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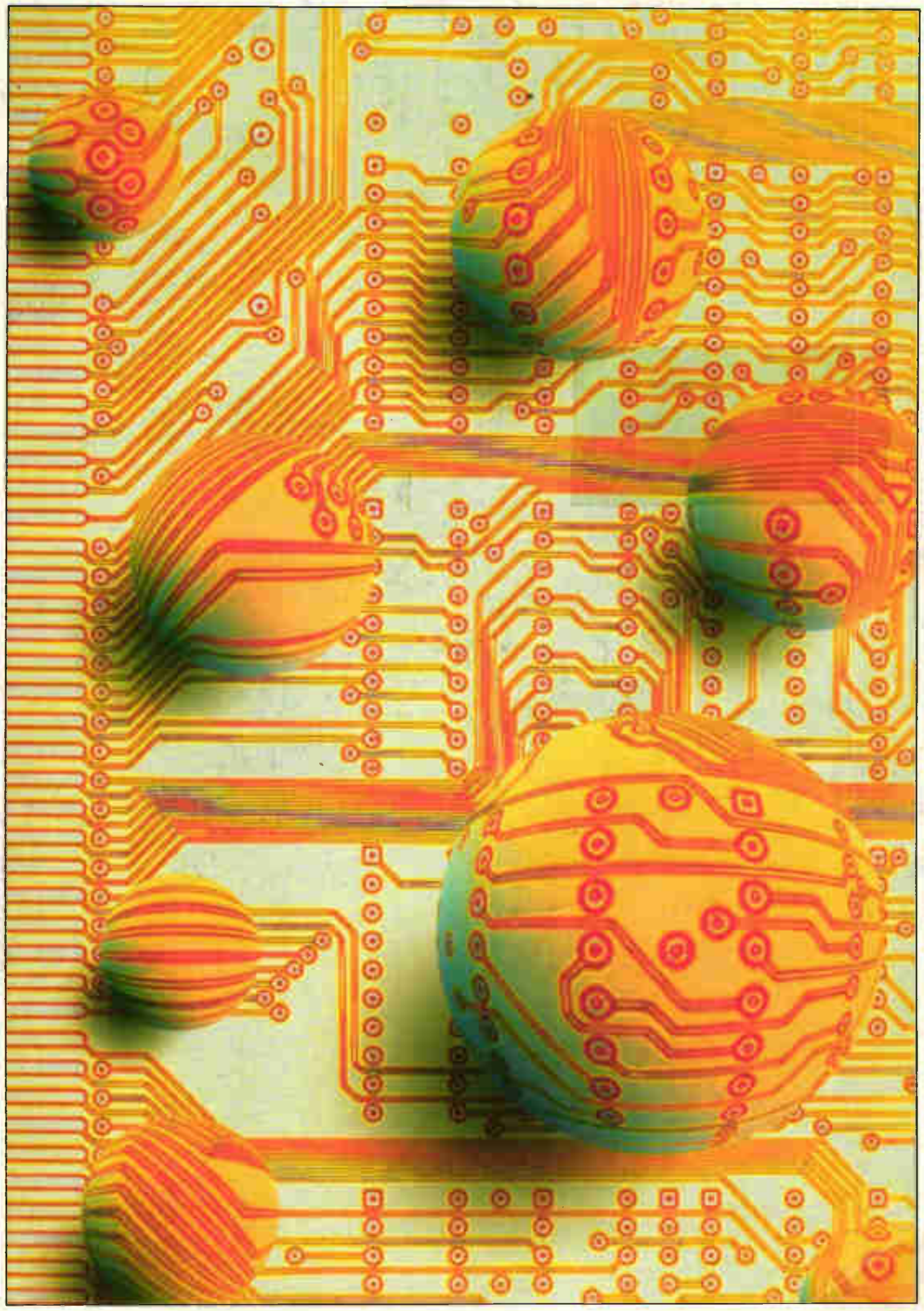

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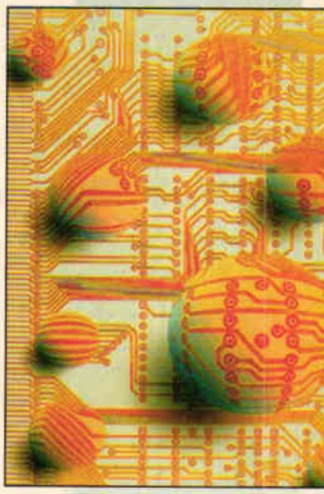

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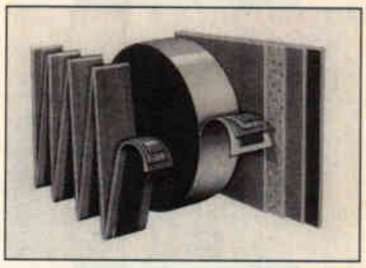
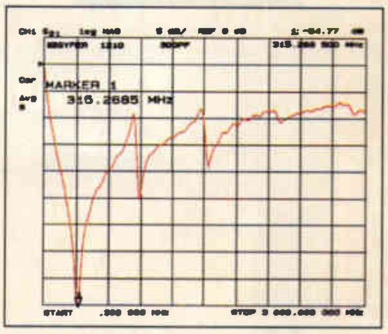
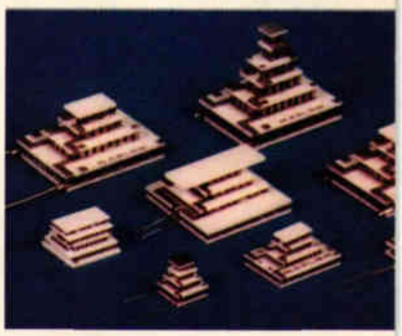


Polymer cells like these could replace you car battery, fitting between the roof and the head liner: page 302.



A ceramic capacitor's multiple high-frequency resonances – find out more about them on page 324.



Piling Peltier coolers one on top of the other increase the hot-to-cold face difference. Cyril tells where to find more on this and other Internet topics: See page 274.

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REED BUSINESS INFORMATION

Phone Bill

There is more than a touch of a John Ford western in the latest move by Intel, Compaq and their chums. The PC firms are usually willing to present themselves as the cavalry saving the day in the nick of time.

The target now is the home PC user and their need to communicate with the rest of the globe over the World Wide Web. Or more particularly the technical means of transmitting the equivalent of between one and half to eight million bits of data a second down one of BT's telephone lines.

Not that BT is a specific target – unless of course you believe the rumours that Microsoft may have designs on the UK telephone operator. The aim is to kick into life a telecommunications market which could present attractive product opportunities for the PC manufacturers.

Oh, blatant commercial self-interest I hear you say. When was it otherwise? But this time Intel *et al* may have a point.

Surprise, surprise. None other than BT – "It's good to talk" – has swallowed its telephone technology pride and joined the telecoms initiative masterminded by PC technology firms like Microsoft, Compaq and Intel. Soon the Phone Shop will be selling PCs and you'll be able to pay your phone bill at PC World.

This is heady stuff. BT has joined the ADSL working group, made up almost entirely of US firms, which is working to create an international technical standard which will provide1.5Mbit/s down phone lines to PCs and Internet users in the home.

BT is already planning the introduction of ADSL technology in some of its exchanges on a trial basis. It is keen to ensure compatibility of its network technology with anything being proposed by the PC firms who look set for the first time to dictate a high speed telecoms interface standard. What is being proposed is over 20 times faster than the ISDN digital telephone lines currently available over BT's network.

The means of transmitting 1.5Mbit/s of data down an analogue telephone line commonly known as ADSL, is nothing new. BT tested its first ADSL hardware as long a go as 1992. Semiconductor firms have been selling the necessary ADSL chipsets for almost as long and still we are yet to see the interactive TV service.

Video to the home over the telephone line has proved too expensive and technically troublesome to introduce in more than just a handful of prototype trials.

As one customer service engineer at BT said recently when asked how long it would take to obtain an ADSL line in Birmingham; "Three to five years, I should think."

The telecoms firms have flapped around with ADSL for long enough. The PC market is looking for broadband communications direct to the home user and so the PC firms have decided that the time is right to step in and make ADSL happen once and for all.

If the telephone companies can benefit from the results, then fine. If some of the telecoms hardware companies find themselves losing out to bigger and smarter PC firms, well tough. The telecoms guys have had five years and plenty of opportunity to make ADSL fly, they cannot start crying when Intel and its PC pals move in on their patch because they can do a better job.

Bill Gates is a canny enough operator to realise that there may be some, perhaps as yet indefinable, business advantage to be gained from linking Microsoft's empire commercially with a phone company.

And the PC boys are serious.

Take the throw away comment of Microsoft boss Bill Gates given quite literally while he was braving custard tarts in Brussels. Gates may have stated categorically that he had held no discussions about buying a phone company. However, such a comment conceals a little more than it says.

For one, why was someone even asking the question about Microsoft's interest in BT in the first place? Denying talks is one thing, but it is a very long way from denying any interest in said phone company.

Bill Gates is a canny enough operator to realise that there may be some, perhaps as yet indefinable, business advantage to be gained from linking Microsoft's empire commercially with a phone company.

It seems unlikely that the well-fed bosses of Europe's monopolistic telephone companies have resorted to throwing custard pies just yet. However, there can be little doubt now that what lies ahead in the battle for high speed data communications to PCs in the home and at work is a shoot-out of John Ford proportions between the PC world and BT with its fellow telephone companies.

The cavalry may have appeared on the horizon, but that battle, if already begun, has some way to go. But custard pie makers should be warned the fight could still get just a little rough. After all there are billions of pounds worth of our phone bills at stake. ■

Richard Wilson

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UP DATE

US firm buys Marconi Instruments

Marconi Instruments (MI) has been sold to IFR Systems of Kansas for £65m. The sale includes the MICAS test services business. The Marconi name will be dropped, but the management team will stay in place and managing director Peter Smith says there will be no redundancies.

"This transaction doubles our size," said IFR's founder, chairman and CEO, Alfred H Hunt III, "it's a once in a lifetime opportunity for IFR - it's quite remarkable that two companies like this have the opportunity to merge on this basis. It's exciting for both sides."

Last year, the 29 year-old company had revenues of $103m for profits of $6.7m. The combined company will have annual sales of $230m making it one of the world's top eight test companies. Hunt said that MI had similar profit margins to IFR. He saw cost savings in advertising, trade show costs and from the combined group's increased purchasing power for components.

Asked by EW if the takeover would mean redundancies in the UK, Smith replied: "With two substantially growing businesses, the problem is recruiting people. We need more people." In the last five years IFR has grown from $39m to $103m.

IFR has two major divisions: the RF division and the fibre-optics division.

The fibre-optics division, based in Oregon with a UK operation at Chandlers Ford in Hampshire, makes test equipment for optical fibre and component manufacturing, for telecommunications, and for r&d purposes. IFR also recently acquired Romsey-based York Sensors.

MI will be merged into the rf division which had $69m in sales last year . It makes test instruments for avionics, wireless communication systems and for general test applications.

MI staff will be relieved to hear that, in 1995, IFR Systems won a trophy from a local tv station for being "the best place to work in Kansas," but the Stevenage dress code may change. Hunt favours blue jeans and cowboy boots.

Here at last – a 56k modem standard

A single 56kbit/s modem technology standard has been finally ratified by the ITU effectively unifying the incompatible K56flex and X2 rival modem technologies.

The standard, called V.90, is being supported by all the major modem manufacturers and the first V.90 modems are expected on store shelves by March. Existing modem users will also be able to upgrade their modems to V.90 through a free software upgrade.

Modem companies are rushing to test their modems against each other to make sure that there are no compatibility problems.

"Lucent has been conducting some of the earliest interoperability tests of these high speed modems, and we've made considerable progress toward ensuring that modems based on our chips can communicate with those based on other chips. This interoperability testing is key to guaranteeing that products are truly in compliance with any new standard," said Craig Garen, general manager of modem integrated circuits for Lucent's Microelectronics Group.

Modem chip companies and modem vendors hope that the V.90 standard will help boost modem sales which have slowed over the past few months as customers postponed purchases until the single standard was available.

Sony views MPEG as the way forward

Sony Broadcast & Professional Europe announced its commitment to MPEG, the internationally accepted standard, in response to growing market demand for non-proprietary environments. Sony's future product ranges will be in line with the MPEG standard, the first of which will be announced at NAB '98 in the US. Sony's SX product range is already MPEG based. "Vendors must break away from format-centric thinking and focus on what the customer is asking for - freedom of choice at competitive prices," said Miles Flint, president of Sony Broadcast & Professional Europe.

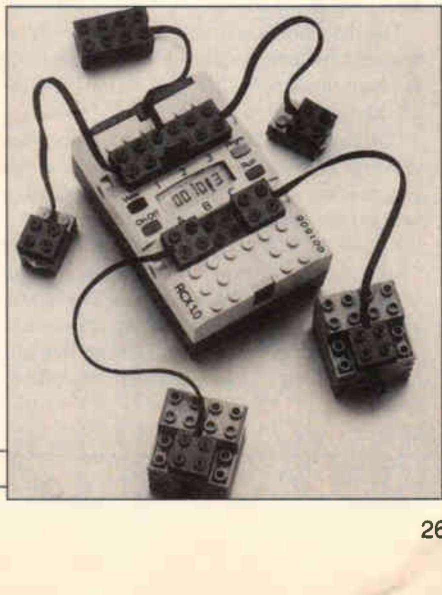
Processor in a brick

Toy manufacturer Lego has added processors to its bricks. Its Technic CyberMaster links a pc to a model using a radio-frequency carrier and allows the model to be controlled by cursor keys.

A cd rom carries on-screen construction information, conveyed to the user through animated characters as well as interactive games. The more advanced Lego Mindstorms Robotic Invention System uses a more powerful Lego block processor, the RCX, in autonomous Lego creations.

Programming is done on the pc using a simple behavioural language; programmes are downloaded to constructions through an infra-red link.

Optical, pressure and other sensors are included to enable models to interact with their environment. Mindstorms' computer was developed in the US by MIT and in trials a nine-year old girl made a bird-feeder digitally linked to a camera which photographed birds as they landed. The products will be available in August.



Subsidised set-tops for summer DTV launch

BSkyB will subsidise the first batch of set-top boxes for its summer launch of digital satellite television, until British Interactive Broadcasting (BIB) receives clearance from the European Commission for its launch of complementary interactive services.

"Sky might provide some of the subsides until BIB launches its service in winter of 1998. BSkyB may retrieve the money back from BIB later," said a spokesperson.

BIB, the joint venture between BT, Matsushita and Midland Bank, was initially set up to provide the interactive digital satellite services such as home banking, home shopping, access to the Internet, E-mail and special community information services among others. It was also going to subsidise the set-top boxes to bring the prices to an affordable level and hence encourage the take up of the new type of television.

However, as BIB is currently being scrutinised for regulatory approval by the DG IV of the European Commission, it will initially miss BSkyB's service launch that is confirmed for June 1998.

As the summer service kicks off, the initial batch of set-top boxes will not offer all those services promised by BIB. The next generation boxes will have greater functionality.

"Boxes will always be upgraded as they go along. It's the same as with mobile phones," said the spokesperson. But the spokesperson did add that the first boxes will be ale to receive the future BIB interactive services in addition to their capability to offer video-on-demand.

Out of the four manufacturers that BSkyB appointed to supply the boxes - Matsushita, Grundig, Pace Micro Technology and Amstrad - only Pace and Amstrad have confirmed that they will supply the initial quantities in time for the June launch. Last week Pace's shares dropped by 1.5p to 39p per share and Amstrad's shares stayed static at 32p per share.

BSkyB and BIB are both hoping to have a fully functional digital satellite service in time for Christmas 1998.

Scottish bodies making the most of product ideas

The industry body Scottish Electronics Forum (SEF) and Scottish Enterprise (SE) are jointly to support the Scottish electronics industry under the banner Realising the Potential.

The deal signed extends a five year-old venture that has seen industry and academia collaborating to take Scottish ideas out of the universities and into production.

"Quite a few developments, including a method of creating production masks and a heart monitor, are progressing well," said a Scottish Enterprise spokesman about the existing scheme.

Not all the achievements lead directly to products. "Among other things, the group has recognised a need for mechatronics training and the universities have altered their courses to suit," the spokesman said.

Shown are SEF chairman George Bennett of Motorola (left) and Ray Macfarlane, MD of operations at SE.

Bandwidth boost for Web access

Lucent Technologies has unveiled an optical networking system with sufficient capacity to handle the Internet's entire data traffic using a single optical fibre.

The company claims that its WaveStar TM OLS 400G system, with an optical fibre capacity of 400Gbit/s, offers five times the bandwidth of competing solutions. For a system combining up to eight optical fibres, 3.2 terabits of data per second is achieved.

"Lucent is driving down the ongoing cost of transmitting a bit of information close to zero while pushing capacity towards infinity," boasted Gerald Butters, president of Lucent's optical networking business.

The huge bandwidth is achieved using dense wavelength division multiplexing (DWDM) technology. Here, multiple light beams of different frequencies are carried within a single optical fibre. At the receiving end, DWDM acts like a prism, separating each wavelength, which carries its own channels of data communications.

Tax will hit small and medium firms

Small-to-medium-sized businesses (SMEs) face increasing financial hardship if the government's proposed changes to corporation tax go ahead, the CBI has warned.

Douglas Godden, head of the CBI's economic policy group, told EW that although he welcomes the abolition of advanced corporation tax (ACT) - a slice of the full tax taken by the Inland Revenue at the financial year's end - other changes could adversely affect SMEs. "On its own the abolition of ACT is quite good news," he said. "But at the same time, they've brought forward the rest of the corporation tax payments."

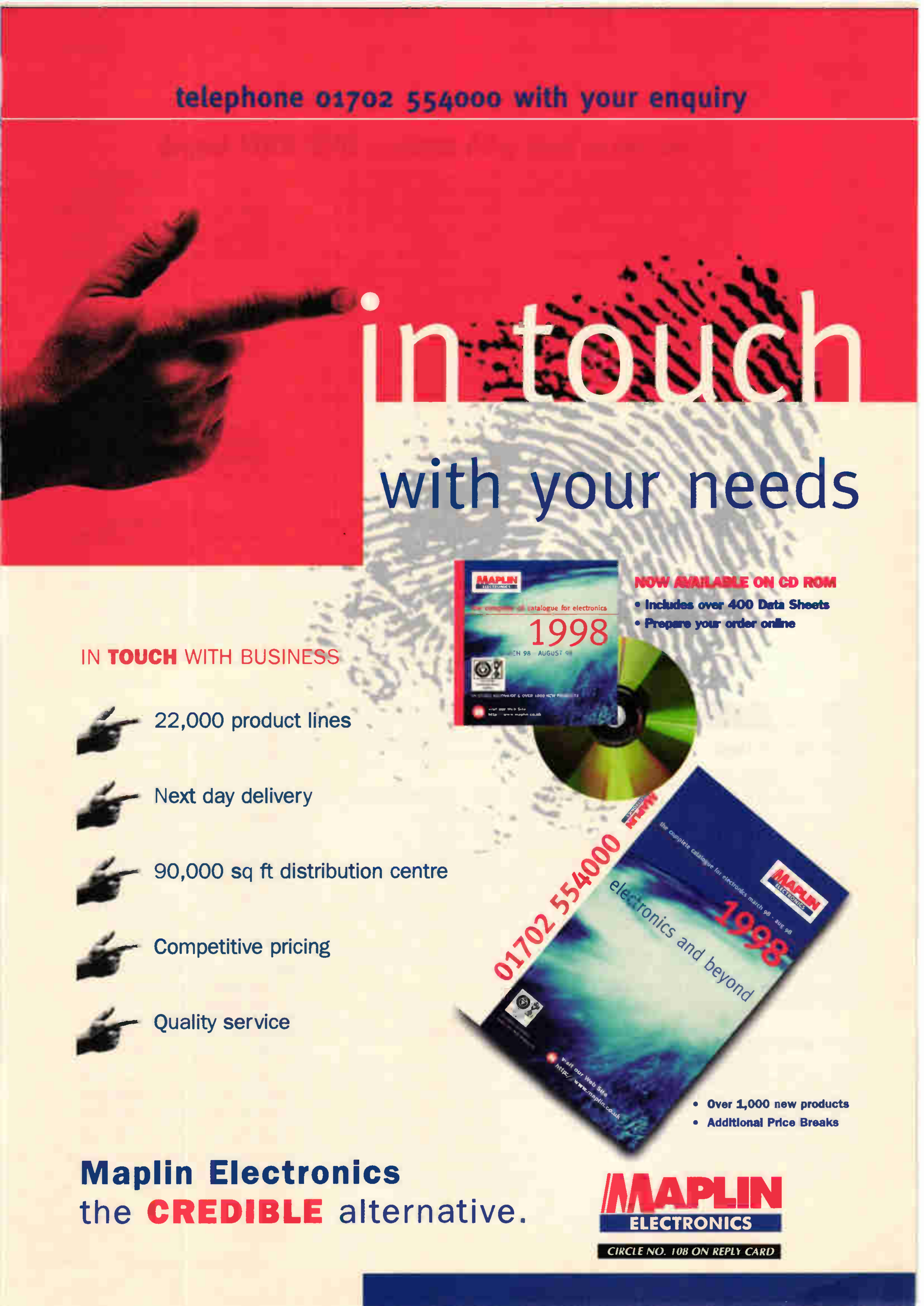
According to Godden, companies most seriously affected by the changes will be those classified as "medium-sized" with annual profits between £300,000 and £1.5m. Whereas small firms, with profits below £300,000 will not have to pay their tax until nine months after the end of the financial year, medium-sized ones will be required to pay in quarterly installments. "For SMEs, cash flow is very important," he said.

Cacheless Pentium II plans

Intel is preparing to introduce a cacheless version of its Pentium II microprocessor, representing a departure from its efforts to offer Pentium II microprocessors in Slot1 modules which also contain cache chips.

The Pentium II version, codenamed Covington, is expected in April and will enable Intel to compete against low priced Pentium-compatible microprocessors from National Semiconductor, AMD and Centaur.

Covington represents Intel's first entry into the low priced PC market.



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Hitachi takes lead with summer DVD-RAM launch

Panasonic is delaying the European launch of DVD-RAM drives until the summer, while competitor Hitachi is expected to launch its drive later this month.

According to Sakon Nagasaki, Panasonic/Matsushita's director of DVD business development, the DVD-RAM delay is linked to hardware compatibility issues and software testing.

"We are cautious in testing the software and compatibility with the existing hardware," said Nagasaki.

DVD-RAM drives from Panasonic were first expected last month. However, its launch in the US and Japan is now scheduled for the spring, followed by a European launch this summer. Meanwhile, Hitachi is believed to be launching its 2.6Gbyte per side DVD-RAM later this month.

Panasonic's drives will be able to read from and write data to DVD-RAM discs with capacities of 2.6Gbytes per side. The read/write head arrangement was developed by Matsushita, Panasonic's parent company, along the drive's signal processor and 32-bit Risc processor. A 4.7Gbyte per side DVD disc version is expected a year later, says Matsushita.

According to Matsushita, up to 500,000 DVD-RAM drives will be shipped worldwide this year, rising to 5 million next year. This growth is based on DVD-RAM's increased usage in such applications as notebook PCs, DVD camcorders and still cameras, and in-car navigation to store downloaded information.

Insight

While equipment makers may boast about the 1.5 million DVD units sold worldwide, DVD technology as a whole has suffered because of a lack of co-ordination between the hardware and software developers, and problems with standards.

DVD Video players have been selling in the US and Japan since last year. However, only now are DVD-RAM drives beginning to appear.

Is your Web site making money?

UK companies are failing to earn money from their Web sites, according to a new report.

The survey from Fletcher Research shows that fewer than 35 per cent of the 250 UK companies analysed are trying to generate revenue from their Web sites. Of these, the most popular method is through advertising, followed by on-line sales of consumer products and subscriptions.

Most UK Web sites display information only. There is little evidence of companies using their sites for business-to-business commerce.

Few of the companies surveyed are involved in electronics. "There simply aren't that many prominent business-to-business sites in electronics," said William Reeve, one of the report's authors.

One site highlighted was GEC's, described as a "professional glossy corporate brochure" for the company. In contrast, US firm General Electric uses its Web site as the sole way for companies to tender supplies

IN BRIEF

Fuel fraud in check

Fleet operators can now remotely check the amount of fuel going into their vehicles thanks to an rf technology developed by the UK-based firm New Tech.

"This is an electronic system that prevents fuel fraud." said a company spokesman. Each pump nozzle is equipped with an rf tag which communicates the amount of fuel requested to the petrol station attendant and the vehicles' on-board computer.

In turn, the on-board computer collects the information to be later processed by the fleet management. The Manchester firm plans to roll out its technology in the UK nationally twelve months after the tests have been completed.

New source of c-mos image sensors

Rockwell Semiconductor has announced its first c-mos imaging sensors aimed at digital cameras. The four device family ranges from 352 by 288 pixels up to 960 by 720 pixels. The former is aimed at pc cameras, toys and security, while the latter, from vga resolution and up, are for consumer stills cameras and video applications.

"We've been making our own c-mos sensors for about 15 years," said Ian Olsen, vice-president of Rockwell's Personal Imaging division. "But within the last year, we decided to commercialise this technology."

Pace to cut costs – and maybe jobs

Pace Micro Technology has begun a consultation process with employee representatives on how to achieve cost savings of £5m in its 1998/99 financial year. The process may involve up to 150 redundancies across the group, with as many as 90 in the UK where Pace plans to cut overheads by £2m per annum.

Three-level quantiser announced by Japanese researchers

If the goals of IC design can be simply stated, it is to advance the speed performance of circuitry while minimising the resulting transistor count.

This is the aim of work at Japan's NTT System Electronics Laboratories. It has created a novel 10GHz, three-level logic quantiser based on extremely fast (1.5ps) resonant-tunnelling diodes (RTDs).

Using the circuit as a building block, the labs claims it can make a 10GHz, 4-bit flash a-to-d converter requiring only a quarter of the active components required for a conventional flash converter.

This multi-valued logic block will form the basis for the 4-bit flash a-to-d converter. Here NTT has used four-level quantisers as well as binary encoders, multiplexers and D-type bistables.

The resulting circuit requires only a quarter of the components needed for a conventional 4-bit flash a-to-d. The main reason for the savings is that five 4-level quantisers are needed rather than the 15 binary comparators employed in a standard converter.

Using simulations, NTT believes that the 10GHz performance will be achieved with a power consumption comparable with conventional flash a-to-d converters.

Asian crisis may help d-ram market

Texas Instruments has told analysts that it expects the financial crisis in Asia to provide a silver lining for the troubled d-ram market which has been holding back growth in world-wide semiconductor markets. TI's chief economist Vladi Catto said that South Korean chip companies have "substantially" reduced their purchase of chip production equipment that will help reduce the current glut of memory chips. With Korean memory manufacturers unable to ramp up new fab lines, they won't be able to flood the market with cheap chips. About 20 per cent of TI's revenues come from d-ram sales. ■

Mike J. Renardson's simple enhancement of a standard class-B output stage dramatically reduces crossover distortion. It can also remove the need for accurate setting and thermal compensation of quiescent current – making the design desirable for volume production.

Class B in a new class

My new idea involves using one half of the class-B output stage as a feedforward error correction amplifier for the other half. It may seem surprising that this is possible, because in class-B each half is normally switched off for most of one half-cycle. The usual requirement for an error amplifier is that it should remain linear at all times.

In this circuit, however, one side can remain in class-A throughout the whole signal cycle. Consider first a widely used circuit giving an output across its load equal to the difference between its inputs, **Fig. 1**. This arrangement is used as the top half of the class-B circuit in **Fig. 2**. In this case op-amps are used to drive output power transistors, but the technique can also be applied using discrete transistors.

The top half of the output stage subtracts the output of the other half, V_3, from the input voltage V_1. It produces a voltage $V_2-V_0=V_1-V_3$ across its output resistor to compensate for the error from the bottom half.

Both halves are biased onto the linear parts of their characteristics in the quiescent state by V_{bias}, and so on negative half-cycles if the bottom half provides the entire output required with no error, i.e. $V_3=V_1-V_{bias}$, then the top half will not have any change in its output across the 1Ω output resistor, i.e. $V_2-V_0=V_{bias}$. It remains operating at a constant current determined by the bias voltage and will not be cut off as in a conventional class-B circuit. On positive half-cycles the bottom half will eventually cut off, at which point, the top half takes over to provide the whole output current.

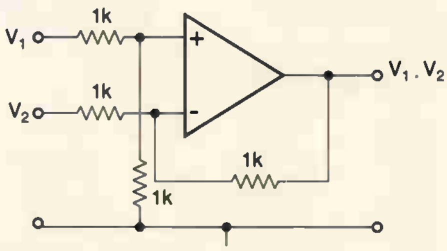


Fig. 1. This familiar circuit gives an output across its load equal to the difference between its inputs.

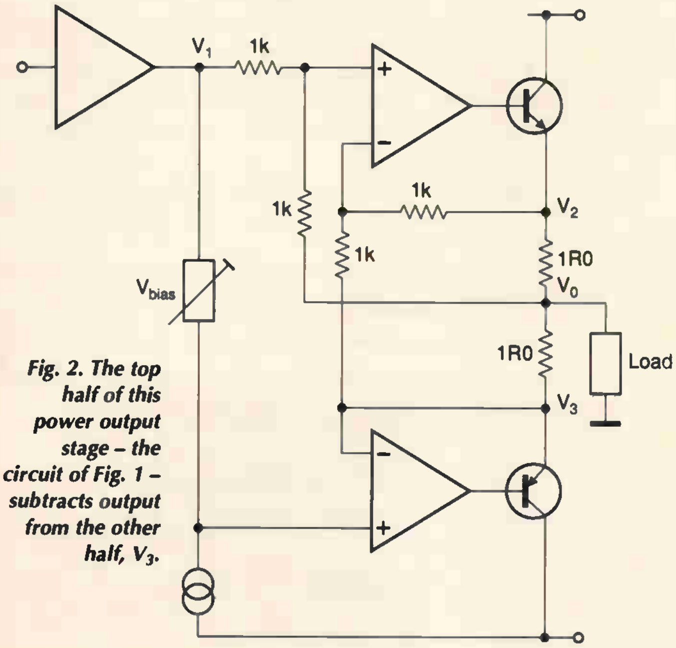


Fig. 2. The top half of this power output stage – the circuit of Fig. 1 – subtracts output from the other half, V_3.

A more convenient arrangement

There is another arrangement which is more convenient for a discrete component output stage, but again, for clarity, it is shown with op-amps in **Fig. 3**.

Operation of this configuration is perhaps more difficult to understand than in the previous example. One way to understand how it works is to observe that the top half functions as an inverting amplifier for the signal at the emitter of the lower, p-n-p power transistor.

The inverted signal is added to the original signal through the 1Ω resistors so that cancellation occurs. Output across the load is therefore independent of the output of the lower, class-B half, and is determined only by the top half of the stage which operates in class-A and can be made highly linear.

The lower half must of course still provide most of the output current on negative half-cycles, otherwise the top half will cut off in an attempt to correct the error. **Figure 4** shows how the currents vary in the two halves of the output stage with a sine-wave signal. The peak output current is I_P and quiescent current I_Q.

Figure 5 shows a complete circuit which has been built and tested. A switch is included which in one position, as shown in the diagram, gives the improved circuit, while in the other position it gives something not too far removed from a standard class-B arrangement for comparison purposes.

Distortion measurements were initially carried out at 1kHz, but in both configurations, with any quiescent current above about 10mA, the overall negative feedback was sufficient to reduce the distortion below the noise level of the test instrument used. This can easily be overcome by taking the feedback from the output of the op-amp so that the output stage is outside the feedback loop and open-loop distortion can be observed.

Doing this confirmed the successful reduction of crossover distortion in the improved circuit and the insensitivity to changes in quiescent current. The tests were repeated at 20kHz, and here the closed-loop distortion did become clearly visible giving the results shown in **Fig. 6**.

The tests were carried out with a 8Ω load, and a 20kHz, 100mV rms input signal. The distortion plus noise was extracted and displayed at various levels of output stage quiescent current for both circuit configurations. At 6mA both circuits produced similar crossover spikes shown in Fig. **6a**. The standard circuit distortion reduced to a minimum at around 10mA, as in Fig. **6b**, and increased at higher currents, and is shown in Fig. **6c** at 60mA.

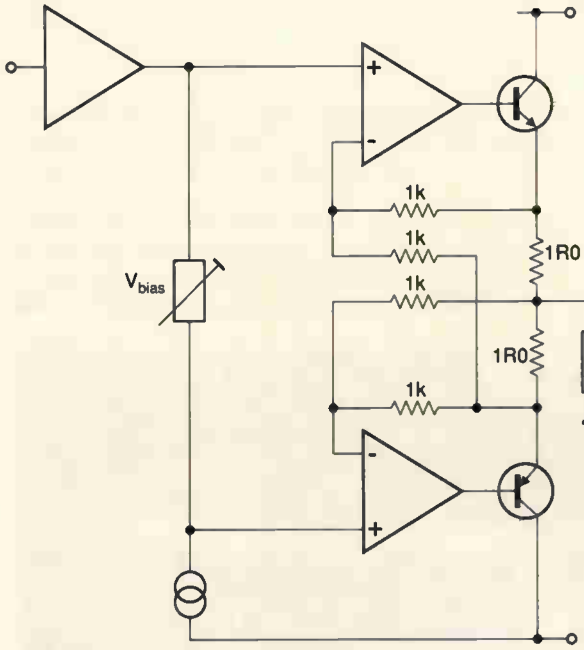


Fig. 3. This alternative arrangement lends itself to implementation in discrete components.

Adjusting for minimum distortion

The adjustment for minimum distortion was quite critical. Even a small variation from the optimum value gave an obvious increase. The improved circuit gave a similar result at 10mA, but at any quiescent current from 15mA upwards – a maximum of 120mA was used in the tests – the distortion fell out of sight below the noise level, as in Fig. **6d**.

Reducing the input signal to zero gave no visible change in the 'distortion' observed, which appeared to be entirely noise with or without an input signal present. The contrast with the standard circuit was dramatic. Above the minimum

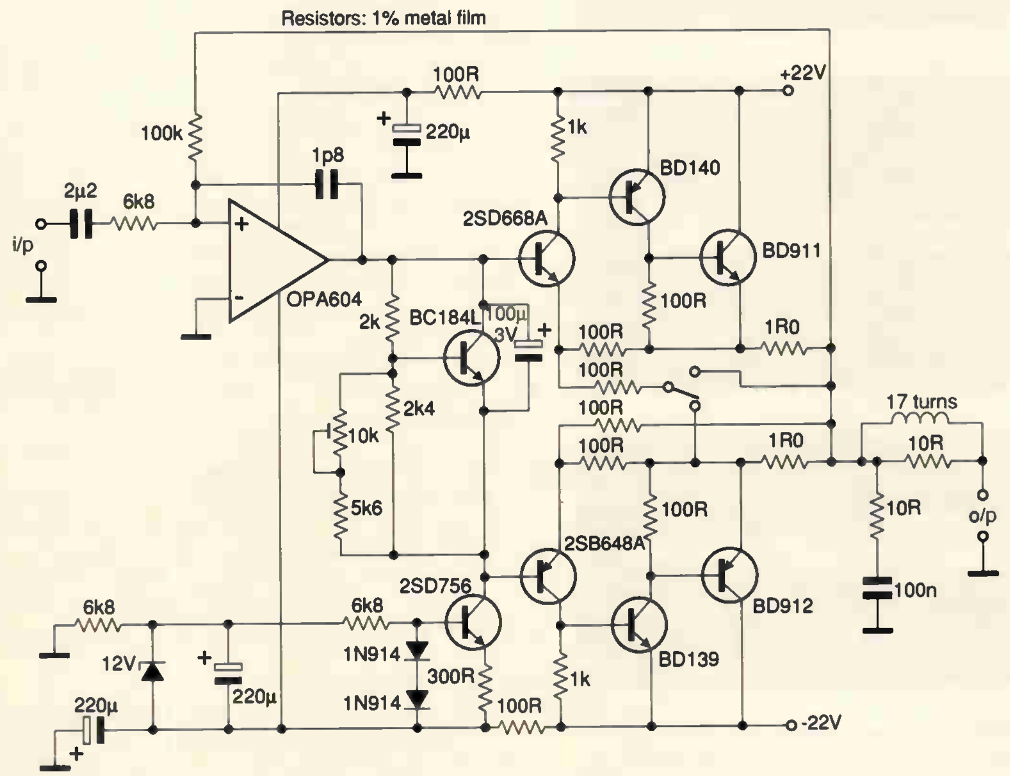


Fig. 5. Complete circuit of the improved class-B stage. A switch is included so that you can compare performance of the improved configuration with traditional class-B.

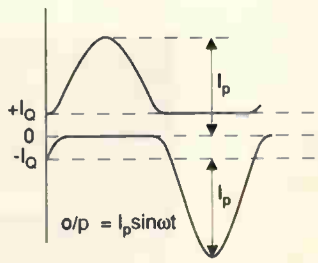
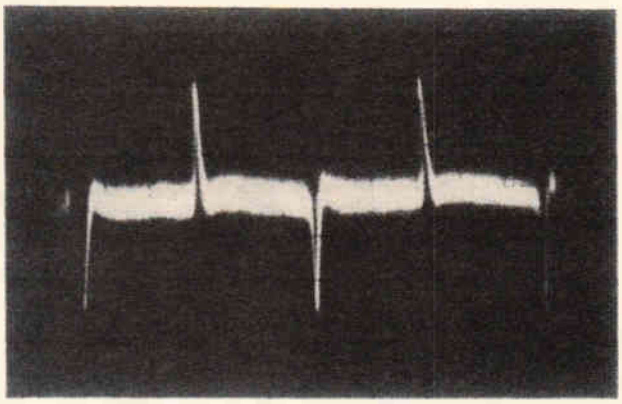
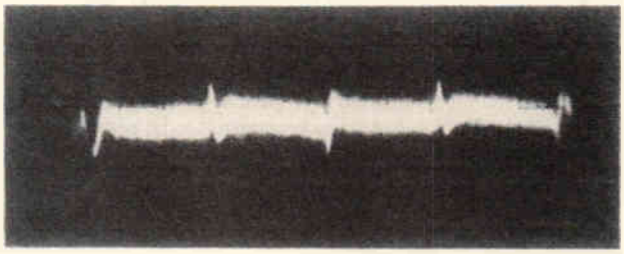


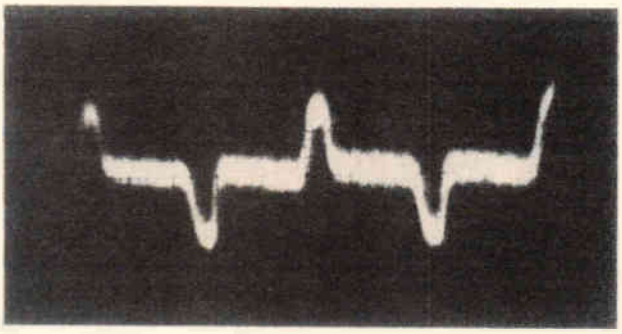
Fig. 4. How currents vary in the two halves of a class-B stage with a sine-wave signal.

level of quiescent current required by the improved circuit, flicking the switch between the two simply gave the impression of switching the crossover distortion on and off.

I am not claiming that the distortion is unmeasurable. Distortion components well below the noise level can still be measured using narrow bandwidth wave analysers, and in any case the noise levels in the test instruments used were far from ideal. My only intention was to demonstrate that crossover distortion really is significantly reduced, so more precise measurements or attempts to further improve or optimise the circuit were not made.


a)


b)


c)

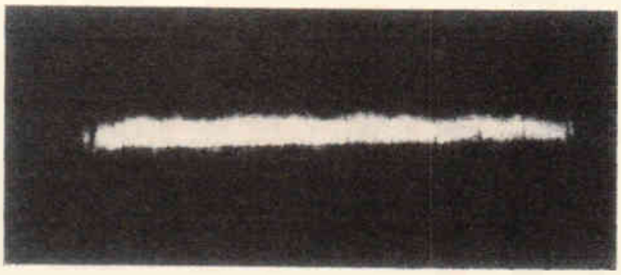

d)

Fig. 6. At 6mA quiescent current, a), both the new and existing circuits showed similar crossover spikes. Distortion in the standard circuit was minimal at 10mA, as in b), but rose again as current increased, as is evident in c) at 60mA. For any quiescent current above 16mA, distortion with the new circuit remains below the noise level.

It may be worthwhile adjusting component values in the output stage to give accurate nulling of the distortion, but in the prototype low distortion was achieved using fixed 1% tolerance resistors. The component values shown are actually not theoretically correct for accurate nulling because the lower 1Ω resistor is in parallel with two 100Ω resistors in series and so more current is fed to the output via these resistors.

A 200Ω resistor connected in parallel with the upper 1Ω resistor would correct for this, but the error is less than the tolerance of the components used, so this is a fairly minor inaccuracy. The finite open-loop gain of the output triples adds further inaccuracy. The four 100Ω feedback resistors in the output stage could be reduced to increase the loop gain, and in a higher power version this may be necessary because of the fall in current gain in the output power transistors at high currents.

The 1.8pF capacitor determines the level of overall negative feedback, giving 34dB at 20kHz and increasing at 6dB per octave at lower frequencies. Square wave ringing with a 2μF capacitive load was considered just about adequate with this level of feedback.

Driving the output stage

One aspect of this design, which is also important in other class-B amplifiers, is that the output stage has a non-linear input impedance. If the driver stage has a high output impedance, the loading effect of this non-linear impedance may add significant distortion at this point in the circuit.

The *OPA640* op-amp used in the present design has a specified open-loop output impedance of 25Ω, while the output triples will have an input impedance considerably higher than this. As a result, distortion from this source should be acceptably low.

Another source of distortion in conventional class-B amplifiers is known as 'dynamic crossover distortion'. The varying signal level causes changes in the output stage transistor temperatures which then cause a change in quiescent current. The thermal compensation circuits used inevitably involve some time delay and some inaccuracy, so that the quiescent current may differ significantly from its optimum value for much of the time, giving increased crossover distortion.

In the improved circuit the quiescent current is not critical, and even for wide variations crossover distortion should remain low.

The *BC184L* transistor in Fig. 5 was used to provide thermal compensation of the quiescent current in the normal way, and was glued to the *2SB648A*. The quiescent current was set so that when switched on from cold it started at 80mA, and was then found to slowly drift up to a little over 100mA after a few minutes. Additional stabilisation of I_Q was not necessary in this example, even though the current stability is poorer than in the standard circuit.

Adding stabilisation

Assuming you want it, there is a fairly simple way to add stabilisation by detecting the minimum current through the upper, class-A side of the output stage, and controlling the quiescent current to keep this minimum value above a certain level. The use of such stabilisation circuitry can avoid the need for any setting up of quiescent current and make

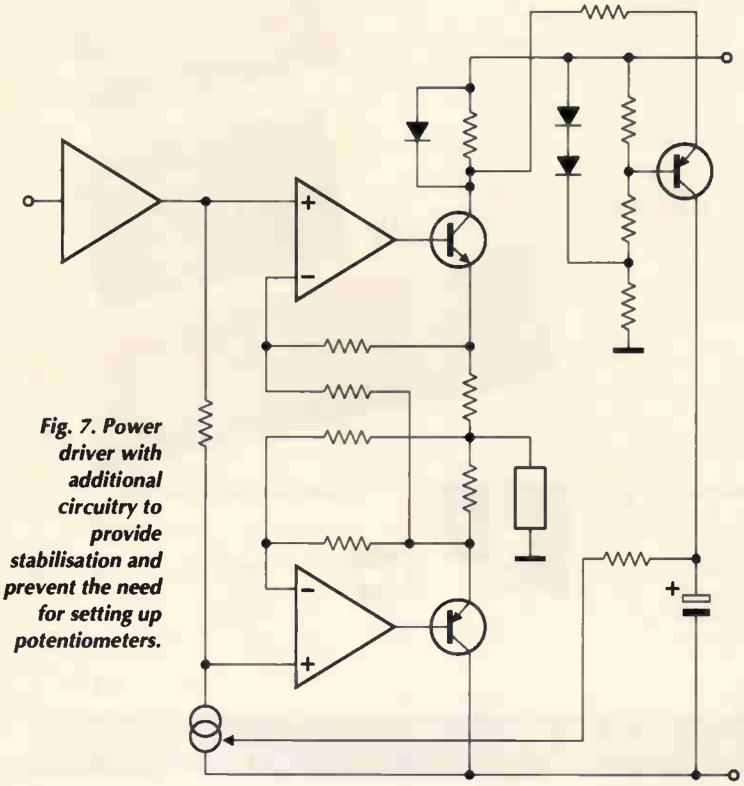


Fig. 7. Power driver with additional circuitry to provide stabilisation and prevent the need for setting up potentiometers.

possible an amplifier with no variable resistors requiring adjustment.

An example of the sort of stabilisation circuit possible is shown in **Fig. 7**. The controlled current source could be just a single npn transistor, but something a little more elaborate may be worthwhile so that the current is already set fairly accurately and the control circuit merely 'tweaks' it to keep it from drifting too far.

Although a quiescent current of 15mA was found to be adequate for low distortion in the prototype, a higher value is needed for two reasons. First, the observed drift from 80 to 100mA when the amplifier is warming up is not guaranteed to be the absolute maximum range of variation under all possible conditions, and some long term drift in component characteristics could add further to the range.

Secondly, even with 1% tolerance resistors in the output stage there could still be sufficient error on negative output currents to call for a little correction from the 'error amplifier', and the maximum correction current available in the negative direction is equal to the quiescent current.

A choice of 100mA seems a reasonable compromise. The maximum correction current required could be reduced by adjustment of the ratio of the two lower 100Ω resistors, i.e. those connected to the emitter of the *2SB648A*, but this is not a particularly simple adjustment to carry out and is unlikely be a great benefit to the performance.

Emitter resistor choice

The choice of 1Ω emitter resistors for the output power transistors is a compromise between quiescent current stability and power loss when feeding impedances below the nominal 8Ω. There are many loudspeakers available with highly inconvenient impedance characteristics, and general purpose audio amplifiers need to be able to cope with these without unacceptable deterioration in performance.

For such general use the present design may be considered inadequate, but being fortunate enough to use speakers with an undemanding impedance I chose to use 1Ω resistors. Even so, the resulting reduction in available output voltage at 4Ω load compared to 8Ω is under 1dB, so the effect is hardly a major concern.

For use with speakers with impedance falling much below 4Ω the resistors could probably be reduced to 0.5Ω without having to resorting to the sort of quiescent current stabilisation circuit mentioned earlier. Resistors with a value of 1Ω or 0.5Ω, 1% tolerance and a power rating of 2W or more are not readily available. Those used in my prototype were actually parallel combinations of higher values, i.e. four 3.9Ω, 0.6W 1% metal film components in parallel. Exact values of these two output resistors are of no importance – only their equality.

The inductor in series with the output is to reduce the effect of capacitive loads on loop stability. The component used was 17 turns, 8mm in diameter and 15mm long.

I chose the Burr-Brown *OPA604* because it has a maximum voltage rating of ±24V and I had a power supply available providing ±22V, giving maximum output power about 20W into 8Ω. I also find it a very good op-amp, designed for high-quality audio applications, with distortion rated at 0.0003% at 1kHz at unity gain. Its noise output is low, and it has a low open-loop output impedance as mentioned earlier. Gain bandwidth of the device is 20MHz.

In this application, output current is only taken from the op-amp in one direction, so the fact that it has a class-B output stage may not be a problem. No contribution to the distortion from this source was apparent in the tests.

The discrete option

A discrete input and driver stage might improve a little on the op-amp in some respects, but there seems little point in increasing circuit complexity to achieve even lower distortion if the distortion is of an innocuous variety which will have no audible effect. For higher output powers an op-amp input becomes less convenient and a discrete design more justifiable to achieve the required voltage swing.

The circuit presented here solves the most serious problems of class-B output stage design. I have only built a single prototype to demonstrate the operation of the output stage. Designing a commercial product would require more work, taking into account component tolerances to ensure repeatable results, and more detailed measurements.

One major advantage of the present design, which I hope any reader who knows otherwise will correct if I am mistaken, is that it is not the subject of any patent protection. It is therefore free to use by any amplifier manufacturer. As far as I am aware it was first published by myself in September 1996 in a low circulation magazine called 'Innovation and Speculation in Audio, Electronics and Physics', which regrettably is no longer available.

