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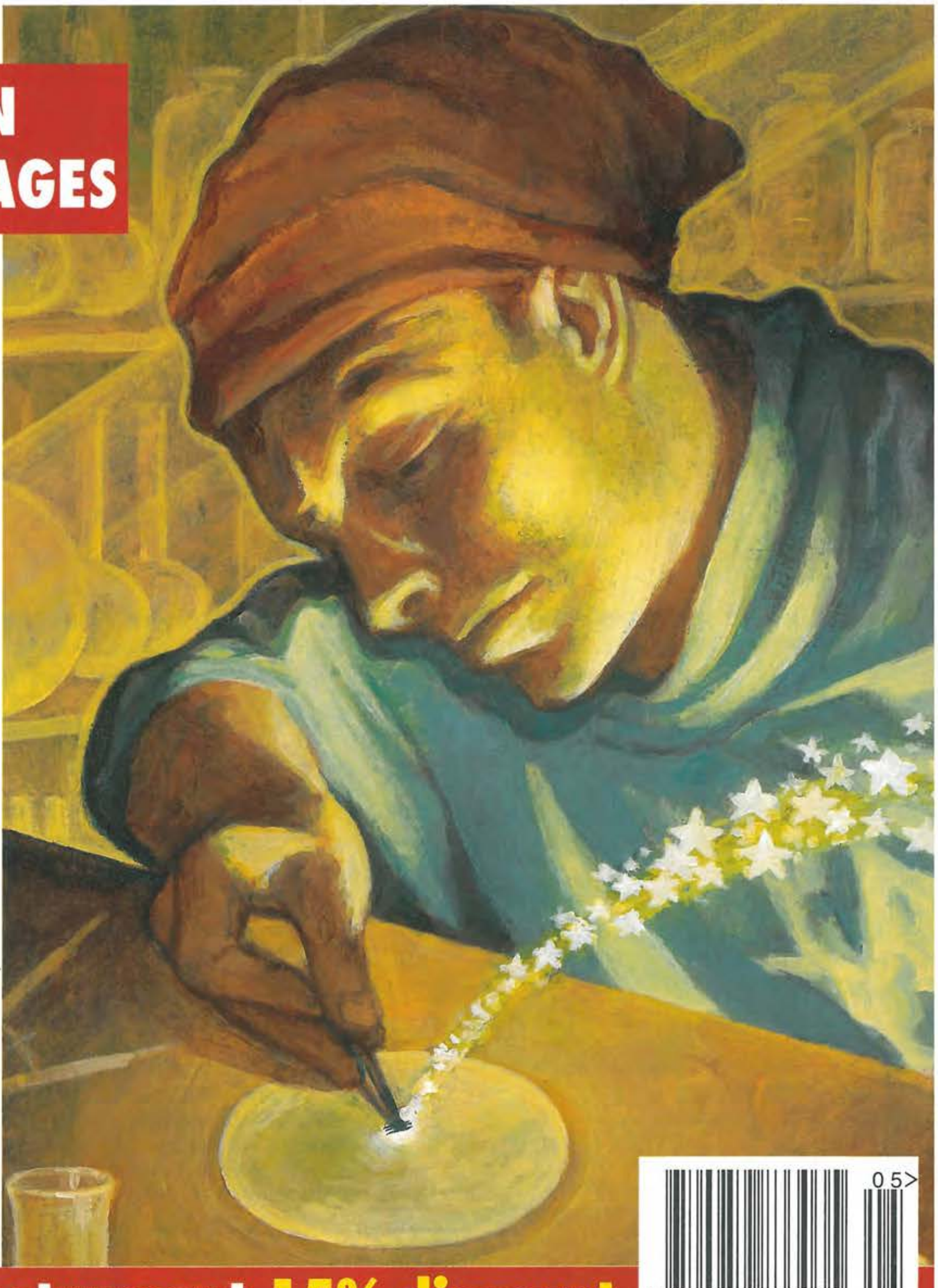

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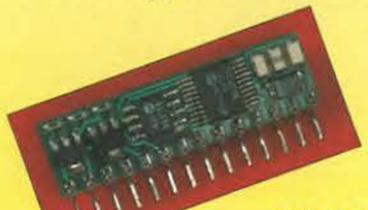

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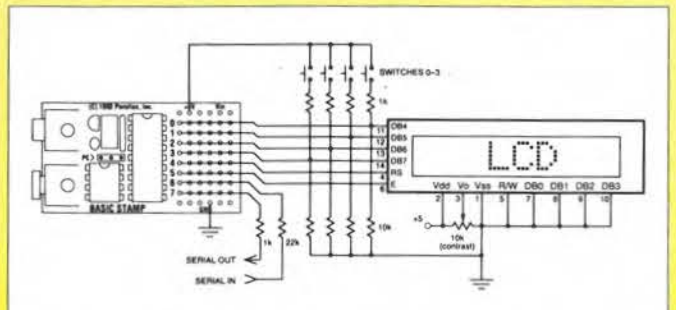

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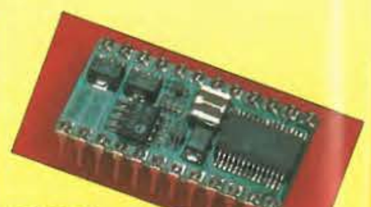

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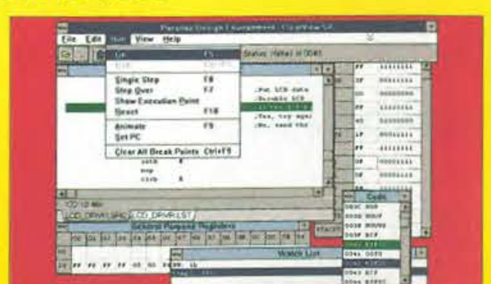

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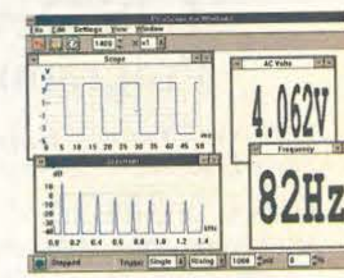

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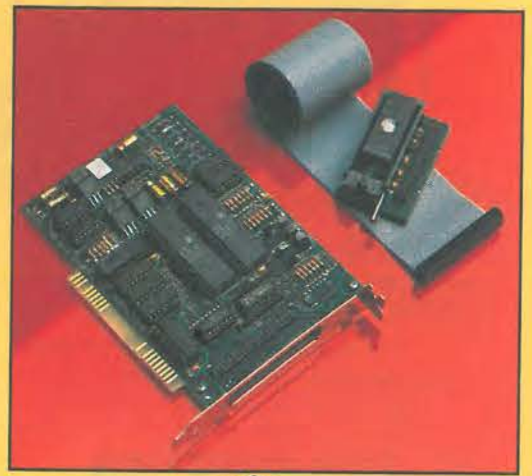

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REED BUSINESS INFORMATION

Who needs robots?

There is a problem at work. Computer power has fallen into the hands of petty authority, the directors and managers of private companies, giving them such power over controlling the minutiae of people's working lives that it could be labelled oppressive. For some, this has meant a radical change for the worse. Victorian mill-owners would have recognised what is being done today to workers and would have been been envious of the ease with which employees' lives can be controlled with the pc.

This would be acceptable if managers had an enlightened attitude to their workers and if integrity was their watchword. But employees are treated as commodities like machine tools or office furniture and are manipulated by managers who mix greed and fear to run their companies. Let me explain how with two examples I have seen.

First, an articled clerk in a medium-sized firm of solicitors had what she felt was an interesting job. There was plenty of social interaction with colleagues and clients and some leeway to pursue cases as she saw fit. Much of her work was done on the phone. Then a pc was installed to monitor her working day. The number of tasks, especially telephone contacts, per hour was set. After collecting the information, the bosses tried to get everyone to conform to a norm, and penalise those who fell below it. Then they slowly increased the pace. Now this person describes her job as detestable. She feels the computer is looking over her shoulder all day, which it is, and the quality of her work is being sacrificed for quantity. She can't wait to get out.

In the second case, a firm had about 40 technicians doing fault-finding on complex mechanical-electronic equipment in the field. As dual mechanical and electronic skills were needed, ready-made high-calibre technicians were hard to recruit. The best were those who had learnt the hard way with many years experience. They did about three field calls a day. A computer was installed to log what they were doing. There was a computer input code for every activity. For a time, the field force carried on with this extra overhead of computer bureaucracy, at a rate of three calls a day, feeding in the vast amount of data required. There were few complaints because this was the time of Thatcherism and everyone feared for their jobs.

Soon the managers tried to impose norms and increase the number of calls per day, to eliminate what they saw as non-productive time and link pay to computer-generated performance figures. What followed was like the famous Charlie Chaplin film sequence where the factory's production-line conveyor belt was speeded up. It would have been equally comical had it not been so serious for those at the receiving end. They had to skimp on engineering quality to do more calls. They no longer had time for reading service manuals or modification notes, and brushing up their technical skills. There was no time for exchanging information between technicians.

Some worked longer hours – up to 60 a week – but still declared 38 hours. Others gave the computer false data. Some found ways to defeat the system. After two years, the computer print-out looked fine, the management was delighted, and the accountants were pleased, but the reality was a disaster. The firm had 40 disgruntled technicians and hundreds of dissatisfied customers whose machines were no longer up to scratch. Technicians who could were already leaving for other companies. These were the better technicians whose qualifications enabled such mobility. And it was even harder to find replacements because the word had spread about what was going on.

Hardware and software engineers who provided the material for these instances are to be criticised. The computer systems were far too intrusive and overbearing. The software did not record customer and job satisfaction, and was used simply to bash workers.

As for the unions, the technology took them unawares. What is needed is a political initiative, not just to stop intrusive computer surveillance systems interfering with private lives, as is presently supposed, but to prevent them being implemented willy nilly in the workplace.

Designing such oppressive software should be an offence as providing the means of close surveillance of ordinary people is an affront to human rights. Laws to govern such practice would not be easy to frame or implement. But the time has come to look closely at such an idea.

There is another ghastly spectre. As far as I know, nobody has linked camera surveillance of a workforce to computer performance monitoring, but this will come. A type of electronic tagging could even be introduced like that being tried on criminals. Employers could track their workers to see where they are. A type of tagging already exists for some employees with pagers. Field staff in many firms already have to carry them and are not allowed to turn them off in working hours. It's a small step from a pager to an electronic tag. Mobile phone users can also be tracked to within a few feet, even when not using their phones. This could also be adapted and misused in a similar way.

If this deluge of electronic surveillance in working hours is allowed to be implemented without legislation to control it, Big Brother will surely have arrived. **Simon Wright**

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UPDATE

Cebit launch of rewritable cd and flat 14.5in lcd monitor

Philips exhibited at Cebit its 14.5in flat LCD monitor in the Brilliance range, which is only 6.3cm thick. "The market for flat TV is expected to reach 1m unit sales in 2000," said Doug Dunn, president & CEO of Philips Sound & Vision.

Philips used the German technology show Cebit '97 to launch what it claims to be the world's first cd-rewritable (CD-RW) drive.

The drive, *CDD3610*, will let users read and write up to 1000 times data on their own cds and will cost around $900.

Based on recording phase-change technology, the cds offer a storage capacity of 650Mbyte and can be read on CD-recordable, CD-RW and multi-read CD-ROM drives, which accounts for only 12 per cent of the market. However, CD-RW discs will be forward compatible with future DVD digital-video disk drives.

This announcement ties in with Philips' plans to launch a $600 DVD video player in the US in spring. Two European models will follow at the Berlin IFA consumer show in September, costing up to $1000.

DVD-ROM drives will be available to oems at the end of this summer. DVD-recordable and DVD-RAM specifications are expected soon.

"With respect to DVD-recordable and DVD-RAM, the discussions are ongoing," said Jan Oosterveld, president of Philips Key Modules.

"We are close to a final format proposal. Version 0.9 for DVD-recordable was issued last week and the final version of the DVD-RAM specifications is expected in the coming months."

There is also talk of a hybrid DVD/CD format.

Svetlana Josifovska,
Electronics Weekly

Motorola rail-to-rail op amp at 1V

Motorola has introduced a rail-to-rail input and output op amp that operates from 1V.

A key part of the *MC33502* is an input stage using variable threshold depletion-mode n-channel mosfets.

The conventional way to design a rail-to-rail input for an op amp is to incorporate two input stages in parallel. A pnp or p-type input to handle inputs near the negative rail and an npn or n-type stage to operate near the positive rail. The next stage combines signals from both stages.

This approach works well but is complex; the open loop gain tends to waver at the handover between n and p inputs.

Depletion-mode n-channel mosfets are 'on' when their gate is at the same potential as their source. This means the new input stage inherently includes the 0V rail.

Under normal circumstances this would also mean they would saturate and become useless once the inputs were only a little above the 0V rail.

By connecting the bodies of the mosfets to 0V, rather than their sources (which is the norm), the threshold voltage of the mosfet is made to follow the input voltage. They slide smoothly into enhancement mode as the input increases, keeping them from saturating even with the input connected to the positive rail.

The final *MC33502* device operates from 1 to 7V with an input offset of 0.5mV. Gain bandwidth is 4MHz at 1V and the output can drive 10mA at 1V and up to 50mA at 5V.

Researchers at the University of Wales in Bangor have developed an electronic sensor that can identify bacteria types. It acts like the electrodynamic equivalent of a linear induction motor. Polyphase signals are applied to either side of a track along which the cells can travel. The cells polarise in the field and go in one or other direction. The photo shows a junction separating two bacteria types.

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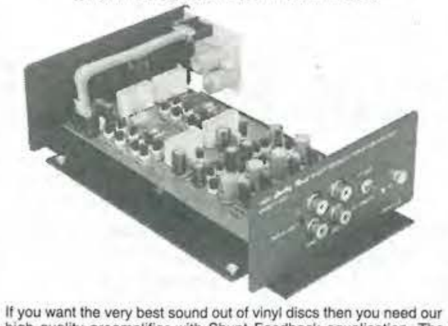

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CIRCLE NO. 107 ON REPLY CARD

Anger over EC ram interference

Semiconductor producers with memory plants in Europe were gearing up production before the return of the EC reference price (RP) on dynamic ram in March.

Meanwhile, equipment makers were angry with what they saw as unjustified interference in a commodity market that will raise their costs. PC prices are bound to rise.

"RP has been suspended for as long as legally possible", an EC official said. "The suspension ends on 8 March and that means there will be a return to the old measures. That is the law."

"It's a waste of space," said Sadru Nanji, the director at ICL responsible for procurement. "Memory is a commodity market. Intervention is counter-productive."

Two fixed prices were applied on d-ram on 10 March – RP for Japanese d-ram, calculated on a weighted average of Japanese costs, and a minimum price for Korean companies based on the individual costs of each company. This may vary.

Prices in contracts signed for Japanese d-ram before 8 March stay valid, but the Korean companies cannot get advice from the EC as to whether this also applies to them.

The level of the fixed prices remains secret, however. A manufacturer said the price would be as high as $5.5 for a 4Mbit d-ram and $13 for a 16Mbit. That is ahead of the expectations of Siemens, which said recently that it expected a $10 RP.

Both estimates contrast sharply with a current contract price of $7.50 for 16Mbits and a spot market price of $9. That is up sharply on sub-$6 prices for 16Mbits in early January, and results from Samsung and LG deciding not to take further d-ram orders.

The return of RP will help the 64Mbit d-ram. Manufacturers need an introduction price of $50 for the 64Mbit if they are to recoup their investment.

David Manners, *Electronics Weekly*

Movie star was first frequency hopper

Military radio specialists have something in common with Hedy Lamarr, the Hollywood movie icon of the 1930s and 1940s.

During World War II Lamarr, who starred opposite Clark Gable and Spencer Tracy, was granted a US patent on a "secret communications system". This, it turns out, was the first manifestation of a frequency-hopping radio protocol.

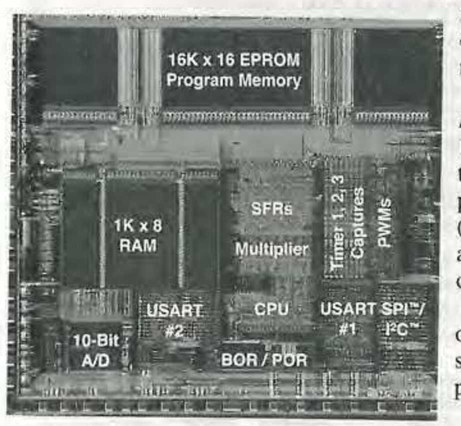
Using the 88 keys of the piano, Lamarr proposed that a radio signal transmitted in parts across random frequencies would be difficult to jam. According to a history professor in Hamburg, her collaborator on the proposals was composer George Antheil, a future husband.

Lamarr made a mistake in trying to sell the idea to the US navy for control of its torpedoes. It was ignored and only in 1962 was frequency-hopping radio first used – three years after Lamarr's patent expired.

PIC of the bunch microcontroller

Microchip Technology has announced the highest performance PIC-series 8-bit microcontroller family yet.

The first family member, the *PIC17C756*, is an 8.25Mips device with a 10-bit a-to-d converter. It is aimed at the upper end of the 8-bit market to compete with Motorola's *68HC11* and Hitachi's *H8* families. Applications may include set-top boxes, motion and process control, printers, airbag controllers and modems.



The *PIC17C756* has 16k×16 of one-time programmable (OTP) memory and 902×8-bit of user ram.

Joe Connelly of Microchip said: "We have put plenty of ram on-chip for compatibility with C compilers."

Like other family members it has a single cycle 8×8 hardware multiplier.

"It is a straight forward multiply, rather than a multiply-accumulate (MAC). We could make a device that was more DSP-like in future by making this a MAC," said Connelly.

A 10-bit pulse width modulator is included to let the processor set the speed and position of motors in motion control applications.

Microchip has a flash memory process, but has chosen to release this chip with OTP rom.

"We have been surprised that flash-based versions of previous microcontrollers have been less popular than we thought," said Connelly. "Masked rom used to be choice for production, but the masking lead-time is now often a significant part of a product's lifetime, which can be as low as nine months."

Sharefun may be no fun for e-mail receivers

US anti-virus firm McAfee says it has discovered the first computer virus that searches through a user's e-mail and automatically generates and transmits e-mail with virus attachments.

However, many e-mail users are claiming this is the latest in a long line of e-mail virus hoaxes.

McAfee says the virus, called Sharefun, affects users of Microsoft's popular Microsoft Mail application.

A user becomes infected by opening a Microsoft Word document that contains Sharefun. The virus looks for MS Mail, finds three e-mail addresses at random and generates an e-mail with the subject heading "You have GOT to read this!"

Then it attaches the same Sharefun-infected Word document, transmitting the infected e-mail to the three addresses. McAfee has a detector program for Sharefun posted on:
www.mcafee.com/corp/press/press.html

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HP141T+8552B IF + 8555A 10Mc/s–18GHzS – £1200.
HP ANZ Units Available separately – New Colours – Tested
HP141T Mainframe – £350.
HP8552B IF – £300.
HP8553B RF 1KHz to 110Mc/s – £200.
HP8554B RF 100KHz to 1250Mc/s – £500.
HP8555A RF 10Mc/s to 18GHzS – £800.
HP8556A RF 20Hz to 300KHzS – £250.
HP8443A Tracking Generator Counter 100KHz–110Mc/s – £300.
HP8445B Tracking Preselector DC to 18GHz – £350.
HP3580A 5Hz – 50KHz ANZ – £750 – £1000.
HP3582A .02Hz to 25.6KHz – £2k.
HP8568A 100Hz–1500Mc/s ANZ – £6k.
HP8569B 10Mc/s–22GHz ANZ – £6k.
HP Mixers are available for the above ANZ's to 40GHz
TEK 492 – 50KHz – 18GHz Opt 1+2 – £4k–£4.2k.
TEK 492 – 50KHz – 18GHz Opt 1+2+3 – £4.5k.
TEK 492P – 50KHz – 21 GHz Opt 1+2+3 – £5k.
TEK 494AP 1KC/S – 21GHz – £7k.
TEK 496P 1KHz–1.8GHz – £4k.
TEK 5L4N 0–100KHz – £400.
TEK 7L5 + L1 – 20Hz–5Mc/s – £700.
TEK 7L5 + L3 – Opt 25 Tracking Gen – £900.
TEK 7L12 – 100KHz–1800Mc/s – £1000.
TEK 7L18 – 1.5–60GHzs – £1500.
TEK 491 10Mc/s–12.4GHzs–40GHzs – £750. 12.4Ghzs–40Ghzs with Mixers.
Tektronix Mixers are available for above ANZ to 60GHzs
Systron Donner 763 Spectrum ANZ + 4745B Preselector .01–18GHz + Two Mixers 18–40GHz in Transit Case – £3k.
HP8673D Signal Generator .05–26.5GHz – £20k.
Systron Donner 1618B Microwave AM FM Synthesizer 50Mc/s 2–18GHzs
R&S SWP Sweep Generator Synthesizer AM FM 4–2500Mc/s – £3.5k.
ADRET 3310A FX Synthesizer 300Hz–60Mc/s – £600.
HP8640A Signal Generators – 1024Mc/s – AM FM – £800.
HP3717A 70Mc/s Modulator – Demodulator – £500.
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HP5316B Universal Counter A+B.
HP6002A Power Unit 0–5V 0–10A 200W.
HP6825A Bipolar Power Supply Amplifier.
HP461A–465A–467A Amplifiers.
HP81519A Optical Receiver DC–400Mc/s.
HP Plotters 7470A–7475A.
HP3770A Amplitude Delay Distortion ANZ.
HP3770B Telephone Line Analyser.
HP8182A Data Analyser.
HP59401A Bus System Analyser.
HP6260B Power Unit 0–10V 0–100 Amps.
HP3782A Error Detector.
HP3781A Pattern Generator.
HP3730A+3737A Down Convertor Oscillator 3.5–6.5GHz.
HP Microwave Amps 491–492–493–494–495–1GHz–12.4GHz – £250.
HP105B Quartz Oscillator – £400.
HP5087A Distribution Amplifier.
HP6034A System Power Supply 0–60V 0–10A–200W – £500.
HP6131C Digital Voltage Source+ – 100V ½ Amp.
HP4275A Multi Frequency L.C.R. Meter.
HP3779A Primary Multiplex Analyser.
HP3779C Primary Multiplex Analyser.
HP8150A Optical Signal Source.
HP1630G Logic Analyser.
HP5316A Universal Counter A+B.
HP5335A Universal Counter A+B+C.
HP59501B Isolated Power Supply Programmer.
HP8901A Modulation Meter AM – FM – also 8901B.
HP5370A Universal Time Interval Counter.
Marconi TF2370 – 30Hz–110Mc/s 750HM Output (2 BNC Sockets+Resistor for 500HM MOD with Marconi MOD Sheet supplied – £650.
Marconi TF2370 30Hz–110Mc/s 50 ohm Output – £750.
Marconi TF2370 as above but late type – £850.
Marconi TF2370 as above but late type Brown Case – £1000.
Marconi TF2374 Zero Loss Probe – £200.
Marconi TF2440 Microwave Counter – 20GHz – £1500.
Marconi TF2442 Microwave Counter – 26.5GHz – £2k..
Marconi TF2305 Modulation Meter – £2.3k.
Racal/Dana 2101 Microwave Counter – 10Hz–20GHz – £2k.
Racal/Dana 1250–1261 Universal Switch Controller + 200Mc/s PI Cards.
Racal/Dana 9303 True RMS Levelmeter+Head – £450. IFFE – £500.
TEKA6902A also A6902B Isolator – £300–£400.
TEK 1240 Logic Analyser – £400.
TEK FG5010 Programmable Function Generator 20Mc/s – £600.
TEK2465A 350Mc/s Oscilloscope – £2.5k + probes – £150 each.
TEK CT-5 High Current Transformer Probe – £250.
TEK J16 Digital Photometer + J6523–2 Luminance Probe – £300.
TEK J16 Digital Photometer + J6503 Luminance Probe – £250.
ROTEK 320 Calibrator + 350 High Current Adaptor AC–DC – £500.
FLUKE 5102B AC–DC Calibrator – £4k.
FLUKE 1120A IEEE – 488 Translator – £250.
Tinsley Standard Cell Battery 5644B – £500.
Tinsley Transportable Voltage Reference – £500.
FLUKE Y5020 Current Shunt – £150.
HP745A+746A AC Calibrator – £600.
HP8080A MF + 8091A 1GHz Rate Generator + 8092A Delay Generator + Two 8093A 1GHz Amps + 15400A – £800.
HP54200A Digitizing Oscilloscope.
HP11729B Carrier Noise Test Set .01–18GHz – LEF – £2000.
HP3311A Function Generator – £300.
Marconi TF2008 – AM-FM signal generator – also sweeper – 10Kc/s – 510Mc/s – from £250 – tested to £400 as new with manual – probe kit in wooden carrying box.
HP Frequency comb generator type 8406 – £400.
HP Vector Voltmeter type 8405A – £400 new colour.
HP Sweep Oscillators type 8690 A & B + plug-ins from 10Mc/s to 18GHz also 18-40GHz. P.O.R..
HP Network Analyzer type 8407A + 8412A + 8501A – 100Kc/s – 110Mc/s – £500 – £1000.
HP Amplifier type 8447A – 1-400Mc/s £200 – HP8447A Dual – £300.
HP Frequency Counter type 5340A – 18GHz £1000 – rear output £800.
HP 8410 – A – B – C Network Analyzer 110Mc/s to 12GHz or 18GHz – plus most other units and displays used in this set-up – 8411a – 8412 – 8413 – 8414 – 8418 – 8740 – 8741 – 8742 – 8743 – 8746 – 8650. From £1000.
Racal/Dana 9301A – 9302 RF Millivoltmeter – 1.5-2GHz – £250-£400.
Racal/Dana Modulation Meter type 9009 – 8Mc/s – 1.5GHz – £250.
Marconi RCL Bridge type TF2700 – £150.
Marconi/Saunders Signal Sources type – 6058B – 6070A – 6055A – 6059A – 6057A – 6056 – £250-£350. 400Mc/s to 18GHz.
Marconi TF1245 Circuit Magnification meter + 1246 & 1247 Oscillators – £100-£300.
Marconi microwave 6600A sweep osc., mainframe with 6650 PI – 18-26.5GHz or 6651 PI – 26.5-40GHz – £1000 or PI only £600. MF only £250.
Marconi distortion meter type TF2331 – £150. TF2331A – £200.

Tektronix Plug-Ins 7A13 – 7A14 – 7A18 – 7A24 – 7A26 – 7A11 – 7M11 – 7S11 – 7D10 – 7S12 – S1 – S2 – S6 – S52 – PG506 – SC504 – SG502 – SG503 – SG504 – DC503 –DC508 – DD501 – WR501 – DM501A – FG501A – TG501 – PG502 – DC505A – FG504 – 7B80 + 85–7B92A
Gould J3B test oscillator + manual – £150.
Tektronix Mainframes – 7603 – 7623A – 7613 – 7704A – 7844 – 7904 – TM501 – TM503 – TM506 – 7904A – 7834 – 7623 – 7633.
Marconi 6155A Signal Source – 1 to 2GHz – LED readout – £400.
Barr & Stroud Variable filter EF3 0.1Hz – 100kc/s + high pass + low pass – £150.
Marconi TF2163S attenuator – 1GHz. £200.
Farnell power unit H60/50 – £400 tested. H60/25 – £250.
Racal/Dana 9300 RMS voltmeter – £250.
HP 8750A storage normalizer – £400 with lead + S.A or N,A Interface.
Marconi TF2330 – or **TF2330A** wave analysers – £100-£150.
Tektronix – 7S14 – 7T11 – 7S11 – 7S12 – S1 – S2 – S39 – S47 – S51 – S52 – S53 – 7M11.
Marconi mod meters type TF2304 – £250.
HP 5065A rubidrum vapour FX standard – £1.5k.
Systron Donner counter type 6054B – 20Mc/s – 24GHz – LED readout – £1k.
Racal/Dana 9083 signal source – two tone – £250.
Systron Donner – signal generator 1702 – synthesized to 1GHz – AM/FM – £600.
Tektronix TM515 mainframe + **TM5006** mainframe – £450 – £850.
Farnall electronic load type RB1030-35 – £350.
Racal/Dana counters – 9904 – 9905 – 9906 – 9915 – 9916 – 9917 – 9921 – 50Mc/s – 3GHz – £100-£450 – all fitted with FX standards.
HP4815A RF vector impedance meter c/w probe – £500-£600.
Marconi TF2092 noise receiver. A, B or C plus filters – £100-£350.
Marconi TF2091 noise generator. A, B or C plus filters – £100-£350.
Marconi 2017 S/G 10Khz – 1024MHz.
HP180TR, HP182T mainframes £300-£500.
Philips panoramic receiver type PM7900 – 1 to 20GHz – £400.
Marconi 6700A sweep oscillator + 18GHz PI's available.
HP8505A network ANZ + 8503A S parameter test set + 8501A normalizer – £4k.
HP8505 network ANZ 8505 + 8501A + 8503A.
Racal/Dana VLF frequency standard equipment. Tracer receiver type 900A + difference meter type 527E + rubidium standard type 9475 – £2750.
HP signal generators type 626 – 628 – frequency 10GHz – 21GHz.
HP 432A – 435A or B – 436A – power meters + powerheads – Mc/s – 40GHz – £200-£1000.
Bradley oscilloscope calibrator type 192 – £600.
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HP8616A signal gen 1.8GHz – 4.5GHz, new colour £400.
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HP 8683D S/G microwave 2.3 – 13GHz – opt 001 – 003 – £4.5k.
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HP 86290B Sweep PI – 2 – 18GHz – £1250.
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HP 8569B Analyser .01 – 22GHz – £5k.
HP 3580A Analyser 5Hz – 50kHz – £1k.
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HP 3437A System voltmeter – £300.
HP 3581C Selective voltmeter – £250.
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HP 3770A Amp delay distortion analyser – £400.
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HP 3781B Pattern generator (bell) – £300.
HP 3782A Error detector – £400.
HP 3782B Error detector (bell) – £300.
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HP 8016A Word generator – £250.
HP 8170A Logic pattern generator – £500.
HP 59401A Bus system analyser – £350.
HP 59500A Multiprogrammer HP – IB – £300.
Philips PM5390 RF syn – 0.1 – 1GHz – AM + FM – £1000.
S.A. Spectral Dynamics SD345 spectrascope 111 – LF ANZ – £1500.
Tektronix R7912 Transient waveform digitizer – programmable – £400.
Tektronix TR503 + TM503 tracking generator 0.1 – 1.8GHz – £1k – or TR502.
Tektronix 576 Curve tracer + adaptors – £900.
Tektronix 577 Curve tracer + adaptors – £900.
Tektronix 1502/1503 TDR cable test set – £1000.
Tektronix AM503 Current probe + TM501 m/frame – £1000.
Tektronix SC501 – SC502 – SC503 – SC504 oscilloscopes – £75-£350.
Tektronix 465 – 465B – 475 – 2213A – 2215 – 2225 – 2235 – 2245 – 2246 – £250-£1000.
Kikusui 100Mc/s Oscilloscope COS6100M – £350.
Nicolet 3091 LF oscilloscope – £400.
Racal 1991 – 1992 – 1988 – 1300Mc/s counters – £500-£900.
Fluke 80K-40 High voltage probe in case – BN – £100.
Racal Recorders – Store 4 – 4D – 7 – 14 channels in stock – £250 – £500.
Racal Store Horse Recorder & control – £400-£750 Tested.
EIP 545 microwave 18GHz counter – £1200.
Fluke 510A AC ref standard – 400Hz – £200.
Fluke 355A DC voltage standard – £300.
Wiltron 610D Sweep Generator + 6124C PI – 4 – 8GHz – £400.
Wiltron 610D Sweep Generator + 61084D PI – 1Mc/s – 1500Mc/s – £500.
Time Electronics 9814 Voltage calibrator – £750.
Time Electronics 9811 Programmable resistance – £600.
Time Electronics 2004 D.C. voltage standard – £1000.
HP 8699B Sweep PI YIG oscillator .01 – 4GHz – £300. 8690B MF – £250. Both £500.
Schlumberger 1250 Frequency response ANZ – £1500.
Dummy Loads & power att up to 2.5 kilowatts FX up to 18GHz – microwave parts new and ex equipt – relays – attenuators – switches – waveguides – Yigs – SMA –APC7 plugs – adaptors.
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CIRCLE NO. 108 ON REPLY CARD

Eye robot, I can see you

Robots may soon be able to see in the same way as humans.

Researchers in Zurich plan to mimic in silicon the way the human brain assimilates data sent to it from the eye.

This should pave the way for robots that make sense of their surroundings.

Rodney Douglas, at the Institut für Neuroinformatik, said his team had already connected an artificial one-dimensional retina to an integrated circuit (IC)-based brain. This proved the artificial brain could follow what the artificial retina was seeing, though only in one dimension.

"Now," said Douglas, "we have developed a 2D retina using the same communication protocol."

An interface is being built to link the latest retina to the artificial brain. Douglas expects this to be ready soon.

The fingernail-sized artificial retina behaves like the human retina by reporting the contrast between an image'§ pixels rather than luminous intensity levels, as used in traditional cameras.

The institute's brain is biologically inspired, modelled on the routeings between the brain's neurons. It reflects these paths by using analogue vlsi devices linked via a serial bus.

Neighbouring ICs have many connections between them, while those further away use fewer ones. However, it has vastly fewer connections than the human brain.

"We do in time what the brain does in space," said Douglas, explaining that his system worked at 10MHz while the brain managed 1kHz.

Mercedes-Benz is threatening to remove the need for a steering wheel with its latest idea, the F200 Imagination. The car's control system dispenses with mechanical and hydraulic linkages, transmitting driver commands electronically using computerised drive-by-wire technology. To steer the car, the driver moves a joystick left and right, controlling acceleration and braking by moving it forwards and backwards.

NEWS IN BRIEF

Researchers at Cornell University in the US claim to have produced a universal substrate that may prove the key to making **blue and UV semiconductor lasers**.

Single crystals of one semiconductor will not grow on another if there is a difference in crystal lattice spacing. For instance, GaAs will not grow on silicon and there is only a four per cent difference in spacing.

"We have made an elastic layer on top of a GaAs substrate, which allows mismatched semiconductors to be grown on top," said Yu-Hwa Lo, a Cornell researcher. "We have grown InGaPh with a one per cent mismatch, GaSb, with eight and InSb with a huge 15 per cent."

US firm Floating Images has introduced an inexpensive system that provides **3D images on a tv** set or monitor without special goggles.

The firm's Real-Deph technology provides horizontal and vertical binocular parallax, letting viewers peer around objects to view images previously hidden. The company claims that unlike other stereoscopic systems, its technology does not produce headaches or eye strain.

To view Real-Deph images, a TV has to be fitted with an adapter costing as little as $25 to make in quantity. The images to be displayed must be reformatted with special software. The company is working on a way to reformat 2D images in real-time to the Real-Deph format.

Motorola will soon make available beta releases of its field **programmable analogue array** (FPAA).

It is based on technology licensed from UK firm Pilkington Microelectronics.

Bill Altonen, programme manager for FPAAs at Motorola, said: "We will be supplying selected customers in April. We are keeping the numbers to a minimum to allow us to provide all the support needed."

ADSL **high speed Internet** over the telephone line is the focus of an alliance between chipset supplier Motorola and Internet services software specialist Sourcecom.

The intention is to combine Motorola's ADSL (asymmetric digital subscriber line) transceiver chipset with Sourcecom's Internet software to create an off-the-shelf reference design for a high speed Internet product.

The University of Washington in Seattle has made a **micromachined fluid pump** with only one moving part, based on ideas patented 75 years ago by the eccentric inventor Nikola Tesla.

The pump consists of a silicon wafer with a shallow circular hollow etched into its surface.

What's in a name? Well, you should **choose your company name carefully** or you may suffer the fate that befell US X Windows graphics software firm X Inside.

It seems too many people confused the company with the many hundreds of companies offering adult images on the Internet and which make prolific use of the letter X in their names. Company president Thomas Roell said that the firm received numerous requests for pornographic materials and in one instance, an aspiring female model showed him a portfolio of suggestive photographs.

The firm's new name is Xi Graphics, which might solve the problem. Then again, it might not.

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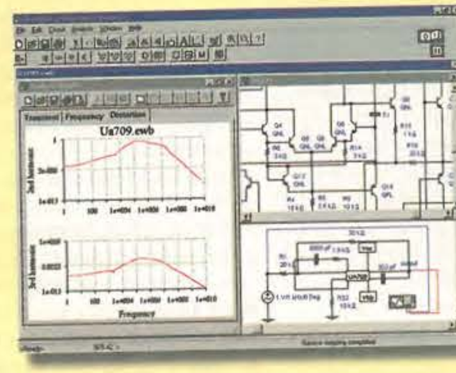

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RESEARCH NOTES

Jonathan Campbell

Laser measures wind turbulence

A prototype, non-Doppler optical sensor that makes inexpensive, accurate measurements of cross-wind speeds over long distances could hold promise for chemical manufacturing, aviation safety and meteorology. The single-ended, long-path laser wind sensor, developed by researchers at Georgia Tech Research Institute, registers faint wind movements that an anemometer cannot measure.

"The sensor is more sensitive and accurate than mechanical anemometers, and it may provide an advantage when monitoring winds over a wide area, by providing a low-cost alternative to complex beam arrays of traditional sensors," explains Mikhail Belen'kii, principal research scientist in the Electro-Optics, Environment and Materials Laboratory of the GTRI.

Its design is simple. An inexpensive helium-neon laser, about 50mm in diameter, projects a beam of light 30m onto a target made of retroreflective materials used on highway signs. A telescope then collects the laser light reflected by the target, and sends it through a series of optics, including two small, horizontally-separated detectors, each of which monitors a spot on the target inside the laser beam. Using a laser beam phenomenon known as the residual turbulent scintillation effect, the detectors pick up shadowy waves, or fringes, moving across the laser beam – just like the shadows of waves created on the bottom of a swimming pool on a sunny day.

Each of the two detectors in the sensor registers the moment at which a dark fringe passes its view. By digitising the points at which detectors pick up a single wave, a computer can measure time and separation, and can work out the average velocity of a massive column of air crossing the laser beam.

Even though air may be flowing erratically – some going in one direction at one end of the beam and some going exactly the opposite direction – the researchers say the method can still be used to get a net flow across the laser beam.

So far the sensor is reported to have correlated extremely well with anemometer readings in test results.

The researchers point out that the sensor is easier to use than Doppler systems, and measures wind across the beam of light instead of along the beam. And, unlike conventional systems, this sensor can pick up turbulence.

Next step is to test the sensor with technologies that measure airborne pollutant concentrations at a real refinery plant. Measuring concentration and cross wind at the same time gives a good idea of the rate at which a pollutant is leaving a plant. But because the sensor measures average wind directions over long distances, the researchers believe it might also have additional applications in aviation, meteorology, or aerosol dispersion studies.

Contact: Dr Mikhail Belen'kii, Georgia Institute of Technology, 223 Centennial Research Building, Atlanta, Georgia 30332-0828, USA. e-mail: mikhail.belenkii@gtri.gatech.edu

Tall tales promise better computer storage

A University of Wisconsin-Madison engineer has fixed on an unusual parameter in the quest to make functional micro-machines. Rather than just smaller, he makes them taller.

Engineer Henry Guckel has been pursuing the 'taller is better' premise for years in his applied micro-electronics laboratory. Now his techniques, that also rely on deep X-ray lithography techniques on metal rather than the standard silicon, are beginning to make their way into commercial products.

Guckel says that the added dimension of his micro-machines dramatically increase their power storage capability and make them more functional for devices such as actuators and sensors. Most parts are less than 75μm wide – but they can be as tall as 1000μm.

The deep X-ray lithography process also offers some advantages over making parts with silicon. By using metals such as nickel, copper and iron alloys in micro-motors, the machines are driven by magnetic rather than electrical fields. Metal can also be layered faster than silicon, so parts can be made taller within industry's normal time demands.

One strong market for metal micro-motors is actuators, the mechanical devices in systems that transmit energy to control precise functions such as computer memory.

Micro-motors could control the movement of magnetic recording heads more precisely, expanding the amount of information stored on computer disks. Current technology can read information on computer disks within 2μm. But a micro-mechanical actuator could read the disk within one-tenth of a micrometre – a factor of ten improvement.

Electronic draw-bar links lorry road train

Daimler-Benz researchers are to trial later this year, an 'electronic draw-bar' that will allow individual commercial vehicles to be joined up into a train-like convoy on the roads. The aim is to enable lorries be able to drive in a densely packed formation, on the basis that cars will be able to over-take briskly, on more efficiently-utilised roads. In addition, the lower air resistance acting on the individual lorries will cut their fuel consumption and so also reduce the burden on the environment.

