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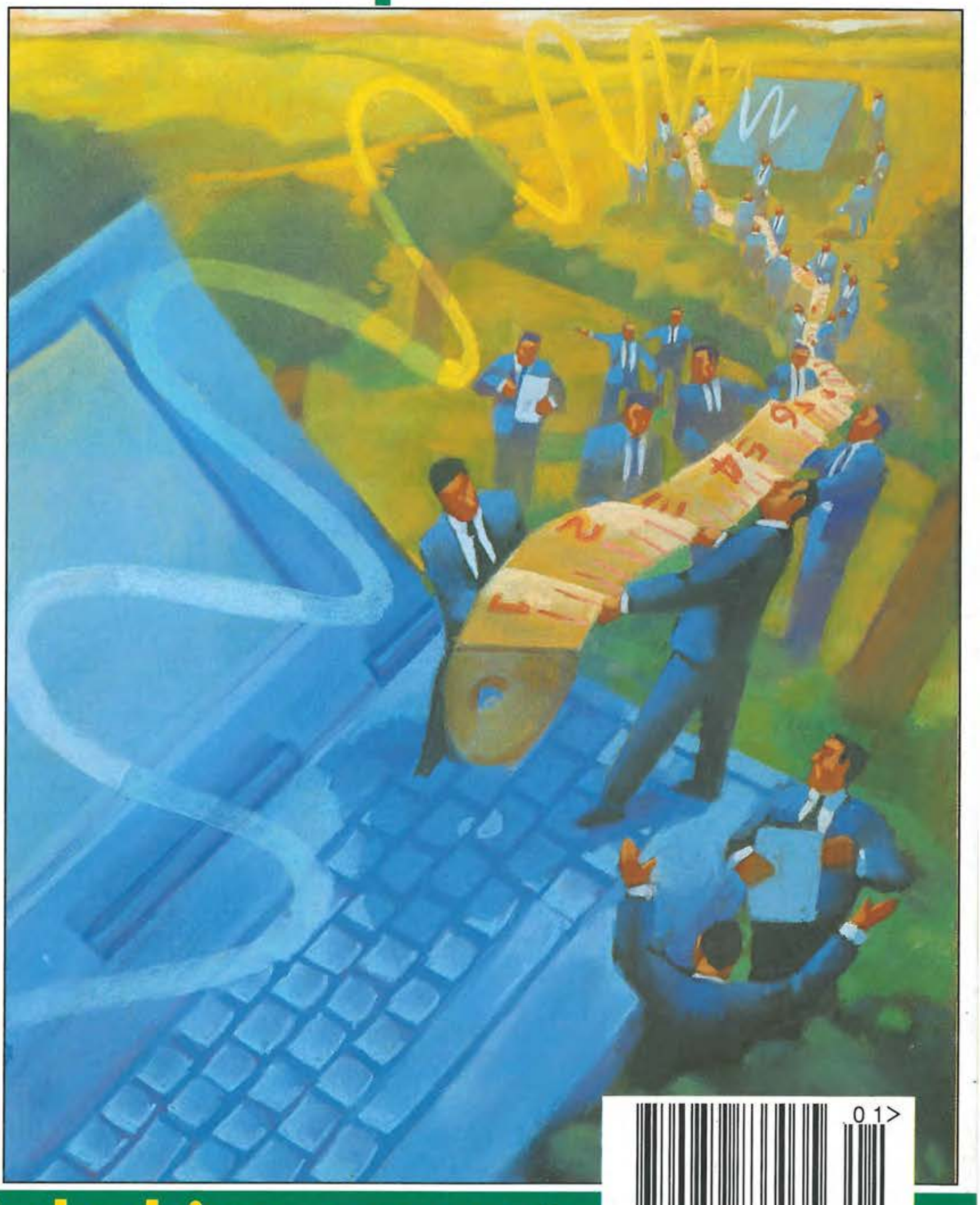

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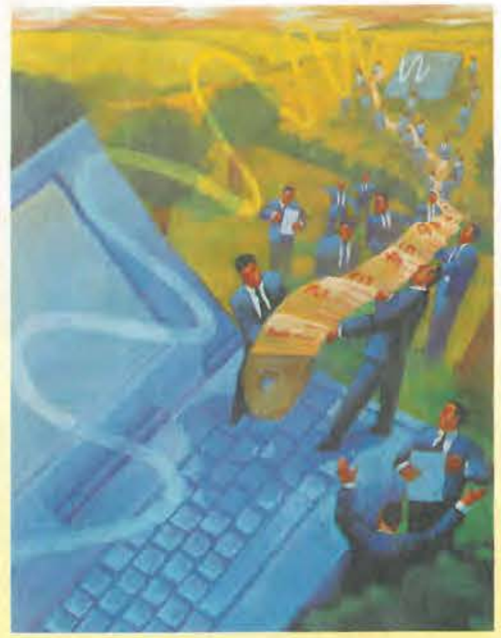

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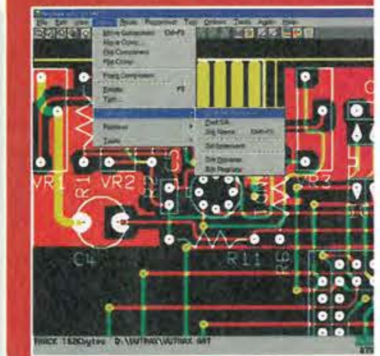
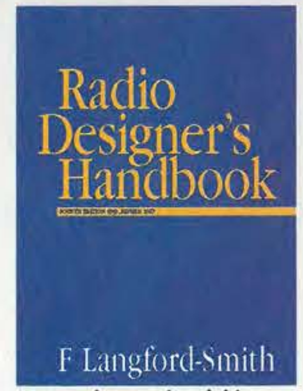
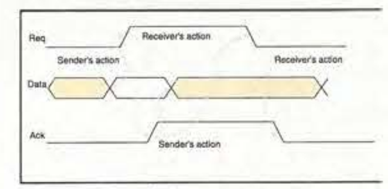




Newnes has reprinted this invaluable and comprehensive work from the valve era. *See page 21.*



Strange though it may seem, a digital system can work more efficiently using asynchronous inter-circuit communication as opposed to the traditional global clock – *see page 10.*

FEBRUARY ISSUE ON SALE 2 JAN

Microchips and megadeaths

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"Then I was shocked by the feeling that the skin of my face had come off. Then, the hands and arms, too. Starting from the elbow to the fingertips, all the skin of my right hand came off and hung down grotesquely. The skin of my left hand, all five fingers, all came off... Hundreds of people were squirming in the stream. I couldn't tell if they were men or women. They all looked alike. Their faces were swollen and grey, their hair standing up. Holding their hands high, groaning, people were rushing to the river... Under the bridge were floating, like dead dogs or cats, many corpses, barely covered by tattered clothes. In the shallow water near the bank, a woman was lying face upward, her breasts were torn away and blood spurting... By my side many junior high school students were squirming in agony. They were crying insanely 'Mother! Mother!' They were so severely burned and bloodstained that one could scarcely dare to look at them. I could do nothing for them but watch them die one by one, seeking their mothers in vain."
(Eyewitness account, Hiroshima, 6 August 1945)

Engineers played their part in the making of these events. Thirty-five years later their role has become central, for the technology of delivering death has been greatly improved. We no longer have to rely on manned aircraft to drop atomic bombs but send them as the warheads of self-guided missiles. This is where electronic engineering makes its particular contribution to slaughter, in the design of the guidance system. Consider, for example, the Trident and the Tomahawk, the two nuclear missiles which the UK Government, without benefit of open Parliamentary debate, has swung on a reluctant nation. Both of these have guidance systems which rely on advanced digital microelectronics to update an inertial navigator. In the Trident, a submarine-launched ballistic missile intended as Britain's independent nuclear weapon, the electronic system receives reference information from the optical pattern of the stars. The Tomahawk, part of a NATO arsenal that will be owned and operated by US military forces, is a cruise missile; here the electronic system receives reference information on the geographic contours of the desired route from a magnetic-core memory and information on the actual contours over which it is travelling from a radar altimeter. And such is technical progress that as we get more and more devices on a single silicon chip so we are able to kill more and more people

with a single missile.

Through work on such weapons electronics engineers in the East and the West have put themselves in the service of politicians, generals and industrialists who have become monomaniacs; who seem to see no way out of the self-perpetuating system of threat and counter-threat into which they have locked themselves and, like drug-addicts, desperately go on with it. The only thing likely to drag them out of their dementia is a threat from another direction – a concerted threat of rebellion from the trapped populations.

It becomes increasingly clear, as our distinguished American contemporary *Science* has said, "that deterrence cannot ultimately be stable, and that the civilian populations of the world are no longer defended by the armed forces for which their taxes pay, but are merely hostages to them."

None of us can be proud to serve a technology which is being used in the name of "defence" as a means to attain immense human suffering. Because we know what this technology can do we should be among the leaders of dissent." ■

This remarkable leader first appeared in *Wireless World*, November 1980. It was written by Tom Ivall, then editor of *Wireless World* and one of the most polite, considerate and intelligent men I have had the pleasure to work with. He died on 12 October 1997, but since he taught me all I know about producing the magazine, his influence lives on.
Martin Eccles.

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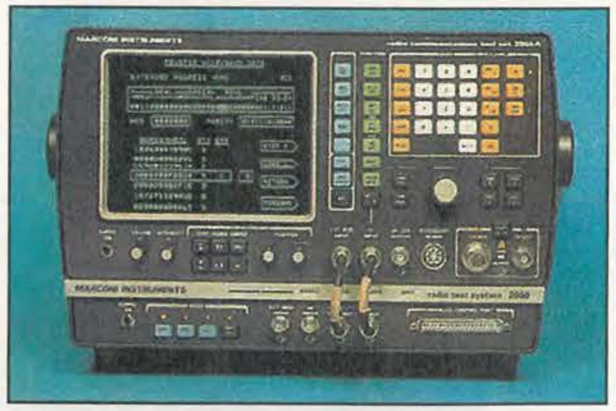


UP DATE

New turn in DVD-RAM spec battle

The DVD Forum led by Toshiba, Matsushita and Hitachi has agreed on a new DVD-RAM specification. It combines the best features of the previous DVD Forum proposal with that of DVD+RW promoted by the Sony/Philips/Hewlett-Packard team.

According to Nick Sundby from Hitachi, neither existing format met the requirements of the Forum. "Sony/Philips format is more oriented to streaming data rather than random data but their proposal had a better error rate scheme," said Sundby.

However, the Sony/Philips/HP faction is determined to sell its own format (DVD+RW) despite its incompatibility with other DVD equipment and despite the companies' presence in the DVD Forum. This has caused the Forum to rebel against the name DVD+RW and force Sony/Philips/HP to rename it to phase-change/read-write (PC-RW).

"The Philips/Sony technology is not DVD. It has now been labelled Phase-Change/ Read Write, and so can no longer be classed under the same umbrella as DVD," said one Toshiba top executive.

Meanwhile, six of the ten companies holding patents on the initial DVD technology will start a joint licensing operation different to the one established earlier by Sony, Philips and Pioneer. The programme's objective, as before, is to act as a one-stop-shop for DVD technology to new parties, but now it will be operated by Toshiba.

The delays introduced as a result of such issues have caused anger across the industry. "We should have been in this position a year ago," said Peter Scatchard of Hitachi.

US trials 1.5Mbit/s phone link

California's telephone company Pacific Bell is planning a major trial of digital subscriber line (DSL) services in Silicon Valley.

The trial is being viewed as the opening shots in a battle between cable TV and telephone companies over the lucrative high bandwidth Internet services market.

Pacific Bell will make a formal announcement later this month and the trial will involve tens of thousands of customers. The company will offer several levels of DSL services, enabling users to download data at up to 1.5 Mbit/s rates.

The Silicon Valley area is also the headquarters of @Home Network, a start-up funded by cable TV companies and operating a local trial cable modem service. Pacific Bell is targeting Silicon Valley because of its high numbers of PC users. If the trial is successful it will be offered across California.

Other US telephone companies are also stepping up with DSL trials. Arizona-based US West Communications, last week introduced DSL services to Phoenix area customers.

"This is a real breakthrough. No longer does getting on-line mean having to wait in line," said Solomon Trujillo, CEO of US West.

Industry analysts have mixed opinions over which technology, cable modems or DSL, will succeed in establishing itself as the dominant technology for faster Internet access.

Both technologies have developed more slowly than expected but increased competition between the two camps should spur the development of cheaper and better products.

Engineers' pay up 20%

Chartered electronics engineers' salaries have grown 20.5 per cent since 1995.

According to the Engineering Council's 1997 'Survey of Professional Engineers and Technicians', chartered engineers working within the electronics and telecommunications fields earn, on average, £42,631 per annum. This compares to £40,131 for chartered engineers in general.

Mike Heath, director general of the Engineering Council, said that the survey "combats the myths and misperceptions that still persist about the engineering profession - that it is badly paid, insecure, and does not offer a route to the top."

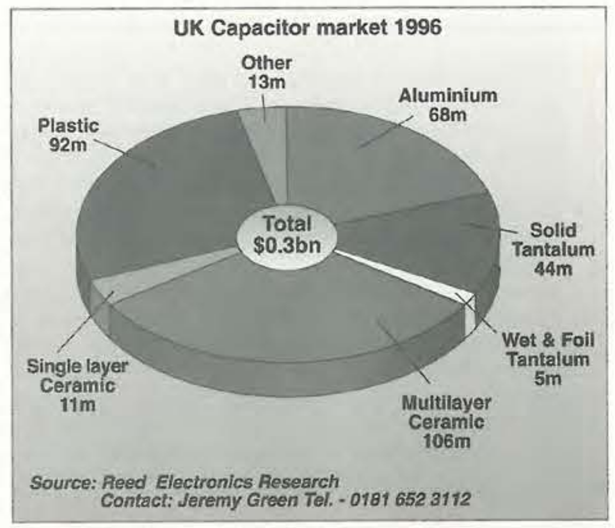
Marathon runner lost in car park

All athletes get injuries from time to time, but few get bent axles. The debut of Professor Kevin Warwick's half marathon-running robot Rogerr was spoiled by a pre-race accident.

"For safety reasons, the race referee asked us to demonstrate Rogerr's collision avoidance performance in a car park just before the race. Unfortunately, half way round, its infra-red sensors locked onto the Sun. Rogerr then tried to keep this huge infra-red source two metres ahead of itself and consequently shot off at full speed in a straight line – until it hit a kerb," said Warwick: "After that it could only go in circles."

The low winter sun came as a surprise. "We were expecting the usual English overcast weather; the robot had worked fine at eight o'clock before the sun came up."

According to Warwick, the mechanical repairs are trivial. "Rogerr tracks an infra-red source on my back and a laser tracker would solve the problem. But we haven't got any spare cash for design changes," commented Warwick.



Internet mobile phones to be developed in UK

UK firms The Technology Partnership (TTP) and STNC Enterprises are to develop a compact Internet-enabled GSM mobile phone.

"It will have similar functionality to the Nokia 9000, but smaller," said Ran Mokady, MD of STNC.

Called the Webwalker, the device "will be no bigger and of similar cost to today's [most compact] phones."

But the intention isn't to clone the 9000. "It will be capable of browsing the Web, but we see it being used with direct services,"said Mokady.

Direct services, says Mokady, are provided for a specific function, for instance a diary that can be remotely updated by a secretary, but read by the user. "Perhaps you come out of a cinema and want a chinese meal," he said: "You would use Webwalker to read a restaurant guide."

STNC's contribution is its 'lightweight' Web access software which is claimed to require less processing to execute while losing "almost nothing" in browsing performance.

TTP consultancy has developed digital mobile phone technology for the likes of Hitachi and Analog Devices. According to Mokady, the Webwalker will use existing TTP-designed chips and will be offered to phone makers to buy off the shelf and have in production "in months". He refused to say if a customer is lined up, but said that the project is funded solely by the two firms.

Is the Webwalker too late to market with the Nokia 9000 available and Ericsson soon to release a communicator? "They may be a bit ahead to market," said Mokady: "but we are a bit ahead on technology: our product will be smaller."

Television censor sensor could appear in US PCs

US pcs may soon be shipped with the controversial V-chip which blocks violent scenes in television programs.

The US Federal Communications Commission (FCC) says it is looking at federal laws that apply to the V-chip and will issue a ruling on whether pcs should include the IC. Several companies already produce hybrid pc/tv systems, and Gateway 2000 says it is making plans to include the V-chip in its Destination pc/tv system if there is an FCC ruling.

The FCC published a report recommending that the law be applied to pcs. "We believe that the program blocking requirements we are proposing should apply to any television receiver – including pcs – meeting the screen size requirements," it said.

However, many computer and Internet companies oppose any attempt to censor Internet content and have fought against proposed laws that would make it illegal to transmit sexually explicit images.

The V-chip is designed to screen out violent content but it could also be used to filter sexual images. According to a US law, all tvs must carry a V-chip within the next two years.

The V-chip could even become available early next year in special converter boxes attached to televisions. The V-chip relies on tv broadcasters tagging violent shows with codes that enable parents to decide which programmes their children can watch.

Will single Euro currency boost UK r&d spend?

A former secretary of the DTI's Innovation Advisory Board believes early adoption of a single European currency will help boost investment in r&d by UK firms.

John Chapman says the relatively high cost of raising funds for research and development is a possible cause for the UK's low standing in international R&D league tables. The cost of investment funds is attributed to the burden of dividend and interest payments.

Writing in the *The Independent*, Chapman said that the convergence of interest rates across Europe resulting from monetary union could reduce the cost to companies of raising funds for investment, including r&d. He called it a potential solution that "shines out like a beacon."

With figures derived from the DTI's latest r&d Scoreboard, Chapman calculates that the average cost of raising funds for r&d to the UK's two largest electronics r&d spenders, GEC and Racal, is higher than international and European averages. In the United Kingdom, the average cost is 5.8 per cent of company sales compared to 2.6 per cent in Europe.

Pocket cinema... Samsung has made what it claims is the world's smallest, lightest DVD. Called *P-Theater*, it measures 200 by 160 by 55mm and weighs 0.9kg. Outside the home, it connects to a head mounted display said to give the sensation of looking at a 40in to 70in screen. When operating the optional head-mounted display the unit's lithium-ion batteries will run for two hours. It can also be connected to a regular tv or projection tv for use in the home. The new product can play DVD titles as well as video cds and regular cds. It supports the conventional 4:3 width-to-height ratio screen and the 16:9 wide screen with definition of 544 horizontal lines or more, as well as 5.1 channel digital 3-D sound. In addition to personal portable use, Samsung intends it to be used to supplement automobile audio-visual systems or as an external DVD-rom unit for pcs.

Britain declared 'most competitive'

"Europe is not competitive - except for Great Britain," said Pasquale Pistorio, president and CEO of SGS-Thomson Microelectronics at the Future Horizons conference. "We need an open, liberal, economic system like the American or the British system," added Pistorio, who has just been given the Italian equivalent of a knighthood by Italy's prime minister.

Electronic nose for faster infection detection

Infection detecting applications for smell sensor technology from the UK's Aromascan could lead to a tie-up with medical companies.

"I won't say who but we are in discussions with several medical equipment companies,"said Sue Kennerley, Aromascan's product manager.

Manchester's Withington Hospital has been using Aromascan's technology since before it was formed as a separate company. Through the collaboration, it has been discovered that volatile chemicals given off by bacteria can be detected and used to diagnose infection. "It can do it much more quickly than conventional techniques which can take between 48 and 72 hours," says Kennerley.

Infection in burns, and the bacteria that cause skin ulcers in the elderly have also been detected, claims Kennerley, as well as an infection that some now think may be a cause of late-term miscarriages in pregnancy.

A hospital at the University of Pennsylvania also has an Aromascan machine and is using it to detect pneumonia by sensing the breath of intensive care patients.

Work is being carried out using the company's laboratory-based machines. "We don't know what form equipment specifically made for medical use might take," said Kennerley. "Eventually a hand-held unit may be taken to the patient."

Fast Internet trial in US

A high speed version of the Internet being developed for US researchers has shown its potential in a series of impressive demonstrations.

The Internet2 (I2) was shown enabling high performance computer processing over remote links between US research centres. These included a three dimensional mapping of a live brain, operating a scanning electron microscope remotely, weather modelling and spectral microscopy.

Many leading US high-tech companies have donated funding and resources to the I2 project which links more than 100 research centres.

Hackers remains a threat

An influential presidential commission has warned Bill Clinton about the threat to vital US computer banking, communications and military centres from hackers and foreign agents.

The commission's report is classified but the published conclusion states: "The commission has not discovered an imminent attack or a credible threat sufficient to warrant a sense of immediate national crisis. However, we are convinced that our vulnerabilities are increasing steadily."

The panel recommends that the US government step up efforts to prepare computer defences and spend as much as $1bn on safety measures by 2004. Part of the money should be given to universities and companies for the development of technology to safeguard computer systems, recommends the report, which also argues for tight controls on powerful encryption technologies.

In the latest example of hacker chaos, messages were placed on New York's subways signs displacing the standard "Watch your step" and "Have a great day".

In brief

Micromachines help spine injuries

Electronic engineers at the University of Sheffield are to team up with the Northern General Hospital to look at ways to tackle spinal injuries using micromachines.

"The project is completely open," said Dr Rob Yates, head of the University's micromachines group: "In the long term, we would like to give surgeons the ability to stitch nerves together."

Yates emphasised that the project is still in its infancy: "Our contribution could be as simple as sensors for a smart bed to help prevent bed sores." But there are other possibilities: "Surgeons may need some mechanism, like nerve-fibre sized tubes, to guide and align nerves so that they can be joined," said Yates. Other ideas include electrically interconnected sensor arrays to bridge the gap between broken nerve ends.

The Sheffield Group is interested in hearing from parties wishing to collaborate. Tel: 0114 222 5854.

Anglo interference

Radio communication equipment in Northern France could be affected following the introduction of digital radio services along the south coast of England.

French defence communications and equipment such as automatic garage doors use the same frequencies as digital audio broadcasts – around 220MHz.

A BBC spokesperson said:"We are aware of that issue, but we are working on plans for the transmitters."

This involves angling the dishes on some transmitters away from France. Those that cannot point away will have their power levels reduced.

The DTI does not see the frequency clash as a problem. "The French have got the same frequency allocation for their digital radio as we have," said a spokesperson for the Radiocomms Agency at the DTI.

Funding sought for UMIST self-drive project

A University of Manchester(UMIST) project involving self-driving vehicles is setting its sights on a large slice of the DTI's £5m Foresight Vehicle LINK pie.

The project, involving UMIST's Panos Liatsis along with automotive specialists Lucas Verity and Rover Group, concerns vision systems for cars that can handle myriad road environments including towns and cities (see Electronics Weekly, April 2).

"This will not be like the experimental systems available today - which are aimed at motorways, where the traffic environment is more constrained and where there is a unidirectional flow of vehicles," said Liatsis.

The latest technology development involves separating light received by an automatic vehicle into reflected light - from the road itself, road markings and obstacles - and ambient light, from the Sun, Moon, street lighting or other sources. This enables much clearer and sharper edge detection, aiding the vehicle's ability to see objects.

Liatsis, who is also negotiating with Nissan, is submitting his application for funds in mid-November. "I am not aware of other groups working in this area," he said. "So, hopefully, we will be able to start a new project continuing on the same lines where we could get out real products rather than just concept demonstrators."

Time for a new clock

Roy Rubenstein reports on the benefits of getting rid of the global clock in digital designs and replacing it with decentralised asynchronous circuitry that works at its own pace.

Digital designers are creatures of habit. For logic, this means working to the beat of a clock. Designers care little about the logic's state during a clock cycle as long as it is stable at the clock edge. This synchronous view of the world has proved remarkably successful; you only have to note the prominence given to the microprocessor's clock speed in pc advertisements.

Yet despite this predominance, there is a growing community of designers – albeit still small – who eschew a global clock in favour of asynchronous circuitry – logic that works at its own pace.

Interest in asynchronous design stems from the promise of a reduced power consumption. Complementary mos, or c-mos, circuits dissipate energy on switching. Clocked circuitry switches irrespective of the processing load, but asynchronous logic switches only when useful work takes place.

Asynchronous design offers other benefits. Since self-timed logic works at its own pace, it can be designed for the most commonly encountered processing loads; on the rare

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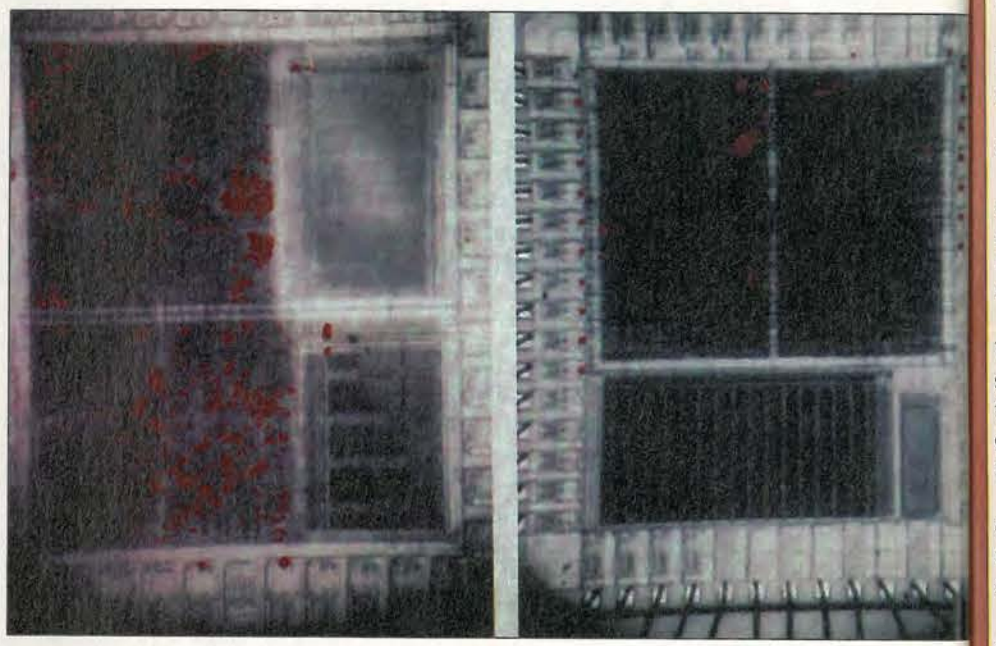
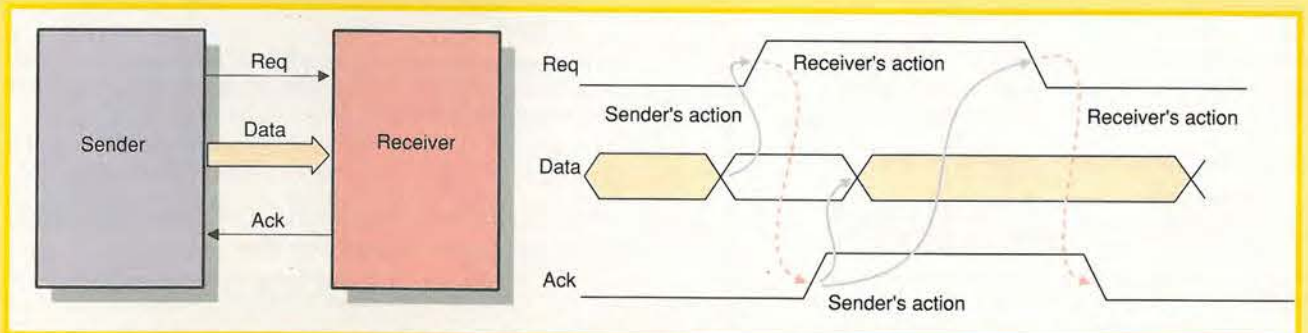


Here are two ICs capable of carrying out similar tasks developed by Philips Research Labs and Eindhoven University. The red colour indicates power dissipation. The IC with less red is the asynchronous version, demonstrating the power consumption benefits of the design approach.



Operation of the simplified circuit on the left is straightforward. The sender prepares the data and then notifies the receiver data is available by asserting the request signal. The receiver consumes the data and asserts the acknowledge signal to complete the transfer. Several timing protocols are possible for the implementation of the handshake signalling. One such protocol is the four-phase 'early' protocol as shown in the timing diagram.

occasion when the worst-case condition arises, the circuitry simply takes that bit longer. In contrast, synchronous designs must always satisfy the worst case condition. This makes it more complex, even though the extra hardware is only occasionally fully exercised.

Avoiding a common clock also significantly reduces the circuit's electromagnetic radiation. "Going asynchronous is akin to the tradition of an army breaking step to cross a bridge to avoid exciting resonances in the structure," explains Professor Steve Furber, a leading proponent of asynchronous design.

It is this trait which is most attracting the synchronous community's attention.

ASYNCHRONOUS DESIGN AT A GLANCE

- Asynchronous design offers inherent power savings since each transistor or gate switches only when doing useful work. This makes asynchronous circuits ideal in standby circuits or mobile communicators where idle times are significant.

- Since there is no global clock switching the logic in unison, large current spikes are avoided. This reduces electromagnetic radiation, simplifying the shielding and aids the long term prospects of a single radio IC.

- A huge momentum has built up for synchronous design. Self-timed circuitry must demonstrate significant advantages before EDA companies consider it worthwhile to develop the required asynchronous tools.

Self timed circuitry

The most successful recent demonstration of the merits of asynchronous design is from UK start-up, Cogency Technology. It has developed an asynchronous version of a commercially available 16-bit digital signa processor, enabling the first direct comparison between the two design philosophies to be made.

Cogency developed the asynchronous signal processor for LGSemicon after the Korean firm wanted to reduce the power consumption of its fax/modem dsp chip. The results are impressive: the asynchronous device's power consumption is almost halved, while its overall transistor count is a fifth less than the original dsp.

The reduced transistor count is surprising since asynchronous designs require additional handshaking logic to interface the various self-timed logic subsystems. "The saving is due to the [processor's] decode logic being considerably simpler," said Warren Lien, president of Cogency.

With the asynchronous dsp chip, all the functional units connect to a common bus. The decoded instruction is placed on the bus and noted by all the units. Those idle during the instruction immediately respond with an 'acknowledge' signal. The other units begin their operations, and on completion signal an acknowledgement. If one unit depends on the results of a second, its stalls until the latter signals its completion.

Cogency used its application-specific integrated architecture, known as ASIA, tool to design the dsp. "The idea of ASIA is to isolate the user from self-timed design," said Lien. Cogency is currently developing a second self-timed device for an unnamed customer. It is aimed at the mobile product market.

Another firm – anything but a start-up – increasingly enamoured with asynchronous design is Philips. It has designed a digital compact cassette error correction logic block that consumes a fifth of the power of an equivalent clocked version. But the resulting silicon area is 20 per cent larger.

Earlier this year Philips Research Laboratories detailed several asynchronous designs for the mobile communications market with a power consumption ranging from a third to a tenth that of equivalent synchronous circuits.

A tenth the power – yet faster

One of the Philips designs – a standby circuit for pagers – is not only ten times more power efficient, but faster as well. This enables the voltage supplies to be lowered, reducing power consumption further. The devices were designed using Philips' Tangram – a silicon compiler for asynchronous design.

Philips is now in the process of migrating its asynchronous expertise from its research laboratories to parts of Philips Semiconductor. "Applications under consideration are in battery-powered, hand-held consumer products," said Kees van Berkel, senior scientist at Philips Research Laboratories, Eindhoven.

A further group trailblazing asynchronous design is one led by Professor Furber at Manchester University. His group has been developing its third generation asynchronous 32-bit ARM risc processor.

The latest Amulet3 will implement Advanced Risc Machine's 16-bit Thumb instruction set. The project's focus is to embed the Amulet3 core in a range of applications, much in the way of existing ARM products.

The current most advanced asynchronous ARM is the Amulet2e. "The Amulet2e has been very successful. It is competitive with clocked ARMs in area, performance and power efficiency," said Furber.

The reason Amulet is still a research project is that that the 2e is still not a full product specification: "It isn't sufficiently testable for volume production." This is something Furber hopes to put right with the Amulet3.

Is it time for asynchronous?

Digital Semiconductor is the company probably best placed to speak for the synchronous logic community. After all, distributing a 600MHz clock across its 209mm^2 Alpha 21164 is no mean feat; the clock skew needs to be below 90ps.

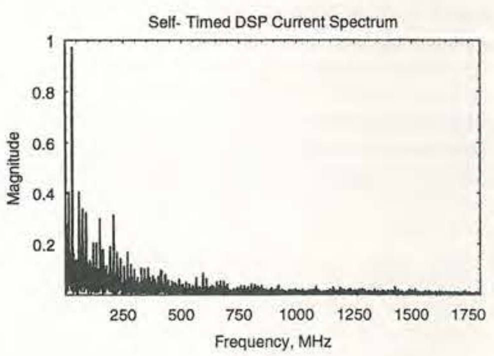
Interestingly, for its next Alpha, the 21264, Digital will use several clocks, each confined to specific regions of the chip. The reason for this, claims Digital's Aaron Bauch, is as much to do with dynamic power management – switching off areas of the die to reduce power consumption – as the issue of clock skew. That said, the company sees plenty of mileage in further increasing the clock speed.

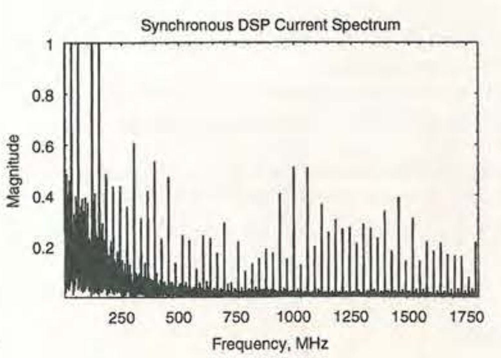
For Bauch, the key issue for the success of asynchronous is one of design automation tools. "Future microprocessors will have tens of millions of transistors. It will not be possible to do such designs without significant automation." Such tools are simply not available in the asynchronous world.

Tudor Brown, chief technology officer at ARM, agrees: "The biggest issue is design time and the issue of productivity. The challenge for asynchronous is to turn out designs just as quickly."

All proponents of asynchronous design agree that its most promising benefit is its reduced radiation. This coupled with its low power characteristics makes it ideal for embedded designs such as mobile phones.

Furber has noted another interesting development: "Intel has been recruiting asynchronous designers a lot recently." It is widely believed that portions of the floating point units in Pentium processors use asynchronous

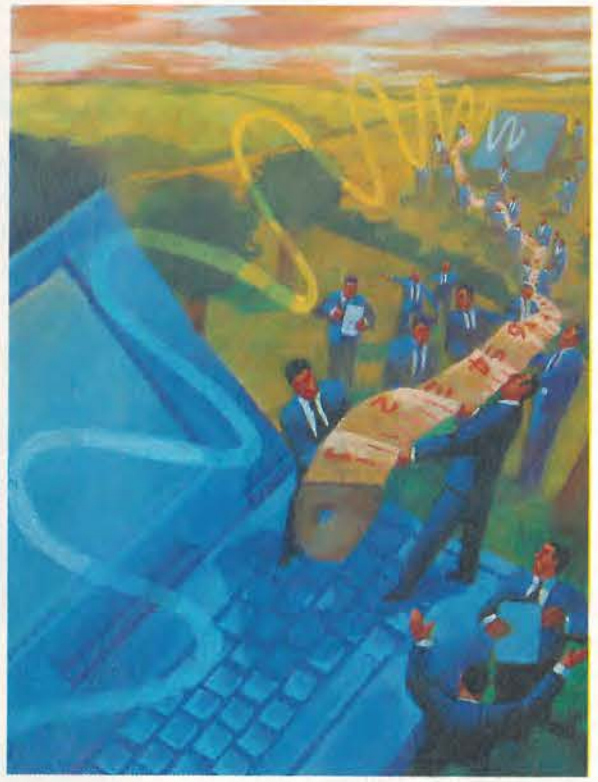




These are spectral plots of rf emissions of the asynchronous and synchronous versions of the new digital signal processor. The synchronous clock spreads spurious components across the radio spectrum.

Analogue input without wires

This wireless data logger has a transmitter with an eight-bit a-to-d converter and a receiver that interfaces directly to the pc's printer port. Its designer, Pei An, has also included a sample Turbo Pascal listing illustrating how to read a serial data stream into the port.



A typical data logging system consists of sensors, analogue-to-digital converters and pc interfacing circuits. The sensors convert physical quantities such as temperature, pressure, sound or light intensity into analogue voltages. The a-to-d converters then convert the voltages into digital data ready to be fed into the computer via an interface.

In many situations, electrical isolation is required between the sensor and the computer. For example, measurements of bio-electrical signals from the human body require safety isolation while measurements involving high voltages need to be isolated in order to protect the computer. Sometimes, the objects to which the sensors are attached are in motion so a conventional wire link is not practicable.

One form of data logger with the potential to meet these requirements is a radio-linked system. This article describes a simple radio-linked data logging system, **Fig. 1**. It consists of an eight-bit a-to-d converter with radio transmitter unit and a radio receiver that feeds a pc interfacing unit.

Output from a sensor at a remote location, after any necessary signal conditioning and amplification, feeds the transmitter unit. This unit is made up of an a-to-d converter, digital control circuitry and the radio transmitter. The transmitter unit constantly converts analogue voltages into digital data and 'broadcasts' the data through its radio transmitter.

The receiver unit plugs into the Centronics printer port of a standard pc. It receives the radio signal, converts it into a serial digital data that is fed directly into the printer port.

Since the radio link relies on low power 418MHz fm radio transmitter and receiver modules that are type-approved for UK use, further type approval for the finished design is not necessary. Indoors, the communication distance is about 50m, while outdoors it extends to 200m, depending on the terrain.

As the system communicates via the printer port of a pc, it can be used with desk-top pcs as well as laptop or even palm-top pcs. It is also possible to modify the receiver unit so that it can use the RS232 port or the game port.

How the link works

Analogue voltage generated by a sensor is amplified, conditioned and then fed into the transmitter unit. Inside the unit, an eight-bit serial i/o a-to-d converter turns the analogue voltage into a stream of serial digital data.

A timing circuit controls the operation of the a-to-d converter. This circuit also adds a start marker at the beginning of the serial data bits to indicate the start of a data transmission. Serial data is fed into the fm radio transmitter module. A frequency-modulated 418MHz radio signal is generated and transmitted to the surroundings via an antenna, **Fig. 2a)**.

Inside the receiver unit, the radio receiver module picks up the radio signal and converts it back to the serial digital data. The data is then read into a computer via the Centronics printer port. Software on the pc finds the start marker first, then finds eight serial data bits. Finally, the serial data bits are converted into parallel data, Fig. **2b)**.

Circuit diagrams of the transmitter and the receiver units are given in **Figs 3** and **4**.

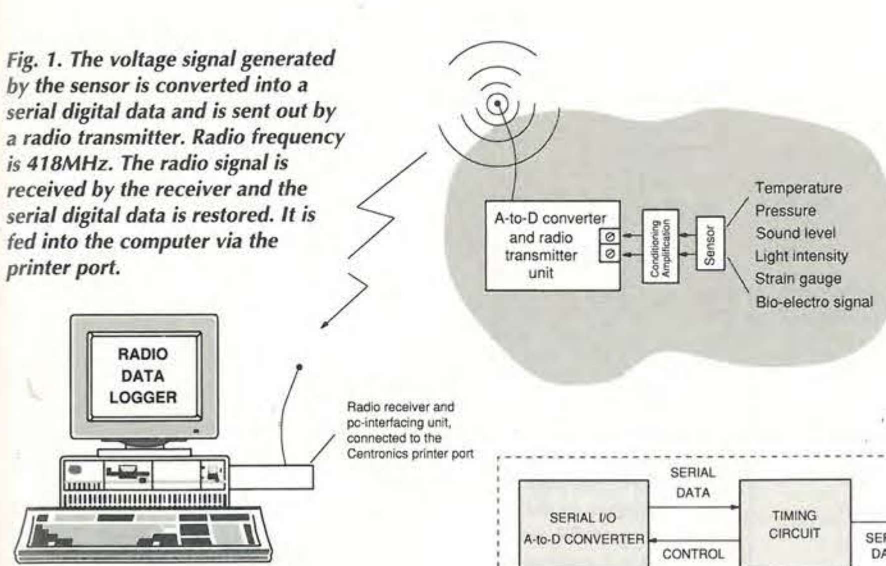


Fig. 1. The voltage signal generated by the sensor is converted into a serial digital data and is sent out by a radio transmitter. Radio frequency is 418MHz. The radio signal is received by the receiver and the serial digital data is restored. It is fed into the computer via the printer port.

Relaying data off air

The transmitter unit has four blocks: the a-to-d converter, which is a *TLC548* with eight-bit serial i/o, the timing unit, based on a *CD4060, CD4017* and *74HC157* logic ICs, the radio transmitter module *TXM-418-A* and a power supply. The functions of the blocks are outlined below.

A-to-D conversion. The *TLC548* is a LinCMOS eight-bit switched-capacitor a-to-d converter using successive-approximation for conversion. It has an on-board sample-and-hold circuit, an on-board 4MHz clock and a serial i/o interface. The device is capable of sampling 45 500 times a second.

The *TLC548*'s power supply voltage range is 3 to 6V and it has a typical current consumption of 1.9mA. Pins 1 and 3, REF+ and REF–,

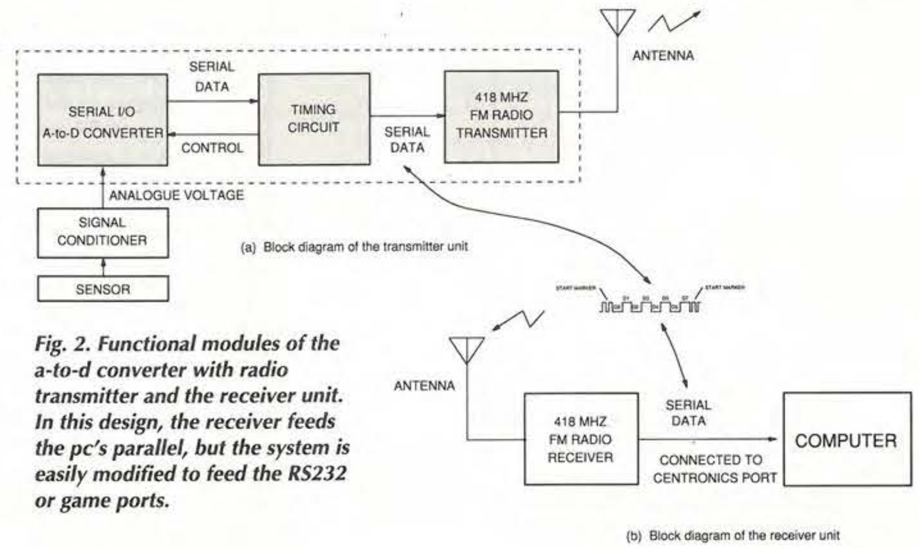


Fig. 2. Functional modules of the a-to-d converter with radio transmitter and the receiver unit. In this design, the receiver feeds the pc's parallel, but the system is easily modified to feed the RS232 or game ports.