The fact that the design is so simple makes it hard to believe that it is a new idea, but at the very least it is not well known or in common use.

Why no listening tests?

Some of you may question the lack of listening tests. Well, the practical design was produced merely to demonstrate the success of the idea in reducing crossover distortion, rather than as a final commercial product. It is certain that if there is any audible difference between the two configurations tested it will be the improved version which will be the more correct. ■

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Construction	Full length ISA card	Full length ISA card
Frequency range	0.5-1300 MHz	0.15-1500 MHz
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)
IF bandwidths	6 kHz (AM/SSB), 17 kHz (FM-N) 270 kHz (FM-W)	2.5 kHz(SSB/CW), 9 kHz (AM) 17 kHz (FM-N), 270 kHz (FM-W)
Receiver type	PLL-based triple-conv. superhet	
Scanning speed	10 ch/sec (AM), 50 ch/sec (FM)	
Audio output on card	200mW	200mW
Max on one motherboard	8 cards	8 cards
Dynamic range	65 dB	85 dB
IF shift (passband tuning)	no	±2 kHz
DSP in hardware	no	yes
IRQ required	no	yes
Spectrum Scope	yes	yes
Visitune	yes	yes
Published software API	yes	yes
Task Manager/scheduler	no	yes
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Advice on buying a Pentium, free CAD, filter design software and Peltier applications information. Cyril Bateman shows you how to find out about these topics and more on the Internet.

Hands-on Internet

While I was writing this piece, most Internet news columns were busy reviewing last year's highlights and making predictions for the coming year.

Intel has already repeated what it did last year by releasing a new processor shortly after the Christmas holidays. The 'Slot 1' *Deschutes* 233MHz processor is scheduled for release on 26 January[1]. *Deschutes* is to be a generic name for Intel's *Pentium II* range, fabricated using 0.25 micron technology.

This change to 0.25 micron fabrication means increased clock speeds of 400MHz and more. Smaller geometry also reduces processor heat dissipation. Portable computers using this *Pentium II* processor are expected to arrive this summer.

The 'Slot 2' version of *Deschutes* is due this summer too, together with the *450NX* chipset, which incorporates support for multi-processor systems with up to four processors. This device is expected to result in the demise of the *Pentium Pro*, leaving Intel producing only processors for Slot1 and Slot2 based systems.

Initially pcs costing under \$1000 represented just 9% of retail sales. But during 1997, the sub \$1000 pc exceeded all expectations, ending the year with 30% share of the pc retail market[2]. This result seriously affected plans to constantly upgrade processor speeds and Level 2 cache capacity. While the processor road-map must continue little changed, expect variations – even removal of the L2 cache – to cater for changing market demands.

A second driving force to emerge during 1997 was low total-cost of ownership, or tco[3]. The term total-cost of ownership refers to the real cost of pc ownership as distinct from its initial purchase cost. In the light of continuing pressure on Microsoft and Intel's near monopolies on software and hardware, total cost of ownership could well come down.

Confused about pc upgrades?

The Advanced Graphics Port, or AGP, was once restricted to Pentium II motherboards, but is now available for the traditional Pentium Socket7, as used for the Cyrix and AMD processors. This increases your options but also adds to the confusing range of choices available when you come to buy or upgrade your pc. If you are considering buying, or you are simply interested in being up to date, you will find information and speculation on the topic at two dedicated web sites.

Tom's Hardware Guide[4] and Jason's Homepage[5] provide regularly updated views of impending changes and their implications, together with details of Intel's stated processor plans. Both pages are worthy of a look, whether you are

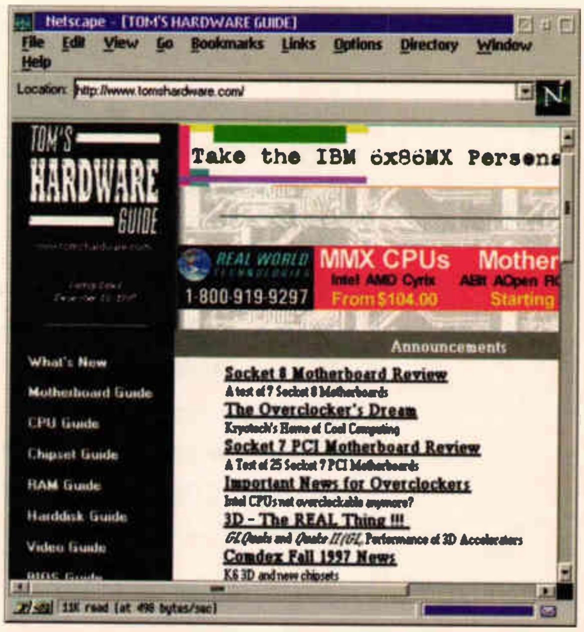


Fig. 1. Tom's Hardware – a collation of independent views and reviews on all computer hardware components.

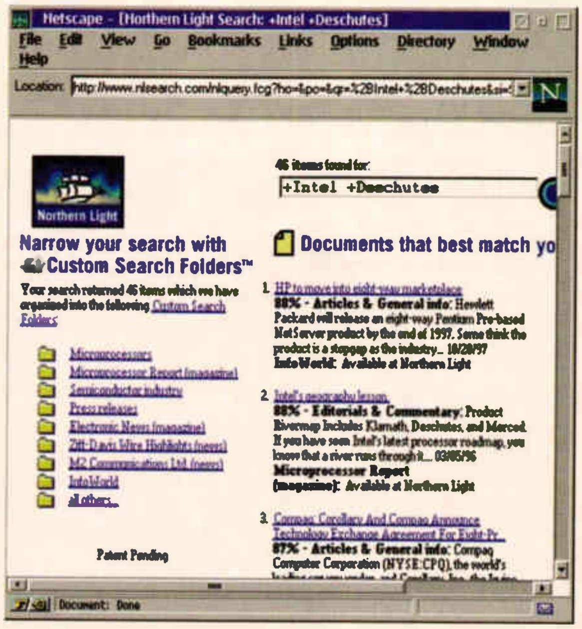


Fig. 2. This search engine, far right, can access both Internet and hard-copy resources. The second hit for my search was from the Microprocessor Report – a hard-copy magazine.

considering upgrading an existing computer or a new purchase, **Fig. 1**.

Microsoft under pressure

After two months of duelling briefs, filings and press releases, the case against Microsoft continues apace. Microsoft and the Department of Justice have both submitted their arguments as to whether Microsoft is in contempt of a court ruling. This case has been fully covered and debated in the Internet news[6] columns.

One possible side effect of this pressure, has caused concern with retail dealers. Will Windows 98 arrive late spring, as scheduled? Or as a result of this case, will the code have to be re-written, causing delays. These dealers[7] fear that impending software upgrades are likely to be delayed until the final court conclusions are reached.

New search options

For many of you, Internet is about finding answers to questions and for downloading software. This is certainly what it is about for me, not forgetting the e-mail aspect for sending messages and relaying drawings and software files.

One new search engine, used exclusively to research for this article, is Northern Light.[8] This engine provides access to information that has been digitised from many hardcopy publications, as well as from Internet resources. It returns your results tabulated and graded in the usual fashion, in addition to being sub-divided into folders by topic or source.

Should the information you seek be available only from its digitised hardcopy files, then a small charge for access is made, refundable if the data is subsequently unsuitable and rejected, **Fig. 2**.

Alternatively, the All-in-One[9] search page may better suit your needs. It provides a single compilation of all the various forms based search tools found on Internet on one large page. In particular, All-in-One should be useful as a last ditch method of finding more elusive data.

Do you use zip files?

Having found your data, it will frequently have been compressed. For years I used the reliable DOS PK204G package with every success, saving files to disk to be unzipped when off-line. If you zip and unzip frequently, there are two more up-to-date packages that are worthwhile evaluating. One is WinZip,[10] which unzips almost transparently and the other NetZip,[11] which is designed as a Netscape or Explorer browser plug in.

Unfortunately, many interesting files are compressed using methods other than 'Zip'. Among the possibilities are 'Arc', 'Arj' and 'Lha' files. De-compression software for all these is available on Internet, from sources detailed on the WinZip pages[10].

Some time ago, in response to a reader's question, I replied that files ending in 'Tar' or 'Gzip' were Unix files and could not be read on a pc. But I have recently found out that these files can be uncompressed on a pc.

The National Geodetic page[12] provides a download called gzip386.exe for dos, and also gives links for other systems. Similarly, NetHelp's page[13] provides the 'tar.exe' utility for dos, together with links and much useful information regarding file compression and de-compression methods.

Fig. 3. Concerned about junk e-mail, Hackers, and in fact any Internet undesirables? Visit the Info-sec.com page for information and advice.

Avoiding junk mail – and worse

Regular Internet users often receive annoying 'junk' e-mails. Unlike mailed junk, which costs the recipient nothing – in UK at least – junk e-mails contribute to on-line costs and waste time. More often than not, my junk mail receipts far exceed genuine mailings.

In addition to the e-mail problem, ActiveX and cookie files can allow outsiders to interrogate your system, which is potentially more harmful.

Sponsored in part by the US Department of Defense, the Info-sec page[14] provides information and solutions to these problems. One well reviewed software package for Win95 users – PC Secure Personal Firewall[11] – claims to provide solutions both for individual users and networks, resulting in an effective yet low-cost 'firewall' against hackers and nuisance e-mailers. For corporate users, the company firewall can be extended to mobile off-site users, **Fig. 3**.

Designing filters?

Continuing last month's theme, I spotted information on designing time-continuous active filters using Burr Brown

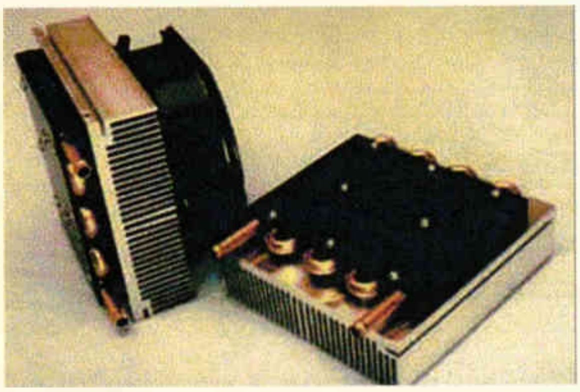


Photo A. Equipped with coolant pipes or as air-to-air, the thermoelectric module transfers heat to or from ambient air, via its heatsink and fan (detailed on page 277).

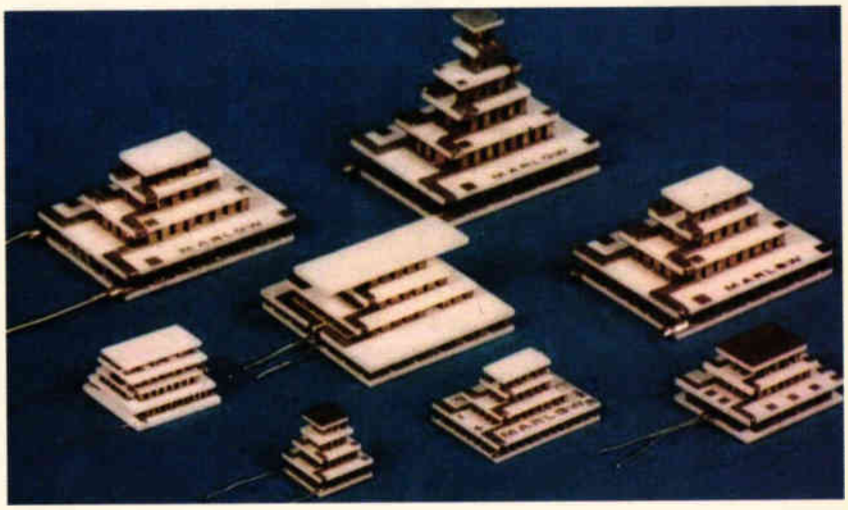


Photo B. Selection of very low temperature capable, multi-tiered modules from Marlow.

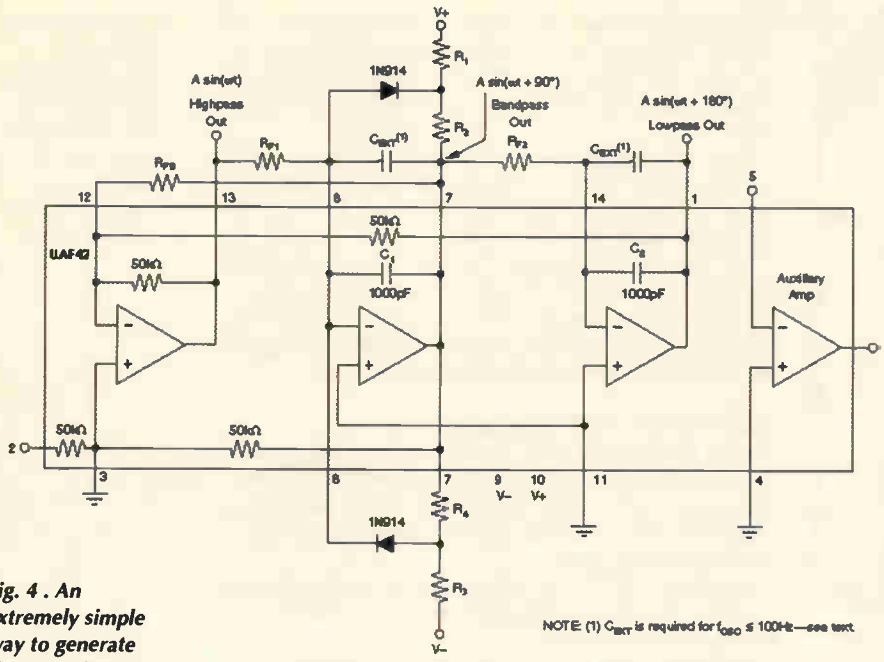
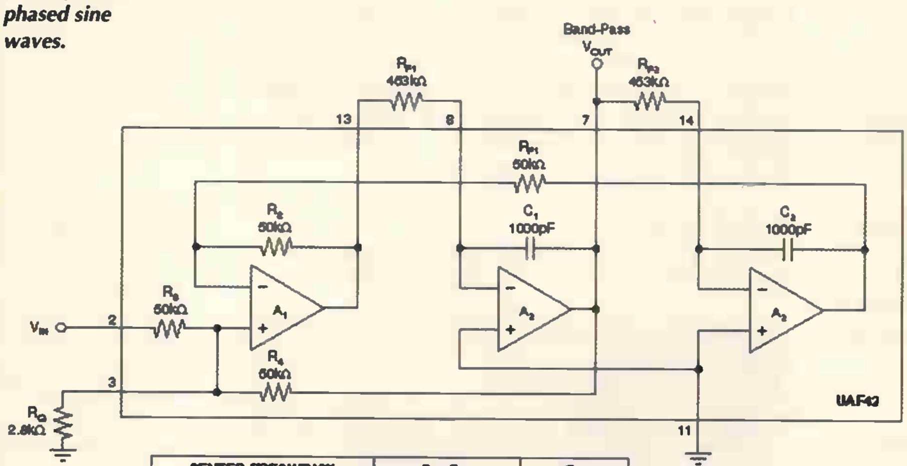


Fig. 4 . An extremely simple way to generate three quadrature phased sine waves.



CENTER FREQUENCY (Hz)	R_{F1}, R_{F2} (Ω)	R_Q (Ω)
350	463k	2.8k
440	365k	2.8k
480	332k	2.8k
620	255k	2.8k

Component values for selected Q = 10 "tuned-circuit" bandpass filters.

Fig. 5. The quickest, easiest route from a square wave to a low distortion sine wave.

Uaf42's integrated circuit.[15] Two downloadable application notes describe how to design a three-phase sine wave generator and a simple method to convert a square wave into a low distortion sine wave.

The three-phase quadrature oscillator, described in application note *AB-096*, is handy in that it provides outputs at 0, 90 and 180°. It does so using only a single integrated circuit, together with seven resistors and two diodes.

Frequencies up to 100kHz are possible, albeit with increased distortion. For frequencies below 100Hz, it is advisable to add two external capacitors, complementing those on-chip, to avoid using excessively large resistor values, **Fig. 4**.

A major problem when designing sine-wave generators is to provide a constant amplitude output without adding significant distortion. One method which completely eliminates the need to control output level is described in the *AB-058* application note.[16] Constant amplitude square waves and even multiple square waves with precise mutual phases can be easily generated.

Using a Q of 10 or better band-pass filter tuned to the square-wave frequency reduces the third harmonic to 40dB below that of the fundamental. This level equals the distortion of many commercial variable frequency sweepers. Higher Q filters reduce this distortion level further.

The circuit as described in this application note has the supreme merit of needing only three resistors, and of course the integrated circuit, **Fig. 5**.

By adjusting the square wave frequency slightly and feeding the filter's output via an on-off switching *HEF4066* IC driven from the square wave, a simple and useful audio frequency 'tone burst' generator is easily built. Although simple, this generator has precise and adjustable zero cross switching.

Simulation

I am often asked about sources of software, and for information on Spice simulation packages. The Spice download page[17] at the Technical University of Hamburg-Harburg is a useful source of information on simulation and is one of the university's most accessed pages.

This page provides links to both European and German sites and lists software not generally found. This includes 'Al's Circuit Simulator' which claims to be the most suitable version of Spice available for distortion simulations.

Perhaps, like me, having completed a printed board design and output its Gerber file, you have felt uneasy at not being able to check this final output before committing it to a photo-plotter.

Claiming to be *the* Internet source for EDA, Ivex provides a free Gerber reader, easily downloaded from the company's page.[18] It is also accessible via the SSS-Mag page[19] previously mentioned, **Fig. 6**.

Ivex also offers two pieces of working software that are free of charge but restricted in capacity. One is *WinDraft*, which is for drawing schematics, and the other is *WinBoard* for producing pcb layouts. Both of these can be upgraded at modest cost to handle more than the 100-pin limit, or larger drawings.

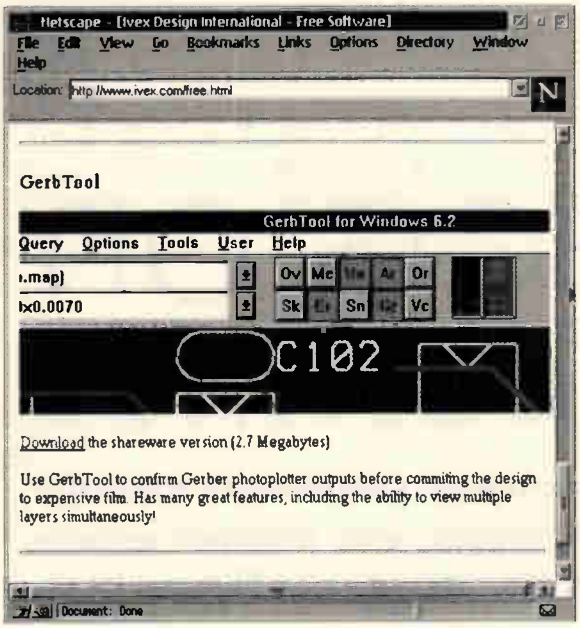
Cool topics

One ever present design problem is removal of heat – either to ensure an acceptable service life for power devices, or simply to minimise circuit noise levels. As Richard Lines' article in the last issue explained, while conventional finned heatsinks can be very effective, the device being cooled will always remain at or above the local ambient temperature. But a thermoelectric cooler can attain lower temperatures.

When a thermocouple junction of two dissimilar metals is heated, the Peltier/Seebeck effect produces a small voltage. Being a reversible effect, current passed through this junction can either raise or lower temperature, depending on the direction of current flow. This forms the basis of a thermoelectric cooler module.

Practical coolers generally use doped 'N' and 'P' type pairs of Bismuth Telluride pellets, connected electrically in series but thermally in parallel, soldered between two electroded ceramic plates. When direct current passes from an N to a P junction, this junction becomes cooled, transferring heat to the P-to-N junction.

Thermoelectric coolers can cool or heat almost anything,



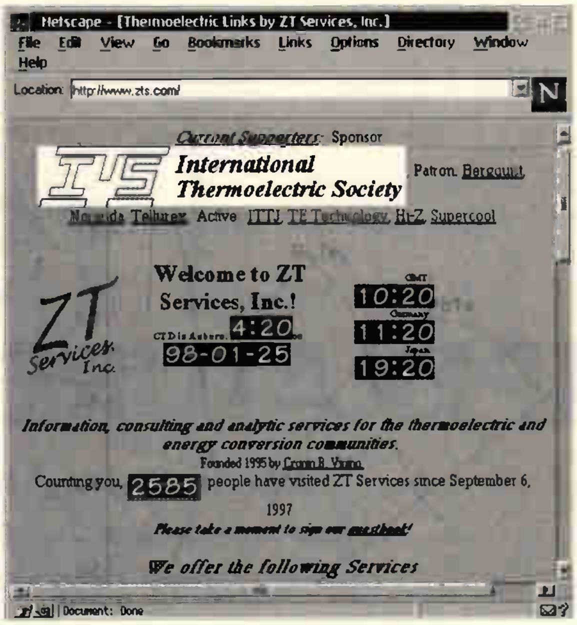


Fig. 6. Banish your 'Gerber' blues. Download this free utility, then read back your Gerber output files before committing photo plots.

Fig. 7. Meeting place for all interested in finding thermo-electric links and information.

whether liquid, gas or solid. They can also generate DC from any source of waste heat.

As with a conventional refrigerator, heat absorbed by a Peltier device must be dissipated into surrounding air via a heat sink. Used to supply heat, the device must similarly be cooled by local air. When generating electricity from waste heat, one face of the device must be maintained at a cooler temperature by ambient air circulation, **Photo A** (p. 275).

Being able to either heat or cool, thermoelectric devices are used to maintain constant process temperatures in medical and laboratory equipment.[20] They can also form a reference 'black body' for use in thermal imaging, night vision, or infra-red thermometry. Used only to cool, they support portable refrigerators, cooled drinks, laser diodes and infra-red or ccd imaging detectors.

Interested reader can start with the International Thermoelectric Society home page,[21] which provides links to many 'cooler' makers together with FAQ's, an FTP download site and a consultancy service, **Fig. 7**.

Due to thermal expansion and contraction of the hot and cold faces, a typical discrete cooler is 50mm square or less with a 30W capacity. For larger capacity, multiple modules can be used.

One demonstration vehicle had its alternator replaced with 76 small modules mounted into a truck silencer, generating 1kW of electricity and releasing engine horsepower.

To reduce the effects of diminishing efficiency with increased hot-to-cold face temperature differential, multi-tiered coolers[22] can be used to provide the lowest cooling temperatures, **Photo B**. ■

Where to surf

1.	Next Pentium II due this month	http://www.news.com/News/Item/0,4,18208,00.html
2.	Sub-$1000 PCs: The future on a budget	http://www.news.com/SpecialFeatures/0,161,17709,00.html
3.	1997 Year in Review	http://www.zdnet.com/pcweek/sr/1222/22prod97.html
4.	Tom's Hardware Guide	http://www.tomshardware.com
5.	Jasons Processor Update	http://www.eisa.net.au/~bozmeg/update
6.	Microsoft's $1 million question	http://www.news.com/News/Item/0,4,18001,00.html
7.	Case against Microsoft causes retail concern	http://techweb.cmp.com/crw/news98
8.	Northern Light Search Engine	http://www.nlsearch.com
9.	All-in-One Search Page	http://www.albany.net/allinone
10.	Niko Mak Computing Inc	http://www.winzip.com
11.	Software Builders International	http://www.softwarebuilders.com
12.	National Geodetic Survey	http://www.ngs.noaa.gov
13.	NetHelp PCTar for DOS	http://www.nethelp.com.au/software/tar
14.	Information Security	http://www.info-sec.com/index.html-ssi
15.	Burr-Brown Corporation	http://www.burr-brown.com
16.	Bateman, C.	'Fazed by phase', *Electronics World*, November 1997
17.	Spice Download Page	http://www.tu-harburg.de/et1/emv/et1kb/download.html
18.	Ivex International	http://www.ivex.com
19.	Spread Spectrum Scene	http://sss-mag.com/swindex.html
20.	TE Technology Inc	http://www.tetech.com
21.	ZT Links	http://www.zts.com
22.	Marlow Industries Inc	http://www.marlow.com
23.	Hi-Z Technology Inc	http://www.hi-z.com
24.	Supercool AB	http://www.supercool.se

Precision TEMPERATURE CONTROL

A versatile, high-performance temperature controller that can be used with thermocouples, platinum sensors or temperature-sensing ICs, designed by Richard Lines

My final article on temperature control describes a comprehensive analogue temperature controller. Outlined in **Fig. 1**, the system is modular, comprising the following units.

Sensor interface. This module is required needed when connecting thermocouples and platinum resistance sensors to bring up the output to 10mV/ºC. For many semiconductor sensors, such as the *LM35*, *LM45* and *AD590*, no interface is required and the sensor can be connected directly to the controller module. Have a look at my article in the February issue for more details on applying these devices.

Controller module. This module generates the set point voltage and supplies a low-level output between 0 and +5V which corresponds to the range fully cold to fully hot. Various alarm signals are made available so you are always aware of the state of the system and the output can be cut off in the event of a sensor fail.

Power module. Low-level output is passed to the power stage which is scaled appropriately for the heater or thermoelectric cooler. Two representative designs are given for unipolar and bipolar outputs, but you will probably have to come up with your own arrangement to suit a particular application in view of the huge range of load parameters possible.

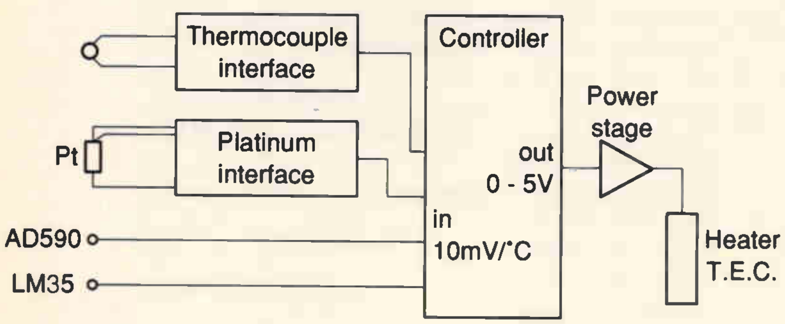


Fig. 1. Outline of the versatile temperature controller.

Circuit details

Figure 2 is the main circuit. The sensor signal scaled at 10mV/ºC is applied to the difference amplifier, IC_1, via R_2 and R_3. A termination network formed by R_2 and C_1 prevents *LM35* type sensors from oscillating if their connection cable is very long.

Resistor R_1 is a 10kΩ, 0.1% tolerance component that returns the sensor input to a precision –2.73V source; this does the scale conversion for a *AD590* sensor, changing 1μA/ºC to 10mV/ºC. It also provides the negative return for other sensors to operate below 0ºC.

A low-pass filter formed by R_3 and C_2 removes rf pickup, preventing random temperature changes. The integrated instrumentation amplifier, IC_1, compares the actual temperature with the set point. Gain is set by the switched resistors $R_{19\text{-}22}$ according to,

$$gain = \frac{50\,000}{R_g} + 1$$

where R_g is the value of the selected gain resistor in ohms. This forms the error signal and it will be 2.5V for zero error since the reference input of IC_1, on pin 5, is taken to the 2.5V line.

Depending on the details of the power stage, it may be convenient to be able to reverse the phase of the circuit. Links between terminals A, B, C and D enable this to be done, swapping over the difference amplifier input terminals. As shown with links A-B and C-D made, there is an inversion over the circuit suitable for a non-inverting unipolar power stage operating a heater. Using the same power stage with a thermoelectric cooler would require links A-C and B-D to be joined.

Together with R_8, diode D_1 clamps the output of IC_1 so that the linear range of error voltage is 0 to 5V. Assuming that gain resistor R_{19} is selected, the gain of IC_1 will be 500×; this means the linear range will be restricted to a sensor deviation of ±0.5º about the set point, over which the system will swing from full heat to full cool. Lower gains will obviously produce a larger linear range.

Monitoring the output from IC_1 gives a very sensitive measure of small temperature changes measured with respect to the set point. The error signal is then passed through a lead-lag filter consisting of $R_{9\text{-}16}$ and $C_{6,7}$. The filter time constants can be adjusted to delay the onset of oscillation as the gain is increased thus improving the stabilisation achieved.

As far as the filter operation is concerned, $C_{6,7}$ are in parallel. If a direct parallel connection were used then the filter output would be held at low voltage while the $C_{6,7}$ charge up through the large series resistor. This can lead to problems with a bipolar power stage since the output will be held at fully hot or cold during the charging process. This can produce large overshoots when the controller is first switched on, or the temperature initially going the wrong way.

Taking one end of C_6 to a relatively low impedance 5V source means that the junction of $C_{6,7}$ will naturally and quickly go to 2.5V – i.e. the centre of the control range – and not produce a false error signal while the initial charging up is in process. This does assume matched capacities and leakage currents. Both C_6 and C_7 must be low leakage tantalum types.

An output buffer, IC_5, produces a low impedance drive for the power stage. Main

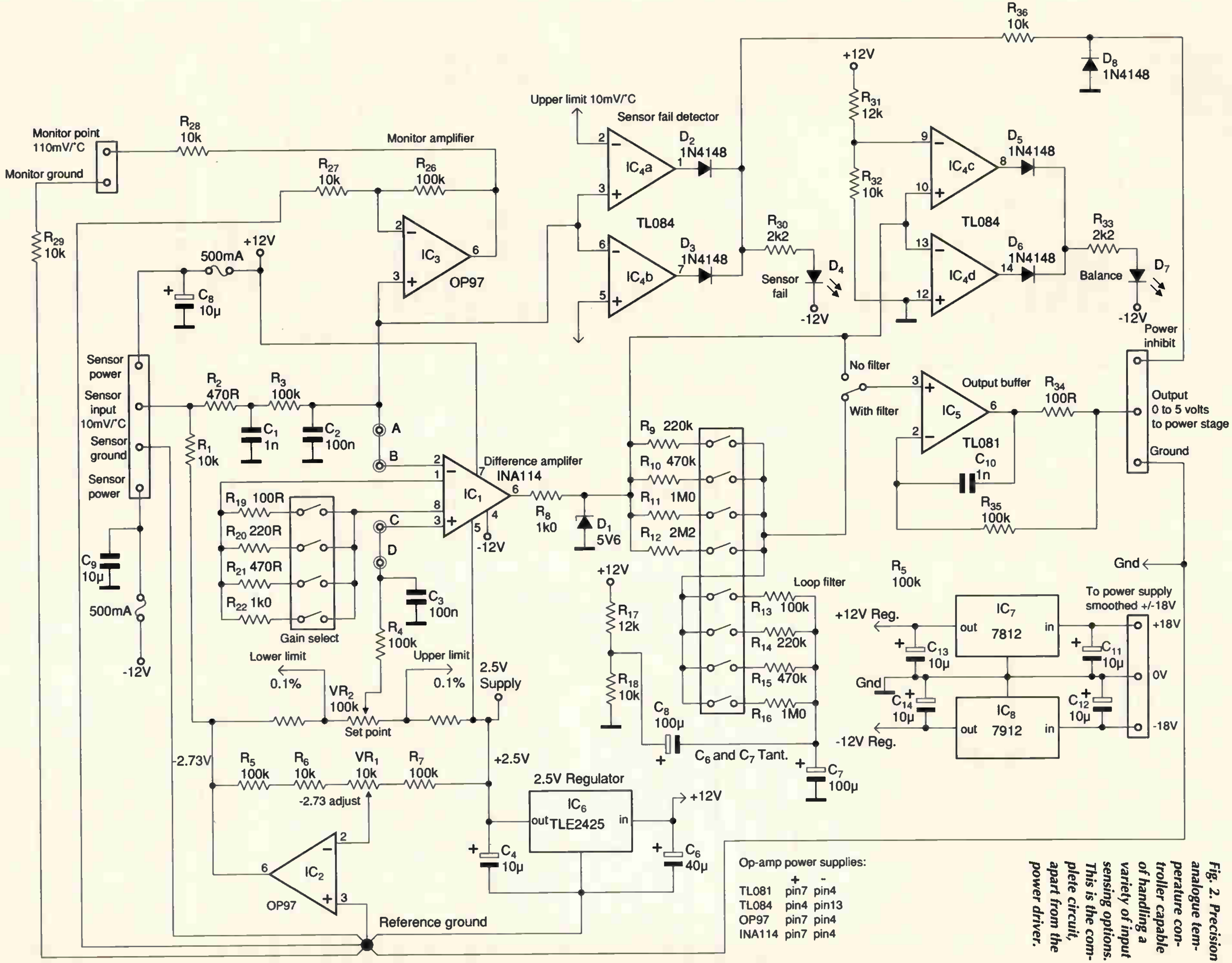


Fig. 2. Precision analogue temperature controller capable of handling a variety of input sensing options. This is the complete circuit, apart from the power driver.

Need more information?

- 'Control Engineering' by W. Bolton, pub. Longman Scientific 1992. This is a very good text on theoretical aspects of control engineering in general explaining the maths clearly.
- National Semiconductor application notes: AN292 Applications of the LM3524 switch-mode controller.A useful single chip control system. AN225 Thermocouple cold junction compensation. AN256 Relative Humidity Measurement. This article contains more information on the stabilised logarithmic amplifier principle described in part 2. AN266 Sample and hold applications. A precision, high efficiency controller is described using a thermistor sensor driven by a high voltage, short duty cycle.
- Linear Temperature Gauge by H. Kuhne, *Elektor*, September 1993. An interface circuit allowing many small diodes/transistors to be used as sensors providing 10mV/ºC.
- Marlow Industries Catalogue and application notes for thermoelectric coolers. Available on request to 7 Laura House, Jengers Rd, Billingshurst, West Sussex RH14 9NZ, tel. 01444 443404.
- *Minco* Catalogue and application notes. Applications for small foil heaters, some come with built in platinum sensors. Useful for ovens etc requiring small surface mounted heaters. Application note 21 especially useful for estimating heater power. Minco components are available from Carel Components, 24 Endeavour Way, London SW19 8UH, tel. 0181 946 9882.
- *Eurotherm* manufacture an excellent range of digital PID temperature and process controllers. 905 series very expensive but very good. Eurotherm Ltd, Unit 8 Sackville Trading Est, Sackville Rd, Hove, East Sussex BN3 7AN, 01273 206641
- *Cal Controls* manufacture temperature controllers, stocked by RS, tel. 01462 36161.
- *Linear Technology Applications Handbook*, 1990. Contains many circuit ideas for platinum and thermocouple sensors, compensation, linearisation.

monitor output is provided by IC_3, which amplifies the sensor output with a gain of 11. This gain is suitable for operation between ±100ºC, before op-amp saturation sets in; if the system is to be used over a wider range then the gain will have to be reduced.

Obtaining best results

For best accuracy, a true differential measurement is recommended with respect to the reference ground indicated on the circuit. The monitor ground is provided for this purpose, rather than the monitor circuit just relying on some gratuitous ground from the main power supply.

Op-amps IC_2 and IC_6 provide the close tolerance +2.5 and −2.73V sources necessary for generating the set point voltage and for the *AD590* current-to-voltage conversion. The set point voltage is taken from a potential divider comprising R_{23}-VR_2-R_{24} strung across the two rails; R_{23} and R_{24} are selected-on-application 0.1% resistors to set the extremes of the operating range on the basis of 10mV/ºC from the wiper of VR_2.

If operation above 0ºC with sensors other than the *AD590* is all you need, then the circuitry around IC_2 can be omitted.

Protection circuitry

System monitoring is provided by IC_4. The four op-amp sections are wired as two window detectors. The first window detector $IC_{4a,b}$, looks at the sensor input voltage, comparing it against the upper and lower set-point extremes.

If the sensor input or ground wire goes open-circuit, the sensor-fail led illuminates and a signal is sent to the power stage to cut off the output. Since this is a very important part of the circuit from a safety aspect, there are a few extra points worth making.

The circuit will not detect shorts to ground if the set-point range includes 0ºC. This is because 0V on the sensor wire will be a valid temperature. Unless your application really demands it, the set point range should not include 0ºC.

There will be instances where, although there is no intention of using the controller at this temperature, it is convenient to make one end of the set-point calibration 0ºC. If this is the case, it is possible to modify the threshold for the window detector to guarantee detection of a short to ground.

It is important not to choose too narrow a set point range in case the temperature at startup is outside this range and brings up a false sensor fail condition. In this case the circuit will never start until the ambient temperature comes in range. Operation at the extreme end of the set point range may cause temporary sensor-failure alarms due to overshoot and poor damping.

The second window detector checks the output from the difference amplifier IC_1 to check whether the circuit is operating within its linear range. If this is not so, it is impossible for the temperature to be controlled. This alarm is illuminated while the temperature is outside the range decided by the gain setting. As a result, expect it to come on when the unit is first switched on and go out as the set point is approached.

This feature is not essential and can be omitted, but I have found it very useful since it can indicate unexpected heat leaks due to missing insulation. It can also be used to inhibit some other process until the temperature is approximately correct.

Driving a heater or cooler

Two representative small output stages are shown, **Figs 3a,b**), for unipolar and bipolar operation respectively. They are included more for completeness than in the expectation of being widely useful since the possible range of applications and powers is too great to come up with a universal design.

Both circuits can be easily scaled up or down to suit. Each is shown with two inhibit controls. Input *INH1* connects to the power inhibit output on the main circuit cutting off the output under sensor fail conditions. Input *INH2* is spare and can be used as protection against some outside contingency.

In the unipolar circuit the op-amp IC_1 needs to have a common-mode range extending to the negative supply, and Tr_1 will need a heat sink of about 5ºC/W. The circuit is a standard current source with Tr_2 providing the current cut-off. The optimum resistance if the heating element is given by,

$$R_{load} = \frac{V_{CC} - 3V}{I_{max}}$$

which is about 17Ω for a 20V supply.

The bipolar design is based on a Howland op-amp current source, using a small power *L165V* op-amp from SGS. These devices can be tricky to keep stable. Components C_4, R_{10}, $C_{7,8}$ are essential for this purpose. Also note the two ferrite beads connected over the leads of R_{11}; conveniently, these can be used to mount the wire-wound resistor clear of the circuit board.

Since the driving voltage is effectively bipolar centred on 2.5V, a special fet optoisolator is used as the current cut-off mechanism. This isolator is a Harris *H11F1*. The fet is symmetrical and operates correctly regardless of polarity. Lamp D_2 provides a visual indication that the output is cut off.

In the context in which the original temperature controllers were used, it was essential not to introduce electrical noise into other parts of the system. Linear circuits tend to be the best bet on this score. The trade-off is low power efficiency.

For a more efficient unipolar switch-mode output stage, take a look at the article 'A simple pwm amplifier' in the September 1997 issue of *Electronics World*. This circuit is quite good at driving very low resistance loads from relatively high supply voltages and has been used to drive a *MI 3040* three-stage thermoelectric cooler requiring 7.3V at 4.5A from a 24V supply with 85% efficiency.

Implementing the controller

When aiming for extremely high temperature stability, it is essential that the controlling electronics is made as free from drifts as possible. Taking the input scale factor of 10mV/ºC, this translates to only 1 microvolt per millikelvin. This determines the level of set-point drift that can be tolerated.

Presumably, the item whose temperature is being controlled will be wrapped up in thermal insulation, but the control electronics will be exposed to changes in ambient temperature.

Referring back to my first article in the January issue, the idea of nesting temperature controllers was described to achieve really good stability. Overall performance is the product of the individual servo gains for each layer of control.

In practice though, it can be difficult to realise the expected performance because drift in the set point becomes more important than lack of servo gain. The obvious answer – space permitting – is to put the control electronics inside the controlled enclosure. In cases where this is not possible, some assessment of the expected drifts must be made.

Starting with the voltage references, the primary source is IC_6, a Texas *TLE2425*. This device has a quoted temperature coefficient of 20ppm/°C. This equates to 50μV/°C at 2.5V and will be scaled down through the set point potential divider.

Reducing drift

For a set point of 50°C the drift is 10μV/°C corresponding to 1mK of set point drift per degree of ambient change. If you need very high stability, it may be worth searching out an improved reference.

Devices in the excellent Maxim *MAX676/7/8* series have quoted drifts down to 1ppm/°C. Unfortunately, a 2.5V component is not available so some redesign would be necessary. Be sure to read the data sheet on the op-amp used here to avoid compromising the performance with layout errors.

The –2.73V source is generated with IC_2, which has been selected for low offset drifts of 0.5μV and 12pA/°C. These numbers should not introduce problems with the circuit resistances used.

Resistors R_{5-7} need to be 0.1% tolerance. The RS ones used in my prototype are quoted at ±15ppm/°C maximum. As these resistors are part of a potential divider, the drift would not matter if all resistors drifted the same way. However, the trimmer potentiometer VR_1 has a drift of 100ppm. This makes it important to keep the value of VR_1 small compared to the total chain resistance. Trimmer VR_1 needs to be a ten-turn device to facilitate accurate and consistent setting of the –2.73V level.

Similar comments apply to the set-point chain R_{23}, VR_2, R_{24}. The set-point control VR_2 is ideally a precision ten or twenty-turn component, for which calibrated knobs are available. Devices in the three-watt *534* series, available from RS, have a temperature coefficient of only 20ppm/°C, which nearly matches the two fixed resistors.

Instrumentation amplifier IC_1, a Burr-Brown *INA114*, has very good drift parameters of 0.25μV and 15pA/°C. Since this device is quite cheap at around £5, it is not recommended to substitute it with a device of lower performance.

Grounding the system

Board layout is not critical with the important exception of the input ground arrangements. All the ground connections important to accuracy are brought back to a common reference point as indicated on the circuit.

Poor grounding arrangements were found to be the largest source of error in the first prototype. It is vital not to pass power supply or load currents through any of the six wires going to this junction.

It is equally important to keep the input circuitry away from sources of heat. Any such sources will introduce unwanted thermocouples in the chip sockets and soldered joints.

The gain select and loop filter switches are conveniently implemented with DIL switch arrays.

Commissioning

On completion of the circuit, fit the links around the difference amplifier for the expected phase. Calculate and fit $R_{23,24}$ for the required set point range.

Leave the heater or thermoelectric cooler initially disconnected, but connect the sensor and interface circuit if appropriate. The sensor fail light should be out.

Check the +2.50V rail with a dvm measuring with respect to the reference ground, and adjust the –2.73V. It should be possible to see the correct set point voltage at pin D. Select the lowest gain setting, R_{22}, and the loop filter out of circuit.

Connect a voltmeter to the output of the main controller circuit. It should be possible to see this voltage swing between 0 and 5V as the set point knob is swung through the expected sensor temperature. This will probably be room temperature if this is included in your set point range.

Connect the voltmeter to the temperature monitor point. Connect up the heater or thermoelectric cooler and see if the system basically works correctly.The balance light should come on until the temperature comes inside the linear range.

After a few minutes the system should settle down to a final temperature. If all is well, this will be somewhere near to the expected set point. But there may be some overshoot in the process.

Leave the circuit a few minutes to check that there are no oscillations. If the system does oscillate on the lowest gain setting then it may be necessary to change the gain selection resistors R_{19-R22}. Alternatively, you can give some consideration to modifying the mechanical arrangements with the heater and sensor to improve their thermal coupling.

If the system is stable then note the set point error. Try selecting lower values of gain resistor so increasing the gain, and check the set point error decreases as expected. With the component values shown the gain roughly doubles for each lower resistor value.

The ringing gets worse as the gain is increased. Note the point at which the system oscillates, and obtain an approximate time for one period. This time is usually seconds to low minutes, depending on the heat capacity, heater power and sensor location.

If the thermal coupling is very good, it may be possible to use the highest gain setting without oscillation, in which case it is probably worth checking the overall performance without the loop filter connected. There are more details on this in the January issue.

Check the operation of the sensor fail circuit by simulating fault conditions. The heater current should fall to zero if the sensor is disconnected.

Adjusting the loop filter

The purpose of the loop filter is to allow the use of a higher gain setting without oscillation. This results in better thermal stability and lower set point error.

The bottom line is that the time constants have to be adjusted empirical basis to see whether more gain can be used without oscillation. However, it will be helpful to quickly outline the principles in the hope that you will have a better feel for which way to set the time constants.

Figure 4 shows the standard temperature control setup, but with the addition of a hypothetical thermoelectric cooler and very low frequency generator. The thermoelectric cooler is to perturb the system. It allows heat to be added or removed from the experiment so that the controller response can be examined as it tries to correct for the added heat.

In particular, you need to look at the amplitude and phase relationships between the heat supplied by the heater and the sensor response as the frequency is varied. There will be a delay between the heat added and the sensor seeing a change due to the thermal capacity and thermal conductivity of the block.

If the generator is set to a very long period – typically several minutes – the delay in the block will be short in comparison and the servo system will keep track. Thus the phase difference is small, and the servo gain will approach the steady state value.

Increasing the generator frequency introduces two effects. The delay time for the heat change to propagate through the block becomes a significant fraction of the cycle time. This introduces a phase shift. Secondly, the temperature changes seen by the sensor become attenuated since there is less heat added/subtracted per half cycle.

These effects reduce the servo gain. There will be a frequency where the heater will be adding heat on the current cycle while the sensor is still seeing a temperature decrease from the last cycle due to the propagation delay. Thus an extra unwanted 180° of phase lag is introduced by the block and the servo will oscillate. This assumes that the electronics can provide enough gain to make the overall loop gain greater than unity.

Low-pass filtering in the loop

To avoid this situation a low-pass function that rolls off the electronic gain as the phase shift inside the metal builds up with increasing frequency needs to be introduced into the loop. This low-pass function makes sure that the loop gain stays less than +1 at all frequencies.

There is the important extra qualification that the low-pass filter should not introduce any extra phase lag; this will simply reduce

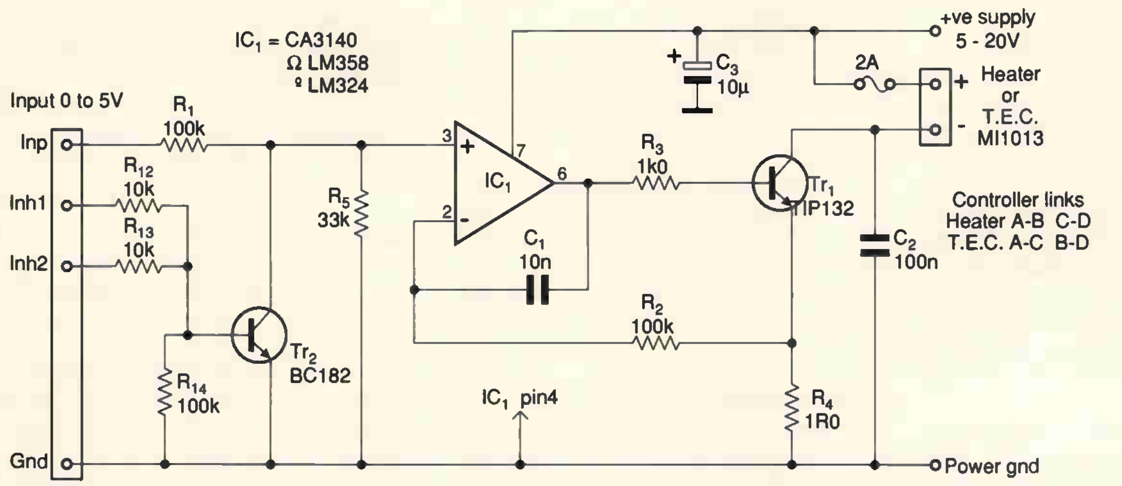


Fig. 3. Circuit a), at the top, is a linear unipolar output stage capable of driving a heater or thermoelectric cooler. The TIP132 is an 8A maximum Darlington capable of dissipating up to 70W. In b) is a bipolar driver for applications where a thermoelectric device is needed to heat or cool. The 165 is a power op-amp with a 3A output rating. It features safe operating area protection and thermal cut-out.