The system makes use of a video camera mounted on a trailing vehicle that homes in on a special pattern marked on the rear of the vehicle in front. An on-board computer determines its distance and direction of travel, along with the relative speed between the two lorries, and the trailing vehicle is precisely controlled on the basis of this information. Fully automatic acceleration and braking procedures are calculated on the basis of these video signals. Infrared light can also be used as the basis of the system – which would enhance the reliability of the image analysis – but this would make the process more expensive.

So what happens if the leading vehicle should suddenly break away or otherwise finds itself in a hazardous situation? Daimler-Benz explains that the leading lorry's driving parameters are continuously transmitted by radio to the trailing vehicle, together with signals for the image processing system. Each vehicle also has its own track recognition system. In theory that means that the following vehicle should be able to react to unexpected situations with practically no delay.

A pilot project currently being planned will involve initial practical tests on two electronically-coupled vehicles. Researchers hope to evaluate how the human driver comes to terms with this type of technology. At the same time, the scientists – in collaboration with the technical inspection association TÜV Rheinland – are drawing up safety guidelines and investigating the legal implications of a market introduction of the electronic draw-bar.

Video cameras, mounted on trailing vehicles to track lorries in front, are at the heart of Daimler-Benz's electronic draw bar.

New angle on sensor design

Researchers have developed a prototype magnetic sensor that offers a wear-free contactless way to measure 360° angle, using a much less complicated method compared to previous devices. Simplicity and cheapness of the c-mos microsystem could make it an attractive option for angular positioning control systems in a variety of automotive and industrial applications.

Traditionally, angle detection systems, such as potentiometers, have been based on detecting the position of an electrode touching a resistive layer. But mechanical wear can degrade performance of these devices.

Alternative non-contact systems are based on magnetoresistive sensors that rely on the dependence of electrical resistance on an applied magnetic field – which can be non-linear – or Hall sensors. Hall sensors give a linear response, but two are needed, mounted orthogonally to provide the 360° measurement.

But the new integrated system developed by Andrea Häberli and colleagues ("Two-dimensional magnetic microsensor with on-chip signal processing for contactless angle measurement," *IEEE Journal of solid-state circuits*, Vol 31, No 12,) consists of a two-dimensional magnetic microsensor, off-set compensation and signal conditioning circuitry in a single chip. Angular measurement is accomplished with the help of a permanent magnet. The sensing element is a lateral bipolar magnetotransistor (lmt) integrated into a cmos chip and.

In contrast to Hall plates, only a single device is required for two dimensional sensing of the applied magnetic field, as the lmt is sensitive to fields parallel to the chip plane.

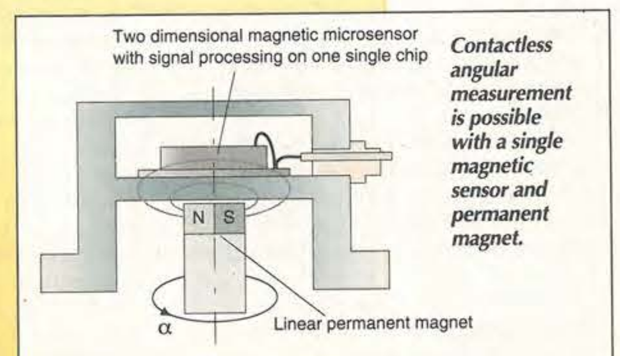
Simplicity of the design gives a number of advantages, and allows the angular position of the permanent magnet to be calculated from the ratio of the two field components detected by the magnetic sensor.

One of the main benefits is that the sensing element has a very small area of 30x30μm^2 allowing the field distribution to be considered as homogeneous, even for small permanent magnets. So angle sensing is accurate and the design allows compact integration of the sensor magnet system.

The galvanomagnetic effect in silicon, on which the operating principle of the lmt is based, also shows a linear behaviour, enabling unrestricted angle measurement to be carried out over the full 360° range.

So far, the system has achieved a 1° angular resolution with 9mW power consumption and permanent magnet of 100mT.

More information contact: A Häberli, Physical Electronics Laboratory, ETH Zurich, ETH Hönggerberg, 8093 Zurich, Switzerland.



Contactless angular measurement is possible with a single magnetic sensor and permanent magnet.

Unravelled electron unzips theories

Long accepted theories on the 'structure' of the electron are being hastily rewritten following work undertaken by Physicists at Purdue University. Science and engineering students have learned for years that the electron has a constant electronic strength. But they not be the case, according to David Koltick, professor of physics at Purdue.

Koltick's research has shown that the electromagnetic force from the electron – or its electronic strength – may increase toward the particle's central core.

According to his data, surrounding the electron's core is a fuzzy "cloud" of virtual particles, which wink in and out of existence in pairs. One particle in the pair is positively charged, the other negatively charged.

The cloud is polarised, and the strong negative charge at the core pushes the negatively charged particle in a pair slightly farther away from the core than the positively charged particle. Polarisation is strongest toward the centre of the cloud.

The polarised pairs essentially cancel each other out so that they do not add any net electric charge to the electron, Koltick says. But the cloud plays a key role in how we perceive the electromagnetic force from the electron.

To obtain their results, Koltick and more than 50 colleagues collided very-high-speed beams of particles at a facility in the Japanese Laboratory for High Energy Physics.

"As we probe into the cloud, getting closer and closer to the core charge, we see less of the shielding effect and more of the core. This means that the electromagnetic force from the electron as a whole is not constant, but rather gets stronger as we go through the cloud and get closer to the core," Koltick says.

Koltick and his colleagues also determined that the strong nuclear force – the glue that holds together elementary particles such as protons – gets weaker closer to the core charge. Other researchers also have seen this effect in the strong force.

"Because the electromagnetic charge is in effect becoming stronger as we get closer and the strong force is getting weaker, there is a possibility that these two forces may at some energy be equal," adds Koltick.

"Many physicists have speculated that when and if this is determined, an entirely new and unique physics may be discovered."

David Koltick, Purdue University, West Lafayette, Indiana, USA. e-mail: koltick@physics.purdue.edu

We can no longer assume an electron is a simple point charge. Instead, the particle may be thought of as being at the centre of a fuzzy cloud, as in this artist's impression, where the electromagnetic force increases toward a central core. The cloud consists of virtual particles, which wink in and out of existence in pairs – one particle positively charged (blue), the other negatively charged (yellow). (Graphic courtesy Dennis Harp, Purdue Physics Department.)

Real life in real time

Three-dimensional medical ultrasound imaging technology being pioneered at Duke University could make current ultrasound scanning techniques obsolete.The process uses a parallel computing to analyse a myriad of reflected sound waves, creating images so quickly that clinicians can view a whole human heart – even as it is beating. Doctors can also electronically 'dissect' the image to remove selected slices of medical interest and display them on a computer screen.

Development of the technique has been led by Olaf von Ramm, a Duke professor of biomedical engineering who also spearheaded real-time, 2d ultrasound technology now used in hospitals around the world.

Ultrasound has a number of advantages. While the pictures are not always as sharp as in other body imaging technologies, no X-rays are used – unlike a computerised axial tomography (CAT) scan, and a powerful magnetic field is not needed as in magnetic resonance imaging (MRI).

Unfortunately, 2d ultrasound has a big limitation: it can scan only a small thickness of the body at one time, while all other imaging technologies require seconds to build up an image of the heart. Since the heart moves, that makes the image blurred.

But the new 3d ultrasound technology being developed at Duke uses an advanced phased-array concept known as Eploso-Scan to take a leap forward on all fronts.

In operation, hundreds of ceramic grain-sized crystals located on the scanning wand emit high frequency sound pulses so that an entire volume of space is swept with sound simultaneously. Hundreds of other crystals then receive the returning echoes, which are converted and processed into digital pictures. Capturing an image of a beating heart or a moving foetus in real time, has been made possible by processing each signal at the same time using massively parallel processing.

As with all diagnostic ultrasound devices, operators first apply a special jelly to provide the electronic wand to get a good acoustical contact with the bare skin. The wand is then moved over a patient's chest or abdomen until the internal feature of interest appears on the viewing screen.

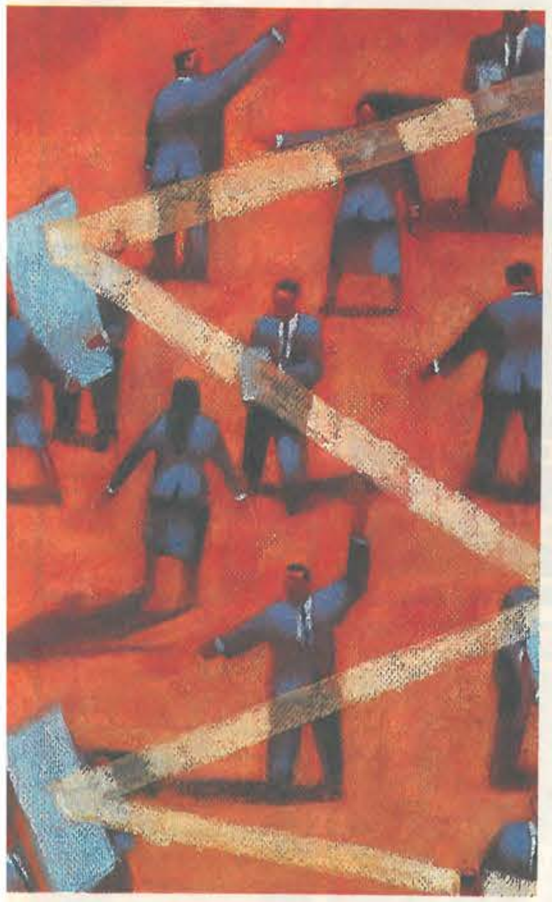
By using a touch pad, doctors can call up views of as many as 16 different slices of the heart or another organ at once. Slices can be at different angles. They can be made to be thicker or thinner.

In addition to viewing them instantly, doctors can also store all the images for later follow up analysis. ■

Contact: Olaf von Ramm, Duke University, Durham, North Carolina, USA

Light-source *for* calibration

Peter Turner describes a PIC-controlled unit for generating a uniform source of illumination with accurately controlled intensity. Peter designed it to measure charge-coupled device linearity and non-uniformity.



The light source described here is one of a pair of modules* we designed for calibrating and characterising a real-time ccd X-ray imaging system being developed in York University's Department of Electronics. Outlined in **Fig. 1**, the light source has been used to accurately asses linearity and device non-uniformity of the charge-coupled sensors incorporated in the imaging system.

Light is provided by a ring of ten ultra-bright leds capable of developing 6500mcd each at a peak current of 50mA. These are Hewlett Packard *HLMT-CL00* types.

A uniform light field was generated at the ccd by two diffusers within the optical path. These were made from drafting film mounted on drainpipe couplers. The sensor is a Texas Instruments *TSL250* light-to-voltage converter although we intend to change this for a Burr-Brown *OPT209* which has a guaranteed nonlinearity of 0.01% of full scale. This change is easily effected electronically – as will become evident.

The sensor is mounted close to the ccd and provides the feedback signal to accurately control the intensity of the leds. We did not require the unit to provide specified levels of light intensity – only very accurate relative values. As a result, there was no need to calibrate the unit as you would normally expect.

Nor was absolute long term stability a part of the specification. We envisaged that the unit would only need to be stable and accurate over several tens of minutes. The voltage reference for the feedback circuit which is used to set the light intensity was required to be accurate to 0.01%.

Further requirements for the unit were the

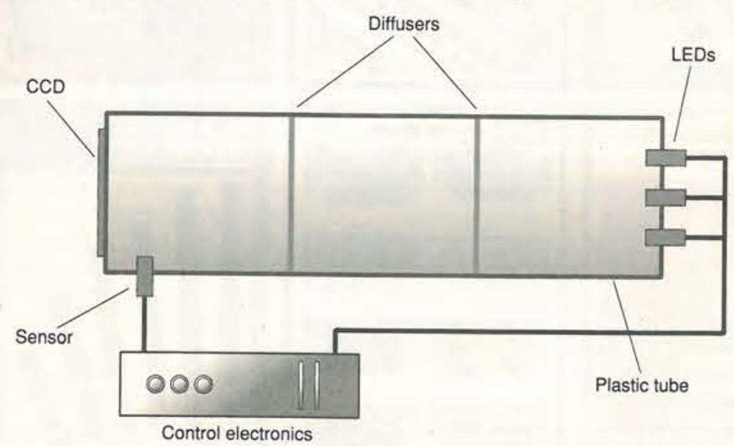


Fig. 1. Calibrating light source showing ultra-bright leds, optical path and position of the TSL250 light-to-voltage converter.

*The second module – an arbitrary waveform generator – is to be described in a subsequent article.

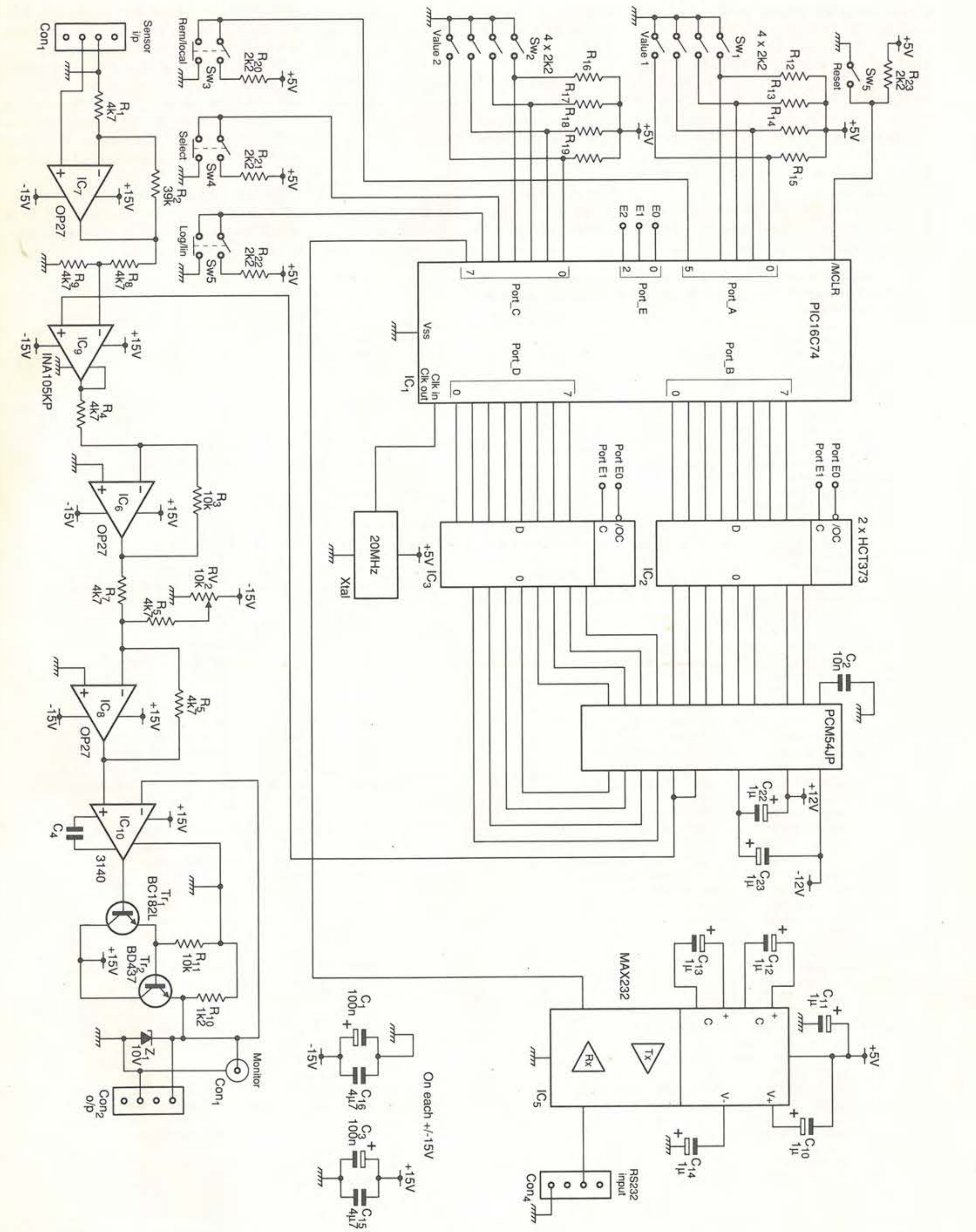


Fig. 2. Full circuit diagram of the calibrating light source, which uses a PIC16C74 to provide accurate relative light intensities.

need to switch between two pre-set light levels easily and to have both logarithmic and linear scales to chose from. The final requirement was for the option to remotely control the light level via an external pc. WeI decided that a 16 bit digital to analogue converter should be used to set the reference level and that the following front panel controls would be useful.

- Two 16-way switches to select one of 16 pre-set light levels.
- Switching to select between two light levels.
- Switching to select between logarithmic and linear scales.
- Switching to select between remote and local.

The idea behind having a logarithmic scale as well as linear was to give the ability to make lots of measurements at the lower light levels within the 16 values as well as covering the full range.

Circuit details

At the heart of the design, **Fig. 2**, is a programmable PIC microcontroller, namely a *PIC16C74*. We chose this uv-erasable PIC because it offered a wide variety of functions which seemed to fit the design perfectly.

This was my first attempt at using a member of the PIC family. The device proved easy to use – despite including relatively complex functions such as RS232 communications driven by interrupts. The limited instruction set with only 35 single instructions to learn is the main reason for this ease of use.

The other key component is the Burr-Brown *PCM54* 16-bit digital-to-analogue converter which has good linearity at reasonable cost.

List 1. PIC code for the light source writes the the d-to-a converter, setting the various light intensities depending on the positions of the switches shown on the main circuit diagram.

```
;Pic code LIGHT SOURCE pt 080296
        list p=16c74
        include "c:\pic\asmicro\p16cxx.inc"
;define various locations
        SWITCH1 equ 0x21
        SWITCH2 equ 0x22
        CURRENT equ 0x23
        W-TEMP  equ 0x30
        STATUS-TEMP equ 0x31
        CURR equ 0x32
;interrupt vector
        org 0004h
        goto receive
;reset vector
        org 0000h
        goto START
;main program
        org 0005h
START   CLRF STATUS
        BSF STATUS, RP0
        CLRF INTCON
;configure ports A and B and E and D and C
        BCF STATUS, RP0  ;SELECT BANK 0
        CLRF PORTB       ;CLEAR  PORTB
        CLRF PORTE       ;CLEAR  PORTE
        CLRF PORTA       ;CLEAR  PORTA
        CLRF PORTD       ;CLEAR  PORTD
        CLRF PORTC       ;CLEAR  PORTC
        BSF STATUS, RP0  ;SELECT BANK 1
        MOVLW 0x00       ;PUT ZERO IN W
        MOVWF TRISB      ;SET PORTB AS O/P
        MOVWF TRISE      ;SET PORTE AS O/P
        MOVWF TRISD      ;SET PORTD AS O/P
        MOVLW 0xFF       ;PUT FF IN W
        MOVWF ADCON1     ;A/D LINES DIGITAL
        MOVWF TRISA      ;SET PORTA AS I/P
        MOVWF TRISC      ;SET PORTC AS I/P
;decide whether to go local or remote
        BCF STATUS, RP0  ;SELECT BANK 0
        BTFSC PORTA, 4   ;TEST REM/LOCAL
        GOTO rem
        GOTO loc
;interrupt service routine
receive     MOVWF W-TEMP
        SWAPF STATUS, W
        BCF STATUS, RP0  ;SELECT BANK0
        MOVWF STATUS-TEMP ;TEMP STORE
        BCF RCSTA, CREN
        BSF RCSTA, CREN  ;RESET OERR BIT
        MOVF RCSTA, 0    ;
        MOVF RCREG,0     ;RECEIVE BYTE IN W
        BTFSC CURR,0
        GOTO hibyte
        GOTO lobyte
hibyte  MOVWF PORTB      ;WRITE HI BYTE
        MOVLW 0x02
        MOVWF PORTE      ;ENABLE LATCH
        NOP
        NOP
        NOP
        NOP
        NOP
        CLRW
        MOVWF PORTE      ;DISABLE LATCH
        BCF CURR,0
        GOTO done
lobyte  MOVWF PORTD      ;WRITE LO BYTE
        MOVLW 0x04
        MOVWF PORTE      ;ENABLE LATCH
        NOP
        NOP
        NOP
        NOP
        NOP
        CLRW
        MOVWF PORTE      ;DISABLE LATCH
        BSF CURR,0
done    SWAPF STATUS-TEMP, W
        MOVWF STATUS
        SWAPF W-TEMP, F
        SWAPF W-TEMP, W  ;GET W/STATUS
        RETFIE
rem     BSF STATUS, RP0  ;SELECT BANK 1
        BSF TXSTA, BRGH  ;HIGH SPEED BAUD
        MOVLW 0x81
        MOVWF SPBRG      ;BAUD RATE TO 9600
        BCF TXSTA, SYNC
        BCF STATUS, RP0  ;SELECT BANK 0
        BSF CURR,0       ;SET HI/LO IDENT BIT
        BSF RCSTA, SPEN  ;ENBLE SERIAL PORT
        BSF INTCON, GIE  ;INTERRUPT ENABLE
        BSF INTCON, PEIE ;PERIPHERAL INTS
        BSF STATUS, RP0  ;SELECT BANK 1
        BSF PIE1, RCIE   ;EN ASYNC REC INTS
        BCF STATUS, RP0  ;SELECT BANK 0
        BSF RCSTA, CREN  ;ENABLE RECEPTION
;loop testing rem/local switch waiting for interrupt
test    BTFSC PORTA, 4   ;TEST REM/LOCAL
        GOTO test
reset   BCF INTCON, GIE  ;DISABLE INTS
        BTFSC INTCON, GIE ;DID IT WORK
        GOTO reset       ;NO TRY AGAIN
;end of remote routine
loc     MOVF PORTC,0     ;READ PORTC IN W
        MOVWF SWITCH1    ;STORE  IN SWITCH1
        MOVF PORTA,0     ;READ PORTA IN W
        MOVWF SWITCH2    ;STORE IN SWITCH2
        CALL display     ;DISPLAY VALUE
swtest  MOVF PORTC,0     ;READ PORTC TO W
        XORWF SWITCH1,0 ;BITS CHANGED?
        BTFSS STATUS,2
        GOTO testrem
        MOVF PORTA,0     ;READ PORTA TO W
        XORWF SWITCH2,0;BITS CHANGED
        BTFSS STATUS,2
        GOTO testrem
        GOTO swtest      ;TEST SWITCHES
testrem BTFSC PORTA, 4   ;TEST REM/LOCAL
        GOTO rem
        CALL display     ;DISPLAY VALUE
        GOTO swtest
display MOVF PORTC,0     ;READ PORTC TO W
        MOVWF SWITCH1    ;STORE IN SWITCH1
        MOVF PORTA,0     ;READ PORTA TO W
        MOVWF SWITCH2    ;STORE  IN SWITCH2
        CLRW
        MOVWF CURRENT    ;RESET OFFSET
        BTFSC SWITCH1,4  ;TEST VALUE SW
        GOTO sw2
```


Starting from the top left of the circuit, there is the reset push-button switch connected to the PIC's /MCLR input. Below that are the two 16-position thumb-wheel switches connected to the four least-significant bits of ports A and C of the PIC.

Shown below the PIC are the remote/local, select and log/lin switches. These are toggle switches To the right of the PIC are two eight-bit buffers whose output enable and clocking functions are controlled by port E on the PIC. These buffers drive the 16 bit data bus for the *PCM54* digital-to-analogue converter.

Output of the *PCM54* connects to the non-inverting input of IC_9. This is a differential amplifier which compares scaled input from the sensor via IC_7 with the output from the *PCM54*.

Resistors R_8 and R_9 are chosen for the particular sensor to be used, the ones shown being suitable for the *TSL250* as described above. Op-amps $IC_{6,8}$ amplify and add offset to the output of IC_9. This offset is trimmed so that the leds are biased into a just-on condition at zero output from the *PCM54*, which is about 1.5V.

Current drive for the ten leds of up to 500mA is provided by the final output stage comprising IC_{10} and $Tr_{1,2}$ etc. The only other components are those associated with the RS232 communications – a Maxim *MAX232* and associated capacitors. This IC takes a simple 5V supply, accepts standard RS232 levels at ±15V and produces ttl logic levels required by the PIC.

Mechanical considerations

We constructed the optical path using simple 70mm drainpipe. Black drainpipe was chosen

```
        GOTO sw1
sw2     MOVLW 0F
        ANDWF SWITCH2,0  ;MASK  TOP 4 BITS
        ADDWF CURRENT,1 ;ADD  OFFSET
        GOTO disp1
sw1     MOVLW 0F
        ANDWF SWITCH1,0
        ADDWF CURRENT,1 ;ADD  OFFSET
disp1   BTFSC SWITCH1,5      ;TEST LIN/LOG
        GOTO log
        GOTO lin
log     MOVF CURRENT,0       ;PUT OFFSET IN W
        CALL table1
        MOVWF PORTB          ;WRITE HI BYTE
        MOVLW 0x02
        MOVWF PORTE          ;ENABLE LATCH
        NOP
        NOP
        NOP
        NOP
        NOP
        CLRW
        MOVWF PORTE          ;DISABLE LATCH
        MOVF CURRENT,0       ;PUT OFFSET IN W
        CALL table1h
        MOVWF PORTD          ;WRITE LO BYTE
        MOVLW 0x04
        MOVWF PORTE          ;ENABLE LATCH
        NOP
        NOP
        NOP
        NOP
        NOP
        CLRW
        MOVWF PORTE      ;DISABLE LATCH
        RETURN
lin     MOVF CURRENT,0   ;PUT OFFSET IN W
        CALL table
        MOVWF PORTB      ;WRITE HI BYTE
        MOVLW 0x02
        MOVWF PORTE      ;ENABLE LATCH
        NOP
        NOP
        NOP
        NOP
        NOP
        CLRW
        MOVWF PORTE      ;DISABLE LATCH
        MOVF CURRENT,0   ;PUT OFFSET IN W
        CALL tableh
        MOVWF PORTD      ;WRITE LO BYTE
        MOVLW 0x04
        MOVWF PORTE      ;ENABLE LATCH
        NOP
        NOP
        NOP
        NOP
        NOP
        CLRW
        MOVWF PORTE      ;DISABLE LATCH
        RETURN
table   ADDWF PCL        ;W=OFFSET
        RETLW 0x00       ;FIRST VALUE
        RETLW 0x10
        RETLW 0x20
        RETLW 0x30
        RETLW 0x40
        RETLW 0x50
        RETLW 0x60
        RETLW 0x70
        RETLW 0x80
        RETLW 0x90
        RETLW 0xA0
        RETLW 0xB0
        RETLW 0xC0
        RETLW 0xD0
        RETLW 0xE0
        RETLW 0xF0
tableh  ADDWF PCL        ;W=OFFSET
        RETLW 0x00
        RETLW 0x00
        RETLW 0x00
        RETLW 0x00
        RETLW 0x00
        RETLW 0x00
        RETLW 0x00
        RETLW 0x00
        RETLW 0x00
        RETLW 0x00
        RETLW 0x00
        RETLW 0x00
        RETLW 0x00
        RETLW 0x00
        RETLW 0x00
        RETLW 0x00       ;LAST LIN VALUE
table1  ADDWF PCL        ;OFFSET IN W
        RETLW 0x00       ;FIRST LOG VALUE
        RETLW 0x7F
        RETLW 0xBF
        RETLW 0xDF
        RETLW 0xEF
        RETLW 0xF7
        RETLW 0xFB
        RETLW 0xFD
        RETLW 0xFE
        RETLW 0xFF
        RETLW 0xFF
        RETLW 0xFF
        RETLW 0xFF
        RETLW 0xFF
        RETLW 0xFF
        RETLW 0xFF
table1h ADDWF PCL        ;W=OFFSET
        RETLW 0x00
        RETLW 0xFF
        RETLW 0xFF
        RETLW 0xFF
        RETLW 0xFF
        RETLW 0xFF
        RETLW 0xFF
        RETLW 0xFF
        RETLW 0xFF
        RETLW 0x7F
        RETLW 0xBF
        RETLW 0xDF
        RETLW 0xEF
        RETLW 0xF7
        RETLW 0xFB
        RETLW 0xFD       ;LAST LOG VALUE
end
```

to minimise internal reflections. The diffusers were made from drafting film attached to drainpipe couplers which pushed into the actual drainpipe.

Experimentation proved that two diffusers were sufficient to produce a good uniform light intensity with acceptable attenuation. The distance between the light source and the plane of the sensor/ccd was approximately 450mm.

Microcontroller details

The *PIC16C74* is a low-cost, c-mos microcontroller. It has a 14-bit instruction bus and an 8-bit data bus. It also has a two-stage instruction pipeline. This allows all instructions to execute in one cycle, except for program branches which execute in two.

With a clock frequency of up to 20MHz it is clear why these devices perform so well. The reduced instruction set of only 35 instructions and the large number of internal registers make the devices easy to use and easy to program.

The device has 4K 14-bit words of program memory, 192 bytes of ram and 33 i/o pins. Peripheral features contained on the chip include three timer/counters, two serial ports, a high-speed eight-channel a-to-d converter etc. Also provided are multiple internal and external interrupt sources.

Writing the software

Software for the PIC is shown in **List 1**. After the initialising statements which set up the registers and define the reset and interrupt vectors, the program reads the remote/local switch and decides which of the two main routines to jump to.

The local routine starts at the label 'loc'. After setting the output at the level defined by the current switch configuration, the program enters the loop at 'swtest' and cycles until it decides a switch has changed state. If the test finds that the remote/local switch has not changed it calls the 'display' subroutine which tests the remaining switches. It then jumps to either the 'log' or 'lin' sections. These in turn select a value from the relevant look-up table and write to the digital-to-analogue converter via the output ports and buffers.

If the remote/local switch is set to remote at any point the program jumps to the 'rem' section where it sets up the asynchronous serial port and the associated interrupt. The program then loops around the 'test' loop. It does so until the remote/local button is changed to local or an interrupt occurs, indicating that the serial port register has received a value. On receipt of the interrupt control passes to the interrupt-service routine at 'receive'.

The interrupt-service routine starts by temporarily storing the W and STATUS registers so that they can be restored to their original values at the end of the routine. The next step is to read the received value and decide whether it is the high-order or low-order byte by testing the CURR register bit 0.

The 'hibyte' and 'lobyte' routines output the value on the relevant port. They also set or reset the CURR register bit 0 so that the program knows which was the last bit displayed when the next interrupt occurs. Finally, when the byte has been output, the W and STATUS registers are restored and the routine returns control to the main program and re-enables the interrupts.

Remote operation

In order to complete the remote operation of the unit a simple program to talk to the PIC was written on a pc in Turbo Pascal, **List 2**. It accepts a four digit hexadecimal value from the keyboard and writes it in two parts – high byte and low byte – to the serial port.

Operation should be fairly self explanatory. The unit 'useful' to which the program refers contains the two procedures for operating on the four-digit hexadecimal value – 'change' and 'stringtochar'. The procedure 'change' converts a valid hexadecimal character into the equivalent decimal integer. The procedure 'stringtochar' converts a unit length string to the equivalent character value.

In summary

The article describes a general purpose test unit for producing very accurate relative light intensities for the testing and characterisation of ccds used in a research project at the University of York, Department of Electronics. The design utilises some of the many features available on the *PIC16C74* microcontroller. It is the author's opinion that use of these and similar devices is becoming virtually essential in the design of all but very simple pieces of digital electronic equipment due to the many advantages over the use of conventional logic which they offer. ■

List 2. Turbo Pascal listing to drive the light source remotely via a pc's serial port.

```
program serial;
uses dos,crt,useful;
CONST
  hexdigits:set of '0'..'F' = "'0'..'9','A'..'F'";
var
  lst:text;
  value:string;
  digit:array"1..4" of string"1";
  i:integer;
  okay:boolean;
  number:integer;
  digit1:char;
  digitchar:array"1..4" of char;
procedure bung;
var
  x:integer;
  y:longint;
  ident:array"1..4" of integer;
begin
  assign(lst,'AUX');
  rewrite(lst);
  for i:=1 to 2 do
    begin
      change(digitchar"i",x);
      ident"i":=x
    end;
  x:=(ident"1"*16) + ident"2";
  write(lst,chr(x));
  for i:=3 to 4 do
    begin
      change(digitchar"i",x);
      ident"i":=x
    end;
  x:=(ident"3"*16) + ident"4";
  write(lst,chr(x));
  close(lst)
end;
begin
  repeat
    okay:=true;
    write('Enter 4 digit HEX value (use CAPS): ');
    readln(value);
    for i:=1 to 4 do
      begin
        digit"i":=copy(value,i,1);
        stringtochar(digit"i",digit1);
        digitchar"i":=digit1;
        if digitchar"i" in hexdigits
          then
          else
           begin
            okay:=false;
            writeln;
            writeln('Not a HEX value.');
            writeln;
            writeln
           end
      end;
   if okay=true
     then
       bung;
  until value='EXIT';
end.
```


BIG surprises... ...small packages

At one end of the surface-mount spectrum, complex digital ICs are becoming so densely pinned that they make prototyping almost impossible. At the other, it is now easy to obtain one logic function or op-amp in a single, minute sm package. While reducing product size, these tiny devices can simplify implementation, improve performance, and even open up new application areas, as Ian Hickman demonstrates.

The surface-mount revolution has been under way for years now, with most products using surface-mount passives. Fixed resistors are migrating from the 1208 size, with dimensions of 0.12 by 0.08in, to 0805, 0604 or even 0402.

Trimmer resistors, with overall dimensions of less than 4mm square are supplied by several manufacturers, including Bourns and Citec. Capacitors are available in a similar range of sizes to fixed resistors, though the larger values such as tantalum electrolytics tend to still be in 1208 or larger format, for obvious reasons.

Trimmer capacitors are available with a footprint of less than 4mm square, from various manufacturers, including Murata. Surface-mount inductors are available in the various formats. Ingenious surface-mount carriers accommodate ferrite toroid cored inductors where higher values of current-carrying capacity or of inductance are necessary – such as in switchmode power supplies – and where the extra height can be accommodated.

But surface-mount passives have been around so long that there is not much new to say about them. So this article concentrates on active devices, and mainly on integrated circuits, which is where the action currently is.

Recent trends

More recently, there has been renewed interest in really tiny devices with eight, five or even just three pins. This format has long been favoured by rf engineers, for uhf and microwave transistors, the consequent reduction in overall size and lead lengths contributing to minimal package parasitics.

Now, the advantages of really tiny devices, which are many, are becoming available also to analogue and digital designers, and this article looks at some of these devices. **Table 1** lists typical examples, giving the package designation – which varies somewhat from manufacturer to manufacturer – the number of pins, a typical example of a device in that package, and its manufacturer, and the maximum overall size of the 'footprint' or board area occupied by a device in that package style. This again varies slightly from manufacturer to manufacturer.

With devices in such small packages, getting the heat away can be a problem. With many of these ICs, though, the difficulty is alleviated due to two aspects. Firstly, many devices such as op-amps, comparators and digital ICs now work from a single supply of 3V or even lower, as against the 5V, ±5V or even ±15V required by earlier generations. Secondly, with improved design techniques, high-speed wide frequency range devices can now be designed to use less current than formerly.

Nevertheless, thermal considerations still loom large in many cases, when applying these tiny devices. This is discussed further in the following sections, which deal with various classes of small outline devices.

Discrete active devices

With discretes such as diodes, in many cases maximum dissipation is a pressing consideration, and package styles and sizes reflect this. Thus the *UDZ* series zeners from Rohm, in the SOD-323 package, **Fig. 1a)**, are rated at 200mW. But *RLZ* series devices, also from Rohm, in the slightly larger LL34 package, Fig. **1b)**, dissipate 500mW, while the *PTZ* series in the even larger PSM package, Fig. **1c)**, is rated at 1W.

With active devices also, special packages are used to cope with the device dissipation. International Rectifier's *IRFD11x* series mosfets for example are mounted in a four pin 0.3in DIL package, **Fig 2a)**. Pins 3 and 4 are commoned and provide not only the drain connection, but also conduct heat to through-hole pads on the pcb, which help dissipate the heat.

These surface-mount devices have a P_{drain} rating of 1.2W. This is actually 20% more than the rating of the *VN10KM*, which is housed in a TO237 package, see Fig. **2b)**. The TO237 is like a TO92 package, but it has a metal tab, connected to the drain, projecting from the top.

The SOT89 is an even smaller package, Fig. **2c)**, measuring just 2.5 by 4mm, excluding leadouts. Nevertheless, the Rohm *BCX53* is rated at 500mW, or 1W when mounted on a suitable ceramic pcb. The wider collector lead, on the opposite side of the package from the base and emitter leads, bends back under the body of the device, providing a large heat tranfser area.

The SOT223 package, not shown, provides a power dissipation of up to about 1.5W at 25°C. The TO252 'D-pak' shown in Fig. **2d)**, housing for example an *IFRF024* 60V, 15A mosfet with 60A pulsed I_d rating, does even better. The device dissipates watts, provided

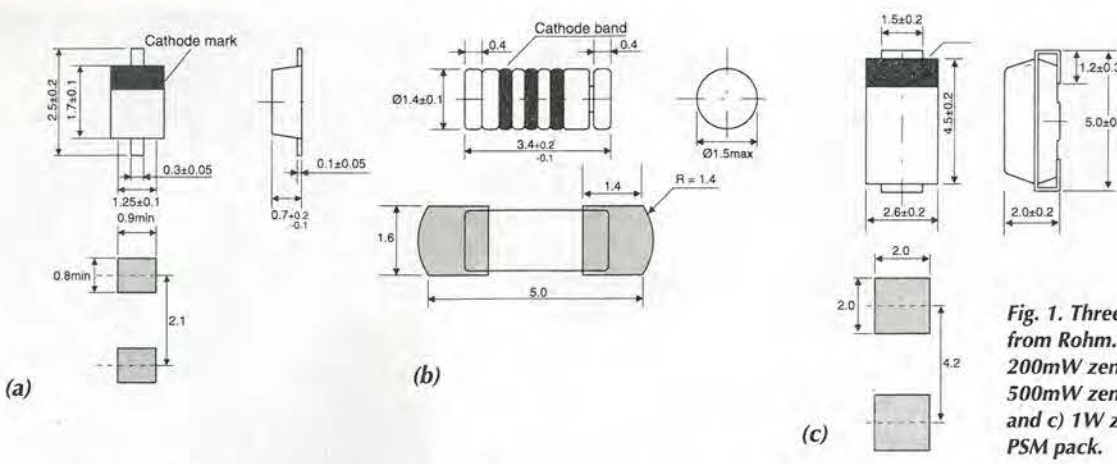


Fig. 1. Three surface-mount diodes from Rohm. Left to right are a) 200mW zener in SOD-23 package, b) 500mW zener in the larger LL34 pack and c) 1W zener in the even larger PSM pack.

that you can keep its case temperature down to 25°C.

For small signal amplifiers, size is less important and transistors are available in packages smaller than SOT23 (SMT3), **Fig. 3a)**. The UMT3 (Ultramold, SOT323) package of Figure **3b)** has a footprint of 2.2mm by 2.2mm overall, including leads, while the EMT3, Fig. **3c)** occupies just under 1.8 by 1.8mm overall, these being the maximum dimensions.

With such very small devices, traditional laboratory prototyping becomes very difficult, not to say tedious.

Analogue ICs

With digital ICs, the tend is to higher and higher levels of functional integration, with an inevitable accompanying inflation in the number of pins per package. In the analogue world however, general purpose functions, such as op-amp, comparator, buffer and voltage reference tend to dominate. The result is that while digital ICs tend to get bigger – or at least not much smaller, due to all those pins – analogue functions are appearing in smaller and smaller packages.

The exception is d-to-a and a-to-d converters with parallel data buses. But these ICs tend to bridge the analogue/digital divide anyway. And even here, devices in tiny eight-pin packages are readily available, thanks to the economy in pin numbers afforded by using serial data input/output schemes rather than bus structures.

While single transistors can be mounted in packages smaller than SOT-23, this is more problematical for the larger silicon die of ICs. So for the most part, the three pin version of SOT-23 is the smallest package used for ICs. An example is the *AD1580* 1.2V micropower precision shunt voltage reference, from Analog Devices.

To the user, the *1580* appears simply as a 1.2V zener diode. But the dynamic output impedance (ac slope resistance) at 1mA is typically just 0.4Ω, resulting in a change in output voltage, over 50μA to 1mA and over –65 to +125°C, of only 500μV typical. Being a two terminal device, pin 3 has no conection, or may be connected to the negative supply.

