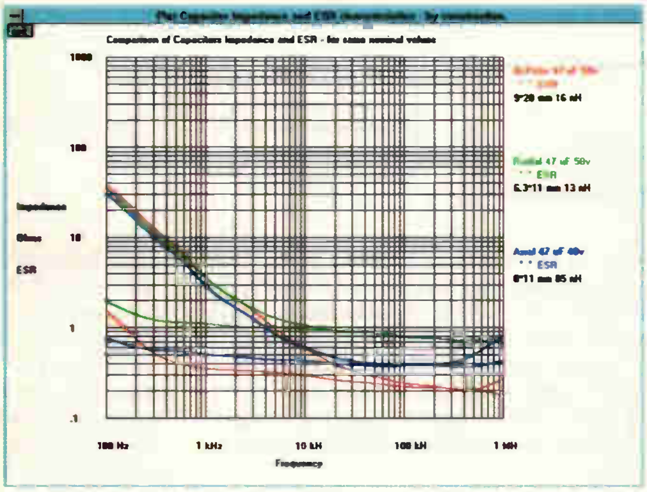

Fig. 9. As for Fig. 8 but this time using 47µF values. Due to the larger size and foil area, the Panasonic bi-polar style easily outperforms the smaller alternatives.

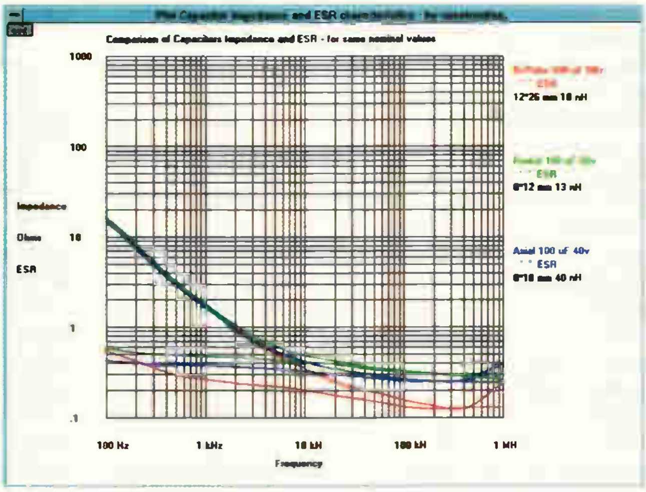

Fig. 10. With 100µF values the Panasonic bi-polar remains a notably better performing capacitor for these tests.

itors, cannot be used to connect the inserted tabs to the electrode foils.

Two common methods to ensure low resistance connections between the electrode foils and connecting tabs are used. Small or low cost capacitors may use a 'stitched' connection. Here, a shaped needle point is burst or punched through the connecting tab and electrode foil. Small 'ears' of tab material are forced out, then turned over and well flattened down. This effectively rivets the two parts together. While simple and crude looking, a good long-life connection results, **Fig. 11**.

A much better technique, called 'cold pressure welding' is used in larger or more professional capacitors. Here, a specially profiled, small press tool is used. It is given enough pressure and time to allow the metals from the electrode and tab foils to interchange, forming a molecular bond. As with a conventional hot weld, the two joined metals cannot be separated.

This excellent joint results from the tool pattern used. By compressing and displacing metal, the pattern removes the oxide layers, allowing direct metal to metal bonds. Cold welds are easily identified by the press tool's outline and inner patterns. No bursting through as for a stitched connection can be seen, **Fig. 11**.

Welding, using either cold pressure or laser, is the preferred method for all aluminium electrolytic capacitor internal connections. However a rivetted connection between connecting tabs and the external solder tags, can be a satisfactory alternative – especially for smaller capacitors.

Aluminium-dielectric systems

The anode electrode foil of an aluminium electrolytic also provides the dielectric oxide which is in intimate contact with the electrolyte – the true second electrode.

Electrical connection to this electrolyte is provided by a second foil, or cathode electrode, which is also covered with dielectric oxide. In other words two foils comprise two distinctly separate capacitances in series, inter connected by the conductive electrolyte.

A polarised aluminium electrolytic possesses two capacitors, the desired one in series with a much larger 'cathode' capacitance. This cathode capacitance has a lower voltage withstand, typically 1.5V and is of opposing polarity to the

Aluminium 'forming'

Placed in a bath of suitable electrolyte and with a positive voltage applied to the foil, aluminium oxide, Al_2O_3, grows on the foil surfaces to a thickness 14 Angstroms for each volt.

As mentioned in the main text, two forms of aluminium oxide can be produced, one porous, the other a non-porous and hard oxide. The hard oxide film is an excellent insulator that can be produced to withstand 600V using super-purity aluminium. Aluminium and electrolyte purity are important, since impurities result in weakness in this insulating oxide film, increasing leakage current.

Pure aluminium can be attacked by pure water at modest elevated temperatures, unless this oxide film has first been hydrated by boiling in water, or is chemically inhibited.

The oxidation process is self regulating. Initially most forming current passes through visible and exposed surfaces. With continued oxide growth, current at these exposed surfaces reduces, slowing local growth and diverting current to less accessible areas. Current flow and oxide growth almost cease when all metal surfaces are insulated.

The aluminium oxide dielectric formed is highly amorphous, except for the outermost surface layer, which may be porous when formed at high voltages. Boiling in very pure water hydrates the oxide sealing this porosity, resulting in very low leakage current capacitors.

Aluminium oxide is visually transparent, but the etched foil surfaces are covered with minute voids and tunnels which absorb light, hence the dark appearance, **Fig. 11**.

In general, foils are etched and formed from wide rolls of aluminium, then slit to the desired widths for capacitor manufacture. Each width has thin exposed raw aluminium edges. These edges, together with any mechanical handling damage, will have the oxide film renewed after capacitor winding, as part of the ageing process.

To avoid consuming excess electrolyte, following impregnation and before final assembly, high voltage capacitors may be 'wet aged' while immersed in baths of electrolyte.

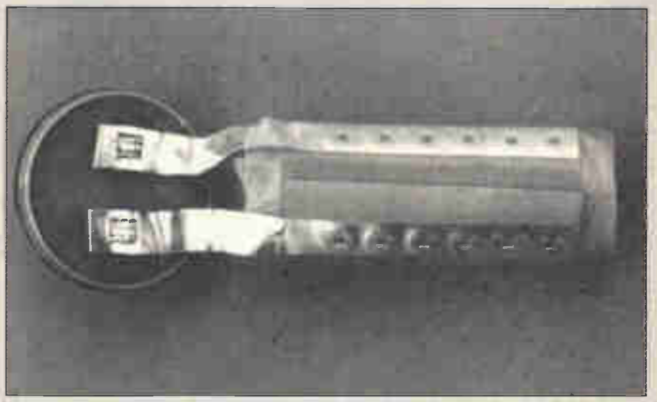

Fig. 11. Photograph of a dismantled and partially unwound capacitor shows a good example of the stitched technique used to connect both anode and cathode foils to their respective tab foils – central tabbed in this example. Also the cold pressure welded joints connecting anode and cathode tabs to the external tag rivets are clearly visible. This excellent joint results from the tool pattern which, by compressing and displacing metal, removes the oxide layers allowing direct metal to metal bonds.

anode capacitor, **Fig. 2**.

A non-polarised, or bi-polar capacitor is usually made using two identical anode foils. This provides two equal value and equal voltage capacitors in series, each double the required capacitance and again of opposing polarity.

Being a semi-conductor, aluminium oxide withstands its formed voltage in one direction only, passing a small 'leakage' current. As for a semi-conductor diode, with reversed polarity it conducts at a low voltage, typically 0.4 to 0.5V while passing a high current.

Regardless of level, this current disassociates some water in the electrolyte, depositing oxygen at the positive electrode, hydrogen at the negative electrode. The oxygen is consumed to generate aluminium oxide.

Many electrolytes contain a hydrogen absorber, to absorb the hydrogen released by normal leakage currents. Even so, with reversed polarity, much of the hydrogen must be allowed to escape.

With the above information, it is now possible to develop an equivalent circuit model of an aluminium electrolytic capacitor, **Fig. 12**.

Reverse bias effects

In a polarised capacitor the 'cathode' foil capacitance, being in series with that of the anode foil, reduces the net or measured capacitance. It also contributes two advantages not present with the tantalum slug styles.

The withstand voltage of the cathode foil's dielectric oxide allows the capacitor to accept small reverse voltages for a significant time, both in service and in approval testing.

This cathode foil's *C/V* product, if equal or greater than that of the anode foil, enables charge displacement from anode foil to cathode foil. The cathode foil in a capacitor subjected to normal charge and discharge currents, does not then become reverse biased with respect to the electrolyte. Such capacitors are described as being charge-discharge proof, **Fig. 12**.

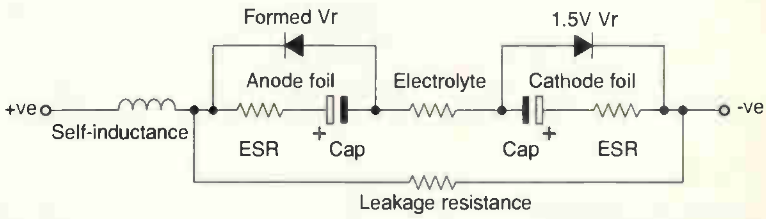

Fig. 12. Equivalent circuit of a polarised aluminium electrolytic capacitor. Since leakage current can be significant, it is shown as a shunt resistance rather than converting to its series equivalent, as more usual for other capacitor types. Simple Spice analysis shows how easily the electrolyte of a polarised capacitor, used to couple irregular or pulse waveforms, can become internally reverse biased to the cathode foil, resulting in early failure.

Many quality aluminium electrolytic capacitors are subjected to a million charge-discharge cycles with rise and fall times of 100ms as part of their approvals testing. For IEC 384-4 charge-discharge approval, capacitance change in this test must be less than 10%.

Note, though, that capacitors that are repeatedly 'crash' charged and discharged, as in photoflash or strobe equipment, require special construction. Commercial capacitors used for these duties – regardless of their voltage rating – have quickly and very dramatically failed.

The current that flows when polarity is reversed can generate sufficient gas to drive electrolyte out of the winding. Pressure builds up in the capacitor case, resulting ultimately in a capacitor failure.

Direct-current polarity reversal at the capacitor terminals is of course easily measured and avoided.

A less obvious problem occurs when the capacitor is used to couple irregular or pulse waveforms. A repetitive charge

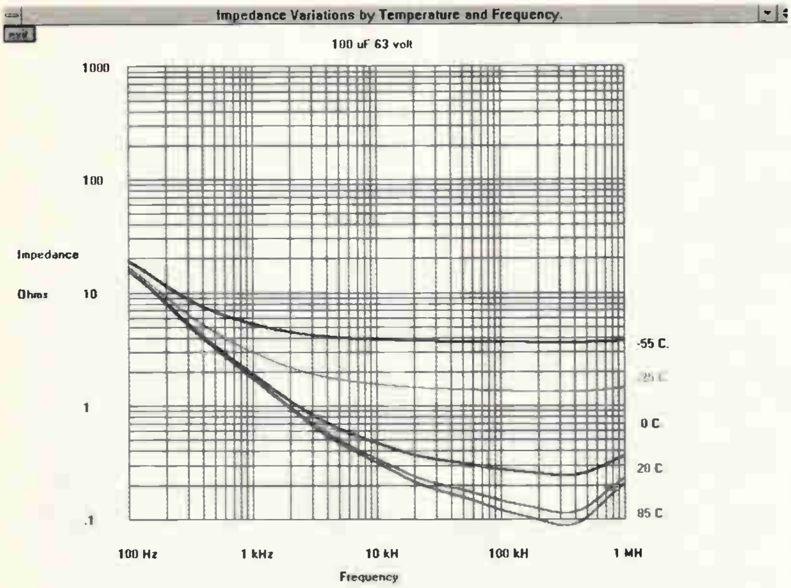

Fig. 13. Plot showing how impedance versus frequency characteristics change with ambient temperature, for a typical 100μF 63V capacitor.

transfer from anode to cathode of a polar capacitor can result in the electrolyte becoming reverse biased relative to the cathode. As a result, a non-polarised or bi-polar capacitor should be used here.

Should this reverse bias exceed permitted levels, oxide growth at the cathode reduces its capacitance. In turn, this further increases the reverse bias voltage. Gas pressure developed drives out electrolyte, resulting in capacitor failure.

Bead or slug tantalum

To produce a capacitor, the formed and sintered tantalum slug only needs impregnating with electrolyte and providing with its connecting cathode electrode.

In most cases the 'dry' electrolyte used is manganese dioxide, while the cathode is a graphite or carbon coating. Electrical connection to this coating is provided by layers of silver loaded ink or epoxy.

As the manganese dioxide is a solid, it is not possible to directly impregnate it. It has to be converted from a suitable liquid medium that can be vacuum impregnated into all voids inside the sintered slug – no matter how small.

Resulting from these voids and passageways, the tantalum slug exhibits a level of self inductance. The device becomes self resonant typically between 500kHz and 5MHz,[10] according to bead size and capacitance.

In many cases, the medium used is liquid manganese nitrate. Easily impregnated into the slug and converted into manganese dioxide by pyrolysis, or thermal decomposition, at 400°C. Many repeated impregnations and pyrolysis cycles are needed to ensure sufficient manganese dioxide.

Contact to this manganese dioxide electrolyte generally uses successive layers of colloidal graphite with silver loaded paints or conductive epoxy resins. This is frequently, finished by dipping in molten solder. The external cathode connection is easily attached either by soldering to the solder coating or with conductive epoxy to the silver loaded paint/epoxy coating.[5]

Since the manganese dioxide electrolyte is directly connected to the capacitor's negative lead, only one polarised dielectric is present. Consequently, reverse voltage should not be applied. The Siemens Matsushita data book for example restricts any reverse voltage application to five applications of a minute maximum each hour.

Two dry tantalum beads can be connected back to back to simulate a bi-polar capacitor. To prevent damage to the reversed polarity capacitor, protection diodes must be placed in parallel with each capacitor. These are detailed in the Siemens Matsushita data book.

Bead tantalums made with liquid electrolytes must not be subject to any reverse voltages. Excess gassing resulting from reversed polarity builds up internal pressure, destroying the capacitor. Silver from the cathode electrode can also transfer to the anode, leading to a short circuit.

Surge currents for all bead tantalums, should be restricted by series impedances of typically an ohm minimum per applied volt.[10]

Aluminium electrolytes

The manganese dioxide used in solid aluminium electrolytics has already been discussed, but for wet or non-solid aluminium electrolytic capacitors, many different formulations exist. These usually comprise a neutralised weak acid in a suitable solvent.

The solvent used must not freeze or boil at the extremes of the capacitor's working temperature range, nor must they attack pure aluminium.

One old but still usable electrolyte is a mixture of ammonia and boric acid dissolved in pure ethylene glycol. Most older electrolytes contained small quantities of water. Prepared using liquid ammonia, they were boiled to reduce water content.

This water content combined with lesser purity foils resulted in early electrolytics having a reputation for leakage current deterioration in storage. They needed re-forming before use.

For the past 30 years, this water attack could easily be inhibited by a chemical additive in the electrolyte, in much the same way that steel is passivated to prevent rust, using phosphoric acid. This, and the use of super purity aluminium foils, has effectively eliminated the need to re-form capacitors.

As an experiment, I recently measured the leakage current of three unused prototype 16 and 25V capacitors, made almost 30 years ago and stored since then in a box in my garage. All three easily passed their catalogued leakage claims without reforming.

Modern electrolytes use many different acid formulations and solvents. These range from simple solutions of ammonium borate or ammonium succinate in glycol to organic acids and solvents such as di-methyl-formamide. In principle, many weak acids or their ammonium salts can be used, leading to a wide choice.

A hydrogen absorber, able to cope with hydrogen released by normal leakage current, is provided in many modern electrolytes.

While very conductive electrolytes are used in the lowest voltage capacitors, such electrolytes cannot be used at high voltages. In this context one must view the separating paper tissues used as being part of the electrolyte system. Thick, low-density 'rag' tissue paper can be used at low voltages, while for high voltage work, higher density thin tissue paper – and even multiple tissues – may be needed.

While most aluminium capacitor electrolytes are innocuous, it is as well to check the makers literature before dismantling a capacitor to investigate its construction. In any case, you are well advised to wear suitable rubber or plastic gloves, and use eye protection against electrolyte splashes.

Formic acid for example stings the eyes and any broken skin. It may also trigger an allergic reaction in hay fever sufferers, as I can well testify from past experience.

Finishing processes

Following final mechanical assembly, both capacitor types are subjected to temperature and voltage ageing or burn in. This allows minor dielectric fault areas to be removed, and leakage current reduction to catalogue specifications.

In aluminium electrolytic capacitors leakage current is reduced by deliberately growing new oxide, replacing any mechanically damaged during the assembly processes. This regrowth consumes some of the oxygen available from the liquid electrolyte.

A level of regrowth continues throughout the capacitor's service life, stabilising and reducing leakage current. Ultimately at the end of service, the capacitor fails usually as a high impedance or open circuit when the available oxygen in the electrolyte has been consumed.

Tantalum and aluminium solid electrolyte capacitors behave quite differently. Fault repair and leakage current reduction does not occur by oxide regrowth, but rather in a similar fashion to metallised film capacitor self healing, by isolating the faulty area.[5]

The increased current due to a minor fault locally heats the manganese dioxide, which spontaneously degenerates to lower oxides. These exhibit much increased resistivity compared to manganese dioxide, effectively isolating this faulty area.

Should a major fault occur in a bead tantalum capacitor used in a low source impedance circuit, excess heat is generated. This excess may be sufficient to locally crystallise the surrounding area of tantalum pentoxide. Energy available from the circuit may then be sufficient to promote an avalanche failure condition,[5] the capacitor failing short circuit. Given sufficient externally supplied power, the capacitor can burn.

In the early days of bead tantalums in order to prevent this avalanche failure mode, a common recommendation was that they should only be used on power rails via a current limiting source impedance of at least 3Ω/V.

Improvements in materials and manufacture of bead tantalums have much reduced this problem – but they have not eliminated it. While not well publicised, many makers offer low-impedance burnt in and even fuse protected tantalum capacitors. However the advice in MIL-STD-198 is best heeded.[10] It says you should ensure a 1Ω/V source impedance minimum together with substantial capacitor voltage de-rating for high reliability.

One hybrid capacitor combining the best aspects of the two aluminium-foil construction with the long shelf and service life characteristics of the solid tantalum is the solid aluminium wound foil capacitor.[11] These characteristics provide reliable service in automotive applications, surviving vibration, temperature extremes and voltage surges. Unfortunately this excellent capacitor style is not usually stocked by mainstream distributors.

Temperature effects

Some capacitance variation with temperature is exhibited both by aluminium oxide and tantalum pentoxide. Their leakage currents will follow the Arrenhuis law, roughly doubling or halving for each 10°C change of temperature.

By far the most notable effect of temperature is the way the

Aluminium etching

The surface gain of modern aluminium electrolytic capacitor foils is attained by electro-chemical etching in a bath of electrolyte. The aluminium to be etched is placed in a highly conductive bath, usually brine and a high current is passed.

In practice, large rolls of aluminium foil are fed at a controlled speed through a very long etching bath. The foil passes over contacting rollers which supply it with the very high etching current needed. While pure dc can be used, dc with a superimposed alternating component is more common.

Depending on the foil's intended formation voltage or end use, different feed speeds, temperature, current density etc., will be used. Low voltage capacitor anodes for example use only thin oxide growth but need the maximum possible capacitance. The foils are etched to maximise surface gain, creating the smallest possible voids and tunnels, while using the thickest foil base, **Fig. 14**.

High surface-gain foils usually have much of their original aluminium removed by this process. Tunnels can be formed connecting both foil faces. While remaining visually opaque, as with a filter paper, they become highly porous to liquids.

If 'formed' to a high voltage these tiny tunnels would become choked with oxide, rendering the foil useless. High-voltage anode foil is etched deliberately to provide much larger tunnels which will not block with oxide, usually on a thinner foil base, **Fig. 15**.