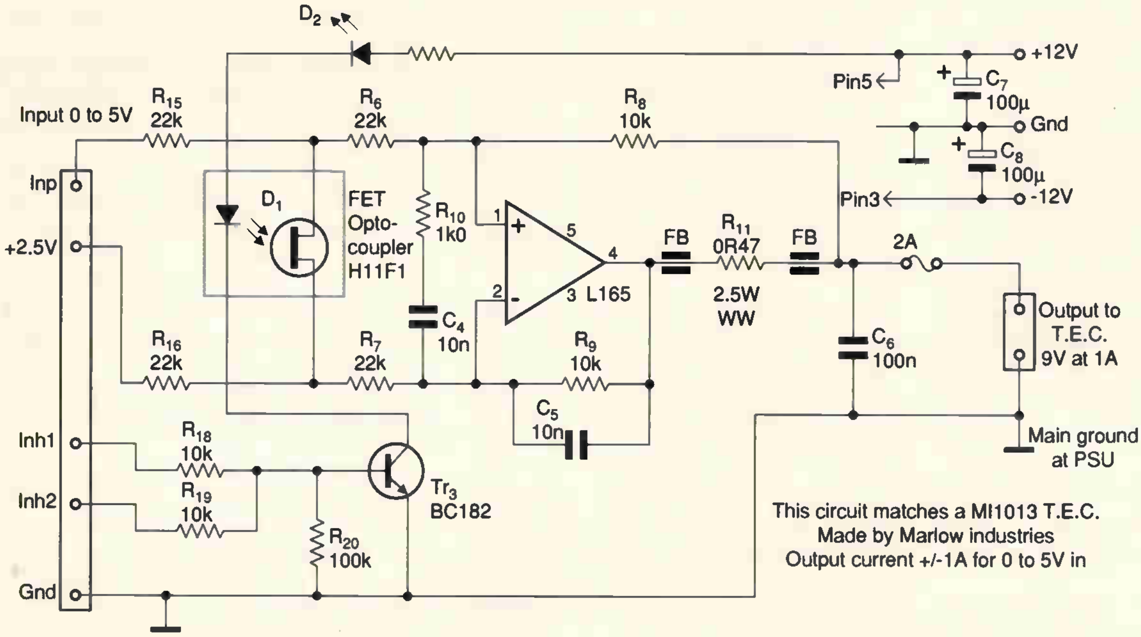


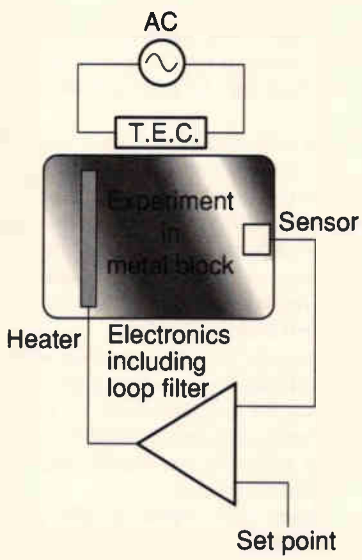
Fig. 4. Standard temperature control, to which is added the hypothetical thermoelectric cooler and very low frequency generator at the top of the diagram.

the frequency at which the loop gain becomes +1. At this lower frequency the metal block introduces less attenuation thus offsetting the extra losses put in by the low-pass filter. If the system still oscillates, nothing has been gained.

For this reason a simple *RC* filter may be no use since the phase lag tends to 90º at high frequencies. The lead-lag version shows more promise since the phase lag starts at almost zero at low frequencies, peaks at some value less than 90º, and then comes back towards zero at high frequencies.

Gain and phase behaviour of the filter is outlined in **Fig. 5.** Frequency f_1 is the initial –3dB cut-off frequency and f_2 is the break frequency for the lead-lag filter, where the filter gain tends to $R_2/(R_1+R_2)$.

The ultimate attenuation with the lead-lag filter is fixed by the ratio of the two resistors. As a result, the extra steady-state gain that can be introduced can be a factor no more than the inverse of the filter loss. On the main circuit



R_1 is chosen from one of the switched resistors $R_{9\text{-}12}$ and R_2 from R_{13} to R_{16}.

Choosing filter time constants

The trick is to choose the two filter time constants such that the maximum filter phase delay happens at a very low frequency, where the phase delay introduced through the metal is still small. Thus at higher frequencies, where the metal introduces a larger delay, the delay in the filter is past the maximum and coming back towards zero.

But now, the attenuation is sufficient to allow the amplifier gain to be increased while keeping the loop stable. Experts among you will recognise the similarity with loop filters for phase locked loops and amplifier compensation.

The options available on the circuit allow a time constant to set the f_1 break point between 40 and 400 seconds, and for f_2, from 2 to 20 seconds. Higher values can give better results, but there is a practical problem realising this sort of time constant reliably, in that electrolytic capacitors must be used.

The leakage current flows through the selected resistor producing a volts drop. For this reason standard aluminium electrolytics are unsatisfactory and low-leakage tantalum components should be used.

A suggested setting up procedure is as follows; recalling the oscillation period when the gain is set too high (filter out), set the f_2 time constant to be of the same order as the oscilla-

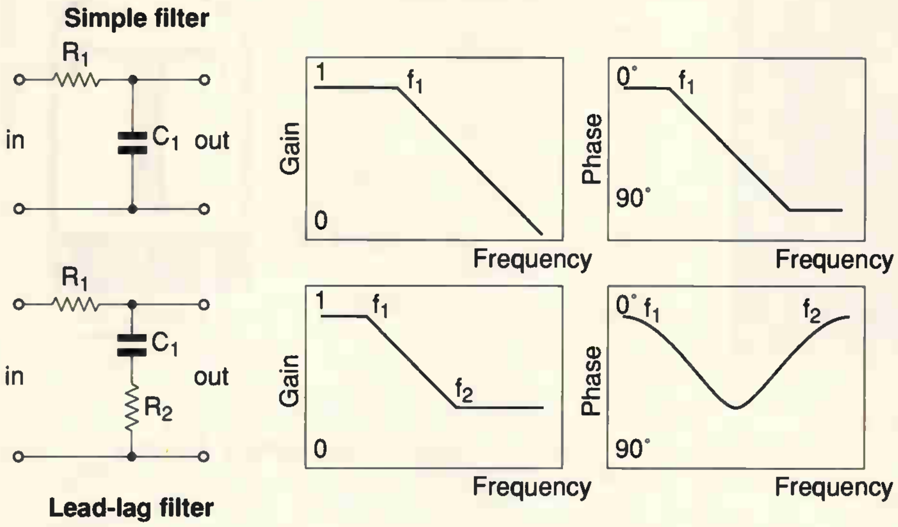


Fig. 5. Comparing gain and phase behaviour of the two low-pass filter options illustrates the benefits of lead-lag option.

tion period using resistors R_{13} to R_{16} on the main circuit. Select a 1MΩ resistor for R_{11}, setting the f_1 time constant to 200 seconds. Switching in the filter will hopefully cause the oscillation to die out and the gain can be increased.

Settling time and overshoot

Having set the gain and the filter it is important to have a look at the settling time and overshoot. This is particularly true if a thermoelectric cooler is being used with an ambient environment that can go above and below the set point.

If the gain is optimised with the thermoelectric cooler cooling hard, there is a risk the system will oscillate when the thermoelectric cooler is required to heat. The best way to check the dynamic performance is to log the temperature every few seconds as appropriate using a pc equipped with a data logging card.

Figure 6 illustrates the stability obtained over a three-hour run cooling a ccd to a set point of –38° C. The system uses the analogue controller described here with a *MI 3040* three-stage thermoelectric cooler driven by a unipolar output circuit capable of producing 0 to 5 amp for 0 to 5V from the controller.

Heat was removed from the thermoelectric cooler with a water loop running at 500cc/min, for a waste heat output of about 30W. Assuming good insulation around the rest of the assembly, the major source of temperature drift is actually the variations in tap water temperature.

The sensor was an *AD590*, and the switch settings selected were;

Gain resistor, R_{19} 100Ω
t_1 time const. resistor, R_{11} 1MΩ
t_2 time const. resistor, R_{14} 220kΩ

You can see that the ccd sensor temperature varied by about 10mK over the run, for a change in water temperature of approximately 1°C. It is difficult to see much in the way of correlation between the two curves. The temperatures were sampled every ten seconds, and directly plotted without any averaging.

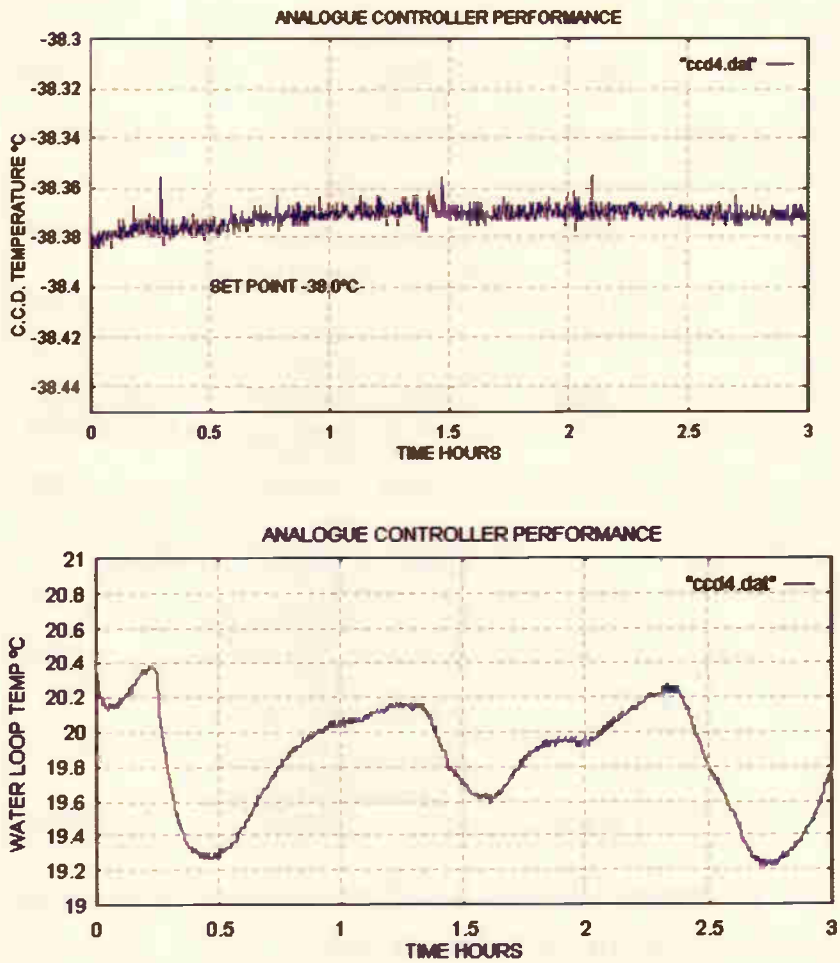


Fig. 6. These two plots illustrate the controller's performance over a three-hour run cooling a ccd to –38°C using a Peltier-effect device. CCD temperature is shown at the top, and cooling water temperature at the bottom. It is difficult to see any relationship between the two.

Final thoughts

If you need to monitor the temperature accurately on a pc then it is worth checking out the thermal stability of the a-to-d card. Some of the cheaper ones are less than wonderful. This can be a real problem as I have found out on occasions.

If there is any doubt, it is worth connecting up the 2.5V, –2.73V and the monitor ground to three spare channels and doing some differential checks with respect to the monitor ground. The two voltages should record the same potential with respect to the monitor ground, give or take one step on the a-to-d converter, all through the test.

The otherwise admirable BBC micro – still in service in many an undergraduate teaching lab – is particularly bad in this respect. It has a nice four-channel a-to-d converter spoilt by the use of three 1N4148 diodes as a 1.8V reference; that's a built-in drift of 6mV/°C! Replacement with a decent bandgap reference solves the problem.

Many items intended for running cold – such as integrating ccds and photodiodes – have a maximum rate of temperature change recommended by the manufacturer. This is easy to exceed, and risks destruction of the device through differential expansion or contraction.

When working with such devices, either turn down the set point knob in stages or it would be wise to engineer some limit to the maximum rate of cooling. ■

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PLUG IN AND MEASURE

TiePie introduces the HANDYSCOPE 2
A powerful 12 bit virtual measuring instrument for the PC

The HANDYSCOPE 2, connected to the parallel printer port of the PC and controlled by very user friendly software under Windows or DOS, gives everybody the possibility to measure within a few minutes. The philosophy of the HANDYSCOPE 2 is:

"PLUG IN AND MEASURE".

Because of the good hardware specs (two channels, 12 bit, 200 kHz sampling on both channels simultaneously, 32 KWord memory, 0.1 to 80 volt full scale, 0.2% absolute accuracy, software controlled AC/DC switch) and the very complete software (oscilloscope, voltmeter, transient recorder and spectrum analyzer) the HANDYSCOPE 2 is the best PC controlled measuring instrument in its category.

The four integrated virtual instruments give lots of possibilities for performing good measurements and making clear documentation. The software for the HANDYSCOPE 2 is suitable for Windows 3.1 and Windows 95. There is also software available for DOS 3.1 and higher.

A key point of the Windows software is the quick and easy control of the instruments. This is done by using:
- the speed button bar. Gives direct access to most settings.
- the mouse. Place the cursor on an object and press the right mouse button for the corresponding settings menu.
- menus. All settings can be changed using the menus.

Some quick examples:
The voltage axis can be set using a drag and drop principle. Both the gain and the position can be changed in an easy way. The time axis is controlled using a scalable scroll bar. With this scroll bar the measured signal (10 to 32K samples) can be zoomed live in and out.

The pre and post trigger moment is displayed graphically and can be adjusted by means of the mouse. For triggering a graphical WYSIWYG trigger symbol is available. This symbol indicates the trigger mode, slope and level. These can be adjusted with the mouse.

The oscilloscope has an AUTO DISK function with which unexpected disturbances can be captured. When the instrument is set up for the disturbance, the AUTO DISK function can be started. Each time the disturbance occurs, it is measured and the measured data is stored on disk. When pre samples are selected, both samples before and after the moment of disturbance are stored.

The spectrum analyzer is capable to calculate an 8K spectrum and disposes of 6 window functions. Because of this higher harmonics can be measured well (e.g. for power line analysis and audio analysis).

The voltmeter has 6 fully configurable displays. 11 different values can be measured and these values can be displayed in 16 different ways. This results in an easy way of reading the requested values. Besides this, for each display a bar graph is available.

When slowly changing events (like temperature or pressure) have to be measured, the transient recorder is the solution. The time between two samples can be set from 0.01 sec to 500 sec, so it is easy to measure events that last up to almost 200 days.

The extensive possibilities of the cursors in the oscilloscope, the transient recorder and the spectrum analyzer can be used to analyze the measured signal. Besides the standard measurements, also True RMS , Peak- Peak, Mean, Max and Min values of the measured signal are available.

To document the measured signal three features is provided for. For common documentation three lines of text are available. These lines are printed on every print out. They can be used e.g. for the company name and address. For measurement specific documentation 240 characters text can be added to the measurement. Also "text balloons" are available, which can be placed within the measurement. These balloons can be configured to your own demands.

For printing both black and white printers and color printers are supported. Exporting data can be done in ASCII (SCV) so the data can be read in a spreadsheet program. All instrument settings are stored in a SET file. By reading a SET file, the instument is configured completely and measuring can start at once. Each data file is accompanied by a settings file. The data file contains the measured values (ASCII or binary) and the settings file contains the settings of the instrument. The settings file is in ASCII and can be read easily by other programs.

Other TiePie measuring instruments are: HS508 (50MHz-8bit), TP112 (1MHz-12bit), TP208 (20MHz-8bit) and TP508 (50MHz-8bit).

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Total Package:
The HANDYSCOPE 2 is delivered with two 1:1/1:10 switchable oscilloscope probe's, a user manual, Windows and DOS software. The price of the HANDYSCOPE 2 is £ 299.00 excl. VAT.

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CIRCLE NO. 116 ON REPLY CARD

CIRCUIT IDEAS

Over £600 for a circuit idea?

New awards scheme for circuit ideas

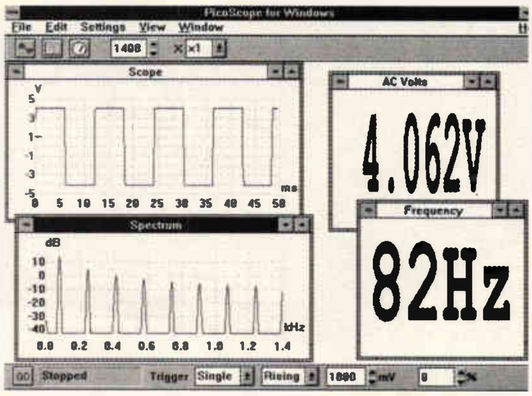
- Every circuit idea published in *Electronics World* receives £35.
- The pick of the month circuit idea receives a Pico Technology ADC42 – worth over £90 – in addition to £35.
- Once every six months, Pico Technology and *Electronics World* will select the best circuit idea published during the period and award the winner a Pico Technology ADC200-50 – worth £586.

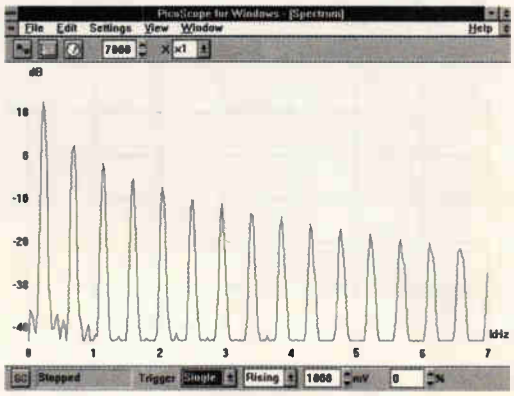
How to submit your ideas

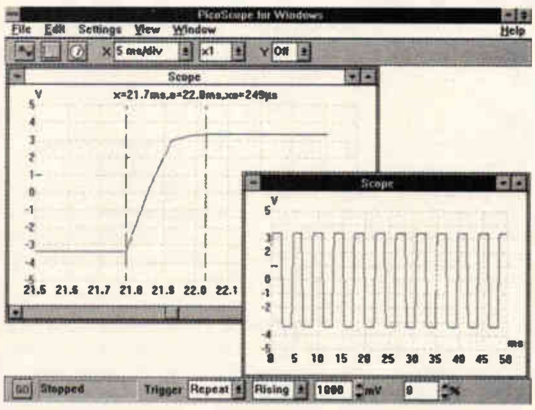
The best ideas are the ones that save readers time or money, or that solve a problem in a better or more elegant way than existing circuits. We will also consider the odd solution looking for a problem – if it has a degree of ingenuity.

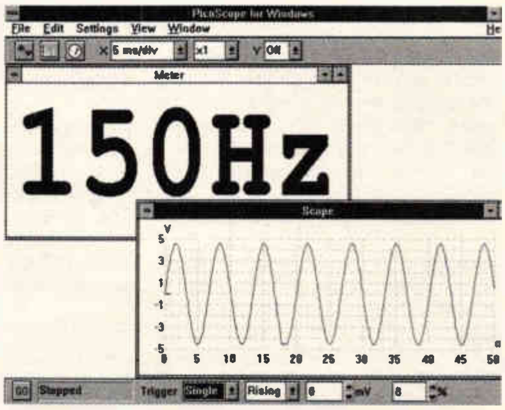
Your submission will be judged on its originality. This means that the idea should certainly not have been published before. Useful modifications to existing circuits will be considered though – provided that they are original.

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.





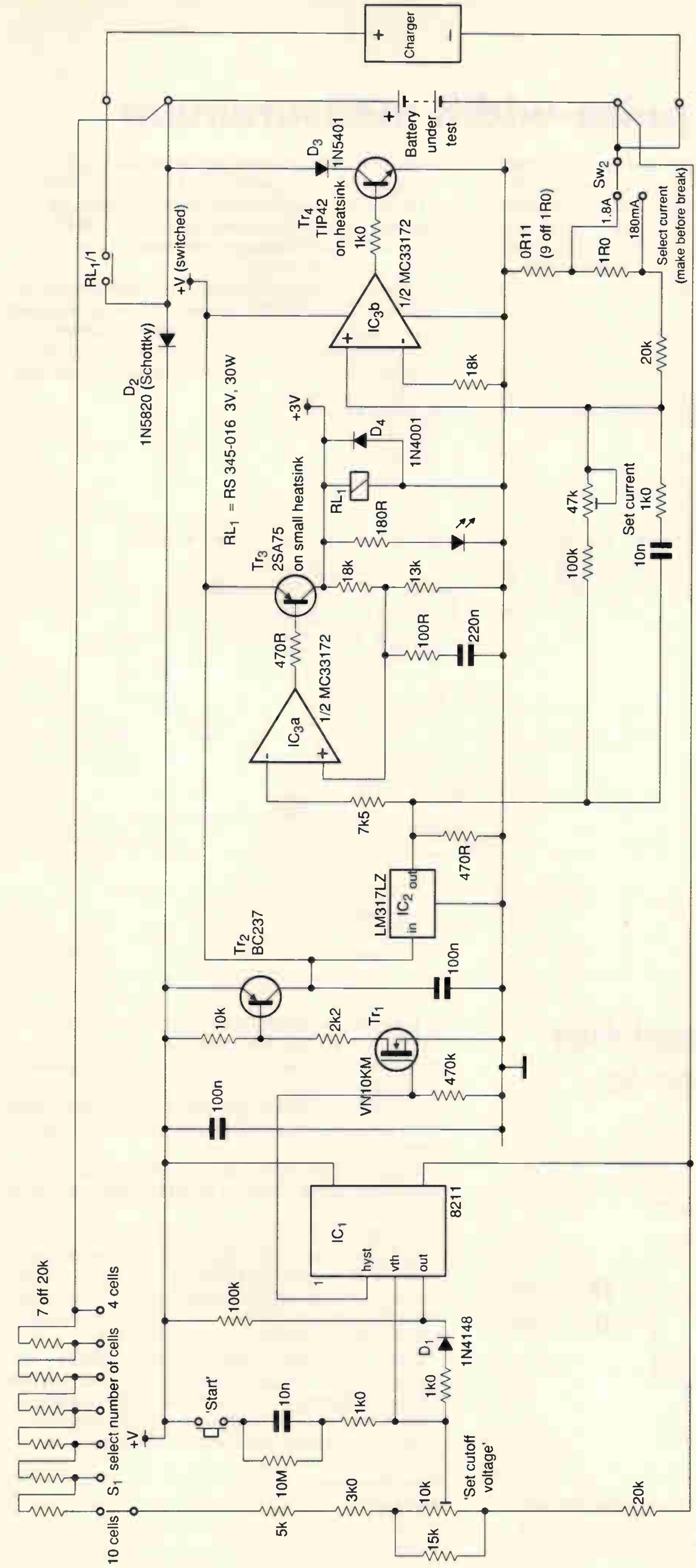




Turn your PC into a high-performance virtual instrument in return for a circuit idea.

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.



ADC-100 WINNER

NiCd capacity meter

Unusually, this capacity meter draws power from the NiCd battery under test. It discharges battery packs of from four to ten cells, a charger being connected after the test. As an example, a four-cell pack applies 3.5V to the circuit after allowing for current sensing and reverse polarity protection.

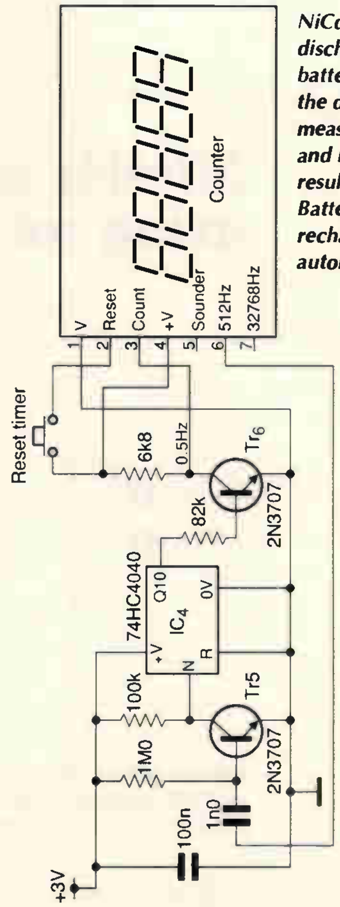
Latching comparator IC_1 starts the discharge when the start button is pressed, discharge ending when battery voltage falls to 1V/cell. In this case, the functions of output and hysteresis pins are reversed, since at low currents their threshold voltages differ by a few millivolts, the output being higher and preventing clean switching. Output current from the hysteresis pin is low and is amplified in Tr_1 and Tr_2, the resulting switch being capable of supplying over 100mA.

The discharge circuit is formed by IC_{3b} and Tr_4 and is switchable to 180mA and 1.8A by S_2. Very good power supply rejection in the *MC33172* (IC_{3b}) renders the discharge current largely immune to battery-voltage changes; the prototype's discharge current varies by less than 0.2% for a battery-voltage change of 4-25V.

To measure discharge time, the *Maplin FS13P* counter accepts its own 512Hz output, divided by the *4040* counter, to increment at two-second intervals while the battery is discharging. This is equivalent to 0.1 or 1mAh per count on the low and high current ranges. Power for the counter timer comes from IC_{3a} and Tr_3, which form a low dropout regulator to supply the relay.

Lower discharge currents may be used; PP9 batteries at 18mA can be handled, but the relay must be omitted. Higher currents or more cells will necessitate a heat sink for Tr_4.

John R Hunt
Middlesbrough



NiCd cell tester discharges batteries, timing the discharge to measure capacity and indicating the result digitally. Batteries are then recharged automatically.

Analogue pulse-width measurement

Measuring the width of a pulse can be done digitally, but needs digital counters or complicated circuits of one kind or another. This analogue arrangement gives a direct reading of pulse width in real time without clocks or digital circuitry of any kind. It is optimised to measure the 1-2ms, low duty-cycle pulses used in model control and, used with a 3.5-digit dvm with a 10Ω impedance, gives a 1μs resolution on the 2V range.

Voltage reference IC_1, a *ZN423* or similar, forms a switched current source with $Tr_{2,3}$, I_c being about 4mA, to charge C_1 during the input pulse. Monostable IC_2, a *555*, turns Tr_4 on once per input cycle, connecting R_c across C_1 to discharge it for the 4ms (t_{ref}). Consequently, C_1 charges each time to,

$$V_{out}=t_{in}I_cR_c/t_{ref},$$

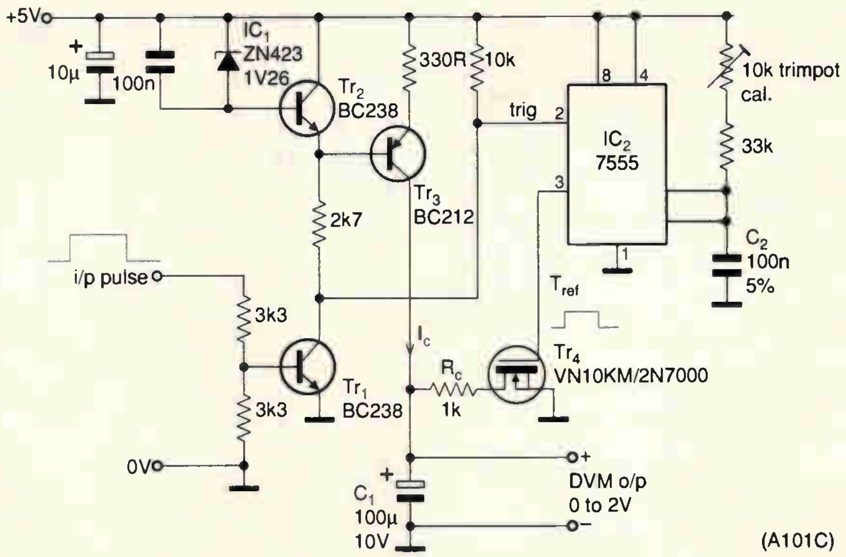
so that the charge gained in each input pulse equals that lost through R_c during every reference pulse from the 555.

The circuit is capable of precise conversion if high-precision components are used, particularly in the C_2 position; a 20-turn trimmer potentiometer to adjust t_{ref} allows adjustment to ±1μs.

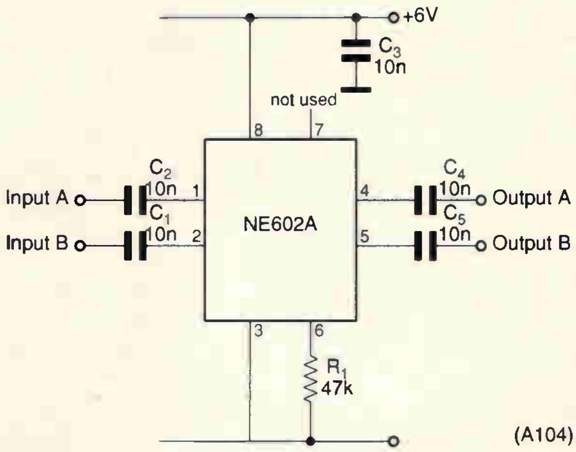
Linearity and supply rejection is 0.1% and accuracy is limited by the dvm itself. The value of C_1 is uncritical and a 10V tantalum bead can be used, although dielectric absorption creates a settling time in addition to the *RC* constant. Both ends of R_c are at high impedance and long pcb tracks should be avoided.

Anthony New
Bristol
(A101c)

Pulse-width-to-voltage converter provides direct readout of pulse width on a dvm, to an accuracy better than that of the dvm itself.



50MHz amplifier using mixer ic



Unusual role for a 602A mixer ic, operating here as a 10.7MHz amplifier, single-ended in this case. There is provision for a balanced mode of operation.

Unusually, this 50MHz amplifier uses the *NE602A* or *SE602A*, intended as a double-balanced mixer. Permanently unbalancing the mixer causes it to operate in linear mode, the advantages being that the circuit is inherently well balanced and matched and has internal bias circuits. It also takes only 2mA and uses six other components.

Inputs are taken to pins 1 and 2, normally used as mixer inputs, and may be connected as single-ended or balanced input.

Since there is no local oscillator, the lo input is biased with R_1, which unbalances the top four transistors in the mixer, so that two are on and two off. The top transistors then act as buffers and the amplifier operates as a cascode type, which has very good reverse isolation and stability. Output comes from either output on pin 4 or 5.

Using transformers at input and output provides a balanced circuit with gain enhanced by 3dB and even-order harmonics suppressed – useful in a wideband amplifier.

My prototype is a 10.7MHz design, giving a gain of 22dB in single-ended mode with source and load resistances of 1.5kΩ.

Peter Goodson
Bracknell
Berkshire
(A104)

Pc-controlled video multiplexer

Using a *MAX498* quad, single-pole/double-throw RGB video switch, which is provided with 250MHz video buffer amplifiers, we blended the RGB signals with composite video and used differential inputs to try to increase the distance we could send the signal over 50Ω cable.

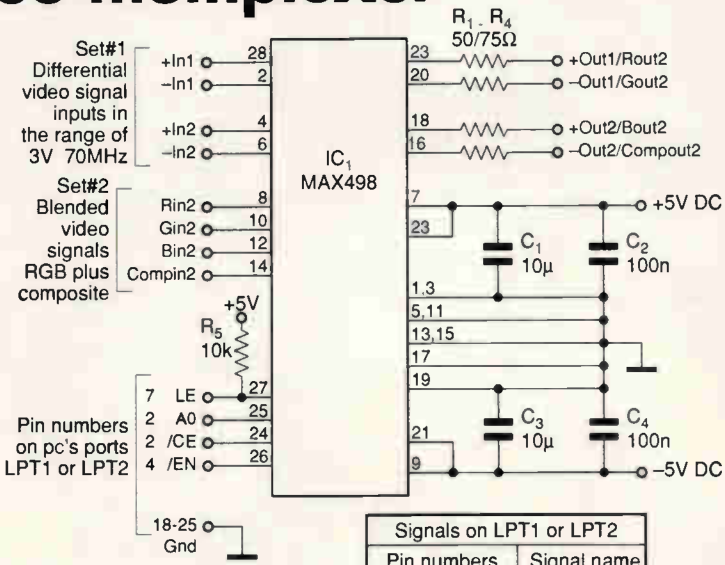
The blending or the use of differential inputs covered most of the video signal types we use. Limitations were that, with differential inputs, we could only multiplex two sets of differential composite video instead of four single-ended signals; and in composite mode we could use three lines for RGB and a composite signal, which coped with half the transmission distance.

Operation is controlled by a pc, to whose LPT1 or LPT2 ports it is connected. A QBasic program controls the output and will put the output in a high-impedance state, so that further signals may be added at the output from, for example, a number of security cameras. Data lines D_0, D_1, D_2 and D_5 at the printer are used.

Power needed is ±5V/100mA, well regulated and with the four capacitors shown to reduce noise. Use resistive matching on all cable terminations – in and out.

We have sent signals to 100m without much degradation up to 70MHz, with a 3Vpk-pk differential input; small signal bandwidth is over 259MHz.

Shyam Sunder Tiwari
Kalpakkam
India



Signals on LPT1 or LPT2

Pin numbers	Signal name
Pin2	D0
Pin3	D1
Pin4	D2
Pin7	D5
Pins 18 to 25	Ground

(A102)

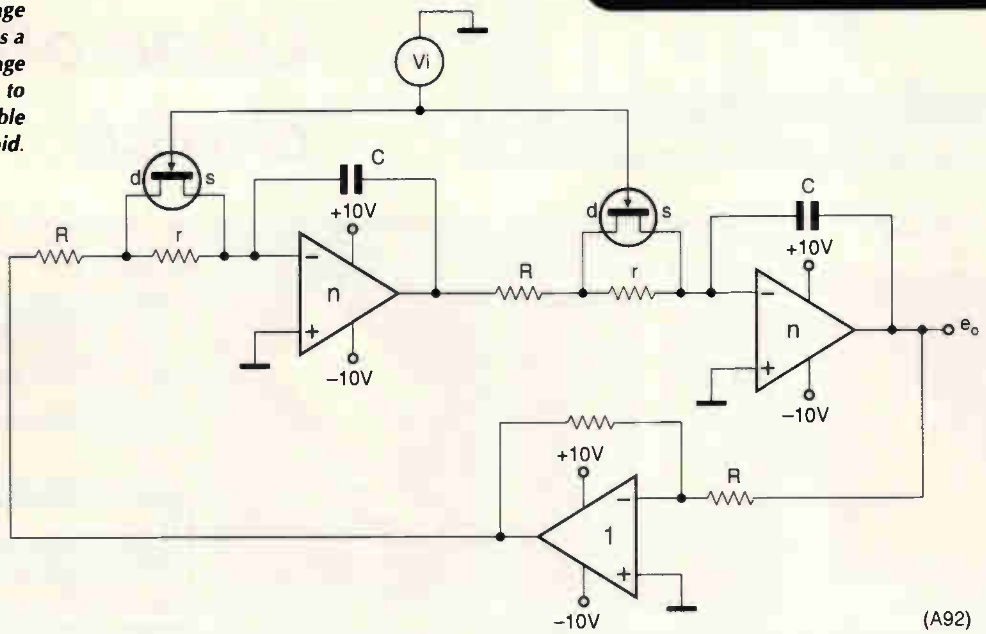
Using a MAX498 to blend RGB signals with composite video or using differential inputs handles most types of video signal, signal being driven about 100m along 50/75Ω cable.

QBasic listing for the 70MHz blend multiplexer based on the MAX498.

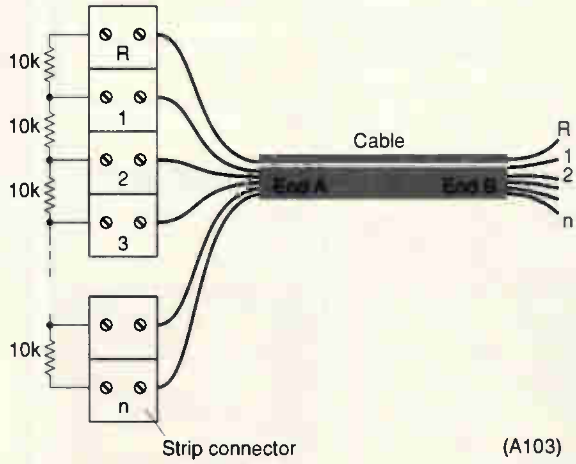
```
        REM Base Address of LPT1 or LPT2
START:  LET BaseAddress = &H278
        REM Base address of the LPT2 port is set to 0278H
        REM Data bits on the LPT1/LP2 Parallel Port
        REM LPT port D7   D6   D5   D4   D3   D2   D1   D0
        REM used as  x    x    LE   x    x    /EN  /CE  A0
        REM x are don't care bits
        REM ***************************************************
        REM Initialize the Port Printer LPT1 or LPT2 to 026H.
INIT:   OUT BaseAddress, &H26: REM initialization completes.
        REM Condition after initialization: A0 (D0) = 0 points
        REM  to the Set#1 of the differential inputs,
        REM  MAX498 chip is disabled i.e. /CE (D1) = 1,
        REM  the output is also disabled i.e. /EN (D2) = 1,
        REM  the latch is active (high) which holds the status
        REM  i.e. LE (D5) = 1, others data bits are don't care.
        REM **************************************************
        REM Activate Set#1 differential signals at the outputs
        REM  and use the LATCHED OUTPUT control mode.
        REM  Latch enable active i.e. LE (D5) = 1. Inputs from
        REM  Set#1, i.e. A0 (D0) = 0 which selects the Set#1
        REM  of the differential inputs. /EN (D2) is set to low.
        REM  /CE is pulsed to low to strobe control signals
        REM  and must remain high.
LOUT1:  OUT BaseAddress, &H22
        OUT BaseAddress, &H20
        OUT BaseAddress, &H22
        REM The outputs are from Set#1. Differential outputs.
        REM The control inputs A0 and /EN are latched.
        REM **************************************************
        REM Activate Set#2 RGB plus composite signals at outputs
        REM  and use the LATCHED OUTPUT control mode.
        REM  Latch enable active i.e. LE (D5) = 1. Inputs from
        REM  Set#2, i.e. A0 (D0) = 1 which selects the Set#2
        REM  of the differential inputs. /EN (D2) is set to low.
        REM  /CE is pulsed to low to strobe control signals
        REM  and must remain high.
LOUT2:  OUT BaseAddress, &H23
        OUT BaseAddress, &H21
        OUT BaseAddress, &H23
        REM The outputs are from Set#2, RGBs and a composite.
        REM The control inputs A0 and /EN are latched.
        REM **************************************************
        REM Activate Set#1 differential signals at the outputs
        REM  and use the TRANSPARENT OUTPUT control mode.
        REM  Latch disabled i.e. LE (D5) = 0. Inputs from
        REM  Set#1, i.e. A0 (D0) = 0 which selects the Set#1
        REM  of the differential inputs. /EN (D2) is set to low.
        REM  /CE is pulsed to low to strobe control signals
        REM  and must remain high. However, /EN = 0 can enable
        REM  the outputs or /EN = 1 disable the outputs.
TOUT1:  OUT BaseAddress, &H02
        OUT BaseAddress, &H00
        OUT BaseAddress, &H02
        REM The outputs from Set#1 differential are now active.
        REM  /EN now can enable or disable the analog outputs.
DIS1:   OUT BaseAddress, &H06 : REM outputs disabled
ENB1:   OUT BaseAddress, &H02 : REM outputs enabled
        REM **************************************************
        REM Activate Set#2 RGB + composite at the outputs
        REM  and use the TRANSPARENT OUTPUT control mode.
        REM  Latch enable inactive i.e. LE (D5) = 0. Inputs from
        REM  Set#2, i.e. A0 (D0) = 1 which selects the Set#2
        REM  of the inputs. /EN (D2) is set to low.
        REM  /CE is pulsed to low to strobe control signals
        REM  and must remain high. However, /EN = 0 can enable
        REM  the outputs or /EN = 1 disable the outputs.
TOUT2:  OUT BaseAddress, &H03
        OUT BaseAddress, &H01
        OUT BaseAddress, &H03
        REM The outputs from Set#2 are now active.
        REM  /EN now can enable or disable the analog outputs.
DIS2:   OUT BaseAddress, &H07 : REM outputs disabled
ENB2:   OUT BaseAddress, &H03 : REM outputs enabled
```

ADC-42 WINNER

Fig. 1. One voltage input controls a two-stage integrator to produce a variable sinusoid.



A handful of anonymous cores in a cable may be sorted out with this simple scheme and a multimeter.



Cable core identifier

A very simple arrangement of resistors allows unmarked cores in a multicore cable to be easily identified.

Connect resistors of equal value to a strip connector at the A end, as shown, with the metallic shield or an external reference to R. At the B end of the cable, assuming the resistors are of 10kΩ each, a multimeter between the shield and one of the cores will read 10kΩ, which shows that it is No 1. A reading of 20kΩ shows No 2, and so on. An absence of any reading indicates an open circuit and if several readings are the same, there are shorts.

Satish Kumar
Shuaiba
Kuwait
(A103)

Op-amp vco and binary fsk generator

The circuit shown can be characterised by the differential equation,

$$d^2y/dt^2+n^2y=0$$

with $y(0)=1$, the solution is $\cos nt$, which is found by the circuit and results in a sinusoidal output e_o at the frequency $f=n/(2\pi)$.

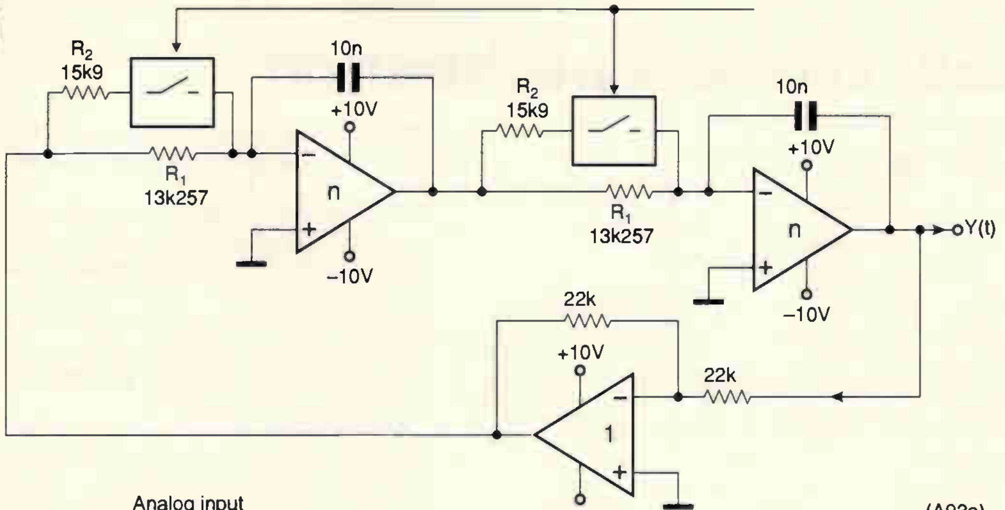
Frequency variation is obtained by varying the drain/source resistance of the two fets, thereby controlling the *RC* constant of each integrator, the resistive component of which is $1/(R_{eq}+R)$, R_{eq} being the parallel combination of r and the drain/source resistance $R_{ds(on)}$.

Available fets have a range of $R_{ds(on)}$ of 100Ω to 100kΩ, achieved by varying the control input V_i to each fet. Since each source is at virtual ground, the same voltage may be used with no cross loading effects.

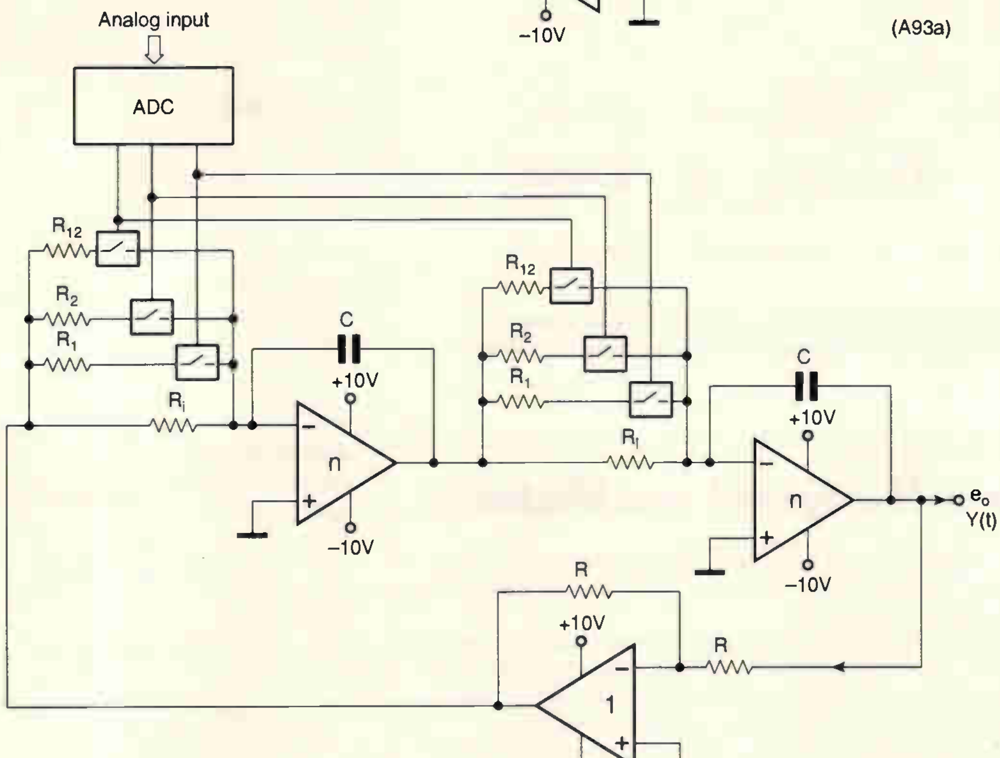
Figure 2 shows a variation on this oscillator which performs binary frequency-shift keying. Here, the fets are replaced by analogue switches controlled by a voltage input. With the switch open, each resistive arm to the integrators is 13.257kΩ and, with the switch closed, 7.23kΩ, giving frequencies at the output of 1.2kHz and 2.2kHz. For start-up, arrange power-on switching to store 1V at C_1 and zero at C_2.

One can envisage the use of more than one switch at each op-amp to give a range of switched frequencies – three switches at each integrator would give a range of eight frequencies and twelve, perhaps controlled by a 12-bit a-to-d converter, would give 4096 frequencies.

K Balasubramanian
NSS College of Engineering
Palakkad
India
(A92/93)



(A93a)



(A93b)

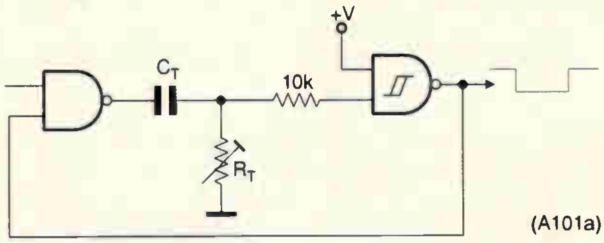
Fig. 2. Replacing fets by analogue switches gives a binary fsk output and the possibility of a switched-frequency generator with as many outputs as desired.

Improving supply rejection in cmos monostables

Using a Schmitt trigger as the output device in a cmos monostable, as in **Fig. 1**, in an effort to obtain clean transitions may cause problems with supply rejection, since threshold voltage in gates such as the *CD4093* do not scale correctly with supply voltage, so that pulse period can vary by up to 20%.

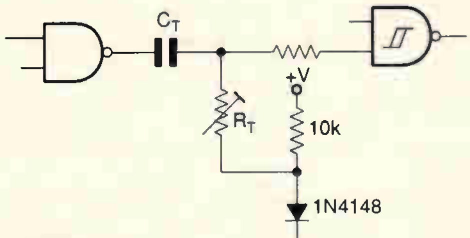
A small constant voltage in series with the timing resistor, shown in **Fig. 2**, compensates for the variation. Ideally, this would be variable to eliminate the period variation completely, but in practice, the forward voltage drop of a diode is about right; using a 4093, the variation was well-nigh perfect over the 4.5-7V range and only 3% out at 10V.