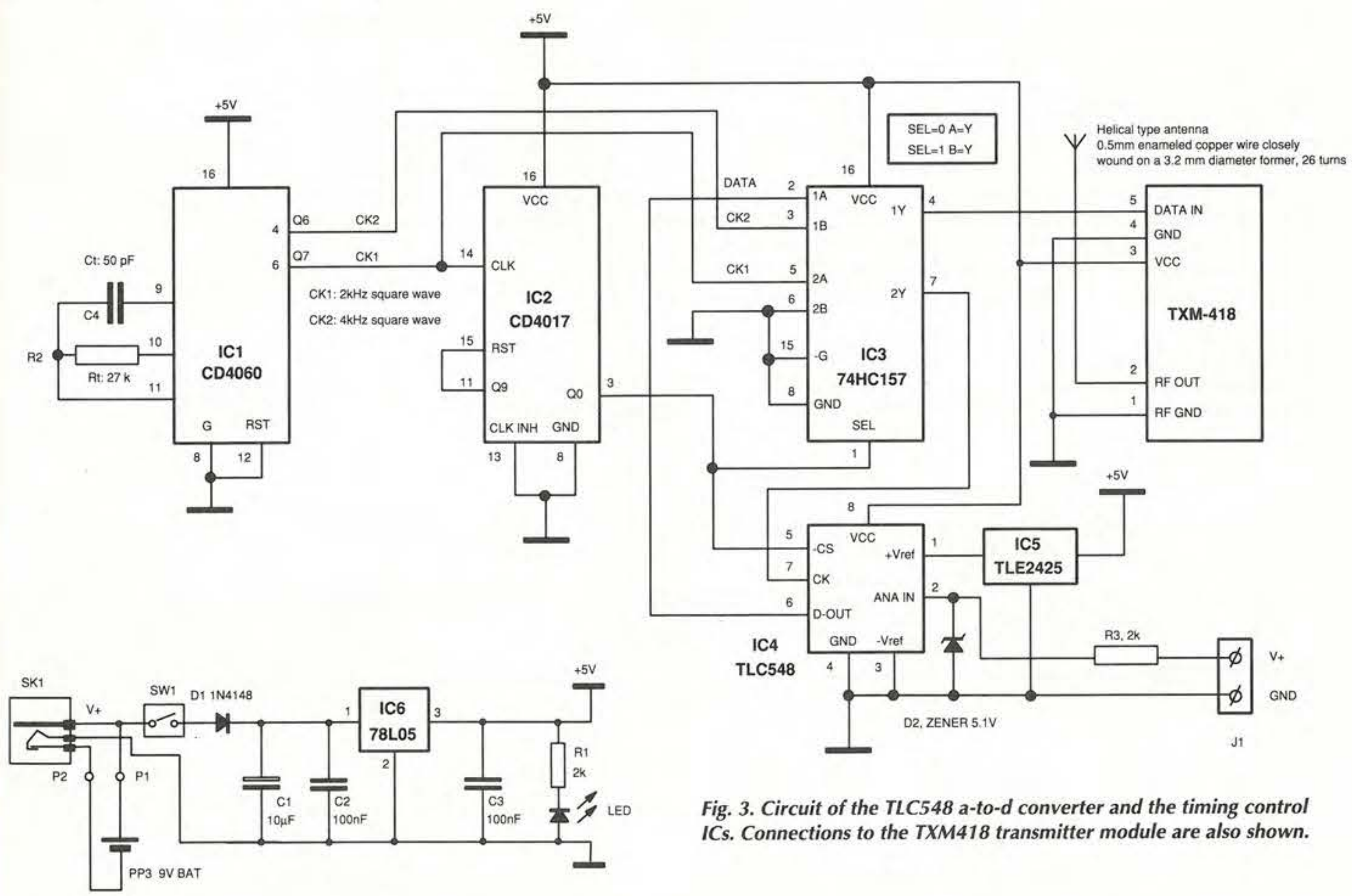


Fig. 3. Circuit of the TLC548 a-to-d converter and the timing control ICs. Connections to the TXM418 transmitter module are also shown.

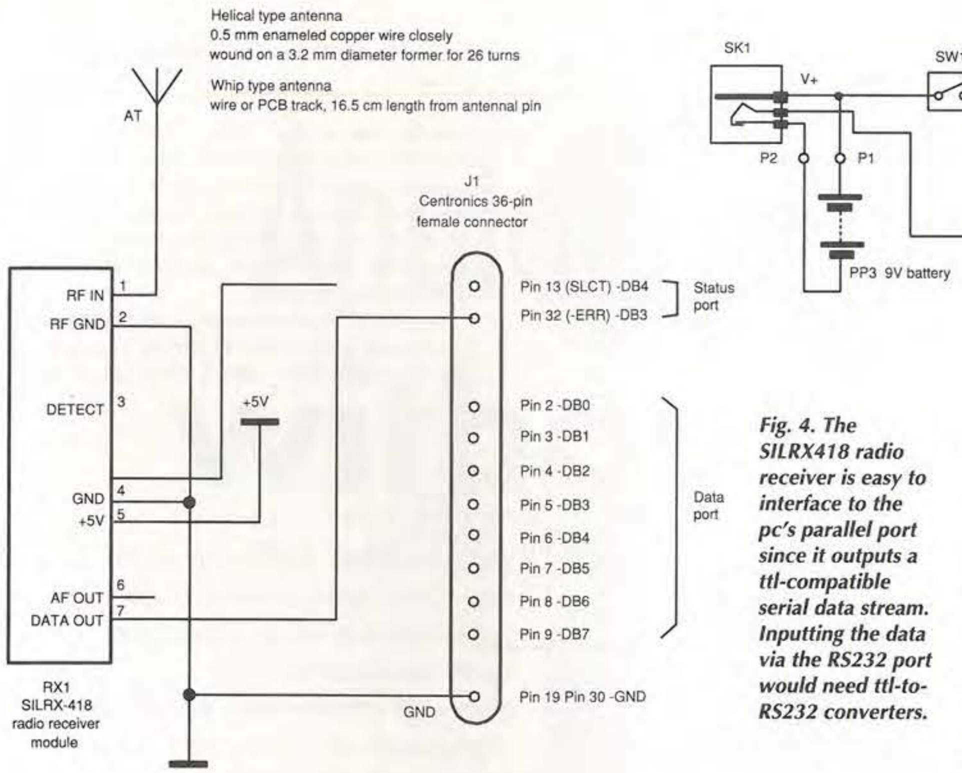


Fig. 4. The SILRX418 radio receiver is easy to interface to the pc's parallel port since it outputs a ttl-compatible serial data stream. Inputting the data via the RS232 port would need ttl-to-RS232 converters.

are connected to an external band-gap voltage reference. Normally, REF– and GND are wired together.

The serial interface of the *TLC548* consists of two ttl-compatible input lines, i/o clock input, pin 7, and chip-select input, –CS, pin 5. Data passes out of the chip via a three-state data output line, pin 6.

The system clock and the i/o clock are used independently and do not require special speed or phase relationships. This simplifies the interfacing with other circuits. Interfacing hardware and software need only initiate the conversion and read the data using the i/o clock and –CS.

The operational sequence is shown in **Fig. 5**. When –CS is high, the data output line is at high-impedance state and the i/o clock is disabled. On –CS going low, the a-to-d conversion cycle is initiated and data output begins.

To minimise errors caused by noise at the –CS input, the internal circuitry waits for two rising edges and then a falling edge of the internal system clock after a high-to-low transition is detected on the –CS pin, before it is accepted. The most-significant bit, msb, of the previous conversion result, DB_7, automatically appears on the DATA OUT pin.

Falling edges of the first four i/o clock pulses shift out DB_{6-3} of the previous conversion result on the data output pin. The on-chip sample-and-hold begins sampling an analogue input after the fourth falling edge of the i/o clock.

At this point, the falling edges of three more clock cycles shift out the remaining three data bits, DB_{2-0} of the previous conversion.

When the eighth and final clock cycle is applied to the i/o clock, the falling edge terminates the sample process and initiates the hold function.

The hold state continues for the next four internal system clock cycles. After that, an a-to-d conversion is carried out during the next 32 system clock cycles. A complete a-to-d conversion takes 36 internal system clock cycles. During the conversion, either –CS must go high or the i/o clock remains low for at least 36 system clock cycles.

Chip select can be kept low during multiple conversion, but special care must be taken to prevent noise from getting into the i/o clock. If noise does get in, the device and its external interface circuit will lose synchronisation. If –CS is taken high, it must remain high until the end of the conversion.

A valid falling edge of –CS causes the device to reset and to abort the conversion in progress.

Timing circuit. Three logic ICs are used here: a *CD4060* 14-stage ripple counter, a *CD4017* decade counter and a *74HC157* 2-to-1 data selector, **Figs 6a-c**).

The *4060*'s on-board oscillator requires a timing capacitor and a resistor, C_t and R_t. It has a 14-stage binary counter. Outputs are pins Q_{1-14}. The *4060* is used as a clock generator which generates two signals, $CK_{1,2}$.

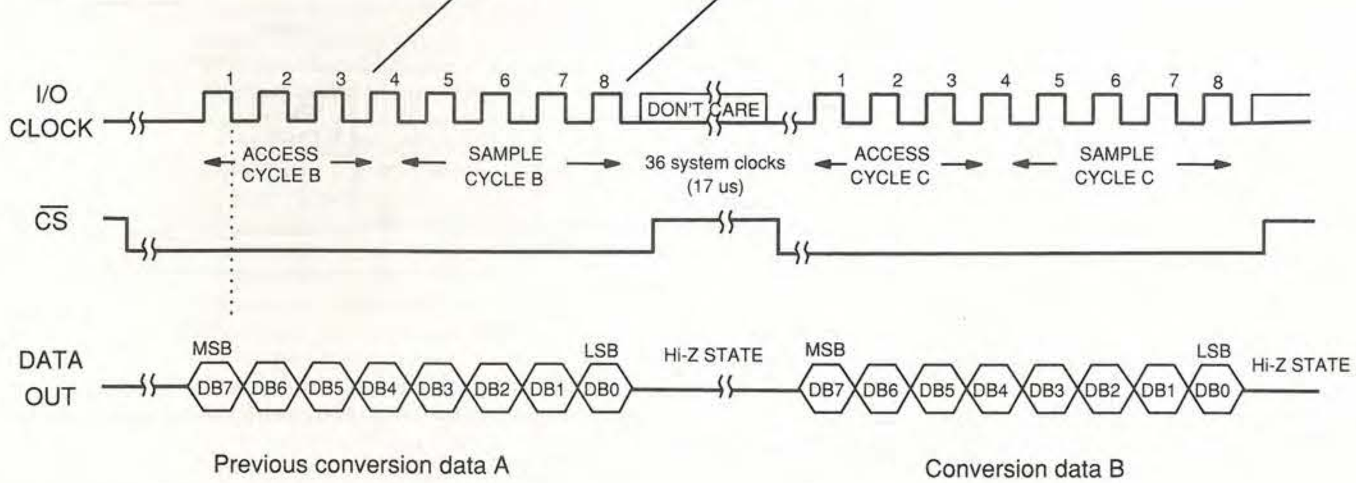
The *4017* decade counter is clocked via pin 14. Its clock-inhibit input, pin 13 must be pulled down to enable the clock. Pin 15 is the reset input and is low-to-high active. Outputs Q_{0-9} are normally low.

After a reset, Q_0 goes high and Q_{1-9} go low. At the first rising edge of the clock input, Q_0 becomes low and Q_1 becomes high. At the second rising edge of the clock, $Q_{0,1}$ both go low and Q_2 goes high. Each output therefore goes high in turn at the rising edge of the clock. On Q_9 going high, the output sequence is restarted. Connecting one of the outputs to the reset pin shortens the sequence.

For the *74HC157* quad data selector, if the select input is low, the respective A inputs feed through to the Y outputs. When SEL is high, the B inputs feed through.

Timing for this circuit is shown in **Fig. 7**. The signal from pin 6, Q_7, of IC_1 is CK_1 which is about 2kHz and the other, from pin 4, Q_6, gives CK_2 which is 4kHz. Clock CK_1 is used for timing while CK_2 is for generating the start marker for the serial a-to-d conversion data.

Fig. 5. Operating sequence of the TLC548 eight-bit a-to-d converter with serial data output.



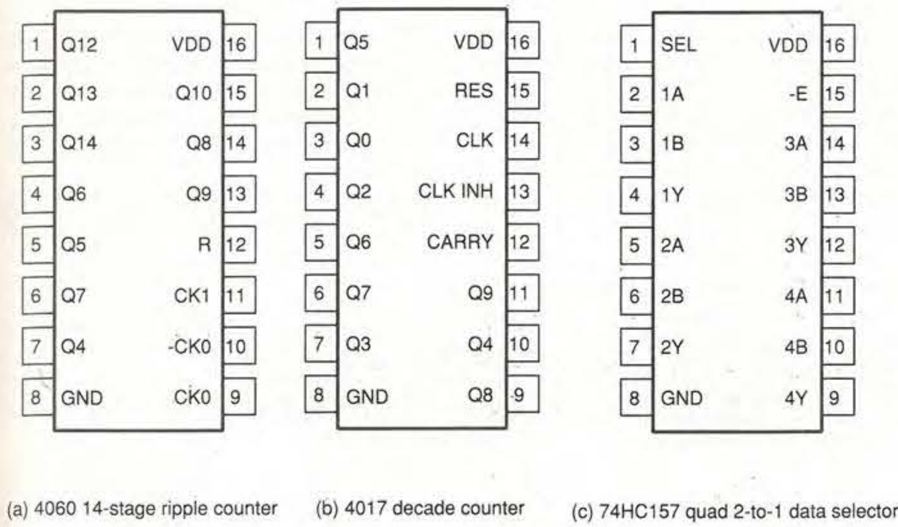


Fig. 6. Pin-outs of the logic ICs used to control the a-to-d converter.

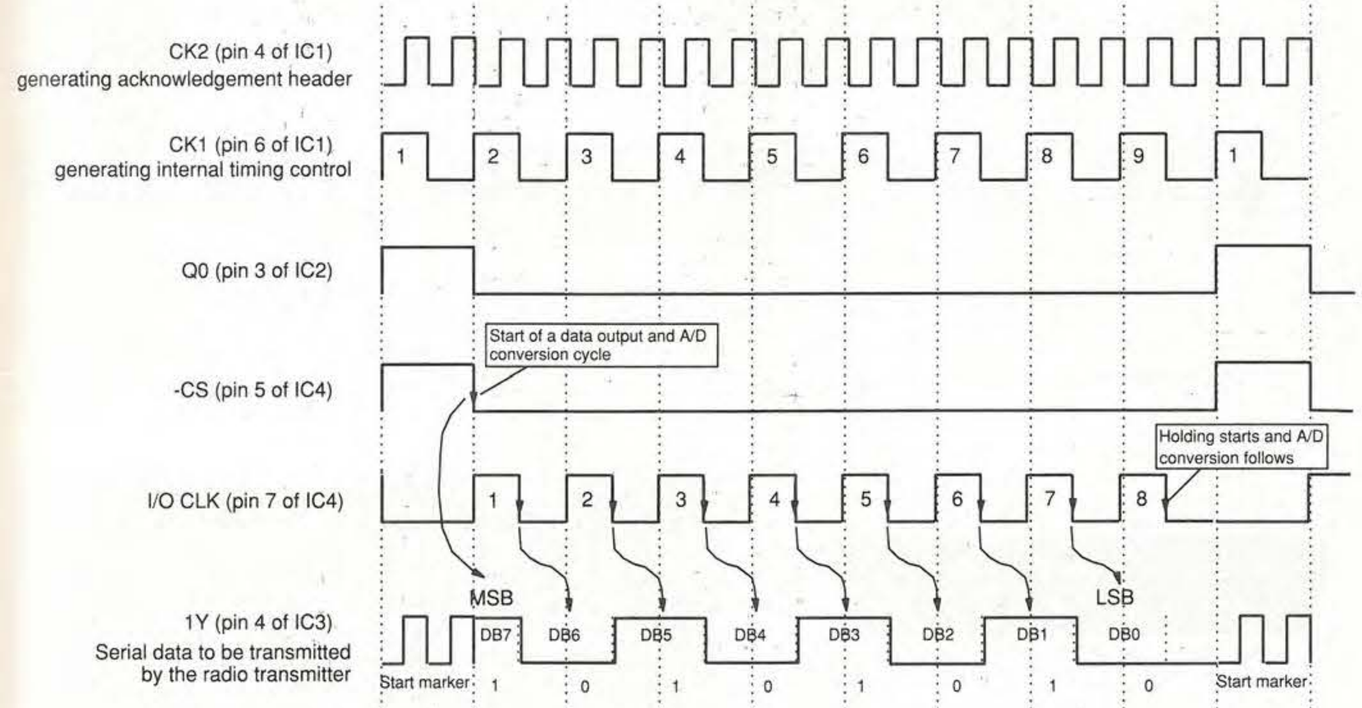
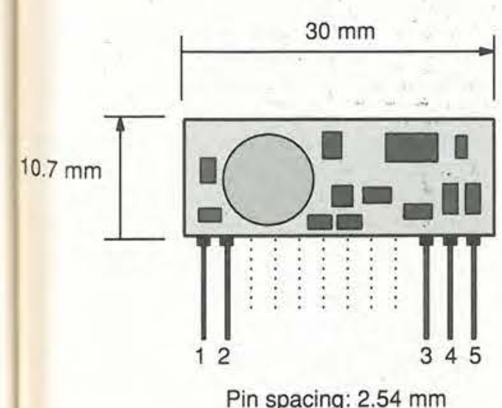


Fig. 7. Timing sequence of the radio transmitter unit.



Pin functions

Pin	Function
1	RF GND
2	RF OUT
3	VCC
4	GND
5	DATA IN

Note: Pin 1 is connected to Pin 4 internally

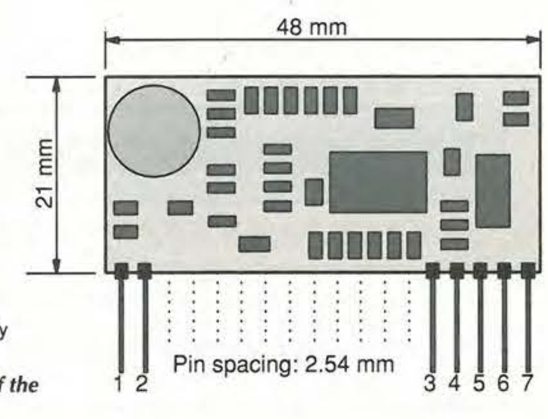


Fig. 8. Connections and dimensions of the radio transmitter and receiver.

The *4017* is configured so that output Q_0, pin 3, is normally low. For every nine low-to-high transitions of the clock applied to pin 14, it goes high for one clock period.

When Q_0 goes from low to high, the select line of IC_3 goes from low to high and this causes CK_2, connected to 1B, to feed through to 1Y of the data selector. It is then transmitted by the radio transmitter as the start marker. This flag tells the receiver the start of a valid serial data transmission.

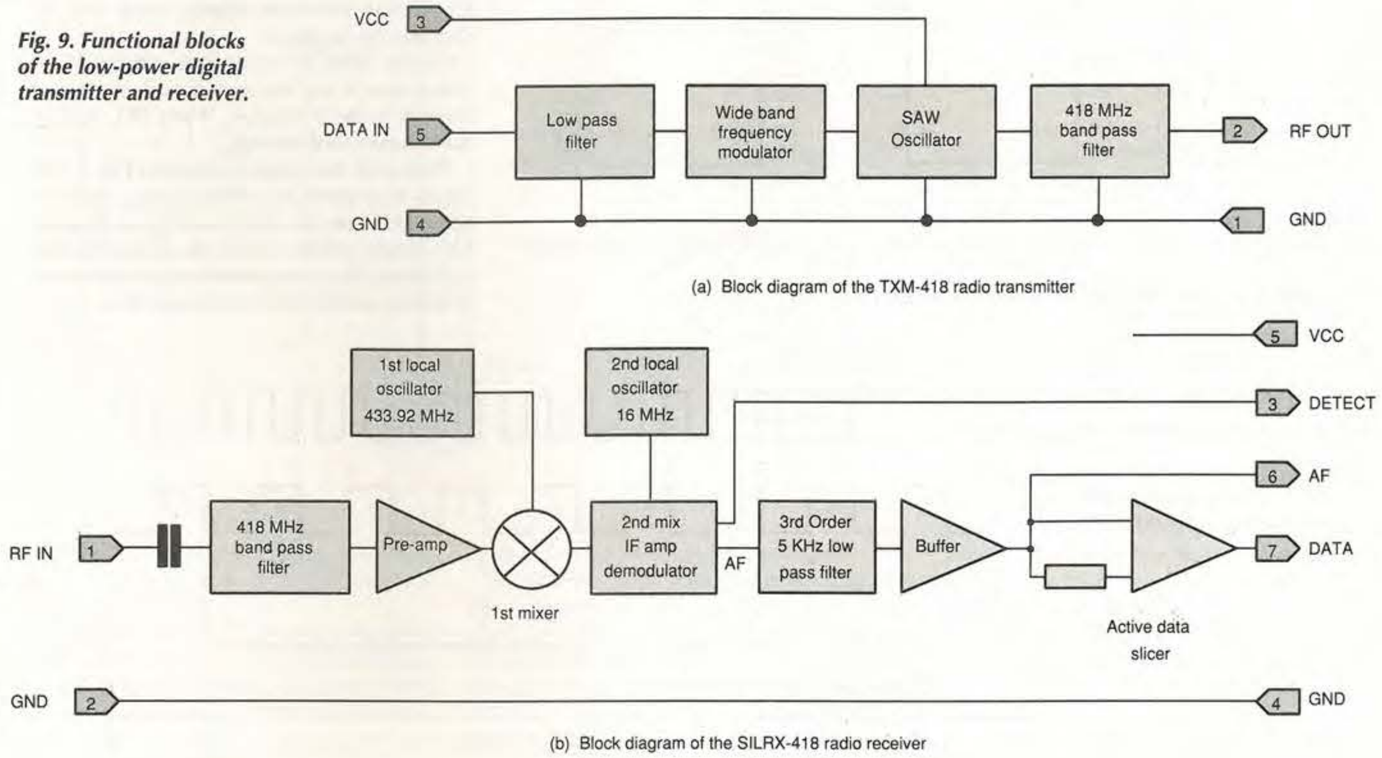
On Q_0 going from high to low, the select line on IC_3 and chip-select on IC_4 both become low. At the falling edge of chip-select on the a-to-d converter, a conversion and data output cycle begins.

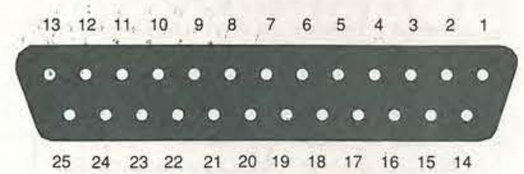
The msb of the previously converted data, DB_7, automatically appears at the data-output pin of IC_4. As the select line of IC_3 is low, 1A, which is connected to the data-out line of IC_4, feeds the radio transmitter. As a result, DB_7 is transmitted.

Output 2Y of multiplexer IC_3 feeds the i/o clock of the converter. At the falling edges of the first to seventh i/o clock, bits DB_{6-0} of the serial data are shift out by IC_4 and transmitted. At the falling edge of the eighth i/o clock, a new a-to-d conversion begins. After that, the i/o clock remains low and $\overline{CS}$ goes from low to high, stays high for one CK_1 period and then becomes low again.

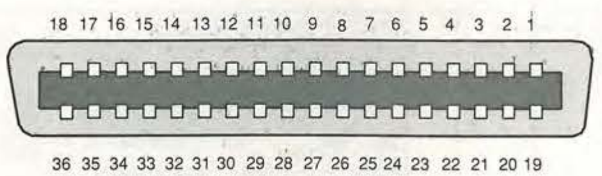
At this point, a new a-to-d conversion/data output cycle begins.

Fig. 9. Functional blocks of the low-power digital transmitter and receiver.





(a) Pin-out of the Centronics connector on pc computers – viewed from the back of the pc
Connector type: 25 pin female D-type



(b) Pin-out of the Centronics connector on printers viewed from the back of the printer
Connector type: 36 pin female Centronics-type

Connector C	P	Direction	Name	Explanation
1	1	C to P	STROBE	Strobe data
2	2	C to P	DB0	Data bit 0
3	3	C to P	DB1	Data bit 1
4	4	C to P	DB2	Data bit 2
5	5	C to P	DB3	Data bit 3
6	6	C to P	DB4	Data bit 4
7	7	C to P	DB5	Data bit 5
8	8	C to P	DB6	Data bit 6
9	9	C to P	DB7	Data bit 7
10	10	P to C	ACK	Data received flag
11	11	P to C	BUSY	Printer busy flag
12	12	P to C	PE	Paper empt flagy
13	13	P to C	SLCT	Printer on line flag
14	14	C to P	LF/CR	Auto LF after CR
15	32	P to C	ERROR	Printer error flag
16	31	C to P	INITIALISE	Initialize printer
17	36	C to P	SLIN	Select/deselect printer
18-25	19-30 and 33		GND	Twisted-pair rtn GND
	18,34		Unused	
	16		Logic GND	Logic ground
	17		Chasis GND	Chasis ground

C = computer P = printer

Fig. 10. Pin functions of the Centronics port connectors

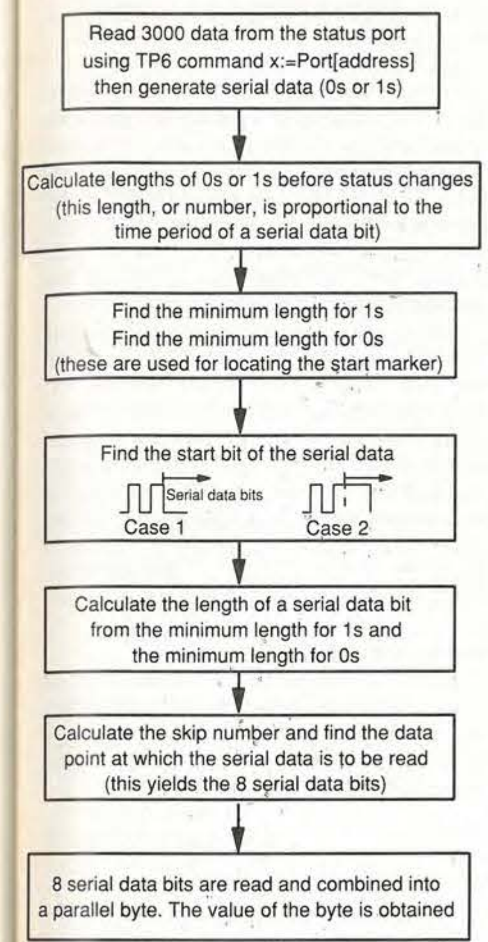


Fig. 11. Flow of the software needed to obtain the eight bits from the a-to-d conversion.

Radio transmitter module. Physical dimensions and pin functions of the *TXM-418* transmitter are given in **Fig. 8a)**. Pins 1 and 4 are connected together internally and form the ground.

The transmitter operates over a wide range from 6 to 12V dc. It consumes typically 6mA at 6V or 14mA at 12V. Its data-modulation input – which requires a c-mos logic level at the same power supply voltage – is applied to pin 5. An antenna connects to pin 2.

Figure 9a) is a block diagram of the module. Digital data at c-mos logic levels is supplied to the data-in line, pin 5. The data first passes through an *RC* low-pass filter, which restricts the bandwidth of the modulation signal to 10kHz at the –3dB point; 10kHz represents the upper limit of the input data frequency.

Filtered data then feeds a wideband frequency modulator which accepts signals of frequencies from dc to 10kHz. The modulator drives a varicap diode whose changing capacitance is used to modify the frequency of the next stage – a radio frequency oscillator.

Centre frequency of the oscillator is accurately set by a surface acoustic wave resonator in the 418MHz band, i.e. 417.90 to 418MHz. The oscillator has a 418MHz band-pass filter to ensure that any spurious emission out of the band is within the limits as specified by the MPT1340 specification. Final filtered rf output appears on pin 2.

The transmitter antenna can have one of three forms – helical, loop or whip. I used the helical form for this design.

Radiocommunication Authority *MPT1340*. The radio transmitter module is type-approved to MPT1340, making it licence exempt for use within the UK. Its applications are specified as telemetry, telecommand and in-building security.

When using the module in a customised application, the requirements issued by the Radiocommunication Authority must be satisfied.

Receiver details

The circuit diagram of the radio receiver unit is given in Fig. 4. The heart of this unit is the radio receiver module *SILRX-418*.

Radio receiver module. Physical dimensions and pin functions of the *SILRX-418* receiver are shown in Fig. **8b)**. The module operates from a 4.5V to 9V dc rail and draws typically 13mA.

Figure **9b** is a block diagram of the module. Incoming radio frequency from the antenna goes to a 418MHz band-pass filter via a capacitor. An rf preamplifier boosts the signal before it enters the first mixer stage.

The first local oscillator runs at a frequency of 433.92MHz, which is produced again by a surface acoustic wave resonator. This signal is mixed with the received 418MHz signal to produce the first intermediate frequency signal at 15.92MHz.

The mixed signal then feeds the second mixer, where a second local oscillator running at 16MHz produces the final intermediate frequency at 80kHz. Final IF is now amplified and demodulated to produce an audio frequency signal. A carrier detect signal is also produced.

To improve the signal-to-noise performance, the audio signal is processed by a third-order low-pass filter. This signal feeds an audio buffer whose output, pin 6, is at half of the power supply. The signal also passes through a data slicer, where the analogue audio signal is converted into a digital signal and is output from pin 7. Logic output is at c-mos levels.

Any of the antenna types previously described in the transmitter section can be used with this module. Criteria for the receiver antenna under MPT 1340 are not as restrictive as those that apply to the transmitter.

As an alternative to the integral antenna as used for the transmitter, you are permitted to use an external arrangement connected by a coaxial feeder.

If you want to optimise the range of the system, other types of antenna may be used. I used a whip type antenna.

Table 1. Input/output addresses of the data, status and control ports.

CPU i/o address	Data port (pc to prn)	Control port (pc to prn)	Status port (prn to pc)
LPT1 to LPT4	Base address	Base address+2	Base address+1

Table 2. Base addresses for LPT1 and LPT2.

	Base address
LPT1 with mono card	956D, (3BCH)
LPT1 with i/o card	888D, (378H)
LPT2 with i/o card	632D, (278H)

Table 3. Memory locations for each of the four printer LPT addresses.

Port	Memory address
LPT1	0000:0408 – 0000:0409
LPT2	0000:040A – 0000:040B
LPT3	0000:040C – 0000:040D
LPT4	0000:040E – 0000:040E

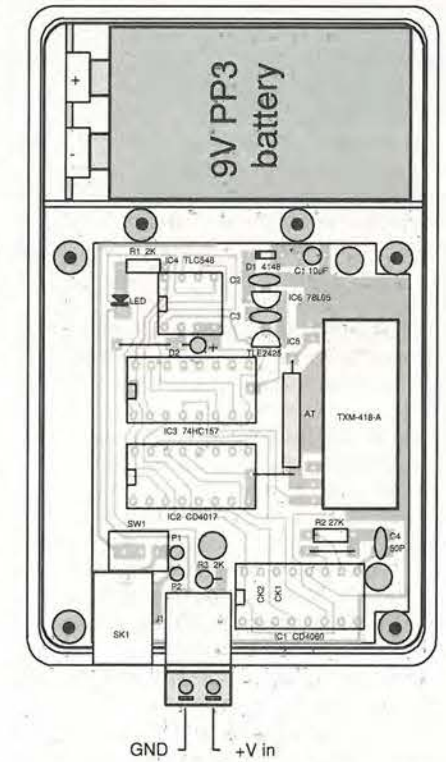


Fig. 12. With careful pcb layout, the entire battery-operated transmitter unit can be made to fit into a small ABS box.

Reading in the data

The receiver has only one line from which the received serial data is output. The data contains the start marker and the serial data, Fig. 7. Consequently, the computer only needs one input line to read the data.

There are several hardware solutions for reading the line into the computer. One is to use the Centronics port. Because all pcs have one or two Centronics ports, this is a universal solution.

As the signal output from the radio receiver module has a ttl/c-mos logic, the Centronics port can be interfaced to the radio receiver directly. An alternative is to use an RS232 port. Again, this is a universal solution, but ttl/c-mos-to-RS232 level converters should be used. A third alternative is the game port.

This system relies on the Centronics port and is powered by external batteries.

The Centronics port

The Centronics interface, also referred to as the printer, LPT or parallel port, is an industry-standard interface designed originally for interfacing with printers. Whether housed in a desk-top case or as a laptop, a pc will have at least one such interface.

If you have a desk-top pc, you can add up to three printer ports using plug-in cards. These ports are given the logical names LPT1 to LPT4.

The connector at the back of the computer and the one on the printer differ, **Fig. 10**. The one on the computer is a 25-pin D-type female connector, Fig. **10a**), while the one on the printer is a 36-pin female Centronics-type connector, Fig. **10b**). Although the connectors are different, their functions are the same.

The i/o lines in the Centronics port are organised into three groups – those handling data, those carrying out control and those signalling status.

Data is sent from the pc to the printer over eight latched lines, DB_{0-7}. Data flow is controlled by an i/o port of the microprocessor inside the pc. The control port controls the operation of the printer. It contains four latched lines, namely -STROBE, -LF/CR, -SLIN and -INITIALISE. These are directed from the pc to the printer. The group is controlled by an output port of the processor. Note that -STROBE, -LF/CR and -SLIN lines are inverted but -INITIALISE is not.

The status group is used by the pc to obtain current status of the printer. It contains five lines, -ERROR, SLCT, PE, -ACK and BUSY. These are directed from the printer to the computer. This group is controlled by an input port of the processor. The busy line is inverted but the other four are not.

Port i/o addresses. The computer addresses of the DATA, CONTROL and STATUS ports can be calculated using the expressions in **Table 1**. You will see that the printer address is the same as the address of the data port. The base addresses for LPT1 and LPT2 are listed in **Table 2**.

Once the base address of the printer port of a pc is known, the addresses of the control and status ports can be calculated, Table 1. There are two ways to obtain this address. One is to check your computer's user manual or to watch the screen carefully after the computer is powered on or reset. A table showing hardware specifications will appear on the screen.

The other way is a convenient one. When the computer is powered on or reset, the bios (basic input/output system) checks all the possible printer addresses. If it finds an installed Centronics port, it writes the addresses of the ports to a special memory location.

For LPT1, this address – a two-byte word – is stored at $0000{:}0408_{16}$ and $0000{:}0409_{16}$. By peeking this memory location, the base address can be obtained. The memory locations for LPT1 to LPT4 are listed in **Table 3**.

Another useful one-byte memory location is $0000{:}4011_{16}$. It stores the total number of installed Centronics interfaces. This information is contained in $DB_{6,7}$ bits, **Table 4**.

Controlling the parallel port

There are three methods to control the Centronics port. The first method is to use printer operation procedures. For example, in Basic, this instruction is 'PRINT'. In Turbo Pascal, it is 'WRITELN(LST)'.

When executing such instructions, the data, control and status groups operate together and can not be used individually. This method is only useful for printer operations. Another method, also used for printer operations, relies on bios interrupt INT 17h.

The third method involves direct i/o access.

Table 4. Location $0000{:}4011_{16}$ tells you how many printer ports the bios thinks you have installed.

DB7	DB6	No of ports installed
0	0	0
0	1	1
1	0	2
1	1	3

This method controls the data, control and status ports separately using direct i/o access instructions. In this case, the Centronics port is treated as three separate i/o ports: two of these are output ports and one is an input. Imagine, for example, controlling LPT1. Assuming that the addresses of the data, control and status ports are 888D, 890D and 889D, respectively, to send data to the data and the control ports, the following procedures are used, in Basic,

```
OUT 888, x and OUT 890, x
```

And in Turbo Pascal,

```
PORT[888]:=x and PORT[890]:=x
```

in which x is the output value in decimal. To read data from the status port, the following procedures are used. In Basic,

```
Y=INP[889]
```

And in Turbo Pascal,

```
Y:=PORT[889]
```

in which Y is the decimal value of the input data.

The software driver. The software of this radio data logger is written in Turbo Pascal 6, **List 1**. In the program, the procedure, *Input_printer_address*, finds the number of Centronics ports installed on your pc, base addresses of these ports and allows you to select a Centronics port.

Radio a-to-d conversion data is read by the function *inputdata:real*. First it reads 3000 serial data from the status port in one go. The serial bits of the a-to-d conversion results are contained in these data and the reminder of the function extracts the useful bits from these 3000 data. The function locates the position of the serial a-to-d conversion data by finding the position of the start marker.

Next, the program skips a certain number of data and reads the serial data. The serial data are finally combined into parallel data, **Fig. 11**. The *inputdata:real* function is repeated three times. Valid received data must satisfy that the square-root error of the three consecutive readings is within a certain value. The valid data is the averaged value of the three readings. If this condition is not satisfied, the above-mentioned procedure is repeated again. The function, *AD_conversion:real*, is used for this purpose.

Implementing the logger

My prototype transmitter unit is constructed on a single-sided pcb and is housed in a small box, **Fig. 12**. The receiver unit is constructed on a piece of strip board, as the circuit is so simple, and is also housed in a small box.

Continued on page 82

Technical support

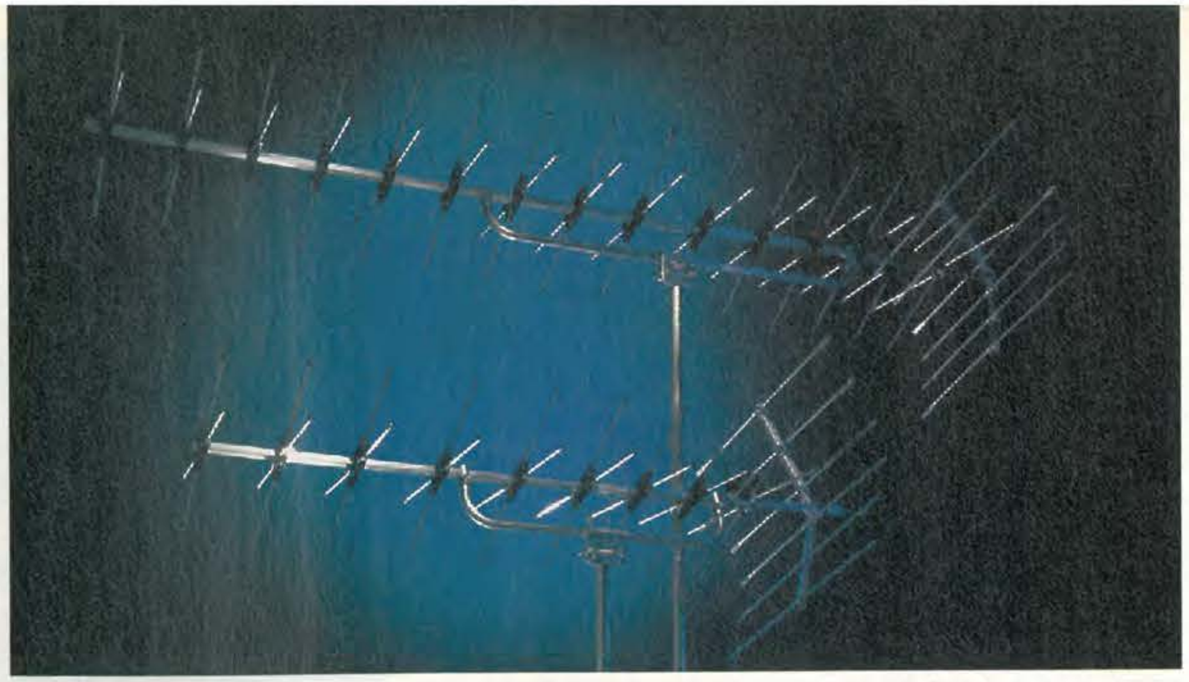
A designer's kit is available from Pei. It consists of all the hardware – pcb plus components – to allow designers to construct one a-to-d converter/radio transmitter unit and one Centronics-port receiver unit.

Source code in Turbo Pascal 6 and EXE file are provided on a 3.5in floppy disk. Please direct your enquiry to Dr Pei An, 11 Sandpiper Drive, Stockport, Manchester SK3 8UL. UK tel/fax/answer: +44-(0)161-477-9583. E-mail: pan@fs1.eng.man.ac.uk.

Tx/Rx modules

The transmitter and receiver modules mentioned in this article are available from Radio-Tech at Overbridge House, Weald Hall Lane, Thornwood Common, Epping, Essex CM16 6NB, tel. 01992 576107, fax 01992 561994, e-mial radtec@radtec.demon.co.uk.

More details of the transmitter and receiver pair were given in the June 1996 issue on page 454.



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Antennas from the ground up

Working from the half-wave dipole to active designs, Ian Hickman explains how antennas function, visiting Yagi and loop aerials along the way.

Appropriately, for a magazine formerly called *Wireless World*, readers are presented from time to time with articles on various aspects of wireless communications – including ones on the crucial topic of antennas. Since such articles have mainly dealt with specific developments in antenna technology,[1] the time is ripe for a broad look at antennas of all types, and their basic principles of operation.

Among other things, this article clarifies some initially puzzling points regarding antennas, such as why communication over a path in free space should require a hundred times as much transmitted power at 100MHz as at 10MHz.

Antenna basics

When considering antennas, the theoretical yardstick is the ideal isotropic antenna. If fed from a transmitter, this antenna radiates power with equal intensity in all directions.

As with any antenna, there will be associated electric and magnetic field lines in space, due to the voltage applied to the antenna elements, and the current flowing in them. In the immediate vicinity of the antenna, these fields are out of phase, and thus represent stored circulating energy: this is called the 'near field' region. This field drops off rapidly with distance, being already small at a distance of λ metres, i.e. one wavelength, from the antenna.

But the antenna also generates electric and magnetic fields which are *in* phase, and these represent a flow of power away from the antenna – the 'far field'. The magnetic field strength corresponds to a magnetising field, measured in amps per metre. This, together with the electric field strength in volts per metre defines the characteristic impedance of free space, which is 120π or 377Ω.

The ideal isotropic antenna looks purely resistive, and as **Fig. 1** shows, this resistance has two components. Resistance R_r is the 'radiation resistance' – a notional non-dissipating resistance representing the 'port' via which power is radiated from the antenna. Loss resistance R_l is the ohmic component of the antenna's total resistance. Clearly the radiation efficiency η_r is given by,

$$\eta_r = \frac{R_r}{R_r + R_l} \qquad (1)$$

In some cases, it is possible to make R_l negligible, but in practice an efficiency well short of 100% must sometimes be tolerated.

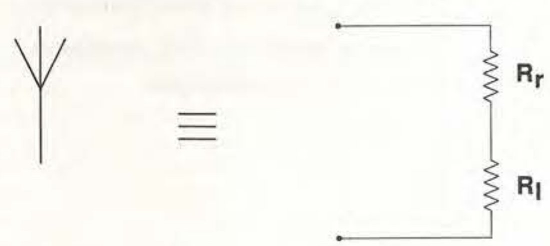
Figure 2 shows an ideal isotropic antenna located in free space, radiating a power P_t. Resulting power density D, in watts per square metre, at a range d, in metres, is given by,

$$D = \frac{P_t}{4\pi d^2} \qquad (2)$$

This assumes that d is much larger than the wavelength concerned, and is of course independent of the frequency. The term $4\pi d^2$ is the surface area of a sphere of radius d, centred on the antenna. The strength of the electric field ε, in volts per metre, in space at any point is given by,

$$\varepsilon = \sqrt{377D} \qquad (3)$$

But if D depends only on P_t and the distance, why does communication over a given distance in free space require a hundred times as much transmitted power at 100MHz as at 10MHz? After all, surely your 1W at 100MHz is as good as the other person's 1W at 10MHz?



At resonance, a tuned antenna looks resistive.

Fig. 1. Loss resistance R_l turns some of the transmitter power output into heat. Radiation resistance R_r does not – it radiates it into space.

Diminishing returns

The answer is all down to the receiving antenna. Consider an ideal isotropic antenna identical to that of Fig. 2 – i.e. also appropriate to the frequency concerned – immersed in the field given by expression (2). It will pick up an amount of power determined by its effective area, known as its 'effective aperture'. The effective aperture A, in square metres, of an isotropic antenna suitable for operating at a frequency fMHz is given by,

$$A = \frac{\lambda^2}{4\pi} \qquad (4)$$

where λ=300/f metres. Since the aperture of an antenna is inherently proportional to the square of the wavelength, in any given field strength, an antenna of any type operating at half the frequency of another of that type will pick up four times as much energy.

Combining (2) and (4), for isotropic transmit and receive antennas, received power P_r is simply D multiplied by A and is given by,

$$P_r = \frac{P_t}{4\pi d^2} \times \frac{\lambda^2}{4\pi} = \frac{P_t\lambda^2}{(4\pi d)^2} \qquad (5)$$

This defines the basic free-space inverse square law for range, i.e. 6dB extra loss for every octave – or doubling – of the range.

It is not possible to construct a simple antenna with an isotropic radiation pattern, so it's time to look at real antennas. There are only as few basic types, but each is capable of various developments and refinements for special purposes. I'll start with the basic types, and move on to the variations later.

Half-wave dipole

The simplest practical antenna is the half-wave dipole. Imagine the last quarter wavelength of an open-circuited balanced transmission line opened out at right angles, to form a tee shape.

Normally, at a point one quarter of a wavelength back from an open circuit, a transmission line would look like a short circuit. But because the opened out arms radiate power, they look like a finite resistance. This turns out to be about 73Ω, so if the half-wave dipole were connected to a balanced feeder of this impedance, all of the incident power would be absorbed and radiated into space.

The absence of reflected power means that the voltage-standing-wave ratio, or vswr, on the line would be 1:1 or unity, a perfect match for the transmitter.

As a receiving antenna, the half-wave dipole may be modelled as a 73Ω source. In accordance with the maximum-power theorem, when connected to a matched load, half the power the antenna picks up from a distant source is delivered to the load; the other half is 'dissipated' internally in the source. But the 73Ω source is the radiation resistance, which unlike a physical resistance, does not turn incident energy into heat. The half of the picked-up power not delivered to the load is reradiated with the usual radiation pattern of a dipole, described below.

So the field distribution in the immediate vicinity of the dipole is the resultant of the incident and reradiated fields. Note that if the centre of the dipole had been short circuited, instead of being connected to a matched feeder and load, all of the power picked up would have been reradiated, locally modifying the field.

The left hand side of **Fig. 3** shows a half-wave dipole, with the current distribution along its length a maximum at the centre, as indicated. The shape of the distribution is a half sine wave, and the shape of the voltage distribution a half cosine wave, i.e. zero at the centre and maximum – in antiphase – at each end.