Various cathode foil etchings and foil thicknesses are needed for differently rated capacitors. The lowest voltage capacitor cathode needs the highest possible surface gain etching. Cathode foil for higher voltage capacitors is much thinner, and modestly etched.

In each case, the object is to provide a cathode *C/V* equal to or larger than that of its anode while minimising size and cost.

Regardless of thickness and etching methods, anode foils use only the highest possible purity aluminium foil – better than 99.99% purity being essential.

By comparison, the cathode foil is not deliberately formed; indeed accidental formation in service is undesirable. Consequently, cathode foils can be made with lower purity material, of typically 99.3% aluminium.

Following etching, the aluminium foil must be most carefully washed using extremely pure water to remove all traces of chloride residues. Minute traces of chloride increase capacitor leakage current and reduces working life.

All foil batches are checked to ensure much less than 1 milligram of chloride contaminant remains, per square metre of foil.

Fig. 14. Large magnification electron microscope view of the surface of a very high 'gain', low-voltage anode foil. Etching and forming methods are optimised for low voltage capacitor use.

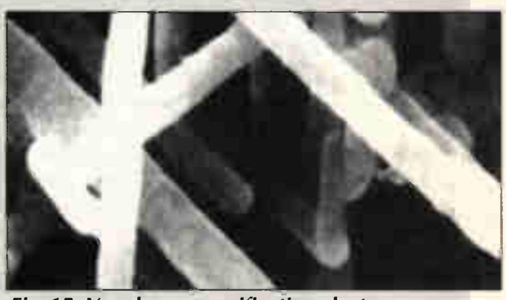

Fig. 15. Very large magnification electron microscope sectional view of low surface gain, very high voltage etched foil. Shows extensive large diameter tunnels able to accept thicker oxide growth needed for high forming voltages. Only the oxide tunnels are seen, the metal having been dissolved away to permit this sectional view.

impedance of wet aluminium electrolytic capacitors changes versus frequency with change of temperature.

The wet electrolyte and separating papers used in aluminium electrolytic capacitors, combined with the minute and tortuous paths into the inner recesses of the anode foils, contributes much of the capacitors series resistance. At 85°C, resistivity of the electrolyte-paper combination is typically half its room temperature value.

At lower temperatures – and especially with higher voltage capacitors – electrolyte viscosity increases more rapidly. By 0°C, resistivity of the electrolyte/paper combination is typically double its room temperature value.

At the lowest temperatures the normal impedance frequency curve becomes a nearly constant resistance plot. This low temperature behaviour fortunately is of no importance for most decoupling applications, which only require the capacitor to be a low impedance path. **Fig. 13**.

The conductivity of solid electrolytes is also temperature dependent. But at very low temperatures, increase of impedance is modest, compared to that of the wet aluminium electrolytic capacitor.

Ripple ratings

These temperature and frequency effects on capacitor esr reflect directly into the capacitors ripple ratings above room temperature. At lower temperatures, any ripple current generated heat serves to warm up the capacitor.

Every capacitor has an internal hot-spot maximum temperature that should not be exceeded. Frequently this cannot be measured, except by manufacturing special test capacitors. A more practical method with aluminium electrolytics is to measure the surface temperature at the aluminium case end, underneath any plastic insulation, using a 0.2mm wire, naked-bead thermocouple with PTFE insulation.

Especially with wet-aluminium electrolytic capacitors, hot-spot temperature directly influences the rate at which electrolyte evaporates through the end seals and thus capacitor life. End-seal materials are chosen to minimise electrolyte losses while allowing excess hydrogen to diffuse out, avoiding undesirable pressure build up.

Makers' ripple current ratings are usually determined in practice by using 100Hz sinusoidal waveforms. Consequently, other frequencies and non-sinusoidal voltage and current waveform requirements must be related to these 100Hz catalogue ratings, using appropriate methods.

Frequently this requires the current for each harmonic component be determined using Fourier transforms.[8] Each harmonic's power is determined from the capacitor's actual esr at that frequency and ambient temperature.[12] The total capacitor power is calculated as the rms sum of all these powers.

Designers interested in a step by step description of this process for calculating capacitor power dissipation for any repetitive waveform will find details in my April 1995 capacitor article.[13]

Alternatively, having calculated the rms current for each frequency, some makers specify suitable frequency correction multiplying factors.[14] The rms sum of these corrected currents is then related to the 100Hz catalogue claims.

In many instances, the waveform may not be sufficiently repetitive, so these methods may not be possible. In that case, a conservative and long-standing rule of thumb is to assume any case temperature rise of less than 5°C should be acceptable. This is subject to the *proviso* that the case operating temperature at the maximum expected equipment ambient does not exceed the maximum service temperature of the capacitor.

Readers interested in this simplistic approach are referred to the IEC 384-4 or CECC 30 300 specifications. These provide tables of permissible case temperature rises for differing ambient temperatures.[12]

Ripple-voltage peaks superimposed on any applied dc level must not exceed the capacitor's rated voltage. The resulting current must not exceed the permitted ripple current at that frequency and with polar capacitors, no polarity reversal is allowed.[8]

While capacitors generally have a small permitted surge voltage level, this is intended to cover equipment switch on conditions. It should not be used when calculating permitted ripple voltages.

Should it become necessary to bank capacitors in series or parallel to provide the needed ratings, voltage and current sharing arrangements are essential to avoid early and dramatic failures.[10]

Whenever possible capacitors should be used de-rated from the permitted catalogue levels. This extendsß service life. All capacitors – and especially electrolytics – provide a much prolonged service life when operated at a reduced operating voltage, and exhibit no adverse side effects.[2]

Washing

Aluminium electrolytic capacitor end seals need to be taken into account when choosing a flux removal fluid so check the capacitor maker's literature. In principle, any seal designed to allow hydrogen to escape might also permit ingress of chlorinated hydrocarbon solvents. Once inside the case, these could supply free chlorides, causing internal capacitor corrosions and leading to failure.

Chlorinated hydrocarbon solvents, used to clean light metals like copper and aluminium quickly become acidified unless frequently replenished. Acidified solvents cause considerable damage to many components, so must be avoided.

Capacitor mounting

It is common practice to conveniently mount large aluminium electrolytic capacitors, terminals down. In many constructions the hydrogen venting seal will then be mounted as the lowest point of the capacitor.

If the capacitor is used well within its voltage and ripple current ratings, having the vent at the bottom should not present a problem. If the device is used incorrectly though, excess gassing can force electrolyte to exude – probably forcibly – from this venting seal. This usually results in corrosion damage to surrounding areas.

Capacitors designed to mount terminals down usually have a vent at the opposite end of the case from the terminals. For satisfactory service life, again the capacitor maker's mounting instructions, voltage and ripple ratings should be observed. ■

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Many apologies. These diagrams did not reproduce properly in Cyril's last article. I sincerely hope that they appear in full this time round, but WYSIWYG doesn't seem to have filtered through to the printing world yet. Ed.

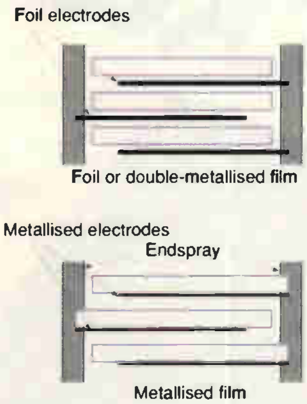

Fig. 1. Having no discrete electrodes, the metallised film capacitor maximises capacitance and voltage for a given size, but has limited current handling. Use of foil electrodes maximises current handling but reduces capacitance. Double metallised film electrodes, described in a German Patent, offer intermediate capacitance and current handling.

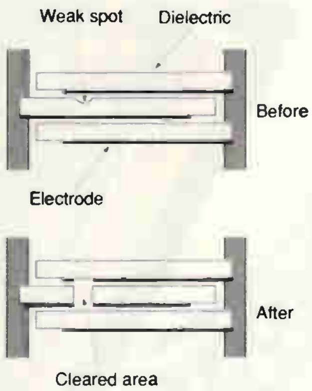

Fig. 2. Metallised film capacitor cross section, illustrating self-healing action. Before and after views show how electrodes and dielectric film are both involved when 'clearing' a fault.

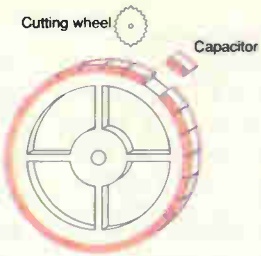

Fig. 3. Stacked film production technology. Large rings of metallised film, of width equal to the final capacitor body length, are wound on 'core' wheels. These rings are metal end sprayed, then sliced to make individual capacitors.

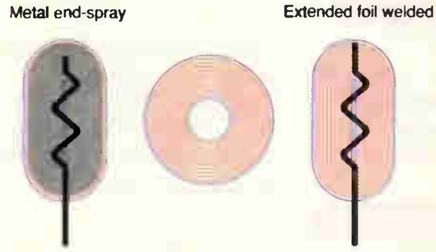

Fig. 4. Winding on a larger diameter mandrel allows the roll to be flattened to form the capacitor element. Both metallised and extended-foil electrode capacitors can take advantage of sprayed-metal end connections. Foil electrodes that extend beyond the dielectric roll allow resistance welded, directly attached lead wires. In the centre is the wound capacitor before compression.

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CIRCLE NO. 130 ON REPLY CARD

Find out the facts about the new Windows versions of *Ranger 2*, and Ultimate's *Challenger* dos/Windows package in Rod Cooper's latest set of reviews.

The route to pcb cad

Challenger is a schematic capture and layout program with two autorouters. Versions with both larger and smaller restrictions than those shown in the availability information panel are available at various prices.

At present, the schematic drawing and capture section, called *ULTIcap*, is only available for dos, but the pcb layout section, *ULTIboard*, is available in both dos and Windows. The external rip-up-and-retry autorouter provided for the Challenger, *ULTIroute GXR*, is designed for Windows. A dos shell, called naturally enough *ULTIshell*, connects the various parts. For a product having sub-programs in dos and Windows, a shell is a good method of linking them.

Figure 1 clearly shows that the shell is the central control not just for starting projects, but also for configuring how the schematic capture and layout programs work, and transferring data between them.

It is quite feasible to do things this way. For example, *CircuitMaker* and *Traxmaker* used to be a successful mixed dos/Windows program until recently – even without a shell interface. Note that Ultimate Technology is developing an all-Windows version of this program.

Both *ULTIcap* and *ULTIshell* run successfully under Windows 95, which is the recommended platform. But the default screen driver is 800 by 600. If you have a non-multisync 1024 by 768 or larger display you may run into difficulties. I found that neither *ULTIcap* nor the shell would run full-screen or in a window with Windows 95 on my 21in monitor. All I got was a blank screen. Transferring to a multisync monitor solved this problem. Nor would these programs run at the higher resolutions in plain dos. The video drivers were incompatible with my graphics card.

I eventually ran *ULTIcap*, *ULTIshell* and *ULTIboard* in dos with a 15in monitor at 680 by 480. This solved a lot of the screen capture problems for the review, and because staying in dos made the system much easier to operate overall.

A dos autorouter is provided as standard with *ULTIboard*, so you could stay in dos for the whole design process. In this mode the program occupies 9Mbyte of disk space.

The dos versions are not limited by the 640Kbyte memory restriction. In fact, the amount of memory can be configured for both sections to whatever your hardware can provide. The amount of memory used can be increased from the 1Mbyte default for *ULTIcap* and 2Mbyte for *ULTIboard* by using the configuration set-up in *ULTIshell*. This is recommended as a first step in operating both programs, otherwise larger circuits will not be accommodated. This is particularly important in

Challenger

Maker Ultimate Technology, The Netherlands.
UK supplier; Ultimate Technology, Unit 1, Lodges Barn, Coxbury Lane, St. Briavels, Gloucester GL15 6QJ. Tel. 01594 810100, fax 01594 810200.
Prices are £395 for 700 pin limit, £775 for 1400 pin limit.

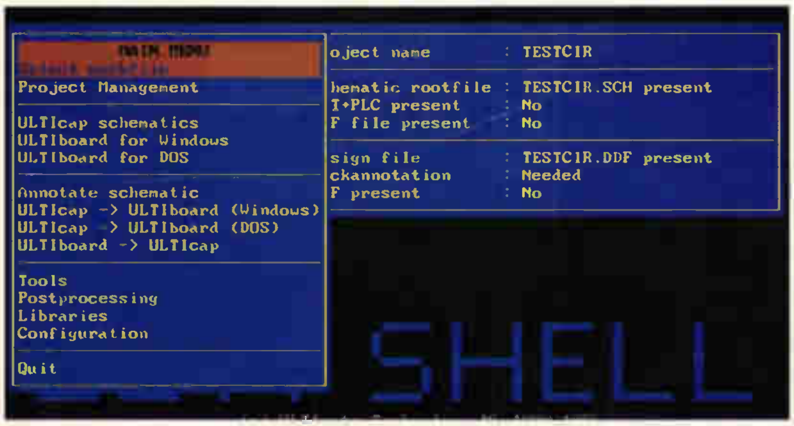

Fig. 1. The Ultimate shell which ties in all the modules, including the parts in Windows as well as dos.

ULTIboard if you are using the autorouter.

There is no recommended hardware set-up, but both *ULTIcap* and *ULTIboard* are 32-bit, so this would indicate a 386DX minimum. Although I used a Pentium for the tests, I tried the program on a 386DX and it worked well.

The extra *GXR* autorouter in Windows is an external add-on. Net-list files have to be transferred via a translator. *GXR* is again a 32-bit program.

Comprehensive literature is provided with the package in the form of four manuals, covering schematic capture, board design, a tutorial and a library reference for pcb footprints. The books are not a native English translation but, given a few errors in grammar and spelling, are understandable.

Some versions of Ultimate software are protected by registration number, but some higher versions have a dongle. For more on dongles, see my first review in the October 1996 issue.

Schematic drawing

ULTIcap is a typical menu-driven dos program. There is no GUI to temper the dos aspect as there is, for example, in *Ranger 2* for dos. One positive result of this is that it gives a good screen drawing area. On a typical 14in monitor you can expect 9.5in by 6.5in of drawing space.

Figure 2 shows a typical screen during the compilation of the same test circuit used in the earlier review. Note a large cross-hair pointer replaces the more usual mouse arrow.

The package revolves around a central database. Whereas *EDWin* calls this plainly "the database," *Ranger 2* calls it "the job", and in *Ultimate* it is referred to as "the project". It is important to realise this before starting off on a design.

The program has autosave, autopan, a 15-level zoom system, and a good electrical rules checker. All these work with typical dos speed and efficiency. The mouse system is simple and logical – i.e left-hand button starts an action, right-hand button escapes or terminates. The system can be set for metric or imperial units.

Initially, symbols are selected from the "external" library in text form. Library accessibility is only fair. For things like resistors and capacitors, there is no immediate way of knowing what the associated footprint is. This is found by cross-checking with the library editor.

The selection of symbols is good. There is also a good selection of pcb footprints to match. Although there is no parts bin, a "local" library system like that in *EDWin* performs a similar duty.

Moving symbols around was easy, but rotating them less so. The method is to crank them round a fixed point using F2 on the keyboard. Again, this is very similar to *EDWin*. Text rotates with the symbol and rotating the text back to readability is cumbersome and takes time. This is an area where things could be improved. Text is fully manoeuvrable so the circuit can be tidied up afterwards to make the circuit look good.

Wiring symbols up is easy, and to make it even easier there is an straightforward auto-wirer. Positive confirmation of connectivity is given when you make connections with component pins. However, I still managed somehow to produce a few "bad" nets using this system. It is possible to draw lines in space and these mistakes have to be deleted manually. Pressing the screen redraw does not clear them up. Moreover, they have to be deleted in segments, which is slow. There is a good connectivity checker which you can

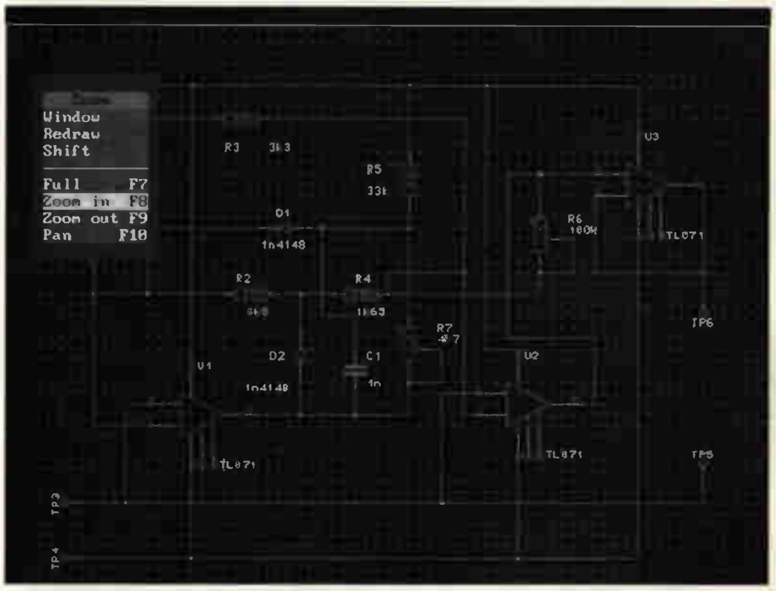

Fig. 2. Schematic of test circuit showing an ULTIcap menu. Note good available screen area for drawing.

Fig. 3. Typical linear rat's nest produced from schematic capture.

Previous review subjects

PCB Designer: Niche Software Ltd, tel 01432 355414 – reviewed September 1996.

PIA:AW Software Ltd, Germany tel +49 89 6915352 – reviewed September 1996.

Easytrax: Protel International pty, Australia. Available from PDSL, tel 01892 663298 – reviewed September 1996.

Ranger 2: Seetrax CAE Ltd, tel. 01705 591037 – reviewed October 1996.

Electronics Workbench: Interactive Image Technologies Ltd Canada, tel. 00141 69 775 550 – reviewed October 1996.

CircuitMaker: MicroCode Engineering USA, UK agent Labvolt, tel 01480 300695 – reviewed November 1996.

Quickroute 3.5 Pro+: Quickroute Systems Ltd, tel 0161 449 7101 – reviewed December 1996.

Propak: Labcenter Electronics, tel 01756 753440 – reviewed December 1996.

Proteus: Labcenter Electronics, Schematic capture and pcb design – reviewed January 1997.

EasyPc Pro XM: Number One Systems, tel 01480 461778 – reviewed January 1997.

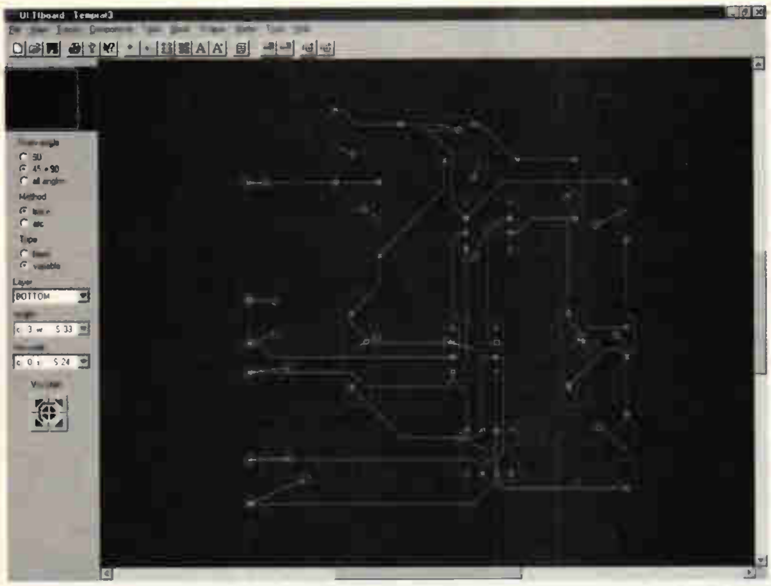

Fig. 4. Ultimate Rat's nest of test circuit after manual sorting and transfer to Windows. Note the force vectors which can aid component placement.