Anthony New
Bristol
(A101a/b)



(A101a)

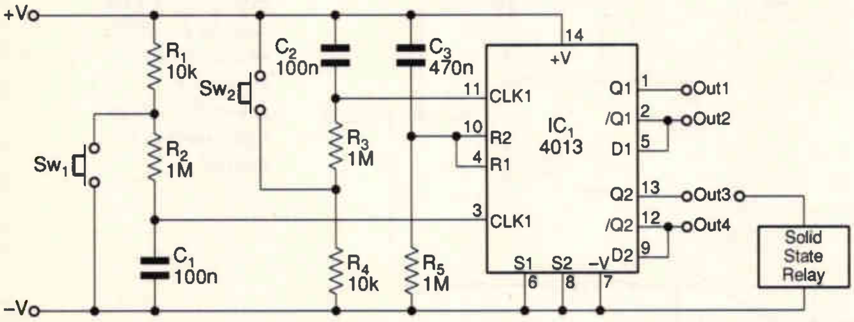
Fig. 1. Schmitt monostable may suffer poor supply voltage rejection

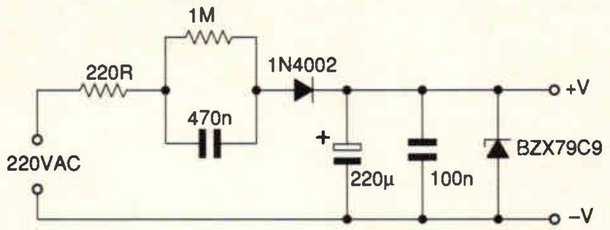


(A101b)

Fig. 2. A constant voltage, here obtained from a diode forward drop, almost completely eliminates the effect.

Antibounced, clickless pushbutton





(A108)

Useful for remote-controlled, independent two-way switching or the switching of power without the clicks associated with ordinary relays, this debounced circuit takes minimal current.

This double push-button power switch is, I believe, cheaper than some designs and uses fewer components.

Two switches, Sw_1 and Sw_2, control a pair of D-type flip-flops contained in a *4013*, $R_{1,4}$ limiting the current and $R_{2,3}$ $C_{1,2}$ forming debounce components. At switch-on, R_5 and C_3 ensure that reset is applied to the flip-flops such that Out1 and Out 3 are forced low. If switch Sw_1 is pressed and released, Out1 and Out 2 change state, while Sw_2 performs the same function for Out3 and Out4.

I have used the Sw_2 section of the circuit to drive the input transformer of an audio amplifier power supply, by way of a solid-state relay, without the audible noise from an electromechanical relay and with the facility of remote control. The circuit draws very little current.

Santo Camonita
Catania
Italy
(A108)

Two lf crystal oscillators

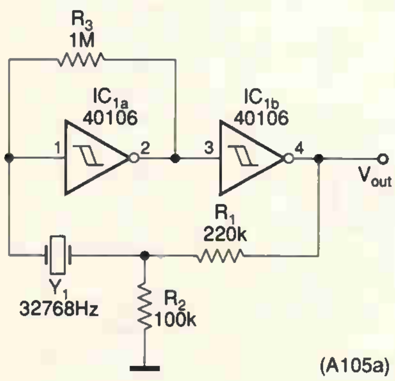
If you cannot lay your hands on the usual *4069* hex inverter, a pair of Schmitts from a *40406* or a *74HC14* make a simple and effective crystal oscillator.

The circuit in **Fig. 1** is for a crystal with a required drive power of 1µW; a watch crystal, for example. Resistors $R_{1,2}$ reduce the drive to this level; increasing R_3 would have the same effect, but would render the circuit sensitive to stray capacitance.

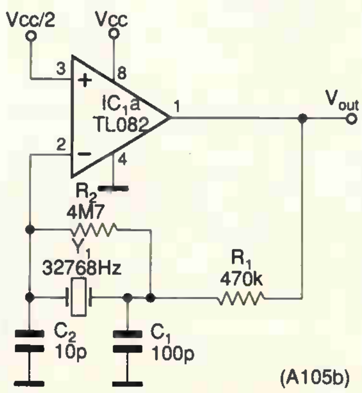
Figure 2 shows an alternative using a fast op-amp such as the *TL082*. Here, R_1 and C_1 set the drive power and R_2 the bias point. In the 40106 circuit, with a 12V supply, R_2 is 100kΩ; for a 5V supply, use 220kΩ.

M Ferrari
Guastalla
Italy
(A105)

Two simple circuits for low-frequency crystal oscillators using low-power crystals.

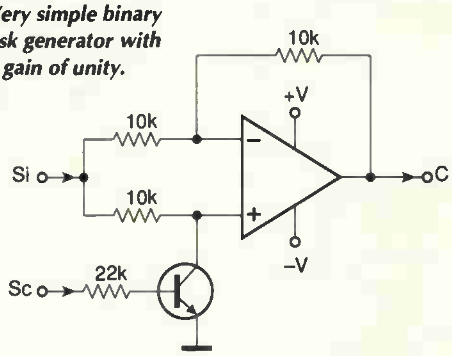


(A105a)



(A105b)

Very simple binary psk generator with a gain of unity.



Binary phase-shift keying generator

In response to a control signal S_c, the output of the circuit shown is either in-phase or out of phase with the input sinusoid, thereby providing a simple means of binary psk.

If S_c is high, the transistor conducts and grounds the non-inverting input, so that the amplifier has a gain of –1 and the output is 180° out of phase with the input. When S_c is taken low, the gain from the non-inverting input is 2 and that from the inverting input –1, so that the effective gain is 1 and the output is in phase with the input.

K Balasubramanian
NSS College of Engineering
Palakkad
India
(A94)

The car electronics market is being driven by convenience and safety as the vehicle manufacturers push for lower power, smaller and more integrated products to interface with the management system. Lana Josifovska reports.

Is your car battery dead?

The increasing complexity and sophistication of in-car electronic systems inevitably means that the vehicle's electrical system requirements are re-evaluated as never before. Convenience and safety are the main forces driving the research into new, and some might say revolutionary, electrical supply concepts such as two batteries instead of one, alternators that supply more than one voltage, on-board electrical systems that are divided into a 12V and a 36V network, belt-less engines with electrically driven ancillary components and new forms of alternators and starters.

The requirement is for a system which provides not only an ever-increasing level of electrical output, but also a stable voltage supply for the growing list of electrical and electronic systems in the modern vehicle.

One way of achieving this is by deploying more than one battery.

Bosch – a pioneer in the development of such systems – is currently working on a vehicle electrical system with two 12V batteries: one for starting the engine and the second to supply the on-board electrical system. This allows an increasing amount of power to be available for starting the engine while ensuring a reliable supply of electricity without voltage drops to the vehicle's electrical system.

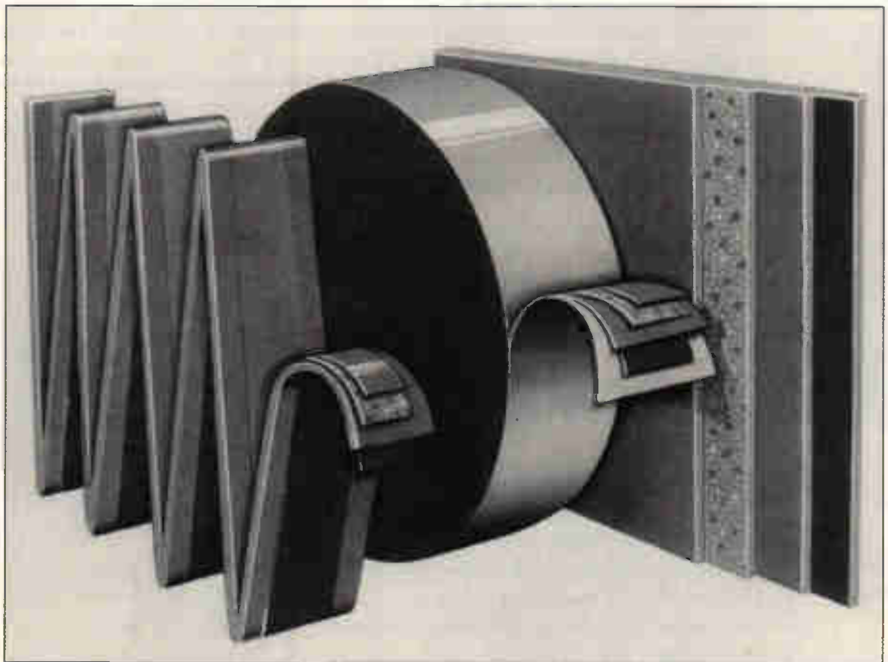
However, the two batteries use different technologies to best fulfil their respective tasks. Their separate power networks are linked to different systems in the vehicle allowing the starter battery, for example, to receive priority or other on-board systems to be supplied with a stable voltage. The use of new Lithium Polymer (LiPo) technology in the dual battery system can even reduce the car's weight compared to a vehicle with a single-battery system.

The drive toward higher voltage systems

There is also a drive for a dual 12V/36V network, to supply high-load consumers via a divided electrical system. A 42V alternator supplies those consumers in the car that have a high electrical power requirement and recharges the 36V battery. The 42V is also known as the 'charging voltage' whilst the 36V the 'starting voltage'. The remaining consumers on the second electrical circuit are supplied via a suitable voltage transformer. This solution presents no problems for the continued use of conventional 12V components.

Even though automotive manufacturers are trying to move towards higher voltages that are typically multiples of the standard 12V, such as 36V, some believe this transition will be slow and arduous.

"The direction that we are being driven by vehicle manufacturers globally is for lower power, smaller and



At the turn of the next century... vehicle batteries will no longer be bulky and cumbersome but light and pliable enough to fit in the smallest spaces in the car. A flat battery could fit in the car headliner (roof) for example. Here are shown the Delphi Freedom battery and the solid, flat and flexible LiPo battery of the future.

more integrated products to interface with the management systems. With so many electrical and electronic components prevalent within vehicle systems - typically 20 per cent of the value of a car - battery power drain is a major problem. Unless there are specific requirements, such as very low temperature starting or heavy duty requirements, then a 12V system will prevail.

Problems with using 24V

A 24V system has heat dissipation issues and running dual voltage systems is cumbersome," said David McCallum, business manager at Panasonic Industrial Car Electronics group.

Panasonic is building sensors to specification for automotive manufacturers and OEM suppliers in the US, Japan and Europe. Panasonic produces switches and sensors amongst many other components and systems used in vehicles.

The spokesperson at Delphi, a major supplier of components to the automotive industry, agrees that the switch to dual-voltage systems may take some time: "A higher voltage system will allow different functions in the car such as electrically heated catalysts, drive-by-wire, electric water pumps and others. The 12V system will be supplied via a DC/DC converter but the idea is to eliminate the 12V even though this may not happen in the foreseeable future, because of all the (negative) issues related to the higher voltage systems such as high heat dissipation and so on."

Delphi is working with a variety of component suppliers and automotive makers to introduce a dual-power system by the year 2005. At present the company is trying to build a dual-voltage test system in Germany.

For dual-voltage systems, the LiPo technology may prove to be ideal. Battery makers such as Delphi and AEA Technology are adapting solid state lithium polymer batteries for use in vehicles.

"For a 36V starting voltage, which is three times the standard 12V battery, using lead-acid can be very sizeable and heavy. LiPo will match up to the requirement of the 36V battery with its energy density but it will also add flexibility and its considerably smaller size," said Gary Ballard, manager of the battery energy department at the Delphi technical centre in Luxembourg.

LiPo – three times the power density

The LiPo batteries that are now under development, offer three times the power/energy density currently available from conventional lead acid batteries.

Lithium ion chemistry with solid polymer electrolyte allows the new batteries to be smaller and lighter, to last the life of the vehicle and to be manufactured to any shape the automotive designer requires.

Polymer state batteries, being very pliable can be cut to shape to fit in the most crammed of spaces such as the rooftop headliner in the passenger cabin or underneath the seats.

"Vehicle manufacturers are looking for flexible packaging, decreased volume resistance to heat and the ability to place the battery in small spaces in the car," said Ballard. ■

Driving power... Battery technology is also paramount in the plans for the environmentally friendly all-electric vehicle. The US is committed to the issue of zero-emission cars and is trying to legislate the issue. As of the year 2002 every US car maker is expected to manufacture zero-emission vehicles to up to two per cent of its total output. Europe has not followed suit but expected to do so sometime in the future. At present there are plenty of European research projects and battery developments concerning electric vehicles. Amongst these are the work of Mercedes-Benz in conjunction with AEA Technology; the UK-based Bluebird Electric project and the University of Sheffield's Electric 3000 racing car, the electric equivalent to the Formula 3000. Pictured with the Electric 3000 racing car are Jordan driver, Damon Hill with Shahnaz Pakravan of BBC TV's Tomorrow's World at the wheel.

Mosfet analogue gating

Bryan Hart explains the problems involved in designing mosfet transmission gates and shows how to get the best out of them.

Operation of a mosfet in a logic gate is relatively straightforward. It normally presents few problems – even to those who have only a sketchy knowledge of mosfet theory. This is because the device can be modelled as a simple, though imperfect, on/off switch.

In the case of a basic cmos inverter stage, a device that is off passes only a small drain-to-source 'leakage' current – typically less than a microamp: for a device that is on, the magnitude of the drain-source voltage does not exceed a few millivolts – typically, 5.

In contrast, the behaviour of a mosfet in a 'transmission gate' is more complex and can cause difficulties. This is because the device can no longer be regarded as a simple switch in this application. The problem is made worse by the superficial treatment given to the topic in the literature.

This review adopts an evolutionary approach and uses a recently proposed mosfet model to clarify the operation of a cmos transmission gate.

Transmission gate characteristics

A 'transmission', 'linear' or 'analogue' gate is one in which a signal at an input terminal appears – ideally – unchanged in shape, at an output terminal during a time interval selected by a gating, or 'strobe', pulse applied to a control terminal. A change in polarity is acceptable and there may be a change in scale through attenuation or amplification.

The transmission gate has been used in a wide miscellany of signal processing applications that includes: pulse radar echo selection; nuclear instrumentation; switch for amplifier gain and polarity control, integrator resetting, digital-analogue converters and active switched-capacitor filters.

Figure 1a) shows a transmission gate connected in series with a load resistor R. The instantaneous values of the input, output and control signals are v_I, v_O, v_C respectively.

Figure **1b)** shows the ideal dc transfer characteristics for the two states of the control signal. Section i) of the illustration corresponds to the off state and ii) to the on state.

Characteristic ii) is shown, arbitrarily, as having a positive slope but it may be negative, indicating signal inversion. Also, the magnitude of the slope can be less than unity, in the case of attenuation, or greater than unity, in the case of amplification.

The general characteristics of a practical circuit are shown, similarly labelled, in Fig **1c**). Characteristic ii) passes through a point V_I=0, V_O=V_P offset from the origin: furthermore the line is no longer straight.

Figure 2 summarises the operation of the gate circuit, for both ideal and non-ideal characteristics when the input signal, Fig. **2a**), is a train of pulses of arbitrary shape and a control or gating waveform, **Fig. 2b**), is applied. Figs **2c**) and **2d**) show the output waveforms for the characteristics of Figs 1b) and 1c) respectively.

In Fig. 2d), the output is distorted and sits on a 'platform' or 'pedestal' of height V_P. Additionally, there are gating transients or 'spikes', S_1, S_2 at the beginning and end of the gating pulse. There are two reasons why these appear. One is capacitive feedthrough from the gate terminal to the output, i.e., via the stray and inter-electrode capacitance C in Fig. 1a). The other is finite rise and fall times and non-coincidence of gating pulse edges in circuit schemes employing complementary gating pulses.

The transmission gate has been implemented using a variety of semiconductor devices. However, the cmos transmission gate is attractive in

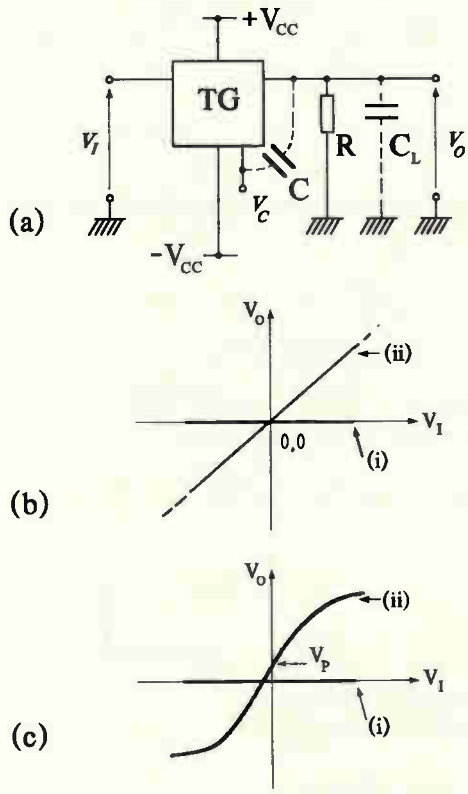


Fig. 1.
a) A series connected transmission gate.
b) Ideal transfer characteristics for a): i) when the transmission gate off, and ii), when the gate is on.
c) Practical counterpart of b).

a wide range of applications because it is easy to drive and is free of pedestal. To understand its origin and operation it is necessary to examine first the operation and limitations of a transmission gate using a single mosfet.

Single mosfet transmission gate

Figure 3 shows a single n-channel enhancement-mode mosfet used in the circuit of Fig. 1a). The electrode potentials are shown as dc quantities for the analysis that follows.

Substrate B is biased so that, under all conditions, V_B is less than V_I and V_O. In that case the source-substrate and drain-substrate junctions of the mosfet are reverse-biased and the mosfet can be represented by the recently proposed large-signal model[1] in **Fig. 4a)**. This applies also for alternating signals if inter-electrode capacitances are added.

Battery V_{TN} models the threshold voltage. Diodes D_F and D_R are shown boxed[2] because they have the ideal I/V characteristics of Fig. 4b).

The device equations for $V_F>0$ ($V_{GS}>V_{TN}$) and $V_R>0$, ($V_{GD}>V_{TN}$) are,

$$I_F=K_NV_F^2=K_N(V_{GS}-V_{TN})^2 \quad (1)$$

$$I_R=K_NV_R^2=K_N(V_{GD}-V_{TN})^2 \quad (2)$$

Since $I_G = 0$,

$$I_D=-I_S=I_F-I_R=K_N(V_F^2-V_R^2) \quad (3)$$

Device design parameters K_N, V_{TN} can be found experimentally from a plot of $\sqrt{I_D}$ versus V_{GS} with V_{DG} at 0V. The slope of the straight line section gives $\sqrt{K_N}$ and the extrapolated intercept on the V_{GS} axis gives V_{TN}.

The model is attractive in showing clearly the device dc operating mode when arbitrary terminal voltages are applied. Once that mode has been established the model can be reduced to a simpler equivalent form, as will become apparent in what follows.

If, in Fig. 3, V_G is connected to a negative bias source $-V_{GG}$, and if V_I is greater than $-(V_{GG}+V_{TN})$ then D_F and D_R in the model of Fig. 4a) are cut off and $I_F=I_R=I_D=0$.

The transmission gate in Fig. 3 is cut off, its idealised equivalent circuit

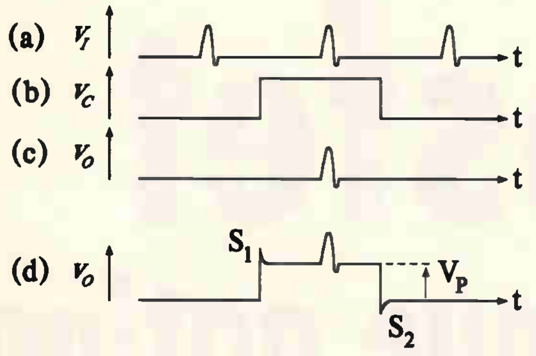


Fig. 2. Transmission-gate operation:
(a) Input pulse train.
(b) Gating pulse.
(c) Ideal output.
(d) Possible practical output. Note that V_P is exaggerated for clarity.

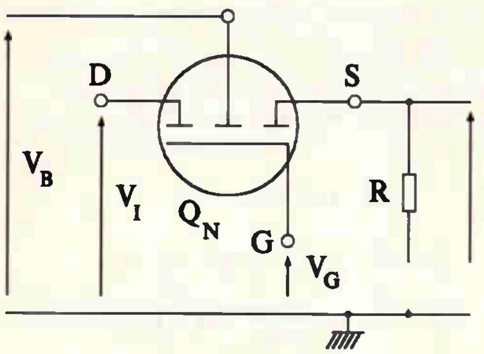


Fig. 3. Transmission gate using an n-channel enhancement-mode mosfet.

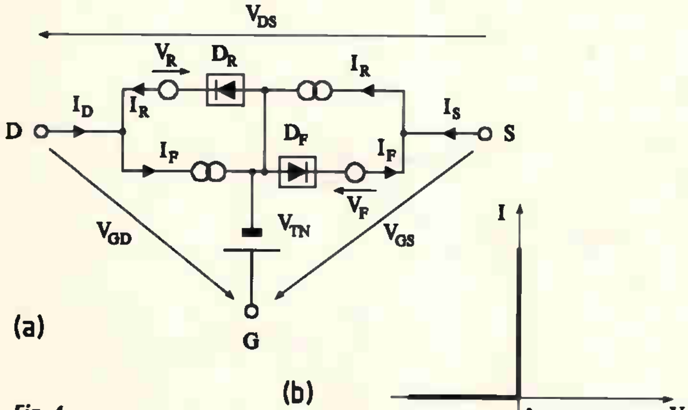


Fig. 4.
a) DC mosfet model for Q_N in Fig. 3.
b) Characteristics of diodes D_F, D_R.

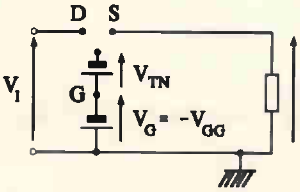


Fig. 5. Transmission-gate equivalent circuit for off state.

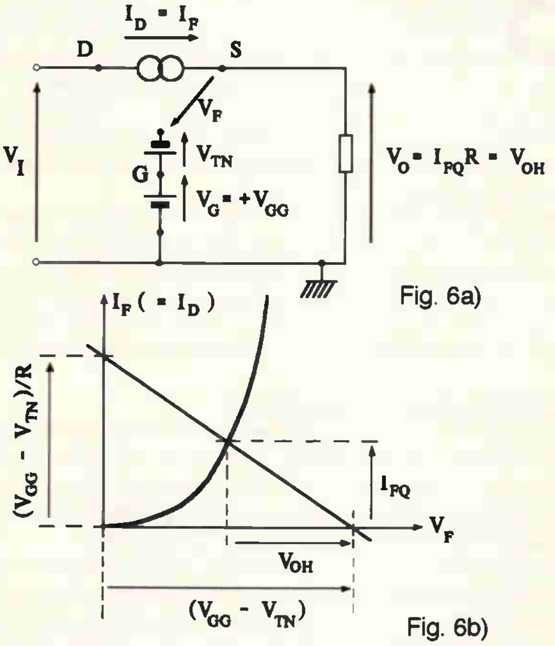


Fig. 6. a) Transmission-gate equivalent circuit for when the gate is on and $V_I>(V_{GG}-V_{TN})$.
b) Load line construction for operating current, I_{FQ}.

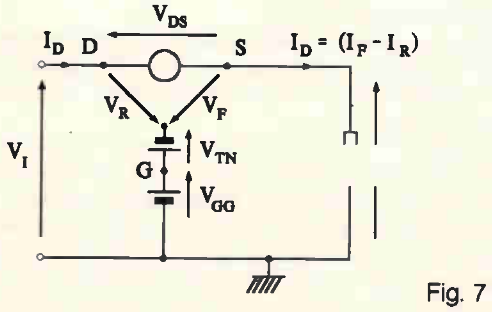


Fig. 7. Transmission-gate equivalent circuit for when the gate is on and $V_I<(V_{GG}-V_{TN})$

being indicated in **Fig. 5**. In a refinement to the mosfet model under discussion, two diodes arranged in series-opposition can be connected between the terminals D and S to allow for the small cut-off leakage current that occurs with real devices.

Consider next what happens in Fig. 3 when V_G is $+V_{GG}$. Using the model of Fig. 4a) it is clear that D_F conducts while V_{GG} is greater than V_{TN}. But there are two cases to consider for D_R. One is that D_R is off when V_I is greater than $V_{GG}-V_{TN}$ and the other is that it is on while V_I is less than $V_{GG}-V_{TN}$.

The equivalent transmission gate circuit for the first case is shown in **Fig. 6a)**. The mosfet behaves as a voltage controlled current generator characterised by equation 1.

The gate-source loop gives,

$$V_O=I_FR=V_{GG}-V_{TN}-V_F \tag{4}$$

Operating current I_{FQ} and V_{OH}, which is equivalent to $I_{FQ}R$, can be calculated using eqns 1 and 4, but the graphical construction of Fig. **6b)** is more instructive.

Equation 4 describes a load line on the plot of I_F, which equals I_D versus V_F. This passes through the points $I_F=0$, $V_F= V_{GG}-V_{TN}$) and $I_F=(V_{GG}-V_{TN})/R$, $V_F=0$. Its intersection with the transfer characteristic of the mosfet gives I_{FQ} and V_{OH}.

Voltage V_{OH} increases with R up to a maximum value os $V_{GG}-V_{TN}$.

When V_{GG} is greater than V_{TN} and V_I is less than $V_{GG}-V_{TN}$, the equivalent circuit is as shown in Fig. 7: diodes D_F and D_R in the model of Fig. 4a) are both on, the device operates in the 'ohmic' (or 'triode') region and equation 3 applies.

Hence,

$$V_O=I_DR=(I_F-I_R)R \tag{5}$$

or,

$$V_O=K_NR(V_F^2-V_R^2)=K_NR(V_F-V_R)(V_F+V_R) \tag{6}$$

But,

$$V_F-V_R=V_{DS} \tag{7}$$

So,

$$V_O=K_NRV_{DS}(2V_F-V_{DS}) \tag{8}$$

Substituting $V_{GG}-V_{TN}-V_O$ for V_F in equation 8 gives,

$$V_{DS}^2-2V_{DS}\left(V_{GG}-V_{TN}-V_O\right)+\left(\frac{V_O}{K_NR}\right)=0 \tag{9}$$

For a chosen value of V_O, you can calculate V_{DS} by solving this quadratic and selecting the root with the smaller magnitude. The other root does not satisfy the condition $V_R>0$. So for each value of V_O, you can find a value of V_I, since $V_I=V_O+V_{DS}$. This procedure allows you to make a plot of the transfer characteristic of the circuit.

Even without specific numerical values for V_{GG}, V_{TN}, K_N you can still make two useful statements concerning equation 9. First, if V_O is 0 then V_{DS} is also 0, because the other root $V_{DS}=2(V_{GG}-V_{TN})$ is inadmissible. So $V_I=0$ and all the transfer characteristics pass through the origin – irrespective of the value of R.

Secondly, if R approaches infinity then V_{DS} approaches 0 because $V_O/K_NR\rightarrow 0$. This means the transfer characteristic approximates to a straight line. **Figure 8** shows the characteristic for this condition and for a finite value of R.

Comparing the practical characteristics with the ideal ones in Fig. 1b) it is apparent that the circuit can handle bipolar input signals but there are deficiencies in two respects, range and linearity.

For a given mosfet type, the range can be extended by making V_{GG} larger but there are limits set by gate voltage breakdown and the magnitude of the supply rail voltage.

Linearity can be improved by having a large value of R but there is a limit set by dynamic considerations. When the transmission gate is switched off, the output terminal has associated with it a time constant $(C+C_L)R$, C_L being the load capacitance. If R is very large the circuit takes

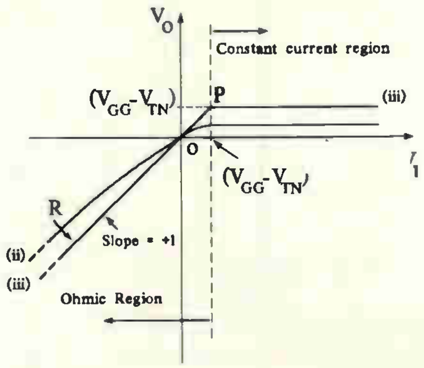


Fig. 8. Transfer characteristics for the circuit of Fig. 3.
i) Gate off
ii) Gate on and R finite
iii) Gate on and R approaching infinity

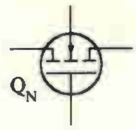


Fig. 9. Alternative symbol for Q_N in Fig. 3.

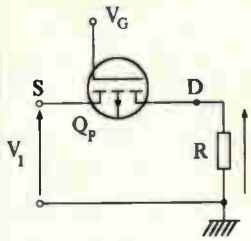


Fig. 10. Transmission gate using a p-channel enhancement-mode mosfet.

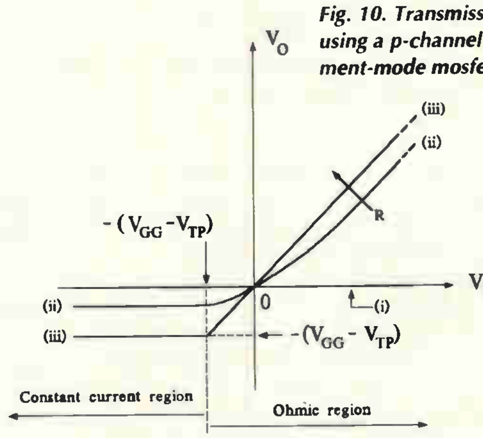


Fig. 11. Transfer characteristics for the circuit of Fig. 10. Compare with Fig. 8, to which the labelling of i), ii), iii) applies.

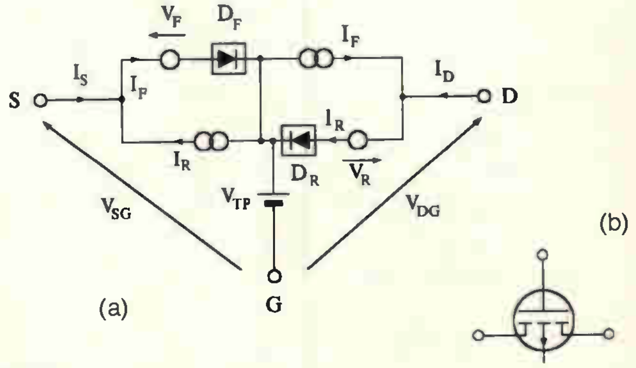


Fig. 12.
a) DC mosfet model for Q_P in Fig. 10.
b) Alternative symbol for Q_P.

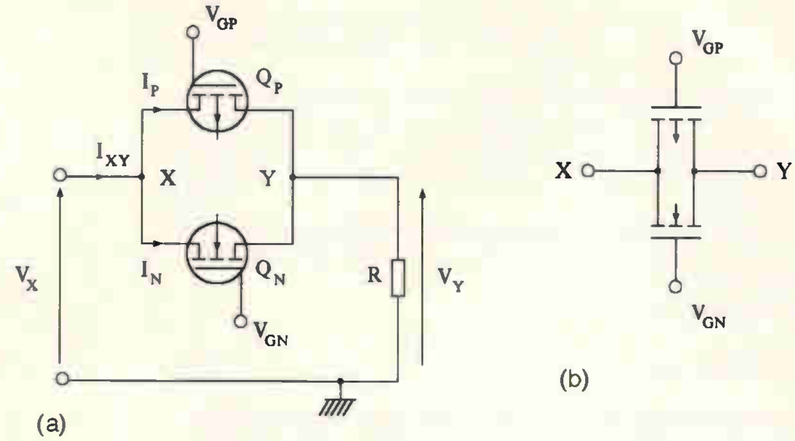


Fig. 13.
a) A complementary mosfet transmission gate.
b) Alternative representation of complementary connection.

a long time to reach the state where V_O is 0.

I will now digress briefly to clarify a point purposely glossed over at the start of this section. It concerns the naming of parts and symbolic representation of the mosfet. The connection of the drain terminal to the input and the source to the output of the circuit in Fig. 3 was an *arbitrary* choice.

Reversing the connections produces exactly the same characteristics as those in Fig. 8. The reason for this is the symmetry in the electrical characteristics of a mosfet. The channel region is uniformly doped in the direction of current flow along it. Also, the device structure is geometrically symmetrical about the plane drawn perpendicular to the semiconductor surface at a point mid-way along the channel. The symmetry is emphasised in the form of the mosfet model.

It is helpful to distinguish between the name given to a terminal and the function performed at that terminal, because a terminal designated 'source' can also function as a 'drain' and *vice-versa*. There is a loose analogy here, in mechanics. What we might call the 'business' end of a double-ended symmetrical spanner is the one being used to tighten or loosen a nut.

When discussing standard circuit configurations, for example a common-source stage and a source-follower, it is preferable to use the symbol shown in Fig. 3. Here, the gate connection is located near what is regarded as the source end of the channel because that aids in the identification of the intended function of the circuit.

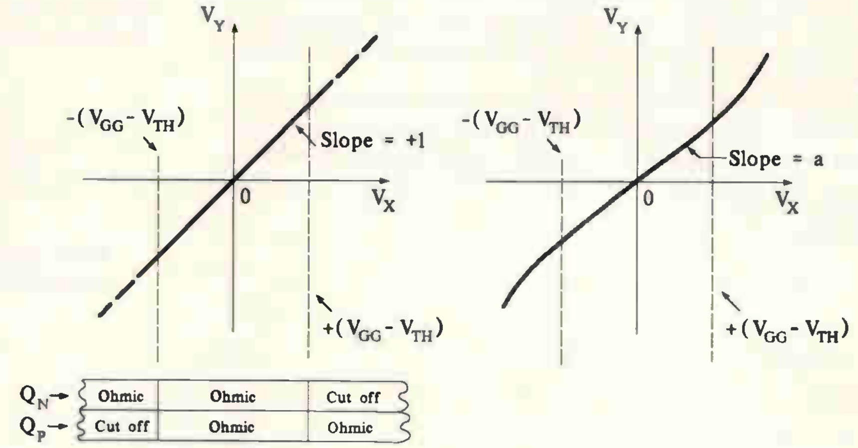


Fig. 14.
a) On characteristic for Fig. 13(a) for R infinite, showing conduction states.
b) On characteristic for finite R. Non-linearity is exaggerated for clarity.

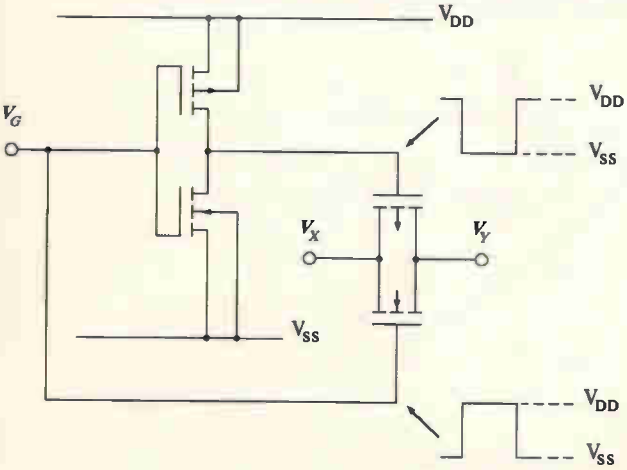


Fig. 15. Data sheet representation of a cmos transmission gate.

However, when neither of the mosfet terminals in the path of signal flow is at a fixed potential, and when there is no preferred direction of current flow through the device, the symmetrical nature of mosfet operation is indicated better by using the standard alternative symbol in Fig. 9. Here, the gate terminal is located centrally.

If a p-channel enhancement-mode mosfet, Q_P, is used instead of Q_N, as shown in **Fig. 10**, then the transmission gate is on for V_G=$-V_{GG}$ and off for V_G=$+V_{GG}$. The resulting transfer characteristics are given in **Fig 11**. These are obtained, by reasoning similar to that already given for a general mosfet. Q_N using the p-channel model of **Fig. 12a)**. Since no new principle is involved, that analysis is not given here.

As with Q_N, the source and drain terminals of Q_P can be interchanged and to reflect this the alternative standard symbol is that in **Fig 12b)**.

A complementary mosfet TG

The shapes of the plots in Figs 8 and 11 suggest that the transfer characteristic of a mosfet transmission gate, when on, might be improved in range and linearity by using two complementary devices in parallel, as shown in **Fig. 13a)** or the alternative representation in Fig. **13b)**. The terminals in the signal path, arbitrarily labelled X and Y, refer to the circuit nodes rather than the device terminal designations.

The transmission gate is off for the conditions V_{GN}= $-V_{GG}$ with V_{GP}=$+V_{GG}$ and on for V_{GN}=$+V_{GG}$ with V_{GP}=$-V_{GG}$. The following discussion refers to the on state. Of special interest are the theoretical conditions K_P=K_N=K and V_{TP}=V_{TN}=V_{TH}.

For a specified R, in order to plot a transfer characteristic for Fig. 13a), V_X can be calculated for assumed values of V_Y following the procedure

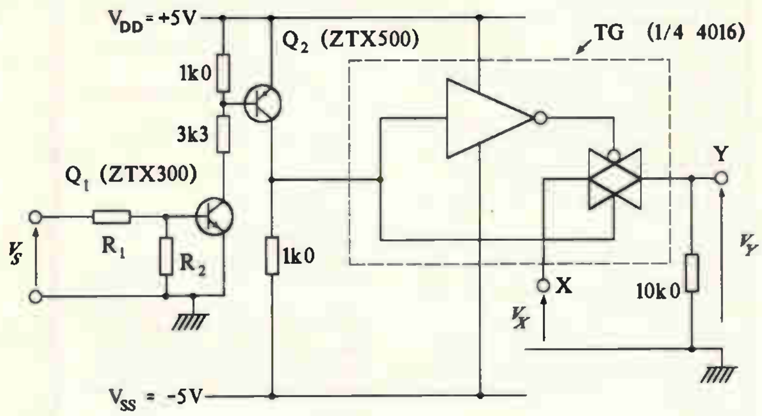


Fig. 16. Driving the cmos transmission gate from ttl-compatible circuitry.

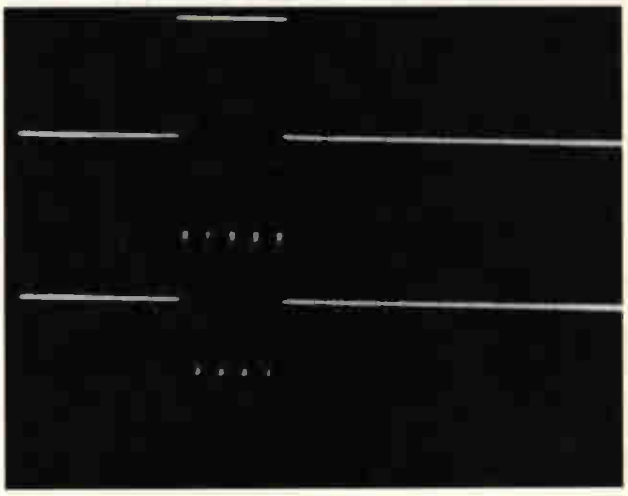


Fig. 17. Oscilloscope display for circuit of Fig. 16. The traces have common time and voltage scales.
Upper trace: v_S, with amplitude 5V and duration 100μs. Lower trace: v_Y, for a sinusoidal input.

described earlier for a single n-channel mosfet. The device equations give I_N and I_P as functions of V_X and V_Y. You can now use the relationship $I_P+I_N=I_{XY}=V_Y/R$.

For *R infinite*, the on characteristic is the straight line in **Fig. 14a**). The conduction states of Q_N and Q_P are determined by using the models of Figs 4 and 12 in the composite connection.

By inspection, Q_N and Q_P both operate in the ohmic region for,

$$(V_{GG}-V_{TH})\geq V_X\geq -(V_{GG}-V_{TH})$$

Mosfet Q_N is cut off and Q_P is ohmic while V_X is greater than $V_{GG}-V_{TH}$). Similarly, Q_P is cut off and Q_N is ohmic when V_X is less than $-(V_{GG}-V_{TH})$.

The theoretical characteristic for a finite value of *R* has the general shape shown in Fig. **14b**). Referring to Fig. 13. it can be shown that for, $+(V_{GG}-V_{TH})\geq V_X\geq -(V_{GG}-V_{TH})$,

$$I_{XY}=4KV_{XY}(V_{GG}-V_{TH}) \quad (10)$$

or,

$$\frac{V_{XY}}{I_{XY}}=R_{ON}=\frac{1}{4}K\left(V_{GG}-V_{TH}\right) \quad (11)$$

Resistance R_{ON} is both the dc and the incremental resistance over the V_X range considered. The transfer characteristic of Fig 14b) is a straight line of slope 'a'. where 'a' is a linear attenuation factor given by,

$$a=\frac{R}{R+R_{ON}} \quad (12)$$

Outside the V_X range considered the transfer characteristic ceases to be linear because R_{ON} is a function of V_X. This is to be expected since one of the devices cuts off.

Whether the composite arrangement of Fig. 13 is implemented in practice using discrete devices or, as is usually the case, using a monolithic cmos structure, there are at least two reasons why the transfer characteristic is non-linear over the whole range of V_X.

One reason is the impossibility of meeting, precisely, the conditions $K_P=K_N$ and $V_{TP}=V_{TH}$. The other is the variation of V_{TP}, V_{TN} with V_X via their dependencies on substrate bias – an effect so far ignored. In a cmos scheme, the second effect is usually minimised by using additional mosfets. These are not always shown on schematic diagrams.

Non-linearity in the transfer characteristic can be assessed from manufacturers' data, which specifies the variation in R_{ON} over a given range of V_X for a given *R*.

If *R* is much greater than R_{ON} and R_{ON} changes by *x*%, where *x* is less than 10, then it follows from equation 12 that 'a' varies by *x*% also.

To conclude this review I will present a practical gate.

The IC cmos gate in action

Figure 15 is a schematic circuit of an IC cmos transmission gate, as shown on a data sheet. The mosfet symbols in the inverter section, which supplies a complementary gating waveform, are different from those used in the signal-path section.

Figure 16 shows part of a circuit for generating 'flashing-ring' symbols for use in oscilloscope tests of human visual perception. The dotted box contains the circuitry of Fig 15: the parallel mosfets are now represented by a symbol comprising two interlocked oppositely directed arrowheads. This highlights the bilateral nature of this transmission gate. The input can be at X, as here, and the output at Y, or *vice-versa*.

Resistance R_{ON} is typically 250Ω for a 4016 type cmos transmission gate operating between V_{DD} = + 5 V, V_{SS} = -5 V. As a result, the choice 10kΩ for *R* meets the condition that *R* should be much greater than R_{ON}.

In this case, the input v_X is a sinusoidal signal. The ttl-compatible drive circuit, comprising Q_1, Q_2 and associated resistors, supplies a gating waveform with the required dc levels. This can be modified for faster operation by connecting a Schottky diode between the collector and base of each transistor, to minimise carrier storage time, and by adding an active pull-down circuit to the collector of Q_2.

Figure 17 shows an oscilloscope display of the time-selection pulse, v_S, applied to the base of Q_1 and the output waveform, v_Y, which is a selected portion of the input v_X. ■

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Amplifier instability

Erik Margan explains why an amplifier's input can be unstable – even when you follow the text-book design method.

The question of instability of an amplifier with reactive source impedance at its input that arose last year triggered memories of some 25 years ago. I remember seeing a very respectable oscilloscope burst into oscillation due to having an inductor connected to its input. I was trying to examine stray magnetic fields from a power transformer at the time.

The effect is due to the amplifier input impedance being negative over a range of frequencies. Here is an example of how even a very simple system, such as a jfet source follower, can exhibit such unexpected behaviour. **Figure 1a)** shows the schematic and **1b)** the equivalent circuit. The impedance seen by the jfet source terminal is,

$$Z_S = \frac{1}{\frac{1}{R_S} + j\omega C_S} + \frac{1}{\frac{1}{R_L} + j\omega C_L} \tag{1}$$

Summing the currents into the source node gives,

$$\frac{V_S}{Z_S} = (V_G - V_S) j\omega C_{GS} + (V_G - V_S) g_m \tag{2}$$

From this, the small-signal transfer function at the source can be written,

$$F(\omega) = \frac{V_S}{V_G} = \frac{j\omega C_{GS} + g_m}{\frac{1}{Z_S} + j\omega C_{GS} + g_m} \tag{3}$$

and the actual output will be scaled down by the source to load impedance ratio,

$$\frac{V_{out}}{V_G} = F(\omega) \frac{1}{Z_S\left(\frac{1}{R_L + j\omega C_L}\right)} \tag{4}$$

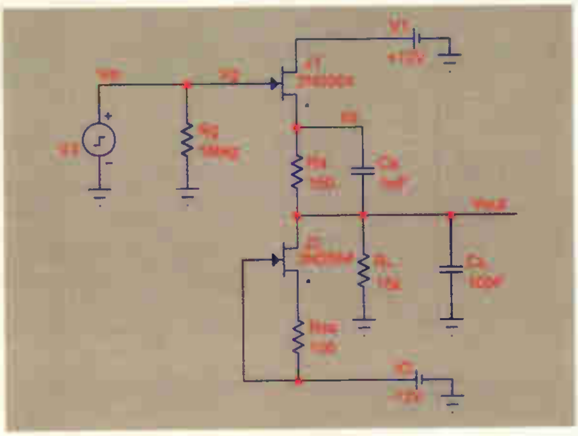
The magnitude, phase and envelope-delay of $F(\omega)$ are shown in **Fig. 2**. Note the high crosstalk in the magnitude plot at high frequencies. A cautious designer will be worried by seeing the phase plot turning upwards and the envelope-delay going positive. But there is more.

The input impedance at the jfet gate can be found by dividing the input voltage by the input current,

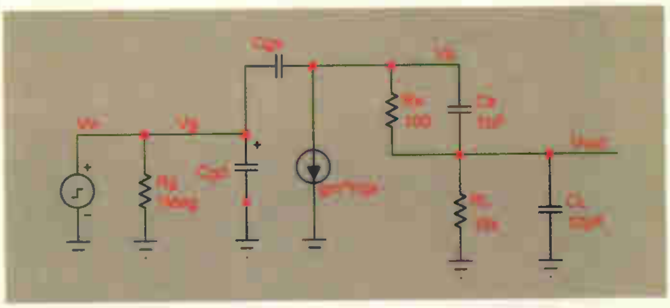
$$Z_G = \frac{V_G}{(V_G - V_S) j\omega C_{GS}} \tag{5}$$

which results in,

$$Z_G = \frac{V_G}{(1 - F(\omega)) j\omega C_{GS}} \tag{6}$$



(a)



(b)

Fig. 1. Typical junction fet source follower, as might be used as an oscilloscope input stage.

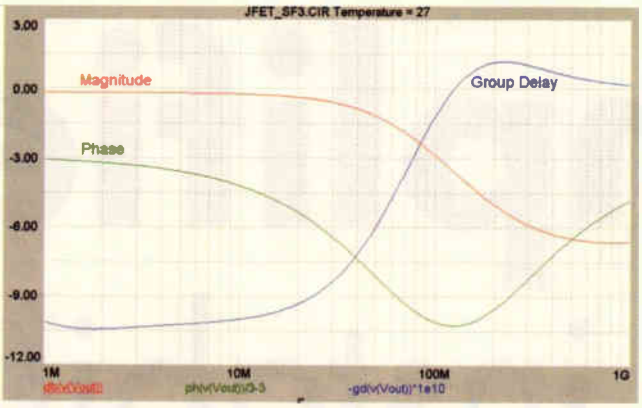


Fig. 2. Response curves of the follower in Fig. 1.

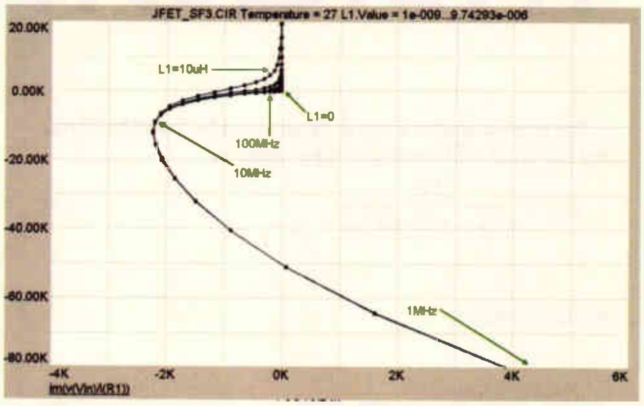


Fig. 3. Junction fet source follower input impedance for several values of input inductance.

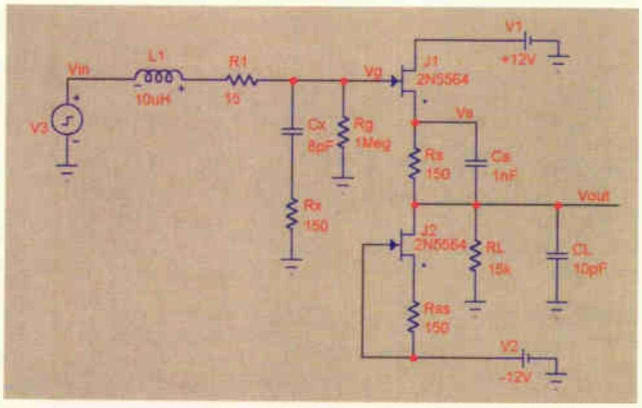


Fig. 4. Improved junction-fet source follower with negative input impedance compensation provided by R_1, R_x and C_x.