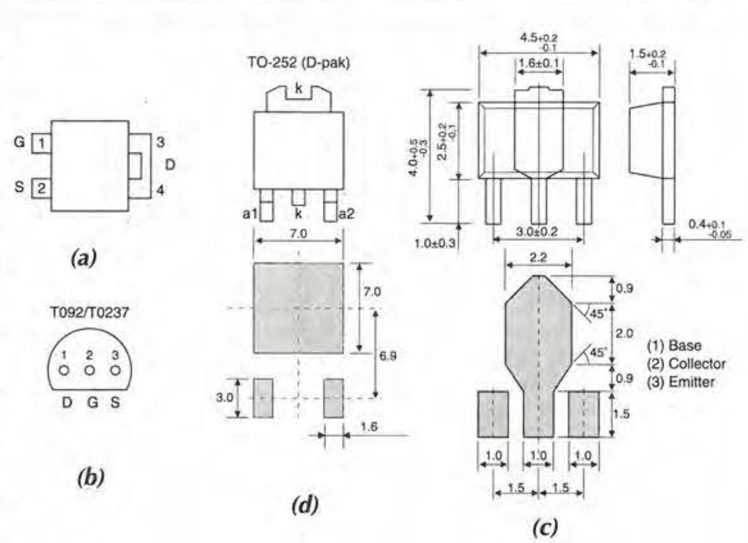
A good example of an op-amp in a small package – also available in an eight-pin DIP – is the *LMC7111*, from National Semiconductor, **Fig. 4**. The leadout arrangement of the five-pin SOT23-5 version is shown in Fig. **4a)**: note the actual size drawing alongside!

Table 1. Some representative devices in small packages, from various manufacturers. 'Small outline' is abreviated to SO.

Style	Leads	Example	Function	Manufacturer	Footprint max
SOD-323	2	1SS356	Diode, band-switch	Rohm	1.35x2.7mm
SOT23-3*	3	LM4040AIM3-5.0	Voltage ref. 5V 0.1%	Nat. Semi.	3.0x3.05mm
SOT23-5**	5	AD8531ART	Op-amp, 5V, 0.25A o/p	Analog Devices	3.0x3.1mm
SO-8	8	MAX840	–2V reg. GaAs fet Bias Generator	Maxim	5.03x6.29mm
SO-14	14	LT1491CS	Quad op-amp, 2-44V supply	Linear Tech.	6.20x8.74mm

*'TinyPak', TM. Also known as TO-236-AB
**JEDEC TO-xxxxx outline definition now due

Fig. 2a). Four pin 0.2in DIP package often used for fets and other smallpower devices. At b) is the TO237 pack is like a TO93, but with a small metal tab extending from the top, c) the SOT89 pack can typically dissipate 0.5-1W and d) TO252 package dissipates watts – provided you can keep the case temperature below 25°C!



The device is a c-mos op-amp with rail-to-rail input and output, operating from a supply voltage V_s of 2.7V upwards to an absolute maximum of 11V). With a gain/bandwidth product, or gbw, of 40kHz with a 2.7V supply, it draws a supply current I_s of around 50μA. Its bipolar stablemate, the *LMC7101*, offers a 0.6MHz gbw and 0.7V/μs slew rate in exchange for an I_s of around 800μA, also at 2.7V.

Need more speed?

Where something a little faster is needed, then in the same package, and from the same manufacturer comes the *LM7131* high-speed bipolar op-amp. This has a gbw of 70MHz, and a slew rate of 100V/μs – even when driving a capacitive load of 20pF. Total harmonic distortion at 4MHz is typically only 0.1% when driving a 150Ω load with a 3V V_s. Even with this level of performance, I_s is only 8mA.

Where blindingly fast speed is necessary, the *LM7121* voltage feedback op-amp, in the same package with the same pinout, has a 1300V/μs slew rate, for an I_s of just over 5mA typical. But note that this is the performance with dual supplies of +15 and –15V. The device works on a single V_s of down to 5V, but the performance is then more modest.

Unusually for an op-amp, this device is stable with literally any level of load capacitance, maximum peaking, up to 15dB, occuring with around 10nF. Other stablemates in the same SOT23 package and with the same pinouts, are the *LMC7211* and *LMC7221* rail-to-rail input comparators, with active and open drain outputs respectively.

Current feedback op-amps are known for their excellent ac characteristics. The *OPA658* is a wideband low power current feedback opamp from Burr-Brown, available in the SOT23-5 pin package. With a unity gain stable bandwidth of 900MHz and a 1700V/μs slew rate, it has a wide range of applications including high resolution video and signal processing, where its 0.1dB gain flatness to 135MHz is exceptional.

Where a circuit requires two op-amps, two separate devices in, say, SOT23-5 packages may be used. This provides the ultimate in layout flexibility and it may even take up less space than a dual. But the dual op-amp will usually be cheaper than two singles.

Figure 5 shows the *AD8532* dual rail-to-rail input and output c-mos op-amp from Analog Devices. Featuring an output drive capability of a quarter of an amp and a 3MHz gain:bandwidth product with a V_s of 5V, it operates from a single supply in the range 2.7 to 6V.

Figures **5a)** and **b)** compare the footprint in the TSSOP, or thin shrink small outline package, and the SO-8 package. Width over the pins is similar, but the TSSOP's pin spacing of 0.65mm, against twice this for the SO-8, results in a package length not much more than half that of the SO-8. For applications where more space is available, the device also comes in the old-fashioned 8 pin DIP package.

Figure **5c)** shows the op-amp's internal circuitry in simplified form. As common in devices with a rail-to-rail input, whether bipolar or fet, complementary input pairs in parallel are used. Likewise, for rail-to-rail outputs, common drain (collector) stages are dropped in favour of common source (emitter) stages.

Figure **5d)** shows the clean large signal pulse response, even at a Vs of just 2.7V. The device is just one of the family of *AD8531/2/4* single/dual/quad opamps, available in a wide variety of package styles.

Another dual op-amp, this time with the exceptional V_s range of 2.7V to 36V, is the *OPA2237*, from Burr-Brown. With its maximum offset voltage of 750μV and its 1.5MHz bandwidth, it is targeted at battery powered instruments, PCMCIA cards, medical instruments etc. It is available in SO-8, and also in MSOP-8, or micro small outline package, which is just half the size of the SO-8 package.

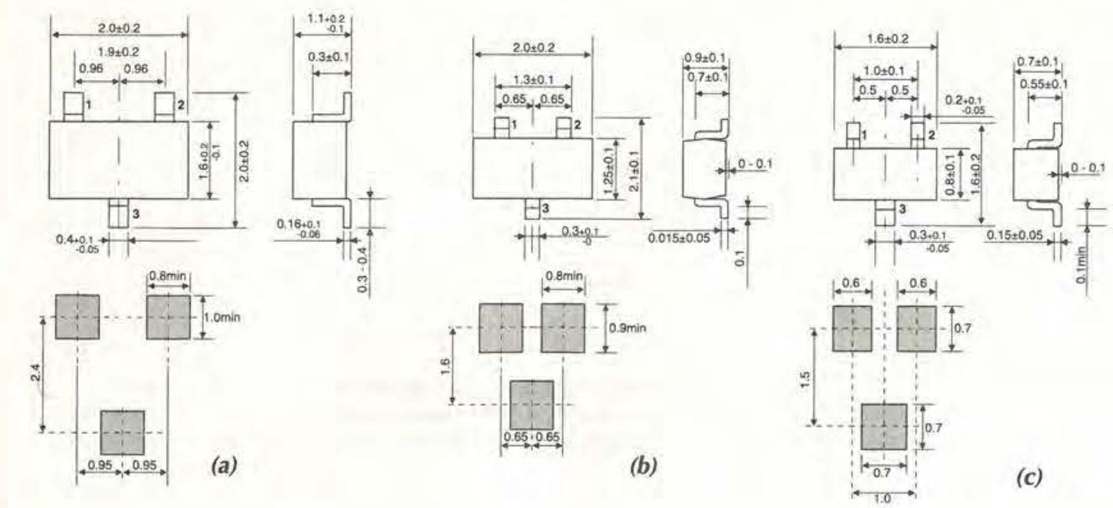
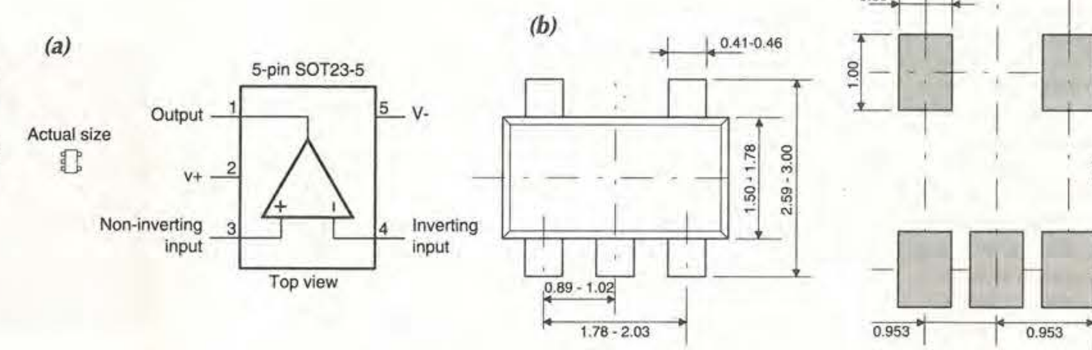


Fig. 3. Three small transistor outlines: the tiny SOT23-3 a) dwarfs the SOT323 at b), which in turn dwarfs the miniscule EMT3 at c).

Fig. 4. LMC7111 from National Semiconductor. a) pinout and actual size b) dimensions of the SOT23-5 package, and of the recomended circuit-board pads.

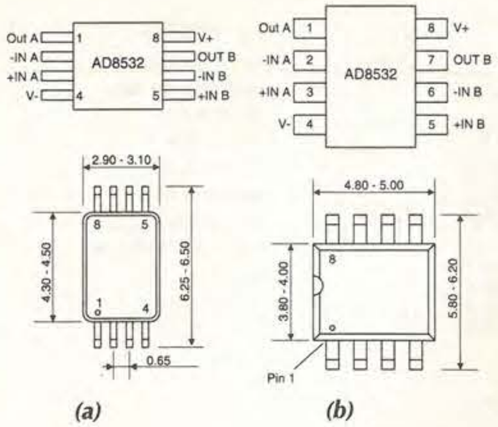


Traditional packaging options

For years, ICs came in just two widths, and a variety of lengths, all with pins on 0.1in centres. Thus 8, 14 and 16 pin dual-in-line DIL devices – whether side brazed ceramic types to military specifications, or commercial plastic moulded DIPs – came with a width between the two rows of pins of 0.3in.

For ICs with 24, 28, 40 or 68 pins however, 0.6in was the order of the day. Even so, there were exceptions, such as 0.3in 'skinny' 24-pin devices. But then, with the appearance of more and more complex ICs, more and more i/o pins were necessary. To accommodate these, square devices with pins on all four sides appeared, such as chip-carriers – both leadless and leaded, J lead devices and plastic quad flatpacks (PQFP) with various pin centre spacings, often only 0.025in or less, and up to 200 pins or more.

To minimise package size, ICs were packaged in 'pin-grid array' packaging, with several parallel rows of pins on the underside of each edge, and again up to 200 or more pins. Yet other formats are SIL/SIP (single in line/plastic) packages for memory chips and surface-mount audio frequency power amplifiers. Audio power amplifiers also appear in through-hole mounting SIPs, with alternate pins bent down at different lengths, to mount in two rows of staggered holes.



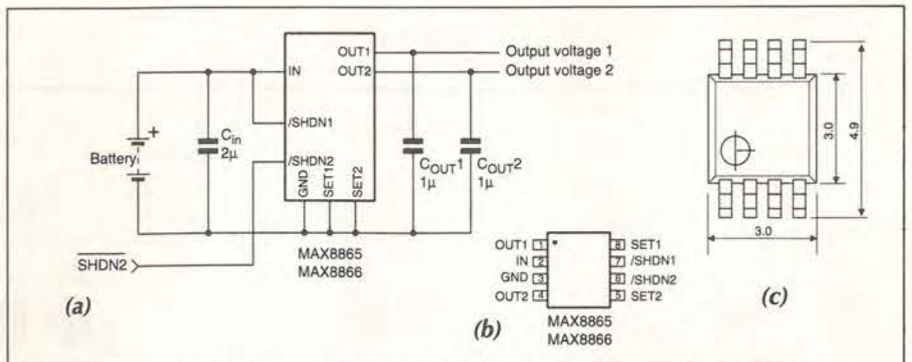


Fig. 6. MAX8865x dual low-dropout regulator a) from Maxim comes in the proprietary muMAX package, with pinout as at b). At 3mm, package length c) is similar to TSSOP, but the width across pins is 1.5mm less, which could lead to its more widespread adoption by other manufacturers.

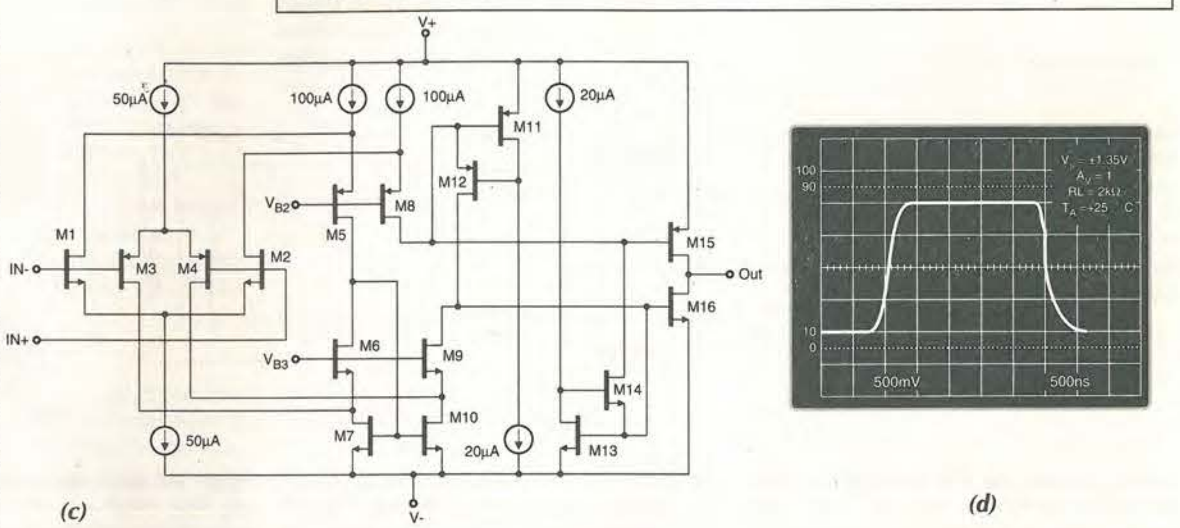


Fig. 5. AD8532 dual op-amp from Analog Devices is available in TSSOP a), SO-8 b) or 8-pin DIP. The parallelled complementary input stages and common source output stages provide rail-to-rail operation at both ends c). The 2V peak to peak response, operating on ±1.35V rails, is shown in d).

Other analogue circuits

Figure 6 shows the Maxim *MAX8865x* dual low drop-out regulator, where suffix *x* is *T*, *S* or *R*, indicating preset output voltages of 3.15, 2.84 or 2.80V respectively. Each output is capable of supplying up to 100mA, with its own individual shutdown input.

Figure **6a)** shows the device conected to supply output 1 continuously, and output 2 only when the /SHDN2 pin is high. If the SET1 or SET2 pin is connected not to ground, but to a voltage divider connected across the corresponding output, the circuit produces whatever stabilised output voltage results in the SET pin being at 1.25V. This assumes of course, that the input voltage,which must be in the range 2.5 to 5.5V, is adequate.

Internal circuitry for each output senses whether the SET pin is at a voltage below or above 60mV, and selects an internal, or the external voltage divider respectively. The pin allocation is as in Fig. **6b)**, while the package dimmensions are given in **c)**. This package is proprietary to Maxim. It is the same length as an eight-pin TSSOP, but with a narrower body, making the width over the pins rather smaller. The *MAX8866* is similar, but includes an auto-discharge function, which discharges an output to ground whenever it is deselected.

Figure 7 shows two other Maxim devices. At **a)**, are shown the *MAX4051* and *MAX4052*, these being single normally-open and normally-closed analogue switches respectively. Mounted in SOT23-5 packages, they are used where a single switch function is needed, providing it in much less space than would be occupied by a quad analogue switch pack.

At **7c)** is shown the *MAX864* dual-output charge pump. This provides outputs of +2Vin and $-2V_{in}$ nominal, for any input V_{in} in the range +1.75 to +6.0V. Two pins, FC0 and FC1, are connected to ground or V_{in} as required, offering a choice of four different internal switching frequencies in the range 7 to 185kHz, assuming that the /SHDN pin is high. The *MAX864* is packaged in a QSOP outline, Fig. **7b)**.

Figure 8 shows a 12-bit d-to-a converter, the *LTC1405*, from Linear Technology. It accepts 12-bit parallel input data and outputs up to 4.095V or 2.048V (pin strappable selection), from a 4.5 to 5.5V supply. The *LTC1450L* provides a 12-bit resolution output of up to 2.5V or 1.22V, from a 2.7 to 5.5V supply.

Figure **8a)** shows the internal workings of the chip, which is available mounted in a 24 lead SSOP package, **b)**, or in a 28 pin DIP. Figure **8c)** shows the companion *LTC1458/1458L*, which is a quad 12-bit d-to-a converter. It is shoe-horned into a 28 pin small-outline package, or a 28-pin SSOP, by using a serial data input scheme, rather than the parallel data input of the *LTC1405/L*.

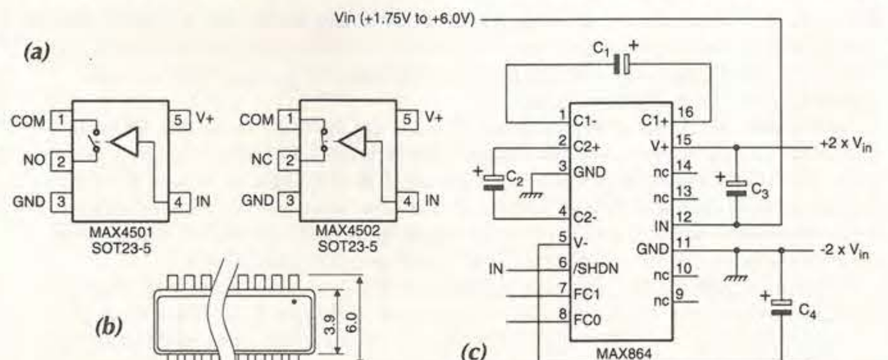


Fig. 7a). Single normally-open or normally-closed analogue switches save space compared to leaving a quarter of a quad pack unused. b) is the MAX861 package and c) is the pin-out and application circuit.

Figure 9 shows another d-to-a converter, this time one which accepts 16 or 18 bit data. It is designed for use in compact-disk systems, MPEG audio, MIDI applications, etc. The *PCM1717E* from Burr-Brown incorporates an eight-times oversampling digital filter, multi-level delta-sigma d-to-a converter and analogue low-pass in each of its stereo output channels. Its selectable functions include soft mute, digital de-emphasis and 256 step digital atttenuation. Using a serial data input, it is supplied in a 20-pin SSOP package, a shorter version of that shown in Fig. 8b).

Digital alternatives

Traditional small and medium-scale integration logic circuits – originally supplied in 0.3in width packages with up to 16 (later, 18, 22 or more) pins – have long ago migrated to the SO and even smaller packages.

Large-scale integration devices with up to 64 or 68 pins came in 0.6in wide packs, but then migrated to a variety of package types, including leaded and leadless chip carriers, J-lead packs, pin-grid arrays etc. The latest development in packaging is ball-pin arrays.

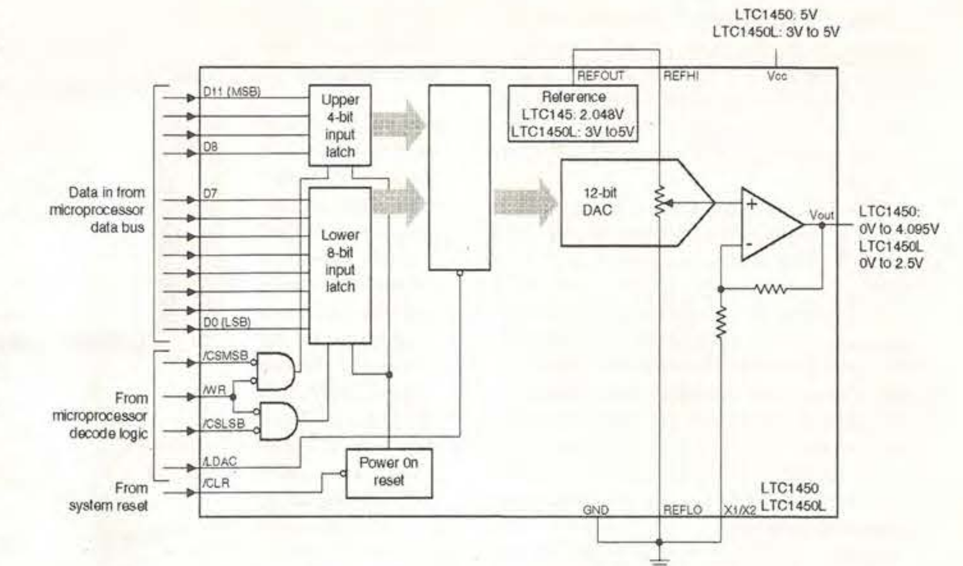
But processors, dsp chips and the like tend to require so many leadouts that they hardly come under the heading of tiny devices, even though truly small considering the number of pins. This is illustrated in **Fig. 10**, which shows packages with a modest 44 pins, **c)** and **d)**; 52 pins, **b)**; and 240 pins, **a)**. This latter package even comes in a version with 304 pins.

In addition to processors, dsp chips etc, package types with a large number of pins are also used for custom- and semi-custom logic devices, and programmable arrays of various types. These enable all the logic functions associated with a product to be swept up into a single device, reducing the size and cost of products which are produced in huge quantities.

But this approach is not without its drawbacks, often leading to practical difficulties at the layout stage. For example, on a densely packed board, the odd logic function such as an inverter, AND gate or whatever, may be required at the opposite end of the board from that at which the huge do-it-all logic package is situated. This forces the designer either to accept long digital signal runs right across the board, or to include a quad small-scale integration package, of which only a quarter is used, or to seek some other solution.

Discrete logic

Such a solution is now at hand, right at the other extreme from multi-pin packs, or even 14-pin small-scale integration quad-gate packs. For example, a simple resistor/transistor logic, or rtl, inverter can be implemented with a 'digital transistor' as shown in **Fig. 11 a)**, using a surface-mount resistor as collector load.



(a)

These digital transistors, from Rohm, are available in the tiny three pin packages shown in Fig. 3, with a variety of values for R_1 and R_2. For example, type *DTC144ExA* is an npn transistor where R_1=R_2=47kΩ. Suffix *x* is a code indicating which of the three

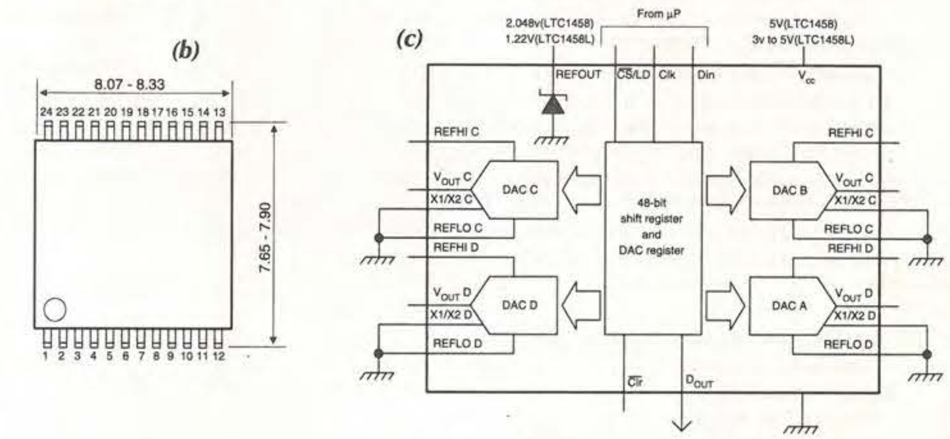


Fig. 8. LTC1405, a), from Linear Technology is a 12-bit d-to-a converter with parallel data input. This reqires a 24-pin package, b), but the SO small-outline pack is still much smaller than the corresponding DIP.
c) shows a block diagram of the internal workings of the LTC1458, from the same manufacturer. This quad d-to-a converter comes in an SO pack, or the even smaller SSOP. Both have only 28 pins, achieved by using a 48-bit serial data input stream.

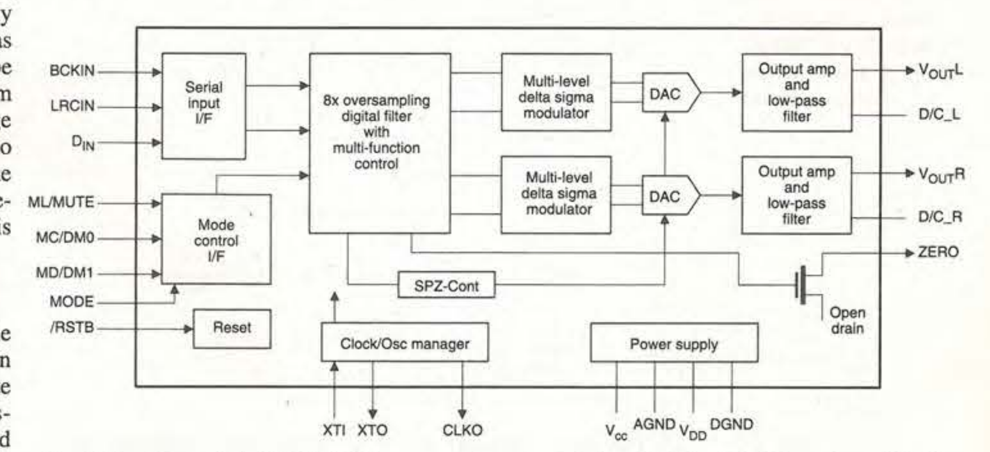


Fig. 9. Burr-Brown's PCM1717E d-to-a converter accepts 16 or 18 bit serial data, and provides L and R stereo output channels. With numerous facilities, aimed at cd systems, MPEG audio, MIDI applications etc.

These digital transistors, from Rohm, are available in the tiny three pin packages shown in Fig. 3, with a variety of values for R_1 and R_2. For example, type *DTC144ExA* is an npn transistor where R_1=R_2=47kΩ. Suffix *x* is a code indicating which of the three packages of Fig. 3 applies.

Adding another such transistor connected to the same collector load provides the NOR function, while connecting them as in Fig. **11b)** gives the inverse EXOR or exclusive NOR function. With three separate components, this provides just about the most flexible layout possibilities that could be devised.

However, a single component solution is also possible. Nearly all the functions which are available in quad small-scale integration packs are also available as singles in the SOT23-5 pack. One example has already been illustrated in Fig. 7a).

Suppose for example that an EXOR gate were required, this is readily available in c-mos as the *NC7S86M5*, see Fig. 11c) and d), from National Semiconductor, along with AND, NAND, OR, NOR gates etc. The device quoted operates from supplies of 2V to 6V, sinks or sources 2mA and has a typical propagation delay T_{pd} of 4.5ns.

As well as the large packages of Fig. 10, special purpose digital ICs are available in the smaller packs discussed here. A good example is the *REG5608*, which is an 18-line SCSI (small computer systems interface) active terminator chip from Burr-Brown, **Fig. 12.** On-chip resistors and voltage regulator provide the prescribed SCSI bus termination, while adding only 2pF per line – important for SCSI FAST-20 operation.

All SCSI terminations can be disconnected from the bus with a single control line. The chip output lines then remain in a high impedance state with or without power applied. This is important for 'hot socket' equipment plugging. The device is available in both 28-pin SOIC and fine pitch SSOP packages.

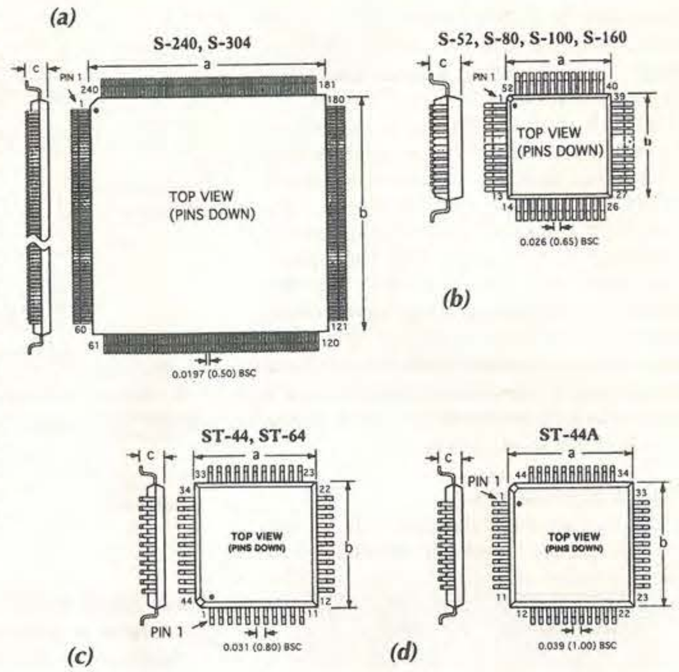


Fig. 10. Digital ICs come in packs with up to 300 pins – or more. a) shows the 240-pin PQFP (plastic quad flatpackage) S-240. The slightly wider pin spacing of PQFP packs with up to 160 pins, b), is more manageable. There are traps for the unwary! The two 44-pin TQFP (thin quad flat package) packs in c) and d) look very similar, but the pin spacing is different.

Technical considerations

When using the very small types of components discussed here, a somewhat different approach is called for, compared with ICs in DIPs and other easily handled parts.

The practical difficulties of conventional breadboarding have already been mentioned. With these very small parts, designers often go straight to pcb design from simulation to avoid the difficult job of prototyping. In any case, if the circuit involves one or more of the fine pin-pitch multi-pin devices, some of which are illustrated in Fig. 10, then a circuit-board layout will be required at the outset anyway.

Simulation is eased by the availability of Spice models for many of these devices; even if not, an op-amp model using just the input capacitance, first and second breakpoints and the output resistance may prove adequate.

It is useful to add a few strategically placed pads or plated-through holes to provide test-points for use in evaluation and debugging. This is safer than trying to probe pins which are spaced a millimeter or less apart.

Manufacturers face various problems producing very small parts. One concerns packaging, where the package dimensions may not be much larger than the basic silicon chip itself. For example, the *LT1078/9* and *LT1178/9* family of single-supply op-amps in standard DIP format from Lineear Technology are justly popular. They exhibit very low supply currents of 55μA and 21μA per op-amp respectively. But the same devices in the surface mount SO outline exhibit worse maximum input offset voltage V_{os}, and offset voltage drift. This is because the plastic surface mount packages, in cooling, exert stress on the top and sides of the die, causing changes in the offset voltage.

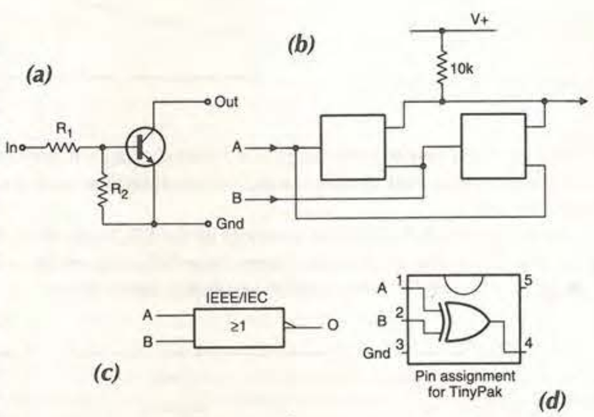


Fig. 11. Digital transistors, a), from Rohm, are available in SOT 23 packs, Fig. 3, with a variety of values for R_1 *and* R_2*. Two such transistors connected as in b) give the inverse EXOR or exclusive NOR function. A single component solution is also possible, being readily available in c-mos as the NC7S86M5, c) and d), from National Semiconductor*

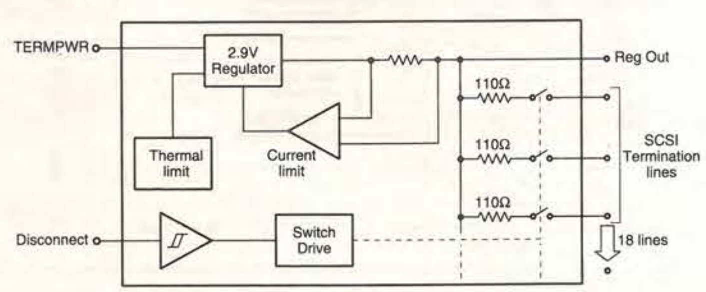


Fig. 12. REG5608 is an 18-line SCSI (small computer systems interface) active terminator chip from Burr-Brown. On-chip resistors and voltage regulator provide the prescribed SCSI bus termination. A single control line open circuits all the terminations, important for 'hot-socket' equipment plugging. The device is available in both 28-pin SOIC and fine pitch SSOP packages.

continued on page 404

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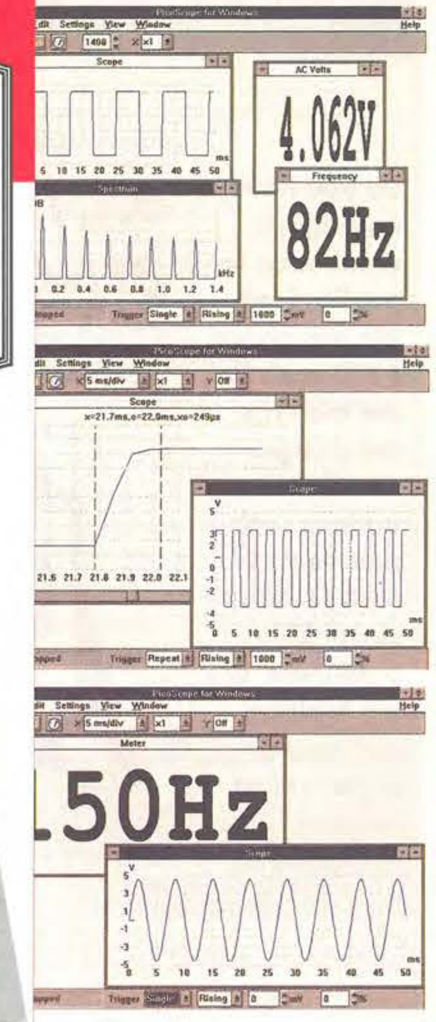

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The ADC40 has 8-bit resolution and is suitable for a wide range of dc and ac measurements and analyses. Resolution of the ADC-42 is 12 bits, making it more suitable for applications where detection of small signal changes is needed.

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Spectrum analysis	100Hz to 10kHz
Max sampling	20ksample/s
Voltage ranges	±5V
Resolution	8 bit for ADC40
	12 bit for ADC42
Channels	1 BNC
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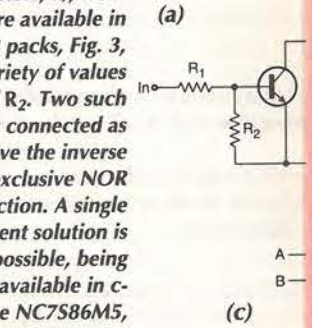

Fig. 11. Digital transistors, a), from Rohm, are available in SOT 23 packs, Fig. 3, with a variety of values for R*₁ *and R*₂*. Two such transistors connected as in b) give the inverse EXOR or exclusive NOR function. A single component solution is also possible, being readily available in c-mos as the NC7S86M5, c) and d), from National Semiconductor

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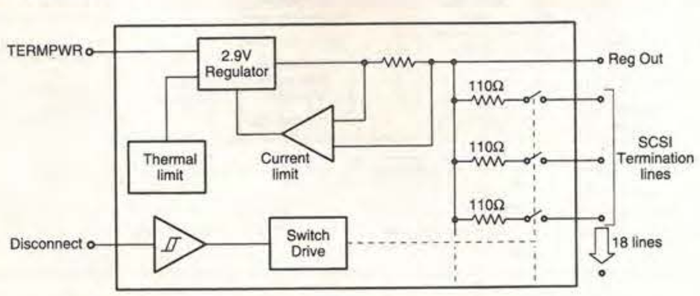

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continued on page 404

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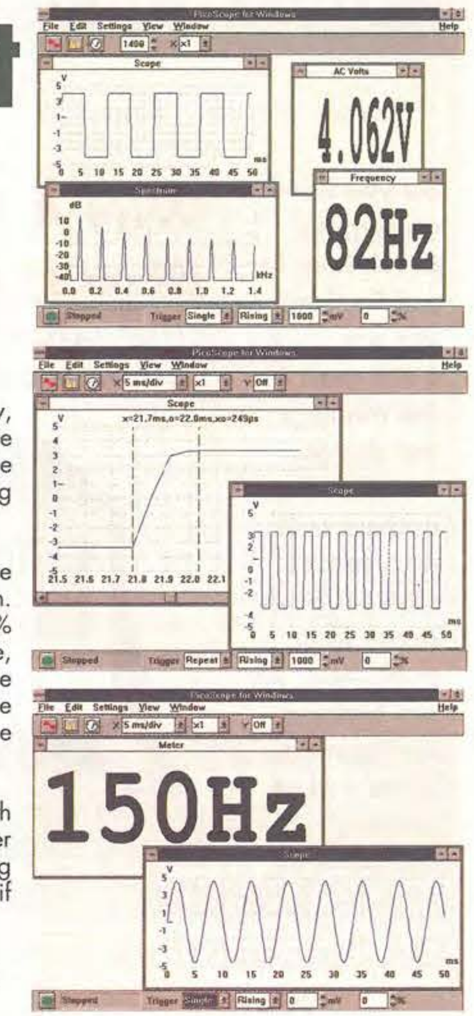

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All the storage and display features are available with the spectrum analyser function. These include pre and post triggering in 1% steps. Seven windows types are possible, together with signal averaging and all the trigger functions available in oscillscpoe mode. And rulers are available for amplitude and frequency measurements.

In addition, you have a true rms voltmeter with decibel range, and a frequency meter capable of reading to 5kHz. Data-logging software is available for an extra £10 if purchased with an ADC40 or 42.

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The ADC40 has 8-bit resolution and is suitable for a wide range of dc and ac measurements and analyses. Resolution of the ADC-42 is 12 bits, making it more suitable for applications where detection of small signal changes is needed.

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Voltage ranges	±5V
Resolution	8 bit for ADC40
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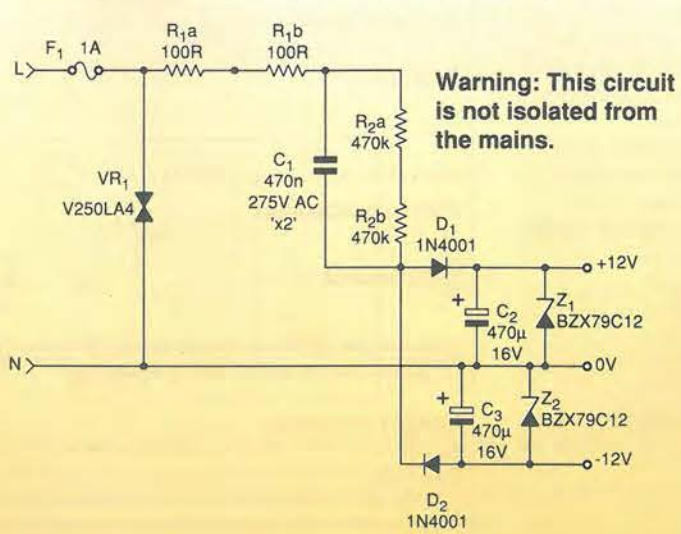
Current from the mains live goes via C_1 to the rectifier $D_{1,2}$ to $C_{2,3}$, C_1 being an X-rated component designed for this purpose, such as the *Arcotronics* or *Philips* 275Vac X2 types. Zeners $Z_{1,2}$ limit the rectified output under no load to suit the application; for example, to ±12V for op-amps.

With the values shown, that of C_1 in particular, the circuit will supply about 12mA from each rail before regulation is lost and the 50Hz ripple rises to around 350mVpk-pk. Voltage rises slowly at switch-on, since C_1 dumps only about 300 microcoulombs into $C_{2,3}$ in each cycle. Short-circuit current is about 15mA; and C_1 acts as a current source dissipating virtually no power, whereas a resistor having to drop 220V and supply ±15mA would lose 7W. Resistor R_1 limits current in the presence of mains spikes; a varistor and fuse can be used for extra safety.

This circuit is not isolated from the mains. If mains neutral becomes disconnected or connections transposed, the output terminals are live. Only use the circuit in an insulated and touch-proof enclosure with no exposed conductors.