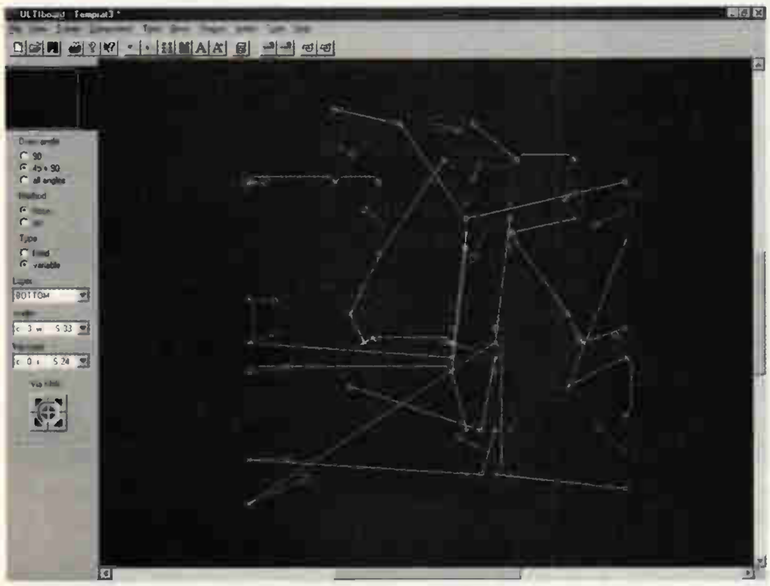

Fig. 5. The above rat's nest of Fig. 4 after routing with the GXR autorouter.

use before capture.

Autopanning makes it quite easy to get lost during the learning phase. Although there is no map showing where the circuit is in relation to the drawing area, you can recover the work by selecting "zoom, all." Multi-sheet schematics are possible, and the hierarchy system already described for Labcenter's products in *Electronics World* December 1996 issue is the preferred method.

There is auto-annotation of symbols, but this is not done in the schematic section but from *ULTIshell* – a strange system not found in other programs. Returning to *ULTIcap* shows the results.

Pcb layout

It is possible to draw tracks manually in *ULTIboard*, but the emphasis is on automated pcb design. The normal route is to draw the schematic, capture it and transfer the data to *ULTIboard*, where it appears as a rat's nest. The transfer is done via *ULTIshell*.

The net list from *ULTIcap* appears automatically as a linear rat's nest with the component outlines placed along the top of the board outline, **Fig. 3**. This method is similar to that used in *QuickRoute* and *EasyPC*, but with the advantage that the components are not so spread out, so subsequent arrangement into a logical order is relatively easy.

The rat's nest is re-organised by picking up each component in turn with the mouse and placing it inside the board outline. Rat lines disengage during this process so it is necessary to press "redraw" if you want to see them again. This slows down what would otherwise be a rapid process. Component rotation is done by the same keyboard method as used in *ULTIcap*, i.e. using F2. Other controls like zoom, pan etc are all in the same style as *ULTIcap*.

Figure 4 shows the design after component placement, ready for autorouteing. Note the force vectors, similar to those in *Proteus*, which are used to aid component placement. Any manual routeing of critical tracks is done by a modified rubber-banding method which I found less difficult than others I have tried.

There is design-rule checker which operates in real time, which is very useful if you are manually routing. It is more usual to have the design-rule checker as a post-routing verification. However, setting up some aspects of the design rules is more convoluted than in other programs.

Selecting track width and clearances is different from other other programs. In *ULTIBoard* you have a choice of 32 preset track widths combined with a clearance. These are configurable to suit your own preference.

The standard dos autorouter is a gridless type – although a nominal grid is actually selected. It has no rip-up-and-retry and cannot autoneck, limiting its routeing power. It is on about the same level as the standard dis autorouter in *Ranger 2 and*, like that autorouter, is fast but often gives lass than 100% completion.

Rat's nest placement takes considerable effort if you want to get high completion rates. This autorouter worked best on double-sided and multilayer boards. Although it can attempt single-sided boards, the results are predictably, nearly always disappointing.

On the test circuit, **Fig. 5**, this autorouter fell into category C – a category for autorouters incapable of being able to route the test circuit completely, as defined in my first set of reviews. Category A routers are able to route the circuit while those in category B can route the circuit given some relaxation of the design rules. There is modest amount of pre-run configuration that can be done on this autorouter.

The second autorouter, *ULTIroute GXR*, is a gridded autorouter with rip-up-and-retry capability. This gives better completion rates, frequently achieving 100% on double-sided boards. To use it you will of course have to quit dos and go into Windows. Transferring a net list to it via the translator mentioned earlier was easy.

The *GXR* has about the same configurability as the standard autorouter, and it can be set up to do a single-sided board. It managed to complete the test board putting it into category A, but considerable manipulation was needed in setting up in order to do this. A costing method is used to set up strategies whereby cost penalties are attached to various autorouter actions such as routeing a track away from a direct line. This is similar to Ranger 2's rip-up-and-retry autorouter, but much simpler.

As well as the usual output for generating Gerber files, *Ultiboard* has its own plotter drivers for HP, Houston and Calcomp. Setting up the printer/plotter was somewhat involved. An intermediate plot/print file is generated in *ULTIboard*, then this is closed and *ULTIshell* restarted.

Hard copy generation is handled by a post-processing subprogram from the shell. There is plenty of choice of both printers and plotter types and the set-up is comprehensive – albeit a little difficult to access – and includes a function for obtaining a mirror image.

The results produced by laser printers and pen plotters was good. Control over plotter pen speed and width was also good.

In summary

The dos versions of *ULTIcap* and *ULTIboard* are logical in operation, but slow in some aspects. Functions such as deleting lines and re-rotating text for example are awkward in comparison to competing programs. The libraries are not very user-friendly. The standard autorouter would be acceptable for non-critical double-sided boards.

The more capable *GXR* is the autorouter most users would probably select. The overall learning curve is steep – almost on par with that of *EDWin*. On the positive side, this product is a flexible, fully-featured program, offering a better screen area than any of the Windows programs.

I hope to review the all-Windows version of this program as soon as it becomes available.

Ranger 2

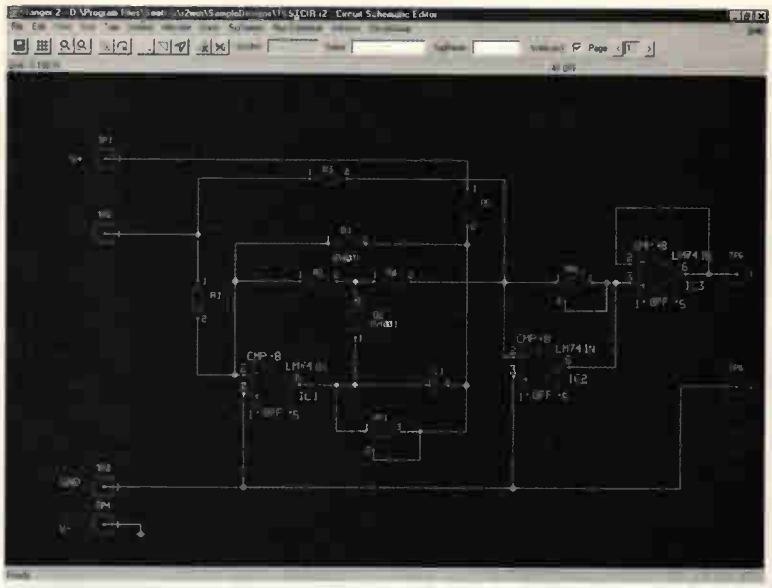

Fig. 1. The completely different schematic drawing controls of Ranger2 for Windows – but note the familiar symbol style.

Ranger 2 for Windows

Maker and supplier; Seetrax CAE Ltd, Old Buriton Limeworks, Kiln Lane, Buriton, Petersfield, Hants GU31 5SJ. Price for the standard version is £250, or £500 with Cooper and Chyan Specctra SP2. Tel. 01730 260062, fax. 01730 267273.

In my first set of reviews, *Ranger 2* for dos was recommended for its good graphical user interface, overall competence, good pricing, and for offering a choice of autorouters. The subject of this review, *Ranger 2* for Windows, has retained much of the character of the previous version, but the appearance has changed out of all recognition. Gone are the full-screen menus, the buttons labelled addpoint, libvol etc.

This approach to conversion to Windows contrasts with Labcenter's *Proteus*, which retained its dos format in Windows to the extent that dos users could convert immediately. Nevertheless, previous users of *Ranger 2* will still recognise the characteristic style of Seetrax products so conversion to the Windows version should not be a problem.

Ranger 2 contrasts sharply with programs such as *EDWin* in its use of icons. While *EDWin* uses a lot of icons, *Ranger 2* uses relatively few. The main access to functions is via Windows 95-style menus, supported by a handful of buttons. The screen of **Fig. 1** shows a typical *Ranger 2* screen and menu. I found the bias towards menus worked very well and was a refreshing change from over-iconed programs.

This new version is intended for use with Windows 95 or NT. It comes on two 3.5in disks and you will need about 5Mbyte to install it.

The operator's manual is in the same style as the original, but much shorter and with a slightly better subject order. Of course, in a Windows program there is often a shorter manual because much more assistance is available in the Help menu. But in the version I tested the Help menu was empty. I am told that the Help text is being prepared and will be available eventually.

The new *Ranger 2* is still formulated around Seetrax's

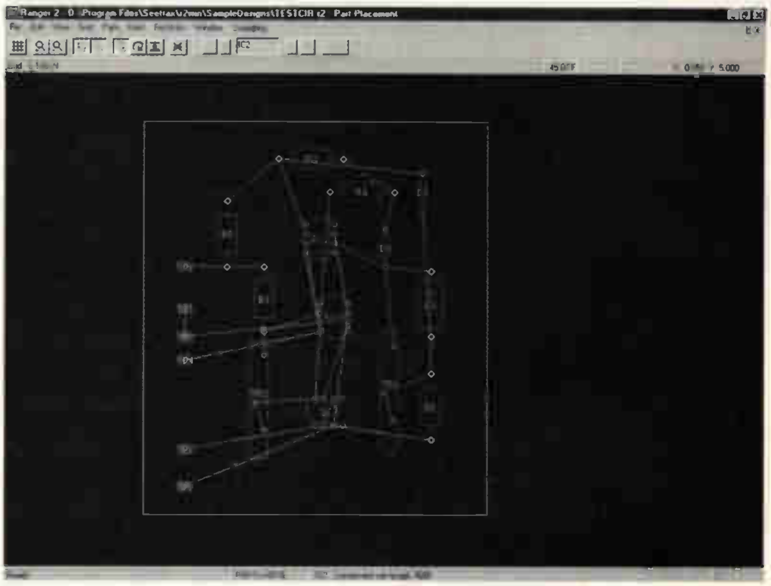

Fig. 2. Rat's nest produced from the test circuit.

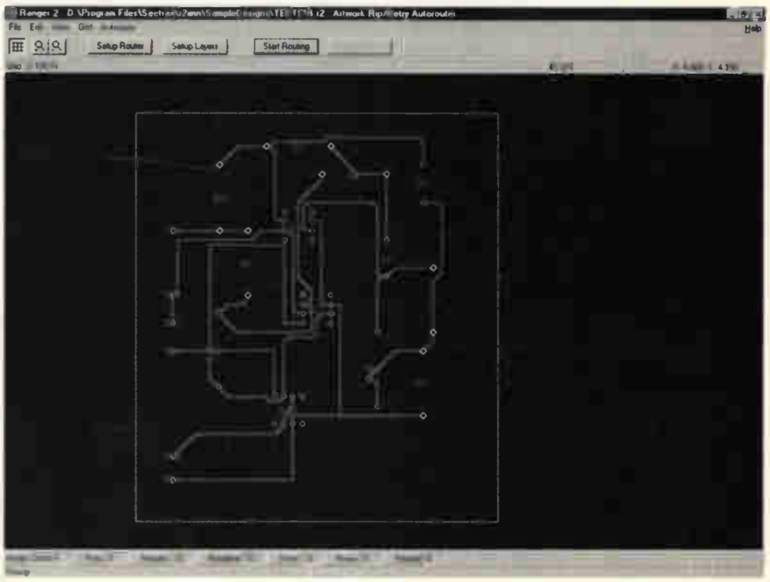

Fig. 3. Results of the Ranger2 rip-up-and-retry autorouter. Compare this with the Specctra results in Figs 5 and 6.

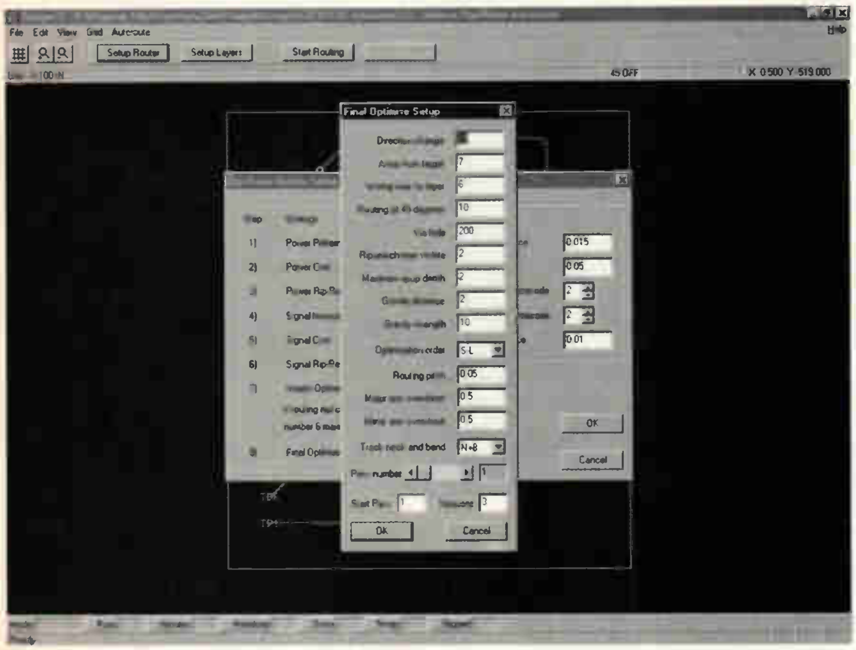

Fig. 4. Typical set of menus for configuring the rip-up-and-retry autorouter. Setting up for optimum results takes some time.

well-grounded concept of "the job" so previous dos users will quickly pick up the threads of operation from this.

Schematic drawing

Figure 1 shows that although the screen layout has changed completely, the symbols and drawing style have not. Note the lack of scroll-bars. Panning is done by selecting this function on the edit menu and pointing with the mouse. Alternatively, you can use the space bar combined with the mouse pointer. Both methods work well. Multi-sheet schematics are supported.

There is no map showing where your circuit is on the page so it quite easy to get lost, but you can recover the circuit with the "view, full" from the view menu. There is no autosave facility yet: this is said to be under development.

Accessing the libraries is done by menus and is straightforward. Symbols are in text only, but if you want to see exactly what you are getting in graphical form before you use them, it is now possible to open library volumes and check. You can also check the associated pcb footprint with this method.

Library volumes are numbered, so if you want to know what they contain without going to the trouble of opening them, you have to refer to the manual. This could be improved. For example, the resistor volume could be labelled "vol 40 – resistors" instead of the cryptic "defndl.v40." The size and content of the libraries appear to be about the same as the former version, so are adequate for general purpose use.

Placing symbols is done via a parts bin – or tray in Seetrax language – and is smooth and well controlled. The tray can be toggled off to increase drawing area, which is a good idea. In fact, with no scroll bars, no map and the on/off parts tray, the screen's drawing area is good for a Windows program, at approximately 9.5 by 5in on the 14in screen.

Symbol text stays upright when symbols are rotated but is non-manoeuvrable. Other text, such as power and signal references, can be moved.

Wiring symbols up is easy due to a strong 'snap-to' function combined with a good orthogonal system. Drawing lines in space is inhibited, but it is possible to draw 'illegal' lines i.e. lines not connecting two component pins. This was prevented automatically in the dos version of *Ranger 2* without the need to go back and delete, although it was not always perfect in operation. The lack of this feature is not welcome in the Windows version.

Capture of the schematic is very much as before, with a parts list and wiring list being created. The lists are in plain text – i.e. easily readable. You can export a PSpice list if you want to connect to a simulator, and import a Futurenet list.

Laying out a pcb

Having generated parts and wiring lists, the next step in *Ranger 2* is to draw a pcb board outline and this is done with a simple sub-program in the artwork section. Both metric and imperial units are supported. Parts are interactively placed inside this area, and the method is to place the components by hand as they are automatically fed in one-by-one.

You will see from **Fig. 2** that the current component being placed is listed in a box next to the icons. This remains a superior system to having component footprints dumped in a line or in a heap. However, *Ranger 2* is oriented towards being an automated system so Seetrax should be considering adoption of an autoplacement system in the near future, to keep pace with the competition like *Proteus, EDWin, TraxMaker*. Force vectors should be considered as well.

A useful feature of *Ranger 2* is that the power nets are shown in a different colour from signal nets – see Fig. 2. You might choose to route the power nets manually, as these often take routing precedence over signal routes, and having them shown separately makes this task easier. You could of course autoroute them on their own as well, particularly if there are a lot of them. Having them identified separately also helps to decide which method to use.

Manual routeing of uncompleted nets is by the rubber-banding method. This is still not easy in the Windows version, and is an acquired skill. Although possible, manual track placement is not given much priority in the *Ranger 2* scheme. If you want to construct a board entirely using manual methods, I would recommend using a simpler program, such as PIA.

The rat's nest automatically optimises as the components are moved round. When a satisfactory placement has been achieved, you have to digitise before going on to the autorouters. If you want to manually route any tracks left uncompleted by the autorouter after it has finished, it is necessary to de-digitise them with the 'Mroute' function. These extra steps of digitising and de-digitising remain over from the dos version and are not found in other programs.

Autorouter options

Ranger 2 for Windows has three autorouter options. The standard autorouter is a window-ised version of the one provided in the dos version of *Ranger 2* and will route double-sided boards. It will attempt single-sided boards but the test results put it into category C. The advantage of the standard autorouter is that it is easy to set up and quick to run.

The second autorouter is a Windows version of Seetrax's former 386 rip-up-and-retry product. This is now referred to as the 'Rip/retry' autorouter in the program menus. With suitable configuration this could route the test board, as shown in **Fig. 3**, putting it in category A.

This router is capable of routeing a single-sided board, or from two to six layers, and it can autoneck. Like the Dos version of this autorouter, setting up is comprehensive. A costing system is used, so you can set the autorouter a penalty for routeing away from the direct route, making a change of direction, putting in vias etc.

Figure 4 shows a typical menu, and this is repeated in various forms for each of the strategies listed 1 to 8 in this screen shot. Setting up may take longer than other autorouters, but to outweigh that, you have more control over the way that the autorouter functions.

Some abstract ideas such as axis overshoot, bend, and gravity strength and distance, are used in the strategy set-up. These are best read about in the manual; they take a little time to get accustomed to.

The third autorouter is Cooper & Chyan's *Specctra SP2* for Windows. This is the most capable of the three and is also in category A. All the comments already made in my first review about this autorouter apply, and it should be noted that the track-spreading option is included, and is no longer an extra. However, the security system for this autorouter still uses a dongle, and in addition, a you need a password.

Seetrax has provided another of its excellent short-form interfaces with *Specctra*, so it is not really necessary to enter the *Specctra* program fully unless you especially want to. With this interface you can avoid learning another, very different program.

However, if you want master the intricacies of *Specctra*, you can run it separately. Three chunky *Specctra* operators manuals are provided should you wish to learn how to do this. I expect that most users will go for the soft option and stick to the *Ranger 2* interface. This highlights the difference between the *Specctra* and the Seetrax Rip/retry autorouter. The latter needs considerable input from the operator to configure it, whereas the *Specctra* is easy and quick to set up with the *Ranger 2* interface and can still produce neat, very well-routed boards.

Hard copy of the routed board is obtained via the Windows printer drivers or from Seetrax's own plotter driver. It is now possible to mirror the output from either. In the dos version this had to be done by means of Gerber input/output files.