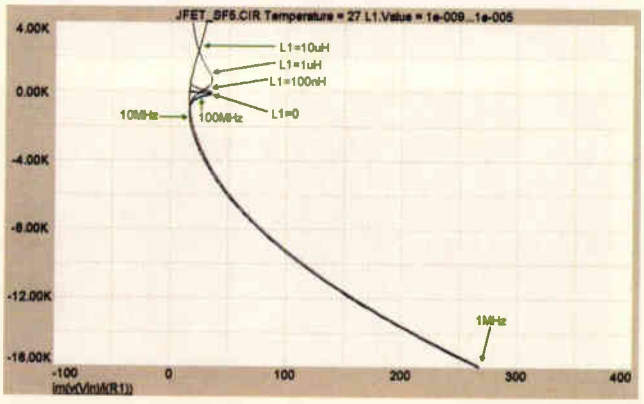


Fig. 5. Junction fet source follower input impedance for several values of L_1, compensated by R_x and C_x.

Note the negative sign, which indicates a potential source of trouble!

In parallel to Z_G there is the gate bias resistor and the gate-drain stray capacitance and any source inductance is added to the whole,

$$Z_{in} = \frac{1}{\frac{1}{Z_G}+\frac{1}{R_G}+j\omega C_{GD}} + j\omega L_1 \qquad (7)$$

The imaginary versus real part of Z_{in} is plotted in **Fig. 3**, showing that beyond 3MHz there is a negative real part, as predicted by the negative sign in the Z_G denominator. With an inductance at the input, even as low as that of a piece of wire, the circuit will oscillate at the frequency at which the Z_{in} plot crosses the negative part of the real axis.

But be warned that in too many text books Z_G is oversimplified to,

$$Z_G = Z_S + \frac{g_m Z_S + 1}{j\omega C_{GS}} \qquad (8)$$

which – since all terms are positive – shows no cause for alarm.

How to cure it

There are four solutions to this instability problem.

a) As noted in previous correspondence, the simplest method is to insert a resistor of a large-enough value – about 3kΩ – in series with the amplifier input, but this will lower the system bandwidth.

b) Choose such values of jfet source loading and transconductance, that Z_{in} remains positive, i.e. instead of using the 2N5564 one can insert the 2N5911 which has lower stray capacitances and higher source resistance, but this will also need some 300Ω in series.

c) If the input signal will come from a transducer of known inductance, it is possible to insert a small resistor in the drain of the upper jfet and thus allow some negative feedback through the gate-drain capacitance (Miller effect). This works for a limited range of inductances.

d) The best way is to add a series C_xR_x network across the input, as in **Fig. 4**, to compensate for the negative input impedance. **Fig. 5** shows the result for inductances up to 10μH. Now the series input resistance can be safely reduced to some 15Ω, without influencing the bandwidth.

With bipolar transistors, the negative input impedance is greatly reduced by the base resistance. Even so, some circuit configurations can still become unstable with input inductance. The series C_xR_x compensation can also be used here, as described for the fet. ■

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Designed to reduce jitter in a digital video disk measurement system, Chris Hancock's precision integrator achieves a repeatability of 0.1% on successive ramps at very high speeds.

Fast and precise integrator

I designed this precision analogue integrator to produce linear dV/dt voltage ramps clamped between +1.5V and −1.5V. The output ramps were produced from the edges of data pulses with a width of between three and fourteen multiples of a data clock running at 27MHz.

Rise and fall times of the data-pulse edges are symmetrical, with a typical value of 1.5ns. The linear slope of 3V, i.e. ±1.5V, is produced in a nominal time of 24ns. This is achieved using a novel feedback loop system, which makes use of state of the art 1.2GHz bandwidth operational amplifiers and 274MHz bandwidth clamp amplifiers.

Results show that the circuit can be used to produce linear precision clamped voltage ramps from the edges of the data pulses, whose widths are changing randomly over the entire range. Repeatability measurements indicate that it is possible to achieve a repeatability of 0.1% on successive ramps.

Background

The motivation designing this integrator came from the need to measure timing jitter between signal data and a synchronised data clock in a digital video disk measurement system. The data clock was running at 27MHz with a 1:1 mark-space ratio, and the width of the data pulses was between three and fourteen clock cycles, i.e. 111ns to 518ns. Voltage levels for both clock and data signals was ttl.

The rise and fall times of the data pulses were symmetrical, with typical values of 1.5ns. For the purpose of this work, the data pulses can be considered as being generated continuously, with random width between three and fourteen clock cycles.

We needed to produce a precision dV/dT ramp from the edges of these data pulses, and from this, voltage levels were to be sampled to enable the timing jitter between the rising edge of the clock and the data signal to be calculated.

Voltage levels taken from the ramp were fed into a fast eight-bit Phillips *TDA8703* a-to-d converter and stored in an *IDT7201* first-in-first-out memory, ready for subsequent processing. This article concerns only the analogue integrator.

We wanted the measurement system to run continuously, so it was vital that no charge remained on the integration capacitor at the start of the next signal integration. In addition, the clamped levels needed to be identical for each data pulse width.

Because we needed identical clamped levels, a basic active integrator using a single operational amplifier with a feedback capacitor was not an option. We needed a solution whereby the output from the integrator would remain constant at either the positive or negative clamp amplitude level for a predetermined time after the ramp level has changed between clamping voltage rails, ready for the next data edge to arrive.

Dr Hancock is involved with research and development at Gyrus Medical Ltd in Cardiff.

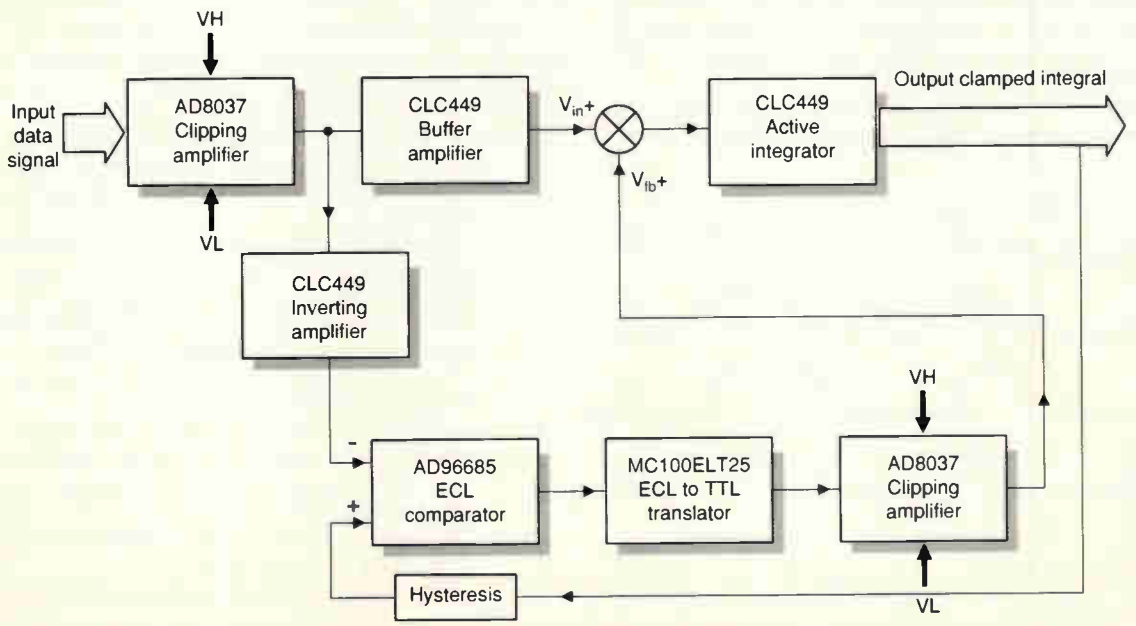


Fig. 1. Fast integrator in outline. Unlike other integrator designs, this one compares an inverted version of the clamped data pulse to a ramp level produced from a basic CR integrator. This CR integrator's forward input comprises the clamped version of the incoming data pulse. Comparator output is again clamped and then summed with the integrator's forward input.

When the integrator is used in the complete measurement system, the contents of the a-to-d converter can be interrogated at any time using an on-board digital signal processing system. After interrogation, the contents of the a-to-d converter, and hence the first-in-first-out memory, would be cleared, ready for the next level to be fed in. This means that the integrator should be running continuously, and since the width of the data could be between 111ns and 518ns, in 37ns increments, the integrator must be able to clip between +1.5V and -1.5V with a ramp time such that a timing jitter error can be accurately discerned from the ramp.

The ramp was designed to be 24ns long; this aspect of the design is detailed later. Since, theoretically, the maximum time that the level is to stay constant –i.e. 'held' – is 494ns, or 518ns–24ns, the level of voltage change during this time must be negligible. In other words, the level must be held constant at either the positive or negative clamp level.

Also, slope and measurement errors, due to temperature variations and delays through the active devices used in the feedback loop, had to be removed by carrying out a system calibration routine before each measurement run. Basically, this involved a fixed delay being introduced in the data clock signal, i.e. a clock skew, to compensate for the delays, together with an additional delay to ensure that the measurement is taken on a truly linear region of the ramp.

Circuit details

This integrator uses a novel method whereby the inverted version of the clamped data pulse is compared to a ramp level produced from a basic *CR* active integrator. This integrator's forward input consists of the clamped version of the incoming data pulse.

Comparator output is again clamped and then summed with the forward input of the integrator. From Fig. 1, you can see that the incoming data signal is first clamped using an Analogue Devices *AD8037* low distortion, wide bandwidth voltage feedback clamp amplifier.[1] This device conditions the input data signal to provide precise ±1.5V drive levels by setting up a closed-loop gain of 2 and then multiplying this by precise input voltage reference clamp levels, V_H and V_L.

These reference levels were derived from a National Semiconductor *LM368* precision 2.5V voltage reference, and a potential divider network, buffered using an National Semiconductor *LM358* quad operational amplifier.

At this point, the clamped data signal is split into two paths; one feeding a buffer amplifier, which consists of a Comlinear *CLC449* 1.2GHz ultra-wideband operational amplifier[2] configured to have a closed-loop gain of 6. Output from the buffer feeds one of the summing inputs to the active *RC* integrator, which again uses a *CLC449*.

In the other signal path, the clamped data signal is first inverted to provide a gain of –1 using a further *CLC449* and then fed into the negative input of an Analogue Devices *AD96685* ultra-fast comparator.[3] The positive input to the comparator is taken from the output of the active integrator.

Some hysteresis is introduced in this path to prevent changeover uncertainties from occurring. Output from the comparator is then translated from its ecl level to ttl, using a Motorola *MC100ELT25* translator[4]. the resulting signal feeds a second *AD8037* clamp amplifier, which is configured identically to the first clamp amplifier.

Note that we used an ecl comparator because it is the fastest device available for comparing two signals and changing the output state – even with the translator stage included.

Input clamp levels V_H and V_L are derived from the same precision voltage reference circuit as that used for the first clamp amplifier. The signal from the second clamp amplifier is then summed with the clamped and buffered version of the initial data signal at the input of the active integrator; hence closing the feedback loop.

If you assume that the two summing resistors are balanced and the clamp voltages are identical, then the output from the active integrator, with the loop closed, will ramp up or down, depending on the input voltage polarity. When the two voltages are 180° out of phase – i.e. inverted – the output clamped voltage is held and remains constant.

Figure 2 illustrates the operation of the circuit in terms of timing sequences. The first waveform shows the input data after it has been precision clamped. The ramp output from the active integrator is shown in waveform 2, which is compared with the inverted version of the clamped data, waveform 3.

When the ramp's amplitude reaches the inverted data input level, the output from the comparator, waveform 4, changes state. This signal is then conditioned by the second clamping amplifier, waveform 5, and added to waveform 1 to hold the output level constant at either the positive or negative clamp voltage levels.

Putting it into practice

The whole circuit, Fig. 3, was laid out on a multilayer board. Layer 1 contains the signal tracks, layer 2 the ground plane and layer 3 the power supply tracks.

All signal tracks were kept less than 3cm long, i.e. less then the wavelength of the highest possible frequency component contained in a data pulse. Power supply decoupling capacitors were placed as close as possible to the power pins; tantalum capacitors were used for the polarised values since these offer the lowest series resistance.

Surface mount capacitors and resistors were used throughout the design to minimise lead inductance. We ensured that low impedance power tracks were kept as far away as possible from sensitive device inputs to prevent the formation of leakage resistors, which can result in leakage currents often of higher magnitude than the device bias currents.

The devices were mounted directly onto the circuit board, since the extra inductance introduced by IC sockets is significant when operating at such high frequencies.

Resistors of 100Ω were placed in series with the output of the *AD8037* and *CLC449* devices to prevent oscillation and overshoot from occurring. Also, 50Ω termination resistors were used at the inputs to prevent pulse reflections and ringing.

The incoming ttl level data signals were ac coupled, using a 100nF ceramic capacitor, C_{60}. This enabled the signal to swing positive and negative around zero to enable it to reach the precision clamp levels, V_H and V_L. Resistor R_9 provides a dc path for bias currents; this prevents C_{60} from charging up to a dc level and superimposing a dc offset on-top of the data signal.

The V_H and V_L clamp levels, obtained using pre-

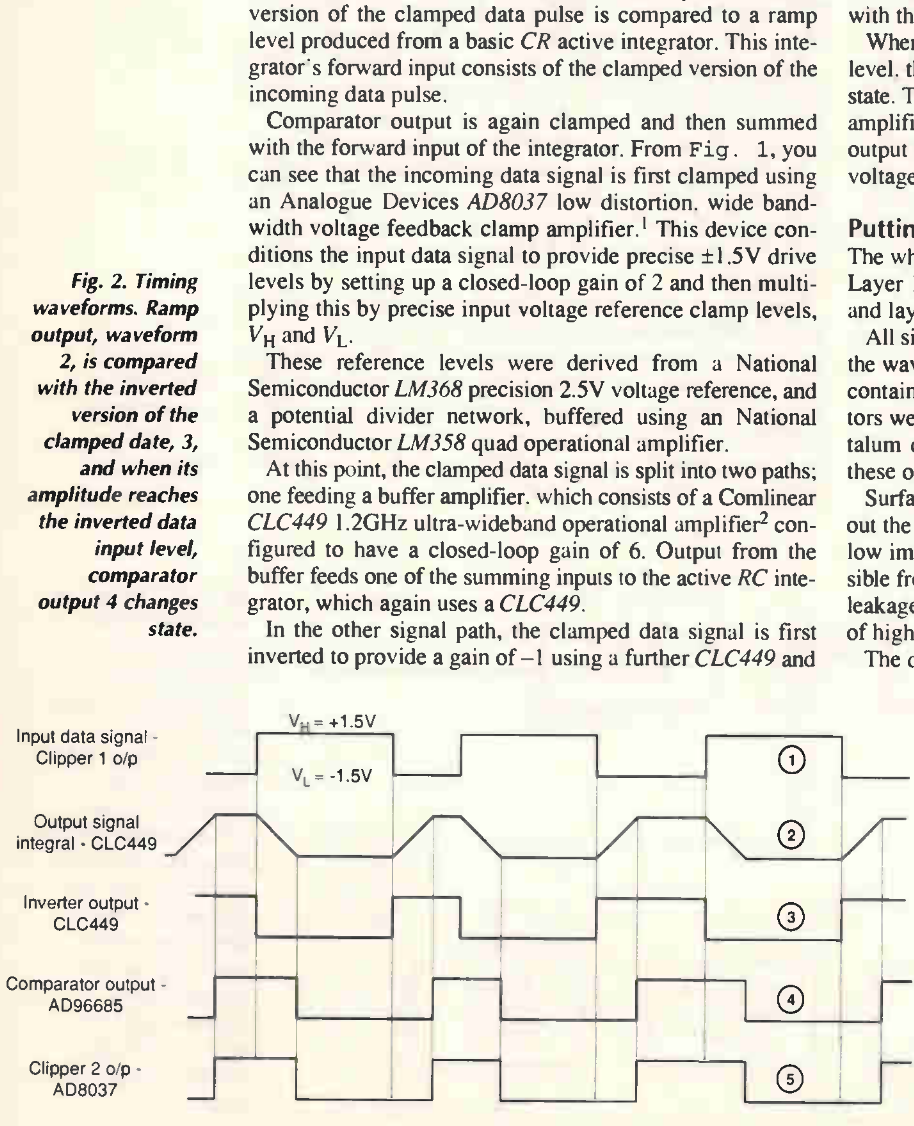


Fig. 2. Timing waveforms. Ramp output, waveform 2, is compared with the inverted version of the clamped date, 3, and when its amplitude reaches the inverted data input level, comparator output 4 changes state.

cision reference IC_8, were followed by a potential divider network and a quad operational amplifier, IC_9, used to buffer and invert the voltage taps from the potential divider.

The precision voltage reference circuit was tested for stability using a Keithley *195* precision digital multimeter. We found that each of the four output voltages varied by less than 1 part in 10,000 – i.e. 0.01% – when monitored separately over an eight hour period.

Adjustable-gain buffer IC_2 was set to provide a gain of 6 to compensate for the attenuation due to the three 100Ω output and 50Ω input matching resistor networks. Each of these introduced a signal attenuation of a third.

The *RC* time-constant of the active integrator, IC_3, is such that the output signal changed from one clamp level to the other in 24ns, whereby a positive or negative going linear ramp was produced when both the summing inputs change to

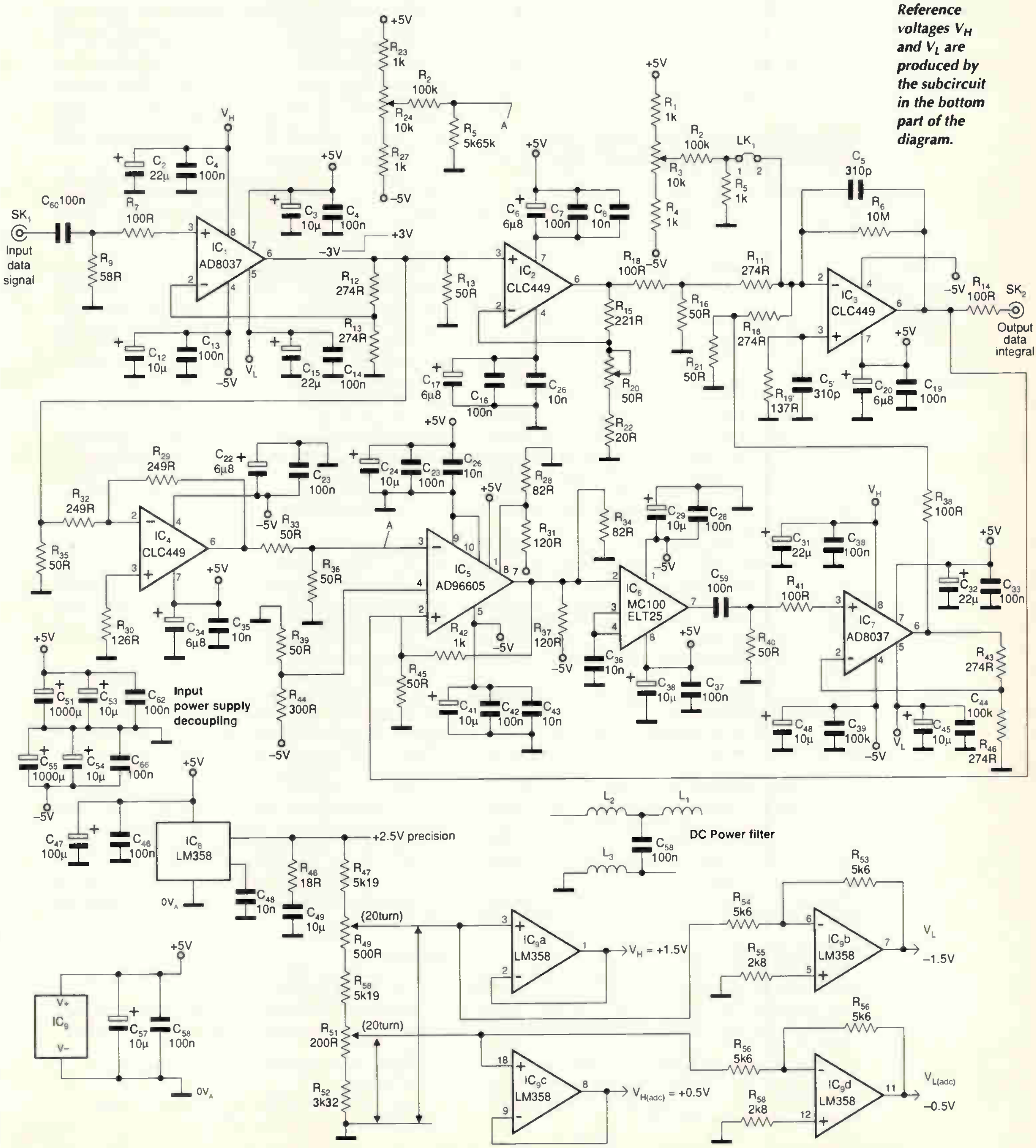


Fig. 3. Full circuit diagram for the precision integrator. Reference voltages V_H and V_L are produced by the subcircuit in the bottom part of the diagram.

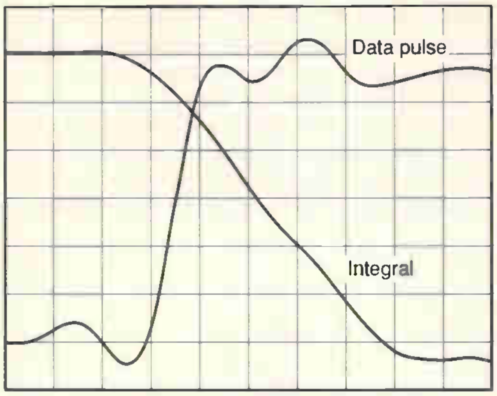


Fig. 4. Rising edge of data pulse with corresponding clamped integral at 1ns/div. The integral is 200mV/div while the data pulse is

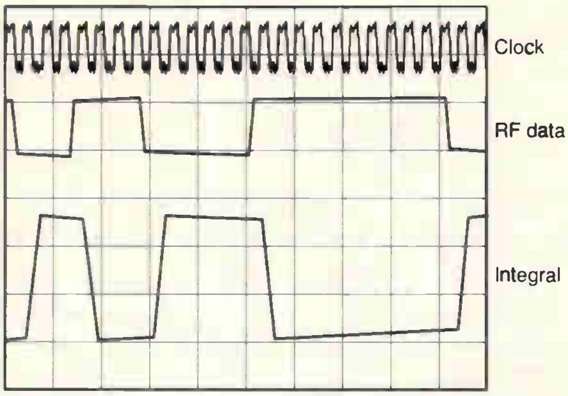


Fig. 5. Clock signal, input data signal and integrated-data signal at 100ns/div. Clock 5V/div, rf data 2V/div and integral 1V/div

a high or low state respectively, Fig. 2. It is most important to ensure that the summing resistors, R_{11} and R_{19}, are balanced.

Final circuit values were tuned for optimal performance, from which the following values were obtained: 310pF for C_5 and $C_{5'}$, 274Ω for $R_{11,19}$ and 137Ω for $R_{19'}$. The integration capacitors are silver-mica and the resistors were precision 0.1% tolerance metal film.

Resistor R_6 is included to limit the dc gain of the integrator. Some input offset voltage adjustment capability is provided via a network consisting of resistors $R_{1\text{-}5}$. The output from the active integrator is fed back to the positive input of comparator. IC_5. Resistors R_{42} and R_{45} were included in this path to introduce approximately 38mV of hysteresis. This prevents indeterminate conditions from occurring at the output of the comparator.

The negative input to the comparator is fed with an inverted version of the clamped data signal, obtained using IC_4, R_{29}, R_{32} and R_{33}. A dc offset adjustment is also provided at the input to the comparator via a network consisting of resistors $R_{23\text{-}27}$.

Output from comparator IC_5 is converted to ttl using IC_6, an *MC100ELT25*, whose output is again converted into a positive and negative swing around zero via C_{59}. This provides the appropriate levels for clamping. The signal is then also precisely clamped to ±1.5V using IC_7, which is configured in exactly the same way as IC_1.

Finally, the feedback loop is closed by summing output from IC_7 with the buffered data signal at IC_3's inverting input.

And how did it perform?

The operation of the circuit was evaluated using data signals. in the format described earlier. Output from the circuit was monitored using a Tektronix *TDS540A*, which is a 500MHz digitising oscilloscope capable of sampling at a rate of 1Gs/s.

Example test results are given in Figs 4 and 5. Figure 4 shows the rising edge of a data pulse and the corresponding integral after the delay through the system has been removed using a Dallas *DS1020* programmable eight-bit delay line, which has a resolution of 0.15ns.

You can see that the slope of the output signal is linear, making it possible to discern the necessary data at various points on the slope using a fast a-to-d converter. It should be pointed out here that an extra delay needs to be introduced during the calibration routine to ensure that the measurement points are taken over the linear region of the ramp, i.e. only using the range 1/4 to 3/4. Here, the slope can be characterised by the equation for a straight line, from which the timing jitter can easily be calculated.

Note that the full detail of this procedure is out of the scope of this article. I should also point out that although the bandwidth limit of the oscilloscope is being approached, Fig. 4 shows that it is still possible to resolve the ripple and overshoot of the data signal. If ripple was present on the integral, then it would be shown superimposed on top of the output slope. This gives us confidence that the slope is in fact truly linear and the waveform shown is not a manifestation of an oscilloscope bandwidth limitation.

You can also see that the output voltage levels change from one clamp level to the other in around 24ns, as required. Figure 5 shows the 27MHz data clock, together with two data pulses of width I_3 and I_{10}, and the corresponding output from the integrator.

The diagram shows that there is a slight voltage drift from the negative clamp level during the 'hold' time. This slight drift is non-effectual in the overall results from the system. Although it shifts the negative clamp level to a level somewhat closer to zero than the desired –1.5V, we would sample at between the 1/4 and 3/4 amplitude points on the slope.

During recent calibration and test measurements on the circuit, we found that we were able to completely cancel the non-constant 'hold' level by carefully adjusting the input offset voltage at the inverting input to the active integrator, via adjustment of potentiometer R_3. Also, the clamping levels may be changed by adjustment of the dc offset voltage at the inverting input to the comparator, which is controlled using potentiometer R24.

In summary

Our precision analogue integrator uses the fastest analogue integrated circuits currently available. We have shown that the feedback loop design effectively controls the precision clamping of the output levels. Recent measurements carried out on the circuit indicate that it is possible to achieve a repeatability of 0.1% on successive ramps.

We found it difficult to obtain long-term stability over a temperature range outside room temperature since this is determined by individual device voltage variation with temperature, which is outside our control. This variation is not detrimental to the overall operation of the system since the circuit will be calibrated prior to each measurement run to compensate for all device delays and circuit variations.

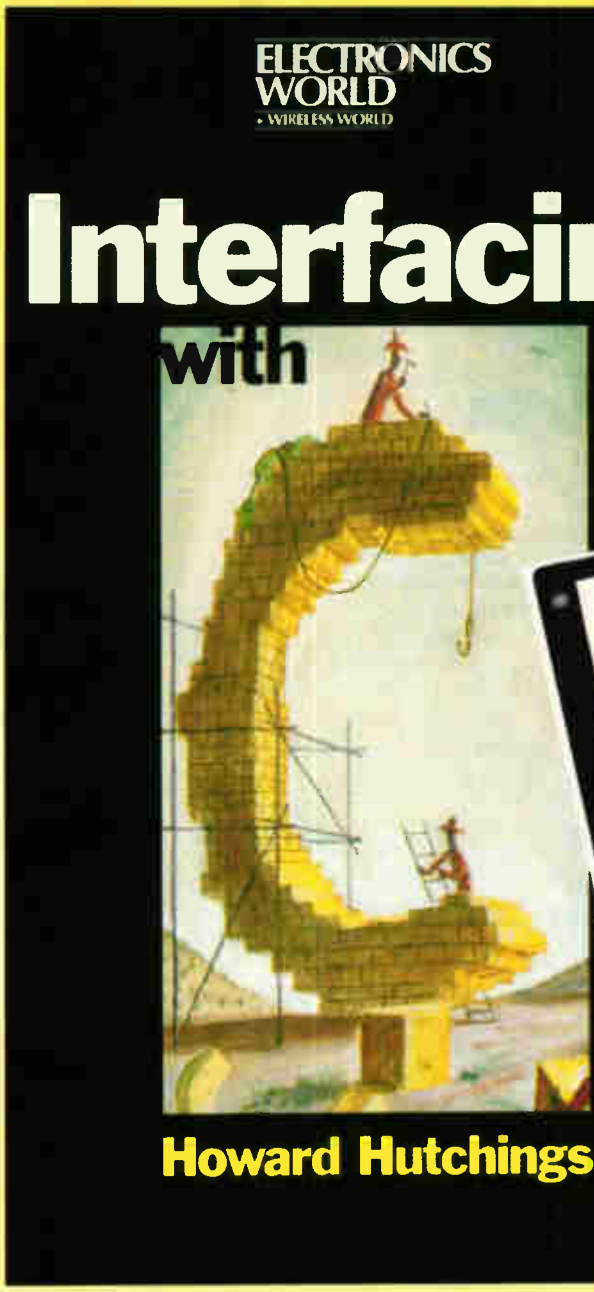
The calibration routine also ensures that the measurements are taken only on the truly linear region of the ramp.

I would like to thank Richard Amey and Dr Adam Hoare, of Aerosonic Ltd, for allowing me to carry out the work described in this paper. Thanks also to Jim Lusk, also of Aerosonic Ltd, for help in producing the circuit diagram and prototype pcbs. ■

References

1. Analogue Devices AD8037 Low Distortion, Wide Bandwidth Voltage Feedback Clamp Amps – 1997 Data sheet.
2. Comlinear CLC449 1.2 GHz Ultra-Wideband Monolithic Op-Amp – National Semiconductor Operational Amplifiers Data Book 1997.
3. Analogue Devices AD96685 Ultrafast ECL Comparator – 1997 Data sheet.
4. Motorola MC100ELT25 ECL to TTL Translator – High Performance ECL Data Book 1997.

Interfacing with C



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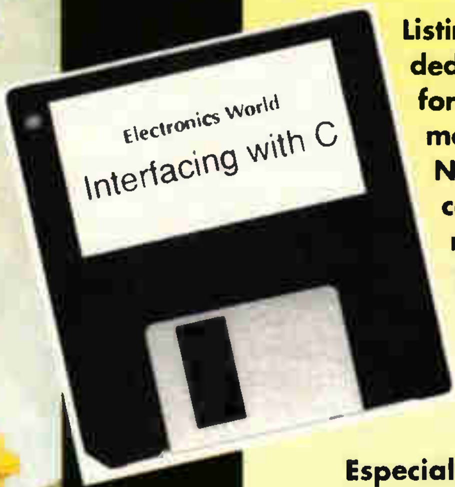
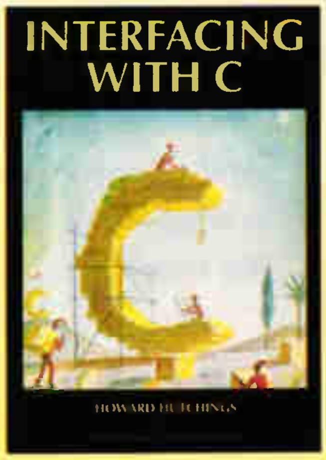


Listings on disk – over 50k of C source code dedicated to interfacing. This 3.5in PC format disk includes all the listings mentioned in the book Interfacing with C. Note that this is an upgraded disk containing the original Interfacing With C routines rewritten for Turbo C++ Ver. 3. Price £15, or £7.50 when purchased with the above book.

Especially useful for students, the original Interfacing with C, written for Microsoft C Version 5.1, is still available at the special price of £7.50. Phone 0181 652 3614 for bulk purchase price.



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Recharging the unchargeable

Our fears about charging zinc-carbon and alkaline cells are based on fifties battery technology and are no longer justifiable, argue Michael Slifkin and Yaakov Levy. They also present a simple and cheap charger that works much faster than the traditional 'dirty-dc' solution.

Although most people believe that recharging zinc-carbon and alkaline primary cells is both difficult and dangerous, our experiments have indicated that this is not the case. Older readers among you may remember wireless sets available in the UK in the fifties. While using ordinary throw away batteries, these sets also had a recharge switch for extending batterty life.

In recent years, a number of devices for 'recharging' primary cells and batteries has appeared on the market. But literature on this topic is hard to come by. Most present-day solutions appear to be based on a charger described by Alan Tong in *Everyday Electronics*.

However Alan stated in his article that to prevent batteries from leaking or exploding, it is necessary to charge using pulsed positive charges with the occasional negative going pulse – a technique known as pcr, or pulse charge reversal.

The rationale behind this, is that it is known from the electroplating industry that metals plated by pcr have a much denser plating layer. Plating using pure dc can produce spongy layers that occupy a much larger volume. If this should happen inside a battery, then the spongy layer's additional volume would cause a marked increase in internal pressure, leading to leaks or explosions.

Folk lore

Although there is little literature on the subject, there is a certain amount of folk lore. In earlier times, when the cost of batteries was relatively more expensive than nowadays, various methods were tried to increase their life.

A favoured method, which works to some extent, was to put the failing battery in a warm oven overnight. Other people would apply a dc voltage to their batteries hoping to revive them. But this only worked if the battery had not been completely exhausted, leading people to consider this method a failure.

It is possible to extend the life of a button battery used in watches and calculators by applying a 2.5V direct current until the battery gets warm. But again, this only works when the battery is not completely flat.

A very old article, in the October 1955 issue of *Wireless World*, refers to the use of "dirty dc" in recharging these batteries. The dirty dc charger consisted of a half wave rectified supply with a small amount of alternating current superimposed by using a 200Ω bypass resistor around the rectifying diode This is probably the simplest charger you can build.

After we completed our research work into chraging cells, a very detailed article by Rod Cooper appeared in the March 1997 of *Electronics World*. In this article, he described a fairly elaborate series of experiments and a charger based on this pcr principle. In addition Rod said that it is necessary to adjust the charging current to the voltage so that as the

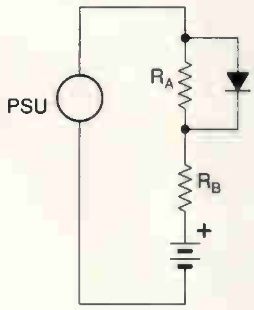
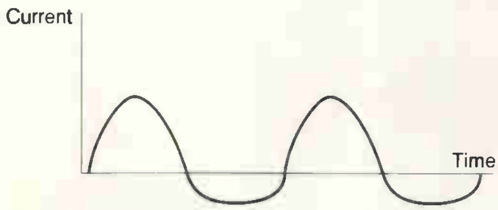


Fig. 1. 'Dirty dc' charger has a diode, which passes all of the positive half cycle, with a series resistor to let through a small portion of the negative half cycle.



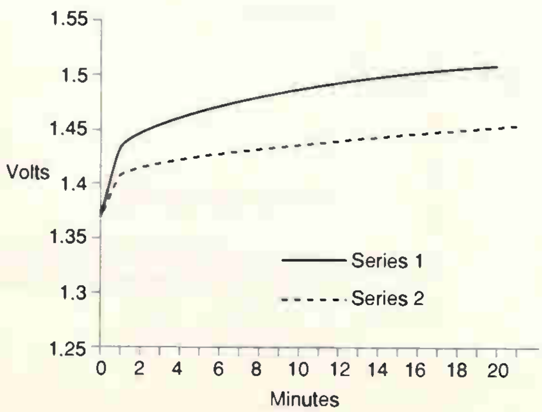


Fig. 2. Charging characteristics of AA zinc-chloride batteries at different frequencies. Series 1 at 50Hz, series 2 at 1kHz.

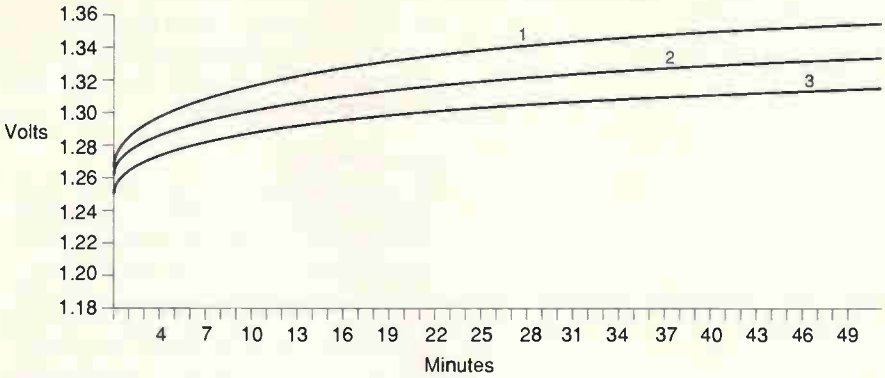


Fig. 3. Successive charging characteristics of AA manganese-alkaline batteries. Curve 1 at 50Hz, curve at 2 at 1kHz, curve 3 at 50Hz.

Table 1. Resistors suitable for charging other battery sizes using the 'dirty dc' charger.

Battery type	D	C	AA	AAA
R_A (Ω)	68	220	560	1500
R_B (Ω)	6.8	18	47	120
Power supply (W)	5	2	1	1

battery's voltage rises, the current must be decreased.

Dirty dc chargers suffer from having low duty cycles of not more than 30 %. The intention of our work was to find the optimum conditions for recharging so that we could build an efficient fast charger – hopefully with a very high duty cycle and for a cheap price.

In this article we describe a number of experiments we carried out on extending the life of primary batteries. At the end, we present a description of a simple apparatus incorporating our results.

Our experiments were carried out over a period of about a year using in all about 100 batteries – mainly of the AA type. All our experiments involved 4.5V peak pulses. Negative pulses were fixed at 1.5V.

The important factor is the charging current. We used about 50mA for AA type batteries at a duty cycle of 50%. Higher duty cycles have proportionately higher currents.

It is interesting that Rod suggested that the maximum charging current for batteries should be about the total capacity divided by 30 or 40 – which would be about 50mA for the AA battery.

In our experiments, batteries were discharged in a controlled manner to about 1V before recharging. We considered a battery to be recharged when the final voltage showed no change for about five minutes.

For new batteries this steady charged state usually occured at between 1.5 and 1.6V. Older batteries never reached this voltage. In common with earlier authors, we found that very old batteries, or batteries that had been completely discharged, were impossible to revive.

Is frequency important?

Our first series of experiments involved a pulsed supply of variable frequency to see whether or not frequency was important. In the electro-plating industry, it is accepted that very low frequency pcr of about 10Hz gives much more compact layers than higher frequencies.

First we used a simple dirty dc charger, as in **Fig. 1**. Results obtained for zinc-chloride batteries are shown in **Fig. 2**. Clearly, much faster charging occurs at 50Hz than at 1kHz.

An interesting set of results are shown for a manganese alkaline battery that has been discharged and recharged three times. The first and third cycle is at 50Hz and the middle cycle is at 1kHz. The superiority of 50Hz charging is amply demonstrated.

A similar experiment is shown for another manganese cell, but in this case, the cell has been subjected to a deeper discharge. Again the superiority of the 50Hz charging regime is demonstrated.

And duty cycle?

The next experiment compares the charging effect of square – i.e. 50% duty cycle – pulses as opposed to the sinusoidal pulses. **Figure 3** shows that cells charged with square pulses reache a slightly higher voltage – even though they start from a much lower discharged voltage. In general, we found that higher duty cycle pulses are more efficient in charging batteries than lower duty cycle sinusoids.

Next we looked at increasing the duty cycle. A typical plot is shown in **Fig. 4**. It shows that the 95% duty cycle is superior to that of the 75% duty cycle, reaching a higher terminal voltage.

Is the discharge cycle needed?

Next we looked at the effect of charging using only positive pulses. We compared the results of positive-pulse-only charging directly against a charging regime using alternate 1.5V negative pulses and 4.5V positive pulses, while maintaining the same duty cycle.

We found that the absence of a negative pulse had no effect on charging efficiency. Even more surprisingly, we also got a slightly higher end point voltage using only positive pulses and the same duty cycle.

After this finding we carried out many charge/discharge cycles on zinc-carbon and alkaline batteries, but found no signs of distress. It became clear to us that pcr is not a requirement for the revitalisation of these batteries.

So is 'dirty dc' really necessary? While our only case of a leaking battery was one that we charged using pure dc, the leak only occurred after many hours. We found that, provided the state of the battery being charged is monitored, and that the battery is taken off charge when its voltage does not change after a few minutes, other precautions were unnecessary.

Later, we present a charger with 85% duty cycle that is easily varied. We added this variable duty cycle feature because most other people in this field believe that pure dc should not be used – a fact we feel needs to be proven. Having a circuit with a variable duty cycle allows you to experiment for yourelves.

Useful – but not immortal

We should stress that during our experiments, we never managed to restore a battery to its original capacity. And each recharge cycle

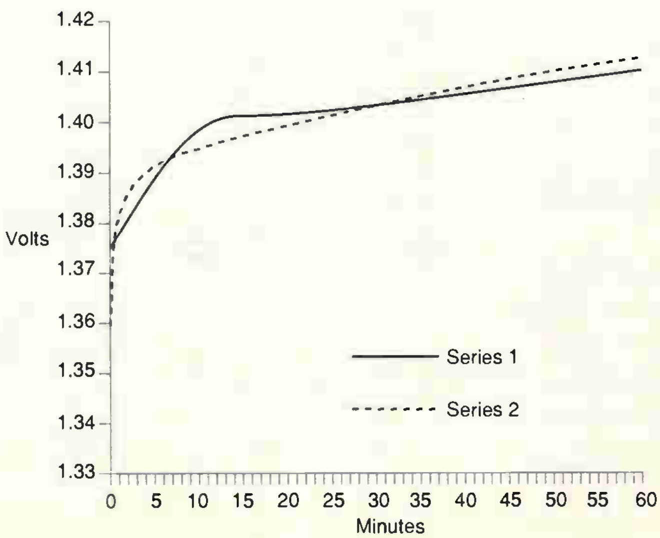
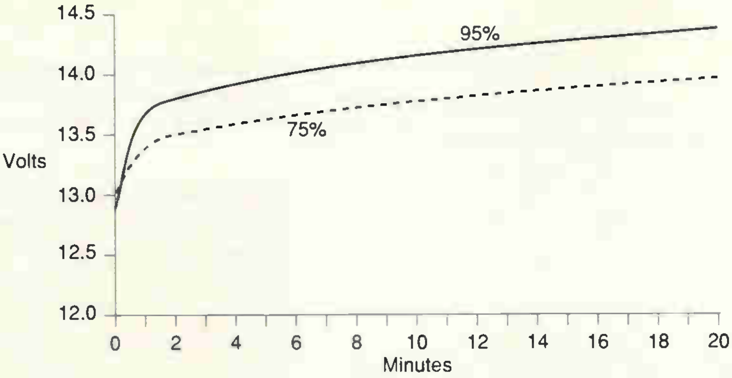


Fig. 4. Charging AA manganese-alkaline battery with sinusoidal versus square pulses. Curve 1 is sinusoidal pulses, curve square.

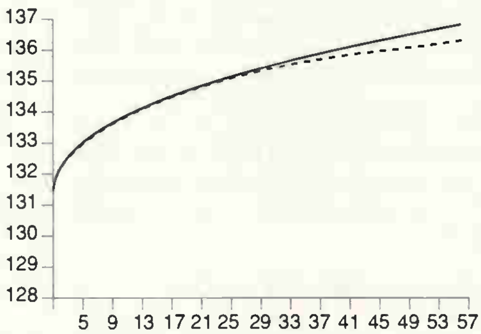
Fig. 5. Effect of charging AA manganese-alkaline battery with high duty cycles.



gave us less capacity.

We found that the internal resistance of the battery increases after each charging. After about ten cycles or so the internal resitance reaches a value that is too high to allow recharging. There was also considerable diffenence in recharging capacity between batteries of the same type.

The word recharge is a misnomer. The situation with these primary batteries is quite different from that with secondary batteries, such as nickel-cadmium types. With secondary cells, you get a complete recharge of the battery for perhaps hundreds of cycles, which suggests a complete reversal of the discharge process. This obviously does not happen with the zinc or alkaline batteries. Revitalise is a better description than recharging.



One point worth noting is that much of the folklore relating to the revitalising of primary cells originated very many years ago. Nowadays modern batteries from reputable firms are manufactured with much better quality control and have far better seals. Leakage from old batteries is far less likely now than it was 50 years ago.

Earlier batteries had a different structure from their modern counterparts. Warnings about attempting to recharge earlier batteries were no doubt valid at that time, although as we have already noted, in the fifties, wireless sets were available with a recharger for extending the life of zinc-carbon batteries.

A commercial charger

We mentioned earlier that commercial chargers are now available on the market. We were able to obtain and examine one called the '*Smart Charger*'. This product allows four batteries ranging in size from AAA to D to be charged simultaneously.

The charger has a three-colour led for each cell. Red indicates that the battery is either in the holder the wrong way or is too depleted to be revitalised. Amber indicates that the battery is being charged and green indicates that the battery is fully charged and no current is being applied to the battery.

With old batteries, it never proved possible to arrive at the green colour – even though the battery still had useful

Fig. 6. Output waveform of the Kenwill Smart Charger switches between 4.8V and 1.5V at a 50Hz repetition frequency. Its peak current is about 60mA.

energy. It is clear from our experiments that older batteries do not reach the final voltage of a new battery after multiple recharging even though they still deliver a useful capacity.

The charging pattern of the charger is shown in **Fig. 6**. It is interesting that the designer of this charger uses only one half of a mains cycle and splits it into four equal rectangular pulses.

This gives an overall duty cycle of only 35%.

Each pulse is 4.8V and the base line is 1.5V. This charging process should be much slower than ours with its 85% duty cycle. The designer has also realised that pcr is not a necessary requirement for this charger.

Is it worth it?

The main question is, is this process worth doing? Although it is possible to recharge up to about 40 times under controlled conditions, each cycle gives successively less capacity.

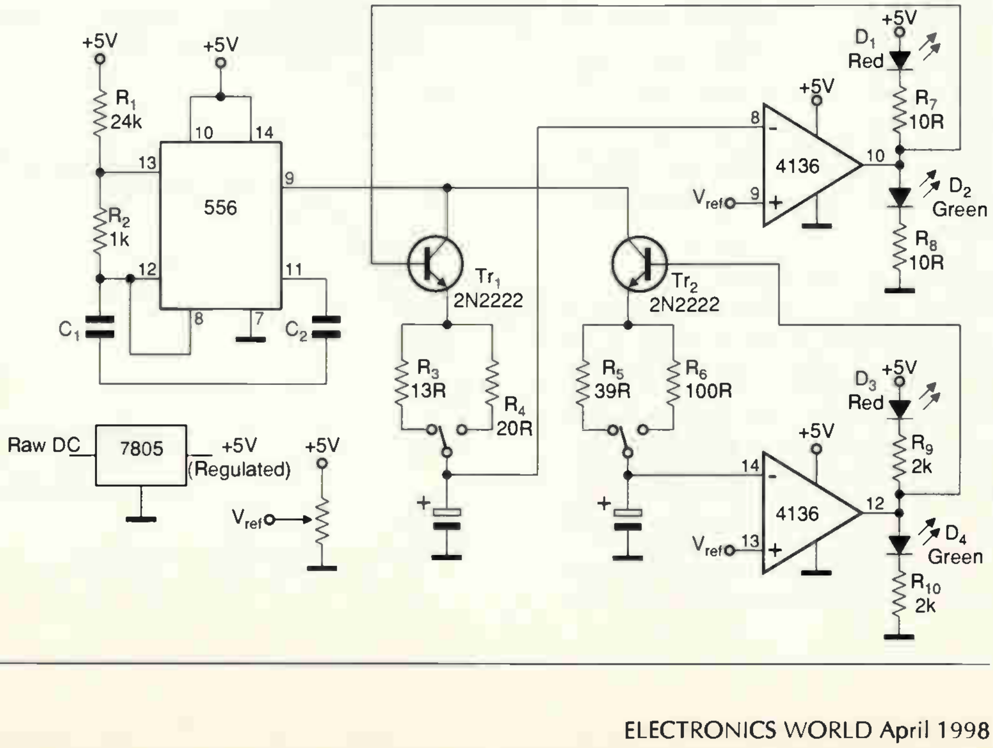
Overall, we estimate that you can only increases the life of the best of batteries by five times under optimal conditions. In general – even under optimal condition – three times the life is more common.