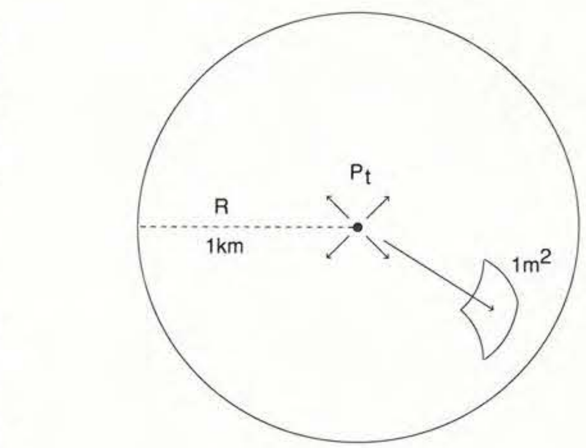


Fig. 2. An ideal isotropic antenna radiates energy into space with equal intensity in every direction.

Note that this diagram is not to scale.

$$\text{Power density D at range R} = \frac{P_t}{4\pi R^2}\ \text{W/m}^2$$

$$\text{e.g. if } P_t = 1\text{W, then at R} = 1\text{km, } D = \frac{1}{4\pi 1000^2} = 0.078\,\mu\text{W/m}^2 = -71\text{dBW/m}^2$$

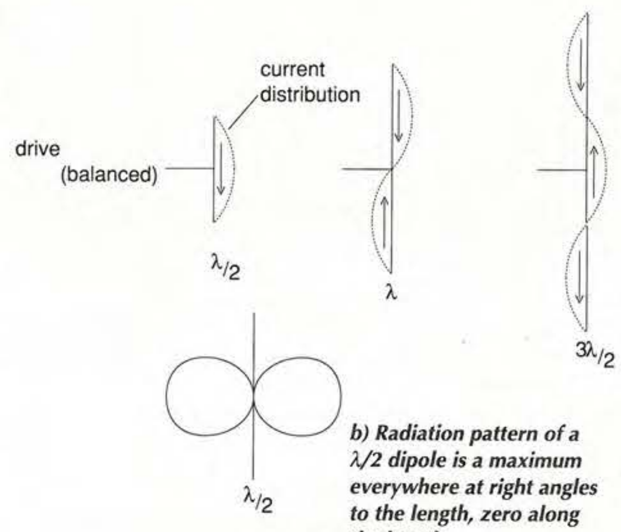


Fig. 3a) Current distribution along a half-wave, or λ/2, dipole, left, is maximum at the feed point and zero at the ends – the very opposite of the voltage distribution. Also shown, the current distribution on dipoles of length λ and 3λ/2.

b) Radiation pattern of a λ/2 dipole is a maximum everywhere at right angles to the length, zero along the length.

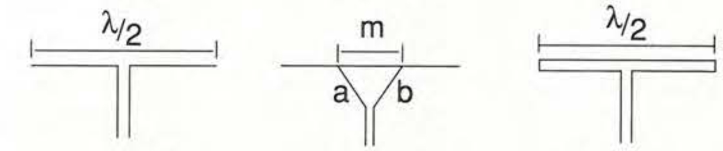


Fig. 4a) A λ/2 dipole presents a source or load impedance of 73Ω resistive, balanced. b) It may be connected without mismatch to a 300Ω balanced feeder using a 'Δ-match'. Feedpoint separation m and length of the sloping members a, b are parameters which may adjusted to suit the particular dipole length to diameter ratio and feeder impedance. c) Impedance of a folded dipole is four times that of the basic dipole, providing a good match to a 300Ω feeder.

With the feeder connected at a point of current maximum and voltage minimum, the impedance is much lower than the 377Ω characteristic impedance of free space. Shown at the bottom of Fig. 3 is the radiation pattern, which is zero along the length of the dipole. At right angles to the dipole, it is at a maximum everywhere, i.e. in three dimensions it exhibits a toroidal or 'doughnut' shape.

Since the power is concentrated mainly at right angles to the dipole, with zero radiation off the ends, the maximum field is greater than it would be for an isotropic antenna fed with the same power. As a receiving antenna, the same pat-

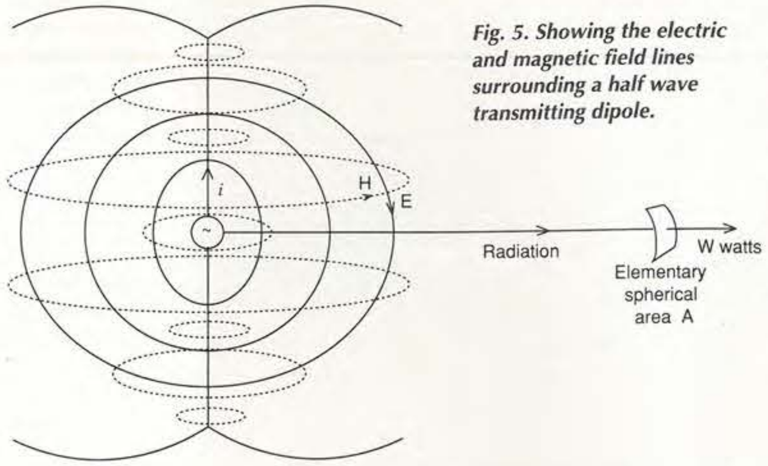


Fig. 5. Showing the electric and magnetic field lines surrounding a half wave transmitting dipole.

tern applies – maximum sensitivity to signals from any direction at right angles to the dipole, zero sensitivity to signals arriving end-on.

The half-wave dipole antenna's effective aperture A is given by,

$$A = \frac{G\lambda^2}{4\pi} \qquad (5)$$

where G is the power gain relative to isotropic; it follows that G is a function of the orientation of the antenna relative to the incident field. In the maximum direction, anywhere at right angles to the dipole, G is 1.65.

Thus, a half wave dipole exhibits a maximum gain of 10log(1.65) or +2.15dB relative to isotropic. Any logarithms mentioned from here on are to base 10 by the way.

Note that for a half-wave dipole, the 'figure-of-eight' cross section of the toroid is not like two circles. For an electrically short dipole – length much less than $\lambda/2$ – the cross section is circular, but such an antenna is not resonant. Even if brought to resonance by tuning out the reactance, the radiation resistance R_r is very small – often much smaller than the loss resistance R_l.

Short dipoles have their uses

So an electrically short dipole does not make a good radiator, or even an efficient receiving antenna. But it is useful as an 'E-field probe', if connected to an amplifier with a very high input impedance.

With such an open-circuit load, the antenna outputs a voltage given by $E_{oc}=l_e \times \varepsilon$, where ε is the field strength in volts per metre and l_e is the effective length of the dipole. If the physical length of the dipole is much less than λ, then the effective length equals the physical length.

Two crossed half-wave dipoles will receive signals from all directions, while three such dipoles, oriented along the three orthogonal axes and the signals again suitably combined, can effectively provide an isotropic pattern. This scheme is in fact used in special instrumentation antennas for emc measurements.

Figure 3 also shows the current distribution on dipoles of length λ and $3\lambda/2$. Here, the feedpoint impedances are respectively high and low relative to the impedance of free space. The direction of maximum radiation is at 54° and 42° to the dipole respectively, as against 90° for the half wave dipole.

Figure 4 shows three ways of connecting a dipole to a feeder. In a), the dipole impedance of 73Ω is about right to connect to a 75Ω coaxial cable, such as a tv down lead. But the latter is unbalanced, whereas the antenna is balanced.

A balun – i.e. a balanced-to-unbalanced transformer – can be used, but it is frequently omitted. The balanced-to-unbalanced conversion then takes place along the first few wavelengths of the feeder, which acts as its own balun.

With some signals, for example Band II broadcast fm, a 300Ω balanced feeder is commonly used. This can be matched to the lower impedance of a half-wave dipole using the delta match shown in Fig **4b)**. Alternatively, the folded dipole of Figure **4c)** can be used. The two closely-spaced dipoles act as a 2:1 ratio – or 4:1 impedance ratio – transformer, transforming the 73Ω dipole impedance to 292Ω.

Quarter-wave whip

Probably the next most important antenna is the quarter-wave monopole or whip. To illustrate its mode of operation, consider first a dipole.

Figure 5 shows the electric and magnetic fields E and H in the vicinity of a half wave dipole. In space, the fields are distributed as shown, being everywhere mutually orthogonal, i.e. at right angles to each other.

In time, the fields in the immediate vicinity of the antenna – the near field – are in quadrature, as noted earlier. This means that they represent stored energy as in a tuned circuit; a half-wave dipole antenna is resonant. In the far field, being in phase, they represent a flow of real power away from the antenna.

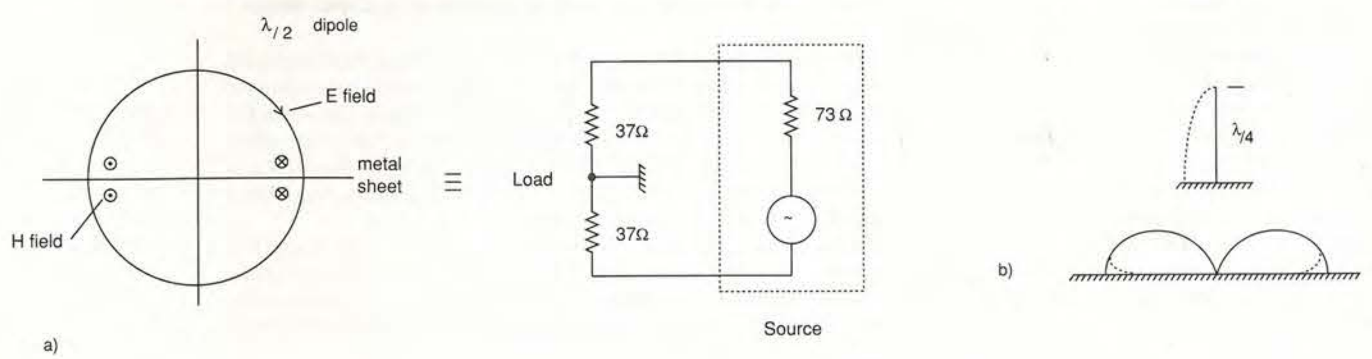
As I mentioned earlier, the impedance of the dipole is 73Ω balanced. Now imagine a flat sheet of copper, of infinite extent, inserted between the two halves of the dipole, as illustrated at the left hand side of **Fig. 6**.

The magnetic flux lines do not cut the conducting sheet anywhere: they are completely unaffected by its presence. Likewise, the electric lines are unaffected, because they meet the sheet everywhere at right angles. Thus the antenna behaves as two separate antennas, one radiating half the energy provided into the upper half sphere, the other into the lower, as in the right hand side of Fig. **6a)**.

Thus a vertical whip antenna over an ideal ground plane presents an input impedance of 37Ω, and has a radiation pattern which is omnidirectional in the horizontal plane. The pattern in the vertical plane is a doughnut sliced in half, being simply the upper half of the pattern of the $\lambda/2$ dipole in Fig. 3.

Since an ideal suitably matched vertical quarter-wave whip

Fig. 6. Showing how the performance of a λ/4 whip antenna over a ground plane is derived from that of a λ/2 dipole.



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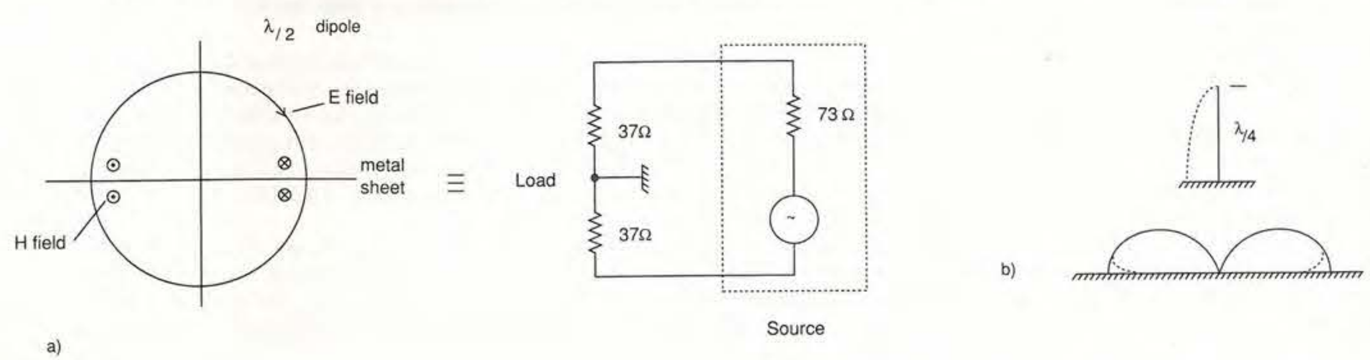
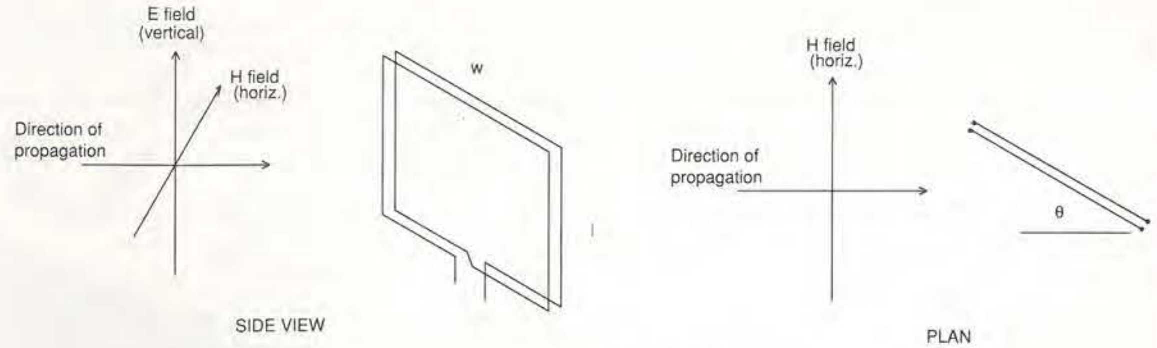
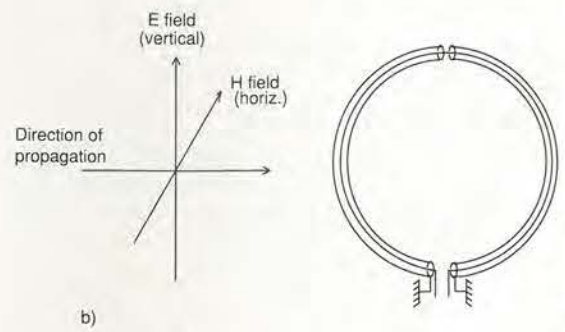


Fig.7a) Illustrating the performance of the frame or loop aerial.



b) A screened loop provides improved rejection of unwanted electrostatic pick-up.



antenna radiates all the energy in the upper half sphere, it provides a maximum directional gain – in the horizontal direction – of 3dB relative to a half-wave dipole, or 5.15dB relative to isotropic.

In practice, the ground-plane is not perfect. Where this is the case, the radiation pattern does not extend down to the horizontal. The direction of maximum radiation is squinted slightly upwards, as represented by the dotted lines in the lower portion of Fig. 6.

In fact, the ground-plane is often very imperfect, being simply the rest of the transceiver, as in the case of a mobile telephone.

The loop antenna

In addition to the half-wave dipole and the quarter-wave whip, the remaining basic antenna type is the loop. **Figure 7a)** shows a vertical loop of height *h* and width *w*, facing a vertically polarised field at an angle θ – less than 90° – as shown in the plan view.

If the vertical electric field is *E* volts/m, then a voltage *E*×*h* will be induced in each vertical conductor. The voltage induced in a vertical conductor nearer the transmitter will lead that in one further, by a phase angle,

$$\phi = \frac{2\pi}{\lambda} w \cos\theta \text{ radians}$$

corresponding to the time for the wavefront to travel the extra distance *w*cosθ metres. If this distance is small compared with λ, the two voltages, which act in opposite directions round the loop, will almost cancel out. The net resultant per turn will be approximately,

$$Eh\phi = \frac{Eh2\pi w\cos\theta}{\lambda} = \frac{EA2\pi\cos\theta}{\lambda}$$

where *A* is the area enclosed the loop.

The results quoted are only valid for the case where *w* and *h* are both small compared to λ. If they are not, standing waves are set up on the conductors, complicating the theory.

More half-wave dipole bandwidth

Many variations on the basic half-wave dipole have been developed.

One disadvantage of this type of antenna – called a 'doublet' in the USA – is its comparatively narrow bandwidth. This can be increased by using fat conductors for the two arms; these may be tubes or parallel wires supported on circular spreaders. In this case, the 'end effect' is increased.

End effect means that the optimum half wave dipole is actually a percent or two shorter than λ/2. If *h* is the dipole half-length and *a* the cross section area of the tube, then for values of *h*/*a* less than 100, the shortening required may exceed 10% or thereabouts. But this figure depends strongly on how the tubes are tapered down to connect to the feed-point.

Yet more bandwidth

If you need a greater bandwidth than can be obtained in this way, resistive loading can be added to the antenna. This increases loss resistance R_l, the value of which will vary with frequency.

The 'Australian dipole' is an example of a loaded dipole, see **Fig. 8**. This is popular with amateur radio enthusiasts as it permits operation on all the hf bands. In a measurements application, involving receiving only, efficiency is not quite so important, and more complicated distributed loading regimes may be adopted.[1]

Similarly, measures can be adopted to reduce the overall dimensions of a whip antenna. The length of a quarter wave whip may be reduced, leaving it looking capacitive, and a base 'loading coil' added to bring it back to resonance.

Alternatively, a loading inductor may be mounted half way up the shortened whip, or top capacity loading may be added. A combination of these two is also an option.

In one popular scheme, the conductor is coiled over a suitable – often flexible – non-conducting armature. Capacitance at the base may be used to bring the antenna to resonance. All these arrangement reduce the bandwidth of the antenna, by raising its *Q*. They also reduce the the radiation efficiency, by increasing the ratio of R_l to R_r.

For reception only, a short top-loaded whip can form an active antenna, operating as an *E*-field probe. The output is connected to a high-gain preamplifier located at the base of the whip. The arrangement is very convenient in the hf band, where external noise is so high as to mask that of the amplifier, despite the low level of the signal at the amplifier's input.

Fig. 8. A popular 300Ω balanced horizontal dipole covering all of the amateur bands. For general use, it should be mounted as high as possible. The two LR-loaded networks are 16μH and 300Ω.

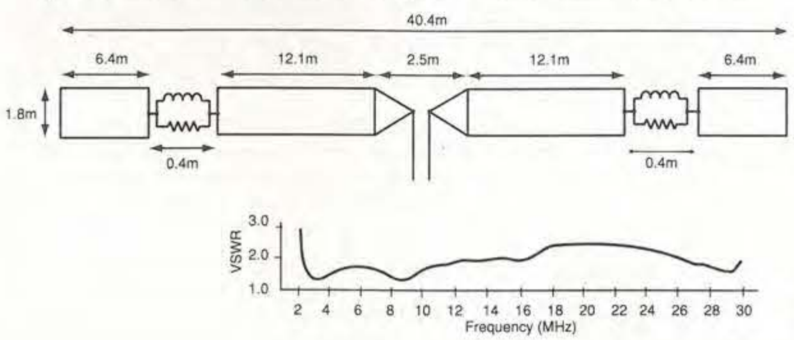


Figure 9 shows a suitable amplifier. It is connected to a vertical wire some 400mm high, inside a supporting plastic tube, topped by a metal disc 250mm in diameter. Details of the excellent electrical performance of the complete active antenna can be found ref. 2.

Loopy alternatives

Where it is not convenient to have an antenna projecting from the case of, for example, a hand-held device, the loop antenna mentioned earlier may be used. In the form known as a 'frame aerial', these have been employed since the earliest days of wireless.

Basically, the loop antenna consists of an air-cored coil, although in practice it might sometimes be more accurate to describe it as 'transceiver-cored'. Unlike the half-wave dipole and the quarter-wave whip, both of which are excited by the *E* field of a passing wavefront, frame aerials respond to the H field.

Generally, the term loop antenna is used to denote a loop aerial with but a single turn, while the term frame aerial is used for a multi-turn loop. The mode of operation has been described earlier.

Figure 7 shows a two turn loop or frame aerial. One use for such antennas is in direction finding. In this application, a loop with two or more turns side by side, as in Figure **7a)**, will not give a complete null when broadside on to the signal. For the loop then has a finite depth *d* in the direction of propagation. This makes it the equivalent of a loop of *N* coincident turns in series with a narrow single-turn loop of width *d*. So for direction finding, a frame aerial should be wound 'pancake' fashion, with spiral coplanar turns.

Of course, direction finding using a loop leaves a 180° ambiguity, for a null is obtained either way round. So arrangements can be made to temporarily add in the signal from a whip. This will be either adding or subtracting, thus resolving the ambiguity.

For best results from a frame aerial, the output should be connected to an amplifier with a balanced input. For if one end of the winding is grounded, different capacitive currents to ground flow in the first and last turns, resulting in some pick-up as from a short vertical whip. This not only upsets the null, but results also in greater susceptibility to pick-up of electrostatic interference. Ideally, the loop responds to the H field component of the signal only.

Where heavy electrostatic interference is experienced, the wires of a loop antenna can usefully be enclosed in a screening tube, as shown in Fig **7b)**. Such antennas are available commercially. The tube will of course have a small gap at one point to bring out the connections. A gap is necessary to prevent any circulating currents in the screen, which would then provide magnetic screening as well, preventing any wanted pick-up!

DXing

Frame aerials are popular with medium-wave 'DX' enthusiasts for a number of reasons. Firstly, such an aerial is directional. This means that when receiving a weak signal from a distant station, the loop provides useful discrimination against unwanted signals. Such signals may be on the same, or an adjacent frequency, arriving at a different azimuth angle to the wanted.

Secondly, the frame antenna is easily tuned, adding more selectivity to that provided by the receiver to which it is connected. Frame aerials are also convenient for reception in the hf band.

Figure 10 shows an active loop antenna. Using a three-turn 15in diameter coil of 8AWG wire with half-inch turn spacing, it covers 4.4-16MHz. The switch provides a choice of 8dB or 20dB gain. Further details can be found in ref. 3.

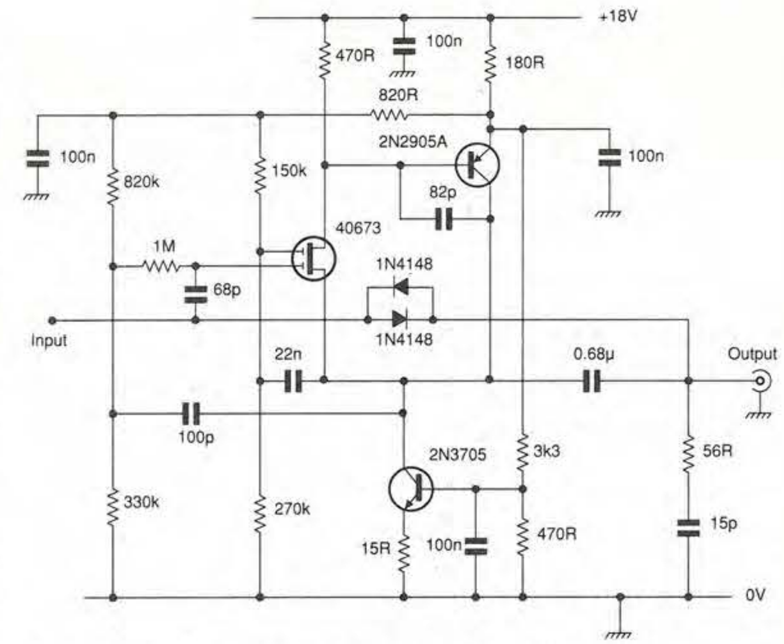


Fig. 9. In conjunction with a short top loaded whip (see text), this circuit provides excellent reception over the band 100kHz-30MHz.

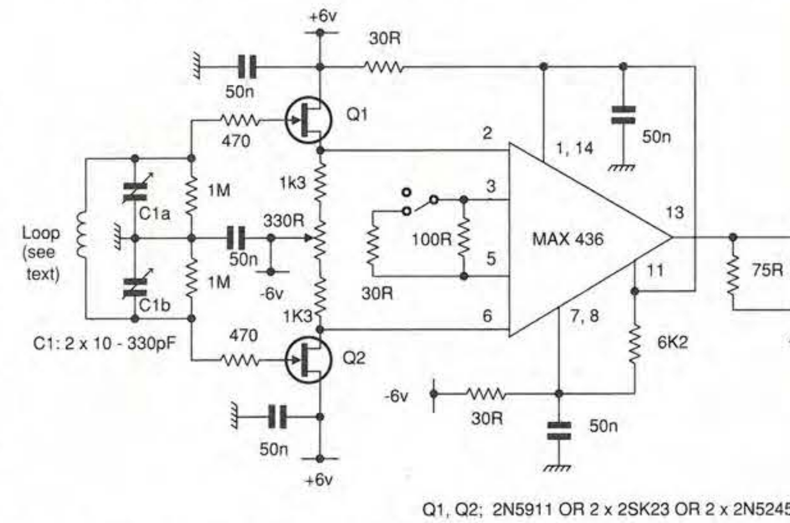


Fig. 10. This active frame aerial (see text) covers most of the hf band.

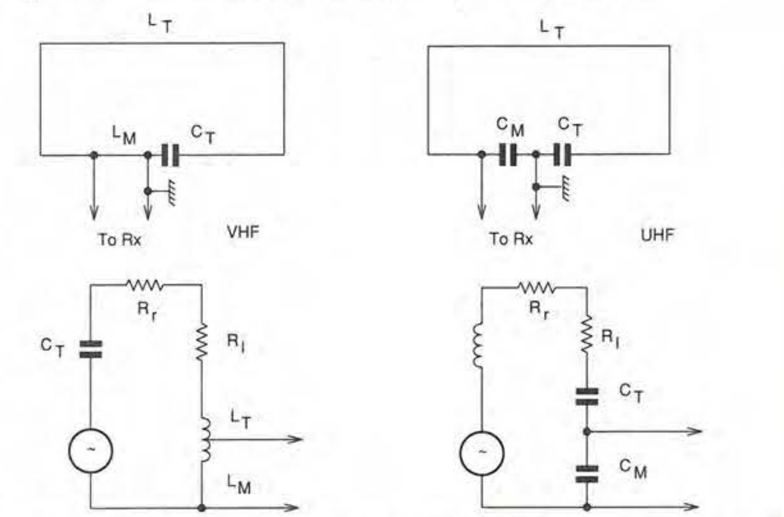


Fig. 11. Examples of arrangements commonly employed for small loop aerials at vhf and uhf.

Loops for vhf and uhf

At vhf and uhf, the loop usually consists of but a single turn, **Fig. 11**, with matching arranged as indicated. The loop may be of wire or strip, or it may be a track printed on a circuit board.

Design equations for such loop antennas involve the thickness and width of the conductor as well as the loop dimensions, and may be found in various references including refs 4 and 5. Remember though that design equations only provide a starting point, since the effects of the presence of such things as a circuit board, battery and outer casing need to be taken into account.

Any of the antennas mentioned – except of course the active ones – may be used for receiving or transmitting. It must be accepted that those types which are physically small compared to a wavelength – the coiled whip and particularly the small loop – exhibit reduced radiation efficiency. In both receiving and transmitting role, their performance may be up to 20dB below isotropic. Note that this is due solely to the losses; the effective aperture of a physically small antenna is theoretically independent of its actual dimensions.

But in practice, the radiation resistance becomes exceedingly small relative to both the loss resistance and the reactance. Tuning and matching then become impracticable. Note however that, depending on the radiation pattern, the antenna will still show a directional gain relative to an equally lossy isotropic antenna.

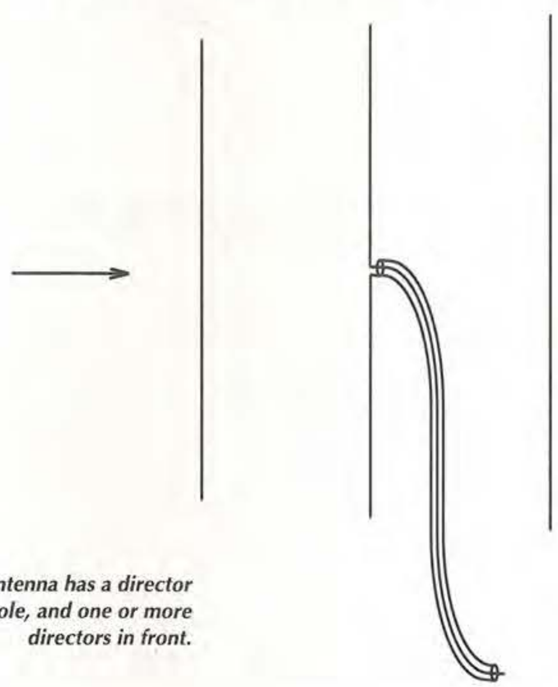


Fig. 12. A Yagi antenna has a director behind the $\lambda/2$ dipole, and one or more directors in front.

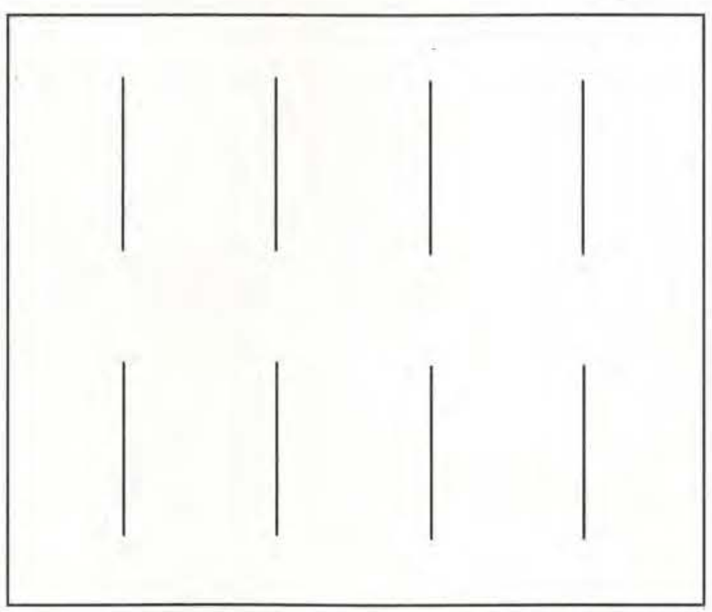


Fig. 13. An array of dipoles provides increased sensitivity in a given preferred direction, usually at right angles to the reflecting sheet. With a variable phase shifter for each element, the beam may be made electronically steerable.

Adding elements

Where space is not a constraint, e.g. in a fixed installation such as a telemetry link from a reservoir to the water board offices, the link may be successfully operated with less transmitter power, if antennas with a gain *G* greater than obtainable with a half-wave dipole are employed.

Such an antenna results if parasitic elements are added to a half-wave dipole. Imagine an element a little longer than half wave, placed close to the half-wave dipole and on the side away from the transmitter, **Fig. 12**. This element – called a reflector – reradiates all the energy that it picks up, but its phase will be leading the incident energy. By the time this reradiated energy reaches the half-wave dipole, it will be in phase with the direct radiation there, reinforcing it.

Similarly, a slightly shorter element, called a director and place in front of the half wave dipole, will reradiate the energy it picks up, in the appropriate phase to reinforce that picked up directly by the dipole. Further shorter directors may usefully be added, further ahead of the half wave dipole, but extra reflectors do not help.

Such 'Yagi' antennas with up to 18 elements are employed and can provide a gain of up to 15dB relative to isotropic. With this extra performance at each end of the link, the required power is only about one four hundredth of that which would be needed with simple half wave dipoles.

Another way to increase gain

An alternative way of achieving greater gain is to combine the signal pickup from a number of antennas. **Figure 13** shows an array of eight half-wave dipoles in front of a conducting sheet – which might in practice be chicken wire.

Using an eight-way hybrid combiner, the outputs of all elements are combined in phase. With such an array, the effective aperture approximates to the physical area of the array, and so can exceed λ^2. For even greater gain, the individual elements could each be a multi-element Yagi.

Such highly directional antennas provide the additional gain at the expense of reduced gain in other directions. To visualise why this is so, you only need consider using such an antenna for transmitting. The power supplied can only be concentrated in one direction at the expense of other directions. As a receiving antenna, this is an added bonus. Not only is the receiver more sensitive to the wanted signal, but unwanted signals from other directions are reduced relative to using a half-wave dipole, and reduced even more relative to the (enhanced) reception of the wanted signal – a double bonus. ■

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Digital storage analogue performance

Terry Marrinan of Tektronix describes a new dso feature that allows very rare events and glitches to be captured without winding up the brightness.

Although digital technology has brought a steady stream of increased functionality to digital storage oscilloscopes, or dsos, a major shortcoming has been the small fraction of time they actually spend capturing waveforms. The human eye gets the impression of rapid waveform capture when a dso updates its display 60 times a second, but in reality the dso misses large blocks of data.

For example, if the oscilloscope is set at a sweep speed appropriate for displaying a 10MHz clock, each display refresh will show about five cycles of this clock, or 500ns. Observing 500ns 60 times per second means acquiring 30ms out of every second, or 30 parts per million of real-time.

The best analogue oscilloscopes can refresh the screen several hundred thousand times a second, but they may still have trouble presenting rare events. The writing rate of the phosphor used in crts is too slow for a single glitch to be observable by a user, even with the use of a viewing hood.

With an analogue oscilloscope, the only way to see rare events is to use a crt that includes an electron multiplying plate between the deflection system and the screen phosphor. The addition of this microchannel plate gives oscilloscopes the capability to capture single-shot events at high sweep speeds, and make them visible to the operator.

New acquisition techniques

A new acquisition technology known as InstaVu has been developed to give dsos better fault-finding capabilities than the best analogue oscilloscopes. When used to verify clock and signal integrity, instruments based on this technology can acquire and display rare events and jitter without blooming, and can show aberrant signals with fine detail without the need of a viewing hood.

The oscilloscope acquisition process can be stopped when a glitch appears on the display, and a hard copy can be generated; alternatively, advanced triggers can be set to trigger on the event once its presence and shape have been determined, **Fig. 1**.

For debugging high-speed digital systems, InstaVu acquisition instantly shows a true picture of crosstalk, jitter and signal interference in a way that was previously possible only with microchannel plate-based analogue oscilloscopes. The accessibility of InstaVu acquisition through a single front-panel button allows quick clarification of the confusing displays often presented by previous dsos.

The new feature is also useful for evaluating high-speed digital communication signals. It can quickly capture intermittent signal anomalies, and this near-instantaneous feedback speeds the debugging process and boosts confidence in verifying the compliance of communication signals to industry standards.

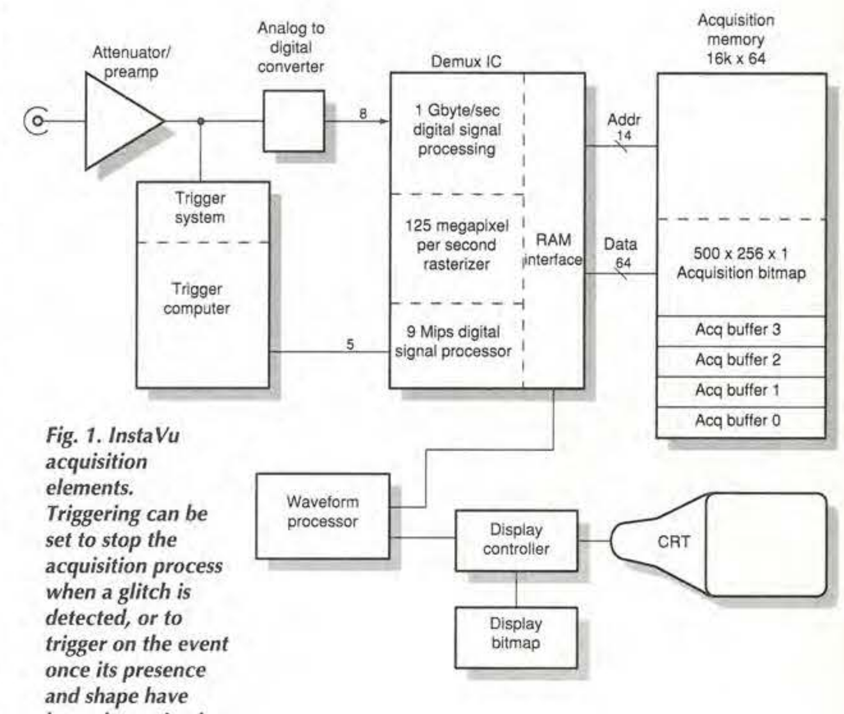


Fig. 1. InstaVu acquisition elements. Triggering can be set to stop the acquisition process when a glitch is detected, or to trigger on the event once its presence and shape have been determined.

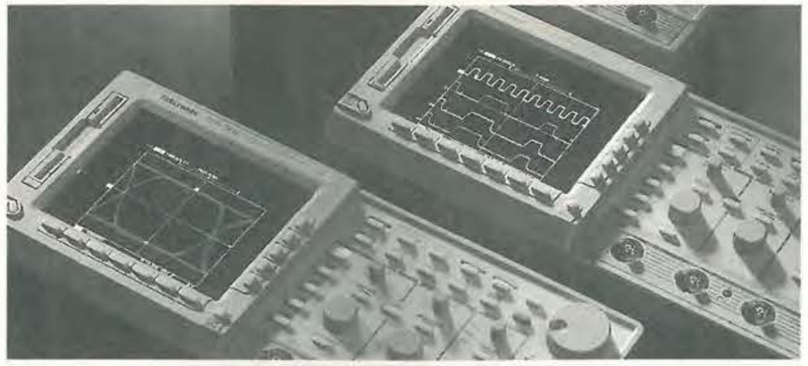
Aliasing and modulation instantly appear in their true form with InstaVu acquisition, and complex waveforms such as video, communications and radar signals can be visually assessed. This was previously possible only with an analogue oscilloscope. InstaVu acquisition is useful in any situation where enveloping or colour-graded displays of timing or amplitude jitter are necessary.

In system design or manufacturing applications, these acquisition oscilloscopes can rapidly display everything present in a signal, providing the user with an expedient and powerful means to analyse and optimise system operation.

A new architecture

To achieve 'live' digital acquisition, combining the display performance of an analogue oscilloscope with the capture and measurement capabilities of a digital oscilloscope, major revisions have been made to dso architecture.

First, the rasterisation capability of the display system has been duplicated in the acquisition system. Next, the rasteriser has been redesigned to use a portion of the high-speed acquisition memory to build display images.



The acquisition hardware itself now starts acquisitions without instrument firmware intervention, and calculates its own trigger positions. Finally, the instrument firmware and user interfaces have been adapted to the new form of data produced by the acquisition system, while supporting many of the dso's conventional functions.

With these changes, a dso now can offer acquisition rates that are the same or faster than those of analogue oscilloscopes. By combining high-speed acquisition memory with high-speed rasterisation, a radical increase in acquisition performance is obtained. When this mode is enabled, the data moved from the acquisition system is a complete, rasterised image of many triggered acquisitions of a user's signal.

Transferring this pixel map requires more data to be moved between the two systems, but the raster is only moved at the refresh rate of the oscilloscope's display and contains information from tens of thousands of acquisitions. Ten thousand 500-byte acquisitions moved to the display every 32ms would require a data rate of 167Mbyte/s. By comparison, one 500 by 200 by 1 bit per pixel raster moved to the display every 32ms equals a data rate of 417kbyte/s.

Besides displaying many acquisitions as a single raster image, InstaVu acquisition achieves its rapid acquisition rates by allowing the system to rearm itself and acquire as soon as it has completed one acquisition, rather than having the instrument firmware intervene on an 'acquisition-by-acquisition' basis. The instrument firmware only occasionally shuts down the acquisition system – once every 32 ms – and copies out the raster that shows the behaviour of the signal over the last 12000 acquisitions.

Common dso trigger rates – a measure of how quickly an oscilloscope can recognise trigger events – are much faster than their waveform capture rates. However, with the new acquisition technique, the trigger rate is actually the limiting factor. So the maximum waveform capture rate occurs with real-time acquisitions.

Without sufficient sample rates, repetitive sampling techniques would have to be used, which would slow down the waveform capture rate by a factor of ten or more, because each waveform would be constructed of several triggered acquisitions. Since the InstaVu instruments have high real-time sample rates – up to 4Gsamples/s – they can be used to acquire signals at the maximum waveform capture rate up to the scope's analogue bandwidth.

Unique demultiplexer

Much of the hardware necessary to implement InstaVu acquisition is integrated into one highly-integrated demultiplex chip.

Normally, the only function of this IC would be to demultiplex data from the analogue-to-digital convertor, and then store that data in high-speed static ram. One-third of this IC is devoted to this purpose, the remainder being split evenly between a high-speed rasteriser and a digital signal processor.

The digital processor is included for local programmability, mathematical algorithms and trigger position computations. The rasteriser is the primary enabler of the high live-time dso. The rasteriser is designed to make efficient use of available memory bandwidth while operating on a 16ns clock. It is able to draw four acquisitions at once into a 500 by 256 by 1 bit map. The bit map is organised as vertical lines of 256 pixels – 512 pixels in high-resolution mode – so that adjacent bits in the memory correspond to vertically adjacent bits. Thus drawing is done in a top-to-bottom, then left-to-right fashion, so that each data point in an acquisition need be fetched only once.

On the first pass through the bitmap, the rasteriser clears the contents of memory while turning on pixels that correspond to the voltage levels of each of the four acquisitions it is rasterising. On subsequent passes through the bitmap, it reads the previous contents of the bitmap, ORs in new pixels, and writes the result back into the same section of memory.

The read/modify/write, or rmw, cycles operate on 64 pixels at a time, and each cycle is 32ns long. Data for each acquisition that is being rasterised is fetched eight bytes at a time as needed. These data reads are allowed every 16ns, and are interspersed among the rmw cycles, at times sufficiently ahead of when the corresponding columns are modified, so that the data can propagate through several pipeline stages.

Dot or vector mode?

Each waveform is drawn into the raster in either dot or vector mode; in dot mode, a single pixel is turned on at the time and voltage level corresponding to a single point in an acquired waveform. In vector mode, all the pixels between a lower and upper limit in a single column of the raster are turned on. This operation does not slow down the waveform capture rate.

If the four acquisitions lie in the same quadrant of the screen for a given sample time, 32ns is required to update that time's raster column. On the other hand, if each of the four samples is in a different quadrant, or if the rasteriser is operating in vector mode and a signal edge is being rasterised, four rmw cycles may be necessary, and it would take 128ns to update that particular column.

If you are looking at a logic signal, few of the time columns will require more than a single rmw. If it is assumed that 5% of the time slots require four rmw cycles, the rmw time for four 500-point acquisitions is 20.8µs.

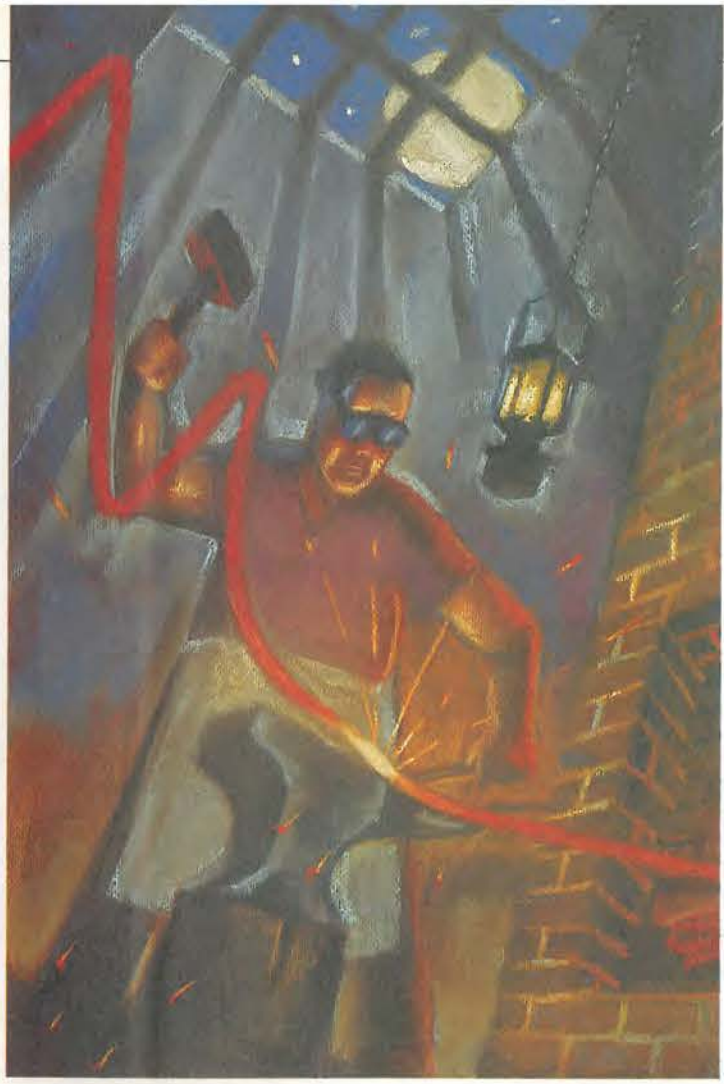
The time required to read the data is 500/8 by 4 by 16 ns – or 4µs – and the total time for a single rasterisation is 24.8µs. This corresponds to 6.2µs per acquisition.