The plotter driver is suitable for HP or compatibles and Houston machines. There is no direct output to the plotter but a text file is produced for sending to the plotter. Output is done from the dos prompt but this is not made clear in the manual, and there are no specific instructions for doing this. For those of you who are rusty on dos, I used the following command to get the plotter to run;

```
C:\PRINT /D:LPT1 A:\TESTCIRC.TXT
```

The plotter in this instance was on LPT1, which is normally the printer port, and I copied the plotter file TESTCIRC.TXT in advance to a floppy disk, drive A:, to avoid a long path entry.

This is a roundabout method way of achieving plotting but it works. Seetrax could improve this aspect, and should include extra advice in their manual about it. The results of using this plotter driver were good.

Printing out the artwork using the Windows laser printer drivers also gave good results. The position and size of the artwork in relation to the page is shown on-screen before printing, so adjustment is particularly easy.

In summary

As most users are following the inevitable Windows trend, this version of *Ranger 2* will undoubtedly attract buyers who would otherwise have passed by the dos version. No major new features over the dos version have been added. Seetrax has simply brought *Ranger 2* into the Windows era and incorporated several items that were optional extras in the dos version into one package. The attractions are obvious; this is the lowest-priced unrestricted Windows program that offers a user-friendly schematic drawing program and a capable rip-up-and-retry autorouter.

The added incentive of a clear autorouter up-grade path to the very able *Specctra* remains, if you eventually find you need more power.

Sensibly, Seetrax still offers its dos version of *Ranger 2* at £150 for the those on a tight budget or who prefer dos. If you are already using *Ranger 2* for dos and want to transfer to the Windows version, you can upgrade for £90. ■

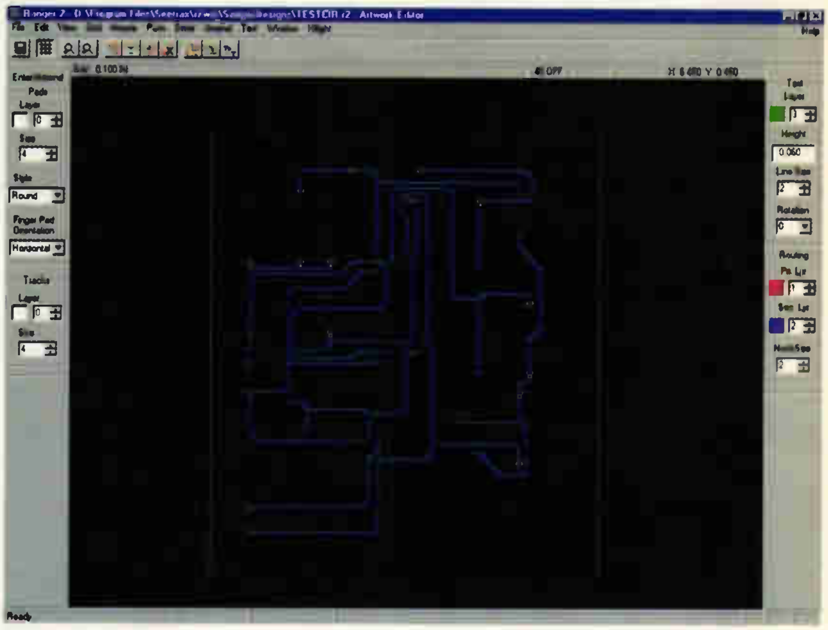

Fig. 5. Results of the Specctra autorouter on the same rat's nest.

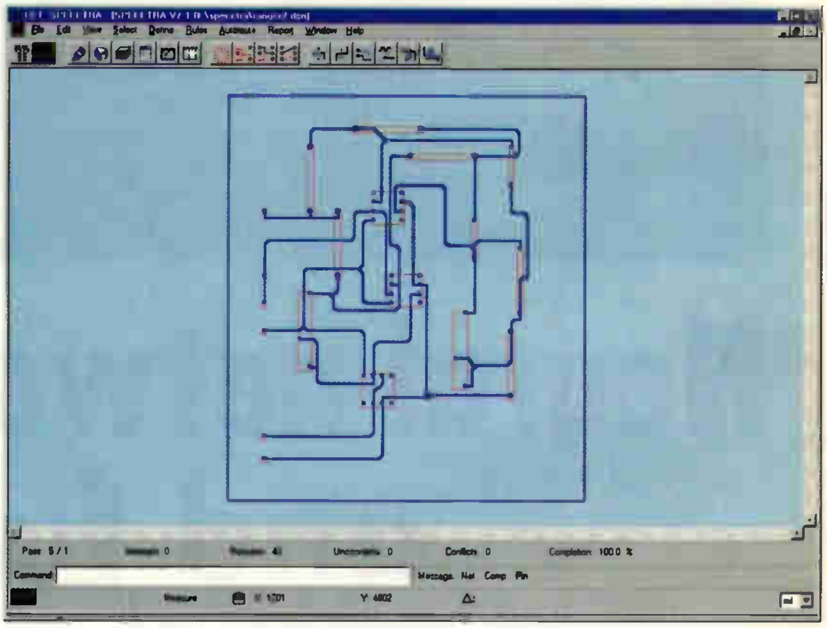

Fig. 6. A view from Specctra. The results of using the clean, spread and centre strategies from within Specctra.

What's next?

Traxmaker and *EDWin* are to be reviewed in Rod's next article. *Proteus IV* was reviewed in last month's issue.

Neural networks and fuzzy logic

"What can neural networks and fuzzy logic tell us about the future of digital electronics?" ask Chris MacLeod and Grant Maxwell.

Artificial neural networks and fuzzy logic are two popular modern techniques used to improve the performance of intelligent electronic systems.

At first sight, the two appear quite different from each other, and also from standard digital electronics. This article explains why the technologies are actually very similar – but complementary – to each other. It also looks at the possible future of intelligent electronics as a combination of the best of these existing techniques.

Artificial neural networks were developed during the fifties and are an attempt to mimic the structure and behaviour of the biological brain. Their most important attribute is that they are capable of learning. Such a network may be trained to act as a controller or to recognise characters for example.

Artificial neural networks achieved some early successes in fields such as pattern recognition, but it was quickly realised that they did not live up to the initial promise of real intelligence. During the early eighties, interest revived, and again neural nets are a popular subject of research. However, they have consistently failed to find their way into domestic or consumer products, and most design engineers are not aware of their possibilities.

Fuzzy logic was proposed in a paper by Lotfi Zadah[1] in 1965. It was received with scepticism by the engineering control community, typical comments being that it was common sense and 'just' statistics.

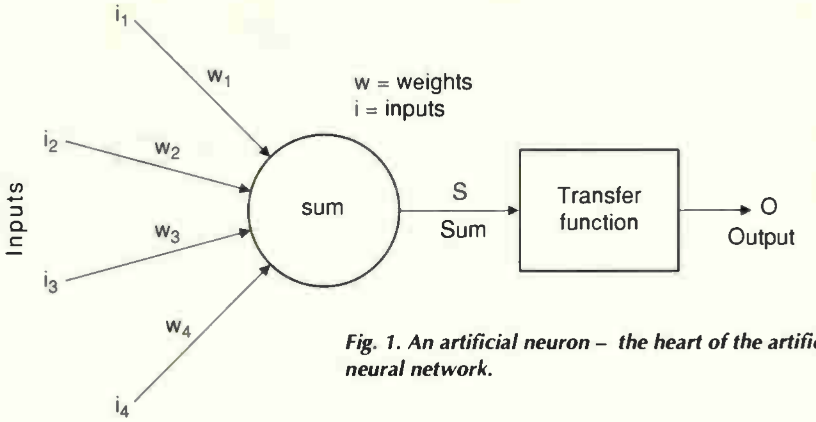

Fig. 1. An artificial neuron – the heart of the artificial neural network.

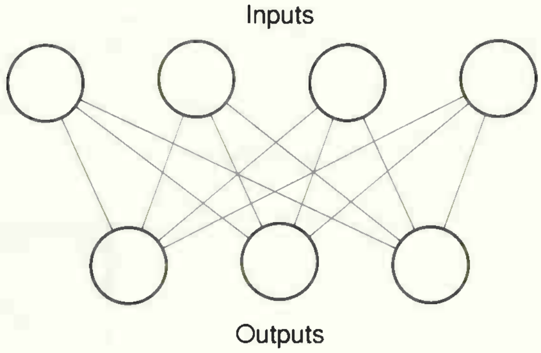

Fig. 2. A neural network seeks to mimic the brain by connecting together the neurons as shown in Fig. 1.

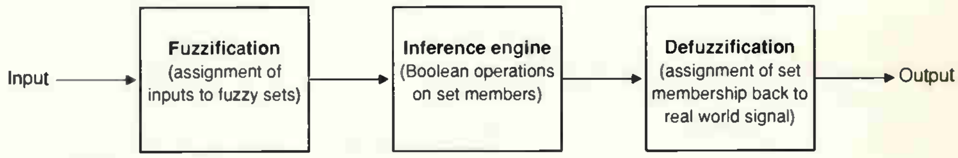

Fig. 3. Elements of a fuzzy system.

Since the initial reaction, fuzzy logic has steadily gained in acceptance and popularity, especially in Japan. It has found its way into many domestic and industrial products[2].

In this article, I take a look at these technologies and show that they are actually very similar, in that they are all mapping methods. First, let's consider artificial neural nets and fuzzy logic in a little more detail.

Artificial neural networks

The artificial neural network was proposed by McCulloch and Pitts[3] in 1943. The first such network was later built by Rosenblatt[4] in 1957.

Artificial neural networks are a network of very simple processing units. Each unit in itself has little power; it simply weights its inputs and sums them together[5], then applies a transfer function to the result, **Fig. 1**.

When a network of these simple units is connected together in a manner similar to that shown in **Fig. 2**, we call the result a neural network.

Networks may have discrete (ie digital) or continuous (ie analogue) outputs. We will be considering those with continuous outputs.

For the network to learn, it requires a training algorithm. This algorithm simply changes the weights of the network until the required outputs are achieved. There are several different methods for achieving this, the most important is called back propagation. This algorithm calculates the difference between the output required and the output that actually occurs and then changes the weights so as to minimise the difference between the two.

Fuzzy logic

Fuzzy logic is most often applied to control systems. The term 'fuzzy' arises from the fact that the system uses continuous as opposed to discrete inputs.

The fuzzy logic system is basically an expert system, in which a continuous input is assigned a membership of several sets. These values are then operated on by inference engine and the result, which is also continuous, is used to control the system, **Fig. 3**.

Fuzzy logic systems show several advantages over conventional control systems. These include ease of design, and better performance in many systems such as speed controllers in lifts and vehicles. In a speed controller, the continuous nature of Fuzzy logic causes the motor to speed up and slow down more smoothly.

Neural nets, fuzzy logic and digital electronics

These systems all appear quite different from one another, but are actually very similar. In the case of combinational digital electronics the binary inputs are mapped onto binary outputs, **Fig. 4**.

The user designs the circuit using Boolean algebra or Karnaugh maps to achieve this mapping. In the case of fuzzy logic or neural networks the mapping is of a continuous input to a continuous output, **Fig. 5**. The system knows how to map either because it has learned the relationship, as in the case of neural networks, or because the relationship has been encoded in the inference engine by the user, as in the case of fuzzy logic.

In many books, the operation of the neural net is viewed as dividing up data space with separators. This causes the network to have certain mapping limitations which are beyond the scope of this article. For more on this, see reference 6.

Although the networks shown above are combinational logic and feedforward networks, the similarities also apply to sequential logic which is equivalent to recurrent neural networks such as the Hopfield net[7].

The relationship between these systems should now be more obvious. Digital logic is simply a special – discrete – case of a general, or continuous, mapping system. To demonstrate this, let's make up a new entity, the fuzzy gate; this is simply a digital gate which accepts a continuous input rather than a discrete input – as in the case of 'normal' digital electronics. First, consider the functions of normal logic represented as Venn diagrams, **Fig. 6**.

In terms of normalised voltage levels these represent MAX, MIN, and COMPLEMENT functions, **Fig. 7**. These functions are actually used within the fuzzy logic inference engine.

Such a simple system, operating on continuous voltage levels, has important advantages:

- It decomposes to normal logic if the inputs are 1 and 0. Consider the Boolean expression A+B(A+B) which reduces using normal methods to A+B, then take a look at **Table 1**. You can see that the continuous logic gives the same answer as the standard discrete logic; this can be shown in every case.

- It may be arithmetically manipulated in the same way as normal logic. The discussion in

Table 1.Using normal methods, the Boolean expression A+B(A+B) reduces to A+B.

A	B	A	B	A+B=D	D.B=E	E+A	A+B
discrete	discrete	cont	cont	MAX	MIN	MAX (answer)	Boolean answer
0	0	small	small	small	small	small	0
0	1	small	large	large	large	large	1
1	0	large	small	large	small	large	1
1	1	large	large	large	large	large	1

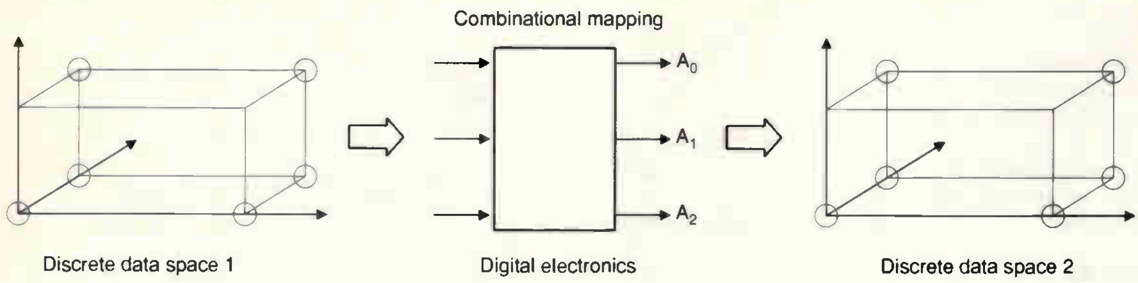

Fig. 4. Conventional digital electronics involves discrete mapping, where the binary inputs are mapped onto binary outputs.

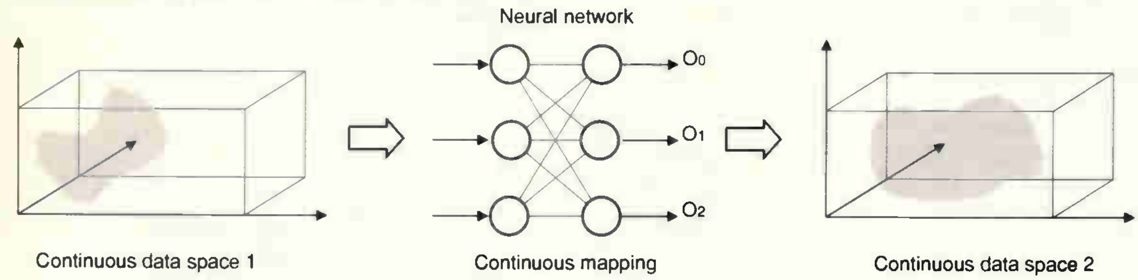

Fig. 5. In fuzzy logic or artificial neural networks, a continuous input is mapped to a continuous output. The system has either learned how to map, or has been told how to map by the user.

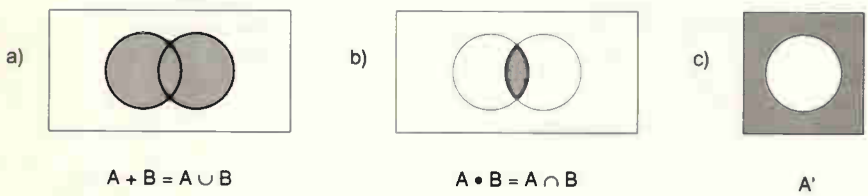

Fig. 6. Venn diagram representation of normal logic building blocks.

the example above shows that the continuous logic reduces to discrete logic if the levels are forced to their maximum and minimum values, in general:

$A \cup B = A+B \qquad A \cap B = A \bullet B \qquad A' = \bar{A}$

and they may be manipulated as such.

- It degrades gently. Consider the case A+B, where a high input to a continuous logic gate is 0.7 and low 0.2

A	B	A + B
0.2	0.2	0.2
0.2	0.7	0.7
0.7	0.2	0.7
0.7	0.7	0.7

- Mathematical logic circuits continue to fulfil their functions. Consider the half adder in **Fig 8**. The XOR is implemented as $\bar{A} \bullet B + A \bullet \bar{B}$. Consider,

A	B	$\bar{A}$	$\bar{B}$	$\bar{A}B$	$A\bar{B}$	$\bar{A}B+A\bar{B}=S$
0.4	0.5	0.6	0.5	0.1	0	0.1

$A \bullet B = C = 0.4$

Output is therefore $(0.1\times2^0)+(0.4\times2^1)=0.9$. To clarify this, one of the operations involving the complement function is shown in **Fig. 9**.

As you can see from Fig. 9, because of the NOT or complement function a true fuzzy digital output requires a number pair to represent it. In the case of the function above 0.4, 0.5 ie lower limit, upper limit; for complex functions there could be several number pairs.

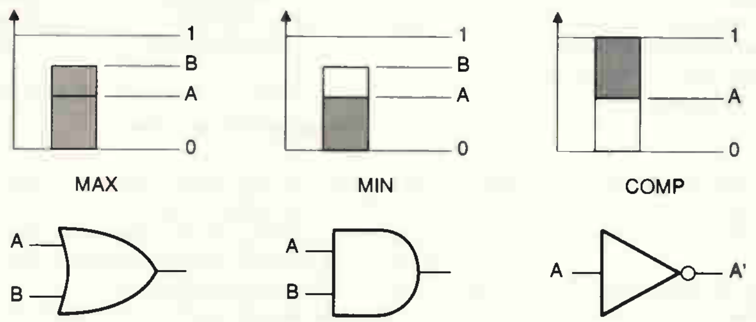

Fig. 7. Boolean functions as continuous functions.

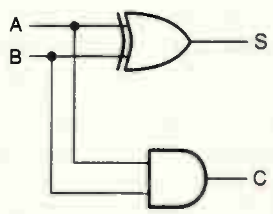

Fig. 8. Half adder using AND and exclusive-OR gates.

Using such techniques it is possible to design, for example, a pattern recognition system which gives similar results to a neural network system. However, the fuzzy gate system has the advantage that it can be manipulated and proved using Boolean algebra.

- The final advantage is that such a system interfaces naturally with the analogue world. This type of system has the advantages already pointed out for fuzzy logic, because it produces intermediate outputs. When used as a controller, it allows outputs to change gradually.

In summary

Looking at the modern techniques, neural networks and fuzzy logic, these technologies give two important advantages over traditional Boolean logic gates.

One advantage is that neural networks can learn from example, so no detailed design is involved. Secondly, neural networks and fuzzy logic operate with continuous signals.

As we pointed out earlier, the continuous nature of signals gives the system certain useful properties. It causes the system to degrade gently in the presence of noise. It means that systems such as lifts slow down and speed up gently – i.e. it interfaces better with the natural world. The continuous nature of the signal also makes the system somewhat more fault tolerant.

As you have seen, standard digital logic is a special case of fuzzy digital logic. This means

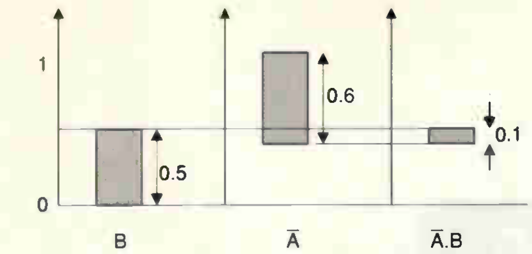

Fig. 9. An operation using the complement function.

that systems designed using fuzzy digital electronics behave in a very similar way to neural networks or fuzzy logic, and further that 'hard' digital logic is a special case of 'fuzzy' digital logic.

The other main difference between standard logic and neural networks is that neural networks learn mappings through a training cycle; the other two systems have a designed structure. However it is possible to design fuzzy digital logic systems that can learn.

One way this can be achieved is by implementing a training algorithm that adjusts the threshold of the MAX and MIN functions of the gates to minimise the output error. You can therefore envisage a compound system *neuro-fuzzy digital electronics*. This would have all the advantages of Boolean logic – for example, easy mathematical manipulation – but have continuous outputs and be trainable in the manner of a artificial neural network.