As mains electricity is very cheap indeed compared to the price of electicity from a battery – by a factor of about 1:500 – and the cost of such a recharger is no more that about £25 then clearly a charger is worthwhile for any company using batteries regularly.

Even for domestic users, who are not going to measure the end voltage of their batteries, there are worth-while savings to be made. One of us used the charger on batteries at home, mainly from clocks, torches and radios, for around 18 months. The batteries were only recharged when the equipment concerned ceased to function properly, but they were replaced without delay.

Fig. 7. Comparison between charging with and without pulse charge reversal using the same duty cycle. Upper curve without pcr illsutrates that the negative discharge pulse is not necessary as far as efficient charging is concerned.

Fig. 8. Primary-cell charger is kept simple since it involves no negative cycle. Our prototype had two of these charging circuits, handling up to four cells.



The charger

The circuit of the charger giving a duty cycle of about 85% is shown here. Resistors R_5 and R_9 set the current for D type batteries, R_6 and R_{10} for C type, R_7 and R_{11} for AA type and R_8 and R_{12} for A-AA type. These can be varied if you want to experiment with the values of the charging current. Power for the charger is provided by an easily available regulated ac/dc adapter rated for 500mA dc and using the 9V output.

To the right hand side of the diagram is a circuit for operating two leds per cell, red and green, which switch on according to the voltage on the battery. This is not very useful though because, as we pointed out in the main text, the end voltages of the batteries vary considerably.

With component values prescribed, the red led comes on when power is applied to the circuit. When a battery is inserted in the holder with the correct polarity, the green led lights. This green light goes out when the preset level, of around 1.6V is reached with the components shown.

If an exhausted battery is inserted, the green light does not come on because the preset voltage is never reached. Such a depleted battery can be left in place for about eight hours. However we have frequently left depleted batteries for more than 24 hours without any problems.

Batteries showing signs of corrosion should not be used under any circumstances.

During the period of this study, not a single battery needed to be replaced. This equates to a saving certainly equal to the cost of such a charger, and a more effective use of environmentally-unfriendly materials.

The alternative to charging alkaline batteries is to use rechargeable nickel-cadmium batteries. These are much more expensive than alkaline batteries and suffer from having a lower voltage, of 1.2V as opposed to 1.5V. Because of this, they are not always a replacement for alkaline or zinc batteries.

Further, the capacity of NiCd batteries is very much lower than that of alkaline batteries – at about an eighth – so you have to recharge at much smaller intervals, which can be annoying.

In summary

It appears that the recharging of zinc carbon and alkaine batteries is perfectly feasible using quite simple charging methods, without resorting to pulse-charge reversal. We consider that the dangers of these techniques have been greatly exaggerated.

While the extension of the battery life is less than that claimed in the advertisements we have seen for commercial chargers of this kind, it is nevertheless useful. ■

Watch out

Do not let what the authors say here make you complacent. Primary cells will explode or leak under adverse conditions, so please take care. Use at least eye and skin protection while experimenting and don't risk building a primary cell charger into expensive equipment – Ed.

Understanding CAPACITORS

Ceramic capacitors – how they are made, how they perform and where you should use them – by Cyril Bateman.

Ceramic capacitors possess two significant features not present in any other capacitor type. These are the ability to control the dielectric's temperature coefficient and its dielectric constant, K.

Variation of the dielectric formulation used can provide either the widest possible range of dielectric constants – and hence capacitance-to-size ratio – or alternatively the widest possible range of controlled temperature coefficients.

As with domestic pottery, these dielectric formulations can be used to produce an unlimited variety of sizes and shapes. When valves dominated electronic design, many obtusely-shape, and sometimes very large, capacitors were produced by first casting a basic shape and then machining it to size, prior to firing. Such custom designs are now very rare.

This unique ability to tailor dielectric formulations provides almost any desired temperature coefficient of capacitance,

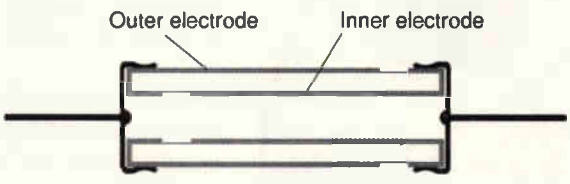


Fig. 1. Cross section of an axial leaded tubular ceramic capacitor looks and behaves exactly like a co-axial transmission line. Depending on the physical dimensions and its dielectric constant K, it can have a very low characteristic impedance. Its electrical length may be 50 - 100 times greater than its physical length, when made with high 'K' ceramic.

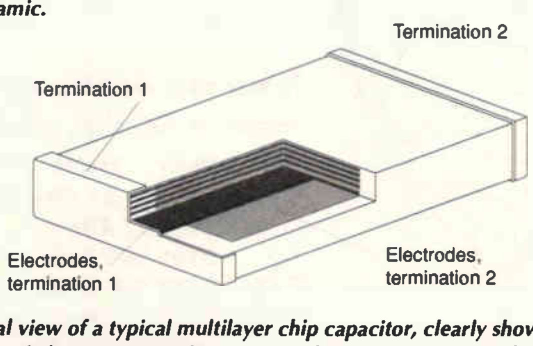


Fig. 2. Sectional view of a typical multilayer chip capacitor, clearly showing how alternate electrode layers connect in turn to each termination. Using high-K ceramic and thin dielectric layers, for maximum capacitance possible in a small size chip.

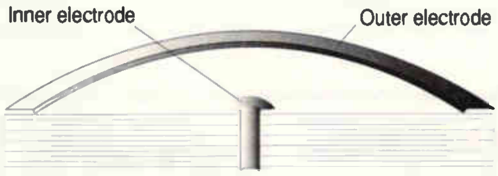


Fig. 3. Developed from the basic multilayer, this discoidal construction provides high capacitance in a small size. The lead-through centre hole and multiple grounded electrodes ensure electrical screening from top to bottom faces.

from minus 5600 to plus 100 parts per degree centigrade. Alternatively, dielectric tailoring can provide an extremely high K constant, up to 12000, giving increased capacitance. Both features resulted in the ceramic capacitors strategic importance in WWII valve operated equipments. Initially based on a tubular construction, the first ceramic capacitors were assembled using production techniques derived from established resistor manufacturing methods.

The Erie company implemented volume production of ceramics at its Pennsylvania plant shortly before WW2. Subsequently, production commenced at Erie Resistor Ltd, the company's UK subsidiary.

Using Type1 dielectric, capacitance as low as 0.5pF was possible, and with high-K dielectric up to 10nF. Both of these values could be produced with a 500V working rating. Uniquely, this tubular construction could be automatically 'silvered' to value, permitting very close tolerances, **Fig. 1**.

The disc alternative

This tubular capacitor format remained dominant until the pressed disc capacitor was developed some years later. Designed for increased volume and reduced costs, this construction resulted in a capacitor then costing little over one old pence. Capacitances from 0.5pF up to 10nF and 500V working in a 9mm diameter disc size were produced.

By pressing thicker discs, much higher voltages could be supported – even up to 30kV. Increased disc diameter up to 31mm, substantially increased the capacitance attained at any voltage. The larger disc capacitor makers were able to select from as many as fifty production-proven dielectric formulations.

Increasing use of transistors for portable radios demanded lower-cost, higher-capacitance and lower voltage disc ceramics. To cater for these needs, the so-called *Transcap* or barrier-layer capacitor was developed. On a 15mm disc, this construction could provide 0.22μF at 10V.

The sixties saw the development of the first solid-state computer, the IBM *360*, which was used for the MinuteMan missile. It required even larger capacitances in even smaller sizes – and high reliability. The recently invented multilayer[1] ceramic capacitor was seen as the ideal solution. Initially only produced in the USA, the multilayer capacitor soon became established in the UK. By 1964, volume production of UK manufactured multilayer capacitors, was established at the Erie UK factory.

Subsequent developments in the multilayer capacitor's construction and manufacturing process ensured its world-wide domination[2]. Ii was initially produced as a leaded product, dipped in encapsulating resin for commercial use, or transfer moulded for more demanding environments.

Capacitors without leads

In 1969, the basic multilayer component 'chip' was marketed without lead wires or finish, for use in the production of the new thick-film hybrid assemblies. Developments in the termination metals to prevent solder leaching and in the construction to reduce the effects of thermal shock resulted in the first generally used surface mounting capacitor, **Fig. 2**.

The military missile and NASA Apollo programmes of the late sixties demanded smaller, lighter, very-high-frequency capable, emc filters. For both requirements, these filters had to survive extremely high 'G' forces and vibration conditions, as well as temperature extremes.

To suit these needs, a special multilayer device called the discoidal capacitor was developed. It comprised a circular multilayer chip with one termination at its central lead-through hole and the other formed by the disc's periphery.

This discoidal construction provided three unique benefits. Electromagnetic signal currents presented at the 'hot' centre hole, rapidly dispersed into the 360° electrode system. Since connections extended all round the circumference, providing an almost zero inductance ground path, the discoidal provided high attenuation at the highest frequencies.

Soldered into a tubular metal casing, these grounded electrodes provided almost perfect high-frequency isolation between the 'dirty' input and the 'clean' output sides. Today this discoidal construction remains dominant for high-frequency lead-through emc filters and similar applications. It is used whenever screening between a dirty input and clean output is needed, **Fig. 3**.

Planar array ceramics

Developed from these early discoidal concepts, the latest multilayer capacitors include multiple-lead-through emc filter capacitors known as planar arrays. These are produced with mechanical dimensions compatible with commercial and military connectors, or as a custom ordered part having almost any desired shape or size.

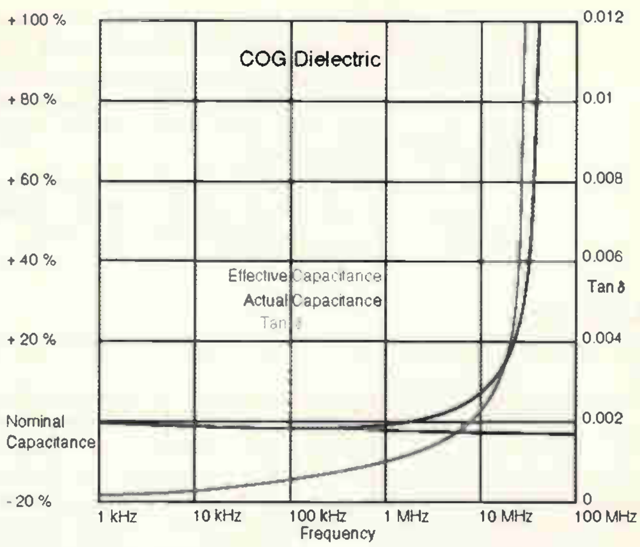


Fig. 4. Since the K of COG dielectric is almost constant with frequency, the sharp increase in measured or effective capacitance, also tan δ as resonance approaches, can be seen. This increase results from the combined reactances of the actual capacitance and the capacitor's self-inductance.

While covering the main stream ceramic products, this brief overview deliberately skips over many styles that were important at their time but that are now redundant. These include the lead-less disc, the wedge capacitor, the very-thin single-layer 'Weecon', the colour-television tripler/quadrupler capacitor, the very-large and thick extra high voltage cupped capacitor, and the wound or 'rolled' tubular multilayer ceramic. The wedge capacitor was common in early vhf amplifier and tuner assemblies while the wound tubular multilayer design was popular in the early sixties in USA for use in emc filters.

Shocking soldering

In the seventies, as more sophisticated or leadless ceramic capacitors became generally used, thermal shock during soldering was a major discussion topic and the cause of most early capacitor failures.

At that time, a second serious soldering problem using conventional methods was that the terminations of these non-leaded parts 'leached' away. Unless the soldering time and temperature were minimised, the precious metal termination would simply dissolve into the molten solder, resulting in a visible capacitor failure. Unfortunately, steps taken to minimise thermal shock often aggravated the leaching problem.

Following significant improvements in termination materials and electroplating, the amount of time that the solder can remain liquid has increased ten-fold. Electrode and termination leaching has been eliminated and most capacitor makers now publish clear soldering guidelines.

Thermal shock creates minute cracks within the ceramic

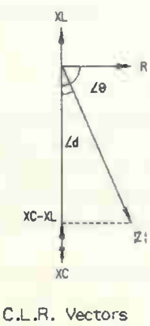


Fig. 5. This vector diagram shows the relationship between tan δ and cos θ. Both are essentially equal at small loss angles. The interaction between the capacitance X_c and self inductance X_l vectors, which causes series resonance, is also visible.

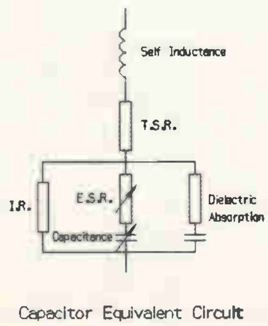


Fig. 6. This equivalent circuit for a practical capacitor is valid for all frequencies up to self resonance and perhaps one octave higher. When mounted onto a microstrip line, an additional small capacitance, not shown, between the capacitor mounting points to earth will also be present.

dielectric, generally close to its interface with the end termination materials. Initially these cracks might be small, and the capacitor will appear undamaged. But with time, operating temperature fluctuations and mechanical stress due to printed board flexure, these cracks grow. Ultimately the capacitor fails, usually as a short circuit.

You might question why ceramic capacitors are so prone to this damage. After all common household ceramics or pottery seem able to survive considerable temperature stress.

Self inductance

Some readers have queried my description[9] that physical size equates to self inductance.

The self inductance of a single straight wire or a track on a printed circuit board can me measured and calculated. Similarly, self inductance for a ceramic chip capacitor can also be estimated, given the capacitor's electrode shape and size. However this self inductance can more easily be measured.

Measurements of the K value by frequency of a test block of dielectric in a dielectric test jig[12] confirms that COG types have a K value almost constant, reducing only slightly with frequency. But high-K materials exhibit a substantial reduction.

Consequently, using COG dielectric, the observed series resonance frequency together with the 1MHz capacitance value can be used to determine the capacitor's self inductance with acceptable accuracy, **Fig. 4**.

Conventional measurements of impedance magnitude – usually published in makers data in logarithmic format – show impedance reducing with frequency to a sharp minimum at resonance, then increasing with further frequency increase. Since most designers prefer to use capacitors below the series resonance, published plots usually stop soon after this resonance has been shown.

Measurement of the chip capacitor's impedance as a vector of resistance (esr) and reactance (j*X*) initially follow exactly this same pattern. But at this series resonance frequency, they clearly show reactance crossing the zero line – a change from negative to positive phase angles with increase of frequency, **Fig. 7**.

For these measurements to be valid, it is essential to eliminate all test jig parasitics, both capacitance and inductance, by applying open, short and load calibration of a vector network analyser. This must be done at each measurement frequency, and at the precise point in a transmission line where the test capacitor will be mounted. Next, 12-term, also called 'full 2 port' error corrections must be applied.[8]

When it is not feasible to apply the open, short load, calibration as above, calibrate at the ends of the test coax cables, which attach the test jig and apply 12 term correction. Characterise the test jig by measuring it with an open, short and known impedance, in place of the test capacitor, at each frequency of interest. These measured error values are then applied using established equations as a means of error reduction. Remember though that each of the measurement and error terms described by the correction equations are vectors or complex numbers and must be calculated accordingly.[13] For open short and load,

$$Z_{dut} = \frac{Z_{standard.true}(Z_{jig.open} - Z_{standard.measured})(Z_{dut.measured} - Z_{jig.short})}{(Z_{standard.measured} - Z_{jig.short})(Z_{jig.open} - Z_{dut.measured})}$$

While I plan to detail to these capacitor measurement techniques in a later article, to demonstrate this self inductance, I took four capacitors, previously characterised up to 1MHz. Using an HP4815 vector impedance meter, I measured all four capacitors at frequencies from 1MHz to 100MHz. and applied the 'open/short/load' error-correction algorithm shown to the measured values, **Fig. 7-10**.

To simplify the calculation and plotting of these results – which involved many complex numbers – I stored the measured jig correction values by frequency in a database and used my custom written, dedicated 'open/short/load' error correction calculation and plotting software.

From this resonant frequency and the inductive reactance with increasing frequency, the capacitor's self inductance can be calculated.

The capacitor ceramic formulations used are quite different, comprising in most cases a mixture of many materials. While the high K formulations may be based on barium titanate, small quantities of many oxide or rare earth additives, together with 'frits' or glasses will be used.

During firing, or sintering, the dielectric is raised in a controlled manner to very high temperatures. Firing temperatures are deliberately chosen to avoid notable melting or alloying of these constituents, or any significant crystal growth.

A multilayer capacitor is essentially a multitiered sandwich of alternating layers of thin ceramic dielectric and precious metal electrodes. The whole laminate of metal and ceramic is co-fired together, the electrode metals must not oxidise or melt at the firing temperatures needed. As a result, precious metals are used.

Dielectric ceramic and the precious metals used have quite different expansion coefficients and thermal conductivity. Temperature differences within the capacitor can result in mechanical stress due to differential expansions, both within the basic ceramic materials and at the ceramic metal interfaces.

Development of sintering methods, dielectric formulations, electrode and termination materials, have produced dramatic improvements, resulting in today's ceramic chip capacitor found on almost every surface-mount circuit board.

But even with today's much improved products and production methods, a degree of thermal shock hazard remains. Since the capacitor ceramic is extremely hard, brittle and easily chipped, the most common problem results from mechanical cracking. This is induced by poorly adjusted assembly and board testing equipment. Each surface mounting ceramic chip capacitor maker provides handling, soldering and mounting recommendations that must be observed.

How is the dielectric made?

So just how are these ceramic dielectrics manufactured?

All ceramic and precious metal-electrode ink formulations are closely guarded, proprietary secrets. In addition, these formulations, and the capacitor processing methods used, are intimately interdependent and so I can only discuss them in broad outline.

With many dielectric formulations, minute percentages of additives and dopants are be used. Consequently, levels of impurities in each ingredient are important.

Having satisfied conventional chemical and physical analysis, each new delivery of a material is used to make a batch of test capacitors. Performance of these capacitors is verified against the test specifications, including microscope examination of capacitor sections.

Each ingredient is weighed out, added to a solvent and placed into a rotating 'ball' mill, together with the correct weight and sized balls of flint or borundum.. The mill is rotated for many hours, even days, to blend together all ingredients and achieve any needed ceramic particle size reduction.

Having removed the grinding balls, these ingredients, now finely dispersed and suspended in the solvent, are injected as a fine dry powder, needing only analysis to confirm correct particle size. The tower needed to dry a typical 50kg charge of powder may be as tall as 20ft.

Traditional ceramic capacitors

The powder can be used to make capacitor dielectric in several ways. For tubular capacitors it is mixed with solvent and plasticiser and machine kneaded to an air-free, stiff, dough-like consistency. This dough can be pressure extruded, like toothpaste, to the desired tubular dimensions. Dried and cut to length, it is ready for firing.

For disc capacitors, the powder is used dry, being fed into

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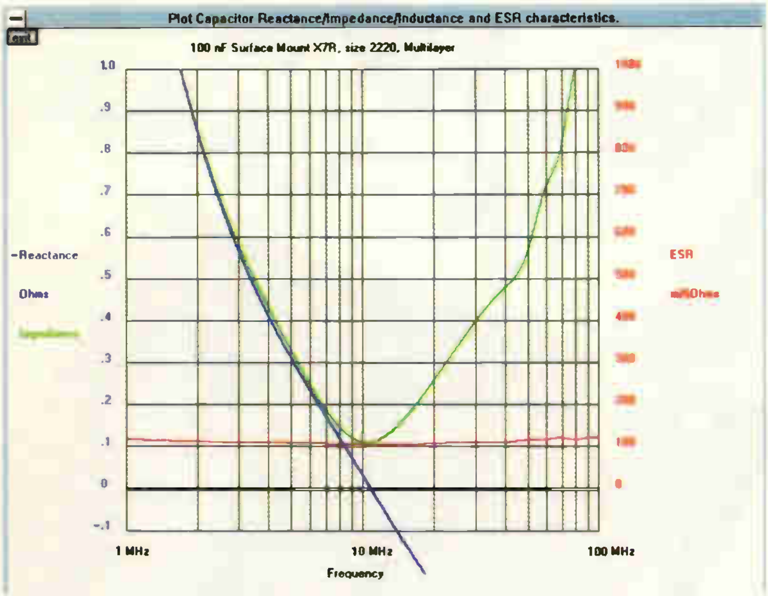

Fig. 7. This specially produced plot of measured reactance, impedance and esr of a typical 100nF X7R chip, shows reactance changing from negative to positive, at resonance. The impedance plot increases above resonance as expected, from a minimum value, equal to the measured equivalent series resistance. For clarity only, the reactance scale has been inverted, having negative values in the upper plot.

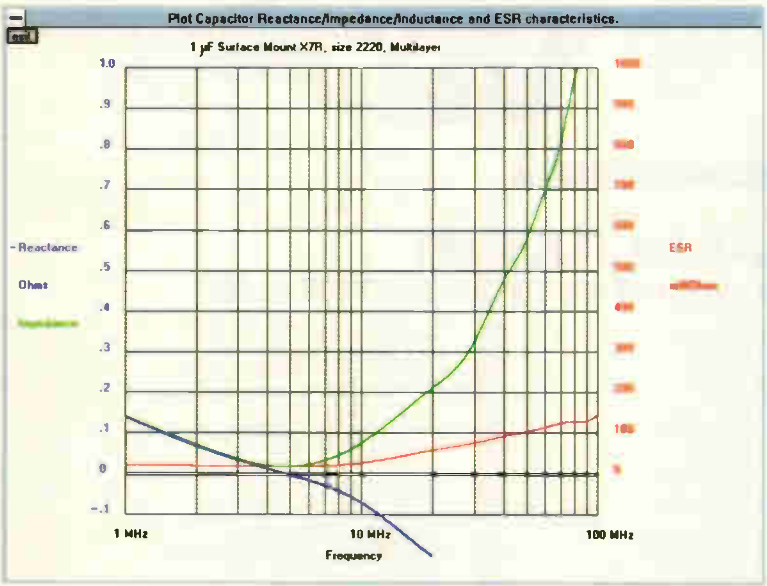

Fig. 8. Similar to Fig. 7 – same size chip and typical 1μF X7R capacitor. As expected the resonance frequency has reduced, but otherwise all curves follow the same trends.

tungsten carbide press tools and subjected to very high pressure presses. High volume production discs can be pressed using multi-station, high speed, single action rotary turret presses, similar to those used to make medicinal pills. Larger discs and pressed tubes however need double-action presses, usually single station, equipped with both top and bottom acting press tools.

Multilayer ceramic capacitors

Early multilayer capacitors were made using the 'dry film' method.[1] The required thin dielectric film was produced by passing a plastic film carrier through a plasticised ceramic 'slip' at controlled speed, to emerge coated with ceramic. After drying, this heavily plasticised ceramic could be stripped off as a thin, self supporting, flexible ceramic film.

To produce the alternating electrode patterns needed for a multi-layer device, the ceramic was screen printed with electrode ink. The dry electroded ceramic film was stacked, layer on layer, to produce a 'stick' of capacitors. Compacted under high pressure, these sticks were then diced into individual capacitor chips, ready for firing.

Later developments allowed similarly thin ceramic films to be continuously cast onto a moving polished stainless-steel belt. This eliminated the plastic carrier film and needed smaller amounts of plasticiser.

The reduction of plasticiser content is important. While, as with domestic pottery, all ceramic shrinks during firing, shrinkage increases with increased quantities of plasticiser and organic binders.

Introduced in the early eighties, the modern 'wet-film' method[3] involves screen printing, in turn, unplasticised ceramic and a matrix of electrode ink patterns. These are precision printed on to a dimensionally stable carrier.

After hot-air drying, the carrier is returned to the printer head and further layers of ceramic or electrode patterns are added until the capacitor is complete. Following final drying, these sticks of capacitors are separated into individual capacitor chips, ready for firing.

With either assembly method, the electrode inks used depend on the ceramic formulations and their firing temperature. For lower firing temperatures, a combination of palladium with silver is often used. At higher firing temperatures, the silver would melt and globulate, so it must be replaced by more costly gold or platinum, depending on the temperature.

Firing the ceramic

Two main methods of firing the ceramics are used. Traditional ceramic capacitors produced in large volumes are often fired in long, temperature controlled multi-zone, continuously operated 'sagger push' kilns.

The pusher kiln is built using high-density refractory bricks and can take days to bring up to temperature and stabilise. More modern, smaller and lighter batch kilns are quite different. Built using extremely low density refractory linings, they allow firing cycles with pre-set individual time and temperature profiles.

As with conventional household pottery, the fired ceramic shrinks significantly from its unfired or 'green' size. While principally dependant on the amounts of organic binders and plasticisers used in the ceramic formulation, change of the firing profile also affects shrinkage.

Organic materials and plasticisers in the raw ceramic have to be removed by a slow pre-fire before the final firing. A complete firing cycle may take as long as two or three days.

The 'wet film' multilayer and the pressed disc processes use the least quantities of organic and plasticiser so shrink least of all. Shrinkage is especially important where a precise finished size ceramic is needed, as for the planar array, multi-hole, capacitor designs.

For both kiln types, very precise control of temperature and the time/temperature profile is essential, to standardise ceramic shrinkage and optimise capacitor yield.

Adding a means of connection

The fired multilayer ceramic capacitor needs only the application of external precious metal ink terminations, to contact the electrodes to produce a capacitor chip.

Disc or tubular ceramics first have to have their precious metal electrodes applied. This is usually silver ink. The external terminations and the silver electrodes are fired for a much shorter time and at lower temperatures than the ceramic.

Continued on page 330

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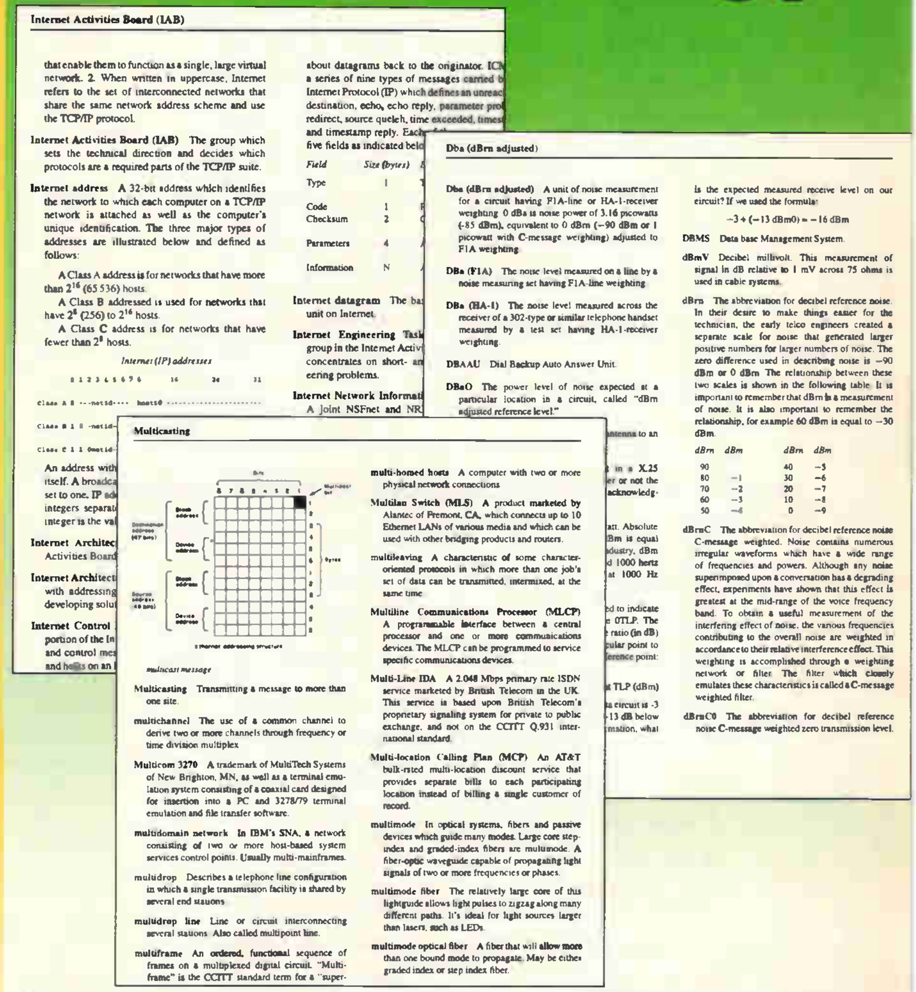

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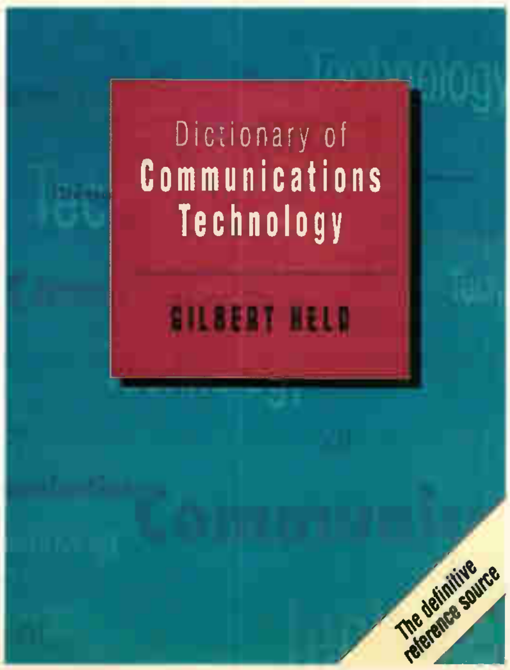

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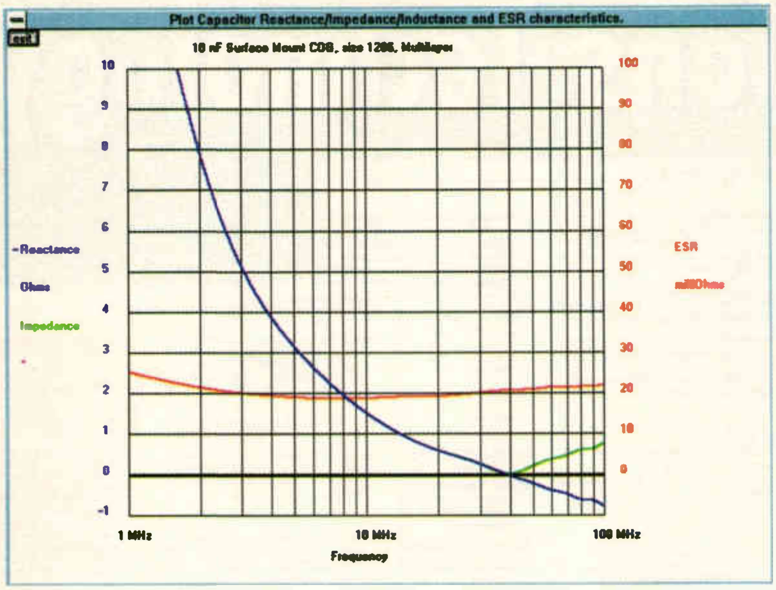

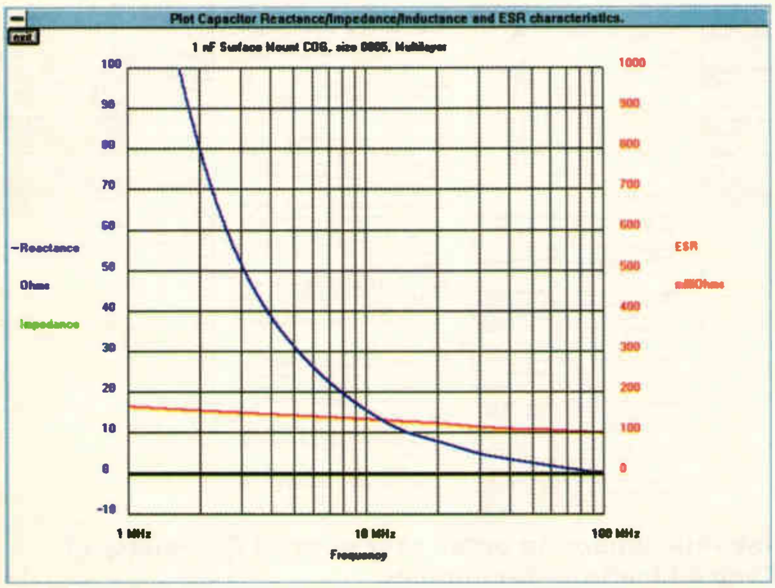

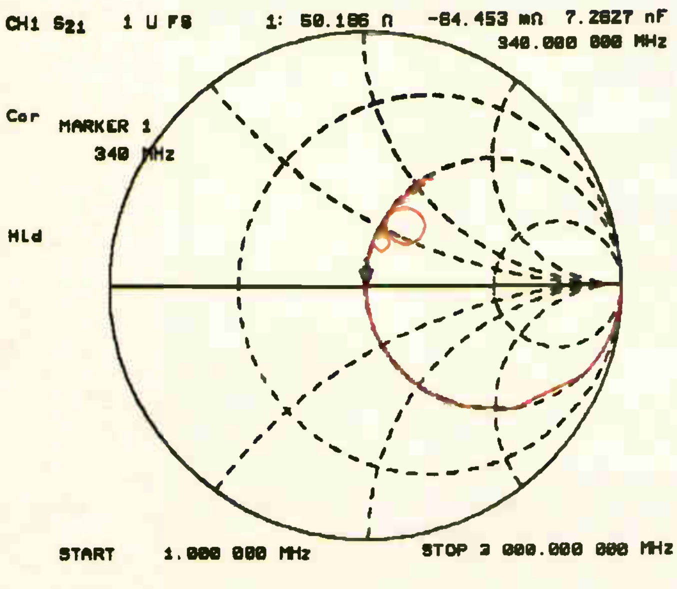

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Top: Fig. 9. This plot of a 1206 sized 10nF COG chip shows a much higher resonant frequency, resulting from its smaller physical size, hence less self inductance, and its reduced capacitance.

Middle: Fig. 10. A similar plot of a 1nF COG chip shows resonant frequency exceeds the 100MHz measured. This plot was deliberately taken to demonstrate the effectiveness of the error correction method. Maker's data suggests resonance should be 150MHz. Uncorrected resonance, due to jig parasitics, was measured at 56MHz.

Bottom: Fig. 11. This Smith chart S21 impedance plot of a 300pF chip mounted as 'shunt' to ground, on a 50Ω microstrip line jig, shows the trace changing sign at series resonance. Above this series resonance frequency, other resonances can be seen which remain in the upper inductive half of the chart. The low frequency straight sections result from values changing rapidly between measured frequency points. Courtesy Syfer Technology.

The barrier layer or Transcap capacitor differs in that the basic ceramic discs are fired twice. Following the conventional oxidising firing in air, these capacitors undergo a second firing in a reducing atmosphere. This renders the ceramic material conductive, with only a few ohms resistance.

Once the termination silver has been fired in air, two capacitors, one on each face, are formed about the conducting central core. Their dielectric is extremely thin. This reduces the working voltage, but increases the capacitance tenfold.

After termination or electrode firing, the capacitor is ready for any final assembly processes. Lead wires can be attached and the capacitor encapsulated. Surface-mount chip capacitors may have their end terminations hot solder coated, electro-tinned, nickel plated or silver-palladium coated depending on the application they are intended for.

In manufacture, all heat process stages are strictly profiled and controlled. To ensure this, each stage of production may be subject to statistical process control analysis of yields.

All soldering operations use a controlled pre-heating stage. Terminations or electrodes incorporating fired silver-bearing inks are soldered using an alloy pre-saturated with silver to minimise leaching. One such alloy is LMP or SN62. This product is tin, lead and silver in the proportions of 62/36/2. It is a true eutectic, i.e turns immediately from solid to liquid, which minimises the possibility of dry joints.

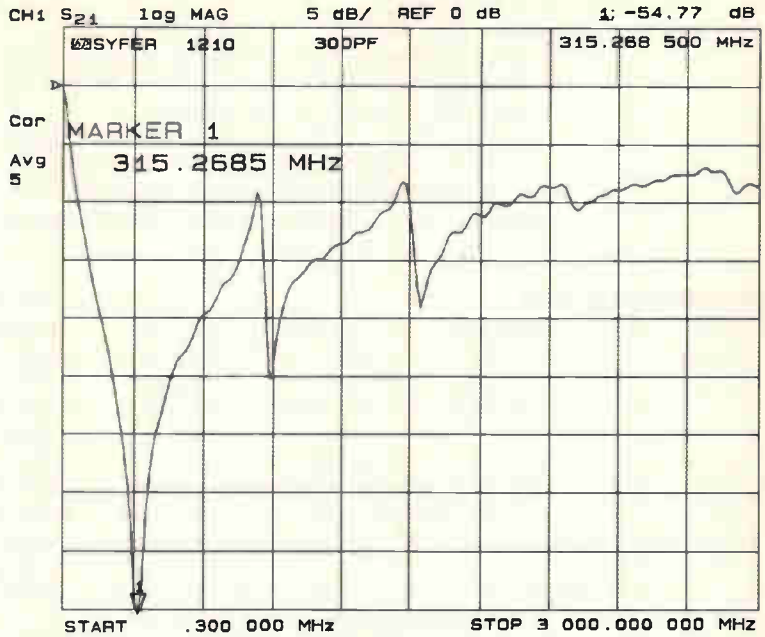
Other metals can dissolve or leach into molten solder so the choice of solder alloy is important. If fine copper wires need tinning, for example, copper loaded solder is a good choice.

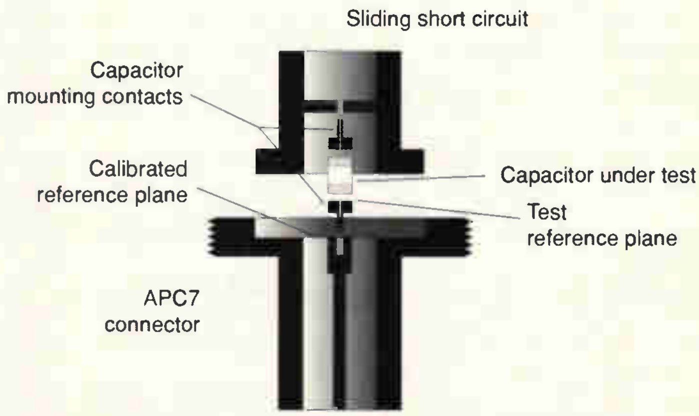
Dielectric characteristics

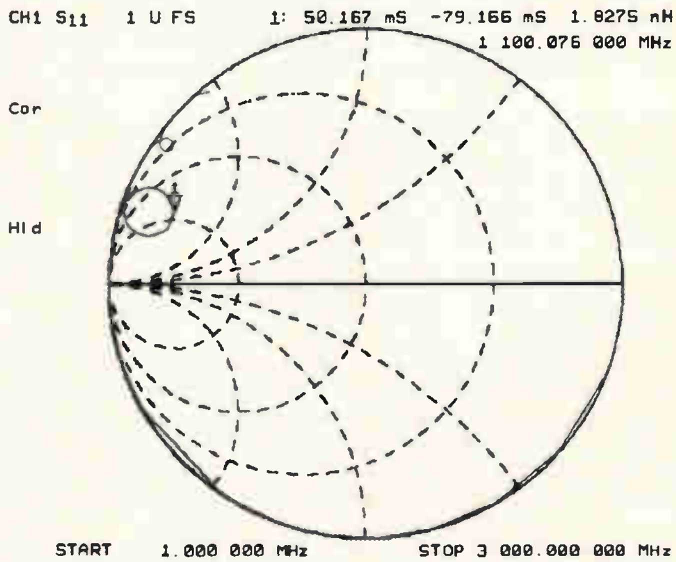
EIA Class I, ultra stable, COG. For almost all circuit needs, the conventional COG ceramic provides the nearest perfect capacitor, having better characteristics than mica or any film type, except perhaps for PTFE. It has a temperature coefficient of ±30ppm, negligible voltage coefficient, capacitance stable over time, negligible capacitance variation at high frequencies, minimal dielectric absorption and a 'Q' approaching 1000 at 1MHz, **Fig. 4.**

A capacitor's effective or measured capacitance rises sharply at series resonance. This is the effect of the positive and increasing reactance of its self inductance, approaching then exceeding that of the negative and reducing reactance of the actual or true capacitance at the resonant frequency.

Apart from the steady decline in K which most dielectrics exhibit with frequency increase, the K value of the dielectric is unchanged at resonance, **Fig. 5.** The panel on capacitor self inductance provides more on this topic.







Top: Fig. 12. This rectangular plot of Figure 11 clearly shows multiple high-frequency harmonic resonances occurring, having smaller amplitude nulls than that of the series resonance. These amplitudes reduce with frequency, as capacitor losses increase. Courtesy Syfer Technology.

Middle: Fig. 13 Schematic section of the Hewlett Packard 16091A, low cost coaxial test jig. Each capacitor chip being pre-soldered to two contact pins. The 1mm displacement between calibration and measurement planes can be mathematically corrected, eliminating all jig parasitics.

Bottom: Fig. 14. Plot of an actual measurement of a COG 600pF 1210 chip using the Hewlett Packard 16091A coaxial test jig. The capacitor is shown as an inductance of 1.8275nH while at a higher frequency resonance. Its series resonance is at 185MHz. For ease of calculating corrections, this admittance plot is preferred to the more common impedance presentation. Courtesy Syfer Technology.

Since the surface-mounted ceramic multilayer COG chip provides a large capacitance in a small size, together with low self inductance at low cost, it is invaluable for high frequency circuits. At the high end of the rf spectrum, especially when stripline circuits are involved, an improved high-Q version[3,4,5] is also available. For low capacitance values, this device can offer a Q of around 1000 at 100MHz.

There are two further types of capacitor that maintain these high Qs to even higher frequencies, albeit at higher cost. These are the ACCU-F[6] – a thin-film capacitor made using silicon dioxide and silicon nitride – and the porcelain multilayer chip[7] capacitor.

Using the best available *LCR* meters at 1MHz, differences in Q between these styles cannot be measured with certainty. Such performance differences can only be measured at higher frequencies, using the best possible jig techniques, together with a 12 term error corrected vector network analyser.[8]

High-K dielectrics

Higher capacitance values are available using high-K X7R, Y5V and Z5U dielectrics. These materials provide the best capacitance, voltage and size combination of all, by trading off temperature coefficient, capacitance loss with frequency and voltage coefficient.

The stable X7R, EIA Class II and the general purpose Z5U/Y5V, EIA Class III grades, are controlled according to windows of temperature coefficient. While each maker's capacitors will comply with these windows, different makers' temperature coefficient curves may follow quite different contours. These high-K capacitors exhibit change of capacitance with voltage, frequency and temperature.

Based on barium-titanate, which is a polar material, high-K capacitors also exhibit a modest level of charge retention. This is called dielectric absorption, and I hope to explore this topic in a future article[9].

The crystalline structure of ceramics based on barium-titanate changes on being heated above the Curie point, which is notionally 125°C. On cooling, this domain structure relaxes with time, reducing the K value in a logarithmic manner. It effectively becomes stable after 1000 hours. As a result, capacitors made using these dielectrics are manufactured to age down to nominal capacitance, having rested or aged for 1000 hours following their last heat cycle.

Any subsequent soldering process that takes the dielectric over 125°C restarts this ageing cycle. Full details of the ageing factors to be applied will be found in makers' data sheets.

Capacitors based on high-K ceramics provide the best possible capacitance versus volume performance while maintaining stability within defined limits. The small physical lengths permitted ensure very low self inductance, extending the upper limits of the high-K capacitor's frequency range.

You may wonder just why a low self inductance value matters. At some frequency, the reactance of this self inductance becomes equal in value to that of the capacitor's reactance. But since the two reactances are opposite in phase, a series-resonance effect results.

At higher frequencies, while still blocking dc, the capacitor otherwise behaves as though it were an inductor, thus limiting its upper usable frequency for most circuits.

As the clock frequency of logic chips becomes ever higher, the decoupling of switching spikes becomes increasingly difficult. Any self inductance in the decoupling capacitor slows or impedes the rate at which the capacitor is able to charge or discharge in order to decouple these spikes.

An easy way to lower a capacitor's self inductance is to reduce its physical length or increase its width. But reducing the inductance further is only possible using specially designed capacitors. One company particularly active in reducing capacitor self inductance, AVX, offers several application notes via its Web page and on cd-rom.[10]

Allowing for all these variants, a capacitor's behaviour at

Measurements at rf

While the 100MHz measurements suffice for general applications, for rf design they are little better than dc measurements. While specialist capacitor measuring equipment is available[14,15] to measure up to 1 and 1.8GHz, it is very expensive and not available in many laboratories.

Two suitable instruments more commonly available are the *HP8510*, which measures from 45MHz to 110GHz, and the *HP8753* which measures from 300kHz to 6GHz. I have used each of these vector network analysers. Both have 12-term error correction.

The Smith-chart presentation provided by these analysers is particularly relevant when demonstrating the self inductance of a capacitor. It displays the measured impedances by frequency, in terms of R+–jX, where jX=j(X_C–X_l).

When jX has a negative net value, i.e. a capacitive reactance, it will be plotted below the horizontal centre line. When jX has a positive net value, or an inductive reactance, it is plotted above the centre line. When jX is zero, i.e. resistive, it is plotted along the central horizontal line.

On this impedance chart, the right-most limit of the central line represents infinity while the left most limit represents a short circuit. Consequently, when measuring a capacitor, the frequency at which the trace crosses the central line represents the capacitor's self resonant frequency.

In the example shown, this occurred slightly below 340MHz. The small circular loops, or whirls, represent a resonant behaviour, but since the trace remains above the centre line, the capacitor has continued to present an inductive reactance, **Fig. 11**.

Obviously, the highest frequencies pose considerable problems when it comes to designing and characterising test jigs. While commercial jigs are available from Wiltron[16] and Hewlett Packard[17], they are extremely expensive. However up to perhaps 3GHz, and ignoring phase, relatively simple jigging can be used to measure resonance, and hence self inductance of low value capacitors. The same jigging can be used to observe transmission line behaviour.

S_{21} measurements. For magnitude parameters only, the easiest method is to measure the chip's insertion loss when mounted as a shunt to earth in a good 50Ω microstrip line. While for a quick looksee FR4 board could be used, its characteristics change dramatically with frequency, making it impossible to produce a good wideband 50Ω line.

For these measurements, I fabricated accurate lines using Rogers *Duroid 5880* PTFE board[18], gold plated, with *APC3.5* surface launching connectors.

You may question why this insistence on using a good 50Ω line. Almost all modern measuring systems have a 50Ω input and output impedance, as does suitable quality connecting cables. Using a jig of impedance other than 50Ω produces a reflection of power at the point where the jig is connected to the 50Ω system. This reflected power reduces that incident on the test capacitor, giving misleading results.

For insertion loss, or S_{21}, measurements this jig mismatch also causes reflections, hence ripples in the receiver measurements. Unless the vector network analyser used has capability for 'Thru/Reflect/Line' jig calibration,[19] these errors cannot be removed. As a result, they must be avoided when designing and characterising any test jig.

The rectangular plot of S_{21} insertion loss of a capacitor in a 50Ω microstrip line, mounted from line to ground, clearly shows the capacitor self resonant at 315.2685MHz. A second, parallel resonance, resulting from stray capacitances with the chip self inductance might also be expected at a very much higher frequency.

Other resonances are clearly visible, but since the net reactance shown in the Smith plot remains inductive, these cannot result from either series or parallel circuits. These resonances exactly mimic those seen when measuring low impedance, high capacitance transmission lines that are grossly mismatched. As frequency increases, the chip's losses increase, reducing the amplitudes of the higher frequency resonances, **Fig. 12**.

S_{11} measurements. While the above measurements can characterise all four possible S-parameter characteristics, S_{11}, S_{21}, S_{12} and S_{22}, of the capacitor in a two port system by frequency, an alternative single-port characterisation of S_{11} only is also possible. This measurement can determine resonant frequencies, capacitor self inductance, equivalent series resistance and capacitance by frequency.