This rasterisation rate allows only about 100000 acquisitions per second, which is still short of the maximum rate attainable by the best analogue oscilloscopes. However, an analogue oscilloscope's maximum waveform capture rate is only attainable when a single channel is used. When a four-channel InstaVu acquisition oscilloscope acquires a single channel, greater live time is achieved by allowing each of the four channels to take turns acquiring a single input.

As a result, three rasterisations can be run while the input continues to be monitored. In addition, three channels can continue to take turns acquiring a single input, while firmware is unloading the raster in the fourth channel.

With this technology, dsos can perform up to 400000 full-screen, i.e. 500-point, acquisition/rasterisation cycles per second on a single channel. This rate works out to 220 million pixels per second, and is limited as much by the trigger system rearm speed as by acquisition system graphics performance.

Once the acquisition system is producing pixel maps, the instrument firmware can collect them and provide infinite and variable persistence displays. Thus, these oscilloscopes can be left on overnight if a few tens of billions of acquisitions are needed, or can be operated interactively to produce colour-graded displays. ■



By looking at distortion residuals and classifying them, Doug Self offers a way to analyse your amplifier's performance with a view to improving it.

Diagnosing distortions

In recent years, some audio commentators have been rudely dismissive of the simplest and most basic kind of distortion measurement – the total-harmonic distortion, or thd, test.

Because thd measurement has a long history, it is easy to imply that it is outdated and used only by the clueless. This is not so. Many other distortion tests exist, but none of them allows instant diagnosis of audio problems with one glance at an oscilloscope.

The test requires an oscillator with negligible distortion, feeding the unit under test. A notch-filter then completely suppresses the fundamental, to reveal the distortion products that have been generated. What remains after the fundamental is removed is not unnaturally called the thd residual.

In several previous articles I have described the various distortions that afflict audio power amplifiers. In the generic circuit, these are relatively few in number as is evident from the panel entitled 'Distortion mechanisms.' Here I will show what some of these distortion residuals actually look like. Distortions 1, 2, 4 and 8 are not very informative visually, being essentially second or third harmonic, so I have omitted them to make room for the more complex waveforms specific to Class B.

The blameless amplifier concept

A Blameless amplifier results when the known distortions in the panel on distortion mechanisms have been either minimised or reduced to below visibility on the thd residual. It is so-called because it achieves its superb linearity not by startling innovation but simply by avoiding a series of possible errors. Avoiding them is straightforward once they are identified.

The concept of a Blameless amplifier has proved extremely useful. Such an amplifier has surprisingly low thd, despite its conventional-looking circuitry, but its greatest advantage is its defined performance, only weakly dependent on component characteristics.

If an amplifier does not perform to Blameless standards of linearity, then there is something fairly simple wrong with it, and to attempt to improve it by adding extra circuitry or turning up the bias into Class AB misses the point totally.

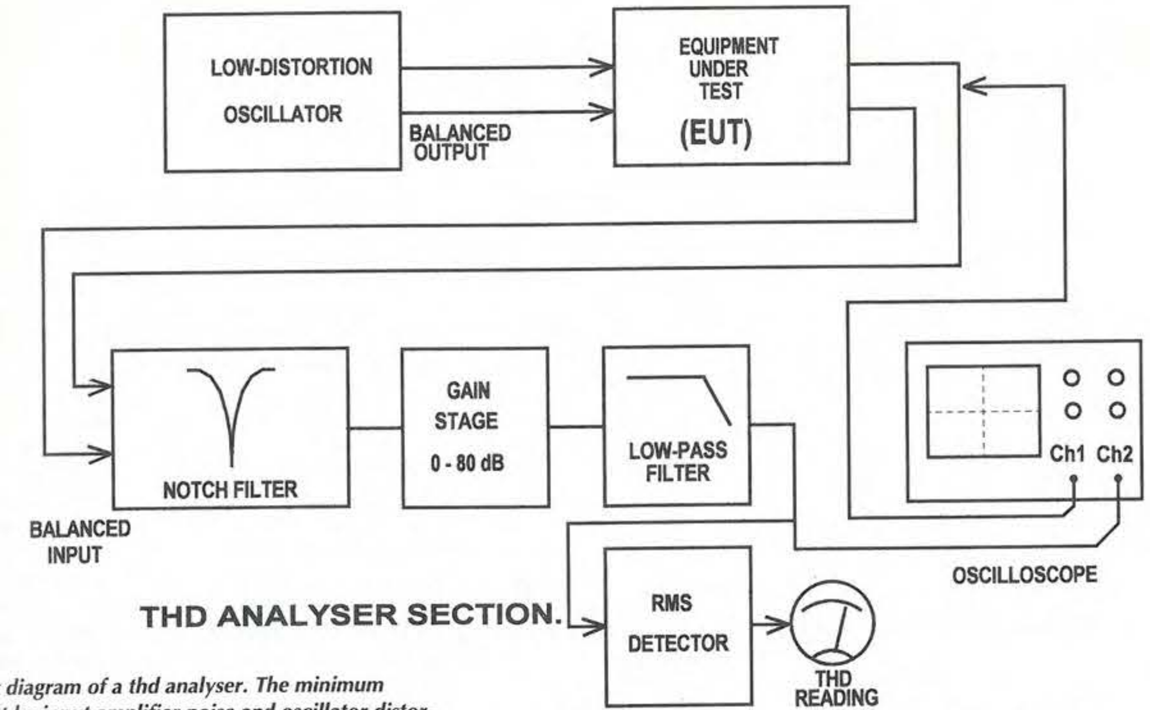


Fig. 1 .Block diagram of a thd analyser. The minimum reading is set by input amplifier noise and oscillator distortion rather than the filter auto-tuning.

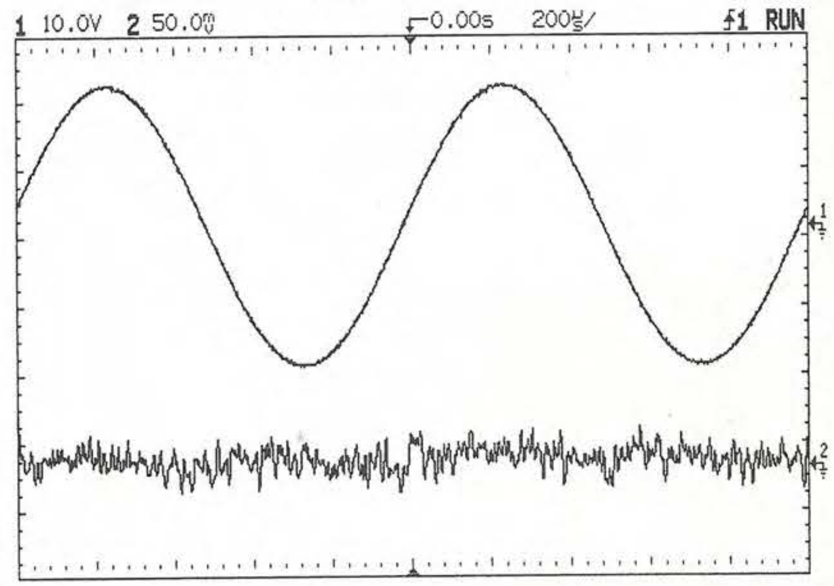


Fig. 2. The thd residual from an optimally-biased Blameless power amplifier at 1kHz, 25W/8Ω is essentially white noise. There is some evidence of artifacts at the crossover point, but they are not measurable. THD 0.00097%, 80kHz bandwidth.

Making distortion measurements

Total harmonic distortion is the rms sum of all the distortion components generated by the path under test. It is usually quoted as a percentage of the total signal level.

The rms calculation – taking the square-root of the sum of the squares of the harmonics – emphasises spiky distortions, but whether this helps to mimic human perception of distortion is unclear. The peak capability of true-rms circuitry is limited, and this may well lead to under-reading of crossover spikes and such.

I hold that the best method is to observe the residual simultaneously, and time-aligned, with the output sinewave as in **Fig. 1**. If you are testing similar pieces of equipment, then the gain of the oscilloscope's second channel for the residual can be kept at the same setting. This allows linearity to be assessed at a glance.

In contrast, it is wiser to connect the actual output to channel 1, rather than an auto-scaled version from the analyser, as this prevents parasitics, etc, from being filtered out by the analyser input circuitry.

The beauty of thd testing is that the error is isolated; in essence, the residual is the difference between perfection and reality. When viewed time-aligned with the output sinewave,

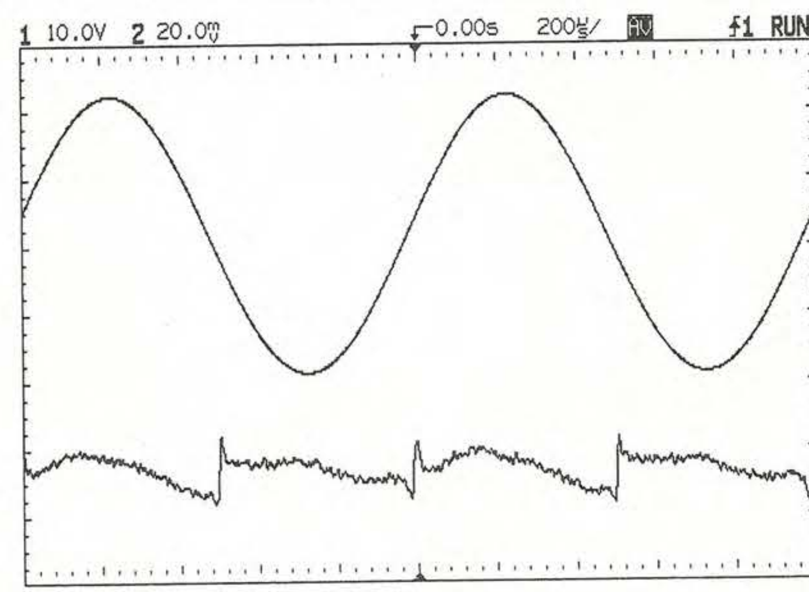


Fig. 3. Averaging the Fig. 2 residual 64 times reduces the noise by 18dB, and crossover discontinuities are now obvious. The residual has been scaled up by 2.5 times from Fig. 2 for greater clarity.

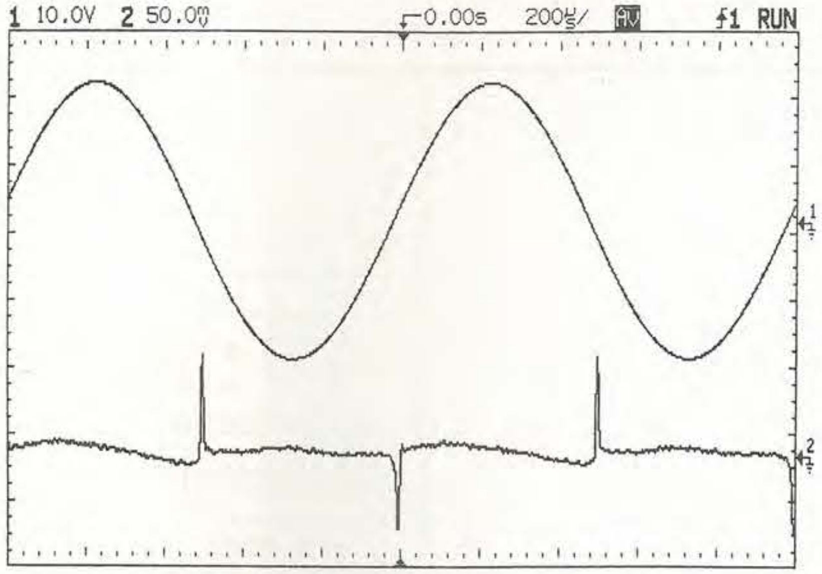


Fig. 4. Results of mild underbias in Class B.

crossover distortion can be diagnosed immediately as it occurs at the zero-crossings. On the other hand, non-linearity confined to one peak is probably due to something running out of voltage swing or current capability.

Two technical challenges...

Figure 1 shows the basic thd measuring system. There are two major technical challenges to be overcome. The signal source must be extremely pure, as any oscillator distortion puts an immediate limit on the measurement floor; it must maintain superb performance at least over the range 10Hz to 20kHz. A balanced output is highly desirable.

In the analyser section a balanced input is essential. Very great attenuation of the fundamental is required – about 120dB if you are going to measure down to 0.0005%, making notch tuning is extremely critical. This cannot be attained by fixed-tuned filters, and manual tuning, requiring at least six controls, is about as much fun as picking oakum. In modern thd equipment both frequency and phase are continuously adjusted by a twin servo-loop that optimises the cancellation.

An addititional low-pass filter defines the measurement bandwidth. Usually, 80kHz is a good compromise, retaining most of the important harmonics while reducing noise. A switchable 400Hz high-pass filter is often fitted, allowing measurements at 1kHz and up, in the presence of hum. Such a filter should be used only in exceptional cases, for thd often rises sharply at low frequencies, and this would be missed.

While frequently advocated as a more searching examination of an audio path, twin-tone intermodulation tests are almost useless for circuit investigation. They give very little information about the source of the non-linearity as the phase relationship between the test signal and the result is lost. It is often claimed they give a better measure of audible degradation in real use, but a test using two or three tones is still a long way away from music that has tens or hundreds of simultaneous frequencies. Intermodulation tests can often dispense with very-low-thd oscillators, but this in itself is not much of a recommendation.

If real subjective degradation is the issue, a test signal much closer to reality is required. This can be either pseudorandom noise as in the Belcher test,[1] or real music, as in the Baxandall[2] and Hafler[3] cancellation tests.

Returning to harmonic distortion, much better correlation between thd measurements and subjective impairment is possible if the harmonics are weighted so that the higher order components are emphasised.

Weighting by $n^2/4$, so that the second harmonic is unchanged, the third increased by 9/4, and so on, is generally accepted to be roughly correct.[4,5] I was surprised to find that this approach goes back to 1937 and before.[6] I doubt however whether this can be applied to crossover distortion.

When the thd residual is displayed on an analogue oscilloscope, artifacts in the noise are easily detectable by the averaging processes of our vision, but they remain unavailable to conventional measurement. A digital scope can perform even more effective averaging by computation, making submerged distortion artifacts both visually clearer and readily measurable, though an rms mode may not be available.

If a noisy signal is averaged two times, by combining two sweeps, the coherent signal stays at the same level, while the uncorrelated noise decreases by 3dB. Averaging 64 times performs this process six-fold, so noise is then reduced by 18dB. The oscilloscope used here was a digital *HP54600B* 100MHz digital storage; an excellent instrument. This choice will not come as a surprise to alert readers.

Although sometimes invaluable, digital oscilloscopes are often not the best choice for audio thd testing and general amplifier work; in particular the problems of aliasing make the detection and cure of hf oscillations very difficult.

To create the residuals shown here, a Blameless amplifier was used essentially identical to that published in reference 7. Output was 25W into 8Ω, or 50W into 4Ω. The Blameless amplifier concept is outlined in a separate panel.

Crossover distortion

Crossover distortion is only one of the three components that make up Distortion 3 but is often the dominant one. Blameless amplifiers show only crossover distortion when driving 8Ω or more, and at low and medium frequencies it should be below the noise. This remains true even if the amplifier noise is within a few decibels of the theoretical minimum from a 50Ω source resistance.

Figure 2 shows the thd residual from such a Blameless power amplifier, with optimally biased in Class-B. Since this is a record of a single sweep, the residual appears to be almost wholly noise. The visual averaging process is absent and so the crossover artifacts are actually less visible than on an analogue scope in real time.

In **Fig. 3**, 64 times digital averaging is applied, which makes the disturbances around crossover very clear. A low-order component at roughly 0.0003% is also revealed, which is probably due to very small amounts of Distortion 6 that were not visible when the amplifier layout was optimised.

Figure 4 shows Class B mildly underbiased to generate crossover distortion. The crossover spikes are very sharp, and

Distortion mechanisms

My original series on amplifiers[7] listed seven independent distortions inherent to the generic/Lin Class-B amplifier, and whose existence is not dependant on circuit details. I have now increased this to eight.

Distortion one

Input-stage distortion. Non-linearity in the input stage. If this is a carefully-balanced differential pair then the distortion is typically only measurable at high frequenies, rises at 18dB/octave, and is almost pure 3rd harmonic.

If the input pair is unbalanced – which from published circuitry it usually is – then enough second harmonic is produced to swamp the third. Hence the hf distortion emerges from the noise at a lower frequency, rising at 12dB/octave.

Distortion two

Voltage amplifier stage distortion. Surprisingly, non-linearity in the voltage-amplifier stage does not always contribute significantly in the total distortion. If it does, it remains constant until the dominant-pole frequency P1 is reached, and then rises at 6dB/octave. In the generic configuration discussed here it is always second harmonic.

Distortion two

Output-stage distortion. Non-linearity in the output stage – the most obvious source. This has three components: crossover distortion (3a) usually dominates for Class-B into 8Ω, generating high-order harmonics rising at 6dB/octave as global negative feedback decreases. Low-order large-signal nonlinearity (3b) appears with 4Ω loads and worsens at 2Ω. Distortion 3c stems from overlap of output device conduction and only appears at high frequencies.

Distortion four

Voltage-amplifier loading. Loading of the voltage-amplifier stage by the non-linear input impedance of the output stage.

Distortion five

Rail decoupling distortion. Non-linearity caused by large rail-decoupling capacitors feeding the distorted signals on the supply lines into the signal ground. This seems to be the reason that many amplifiers have rising thd at low frequencies.

Distortion six

Induction distortion. Induction of Class-B supply currents into the output, ground, or negative-feedback lines. Almost certainly the least understood and so most common distortion afflicting commercial amplifiers.

Distortion seven

Negative-feedback take-off distortion. Non-linearity resulting from taking the negative feedback feed from slightly the wrong place near the point where Class-B currents sum to form the output.

Distortion eight

Capacitor distortion. Rising as frequency falls, capacitor distortion is caused by non-linearity in the input dc-blocking capacitor or the feedback network capacitor. The latter is more likely.

Distortions *x*

Non existent or negligible distortions. Common-mode distortion in the input stage and thermal distortion in the output stage – or anywhere else.

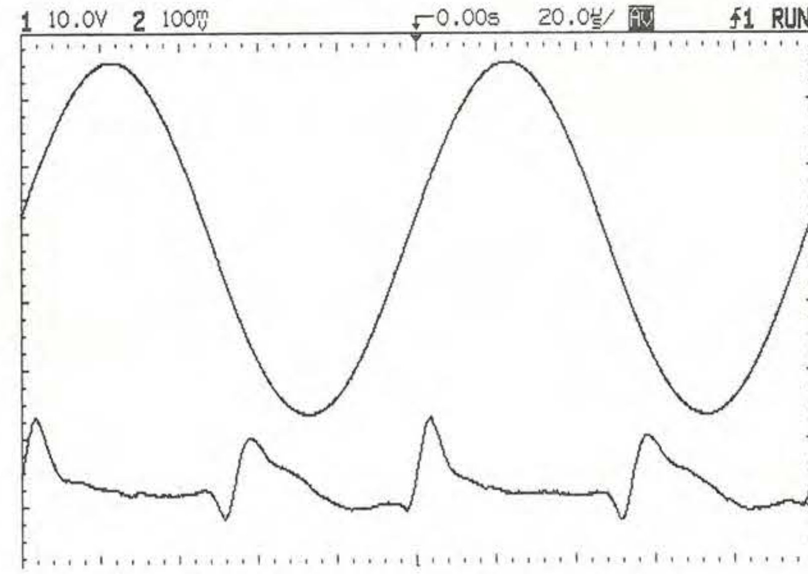


Fig. 5. An optimally-biased Blameless power amplifier at 10kHz. THD is around 0.004%, bandwidth 80kHz. Averaged eight times.

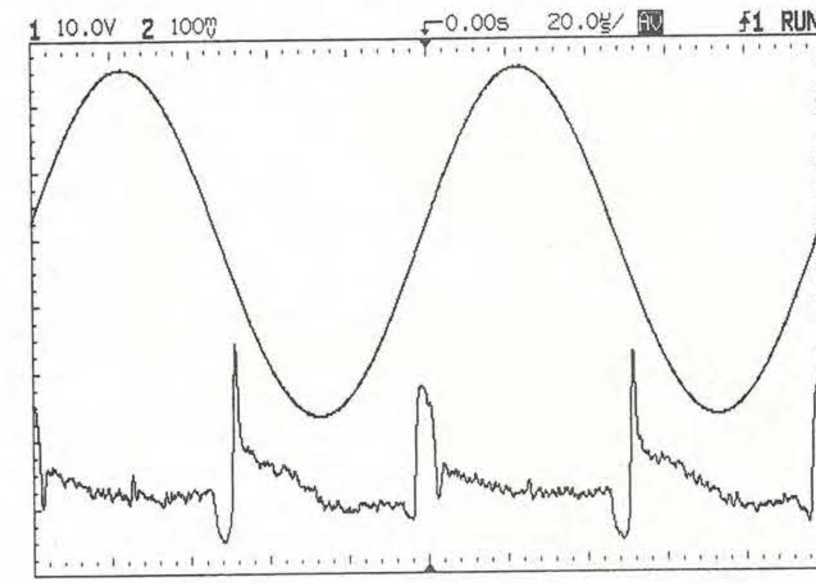


Fig. 6. As Fig. 6, but in 500kHz bandwidth. The distortion products look quite different.

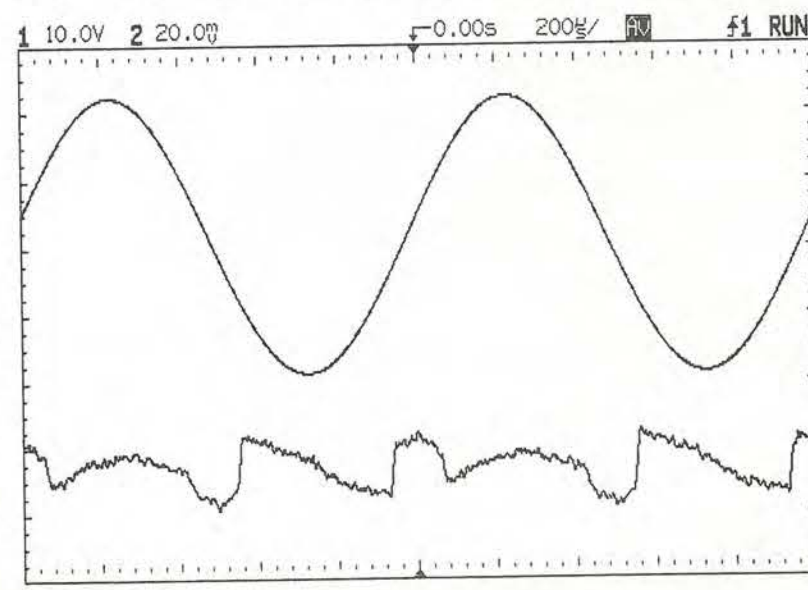


Fig. 7. The g_m-doubling distortion introduced by Class AB. The edges in the residual are larger and no longer at the zero-crossing, but displaced either side of it.

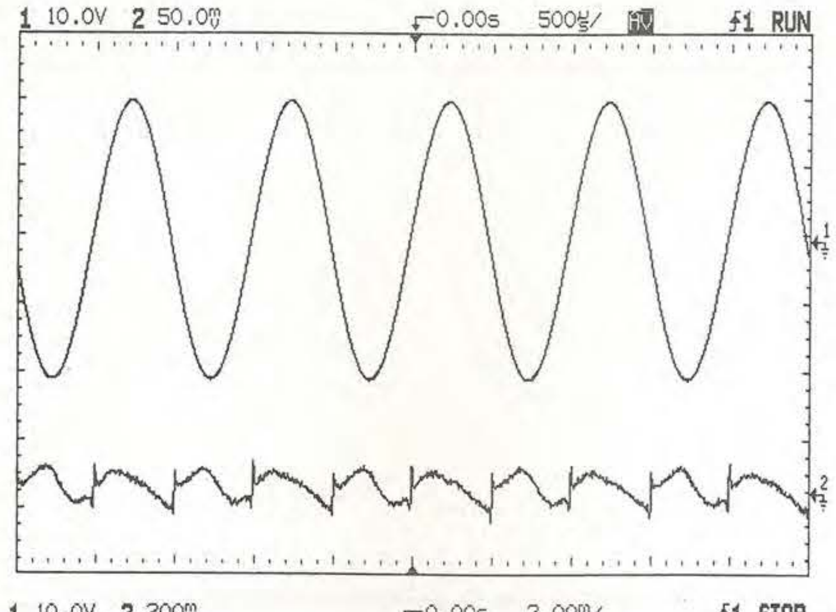


Fig. 8. Large-signal nonlinearity, driving 50W into 4Ω, and averaged 64 times. The extra distortion appears to be a mixture of third harmonic – occurring as a consequence of the compressive nature of beta-loss – and second harmonic arising because the beta-loss is not perfectly symmetrical in the two halves of the output stage.

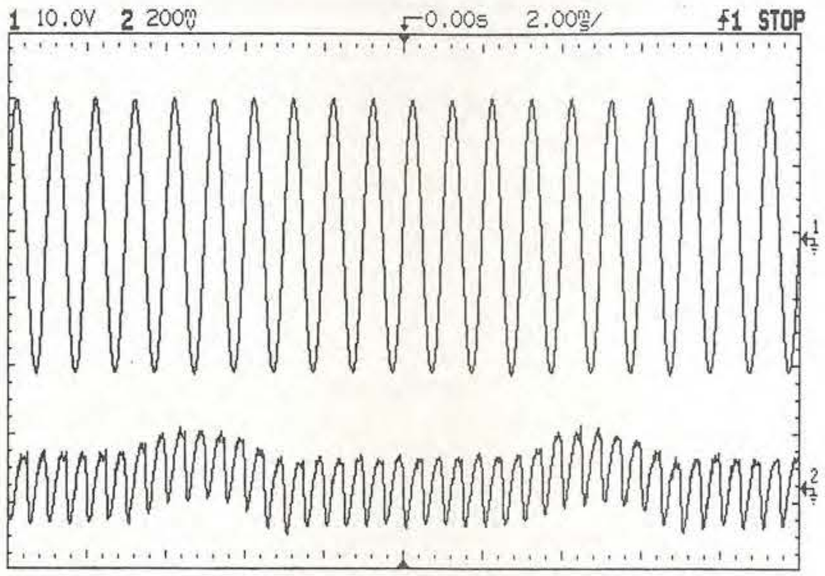


Fig. 9. Distortion 5 revealed. Connecting the rail decoupler to input ground increases thd eight-fold from 0.00097% to 0.008%, mostly as second harmonic. 100Hz ripple is also visible. No averaging.

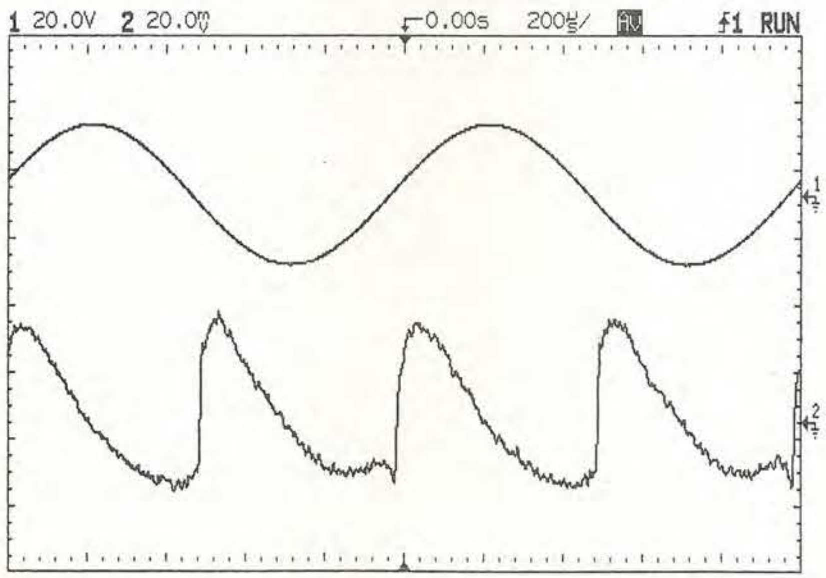


Fig. 10. Distortion 6. Induction of half-wave signal from the negative supply rail into the negative feedback line increases thd to 0.0021% Averaged 64 times.

their height in the residual depends critically on measurement bandwidth. Their presence warns immediately of under-biasing and avoidable crossover distortion.

In **Fig. 5** an optimally-biased amplifier is tested at 10kHz. The thd has increased to approx 0.004%, as the amount of global negative-feedback is 20dB less than at 1kHz. The crossover events appear wider than in Fig. 3. The higher thd level is above the noise so the residual is averaged eight times only.

The measurement bandwidth is still 80kHz, so harmonics above the eighth are lost. This is illustrated in **Fig. 6**, which is Fig. 5 rerun with a 500kHz bandwidth. The distortion products look very different.

The 80kHz cutoff point is something of *de facto* standard, which is reasonable as it seems highly unlikely that ultrasonic harmonics can detract from one's listening pleasure. This does not mean thd testing can stop at 10kHz, as there might be an area of bad intermodulation in the top octave.

My practice is to test up to 50kHz, to check that nothing awful is lurking just outside the audio band; this is safe for moderate powers, and short durations.

Classes B and AB

I showed in my series on power amplifier distortion[7] that Class AB is not a true compromise between Class A and Class B operation. If AB is used to trade off efficiency and linearity, its linearity is superior to B since below the AB transition level, it is pure Class A.

The Class-A region can – and should – have very low thd indeed, below 0.0006% up to 10kHz, as demonstrated in ref. 8. However, above the AB transition level thd abruptly worsens. This is due to what has been called 'g_m-doubling', but is better regarded as a step in the gain/output-voltage relationship. Linearity is then inferior not only to Class-A but also to optimal-bias Class-B.

It is possible to make Class AB distortion very low by proper design. Basically, this means using the lowest possible emitter resistors to reduce the size of the gain step.[9] Even so, thd remains at least twice as high as Class-B.

Tweaking up the bias of a Class-B amplifier most certainly does *not* offer a simple trade-off between power dissipation and overall linearity, despite the constant repetition this notion receives in some parts of the audio press. The real choice is: very low thd at low power and high thd at high power, or medium thd at all powers. The electricity bill is another issue.

Figure 7 shows the gain-step distortion introduced by Class AB. The undesirable edges are caused by gain changes that are no longer partially cancelled at the crossover; they are now displaced to either side of the zero-crossing. No averaging is used here as the thd is higher and well above the noise.

Large-signal non-linearity

When the load resistance falls below 8Ω, extra low-order dis-

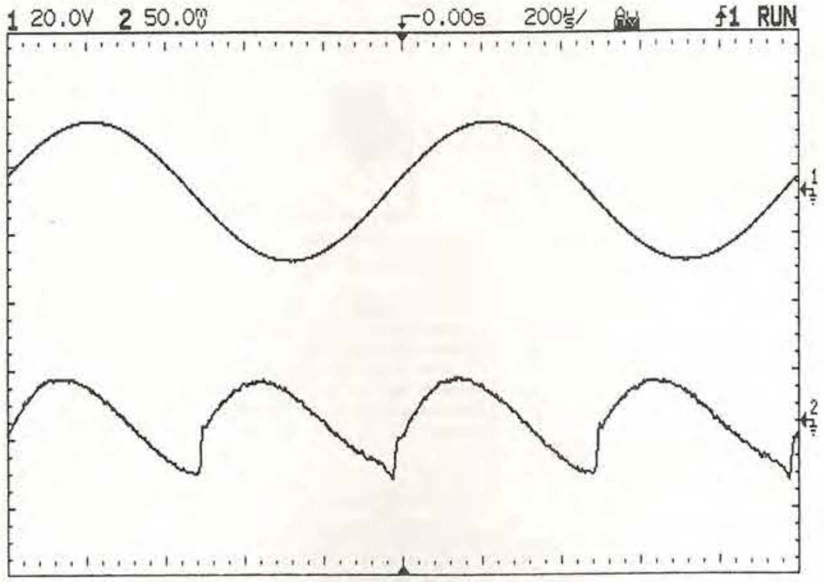


Fig. 11. Distortion 7, caused by choosing an negative feedback take-off point inside the Class-B output stage rather than on the output line itself. THD increases from 0.00097% to 0.0027%, by taking the negative feedback from the wrong end of 10mm of very thick resistor leg. Averaged 64 times.

tortion components appear. This is true for most or all modern power bipolar junction transistors, but with old devices like 2N3055, some large-signal nonlinearity may appear at 8Ω. This is a compressive non-linearity, ie gain falls as level increases, 'squashing' the signal, and is due to fall-off of transistor beta at high collector currents.

Figure 8 shows the typical appearance of large-signal non-linearity, driving 50W into 4Ω, and averaged 64 times. The extra distortion appears to be a mixture of third harmonic, due to the basic symmetry of the output stage, with some second harmonic, because the beta-loss is component-dependant and not perfectly symmetrical in the two halves of the output.

Other distortions

Of the distortions that afflict generic Class-B power amplifiers, 5, 6 and 7 all look rather similar in the thd residual. This is perhaps not surprising since all result from adding half-wave disturbances to the signal.

Distortion 5 is usually easy to identify as it is accompanied by 100 Hz power-supply ripple; 6 and 7 introduce no ripple. Distortion 6 is easily identified if the dc power cables are movable, for altering their run will strongly affect the quantity generated.

Figure 9 shows Distortion 5, provoked by connecting the negative supply rail decoupling capacitor to the input ground instead of giving it its own return to the far side of the star point. Doing this increases thd from 0.00097% to 0.008%, mostly as second harmonic. Ripple contamination is significant and contributes to the thd figure. It could be easily filtered out to make the measurement, but this is just brushing the problem under the carpet.

Distortion 6 is displayed in **Fig. 10**. The negative supply rail was run parallel to the negative-feedback line to produce this diagram. Although more than doubled, thd is still relatively low at 0.0021%, so 64-times averaging is used.

Figure 11 shows a case of Distortion 7, introduced by deliberately making a minor error in the negative feedback take-off point.

If it is attached to a part of the Class-B output stage so that half-wave currents flow through it, rather than being on the output line itself, thd is increased. Here it rose from 0.00097% to 0.0027%, caused by taking the negative feedback from the wrong end of the leg of one of the output emitter resistors, R_e.

Note this was at the right end of the resistor, otherwise thd would have been gross, but 10mm along a very thick resistor leg from the output line junction. Truly, God is in the details.

Diagnosis

The rogue's gallery of real-life thd residuals portrayed here will hopefully help with the problem of identifying the distortion mechanism in a misbehaving amplifier. There is no reason why the generic/Lin configuration should give measurable thd at 1kHz, or more than say, 0.004% at 10kHz when driving 8Ω.

It is important to be sure that you are measuring a real distortion mechanism, and not the results of parasitic oscillation upsetting circuit conditions; the oscillation itself may be outside the scope bandwidth. Parasitics usually vary greatly when a cautious finger is applied to the relevant section of the circuitry. Real distortion changes little, though the thd reading will probably be increased by the introduction of hum.

I hope I have shown that thd testing gives an immediate view into circuit operation that other methods do not, however useful they may be in other applications.

It cannot be stated too strongly that to attempt amplifier design and diagnosis without continuous visual observation of the thd residual is to work blind. You *will* proverbially fall into the ditch. ■

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5. Shorter, D, 'Influence of High-Order Products in Non-Linear, Distortion', *Electronic Engineering*, April 1950, p. 152
6. Callendar, M, Letter, *Electronic Engineering*, Oct 1950, p. 443.
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8. Self, D, 'Distortion In Power Amplifiers, Part 8,' *Electronics & Wireless World*, March 1994, p. 225, Class A amp.
9. Self, D, 'Trimodal Audio Power,' *Electronics World*, June 1995, p. 462. Effect of R_e in Class AB.

Cyril Bateman tells where to find information on browser wars, LM10 applications and simulation and design software. He also offers tips to help save you time when searching and downloading.

Hands-on Internet

Internet's browser wars escalated, following mixed reactions to the 1 October final release of Internet Explorer 4 by Microsoft. Some reviewers simply recommended waiting for Windows98, which consolidates a browser into its operating system. Meanwhile, to cope with anticipated demand, Microsoft increased its download capacity to 6.1 terabytes, permitting up to 450 000 Explorer 4 downloads a day.

Internet Explorer 4 takes a long time to download and install, so if you are interested in obtaining it, I advise you to read the on-line set-up and pre-installation instructions.[1] Non-US users should consider using one of the local mirror sites. You might prefer to order this software on cd-rom, available for a nominal fee, rather than attempting to download the software.

Java – once the hope for a universal cross platform system – has now fragmented, leaving software developers in a quandary[2] as to how best to proceed. According to Sun, this is due to Microsoft's specific implementations of Java in its Explorer 3.x and 4.0 browsers. After a long-running un-resolved dispute, Sun announced legal action against Microsoft on 7 October for allegedly violating its Java licence agreement, **Fig. 1**.

Following the Explorer 4 release, many Web sites reported operating difficulties[3] from people using Windows 95. Adobe *PhotoDeluxe 2.0* and *Norton Utilities* for Windows 95 are reported to display blank screens when run on a system after Explorer 4 is installed. Both companies are working to supply patches to correct this.

These problems are widely reported in the on-line news pages and Microsoft has issued open letters and

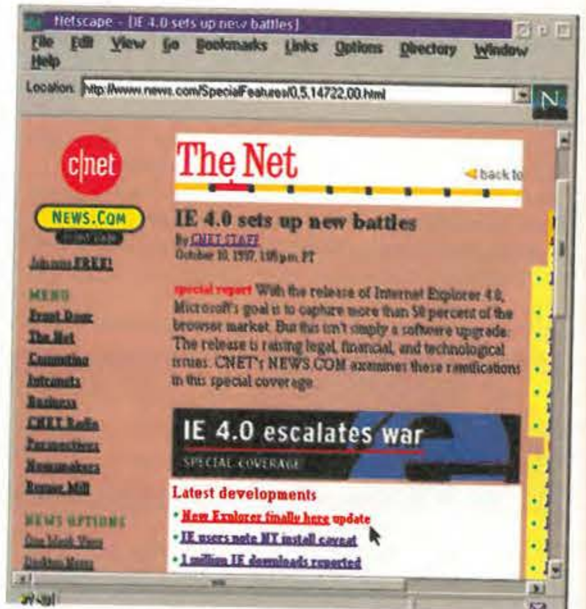


Fig. 1. Internet Explorer 4 is more than a simple software upgrade. It has triggered legal as well as Internet battles.

Where to surf

1.	Internet Explorer 4.0.	http://www.microsoft.com/ie/ie40/download/win95.htm
2.	News.Com	http://www.news.com/SpecialFeatures/0,5,14722,00.html
3.	American Weather Concepts	http://www.weatherconcepts.com/bugs_awc.html
4.	Adobe Acrobat Reader 3.01	http://cgi1.adobe.com/acrobat/download5.cgi
5.	Aladdin Ghostscript 5.03	http://www.cs.wisc.edu/~ghost/aladdin/get503.html
6.	GSview 2.3	http://www.cs.wisc.edu/~ghost/gsview/new.html
7.	Intusoft.	http://www.intusoft.com
8.	Microsoft Word Viewer 97.	http://www.microsoft.com/msword/internet/viewer/viewer97_16
9.	LM10 Application.	http://www.linear.com
10.	Conversions and Calculations	http://www.woodbas.demon.co.uk/calcs/calcs.htm
11.	Power Innovations.	http://www.powinv.com
12.	Macquarie University	http://www.mpce.mq.edu.au/elec/microelec/distortion/distortion_abs.html
13.	Analogy Inc.	http://www.analogy.com/default.htm
14.	California Institute of Technology	http://www.leonardo.caltech.edu
15.	Protel International.	http://www.protel.com
16.	The Major Search Engines	http://www.searchenginewatch.com/major.htm

FAQs in response.

Netscape browser and server software supports some 70% of all Web accesses. Validated by NIST, it is the only software meeting the FIPS 140-1 security standard required for the US Government. The 8 October licensing agreement between the US Department of Defence and Netscape has the potential to add two million licences, further increasing Netscape's dominance.

PDF and PostScript

A wealth of component data and circuit application notes can be found on Internet, generally in the form of .PDF files. These require use of the Adobe Acrobat 3.01 file reader, freely available by download from the Adobe[4] site.

Unfortunately this format is not universally used. Frequently, at locations driven by UNIX, only PostScript files are available. And many universities and colleges use Unix. This is fine if you are using a PostScript printer, but otherwise seemingly useless.

One possible solution requires a file conversion utility. These work with files comprising text, but not equations or drawings. Searching on 'file utilities' will result in a variety of solutions.

One universal solution, able to display PostScript files on screen and print to non-PostScript printers can be found in the *Ghostscript* interpreter. This is a command line utility that can be downloaded as source files from the Aladdin[5] *Ghostscript* site.

Fortunately, *Ghostscript* is also available as pre-compiled executables for OS/2, MS-DOS, Windows and Macintosh. Combined with the *GSview*[6] graphical front end for OS/2 or Windows, it produces an easily used but large installation, able to display or print text, equations and graphics.

Alternative file formats

Some interesting and useful simulation files are only available in the .RPL format, as used on the Intusoft site and the company's cd rom. A suitable reader, which can be downloaded from the Intusoft[7] page, is included on the cd.

Files written in various MS Word formats can be read using the freeware 'Word Viewer for Windows 16-bit reader', available on the Microsoft page.[8] This permits reading, printing and copying sections to the clipboard, but not editing.

Especially for larger .PDF files, I prefer to download and save the file to disk for off-line viewing. This is because I find that Adobe's 'plug-in' used with my Netscape browser slows down my system's download rate.

Many application files can be read on-line using HTML in conventional browsers. I found one such file – a novel application for the *LM10* op-amp – on Linear's page[9].

For those of you not familiar with the *LM10*, it is usable with 1.1 to 40V supplies. It combines a quality low-frequency op-amp, on chip 0.2V precision reference and an adjustable reference booster. Designed by Robert Widlar, when it first became available some eight years ago, it was voted IC of the year, **Fig. 2**.

If converting material quantities between formats, calculating basic physical parameters or typing in the recent program to calculate parallel resistor values, causes problems, try out the conversions and calculation[10] pages on Demon. Here you will find usable real-time on-line calculators answering many needs, **Fig. 3**.

Power Innovations[11] offers new databooks for discrete bipolar devices. This company acquired Texas Instruments' business involved with designing, making and marketing the TI range of discrete bipolar devices. It continues to operate from the Manton Lane, Bedford plant, **Fig. 4**.

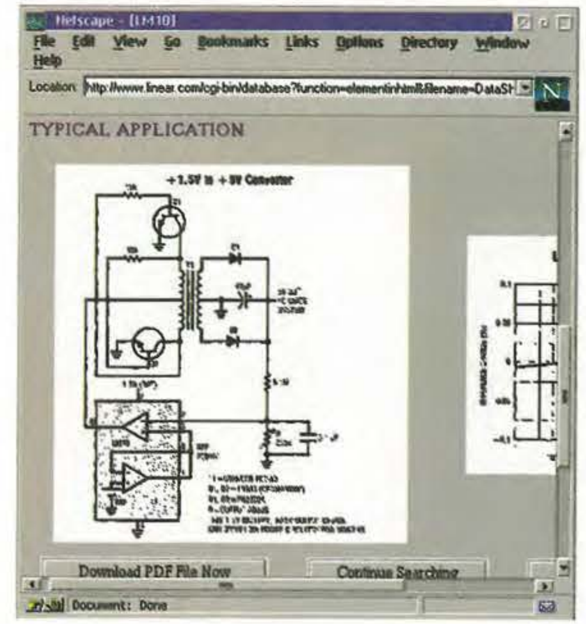


Fig. 2. A stable 5V supply from a single 1.5V battery. A unique application of a unique chip.

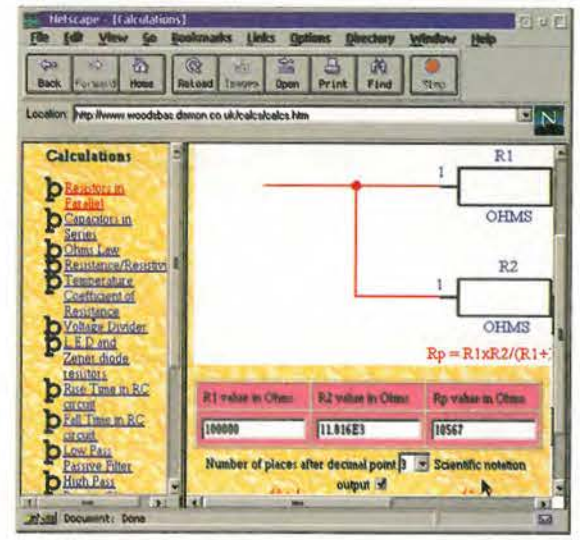


Fig. 3. An easy route to solving many calculation and conversion problems. Also stores many physical constants definitions.