Looking even further ahead, the design of the neural processing elements is based somewhat loosely on the structure of biological neurones. This association has caused many workers to only consider models which are biologically feasible and this may have hamstrung researchers somewhat – in fact it has almost become an ideology.

Future networks may take a much less rigorous view of neural units, and allow them to have designed or evolved processing capabilities. The end result may be a general technology – a new electronics, one stage beyond neuro-fuzzy digital electronics – which has all the advantages of the previous examples without any of the disadvantages. ■

For more on this topic, visit the author's web page: http/www.eee.rgu.ac.uk/staff/cmhome.htm

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Trac competition winner

Ben Sullivan's winning entry from the Fast Analog Solutions *TRAC* design competition is a synchronisable 400Hz three-phase generator with out-of-lock indicator for testing aircraft equipment.

400Hz in three phases

In the main, modern aircraft electrical systems use 200V three phase supplies running at 400Hz. When development, testing and repairing of such instruments, there is a need to power them on the bench, away from their normal supply.

In the early days, a 400Hz alternator, driven by a mains electric motor was used. But electronic supplies have long been preferred on various grounds. These include flexibility, reliability, the option of synchronising the source with an external 400Hz pilot signal, and of course, the absence of acoustic noise.

An electronic 400Hz three phase supply consists of a bank of three high power audio amplifiers plus their associated dc supplies, driven by a suitable exciter. This note describes the implementation of a 400Hz three phase exciter, using the *TRAC* device.

Circuit description

The exciter uses a state-variable filter as the basis of a 400Hz oscillator. The filter section consists of auxiliary elements at pins 9, 13 and 17, providing high, low and band-pass outputs HP, BP and LP at pins 11, 15 and 19 respectively, **Fig. 1**.

The band-pass output is attenuated by 20dB – i.e. a factor of ten – in Aux 14 and applied as the damping term via pin 4 and two adders back to the input of the highpass stage, pin 9. This sets the filter Q at 10.

The filter is driven by a limited version of its own output, applied as positive feedback via pin 6 and one adder, again to the input of the high pass stage. Element Aux 20 amplifies the bandpass signal by 26dB (×20), producing a near squarewave, of peak to peak amplitude limited to the maximum swing permitted by the supply rails.

This signal is attenuated by 34dB – a factor of 50 – and is then negated to produce the correct phase for positive feedback to pin 6. The result is a low-distortion sinewave at the band-pass filter output, due to the operating Q of the circuit.

The Q enhances the fundamental component of the positive feedback signal by 20dB, while the action of the integrator Aux 13 attenuates the third harmonic component by 9dB. As the third harmonic content of the positive feedback signal is in any case only one third the amplitude of the fundamental, the harmonic content of the filtered sinewave at the BP output is only around 1%, the harmonic content at the LP output being of course lower still.

Aux 25 scales the BP output down by 50%, while Aux 31 produces an output of 86.6% of the LP output. Combining these in Add 39/40 produces an output at 120° to the BP output. Similarly, combining a negated version of the output of Aux 31 with the output of Aux 25 provides the third phase. The outputs at pins 12, 37 and 41 – scaled to an appropriate level if necessary – drive the inputs of the three high power audio amplifiers.

In hardware, the three-phase oscillator runs as described. In simulation, a small input applied to pin 1 is neces-

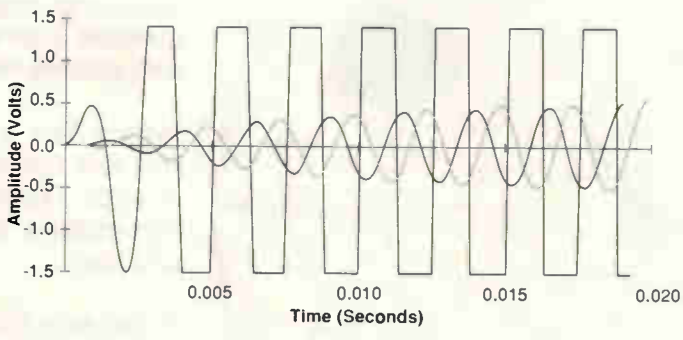


Fig. 1. Phases A and C of the output. Also shown is the clipped version of output B, which forms the input to drive the filter. The 0°, 120°, 240° phasing is clearly shown.

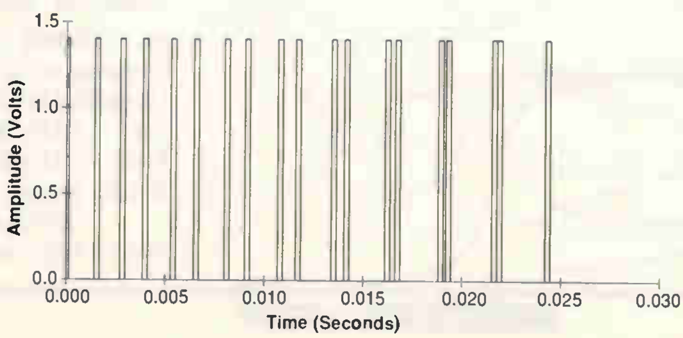


Fig. 2. The out-of-lock detector output, shown for the case where the oscillator frequency is not locked to the injected synchronising frequency. The equal 180° spacing of some pulses is evident. In lock, the pulse spacing will not exceed ±90°, as in the right hand half of the trace.

sary to start oscillation. The amplitude of the positive feedback signal stabilises within a few cycles of start-up, due to the large gain in the limiting amplifier Aux 26.

The three-phase outputs, however, build up more slowly, at a rate determined by the Q of the filter. In hardware, pin 1 may also be used to inject a low-level locking signal, either sine or square. This may be used to synchronise the three phase output with other equipment in use at the same time. Provided the frequency difference is not excessive, the state variable filter oscillator will pull in to the injected external frequency.

Increasing the amplitude of the injected signal increases the frequency range over which lock can be maintained. Designing the oscillator to operate at a lower Q will also increase the frequency range over which lock can be maintained, but at the expense of an increase in harmonic content.

Locked, or not?

When operating with the supply synchronised with an external signal, it is important to know whether the circuit is properly synchronised, or whether lock has been lost. This is achieved in conjunction with a dual retriggerable monostable, not shown.

The external sync signal is applied to log stage 45, which produces a squared up version, approximately independent of its actual amplitude. This is attenuated by 36dB by Aux 49, differentiated by Aux 53 and the positive spikes selected by Rec 55.

The BP signal is processed similarly by items 46-56, and the two sets of spikes appear combined at pin 59. The characteristics of this pulse train provide a means of detecting the 'out-of-lock' condition as described below.

When the frequency of the external input equals the free running frequency of the oscillator, the BP output will be in phase with the external input. When there is a small difference, lock will be maintained, but with a standing phase shift between the two. As the frequency difference increases, so does the standing phase shift, up to a maximum of 90°. With a larger frequency difference, lock is lost, and a beat frequency appears.

The period of the 400Hz waveform is 2.5ms. When the oscillator is in lock, the phase shift between its BP output and the locking signal cannot exceed 90°. There will therefore be 'pairing' of the pulses at pin 59, the gap between pairs being at least 1.875ms – i.e. 75% of 2.5ms.

Positive going spikes at pin 63 are applied to a retriggerable monostable with a period of 1.6ms. As a result, it will always time out while the oscillator is in lock.

The output of the 1.6ms monostable device is applied to a retriggerable mono with a 2.8ms period. This monostable remains permanently triggered while the oscillator is in lock. But if lock is lost, as the two frequencies drift, a point will be reached where the spikes at pin 63 are evenly spaced at 1.25ms intervals. Therefore the 1.6ms mono no longer times out every cycle. Hence it no longer keeps the 2.8ms monostable triggered, and timeout occurs. This may be used to drive an interrupt on a controlling micro, or may be latched to illuminate an out-of-lock indicator lamp, sound an audible warning etc.

Apart from the usual passive components associated with AUX blocks, and a dual retriggerable mono with its timing components, the exciter is realised entirely with Zetex TRAC020 devices. ■

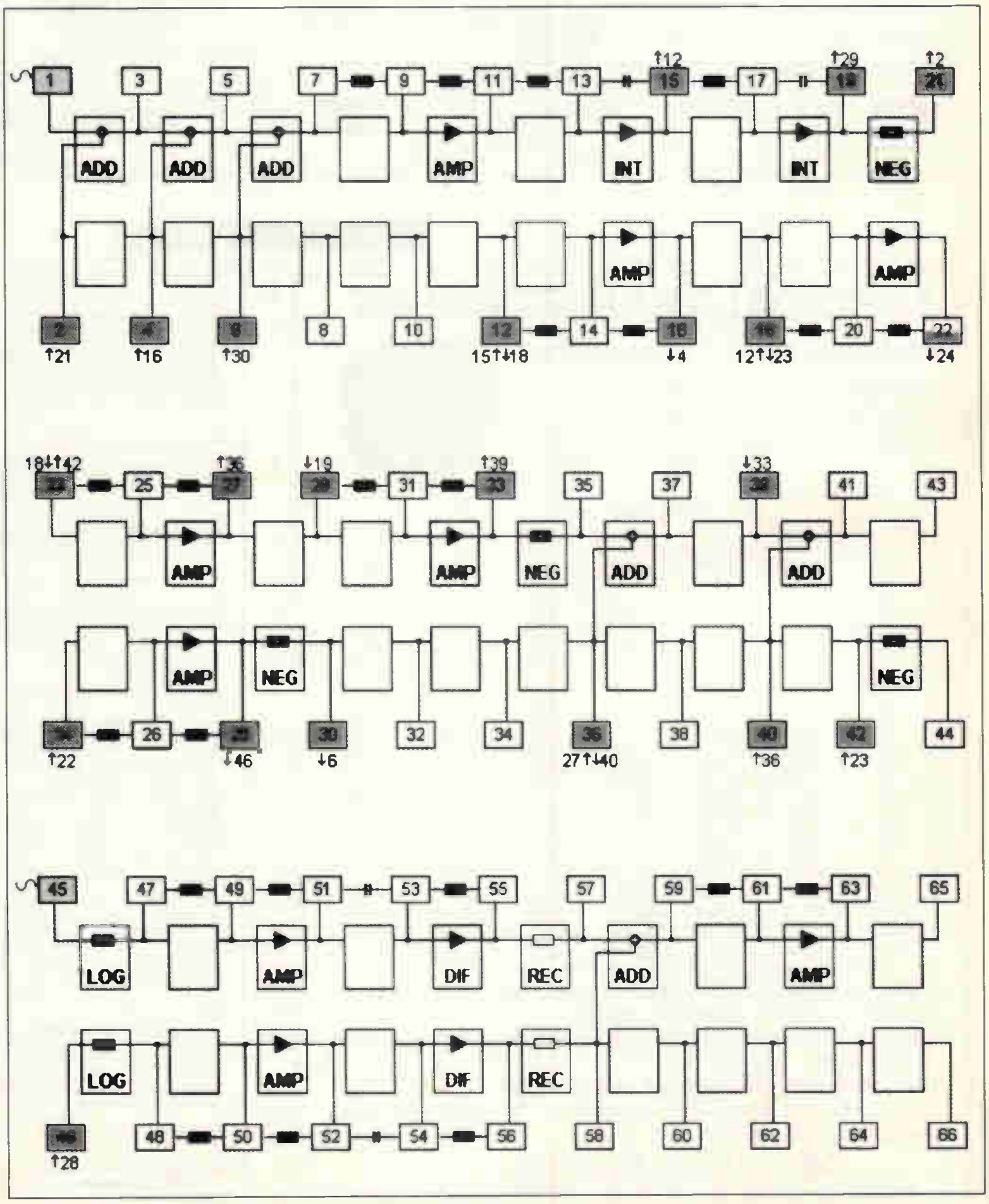
Inputs representing the simulation stimullii applied to the exciter to get the results shown in Figs 2, 3 and a list of the values applied to the auxiliary functions.

Inputs

Pin	Generation	Amplitude	Frequency
1	Sinewave	0.02	400
45	Sinewave	0.02	400

Auxiliaries

Pin	AUX function	Value
9	Amplify	1
13	Integrate	2513
14	Amplify	0.1
17	Integrate	2513
20	Amplify	20
25	Amplify	0.5
26	Amplify	0.02
31	Amplify	0.866
49	Amplify	0.02
50	Amplify	0.02
53	Differentiate	0.1
54	Differentiate	0.1
61	Amplify	1



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1GHz power transistor. Ericsson's *PTF10007* n-channel, laterally diffused mosfet shows a flat gain curve of at least 13dB at up to 30W and is meant for use at frequencies to 1GHz. It is designed for a 28V supply and has a maximum drain/source breakdown of 65V. The device has an advantage over bipolar types in that intermodulation reduces as power is reduced and, at 20W in class AB, the two-tone *IMD3* is –36dB. The *PTF10007* is an alternative to the *MRF182*, which is in an identical package. Ericsson Components AB. Tel., 01793 488300; fax, 01793 488301.
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Linear integrated circuits

Power op-amps. MSK's *MSK0021/41* are fet-input amplifiers producing output currents of 1A/2A, suitable for use in video amplification, comparators, line drivers and fast integrators. Each is a replacement for a military component, the *0021* in a TO-3 package meets the DESC SMD 51651-850088 requirement and the *0041* in TO-8 meets DESC SMD 51651-85087. Ashwell Electronics Ltd. Tel., 01438 364194; fax, 01438 313461.
Enquiry no 502

Quad, fet-input audio amp. Very low distortion of 0.00008% thd and voltage noise of 8nV/√Hz are claimed for Burr-Brown's *OPA4134*, the quad version of the *134* and *2134*. A fet cascoded input stage prevents input bias current varying throughout the common-mode range and the output swing is to within 1V of the rails. It is unity-gain stable and shows no phase inversion or overload problems. Characteristics are: 8MHz gain bandwidth, 20V/µs slewing, 5pA input bias and a supply range of ±2.5V to ±18V. Burr-Brown International. Tel., 01923 233837; fax, 01923 233979.
Enquiry no 503

Quad switch. Siliconix has the *DG213* low-cost quad analogue switch that has two normally closed and two normally open switches. It is intended for analogue signal processing and replaces the more expensive *DG403/413*. On resistance is 45Ω, a current consumption of 1µA and supply tolerance of 3-22V rendering the device very suitable for use in battery-powered equipment. Packages on offer are TSSOP-16, SOIC-16 and DIP-15. IEC Micromark Electronics Ltd. Tel., 01628 76176; fax, 01628 783799.
Enquiry no 504

Memory chips

Mtp flash. Many-time programmable flash memory by Silicon Storage Technology combines the functions of flash and the cost of eprom; each mtp may be electrically erased and reprogrammed at least 1000 times by the use of an external programmer. Lower cost is obtained by means of the company's *SuperFlash* technique which uses a thick oxide process to give better retention and reliability and also by reducing peripheral circuitry such as single power supply and in-system programming, although the faster write performance of *SuperFlash* is retained. The devices are pin-compatible with standard eproms and one-time programmable types. First to appear is a 1Mb device, organised as 128K by 8, having the option of reading at either 5V or 2.7V, programming and erase at 12V, access at 55ns and program time 70ns. Silicon Storage Technology Ltd. Tel., 01784 490455; fax, 01784 490512.
Enquiry no 505

512K×32, 3.3V sram. EDI has a new, fast, 3,3V 16Mb sram having access times of 15, 17 and 20ns. The asynchronous *EDI18L32512V* is byte-addressable and provides a single-chip data memory for the TI *TMS320LC31* processor or, since the device may be arranged as a 1Mbx16 memory, for the Analog Devices *SHARC* dsp. EDI (UK). Tel., 01276 472637; fax, 01276 473748.
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Microprocessors and controllers

C-programmable controller. Z-World's *PK2400* is a C-programmed controller with a built-in graphic display and keypad, meant for use in machine control, embedded systems and any applications needing operator input. There are nine protected digital inputs, two conditioned analogue inputs for a variety of sensors, six high-current outputs for solenoids and relays, 2A relays, two serial ports and 128K of non-volatile flash memory. Programming is carried out using the company's Dynamic C version of standard C, which is optimised for real-time, multi-tasking and has an easy editor, compiler and interactive debugger. It also has a library of functions and software drivers. A development kit is available. Impulse Corporation Ltd. Tel., 01543 466552; fax, 01543 466553.
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Motors and drivers

Microstepper drive. Digiplan's *OEM750* microstepping drive has a 7.7A output, yet is contained in a 127 by 91 by 41mm module. Outputs are optimised, for 57 and 83 frame size steppers three-state current control being used for efficiency and cooler motor operation. Operation may be from any voltage between 24V dc and 75V dc, due to the use of current-loop gain control, which also allows the range of motor inductances to lie between 0.2mH to 80mH – in other words, most of the available motors. With Parker motors, holding torque may be up to 3Nm and resolution is selectable from 200 to 50800 steps per revolution. Parker Hannifin plc, Digiplan Division. Tel., 01202 699000; fax, 01202 695750.
Enquiry no 508

Programmable logic arrays

24000-gate pld. Altera announces the *EPF6024A*, the newest and densest recruit to the *FLEX 6000* family of programmable logic devices, this one having 24000 gates and operating from 3.3V. It uses a 0.35µ, triple-layer metal sram process and incorporates the company's *MultiVolt* feature that allows the i/o to work with 5V, 3.3V and 2.5V interfaces. All members of the family meet PCI specifications. Altera UK Ltd. Tel., 01494 602020; fax, 01494 602021.
Enquiry no 509

Power semiconductors

7mΩ power mosfets. Power mosfets in the DPAK by Siliconix, the *SUD50N03-07* and *SUD45P03-10* are made using the Siliconix Trench technique and offer 7mΩ and 10mΩ on resistance respectively and provide about one-third more current at 20A and 15A. Siliconix/Temic Marketing. Tel., 01344 707300; fax, 01344 427371.
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PASSIVE

Passive components

Minute mains transformer. Stontronics believes it has the world's smallest, printed-board-mounted mains transformer, the *Mini EE 20/4*, which weighs 18g and is 12mm high. It has a double-section bobbin for maximum insulation and is rated for use at 70°C, being approved to UL, CSA and VDE 0551/EN 60742; it is also short-circuit-proof. All versions in the range take 230V ac input or 115V ac for the USA. Output voltages are in the range 6-18V ac and the transformers are rated at 0.08VA. Stontronics Ltd. Tel., 0118 9311199; fax, 0118 9311145.
Enquiry no 512

Inductors. *PKH* radial lead inductors from Total Frequency Control are wound on flat cores and provide twice the inductance and *Q* values of axial lead types, as well as taking up less board space. Values range from 1µH to 4700µH to within ±10% of nominal. Total Frequency Control Ltd. Tel., 01903 745513; fax, 01903 742208; e-mail, eddie@tfc.co.uk.
Enquiry no 513

Thin-film delay lines. Thin-film lines in the *CL*, *GL* and *DL1L5* series by Thin-Film Technology Corp. are based on microstrip, stripline and multiconductor transmission lines. All work at frequencies to 5GHz, the *GL* chip lines having delays of 20-200ps in 20ps steps. Tolerance is ±10ps, impedance 55Ω and rise time under 10ps, with insertion loss at 0.5dB at 1GHz. *GL* surface-mounted models delay in the 0.1-4ns range with 200ps transition times and the microstrip *DL1L5* type covers the delay range 0.1-5.5ns with transitions of under 1ns. Three-pin types in the *DS1L5*

Cameras

Miniature camera module. *GMC-810* by Fujitsu is a ccd camera measuring 38 by 43 by 20mm and intended for use in security or multimedia work. It generates a standard composite video output and there is a dc-to-dc converter to supply the logic and ccd. A 0.25in ccd array has 440 000 pixels and handles both Pal and NTSC formats, with an s:n ratio of 48dB with agc off. Automatic sensitivity control has a range of 300:1 and the camera will work with a minimum illumination of 1 lux at f1.2 (6 lux at f2). Through-the-lens sampling effects auto white balance. Adital Ltd. Tel., 01296 337755; fax, 01296 398085.
Enquiry no 511