One special benefit of measuring only S11 is that a commercial test jig is available for smaller chip capacitors, at a reasonable cost. The *HP16091A* jig requires the chip to be soldered to two contact header pins, which permit the capacitor to be inserted into the central conductor of either a 7mm or 10mm variable length coaxial airline.

The point of connection of this jig to the measuring system can be fully calibrated using *APC7* open, short and load impedance standards, **Fig. 13**.

The header pins displace the test capacitor by some 1mm from the calibrated reference plane and introduce parasitics of resistance, inductance and capacitance. In the measurement shown, this test jig was used, so the on-screen display as plotted, includes jig parasitics.

Similar looping but inductive resonances are clearly visible and the self inductance of the jig mounted capacitor, is shown as 1.8275nH. While the Smith chart default is to display impedance, the S_{11} plot of admittance values shown, facilitates calculation of the error corrected results, **Fig. 14**.

Using suitable open and short-circuit dummy capacitors, these test jig strays can be separately quantified and used for error reductions using a computer. This results in an accurate measurement of S_{11}, for the capacitor being tested.[20]

$$R_x = \frac{G_m - G_o}{((G_m - G_o)^2 + (B_m - B_o)^2)} - R_o$$

$$X_x = \frac{B_o - B_m}{((G_m - G_o)^2 + (B_m - B_o)^2)} - X_o$$

Where,

$G_m + jB_m$ = measured admittance of DUT in siemens

$R_o + jX_o$ = measured jig 'shorted' in ohms.

$G_o + jB_o$ = measured jig 'open' in siemens

This error corrected S_{11} measurement, can easily be converted into R+–jX, hence any other desired impedance parameter, exactly as described in the December 1997 article.[9]

frequencies below the series resonant point – and perhaps to one octave above it – is described exactly by the common equivalent circuit shown earlier.[9]

Simple measurement of impedance magnitude suffices to demonstrate this behaviour. All makers' data books include similar impedance frequency plots, permitting easy comparisons of resonant frequency and impedance minima between capacitor styles. **Fig. 6.**

However at higher frequencies, this simple circuit does not equate with the observed behaviour of practical capacitors.

With increasing frequency, the capacitor may exhibit either a single higher frequency parallel resonance or multiple resonances – exactly as are found with a mismatched transmission line.[11] These characteristics will depend on the dielectric's behaviour and the electrode resistances.

This behaviour is clearly measurable with porcelain and high-Q ceramic multilayer chips. Consequently Dielectric Laboratories[6] – a specialist maker of very high frequency capacitors – provides the transmission line simulation program *CapCad* for use with its products ■

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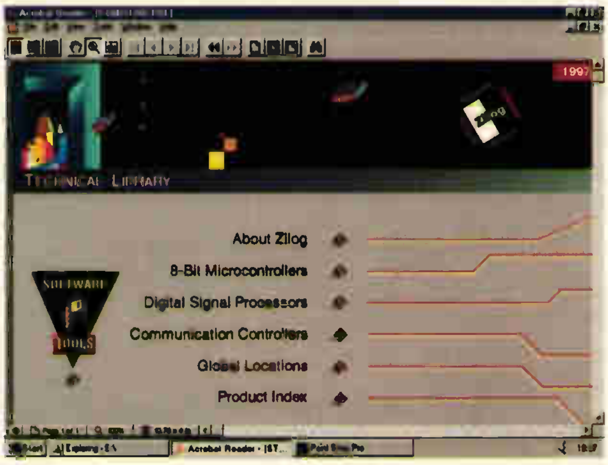

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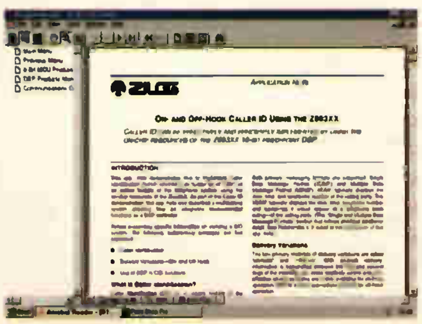

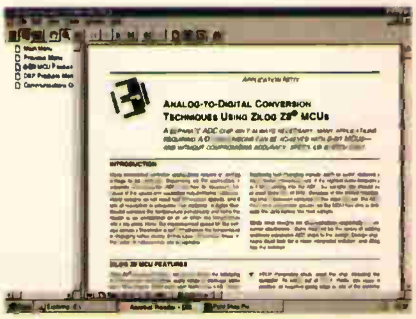

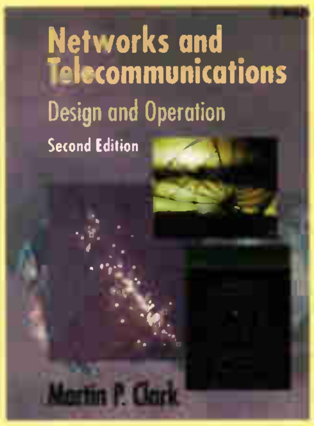

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Telecommunications network design and operation is now widely regarded as an issue of business management as well as electrical engineering. In this updated edition, Martin Clark, a pioneer of this perspective, applies it to the increasing complex and diverse realm of voice, and data and multimedia networks.
Written in an accessible style and clearly illustrated throughout, this is a basic, practical and intuitive insight into modern network engineering and sections including:

- Technical accounts of modern voice, data and multimedia networks
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- Practical aspects of running and setting-up networks
- Running a business based on telecommunications

A text specifically for readers new to the whole subject of telecommunications, and professional telecommunications managers who need an introduction and reference work on all aspects of technology, operational techniques and regulation.

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Introduction to Digital Mobile Communication

Yoshihiko Akaiwa

A comprehensive treatment of the digital technologies that make personal mobile communication a reality

Although today's mobile communication engineers and designers can build upon the advances in digital telecommunications, specific technical requirements - robustness against fast fading,

spectrum and power efficiency, and the demand for low-priced equipment- post new technological challenges that demand creative solutions.

Introduction to Digital Mobile Communications is a comprehensive treatment of the digital technologies that are rapidly spawning new advances and applications. Written by a pioneer in the field, this book covers all the important concepts, from the fundamentals of signal analyses and digital communication to descriptions of the latest transmission systems. Rich in detail and broad in its coverage, this remarkable book:

- Describes equipment and circuit implementation methods and their performance characteristics
- Discusses elements of and methods for digital modulation and demodulation schemes
- Provides practical designs and circuits for spectrum-efficient modulations
- Covers mobile radio channels and digital mobile radio systems
- Includes extensive mathematical treatments and mathematical models
- Presents the latest research results with detailed references

This valuable resource provides a solid introduction to mobile radio communication for the students or professional in related fields. Most important, for design engineers and equipment manufacturers, its up-to-date findings will stimulate new research and creative design and system development efforts.

Yoshihiko Akaiwa is a leading researcher in the digital mobile communication field. Currently a professor at Kyushu University, he worked as a researcher for over twenty years at the NEC Corporation

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Cellular Mobile Radio Systems

Designing Systems for Capacity Optimization
Husni Hammuda, Ericsson (UK) Ltd

Rapidly increasing demand for mobile radio frequency subscription is already pushing cellular networks to the point of overload. Of the various methods which are being explored to tackle this problem one of the most notable is the integration of advance modulation and multiple access techniques. In this book, Husni Hammuda, a pioneer of this hybrid, shows how it can be applied in practice to optimise the efficiency of mobile radio cells.

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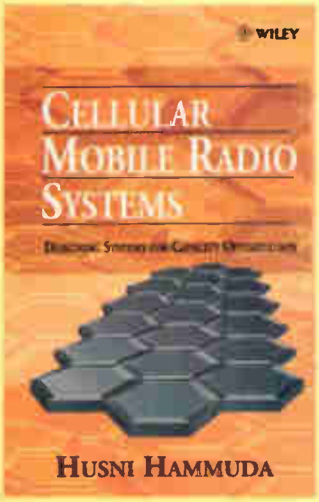

the evaluation of combinations of modulation and multiple access techniques
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- Covers recent innovations in personal communications

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A Practical Guide to Acquisition, Storage, Conversion, Compression and Transmission
Gilbert Held, 4-Degree Consulting, Macon, Georgia, USA

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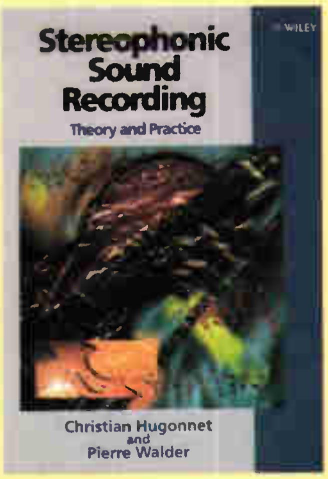
Stereophonic Sound Recording

Theory and Practice
Christian Hugonnet and Pierre Walder

Recent advances in digital audio have heralded substantial innovations in sound recording techniques and increased the importance of applying the latest microphone techniques. The authors of this book focus on these innovations, giving numerous examples of their use within the framework of an analysis-based recording



engineering theory.
The book provides a complete overview of well-known sound recording procedures practised world-wide, whilst also presenting a methodology that will provide the reader with an efficient approach to sound recording of classical music, rock and pop music, drama and speech. The widely illustrated theoretical knowledge is presented in clear and simple language.
Building on their considerable experience of creating innovative recording techniques, the authors provide an authoritative analysis of the subject that offers valuable, practical guidance that will aid the development of new recording methods. Their inside knowledge of the requirements of the phonographic, broadcasting, film and other media industries ensures expert coverage of new products and approaches including:

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For professional audio engineers, this manual provides systematic advice for getting optimal performance from studio equipment. For students of audio engineering it will form a comprehensive introduction to the area of stereophonic recording, backed up by real-world case studies and a wealth of practical experience.
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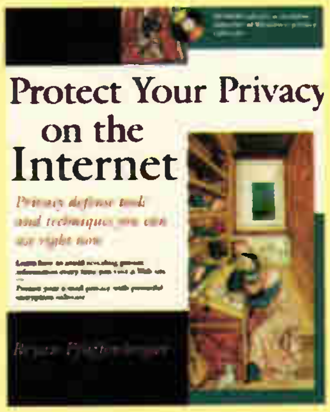

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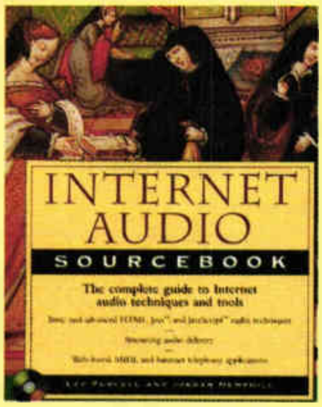
Internet Audio Sourcebook



The complete guide to Internet audio techniques and tools
Lee Purcell and Jordan Hemphill

"The World Wide Web has spouted vocal cords, gained a voice, and begun to sing." – Lee Purcell and Jordan Hemphill

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Finally, the authors show you how to get the most out of new audio technologies, including streaming audio, MIDI applications, voice synthesis and recognition, and Internet telephony, as well as covering hot new products like the Headspace Beatnik audio engine and Liquid Audio.

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You get fully functional demos of top-of-the-line sound processing applications, including Sonic Foundry's Sound Forge and Hohner Midia's Samplitude Studio. Authoring tools such as Symantec's Visual Café, Aimtech's Jamba, and Acadia Software's Infuse JavaScript editor are also included.

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A Complete Guide to Web Security Threats and Solutions
Aviel D. Rubin, Daniel Geer, and Marcus J. Ranum

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SPEAKERS' CORNER

AUDIO DESIGN

How can two loudspeakers with almost identical frequency response curves sound so different? Its all to do with time response, explains John Watkinson.

One important way of making progress in audio system design is to test prototypes according to some objective measurement. The objective measurement criteria must be psychoacoustically based, otherwise the test will be either too stringent or not stringent enough.

Objective test results need to be carefully interpreted because they generally test only one aspect of performance. This is why good designers always combine objective tests with subjective tests. Subjective tests check everything unreliably, whereas objective tests check one thing precisely. Just because a loudspeaker – or any other device for that matter – passes a particular objective test with flying colours does not mean that it won't fail a subjective test spectacularly.

The subjective test failure indicates that a new objective test needs to be designed, to quantify the subjective problem. Unfortunately most loudspeaker designers appear to prefer a more exciting approach. While these are strong words, the supporting evidence is all around. I find it astonishing that the vast majority of today's loudspeakers sound so different from one another – and from real sounds. This can only be because the necessary test criteria are virtually unknown.

One of the ways that one musical instrument is distinguished from another is via the relative amplitude of the harmonics which determines timbre. An accurate – i.e. uniform – frequency response and low harmonic distortion are important in an audio system. Without them, the timbre would be changed. It is well accepted that a wide and flat frequency response is an essential characteristic of a loudspeaker. Naturally, such a response is worth striving for – but not in isolation.

The problem with so many loudspeakers is that in the struggle to achieve a wide and flat frequency response, techniques are used that do active harm to other aspects of the loudspeaker's performance. As these aspects are not tested, the designer is satisfied, but the listener hoping for realism is disappointed.

Identical response, different sound

How can two loudspeakers with an identical frequency response sound so different? Well, there may be many reasons, but often there will be a difference in time response and if this is impaired the realism is lost. Poor time response causes linear distortion of the waveform and is to a transient what harmonic distortion is to a sine wave.

Time response is linked through the Fourier transform to frequency response – but only if the phase response is known. The frequency response can be fanatically accurate, but it tells nothing about the time or phase response or what the speaker will sound like.

In the same way that a system which is free of harmonic distortion is said to be linear, one which is free of linear distortion is said to be phase linear. If a loudspeaker is to be transparent – i.e. add nothing of its own character to the sound – it must be phase linear. If it isn't, each transient will carry a description of the loudspeaker instead of a description of the sound source. This destroys realism and impairs localisation in stereo.

There is nothing new in this requirement. Peter Walker, founder of Quad, has been saying for a number of decades that loudspeakers should be aperiodic.

Fig. 1. Pressure-step testing. Input waveform a) should result in pressure waveform b). Most loudspeakers cannot achieve such a response.

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W&G	MU3 Test Point Selector	£POA
W&G	PJM-1 PCM Jitter Meter	£250.00

SIGNAL ANALYZERS

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B&K 2033	Signal Analyzer	£2,000.00
HP182T+HP8557A	Spectrum Analyzer 01-350MHz	£800.00
HP182T+HP8558B	Spectrum Analyzer 100KHz-1.5GHz	£1,350.00
HP334A	Distortion Analyzer	£250.00
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HP3582A	0.02Hz-25.999KHz Spectrum Analyzer	£1,500.00
HP8903B	Audio Analyzer	£3,200.00
Marconi 2305	Modulation Meter	£2,000.00
Marconi 2382+2380	400MHz Spectrum Analyzer	£3,500.00
Marconi 2601	True RMS Voltmeter	£500.00
Marconi 2955	Test Set+2960 TACS Unit	£2,400.00
Marconi TF2370	110MHz Spectrum Analyzer	£600.00
R&S	SMOU Z1	£400.00
R&S	Ure RMS Voltmeter	£800.00
R&S CMS52	Comms. Service Monitor	£6,500.00
Racal 9302	RF Millivolt Meter	£350.00
Spectral DY. SD37	FFT Analyzer	£1,800.00
Tek 7L12	100KHz-1.8GHz	£800.00
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Farnell	E350 Stabilised Voltage Supply	£120.00
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Ferrograph	RTS2 Recorder Test Set	£200.00
Fluke	8520A Digital Multi Meter	£250.00
Fluke	8860A Digital Multi Meter	£150.00
Giga	GR1101A 12-18 GHz Microwave Signal Generator	£150.00
Giga	GU1328A 2-8GHz Microwave Signal Generator	£150.00
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HP	1741A 100MHz Oscilloscope	£300.00
HP	1742A 100MHz Oscilloscope	£275.00
HP	181A Main Frame c/w 1840A+1825A	£125.00
HP	3400A RMS Volt Meter	£120.00
HP	3570A Network Analyser 50Hz-13MHz	£150.00
HP	4333A Distribution Analyser	£300.00
HP	435A Power Meter	£175.00
HP	435B Power Meter	£250.00
HP	489A Microwave Amplifier 1-2GHz	£125.00
HP	5315A Universal Counter 100MHz	£200.00
HP	5328A Universal Counter 100MHz	£120.00
HP	5363A Time Interval Probes	£150.00
HP	8412B Phase Magnitude Display	£175.00
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HP	8552B IF Section	£250.00
HP	8553B RF Section 0-100MHz	£200.00
HP	8553L RF Spectrum Analyser 0-110MHz	£200.00
HP	8556A LF Spectrum Analyser	£200.00
HP	88001A RF Section 0.1-110MHz	£300.00
HP	8750A Storage Normalizer	£200.00
HP	491C Microwave Amplifier 2-4GHz	£125.00
HP	8413A Phase Gain Indicator Unit	£175.00
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Keithley	192 Programmable Digital Multimeter	£200.00
Marconi	6056B Signal Source 2-4GHz	£150.00
Marconi	TF2337A Auto Distortion Meter	£150.00
Polarad	1105E/FT 0.8-2.4GHz Signal Generator	£250.00
Racal	9104 RF Power Meter	£200.00
Racal	9300 RMS Volt Meter	£175.00
Racal	9301A RF Millivoltmeter True RMS	£180.00
Racal	9921 UHF Frequency Counter	£275.00
Racal	MA1720 Drive Unit	£300.00
Schlumberger	7055 Micro Volt Meter	£150.00
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Tinsley	5761 Resistance Bridge	£200.00
W&J	3737A/10 HF Receiver	£120.00
W&J	DMS107 Demodulator	£120.00
W&J	565 Receiver	£200.00
W&J	DM/112 Demodulator	£250.00
W&J	MTF/106A Microwave Tuning	£150.00
W&J	WJ9518AE/9 FDM Demodulator	£250.00
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Adret	6303B Signal Generator	£75.00
Advance	TC17 Timer/Counter	£15.00
Aiken	SH/2400P 20-100MHz	£75.00
AMF	7736 Digital Counter	£15.00
Aplab	LVE30/2 Regulated Power Supply 0-30V 0-2A	£25.00
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Datron	1030A RMS Volt Meter	£25.00
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Farnell	LFM2 Sine/Square Oscillator	£65.00
Farnell	SGIB Signal Generator Interface	£30.00
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Fluke	7261A Universal Counter/Timer	£75.00
Fluke	8000A Digital Multimeter	£50.00
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Gen. Radio	1232/A Tuned Amplifier & Null Detector	£25.00
Gould	J3A+B Signal Generator	£50.00
Gould	TC314 Timer/Counter	£20.00
HP	10254A Serial-Parallel Convertor	£25.00
HP	11859B Amplifier Switch + HP8709B Synchroniser	£75.00
HP	1600A Logic State Analyser	£40.00
HP	203A Variable Phase Generator	£25.00
HP	3465B Digital Multimeter	£40.00
HP	3710A IF/BB Transmitter	£75.00
HP	3720B IF/BB Receiver	£75.00
HP	431C Power Meter	£50.00
HP	434A Calorimetric Power Meter	£75.00
HP	461A Amplifier	£50.00
HP	5004A Signal Analyser	£30.00
HP	5150A Thermal Printer	£60.00
HP	52454 Plug In Unit & HP5245L Counter	£50.00
HP	5245L + 5255A Counter/Converter	£50.00
HP	5246L Electronic Counter	£30.00
HP	5302A 50MHz Universal Counter	£40.00
HP	5304A Timer/Counter	£40.00
HP	59301A ASCII-Parallel Converter	£30.00
HP	59313A A/D Converter	£30.00
HP	59501A HP/IB Isolated D/A/Power Supply	£40.00
HP	6111A DC Power Supply 0-20V 0-1A	£75.00
HP	6130C Digitally Controlled Voltage Source	£60.00
HP	8403A Modulator	£60.00
HP	8406A Frequency Combined Generator	£75.00
HP	8418A/H01 Auxiliary Display Holder	£90.00
HP	8717B Transistor Bias Supply	£50.00
Iwatsu	SS/5416A DC-40MHz Synchroscope	£50.00
Keithley	177 Millivolt Digital Multimeter	£75.00
Kemo	DP1 1.0Hz--100KHz Phase Meter	£70.00
Kepco	E30/2B 30V 2A Power Supply	£25.00
KSM	T25 Pulse Generator	£30.00
Levell	TG66A Transistor Decade Oscillator	£50.00
Levell	TM3B AC Microvolt Meter	£50.00
Lyons	PS22 Pulse Generator	£25.00
Lyons	PG2B Pulse Generator	£30.00
Lyons	PG71N Pulse Generator	£30.00
Lyons	PG73N Pulse Generator	£25.00
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Malden	1000F Mk II 10Hz-32MHz P.O. Frequency Counter	£10.00
Marconi	2304A 80MHz Digital Frequency Meter	£30.00
Marconi	2437 100MHz Universal Timer Counter	£45.00
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Marconi	6550B + /1 Programmable tft Power Meter	£20.00
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Marconi	TF2102M AF Oscillator 3Hz-30KHz	£30.00
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Marconi	TF2169 Pulse Modulator	£20.00
Marconi	TF2173 Digital Synchronizer	£60.00
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Marconi	TF2424 Frequency Counter	£60.00
Marconi	TF2430 80MHz Digital Frequency Meter	£50.00
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Marconi	TF2600B Video Voltmeter	£30.00
Marconi	TF2604 Electronic Voltmeter	£25.00
Marconi	TF2700 Universal Bridge	£75.00
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Montex	MS/500 Oscilloscope	£20.00
Philips	6302 RCL Bridge	£75.00
Philips	7832 SWR Meter	£50.00
Philips	7841 Power Meter	£50.00
Philips	PE1510 Power Supply 0-35V 1A	£20.00
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Philips	PM2423 Digital Multimeter	£15.00
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Philips	PM5108 Function Generator 1Hz-1MHz	£95.00
Philips	PM5142 0-01Hz LF Synthesizer	£60.00
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R&S	Scud Radio Code Test Set	£50.00
Racal	1200 Series Universal Switch Controller	£25.00
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Racal	9059 Frequency Period Meter	£50.00
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Racal	9902A 50MHz Universal Counter	£60.00
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Racal	9906 Universal Counter Timer	£50.00
Racal	9910 Frequency Counter	£50.00
Racal	9911 Frequency Counter	£50.00
Racal	9912 Frequency Counter	£50.00
Racal	9913 Frequency Counter	£50.00
Racal	9914 UHF Frequency Counter	£60.00
Racal	9914A 200MHz Frequency Counter	£50.00
Racal	9915 UHF Frequency Meter	£90.00
Racal	9916 UHF Frequency Meter	£70.00
Racal	9918 UHF Frequency Meter	£75.00
Racal	9960A 200MHz Universal Counter	£65.00
Racal	GPIB Interface 110	£40.00
Racal	9905 Universal Counter Timer	£50.00
Racal	9932 Instrument Interface	£20.00
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SE Labs	SM205B Automatic Counter Timer	£20.00
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Tek	453 Oscilloscope	£60.00
Tek	Time Mark Generator	£50.00
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Telequipment	D83 Oscilloscope System	£40.00
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Thander	TA2080 20MHz Logic Analyser	£75.00
Thurlby	OM358 Scope Multiplexer	£45.00
Trio	CS/1566A 20MHz Oscilloscope	£60.00
W&G	SMP/31 Level Meter 200Hz-620KHz	£80.00
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W&J	TFC103 Telephone Frequency	£30.00
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W&J	UH15 Tuning Head	£40.00
W&J	VH103 VHF Tuning Head	£75.00
W&J	521A Receiver	£75.00
W&J	SM/6108 Signal Monitor	£30.00
W&J	TDM/110 Basic Group Demodulator	£10.00
Wavetek	147 HF Sweep Generator 0.0005Hz-10MHz	£50.00
Wier	460 Power Supply 0.3A 0.60V	£50.00
Wiltron	501 Logic Level Meter	£60.00

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And there is an increasing body of psychoacoustic research that supports the view.

Time response testing is remarkably easy, as Fig. 1 shows. Voltage step a) is fed into the speaker. An ideal speaker will reproduce the leading edge of the step, but the raised pressure caused by the step cannot be maintained in the surrounding air. The waveform due to the pressure equalisation can be calculated, and is shown in b). Any speaker that cannot reproduce something like b) is suffering from linear distortion.

Damaging techniques

The most damaging techniques employed for bandwidth extension in loudspeakers are often found at the low frequency end. The transmission line speaker uses a long duct leading from the rear of the driver to a port. At the frequency where port augmentation is required, the duct causes a delay of half a cycle. Figure 2 shows that for a sinusoidal input, a delay of half a cycle is the equivalent of an inversion, and the port output is now in-phase with the radiation from the front of the driver.

Unfortunately the assumption that a time delay is the same thing as an inversion is fallacious. Figure 2b) shows that the step response is a double pulse, which describes the design of the speaker to the listener. Transmission lines speakers are not phase linear and cause linear distortion.

By definition, any loudspeaker employing in-band tuning or resonance cannot be phase linear. This includes reflex tuning, auxiliary bass radiators and band-pass enclosures. All of these techniques cause linear distortion and cannot reproduce a step. A generalisation, which is none the less true, is that any loudspeaker enclosure having a port will suffer from linear distortion.

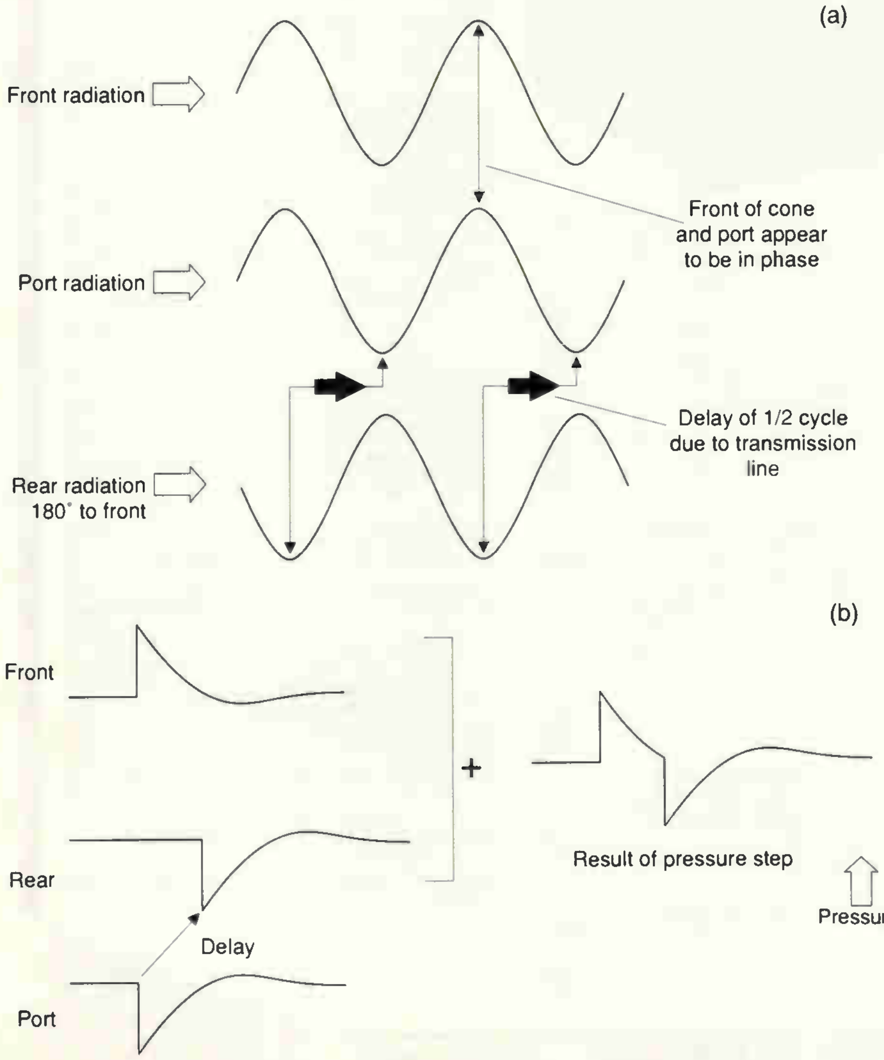
This saves a lot of unnecessary testing because you can tell it's not going to work just by looking. Figure 2c) shows the result of a step test on a highly regarded reflex loudspeaker. It's basically out of control.

The only low-frequency transducer of practical size which can be made phase linear is the sealed box. The driver has a fundamental resonant frequency. But with a suitable damping factor there is no resonant behaviour and the speaker behaves like a high-pass filter.

With a suitably accurate signal processor, such as those described from time to time in this magazine, the response of the loudspeaker can be extended below the natural resonance, making large enclosures unnecessary.

Active loudspeaker techniques allow wide, flat frequency response without impaired time response. The audible result is so superior that alternatives are really not worth considering where accuracy is paramount.

Resonant techniques had a place in earlier times, when electronic signal processing was expensive and unreliable and amplifier power was limited. We should recognise and respect the achievements of those pioneers who made these technologies work within the restrictions of the time, but we must also recognise that those restrictions no longer apply. ■



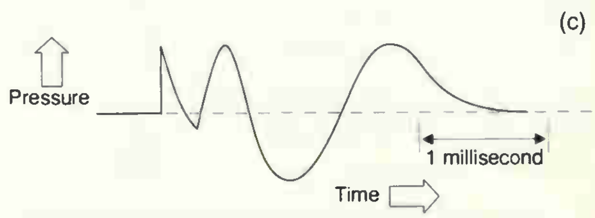


Fig. 2. Transmission loudspeaker assumes that a delay is the same as an inversion, a). This is only true for a sine wave. The result is linear distortion of transients. Waveforms in b) show that the transmission line speaker fails the pressure step test. Waveforms c) represent a typical pressure step test on a reflex loudspeaker. Note repeated transients due to resonant low-frequency design and crossover.

NEW PRODUCTS CLASSIFIED

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ACTIVE

Linear integrated circuits

3V rf receivers. Two GaAs rf receiver ics by TriQuint, the *TQ9225* and *TQ9228*, provide makers of CDMA PCS telephones with a means of meeting the IS-95/98 standards, both containing a multi-stage down-converter chain and having a local-oscillator buffer to drive the transmitter circuitry. Conversion gain is 28dB for the 9225 and 27dB in the 9228, noise figure 2.3dB (2.6dB) and input third-order intercept point at –9dBm. Local-oscillator frequency range in the 9225 is 1710-1790MHz (2015-2075) and if range 210-212MHz (84-86MHz). Pronto Electronic Systems Ltd. Tel., 0181 5545700; fax, 0181 5543222.
Enquiry no 501

Charge-sensitive preamp. Using the newest j-fet techniques to obtain an open-loop gain of 85dB at 200MHz and low noise, the *IFPA-300* is intended to buffer high-impedance, low-level signals into 50W. The device consists of eight-channel, epitaxial, diffused-gate j-fets to form an inverting transimpedance amplifier to handle detector capacitance input of 100-1000pF. Input referred noise figures are 3nV/√Hz and 0.6nV/√Hz at 10Hz and 10kHz respectively. *IFPA-300* is packaged in a 4.2mm high, 9mm diameter TO-99 and the 301, which is electrically the same, comes in an SOIC-8 package. MCP Electronics Ltd. Tel., 0181 5182000; fax, 0181 5183222.
Enquiry no 502

Discrete active devices

Rf power. Rf power transistor *PTB20245* by Ericsson puts out 35W pep in the 2.1-2.2GHz frequency band. It is a Class AB type, meant for use as a final or driver stage in wideband CDMA or TDMA application. Minimum gain figure is 7.5dB (typical 8.2dB) at 10W output power and gain linearity 0.3dB. At 1dB compression, power output is 35W at a collector efficiency of 40%. *PTB20245* has an intermodulation distortion figure of –38dB at 28W pep and tolerates a load mismatch of 5:1. Ericsson Components AB. Tel., 01793 488300; fax, 01793 488301.
Enquiry no 505

Memory chips

3.3V srams. 32-bit wide, 3.3V asynchronous srams from IDT, the *IDTMPV4060/4145* have standard pin arrangement and are pin-compatible with each other and with future devices with larger densities. The *4060* has four 128K by 8 srams in plastic small outline J-lead packages (SOJ), while the *4145* has eight 256K by 4 srams in SOJ, both being 72-pin simms taking up 1sq.in of board space. Integrated Device Technology. Tel., 01372 363339; fax, 01372 378851.
Enquiry no 503

Sram+flash. With an eye on the GSM market, Toshiba has introduced the *MCP*, a 1sq/cm ball-grid array device that combines Nor-flash memory and sram, so taking up around 70% less board space than separate components and reducing wiring. Current devices combine 1Mbyte or 2Mbyte sram with 4Mbyte or 8Mbyte Nor-flash memory, the sram being used for functional control and the flash for program storage. Toshiba Electronics UK Ltd. Tel., 01276 694600; fax, 01276 694800.
Enquiry no 504

Microprocessors and controllers

Eight-bit microcontroller. NEC's *78K/0S* series of low-cost, general-purpose, 8-bit microcontrollers give improved performance over and many of the features of the established 78 series. These devices are not risc-based and provide good event capture, i/o manipulation and have extensive interrupt resources and timing. Instruction process time is 0.4ms with a 5MHz clock and supply needed is 1.8-5.5V. In some models in the range, prom and flash memory can be specified. NEC Electronics (UK) Ltd. Tel., 01908 691133; fax, 01908 670290.
Enquiry no 506

Inexpensive microcontroller. Canadian company Intec Inoventures has introduced the *Freedom 16* microcontroller, which is said to provide more power, versatility, memory and i/o than some others while costing less. The processor is a *68HC16* 16-bit type and there are 30 i/o ports, a-to-d ports, RS485 and

debug ports, expansion bus and lcd and keypad port; memory is up to 256K of ram and 256K of flash eprom. Programming is in C and an assembler, editor, communication and debug software is available. The board is available in two versions: the *16V 1.2* at £299 or £499 with C compiler and library; and the *16V 2* at £199 or £359 with the software. Electroustic Ltd. Tel., 01264 333664; fax, 01264 333665.
Enquiry no 507

Mcu with USB facility. Mitsubishi's *Slim 740* microprocessor pack includes development tools and an evaluation board. Two types are available with double-speed mode to give the option of a full Universal Serial Bus or a low-cost, non-USB type. Both use the company's 8-bit 740 controller with integrating serial i/o, 8-bit timers and a 10-bit a-to-d converter. There is 8K of rom and 256byte of ram; minimum instruction time is 0.34ms or 0.5ms in the low-cost devices. All types have 69 basic machine language instructions and 29 programmable i/o ports. IEC Micromark Electronics Ltd. Tel., 01628 676176; fax, 01628 783799.
Enquiry no 509

Z8 controller with eeprom. *Z86144* from Zilog is an 8-bit type having a 16byte user-addressable eeprom, 1Kbyte of rom, 60byte ram , two analogue comparators, a pair of counter-timers and a watchdog timer. The eeprom may be used to hold encryption seeds, calibrations, ID numbers, data codes and the like that can be erased and written in circuit. Zilog can provide a remote, keyless-entry, rolling-code reference design using the eeprom to hold values for the encryption of signal between transmitter and receiver and includes resynchronisation. Zilog UK. Tel., 01628 639200; fax, 01628 781227.
Enquiry no 553

Linear integrated circuits

Audio power. Philips claims its *TDA8551* audio power amplifier to be the first of this type to possess a digital 80dB volume control operated by push-button. From a 5V supply, the device puts out 1.4W into an 8Ω bridge-tied load; immunity to rf interference from nearby sources is low. Its single volume-control pin may also be driven by a digital i/o port and another pin mutes the amplifier or puts it into standby. Thd is less than 0.15% at 0.5W, ripple rejection over 50dB and there is short-circuit, electrostatic discharge and thermal overload protection. Philips Semiconductors (Eindhoven). Tel., 00 31 40 2722091; fax, 00 31 40 2724825.
Enquiry no 508

Optical devices

Better ccds. Sony has a new generation of charge-coupled-device image sensors that use Super HAD (Hole Accumulating Diode) techniques to provide increased resolution and brightness and smaller

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overall size. This is an improvement over the plain old HAD, in that Sony has improved the shape of the polymer microlens structure over each individual pixel in the new technique to give, in the *ICX206AK* 0.25in ccd with 250,000 pixels for example, a 30% increase in focussing efficiency and consequently a 4dB improvement in sensitivity over earlier units, with no s:n degradation. Saturation signal level at 900mV is 25% higher than before. Sony Semiconductor Europe. Tel., 01256 478771; fax, 01256 818194.
Enquiry No 510

Laser diodes. A new series of Sanyo aluminium gallium arsenide laser diodes are for free-air communications, cd-roms, bar-code scanners and many other uses. Output powers range from 20mW in the *DL-4034* to 150mW in the *DL-8032*, operating at wavelengths from 780nm to 830nm at –10°C to 50°C. There is also a 30mW device, the *DLLS52*, working at 655nm. Semicom UK Ltd. Tel., 01279 422224; fax, 01279 433339.
Enquiry No 511

Three-level led bars. The *TRA T1 Series* of three-level, right-angle leds by Wilbrecht includes versions with one three-level housing or up to 35 such housings in a bar assembly 152mm long. They are available with a variety of led types, including high-efficiency red, green, yellow and orange types and ultra-bright types. Bi-colour and low-current versions are also available, as are light shields to prevent light spillage. TW Electronics Ltd. Tel., 01635 278585; fax, 01635 278122.
Enquiry No 512

Oscillators

Clock oscillator. *Model H-3298* clock oscillator by MF Electronics in New York offers low jitter at up to 170MHz by operation at a high overtone of the natural frequency, as opposed to frequency multiplication and meets the needs of ATM, SONET and SDH circuits, while contained in a package 0.52in square by 0.2in. Jitter is less than 10ps rms, the oscillator takes 10-55mA, depending on frequency and 5mA when a pin is grounded to stop oscillation. Stability is 20ppm over the 0-70°C range and drift ±3ppm or less in the first year. Output is compatible with cmos or ttl logic. MF Electronics Inc. Tel., 001 914 576 6570; fax, 001 914 576 6204.
Enquiry No 513

Power semiconductors

Input rectifiers. International Rectifier has three new device types for input rectification: *SAFEIR* diodes and thyristors protect against surges and spikes; *QUIETIR* soft-recovery diodes that give reduced emitted and conducted noise with low forward drop; and *SMALLIR* moat-processed diodes and glass-passivated scrs rated at up to 25A, 1.6kV, in D-Pak and D^2Pak s-m packages. International Rectifier. Tel., 01883 732020; fax, 01883 733410.
Enquiry No 514

Power mosfets. Temic's newest Little Foot SO-8 power mosfets have 80V and 100V drain/source ratings. There are four new ones: the 80V, single *Si4480DY* handling up to 6A and with an on resistance of 35mW; the *Si4980DY* dual version; the 100V *Si4482DY* and its dual version *Si4982*. IEC Micromark Electronics Ltd. Tel., 01628 676176; fax, 01628 783799.
Enquiry No 515

PASSIVE

Cameras

Three-chip camera. Sony's *MCC-1* 3.3V three-chip set, with the addition of a colour ccd, forms a light, cheap NTSC camera for makers of multimedia equipment. The chipset performs all driving and signal processing and takes Sony ccds from 1/5in types giving 180k pixels to 1/4in units at 250k pixel resolution. The three chips consist of the *CXD3123R* signal processor with a/d and d/a converters to provide both analogue and digital output; the *CXA2006Q* (5V) or *CXA2096N* (3.3V) head amplifier; and the *CXD1267AN* ccd driver for NTSC systems. Sony says it has no plans for a PAL set. Sony Semiconductor Europe. Tel., 01256 478771; fax, 01256 818194.
Enquiry No 516

Passive components

Ethernet transformers. Fast Ethernet transformer modules in surface-mounted packages are designed for use in 10/100 BaseT networks. The series consists of single and multi port devices, the singles (*H1012,1019,1042*) each containing the transformers and chokes for an integrated transmit and receive channel for single-port application. Also available are the two-channel *H1026* and the four-channel *H1044*. These modules work with current transceivers needing a 1:1 turns ratio and there are pinout options to suit customers' needs. All work with unshielded, data-grade, twisted pair cable and have an insertion loss of 0.5dB from 100kHz to 100MHz and cmrr of 30dB from 100kHz to 60MHz. Pulse. Tel., 01483 428877; fax, 01483416011.
Enquiry No 517

Miniature solenoids. Densitron has a range of miniature latching solenoids to provide a holding force, in the smallest model, of 100g and in the largest, 400g. They need 1W pull-in power and 10W to release. The smallest versions are of a size to fit into a camera to operate the shutter. Densitron Europe Ltd., 01959 700100; fax, 01959 700300.
Enquiry No 519

S-m wire-wounds. *WSM Series* wire-wound, surface-mounted resistors by Welwyn can be picked-and-placed by automated equipment at the same time as standard s-m components. They come in 1W and 3W ratings in E24 values from 0.05Ω-900Ω for the 1W types and 0.01Ω-10kΩ for the 3W version and in tolerances of 1, 2, 5 and 10%. Welwyn Components Ltd. Tel., 01670 822181; fax, 01670 827434.
Enquiry No 520

Thermistors. *Cera-Mite* ceramic positive-coefficient thermistors for overcurrent protection come in a range of hold currents from 5mA to 1.5A and in a size range of 4-22mm. The thermistors remain latched in the tripped state, resetting after the removal of voltage, with no hysteresis, so that the initial resistance value is retained. Acal Electronics Ltd. Tel., 01344 727272; fax, 01344 424262.
Enquiry No 521

Connectors and cabling

Optical connectors. Deutsch has introduced the smallest member of its *MC5* range of multi-way optical fibre connectors. The four-way shell No 13 uses 1.25mm diameter zirconia ceramic ferrules and lightweight shells to MIL-C-38999. Spring-loaded optical contacts may be individually inserted or removed for easy assembly and the removable alignment insert allows simple cleaning. Insertion loss is 0.25dB and repeatability better than 0.1dB with 50/125mm fibre. Deutsch Ltd. Tel., 01342 410033; fax, 01342 410005.
Enquiry No 518

Chip electrolytics. Sanyo has announced a new range of 125C chip capacitors, the *CV-PX* family, which contains components in values from 33μF to 680μF and voltage ratings of 6.3-50V. These surface-mounted capacitors are solvent-proof. Inelco Ltd. Tel., 0118 9810799; fax, 0118 9810844.
Enquiry No 522

Communications equipment

DECT line interface. Ericsson's *PBL38571* is a universal speech circuit ic intended for use as a line interface in fully balanced DECT equipment or as a speech interface in electronic telephones. It is designed to be used with a low-impedance microphone and balance to ground is conferred by the differential microphone input; a balanced output is provided for DECT use. For flexibility, line balance, frequency response, side tone level and gain are set by external components, a total of five capacitors and ten resistors. A mute function is provided for use with a dtmf generator during dialling and the device starts up quickly, as there is a temperature-regulated voltage reference to settle the line voltage to its final value rapidly and with little overshoot. Ericsson Components AB. Tel., 01793 488300; fax, 01793 488301.
Enquiry No 523

Dsp for comms. A digital signal processor for use with digital cordless telephones, answering machines and Internet appliances is announced by Lucent. It operates at 100Mips, which is three times as fast as others costing about the same – $4.95 in large numbers. Devices using the *DSP1609* are able to do several things at once, such as caller identification, speakerphone, digital cordless telephone, digital answering machine and modem. The device runs from both 5V and 3.3V and needs only 1mA/Mips; it has 2K of

Please quote *"Electronics World"* when seeking further information

ram and 24Kword of rom. Lucent Technologies. Tel., 0118 932 4299; fax, 0118 932 8148.
Enquiry No 524

Connectors and cabling

Russian plug. Interpower's new mains plug is more or less the same as the European CEE 7/4 unit, but with 4mm pins. The *Schuko* rewirable plug has a 250V/16A ac rating and meets all the major European standards. Interpower Components Ltd. Tel., 01243 842323; fax, 01243 842066.
Enquiry No 525

Crystals

Thin crystals. Tele Quartz has a new range of slimline crystals in the HC-52/U SL leaded or surface-mounted package, which is 2.3mm in height. The crystals are AT cut and come in frequency ranges from 3.5MHz to 360MHz, with temperature stabilities of ±1.5 to ±50ppm, temperature ranges of 0-50°C and –55°C to 105°C. Ageing is less than 3ppm/year at 23°C, drive level 0.1mW and shunt *C* under 7pF. Webster Electronics Ltd. Tel., 01460 57166; fax, 01460 57777.
Enquiry No 526

Displays

Multi-image monitor. *SuperView VGAplus* by RGB Spectrum is a video windowing system for the display of the images from several computers simultaneously on one monitor, designed for use in operations centres, control rooms, presentation and training. It is compatible with pcs and workstations and takes up to six 1280 by 1024 inputs for display either as windows or in the background, Pal or NTSC and S-Video display being an option. Each image may be positioned independently and sized from icon size to full screen and can be panned and zoomed. The equipment is an external standalone computer peripheral and may be controlled from the front panel or by RS-232 port; software for control under Windows is available. Gothic Crellon Ltd. Tel., 01734 776161; fax, 01734 776095.
Enquiry No 528

Literature

Farnell. More than 1000 new products are described in Farnell's *Semiconductor New Product Brochure*, all of them being new to Farnell since October. The brochure is included with the newest 33-page Semiconductor Application Directory, which includes a "getting started with PIC microcontrollers" section, and they are also available in cd-rom form that holds more than 13500 data sheets. Unfortunately, the cd sent my computer to sleep, from which it refused to wake up without the application of Ctrl+Alt-Del. Farnell Components Ltd. Tel., 0113 2636311; fax, 0113 2633411.
Enquiry No 527

Hardware

S-m bobbins. A full range of surface-mounted bobbins for chokes and transformers is now available from BFI-IBEXSA. Specially designed types are available, as are samples. BFI IBEXSA Electronics Ltd. Tel., 01622 882467; fax, 01622 882469.
Enquiry No 529

Slim fans. Papst's *405F/412F/424F* range of 40mm square cooling fans are all 10mm thick and deliver a flow of eight cubic metres/h while running on 5/12/24V dc; a flow of 10m^3/h can be provided by a further range of 20mm deep versions. Motors are electronically commutated, external-rotor types, protected against reverse polarity and with impedance protection against blocking and overload. Special bearings confer low noise. XP plc. Tel., 0118 9845515; fax, 0118 9843423.
Enquiry No 530

Another slim fan. This one is not very big in any direction, being 45mm square, 10mm thick and intended to cool the processor in notebook computers. Sanyo Denki's miniature version of its *San Ace MC* range of coolers is only half the size and weight of its mightier cousins and has a thermal resistance of 3.2°C/W with an acoustic noise level of 28dBA. The unit is protected against reverse connection and locking of the rotor. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.
Enquiry No 531

Test and measurement

Four-channel dso. Nicolet's new *Integra 60* four-channel digital oscilloscope and transient recorder has a colour display, a sampling rate of 100Msample/s and real-time data analysis. The instrument uses 8-bit digitisers to give the sampling rate and to provide 5Gsample/s equivalent real-time sampling at a bandwidth of 200MHz; optional expansion facilities provide for a memory capability of 200ksample/channel. Acquisition modes include multi-shot for rapid storage and recall, autocycle for auto capture and storage to disk, averaging and persistence to simulate a long-persistence crt. Optional analysis includes FFTs, integration and differentiation, histograms, trends and filtering, all at high speed. There is an internal floppy disk drive and the options of a hard disk and thermal plotter. Nicolet Technologies Ltd. Tel., 01908 679903; fax, 01908 677331.
Enquiry No 532

3-phase wattmeter for motors. From Yokogawa comes the *WT1030M*, which is a three-phase digital wattmeter, designed to investigate the overall efficiency of electric motors, being used with a torque sensor to give data on torque and speed and to perform computation to show sync. speed, slippage, mechanical power, motor efficiency and overall efficiency. Accuracy of the power meter is 0.1% over a 300kHz bandwidth, accepting inputs of 15-1000V rms, current to 20A and shunt or transformer input for higher currents. The *WT1030M* integrates power readings to give long-term measurement of varying inputs and there is an analogue output for display of waveforms on an oscilloscope. Martron Instruments Ltd. Tel., 01494 459200; fax, 01494 535002.
Enquiry No 533