Simulation and design software

Recent articles have discussed the merits of using simulators to predict distortions resulting from bipolar and mosfet devices. The Macquarie[12] University of Australia, has an ongoing project to fully investigate, measure and characterise, distortion and intermodulation performance of microwave transistors.

Not all circuit simulation software is based on Spice. As discussed last month, many vendors specialise in providing frequency-domain-only simulation to extremely good effect.

An interesting and quite different system I examined some years ago has resulted in a mature and reliable general purpose simulation engine that is able to provide results when other engines fail. The *Saber* engine from Analogy claims to be world leading technology, providing analogue, mixed-signal, even mixed-technology simulation for electronic design, **Fig. 5**.

This level of simulation technology demands particular care in model development. Readers interested in exploring this simulator can download a 780Kbyte textbook in PostScript format from the Analogy page.[13] Alternatively, readers can register on-line to request copy of an interactive cd-rom.

Fig. 4. Old and new discrete bipolar devices under new management. For TI bipolar devices read Power Innovations.

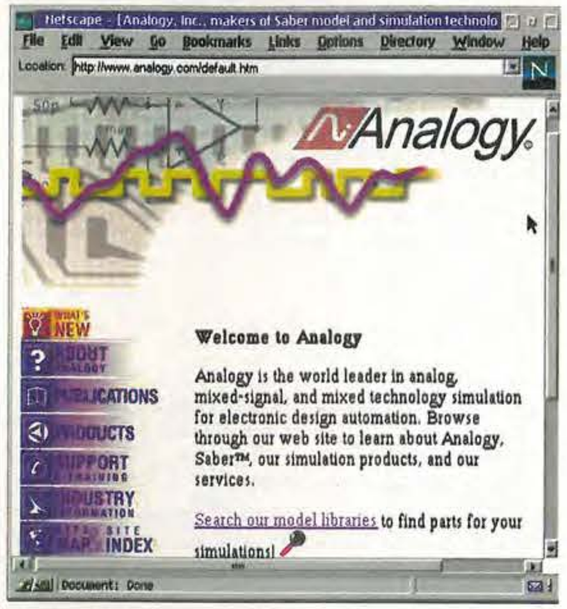


Fig. 5. Saber by Analogy – a mature mixed-technology simulation engine. Try very hard to give it an impossible simulation task.

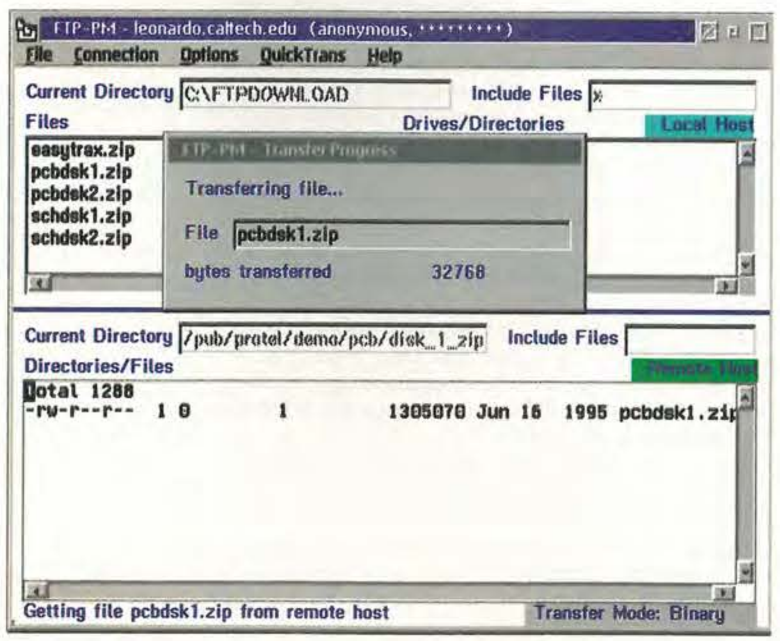


Fig. 6. An FTP client in action – simply open your connection to transfer files. Little more than point and click gives faster more reliable file downloads.

Used and supported on-line at Caltech, the Protel design system can be evaluated by downloading versions for Windows 3.x or 95, either from Caltech[14] or Protel[15]. This system comprises the *SIM 3* mixed-signal simulator, *Advanced Schematic* editor and *Route 3* shape-based pcb autorouter.

Searching the Net

This wealth of application and simulation resource must be unlocked using search techniques. Having found your needed information it can then be downloaded.

As Internet becomes more heavily used, large files download most economically while the USA sleeps and universities and business users are inactive. For UK users this effectively means Saturday and Sunday mornings up to perhaps 2pm. Additionally, when the required file's URL address starts with 'ftp' rather than 'http', for large files, I use a file-transfer-protocol client for transfer, in preference to my Web browser.

By way of example, using my FTP client and starting just after noon on Saturday, I downloaded the 5.2Mbyte for the Protel evaluation, at an average rate of 125kbyte minute via my Motorola 28.8 modem. Total download time was 41 minutes – roughly the cost of a first class stamp. In contrast, a good browser transfer on the same equipment is 90Kbyte a minute. FTP transfer is much less liable to interruptions or dropping out, so maintains a consistent throughput.

Using FTP, you have access to all disk directories designated 'public'. Clicking on a directory/file name, or typing in the desired path, permits change of directory or request of files from the remote host – almost as if it were on your own machine.

If you are unfamiliar with FTP, two points should be noted. FTP servers mostly run UNIX, which uses a forward '/' as a file path delimiter, rather than the '\' used by DOS/Windows/OS2, etc. Two transfer modes are provided – seven-bit text and eight-bit binary. Binary mode must be used to transfer program executables and zipped files. Both these points are clearly visible in the screenshot of my FTP client, **Fig. 6**.

Are you getting garbage?

While performing a search, you may notice apparently meaningless responses listed. An example is an address followed by a few words telling nothing about the page's content. Alternatively, the page listed seems totally irrelevant. How does this happen?

Most search engines list only the first few words used on the page, regardless of their visibility. So if the page starts with an acknowledgement or worse an apology, that is exactly what appears in the listing.

Many search engines also access keywords that can be hidden in controls called 'meta tags'. These are entered to the page when it is drafted and are read by search engines, but invisible to your browser.

The most successful searches result when your search keywords and the meta tag keywords coincide. To view these meta tags, simply save an interesting page as a file, then view using a text editor. Obviously a page drafted using inappropriate meta tags becomes an irrelevant listing for your search.

Naturally, Internet has a wealth of information and comparative reviews of search engine performance. One interesting page, called search engine[16] watch, presents a good starting point for users interested in exploring these topics further. Regularly updated, since search engines also change and develop, it provides a wealth of information for all search engine users. ■

LETTERS

Letters to "Electronics World" Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS

Distorted power interface?

I read Douglas Self's articles on audio power interface/cables with more than a little concern, finding little with which I could agree. However I write not as a general 'gripe' but rather to draw attention to the most serious points.

In his treatment of the series electrolytic capacitors, he seems to completely overlook that all speaker current must return via the power supply capacitors and not simply his convenient ground symbol.

Introducing a lesser value capacitor will obviously have more effect than one which is larger than the supply capacitors.

While on the topic of capacitors, to state the 'magnetic characteristics of a large electrolytic capacitor case are unknown' is utter nonsense. Given a commercial power supply capacitor, the only chemically permissible metal is Aluminium, clearly magnetically well understood.

Much more important, being made from a seamless 'impact' extrusion, this capacitor's case wall makes a near perfect shorted turn for any inductor wound on its outside, disbelievers can easily try this out for themselves.

As to the inductor's self heating affecting the capacitors life – really? Surely, even assuming 3.5A continuous – i.e. 100W into 8Ω – this heating of 0.1225W remains negligible compared to the surface area of the capacitor involved.

In the section on Zobel networks, Douglas remarks on various measurements, none of which are reported. The graphs shown in Figs 6-10 are PSpice simulations – not measured results.

For those of you unfamiliar with PSpice plots, the giveaway lies in the circle, square, triangle etc., trace identifiers, almost a PSpice trademark, which are used by PSpice both on screen and printed to paper. Since his underlying detail assumptions/models used for these simulations are not stated, these graphs become of no value, whatsoever.

As to the 'models' Douglas devised to simulate his speakers behaviour, it seems these were based only on impedance measurements, ignoring phase. Such simple models cannot then be used to calculate network phase responses with any certainty of accuracy at all.

Douglas makes reference to the output inductor in his Blameless amplifier as being 6μH made using 20 turns of 1.5mm wire.

Like many readers, I too built this amplifier using the purchased printed boards. Both in the Feb. 1994 article and the PCB001.DOC which accompanied these boards, this inductor was described as having ten turns, on 1in diameter of 18SWG wire.

The pcb inductor terminations are spaced at 30mm, and their outline is clearly marked. These factors implied that the ten turns should be spaced to suit.

Having today removed one inductor, I measured only 1.42μH inductance at 1kHz, 11.00mΩ resistance at 1A dc. These differences may explain much. For the record, these same ten turns, compressed to eliminate all spacing, measured 2.3μH. If, as Douglas states, he designed for 6μH, I feel some advice and explanation should be forthcoming from him.

In the October article, justifying his statement "coaxial cable is not made with anything like the weight of copper..." by referring only to RG58, is incorrect. In my article to which he refers, I measured and recommended UR67 cable which does indeed have more than adequate copper measuring only 47mΩ for 5m – less than half the resistance and inductance of Douglas' preferred 13A cable.

UR67 is low cost and freely available in short runs. Its professional brother RG214 has a double thickness outer braid offers even lower resistance.

I notice that Douglas has referred only to the inner core of RG58. But coaxial cable outer braid may well contain much more copper than the inner core, so the combined resistances must be taken into account.

While discussing cables, Douglas had problems with the lack of correlation between copper section *versus* current ratings. With the domestic cables he used, current rating is determined by permissible temperature rise – not voltage drop as he suggests. This permitted temperature rise varies according to insulation used, expected duty and whether housed in trunking or used in free air etc.

As to Douglas' comments regarding amplifier hf instability, this I have experienced using an amplifier having a normal zobel network and an output inductor, when used with a very low impedance experimental cable driving a two way loudspeaker. This cable was built following my last cable article, so has not been publicly described.

Capacitance for 5m of this cable, measured less than 2nF – much much less than the 100nF Douglas suggests is needed. For the record, I left one channel of this amplifier, cable, speaker combination running with the Fireworks track of the *Hi-Fi News*, test cd, at normal listening levels. This channels output transistors fried. The second channel was used my MkII coaxial cable, with no problems whatsoever. I expect to build on this work to identify the mechanisms causing amplifier hf instability, however this is for another day when I find time to recommence my work on cables.

Douglas has at last fallen into the subjectivist camp's trap. Having recently put down the 'magic diodes' concept, to explain behaviour not explained by his inadequate models, he has invented 'magic semi-inductors.' Whatever next?

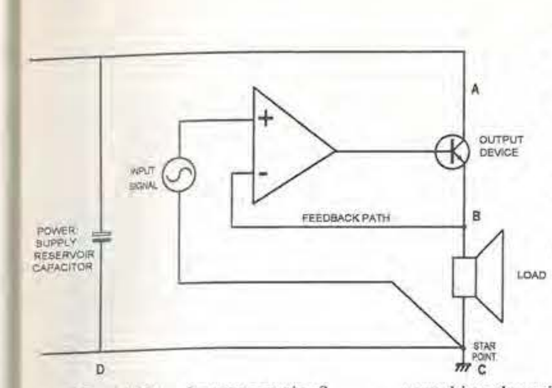
Douglas replies:

I can reassure Mr Bateman (Letters) that I have not forgotten that speaker currents return eventually to the reservoir caps; however this is irrelevant to the purity of the signal applied to the loudspeaker.

My article of Sept 1997 showed that virtually all the distortion disappeared when the output capacitor was removed or made very large. If the reservoir capacitors had a role like the output capacitor, one might have expected the distortion to be halved, but it would not have vanished. I'm afraid this is enough in itself to show that Mr Bateman's thinking is incorrect.

The output current from the loudspeaker does indeed return to the reservoirs. However, as the diagram of the top half of a Class-B amplifier shows, it first goes to the star point C, which is by definition the signal reference. The amplifier, through the beneficent workings of negative feedback, controls the output voltage B so it is clean and undistorted with respect to point C; it is the voltage B-C that matters. The voltage B-D does not. What happens to the current afterwards-well, who cares? Its job is done, and the linearity of the reservoir caps is supremely irrelevant.

You might as well argue that all amplifiers must distort because the return current has to go through the bridge rectifier to get back to the transformer secondary whence it came. And what about the linearity of Ferrybridge A power station?



As this diagram of the top half of a class B amplifier shows, negative feedback helps keep output voltage B clean with respect to point C – and voltage B-C is the one that matters.

A good amplifier is supremely indifferent to noise, hum, and fluctuations on the supply rail at A, and I do not accept that any such amplifier could show changes in quality of reproduction when the supply reservoirs were changed – assuming of course that they were working in the first place.

Electrolytics can and, if you are careless, do generate low-frequency distortion in small-signal circuitry. What this has to do with capacitors that do not have audio voltages impressed across them is unclear. Worrying about the audio properties of power-supply capacitors is quite pointless.

Mr Bateman claims I said "the magnetic characteristics of the capacitor case are unknown," and apparently he thinks this is nonsense. Me too.

What I actually said was "the magnetic characteristics of the capacitor are unknown" which they are; there is at least the possibility of extra losses and heating in the electrolyte. A little more care in quotation would be desirable here.

He makes a good point when he says that the can might act as a shorted-turn; whether this would have any practical effects I do not know, but the obvious danger is that eddy currents will heat up the capacitor further.

I know from experience that even heavyweight output inductors can and do get hot to the touch when an amplifier is working hard into 4Ω, and I cannot for the life of me see why he thinks this is so laughably unlikely. Big power electrolytics are expensive, and wrapping a cosy heating coil around them is not the way to maximise their lifetime.

The magnetic misbehaviour or otherwise of electrolytics used as coil formers is perhaps not worth arguing about, since hopefully no-one does it any more. It is bad practice.

My section on Zobel networks does not make any reference to measurements; I have merely reported stability or otherwise for various cases. Figs 6-10 in the article relate to the output inductor and are nothing whatever to do with Zobel networks. These are indeed PSpice simulations, though why Mr Bateman should think there is something clever in deducing this defeats me; admittedly the axes have been redrawn at the editorial stage, but it is still quite clear that they are not AP plots. The simulation used an ideal voltage source to drive the circuit in Fig. 1 of the article. While this does not include every possible interaction, eg effects of finite slew rate, it shows the major effects- for the first time, I believe. I think this is quite clear from the text, and to say these graphs are "of no value" is as ungenerous as it is incorrect. I must confess to some lack of rigour in specifying the output inductor. I did indeed originally recommend 10 turns of 18SWG wire, the latter because it was easily available from Maplin.

All the prototypes were however made up with 20 turns of 1.5mm wire close-spaced, which does indeed give 6μH. However, as my article showed, while this inductance is an important factor in artificial capacitive load testing, its value is not critical for stability with even highly reactive loads, and 2μH seems to be quite enough. I do try to be totally consistent on every point in every article, but being human, I sometimes fail.

Finally, I am very unimpressed by Mr Bateman's jeers at "magic semi-inductors" (his phrase, not mine). It has been known for many years that the voice-coil impedance of a loudspeaker cannot be accurately represented by a pure inductance, because of eddy-current effects in the pole-pieces, etc. This phenomenon is inconvenient for simplistic theories, but its existence is not in doubt. If Mr Bateman had troubled to read any of the references I gave on this – I repeat two very good ones below – he would have found the matter clearly explained.

1. Vanderkooy, J, 'A Model of Loudspeaker Driver Impedance Incorporating Eddy Currents In the Pole Structure,' *JAES*, March 1989, p. 119.

2. Wright, J R, 'Empirical model for loudspeaker motor impedance,' *JAES*, Oct. 1990, p. 749.

Spark erosion

Michael Ward's perceptive and lucid feature (*EW* Dec '97) showed how electronic ignition can improve engine combustion efficiency, but begged a question inherent in the use of high energy spark ignition systems-spark plug electrode erosion.

The problem has been around for nearly 100 years, and led to the use of noble metals like platinum to reduce ignition timing drift as spark gaps eroded.

There is a way of improving ignition ok lean mixtures which takes advantage of electrode erosion.

Soft electrodes produce larger sparks which contain vapourised metal which acts as a catalyst to promote ignition of weak mixtures. Some work was done in this area in prewar aircraft engines to see if larger sacrificial spark electrodes could improve high altitude performance, but work ceased with the adoption of the exhaust driven turobo-charger.

Perhaps it's time to revisit the sacrificial electrode 'catalytic' spark plug electrode idea now electronic ignition can cope with larger gap capacitance and surface area -especially if the new ignition systems mean ordinary plugs wear out more quickly.

Anthony Hopwood
Holdfast
Worcestershire

Watch which PIR

Ian Hickman has produced a fascinating project with his Video surveillance system – November 1997 issue – one which might just tempt me to build something, after years dreaming up ideas for similar units.

However, if he intends to drive an audible alarm from externally-mounted PIRs intended for lighting then severe problems with false alarms might be expected. These units are usually characterised for sensitive detection without false alarm suppression. Little nuisance is caused due to a light switching erroneously.

The alternative is to use types designed expressly for burglar alarm application, which are easily available through catalogue distributors as well as general wholesalers. Besides giving a switch-selectable number of counts before activation, they have other advantages for this project. The 12V dc power required – already available for the cameras – offers smaller size, easier wiring installation, and the avoidance of mains-voltage interface hazards. An alarm relay contact with separate anti-tamper circuit is standard. The principal disadvantage is that most are intended for indoor installation, requiring sheltered mounting and perhaps a tube of silicone sealant.

R G Newman
Hampton Hill, Middlesex

Misunderstanding capacitors

In the panel on page 1001, *EW* December 1997 I read, "Each practical capacitor also incorporates an inductive element..."

In the article 'Displacement Current', *Wireless World* – the same journal – December 1978 I read, 'Series inductance does not exist. *Pace* the many documented values for series in a capacitor, when the so-called series inductance of a capacitor is measured it turns out to be no more than the series inductance of the wires connected to the capacitor. No mechanism has ever been proposed for an internal series inductance in a capacitor.'

Curiously, each article has identical descriptions of a capacitor. 1978 "...any capacitor has now become a transmission line..." 1997, "What is a capacitor? Any two conducting surfaces, separated by an insulator..."

What's going on? Before "Understanding Capacitors", *EW* needs to understand itself.

Penelope Lyon
Redbourn
Herts

Cyril replies:

Thank you Penelope. Both the Ivor Catt 1978 article and my article of December 1997 are correct. Capacitors certainly do behave exactly as transmission lines and all capacitors *and* all transmission lines possess measurable inductance as well as capacitance, (see page 999, Properties of Capacitors).

Indeed I first proposed in internal company reports that capacitors be considered transmission lines as long ago as 1968 while employed as design engineer for electrolytic capacitors at the old Erie company.

In the company's 1975 catalogue I wrote "Inductance-L. This is due to the lead wires and tab connections to the element winding. With appropriate construction, the inherent inductance of the element will be insignificant in comparison"

As to Ivor's article he is thus perfectly correct in stating in 1978 that a measurement of a leaded capacitor's self inductance is dominated by that of its leads. Indeed with equipment then available and measurement of any physically small, low-K non-electrolytic capacitor, no other conclusion would be valid.

However with the passage of time better measurement equipment and non-leaded capacitors are now commonplace. For example the measured self inductance, reported in the makers data, of a typical 1206 size surface mount ceramic chip capacitor is 1.5nH, i.e. less than its body length of 0.6mm lead wire.

The measured inductance of this same capacitor element with two 6mm long 0.6mm diameter lead wires is around 16nH. The lead wire alone has a self-inductance around 8nH/cm. The difference is accounted for by additional wire length inside the capacitors housing, needed to make connections. While both are extremely low values, clearly the lead wire inductance dominates.

More recently, as an rf engineer, I consider as in my article, that any two conducting surfaces are both a capacitor and a transmission line – albeit possibly of very short electrical length.

You may ask why this transmission line view was not presented in my capacitor article. The December article is the first of a short series and was intended as an overview for all capacitors. Secondly, for low-K, physically small, non-electrolytic capacitors this self-inductance dominates. With high-K capacitors used at high frequency and larger capacitors used at lower frequencies, then transmission line effects can dominate.

By way of example a high-K tubular ceramic capacitor commonly used for emc suppression, having diameter of 4.45mm and electrode length of 7.5mm, becomes open circuit resonant at 200MHz. This effect is clearly visible on its published and measured insertion loss claims.

This discussion will be expanded in a forthcoming capacitor article, in the meantime I hope this answers Penelope's question. ■

Test power valves

Examine how an af power valve performs both statically and dynamically using Theo Argiriadis' valve tester.

This meter performs static and dynamic tests on power pentodes and beam tetrodes such as the EL34, EL84, 6L6, 6V6, 6550 and KT88. It enables you to match valves and diagnoses faults. Although not a precision instrument, it gives a clear picture of the condition of a valve via the following.

Detection of catastrophic faults. If a shorted valve is suspected, a light bulb can be inserted between the valve under test and the ht to act as a current limiter. Screen current I_{SCR} is limited via high voltage transistor circuitry. A miniature speaker detects noisy and mechanically faulty valves.

Emission testing. This is carried out by measuring the anode current I_A at any value of grid to cathode voltage, V_{gk}.

Power testing. The valve is made to deliver a small amount of power, which can be measured. This is the quickest and most effective test in my opinion. Depending on the valve and its condition, this amount ranges from 4.5W for a good 6L6 to 8.5W for a good KT88. The measurement can be used as a comparison standard in fault-diagnosis

Amplification factor. The tester gives the $\mu_{(T)}$ and rms values of output signals that the valve produces under load, $V_{a(T,L)}$) and no load, $V_{a(T)}$, conditions in triode mode. Triode and pentode transconductances $g_{m(T)}$ and $g_{m(p)}$ can be derived from these mathematically.

This tester allows valves to be matched in terms of,

- I_A versus V_{gk} in pentode or triode mode.
- μ, r_a and g_m through triode-mode dynamic testing.

In pentode mode, anode voltage V_A is 320V and screen voltage V_{SCR} is around 250V. Many published output characteristics are taken at this value. In triode mode, V_A is about 250V. In my experience, if a set of valves match both statically and dynamically under these conditions, they will also match at different V_A and V_{SCR} values, whether in pentode or triode mode.

Valve parameter variation from device to device of the same type seldom exceeds 10%. Nevertheless, both anode and screen terminals can be switched to external power supply sources. Test points are indicated to allow you to observe anode and grid signal voltage and cathode current waveforms, and to measure screen current.

This circuit can be modified to test directly-heated power triodes like the 211, the SV811 and the 845 – and even transmitting beam tetrodes like the 4-250A. Components will have to be of higher voltage ratings. Anode, grid bias and drive voltages will also be higher, but the principle remains the same.

Valve failure modes

The following are catastrophic valve faults, which will rapidly lead to failure.

Switch functions

- S_1. Mains switch.
- S_2. Switches ht on and off; can be used to remove the ht while changing valves.
- S_3. In off position, light bulb is in series with anode circuit.
- S_4. Selects between internal or external ht for valve under test's anode.
- S_5. Selects between raw 330V and stabilised, current limited 250V ht.
- S_6. Same as S_4 but for the screen grid.
- S_7. In triode mode, links the screen grid to the anode through a 100Ω resistor.
- S_8. When on, connects a resistive load across the output of the valve being tested.
- S_9 Speaker switch.
- S_{10}. When off, a 560Ω resistor is in series with the screen.

Test points

- $TP_{1,2}$. Voltage across these points divided by R_{12}, at 100Ω, gives screen current in milliamps.
- TP_3. Grid waveform view point.
- TP4. As TP_3, but for the cathode. Also through dividing the dc voltage at TP_4 by R_{13}, which is 10Ω, cathode current can be obtained. This is the $I_{SCR}+I_A$ sum which is necessary when converting triode parameters to pentode parameters.
- TP_5 Allows output waveform to be observed. A true rms meter at this point will produces more accurate results.

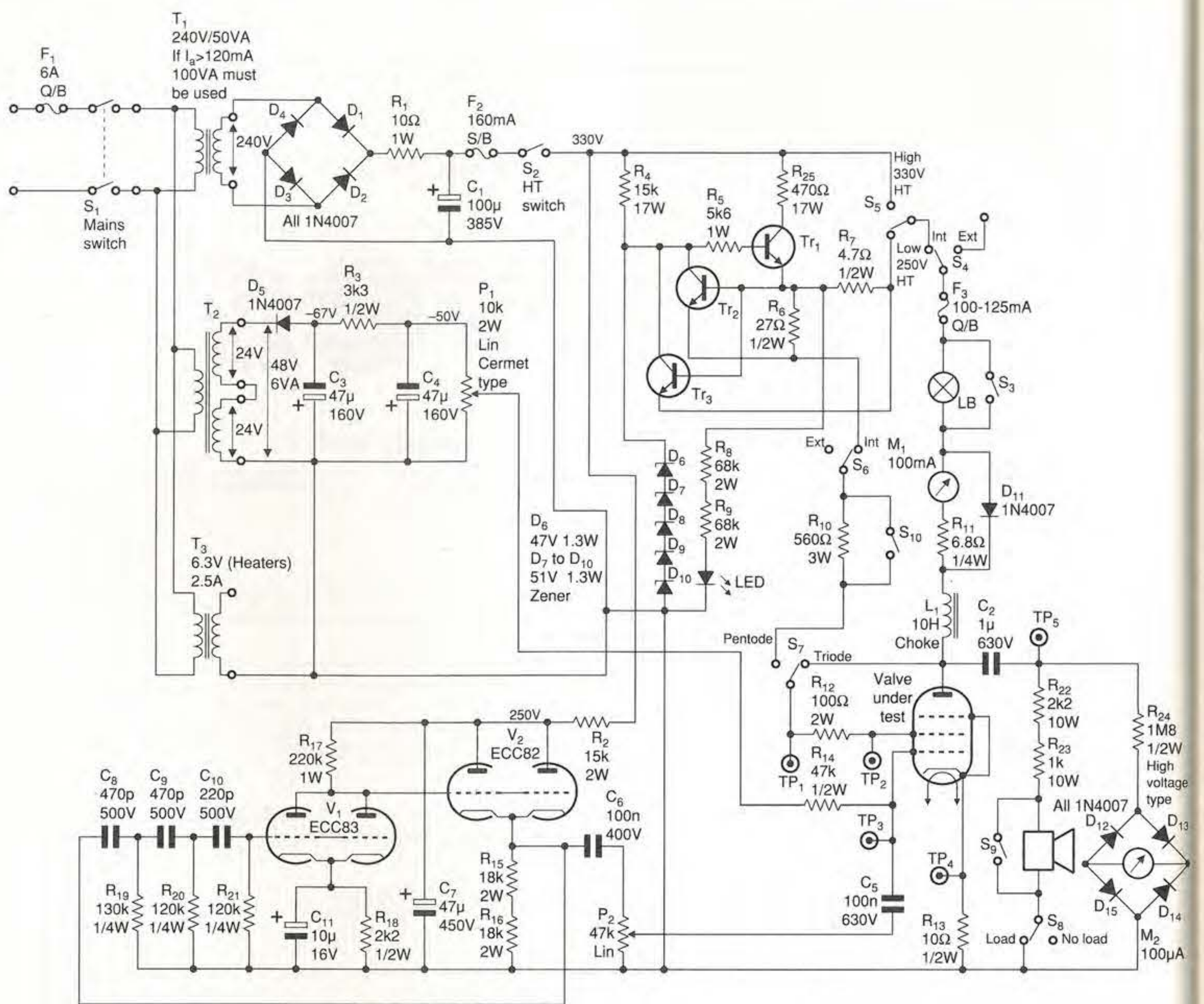


Fig. 1. Complete tester for audio power pentodes examines static and dynamic performance in both pentode and triode modes.

Internal arcing and shorts. High dc and signal voltages can cause internal arcing which may lead to shorts. High temperature thermal cycling combined with mechanical vibration of the elements within the glass envelope – guitar amps in particular – can also contribute to a short.

Control-grid burn out. High voltages and temperatures can cause the control grid to emit large amounts of electrons or, if it becomes too positive, draw excessive amounts of current for its size. Since the grid is made out of very fine wire, it is easily burnt out. Control grid to cathode shorts are also common because the grid is physically very close to the valve. Without the control grid, or without grid biasing the cathode will glow red hot and the valve will soon fail.

Screen-grid burn out. High screen voltage causes high screen current and excessive dissipation, eventually damaging the grid. Without the screen, there is no way of providing the necessary acceleration for the electrons to reach the anode. The valve will hardly conduct any current or produce any signal.

Anode burn out. This is due to high V_A, I_A or over-dissipation P_{DA}.

High peak signal amplitudes combined with the reactive nature of the output transformer/speaker load can damage the anode – especially if the speaker is accidentally disconnected during loud volumes.

Filament burn out.

Air leaks.

Ageing and other faults

These faults are mainly due to valve ageing and cause erratic operation or fall-off in performance.

Emission loss. This fault is due to erosion of the cathode coating material. A valve with low emission takes longer to reach its operating conditions. It also conducts less anode current for a given value of grid-cathode voltage and screen voltage than it would have when new. Emission loss is linked to μ and g_m fall and in amplifiers biased more towards class B will cause crossover distortion.

Rise in r_a, fall in μ and g_m. This fault is mainly due to ageing. Lower μ results in less amplification. Higher r_a and/or lower g_m reduces the amount of signal that a valve produces across a load – ie its power output.

Electrode leakages. Insulation breakdowns cause electrode leakages. Heater to cathode leakage manifests itself as mains hum. Grid to cathode leakage causes the bias to drift to lower values, resulting in thermal runaway. In this case, the valve gets red hot.

Noisy operation. Microphony and crackling noises when the valve is moved are caused by mechanical problems – ie loose elements. Such faults are mainly due to thermal cycling and vibration.

Gassy valve. With power valves, some blue glowing within the anode metal structure is normal. Gassy valves exhibit an excessive blue glow.

Sometimes a valve may survive temporary overheating, arcing or shorting problems and continue operating for a while. Also, a valve with a grid to cathode leakage may be made usable by lowering the operating voltages. It is however advisable to replace such valves because they may damage other parts of the circuit.

How the circuit works

In pentode mode an anode voltage between 320 and 330V with a screen voltage of 250 to 260V is supplied to the valve under test. This assumes that S_4 and S_6 are set to internal, S_5 is set to high and S_7 is set to 'pentode'. When testing triodes, S_7 is set to 'triode' and S_5 set to 250V. This reduces output distortion and consequently gives more accurate measurements.

A variable negative bias is applied to the valve under test's control grid. Meter M_1 measures the anode current I_A. Resistor R_{11} protects the 100mA meter movement when I_A exceeds 100mA by developing a voltage drop of 0.1A×6.8Ω, or 0.68V. This forward biases D_{11} which then shunts the meter. In my prototype, the meter had a dc resistance of 0.8Ω and R_{11} was chosen empirically.

Transistor Tr_1 buffers the zener chain $D_{6\text{-}10}$, providing a stabilised low output impedance 250V source, for the screen and/or the 'low' ht setting. This must be a high voltage transistor with enough gain to stop it seriously loading the zener chain.

Out of a sample of ten *BUX84s*, I measured an average gain of 80 at between 3 and 100mA. The *BU426A* behaved similarly and it has a higher dissipation. The *BUT11AF* has less gain, but it is an isolated case type, which is an advantage in high voltage circuits.

I used a *BUX84* for Tr_1 and *BUT11AFs* for $Tr_{2,3}$.

If a short appears at the low or high current output of this supply, Tr_2 or Tr_3 turn on and remove base current from Tr_1. The output drops to zero and nearly all of the 330V appears across R_4. As a result, this component must be a 17W wirewound type.

If 100mA is drawn from Tr_1 during a triode mode test, R_{25} dissipates 4.7W so I've used a 17W type. The purpose of this resistor is to reduce Tr_1 dissipation and V_{CE}, particularly during high current surges, in case of a short for example. Without it Tr_1 will eventually fail.

The led must be a low current, 2mA, type so that it does not load the screen supply.

The valve under test is connected as an inductively coupled common cathode amplifier. With S_3 on, the ht feeds the anode through a 10H choke. The anode is ac coupled to a resistive load comprising R_{22+23}. If accurate measurements are required, the exact value of this load is important. The series R_{22+23} combination I used was 3.12kΩ, and the μ, g_m,

Analysis

Triode-mode parameters. A triode can be modelled as a voltage generator $\mu_{(T)}v_g$ having an internal resistance $r_{a(T)}$, **Fig. 2.**

With R_L disconnected,

$$v_{a(T)} = \mu_T v_g \quad \text{so,} \quad \mu_T = \frac{v_{a(T)}}{v_g}$$

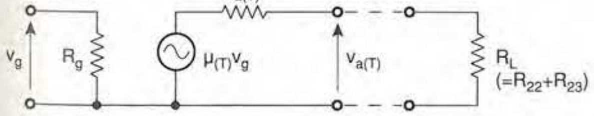


Fig. A. Model of a pentode valve connected as a triode.

When the valve drives load R_L, the anode signal is,

$$v_{a(T,L)} = \mu_T v_g \frac{R_L}{r_{a(T)} + R_L}$$

Solving for $r_{a(T)}$,

$$r_{a(T)} = R_L\left(\frac{v_{a(T)}}{v_{a(T,L)}} - 1\right)$$

and the expression for $g_{m(T)}$ is,

$$g_{m(T)} = \frac{\mu_T}{r_{a(T)}}$$

Pentode mode parameters

Pentode transconductance $g_{m(p)}$ can be derived from $g_{m(T)}$,

$$g_{m(P)} = g_{m(T)}\left(\frac{I_A}{I_A + I_{SCR}}\right)$$

In the case of this valve tester, the value of $g_{m(P)}$ derived from this equation will assume that both screen and anode are at 250V. Certain assumptions are necessary for these conversions so for better accuracy, see reference 3.

A pentode can be modelled as a current generator $g_{m(P)}v_g$ having an internal resistance $r_{a(P)}$. Voltage $v_{a(P)}$ is then,

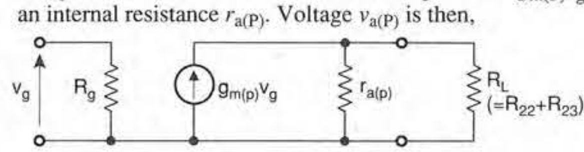


Fig. B. Model of a pentode.

$$v_{a(P)} = g_{m(P)}v_g \frac{r_{a(p)}R_L}{r_{a(p)} + R_L}$$

Solving for $r_{a(p)}$,

$$r_{a(P)} = \frac{1}{g_{m(P)}\left(\frac{v_g}{v_{a(P)}}\right) - \frac{1}{R_L}}$$

For this last equation to work, v_g must not be higher than 1V rms. If it is, the pentode output distortion will introduce serious errors in the reading on meter M_2.

r_a values shown are based on that value.

Choke L_1 is a 10H Maplin ST28 rated at 50mA. At 1kHz, its inductive reactance X_L is high enough to be considered an open circuit in both unloaded triode mode and loaded pentode mode. This choke will work up to 100mA and can withstand 150mA.

A sine wave of around 1kHz is derived from an *RC* phase shift oscillator based on an ECC83. This is fed through an ECC82 cathode follower to the signal control potentiometer that drives the valve under test's control grid.

Both triodes in the ECC83 and ECC82 are connected in parallel. This in effect halves the ECC83's anode resistance and increases its gain. In theory, for this circuit to work the gain must be 29, but in this application I experimentally found out that more gain is required to ensure oscillation under all conditions. The signal across R_{14} may be as high as 100V pk-pk and must be able to enter the valve under test's grid current region.

The ECC82 here works as a low power driver and the parallel arrangement overcomes any grid current problems.

Capacitor C_{10} is only 220pF since the parallel ECC83 Miller capacitance is around 260pF.

Internal shorts, leakages and arcing

If you have reason to believe that a valve suffers from problems such as internal shorts, leakages, arcing, etc, S_3 must be left open so that the light bulb is in series with the anode circuit to act as a limiter. I have recently tested three 6L6s – two with burned control grids and one with control grid to cathode leakage. In the first two instances, the bulb came on at full brightness within 30s. In the third, the valve had to be biased and left on to conduct an I_A of 60mA for five minutes for the leakage to show; gradually the bulb got brighter and the I_A increased to 100mA. An external voltmeter at TP_3 indicated a drop in V_{gk} occurring simultaneously.

Table 1.

Pentode Valve under test		V_{gk} (V)	I_{SCR} (mA)	I_A (mA)	v_g (V rms)	P_{out} W	Condition	Triode connection V_{gk} (V)	I_{SCR} (mA)	I_A (mA)	$V_{a(T)}$ (V rms) no load	$V_{a(TL)}$ (V rms) loaded	$\mu_{(T)}$	$r_{a(T)}$ Ω	$g_{m(T)}$ mA/V	$g_{m(p)}$ mA/V
A	Unused but old	–23	3	30			OK	–20	4	30	95	66	9.5	1.37	6.9	6.1
	Mullard EL34	–18	7.7	70	6	6.6	OK	–15	9	70	102	78	10.2	958	10.6	9.4
B	Old used slimline	–22	2.5	30				–20	3.3	30	100	68	10.0	1.47k	6.8	6.1
	Sovtek EL34	–17	6.5	70	6	6.7	OK	–13	8.5	70	108	81.6	10.8	1k	10.8	9.6
C	Old used Tesla	–28.7	2.5	30				–23.6	4	30	81	54.6	8.1	1.51k	5.3	4.7
	EL34	–21	7	70	6	4.9	*	–16.2	9	70	92	70.6	9.2	969	9.5	8.4
D	Mod. used 'STR'	–18	2.5	30				–14.4	4.8	30	122	72.0	12.2	2.16k	5.6	4.8
	Sovtek EL34	–12	9	70	6	4.5	**	– 7.4	13	70	125	87.0	12.5	1.36k	9.2	7.8
E	Used Chinese	–25	1	30				–22.6	1	30	78.1	47.3	7.8	2k	3.9	3.8
	6L6	–18.5	2.4	60	8.5	4.8	OK	–15.4	3.1	60	85.1	58.4	8.5	1.43k	5.9	5.6
F	Old/used	–25	1	30												
	orig. GE 6L6	–20	2	60	8.5	4.9	***									
G	Almost new	–31	1	30				–28.8	1.5	30	68.9	47.0	6.9	1.45k	4.8	4.6
	Sovtek 6550	–24.1	2	70	7.1	8.7	OK	–20.1	3.4	70	74.9	57.7	7.5	932	8.0	7.6
H	Mod. used, good	–29.7	1	30				–27.5	1.1	30	74.5	50.3	7.5	1.5k	5.0	4.8
	qual. Chinese KT88	–23.2	2.3	70	7.8	8.4	OK	–23.0	1.9	50	76.4	56.5	7.6	1.1k	6.9	6.6
								–20.0	2.6	70	80.0	61.1	8.0	964	8.3	8.0
								–14.9	4.3	100	84.8	67.7	8.5	790	10.8	10.4
I	Very old, used	–33.1	1	30				–30.0	1	30	67.2	45.4	6.7	1.5k	4.5	4.3
	GEC KT88	–24	2.5	70	7.8	6.9	****	–24.4	1.9	50	71.7	53.5	7.2	1.1k	6.6	6.4
								–21.5	2.8	70	73.6	57.2	7.4	895	8.3	8.0
								–16.6	4.5	100	77.8	62.7	7.8	751	10.4	10.0

Notes

* This valve is noisy. Biased for anode current between 30 and 70mA, when it was lightly tapped, both meters deflected and noises were heard from the loudspeaker. Static tests do not indicate emission loss but the power it produces is low compared to valves A and B.
Triode tests show a fall in amplification $\mu_{(T)}$ which results in lower transconductance. According to published data, a triode connected EL34 has a μ_T of 10.5 and a $g_{m(T)}$ of 11.5mA/V for an anode voltage of 250V and anode current of 70mA.

** This valve has lost emission since V_{gk} must be set rather low for the 30mA and 70mA anode current. It also took five minutes for the anode current to reach these values. Output power is low and the triode tests revealed that although the μ_T is high, the $r_{a(T)}$ has nearly doubled. To calculate $r_{a(p)}$ the valve was connected in pentode mode again – S_8 in load position – with 1V rms at the grid.
Anode signal measured 22V rms. Then formula for $r_{a(P)}$ was applied to give an $r_{a(p)}$ of 29.4kΩ under similar operating conditions. Valve A, an EL34, produced a 25V rms output with 1V rms at its grid in pentode mode. This equates to an $r_{a(P)}$ of 18kΩ. It is interesting to note that in valve D, $g_{m(T)}$ and $g_{m(p)}$ have not fallen dramatically due to the high μ_T. This valve however cannot deliver power due to the rise in its anode resistance. This is a common problem in old – and sometimes new – valves. Some commercially available testers fail to detect it because they only test for transconductance and static emission.

*** This valve tested as acceptable in terms of emission and power but it was cutting off intermittently. Anode current would drop to zero at random.

**** Amplification, μ_T, is low in this valve, which arced internally with an anode voltage of 500V and screen voltage of 350V. Switches S_4 and S_6 were set to their external positions.

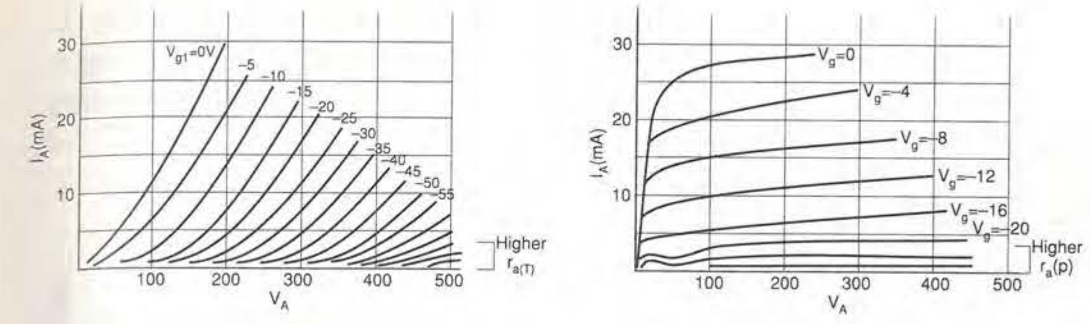


Fig. 2. Characteristics for the KT88 power valve connected as a triode, left, and as a pentode, right. Note that the curves for the pentode only apply for 200V screen voltage. Varying the screen voltage causes the curves to change.

A mains 40W bulb draws 167mA at 240V rms, but it can withstand just over 440V dc. Similarly, a 25W bulb draws 100mA and can handle over 330V dc. With S_4 switched to 'internal ht', a 25W one is a good choice. Fuse F_3 can be a 100mA quick-blow type to offer further protection.

In valves like the KT88/6550 internal arcing does not normally take place at a HT as low as 330V. Switch S4 can then be switched on to an external higher voltage power supply, preferably to the one in the equipment from which the suspected valve came from, with a 40W bulb in series.

If the arcing is so severe that it results in an internal short, the worst that could happen is that F_3 and the bulb may blow.

Note that for high voltage arcing to occur, anode current does not need to be high. Turning P_2 fully clockwise, with S_8 set to 'load' position, can help to trigger arcing if the fault is intermittent. When S_6 is switched to 'internal', if there are any shorts in the screen terminal, the led goes off.

Static emission tests

Potentiometer P_1 must initially be set to maximum i.e. so that V_{gk} is –50V. This voltage must then be gradually reduced while monitoring the anode current.

Emission tests at an anode current of say 30, 60, 70 and 100mA can be carried out for instance every 6 months. The corresponding V_{gk} values can then be recorded in order to monitor the emission fall – if any – of a certain valve over a long period of time.

If any tests are to be performed on EL84 or 6V6 valves the anode current must not exceed 40mA, and switch S_5 must be set to the 'low' 250V position. I would also recommend a 560Ω screen limiter resistor R_{10}, for these valves, unless they are only to be tested in triode mode.

When P_1 is set so that each valve under test produces an anode current of 30mA, the time it usually takes for a good valve to reach this value is as follows,

EL34 slimline/6L6/EL84	30-40s
KT88/EL34 (6CA7 'STR' type)	50-60s
6550	60-70s

Testing for power

When testing power, S_8 must be set to 'load' position. For the output power tests in **Table 1**, anode current was set to 70mA for the EL34/6550/KT88s and 60mA for the 6L6. Ten valves of each type that were known to be good were then tested. On average, the EL34s produced 6.5W, the 6L6s produced 4.9W and the 6550 and KT88s produced 8.5W. These are approximate values since the purpose of this test is quick diagnosis rather than exact measurement. These power values were then taken as a comparison standard for the measurements in the table.