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range cover 0.1-10ns. Rhopoint Components Ltd. Tel., 01883 717988; fax, 01883 712938.
Enquiry no 514

Connectors and cabling

USB plug-and-play. GTK has a range of Universal Serial Bus connectors to work as 'hot pluggable' units for those applications needing plug-and-play capability; they are available as single or dual connectors and support low and high data rates. *Series A* or *B* types are on offer in through-hole or surface types. GTK (UK) Ltd. Tel., 01344 304123; fax, 01344 301414.
Enquiry no 515

High-temperature DIN connectors. Made by Elco, part of AVX, the *AVX 8557/8577* DIN 41612 connectors are fabricated in a liquid-crystal polymer to withstand up to 15s at 150°C, which allows them to handle the surface-mount process and 20min of pre-process and post-process heat. They are supplied in three-row versions with 48 or 86 contacts in standard and inverted form and in four-row types of 128 contacts, inverted. Hawnt Electronics Ltd. Tel., 0121 784 3355; fax, 0121 7831657.
Enquiry no 516

Emc gaskets. With five new types of D-type anti-emi gaskets, Brandauer is now able to cover all standard 9-50way DIN connectors. As a lower cost alternative to silver-loaded models, these gaskets use beryllium-copper as standard or stainless steel with various finishes. Gasket cross-section is wave-like in form to provide maximum attenuation and improve surface contact between surfaces. Brandauer & Co. Ltd. Tel., 0121 359 2822; fax, 0121 359 2836.
Enquiry no 517

S-m backplane connectors. Siemens believes its *DensiPac* surface-mounted backplane connectors to be the first available. 576 pins are accommodated in a 100mm strip on a pitch of 1.25mm – about 40% more than is available in other designs, and data rate supported is up to 622Mb/s. For even greater density, the connectors can, being surface-mounted, go on both sides of the board. Siemens plc. Tel., 01344 396313; fax, 01344 396721.
Enquiry no 518

Materials

Thermally conductive gel. Raychem announces the new *HeatPath* thermally conductive gels to improve the thermal management of heat-sensitive components, its compliant nature allowing it to fill all gaps between components and heat sinks with little pressure, so avoiding damage and the need for clamps. Three types are available: *CTQ 1000* is a thin gel on woven glass fibre in thicknesses of 0.25mm, 0.5mm, 0.75mm and 1mm; *GTQ 200* is an open-cell reticulate of 1.5-4mm thickness for large gaps; and *GTQ 3000* is cast into a sheet or specified profile in 5-15mm thickness. No heat or curing is needed and the material will not flow away over time. Raychem Ltd. Tel., 01973 572692; fax, 01973 572209.
Enquiry no 533

Crystals

S-m crystals. *HC-49/U-SMX* crystals by Advanced Crystal Technology are said to offer the best value for surface-mounted use. The crystals are AT-cut and come in standard and 2.5mm high form, frequency and temperature stability both being ±50ppm. Frequency range is 3.18MHz to 72MHz, with others available to order. Flat-tab terminations allow the use of reflow soldering. Advanced Crystal Technology. Tel., 01635 528520; fax, 01635 528443.
Enquiry no 519

Displays

Lcd kit. Display Solutions has a kit, assembled from components by various manufacturers, to provide an lcd for mobile computers and vehicle instrumentation. The display itself is a 4.7in or 6in transmissive vga colour type or a 5.3in mono screen in reflective, transmissive and transflective versions. There are options of software-controlled contrast and backlight control and resistive touch-screen control. Add-on modules include processor motherboard, floppy and hard drives, solid-state disks, a modem, a-to-d converters, a sound card, a SCSI port, wireless Ethernet and four-port serial cards. Display Solutions Ltd. Tel., 01480 463377; fax, 01480 468989.
Enquiry no 520

Data logging displays. Two new data logging displays by Lascar have joined the *EasyLog* range: the *EL-1-12BIT* panel-mounted unit and the *EL-2-12BIT*, which is in a small enclosure. Both measure, display, record and control voltage, current and temperature, a 12-bit a-to-d converter producing a 1999 maximum count with a 1 count resolution on the high-contrast lcd. There is an 8000-reading memory and a selectable sample rate up to 12h. Both may be set up using the same EL-WIN software as the earlier 8-bit models. Lascar Electronics Ltd. Tel., 01794 884567; fax, 01794 884616.
Enquiry no 521

Filters

Saw filters. Fourteen members of RF Monolithics' *SF* family of low-loss transversal surface acoustic wave filters cover the frequency range 71MHz for GSM ifs to 459MHz for cable television pilot tones, with centre frequencies of 110.59, 135.5, 254.4, 254.5 and 350MHz, minimum 3dB pass-band widths being 680kHz-1.5MHz. Acal Electronics Ltd. Tel., 01344 727272; fax, 01344 424262.
Enquiry no 522

Hardware

Quiet fans. Papst now produces fans with sleeve bearings for a quieter life, these being sleeve versions of existing *4400, 4300* and *4200* ball-bearing types, which will still be made. The bearings use in the patented Sintec technique, which lasts as long as ball bearings but tolerates a wider temperature range, costs less and makes less noise. Further advantage of the sleeve bearings are that they are less vulnerable to shock and vibration and that the noise does not get worse with age, in contrast to the ball type. Papst plc. Tel., 01264 333388; fax, 01264 332182.
Enquiry no 523

Optical-fibre management. Radiall-Transradio's *Fibre Management Centre* is an aid to the secure routeing and connection of multiple, multi-fibre cables, catering for the use of over 2000 connectors. It allows for the breakout and connection of fibres or re-routeing by splicing and accommodates a number of standard connector interfaces. Four standard steel cabinets can be used, which have a central support bar to carry the cassettes and magazines, there being twelve cassettes to one magazine. Cassettes protect fibres and allow access, each holding 12 connectors. Transradio Ltd. Tel., 0181 997 8880; fax, 0181 997 0116.
Enquiry no 524

Test and measurement

5kV megohmmeter. Danbridge's *DB600* megohmmeter range, now available in the UK, includes three models: the *DB601*, a general-purpose unit for 1kV/25mA working; a capacitor tester, the *DB602*; and the *DB604*, which is able to test at up to 5kV, in 1V steps. All have a large backlit lcd to show variation of leakage current or insulation resistance with time as a bargraph or X/Y curve, the bargraph using either autoscaling or eight fixed two-decade scales. All the instruments carry out timed sequences of up to 60min and can use the results to sort components into bins. RS232 and GPIB interfaces are fitted, as is an optically insulated interface. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.
Enquiry no 525

Digital wattmeter. Yokogawa's *WT110* single-phase power analyser measures voltage, current, effective, apparent and reactive power, power factor and phase angle to an accuracy within ±0.25% over the 0-50kHz frequency range. Six voltage ranges are 15-600V and six for current in the 500mA-20A range. Measurement mode is true-rms or mean and there is an optional harmonic analysis function with output to a printer. External current sensors with voltage output such as clamp-on transformers and Hall devices may be used. RS232 or GPIB interfaces are standard and an integration facility provides measurement of watt-hours or amp-hours. Martron Instruments Ltd. Tel., 01494 459200; fax, 01494 535002.
Enquiry no 526

Panel meter for power. *Multi-DIN* from Howard Butler is a panel meter to measure twelve electrical quantities in the installation and maintenance of power systems, including reactive power, active energy, reactive energy and frequency, any three of which may be displayed at once. The instrument is contained in a DIN 96 case and has a green, backlit screen

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with black figures, a locking transparent cover being provided. There is the option of RS 232 output. Farnell Components Ltd. Tel., 0113 263 6311; fax, 0113 263 3411.
Enquiry no 527

Visual behaviour analysis. Vision Control Systems announces *Testimony*, which is a pc-based instrument able to detect, record and analyse the exact point of observation of a user's eyes, which is of advantage in pilot and driver training and surgery. The instrument combines this technique with a miniature field-of-view camera, *Testimony* using retinal reflection to provide a cursor to indicate precisely where the user is looking. Currently, connection to the pc is by the serial port, but a radio link is also on the cards. Vision Control Systems. Tel., 0161 236 9880; fax, 0161 236 09730.
Enquiry no 528

Picoamp current source. *PA-1* precision picoamp current source from the Indian company Laxtronics is a secondary standard calibrator, providing outputs of 1pA-199.99μA in five ranges, selected by a front-panel rotary switch, with fine and coarse continuous settings. Current set is displayed on a 4.5-digit lcd. Power needed is 16.5mW from mains or batteries, battery power offering freedom from stray pickup and ground-loop leakage and allowing the instrument to 'float', if required. Output resistance is more than $10^{18}\Omega$. Laxtronics. Tel., 0091 253 383737; fax, 0091 253 381873. E-mail: laxtron@bom4.vsnl.net.in.
Enquiry no 529

Literature

Fairchild website. Over 5500 products and more than 10000 pages of information, Fairchild's site is divided into search tools, What's new? Products, Contact Fairchild and Company. New information is posted on the same day and customers receive replies quickly and in private. The search facility looks at the whole site when given one word or even a partial part number to show data sheets, application notes and static html pages, and the search process may be narrowed by asking for details under a number of requirements – part number, number of pins, price, etc. Fairchild Semiconductor Corporation. http://www.fairchildsemi.com.
Enquiry no 530

Data communications. A free catalogue of *Industrial Data Communications Products* is available from Dataforth. There are 60 pages of product data, applications mechanical details and installation information on modems, pc serial i/o boards and rack-mounted modem systems. For systems using RS-232, 422 and 485 and optical networks, the company has products to isolate, protect and extend the operating distance of computers and peripherals. Impulse Corporation Ltd. Tel., 01543 466552; fax, 01543 466553.
Enquiry no 531

Revolutions sensor. Panasonic's semiconductor magnetoresistive sensor detects rotation in the 0-15kHz range at temperatures between –40°C and 150°C. The device is said to be an ideal replacement for variable-reluctance and Hall sensors often used in vehicles when crankshaft and camshaft positions must be sensed, offering improved tolerance of poor conditions and better low-speed detection; it also maintains a constant output with speed and no amplification is needed of the 0.8-4.8V output voltage. Panasonic Industrial (Europe) Ltd. Tel., 01344 853862; fax, 01344 853310.
Enquiry no 540

Motor control. Ericsson's 1998 *Industrial Circuits* short brochure provides an introduction to the company's stepper motor ics and test boards. Applications information includes circuit diagrams and design help, guidance on thermal management and heat sinks. This family of stepper-motor drivers includes single and dual channel drivers with outputs to 1.5A continuous and voltage handling to 60V. Also described is a chip set for microstepping control and drivers for inductive load control. Ericsson Components AB. Tel., 01793 488300; fax, 01793 488301.
Enquiry no 532

Power supplies

Switched-mode, pfc supplies. A large number of single-output, switched-mode power supplies with power factor correction is introduced by Coutant Lambda. The *EWSP* range contains more than 40 models, all in metal cases and designed for use in industrial applications where conditions can be extreme. Outputs are from 100W to 600W at 2-48V, efficiencies being in the 60-84% range, depending on the model. *EWSP-100* and 150 models accept 47-63Hz input and autoselect input voltage of 85-132V ac or 170-265V ac, while the 300 and 600 types take 47-440Hz and 85-265V ac. Overvoltage and overcurrent protection and remote sensing are all standard and the higher-power versions are fitted with fans. Coutant Lambda Ltd. Tel., 01271 856666; fax, 01271 864894.
Enquiry no 534

Zero-standby regulator. Offering low dropout and zero standby current, the surface-mounted *BA00ASFP* voltage regulator from Rohm has p-n-p transistor output and provides an output voltage of 1.5-15V at up to 1A from an input of 4-25V; dropout is 300mV. Protection includes output over-current/voltage and thermal shutdown. Package is TO252-5DPAK measuring 6.5 by 2.3 by 5.5mm. Rohm Electronics UK Ltd. Tel., 01908 282666; fax, 01908 282528.

Radio communications

Agc/demod for digital cellular radio. *RF2667* from Anglia Microwaves is a 5-300MHz if agc amplifier and quadrature demodulator for battery-powered dual-mode cellular radios, spread-spectrum cordless telephones and related equipment. Voltage supply needed is 2.7-3.3V, current taken being 20mA from 3V in receive mode and 20μA in power-down, which is digitally controlled. The dual if stages produce a maximum gain of 50dB with 100dB of control; noise figure IP_3 is –50dB. An evaluation board is available. Anglia Microwaves Ltd. Tel., 01277 630000; fax, 01277 631111.
Enquiry no 535

Air-pressure sensors. Sumitomo high-sensitivity air-pressure sensors stocked by Anglia include the *AS102* basic sensor and the *AP104* complete sensing module. Both are immune to normal pressure changes, which makes them suitable for applications such as air-control systems, door and window opening monitors and intruder detection. Operating temperature is –10°C to 80°C and sensitivity is 0.1mm of water at frequencies down to 0.1Hz. The sensors are in TO-5 cans and the supply is 2.2-10V, the *104* having extra circuitry to make a complete instrument in an inch square metal case, having analogue and digital output. Anglia. Tel., 01945 474747; fax, 01945 474849.
Enquiry no 539

Protection devices

Connector protector. The *EMC* range of hoods, housings and accessories for components and connectors by Wieland is announced. Hoods and housings for multi-pole connectors are corrosion resistant and afford 360° hf protection, with shielding to limit em emission. Standard multi-pole connector protection is about 40dB at 500MHz, against 80dB for the *EMC* components. Versions in 6, 10, 16, 24 and 40/64 ways are available. Wieland Electric Ltd. Tel., 01483 456262; fax, 01483 505029.
Enquiry no 536

Switches and relays

Power relays. New relays handling 25-125A by Crydom come in two versions: the *CMRD Coolpak*, with a heat sink; or the *CMD Propak*,

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without. The relays are fully enclosed, input and output terminals having captive screws for easy installation and maintenance, a safety cover being provided and the units may be panel or DIN rail mounted. A green led indicates the state of the control signal. Output is from an scr with or without zero-crossing switching and all types are available with 240, 480 and 600V ac operating voltages. Crydom. Tel., 01202 897969; fax, 01202 891918.

Surface-mounted reeds. C P Clare has a new group of reed relays, the *SMR Series*. These are lower-cost alternatives to solid-state and GaAs switches for high frequencies. A new process ensures a better hermetic seal and an assurance of internal continuity in the presence of thermal stress. These relays handle 0-1.5GHz and 150ps rise-time pulses and there is a coaxial shield to match a 50Ω line. C P Clare nv. Tel., 0032 12 672002; fax, 0032 12 672003.

Solid-state relays. Teledyne's military/aerospace and *COTS* solid-state relays for surface mounting are optically isolated and are in various packages: the gull-wing *CA* for ac operation and *CD* for dc; a plastic package to military specification *SR75 COTS DC*; and the *SED* and *SGD* ceramic leadless chip carriers. There are also commercial types in the *C60* for ac and dc and the *C45* for ac. All may be mounted on both sides of a board for space saving and lower costs and all are short-circuit protected with status indication. Teledyne Tel., 01236 453124; fax, 01634 863494.
Enquiry no 537

Small rf relays. Ultraminiature TO-5 relays by Teledyne, the *RF300* and *RF303* series are meant for switching duty at over 3GHz. Rf signal insertion loss over the whole 300kHz-3GHz bandwidth is ±0.1dB or better. The relays are in metal enclosures and there is a ground pin, high control/signal isolation and good resistance to esd. The two types are 7mm and 10mm high respectively. Teledyne Electronic Technologies. Tel., 0181-571 9596; fax, 0181-571 9637.
Enquiry no 538

Transducers and sensors

Medical pressure sensor. Intended for use in blood pressure measurement, in infusion pumps and in dialysis machines, the *NPC-107* disposable pressure sensor is compensated and calibrated to the requirements of the Association for the Advancement of Medical Instrumentation. It is calibrated for the –30 to 300mmHg (–0.58 to 5.8lb/in^2) and gives an output of 100mmHg with no pressure. There is a radiation-tolerant sensor die, compensated for low temperature and coming as an array of 120 sensors. A thick-film resistor network is laser trimmed to provide a linearity of better than 1% over the –15°C to 40°C temperature range. Maximum risk current is 2μA and offset drift 1mmHg over eight hours, maximum. Multytech Ltd. Tel., 01747 870079; fax, 01747 871155.
Enquiry no 541

Microphone Inserts. Versions of the *Magnetone* range of microphone inserts are available in sizes down to 6mm diameter and 3.8mm high. A sub-miniature type (9.7mm by 4.5mm) has a coverage to 16kHz at 40dB s:n ratio, the smaller one covering 30Hz-20kHz. Both offer a sensitivity of 58dB. Various terminations are available to order, but solder-pad types are in stock. Anglia Microwaves Ltd. Tel., 01277 630000; fax, 01277 631111.
Enquiry no 542

Rotary encoders. Grayhill *Series 25L* single-deck mechanical encoders are made with between eight and 36 positions and generate a 2-bit quadrature code, the 16-position version also being available with hex. and Gray code output. The encoders are less than an inch square and various shafts, rotation stops and seals may be specified, as may terminations. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.
Enquiry no 543

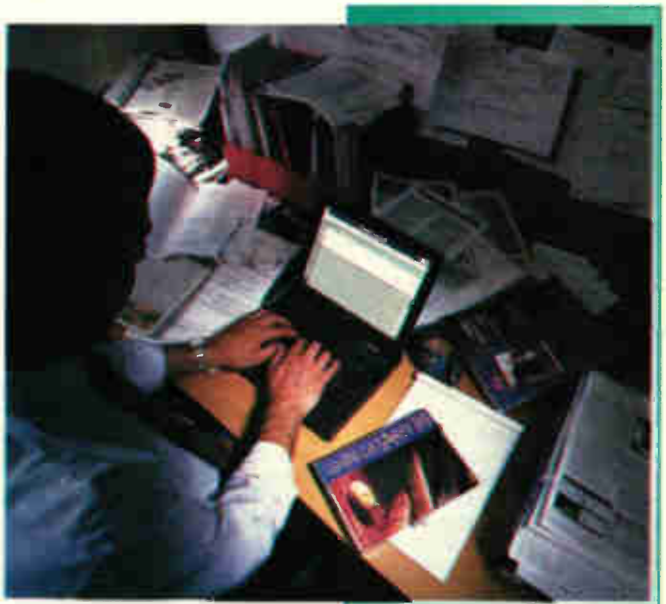
COMPUTER

Computer board-level products

MMX Pentium cpu card. *PCA-6154* from Advantech is the size of an ISA interface card and will take a range of processors including Pentium and MMX, AMD K5/6, Cyrix M1 IDT C6. Being much smaller than usual, this pc motherboard has applications in industrial control and embedded pcs. There is also a 100Base-T-compliant Ethernet adaptor. A socket is provided for an M-Systems *DiskOnChip 200* flash disk of 2-24Mb capacity, with 72Mb soon to arrive. A PCI svga takes a crt and shared system memory of 1-4Mb and two IDE devices may be connected. Power management is on board. Semicom UK Ltd. Tel., 01279 422224; fax, 01279 433339.
Enquiry no 544

Computers

Industrial pc. For use in industry, Blue Chip's *ICON* pc is claimed to be rather more reliable and maintainable than the average computer, being provided with special features with this in mind. There is power rail monitoring that can be fitted with an alarm and warning system and an enhanced systems status display to give cpu diagnostics and internal



temperature monitoring. Hot swap power supplies, accessible filters and two fans with thermistor control further reduce the possibility of serious trouble. Choice of passive backplane is between a 14-slot ISA and ISA/PCI and ATX, providing for the use of 486 or Pentium cpus; alternatively, a segmented backplane can be fitted to take multiple pcs. Two 5.25in and two 3.5in drives can be fitted in a shock-proof cage. The computer is in a stainless steel case, 440mm deep and fitting a 600mm 19in rack, or in a benchtop case. Blue Chip Technology. Tel., 01829 772000; fax, 01829 772001.
Enquiry no 545

Mobile pc. Densitron's *Mobile PC* is exactly that – a single-board pc in the boot of a vehicle and a 10.4in vga tft lcd on the dashboard (passenger side, one hopes). Processing power is a 486 or Pentium and the unit works from 12-24V dc, internal batteries being under constant charge to allow working when away from the vehicle. A 72Mb flash memory looks after the system applications software. Touch-screen operation is provided, so there are no externals, and there are up to four i/o ports, a video card/flat-panel controller, floppy and hard-disk controllers all on the mother board, leaving room for two half ISAbus or standard PC104 expansion cards, with provision for a PCMCIA cardholder. Densitron Europe Ltd., 01959 700100; fax, 01959 700300.
Enquiry no 546

Software

Engineer on a cd. *DAvE* is a digital applications engineer and is of the virtual persuasion, being on a cd-rom. The disk is by Siemens and is meant to assist designers of embedded systems using the company's *C500* and *C166* microcontroller families. He, or rather it, will help an engineer to select the best controller for the job, obtain the relevant configuration and then automatically generate the C code. There are context-relevant links to explanatory information, to user manuals and to function registers by menu-clicking. More such support is to be distributed for new controllers over the web site http://www.smi.siemens.com/DAvE.html, any new files acquired this way working with the cd-rom, even though they will be on a hard disk. Siemens plc. Tel., 01344 396313; fax, 01344 396721. ■
Enquiry no 547

Reference retrieval. *GetARef*, the reference management package by the Swedish company DatAid AB, is now able to gain access to data on the Internet. The system is a Windows-based program that will locate references from various sources, format them and put them into a word processor document, the latest version taking references from other sources including university and public libraries. As well as accessing data from Medline, Silverplatter, CDPlus, Compact Cambridge and Data Star, the system now takes BIDS data from the University of Bath and the dos-based Papyrus, using a new conversion module GarConv, which is available free from http://www.adeptscience.co.uk Adept Scientific Micro Systems Ltd. Tel., 01462 480055; fax, 01462 480213
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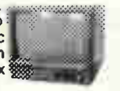

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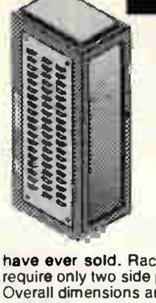

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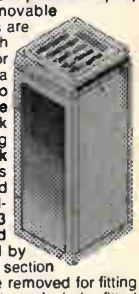

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CIRCLE NO. 135 ON REPLY CARD

LETTERS

Letters to "Electronics World" Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS

Modelling quartz in Spice

This letter is in response to Hands on Internet, February 1998, in which Cyril mentioned Tele Quarz's crystal oscillator design tool.