CJD Catto
Cambridge



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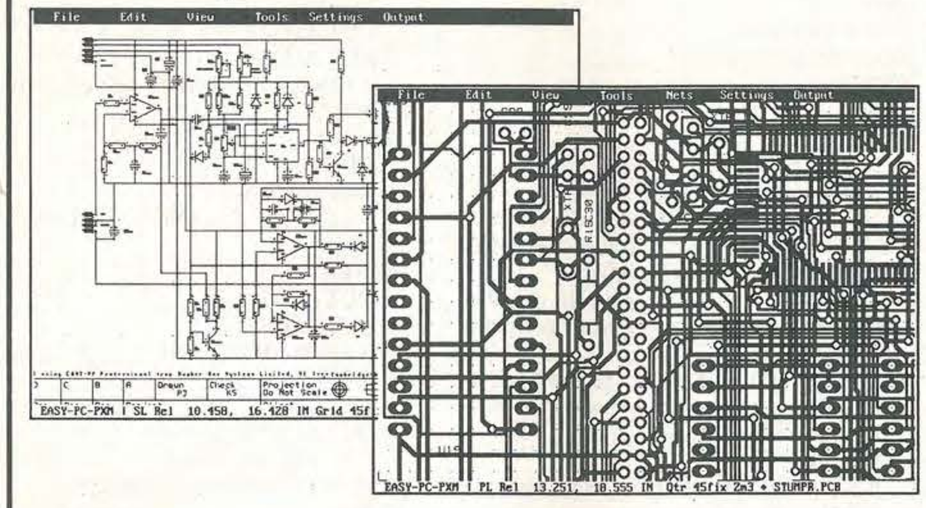

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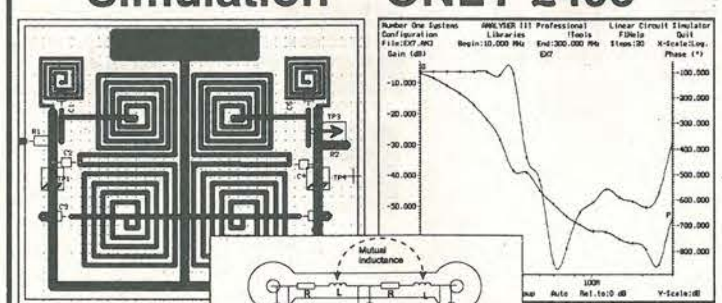

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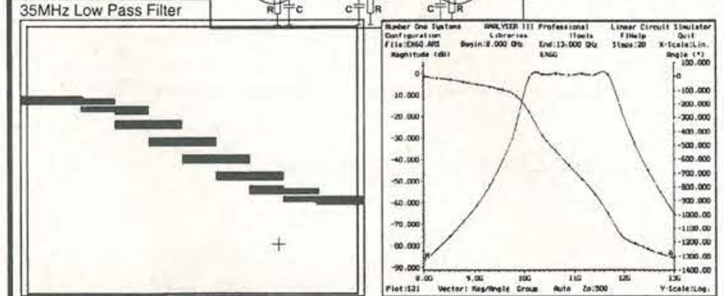

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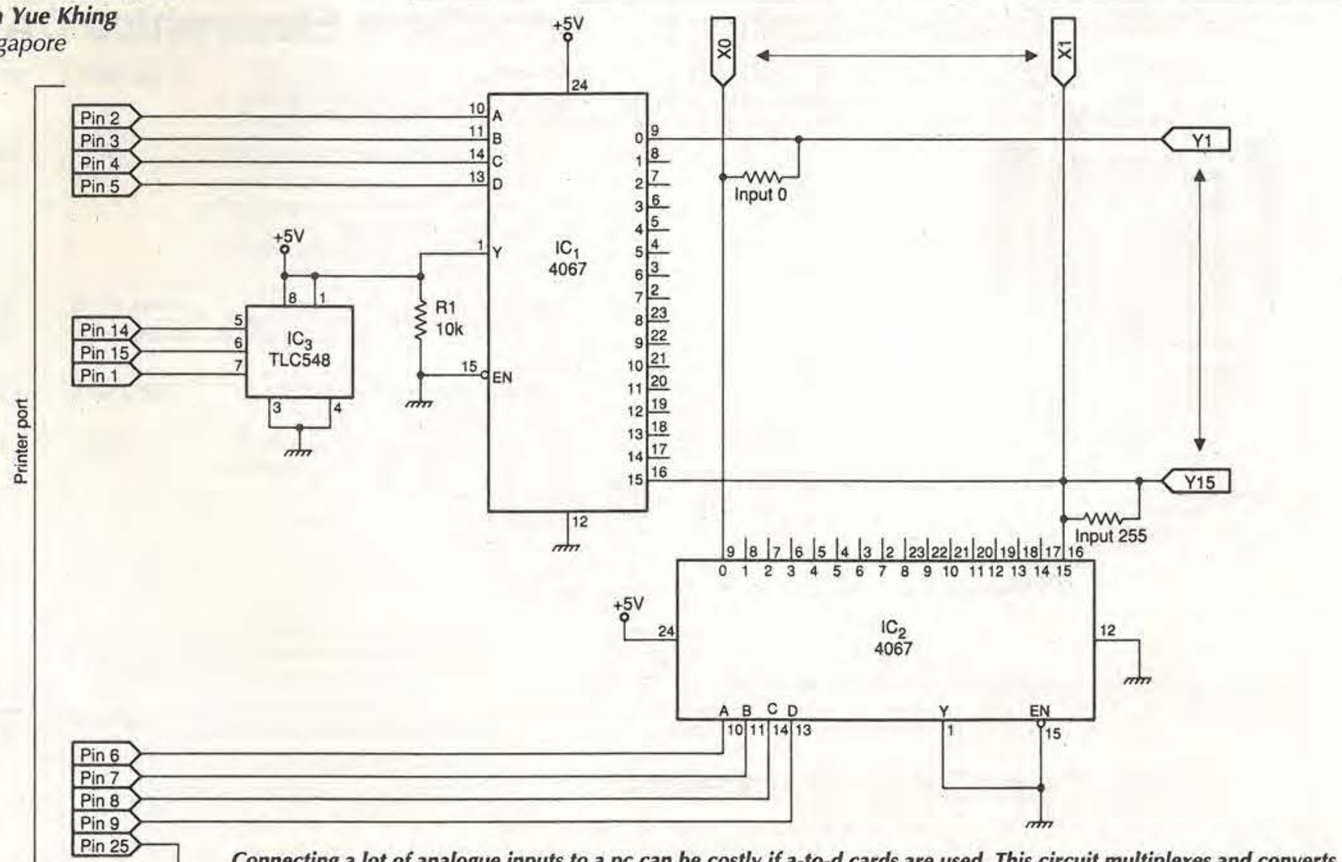
A matrix formed by the analogue inputs is controlled in X and Y directions by the two *4067B* multiplexers, in turn controlled by code from the computer. Firstly, the IC_2 multiplexer connects one end of the selected X input to ground, IC_1 connecting the other end to the input of the *TLC548* a-to-d converter, which sends it in the form of serial binary code to the pc. Other inputs on the same Y coordinate are not connected to ground and do not affect the selected input.

First four bits of address 378_{16} control IC_1, the last four being for IC_2. For example, 00000000 selects X0Y0 and 00000001 selects X1Y0. On address 379_{16}, the multiplexer transmits the binary back to the computer; $37A_{16}$ controls the clock and clears the converter.

Toh Yue Khing
Singapore

Qbasic listing for controlling the printer port analogue capture system.

```
CLS
DO
x=0
y=0
FOR x = 0 to 15
LOCATE 1,4*x+5
PRINTx
   FOR y = 0 TO 15
   PRINT y
   LOCATE 21, 21
   port=y+x* 16
   OUT &H378, port          'sent address to
                             analog multiplexer
   'code for getting data for ADC
   OUT &H37A, 1             'clear the chip data
   OUT &H37A, 3             'sent a clock pulse
   OUT &H37A, 2
   a = INP(&H379) AND 8     'get the first bit
   a=a* 16
   OUT &H37A, 3
   OUT &H37A, 2
   b = INP(&H379) AND 8     'get the second bit
   b=b*8
   a=a+b
   OUT &H37A, 3
   OUT &H37A, 2
   b = INP(&H379) AND 8     'get the third bit
   b=b*4
   a=a+b
   OUT &H37A, 3
   OUT &H37A, 2
   b=INP(&H379) AND8        'get the four bit
   b=b*2
   a=a+b
   OUT &H37A, 3
   OUT &H37A, 2
   b=INP(&H379) AND 8       'get the fifth bit
   a=a+b
   OUT &H37A, 3
   OUT &H37A, 2
   b = INP(&379) AND 8      'get the six bit
   b=b/2
   a=a+b
   OUT &H37A, 3
   OUT &H37A, 2
   b = INP(&H379) AND 8     'get the seven bit
   b=b/4
   a=a+b
   OUT &H37A, 3
   OUT &H37A, 2
   b = INP(&H379) AND 8     'get the eight bit
   b=b/8
   a = a + b                'compute the result
   OUT &H37A, 3
   LOCATE y+2, 4*x+5
   PRINT CHR$(255); CHR$(255),
CHR$(255);                  CHR$(255) 'clear
the previous number
   LOCATE y+2, 4 * x+ 5
PRINT INT(a/255 * 5 * 10) / 10 'round off to 1
decimal place
   NEXT y
NEXT x
LOOP UNTIL INKEY$ = CHR$(27)   'loop
   until "ESC" key is pressed
```



Connecting a lot of analogue inputs to a pc can be costly if a-to-d cards are used. This circuit multiplexes and converts 256 inputs, sending them to the pc's printer port.

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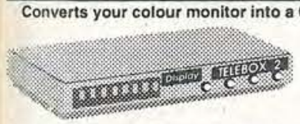

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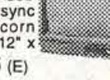

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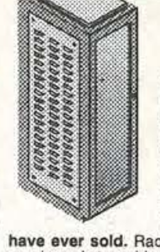

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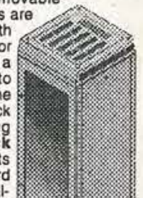

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£75 WINNER

Contactless ignition with electronic advance for motor-bikes

Electronic ignition for motor-cycle engines advances electronically and has an electronic equivalent of vacuum advance during acceleration.

An optical sensor and rotating shutter on the crankshaft control coil drive and the circuit effects advance electronically; this is primarily meant for two-stroke engines or twin four-strokes with twin coils.

OPB625 opto-sensor, a latch formed by half of *4013* IC_4 and Tr_1 turn coil driver Tr_2 on at bottom dead centre and interrupt coil current at top dead centre by way of IC_{3f}, C_1 and D_2 at engine speeds of 0-1500rev/min, led D_1 indicating TDC. At TDC, the *4024* divider IC_2 and *4046* phase-locked loop IC_1 are synchronised to give an output frequency of 128×rev/min to drive the *40161* counters $IC_{5,6}$; each cycle represents 2.8° of crankshaft rotation.

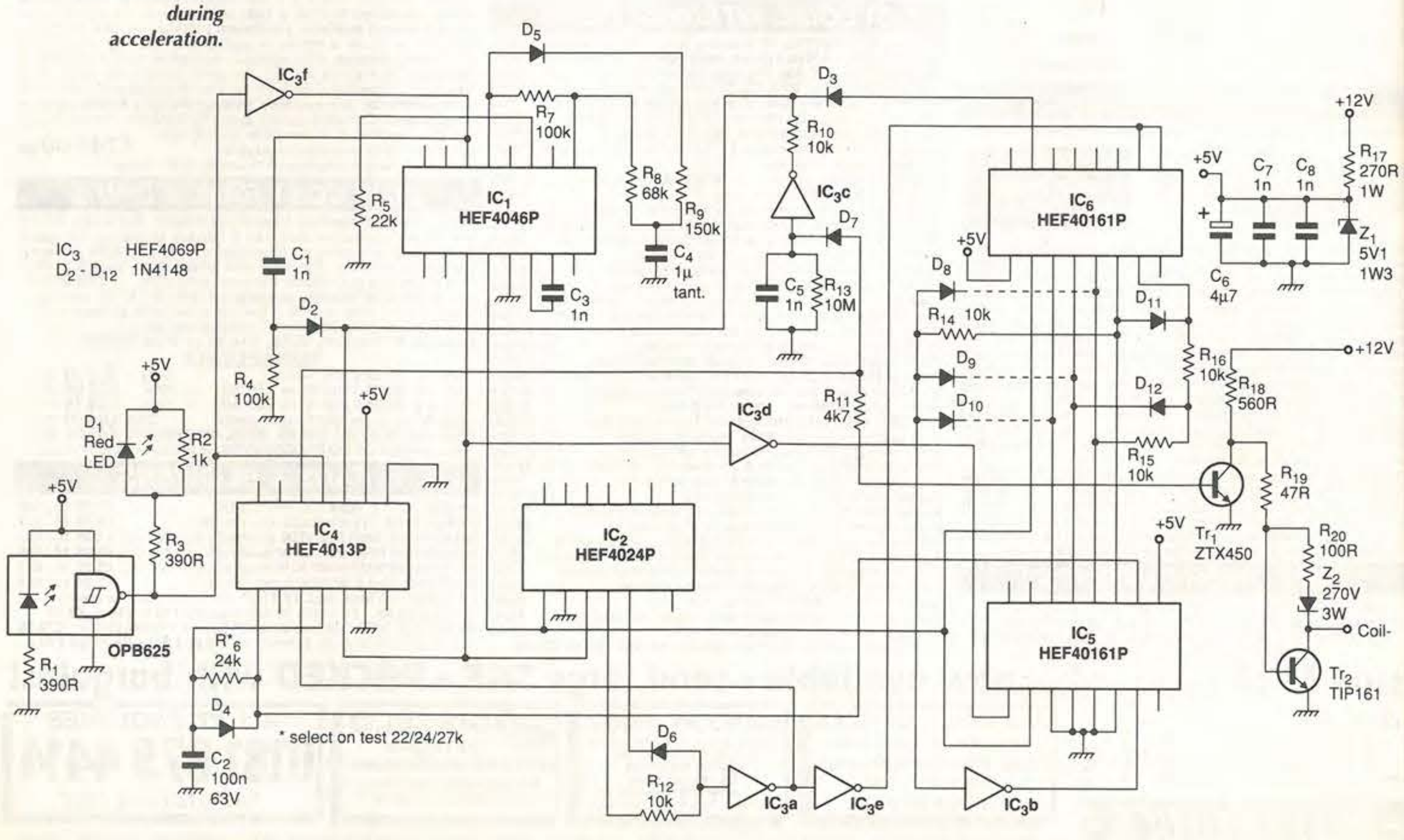
The other half of IC_4 is a monostable device whose 1.9ms pulse is obtained by selecting R_6 and which is triggered at BDC by IC_2 via IC_{3a}, enabling IC_5 to count the output of IC_1. As this count reaches 6, which represents 1500rev/min, the And/Or network $R_{15,16}$, $D_{11,12}$ sets IC_6, which now counts up from the value reached by IC_5 when enabled by IC_{3e} at 45° before TDC. As it reaches 15, it resets the IC_4 latch via D_3, interrupting coil current at 20° advance (45–(9×2.8)°). IC_{3d} resets IC_5 at TDC.

Nand circuit IC_{3b}, R_{14} and diodes selected from $D_{8\text{-}10}$ determine the maximum count reached by IC_5; using R_{14} and D_8 alone, for example, the count stops at 12, which is the value reached at 3000rev/min. Since IC_6 is now only required to count to 3 before resetting the IC_4 latch, coil current is interrupted at 37° advance (45–(3×2.8)°). Advance angle therefore increases in six steps of 2.8° from 1500 to 3000rev/min, and then stays constant. For other maximum advance values, select different diode combinations.

Circuitry around D_7, R_{13}, C_5 and IC_{3c} only lets coil current flow when the engine rotates and reset the IC_4 latch about 5s after it stops. A filter, D_5, R_9 reduces pll phase lag during acceleration, which retards ignition proportionately, this being equivalent to vacuum advance which is not used in the type of engine considered here.

For operation on a 6V battery, R_{17} becomes 33Ω and R_{18} 270Ω.

H Maidment
Wilton
Wiltshire



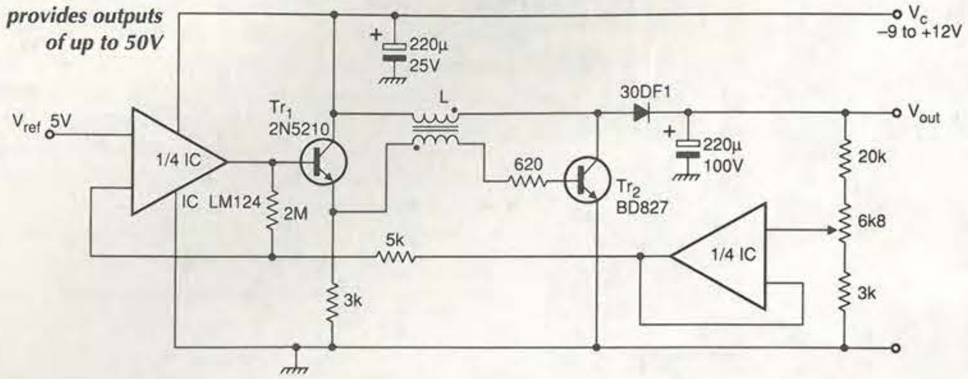
Self-oscillating step-up converter

With a supply voltage of around 9-12V, this stabilised converter provides outputs of up to 50V

With the additional feature of good stability, this oscillator provides an output voltage several times higher than its supply rail.

Transistor Tr_2 and the transformer form the oscillator, whose mark:space ratio depends on the emitter voltage of Tr_1; feedback from the output to that point via the voltage divider and the two op-amps confers stability.

Transistor Tr_1 should have a h_{FE} of several hundred or be a Darlington stage, while Tr_2 must handle reverse voltages of around 10V; a heat sink may be needed for this stage. Design the transformer carefully to avoid saturation; collector and base windings may be equal or, with higher supply voltages, the base winding can be about three-quarters of the collector one.



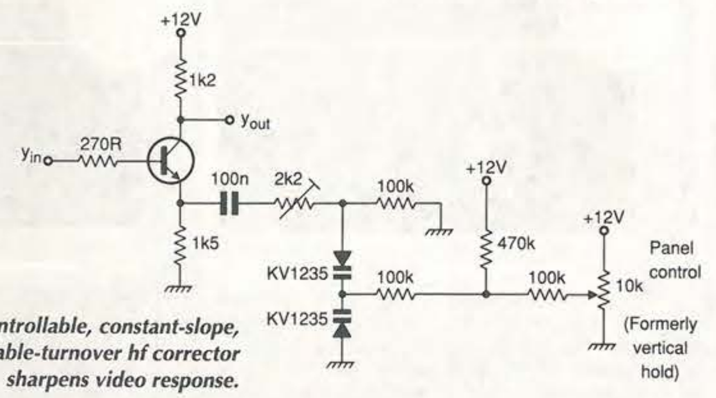
G Mirsky
Moscow
Russia

Luminance hf corrector sharpens video

A couple of varicaps yield a method of varying the hf response of a video luminance amplifier.

A new video circuit is built around the Sony XE3 chassis's luminance amplifier transistor, which is fed with the Y signal, delayed and minus chroma. Control to allow adjustment of turnover frequency is a 10kΩ pot., which was the little used vertical hold control.

Some 'sharpener' circuits give a variable-slope, fixed turnover frequency characteristic, but here the reverse happens and the circuit can be set to compensate for droop from any frequency. Negative feedback generated by the previously undecoupled emitter resistor is varied by the new control, the 2.2kΩ potentiometer preventing gain becoming too high at very high frequencies.



Controllable, constant-slope, variable-turnover hf corrector sharpens video response.

Norman J McLeod
Brighton

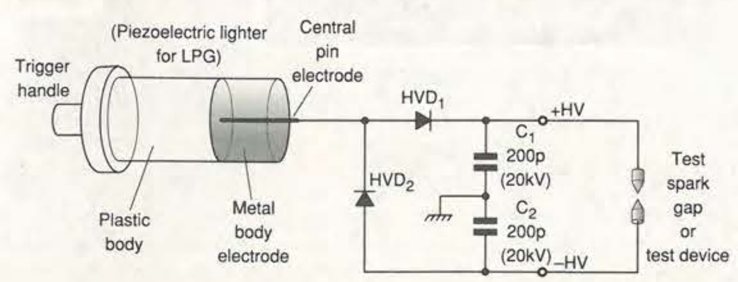
Man-powered high-voltage tester

A gas-stove lighter makes a good 20kV source, but be careful!

You can make a 20kV, low-current source for simple insulation or gap testing from the kind of piezoelectric gas lighter operated by a trigger and a few extra components; no other power is needed.

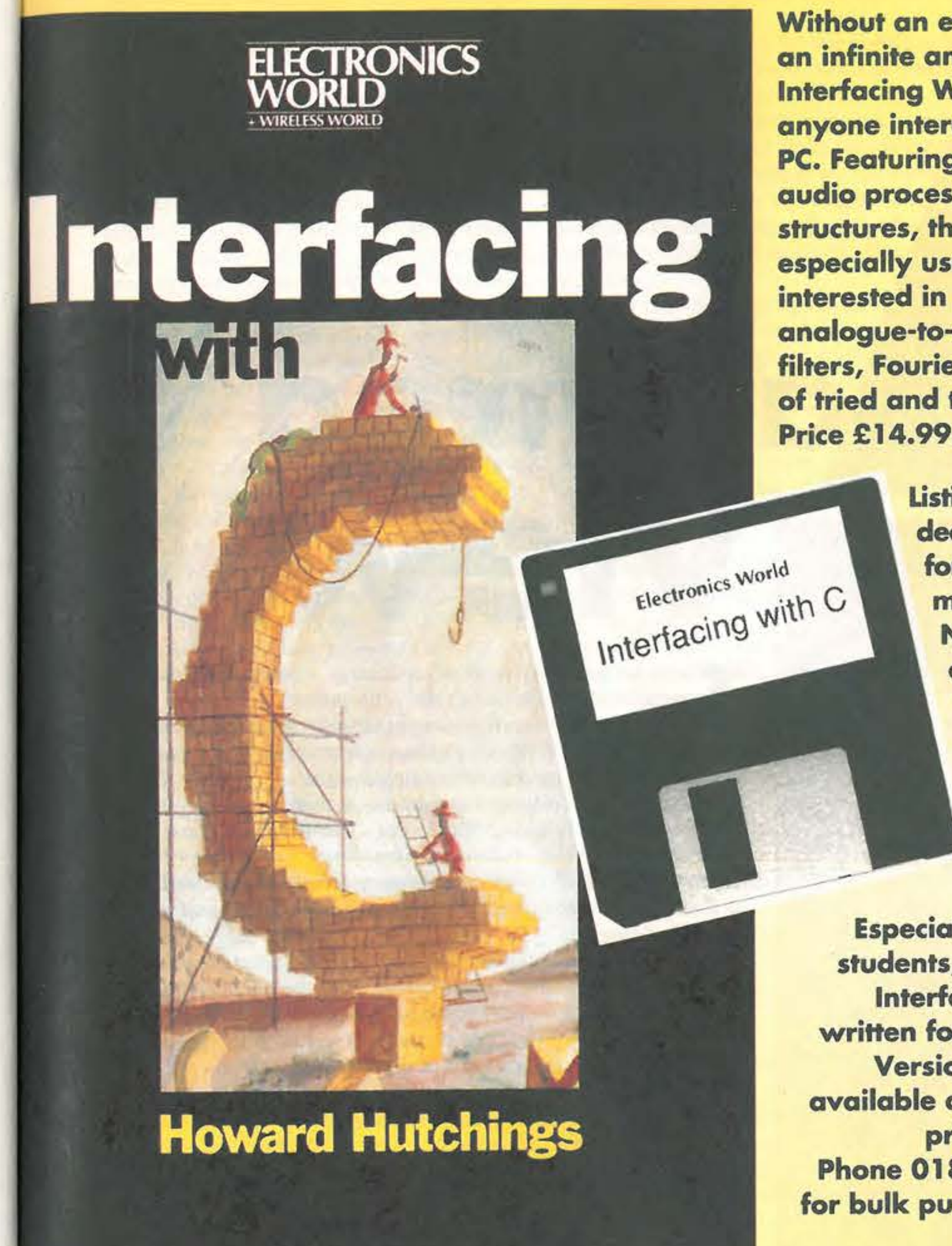
The lighter generates an oscillatory 20kV pk-pk waveform which is rectified by high-voltage diodes, the resulting dc being stored on 200pF capacitors in series to give 100pF, which represents 2microcoulombs at 20kV. This is probably safe, but **increasing the capacitance to give more charge could, if applied in the wrong place, be very dangerous. If you do increase it, take thorough precautions!** It might also be necessary, if the capacitance is increased, to trigger the lighter several times. Apply the output to the spark gap and observe the corona.

In the original, the lighter electrodes were covered with Teflon tape to stop corona discharge in the lighter's spark gap cavity.



Shyam Sunder Tiwari
Kalpakkam
India

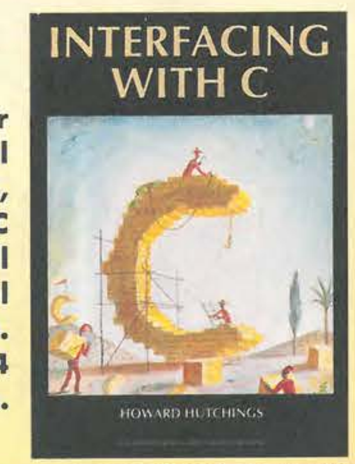
Interfacing with C



Without an engineering degree, a pile of money, or an infinite amount of time, the revised 289-page Interfacing With C is worth serious consideration by anyone interested in controlling equipment via the PC. Featuring extra chapters on Z transforms, audio processing and standard programming structures, the new Interfacing with C will be especially useful to students and engineers interested in ports, transducer interfacing, analogue-to-digital conversion, convolution, digital filters, Fourier transforms and Kalman filtering. Full of tried and tested interfacing routines. Price £14.99.

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

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



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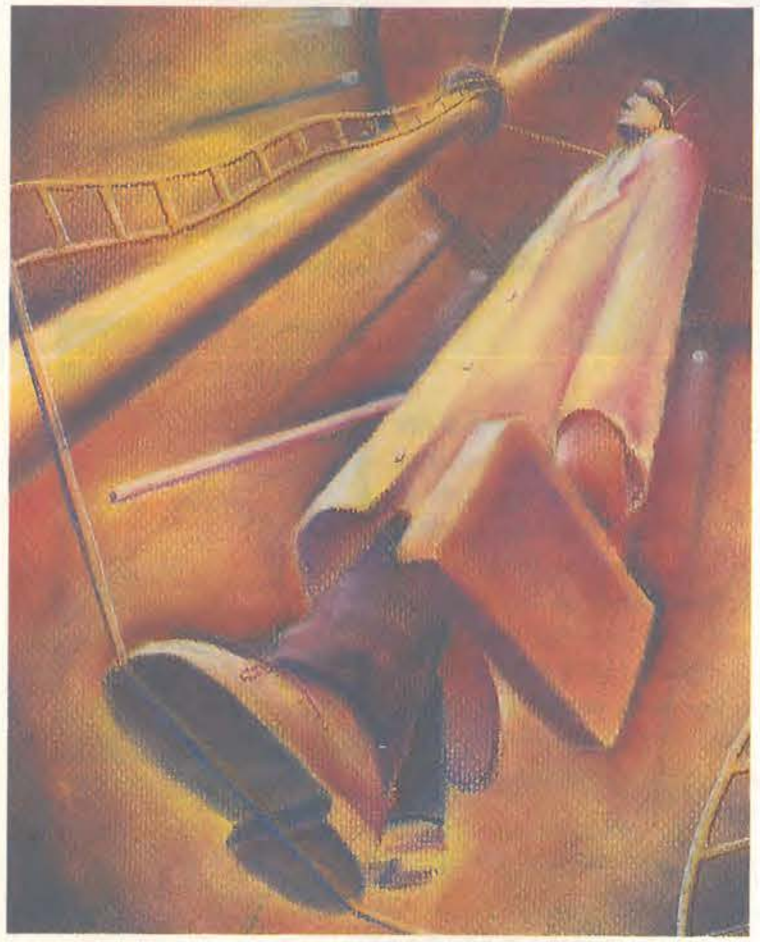
A balanced view



There's a multitude of balanced line input topologies to choose from. Here, Douglas Self looks at a selection of the more useful ones, and explains where and how they should be applied.

There are only two kinds of input stage – unbalanced and balanced. For interconnection this is the primary distinction. Apart from balancing requirements, a line-level input, as opposed to a microphone input, is expected to have a reasonably high impedance to allow multiple connections to a single output.

Traditionally, a 'bridging impedance' – ie high enough to put negligible loading on historical 600Ω lines – was 10kΩ minimum. This is still appropriate for modern low-impedance outputs. However, a higher impedance of 100kΩ or even more is desirable for interfacing to obsolete valve equipment, to avoid increased distortion and curtailed headroom.

Another common requirement is true variable gain at the balanced input, as putting the gain control further down the signal path means that it is impossible to prevent input amplifier overload. Thus you need a balanced stage that can attenuate as well as amplify, and this is where the circuit design starts to get interesting.

In the following circuitry, small capacitors often shunt the feedback elements to define bandwidth or ensure stability. These are omitted for clarity.

Unbalanced inputs. These are straightforward; variable-gain series-feedback stages are easily configured as in **Fig. 1**, providing a minimum gain of unity is acceptable; R_2 sets the gain law in the middle of the pot travel.

It is also simple to make a stage that attenuates as well as amplifies. But this implies a shunt-feedback configuration as in **Fig. 2**, with a variable input impedance. The minimum input impedance R_1 cannot be much higher than 10kΩ or resistor noise becomes excessive.

For a series-feedback stage, the input impedance can be made as high as desired by bootstrapping; an input resistance of 500kΩ or greater is perfectly possible. This does *not* imply a poorer noise performance, as the noise depends on the source resistance and semiconductor characteristics.

To ram the point home, my own personal best is 1GΩ, in a capacitor microphone head amplifier. Although the input impedance is many orders of magnitude greater than the 1 to 2kΩ of a dynamic microphone preamp, the E_{IN} is –110dBu, ie only 18dB worse.

Naturally, any unbalanced input can be made balanced or floating by adding a transformer.

Balanced inputs. A standard one-op-amp differential input stage is shown in **Fig. 3**. Unlike instrumentation work, a super-high cmrr is normally unnecessary. Ordinary 1% resistors and no trimming will not give cmrr better than 45dB; however this is usually adequate for even high-quality audio work.

It is never acceptable to leave either input floating. This causes serious deterioration of noise, hum etc. Grounding the cold input locally to create an unbalanced input is quite alright, though naturally all the balanced noise rejection is lost.

The hot input can be locally grounded instead. In this case, the cold input is driven, to create a phase-inverting input that corrects a phase error elsewhere, but this is not good practice: the right thing to do is to sort out the original phase error.

Table 1. Differential amplifier input impedances.

Case	Conditions	Hot i/p Z	Cold i/p Z
1	Hot only driven	20kΩ	Grounded
2	Cold only driven	Grounded	10kΩ
3	Both driven balanced	20kΩ	6.7kΩ
4	Both driven cm, ie together	20kΩ	20kΩ
5	Both driven floating	10kΩ	10kΩ

Balanced input technologies

There are many, many ways to make balanced or differential input amplifiers, and only the most important in audio are considered. These are:

- The standard differential amplifier
- Switched-gain balanced amp.
- Variable-gain balanced amp.
- The 'Superbal' amp.
- Hi-Z balanced amp.
- Microphone preamp plus attenuator
- Instrumentation amp.

Standard differential amplifier. The standard one-op-amp differential amplifier is a very familiar circuit block, but its operation often appears somewhat mysterious. The version in Fig. 3 has a gain of R_3/R_1. ($=R_4/R_2$) It appears to present inherently unequal input impedances to the line; this has often been commented on[1] and some confusion has resulted.

The root of the problem is that a simple differential amplifier has interaction between the two inputs, so that the input impedance on the cold input depends strongly on the signal applied to the hot input. Since the only way to measure input impedance is to apply a signal and see how much current flows into the input, it follows that the apparent input impedance on each leg varies according to the way the inputs are driven. If the amplifier is made with four 10kΩ resistors, then the input impedances Z are as in **Table 1**.

Some of these impedances are not exactly what you would expect. In Case 3, where the input is driven as from a transformer with its centre-tap grounded, the unequal input impedances are often claimed to 'unbalance the line'. However, since it is common-mode interference we are trying to reject, the cm impedance is what counts, and this is the same for both inputs.

The vital point is that the line output amplifier will have output impedances of 100Ω or less, completely dominating the line impedance. These input impedance imbalances are therefore of little significance in practice; audio connections are not transmission lines (unless they are telephone circuits several miles long) so the input impedances do not have to provide a matched and balanced termination.

As the first thing the signal encounters is a 10kΩ series resistor, the low impedance of 6.7kΩ on the cold input sounds impossible. But the crucial point is that the hot input is driven simultaneously. As a result, the inverting op-amp input is moving in the opposite direction to the cold input, due to negative feedback, a sort of anti-bootstrapping that reduces the effective value of the 10kΩ resistor to 6.7kΩ.

The input impedances in this mode can be made equal by manipulating resistor values, but this makes the cm impedances (to ground) unequal, which seems more undesirable.

In Case 5, where the input is driven as from a floating transformer with any centre-tap unconnected, the impedances are nice and equal. They must be, because with a floating winding the same current must flow into each input. However, in this connection the line voltages are *not* equal and opposite: with a true floating transformer winding the hot input has all the signal voltage on it while the cold has none at all, due to the internal coupling of the balanced input amplifier.

This seemed very strange when it emerged from simulation, but a reality-check proved it true. The line has been completely unbalanced as regards talking to other lines, although its own common-mode rejection remains good.

Even if perfectly matched resistors are assumed, the common-mode rejection ratio of this stage is not infinite; with a *TL072* it is about –90dB, degrading from 100Hz upwards, due to the limited open-loop gain of the op-amp.

Switched-gain balanced amplifier. The need for a balanced input stage with two switched gains crops up frequently. The classic application is a mixing desk to give optimum performance with both semi-professional (–7.8dBu) and professional (+4dBu) interface levels.

Since the nominal internal level of a mixer is usually in the range –4 to 0dBu, the stage must be able to switch between amplifying and attenuating, maintaining good cmrr in both modes.

The obvious way to change gain is to switch both $R_{3,4}$ in Fig. 3, but a neater technique is shown in **Fig. 4**. Perhaps surprisingly, the gain of a differential amplifier can be manipulating by changing the drive to the feedback arm (R_3 etc) only, without affecting the cmrr. The vital

Table 2.

	Capacitive 1mA	CMRR
Conventional	–20dBv	–46dB
Impedance-bal 99Ω	–60dBv	–101dB
Impedance-bal 100R	∞	–85dB
Impedance-bal 101R	–60dBv	–79dB

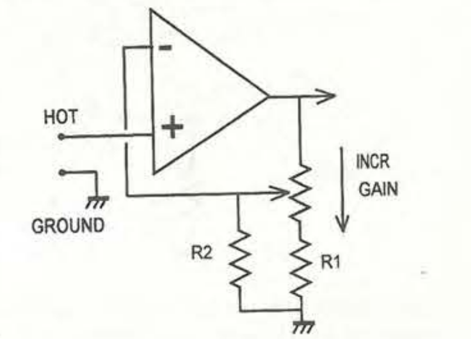


Fig. 1. variable-gain series-feedback unbalanced input stage. Resistor R_2 sets mid-position gain.

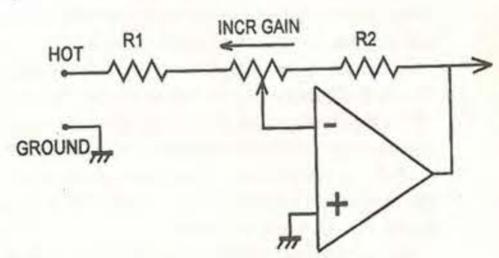


Fig. 2. Shunt-feedback configuration, with a low and variable input impedance.

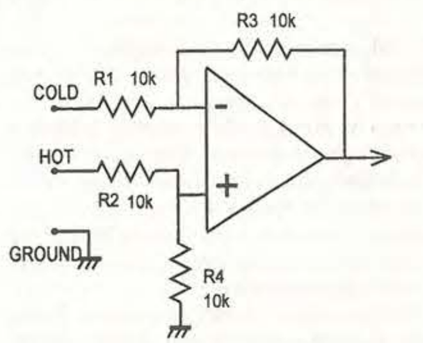


Fig. 3. Standard one-op-amp differential amplifier, arranged for unity gain.

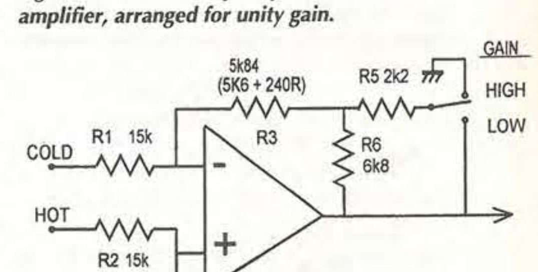


Fig. 4. Switched-gain balanced input amplifier. The values shown give gains of –6dB and +6.2dB, for switching between pro and semi-pro interface levels.

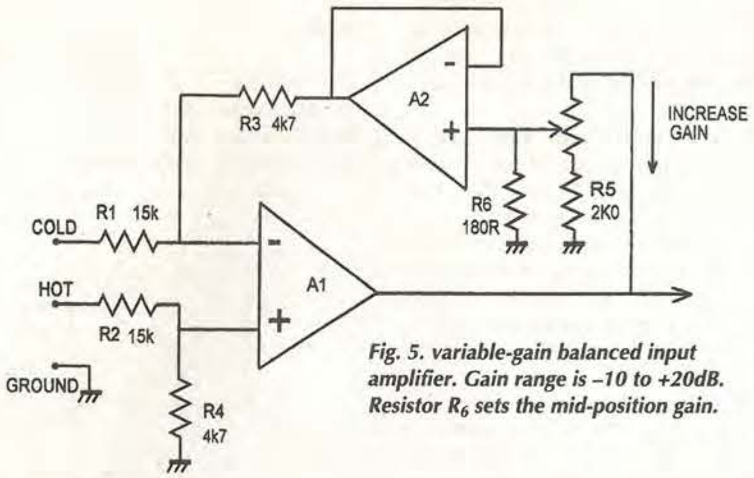


Fig. 5. variable-gain balanced input amplifier. Gain range is –10 to +20dB. Resistor R_6 sets the mid-position gain.

point is to keep the resistance of this arm the same, but drive it from a scaled version of the op-amp output.

Figure 4 uses the network $R_{5,6}$, which has the same 2kΩ output impedance whether R_4 is switched to the output (low gain) or ground (high gain). For low gain, the feedback is not attenuated, but fed through $R_{5,6}$ in parallel.

For high gain, $R_{5,6}$ become a potential divider. Resistor R_3 is reduced by 2kΩ to allow for the $R_{5,6}$ output impedance. The stage can attenuate as well as amplify if R_1 is greater than R_3, as shown here. The nominal output of the stage is assumed to be –2 dBu; the two gains are –6.0 and +6.2 dB.

The differential input impedance is 11.25kΩ via the cold and 22.5kΩ via the hot input. Common mode input impedance is 22.5kΩ for both inputs.

Variable-gain balanced amplifier. A variable-gain balanced input should have its gain control at the very first stage, so overload can always be avoided. Unfortunately, making a variable-gain differential stage is not so easy; dual potentiometers can be used to vary two of the resistances, but this is clumsy and will give shocking cmrr due to pot mismatching. For a stereo input the resulting four-gang potentiometer is unattractive.

The gain-control principle is essentially the same as for the switched-gain amplifier above. To the best of my knowledge, I invented both stages in the late seventies, but so often you eventually find out that you have re-invented instead; any comments welcome.

Feedback arm R_3 is of constant resistance, and is driven by voltage-follower A_2. This eliminates the variations in source impedance at the potentiometer wiper, which would badly degrade cmrr. As in Fig. 1, R_6 modifies the gain law; however, the centre-detent gain may not be very accurate as it partly depends on the ratio of potentiometer track (often no better than ±10%, and sometimes worse) to 1% fixed resistors.

This stage is very useful as a general line input with an input sensitivity range of –20 to +10dBu. For a nominal output of 0dBu, the gain of **Fig. 5** is +20 to –10dB, with R_6 chosen for 0dB at the central wiper position.

An op-amp in a feedback path appears a dubious proposition for stability, but here, working as a voltage-follower, its bandwidth is maximised and in practice the circuit is dependably stable.

The 'Superbal' amplifier. This configuration[2] gives much better input symmetry than the standard differential amplifier, **Fig. 6**. The differential input impedance is exactly 10kΩ via both hot and cold inputs. Common mode input impedance is 20kΩ for both inputs. This configuration is less easy to modify for variable gain.

High-Z balanced amp. High-impedance balanced inputs, above 10kΩ, are useful for interfacing to valve equipment. Adding output cathode-followers to valve circuitry is expensive, and so the output is often taken directly from a gain-stage anode. Even a light loading of 10kΩ may seriously compromise distortion and available output swing.

All of the balanced stages dealt with up to now have their input impedances determined by the values of input resistors etc, and these cannot be raised without degrading noise performance. **Figure 7** shows one answer to this. The op-amp inputs have infinite impedance in audio terms, subject to the need for *R,R* to bias the non-inverting inputs.[3]

Adding R_g increases gain, but preserves balance. This configuration cannot be set to attenuate.