The amount of signal drive at the control grid was adjusted so that during the power test, anode current increased by 10mA and screen current by 2 to 3mA. This increase is due to the second harmonic component of the output signal.[3]

If drive signal v_g increases further so that its peak amplitude becomes greater than V_{gk}, I_A will start decreasing. This is due to grid clipping – caused by grid current – during the peak of the positive cycle of the drive signal. The negative cycle is then higher in amplitude and introduces a negative dc component at the grid. This action reverse biases the valve and is an additional, quick test since any good valve must behave this way.

Measuring in triode mode

To measure μ_T and the no-load output signal $v_{a(T)}$, S_8 must be off. Most triode connected power pentodes have an $r_{a(T)}$ of between 700 and 1500Ω. At 1kHz the reactance X_L of a 10H choke is 62.8kΩ, and the effect that it will have on the measurements will be negligible.

Even if the inductance of the choke falls to say 5H due to core saturation at 100mA anode current for instance, X_L will still be 20 times more than the highest $r_{a(T)}$ of a good valve. In pentode mode the anode resistance is too high and X_L introduces a serious error.

Also, in triode mode, valves produce a more accurate, less distorted sine wave. In pentode mode, the distortion is too high and will introduce calibration errors in meter M_2.

Finally, triode mode parameter testing can be useful since most hi-fi amplifiers incorporate triode or ultra-linear connected pentodes.

KT88 tests

As you will see from the table, I tested two KT88s at four anode current values to demonstrate the fact that both $g_{m(T)}$ and μ_T increase at higher anode currents. Similarly the anode resistance increases at low anode currents because of the reduction in the,

$$\left.\frac{\Delta V_A}{\Delta I_A}\right|_{V_{gk} = const}$$

anode-voltage gradient in both triode and pentode mode, **Fig. 2**. According to data, a good KT88 has an $r_{a(T)}$ of 670Ω, a $g_{m(T)}$ of 12mA/V and a $\mu_{(T)}$ of 8 at an anode voltage of 250V, but at the high anode current of 140mA. ■

Further reading

1. *Vacuum Tube Valley* magazine, 1095 E. East Duane Ave, 106 Sunnyvale, CA94086.
2. *Glass Audio* magazine, Dept NW6, Audio Amateur Publications, Peterborough, HH03458-0576.
3. Langford-Smith, F, *'Radio Designer's Handbook'*, pub. Newnes.

In this, the first of three articles, Richard Lines looks at problems and solutions relating to keeping objects at a particular temperature.

Precision temperature control

This article – and its two follow-ons – arose out of development work we were carrying out involving astronomical instruments for studying the sun. One of the key design criteria was that the instrumentation must be very stable over time.

Since the equipment is located at unmanned sites that are open to the weather, the instruments experience a daily temperature change of typically 20°C. To this must be added seasonal changes, so the instrumentation has to be shielded from a 30 to 40°C variation throughout the year. Without suitable regulation, the drifts induced would have ruined the data and made analysis difficult, and maybe even impossible.

There are many other applications in science and industry where precision temperature regulation is an important, albeit secondary criterion. If you are to become involved with temperature control – particularly if you have not have previous experience of the subject – you should find these notes useful. The emphasis in here is very much on practical aspects; others have already written excellent textbooks on control theory. On that topic, I found 'Control Engineering' by W. Bolton particularly useful.

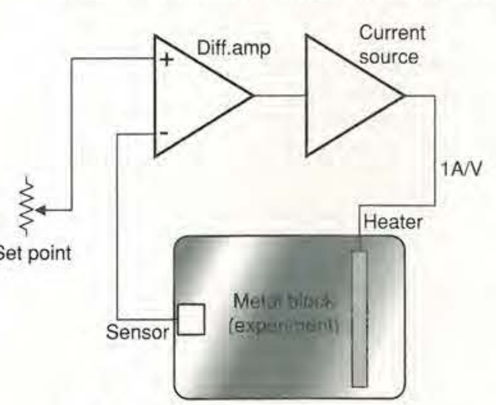


Fig. 1. Possibly the simplest form of temperature controller in which the sensor voltage is compared with a reference. If the sensor voltage indicates that the metal block is too cool, the output of the differential amplifier drives up current from the source.

Expected performance

The term 'precision' deserves some explanation. The achievable temperature stability clearly depends on how well heat that leaks into and out of the item concerned can be controlled – i.e. the quality of the thermal insulation – as well as the ability of the controller to correct.

It can be easier to hold small items to 0.001°C in a stable room environment than to control a larger mass to 0.1°C outdoors. In the context of our original work with astronomical instruments, it has been possible to control photodiodes, ccd detectors and small ovens to 0.01°C rms over 24 hours where the outside temperature varies several degrees. In the examples given here, it has been more important to guarantee stability than to know the absolute temperature exactly; indeed, the actual temperature is often varied to optimise some other parameter. The intended temperature is commonly referred to as the set point.

There are three articles in this set. Part one outlines basic principles and compares the performance of a very simple design against a more sophisticated commercial module. In this way I hope to illustrate what is feasible with easily available components.

Part two will describe sensors in some detail, with suitable interface circuits. Part three will discuss heaters and thermoelectric Peltier coolers, and provide circuit details of a good analogue temperature controller.

Types of controller

There are two basic types of temperature controller – switching types employing hysteresis and those using a proportional feedback system. In the switching group, I include simple systems using bimetallic strips or schmitt triggers to switch heaters/coolers fully on/off as threshold temperatures are crossed. As these systems invariably oscillate over a degree or two, I do not consider them as part of the remit of this article and will not discuss them further. Instead I will concentrate on the proportional feedback systems which do not oscillate – provided that they are correctly set up.

Simple proportional system

Figure 1 illustrates what is probably the simplest possible arrangement. The item whose temperature is to be controlled is encased in a metal block along with a temperature sensor and heater.

Sensor output – here 10mV/°C – is compared in a difference amplifier with a voltage corresponding to the desired temperature. Amplified output, corresponding to the temperature error, is then applied to a power output stage which drives the heating element.

The power stage can be configured to produce either a voltage or a current output. In either case the heat supplied is the square of the error temperature. This assumes that the system is linear and that the heater element obeys Ohm's law. The current source has the practical advantage of being automatically short-circuit proof. This has proved to be a valuable asset in our case.

As the metal block warms up, the sensor output approaches the set-point voltage and power supplied to the heater is reduced. Eventually, a balance will be achieved and the system will in a large measure correct for changes in the surrounding environment; if the ambient temperature drops the sensor will cool slightly causing the error voltage to increase and more power will be supplied to the heater.

This simple setup can work well, but it suffers from two problems;

Set-point error. As shown, the actual temperature can never be exactly equal to the wanted set point since a temperature error is required to 'finance' the power delivered to the heater. If the error is zero then the heater current will also be zero.

Set point error may or may not be a real problem depending on the context. It can be improved by increasing the gain of the difference amplifier; doubling the gain halves the error. But there is a limit to how far the gain can be increased before the next problem begins – namely that of oscillation. The set-point error increases in a non-linear fashion as the heater current increases.

I should mention here that the set point error can be forced to zero for any one temperature by simply adding a dc offset. This may be a satisfactory solution if you do not need to vary the set point.

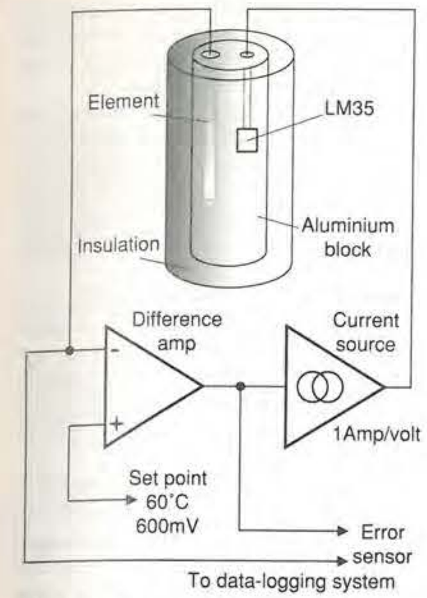


Fig. 2. This diagram is typical of set-ups used to control the temperature of small optical items or chemical vessels.

Oscillation. If the gain of the difference amplifier is increased too far, the system will begin to oscillate. This is due to a phase lag introduced by the thermal characteristics of the metal block, heater and sensor. The heat capacity of the metal block produces a natural integrating function and thus a phase lag,

$$\Delta\theta = \frac{1}{s}\int Q(t).dt$$

where D_q is the temperature change, S the heat capacity and $Q(t)$ the heat flow as a function of time.

This equation is strictly true going from one steady state to another, i.e. it assumes that the heat flow has time to come to equilibrium so that all parts of the metal are at the same temperature. The real situation is much more complex since the metal has a thermal resistance; a pulse of heat applied at one side of the block takes time to permeate through to the sensor. This produces an extra unwanted phase delay.

The general result is that the metal block introduces a low-pass function into the loop with a phase delay. At the point where the phase delay is 180°, if the gain provided by the sensor and electronics is greater than the attenuation through the block, the system will oscillate. And calculating heat flows in the metal is not a trivial task – especially if the shape is not simple.

There are computer programs available to do this, but fortunately it is not necessary to go to such lengths, except in the most extreme cases. A little forethought on the mechanical arrangements, combined with some experimentation is usually enough.

Putting it into practice

Considering this real example should help you put the previous discussion into context. **Figure 2** illustrates an aluminium block 25mm in diameter and 85mm long. This was fitted with a 24V, 25W soldering-iron element to provide the heat, and an *LM35* temperature sensor whose sensitivity is 10mV/°C.

The arrangement is typical of those used to hold small optical items or chemical vessels. The assembly was fitted with a 3mm layer of neoprene rubber insulation and an overcoat of bubblepack.

This construction illustrates the two most important points in ensuring accurate temperature control.

Thermal contact between the heater, sensor and item being controlled must as close as possible. Bear in mind that the only thing whose temperature is controlled is the sensor itself. All other parts of the assembly will show greater variations depending on the relative thermal resistance between the sensor and its environment. Note that the sensor is buried deep in the metal. The sensor wires form a significant heat leak and here they have to be run through the block; the sensor thus accurately reflects the temperature of the block.

Insulation needs to be as good as it possibly can be. This is important not just in stopping heat leaks, but also in preventing sudden changes in temperature due to draughts. Servo systems always cope better with slow changes.

These points should be obvious; the problem in my experience is that many experiments are

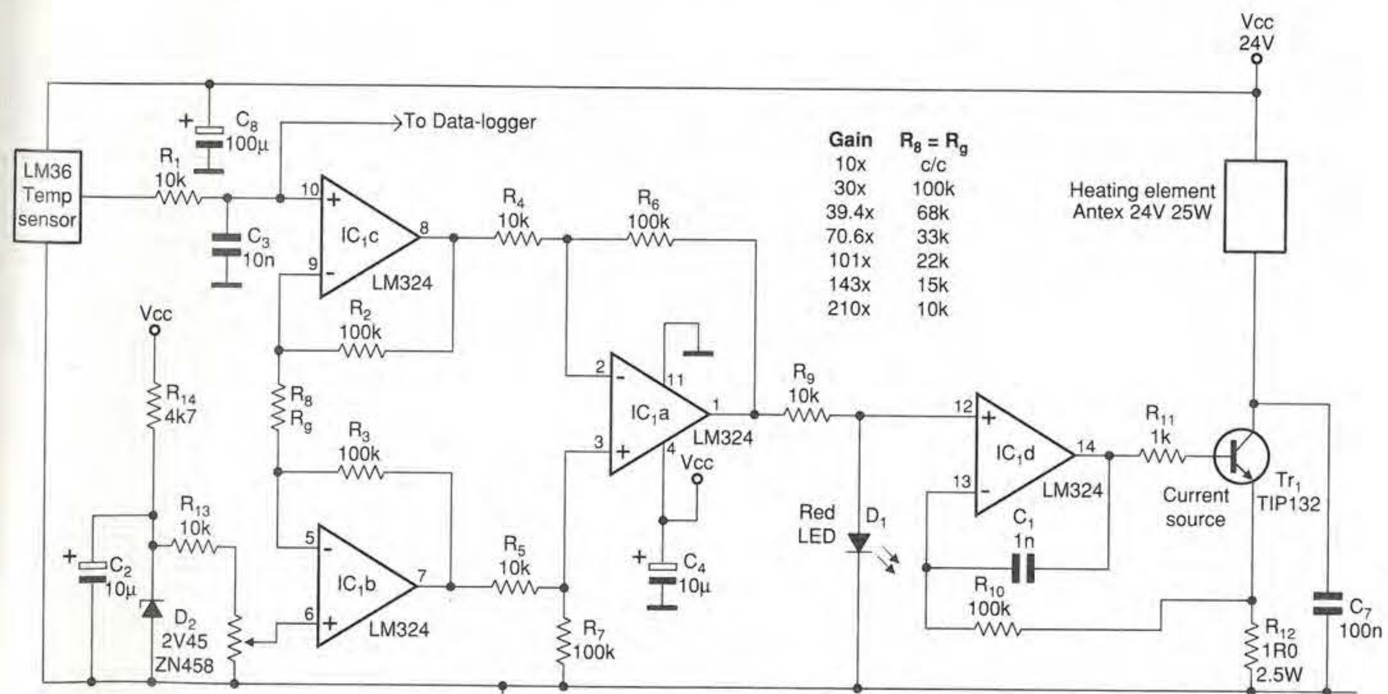


Fig. 3. Simple temperature controller based on the idea shown in Fig. 1. Simpler circuits are possible but the instrumentation amplifier has the benefit that it presents very little load to the sensor and set-point circuit.

designed by committee and the requirements for temperature control are only realised after all the mechanical arrangements have been worked out. If the sensor and heater are fitted as an afterthought, they tend to go in whatever space is available instead of where they need to be.

The circuit used for the experiment is shown in **Fig. 3**. It consists of an instrumentation amplifier whose gain is adjusted with R_8. It is possible to devise simpler arrangements, but the instrumentation amplifier is very convenient since it presents a high impedance to both the sensor and the set point voltage.

Gain is very easily changed with only one resistor. The amplified error signal is limited to about 1.2V maximum by the led, D_1, and fed to the current source scaled for 1A/V. Thus the maximum output current is 1.2A at the collector of Tr_1.

Not much more explanation is needed. The set point voltage is derived from D_2, a *ZN458* band-gap reference having a low temperature coefficient. This is essential. Without it, the set point will drift as the controller circuitry warms up; this topic will be expanded on in my third article.

Capacitors $C_{1,7}$ were needed to keep the current source unconditionally stable; the values are not critical. This circuit is not recommended for use as a general purpose controller since it does not protect against sensor failure. If the sensor fails there is a chance that the output will go to full power – with disastrous consequences for whatever is in the block. A fuller design with several useful extra features will be given in part 3. However this simple circuit may be considered where the sensor cannot be accidentally disconnected.

The block was placed in an environmental chamber so the performance of the system could be checked with changes of chamber temperature. Block temperature, chamber temperature, error voltage and heater current were logged using a pc-controlled data logger. Performance obtained is best described with reference to the graphs produced from the data logging system illustrated in **Figs 4** and **5**.

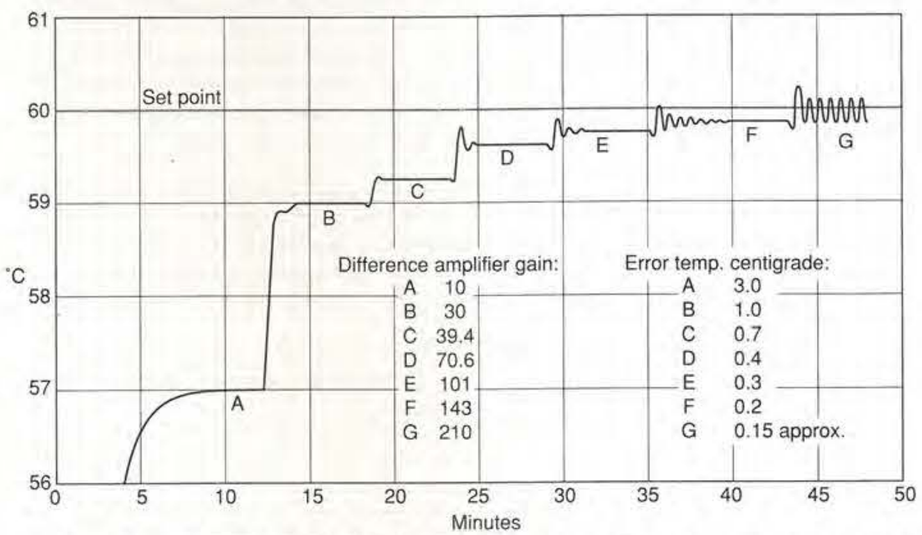


Fig. 4. Effects of increasing gain of the difference amplifier in the simple temperature controller. Improvements with increasing gain are clear, but you can see that increasing gain too much causes instability that starts to become evident as overshoot at step D, where gain is 70 or so.

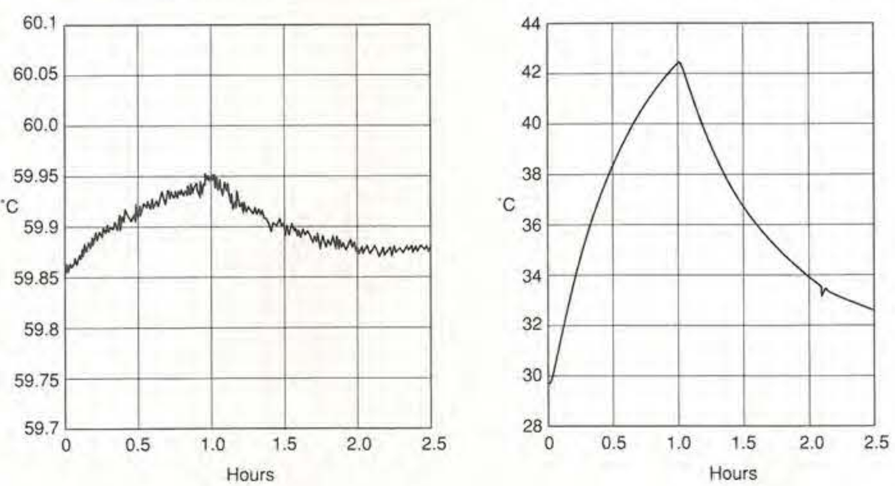


Fig. 5. Illustration of how well the temperature controller stands up to wide swings in ambient temperature. Readings taken directly from the block being controlled are in the left-hand chart. Ambient temperatures swing over the same period is shown in the right-hand curve.

Figure 4 illustrates the effects of increasing the gain of the difference amplifier. Initially, the gain was set to 10 by selecting R_8 on the circuit and the set point was 60°C.

After about five minutes, the system had come to its steady state condition at point A in the graph. You can see that the approach to equilibrium was rather sluggish but with no overshoot, i.e. overdamping. The final temperature is only 57°C, giving an unacceptable 3.0°C steady-state error.

Increasing the gain improves the situation; point C is the critically damped case which gives the shortest settling time and a temperature error of 0.7°C which is acceptable in most circumstances. Further increases in gain produce overshoot and ringing (D,E and F) until oscillation set in at point G.

The real test is of course how well the block temperature is held constant as the external temperature changes. The environmental chamber containing the block assembly was heated and cooled over a period of two hours and the temperatures logged every five seconds.

Figure 5 illustrates the results obtained with gain setting D; the difference amplifier set to ×70.6. The peak-to-peak temperature swing in the block is approximately 0.1°C for a change in ambient temperature change of 13°C.

There is a useful concept known as servo gain, which is defined as ambient temperature change divided by change in temperature of the block. The higher this value is, the better the controller can cope with external changes.

Ideally the servo gain would be infinite, and this can be achieved in theory with the PID type of controller described later. In this example the servo gain would be 13 divided by 0.1 or 130. This is a typical result.

Note that the concept is only really valid when measured between two steady-state conditions. Otherwise the servo bandwidth must be considered. In practice this can mean a long wait and it is usually enough to make sure the controller is tracking the changes.

The optimum choice of gain for the difference amplifier depends on the context. It may be acceptable to live with a small permanent oscillation in order to obtain improved long-term stability. But if the unit is part of an experiment involving the generation of power spectra using Fourier analysis, oscillations will be unacceptable if there is a risk of them influencing the results.

Having reviewed the basic concepts I will move on to look at methods for improving the performance.

PID controllers

The proportional-integral-differential, or PID, controller is a mechanism for solving the steady-state error problem. By including an integrator in the circuit the steady-state error can be reduced to zero for all set-point values – at least for steady state conditions.

The trick is to include an integrator as part of the loop as in **Fig. 6**. This design is based on the simple controller with the addition of two new blocks and a summing amplifier. If the integral and differential blocks are removed, the system behaves exactly the same as the simple controller in that a small difference between the set point temperature and the actual temperature is necessary to drive the system.

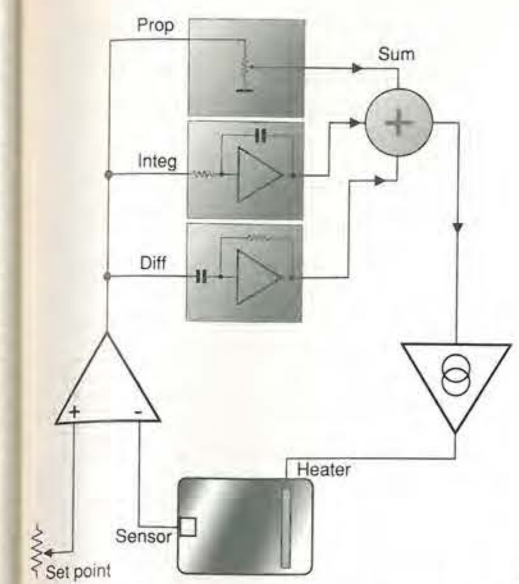


Fig. 6. In a proportional-integral-differential controller, the integrator plays a key role in reducing set-point error.

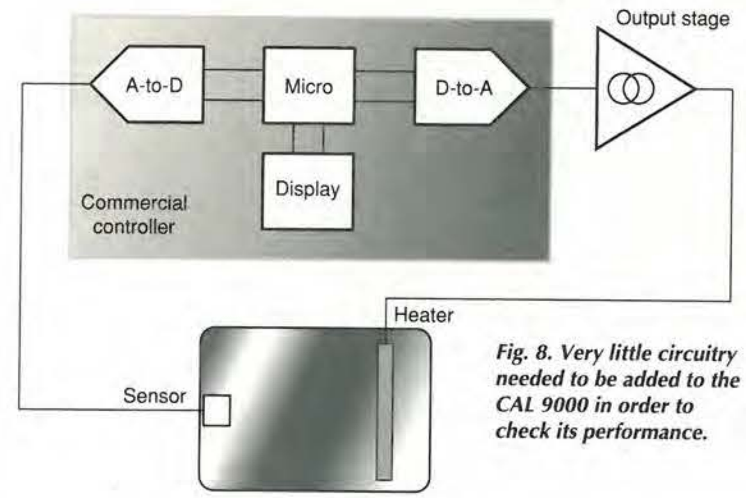


Fig. 8. Very little circuitry needed to be added to the CAL 9000 in order to check its performance.

The main property of an integrator is that a constant non-zero input will produce a linear ramping voltage at the output – positive or negative. Thus any residual setpoint error results in a correction which will go on increasing with time until the error is removed. Put another way, the only condition in which the integrator will produce a steady-state output is if the input to the integrator is zero.

Integrator input is zero when there is no set point error and the block temperature is exactly equal to the set point. This means that the controller can maintain the same temperature if the ambient conditions change, albeit slowly, from one steady-state condition to another. As a result, the servo gain tends to infinity. So remember that the gain of an integrator goes to infinity at dc.

There is but one fly in the ointment. Integrators are inherently sluggish devices and introduce extra unwanted phase lag. A sinewave input is always delayed by 90° and has an amplitude inversely proportional to frequency. Thus the oscillation problem can be made worse if the gain roll-off is not sufficient to compensate for the extra phase shift.

The way around this is to add in a differential component to counteract the phase effects of the integrator. By mixing proportional, integral and differential components it is possible to obtain a critically damped system with no set point error.

I have attempted to build analogue PID controllers on several occasions with varying degrees of success. With such controllers, you generally run into two problems.

Analogue PID drawbacks

The time constants associated with the integrator and differentiator tend to be long – at least high seconds or low minutes. This usually involves high value resistors and electrolytic or tantalum capacitors. There are invariably problems with leakage current, op-amp input current *et cetera*.

There are two time constants and a gain setting to adjust which demands a certain amount of patience if the thing is to be made to work empirically. There are analytical techniques which look at the system response to a pulse of heat and calculate the necessary time constants using Laplace transforms, but using these can be time-consuming.

Fortunately there are now available a wide range of commercial control units from manufacturers like Eurotherm and Cal Controls. These modules are readily available from RS and others, and can take out most of the grief.

All such modules operate digitally; an analogue-to-digital converter reads the sensor and supplies data to a microcontroller, **Fig. 7**. The microcontroller does all the computational work and supplies an output to a digital-to-analogue converter.

Since these units are general purpose you still have to provide a suitable output stage to suit the load, and a case and power supply.

A wide range of options and features is supported; the following are typical.

Choice of sensor. Various types of thermocouple and platinum sensors are supported.

Self-tuning facility. There is invariably an autotune mode which allows automatic setting of the integral and differential time constants. A pulse of heat is applied and the unit moni-

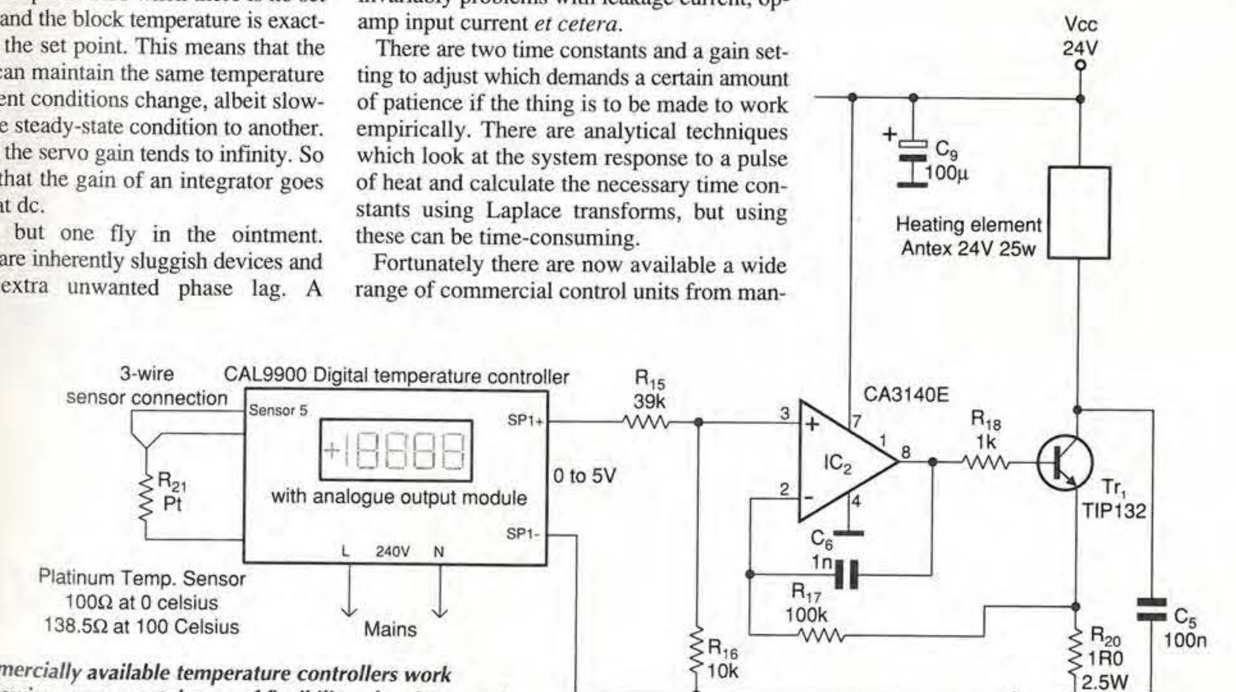
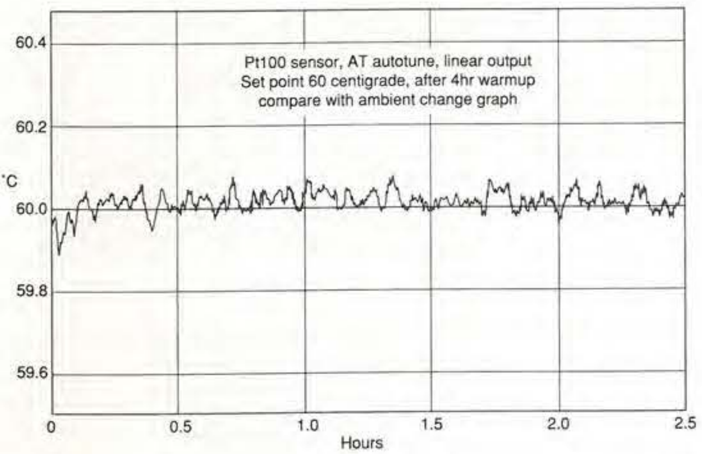


Fig. 7. Commercially available temperature controllers work digitally, allowing a you great degree of flexibility when it comes to tailoring the controller to suit your application.

Fig. 9. Performance of a CAL 9000 temperature controller over the ambient temperature swing shown in Fig. 8 reveals excellent long-term performance.



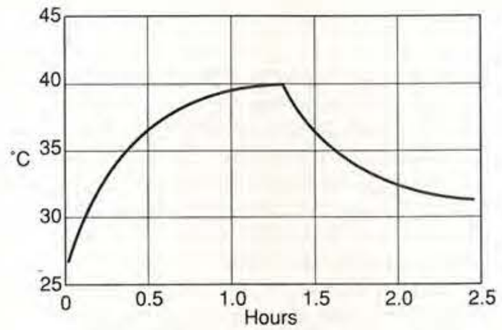


Fig. 10. Record of the ambient temperature changes used to evaluate the CAL 9000 controller.

tors the rate of temperature rise. After the pulse the cooling characteristics are recorded and the time constants determined.

Built-in sensor fail detection. This vital provision will switch off the output and provide a visible warning if the sensor becomes disconnected.

Choice of output format. Linear output is usually 0 to 5V, corresponding to full-on cooling via zero output to full heat. Sometimes there are separate heat and cool outputs. Relay outputs are usually an option for switched operation (non linear).

RS232 links. These enable monitoring and control of the set point by computer. A calibrated digital display is available which will show the set point, current temperature and sometimes the programming parameters.

As a demonstration of how these units perform, a CAL Controls model *9900* was set up with a linear power stage to heat the aluminium block used previously to 60°C. A platinum sensor was fitted close to the existing *LM35* sensor which in turn was used for monitoring the block temperature.

The controller was first allowed to warm up and the tuning procedure was used to obtain optimum results. Circuitry used is shown in **Figs 8** while **Fig. 9** shows the results of a 150 minute run. Changes in ambient temperature are shown in the smaller graph, **Fig. 10**.

Note how well the set point is maintained. Using the naked eye, it is impossible to see any reflection of the 12°C changes in ambient temperature swing in the block. There is no discernable set-point error beyond the calibration error between the two sensors.

For many applications this performance will be ideal and, at a cost of about £150 in one-off quantities, such a controller represents good value in view of the flexibility built into these units.

Only a minimum of support electronics is needed in the power stage and power supply, and the autotuning facility allows fast and accurate resetting for different loads and conditions. By the time the unit is mounted in a box with output stage and linear power supply, the cost will probably be around £200.

Need you buy a controller?

This level of performance will usually be more than adequate, but there are a few drawbacks which may need to be considered. The first is that of cost. If only one controller is required then there would usually be no problem, but if your experiment needs ten individually controlled items – as ours did – then the costs soon become quite expensive.

A second point concerns the stability. As already noted, the stability of the PID controller is very good over a number of hours hours. But there are substantial short-term variations of 50 to 100mK, i.e. 0.05 to 0.1°C, over time scales of minutes. This is due to quantisation error at both the input and output of the module. In this respect the simple analogue unit performs much better. Although it copes less well with external changes the thermal noise is much less.

Referring back to the stability plots, you will see that the simple controller has less than 10mK thermal noise compared to 50–100mK for the PID module. I should point out here that some of the more expensive PID modules allow you to set the temperature range over which the internal a-to-d converter operates thus the converter step size can be made smaller.

A third point concerns temperature monitoring. Some experiments require constant recording of the temperature, if only to correlate otherwise unexplained drifts in the results. Where a monitor output is provided it is usually of the RS232 variety, combined with remote programming facilities for the controller. Most pcs only provide two serial ports as standard and more often than not, these are taken up by a mouse and a modem.

If remote programming is not required then straightforward analogue monitor points on PID controllers can be more convenient if a multichannel datalogger is available. Such is often the case for collecting data from the main experiment.

Controlling to millikelvins

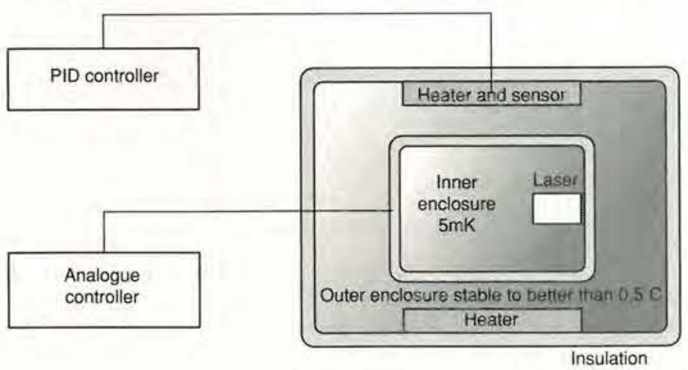
Certain applications demand temperature stability at the millikelvin level. One example is tuning a semiconductor laser diode to match an atomic absorption line. For this application it is necessary to keep the laser diode temperature constant to better than 5mK. This rather stringent requirement cannot be met by either of the controllers described so far.

But it can be done by building a two-stage control system, as outlined in **Fig. 11**. An outer enclosure is constructed whose temperature is controlled by a PID unit. With care, the temperature anywhere inside this enclosure will not vary by more than 0.5°C, provided that it is well insulated and not too big.

The laser is built into an inner enclosure whose temperature is independently controlled by an analogue controller. This controller's function is to remove the residual 0.5°C variations left due to convection in the outer enclosure and the quantisation errors.

Provided the servo gain of the analogue unit is greater than 100, the 5mK specification can be met. ■

Fig. 11. When controlling temperature down to a few millikelvins, techniques discussed so far are no use. But this two-stage design is.



Richard's next article looks at temperature sensors in detail.

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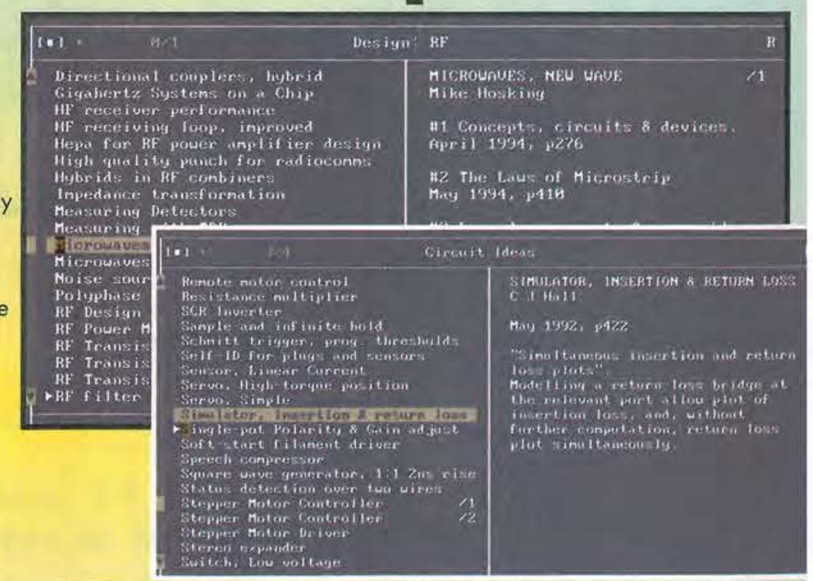

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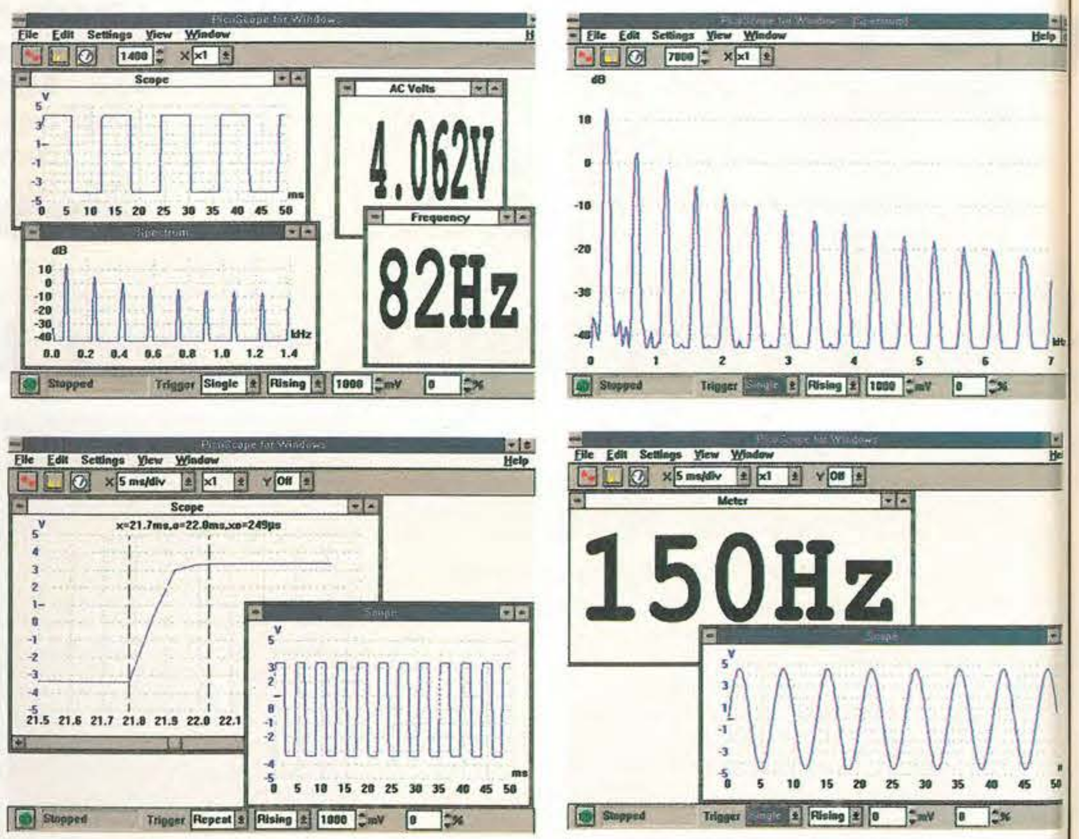


Turn your PC into a high-performance virtual instrument in return for a circuit idea.

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.

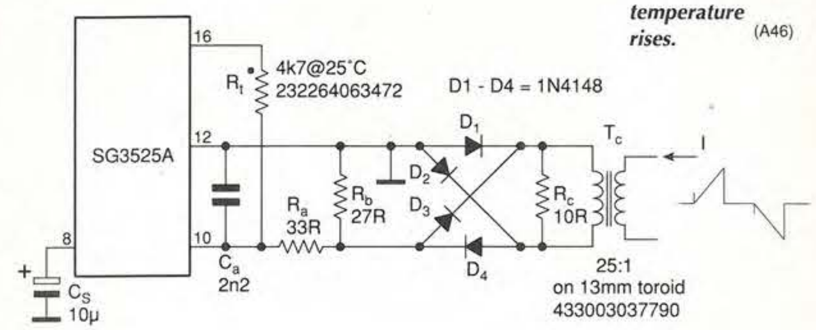
ADC42 is a low-cost, high-resolution a-to-d converter sampling to 12 bits at 20ksample/s. This single-channel converter benefits from all the instrumentation features of the ADC200-50.

A new way to protect smps controller

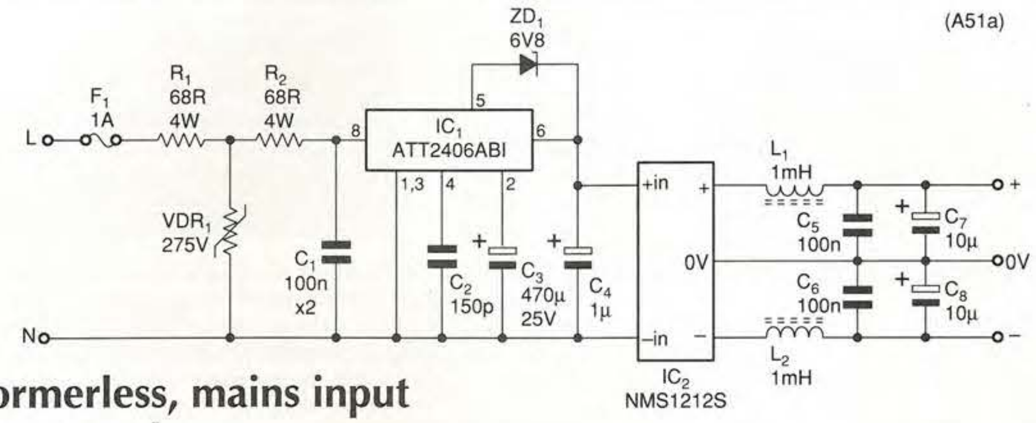
Switched-mode power supply control ics, of which the *SG3525A* is an example, provide a slow-start facility, which is implemented by connecting a capacitor (C_S) to pin 8, causing the pulse width to increase from zero as the capacitor charges until the normal feedback loop takes over. In addition, there is a pulse-by-pulse current limit using the shutdown facility on pin 8, which may be biased to take effect early or late in response to specified conditions.

As an example, the circuit shown reduces the current limit as ambient or heatsink temperature rises. Normally, current limiting does not take effect until pin 10 is about 1V positive with respect to the local 'ground' at pin 12, the current through the power mosfets being sensed by the current transformer, which has bridge-rectifier diodes on its secondary. As temperature rises, the ntc thermistor R_t pulls pin 10 more positive and causes the current limit to come in earlier, at a lower current; the diodes also possess a small negative temperature coefficient. Pin 16 has a convenient 5V reference and C_a filters out spikes and noise, since pin 10 is vulnerable to these effects.

C J D Catto
Cambridge
(A46)



Shutdown pin on a switched-mode controller may also be used to vary current limiting, in this case when temperature rises.



Transformerless, mains input dc power supply

The *ATT2406ABI* off-line power supply ic gives transformerless dc from a mains input. It provides variable voltage output at up to 100mA, short-circuit protection and regulates at inputs as low as 32V rms. A disadvantage is that there is no mains isolation but, using the output to drive a dc-to-dc converter with good isolation, such as the Newport *NMS12012S* overcomes the problem and gives dual outputs at up to 1W.

Output voltage of the off-line supply is set by ZD_1, 6V here, plus 5V. The voltage-dependent resistor protects against transients and C_2 shuts down the ic during transients. The *NMS* converter provides 6kV isolation, the *LC* networks removing residual noise. Minimum load to maintain regulation is 10%.

Nigel K Goodman
Westfield Near Hastings
East Sussex

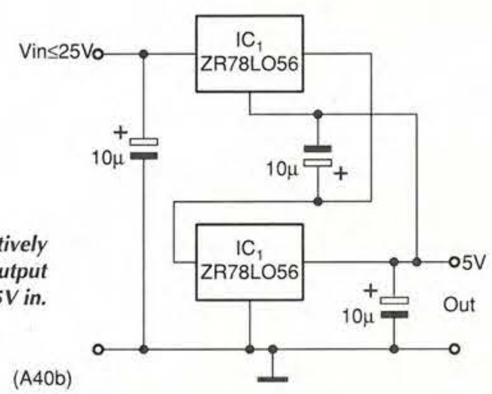
Warning

Please don't attempt to implement this circuit in any form unless you are fully conversant with your country's regulations regarding mains circuitry and isolation requirements. Component failure could result in fire, or lethal live mains voltage appearing at the circuit output.