I still regularly use negative resistance models to establish the activity of quartz crystal maintaining circuits, as described by Tele Quarz. This company's design tool, downloadable from its web site, is particularly valuable for SC cut crystal oscillators, where it is necessary to ensure that only one of two closely spaced modes is activated.

You may find the *Wireless World* article "Quartz crystal oscillator circuit without inductors" interesting. Published in the October 1969 issue, it was written by me, J. Roberts and G. Haynes. It involves the same negative resistance method, without benefit of PSpice, to design a maintaining circuit for a 2.5MHz fifth-overtone crystal. This design provides only sufficient activity to set the required overtone mode into oscillation.

Douglas Dwyer, Frequency Precision Ltd
101505.3427@compuserve.com
Okehampton
Devon

Are *your* ears flat?

Current hi-fi equipment reviews seem to favour systems with a 'warm, rich and detailed' sound. It has been said that the 'warm rich' sound results from a slightly enhanced – but smooth – lower-mid and bass response.

The extent to which the 'warm, rich' sound is liked can be measured by the growing number of new advertisements for speakers, valve amplifiers and valve d-to-a converters specially designed with this attribute in mind.

Ironically the warm, rich qualities are usually appreciated best when listening at low levels – precisely where the human ear starts to become less sensitive at the lower frequencies. This characteristic is a well established fact, illustrated graphically by the famous Fletcher & Munson equal loudness contour curves. These curves show the bass and treble lift required to obtain the same subjective, or perceived, loudness level as that of a given reference level, say, at 1kHz.

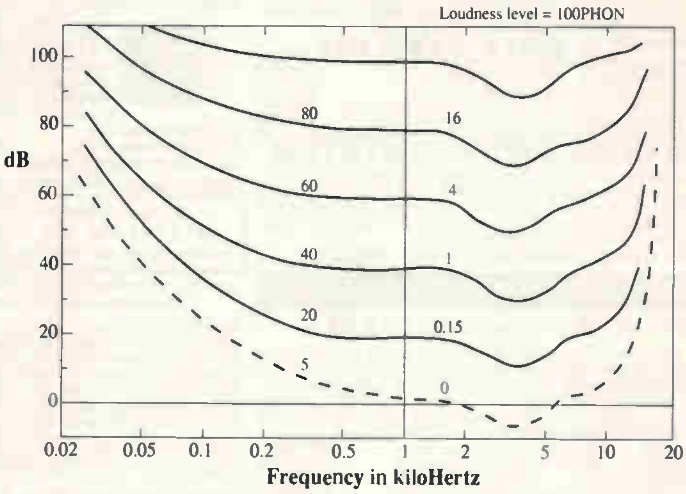
For example, at a sound-pressure level of 60dB, around +10dB of bass lift is required at 100Hz to produce the same subjective level as at 1kHz.

At higher pressure levels, for example 90 to 100dB, very little or no bass lift is needed, confirming the level dependent characteristic of our hearing. In other words, our hearing becomes 'flatter' at higher sound levels.

Amplifier designs of the late fifties and sixties almost always included a loudness control to compensate for the ear's non-linearities. Why is it no longer considered a useful addition? A loudness-override switch would keep those with 'flat ears' happy.

Could it be that some manufacturers already provide the required lift, mainly in d-to-a converters, on the quiet to achieve the sound qualities that earns them the best-buys and four/five star votes in the reviews?

George Evans
Lymington
Hants



Fletcher-Munson perceived loudness curves show that you need more than just a perfectly linear audio system to experience faithful reproduction at all listening levels.

Confused? Read this...

In the letters section of *Electronics World*, February 1998, Malcolm Bloor stresses the need to fastidiously check code, which I totally agree with. However, he also suggested that assemblers can kill: this is worrying since the assembler he trusts is at fault.

Depending on how the Small Assembler describes its instructions' semantics, it would appear that the issue of the 80x86 instruction format is causing confusion. Let me explain.

The comparison instruction for the AX register will always take the form '0x3D,0xNN, 0xMM'. Here, NN and MM are the two hexadecimal digits of the 16bit numbers being compared with AX.

However, for the BX, CX and DX registers, the normal format of the instruction is a bit more complicated. It is a four byte string and sometimes three bytes. It is four bytes when the instruction is comparing 0xFF7F *(–0x81)* to 0x0080 *(+0x80)* and three bytes when it is comparing in the range 0xFFFF (–0x01) to 0xFF80 (–0x80) AND 0x007F (+0x7F) to 0x0000 (+0x00).

In other words the most-significant bit, bit 16, indicates the sign. If the number is between –128 and +127 then the instruction is a three byte one, possibly due to Intel wanting to use less program space on frequent comparisons. Let me demonstrate.

```
mov   AX,   0x1234   gives   0x3D   0x34   0x12
mov   AX,   0xFF34   gives   0x3D   0x34   0xFF

mov   BX,   0x1234   gives   0x81   0xFB   0x34   0x12
mov   BX,   0xFF34   gives   0x81   0xFB   0x34   0xFF
mov   BX,   0xFF81   gives   0x83   0xFB   0x81

mov   BX,   0x0062   gives   0x83   0xFB   0x62
mov   BX,   0x0090   gives   0x81   0xFB   0x90   0x00

mov   BX,   0x62     gives   0x83   0xFB   0x62
mov   BX,   0x90     gives   0x81   0xFB   0x90   0x00
```

In the case of the Small Assembler, it produces '0x83, 0xFB, 0xFF' for a 'CMP BX, 0xFF' where it should actually be producing '0x81, 0xFB, 0xFF, 0x00' (see line 29 of the SMALT.ASM listing from the original letter). Also in error is line 28 of the SMALT.ASM listing for similar reasons. MASM, DEBUG, and the like all produce the correct results, although the following are equivalent and would work just as well.

```
cmp   BX,   0x7f     gives   0x83   0xFB   0x7F
cmp   BX,   0x7f     gives   0x81   0xFB   0x7f   0x00

cmp   BX,   0xff80   gives   0x83   0xFB   0x80
cmp   BX,   0xff80   gives   0x81   0xFB   0x80   0xFF
```

I hope that has cleared things up. If anyone is interested there is more information like this, and also about electronics and obsolete computer systems on my web site at http://www.tardis.ed.ac.uk/~lucien.

Lucien Murray-Pitts
Edinburgh

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RFI sources and solutions

In today's crowded radio spectrum, it is all too common for radio users to experience interference, the causes of which can be complex. Alan Wood details potential sources of interference in telemetry applications, and suggests possible solutions.

Interference on radio links falls into three categories. There is interference internal to the radio and associated electronics, interference classed as 'in-band' and interference that is described as being 'out of band'.

Reference timing clocks in associated equipment are a source of internal interference. Modern electronic equipment invariably contains a clock for timing purposes. Quite often such clocks are derived from a high-frequency crystal that is divided down to match the lower frequency of the circuitry in use.

It is possible for higher multiples or harmonics of the clock source crystal to fall on the exact frequency that a radio link is attempting to operate on.

Modern high-quality radios use a heterodyne or mixing technique to convert the high frequency uhf or vhf signals to a low frequency that can be filtered and processed more readily. Typically this will be a conversion to 45MHz, 21.4MHz or 10.7MHz, followed by a second conversion to 450kHz, 455kHz, 470kHz or similar. A clock sub-multiple at any of these frequencies will cause problems for the radio user.

Radio-related sources

In its conversion processes, a radio receiver needs a locally generated frequency source to mix with the incoming signal. The difference between the frequencies of the incoming rf signal and the local oscillator is the intermediate frequency and, as detailed above, is usually a commonly accepted frequency. The second conversion process is likewise.

For each conversion, the local oscillator signal could be a source of self-interference. This occurs when, perhaps, a multiple of the second local-oscillator signal falls on the frequency being listened to.

For example, 20.945MHz is often used in receivers to convert 21.4MHz to 455kHz. The 21st harmonic of 20.945MHz is 439.845MHz and therefore could cause a problem if this is the frequency being used.

At my company, we use test procedures to bring such problems to light and cure them before any product is issued.

Power-supply ripple

Modern equipment often has a switch-mode power supply to ensure maximum power efficiency. Switch-mode power supplies are electronic chopper circuits that create an ac waveform from a dc input to allow it to be transformed to a new potential where it can be rectified and smoothed back to dc.

This process causes switching noise at the chopper frequency. A great deal of care is needed to ensure that this noise is fully filtered and suppressed from the power supply lines to the radio module. Also, careful choice of switching frequency is needed.

If, for instance, the switching frequency was 25kHz and this were to pass through into the transmitter, it could cause spurious sideband signals at 25kHz intervals around the main carrier output. Any receiver listening to adjacent channels would hear these as interference signals. Likewise, a 25kHz ripple on the receiver local-oscillator signal would create a whole array of mini local-oscillators on the adjacent channel frequencies, causing overlaying of many channels onto the wanted one.

Careful filtering of the supply to the modules is therefore essential. All Wood & Douglas modules accept an unstabilised supply since they incorporate stabilisation and filtering. However, it is difficult to ensure total immunity to all possible frequencies that could be carried on the power supply wiring.

In-band interference

In-band interference can be defined as problems due to signals within the rf pass-band of the module in question.

On-channel interference is the worst and most obvious form of interference. It occurs when one user has chosen exactly the same frequency as another user. It can be most frustrating to verify this type of interference if the other signal is intermittent, or if it has a long duty cycle and leads to data loss on the required link for no apparent reason.

A site radio survey is needed to identify the existence of this type of interference, to locate its source and, if necessary, its legality. Data corruptions can then be logged and plotted to determine when and how often the problems occur. Is it only during working hours? Is it at the same time each day or hour, etc?

Having traced the source and its legality, there may be potential to negotiate with the other user to modify their power output levels or the antenna configuration to reduce the impact on your service. It may also be possible to improve your receiving site's immunity to the interference by use of directional antennas, or by changing from vertically polarised antennas to horizontal or vice versa.

Remember, though, that this type of signal is always going to be there and is not a characteristic of the equipment being used. Of course, the technical specification of the radio may be too good and cause it to be too sensitive. Attenuation in the antenna lead, which would reduce sensitivity and transmitted power, might help.

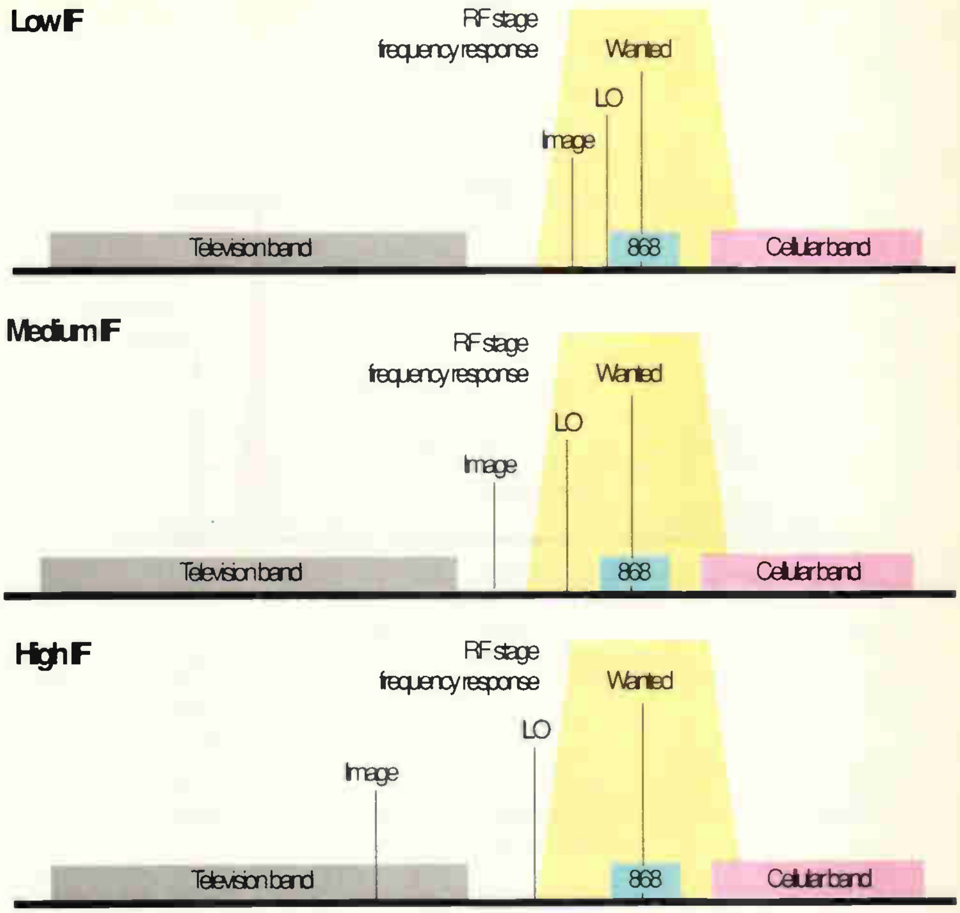
Radio telemetry modules that we make are capable of letting you hear what is on the chosen frequency by simply connecting to the recovered audio output with a high impedance ear piece or small speaker driven amplifier. This is a simple method of monitoring the frequency but very effective.

Adjacent-channel interference. The vhf and uhf radio spectrum is divided up into channels spaced either 12.5kHz or 25kHz apart. The channel either side of the one you have chosen is the adjacent channel.

The ability of your receiver to only listen to the signal on the required frequency, while ignoring any signals on the adjacent channels and beyond, is called the selectivity of the receiver, or adjacent channel rejection.

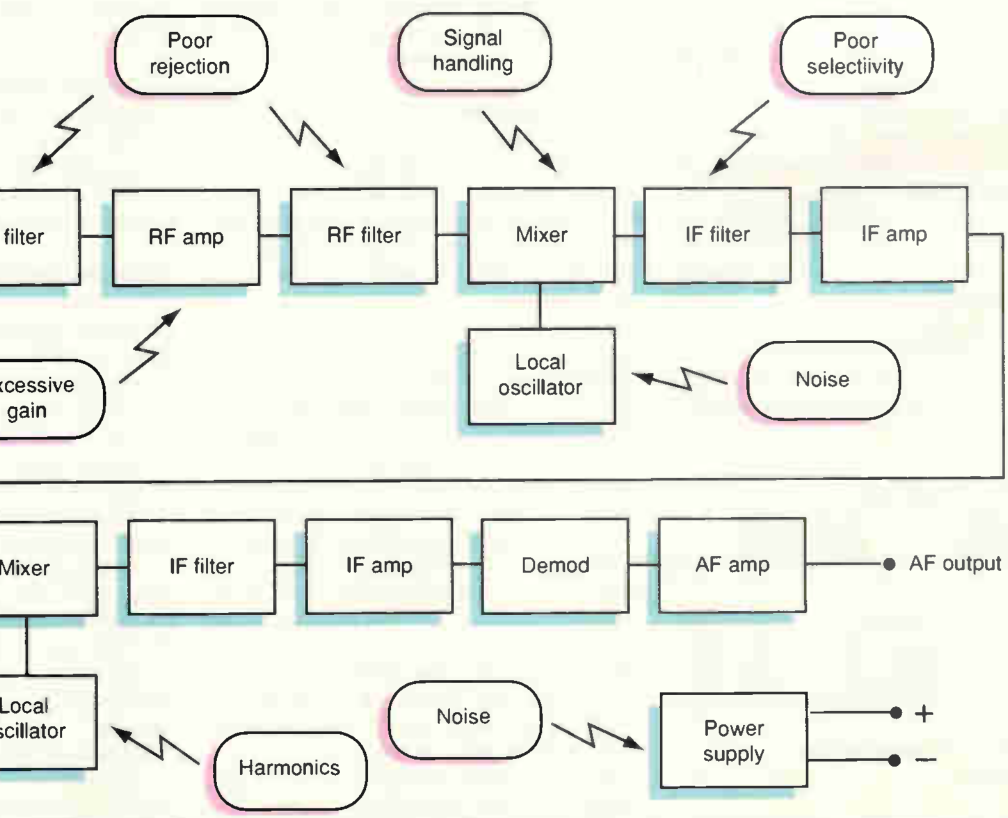
The ability of the receiver to discern only the channel you wish is a reflection of the filtering of the first and second intermediate conversion frequencies. More filtering will ensure better rejection of signals on adjacent channels and typically, figures of 60dB should be sought to ensure relatively trouble-free operation.

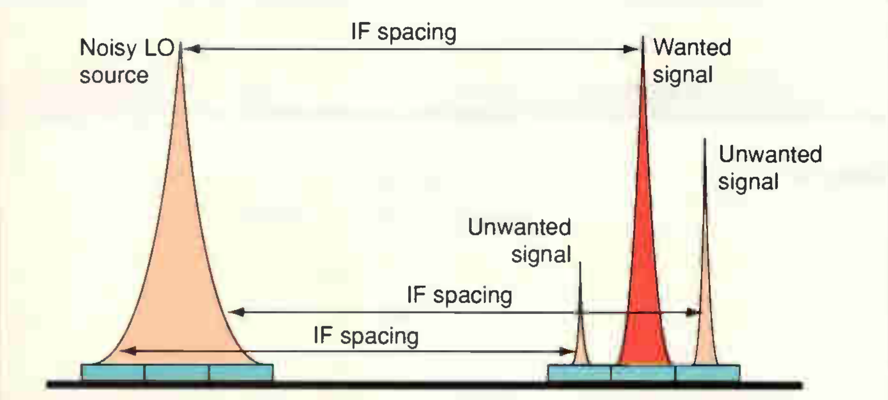
What does this mean in practice? It simply means that a signal on the adjacent channel has to be one million times bigger than the



Impact of intermediate frequency and front-end filtering choices.

Receiver block diagram showing possible areas of interference.