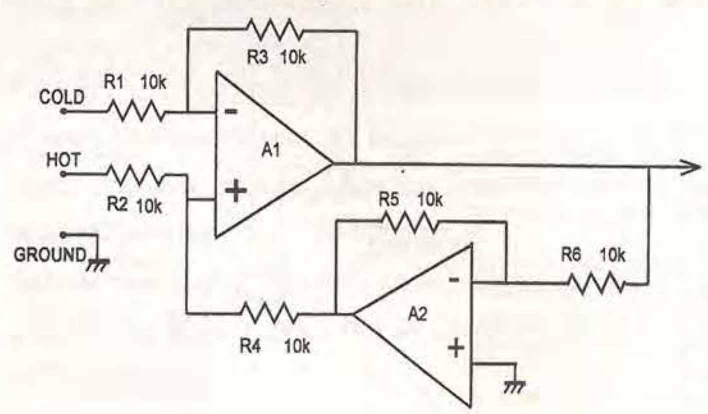


Fig. 6. The 'Superbal' balanced input stage; input impedance on hot and cold are equal for both differential and common mode.

Microphone preamp with attenuator. It is often convenient to use a balanced microphone preamp as a line input by using a suitable balanced attenuator, typically 20 to 30dB. The input impedance of the microphone input stage will be 1 to 2kΩ for appropriate mic loading, and this constrains the resistor values possible.

Keeping the overall input impedance to at least 10kΩ means that the divider impedance must be fairly high, with a lot of Johnson noise. As a result, the total noise performance is almost always inferior to a dedicated balanced line-input amplifier. Common-mode rejection ratio is determined by the attenuator tolerances and will probably be much inferior to the basic microphone amp, which usually relies on inherent differential action rather than component matching.

Figure 8a shows a bad way to do it; the differential signal is attenuated, but not the common-mode, so cmrr is degraded even if the resistors are accurate. **Figure 8b** attenuates differential and common-mode signals by the same amount, so cmrr is preserved, or at any rate no worse than resistor tolerances make it.

Instrumentation amplifier. All the balanced inputs above depend on resistor matching to set the cmrr. In practice this means better than 45dB is not obtainable without trimming. If a cmrr higher than this is essential, an IC instrumentation amplifier is a possibility.

Common-mode rejection ratio can be in the range 80 to 110dB, without trimming or costly precision components. The IC tends to be expensive, due to low production volumes, and the gain is often limited in range and cannot usually be less than unity.

In audio work, cmrr of this order is rarely if ever required. If the interference is that serious, then it will be better to deal with the original source of the noise rather than its effects.

Input/output combinations

Taking five kinds of output – the rare case of floating output transformers being excluded – and the two kinds of input amplifier, there are ten possible combinations of connection. The discussion below assumes output R_s is 100Ω, and the differential input amplifier resistors *R* are all 10kΩ, as in Fig. 3.

Unbalanced output to unbalanced input. This is the basic connection. There is no rejection of ground noise (cmrr=unity) or electrostatic crosstalk; in the latter case the 1mA notional crosstalk signal yields a –20dBv signal as the impedance to ground is very nearly 100Ω.

Unbalanced output to balanced input. Assuming the output ground is connected to the cold-line input, then in theory there is complete cancellation of ground voltages. This is true, *unless* the output has a series output resistor to buffer it from cable capacitance, – which is almost always the case – for this will unbalance the line.

If the output resistance is 100Ω, and the cold line is simply grounded as in Fig. 8a, then R_s degrades the cmrr to –46dB even if the balanced input has exactly matched resistors.

The impedances on each line will be different, but not due to the asymmetrical input impedances of a simple differential amplifier; hot line impedance is dominated by the output resistance R_s on the hot terminal (100Ω) and the cold line impedance is zero as it is grounded at the output end. The rejection of capacitive crosstalk therefore depends on the unbalanced output impedance. It will be no better than for an unbalanced input, as for the unbalanced output to balanced input case. The main benefit of this connection is ground noise rejection, which solves the most common system problem.

Impedance-balance out to unbalanced in. There is nothing to connect the output cold terminal to at the input end, and so this is the same as the ordinary unbalanced connection for the **unbalanced output to balanced input** configuration.

Impedance-balance out to balanced in. In theory there is complete cancellation of both capacitive crosstalk and common-mode ground voltages, as the line impedances are now exactly equal.

Table 2 shows the improvement that impedance-balancing offers over a conventional unbalanced output, when driving a balanced input with exactly matched resistors.

The effect of tolerances in the impedance-balance resistor are also shown; the rejection of capacitive crosstalk degrades as soon as the value moves away from the theoretical 100Ω, but the cmrr actually has its point of perfect cancellation slightly displaced to about 98.5Ω, due to second-order effects. This is of no consequence in practice.

Ground-cancelling out to unbalanced in. There is complete cancellation of ground voltages, assuming the ground-cancel output has an accurate unity gain between its cold and hot terminals. This is a matter for the manufacturer.

Ground-cancelling in this way is a very efficient and cost-effective method of interconnection for all levels of equipment, but tends to be more common at the budget end of the market.

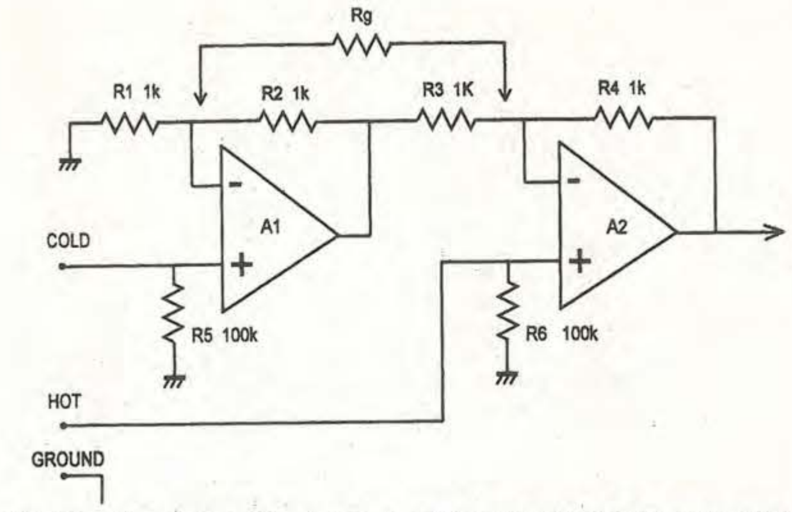


Fig. 7. High-impedance balanced input stage; R_5 and R_6 set input impedance, and can be much higher. Add R_g to increase gain.

Ground-cancelling out to balanced in. This combination needs a little thought. At first there appears to be a danger that the ground-noise voltage might be subtracted twice, which will of course be equivalent to putting it back in in anti-phase, gaining us nothing.

In fact this is not the case, though the cancellation accuracy is compromised compared with the impedance-balanced case; the common-mode rejection will not exceed 46dB, even with perfect resistor matching throughout. Capacitive crosstalk is no better than for the 'Unbalanced output to balanced input' ie approximately –21dB, which means virtually no rejection. However, this is rarely a problem in practice.

Balanced output to unbalanced input. This is not a balanced interconnection. There is nowhere to connect the balanced cold output to; it must be left open-circuit, its signal unused, so there is a 6dB loss of headroom in the link. The unbalanced input means the connection is unbalanced, and so there is no noise rejection.

Balanced out to balanced in. A standard balanced system, that should give good rejection of ground noise and electrostatic crosstalk.

Quasi-floating out to unbalanced in. Since the input is unbalanced, it is necessary to ground the cold side of the quasi-floating output. If this is done at the remote (input) end then the ground voltage drop is transferred to the hot output by the quasi-floating action, and the ground noise is cancelled in much the same way as a ground-cancelling output.

However, in some cases this ground connection must be local, ie at the output end of the cable, if doing it at the remote (input) end causes high-frequency instability in the quasi-floating output stage. This may happen with very long cables. Such local grounding rules out rejection of ground noise because there is no sensing of the ground voltage drop.

Perhaps the major disadvantage of quasi-floating outputs is the confusion they can cause. Even experienced engineers are liable to mistake them for balanced outputs, and so leave the cold terminal unconnected. This is not a good idea. Even if there are no problems with pickup of external interference on the unterminated cold output, this will cause a serious increase in internal noise. I believe it should be standard practice for such outputs to clearly marked as what they are.

Quasi-floating out to balanced in. A standard balanced system, that should give good rejection of ground noise and electrostatic crosstalk.

The hot and cold output impedances are equal, and dominate the line impedance, so even if the line input impedances are unbalanced, there should also be good rejection of electrostatic crosstalk.

Wiring philosophies

It has been assumed above that the ground wire is connected at both ends. This can cause various difficulties due to ground currents flowing through it.

For this reason some sound installations have relied on breaking the ground continuity at one end of each cable. This is called the one-end-only, or oeo, rule.[4] It prevents ground currents flowing but usually leaves the system much more susceptible to rf demodulation. This is because the cable screen is floating at one end, and is now effectively a long antenna for ambient rf.

There is also the difficulty that non-standard cables are required. A consistent rule as to which end of the cable has no ground connection must be enforced. The oeo approach may be workable for a fixed installation that is rarely modified, but for touring sound reinforcement applications it is unworkable.

A compromise that has been found acceptable in some fixed installations is the use of 10nF capacitors to ground the open screen end at rf only; however, the other problems remain.

The formal oeo approach must not be confused with 'lifting the ground' to cure a ground loop. Unbalanced equipment sometimes provides a ground-lift switch that separates audio signal ground from chassis safety ground; while this can sometimes be effective, it is not as satisfactory as balanced connections. Lifting the ground must **never** be done by removing the chassis safety earth; this removes all protection against a live conductor contacting the case and so creates a serious hazard. It is also in many cases illegal.

The best approach therefore appears to be grounding at both ends of the cable, and relying on the cmrr of the balanced connection to render ground currents innocuous. Ground currents of 100mA appear to be fairly common; ground currents measured in amps have however been encountered in systems with serious errors.

A typical example is connecting incoming mains 'Earth' – which is actually 'Neutral' in many cases – to a technical ground such as a buried copper rod. Take a look the section headed 'Electrical Noise' in last month's article for more details.

Ground currents cause the worst problems when they flow not only through cable shields but also the internal signal wiring of equipment. For this reason the preferred practice is to terminate incoming ground wires to the chassis earth of the equipment. This keeps ground currents off pcbs, where the relatively high track resistances would cause bad common-impedance coupling, and preserves rf screening integrity.

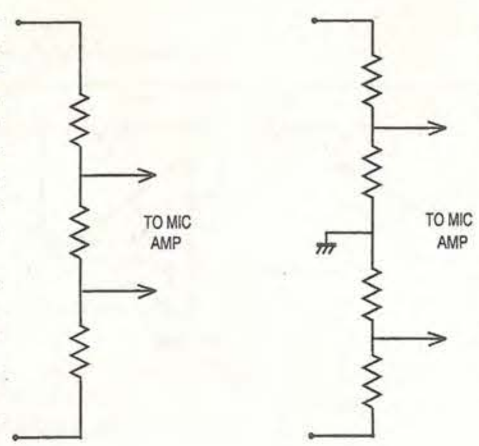


Fig. 8. At a), balanced attenuators convert a microphone preamp to line input. Circuit b) is superior as both differential and common-mode signals are equally attenuated, so common-mode rejection is not degraded more than necessary.

Grounding is simplified for source equipment that has no other connections, such as double-insulated compact-disc players. These carry a 'square-in-a-square' symbol to denote higher standards of mains insulation, so that external metalwork need not be grounded for safety. Such equipment often has unbalanced outputs, and can usually be connected directly to an unbalanced input with good results, as there is no path for any ground currents to circulate in.

If a balanced input is used, then connecting the hot input to cd signal and the cold to cd 'ground' leaves the cd player ground floating, and this will seriously degrade hum and rf rejection. The real ground must be linked to cd player common.

I think this article shows that balanced line interconnections are rather more complex than is immediately obvious. Having said that, with a little caution they work very well indeed. ■

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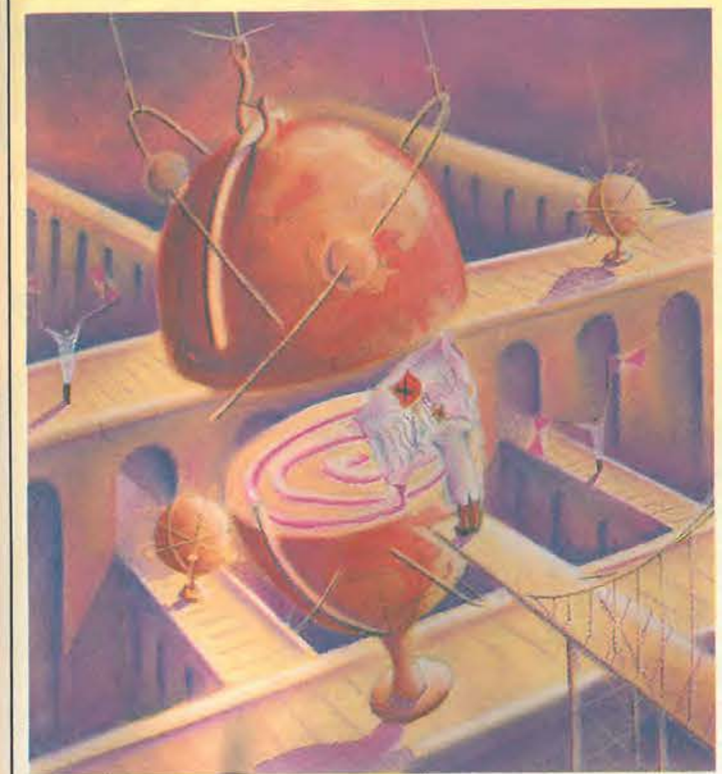

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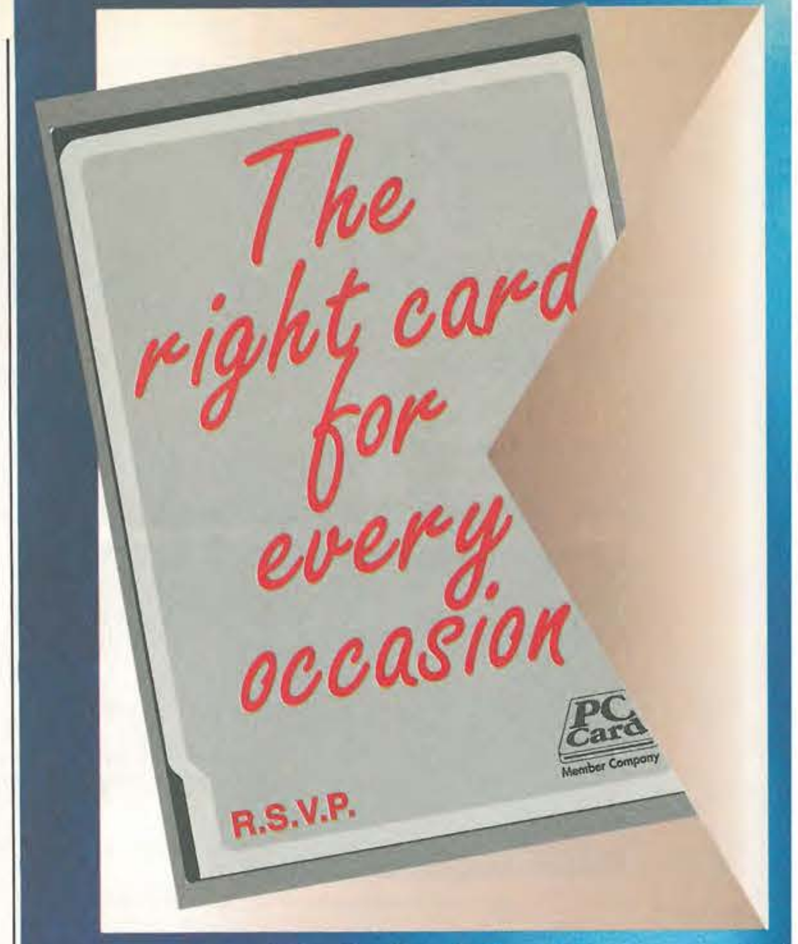

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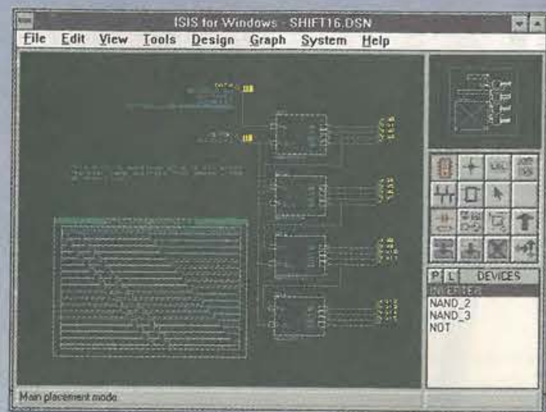

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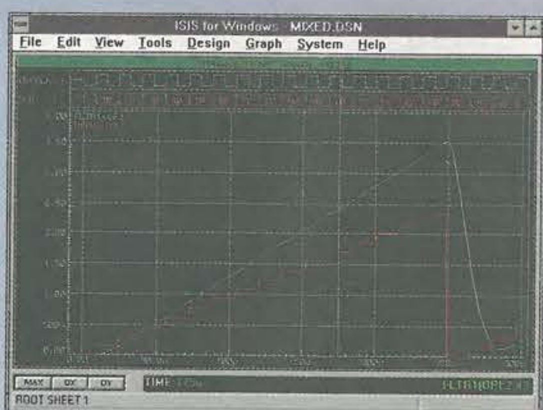

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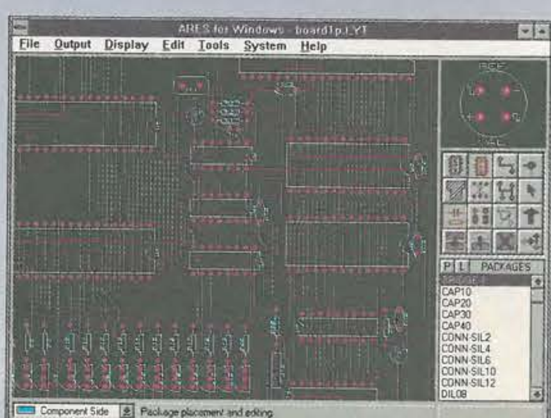

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Programmable logic

Geoff Bostock looks at the steps involved in designing field-programmable logic arrays.

The one thing that all the devices described in these articles have in common is that they are all logic devices. Because of this, they all share a common approach to the design process.

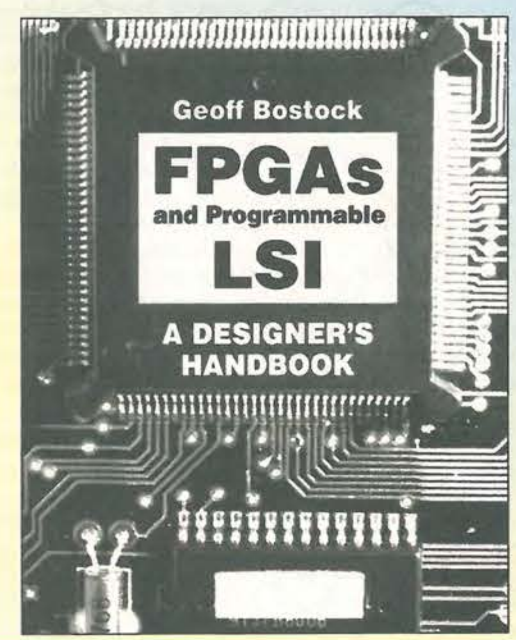
In principle, any logic circuit will fit into any fpga within the restraints of logic content and connectivity. Crudely, as long as there are enough gates and enough i/o lines, the choice of fpga does not affect the way in which the logic is defined. In practice this is not quite true. I will explain dedicated approaches to the various families in a later article. This article cover the general points of design.

There are only two general ways to design logic; the required function may be described in terms of some written language, or it may be drawn in some symbolic manner. Written methods include logic equations, state equations and hardware description langnages; symbolic descriptiom are covered by circuit diagrams and state diagrams.

All design packages make use of one or more of these categories of input to define the logic. Once the logic has been defined it is usually not dependent on any one target device or architecture, so this may be looked on as just the first stage in a design. The second stage is to ensure that the defined logic does the job which it is intended to do. This is achieved by simulating the design; that is, applying inputs to a software model of the design and checking that the outputs are as expected.

At this stage the target device can be considered. A translation from the general logic definition to specific architectural units is undertaken; abstract logic is mapped onto the physical components of the target device and

This article is derived from Geoff Bostock's new book 'FPGAs and programmable LSI – a designer's handbook'. The work covers designing FPGAs, large PAL structures, RAM and antifuse-based FPGAs and FPGA selection. Comprising 215 pages, this book is available by sending a postal order or cheque with a request for the book to Electronics World, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS. The fully-inclusive price is £27.50 UK, £30 Europe or £33 rest of world. Alternatively, fax your full credit card details and address on 0181 652 8956 or e-mail jackie.lowe@rbp.co.uk.



Geoff Bostock runs his own FPGA/PLD Design Consultancy, and may be contacted on 01380 828241, or by e-mail at geoff.bostock@zetnet.co.uk

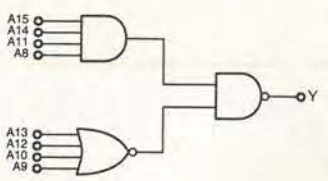


Fig. 1. The most common application for pals was address decoding in microprocessor circuit, where one pal replaced two discrete logic ICs.

potential internal connections decided. This third stage – device fitting – shows whether the target device has the capability to contain all the logic functions in the abstract design, but does not guarantee that the programmed device will work in the designers circuit. This can be taken a step doser by post-layout simulation.

Now that a real device is involved, with internal components and connections of predictable performance, the time-dependent factors can be added to the simulation result. Not only can we confirm that output Y = input A AND input B, but we can also predict that signal Y will go high within 5ns of both inputs A and B going high. Thus it is possible to program a device and plug it into a circuit with a high confidence of success – provided that all critical eventualities have been covered in the simulation.

We can now examine each design method in detail and see how they measure up to the requirements outlined above.

Logic equations
Input methods

Logic equations have been the standard method for designing plds ever since their introduction to the market in the late 1970s. As I showed in an earlier article, the early pals were simple sum-of-product devices whose outputs depended on a straightforward logic relationship between the inputs, often without any feedback, or other complications. Logic equations are the most common way of defining the logic content of classical plds.

Compilers for logic equations commonly consist of four sections. These are an introduction, a pin-out definition, an equations section and a simulation segment.

The introduction can include the drawing number, designer's name and company and a brief functional description plus other relevant information. It needs no further discussion as it merely annotates the design with information needed for future reference.

The pin-out definition section is also self-explanatory. It lists the signals used in the design and allocates them to the device i/o pins.

Simple examples are a good way of illustrating the logic equation part of the design input. The most common application for pals was as address decoders in microprocessor circuits. In this application, standard ttl parts did not have the input width or flexibility to provide an economical solution.

For example, an equation such as:

!Y = A15 & A14 & !A13 & !A12 & A11 & !A10 & !A9 & A8

was – and still is – commonplace in pal design sheets. It fits comfortably into one eighth of a standard pal whereas it would need the best part of two gate packages to implement it in discrete logic.

It is also easy to understand what is meant by this equation; when the processor puts out address 36xx, this output is taken low and will, presumably, enable some peripheral chip. **Figure 1** shows how this circuit may be implemented in discrete logic. Although not completely incomprehensible, it is not as immediately apparent as the logic equation.

If there were six or seven decoded outputs, all looking like Fig. 1, there might well be some confusion, especially without some annotation on the drawing. Annotating equations is straightforward enough; our address decode may be commented as:

!Y = A15 & A14 & !A13 & !A12 & A11 & !A10 & !A9 & A8;
decode of 36xx

the semi-colon delimiting the comment from the actual equation.

More complex functions may be defined as logic equations, with equal clarity, by using various shorthand techniques. For example, a four-bit identity comparator, which is constructed from four exclusive-OR gates and an AND gate may be defined in the following way:

EQ0 = A0 & B0 # !A0 & !B0; 'zero' bits equal
EQ1 = A1 & B1 # !A1 & !B1; 'one' bits equal
EQ2 = A2 & B2 # !A2 & !B2; 'two' bits equal
EQ3 = A3 & B3 # !A3 & !B3; 'three' bits equal
EQ = EQ0 & EQ1 & EQ2 & EQ3; all bits equal

Using the exclusive-OR symbol '#' instead of' the expanded logic definition would make the equations even clearer, and would be understood by most logic compilers.

You can also define a state machine with logic equations. Consider a four-digit identification number detector for use in a keypad combination lock. Its state diagram is reproduced in **Fig. 2**.

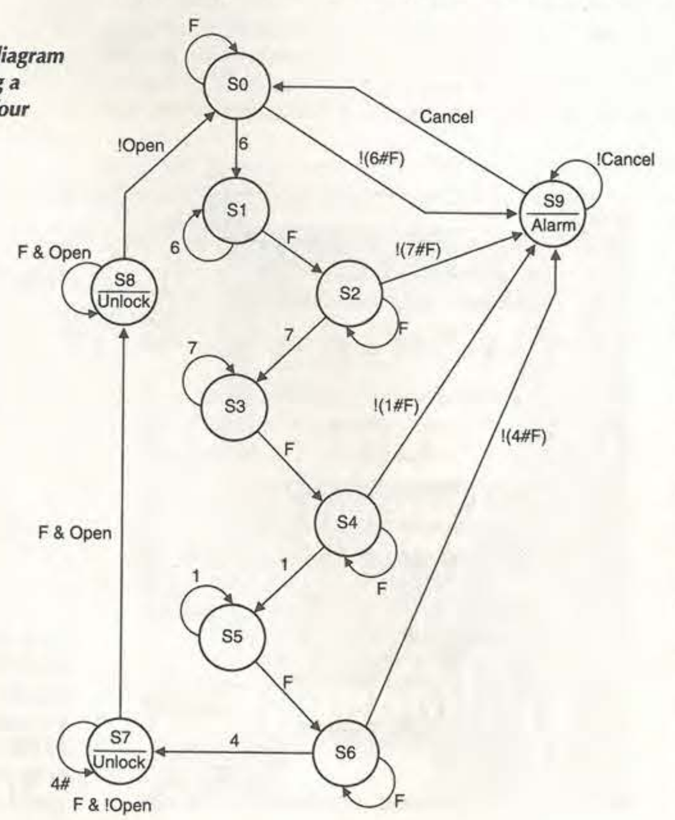


Fig. 2. State diagram for dectecting a sequence of four numbers.

The lock is unlocked by entering the code '6714'; any other code sets off an alarm which can be cancelled with a remote reset. The machine is also reset when it detects that the door has been dosed after a successful entry. Physically, the state register has four bits, outputs Q_{3-0}, with four inputs I_{3-0} for the code entry. Signal DOOR closes the door while ALARM cancels the alarm.

The keypad decoder sends out a digit encoded by I_{3-0} or 'F' if no key is being depressed. System outputs are UNLOCK to enable the door, and SOUND to set the alarm.

First, it is necessary to define the states and the 'input codes, as follows:

S0 = !Q3 & !Q2 & !Q1 & !Q0
S1 = !Q3 & !Q2 & !Q1 & Q0
S2 = !Q3 & !Q2 & Q1 & !Q0
and so on through
S9 = Q3 & !Q2 & !Q1 & Q0

INF = I3 & I2 & I1 & I0
IN7 = !I3 & I2 & I1 & I0
IN6 = !I3 & I2 & I1 & !I0
IN4 = !I3 & I2 & !Il & !I0
IN1 = !I3 & !I2 & !Il &I0

If you are designing for a device with D-type bistable devices, the state transitions must be defined by noting which bits have to be set high for each state transition, as follows:

Q3.D = S7 & INF & !DOOR; transition S7 to S8
# S8 & !DOOR; hold in S8
# SO & !(IN6 # INF); transition SO to S9
# S2 & !(IN7 # INF); transition S2 to S9
# S4 & !(IN1 # INF); transition S4 to S9
# S6 & !(IN4 # INF); transition S6 to S
# S9 & !ALARM; hold in S9
Q2.D = S3 & INF; transition S3 to S4
# S4 & INF; hold in S4
# S4 & IN1; transition S4 to S5, etc.

The outputs are simply defined by:

UNLOCK = S7;
SOUND = S9;

The notation 'Q3.D' implies that this function is applied to the D-input of the bistable device driving the Q_3 output. At the active clock edge, the output will be set high if the function is true.

From the first few lines, it is apparent that equations do not give a transparent view of the function being implemented. However, by considering each transition it is possible to generate a set of equations which, when mapped onto the device, will produce a working part. The logic compiler minimises each equation and, if there are enough product terms driving each bistable device, the mapping will be successful.

The next section shows how state machines can be specified in a way which is directly associated with their function.

Simulating the logic design

The final stage of a logic design is usually a simulation. This has two functions; it checks that the logic will operate as intended, and it produces a set of test vectors for performing functional tests on the device after it has been programmed. Simulation can be defined by means of a text-based entry, or with a truth table, in most logic compilers.

I will show how the door lock can be simulated with an example of each method. The PALASM format would yield the following:

```
SIMULATION; keyword for the simulation
   segment
TRACE_ON I3 I2 I1 I0 DOOR ALARM Q3 Q2 Q1 Q0
   RESET CLK ;defining the signals we wish
   to view on the simulation output. Reset
   is added to initialise the state machine
SETF I3 I2 I1 I0/DOOR/ALARM RESET; initial
   values of the inputs at reset
SETF/RESET; release the reset condition
CLOCKF CLK; clock the state machine
CHECK/Q3/Q2/Q1/Q0; check that it is state
   '0' (an alternative notation is CHECK S0)
SETF/I3 I2 I1/I0 ;put '6' onto the inputs
CLOCKF CLK; clock it again
CHECK S1
CLOCKF CLK
CHECK S1; check that it holds in S1

TRACE_OFF ;end trace at finish of simulation
```

The same test sequence in truth table format would appear as in **Table 1**. Although the truth table requires more typing effort, the result is clearer in terms of the device operation. Also, the output is automatically checked on each line, not just when specified as in the text-based method.

Simulation results may usually be viewed in tabular form or as a waveform display after compilation of the logic equations. Any discrepancies between the expected result of the simulation and the actual result will show up and allow modification of the logic equations to give the fiesired logic function.

Compilation of logic equations, together with the simulation segment will give a programming file with test vectors where this option is possible. Most PAL-type devices will accept programming files with vector testing; most fpgas are either configured in-circuit or programmed on dedicated programmers which do not allow for vector testing. In-circuit testing, using JTAG protocols, is more usual for fpgas.

Logic equation shortcomings

There are two problems with using logic equations in fpga designs. Firstly there is no global standard for the symbols and syntax; in these articles I have standardised on the ABEL symbols, but other standards, such as PALASM are equally valid. This is not an obstacle in itself for these different standards have worked very well for classical pld designs, but it would be very useful to have a universal standard which would be accepted by any fpga design system.

More important is the scale of fpgas compared with plds. A fairly complex pld such as the 22V10, contains over one hundred product terms, each of which can be defined by an equation similar to the decoder example above. To try to design with equations, even at this scale, can make a design extremely difficult to comprehend. Writing down the equations is a time consuming task in itself. But trying to decide where modifications should be made in the event that simulation throws up a mistake, for example, may prove even more arduous.

The architecture of fpgas dictates that logic equations are not the best way to define their logic content. Complex plds have a fixed structure which means that a logic equation maps directly into the AND-array of the logic cell. The only variable is the way in which signals are routed from the i/o lines into the logic blocks. But even this task can be approached in a fairly mechanical way, for there are fixed paths for the signals to travel inside the device.

In an fpga, the structure is much freer. For a start, there is not a fixed two-level AND-OR structure for a one-to-one association with the typical sum of products logic equation.

Table 1. Test sequence in truth-table format for simulating the combination lock.

[I3	I2,	I1	I0,	DOOR,	ALARM,	RESET	CLK]→	[Q3,	Q2	Q1,	Q0]
[1,	1,	1,	1,	0,	0,	1,	0]→	[0,	0,	0,	0]
[1,	1,	1,	1,	0,	0,	0,	C]→	[0,	0,	0,	0]
[0	1,	1,	0,	0,	0,	0,	C]→	[0,	0,	0,	1l
[0,	1.	1,	0,	0,	0,	0,	C]→	[0,	0,	0,	1]

and so on...

This may have to be broken down into a chain of gates in order to achieve the required input connectivit. If the same logic expression is used in two different equations the inference is that it is recreated at each location in the fpga where the overall equations are implemented. It may be more efficient to generate it only once and route it to the second area of the fpga. This will save logic modules but use routing resources.

I have stressed that fpga structures are more like an integrated form of discrete logic. It is most unusual to design pcbs for discrete logic with logic equations. While logic equations are useful and help visibility for pal-type plds, they are not the ideal way to design fpgas.

State machine basics

A state machine is a system which, as its name implies, can exist in a number of stable states. Each state is usually defined by a unique number stored in a set of bistable devices called the state register.

Inputs to a state machine are those signals which can influence the sequence in which the states are entered. There is also a clock to define the time intervals at which the inputs are sampled and the decision made as to whether the state register is changed, and which state should be entered next.

Outputs from a state machine may depend on the state register only, in which case it is called a Moore machine. Alternatively, they may be a logical combination of inputs and state register, when it is known as a Mealy machine.

A state machine may be described by a state diagram; the door lock in Fig. 2 is an example. Each of the ten possible states is represented by a circle labelled with the value of the state register for that state. Transitions between states are represented by arrows labelled with the logic condition which enables that transition.

The 'main sequence' runs vertically downwards from S0 to S8, wrapping around back to S0 when the sequence ends with the door closing again. Transitions to S9 are triggered by incorrect key depressions. Each state has a hold condition which is shown by an arrow wrapped round back to the same state. Thus, keying '6' triggers a jump from S0 to S1 but, if key '6' remains depressed for more clock cycles the state machine remains in S1. Releasing the key sends 'F' to the inputs and triggers the jump to S2, ready for the next key push.

The outputs are decoded directly from the states, making this design a Moore machine.

State equation syntax

Most pld and fpga logic-entry systems allow logic equations and state equations to be mixed in the same design file. The syntax for entering state equations varies from system to system but the commonest methods use if-then-else or case statements. With either, it is mandatory to define the states in terms of the state register elements. This may be done with the following syntax:

```
[Q3, Q2, Q1, Q0]
S0 = 0000b;
S1 = 0001b;
S2 = 0010b;
```

and so on, or by:

```
STATE
S0 = !Q3 & !Q2 & !Q1 & !Q0;
S1 = !Q3 & !Q2 & !Q1 & Q0;
S2 = !Q3 & !Q2 & Q1 & !Q0;
etc.
```

The if-then-else structure may be written as:

```
WHILE [S0]
  IF I3 & I2 & I1 & I0 THEN [S0]
  IF !I3 & I2 & I1 & !I0 THEN [S1]
  ELSE [S9] WITH SOUND
WHILE [S1]
  IF !I3 & I2 & I1 & !I0 THEN [S1]
  IF I3 & I2 & I1 & I0 THEN [S2]
  .
  .
  .
WHILE [S9]
  IF ALARM THEN [S0]
  ELSE [S9] WITH SOUND
```

The else statements, above, have different effects. In the [S0] statement, the 'else' sends the machine to state [S9] if the input is not 'F' or '6'; in the [S9] statement it defines the hold condition.

The 'with' operator defines a combinatorial output. In a Moore machine, as in this case, an output is always associated with the same state; in Mealy machines, output conditions may depend on the path by which a state is reached.

The 'case' construct defines conditions to be tested, and the action to be taken when the condition is true. In the door-lock example, you could define the state machine with the following case statement:

```
CASE (ALARM, DOOR, Q[3..0])
BEGIN
   #h00:          CASE     (I[3..0])
      BEGIN
      #hF:        BEGIN  Q[3..0] = #h0 END
      #h6:        BEGIN  Q[3..0] = #h1 END
      OTHERWISE:         Q[3..0] = #h9 END
      END
   #h01=          CASE (I[3..0])
      BEGIN
      #hF:        BEGIN  Q[3..0] = #h2 END
      #h6:        BEGIN  Q[3..0] = #h1 END
      END
.
.
.
   #h09:          BEGIN  Q[3..0] = #h9 END
   #h29:          BEGIN  Q[3..0] = #h0 END
OTHERWISE:               Q[3..0] = #h0 END
END
```

This illustrates the use of nested case statements. In this example, it is necessary to allow the default jumps to [S9]; if the input conditiom were not nested within each present state 'case', every possible input combination would have to be specified to define the jump to the error state. Nesting, the 'otherwise' operator takes care of defaults as the else operator does in the if-then-else construct. ■

The next article in this series will cover hardware description languages – including VHDL

PWM for small motors

Peter Hale presents two simple but efficient pwm schemes for driving and controlling small motors.

There are two pulse-width modulated drives described here. Each controls the speed and direction of a low voltage dc motor via a single potentiometer.

Pulse width modulation, or pwm, is a method by which a rectangular pulse has its mark to space ratio varied thus controlling the average value of voltage 'seen' by the motor. These pwm drives reverse the polarity of the average value of voltage from zero volts thus enabling a smooth change of direction of the rotor.

Pulse-width modulated drives have

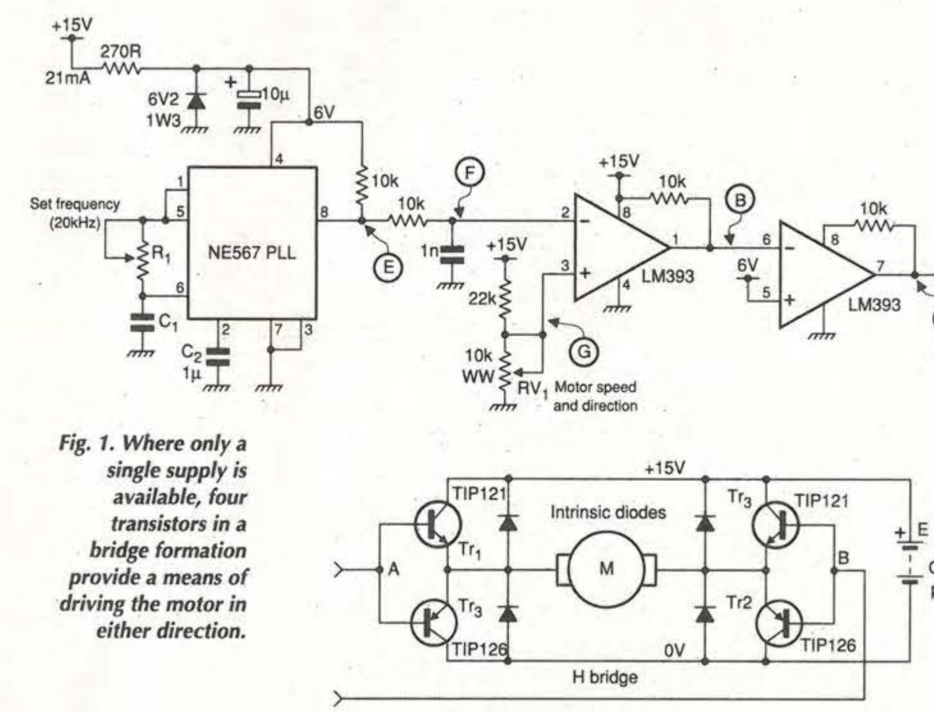


Fig. 1. Where only a single supply is available, four transistors in a bridge formation provide a means of driving the motor in either direction.

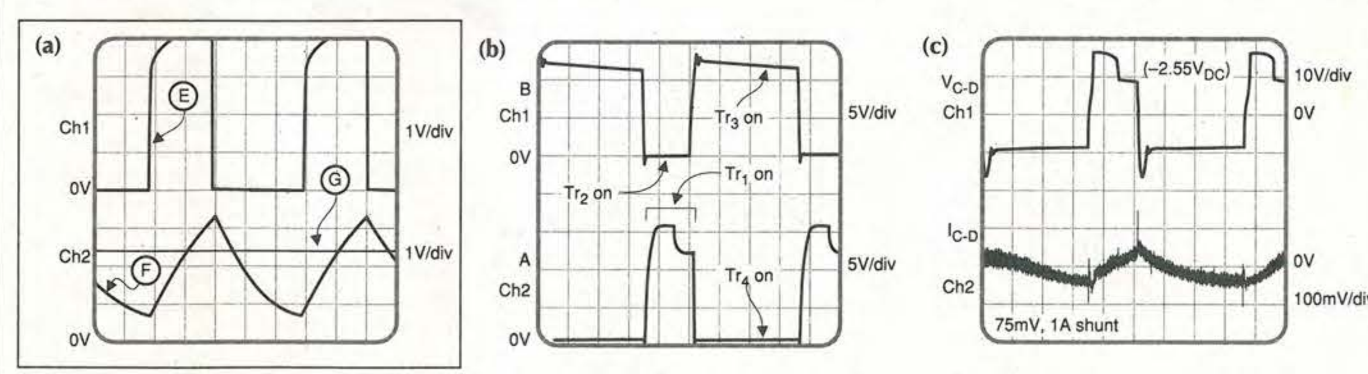


Fig. 2. Waveforms associated with the circuit of Fig. 1. Triangle wave F causes switching as it rises above and below reference G, resulting in pulses at B. Varying the reference causes variations in pulse width. Degradation of waveform A, which is B inverted, is due to using the second comparator in the dual package. Motor speed is 1000rev/min, supply is 15V and horizontal deflection is 10µs/div.

an advantage over linear alternatives in that the heat loss in the transistors is greatly reduced. This is because they are conducting only in the saturation region when V_{ce} will be about 1V or even less.