Increased voltage regulator input

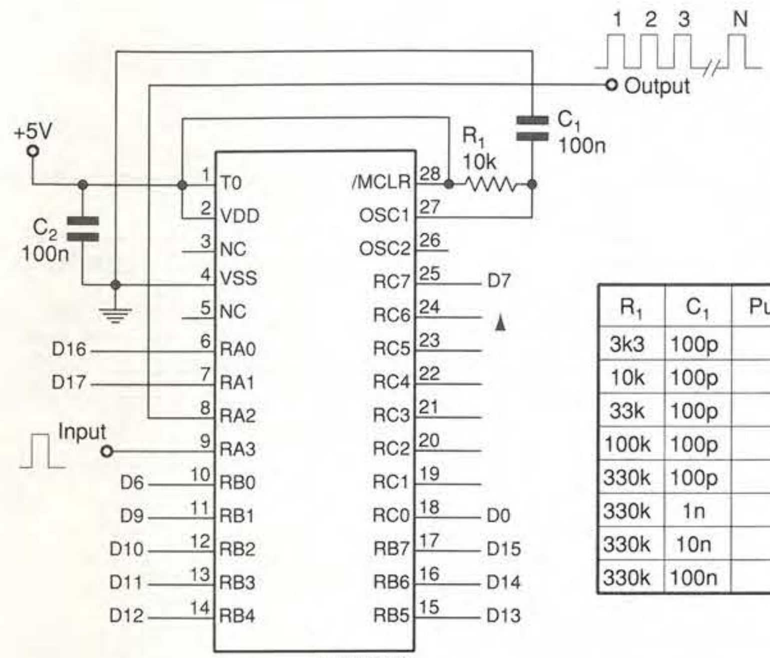
If you have a supply of up to 25V dc and need a regulated 5V out, two regulators may be connected as shown, with inputs and outputs in series and the ground of the first taken to the second output.

Kamil Kraus
Rokycany
Czech Republic
(A40a)



Two regulators, effectively in series, give a 5V output for up to 25V in.

ADC-42 WINNER



A microcontroller controlled by a binary input produces a programmed sequence of pulses in a train when it detects a trigger signal.

R_1	C_1	Pulse frequency
3k3	100p	47.6kHz
10k	100p	14.2kHz
33k	100p	6.25kHz
100k	100p	2.2kHz
330k	100p	660Hz
330k	1n	69Hz
330k	10n	7.4Hz
330k	100n	0.81Hz

(A45)

Programmable pulse train

This *PIC16C55* generates a fixed number of pulses in response to a trigger signal, the number of pulses in the train being programmable and the frequency determined by a *CR*.

Port B, Port C and two bits of Port A read an 18-bit binary number that determines the number of pulses, while the other two Port A pins take the trigger input, RA_3, and the output on RA_2.

In the listing, the trigger input is tested continuously, a logic high causing the binary input to be read and stored in the three software counters. Each counter loops, its count being decreased by one each time until it reaches zero, after which the program returns to start and retests for the next trigger input.

Yongping Xia
Torrance
California
USA

PIC assembly code for the pulse-string generator.

```
        LIST    P=16c55

STATUS          equ     0x03
PORTA   equ     0x05
PORTB   equ     0x06
PORTC   equ     0x07
z       equ     0x02
cnt_1   equ     0x08
cnt_2   equ     0x09
cnt_3   equ     0x0a
        org     0x0
main    clrf    PORTA
        clrf    PORTB
        clrf    PORTC
        movlw   0xfb
        tris    PORTA
        movlw   0xff
        tris    PORTB
        tris    PORTC
start
        btfss   PORTA,  3
        goto    start
        movf    PORTA,  0
        andlw   0x03
        movwf   cnt_3
        movf    PORTB,  0
        movwf   cnt_2
        movf    PORTC,  0
        movwf   cnt_1
loop
        movf    cnt_1,  0
        btfsc   STATUS, Z
        goto    next_1
        decf    cnt_1,  1
send    bsf     PORTA,  2
        nop
        nop
        bcf     PORTA,  2
        goto    loop
next_1
        decf    cnt_1,  1
        movf    cnt_2,  0
        btfsc   STATUS, Z
        goto    next_2
        decf    cnt     1
        goto    send
next_2
        decf    cnt_2,  1
        movf    cnt_3,  0
        btfsc   STATUS, Z
        goto    start
        decf    cnt_3,  1
        goto    send
        end
```

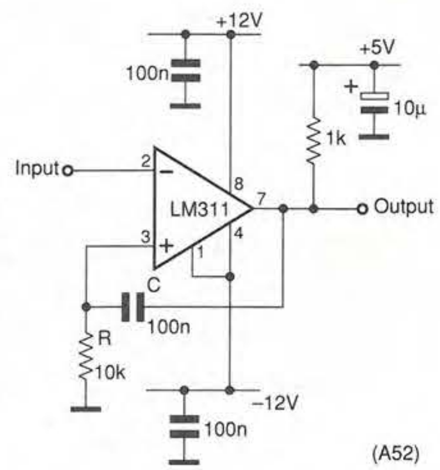
Accurate zero crossing detector

Positive feedback is needed by a comparator to avoid oscillation around the switching point, and resistive feedback is commonly used. This, of course, unavoidably causes hysteresis, resistor values are tedious to calculate for an open-collector comparator and the switching level depends on the exact value of the pull-up rail voltage.

If, however, ac feedback is used, as shown in the diagram, all these problems disappear. This particular circuit is a zero-crossing detector for 50Hz input, but it should work at frequencies up to the propagation-delay limit of the comparator.

Perfect switching at zero occurs and the large overdrive to pin 3 causes switching in under 30ns even with slowly varying inputs; delay between a positive-going crossing of a 50Hz input of 5V pk-pk and the 90% level at the output is under 100ns.

Feedback *RC* time constant must be small when compared with the time between zero crossings to let the switching threshold reach ground. Decouple the 5V logic rail well to avoid switching noise causing inaccurate triggering when the output is high and drive the circuit from a low-impedance source.



(A52)

A Lloyd
Chester
Cheshire

Bootstrap ramp/triangle generator

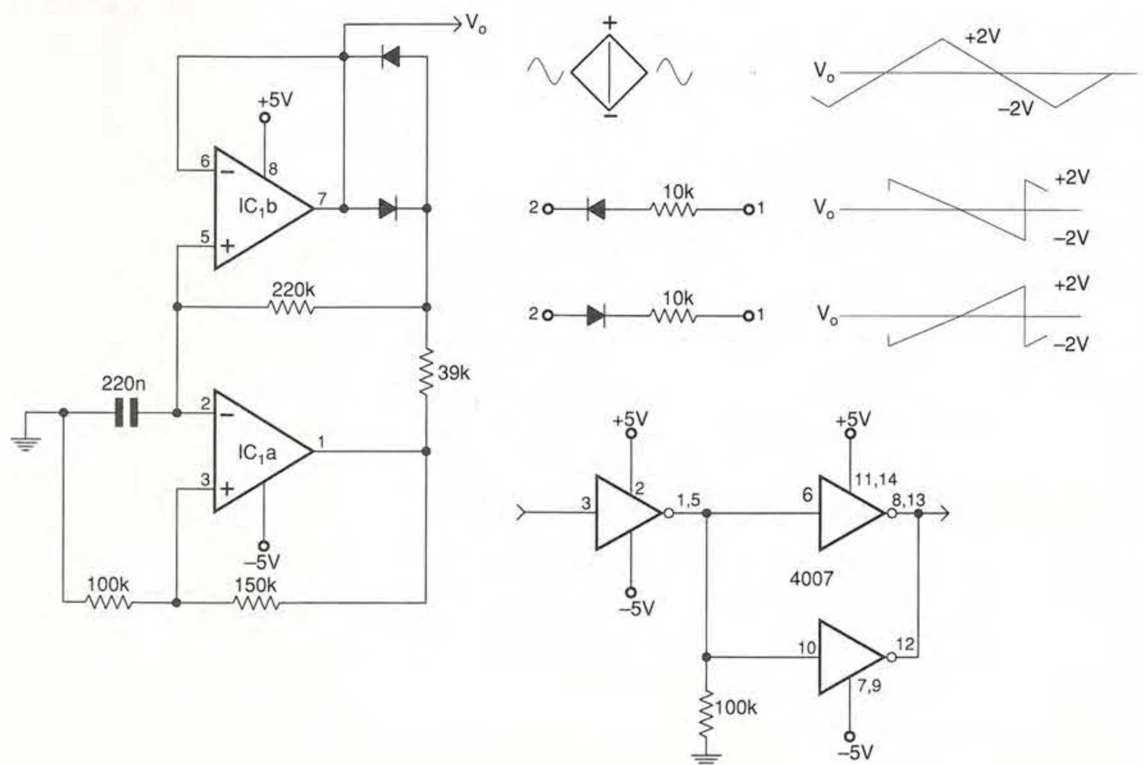
Using a comparatively small capacitor of 0.22µF, this linear ramp generator provides periods of up to 800ms, and the use of a dual op-amp allows low-impedance outputs of both ramp and associated square wave.

Current through the 39kΩ resistor provides bias for the diodes during charge and discharge of the capacitor and, although not constant, forces an essentially fixed offset of about 500mV across the 220kΩ resistor and maintains constant voltage change across the capacitor. Amplifier b follows the capacitor voltage to produce the output v_o.

Assuming that pin 1 has a rail-to-rail swing, the 150kΩ/100kΩ divider imposes a 2V offset on pin 3. Such a swing provides equality of switching current to the diode source/sink; if the op-amp in use does not have this capability, a cmos switch (*4007* shown) may be interposed between the op-amp output and the point now shown as pin 1.

Replacing the diodes by a shorted bridge rectifier effectively puts two diodes in series for each phase and therefore doubles the frequency. To obtain a sawtooth waveform, a rapid charge or discharge is given by placing the series diode/resistor between pins 1 and 2.

John A Haase
Colorado State University
Fort Collins
USA
(A48)



Electronic fuse has manual reset

After this fuse has disconnected the circuit from its supply, operation can only continue after the load is removed and the reset switch momentarily operated.

Both negative and positive forms of the circuit may be made, as shown in the diagrams. To take the positive form, the inverting input of the *LT1366* rail-to-rail output op-amp reflects the drop across the power mosfet $R_{DS(on)}$ and the non-inverting input receives the feedback from the op-amp output, reduced by $R_{1,2}$. The op-amp output drives the fet directly, normally providing a positive gate bias and allowing current to pass to the load.

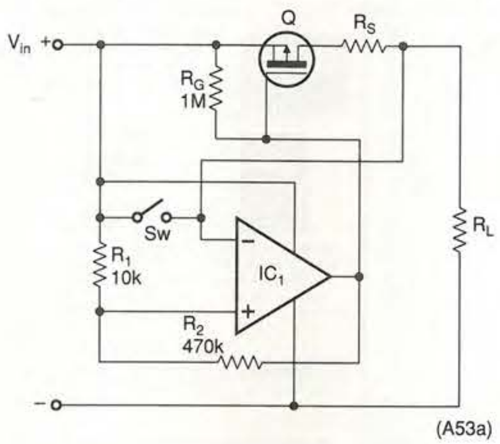
In the presence of increasing current, the input to the non-inverting input will eventually exceed the fed back voltage and trigger the op-amp into its alternate state, imposing a negative gate voltage and cutting the fet off.

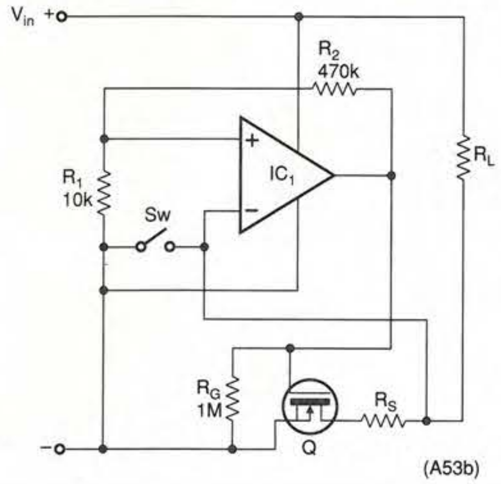
The value of current to cut the fuse off is given by,

$$I_L = \frac{R_1}{(R_1 + R_2)(R_s + R_{DS(on)})} V_{in}$$

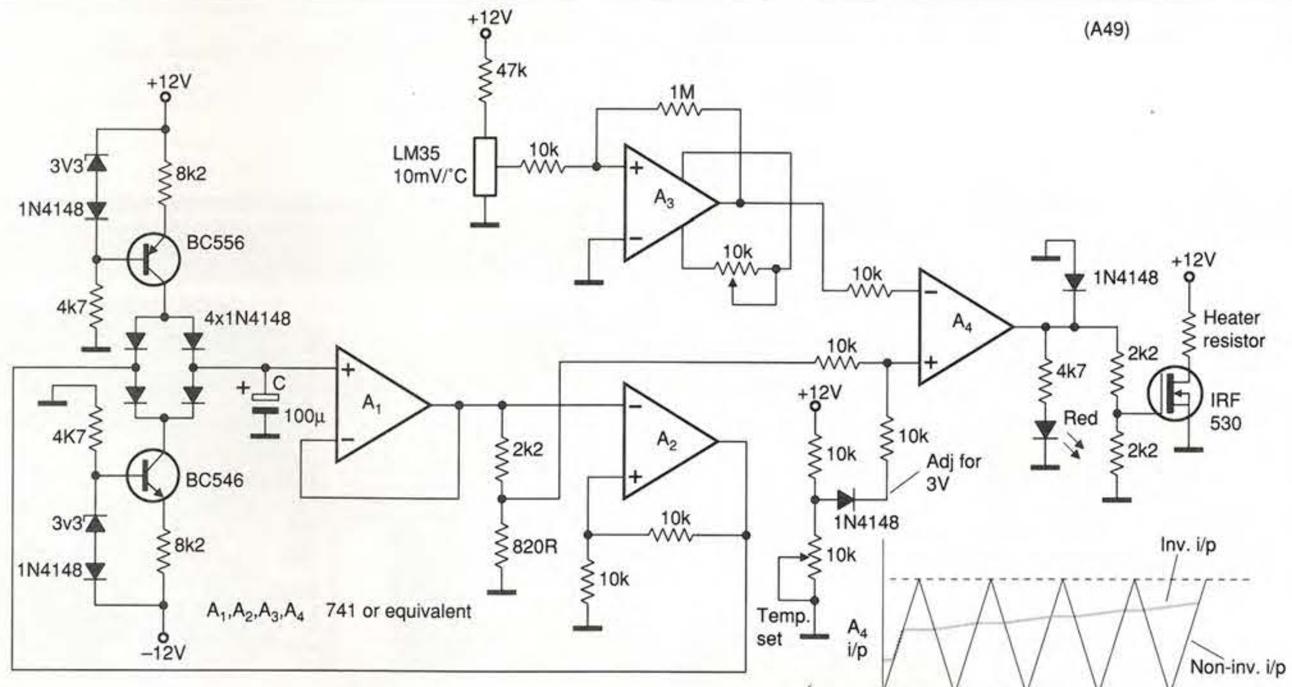
The negative equivalent of this is the second circuit and is virtually identical, except for the n-channel fet.

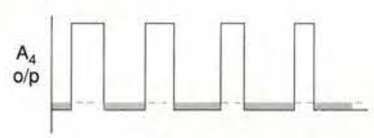
Rae Perälä
Helsinki
Finland





Electronic fuse, resettable by removing the load and operating the reset button.





Temperature controller

Originally used to control the temperature of a small box in which to keep temperature-sensitive components, this controller clearly has many other applications.

The two transistors, zeners and diode bridge form a constant-current source and sink, which charges and discharges capacitor *C* linearly. After buffering in op-amp 1, the capacitor voltage goes to op-amp 2, which switches the charge/discharge circuit to the opposite state when op-amp 2 output exceeds 6V dc or –6V dc.

Op-amp 4 takes a potted-down version of the resulting triangular wave at its non-inverting input, together with an adjustable direct voltage, set to 3V for a 40°C temperature. Op-amp 4's other input sees an amplified version of the output from the *LM35* temperature sensor, the two inputs being compared to produce a width-modulated pulse at the output of op-amp 4, which narrows as temperature rises; the led indicates that the heating element is on. If temperature continues to rise beyond the setting, op-amp 4 output clamps at –0.6V until the temperature is again within the control range.

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(A49)

Prototyping with smds

Nick Wheeler looks at economical ways of prototyping with surface-mount devices, and describes an adaptor that allows you to make use of the growing number of chips that not available as through-hole versions.

Surface-mounting is rapidly coming to dominate production electronics, both because of its small size and because of the relative ease with which surface-mount assembly can be automated. As a result, an increasing number of parts is becoming available *only* in surface-mount format. Unless you can master the technique for prototyping, certain areas of experimentation will become inaccessible to you.

The methods described here are usable without difficulty for packages with leads on 0.05in centres such as SOIC, FK and many others. They are also suitable for most of the available packages for discrete semiconductors and passive components.

Software

For components based on 0.05in centres, the precision with which pads and traces need to be positioned should be 0.01in or better. This takes into account traces passing between adjacent pads.

Such precision can be achieved in one of two ways. One method is to draw the pcb layout using ordinary, manual draughting methods at many times actual size, and photographically reduced to the required size. The other is to use computer-based graphics software, when the layout can be expanded or reduced at will. This by far the most practicable method.

I looked at Windows Draw V3 and Serif Draw Plus V2. These are similar in many respects, but Serif has several significant advantages. Although it notionally operates on a spatial increment of 0.01in you should find it quite easy to interpolate to a precision at least twice as good as this, even though the numerical position indicator only moves in steps of 0.01in.

Serif also includes adjustable X and Y 'rulers' that can be positioned with the enhanced accuracy noted above. They take the form of red guide lines across the whole screen that can be switched on or off – without being lost when off – and which do not appear in any printout.

Later I describe a simple SOIC-8 to eight-pin DIL adaptor to illustrate the use of Serif Draw.

Circuit boards

I assume here that you will be using commonly available photographically coated, double sided, 1oz, 1.6mm thick GRP board. On such boards, traces of one typographic 'point' (0.0138in) width can reliably be produced. Satisfactory production of 0.5 point traces is not difficult, but requires much more attention to cleanliness, photographic exposure times and development times.

A feature of narrow traces produced by typical computer-aided draughting programs is that unless they are strictly vertical or horizontal they can be seen to consist of a series of very small steps. Such traces can break up into dotted lines unless very close attention is paid to the points noted above. Cost savings can be made if you can foresee that single-sided board is adequate for the requirement.

Drilling of circuit boards is best done with a tungsten-carbide drill. They leave a cleaner hole which can be important with the small clearances involved. But take care since these drills, although extremely hard, are also very brittle – and expensive. I use a miniature drill press.

Any holes that must mate accurately with, say, a 0.1in matrix chip must be accurately located. Drill the first as accurately as possible, and use this to position a drilling guide. This jig should be made of metal. For a 0.1in matrix, the jig can be conveniently marked out using a piece of Veroboard and a 1mm drill.

Drilling must be done before soldering. Solder may cause deflection of the drill and might affect the performance of the drill jig.

Soldering

There are several soldering processes used industrially, such as wave tanks and vapour phase, in which a suitably hot vapour is directed over the populated board. The only two that are relevant in the context under discussion are the use of a small conventional soldering iron

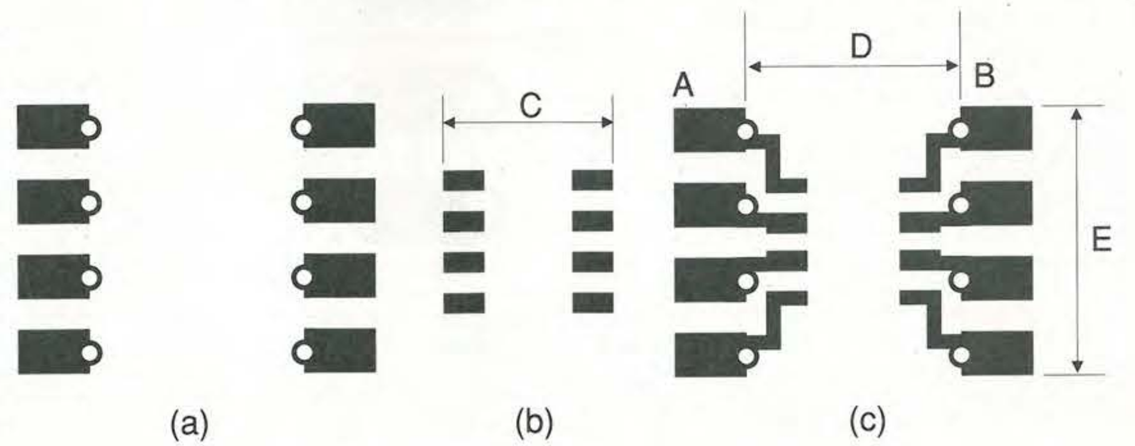

Fig. 1. More and more ICs are becoming available exclusively as surface-mount options. This adaptor that allows you to use eight-pin SOIC surface-mount chips as through-hole DIL devices. It is easiest to draw the DIL and SOIC pads, as in a) and b) respectively, then amalgamate the two, as in c).

– and I do mean small – or infra-red heating.

A soldering iron can be used with cored wire solder, 24SWG being suitable. Alternatively, small dabs of solder paste may be applied to the pads for both these processes. The paste is usually obtained in hypodermic syringes (RS, Maplin) and a 21 gauge needle should be used. A plastic nozzle is usually included in the solder pack, but this is quite clumsy.

The needle should be ground down so that it is about 1 cm long. Bearing down on the syringe plunger will produce a minute sausage of paste at the needle tip. About 1mm of this is enough for each leg of a typical smd part. Unless you have remarkable eyesight, you will find a 'hands-free' magnifier necessary. You will certainly need to practice the technique of applying the paste on an easy-to-get part or two.

Small soldering irons can be fitted with 'needle' bits. These are easy to use, but have a limited life. And their failure mode is exasperating. What happens is that the iron coating develops a pinhole, after which the copper inside erodes leaving the bit apparently intact but scarcely able to conduct heat to the work.

An alternative, which I recommend, is always to retain bits which have failed as described above. They can then be drilled – not while still fitted to the iron – and a length of 1mm tinned-copper wire inserted, protruding about 1cm. This will last some time and can be replaced at negligible cost.

I find that the largest size of bit for an Antex 25W iron gives good results using this method. The extra power ensures that enough heat gets down the thin wire to the work.

I made an infra-red heater as follows. An industrial bar heating element with an area about 35cm by 5cm was under-run at about 700W from a 65V, 10.7A supply, producing a dull red glow. Placed 4cm above a board, with solder-paste dabs in position, resulted in the paste melting after a minute. Obvious solder melting occurred after three minutes. The power was then switched off and the board allowed to cool naturally.

Devices – prepositioned on the solder paste dabs, of course – have been successfully attached by this method and they worked properly afterwards. If you put a solder-paste dab of representative size on a suitably sized pad, say 0.05 by 0.025in, where it is easy to observe, it is easy to see when the heater should be switched off.

Note that the power that needs to reach each square centimetre of the board is only a few watts, so in many cases a less powerful heat source will suffice. A 250W domestic reflector-type bulb, for example, melted solder over a 1.5in^2 area in 2.5 minutes at a range of 2cm from the face of the bulb. Experiment with what you have available.

Solder paste has a short shelf life and should also be kept refrigerated at 0-10°C by the way.

Producing a soic to dil adaptor

Such an adaptor enables a part which is only available in SOIC packaging to be applied to a 0.1in matrix strip board for experimentation. Being a simulated DIL part, it can also be desoldered for use elsewhere.

So far I have used this technique successfully with 8, 14 and 16 pin parts. If you don't want to make one, you can buy them from Winslow, whose address is given in the 'Further reading' panel. But note that the company has a £50 minimum order charge. A limited range of Winslow parts is stocked by RS, but not the equivalent of the one described below.

If you can make the item I will now describe, you will have little difficulty in constructing circuits on a 0.05in matrix with interconnecting traces as narrow as 0.0138in across.

Figure 1 is an enlargement of the artwork for an eight-pin adaptor. The following description of how to produce the adaptor assumes that you have Serif Draw Plus V2.0, If you are using a different package, with a little experimentation, it should be easy to translate the process.

From the point of view of circuit layout, the operation of the 'pointer tool', accessible at the top of the left tool bar, is centrally important. You should be able to deduce what it does from what follows.

1. After obtaining the initial display – an A4 page surrounded by margins – select 1000% magnification. Press the View button and select Guides, Rulers and Grid. Use the mouse to draw two X-rulers separated by 0.3in, dimension B, adjusting them carefully to coincide *exactly* with divisions on the horizontal scale at the top of the display area, and the grid. The numerical position indicator at the bottom right-hand corner of the screen gives only approximate accuracy.

2. Draw a Y-ruler, exactly on a major division of the vertical scale. The intersections of this with the two X-rulers are the positional datum for all that follows. Shifting these rulers to the exact positions is achieved by using the pointer tool and mouse.

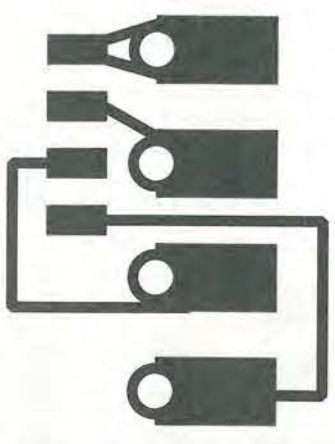

Fig. 2. Extending the adaptor for use with 14-pin devices involves some fine pcb work, but with care ,a good draughting package and a fine soldering iron, you should be able to manage it without resorting to expensive, specialist surface-mount tools.

3. Using the 'quick-shape' tool, draw two circles of 1mm outside diameter and 0.5 point line width at these two intersections. These are the guide points for the drilling of the 1mm holes which will eventually accommodate the DIL-spaced output pins.

4. Using the rectangular quick shape tool, draw two pads 'A' and 'B', 0.06in high and 0.1in wide with their inner edges coinciding with the original two Y-rulers. Serif indicates coincidence by a colour change. These pads will over-write the two circles as indicated in Fig. **1a**).

5. Now use the 'replicate' function, found by using the 'Effects' button on the top toolbar, to reproduce the two pads and location loops four times vertically spaced on 0.1in centres. You may experience some difficulty in getting this exactly right. Dimension E must be 0.3in. Don't worry. The pointer tool enables you to 'lasso' a misplaced pad and drag it to the right place.

   If you make a mistake, press the edit button and undo the last move before proceeding.

6. You will now have eight pads appropriate to DIL spacing with eight indications as to where to drill for through pins, Fig. 1a).

7. The SOIC pads are now created elsewhere in the drawing area. They are 0.05in wide and 0.025in high, on 0.05in vertical centres. The important fact is that the horizontal footprint of these pad pairs does not exceed 0.228in wide, dimension 'C' on Fig. **1b**).

8. This group of four pad-pairs is now lassoed with the pointer tool and moved to a symmetrically placed position between the four pairs of DIL pads.

9. Finally, using any method you like, line tool or successive rectangular shape tool operations, the SOIC pads are connected to the DIL pads. Choose a route which does not produce a reduction in clearance between existing pad areas, Fig. **1c**).

At 0.5 by 0.36in, the resulting pattern is very small and can be replicated seven or eight times at the very top of the page area, between the violet and black lines, for printing. Note that although mistakes appear to be deleted when you overwrite them in white, the replicate process does not recognise this as a deletion. The deleted parts will appear on the replicas.

This description may seem lengthy, but paying attention to it could save you a lot of time. The process by which the pattern was enlarged for reproduction for this article, also obtained using the pointer tool, is not exact. Figure 1 has been modified manually and should not therefore be scaled.

Printing

The pattern is now printed on foil such as HP transparency film. Patterns that are not symmetrical left to right must be flipped left-for-right, so that the side of the foil with the actual ink ends up in contact with the circuit board. This is essential as the fine detail will otherwise be blurred or lost. Ultra-violet exposure, development and etching will depend upon the equipment and chemicals you prefer.

Pinning

There is not enough space on such a tiny board for ordinary 1mm Vero board pins, as the clearance between a 1mm round pin and the outer ends of the SOIC pads is dangerously small.

The problem is solved by using Vero wire-wrapping pins, sold as Vero 18.56067, Maplin FL80B or RS 434-093. Placed on 0.3in centres, dimension D, they leave a clearance between the two columns of 0.274in, when oriented correctly, since they are not round.

Dimension C is 0.228in, which means that there is just 0.023in between these flat pins and the SOIC pads. This extreme situation only arises with 14-pin devices, but is workable.

Fig. 3. Bending down the leads of a SOIC package gives you an extra 0.02in of clearance on either side of the pin.

Figure 2 is a strictly diagrammatical indication of how a 14-pin device, or 16-pin device, can be accommodated. There is another solution to this problem. If the gull-wing leads of the devices are straightened, another 0.02in of clearance becomes available on either side of the pins, **Fig. 3**. This straightening is in any case recommended if you intend using a soldering iron and wire solder.

The pins are a tight fit in 1mm holes. The board needs to be supported on a metal block or sheet drilled with 2mm holes on a 0.1in matrix. This allows the wider part of the pin to protrude slightly below the board without fouling. Firm pressure is required for insertion.

Guide rings for drill positioning are 1mm diameter and should disappear as the holes are drilled. The area should be carefully inspected and any unwanted traces of copper around the holes removed.

In summary

It is possible to work on a 0.05in matrix using the procedures described. The adaptor is a worst-case example because of the limited clearances involved.

However, the chance of being able to use some of the smd parts that are not available in through-hole versions is well worth the effort. ■

Technical support

You might find the following useful:

'The World of Surface Mount Technology' RS part number 436-588.

Winslow's catalogue is available from Winslow International, Brecon Enterprise Park, Powys LD3 8EF, tel 01874 625555.

See Wheeler, NPE, 'RF Active Probe.'

Serif is at PO Box 15, Nottingham NG7 2DA, tel. 0800 924925 for sales.

Electronics World, Aug. 1995 – if only to see how much better can be done now.

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ACTIVE

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

A-to-d for load cells. Analog's *AD7730* is a complete, 24-bit analogue front end for use with load cells and pressure transducers. Resolution is 220 000 counts and there are self and system calibration facilities. Offset drift is 5nV/°C, gain drift 2ppm/°C. Features include digital filtering that accommodates fast changes, two differential inputs accepting eight voltage ranges from 10mV to 80mV single-ended or differential, a 6-bit converter to remove the tare voltage and clock signals to synchronise bridge excitation. Analog Devices Ltd. Tel., 01932 266000; fax, 01932 247401.
Enq No 501

Discrete active devices

Rf power fet. *PTF10031* is a common-source, n-channel enhancement-mode fet rated at 40W minimum output power, with applications to 1GHz. It is a laterally diffused mos (ldmos) device, a process that produces a smaller drain-gate capacitance to improve gain and confer better stability and reduced inductance by the elimination of bonding wires. Compared to a 45W bipolar transistor, the ldmos fet exhibits better intermodulation distortion at power levels below 30W. Maximum drain/source voltage is 60V, gate voltage 20V and maximum operating junction temperature 200°C. Package is either 20222 or the flangeless 20235. Ericsson Components AB. Tel., 01793 488300; fax, 01793 488301.
Enq No 502

Igbt plus diode. International Rectifier has produced new surface-mount *D-Pak* and *D-CoPack* versions of its insulated-gate bipolar transistors, in which the igbt and a Hexfred fast recovery diode are combined in the one package, an industry first, according to IR. These devices will cut losses by up to 60% when compared with mosfets, with better switching performance and often reduced heat sinking requirements. International Rectifier. Tel., 01883 732020; fax, 01883 733410.
Enq No 503

Memory chips

"Fastest" shared-port ram. QSI's *QS75436* 3.6Gb/s ram consists of two independent blocks of 2K by 32bits of sram, accessible from either port using independent control pins. Each port has a clocked interface taking addresses, data and control on the rising edge of the relevant clock signal, giving block transfers at 3.6Gb/s in burst mode with both ports active. Block contention is resolved by a busy flag. Silicon Concepts Ltd. Tel., 01428 751617; fax, 01428 751603.
Enq No 503

Microprocessors and controllers

C-programmable with graphics. From Impulse Corporation comes the *PK2240* C-programmable controller, which is designed for use where a keypad and display are needed. It is provided with 16 protected digital inputs and 14 high-current digital outputs for driving actuators, an 18.432MHz processor, 256Kbyte of flash eprom and 512Kbyte of static ram. An on-board RS-485 port allows more i/o or connection in a network. Programming is in Dynamic C, which is optimised for real-time control and which is integrated, with editor, compiler and debugger. Geometric shapes and text can be drawn on the graphics display to represent systems, alarms and various components such as pumps and valves. A development kit is available. Impulse Corporation Ltd. Tel., 01543 466552; fax, 01543 466553.
Enq No 504

Optical devices

Leds in a block. Lumex has a block of four rectangular leds standing only 0.47in high, each led being enclosed on five of its sides to stop light leaking out. *SSF-LXH534xx* series leds have pins on 0.1in centres and the leds themselves come in all colours and combinations. A & E Marketing Ltd. Tel., 01592 873888; fax, 01592 874555.
Enq No 505

Oscillators

Network clock. AMI's *FS6017* cmos clock synthesiser ic eliminates the need for multiple clock oscillators in a network computer by using one crystal oscillator and five independent phase-lock loops to produce 1.28MHz, 80/3.68MHz, 40/48MHz and 56/64MHz, the choice in the case of dual output frequencies being made by a select pin. Amega Technology Ltd. Tel., 01256 305340; fax, 01256 305348.
Enq No 506

PASSIVE

Passive components

Bipolar electrolytics . *RB Series* bipolar electrolytics by Nover are for applications in which they are subject to reverse voltages, such as audio amplifiers. Values are in the range 1-220μF at 50V. others being available to order. Tolerance is ±20%, surge rating 63V and tanδ of 0.12 maximum. The largest can size is 12.5mm diameter by 25mm and the smallest about half that. Anglia. Tel., 01945 474747; fax, 01945 474849.
Enq No 507

Spring coils. Air-cored *SC* spring coils can be produced by Total Frequency Control to customers' requirements, which may include diameter, length, wire diameter and style of winding, of which a variety is available. All coils use enamelled copper wire with tinned connections and delivery is within ten days. Total Frequency Control Ltd. Tel., 01903 745513; fax, 01903 742208; e-mail, eddie@tfc.co.uk.
Enq No 508

Suppression capacitors. Konex *MKP12* X2 polypropylene suppression capacitors are made in the range 0.01μF to 2.2μF with tolerances of ±10% and ±20%. Rating is 275V ac and operating temperature range –40°C to 85°C. Standard leaded versions, cropped and insulated types are available and the boxed construction is sealed in UL94V-0 flame-retardant epoxy resin. Components Bureau Ltd. Tel., 01480 496565; fax, 01480 496480.
Enq No 509

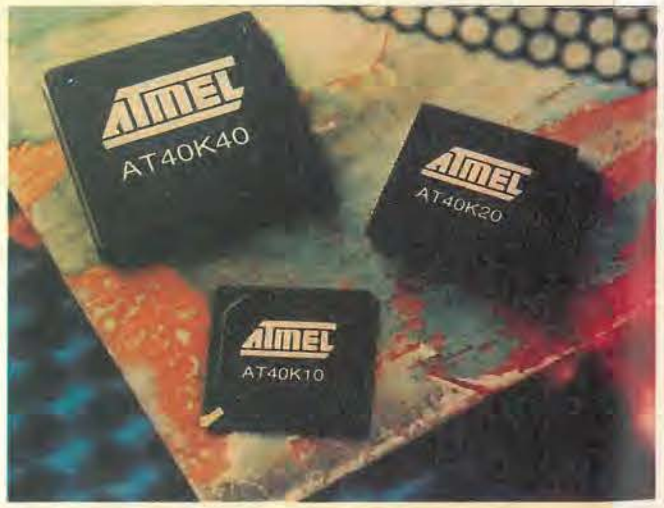
Chip inductors. Two types of chip inductor are available from Total Frequency Control: The *Type FD*, for standard needs and *Type ND* for use when high self-resonant frequencies are needed. The range of values is 0.008μH to 4.7μH. Leadouts are phosphor-bronze, tin-plated for easy reflow or bath soldering, the case is epoxy resin and the coil is of polyurethane enamelled-copper wire on ferrite. Total Frequency Control Ltd. Tel., 01903 745513; fax, 01903 742208; e-mail, eddie@tfc.co.uk.
Enq No 510

Connectors and cabling

USB connectors. Connectors for the Universal Serial Bus have been introduced by Methode. Two non-interchangeable kinds are available: Series A connectors are for use

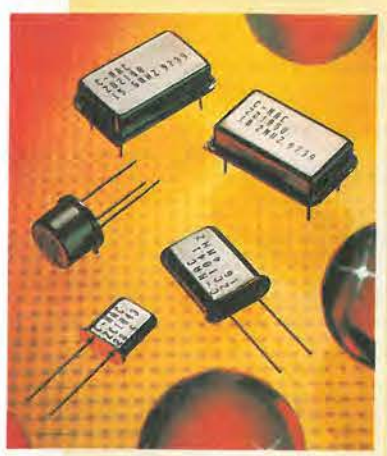
Arrays

Co-processor fpgas. From Atmel, the *AT40K* family of dynamically reconfigurable co-processor field-programmable grid arrays, in densities of 8000-50 000 gates. They are sram-based, and have an array of eight-sided look-up-table-based cells with single or dual-port sram distributed throughout the array and extensive PCI-compliant i/o and bussing options. Since, says Atmel, multipliers are the basis of high-performance computing, the eight-sided lut-based cell (for direct connection to eight adjacent cells without the use of the bus) is optimised for large array multipliers that need no routeing and the distribution of sram allows a variety of sram structures to be made without the use of logic cells. Software support comes in the form of Atmel's *FPGA Designer 5.0.* Atmel UK. Tel., 01276 686677; fax, 01276 686697.
Enq No 500



where the cable is permanently attached to a peripheral such a keyboard, mouse or hub; and Series B to be used when used with detachable peripherals such as printers, scanners and modems. The bus carries both power and data, supporting data at 1.5Mb/s and 12Mb/s, and the cable therefore consists of a 20/28awg pair for power and a 28awg twisted pair for data. Series A and B plugs and receptacles are available in selective gold flash and 0.03in gold-plated finishes and as single and double stacking form. Methode Electronics Europe Ltd. Tel., 01389 732123; fax, 01389 732777.
Enq No 511

Flexible cables. Alpha Wire announces *XTRA-Guard* continuous-use flexible control cables, which are able to withstand continuous bending in cases where machinery is in motion. The cables' flexibility conforms to MIL-C-13777G and are resistant to oil and chemicals. Small quantities are available and Alpha Wire can supply a comprehensive guide to this range. Alpha Wire Ltd. Tel., 01932 772422; fax, 01932 772433.
Enq No 512



Crystals

Custom crystals. Crystals and oscillators to customers' requirements can now be obtained within a guaranteed five or ten days or, in desperate cases, less from C-MAC Quartz Crystals; the company has a service desk via which you can talk to the people at the sharp end of manufacture. This service refers to the company's *ZC* clock oscillators with ttl or Hcmos/LS ttl output (750kHz-70MHz) and to the HC49/U, HC45 and TO-5 crystals (4-60MHz). QuickQuartz service. Tel., 01522 883520; fax, 01522 993521.
Enq No 514

Processor connectors. Beacon has produced in-circuit emulator adaptors for fine-pitch, surface-mounted processors that eliminate the nêed for alignment and without the use of glued studs or other devices, leaving no trace of the connection on the board. *Wedge* adaptors are available for both QED and Checkmate types of emulator and support the Am186 and Intel 386. Beacon Development Tools Ltd. Tel., 0117 9870444; fax, 0117 9860401.
Enq No 513

Hardware

Collapsible conduit. *Raceway* is plastic conduit for the electrical industry that avoids the difficulties of packaging and threading by arriving wound flat on a reel. On being unwound, it can then be folded along existing creases to form a rectangular section which has foam strip with adhesive for fixing to plaster, plastic or wood surfaces. Joints are made simply with a knife. When the conduit is folded, an adhesive strip seals the tube and may be re-used when further cables must be inserted. Richco International Co., Ltd. Tel., 01474 327527; fax, 01474 327455.
Enq No 515

Boxes for hand-helds. Boss's *K Box* division has the facility to construct 'T-case' housings for hand-held instruments from flat sheet plastic, no tooling or moulding being necessary. The cost is thereby much reduced and waiting time is a matter of a few days; modifications can even be incorporated during a production run. Material is 2-6mm in thickness, is UL Listed for flame retardation and may be internally coated for emi/rfi reduction. Boss Industrial Mouldings Ltd. Tel., 01638 716101; fax, 01638 716554.
Enq No 516

Knobs. Rogan *Pure Touch* round clamp knobs have a soft outer surface to make them feel better and to give a non-slip grip. They come in sizes from 19mm to 47mm diameter and may be printed with specified legends on request. Under the soft plastic surface is a harder one and under that a plated-steel threaded insert. There is a range of colours to choose from and you can have each half of the knob in different colours. The surface is treated to stop them becoming dirty. Bulgin Components plc. Tel., 0181 5945588; fax, 0181 5916913.
Enq No 517

Air filters for Armagard. *Armagard* Computer enclosures by Intek in stainless steel and mild steel, designed to protect the equipment in factory use, are now available with a range of filters to prevent the ingress of dust particles down to 5µm in size, carbon filters to remove smoke and smells, some to stop carbon black, flour and ceramic dust and types to cut out oil mist, diesel fumes and other gases. If all else fails, the enclosures can be air purged to expel the atmosphere completely. Intek Electronics Ltd. Tel., 01352 810603; fax, 01352 810403.
Enq No 518

Test and measurement

Micro-ohmmeter. Tinsley's battery-powered digital micro-ohmmeter uses a four-wire measurement technique to eliminate lead resistance and has six ranges of 600µΩ to 60Ω to a resolution of 0.1µΩ. Power comes from built-in rechargeable batteries (20h), but mains power may also be used. There is a 20mm led display, 10A of measurement current, 415V input protection, forward and reverse current measurement with auto averaging and digital calibration. A PT100 temperature probe provides temperature compensation for measurements on copper and aluminium cable. The instrument is contained in a plastic carrying case. Tinsley & Co. Ltd. Tel., 01689 800799; fax, 01689 800405.
Enq No 519

Emi receiver. The *PMM 9000* receiver measures conducted and radiated interference at frequencies from 9kHz to 1.2GHz to CISPR 16 standard. It simultaneously measures peak, quasi peak and average values to arrive at results very quickly, limit setting, measurement, saving and printing the results taking only a few keystrokes. PMM 9000 has a large colour display and its own pc to control peripheral equipment, take worst-case readings and print only those readings outside the set limits. There are RS-232 and GPIB interfaces and an 8-bit user port, hard and floppy disks, a VGA output, an internal speaker and a 'phone output. The instrument has correction facilities for any antenna and many accessories are available. Martron Instruments Ltd. Tel., 01494 459200; fax, 01494 535002.
Enq No 520

500Msample/s oscilloscope card. *CompuScope 8500/PCI* by Strategic Test is claimed to be the fastest a-to-d converter board for the PCI bus, able to sample one analogue input at up to 500Msample/s with 8-bit resolution. Data may be stored either in on-board memory or in the pc's memory via the bus at rates of 100Msample/s. For more than one channel, more boards may be used as master/slave when common clock and triggering is needed or independently, in which case different boards may have different sampling rates and memory depth. Accompanying software allows the boards to behave exactly as a digital oscilloscope with no programming needed at all and to store, analyse and print the results. A CompuScope catalogue is available. Strategic Test and Measurement Systems Ltd. Tel., 0118 9795950; fax, 0118 9795951.
Enq No 522

Hand-held dso. Looking more like an engraving tool than an oscilloscope, the *OsziFox* from Pico nevertheless has a sampling rate of 20Msample/s and features generally found on a bench-top instrument. There is a small lcd to show both waveforms and digital voltage/current measurement. The dso may also be connected to a pc (cable supplied) for screen display and to allow the signals to saved to disk. A demonstration may be downloaded from www.picotech.com. Pico Technology. Tel., 01954 211716; fax, 01954 211880.
Enq No 525