Literature

Enclosures. West Hyde's catalogue of enclosures, showing more than 30 new products, is just published and is now available on cd-rom for Windows 3.x/95, AppleMac and Unix. The company also has its own website at http://www.westhyde.co.uk (e-mail boxes@westhyde.co.uk). West Hyde Enclosures. Tel., 01453 836789; fax, 01453 836444.
Enquiry No 535

Materials

Hot adhesive. Newly introduced by Cotronics is the *E098* adhesive, which will withstand temperatures up to 1650°C continuously. It is based on ceramic epoxy, cures at room temperature and was developed for cementing, insulating and embedding coils, filament windings etc. and may be used on metals, glass, ceramics and plastics. Thermal conductivity is good, electrical resistance is high and the material is resistant to solvents and other chemicals. Symonds Adhesives Ltd. Tel., 01707 372564; fax, 01707 372573.
Enquiry No 536

Power supplies

Dc-to-dc converters. *SIP20* non-isolated, 20W converters are single in-line devices to convert 5V to 3.3V dc and lower. Power density is 60W/cubic inch and efficiency up to 90%. Facilities include adjustable output voltage, fixed 500kHz frequency operation, undervoltage lockout and no minimum load. Computer Products, Power Conversion Ltd. Tel., 00353 24 25272; fax, 00353 24 93510.
Enquiry No 534

Wrapping. *Barnier* adhesive tapes for the wrapping of transformers, relays and other wound components are produced by Scapa Tapes. Core and interlayer insulation is provided by polyester film types rated at 130°C, which are resistant to solvents, freon and oils. Double-sided film and glass cloth or fabric are used for taping the ends of windings. For outer layer protection, adhesive acetate, which may be printed, is used. Scapa Tapes. Tel., 0161 3364433; fax, 0161 335 0104.
Enquiry No 537

Power supplies

Bench supply. Power supplies in TTi's *TSX* range are said to give a high performance at a lower cost than others in the same class. There are models to provide outputs of 35V/10A and 18V/20A in standard and programmable versions, the standard type having ordinary panel controls and the programmable version keyboard control, RS-232 and IEEE-488 interfaces. The new series uses a new regulator design in which switched-mode pre-regulation is combined with linear post-regulation, the pre-regulator presenting a low input-to-output capacitance and therefore reducing switching noise. All units in the series work in both constant-*V* and constant-*I* modes with automatic crossover and mode indication. Digital readout of V and I is provided. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.
Enquiry No 538

Pwm controller. Cherry's *CS-5106* pulse-width modulation controller has two separate pwm drivers and many monitoring, protection and control facilities. There are a dual nfet driver as the main controller, driving two external nfets and a single nfet driver for an auxiliary supply, operating independently of each other after start-up, and switching in around 80ns for high efficiency. Clock frequency is 512kHz and input/output sync. pins allow master/slave working with other supplies. Clere Electronics Ltd. Tel., 01635 298574; fax, 01635 297717.
Enquiry No 539

18A step-down voltage converter. The *Voltec* step-down dc-to-dc converter takes in 24V dc found in cars, boats and industrial work and puts out regulated 12V at 18A continuously, or 20A for five minutes when surges are to be expected. Efficiency is 90% and an idle circuit reduces standby current to 27mA; there is also provision to shut the unit down when the input falls to 20V dc. Full protection against all the usual disasters is provided and there is a led display to indicate health or sickness. Merlin Equipment. Tel., 01491 824333; fax, 01491 824466.
Enquiry No 540

Protection devices

Watchdog timer/monitor. BVM's *BVME290* is a watchdog timer/system monitor to protect VMEbus systems. The unit monitors the software by demanding a wake-up command after a selected interval of 0.1-60s, and also watches eight external signals from sensors such as temperature switches and interlocks in the system, and from supply rails; SYSFAIL and ACFAIL lines are also monitored. If an alarm triggers the monitor, it either enables the SYSRESET driver, generates an IRQ7 interrupt or sounds an alarm. A front-panel switch allows the device to be disabled for maintenance. BVM Ltd. Tel., 01489 780144; fax, 01489 783589.
Enquiry No 541

Thermal watchdog. An improved thermal watchdog for cpu protection is announced by Dallas. The *DS75-2* two-wire device is a replacement for the National *LM75* but with advantages. Instead of 9-bit resolution, the *DS-75* may select 10, 11 or 12 bits instead of the default 9 by changing the sram configuration register; at 12-bit resolution, the smallest temperature change registered by the watchdog is 0.0625°C, resolution and accuracy being preserved over 2.7V-5.5V supply range. Dallas Semiconductor Corporation. Tel., 0121 782 2959; fax, 0121 782 2156.
Enquiry No 542

Thermal cutout panel switches. Siemens *W28* thermal cutout switch is a single-pole, series trip, available for a range of cutout currents from 0.25A to 20A at up to 32V dc or 250V ac, an optional, switchable version having the reset function inside the switch. Time to reset is 180s maximum for models handling up to 2A and 10-60s for 3-20A versions. The switches snap in to a 15.9mm panel cutout and have quick-connect terminals. Easby Electronics Ltd. Tel., 01748 850555; fax, 01748 850556.
Enquiry No 543

Video

Video scan converter. The *RGB*/Videolink 6U VME PAL/NTSC video scan converter transforms computer output to broadcast standard video in real time. It automatically synchronises to all computer signals up to 1600 by 1200 pixels and provides composite video, S-video (S-VHS/Hi-8) and component analogue video for Betacam/M11 outputs for recording. There is dsp circuitry to eliminate flicker so that even thin horizontal lines are stable, and zoom is present. Gothic Crellon Ltd. Tel., 01734 776161; fax, 01734 776095.
Enquiry No 544

Transducers and sensors

Magnetoresistive sensor ics. Magnetoresistive integrated circuits by Honeywell are designed for use in navigation systems, proximity detection, virtual reality and medical monitoring. *HMC1020/2* are single and dual axis versions, both being wideband devices to sense magnetic fields up to ±6gauss, resolving to 85mgauss with low noise. Honeywell says that these devices have the edge over fluxgates by their smaller size, better reliability and overall toughness. The *1021* is in an 8-pin sip and the *1022* in a narrow-body SOIC; both use a 3/10V supply. Inertial Aerosystems Ltd. Tel., 01252 782442; fax, 01252 783749.
Enquiry No 545

Flying-lead optical encoder. Grayhill's *Series 62A* rotary optical encoder is a compact unit having a flying lead for easy installation, which is also eased by the size: 12.7 by 13.72mm of board space, 13.34mm deep. The unit can be provided with a pushbutton and there is the choice of 16, 24 or 32 detent positions. Output is standard quadrature 2-bit, compatible with cmos, ttl and Hcmos logic. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.
Enquiry No 546

COMPUTER

Data acquisition

Daqs and a GPIB for USB. National has three data acquisition devices and a GPIB interface for Windows pcs using the universal serial bus. *DAQPad-4350* is a temperature and voltage measuring instrument; *DAQPad-6020E* a 100ksample/s, 12-bit E Series-based multifunction data acquisition device with 16 analogue inputs, two analogue outputs, eight digital i/o and two counter/timers; and the *DAQPad-6057* 96-line digital i/o, all the above being for USB; and then there is the IEEE 488.2 controller. The usb data acquisition devices include the *NIDAQ* driver software and the GPIB interface the NI-488.2M driver software. All these devices are compatible with National's virtual instrument software. National Instruments UK. Tel., 01635 572400; fax, 01635 524395.
Enquiry No 547

Data communications

Single-chip V.22bis. TDK's *73K224BL* modem ic combines analogue, switched-capacitor array functions, a risc dsp and hybrid drivers in one 32-lead PLCC pack. It will interface directly with any standard microprocessor to form a complete V.22bis-compatible modem to give full-duplex operation at 2400bit/s on dial-up lines. In addition to the 2400bit/s quam, 600/1200bit/s dspk and 300bit/s fsk mod/demod, it also provides sync and async converters, scramble/descramble, call progress tone detection, a dtmf tone generator and handshake pattern detectors. Pronto Electronic Systems Ltd. Tel., 0181 5545700; fax, 0181 5546222.
Enquiry No 548

Computer peripherals

17-in flat crt. *PanaSync/Pro P70* by Panasonic replaces the *Pro 5G* monitor, this one having a 16-in viewable diagonal, a 0.25mm dot pitch and a 95kHz scanning rate. Accuracy, particularly of vertical edges, is improved and power consumption reduced to 110W, or 4W standby. Resolution is 1600 by 1200 at 75kHz vertical refresh rate and full VESA plug-and-play is incorporated. Industry standards met include Energy Star, MPR II and TCO'95 and other features are Moire compensation, colour temperature adjustment and image rotation, all assisted by an on-screen menu – in five languages. Panasonic. Tel., 0500 404041. Web http://www.panasonic.co.uk.
Enquiry No 549

Identification for plug/play monitors. Rohm has the *BU9881* ID rom ic that stores the identification information to allow a pc to recognise monitor characteristics on connection to the pc. It contains a 128 by 8bit serial eeprom and takes 2mA, or 10mA in standby mode. The device offers transmit-only and bidirectional, two-wire serial I^2Cbus working, operates at up to 400kHz and runs on 2.7-5.5V. There is auto-erase and auto-complete during write operation. Rohm Electronics UK Ltd. Tel., 01908 282666; fax, 01908 282528.
Enquiry No 550

Software

Colour from greyscale hardware. *NI-IMAQ* driver software allows the acquisition of 24-bit colour images at up to 10frames/s while using 8-bit image acquisition hardware of the type normally used for greyscale image capture. This is achieved by the use of National's virtual instrument software. National Instruments UK. Tel., 01635 572400; fax, 01635 524395
Enquiry No 551.

Development and evaluation

PIC emulator. An entry-level emulator for the Microchip *PIC16C5X* and *12C5X* ranges of microcontrollers is announced by RF Solutions, who say that it offers the best ever price/performance ratio. The non-real-time *SIMICE* i/o emulator provides a full hardware interface to Microchip's MPLAB/MPSIM software simulator, with the availability of the associated debugging features that eliminate the need to write i/o lines. RF Solutions Ltd. Tel., 01273 488880; fax, 01273 480661.
Enquiry No 552

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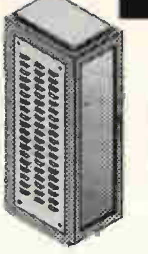

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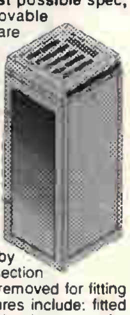

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LETTERS

Letters to "Electronics World" Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS

Direct conversion

I read the article on the direct conversion receiver in the February issue with interest. Concerning the vfo. I have the following comments.

The phase noise can be improved by 3dB by replacing the IN4148 diode connected between the gate and ground of the fet used in the vfo, with a trimmer capacitor adjustable from 2 to 10pF.

By adjusting the trimmer capacitor, the gate to source conduction can be minimised. This enables you to make use of a higher value gate to ground resistor.

The drain dc voltage can be used to indicate whether the oscillator oscillates or not. I measured the phase noise of the modified vfo and it is 3dB better than the original.

My explanation for this result is that a forward biased diode is a current controlled capacitor causing am to fm conversion caused by the always present $1/f$ noise current.

By the way, the phase noise level of the vfo is very low. The requirements are high when operating on 28MHz. I measured the phase noise with valve equipment – a vacuum diode is not a current controlled capacitor. This emphasises how important it is to isolate the load from the vfo. In Coetzee's design however, this point was very well addressed.

Wim de Ruyter
Oudkarspel
The Netherlands

Hannes comments:
Since my article was written, an article by Ulrich L Rhode h as been published in *Communications Quarterly*. In that article, Ulrich analysed the effect of the diode between the gate and ground of the fet on the phase noise of a vfo. I do not have access to the article, but if I remember correctly, Ulrich came to more or less the same conclusion as Wim.

Hot topic

I read Richard Lines' article on sensing temperatures in the March '98 issue with interest. He did not mention that some ranges of thermistors are made specifically for temperature measurement and that these may have 'matched' characteristics. If you add a carefully chosen fixed resistor R_1 in parallel with the thermistor shown in Fig. 9, you can obtain a very considerable degree of linearisation of the temperature/voltage characteristic of the thermistor T_h with a reciprocal function. $1/R_{tot}=1/T_h+1/R_1$.

There were a few errors in the description of diodes and transistors as temperature sensors. The upper temperature limit is determined by the reverse leakage current of the device. The voltage/temperature linearity falls off when the leakage current becomes a significant fraction of the forward bias current. The lower temperature limit may be determined by the case, but for a glass signal diode it is about –220°C.

Schottky diodes have a lower sensitivity, but a more linear characteristic. I have used both regularly in cryogenic applications down to –196°C without any problems. Thanks to the accuracy of current fabrication techniques, the difference between individual devices from the same manufacturer is usually quite small.

Transistors give the most linear characteristic, but usually have a bulky case and the gain falls off below about –50°C, leading to bias problems. Physical size, power dissipation, response and thermal capacity are also important parameters when choosing a sensor.

The sensitivity increases and the noise decreases as the junction bias current is reduced. The *BC182* graph should show different slopes as well as different intercepts for currents. Richard's formula for the "differential resistance" does depend on the particular junction – I is the current per unit area of the junction, not the absolute current.

Junction thermometers have an extremely high sensitivity to power dissipation ratio. What other device will give you 2.5mV/°C at 5µW? Silicon resistors with linear resistance/temperature characteristics are also available.

CJS Chapman
Upwaltham
Sussex

Richard replies:
Thank you for your letter. You raise some worthwhile points that I will attempt to answer in the order you mention them.

All your complaints relate to the sections on thermistors and diodes/transistors. It may help to explain the context in which this article was written. It forms part of a series which was originally written in three parts for publication, the whole of which was intended to be a fairly comprehensive review of the practical aspects of temperature control.

However, when I completed the first draught, it was obvious there was far too much material – running to some 45 pages. Something had to be cut down. There was a general background theme in that much of the material was designed to be compatible with an analogue controller which will be published in a later issue. You will find a reference to this in the opening paragraph on page 165 of the February issue.

Thus material that was directly relevant to this controller was given priority over that which was not. The hardware I built used thermocouples, platinum sensors and the various IC sensors described.

The interface circuits illustrated concentrate on producing a standard linear 10mV suitable for the controller input, which gives a common linear calibration on the controller. Hence they feature more prominently in this article at the expense of thermistors and diodes.

If you read the article in isolation as a general and comprehensive review then I agree I have failed in the areas you point out; the sections on thermistors and diodes are more superficial than I would usually like. Even the cut down script I submitted will run to four parts – it was intended to be published as three.

Now, to answer your remarks more specifically. You are of course perfectly correct in that there are some fairly simple circuits that will linearise thermistor response. However, the response will never be perfectly linear with the simple parallel arrangement, and the component values will always be

FM transmitter please

Can any of your readers suggest a circuit for a modulator to produce a frequency modulated stereo signal in the 88-108 MHz band from left and right audio channel inputs?

J K Carter
Maidstone
Kent

Welding and emc

With reference to Ian Johnson's letter in the February 1998 issue, flicker caused by arc-welding equipment used in private households is an increasing problem, and it is being addressed under emc control measures. The problem is that the whole business of disturbances on the supply mains is very complex – far more complex than was realised when the relevant standards EN61000-3-2 (harmonics) and EN61000-3-3 (voltage changes and flicker) were prepared.

Some remedial work is under way, but, according to the latest decisions, the standards do not come into force until 2001.

There are existing, current standards EN60555-2 and EN60555-3, but their scopes are restricted. It would be necessary to have arc-welders officially classified as 'domestic appliances' for these standards to be applicable, and then EN55014, for hf emissions, would also apply. None of them probably meets that, so they would all have to be taken off the market!

The problems with arc-welders are particularly dire. For example, welders exist with a 13A plug fitted that actually draw more than 13A from the mains. Can we reasonably write an emc standard for a product that appears to be capable of overloading its power supply?

Meanwhile, Mr Johnson should contact the technical department of his local electricity supplier. Apart from emc regulations, there are supply authority regulations and bye-laws that can apply in such cases as his.

John Woodgate
Essex

specific to one type of thermistor.

When I came up with the controller system, there simply wasn't an immediate requirement for a thermistor sensor, since over the temperature range I was working, namely –50 to +200°C, one of the other sensors fitted the bill better.

It's not that it is impossible to design an interface for a given thermistor. I simply never needed to do it. This point is alluded to in the final few sentences on thermistors on page 169. The references to be published with the final article contain some thermistor ideas.

Moving on to diodes and transistors, you mention the properties of physical size, dissipation (self heating?) response time and thermal capacity. May I draw your attention to my opening paragraph on page 165. You will find all of these covered – albeit with different words in some cases – in the headings with bullets. I also bring in a few other important considerations like long term stability and calibration accuracy.

It is against all the engineering requirements that the suitability of a sensor for a given application must be judged, and, when considered in this light, some of your comments may be a little simplistic.

You state that the upper temperature limit is decided by the increasing leakage current with respect to the bias current. This is fine as far as it goes – without numbers. But it may be that at that temperature, you already have problems with long-term calibration drift or high failure rates due to thermal cycling when the equipment is switched on and off. Or there may be a problem simply stopping the solder on the wires from melting. There is a solder available which melts at 221°C compared to the usual 183°C by the way.

As for the lower limit, I suspect you have more experience at –200°C than myself. However, a colleague of mine-uses diodes of the 1N4148 variety as sensors at this sort of temperature. They certainly work, but there have been failures due to differential expansion/contraction with thermal cycling – not often, but certainly often enough to be far above the expected failure rate for a diode and to pose a genuine problem.

Manufacturers making semiconductors intended to be run at this sort of temperature – an astronomical ccd at –100°C for example – generally specify a maximum rate of temperature change, typically 15°C a minute.

With the above considerations in mind I specified a range of –50 to +150°C in the text because this is typical of the maximum storage temp range quoted by most manufacturers. For metal can devices maybe the range is a tad wider, at –60 to +200°C. Inside these limits it is unlikely people will run into problems with basic reliability. I concede I was probably a little too dogmatic; it is quite valid to experiment outside the manufacturer's parameters using components in ways they were never intended to be used. But there is an extra responsibility for making sure the equipment meets all aspects of the design specification reliably and in a reproducible manner.

This is especially true when the failure of a 5p diode can cause problems with other components costing hundreds or, in some cases, thousands of pounds.

I have not tried Schottky diodes, so will defer to your greater experience!

I would generally agree with your remarks concerning repeatability. Diodes or transistors from the same batch by the same manufacturer usually show excellent consistency, the calibration generally holding to 1 or 2°C over a wide range. However, between different manufacturers it is a different story. At this moment I have on my bench two batches of BC182 transistors, one by NS and the other by Motorola. The gain term on both is about –2.1mV/°C at 100μA bias but the offset term differs by about 30 mV – a calibration error of about 14°C.

Your final paragraph I find somewhat puzzling. You state the sensitivity of a diode/transistor increases with decreasing bias current. Over the years I have built several temperature controllers using transistors at various bias currents and the sensitivity has always been approximately the same.

I will do some practical checks when I get time to establish the exact relationship. Maybe I've never altered the current over sufficient orders of magnitude to notice the effect. I will reserve further comment until I have some real numbers from real devices. You are probably right in that there is some small slope change.

Your final comment concerning the low self-heating is a good one which deserved to be mentioned.

Silver's not sterling

Regarding the letter from Doug Self, November 1997 issue of *Electronics World*, he is quite correct asserting that silver forms a silver sulphide surface, and that this surface is non-conductive. I disagree though about the atmosphere as a source of this pollutant. Tarnish has been around a lot longer than diesel engines.

Because of the sulphide layer, silver is not a panacea for improving conductivity of copper microwave and rf components. A little over 20 years ago, the microwave community phased out the coin-silver waveguide for the same reason. Initially, coin silver provided a lower loss circuit than unplated copper. But with time, and a bit of tarnish, the situation would reverse.

I recall when a manufacturer of military hardware began substituting copper waveguides for silver. His product was installed in a container that was pressurised with water-pumped dry nitrogen. Over the course of several years, this dew-point would slowly rise due to atmospheric penetration. I believe Dalton's gas law was the culprit here.

When the product was retuned for refurbishment, the loss through the waveguide pieces had usually increased. The problem was ameliorated when the waveguide was changed to unplated copper. Presumably atmospheric sulphur as well as water vapour entered the container.

On another occasion, I was advising a business that had several high-Q resonators in a coupler. More Q was required, so silver plating was in order. I believe they employed a rhodium flash coating to combat the sulphide corrosion problem.

Gold is often employed as a finish for microwave connectors. In my opinion, this is more for aesthetic purposes rather than improving performance. The conductivity of gold is lower than that of copper. Often, these connectors are attached to an aluminum chasis, which eventually forms an absolutely horrid connection in the presence of a salt-water atmosphere. Stainless finish connectors are preferred in this application.

Today, unprotected silver waveguides and waveguide components are only found at flea markets.

Jim Ussalis
Florence
USA

Assembler error

I'm writing to you in response to Malcolm J Bloor's letter 'Debug your assembler first,' February 1998 issue, p. 157. I would like to point out that the error he believes to have discovered in various 80x86 assemblers is not an error at all.

Several 80x86 instructions can be coded in alternative ways. A quick look at the instruction code table clarifies the situation. Take for example the CMP instruction depicted by Mr Bloor.

```
CMP BX, 0FF80H.
```

The binary coding of the relevant CMP BBX. (immediate data) instruction is the following:

```
[1000 00sw] [1111 1011]
[data] [data if s:w=01]
```

where bits s.w have the following meaning:

```
if s.w=01 then 16 bits
of immediate data form
the operand,
```

while if s:w=11 then an immediate data byte is sign extended to form the 16-bit operand.

The more straightforward way of coding the CMP BX. 0FF80H instruction is to choose s:w=01 and then use the 16-bit immediate data. This results in the byte sequence 81 FB FF 80, which is exactly what 'Small Assembler' produces.

There is, however, a more tricky way to achieve the same result – and save a byte. As 80_{16} sign extended gives $FF80_{16}$, choosing s:w=11 and 80_{16} as immediate data will have the same effect. The resulting byte sequence is now 83 FB 80, which is what Microsoft Macro Assembler and A86 will produce.

The two ways of coding the same instruction will have the same effect, but the second way is one byte shorter, which means that the resulting program will run faster. More sophisticated assemblers try to optimise the code and this is what causes the difference between the two object files. Mr Bloor can rest assured that nothing and nobody will be killed as long as that's the only 'error' in his programs.

Laszlo Gaspar
Biggleswade
Beds

● Malcolm Bloor in a letter in the February 1998 issue complained that two out of three X86 assemblers tested did not generate the expected code, quoting various examples of the instruction:

```
cmp bx, <immediate
data>.
```

Perhaps the majority of assembler results were correct, and Mr Bloor's expectations were misguided. I tried the source example using Borland's TASM V2.01 Assembler, and got similar results to the Microsoft .ASM listing shown.

It seems that these assemblers detect the most significant FF_{16} or 00 of a suitable two byte operand,

and generate an optimised instruction variant (opcode 83 FB) which uses just the least significant byte as immediate data, but sign-extends it to a two-byte operand. There should be no objection to saving one byte, and the instruction result should be identical to that produced by the normal form of the instruction (o pcode 81 FB).

More serious is the example shown for immediate data $0FF_{16}$, which is interpreted incorrectly by the Small Assembler as $0FFFF_{16}$, and results in the optimised code 83 FB FF. The other assemblers all produce code 81 FB FF 00, corresponding to operand $00FF_{16}$. Erroneous instruction results would appear in the former case.

It is unfortunate that these assemblers do not offer an alternative convention to indicate hexadecimal notation, but insist on the leading decimal digit and trailing 'h'. This appears to be the main source of confusion.

A single non-numeric symbol equivalent to the 'S' prefix used in Pascal, would be much preferred. In the absence of this option, the best practice would seem to be always to write out immediate data to the full number of hexadecimal digits implied by the size of the destination, adding the leading zero prefix when required.

Tony Purser
Portsmouth
Hampshire

Room for resonance

I have some practical experience in enhancing low-frequency room response, and am currently analysing servo subwoofer designs. In comment on the difficulties of Alan Frobisher re bass boom, February issue, I fear that of the electronic options he lists, dsp may be the only one to offer real benefit.

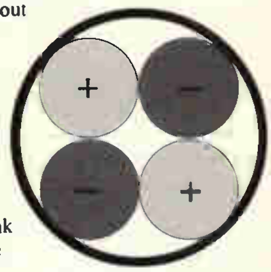
In my opinion, analogue filtering is of limited value. Even the elimination of a single resonance by analogue notch filtering will radically alter such system properties as transient response and phase behaviour. Neither will analogue phase-shifters alter 'room boom' unless it is the result of subwoofer interaction with the main speakers.

The problem primarily lies in trying to correct for complex time-domain interactions – dependent on the position of the subwoofer and the room boundaries, which are near-field at low frequencies – with the use of simple frequency-domain filtering. Since time and frequency-domain behaviours are inextricably linked in analogue systems, the appropriate corrections cannot be completed.

With dsp, such restrictions are shed, but the different complexities of design are, I guess, beyond the time commitment available to most readers.

Cable sonics

With reference to your article 'Cable sonics?' in the October 1997 issue, I would point out another interesting speaker cable configuration: the four-core one shown. This was suggested to readers a few years ago by the Italian hi-fi magazine *Audio Review*.



Using the geometric explanation of your Fig. 3, this configuration has null effective conductor spacing and consequent very low inductance. I think that there is some capacitance increase though.

I found an economical and flexible 4x2mm^2 type with orange external sheath. It was recommended for professional tools with 'three-phase plus ground' supply.

Marco Bandiera
Bologna Italy

Mr Frobisher's discussion shows that he is well aware of room dimension effects, and although I have not examined the dimensions beyond seeing that they include a hazardous 2:1 ratio, the calculation of all possible resonances is unlikely to be very informative. The way in which each of these resonant modes is driven is more important. Thus the 'suck it and see' approach is the best, and I am afraid that I am going to immediately suggest that the subwoofer must be moved from the corner!

To reiterate some old knowledge in my own terms, the corner is the position most strongly acoustically-coupled to the room, including coupling to all the room resonance modes. To describe it another way, the room corner is like the mouth of a very poorly shaped horn. Consequently this horn has a wild and varying impedance characteristic – and hence efficiency – as the driving frequency is altered.

The introduction of a wide-bandwidth linear driving source of low distortion (the Breden subwoofer) has simply illuminated the non-linearity of this corner coupling. Note that low-frequency room resonances are very hard to damp mechanically without using very large acoustic absorbers in the room. But if you have to, start in the corners.

The use of notch filters or equalisers undoes many of the design advantages of Mr Breden's excellent work, as well as being unable to correctly compensate for these room effects.

Since the way in which a room resonance will be excited depends on where the driving source is located with respect to the nodes and antinodes of that resonance, that mode may be suppressed by choosing a suitable place for the subwoofer. To cut a long story short, a position is required to minimise the effect of all the possible room modes, which is typically out in the room and away from the walls.

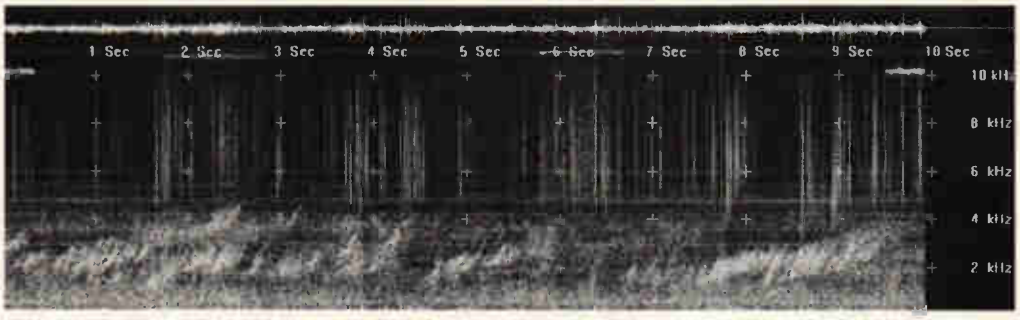
As in the case of dipole speaker in a simply shaped room about a third of the way along room diagonals – including being raised, if possible – is often a very neutral starting position. There is minimal emphasis of particular modes, since this position removes the simple multiples of dimensions appropriate to reinforcement – avoid halves and quarters of room dimensions

Anyone wishing more complexity on this topic should examine the theory of quadratic residues, as applied to sound studio design, etc.

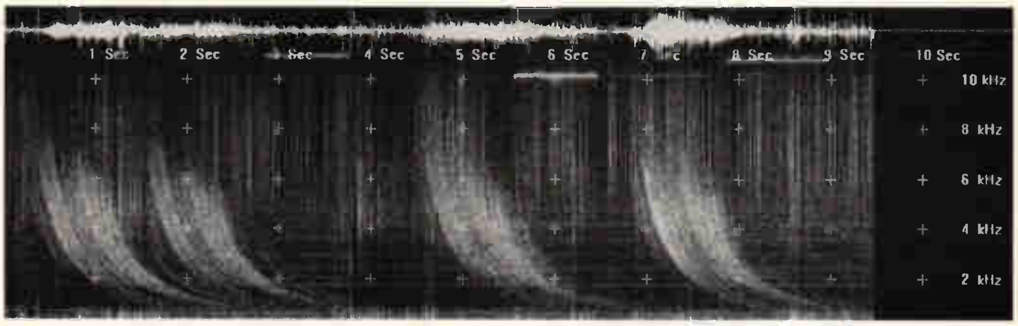
With the move out from the corner, you will also notice a drop in apparent efficiency, and will likely have to increase the subwoofer gain relative to the main speakers. Beware of the tendency to have the subwoofer volume set too high. Listen to the result, and then experiment further with positional tuning.

A little laterality in presentation will be required next to solve the 'out-of-sight' requirement. The subwoofer may perhaps be disguised as a coffee table or tv stand, or could be placed under the sofa – with the potential for other exciting effects. If the subwoofer looked like a Chippendale or a piece of sculpture, it might be given a place of honour. Unfortunately for Alan, and his wife, I do firmly believe that repositioning is the realistic solution to this dilemma, and unless you are a digital whiz, also the only reasonable solution.

Richard H Barton
Bramley
Guildford



These 'risers' were recorded during a minor magnetic storm. Short bursts of rising frequency-packages embedded in auroral noise are evident. Played through a loudspeaker, this signal resembles birdsong! A weak arc of Aurora was seen low in the north during this recording.



Four huge – and noisy – whistlers shown in the frequency domain. Horizontal lines show the overtones of the mains. Vertical lines are individual lightning strokes. The top graph displays the signal amplitude.

Rumblings

If you extend the frequency response* of Dominic Di Mario's electric-field sensor described in *EW* October 1997, you will access even more amazing signals – especially in the range of the human ear. They originate in the ionosphere.

Known as very low frequency ionospheric emissions, or vlf-ie, these signals are created by two different processes. One is lightning, the other is associated with Aurora Borealis.

If you sense the electrical field, there is a good chance that you will hear a tone with a descending frequency; this is a whistler. The magnetic path in the ionosphere acts like a prism. The higher frequencies arrive first, followed by the lower frequencies.

The time-span of a whistler varies from parts of a second to several seconds. There are variants of the phenomenon. The frequency can ascend or divide in two – or more – branches. It can be more or less broadband noise.

While there is a great Aurora above, the ionosphere is choked with electrical currents. They are induced by plasma ejected from the Sun.

The impact of this effect of radio wave propagation is well known. Some of the effect has frequency components in the range of interest here.

Currents flow along the magnetic lines equivalent to the visible streaks of the Aurora proper. There are hisses of various frequencies and bandwidths. Singing tones like birds or whales may be heard, and/or sounds of any fashion, shape or form. Sometimes many echoes can occur.

Choice of listening site is important. Although it is possible to retain the notch for the fundamental frequency of the mains 50Hz, the power grid emits a huge number of overtones that are difficult to cope with. Unfortunately, they are most frequent in the range 20Hz-16kHz. The best solution is to find a listening site well out of sight of all power lines.

A clean spectrum of vlf-ie is not conceivable within an urban area. But you can train your ears to distinguish interesting signals among trivia in a hum-filled environment – at the price of a few head-aches.

Another major consideration is the geographic latitude where this exercise is to take place. The farther north you are, the greater chance that you will hear something more than the lightning strokes. If you live near or farther north than the arctic circle, it is likely that vlf-ie will occur on a daily basis.

I live at 59° in Sweden. I hear whistler activity of some significance once a fortnight, on average. During 1997, Aurora generated sounds occurred once a month.

If you live at a latitude in the lower 40 to 50° range, you will probably have to wait a long time before you encounter even a whistler. But solar activity should pick-up in the coming few years. This will certainly increase the occurrences of sounds from both modes of vlf-ie by an order of magnitude.

If you want to keep track of Aurora activity, you can assemble your own magnetometer. See *EW+WW* March 1992, Letters.

Keywords to look for on the Internet are natural radio, whistler, Aurora Borealis and CME.

Two spectrograms of VLF-IE are shown here.

Göte Flodqvist
Farsta
Sweden

* Or, tap the signal at the output of the front end amplifier at point 6, i.e. the *OPA124P* output. This signal should be sufficient to feed to the microphone input of a cassette recorder. The antenna in this case should be a vertical 2 to 5 meters long wire.

Hit the road jack

I recently spoke to a very concerned engineer who was worried because his employer had instructed him to make up a set of connecting cables for a customer's PA system. This involved fitting jack plugs to the end of loudspeaker leads. I explained my views on this dangerous practice to him.

I appreciate that a number of big

manufacturers use jack plug loudspeaker connections – especially those in the guitar amplifier sector – and it is unlikely that this situation will ever change. My argument revolves around the application of EN60065; in particular, clause 9 "Shock hazards under normal operating conditions". Clause 9.1.1 is, I feel, quite specific in its definition of whether a contact terminal is live or not, and as such constitutes a potential hazard;

Clause 9.1.1
The part or terminal is not live if;
b) from each other part or contact, the current measured through a non-inductive resistance of 50000Ω does not exceed 0.7mA (peak) or 2mA dc.

Essentially, if the terminal can provide in excess of 34V peak, then there is a real danger of electric shock. My own personal view of this is that for safety reasons, it must not be possible to touch the output terminals of any amplifier over about 50W.

If we agree up to this point, why does the musical-instrument industry continue to use jack and Cannon plugs as connectors for the loudspeakers? I suppose you could argue that Clause 4.2.6 (for the purposes of testing, the amplifier is set up to produce 0.125 of its rated output), could be invoked to establish the loudspeaker terminal voltage, but I cannot condone this. You could also argue that the equipment should be switched off while connections are being made, but this is simply hoping that an accident never occurs.

I have never had to have to have this point tested by an accredited test laboratory. All the amplifiers I have had tested have been under 40W output, and the only one with an accessible loudspeaker is a little 10W practice amplifier.

I feel that as an industry, we should have a unified interpretation of this and I would be extremely grateful for your thoughts on this matter.

Kindest regards.

K Aston
Leeds

Low voltage circuits

With reference to Ken Hawes' query in the December issue letters pages regarding circuits operating at 1.3V, Ken might benefit from reading 'The Art of Electronics' second edition. On page 917 there starts a chapter on the art of low and micro-power design. This explains in some detail how to go about designing very low voltage power devices. The book is packed with useful information for the electronics designer.

May I say with regard to the content of the letters page I and many others I am sure are fed up with reading about the finer points of audio design. Why can't some other people be given space to air their point of view or to request technical help? Surely most people don't really care about whether there is 0.00001% or 0.0001% distortion from an amplifier at some frequency they can't hear any way? More on rf design instead please.

Ian Johnson
St. Albans Herts

I counted just two letters on audio in the January, February and March issues. I agree with your comments about A of E though – Ed.

Oops...

Fields, lines and tremors. On page 806 of the September issue, Fig. 4, both negative and shield should be connected to ground. In Fig. 8, the 6μ8 capacitor is actually 6μ8 and the rating of the instrument is 200μA and not 200μA as shown.

Frugal buzzer. This circuit, on page 767 of August issue, showed three loose wires. These should be connected together for proper operation.

Apologies to readers.

D. Di Mario

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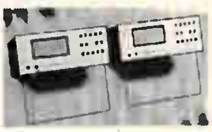

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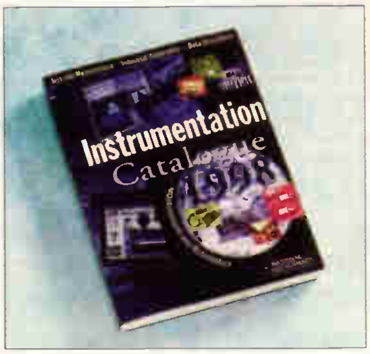

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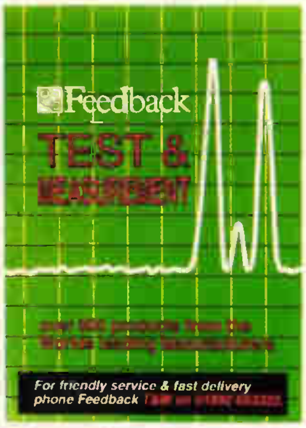

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New Special Offers

Mini waterproof TV camera 40x40x15mm requires 9 to 13 volts at 120mA with composite video output (to feed into a video or a TV with a SCART plug) it has a high resolution of 450 TV lines Vertical and 380 TV lines horizontal, electronic auto Iris for nearly dark (1 LUX) to bright sunlight operation and a small lens with a 92 degree field of view, it focuses down to a few CM It is fitted with a 3 wire lead (12v in gnd and video out) Now also available with wall mount tilt and swivel case (at the same price) .... £93 57 + VAT = £109 95 .... or 10+ £89.32 + VAT = £104 95

Board cameras all with 512 x 582 pixels 4.4 x 3.3mm sensor 9-13 volts power supply and composite video out All need to be housed in your own enclosure and have fragile exposed surface mount parts 47MIR size 60 x 36 x 27mm with 6 infra red leds (gives the same illumination as a small torch would) .... £50.00 + VAT = £58 75

40MP size 39 x 38 x 23mm spy camera with a fixed focus pin hole lens for hiding behind a very small hole ..£57.00 + VAT = £66 98
40MC size 39 x 38 x 28mm camera for 'C' mount lens this gives a much clearer picture than with the small lenses .... £68.78 + VAT = £80 83
Standard 'C' mount lens F1.6 16mm for 40MC .... £26.43 + VAT = £31.06

High quality stepping motor kits (all including stepping motors) 'Comstep' independent control of 2 stepping motors by PC (through the parallel port) with 2 motors and software .... Kit = £67.00 . .Ready built = £99 00
Software support and 4 digital inputs kit £27 00
Power interface 4A kit £36 00
Power interface 8A kit £46 00

Stepper kit 4 (manual control) includes 200 step stepping motor and control circuit £23 00
DTA30 Hand held transistor analyser It tells you which lead is the base, the collector and emitter and if it is NPN or PNP or faulty (NEW VERSION does not say FETs & SCRs are transistors) DTA30 .... £38 34
HMA20 hand held MOSFET analyser identifies gate drain and source and if P or N channel HMA20 .... £38 34

Speaker cabinets 2 way speaker systems with Motorola tweeters

Speaker dia	15"	12"	8
Power rating	250WRMS	175WRMS	100WRMS
Impedance	8ohm	8ohm	8ohm
Frequency range	40hz-20khz	45hz-20khz	60hz-20khz
Sensitivity (1W/1M)	97dB	94dB	92dB
Size in mm	500x720x340	450x640x345	315x460x230
Weight	21.1kg	16.8kg	7 4kg
Price each for black vinyl coating	£139.95	£99.99	£54 94
Grey felt coating	£159.97**	£119.97**	£64 99

(** = not normally in stock allow 1 week for delivery)

Power amplifiers 19" rack mount with gain controls & VU meters
STA300 2 x 190Wrms (4ohm load) 11kg £339 00
STA900 2 x 490Wrms (4ohm load) 15kg £585 00

LED's 3mm or 5mm red or green .... 7p each .... yellow 11p each

Cable ties .... 1p each ... £5 95 per 1000 ... £49.50 per 10.000

Rechargeable batteries

AA (HP7) 500mAH	0 99	AA 500mAH with solder tags	£1 55
AA 950mAH	£1 75	C (HP11) 1 2AH	£2 20
C 2AH with solder tags	£3 60	D (HP2) 1 2AH	£2 60
D 4AH with solder tags	£4 95	PP3 8 4V 110mAH	£4 95
1/2AA with solder tags	£1 55	Sub C with solder tags	£2 50
AAA (HP16) 180mAH	£1 75	1/3 AA with tags (philips CTV)	£1 95

Nickel Metal Hydride AA cells high capacity with no memory. If charged at 100ma and discharged at 250ma or less 1100 mAH capacity (lower capacity for high discharge rates) £3 75

—Special offers please check for availability.—

Stick of 4 42 x 16mm nicad batteries 171mm x 16mm dia with red and black leads 4.8v £5.95
5 button cell 6V 280mAH battery with wires (Varta 5 x 250DK) £2.45
Orbitel 866 battery pack 12v 1 6AH contains 10 sub C cells with solder tags (the size most commonly used in cordless screwdrivers and drills 22 dia x 42mm tall) It is easy to crack open and was manufactured in 1994 .... £8 77 each or £110 50 per box of 14

BCI box 190 x 106 x 50mm with slots to house a pcb the lid contains an edge connector (12 way 8mm pitch) and screw terminals to connect to wires and 5 slide in cable blanks £2.95

7 segment common anode led display 12mm 0.45

GoAs FET low leakage current S8873 .... £12.95 each .... £9.95 10+ £7.95 BC547A transistor 20 for £1.00

SL952 UHF Limiting amplifier L.C. 16 surface mounting package with data sheet £1.95

DC-DC converter Reliability model V12P5 12v in 5 v 200ma out 300v input to output Isolation with data .... £4.95 each or pack of 10 £39 50

Airpax A82903-C large stepping motor 14v 7.5 step 27Oohm 68mm dia body 6.3mm shaft . £8 95 or £200.00 for a box of 30

Polyester capacitors box type 22 5m lead pitch
0 9uf 250Vdc .... 18p each, 14p 100+, 9p 1000+
1uf 250Vdc .... 20p each, 15p 100+, 10p 1000+
1uf 50v bipolar electrolytic axial leads .... 15p each, 7.5p 1000+
0 22uf 250v polyester axial leads .... 15p each, 7.5p 100+

Polypropylene 1uf 400vdc (Wima MKP10) 27.5mm pitch 32 x 29 x 17mm case .... 75p each, 60p 100+

Philips 123 series solid aluminium axial leads 33uf 10v & 2 2uf 40v .... 40p each, 25p 100+

Philips 108 series long life 22uf 63v axial .... 30p each, 15p 1000+

500pf compression trimmer 60p

Solid carbon resistors very low inductance ideal for RF circuits 27ohm 2W, 68ohm 2W .... 25p each, 15p each 100+

We have a range of 0.25w, 0.5w, 1w and 2w solid carbon resistors – please send S.A.E. for list.

P.C. 400W PSU (Intel part 201035-001) with standard motherboard and 5 disk drive connectors, fan and mains inlet/outlet connectors on back and switch on the side (top for tower case) dims 212 x 149 x 149mm excluding switch .... £26.00 each, £138 00 for 6

MX180 Digital multimeter 17 ranges 1000vdc 750vac 2Mohm 200mA transistor Hfe 9v and 1 5v battery test £9 95

Hand held ultrasonic remote control £3.95

CV2486 gas relay 30 x 10m dia with 3 wire terminals will also work as a neon light .... 20p each or £8 50 per 100

Verbatim R300NH Streamer tape commonly used on nc machines and printing presses etc, it looks like a normal cassette with a slot cut out of the top .... £4 95 each (£3 75 100+)

Heatsink compound tube 0 95

HV3-2405-E5 5-24v 50mA regulator ic 18-264vac input 8 pin DIL package .... £3 49 each (100+ £2 25)

LM555 timer ic 16p
8 pin DIL socket 6p

All products advertised are new and unused unless otherwise stated
Wide range of CMOS TTL 74HC 74F Linear Transistors kits rechargeable batteries, capacitors, tools etc always in stock Please add £1 95 towards P&P (orders from the Scottish Highlands, Northern Ireland, Isle of Man, Isle of Wight and overseas may be subject to higher P&P for heavy items) VAT included in all prices

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AZ31	6.00	KT88 China	12.00	6AQ5	2.00	6V6GT	5.00
CL33	10.00	N78	6.00	6AR5	20.00	6X4	3.00
E88CC	8.50	OA2	3.00	6AS7G	7.50	6X5GT	3.00
E180F	3.50	OB2	3.00	6AU5GT	4.00	12AT7	3.00
E810F	20.00	OC3	3.00	6AU6	2.00	12AU7	3.50
EABC80	2.00	OD3	3.00	6AW8A	4.00	12AX7	5.00
EB91	1.50	PCF80	2.00	6B4G	22.00	12AX7A	7.50
EBF80	1.50	PCL82	2.00	6BA6	1.50	12AX7W	6.00
EBF89	1.50	PCL85/805	2.50	6BE6	1.50	12BA6	2.00
EBL31	15.00	PCL86	2.50	6BH6	2.00	12BE6	2.00
ECC33	12.00	PD500	6.00	6BQ7A	2.00	12BH7/1	10.00
ECC35	12.00	PL38	3.00	6BR7	4.00	12BY7A	7.00
ECC81	3.00	PL81	2.00	6BR8	4.00	12DW7	15.00
ECC82	3.50	PL504	3.00	6BW6	4.00	12E1	10.00
ECC83	5.00	PL508	3.00	6BW7	3.00	13E1	85.00
ECC85	3.50	PL509/519	10.00	6BZ6	3.00	572B	85.00
ECC88	6.00	PL802	4.00	6C4	2.00	805	45.00
ECC808	15.00	PY500A	3.00	6CB6A	3.00	807	7.50
ECF80	1.50	PY800/801	1.50	6CD6G	5.00	811A	25.00
ECH35	3.50	QQV02-6	12.00	6CL6	3.00	812A	55.00
ECH42	3.50	QQV03-10	5.00	6CG7	7.50	813	27.50
ECH81	3.00	QQV03-20A	10.00	6CH6	3.00	833A	85.00
ECL82	3.50	QQV06-40A	12.00	6CW4	6.00	866A	20.00
ECL86	3.50	U19	8.00	6DQ5	17.50	872A	30.00
ECLL800	25.00	UABC80	1.50	6DQ6B	10.00	931A	25.00
EF37A	3.50	UCH42	5.50	6F6G	6.00	2050A	12.50
EF39	2.75	UCL82	2.00	6FQ7	7.50	5751	6.00
EF40	4.00	UCL83	2.00	6GK6	4.00	5763	6.00
EF86	10.00	UF89	4.00	6J5G	6.00	5814A	5.00
EF91	2.00	UL41	12.00	6J5M	4.00	5842	12.00
EF183/4	2.00	UL84	3.00	6J7	3.00	6072A	6.00
EL33	15.00	UY41	4.00	6JB6A	27.50	6080	6.00
EL34	8.00	UY85	2.00	6JE6C	27.50	6146B	15.00
EL34G	6.00	VR105/30	3.00	6JS6C	27.50	6201	8.50
EL36	5.00	VR150/30	3.00	6K6GT	4.00	6336A	35.00
EL41	3.50	Z759	10.00	6L6G	15.00	6550A	25.00
EL84	2.25	Z803U	15.00	6L6GC	15.00	6883B	15.00
EL95	2.00	2D21	3.50	6L6WGB	10.00	7025	7.50
EL360	15.00	3B28	12.00	6Q7	3.00	7027A	25.00
EL509/519	12.00	4CX250B	45.00	6SA7	3.00	7199	15.00
EM34	15.00	5R4GY	7.50	6SC7	3.00	7360	25.00
EM81/4/7	4.00	5U4G	10.00	6SG7	3.00	7581A	15.00
EN91	7.50	5U4GB	10.00	6SJ7	3.00	7586	15.00
EZ80/81	3.50	5V4G	4.00	6SK7	3.00	7587	20.00
GZ32	8.50	5Y3GT	2.50	6SL7GT	5.00		
GZ33/37	6.00	5Z3	5.00	6SN7GT	5.00		
KT61	15.00	5Z4G	6.00	6U8A	1.50		

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