Both of the drives are bipolar. This means that the instantaneous voltage across the motor reverses polarity – at a rate of 20kHz in this case. But the average voltage appearing across the motor is the net sum of the positive and negative blocks of voltage. Hence the direction and speed of the motor can be precisely controlled.

Both drivers have similar performance. A major consideration in selecting which one is preferable for a given application may be the availability of single or dual power supply rails.

H-bridge using a single rail

This method involves turning switches Tr_1 and Tr_2 on and Tr_3 and Tr_4 off simultaneously, then Tr_1 and Tr_2 off and Tr_3 and Tr_4 on simultaneously, at a rate of 20kHz.

The oscillator generates a rectangular pulse at E, then a low-pass filter integrates to produce waveform F. Mark to space ratio of the driving pulses for the transistor bases at points A and B is set via potentiometer RV_1. Waveform B is an inverted version of waveform A.

Figure 1 shows the circuit diagram and the H-bridge, and **Fig. 2** the associated waveforms.

Using two supply rails

This alternative method has only two switches, Tr_1 and Tr_2, switching at a rate of 20kHz and varying pulse width set by the potentiometer RV_2. Hence the polarity of the voltage across the motor changes at a rate of 20kHz and the motor sees the net sum, i.e. average value, of these two voltage blocks.

Figure 3 shows the circuit diagram and the two-transistor driver while **Fig. 4** shows the associated waveforms. ■

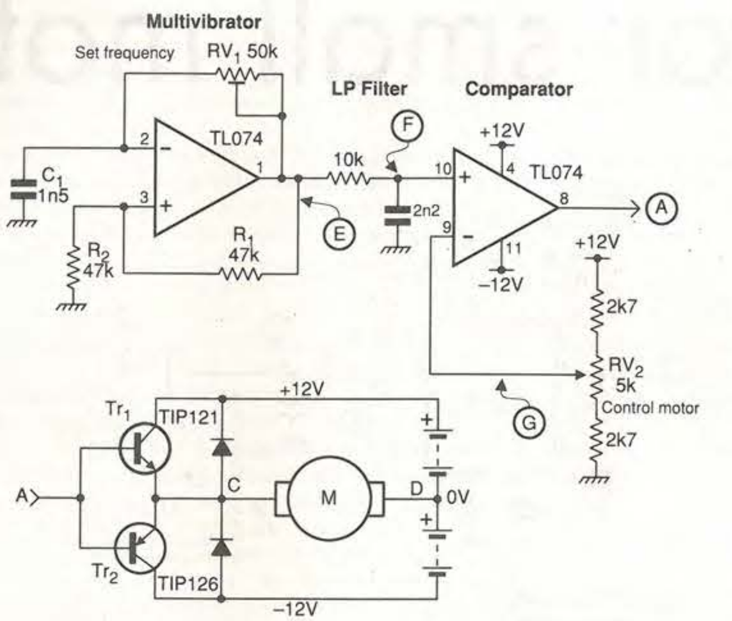


Fig. 3. With a dual supply rail, only two drive transistors are needed to provide forward and reverse motor control.

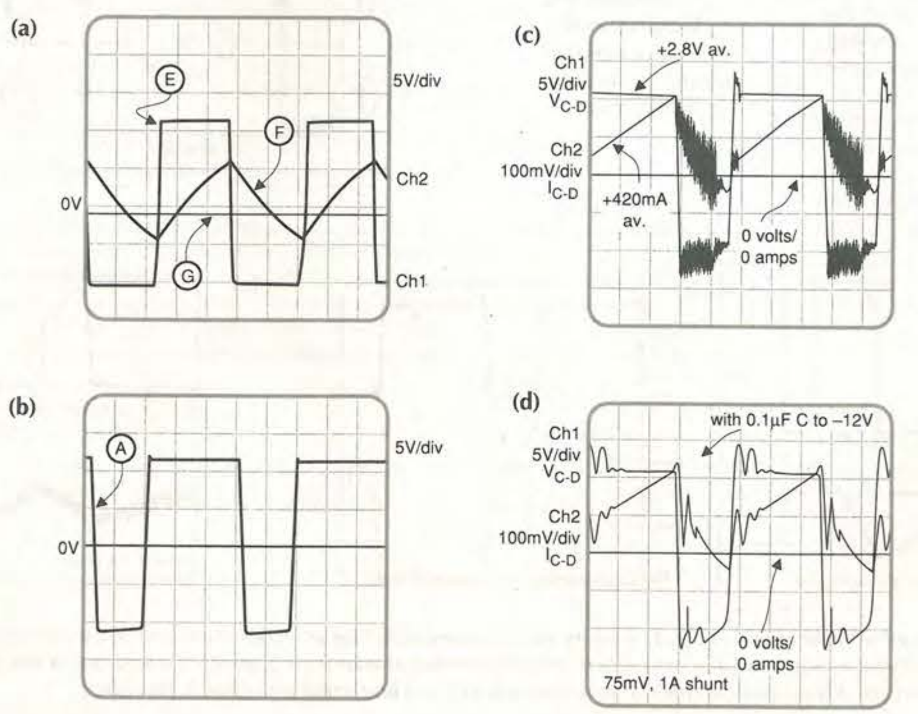


Fig. 4. Waveforms for Fig. 3. Screens a) and b) are control waveforms while those in c) and d) show drive output under load. In d), you can see that adding the 0.1µF capacitor between the negative rail and output significantly reduces the high-frequency noise. Again, motor speed is 1000rev/min and horizontal deflection is 10µs/div.

Hands-on Internet

Cyril Bateman's net discoveries this month include a search tool specifically for scientists and engineers and new simulation software whose demo is perhaps unique in including optimisation.

Based in Norway, FTPSearch[1] is perhaps the best and most popular search engine for locating and transferring software files available on Internet. In early 1997, the FTPSearch statistics page graphs were indicating more than 210,000 searches a day compared to a peak of 65,000 searches one year ago, confirmation indeed of the continued and rapid growth in Internet usage.

As a new surfer two years ago[2], I found two books[3,4] covering the history and development of the Internet most useful in helping me to understand how to look for information. However one consequence of this rapid Internet growth and the resulting changes, is that even the newest Internet books can become dated before publication. Hence the emergence of new, regularly updated, on-line introductions to Internet.

SENN,[5] the Science and Engineering Network News, which is updated monthly, is written by Shari Worthington, a scientist turned engineer, and now editor. It is intended specifically for aiding understanding of the scientific biased Internet resources. While the introductory section on 'Cyber Tools' is designed to help new users navigate the Internet, downloading and printing all three sections will be beneficial even to seasoned users in their searches, **Fig. 1**.

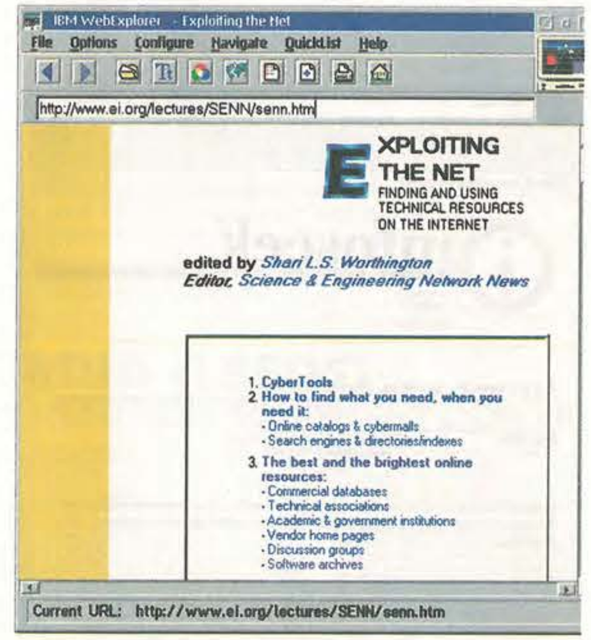


Fig. 1. SENN – an on-line guide dedicated to exploiting Internet Science Archives.

UK-biased search engines?

Being dominated by North American usage and needs, most search engines have a US bias. Recent months however have seen a steady introduction of European or UK biased search tools. UK.Search.Co[6] claims to supply a simple and straight forward Internet access, providing directory information for UK businesses, users home pages as well as Internet search engines, **Fig. 2**.

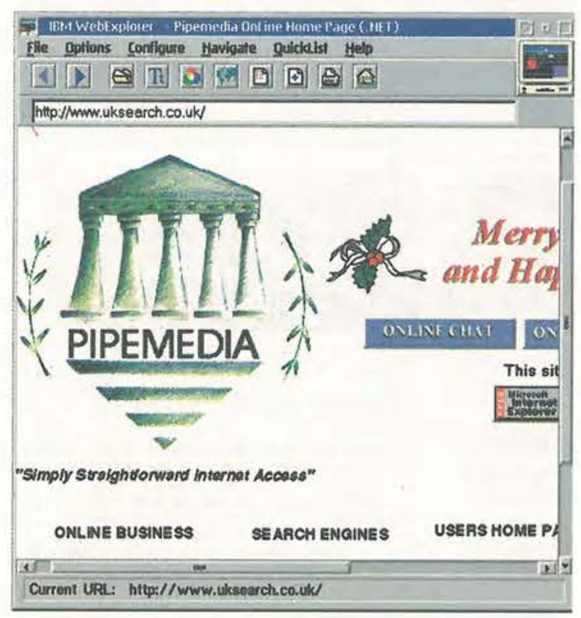


Fig. 2. Searching the Net with a straightforward but UK slant.

Infospace.Com[7] provides similar facilities but with a US bias, having detailed telephone and 'E-Mail' directories for North America, but not Europe. However their access to

Fig. 3. Infospace's Internet innovations. Find E-Mail or Business

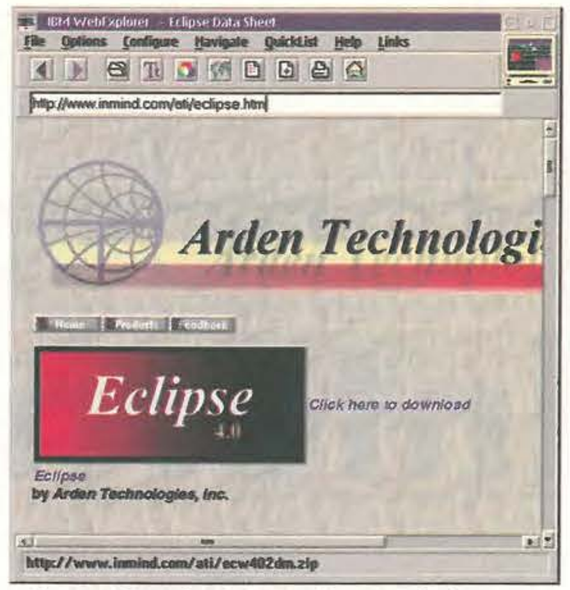


Fig. 6. Arden Technologies Inc. low-cost rf/microwave simulator. If you want to try out optimisation of a circuit, download this demonstration.

Fig. 4. Infoseek.Com Ultra search engine. Rethink of an original and good system.

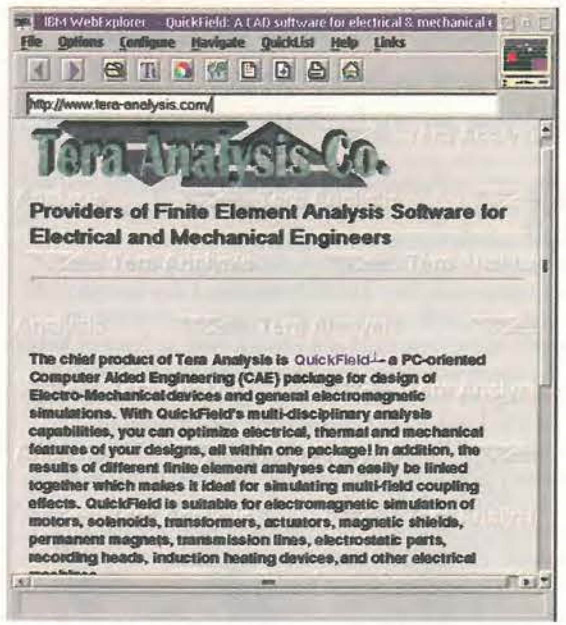


Fig. 7. Tera-Analysis, home of the updated and re-named 'Elcut' software. Simple and low cost finite element analysis for the electronic designer.

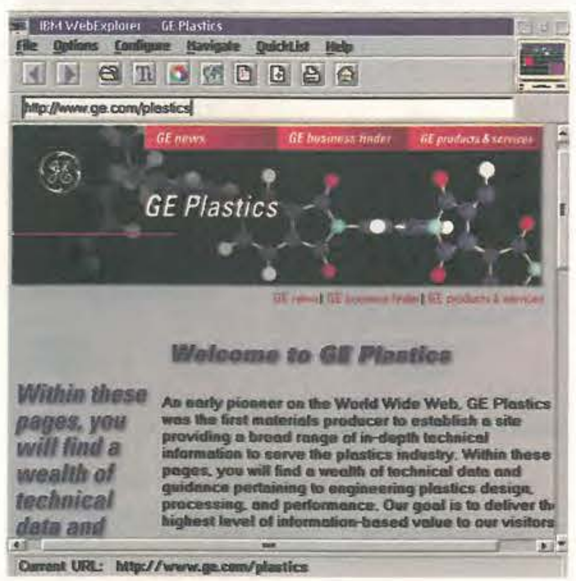


Fig. 5. The GE database of Plastics and Polymers for you to download.

information for USA or Canada is truly first class, **Fig. 3**.

The Infoseek and AltaVista search engines, both featured in the July issue of *Electronics World*, have remained my personal first choices for general Internet searches. Such is the speed of development of Internet needs that Infoseek now has a second generation search tool, Ultra.Infoseek.[8] This tool is claimed to provide the fastest most comprehensive and accurate search engine presently available on the Web.

From recent experience Ultra.Infoseek seems to provide the more acceptable results, if searching for less popular topics, and especially for electronics keywords, **Fig. 4**.

In years gone by, I have spent much time seeking electrical and mechanical data for polymers. GE Plastics[9] has available GE Select, a software database of the company's product range, which can be downloaded for use on a pc or Macintosh. You first have to register a requirement. Then downloading instructions and a registration number will then be sent to you by e-mail, **Fig. 5**.

Simulation software

Searching the software library of the on-line magazine 'SSS', featured last month, revealed an interesting but low cost simulator, 'Eclipse' from Arden Technologies[10]. This 'small signal' simulator, intended for use for rf design, includes the facility to optimise the circuit component values to achieve design targets.

Since components such as capacitors or inductors can also be defined using mathematical expressions and still use optimisation, this simulator could prove useful at lower frequencies when designing bandpass filter circuits. A crippled demonstration version can be downloaded from Arden's homepage, **Fig. 6**.

While electronic circuit simulation is the most popular electronic design calculation short cut, with a variety of software programs now available from many archives and commercial vendors, it is not the only useful design tool for use on a personal computer.

Similar archives of mathematics tools also exist. The Math Archive[11] is a most interesting source of tools for equation solving and plotting also finite element analysis.

While a certain amount of work using electrical/mechanical analogues within Spice is possible assuming suitably specified equivalent models, more general solutions of heat flow, or electrical capacitance, inductance or voltage stress predictions, require use of field plotting techniques or preferably the mechanical engineers 'what-if' tool, finite element analysis.

The useful, low cost and simple to use, finite element simulator, 'Elcut' which I have regularly used to calculate printed circuit board stray capacitance and localised heat gradient behaviour, has been updated, re-named '*Quickfield*'[12] and is now available from Tera-Analysis. A no cost 200 node evaluation version can be downloaded from their page, **Fig. 7**.

The most frequently used circuit simulators are based on a concept developed by the University of California at Berkeley – namely Spice. Enhancements of this package provide amplitude/time displays similar to that of an oscilloscope screen[13]. ■

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10. Arden Technologies.Inc. http://www.inmind.com/ati
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13. http://www.microsim.com

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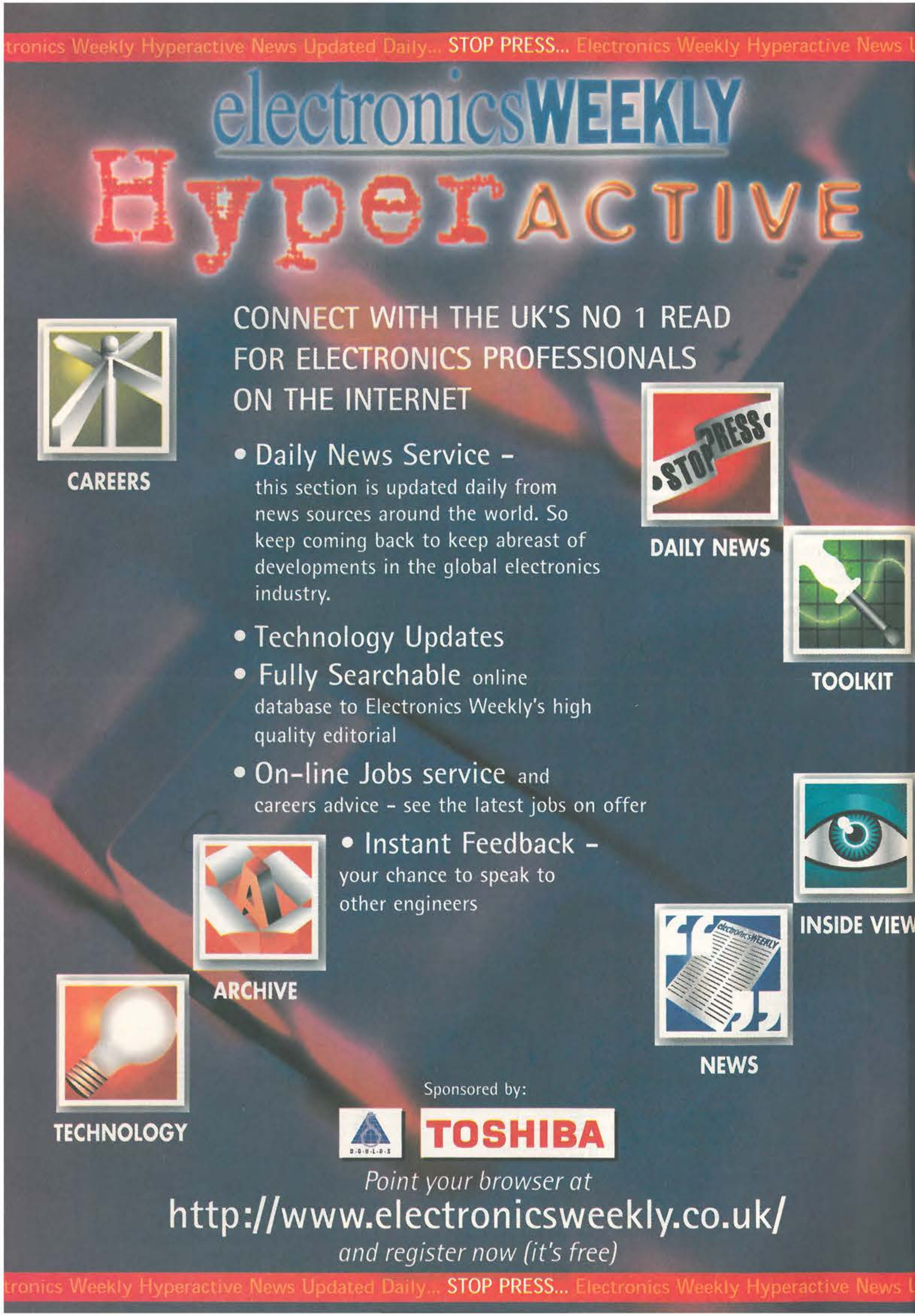

Analysing circuits

Lee Johnston expands on a new circuit analysis technique based on power.

In the article 'Analysing circuits via energy', *Electronics World* October 1995, Andrew Gibson and Bernice Dillon presented a circuit analysis technique based on power. After reading the article and working through the examples shown, I attempted to apply this technique to several simple circuits. The first, a transient response dc circuit shown in **Fig. 1**, introduced a difficulty upon writing the power equation.

According to their technique the power equation would be written as,

$$P(V) = \frac{(V-50)^2}{20} + V \times i_L$$

$$\text{where } i_L = i_o + \frac{1}{L}\int V(t)dt$$

Now the power delivered by the 50V source is a function of V and of time which makes finding the solution considerably more difficult than the standard technique[1] for this type of circuit. As a way around this problem, replace the inductor with a 'resistance' that is defined by,

$$R_L = V/i_L$$

Then setup the power equation as follows,

$$P(V) = (V-50)^2/20 + V^2/R_L$$

and differentiating with respect to V,

$$dP/dV = V-50/10 + 2V/R_L = 0$$

$$V = 50R_L/(R_L+20)$$

After rearranging and substituting the inductor's v–i relationship yields,

$$di/dt + 12500i = 31250.$$

Solving the first-order differential equation gives the transient response,

$$i(t) = 2.5 - 2.5e^{-12500t} \text{ in amps}$$

$$\text{or } V(t) = 50e^{-12500t} \text{ in volts.}$$

These results agree with the standard approach. The second type of circuit I attempted to analyse with this technique involves the use of a dependent current source, and again discovered a problem after writing the power equation, **Fig. 2**.

In this case the power delivered by the dependent current source is a function of node voltage V_1 and V_2 which leads to the power equation,

$$P(V_1V_2) = (V_1-50)^2/2 + V_1^2/4 + (V_1-V_2)^2/5 + V_2^2/20 - V_2(1.7(50-V_1))$$

After following the authors' procedure of taking the partial derivatives, minimising and solving the system of equations leads to a V_1 node voltage of 115.5V and a V_2 node voltage of –130.4V. This result does not agree with the standard node voltage solution of 43.26V and 80.46V respectively[1]. However, this problem can be corrected by replacing the dependent source by a 'resistance' defined by:

$$R_C = -V2/(1.7(50-V_1))$$

Establish the power equation and find the partial derivatives as follows,

$$P(V_1,V_2) = (V_1-50)^2/2 + V_1^2/4 + (V_1-V_2)^2/5 + V_2^2/20 + V_2^2/R_C$$

$$\frac{\partial P}{\partial V_1} = \left(\frac{3}{2}\right)V_1 + \frac{2(V_1-V_2)}{5} - 50 = 0$$

$$\frac{\partial P}{\partial V_2} = \left(\frac{1}{10}\right)V_2 - \frac{2(V_1-V_2)}{5} + \frac{2 \times V_2}{R_C} = 0$$

Substituting the definition of R_C into the equation above and solving the system of equations leads to a V_1 node voltage of 43.26V and a V_2 node voltage of 80.46V which agree with the standard approach. The reason why the first approach does not yield the correct answer is illustrated by examining the power function plot, **Fig. 3**.

Clearly, finding a minimum for this function does not have any physical meaning. This problem also occurs in circuits containing more than one independent source such as the one in **Fig. 4**.

Calculation of the unknown node voltage V may be based on the power delivered by the 20V source or the power absorbed by the 10V

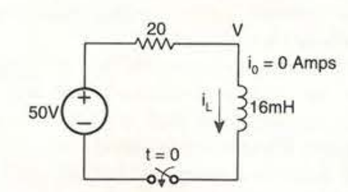

Fig. 1. A transient response dc circuit.

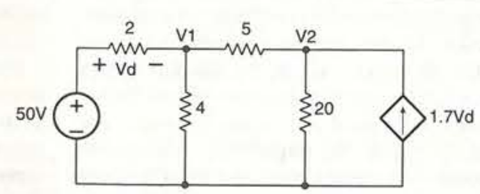

Fig. 2. Dependent current source circuit.

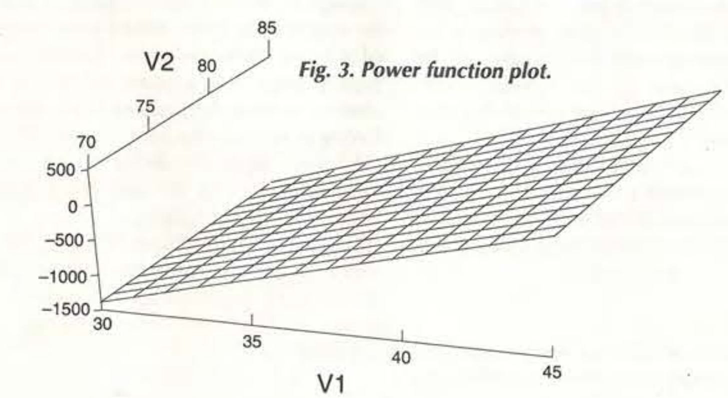

Fig. 3. Power function plot.

source as shown below,

$$P_{10}(V)=(V-10)^2/2+V^2/10+(20-V)^2/4+20((V-20)/4)$$

Minimizing this function gives a node voltage V of 8.82V as can be seen by the plot of the power function, **Fig. 5**.

However, the actual node voltage is 11.765V[1]. This result can be obtained by introducing a 'resistance' defined by

$$R_{20V}=4V/(V-20)$$

Now the power equation is,

$$P(V)=(V-10)^2/2+V^2/10+(20-V)^2/4+400/R_{20V}$$

Minimizing this equation where R_{20V} is treated as a constant leads to the correct value of 11.765V as verified by the node voltage method.

As demonstrated by the above circuits, the authors' approach of minimizing the power function does not always work since the concept of a stationary turning point does not exist in circuits with constraints other than resistive constraints. Therefore, the only way to handle these circuits is to replace the non-resistive elements with a 'resistance' defined by the quotient of its voltage and current. By treating this resistance as constant, just as the real resistances are constant, partial differentiation will lead to the correct gradient and ultimately the correct node voltages. ■

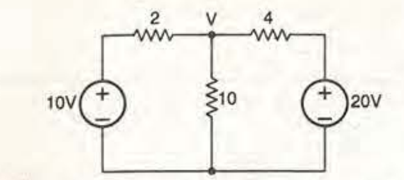

Fig. 4. Circuit containing more than one independent source.

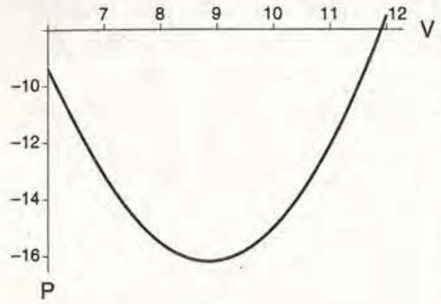

Fig. 5. Power function plot.

Standard analysis techniques

Standard approach to solving the transient response of the first circuit presented:

$$i(t)=i_\infty+(i_0-i_\infty)e^{\frac{-t}{\tau}}$$

where $i_\infty=2.5A$

and $\tau=\dfrac{0.0016}{20}$ sec.

$i(t)=2.5-2.5e^{-12500t}$ amps

Node voltage approach to solving the second circuit presented:

$$V_1/4+(V_1-V_2)/5+(V_1-50)/2=0$$

$$V_2/20+(V_2-V_1)/5-1.7(50-V_1)=0$$

solving this simultaneous system of equations leads to V_1=43.26V and V_2=80.46V

Node voltage approach to the third circuit:

$$V/10+(V-10)/2+(V-20)/4=0$$

$$V=11.765V.$$

BIG Surprises... small packages

contiued from page 374

In response to this problem, Linear Technology has introduced the *LT2078/9, 2178/9* range. These new devices use a thin, approx 50μm thick jelly-like coating, applied before encapsulation, to reduce stress on the top of the die. This results in significantly better V_{os} and V_{os} drift.

Manufacturers also face problems with the marking of these very small parts. The capacitance value is marked on AVC ceramic chip capacitors, for example, in neat clear print. But the print is so tiny it can only be read with the aid of a powerfull eyeglass.

Integrated-circuit designations tend to be quite long, so manufacturers are often obliged to use abbreviated codes to designate a part. For example, the SOT23-5 packaged *NC7S86M5* exclusive OR gate of Fig. 11 is marked simply *'7S86'* on the top, while the similarly packaged *LMC7101BIM5X* op-amp, also from National Semiconductor is marked *A00B*.

Standard sizes?

Figure 10 illustrates another point that you should be aware of when using these devices – watch out for the mechanical dimensions. While the two 44-pin devices illustrated in Figs 10c) and d) look very similar, the pin pitch on the ST44 in c) is 0.8mm, while that on the ST44A in d) is 1mm.

Pin connections are another possible trap. The connections for a single op-amp in the SOT23-5 package shown in Fig. 4a) are the commonest variety, used by a number of manufacturers. But some SOT23-5 op-amps use pin 1 and 3 as inputs, with pin 2 ground, and the output on pin 4.

With today's densely packed boards, multi-layer circuit-board construction is the order of the day. Usually, the inner planes carry power while the signals run on the top and bottom planes. Interconnection between top and bottom planes, often used for mainly horizontal and mainly vertical runs respectively, is by plated-through hole vias. Conections to or between inner layers may be made using 'blind' vias.

Unfortunately, the minimum pitch of conventional plated-through holes is greater than the pitch of the pins on many packages. So adjacent plated-through holes have to be stagered. This takes up more board space, negating some of the advantages of the very small packages.

A more recent development – namely microvias – provides a solution, but at a cost. These vias are so small that they can be located within the land area of each pin's pad, permitting much closer spacing of ICs.

Real benefits

Although more difficult to apply than their larger counterparts, these very small devices benefit the designer in many ways.

For example, two single op-amps in SOT23-5 packages occupy about half the board space of a dual op-amp in an SO-8 pack. Additionally, even more space saving may accrue, due to the greater flexibility afforded by two separate packages. Each can be place exactly where needed, minimizing circuit-board trace lengths.

The problem of needing the odd gate, right across the other side of the board from a bespoke masked logic chip or ASIC containing all the other logic, has already been mentioned. Individual gates and buffers such as that in Fig. 11 clearly supply the answer. But they have another use, no less important.

Single logic elements can be used to buffer the output of an ASIC, found to be over loaded at board evaluation stage. They can even be used to implement a minor last minute logic change, without the cost and delay penalty of having to redesign the ASIC – provided that at the layout stage, the designer took the precaution of leaving the odd spare scrap of board area here and there.

With all their advantages, tiny ICs, both analogue and digital, are destined to play in increasingly important role in today's electronic world, where time-to-market is all important. ■

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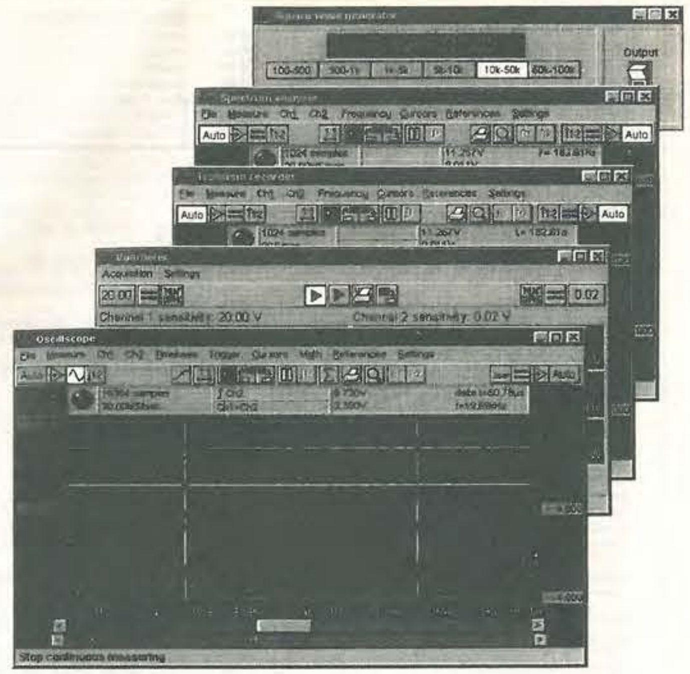

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Foster-Seeley and phase shift

Building on a recent discussion of the Foster-Seeley discriminator, James Diggins provides a more in-depth analysis of how the circuit works.

In the December 1995 issue, Richard Brice presented a discussion of the Foster-Seeley discriminator. It seemed to me that Richard's theme concerned the phase shift through a tuned inductively-coupled transformer, rather than the discriminator itself.

His conclusion that there is not a 90° phase shift between the voltage V_1, across the input capacitor, and the output voltage V_2, is correct; the 90° phase shift is between the input current I_1 and the output voltage. With the aid of a little algebra, I produced the following.

Equivalent circuit **Fig. 1** is the one used by Sturley[1] and Balabanian[2]. The primary and secondary circuits are tuned to the same frequency f_o, by L and C. In accordance with the classical analyses of coupled circuits – and there are plenty of them – I have used Q_k to describe the coupling between the primary and secondary; maximum flatness in the passband occurs when Q_k=1; double humps occur when Q_k>1. Response is narrower and less flat in the passband when Q_k<1.

A little elementary loop analysis allows two equations of interest evolve. At centre frequency f_o,

$$\frac{V_2}{I_1} = -j\left[\frac{M}{\omega_o C^2(R^2+\omega_o^2 M^2)}\right] \quad (1)$$

$$Z_{in} = \frac{V_1}{I_1} = \frac{1}{j\omega_o C} + \left[\frac{M}{\omega_o C^2(R^2+\omega_o^2 M^2)}\right] \quad (2)$$

From equation (1), you can see that output voltage V_2 and input current I_1 are 90° out of phase. Equally clearly, the input impedance is not resistive but consists of the input capacitor C in series with a resistance. Therefore the voltage V_1 is not in phase with I_1, so, V_2 is not 90° out of phase with V_1.

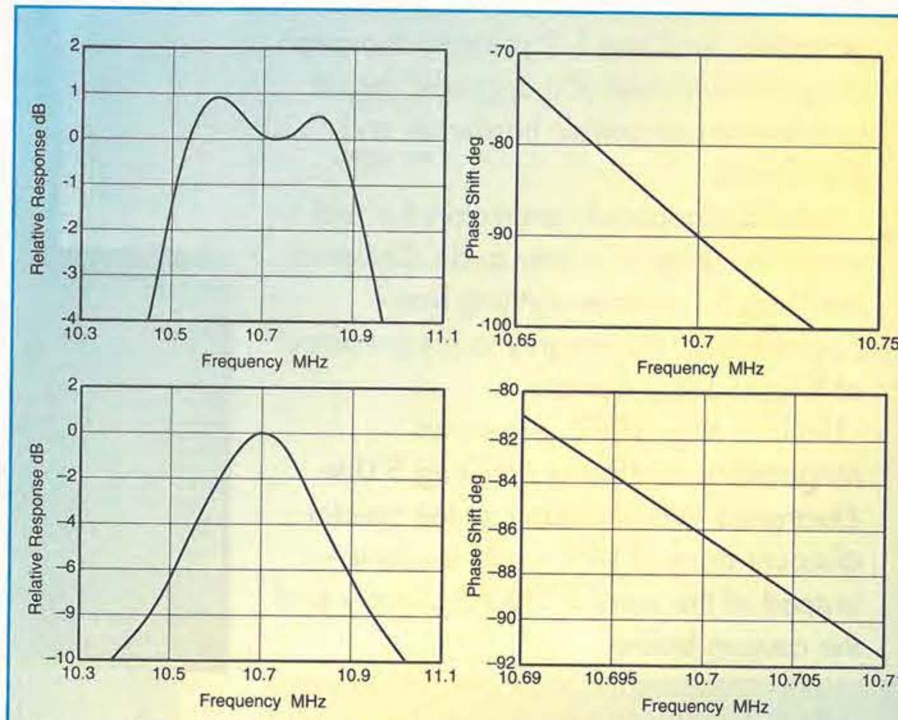
To see how the various characteristics of our circuit behave over a range of frequencies close to resonance, slightly higher mathematics are needed. I set up the transmission matrix shown in the panel for a specific circuit with a 10.7MHz centre frequency, Qs of 50 and 50pF tuning capacitors. The graphs were drawn with the aid of the matrix and graphing facilities of *Mathcad 6*.

I attach no comments other than to say that for any practical value of Q_k that I used, the 90° phase shift holds. I chose Q_k=1.5 just to see the double humps in the frequency response.

Finally, we are left with the question, 'if the phase shift between v_1 and v_2 is not 90° at f_o, then how does the Foster Seeley discriminator work? And does it matter?

References

1. Sturley, K. R., "Radio Receiver Design", Chapman & Hall, 1949.
2. Balabanian, Norman, 'Electric Circuits', McGraw- Hill, Inc., 1994.



Foster-Seeley response and phase-angle curves near resonance for V_2/I_{in}, top, and for V_2/V_1, bottom.

Foster-Seeley transmission matrix

Transmission matrix *m*(f), for Fig. 1 is the product of the component matrices:

$$m(f) = \begin{bmatrix}1 & 0\\ j(2\pi f)C & 1\end{bmatrix}\begin{pmatrix}1 & R\\ 0 & 1\end{pmatrix}\begin{bmatrix}1 & j(2\pi f)(L\pm m)\\ 0 & 1\end{bmatrix}\times\ldots$$

$$\begin{bmatrix}1 & 0\\ \frac{1}{j(2\pi f)m} & 1\end{bmatrix}\begin{bmatrix}1 & j(2\pi f)(L\pm m)\\ 0 & 1\end{bmatrix}\begin{pmatrix}1 & R\\ 0 & 1\end{pmatrix}\begin{bmatrix}1 & 0\\ j(2\pi f)C & 1\end{bmatrix}$$

Mathcad 6 evaluates this product transparently; you do not have to print it out, which is lucky, because it would print out over three pages. Using *Mathcad* Subscript, the elements of *m*(f) can be extracted and inset into the transmission equations to draw the graphs.

f_o = 10.7MHz
Q = 50
Q_k = 1.5
C = 50pF
$L = \frac{1}{\omega_o^2 C}$
$R = \frac{\omega_o L}{Q}$
$M = Q_k\frac{L}{Q}$

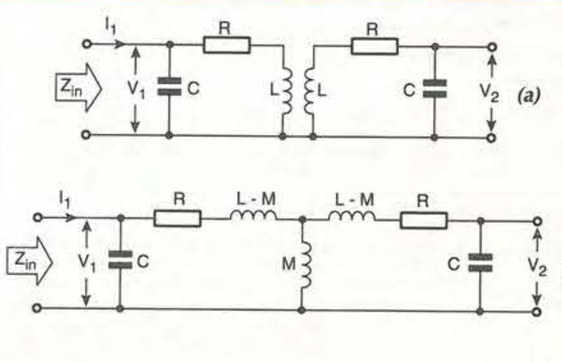


Fig. 1a). Tuned, inductively-coupled transformer, and b), its equivalent circuit.

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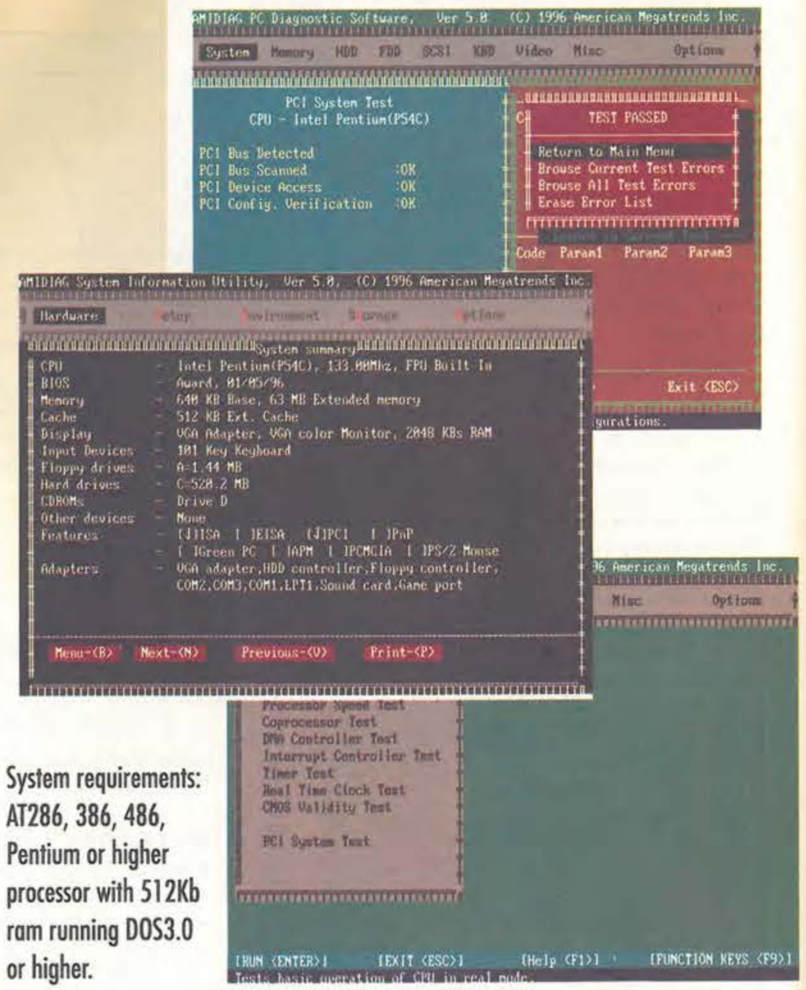

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Enhanced 5V regulator

UK manufacturer Zetex has produced a 5V regulator featuring improved supply rejection and a 350µA quiescent current figure.