Magnified view of reciprocal mixing effect.

signal you are trying to receive, to be heard at the same level. This sounds impressive but such variations in levels are possible and do occur.

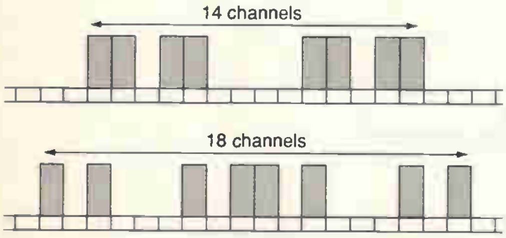
Reciprocal mixing

A more complex cause of adjacent-channel interference can come about due to reciprocal mixing, which is more common in poorly designed synthesised receivers.

Reciprocal mixing can occur when a poorly designed synthesised local oscillator in a receiver doesn't inject a pure signal into the conversion mixer. Imagine it as a signal with a host of noise sidebands gradually diminishing in amplitude as you move away from the centre frequency.

Mathematically, if any of this energy spreads over into the adjacent channels and beyond, it will provide a 'local oscillator' of noise, for any signals on the adjacent channels. This will mix and create a down conversion to the intermediate frequency in use, just as the required signal does.

Because this is an 'accepted' signal to the receiver due to its mathematical relationship of rf to local-oscillator to intermediate-frequency conversion, no amount of IF filtering will make any difference. It's a real signal as far as the receiver IF is concerned and will have an impact in relation to the relative amplitude of wanted to unwanted local-oscillator energy.



Third-order intermodulation-free channel options.

This is an important factor to check with the receiver supplier. Ask what the adjacent power of the receiver oscillator is. It should be specified for a transmitter but rarely is for a receiver. A good figure is 70dB but 80dB is ideal. Figures below this will falsify the adjacent channel rejection figure based purely on poles of IF filtering.

It is worth noting that the accurate measurement of adjacent channel rejection is also subject to how noisy the signal generator being used for the measurement is.

Of course, paying a little more for a quality receiver will help to avoid this type of problem. Very low cost receivers, which have little adjacent channel filtering, could admit the entire band at once if sufficiently strong signals were on every channel.

Intermodulation

Intermodulation can be an in-band or out of band effect. If two signals get mixed together in an overloaded circuit such as a receiver front end, a mathematical combination of spurious frequencies are generated. These are calculated from $2F_A \pm F_B$ where F_A and F_B are the frequencies.

The '+' part is not such a problem as it creates a signal 'miles away' in frequency terms. But the '–' term could be. Suppose you have a signal on 458.650MHz and another on 458.700MHz, this would create a signal on 458.600MHz and another on 458.750. These are real signals. Admittedly they are at a low level, but they could cause a problem for receivers trying to listen on these frequencies.

The effect is obvious when a receiver listening to its required signal has a very strong signal also coming down its antenna lead on another frequency and a secondary signal at another frequency. The combination of these two signals gets churned up in the receiver front end and the product is a signal 'phantom' on the required channel.

Such a situation can arise on a multi-transmitter site where one transmitter is very close to the receiver, causing an overload to occur. The effects are made worse if there are lots of overloading signals. Overloading can be avoided in the following ways:

- Careful choice of frequencies. An optimum intermod bandplan should be considered;
- Careful choice of transmitter/receiver siting and use of the minimum power necessary for error free operation helps avoid overload;
- Careful choice of the receiver module is a major factor.

For a receiver to be able to cope with very large signals in a relatively intermod/overload free manner, the front end design needs attention. The filter on the receiver front end is always more than one channel wide and so will allow lots of signals, including the signal you want, through to the first rf amplifier stage.

To avoid the array of inputs causing the receiver to go into overload, the front end device, and subsequent rf stages, will generally need a significant amount of supply current passed through it.

Therefore, a radio receiver designed for low current operation won't be as effective on strong signal handling and intermodulation performance. This is important to note if you have crowded radio spectrum on site. Typically 5-10mA will be needed in each rf stage to ensure relatively trouble free operation.

Check the manufacturer's total current consumption and ask what the intermodulation specification is; some receivers offered on the market consume only 5mA in total. A figure of 50dB minimum is ideal for a really good product, a figure of 65dB is typical.

Desensitisation. Desensitisation is a dynamic effect caused by strong signals in band, rather than on the operating frequency. These signals totally overpower the front end of the receiver causing it to partially or fully shut down. This leads to a loss in sensitivity and a reduction in the strength of the signals that the receiver is trying to hear. It can be heard as sudden 'silence' on the receiver audio.

An example of this is when MPT1329 telemetry products operating at 458MHz are used by utilities on prime sites which are also using MPT1411 equipment. MPT1411 frequencies are duplex and straddle 458MHz at 457MHz and 463MHz. As up to 10W could be radiated from an MPT1411 device, even a well designed, local MPT1329 receiver would struggle.

The solution, once again, is careful design of the receiver to ensure that the larger signal handling parameters, such as blocking, have been addressed. Again, a very low current receiver will not perform as well as one with a robust front end.

Ask your supplier what the blocking specification is and then decide if it is appropriate for the site being considered. A typical figure would be 80-100dB.

Out-of-band effects

In-band effects are defined as occurring within the input acceptance bandwidth of the receiver. Signals outside this band will be attenuated or rejected by the front end filter components.

If the signals outside the front-end 'window' are significantly large to exceed the attenuation due to the filtering circuitry then they will be just as likely to cause in-band type interference.

In-band and out of band effects tend to overlap somewhat; both depend on signal level and receiver characteristics but in general, out of band effects are accepted as being outside the tuned bandwidth of the receiver front end.

Image rejection. The mixing process in a superheterodyne receiver takes the incoming radio signal and mixes it with a locally generated signal derived from a crystal or synthesised source within the receiver module. The difference frequency between the two is then processed to recover the information being carried by the signal. This difference signal is usually chosen from a long list of standard frequencies of which 45MHz, 21.4MHz and 10.7MHz are the most common.

If an incoming radio signal is on 455MHz and you need a 45MHz intermediate frequency, then 410MHz must be generated locally to create a 45MHz difference.

Alternatively, you could generate 500MHz and still create the required difference. These two possible injection frequencies from the local oscillator give the same result; one is a 'low side mix' and one is a 'high side mix'.

Equally, for any one injection frequency there will also be a second radio frequency to give a 45MHz intermediate frequency. In this example, while 455MHz mixed with 410MHz gave 45MHz, so does 365MHz mixed with 410MHz. This second frequency, which is acceptable to the IF, is called the image frequency.

If the receiver front end filter does not reject any signals on this image frequency, they will pass through and be just as acceptable to the IF processing electronics as the required signals, causing a serious problem.

The image-rejection parameter is specified on radio data sheets and you should look for a figure of 60dB or more. A manufacturer usually achieves this figure by using a high first intermediate frequency. This simplifies the front-end or rf filtering needed.

This is a brief outline of all the more obvious mechanisms for interference that can occur with modern superheterodyne receiver modules and transmitters. The problems of spurious responses in receivers are made worse with crystal derived local oscillators as there is more potential for other frequencies to be present at the mixer of the receiver. Modern synthesised receivers have a local oscillator generated at the final injection frequency and are therefore cleaner.

In summary

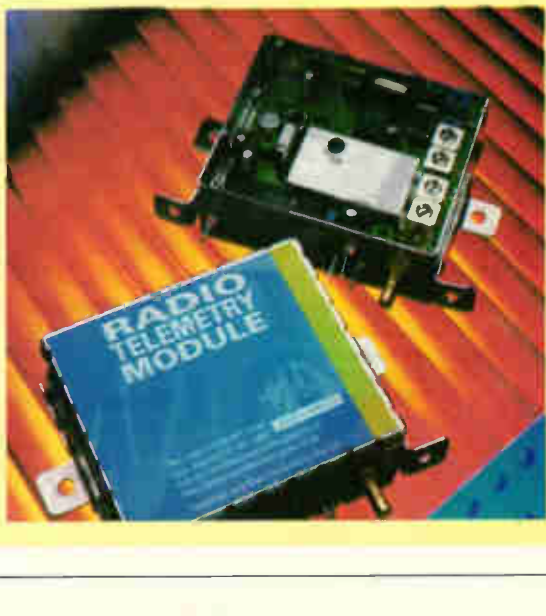
If you want a robust, reliable device to withstand today's demanding telemetry bands, then you must pay attention to and ask questions about those confusing spec sheet parameters such as intermodulation, blocking, desense, image rejection and adjacent channel rejection. The success and reliability of your radio link may depend on it. But so will the price. ■

Technical support

Based in Baughurst, near Tadley, Wood & Douglas is a radio telemetry design and manufacture specialist for the low power radio market. The company provides technical support at both design and after-sales stages. Tel 0118 981 1444, fax: 0118 981 1567. Also e-mail, to info@woodanddouglas.co.uk or see our web site http://www.woodanddouglas.co.uk+info@woodanddouglas.co.uk

Telemetry module examples

Capable of working in a crowded spectrum, this 10-100mW synthesised telemetry transmitter can be set to a new frequency simply by sending a programming sequence down a serial or parallel interface. The module is approved to ETSI 300 683 and MPT1361. It operates on the UHF 430-470MHz band and has a switching bandwidth of 5MHz. Channel spacings are 12.5, 20 or 25kHz.



The main feature of this telemetry receiver is that it provides programmable operating frequency with the performance of a fixed-frequency crystal alternative. It includes an fsk data extractor and has either modem or controller options. Operating frequency is 450 to 470MHz, programmable either by replacing the PIC microcontroller or by downloading serial data from a pc.

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This client, is an internationally renowned consultancy, specialising in communications, especially mobile comms. Continued expansion requires them to recruit a number of principal radio systems engineers (Radio Architecture) to provide a consultancy service to clients throughout the world. You must have a minimum of eight years experience of RF circuit design and/or radio architecture design. You should have experience of GSM or analogue cellular terminal or base station design and have expertise in synthesiser, PLL, LNA, Mixer, Filter, PA, Tx/Rx etc design. Alternatively, you may have developed the above skills in other areas of mobile radio, including DECT, TETRA, PMR, CT2 or similar. It will be an advantage if you also have team leading experience.

ASIC DESIGN ENGINEERS — to £40,000+
Mobile Radio — E. Anglia

This client, is an established, highly successful, design house and consultancy, specialising in innovative and complex solutions for mobile comms products. They are retained by many leading companies throughout the world for their pioneering and creative designs. Continuing success has led to new opportunities within the company and they are now looking to recruit a number of key individuals, at various levels, to combine technical challenge with the opportunity to work directly with clients on a wide variety of products. In particular, the company is seeking a number of ASIC Design Engineers. Candidates should have experience in the design and implementation of Digital, Analogue and/or Mixed Signal ASICs.

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A major international client involved in developing GSM, Dual Band and Satellite phones. They are seeking DSP Software Engineers to design, implement and integrate DSP software. Excellent career opportunities with first class packages. Ideally, you will have at least one year's experience of developing real time embedded DSP software, as well as experience of developing software in a communications environment. This company offers the ideal environment for creative thinking and innovative design.

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Mobile Radio Solutions — Hants

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SOFTWARE ENGINEERS — to £45,000+
Mobile Radio — Various M3/M4

A wide range of opportunities, all all levels with a variety of clients involved in cellular, satellite and cordless phone development. You should have an appropriate level of experience of developing embedded "C" and targeting 8/16 bit processors.

CDMA At the Cutting Edge of Mobile Communications..

Based in Surrey, within easy reach of a number of motorways, our client is the European HQ and the premier design centre of a major international manufacturer of mobile and cordless telephones. At the cutting edge of communications technology, they produce a wide range of cellular products, including GSM, NMT, ETACS, AMPS and CDMA phones. In this last area, they are undoubted world leaders and are currently the only company with CDMA phones in volume production.

With several exciting new projects in the CDMA arena, our client is planning to significantly grow the development team and has asked Lateral Connections to locate the key staff they are seeking.

PROJECT MANAGER **to £40,000 + Car + Bonus**

You should have a strong technical background with product development experience gained in RF, Baseband or Software. Ideally some of this experience will have been gained in a high volume mobile comms environment. However exceptional candidates will be considered from other backgrounds.

SOFTWARE ENGINEERS **to £38,000 + Bonus**

Many of the software functions are common to more than one product and the Software Group is organised mainly in functional rather than project based team. Movement between teams is encouraged and therefore software engineers are frequently exposed to different products and areas of work. Currently the Group is organised into the following teams: **Call Processing, Data Services, MMI & Drivers, Analogue & Cordless Handsets and Software Tools.** Opportunities exist in all of these teams and our client would like to consider applications at all level from Engineer to Team Leader.

A background in mobile radio or telecomms is desirable, but not necessary. However, you should have a minimum of 12 months experience of developing real-time embedded software using "C", with experience of integrating and debugging software on the target hardware. Additionally, you most be self-motivated, be able to work as an individual and as part of a team. You must also have a flexible attitude and be willing to undertake some foreign travel

MECHANICAL DESIGN ENGRS **to £30,000 + Bonus**

Two excellent opportunities for Mechanical Design Engineers at either Engineer or Senior Engineer level. Your brief will be to design and development all mechanical aspects of cellular handsets and cordless phones. It is therefore essential that you have experience of designing for high volume in some or all of the following areas: Electronic Packaging, Plastics, Injection Mouldings, Extrusions, Small Pressed Metal Parts, etc. It would also be useful to have an understanding of radio issues, namely screening.

PCB DESIGNER **to £28,000 + Bonus**

Your opportunity to design PCBs for market leading cellular and cordless phones. You should have experience of designing high density, multi-layer, surface mount PCBs, ideally for high volume consumer products. An understanding of laying out RF boards would be a distinct advantage, as would PADS or AutoCad experience.

RF DESIGN ENGINEERS **to £38,000 + Bonus**

A number of requirements, from graduate to team leader level. Preferably, you will be degree qualified, majoring in RF and Communications during your final year. Depending on your seniority, you will have an appropriate level of experience in designing low power, low cost, high performance RF circuitry up to 2GHz, for commercial, high volume applications. Ideally, you should also be fluent in the use of Smith Chart and familiar with at least one RF CAD package such as Super Compact. Team Leader applicants should also have man management skills and experience of leading a small team, as well as the potential for future development and promotion in

HARDWARE ENGINEERS **to £38.000 + Bonus**

The Hardware Group is responsible for the hardware baseband design of handsets and accessories, the design, modification and de-bug of low-level driver software and the integration of these designs with handset software and RF design. Our client is keen to consider applications from Hardware Engineers at all levels to join this Group. A background in mobile radio or telecomms would be useful, but is not necessary. However, you must have an appropriate level of expertise in embedded 8/16-bit processor development, digital design techniques and working knowledge of "C" and Assembler, ideally gained in a medium to high volume environment

If you are interested in finding out more about our client, please mail, fax or e-mail your CV to Lateral Connections at the address below, quoting Ref: LC/EWW 3745.

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To find out more about these opportunities, please telephone quoting reference number on 01273 820911 until 7.00pm this week or next. Alternatively, post / fax / Email your CV to:

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Ref: ML/10

HARDWARE DESIGN

UK/Europe Excellent Package

A background in digital design will enable you to join this dynamic organisation working on leading edge electronics hardware and software technology for embedded processor design. You will have between 1 to 10 years digital design experience using microcontrollers, FPGA design, ideally knowledge of high level system design tools and be educated to degree level.

Ref: EM/11

SOFTWARE

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Ref: AW/11

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Europe/USA Excellent Package

As a result of significant investment in Europe, a brand new design centre dedicated to microcontrollers has been opened by one of the world leaders in electronics. They now seek experienced design engineers to work on next generation microcontrollers. You will be able to contribute from the outset to the success of future products. Opportunities also exist within Europe/USA.

Ref: ML/11

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International, young rapidly expanding telecoms company needs experienced software developer to lead a team of software engineers to design network planning tool software. Opportunity to work with market leaders across the world. UNIX/C, X Windows, Embedded SQL essential, GSM and team management experience preferred.

Ref: AR/11

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Ref: AR/10

RADIO PLANNING LEADER

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Ref: AW/12

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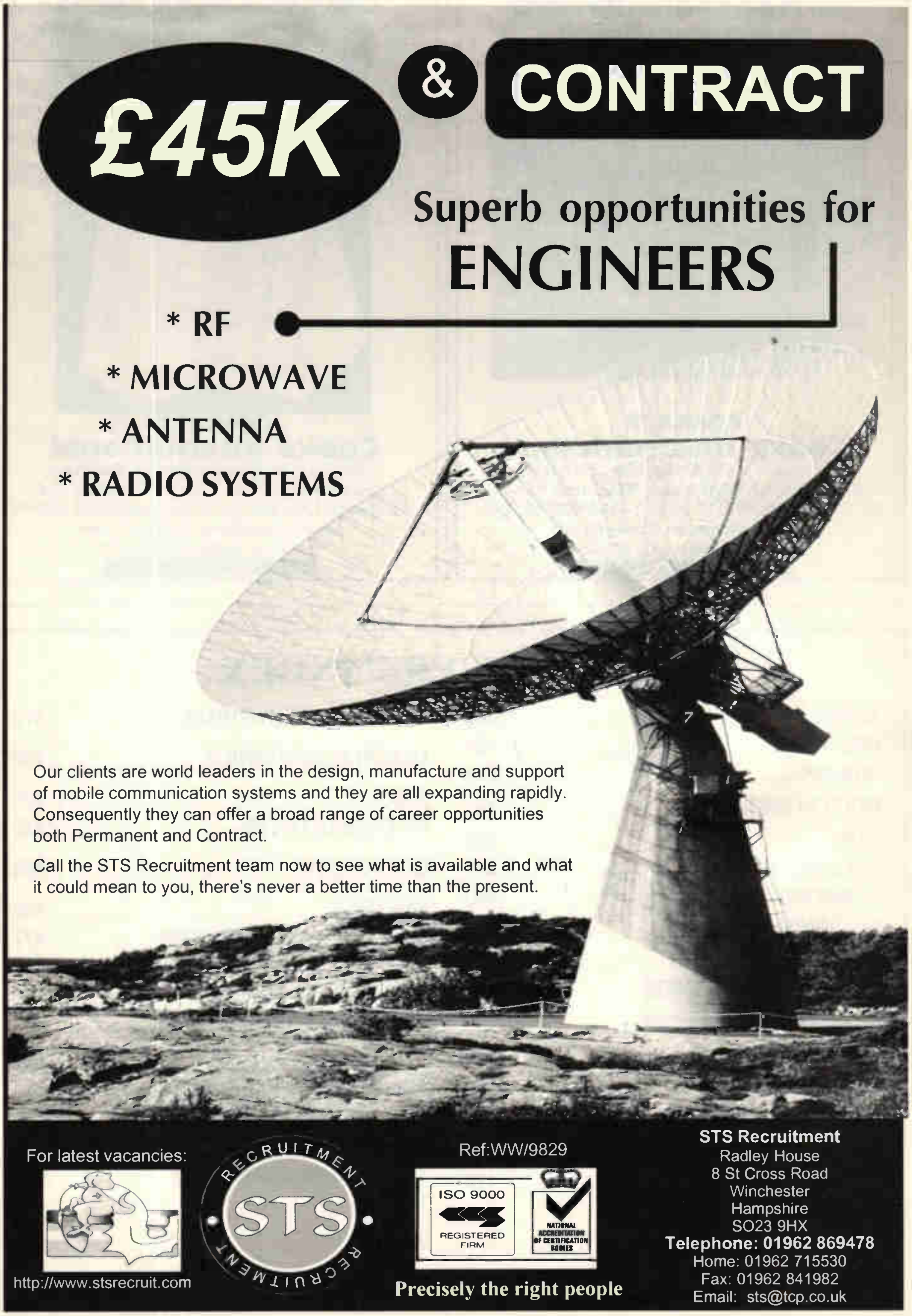
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Modes	AM,SSB/CW,FM-N,FM-W	AM,LSB,USB,CW,FM-N,FM-W
Tuning step size	100 Hz (5 Hz BFO)	100 Hz (10 Hz for SSB and CW)
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Receiver type	PLL-based triple-conv. superhet	
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Max on one motherboard	8 cards	8 cards
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