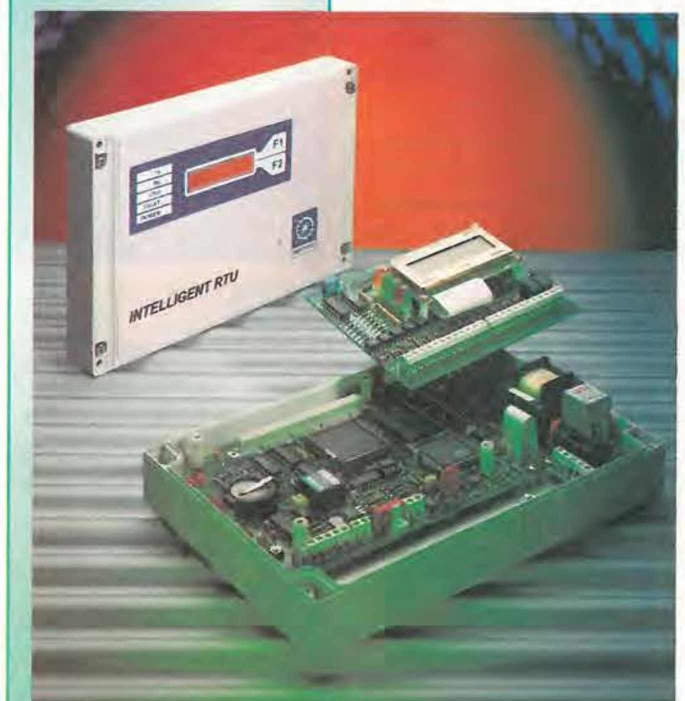
Instruments on a card. Two cards from the French company Multipower form a vhf signal generator (*SG100*) and a fast pulse generator (the *FPG10*); both plug straight into a pc's ISA bus and are software controlled. *SG100* is a vhf generator working between 80MHz and 120MHz in 10kHz steps (others in the series go up to 1250MHz and some have 2.5kHz steps), the output frequency being locked in phase to an on-board crystal reference. Output is 2V pk-pk into 50Ω. *FPG10* produces 3ns rise-time pulses down to 10ns from two independent outputs at ttl level from 50Ω, accompanied by a sync. output, pulse width being software controlled from 10ns to 5s in 10ns steps at rep. rates from 0.2Hz to 12.5MHz. A sync. output allows pre or post sync. Multipower. Tel., 033 0169301379; fax, 033 0169206041.
Enq No 523

100MHz/40Msample/s rso. Two real-time and digital storage oscilloscopes by Hitachi, the *VC-6545* and *VC-6525*, offer bandwidths of 100MHz and 50MHz and sampling at 40Msample/s and 20Msample/s respectively. Memory capacity is 4kword and 2kword and there is provision for connection to a plotter or a pc. Waveforms captured by the acquisition memory of 8kword may be backed up in a 1kword save memory and held for several days when power is switched off. Both instruments are automatically calibrated and a 100MHz counter is provided. Hitachi Denshi (UK) Ltd. Tel., 0181 202 4311; fax, 0181 202 2451.
Enq No 524

Literature

Tektronix. Tek's 1997/8 catalogue is now with us. There are 400 pages of it with 75 new products and "measurement solutions". You can see the company's web site on http:/www/tek/com/measurement, from which you can download VXI software drivers, have a twiddle with the latest oscilloscopes and use the oscilloscope selector. Tektronix UK Ltd. Tel., 01628 403300; fax, 01628 403301.
Enq No 526

Telemetry controller. Infotec has introduced the *Intelligent Controller*, a modular design for use in remote telemetry units or other types of monitor; it is based on the company's *K3* real-time hardware and software core of modules. Hardware includes a controller board, modem and power supply, with smaller boards providing i/o signal conditioning for a variety of sensors, led indicators and a front-panel lcd. Inputs and outputs are easily varied by changing daughter board layout, the company being able quickly to produce variants to measure combinations of voltage, current, temperature and other quantities. Infotec Ltd. Tel., 01530 560600; fax, 01530 560111.
Enq No 521



Power supplies. *Coutant Lambda's* 188-page catalogue, supplied free characterises dc-to-dc converters and both ac/dc switched and linear power supplies. New this time is a 10W version of the SM series of surface-mounted converters and two DIN rail-mounted ac/dc supplies. A set of application notes is available separately on request. Coutant Lambda Ltd. Tel., 01271 856666; fax, 01271 864894.
Enq No 527

Microchip. Two books from Microchip are intended for people working with *PICmicro* field-programmable microcontrollers, secure data products and serial eeproms. The 1866-page *Embedded Control Handbook* is a comprehensive collection of application notes, data and design information on PICs, while the second is the *In-circuit Serial Programming Guide* to the company's micros. Both can be obtained via distributors or from www.microchip.com. Arizona Microchip Technology Ltd. Tel., 01628 851077; fax, 01628 850259.
Enq No 528

Linear actuators. Parker Hannifin has published a 120-page catalogue of the range of *Electro-Thrust* electric cylinder actuators, which are replacements for pneumatic and hydraulic cylinders when programmable and highly repeatable positioning is needed. Frame sizes are 32, 50 and 80mm and stroke lengths 50-1000mm. Smooth control of velocity is possible at speeds up to 1250mm/s, thrust force to 7200N and positioning to within ±0.13mm. Parker Hannifin plc, Digiplan Division. Tel., 01202 699000; fax, 01202 695750.
Enq No 529

CD-rom from IR. International Rectifiers new free CD, which is regularly updated, contains almost 600 data sheets, a concise catalogue, application notes and design tips, sales information and a free copy of the Adobe Acrobat reader to make it easier to use. The CD is available from representatives and distributors and you should specify your version of Windows or Mac. International Rectifier. Tel., 01883 732020; fax, 01883 733410.
Enq No 530

Materials

Conductive fabric. Holland Shielding Systems has a highly conductive fabric for electrical shielding that needs a very low closing pressure to make good contact, thereby reducing the chances of distorting an enclosure and also allowing looser tolerances in manufacture. A flame-retardant version is available. The material comes as shaped gaskets or in rolls in widths from 10mm to 1.4m, applications including use as shielded tents and to cover the walls of Faraday cages. Holland Shielding Systems bv. Tel., 0031 78 6131366; fax, 0031 78 6149585.
Enq No 531

Power supplies

Compatible 'phone supply. *TEA1206* from Philips is a single-chip dc-to-dc converter designed to solve the problem of emi from switched-mode converters in mobile 'phones and other equipment suffering from the interference. It does this by allowing the switching frequency to be synchronised to the 'phone's reference frequency anywhere in the 9-20MHz range. Additional advantages of the device over others in the range include a continuously variable output voltage and a higher switching frequency, which allows the use of smaller reactive components. Quiescent current is 50µA and conversion efficiency over 96%. Philips Semiconductors (Eindhoven). Tel., 00 31 40 2722091; fax, 00 31 40 2724825.
Enq No 532

Adjustable plugtop supplies. Chloride can supply plugtop power supplies that are adjustable from 5V to 24V at 1.75A to 450mA by the manufacturer before dispatch or by oems. Four types are available, complying with EN60742 and EN60950 and CE marked. Voltage and current protection are incorporated and the units are in a double-insulated plastic case. Chloride Powerline. Tel., 01734 868567; fax, 01734 755172.
Enq No 533

Gas pressure sensor. The *UZU2* gas pressure sensor by Matsushita is available in positive and negative pressure form in the range –5MPa to 1MPa and all models will display in various units, including bar, psi, mmHg and inHg. There is a 3.5-digit led display and adjustments include zero point setting, and two set points for upper and lower acceptance window levels, all by keys on the front panel. Digital output is n-p-n or p-n-p and also a 1-5V analogue signal proportional to the pressure. The unit measures 30 by 30 by 24mm, is sealed to IP40 and comes with a mounting bracket. Matsushita Automation Controls Ltd. Tel., 01908 231555; fax, 01908 231599.
Enq No 540

"Smallest" dc-to-dc converter. *NPH15* is claimed by XP to be the world's smallest 15W converter and also costs less than other 15W types. It is the first in a new family of miniature converters of up to 40W output, combining an efficiency of 90% with a 50mm by 25mm package, standing 10mm high. The converters

provide a single output, have a 2:1 input range and 1.5kV isolation. Current limit is non-latching and surge rating is over 25W. Outputs are 3.3V, 5V, 12V and 15V. A sync. pin allows for fine frequency adjustment and may also be used for remote shutdown. XP plc. Tel., 01189 845515; fax, 01189 843423.
Enq No 534

High-voltage supply. Features of Keithley's *Model 248* 5kV, 5mA source include low noise, programmable output filtering, precision current readback and small size – 3.5in half-rack. Control is from the front-panel, over an IEEE-488 interface or by an analogue voltage from a remote source, panel meters indicating *V*, *I* and setting. There are store and recall functions for up to nine complete settings, including trip points. Voltage setting accuracy is ±0.01% of setting + ±0.05%of range, display accuracy ±1V; stabilisation 0.0001% for 10% input change and regulation 0.005% for 100% load change. Current output and display is also closely controlled. Keithley Instruments Ltd. Tel., 01189 575666; fax, 01189 596469.
Enq No 535

Thin power supply. *MPU150* low/medium-power, ac/dc, single and multiple output, switched-mode supplies from Power-One are contained in a U-channel case 1.5in high (1U). Input range is 85-264V ac with power-factor correction, and a later introduction will be a 48V dc input model. To some extent the units are modular, in that components not needed in particular applications may be removed. Outputs are at standard voltages and, with air cooling from an optional fan or from the user's system, will provide a total of 150W, all the usual protection features being present. All relevant standards requirements are met. Power-One Europe. Tel., 01769 540744; fax, 01769 540756.
Enq No 536

Radio communications products

2GHz amplifier. The GaAs *AM50-0006* microwave ic by M/A-COM is a high-dynamic-range, amplifier covering the 1400-2000MHz band for personal communications and having a noise figure of 1.6dB. It finds application in receiver front ends and is also useful as a gain block, buffer, driver and if amplifier in fixed and portable systems. It uses external matching for best noise figure and frequency flexibility. Supply needed is 3-8V at a current of 3-20mA, controllable by an external resistor. BFI IBEXSA Electronics Ltd. Tel., 01622 882467; fax, 01622 882469.
Enq No 537

Switches and relays

Sealed switches. "Blows to the operating button" of Bulgin's *MP0037/8* sealed switches will be thwarted by shoulders in the switch body to limit travel, and there are seals to stop dust and water getting in, to the satisfaction of IP66. These switches are made of stainless steel with a threaded body and a single nut (not the one with the hammer), so there is no need for studs and other bits and pieces. The *0037* has a single-pole, slow, momentary make/break action and the *0038* a single-pole, momentary, snap-action changeover operation. Ratings are 1A, 50V ac (0037) and 5A, 250V ac (0038). Gothic Crellon Ltd. Tel., 01734 788878; fax, 01734 776095.
Enq No 538

High-current ssr. Teledyne's *IGTA* and *IGTD* series of high-voltage, high-current, bidirectional and dc solid-state relays use an insulated-gate bipolar transistor at the output to control ac, dc or bidirectional dc high-voltage loads. *IGTA 1200480R100-L*, for example, is a dc/ac ssr to control 100A at a line voltage of 480V ac or 800V dc, having logic-level control input and a random cross switch with inverse parallel igbts to withstand 1200V transients. Isolation is to 2500V and *dv/dt* rating 500V/μs. Teledyne Literature Centre. Tel., 01634 670820; fax, 01634 863494.
Enq No 539

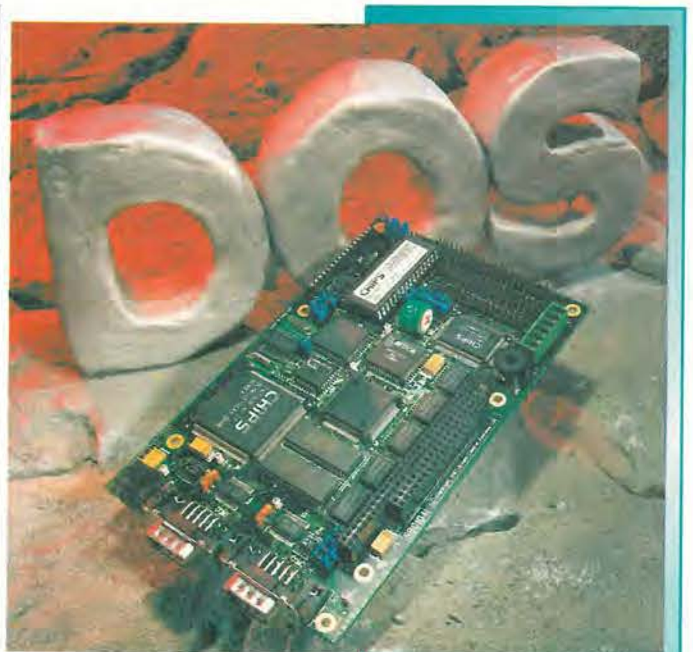
Transducers and sensors

Pyranometer. Made in Holland by Kipp & Zonen, the *SP-LITE* is a silicon pyranometer to measure solar energy received from the whole hemisphere, one of its uses being to determine the power that may be used in solar energy applications. Output can be connected to a voltmeter or data logger. A photodiode provides a voltage output proportional to the cosine of the angle of incidence of the radiation, which, says K&Z, gives consistent and accurate measurement. Kipp & Zonen. Tel., 01727 858098; fax, 01727 842185.
Enq No 541

COMPUTER

Computer board-level products

VME64 processor boards. VI Computer's new range of *PowerPC*-based VME64 processor boards include an entry-level unit, the 100MHz or 200MHz single-slot *Power3e* which is believed to offer the best performance:cost ratio available and a 300MHz single-board computer and a CompactPCI sbc using a 300MHz PowerPC. This is the *Power4e* sbc, said to be the fastest PowerPC VME engine yet seen, which is provided with 256Mbyte of memory and a full set of i/o. The *CompactPCI4e* is the same unit in PCI form. All boards have fast Ethernet facilities. Sirretta Ltd. Tel., 0118 9258080; fax, 0118 9258070.
Enq No 542



Computers

Larger-screen handheld. *Geofox-One* is a compact computer, the size of a paperback, but with a larger screen than is found in other models of this size and a mouse-pad. Its software, running under Psion's *EPOC32* operating system, offers a full complement of word processing and spreadsheet features and a set of personal organiser programs, including a calculator, route planners with maps, digital voice recorder and, for exhausted professionals, games and a crossword solver. It is compatible with Windows to allow file transfer between Geofox and pc and the wp and spreadsheet are compatible with Word and Excel. Memory size is up to 16Mb and there is a full pc card slot; the Professional models also contain a modem. Geofox Ltd. Tel., 01223 425444; fax, 01223 425422.
Enq No 544

Accelerated graphics. Taiwanese company Soyo has a new motherboard for systems designed for the Pentium II processor in Slot 1 and meant for use with the high-speed graphics interface specified by Intel, the Accelerated Graphics Port (AGP). The *SY-6KB* ATX-format board uses the Intel *430LX* PCI chipset and, in addition to the 32-bit, 66MHz (effectively 133MHz since it uses both edges of the clock) AGP, uses the standard PCI and ISA expansion buses. AGP uses a dedicated channel instead of the PCI bus to give direct main memory access. It has four 168-pin dimm sockets for up to 512MB of sdram or 1GB of EDO dram for main memory and Slot 1 allows a Pentium II with 256KB or 512kB of cache. Soyo UK Ltd. Tel., 0181 481 9720; fax, 0181 481 9725.
Enq No 545

Single-board computer. Arcom's *SBC104* single-board computer is a dos-based, embedded device provided with either a 25MHz Intel 386 or a 50MHz Texas 486SX. New features of the computer include a parallel printer port, RS-485 serial port and another site for flash eprom. All models have 2Mbyte or 4Mbyte of dram, 1Mbyte or 2Mbyte of flash eprom and an optional 128Kbyte of battery-backed ram. Each board comes with rom-dos v. 6.22 loaded in flash eprom and using the Arcom *Flash Filing System*, which eliminates the need for disk drives. Expansion is possible by means of a 16-bit PC/104 interface and a range of Arcom and other compatible modules. Arcom Control Systems Ltd. Tel., 01223 411200; fax, 01223 410457.
Enq No 543

Data communications

Radio modem. Radio Data Technology announces the *RM9600* wireless modem, which has on-board RS232 and RS485 serial ports to interface with pc and plc networks, the same unit being used for data logging or IEEE-compliant control. This is a medium-range module working at 9600baud with forward error correction to allow programming and down-loading of data at the speed commonly used by pcs and plcs. 'Listen-before-transmit' and a 10ms turn-round help to avoid interference. The transceiver uses single-frequency, half-duplex mode in bands of up to 32 channels with 12.5, 20 or 25kHz spacing in the 406-470MHz band. Output power is 500mW maximum. No licence is needed. Radio Data Technology Ltd. Tel., 01376 501255; fax, 01376 501312.
Enq No 546

Development and evaluation

Peripheral emulators. The *Ceibo DS-300* peripheral development tool comprises hardware and software to support file generation and emulation for Wafer Scale Integration's *PSD-300* series of programmable microcontroller peripheral chips, which bring together many peripheral functions into one chip. *DS-300* comes with configuration software to allow memory, i/o, bus width and the dpld to be set up in a gui running under Windows. There are 16Kbit of sram and 1024Kbit of eprom, which is also emulated in ram to avoid the need to generate a new file when modification is needed. Great Western Instruments Ltd. Tel., 0117 983 0333; fax, 0117 9860401.
Enq No 547

Data logging

Temperature logger. Lascar has a new range of data loggers, the *EL-SOL-TEMP* series, which measure, record, display and control temperature in the -50°C to 250°C range. The *EL-2-12BIT* version has user-selected sample rates from 5s to 12h between samples and a capacity of 8064 readings. Modules are set by means of a pc running the supplied Windows-based control software, data being then displayed on screen or exported to a spreadsheet for analysis; once set up by the pc, the logger can be disconnected from the pc and left to record. The hand-held or wall-mounted temperature probe is already calibrated and the whole thing comes in a plastic carrying case. Lascar Electronics Ltd. Tel., 01794 884567; fax, 01794 884616.
Enq No 548

Infrared data logging. *EasyLog-HL* is a hand-held programmer and data retrieval unit that works with any of the *EasyLog* data loggers, but which works by infrared, so that the logger may be programmed and the logged data retrieved with no cable being needed. The corresponding data logger *EL-2-IR* is available. An advantage is that no portable pc is required and another is that capital costs are reduced greatly. Sample rates from the logger, which has an lcd and internal battery, are variable between 5s and 12h, with a maximum of 8000 readings. Lascar Electronics Ltd. Tel., 01794 884567; fax, 01794 884616.
Enq No 549

Programming hardware

Prom programmer. Data I/O's *ChipWriter Gang* programs and verifies at voltages down to 3.3V for a range of memories including 8-bit eproms, flash, paged eproms and 8Mb eeproms in a variety of packages. This PC-based or stand-alone unit uses the same pin driver techniques as *ChipWriter* and *ChipWriter Portable*, so there is no need for adaptors, family-specific modules or special software for most dips to 32 pins. Adaptors are available for plccs, SOICs and TSOPs to 32 pins. The instrument gang-programs up to eight devices or, since it has four data buses, will set-program devices with 16 and 32-bit wide data. Eight 1Mb devices can be down-loaded, programmed and verified in 23 seconds. Direct Insight Ltd. Tel., 01280 700262; fax, 01280 700577.
Enq No 550

Software

Basic for PICs. RF Solutions has come with a Basic compiler to speed the programming of Microchip's PIC microcontrollers. The compiler is suitable for use with *PIC16CXX* and *PIC14000* devices, for which the *PicBasic Compiler* converts Basic into hex or binary to be programmed directly to the PIC. Its instruction set is compatible with the Parallax *Basic Stamp 1*, so that a *Basic Stamp* module is not needed. The compiler uses Peek and Poke instructions to give access to on-chip features such as a-to-d converters and i/o with no need to go back to assembler language, and I²C commands allow communication with external devices such as eeproms on the two-wire interface. It runs under dos or Windows and allows the mixing of Microchip's MPASM assembly language with Basic. RF Solutions Ltd. Tel., 01273 488880; fax, 01273 480661.
Enq No 551

Thermal analysis. Flomerics announces version 2 of *Flotherm*, which is a complete redesign of the original thermal analysis software; the new version greatly reduces the time needed to determine the cooling needs of systems. New features include a gui to allow the user to operate in a cad-like manner, editing, creating and manipulating data by mouse; three application windows to provide three different ways of viewing the model and all interactive, the Visualisation window using 3D graphics showing, among other effects, images of particle tracks and heat flux lines. A radiation model automatically calculates views between surfaces to ease the problem of including radiation effects in the model. Flomerics Ltd. Tel., 0181 9418810; fax, 0181 9418730.
Enq No 552 ■

Colour stereo tv in '41

Ray Herbert has discovered that stereoscopic television pictures – in colour – were first demonstrated over half a century ago. And they were produced by a man who some now say made little contribution to modern television.

Few people are aware that both stereoscopic and colour television were demonstrated in 1928 – but not both together. That was to follow on only 13 years later.

On 9 August 1928, the press was invited to see stereoscopic television pictures using the 30-line spotlight system at the Baird laboratories. Briefly, this system involved transmitting two images from the studio alternately, the first as seen by the right eye, and next the left eye view. At the receiving end these images appeared in rapid succession side by side. When they were viewed through a prismatic stereoscope, a single image could be seen with a good perception of depth.

Stereoscopic television pictures were shown at the 1928 British Association meeting held in Glasgow during September, but it seems that no further experiments were carried out here or elsewhere until Baird resumed this work in 1941.

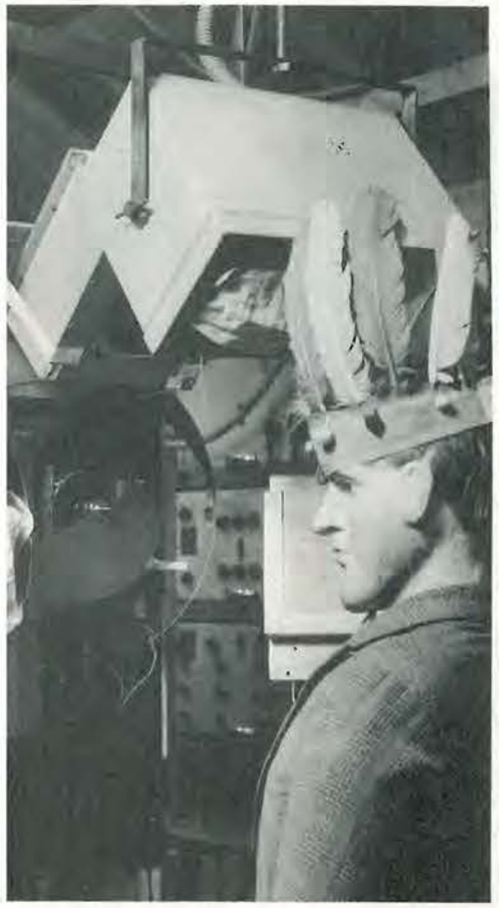
Options for stereoscopic tv

Baird considered several different arrangements for producing stereoscopic pictures. Among these were the use of polarised light and the anaglyphic method which required displaced images of differing colours and tinted glasses for viewing. Both of these alternatives were discarded for reasons of cost or complexity. Instead, Baird opted for an adaptation of his 1928 system but with the added

3D pictures were projected from the television receiver on the right to the field lens in front of the person in the chair – in this case John Baird.

benefit that the viewer would not need to use either a prismatic stereoscope or coloured spectacles.

Electronic colour cameras had not yet arrived and Baird employed the flying spot or spotlight method of transmission which had been used for the low definition public service between 1929 and 1935. The subject to be televised stood in an unlit studio and was scanned in sequential parallel strips by a sharply focused pin-point of brilliant light. As the spot traversed the scene, panchromatic multiplier photocells with a large cathode area recorded the level of reflected light.



Eustace – a tailor's dummy – in front of the stereoscopic scanner. Note the specially shaped shutter for producing the left and right eye perspectives alternately.

Baird's patented tube coating

These special photocells were developed in the laboratories of Baird Television Ltd by Dr A. H. Sommer. He found that the correct characteristics could be obtained from a combination of bismuth, silver, caesium and oxygen. After the war, this patented material came to be used in all camera tubes of the image orthicon type.

For these stereoscopic experiments John Baird had chosen a definition of 500 lines – 100 lines interlaced five times – with a repetition rate of 150 frames per second, and horizontal scanning. This higher definition meant abandoning mechanical scanning and substituting an electronic arrangement using a high intensity cathode-ray tube operating at around 37000 volts.

Projection tubes of this type had been used by the Baird Company in 1939 for large screen television in several London cinemas. They were known as teapot tubes because of their appearance. Cylindrically shaped, the neck entered at an angle, the fluorescent screen being deposited on an aluminium disc instead of the tube itself. Scanning had to be carried out obliquely and this required keystone correction. The light beam was front projected through the optically flat tube face.

In order to obtain the correct stereoscopic effect the subject had to be scanned from slightly different positions, the displacement being equal to the average separation of the eyes. This was accomplished by splitting the beam using pairs of mirrors, a revolving shutter ensuring that the differing perspectives were transmitted alternately. To reduce flicker, which would have resulted from an abrupt changeover of the scanning position, the shutter was specially shaped so that the area obscured in one light beam at any moment was equal to the area uncovered on the other.

Adding colour

Colour discrimination was achieved by placing a disc containing six segments with red, green and blue filters in the scanning beam. This resulted in the subject being scanned successively with red, green and blue light. The red scan produced maximum reflection from those parts of the scene containing this colour. Other pigments absorbed all or most of the red beam and did not produce a signal from the photocells. Similar considerations applied to the green and blue scans.

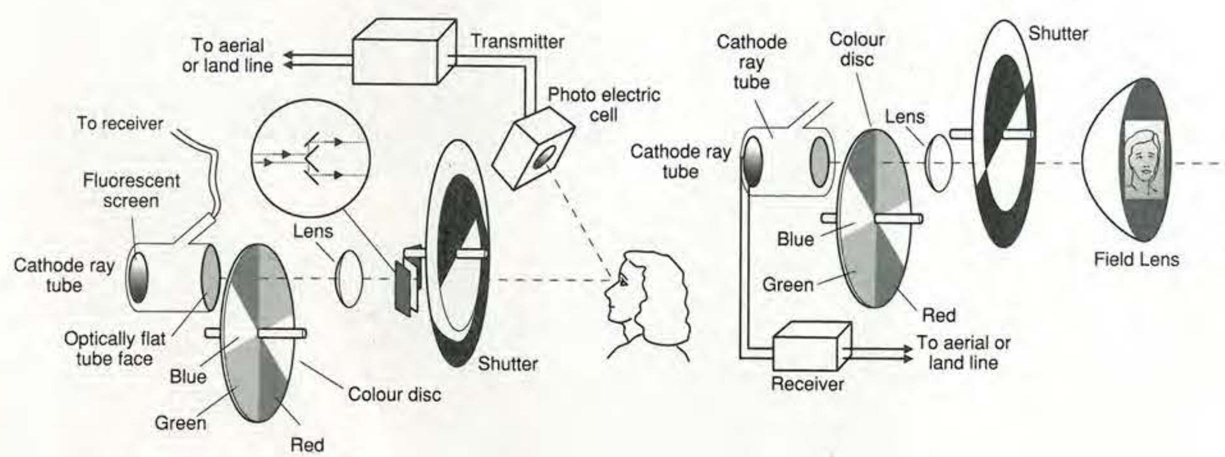
At the receiving end another teapot tube displayed the basic 500-line monochrome picture, colour being added by the rotating colour filter disc which had to be synchronised with the studio scanner. A shutter provided right and left eye perspectives in rapid succession, producing on the image forming lens a stereoscopic picture in full colour.

Although coloured spectacles were not required, this system suffered the disadvantage that the viewer had to be in a fixed position. If not, the stereoscopic effect disappeared.

The wartime demonstrations in Baird's rambling Georgian house were relaxed affairs. Usually, separate sessions were arranged for the newspapers and the technical press. John Baird handed out ham sandwiches and a press release which he had typed himself. It omitted the more subtle design details but retired members of Baird Television Ltd, who visited the laboratory during the war, have since supplied additional information.

First demonstrations

On the afternoon of 18 December 1941, the journalists sat in turn before the image forming lens and witnessed the first stereoscopic television pictures in colour to be seen anywhere in the world. Whatever may have been their thoughts regarding the electro-optical lash-up, they were impressed by the results.



Demonstrated to journalists in 1941, Baird's stereoscopic television system involved complete 100-line frames being repeated 150 times a second and successively scanned through red, green and blue filters at the transmitter end, left. These were interlaced five times resulting in a 500-line picture.

Wireless World reported "If the colour reproduction lacked the ability in this early experiment to differentiate the subtler shades, it dealt faithfully with the bolder colours. The stereoscopic effects were an unqualified success, and when the person being televised reached towards the 'camera' his arm at the receiving end seemed to project out of the lens towards the viewer."

By any standards, these demonstrations represented a remarkable achievement. John Baird had relinquished his position with Baird Television Ltd shortly after the outbreak of war in 1939. He financed the colour development work from his savings, supplemented by a consultancy fee from Cable and Wireless.

At the time of the stereoscopic demonstrations only one assistant remained, but Baird also had the part-time services of a retired glass technologist who kept an antique shop in the Crystal Palace Parade. The enterprising editor of *Electronic Engineering* arranged for a colour photograph to be taken directly from the receiver screen, thus preserving for posterity the result of this important contribution to television progress.

John Logie Baird died on 14 June 1946, leaving his colour work unfinished and, as it has turned out, largely unpublicised. At the time of his death he was experimenting with a special type of cathode-ray tube for stereoscopic television which incorporated an internal revolving fluorescent screen. He was also looking at a colour projection system using three separate tubes.

It has been said that John Baird made little contribution to modern television, but this is clearly absurd when you remember that he produced the first multi-gun colour tube in 1944. This was described in *Wireless World*'s October 1944 issue. He also demonstrated stereoscopic television – in colour – 56 years ago. We have yet to catch up. ■

Stereoscopic pair of images photographed directly from the screen of Baird's receiver on to Dufaycolor film in 1941.

SPEAKERS' CORNER

John Watkinson explains why some woofers are good – and others aren't.

The job of a woofer is relatively simple because it works over a frequency range where the wavelength is much larger than the diaphragm. This means that the whole diaphragm can work as a simple piston yet the radiation from a moderate sized enclosure will still be omni-directional. It also means that the woofer can be mounted magnet outwards without affecting the sound quality.

The goal in a woofer diaphragm is rigidity. Many different approaches have been tried to make rigid diaphragms. The traditional cone shape is used because it is stiffer than a flat plate, but cones still have straight lines which allow flexing. A flared or compound curved 'cone' will be stiffer. Some woofer cones are solid foam plastic, others have a core of foam between skins of aluminium foil. Anodised aluminium gives a rigid cone because the anodised layers are stiff and hard and form a sandwich with the softer aluminium as a core.

Some designs of woofer reduce the stresses on the cone, so that less rigidity is needed. With a traditional small coil in **Fig. 1a)** the drive force is applied a long way from the air load and the cone sees bending moments.

At **b)** a larger diameter coil drives the diaphragm more uniformly as some of the moving mass and air load is the central dome. The maximum distance from any part of the diaphragm to the coil is reduced. A further advantage is that the larger coil can dissipate heat more easily.

The woofer is driven by a motor consisting of a coil in the radial field of a magnet. Some of the flux leaks at either side of the gap and some of the drive force is produced in the gap leakage field.

It is important for low distortion that the leakage field is symmetrical so that the force produced is independent of the position of the coil. This requires a magnetic circuit which is designed to equalise the reluctance above and below the gap. **Fig. 2a)** shows some examples.

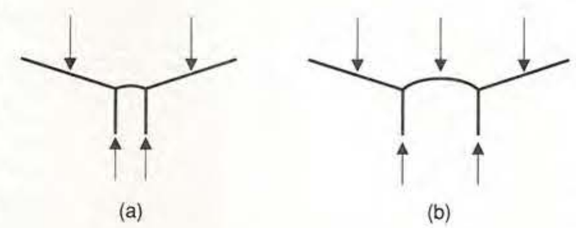


Fig. 1. For a woofer, the benefit of a large coil diameter, as in b), is that it spreads the driving force over a larger area so there's less likelihood of the cone bending under the strain.

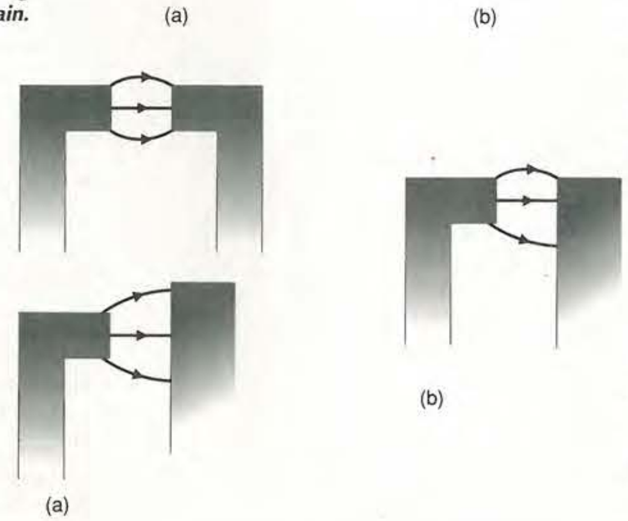


Fig. 2. Good pole design is shown in a). Pole design in b) is cheaper but causes asymmetrical flux pattern.

No such thing as cheap

Unfortunately many cheap woofers use pole pieces which are designed to be easy to make as in **b)**. The result is distortion which renders these units unsuitable for high quality applications.

The loudspeaker magnet should produce a field in the gap, but in practice this is not as easy as it sounds. Although some materials have low reluctance and carry magnetic flux readily, there is no such thing as a magnetic insulator. Actually there is, but superconducting materials are impractical.

As a result, a lot of the flux from the magnet is lost to leakage. Magnetic materials are classified by the available magneto-motive force per unit of length and the available flux per unit area.

Alnico magnets need a long columnar structure as shown in **Fig. 3a)** which naturally goes up the centre of the coil. However, the high cost of cobalt made Alnico very expensive and today most speakers use ferrite.

Watch out for leaks

Ferrite has a poor available flux per unit area and so needs a large cross sectional area **b)**. This means that the magnet has to go outside the coil and there is a serious external leakage because of the large surface area. This leakage flux will distort the picture on nearby crts. Steel chassis are a further problem with ferrite magnets since the steel increases the leakage.

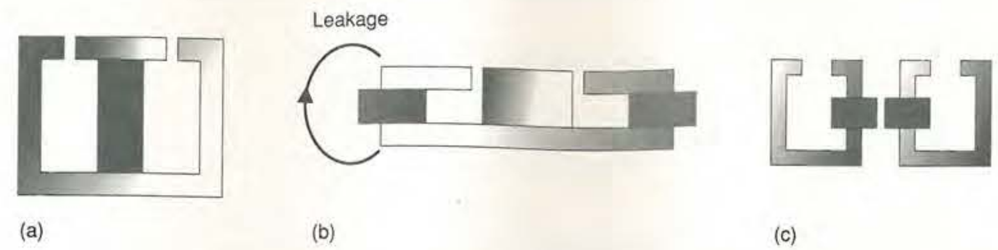


Fig. 3. Columnar Alnico magnet is represented in a) while b) shows how external ferrite magnet leaks flux badly. With the rare-earth magnet in c), no leakage occurs.

More recently rare earth magnets have become available which can be made small enough to fit inside the coil once more **c)**. This allows leakage to be eliminated and a steel chassis is no longer a problem.

According to Newton's laws, when the coil accelerates, there is a reaction which tries to distort the magnetic field. This is called flux modulation and can be a source of distortion. Flux distortion is minimised if the magnetic circuit is electrically conductive to create shorting turns.

If the flux tries to move it will induce heavy currents in the shorting turns. Rare-earth magnets are better than ferrite here because they are electrically conductive whereas ferrites are insulators. In some designs copper shorting rings are incorporated in the magnetic circuit.

Below resonance, the motion of the cone is controlled by the stiffness of the suspension which should be dominated by the spider which locates the coil former. The displacement is proportional to coil current and velocity leads the current. Above resonance the system is mass controlled. The acceleration is proportional to coil current and velocity lags the current. At the resonant frequency the velocity is exactly in phase with the coil current. Because of this phase characteristic the polarity of a loudspeaker is a matter of opinion.

Manufacturers mark one terminal with a red spot or a + sign as an aid to wiring in the correct polarity. However some use the convention that a positive dc voltage – for example from a battery – will cause forward motion of the cone, whereas others use the convention that the positive half cycle of an ac voltage at a frequency above resonance will cause forward motion.

Clearly these two conventions are in perfect opposition. The ac definition makes more sense as that is how the speaker is used, however most manufacturers use the dc definition.

The phase reversal of a moving coil driver as it passes through resonance means that it is incapable of reproducing the input waveform at frequencies near resonance. If it is intended to reproduce the input waveform accurately the fundamental resonance must be placed below the audio band at around 20Hz or signal processing must be used to artificially lower the resonance. ■

John Watkinson, FAES

List 1. Turbo-pascal 6 for controlling the eight-bit resolution radio-linked data logger via the printer port. Although the port normally communicates parallel data, this routine reads the serial coming directly from the radio receiver and thus serves as a general example of how to read serial data via the printer port.

```
Program Centronics_radio_AD;
{TP6 Demonstration program for radio data logger developed by Dr. Pei An, 8/97}
uses
     dos, crt, graph;
var
    bitnumber,outputbyte,dummy:byte;
    P_address:integer;

Procedure Input_printer_address;
{Universal auto detection of printer base address}
{ $000:$0408 holds the printer base address for LPT1
  $000:$040A holds the printer base address for LPT2
  $000:$040C holds the printer base address for LPT3
  $000:$040e holds the printer base address for LPT4
  $000:$0411 number of parallel interfaces in binary format}
var
    lpt:array[1..4] of integer;
    number_of_lpt,LPT_number,code:integer;
    kbchar:char;
begin
       clrscr;
       LPT_number:=1; {default printer}
       number_of_lpt:=mem[$0000:$0411]; {read number of parallel ports}
       number_of_lpt:=(number_of_lpt and (128+64)) shr 6;
       lpt[1]:=memw[$0000:$0408];        {Memory read procedure}
       lpt[2]:=memw[$0000:$040A];
       lpt[3]:=memw[$0000:$040C];
       lpt[4]:=memw[$0000:$040E];
       textbackground(blue); clrscr;
       textcolor(yellow); textbackground(red); window(10,22,70,24); clrscr;
       writeln('Number of LPT installed    :  ',number_of_lpt:2);
       writeln('Addresses for LPT1 to LPT 4:  ',lpt[1]:3,'   ', lpt[2]:3,'   ', lpt[3]:3,'   ', lpt[4]:3);
       write('Select LPT to be used (1,2,3,4)    :  ');
       delay(1000);

       if number_of_lpt>1 then begin   {select LPT1 through LPT4 if more than 1 LPT installed}
         repeat
            kbchar:=readkey;                {read input key}
            val(kbchar, LPT_number, code);   {change character to value}
         until (LPT_number>=1) and (LPT_number<=4) and (lpt[LPT_number]<>0);
                                    end;
       clrscr;
       P_address:=lpt[LPT_number];
       writeln('Your selected printer interface:  LPT',LPT_number:1);
       write('LPT  Address                   :  ',P_address:3);
       delay(1000);
       textbackground(black); window(1,1,80,25); clrscr;
end;

Procedure timedelay;
{A short time delay}
var
    dummy:real;
    i:integer;
begin
      for i:=1 to 40 do dummy:=0;
end;
Function bit_weight(bitnumber:integer):integer;
{find the bit weight}
begin
      if bitnumber=1 then bit_weight:=1;  {find the bit weight of the selected bit}
      if bitnumber=2 then bit_weight:=2;
      if bitnumber=3 then bit_weight:=4;
      if bitnumber=4 then bit_weight:=8;
      if bitnumber=5 then bit_weight:=16;
      if bitnumber=6 then bit_weight:=32;
      if bitnumber=7 then bit_weight:=64;
      if bitnumber=8 then bit_weight:=128;
end;

Function inputdata:real;
{read radio serial data into the computer and find the value of the data}
var
    bytex: array [1..3000] of byte;
    high,low: array [1..50] of integer;
    i,counter,status,minimum_high,minimum_low, minimum_count,start_point:integer;
    total_number,point_found_count, scan_number, width_number, byte_value:integer;
    read_point, bitx :array [1..10] of integer;
    flagx:boolean;
begin
       {read radio serial data into memory 3000 times with a timedelay
       between each read. data assigned to bytex[i], i from 1 to total_number}
       Total_number:=3000;
       for i:=1 to total_number do begin
            bytex[i]:=port[P_address+1];
            timedelay;
                          end;

       {make bytex[i] eith to be 0 or 1}
       for i:=1 to 3000 do begin
           bytex[i]:=round((bytex[i] and 8)/8);
           {if bytex[i]<>bytex[i-1] then writeln;
           write(bytex[i]);}
       end;

       {readln;}
       {analysis of radio serial data}
       {find number of data for serial data bit =0s or =1s before it changes}
       for i:=1 to 50 do begin high[i]:=0; low[i]:=0; end;
       counter:=1;
       status:=bytex[1];
       for i:=1 to total_number do begin
           if bytex[i]<>status then begin counter:=counter+1; status:=bytex[i]; end;
           if bytex[i]=1 then high[counter]:=high[counter] + 1;
           if bytex[i]=0 then low[counter]:=low[counter]+1;
                                  end;

       {for i:=1 to 15 do writeln(i,'  1=',high[i],'  0=',low[i]);}
       {find minimum counts for 1 and 0, They represents the synchronising 1 and 0}
       minimum_high:=250; minimum_low:=250;
       for i:=3 to counter-2 do begin
           if (high[i]>0) and (high[i]<minimum_high) then minimum_high:=high[i];
           if (low[i]>0) and (low[i]<minimum_low) then minimum_low:=low[i];
                         end;

       {writeln('minimum_high=',minimum_high,'   minimum_low=',minimum_low);}
       {find the starting point of a serial data transmission}
       for i:=3 to counter-2 do begin
           flagx:=( high[i+2]>minimum_high+20) or (low[i+1]>minimum_low+20);
           if (abs(high[i]-minimum_high)<=5) and flagx then
              begin
                   start_point:=i;
                   i:=counter-2;
              end;
                              end;

       {writeln('start of data=', point_found[1]);}
       {find the number of scans for a data bit, width_number }
       width_number:=2*(minimum_high+minimum_low);
       {use minimum_high and minimum_low to find the width_number}
       {find the read point}
       scan_number:=0;
       for i:=1 to start_point do scan_number:=scan_number+high[i]+low[i];
       read_point[1]:=scan_number+round(width_number/2)-5;
       for i:=2 to 8 do read_point[i]:=(i-1)*width_number + read_point[1];
       byte_value:=0;
       for i:=1 to 8 do begin
           byte_value:=byte_value+bytex[read_point[i]]*bit_weight(9-i);
                        end;
     {  writeln('skip number=',skip_number);
       for i:=1 to 8 do write(bytex[read_point[i]],'  ');}
       inputdata:=byte_value*2.5/256;
end;

Function AD_conversion:real;
{read data from the radio logger three times, check the data and find the average}
{if the difference amongst the three readings is too big, read the data again}
var
    dummy1,dummy2,dummy3, error, average:real;
begin
      {read data from radio logger three time and the find the average satisfying the
      error requirements}
      repeat
            dummy1:=inputdata;                   {read date the 1st time}
            delay(1);                            {a short delay}
            dummy2:=inputdata;                   {read data the 2nd time}
            delay(1);
            dummy3:=inputdata;                   {read data the 3rd time}
            average:=(dummy1+dummy2+dummy3)/3;   {find the average}
            error:=sqrt((sqr(dummy1-average)+sqr(dummy2-average)+sqr(dummy3-average))/3);
                                                 {find the square root error of the three
readings}
      until (average=0) or (error/average<=0.05);{if error/average<5%, data accepted}
      if average=0 then AD_conversion:=0 else AD_conversion:=average;
end;

   {***********Main program*********}
begin
      input_printer_address;
       begin
            clrscr;
            writeln;
            writeln('                 Centronics radio data logger testing program');
            repeat
            gotoxy(15,10); write('AD conversion from the radio logger
:',ad_conversion:5:3);;
            until keypressed
      end;
end.
```

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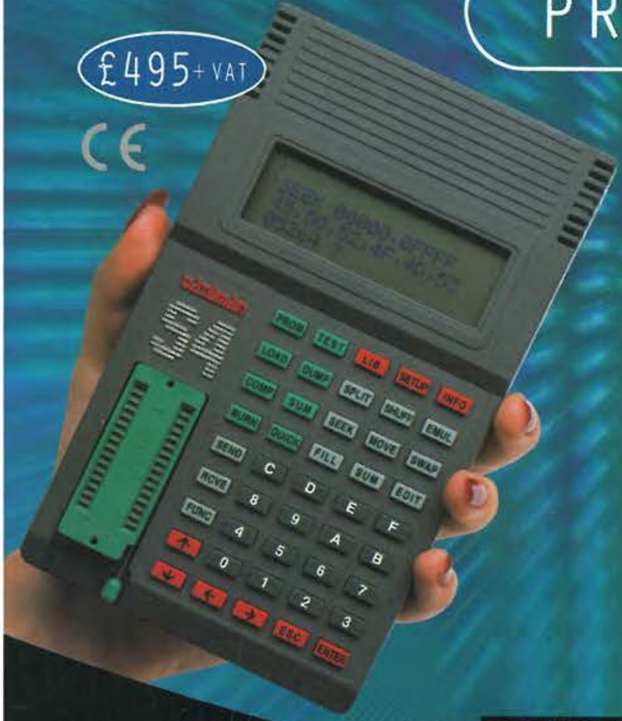

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