The *ZSAT500* is a three terminal 5V regulator with enhanced electrical performance. Despite using a die small enough to fit into TO92 and SOT223 packages, the regulator can supply loads of up to 200mA, frequently allowing it to be used where a TO220 or TO139 part was necessary.

Its quiescent current is typically only 350µA and this current changes little with input voltage or load current. These features allow the regulator to be used in constant current generator circuits or circuits where the output voltage must be trimmed away from standard voltages using external resistors.

Device description

The ZSAT500 was originally designed for use in Satellite receiver low-noise blocks, where a high degree of supply rejection is required to extend frequencies. The device has been improved to give superior performance with ripple rejection of 65dB up to 22kHz, and 40dB up to 200kHz.

Current limit and thermal shutdown circuitry is built in. The *ZSAT500* shows performance characteristics superior to other local voltage regulators. Initial output voltage is maintained to within 2.5% with a quiescent current of typically 350µA. Line and load regulation is superior to that of other devices, with load current up to a maximum 200mA.

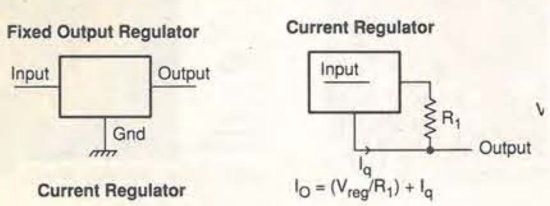
For the lnb application, the regulator is available in surface mount SOT223 packaging – which permits power dissipation up to 3W. Additionally the device can be made available in SO8 surface mount packaging, as well as TO92 for through hole application.

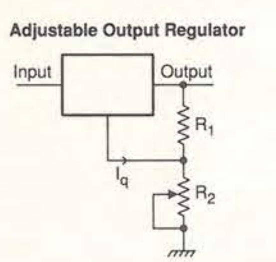
ZSAT500 electrical characteristics

Parameter	Conditions	Min	Typ	Max	Units
Output voltage	I_o=1 to 200mA	4.8		5.2	V
	T_j=–55 to125°C	4.8		5.2	V
Line regulation	V_{in}=7 to 20V		10	40	mV
Load regulation	I_o=1 to 200mA		5	25	mV
	I_o=1 to 100mA		2		mV
Quiescent current	T_j=–55 to 125°C		350	600	µA
Quiescent current change	I_o=1 to 200mA			50	µA
	V_{in}=7 to 20V			100	µA
Output noise	f=10Hz to 10kHz		75		µV rms
Ripple rejection	V_{in}=8 to 18V f=120Hz-22kHz	65			dB
Ripple rejection	V_{in}=8 to18V f=200kHz	40			dB
Min. input to maintain regulation		7	6.7		V
Ave. output temperature coeff.	I_O=5mA T_j=–55 to 125°C		0.1		mV/°C
coeff.	I_{Gf}=500mA			2	V

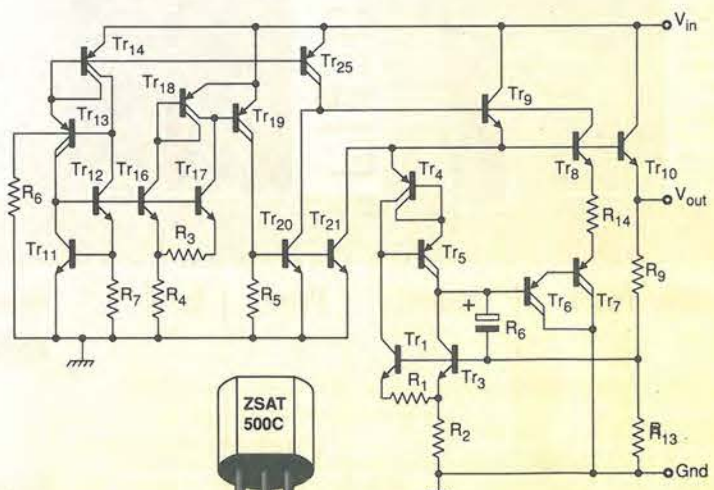
Test conditions, unless stated, are T_j=25°C, I_o=100mA and V_{in}=9V.

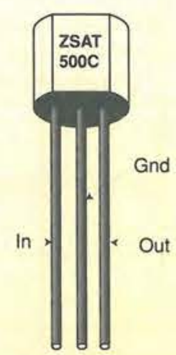
The devices are suited to local voltage regulation applications, where problems could be encountered with distributed single source regulation, as well as more general voltage regulation applications. The device operates over a wide temperature range of –55 to 125°C and needs no external components.





Three typical application circuits for the ZSAT500 regulator. Note the absence of the usual input decoupling capacitor.





Simplified internals of the ZSAT500 enhanced 5V regulator.

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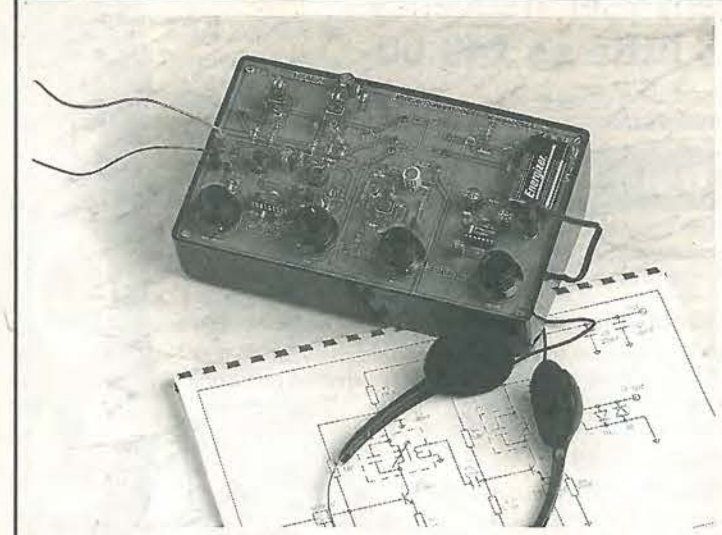

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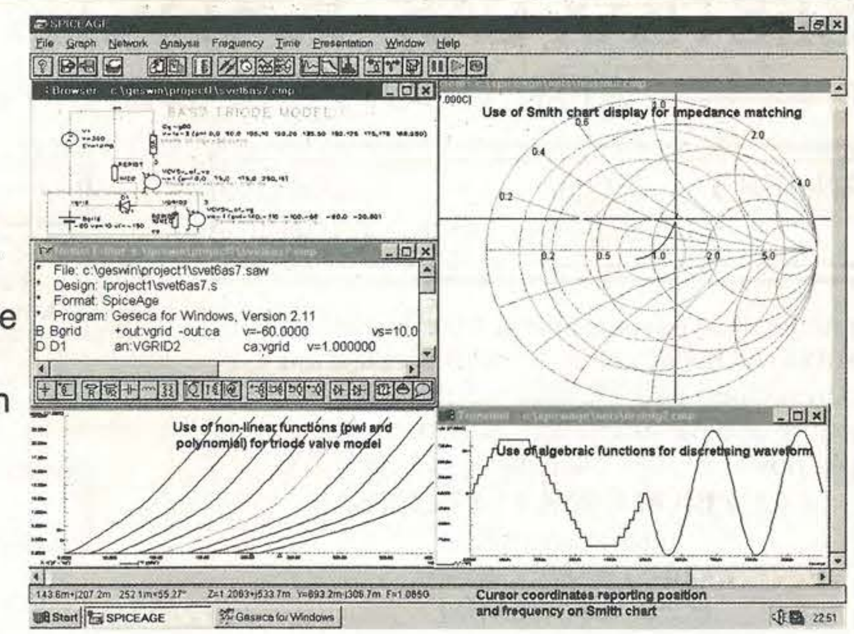

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LETTERS

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

Perfect amplifier? So what?

Having been a reader on and off since 1961, I can say that generally, things have changed significantly. But one thing has remained the same – the ambition of of designers to produce the perfect amplifier.

Had we a perfect means of reproducing the output from their blameless amplifiers – a loudspeaker with no distortion – I would feel that their efforts were less wasted. Surely any distortion is additive?

Robin Froud
Maidstone
Kent

Cable at audio frequencies

I intended to keep out of this wrangle, but in the April issue letters column, John Watkinson seems to have repeated an over-simplification made elsewhere in print. I also feel that Cyril Bateman has not followed it up fully.

Firstly, assuming that what you want to do is to have the loudspeaker sound as the designer intended, then you should use the same cable as the designer used, whatever it was, because he will, if competent, have incorporated any cable effects into the design and his assessment of sound quality. This is not to discount JW's remarks about active loudspeakers, with which I agree.

Secondly, Cyril correctly states the formula for the characteristic impedance of a cable, and this can be evaluated at any frequency. John is over-simplifying when he says that 'loudspeaker cables do not have a characteristic impedance at low frequencies'. It would be more precise to say 'characteristic impedance is not a useful metric of loudspeaker cables'.

The point is that, at low frequencies, the R and ωC terms predominate, and the characteristic impedance is purely imaginary. For PVC insulated cable, G is a function of frequency – even over the audio range. It is often not negligible – even at 10kHz. This introduces a real – as opposed to imaginary – component into the characteristic impedance. At *much* higher frequencies, unless the conductors are very large in diameter, skin effect begins to affect R, but at these frequencies, ωL becomes significant too, and the characteristic impedance tends to become independent of frequency and largely resistive.

I measured some screened balanced cables (not loudspeaker cables) a while ago, and found these results:

Calculated high-frequency characteristic impedances of the cables under test are:

Type M: 143Ω
Type K: 102Ω
In contrast, the values at 1592Hz are:

Type M: j694Ω
Type K: j571Ω

Some of the effects reported are clearly due to the use of test signals whose spectra extend far beyond any reasonable upper limit of the audible range. I would like to see the spectra of the input signals included in future articles on subject.

John Woodgate
Rayleigh
Essex

Shortage of engineers?

According to your report in Update, March 1997 issue, Dr Mary Harris tells us there is "a shortage of engineers".I cannot believe she can be referring to our profession.

Although degree qualified myself, and having design experience extending from television standards converters to switch-mode power supplies, several years ago I, in common with many others, was forced to join the ranks of the reluctant self-employed – and usually unemployed. Why is this I wonder?

A glance at the job advertisements in the trade press might give a clue. Most reveal a narrow, project-oriented focus in which familiarity with a particular brand of CAD software seems more important than one's track record. Some advertisements are even explicit about what sort of employer the applicant is expected to be working for currently – surely a blatant attempt to woo talent from competitors?

Perhaps what we need is not more engineers but more employers able to see beyond the horizon of their current projects, to understand that deftness with a particular software package does not a good designer make, and to recognise that in-house training on the job is a better way of spending cash than poaching talent from a competitor.

We reap what we sow.

E R Lisle

Q & A

Simulating mosfet amplifiers

Can anyone advise me which circuit simulator can realistically simulate class AB crossover distortion?

In the February 1996 issue, Marcel Van de Gevel found his version of PSpice did not appear to model weak inversion – the region where crossover occurs in class AB amplifiers. Another attempt using *Electronics Workbench*, in Circuit Ideas, July/August 1996, did not give a realistic result.

The manual states that mosfets are modelled as square-law devices at low currents – no mobility modulation (velocity saturation) and no weak inversion effect. Is the latest *Electronics Workbench* Version 5 capable of realistic results?

I was hoping to simulate my square-law output stage (September 1995) using *2SJ49/2SK134*s or equivalents but I have not got very far. Will level-three mosfet models do the job?

Ian Hegglun *Australia*

Designing reliable linear power supplies

Rather belatedly, I write to advise anyone approaching this subject for the first time to be cautious when applying Ray Fautley's design procedure (Sept. 1996 issue) to the design of low voltage, high-current supplies, such as the 13.5V 10A bridge rectifier example given.

In my opinion, the procedure, although essentially correct, contains two flaws. These lead to significant errors in transformer sizing, output voltage and diode specification when designing this type of supply.

The first, which affects calculation of R_S/R_L and surge current, is in the derivation of R_{rec}. This is given as $V_{rec}/I_{average}$ when it should really be V_{rec}/I_{peak} where I_{peak} is peak flow through diodes during conduction period.

There is a chicken and egg situation here of course, but assuming a transformer regulation in the normal range of 5 to 15% and a reservoir capacitor selected to give $V_{ripple(rms)}$ about 3 to 5% of V_{dc}, a reasonable estimate of I_{peak} for a full-wave rectifier is 3 to $4I_{dc}$. Using $4I_{dc}$ in the example makes R_{rec}0.9V/40A=0.0225Ω. At these current levels it would probably be nearer 1.1V/40A=0.0275Ω, but the important point is that it is much lower than the 0.18Ω obtained by using $I_{average}$ and that even $2R_{rec}$, does not dominate source resistance.

The corollary is that rectifier surge current at switch on will be much higher than 59A.

The second error is in how allowances are made for diode voltage drops. In step 6 of the example, E_{dc} is given as $E_{dc(load)}+2V_{rec}$=15.3V and R_L=1.53Ω. In other words the combined voltage drop of conducting diodes has been allowed for by adding it to the output voltage. The ac supply can then be regarded as providing a dc supply of 15.3V through perfect rectifiers, and the source impedance becomes simply the transformer impedance ref. secondary, assumed resistive.

This is a valid approach and R_S/R_L calculated on this basis will give good results when used in the tables. Calculating R_S/R_L by also adding $2R_{rec}$ into the source resistance as in step 8 surely cannot be correct, because allowance for diode voltage drops has then been added in twice – their I_R drops on the source side and $2V_{rec}$ on the output side.

If $2R_{rec}$ is added to source resistance then true R_{load} of 1.35Ω should be used to calculate R_S/R_L. Reworking the example using a transformer impedance of 0.0675Ω (5% of 1.35Ω) and assuming V_{rec} of 1V and I_{peak} of $4I_{dc}$=40A, (R_{rec}=0.025Ω) gives the following results.

First method;

R_S/R_L=0.0675/1.55=4.4%,
X=15.5, C=36600µF
(X=14 for C =33,000µF used in tables)
Y=0.85, V_{rms}=15.5/0.85√2=12.9V,
Z=2.56, $I_{rms(diode)}$=2.56×5=12.8A,
$I_{rms(transformer)}$=12.8√2=18.1A,
W=7.81,
I_{peak}=5×7.81=39A.

Second method;

R_S/R_L={0.0675+2(0.025)}/1.35
=0.1175/1.35=8.7%,
X=14.8, C=34900µF,
(X=14 for C=33000µF used in tables)

Y=0.76, V_{rms}=13.5/0.76√2=12.6V,
Z=2.35, $I_{rms(diode)}$=2.35×5=11.75A,
$I_{rms(transformer)}$=11.75√2=16.6A,
W=6.78, I_{peak}=6.78×5=33.9A.

Estimated surge current at switch on is 152A. Results are in fair agreement, with the first method, probably over estimating rms and peak transformer currents. They suggest that a transformer with a no load secondary voltage of 12.9Vrms and current rating of 17A rms would be suitable, e.g. a standard 12V transformer of 7% regulation.

To verify this I built a 5A version of the supply using components to hand; a 240V/12V, 8A transformer with a Z ref. secondary of 0.125Ω (hot), open-circuit voltage 13.2V, full-load voltage 12.2V (hot), a 35A I_{AV}, 400A I_{FSM} bridge rectifier and a 10 000µF, 11A capacitor. Load was 2.75Ω, and the following results were obtained at a mains voltage of 244V rms:

DC output 14V at 5.1A
Ripple 3V pk-pk, approx. 1V rms,
Bridge rectifier volt drop 1.7V

A 15000µF capacitor would have reduced ripple to approx. 0.66V rms. The best estimate of peak current flow, using oscilloscope to determine difference between peak secondary voltage at no load and full load was 15A.

These results show that the 18.34V rms, 14.14A rms secondary specification of the September example transformer is clearly wrong – particularly with respect to voltage. Assuming 18.34V is the open-circuit value and that ref. secondary impedance is 0.0765Ω, (5% of 1.53Ω) this transformer would give a dc output of about 20V at 10A, although its rms current rating would then be exceeded at around 18A. It would safely support a dc output of about 21V at 7.5A.

A dc supply of 13.5V at 10A could be obtained, but only by fitting a resistor of 0.3Ω permanently between transformer and bridge or between bridge and capacitor. Output regulation would then be poor and the resistor would dissipate 60W of heat at 10A dc output.

In calculating diode surge current I_{FSM} at switch on, a gross but safe over estimate can obviously be made by assuming it to be limited purely by transformer impedance. In practice adding $2R_{rec}$ – as calculated using I_{peak} – to transformer impedance gives good results, because although rectifier forward resistance drops to an even lower figure at surge current levels – for example 0.01 to 0.015Ω for a *12F10* diode – capacitor esr has been ignored. The net result is a close approximation to the real situation.

Recalculating surge current for the transformer as specified in the September design, assuming a diode drop of 1V and I_{peak} of 40A, gives,

$$I_{on}=25.9/\{0.0765+2(0.025)\}=205\text{A}.$$

These comments are offered as constructive criticism, and in the hope that they will be seen as a useful modification to Ray's design procedure. I congratulate him on his hard work and excellent article. It is precisely because the tabular data he has produced is so useful, and his explanation of the design process so clear, that I feel the need to write.

Finally, for those interested in this subject, can I also recommend the article 'Simplified design of DC power supplies' by JCS Richards, published in *Wireless World* August 1981. This article adopts a different approach, but using it along with a basic scientific calculator, also enables an accurate dc supply to be quickly designed.

Ralph A E Goold
St Albans
Hertfordshire

No-contact current measurement

In Steve Winder's article 'No-contact current measurement' in the November 1996 issue, the differential gain-setting resistor between the two non-inverting inputs on Fig. 7 was incorrectly marked as 10kΩ. It should have been marked 1kΩ – apologies.

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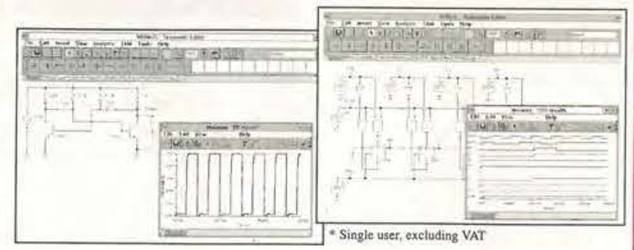

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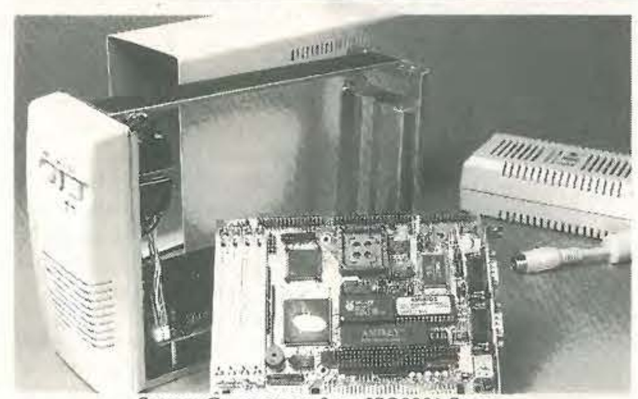

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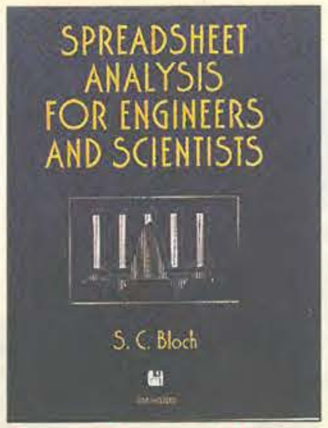

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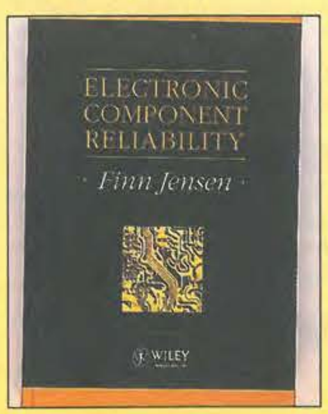

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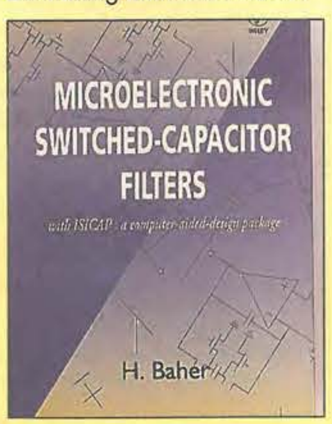

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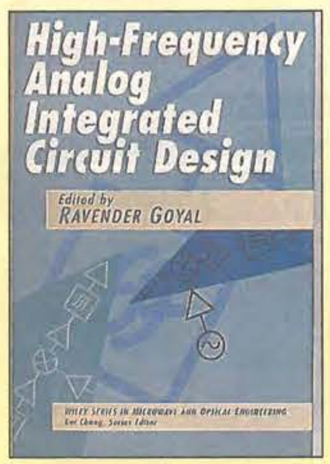

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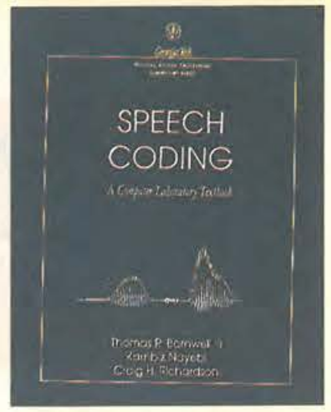

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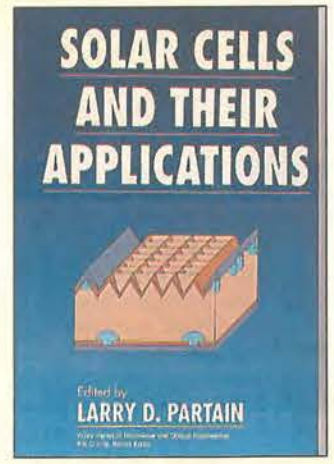

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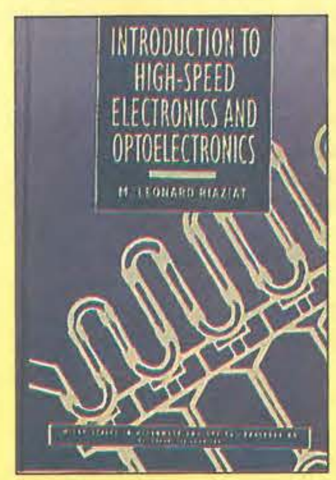

text/reference examines the complementary nature of electronics and optics and emphasizes high-speed technology in which the two fields are less differentiated. Beginning with an overview that develops a perspective and appreciation of analog high-speed technology in general, the book goes on to cover devices and circuits used at microwave and millimeter-wave frequencies, optical components, and opto-electronic integrated circuits and subsystems. Particular attention is paid to applications in the area of high levels of interest in this area and because many of the concepts are applicable in other fields. The book concludes with important coverage of the often-overlooked area of measurement and characterization of high-speed devices. Fully referenced and supplemented with hundreds of helpful illustrations, *Introduction to High-Speed Electronics and Optoelectronics* is equally useful as a professional reference or a textbook for senior undergraduate and first-year graduate courses.

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Birth of broadcasting

Who is the father of broadcasting? Pat Hawker investigates.

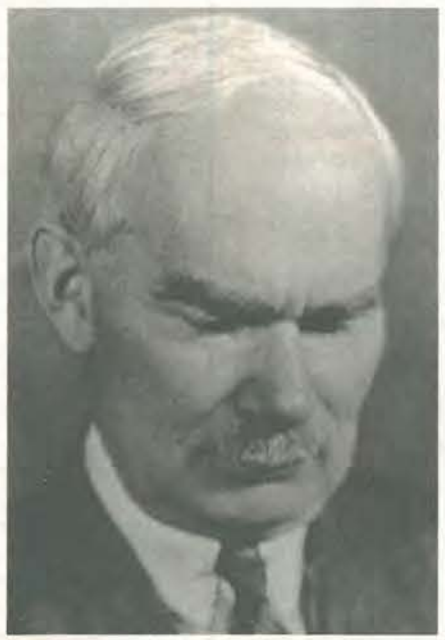
Many sources in the UK and USA credit KDKA at Pittsburgh as being the world's first broadcasting station in November 1920. Alternative claims are made on behalf of WWJ Detroit; other historians point to the 1906 transmissions from Brant Rock made by Canadian-born Professor Reginald Fessenden. Still others proclaim Lee de Forest as the father of radio. He regularly transmitted entertainment programmes from about 1907 onwards, and seems to have been the first person to apply the term 'broadcasting' to radio transmissions directed at the public.

De Forest in 1908 is also credited with having brought radio-telephony equipment to Europe where he transmitted music from the Eiffel Tower. In 1909 in New York he transmitted an appeal on behalf of women's suffrage by Hariot Stanton Blatch whose granddaughter he had married. This was claimed as the first use of radio for propaganda.

In 1910 De Forest broadcast Eurico Caruso from the New York Metropolitan Opera House, another first. But beset by financial problems, his broadcasts ceased in 1911. They were not resumed until 1916 when money received from AT&T, in respect of his valve patents, let him resume his pastime of broadcasting to the many amateurs in New York.

His motives were not entirely altruistic. By November 1916 he was transmitting music nightly from his factory in the Bronx, interspersing Columbia Gramophone records with sales messages lauding his radio apparatus. During that year's presidential elections, de Forest broadcast news items culled from the papers and, as a convinced Republican and optimist, wrongly announced before midnight local time (before the West Coast votes had been counted) that the Republican candidate Charles Hughes had defeated Woodrow Wilson!

The American entry into World War 1 in 1917 brought about another hiatus in de



Lee de Forest, from James Hijiya's book, Lee de Forest and the Fatherhood of Radio.

Forest's broadcast programmes but these were resumed in 1919 from Manhattan. But the following year he was ordered by the US authorities to cease broadcasting on the grounds that he was interfering with navy and commercial radio transmissions, adding, "there is no room in the ether for entertainment".

De Forest died in 1961 penniless, his work often discredited by engineers who believed his many, largely successful, patent actions against Armstrong and others were unjustified. Some also believed that his most famous invention – the triode valve – was brought about less from insight than from his need to circumvent the Fleming diode patent.

Yet, if one excepts the two isolated transmissions by Fessenden on Christmas Eve, 1906, and the following week, New Year's Eve, there can be little question that de Forest has many claims to be recognised as 'a' if not 'the' major pioneer of radio broadcasting.

Entertainment via telephone

The idea of providing entertainment in the home via telephone wires had been implemented in many cities including London in the late 19th century. Almost as soon as radio telephony became practical, visionaries foresaw its possibilities. In 1904 that strange scientist, Nikola Tesla, wrote of a possible device, "which will be very efficient in enlightening the masses... a cheap and simple device which might be carried in one's pocket".

The astute and ambitious young David Sarnoff, while assistant manager of American Marconi, in 1915 wrote an internal memorandum which foresaw the commercial possibilities of a mass market for a 'magic radio box':

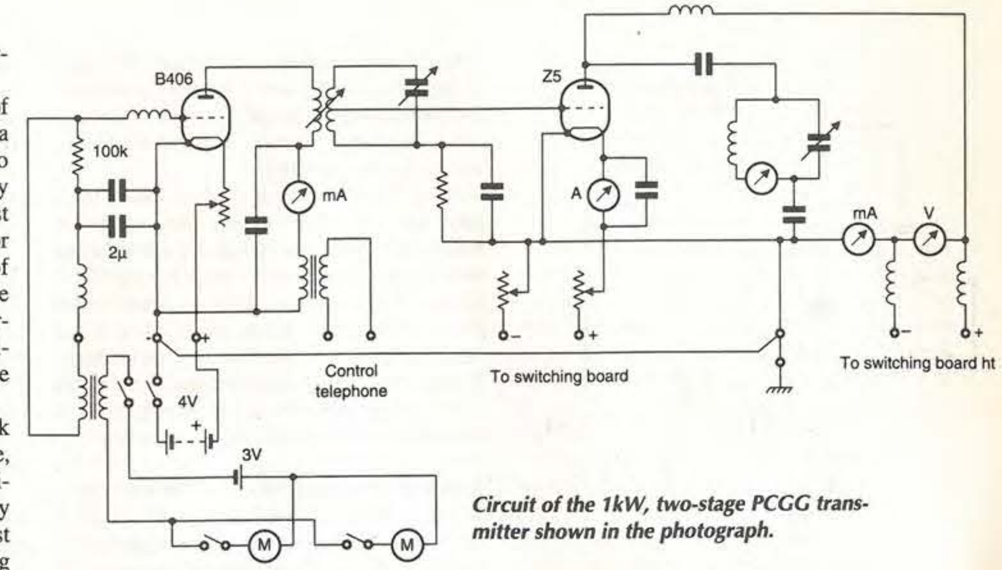
"I have in mind a plan of development which would make radio a household utility in the same sense as the piano or phonograph. The idea is to bring music into the home by wireless. While this has been tried in the past by wires, it has been a failure because wires do not lend themselves to this scheme. With radio, however, it would be entirely feasible...

"The receiver can be designed in the form of a simple radio music box and arranged for several wavelengths, which should be changeable with the throwing of a single switch or pressing of a single button... amplifying valves and a loudspeaking telephone... The manufacture of the radio music box in large quantities would make possible their sale at perhaps $75 per outfit... if manufactured in quantities of 100,000 or so could yield a handsome profit."

Vice President Nally to whom the memo was addressed was apparently not impressed; the memo was simply filed and largely forgotten.

Fessenden whose role in the development of continuous-wave radio has been undergoing a major reappraisal in recent years made two major breakthroughs which opened the way for radio broadcasting. Not only did his first high-frequency 70kHz alternator, made for him by Alexanderson, provide a source of continuous waves rather than spark, but he also developed one of the first detectors (thermal barretter) which by acting as an rf rectifier was suitable for demodulating amplitude modulated, continuous-wave signals.

Soon his work was augmented by the work of others leading to the rapid spread of simple, cat's whisker crystal detectors making possible low-cost receivers that could be built by amateur enthusiasts – providing for the first time the audience on which broadcasting depends.



Circuit of the 1kW, two-stage PCGG transmitter shown in the photograph.

And in the US...

The use of the early radio-telephony transmitters to provide music and speech transmissions for general reception rather than for communication to a specific receiving station soon spread from the USA to other countries. Remember that in the USA, radio transmission by amateurs as well as professional engineers remained unregulated until 1912, whereas in the UK the first Wireless Telegraphy Act was on the statute books in 1904.

It was not until the one-off Marconi Company broadcast of 15 June 1920 by Dame Nellie Melba that it can be claimed that the UK was beginning, if hesitantly, to enter an age of radio broadcasting to the public only to be further delayed. This delay was caused by an unsympathetic Post Office which withdrew the Marconi licence on the ground of "interference with legitimate services".

In Europe, a strong claim for giving birth to the first 'broadcasting' service can be made on behalf of Raymond Braillard and other Belgian enthusiasts who set up a transmitter OTL in the grounds of the Royal Castle at Laeken and broadcast programmes of music for public reception every Saturday at 5pm from 28 March 1914. Braillard was later destined to play an important role in the International Broadcasting Union.

A report in *Le Soir* on 30 March 1914 stressed the intention to transmit regular programmes for the general public (enthusiasts): "To meet the requests from certain radio amateurs who occasionally picked up our irregular experimental transmissions, we decided to devote a special session to them every Saturday at five o'clock."

It has been claimed that soon there were several hundred listeners to these broadcasts in Belgium and northern France. They continued until shortly before the German army entered Brussels when the transmitter was hurriedly dismantled.

Due to unfortunate circumstances, the work of neither de Forest in the USA nor the Belgian enthusiasts – although both undoubtedly represented embryonic radio broadcasting services (unlike the 1906 experimental transmissions of Fessenden) – did not continue unbroken into the true age of radio

This is not the case with the Dutchman Hanso Idzerda who was in practice the first to develop and build transmitters specifically intended for broadcasting music and speech to the public. His transmitters used a patented modulation system that produced a mixture of AM and narrow-band FM. He carried out his transmissions with the clear intention of expanding the sale of crystal sets, valve amplifiers and components made or marketed by his own firm – Nederlansche Radio-Industrie, or NR-I for short.

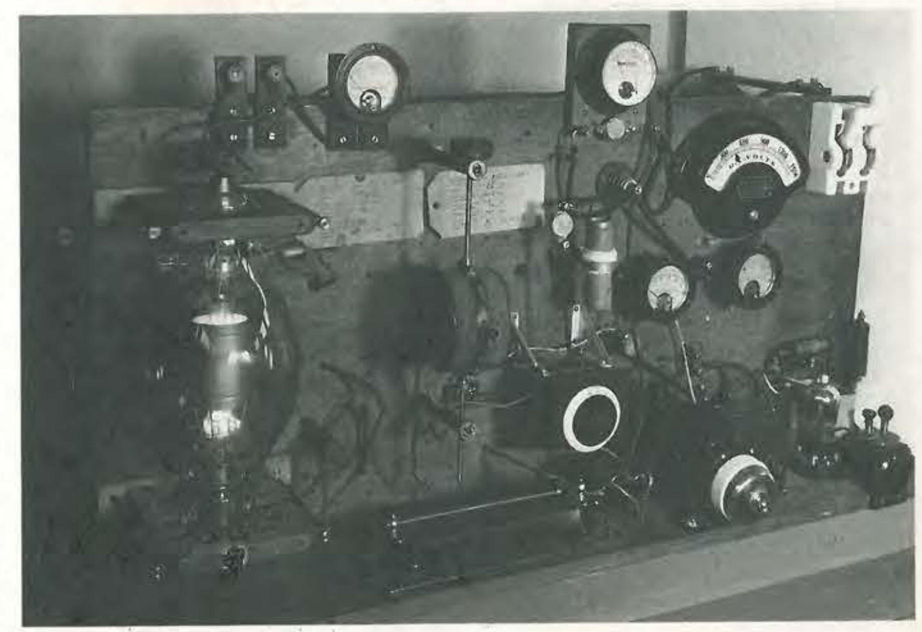
Hanso Henricus Schotanus à Steringa Idzerda, born at Weidum, Friesland in the north of Holland in September 1885, son of a country doctor, was never a man to go along with the crowd; fiercely individualistic he rebelled constantly against established authority. His secondary education was at a school noted for extremely strict discipline, but it failed to make him a conformist; he later graduated with an engineering degree from a German technical university.

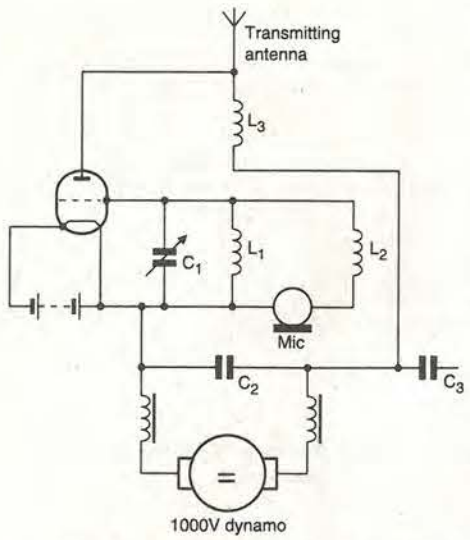
By 1905 Idzerda was trying to build an early aircraft, with what success is not known. But soon he became enthralled with the equally new science of 'wireless'. He became an early experimenter who took advantage of the liberal spirit whereby the Dutch authorities – like the British – permitted wireless telegraphy experiments in an era when many European governments were endeavouring to restrict radio to military and official communications.

In 1913 Idzerda set up a Technical Wireless Bureau in The Hague, helping the still small band of Dutch enthusiasts with whom he soon established a lasting rapport. When war broke out in August 1914, Holland, although remaining neutral throughout, withdrew all experimental radio privileges.

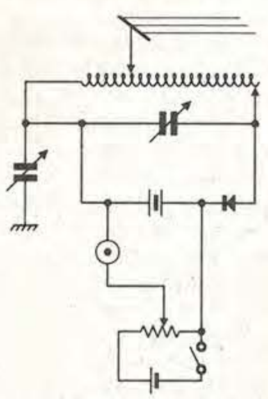
But in September 1917 – still 14 months

Two-stage 1kW PCGG transmitter now in the Netherlands Post Museum – in working order.





Original PCGG transmitter, from 1919, originally ran at 75W but a later increase in anode voltage to 1kV doubled the output.



Circuit of the 'Ontvangtoestel Type Amateur' as featured in the 1918 catalogue of NR-I and offered as a DIY kit. The battery and potentiometer weer used to select the best characteristic of the crystal.

Hanso Idzerda speaking into the PCGG microphone. Part of the large antenna coil is visible on the left.

before the Armistice – the ban was lifted and Holland became for a time the only country in the world where it was legal for the public to listen to radio signals; broadcasting thus became a real possibility.

Idzerda had friends among the Dutch military. He was aware of the progress being made in the use of thermionic valves, although none were available to civilians. Late in 1917, Idzerda tried unsuccessfully to obtain some from the factory in Utrecht where a few valves were then being made for the Dutch forces. These were based on a German Telefunken valve taken from a German aircraft that had landed by error in Holland.

From coffee and tobacco to valves

Idzerda who had obtained one of the original de Forest Audion triode valves took this to NV Philips Gloeilampenfabrieken at Eindhoven, seeking to interest the firm in the manufacture of thermionic valves.

The family firm of Philips, originally a tobacco and coffee processing business, had in 1891 changed to the production of electric lamps. Initially, Philips management was reluctant to enter the valve business, but finally agreed to make near copies of the Audion if Idzerda would contract to buy a minimum of 180 a year.

By early 1918, the first Philips-Ideezet valves were being delivered to him. (Ideezet represents the sound in Dutch of the first three letters of his name). By December 1918 more than 1200 had been sold at 12.50 guilders each.

Philips soon realised there was a promising future in valve manufacture. By July 1919 it was marketing the Philips-Ideezet 'soft' triodes independently of NR-I. It was also producing 'hard' valves for transmitters, although initially its Zendlampe gave only 5 to 10W rf output. Idzerda thus launched Philips into electronics in which it was subsequently to become among the world's giants. But soon this powerful firm was a rival rather than his partner.

By late 1918, with the Armistice signed, Idzerda was experimenting with a home-made continuous-wave valve transmitter. On 7 February 1919 he wrote to the minister of waterways seeking permission to demonstrate radiotelegraphy and radiotelephony at a Netherlands Trade Fair in Utrecht (24 February to 8 March 1919). He also asked to be allowed to make regular transmissions.

Permission for regular transmissions was finally granted on 14 August 1919, subject to tests with the military: NR-I was granted the callsign PCGGW and Philips PCJJ – later to become famous as a pioneer hf broadcasting callsign.

On 1 September 1919, the Dutch monthly magazine *Radio News* was announcing: "Every Thursday evening there will be continuous-wave transmissions from 8 to 10pm." In practice, the first broadcast with a selection of musical items was on 6 November, with details announced on 5 November in the daily newspaper *De Nieuwe Rotterdamse Courant*.

At first PCGG transmitted on 670m but this was soon changed, first to about 800m and then to 1030m. By the end of 1919, PCGG broadcasts were also being made on Sunday afternoons during which Idzerda replied personally to correspondents who reported reception of his transmissions.

1920 – Holland broadcasts abroad

The fact that his broadcasts were being received by British listeners soon encouraged Idzerda to initiate, from April 1920, The Hague Concerts. These were directed specifically at UK listeners. This date is sometimes wrongly given as the start of PCGG broadcasts.

Idzerda's single-valve transmitter output was at first about 75W, but this was increased to about 150W by running the valve with 1000V on the anode; by 1921 a Mullard valve was giving some 250W output.

When, in 1922, the *Daily Mail* sponsored for a time The Hague Concerts, power was increased to about 1 to 1.5kW. I believe that this later two-stage transmitter that was given in 1940 to the Netherlands Post Museum where it remains in working order.

The PCGG aerial was erected on the NR-I building at 8-10 Beukstraat, The Hague. It consisted of three wires each about 40m long, raised 15m high and stretched across a road. The earth comprised some 24ft of iron pipe sunk in a well. At full load, the early transmitter showed an aerial current of about 1.3A on 1000m and about 1.6A on 800m. With The Hague Concerts established, NR-I appointed W Burnham & Co of Deptford as its British agent.

Idzerda claimed a range of about 500 miles but this may have been based on a single report from Aberdeen. However, PCGG could be heard well in south-east England provided Croydon Airport radio was not causing interference. In May 1921, EW Kitchin wrote in "Notes on the reception of the Dutch concerts" *Wireless World* (14 May, 1921):

> *"As regards the strength of PCGG music, the writer gets it ten miles south of London quite nicely, audible with a single valve; and, with three note magnifiers added, it is quite loud on the telephone headgear and can be heard across the room."*

A month earlier, *Wireless World* had written:

> *"The phonographic selections sent out by this station are also interspersed with selections by a small band, and by four mandolin performers; occasionally also some singers take part. The orchestra and singers perform under a large funnel or horn which contains the microphones connected to the transmitting apparatus, so that the voices and music modulate the radiated power in the usual way... These concerts are addressed primarily to British wireless experimenters, as is evidenced by the introductory CW messages addressed to all British amateurs with which they are prefixed."*

The records were played on a wind-up, acoustic gramophone with its horn replaced by a tube in which the microphone was mounted. A second microphone was placed on the table near the gramophone.

In his broadcasts to Dutch listeners, Idzerda was often outspoken and critical of the authorities, leading to difficulties and reprimands. With Philips marketing valves and equipment, the funds for PCGG soon became sparse. In 1921, *Wireless World* appealed to its readers for funds to keep the broadcasts going; in 1922-23 there was the *Daily Mail* sponsorship but, with the BBC now on the air, the newspaper did not receive the promotional rewards it had expected and did not renew the contract.

By 1923 there were other stations broadcasting to Dutch listeners including PCUU and PCKK in The Hague. By about 1925 Hilversum NSF was on 1050m with 1.5kW, with PCGG listed as 1070m with 1kW providing only irregular transmissions.

As the popularity of radio spread, Idzerda was gradually squeezed out. As Professor Swierstra has written:

> *"The pity is that, in his single-minded devotion to his work, he failed to hear or heed the danger signs emanating from circles more commercially minded. Consequently, as the big manufacturers moved also into this field and took control, Idzerda, too proud to adopt the principles of the business world, was on a losing ticket... The memories of those glorious pioneer achievements faded."*

By about 1924, with Europe well launched into the radio era, PCGG broadcasts virtually ended and Idzerda and his family drifted out of the public eye – a situation which his wife found hard to bear. According to his daughter, Hanso bore it with typical Friesen stoicism. It was left to the Dutch radio amateurs to continue to show awareness of his contributions to broadcasting and radio technology.

Then came the second world war with, this time, Dutch neutrality rudely shattered by the German invasion of May 1940. Idzerda did not live to see the liberation. According to Professor Swierstra:

> *"By one of those strange cruelties of fate, he fell victim of his own passionate interest in technical developments... First he was caught listening to broadcasts from London – something strictly forbidden by the occupation authorities and more than enough to put him in their black books. But he followed this by committing an even greater faux pas.*
>
> *In Holland, as elsewhere, the Germans had set up launching pads from which to direct their infamous V1 and V2 rockets at Britain. When Idzerda was found trespassing on a prohibited area, searching for fragments of an exploded V2, he was immediately arrested by the enemy on suspicion of espionage. Probably without so much as a simple trial he was executed by shooting during the night of 3-4 November 1944."*

There may be some doubt whether this was the exact story since the prison where his execution took place was used for hostages rather than suspected intelligence agents. It would seem that, once again, the dice was loaded against this stubborn but dedicated Dutchman who had introduced broadcasting to the UK.

Finally, thanks to Dick Rollema, PAOSE, who supplied the illustrations and helped with technical information. ■

Further reading

Details of Idzerda's life are largely drawn from 'The birth of broadcasting' by M Tj Swierstra *(EBU Review* No 114B, March 1969) with follow-up letters in *EBU Review* No 116B, 117B and 120B providing information on the 1914 Belgian transmissions. Information on de Forest's early broadcasts comes mainly from 'Lee de Forest and the fatherhood of radio' by James A Hijiya (Lehigh University Press, 1992) and David Sarnoff's Magic Box memo from 'David Sarnoff' by Eugene Lyons (Harper & Row, 1966). Information on the Ideezet valves from 'Saga of the vacuum tube' by Gerald FJ Tyne (Howard W Sams & Co Inc, 1977).

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COPPER VAPOUR LASER PLANS Produces 100mw of visible green light. High coherency and spectral quality similar to Argon laser but easier and less costly to build yet far more efficient. This particular design was developed at the Atomic Energy Commision of NEGEV in Israel. £10/set Ref F/CVL1.

VOICE SCRAMBLER PLANS Minature solid state system turns speech sound into indecipherable noise that cannot be understood without a second matching unit. Use on telephone to prevent third party listening and bugging. £6/set Ref F/VS9.

PULSED TV JOKER PLANS Little hand held device utilises pulse techniques that will completely disrupt TV picture and sound! works on FM too! DISCRETION ADVISED. £8/set Ref F/TJ5.

BODYHEAT TELESCOPE PLANS Highly directional long range device uses recent technology to detect the presence of living bodies, warm and hot spots, heat leaks etc. Intended for security, law enforcement, research and development, etc. Excellent security device or very interesting science project. £8/set Ref F/BHT1.

BURNING, CUTTING CO2 LASER PLANS Projects an invisible beam of heat capable of burning and melting materials over a considerable distance. This laser is one of the most efficient, converting 10% input power into useful output. Not only is this device a workhorse in welding, cutting and heat processing materials but it is also a likely candidate as an effective directed energy beam weapon against missiles, aircraft, ground-to-ground, etc. Particle beams may very well utilize a laser of this type to blast a channel in the atmosphere for a high energy stream of neutrons or other particles. The device is easily applicable to burning and etching wood, cutting, plastics, textiles etc £12/set Ref F/LC7.

MYSTERY ANTI GRAVITY DEVICE PLANS Uses simple concept. Objects float in air and move to the touch. Defies gravity, amazing gift, conversation piece, magic trick or science project. £6/set Ref F/ANT1K.

ULTRASONIC BLASTER PLANS Laboratory source of sonic shock waves. Blow holes in metal, produce 'cold' steam, atomize liquides. Many cleaning uses for PC boards, jewllery, coins, small parts etc. £6/set Ref F/ULB1.

ULTRA HIGH GAIN AMP/STETHOSCOPIC MIKE/SOUND AND VIBRATION DETECTOR PLANS Ultrasensitive device enables one to hear a whole new world of sounds. Listen through walls, windows, floors etc. Many applications shown, from law enforcement, nature listening, medical heartbeat, to mechanical devices. £6/set Ref F/HGA7

ANTI DOG FORCE FIELD PLANS Highly effective circuit produces time variable pulses of accoustical energy that dogs cannot tolerate £6/set Ref F/DOG2

LASER BOUNCE LISTENER SYSTEM PLANS Allows you to hear sounds from a premises without gaining access. £12/set Ref F/LLIST1

LASER LIGHT SHOW PLANS Do it yourself plans show three methods. £6 Ref F/LLS1

PHASOR BLAST WAVE PISTOL SERIES PLANS Handheld, has large transducer and battery capacity with external controls. £6/set Ref F/PSP4

INFINITY TRANSMITTER PLANS Telephone line grabber/room monitor. The ultimate in home/office security and safety! simple to use! Call your home or office phone, push a secret tone on your telephone to access either: A) On premises sound and voices or B) Existing conversation with break-in capability for emergency messages. £7 Ref F/TELEGRAB.

BUG DETECTOR PLANS Is that someone getting the goods on you? Easy to construct device locates any hidden source of radio energy! Sniffs out and finds bugs and other sources of bothersome interference. Detects low, high and UHF frequencies. £5/set Ref F/BD1.

WOLVERHAMPTON BRANCH NOW OPEN AT WORCESTER ST W'HAMPTON TEL 01902 22039

ELECTROMAGNETIC GUN PLANS Projects a metal object a considerable distance-requires adult supervision £5 ref F/EML2.

ELECTRIC MAN PLANS, SHOCK PEOPLE WITH THE TOUCH OF YOUR HAND! £5/set Ref F/EMA1.

PARABOLIC DISH MICROPHONE PLANS Listen to distant sounds and voices, open windows, sound sources in 'hard to get' or hostile premises. Uses satellite technology to gather distant sounds and focus them to our ultra sensitive electronics. Plans also show an optional wireless link system. £8/set ref F/PM5

2 FOR 1 MULTIFUNCTIONAL HIGH FREQUENCY AND HIGH DC VOLTAGE, SOLID STATE TESLA COIL AND VARIABLE 100,000 VDC OUTPUT GENERATOR PLANS Operates on 9-12vdc, many possible experiments. £10 Ref F/HVM7/TCL4.

INFINITY TRANSMITTERS The ultimate 'bug' fits to any phone or line, undetectable, listen to the conversations in the room from anywhere in the world! 24 hours a day 7 days a week! just call the number and press a button on the mini controller (supplied) and you can hear everything! Monitor conversations for as long as you choose £249 each, complete with leads and mini controller! Ref LOT9. Undetectable with normal RF detectors, fitted in seconds, no batteries required, lasts forever!

SWITCHED MODE PSU'S 244 watt, +5 32A, +12 6A, -5 0.2A, -12 0.2A. There is also an optional 3.3v 25A rail available. 120/240v I/P. Cased, 175x90x145mm. IEC inlet Suitable for PC use (6 d/drive connectors 1 m/board). £10 ref PSU1.

VIDEO PROCESSOR UNITS?/6v 10AH BATTS/12V 8A TX Not too sure what the function of these units is but they certainly make good strippers! Measures 390X320X120mm, on the front are controls for scan speed, scan delay, scan mode, loads of connections on the rear. Inside 2 x 6v 10AH sealed lead acid batts, pcb's and a 8A? 12v torroidial transformer (mains in). Condition not known, may have one or two broken knobs due to poor storage. £17.50 ref VP2

RETRON NIGHT SIGHT Recognition of a standing man at 300m in 1/4 moonlight, hermatically sealed, runs on 2 AA batteries, 80mm F1.5 lens, 20mw infrared laser included. £325 ref RETRON.

MINI FM TRANSMITTER KIT Very high gain preamp, supplied complete with FET electret microphone. Designed to cover 88-108 Mhz but easily changed to cover 63-130 Mhz. Works with a common 9v (PP3) battery. 0.2W RF. £7 Ref 1001.

3-30V POWER SUPPLY KIT Variable, stabilized power supply for lab use. Short circuit protected, suitable for profesional or amateur use 24v 3A transformer is needed to complete the kit. £14 Ref 1007.

1 WATT FM TRANSMITTER KIT Supplied with piezo electric mic. 8-30vdc. At 25-30v you will get nearly 2 watts! £12 ref 1009.

FM/AM SCANNER KIT Well not quite, you have to turn the knob your self but you will hear things on this radio that you would not hear on an ordinary radio (even TV). Covers 50-160mhz on both AM and FM. Built in 5 watt amplifier, inc speaker. £15 ref 1013.

3 CHANNEL SOUND TO LIGHT KIT Wireless system, mains operated, separate sensitivity adjustment for each channel, 1,200 w power handling, microphone included. £14 Ref 1014.

4 WATT FM TRANSMITTER KIT Small but powerful FM transmitter, 3 RF stages, microphone and audio preamp included. £20 Ref 1028.

STROBE LIGHT KIT Adjustable from 1-60 hz (a lot faster than conventional strobes). Mains operated. £16 Ref 1037.

COMBINATION LOCK KIT 9key, programmable, complete with keypad, will switch 2A mains. 9v dc operation. £10 ref 1114.

PHONE BUG DETECTOR KIT This device will warn you if somebody is eavesdropping on your line. £6 ref 1130.

ROBOT VOICE KIT Interesting circuit that distorts your voice! adjustable, answer the phone with a different voice! 12vdc £9 ref 1131

TELEPHONE BUG KIT Small bug powered by the 'phone line, starts transmitting as soon as the phone is picked up! £8 Ref 1135.

3 CHANNEL LIGHT CHASER KIT 800 watts per channel, speed and direction controlssupplied with 12 LEDS (you can fit triacs instead to make kit mains, not supplied) 9-12vdc £17 ref 1026.

12V FLOURESCENT LAMP DRIVER KIT Light up 4 foot tubes from your car battery! 9v 2a transformer also required. £8 ref 1069.

VOXSWITCH KIT Sound activated switch ideal for making bugging tape recorders etc, adjustable sensitivity. £8 ref 1073.

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PREAMP MIXER KIT 3 input mono mixer, sep bass and treble controls plus individual level controls, 18vdc, input sens 100mA. £15 ref 1052.

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SOUND EFFECTS GENERATOR KIT Produces sounds ranging from bird chips to sirens, Complete with speaker, add sound effects to your projects for just £9 ref 1045.

16 WATT FM TRANSMITTER (BUILT) 4 stage high power, preamp required 12-18vdc, can use ground plane, yagi or open dipole. £69 ref 1021.

HUMIDITY METER KIT Builds into a precision LCD humidity meter, 9 ic design, pcb, lcd display and all components included. £29

PC TIMER KIT Four channel output controlled by your PC, will switch high current mains with relays (supplied). Software supplied so you can program the channels to do what you want whenever you want. Minimum system configeration is 286, VGA, 4.1,640k, serial port, hard drive with min 100k free. £24.99

FM CORDLESS MICROPHONE This unit is an FM broadcasting station in minature, 3 transistor transmitter with electret condenser mic+fet amp design result in maximum sensitivity and broad frequency response. 90-105mhz, 50-1500hz, 500 foot range in open country! PP3 battery required. £15.00 ref 15P42A.

MAGNETIC MARBLES They have been around for a number of years but still give rise to curiosity and amazement. A pack of 12 is just £3.99 ref GI/R20

NICKEL PLATING KIT Proffesional electroplating kit that will transform rusting parts into showpieces in 3 hours! Will plate onto steel, iron, bronze, gunmetal, copper, welded, silver soldered or brazed joints. Kit includes enough to plate 1,000 sq inches. You will also need a 12v supply, a container and 2 12v light bulbs. £39.99 ref NIK39.

Minature adjustable timers, 4 pole c/o output 3A 240v,
HY1230S, 12vDC adjustable from 0-30 secs. £4.99
HY1260M, 12vDC adjustable from 0-60 mins. £4.99
HY2405S, 240v adjustable from 0-5 secs. £4.99
HY24060m, 240v adjustable from 0-60 mins. £6.99

BUGGING TAPE RECORDER Small voice activated recorder, uses micro cassette complete with headphones. £28.99 ref MAR29P1

POWER SUPPLY fully cased with mains and o/p leads 17v DC 900mA output. Bargain price £5.99 ref MAG6P9

9v DC POWER SUPPLY Standard plug in type 150ma 9v DC with lead and DC power plug. price for two is £2.99 ref AUG3P4.

COMPOSITE VIDEO KIT. Converts composite video into separate H sync, V sync, and video. 12v DC. £8.00 **REF: MAG8P2.**

FUTURE PC POWER SUPPLIES These are 295x135x60mm, 4 drive connectors 1 mother board connector. 150watt, 12v fan, iec inlet and on/off switch. £12 Ref EF6.

VENUS FLYTRAP KIT Grow your own carnivorous plant with this simple kit £3 ref EF34.

6"X12" AMORPHOUS SOLAR PANEL 12v 155x310mm 130mA. Bargain price just £5.99 ea REF MAG6P12.

FIBRE OPTIC CABLE BUMPER PACK 10 metres for £4.99 ref MAG5P13 ideal for experimenters! 30 m for £12.99 ref MAG13P1

ROCK LIGHTS Unusual things these, two pieces of rock that glow when rubbed together! belived to cause rain! £3 a pair Ref EF29.

3' by 1' AMORPHOUS SOLAR PANELS 14.5v, 700mA 10 watts, aluminium frame, screw terminals, £44.95 ref MAG45.

ELECTRONIC ACCUPUNCTURE KIT Builds into an electronic version instead of needles! good to experiment with. £7 ref 7P30

SHOCKING COIL KIT Build this little battery operated device into all sorts of things, also gets worms out of the ground! £7 ref 7P36.

FLYING PARROTS Easily assembled kit that builds a parrot that actually flaps its wings and flies! 50 m range £6 ref EF2

HIGH POWER CATAPULTS Hinged arm brace for stability, tempered steel yoke, super strength latex power bands. Departure speed of ammunition is in excess of 200 miles per hour! Range of over 200 metres! £7.99 ref R/9.

BALLON MANUFACTURING KIT British made, small blob blows into a large, longlasting balloon, hours of fun! £3.99 ref GI/E99R

9-0-9V 4A TRANSFORMERS, chassis mount. £7 ref LOT19A

2.5 KILOWATT INVERTERS, Packed with batteries etc but as they weigh about 100kg CALLERS ONLY! £120.

MEGA LED DISPLAYS Build your self a clock or something with these mega 7 seg displays 55mm high, 38mm wide. 5 on a pcb for just £4.99 ref LOT16 or a bumper pack of 50 displays for just £29 ref LOT17.

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ACTIVE

Discrete active devices

Rf power transistors. Ericsson announces six new power transistors for the 1500-1700MHz Inmarsat band. PTB20077/8 and PTB20228 are for use in cw and pep application between 1525MHz and 1660MHz, being rated at 0.7W, 2.5W and 6W respectively. They are all n-p-n devices for common-emitter circuitry, the 20077 being Class A with 12dB of gain and the others Class AB with 11dB. Higher-power types, the PTB20079/80 and PTB20210, produce 10W, 25W and 55W and are intended for pep application in the 1600-1700MHz band. Again, they are all n-p-n devices for common-emitter in Class AB with a gain of around 10-12dB. All six are for 26V dc working. Richardson Electronics (Europe) Ltd. Tel., 01753 733010; fax, 01753 733012.

Voltage references. ZRA250 and ZR4040-2.5, newly announced by Zetex, provide micropower reference voltage of 2.5V with tolerances of 0.5%-2%, or up to 3% in the 250. Bandgap design avoids the use of stabilising capacitors, capacitive loads not affecting performance. The devices handle 50µA to 5mA and 60µA to 15mA respectively and draw a quiescent current of 25µA, performance being held to a maximum of 25mA; transient currents to 200mA are permissible. Stable operation is reached in 10µs and temperature coefficient is 30ppm/°C. Zetex plc. Tel., 0161-627 5105; fax, 0161-627 5467.

1A Schottkys. Zetex's ZCHS1000 Schottky diode is the first capable of continuous 1A working. Part of the Superbat range, it exhibits a very low forward voltage and takes an average current of 2A or 1A continuously. Total power is 500mW and maximum reverse voltage 30V. Zetex plc. Tel., 0161-627 5105; fax, 0161-627 5467.

Mixed-signal ICs

Camera chipset. Sony has the SS-1 three-chip set which, with one of a variety of Sony colour ccds, forms a low-cost addressable Pal/NTSC camera for security and multimedia work. Ccds available range from a 0.2in type with 180k pixels to a 0.5in version with 380k pixels. The set comprises the CXD2163 signal processor which provides both analogue and digital chrominance and luminance output and a choice of communication channels to work with a microprocessor or with a pc via its RS-232 port; the CXA2006 ccd head amplifier; and the CXD2480 ccd driver incorporating timing-control for Pal and NTSC. Horizontal and vertical clock drivers and a shutter function are provided. Sony Semiconductor Europe. Tel., 01256 478771; fax, 01256 818194.

Motor controller. An entire motor control signal chain in a single ic, the ADMC330 by Analog Devices, contains a fixed-point digital signal processor, 4K of program memory and a set of peripheral functions for variable speed control of induction and electronically commutated motors. There is a seven-channel a-to-d converter synchronised to the switching frequency to reduce ripple, two auxiliary pwm timers for power factor correction, set point control and an 8-bit i/o port for expansion. Analog Devices Ltd. Tel., 01932 266000; fax, 01932 247401.

Optical devices

Led for links. Mitel's MF430 led is meant for use in short-distance optical links, wavelength being 865nm. Using GaAlAs, it has an actively aligned ST receptacle for best power coupling to the 62.5µm or 125µm fibre. The device operates at 155Mb/s and bandwidth is 250MHz. Mitel Semiconductor. Tel., 01291 430000; fax, 01291 436389.

PASSIVE

Passive components

Encapsulated transformers. A range of low-voltage mains transformers from Stontronics are encapsulated and mount directly to a pcb. There are six output voltages and fifteen power ratings, all of them accepting 230V input and having two independent output windings. Double-section bobbins provide good insulation and the transformers are approved to UL, CSA, and VDE 0551/EN 60742 standards. Stontronics Ltd. Tel., 01734 311199; fax, 01734 311145.

Varicaps. Philips has a number of new varicaps: two low-voltage types for communications and two for television tuners. First in the new comms family is the BB155, which has a capacitance spread of 45.2-49.8pF at 0.34V and 24.55-26.7pF at 2.82V and can be used as a direct replacement for existing varicaps, albeit with much tighter tolerance. In the tuner family, the BB146/7 have twice the capacitance ratio of standard devices and 'gliding' matching, a gliding sequence of eight diodes providing capacitance matching of under 1.6% for the BB146 and 2% for the 147. For extended uhf tuning, the BB146 gives a minimum capacitance ratio of 17.5 at 1MHz, the BB147 giving a ratio of 40. Philips Semiconductors (Eindhoven). Tel., 00 31 40 2722456; fax, 00 31 40 2724825.

High-res. potentiometer. With 25 turns, Murata's 3106 potentiometer series meets demands for higher resolution and lower costs. In five styles of terminal, the 3106 comes in a range of values from 10Ω to 2MΩ at 0.5W and is rated at 300V, Dielectric strength is 1000V ac and insulation resistance 1GΩ. Murata Electronics (UK) Ltd. Tel., 01252 811666; fax, 01252 811777.

Low-Z capacitors. Nippon Chemi-Con LXVVB series aluminium electrolytics come in values between 12µF and 15,000µF, all of them working at temperatures in the –55°C to 105°C range and having assured load lives at 105°C of 2000-5000 hours, depending on type chosen. Voltage ratings are 6.3-63V dc and tolerances ±20%. Hawnt Electronics Ltd. Tel., 0121 7843355; fax, 0121 7831657.

S-m inductors. ECM Electronics considers that its range of surface-mounted inductors should keep 95% of customers happy. The range covers thin-film types (the 0805), encapsulated wire-wound models in the 1008, 1210 and 1812 versions and a surface-mounted power range. The three wire-wound series have inductances of 0.005-100µH for the 1008, up to 330µH in the 1210 and to 1mH for the 1812. Power inductors come shielded or unshielded in inductances to 1mH with a current rating of 1A. ECM Electronics Ltd. Tel., 01903 892810; fax, 01903 892738.

Electrolytics for audio. CEBNP (non-polar) and CEBBP (bipolar) aluminium electrolytic capacitors by Dubilier are intended for use in audio work, taking charge in both directions. CEBNPs are useful in audio crossover networks, providing minimum capacitance variation with frequency and maximum leakage current of 0.03CV or 3µA. Two ranges offer voltage ratings in the 6.3-100V and 160V-250V ranges in values from 0.47µF to 1000µF. CEBBPs can be used in television line deflection circuitry, possessing a low dissipation factor and maximum leakage of 100µA. Values in the range 1-47µF at 25-50V. Dubilier Ltd. Tel., 01371 875758; fax, 01371 875075.

Electrolytics. The Nover range of aluminium electrolytic capacitors, described in a new catalogue, includes both polarised and bipolar types with ratings of 4V to 450V dc and 350V ac. Values range from 0.1µF to 820,000µF in three temperature grades, these components being principally intended for high-volume consumer application. Anglia. Tel., 01945 474747; fax, 01945 474849.

High-current electrolytics. GW Series electrolytics by Fischer & Tausche provide higher ripple current, lower hot-spot temperature and longer life than those of more conventional design, brought about by a winding design to minimise internal loss and a large base area and thermally conducting pad to give efficient heat transfer, with a thermal resistance of under 2°C/W. The capacitors come in values between

1000μF and 100,000μF in a 40-500V range. Operating temperature is –40°C to 105°C and isolation over 2500V ac. Can design eases mounting and heat transfer. Campbell Collins Ltd. Tel., 01438 369466; fax, 01438 316465.

Audio products

100W amplifiers. Exicon mosfet evaluation amplifier modules come with comprehensive applications data which, with Spice data and models also provided, should help to reduce development costs and time. Modules exhibit a total harmonic distortion of under 0.01%, slew at more than 100V/μs and offer a power bandwidth of more than 100kHz. Power output is 100W to over 3kW. Profusion. Tel., 01702 543500; fax, 01702 543700.

Connectors and cabling

High-temperature connectors. IMC is a family of industrial micro connectors by Deutsch which are made of Ultem, a temperature-resistant composite that allows use at temperatures between –55°C and 175°C. The housings are moisture and dust resistant to IP67. Two sizes are made: Series 100 are 15.25mm in diameter for both plug and socket and has up to four contacts; series 20 measures 19mm in diameter and has up to 12 contacts. Current rating of the gold-plated contacts is up to 13A, voltage rating 750-1000V ac and insulation resistance 1GΩ. Surtech Interconnection Ltd. Tel., 01256 51221; fax, 01256 471180.

Crystals

Crystals for PCMCIA. Crystals by Seiko Epson in the FA356/7/8 series are meant for use in applications such as PCMCIA cards, disk drives and network cards. They are contained in ceramic packages with glass seals and resonate at frequencies from 14MHz to 41MHz with stabilities of ±50ppm or ±100pmm; standard frequencies off the shelf are 35.2512MHz and 40.32MHz. ECM Electronics Ltd. Tel., 01903 892810; fax, 01903 892738.

Displays

Colour lcd. Seiko's G121C colour liquid-crystal display changes colour by means of the electrically controlled birefringence effect in response to a change in voltage applied to the liquid crystal; no filter is used, so that brightness is increased and no back light is needed. The 128 by 128 matrix is composed of 0.46 square millimetre dots, the viewing area being 67.4 square millimetres. Supply is +5V, –15V. Craft Data Ltd. Tel., 01494 778235; fax, 01494 773645.

Hardware

Emi/rfi screens. Tecan Components offers a range of screening products to meet the EC Directive on emc. Shielding cans for board-mounted components are made by photo-chemical etching, providing low cost and speed of manufacture of burr and stress free screens to customers' specifications; fences with sprung lids are also made. Screening mesh is made with up to 120 openings/in^2 in 0.08mm to 0.25mm copper or stainless steel and are available with a supporting framework. For prototypes there is Tecshield a bench modelling material in copper etched sheet. Tecan Components Ltd. Tel., 01305 765432; fax, 01305 780194.

Designer hardware. For all those of us who think that electronic equipment is too pedestrian, West Hyde can provide 1U-6U front panels in glorious Technicolor. Choice of colours is virtually limitless, so if you need a little purple and yellow splash-anodised confection to match a nightclub's decor, you can have it. The surface is smooth, scratch-resistant and withstands heat well. Nineteen-inch panels can be ready punched or drilled and come with four mounting holes. West Hyde Enclosures. Tel., 01453 8367789; fax, 01453 836444.

Board storage rack. Printed-circuit boards in the course of manufacture need protection from electrostatic, physical, environmental and chemical damage; racks from TBA Electro Conductive Products take care of all these factors. They are made in ECP 104 conductive polypropylene, are adjustable to take boards of different sizes and are modular for expansion. TBA Industrial Products Ltd. Tel., 01706 47718; fax, 01706 46170.

Test and measurement

GPS frequency standard. Sematron offers a disciplined GPS frequency reference controller with its receiver, antenna and stabilised oscillator in one weatherproof package. It is meant to be roof-mounted, where it will monitor up to eight GPS satellites, a single coaxial cable taking power to the unit and frequency reference from it. Output frequency is set at the factory and may be 5MHz, 10MHz, 13MHz, 2048kHz G703/10 or 2048Kb/s G703/6. Accuracy is around 1 in 10^{11} and warm-up time 20 minutes. External supply needed is 12-76V dc. Sematron UK Ltd. Tel., 01256 812222; fax, 01256 812666.

Dsos at analogue prices. Reasons for the traditional use of analogue oscilloscopes in education and servicing are addressed by the Tektronix TDS210/20 digital storage instruments. Ease of use comes from the use of a similar user interface, which is multi-lingual, to that in analogue types; bandwidth is 60MHz or 100MHz; and cost is reduced by the use of a flicker-free lcd display, which also shows readouts and menus. A full set of dso functions is provided. Options include a hard copy extension module and provision for RS232 and GPIB communication. Amplicon Liveline Ltd. Tel., 0800 525 335 (free); fax, 01273 570215.

Dso with 4Mbyte storage. Digital oscilloscopes in the VC-75XX series by Hitachi Denshi have PCMCIA interfaces to take memory cards of up to 4Mbyte capacity, which is enough to store 325 waveforms of 8Kword each. There is also an MS-DOS card to allow downloading onto an sram for display on a PCMCIA-equipped pc; pixel data can also be stored for up to three months. A printer is provided, useful if left on overnight to record transient traces, for example. VC-7504/2 are both 150MHz instruments sampling at 100Msample/s on four (two) channels simultaneously. A number of special trigger modes is available and automatic measurement facilities are provided, as is waveform manipulation. Hitachi Denshi (UK) Ltd. Tel., 0181-202 4311; fax, 0181-202 2451.

Digital radio base station tester. Racal's 6113 digital radio tester now has two new test modes. The Base Station On-air Service System now allows the monitoring and measurement of performance while normal two-way traffic is in progress at the base transceiver station, the required sensitivity and selectivity being provided to allow the monitoring of individual signals in the presence of interfering adjacent traffic. Additionally, the 6113 will operate as a monitor/emulator to simulate faults or incipient faults when used in conjunction with the Air Interface Monitor Emulation software. Racal Instruments Ltd. Tel., 01628 604455; fax, 01628 662017.

Literature

Bull Electrical. New from Bull, its catalogue of kits and ready-built equipment, containing most things from an fm transmitter to a wind-up flying parrot, by way of video cameras, night sights and a kit for making chewing gum. Bull Electrical. Tel., 01273 203500; fax, 01273 323077.

Murata on cd. Murata's 1997 short catalogue is now on cd-rom and uses a Windows-based search facility to handle more than 10,000 pages, which include applications. The facility also allows users to abstract parts of the catalogue to use in spreadsheets and word processors. Components covered include filters, sensors, capacitors, thermistors, coils, resonators and piezoelectric audio components. Murata Electronics (UK) Ltd. Tel., 01252 811666; fax, 01252 811777.

Safety and foot switches. Camden Electronics has published a new catalogue of a range of position, safety and foot switches. Ranges include miniature, medium and heavy-duty switches, microswitches and safety switches actuated by tongues, hinge, reset and pull wire. Camden Electronics Ltd. Tel., 01727 864437; fax, 01727 855400.

Ic range on cd-rom. A cd-rom containing comprehensive information on the Cypress range of integrated circuits includes application notes and data sheets on static memory, programmable components and computing devices. The latest version of Adobe's Acrobat document reader software is included on the cd and quarterly updates will be provided. Pronto Electronic Systems Ltd. Tel., 0181-554 5700; fax, 0181-554 6222.

Power supplies

40/60W supplies. New 3.3V and 48V versions of the Astec LPS40 (40W) and LPS60 (60W) single-output supplies are now available. Their size, 76.2 by 127mm and 29 and 41.9mm high, makes them suitable for small spaces and integral remote sensing ensures accuracy of ±2%. Mtbf is 550,000 hours and universal input 85-264V, 47-440Hz and 120-370V dc. Convection cooling allows output of 8A and 12A respectively in the 3.3V versions and 0.9A and 1.3A for the 48V types, these figures increasing by a factor of up to about 1.3 with forced-air cooling. Cased and uncased versions are offered. Powerline Electronics Ltd. Tel., 01734 868567; fax, 01734 755172.

Dc-to-dc converters. Dual-output converters from Newport in the NMJ series are provided with 5.2kV isolation between input and output. The 1W devices comply with BS EN 60950 safety standards and are contained in standard 7-pin SIP packages. From 7-15V nominal input, NMJ converters provide outputs of ±5V, ±9V, ±12V or ±15V with loading split in any proportion between outputs. Switching frequency is 55kHz and zero-load power 100mW. Newport Components Ltd. Tel., 01908 615232; fax, 01908 617545.

Reversible high voltage. In 10ms, the output of the HP2.5RZC zero-crossing power supply changes from –2.5kV to +2.5kV in response to a square-wave signal. It is continuously controllable between the two levels, output near zero being less than 100mV. Voltage input is 24V and the unit delivers 400μA, stability being assisted by both current and voltage monitor signals. Optionally, the unit may be used to sink as well as source current. Applied Kilovolts Ltd. Tel., 01273 439440; fax, 01273 439449.

Vision systems

Camera module demonstrator. Vision offers a demonstrator for its 5400 range of cmos imaging products, which is usable as an evaluation and development device or to demonstrate products. The demonstrator consists of a microcontroller, a-to-d converter, RS232 interface, audio circuitry and microphone and the VV5426 sensor and circuitry and lens, although there is also a socket for an external camera. It supports digital video conversion by analogue and sync. outputs and a digital output stream. Functions are controlled by a push-button menu and lcd or by pc, settings being stored in a non-volatile memory. VLSI Vision Ltd. Tel., 0131-539 7111; fax, 0131-539 7141

Radio communications products

Key-fob transmitter. Low-power Radio Solutions offers a little transmitter for key rings, working at 418MHz or 434MHz. It is contained in a 11.5 by 7.5mm two-pin package and is approved to MPT1340 in the UK and ETS-300-220 in Europe. Power needed is 2.5V-13V at 4.6mA maximum and it uses am to make it compatible with super-regen receivers. Radiated power is up to –6dBm, giving 100m range with a decent receiver and data rates up to 1200b/s are achievable. The only externals needed are a capacitor, a resistor and an encoding circuit such as PIC microcontroller. Low Power Radio Solutions Ltd. Tel., 01993 709418; fax, 01993 708575.

Switches and relays

Sub-miniature relays. HanKuk HR702 relays are single-pole types for use where it is required to handle power in a small space; size is 15.4 by 14.8 by 19.2mm. Ratings available are 7A to carry 10A, and 10A to carry 15A, both with coil voltages from 5V to 24Vdc, dissipating 0.36W. Dielectric strength is 1.5kV for a minute between coils and contacts and 750V between contacts. Inelco Ltd. Tel., 0118 9810799; fax, 0118 9810844.

Flush-mount switches. EAO-Highland's Series 04 range of illuminated push-button switches and indicators can now be supplied in flush-mounted versions, which can be cleaned simply by wiping them down. The devices are modular, using snap-on contact blocks to allow switch configurations to be easily modified. The range includes momentary or maintained action devices, multi-position rotary selectors and multi-led or lamp indicators and buttons in ratings from 5V/10mA to 500V/10A. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

Tough keyboards. Rowland Automation offers custom design and manufacture of data-entry keyboards meant for use in positions where accidental or deliberate damage is a possibility. Brass, stainless steel and aluminium are all used for both keycaps and panel and the units will, it is claimed, survive attack by a hammer and 'hostile substances'. Costs are said to be lower than in standard products because the type of key used is cheaper to make. Rowland Automation Ltd. Tel., 01202 826398; fax, 01202 828205.

Board-mounted switches. EAO-Highland's range of COSMOS switches now includes pcb mounted types with a 'feel', produced with variety of button shapes, colours, lenses and types of illumination. They are in 12.5mm square modules to allow mounting in a 2.54mm matrix and contacts and terminals are proof against dust, fluid splashes, flux and solvents. Single-

Production test equipment

BGA inspection. Cognex UK offers a machine vision package, for use by device manufacturers or oems, to check ball grid arrays for missing, misplaced or deformed solder balls at up to 4000 balls per second. This is a PCIbus plug-in vision processor and software package complete with a Windows-based graphical device description editor for training on a variety of devices. To train the system, the user describes the ball pattern by specifying the ball count, pitch and size, whereupon a point-and-click graphics tool edits the device parameters. The package also calibrates the system to convert pixels to physical units and to correct for camera skew, distortion and scaling. Cognex UK Ltd. Tel., 01707 828018; fax, 01707 828019.

Please quote "Electronics World" when seeking further information

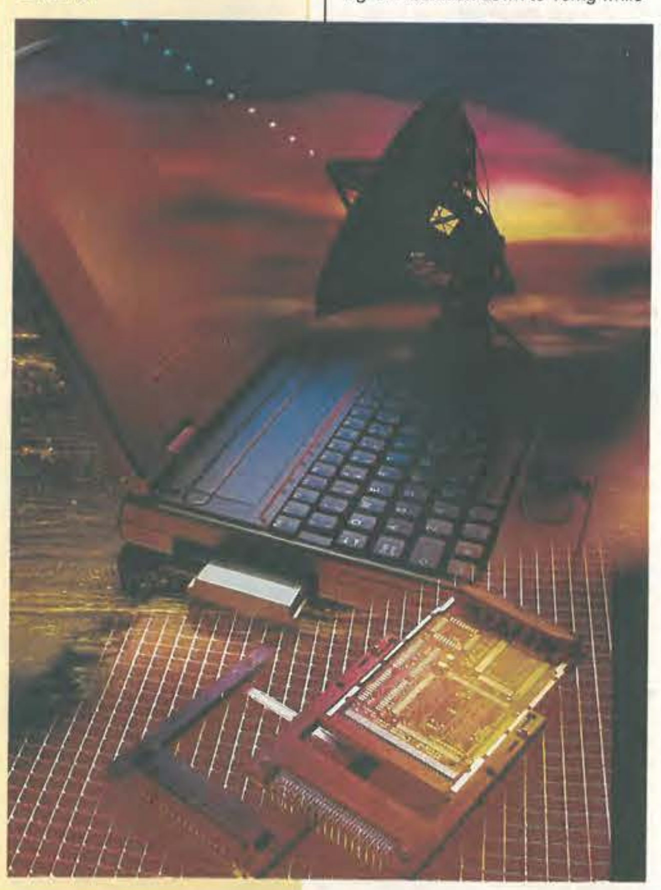
pole, two-position contacts are either 2μm silver plated on 0.5μm gold on nickel. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

Transducers and sensors

Hall sensors. FET Electronics has a range of chopper-stabilised cmos Hall sensors, the US1881/3881/5881, which includes latches and switches for the commercial, industrial and automotive temperature ranges. Operating voltage is 3.5-18V and switched current 20mA. The use of cmos with chopper-stabilisation and switched capacitors eliminates amplifier offset voltage and temperature drift. Activation in the typical applications of speed, position and angular position is by commercially available magnets. The devices are optimised for commutation in 5V and 12V brushless dc motors. FET Electronics Ltd. Tel., 01635 524490; fax, 01635 552244.

Computer peripherals

PCMCIA type III kits. Robinson Nugent can supply card kits, part of the MEMPAK family, to the PCMCIA type III standard. The cards are the same size as types I and II, but are 10.5mm thick, needing only one 68-pin connector to plug into the pc or satellite decoder (applications envisaged are wireless lans and satellite decoding). A number of mounting styles for the connector are offered, the possibility of stacked printed boards being mentioned. Robinson Nugent (Europe) Ltd. Tel., 01256 842626; fax, 01256 842673.



Optical switch. EE-SC1031 by Omron is a dual-channel slotted optical switch for speed and direction sensing of coded disks or linear strips, using two infrared led and phototransistor pairs. Aperture is 0.5mm and slot width 3.5mm. Omron Electronics Ltd. Tel., 0181-450 4646; fax, 0181-450 8087.

Integrated accelerometers. Analog's ADXL150/250 are said to be the first fully integrated, ±50g accelerometers to be made available with signal conditioning on the same chip. Signal:noise ratio is 74dB at 100Hz bandwidth, which permits signal resolution down to 10mg while still measuring ±50g full-scale. Zero-g drift is 0.4g over the standard industrial temperature range and scale factor and zero-g output level follow the supply voltage, so that succeeding circuitry tracks the accelerometer output. The 150 is a single-axis version, the 250 having two axes. Analog Devices Ltd. Tel., 01932 266000; fax, 01932 247401.

Gas sensors. Goldtech Transduction gas sensors are available to detect carbon monoxide, hydrogen, methane, butane, oxygen and alcohol in concentrations down to 20ppm. None of the sensors contains rare metal catalyst; they are therefore proof against poisons and saturation. Complete modules using these sensors and mask-programmed microcontrollers are supplied to order. Anglia. Tel., 01945 474747; fax, 01945 474849.

Laser displacement sensing. Three Schaevitz non-contacting LSL Series laser displacement sensors are announced. DistanceStar sends out a beam from a laser diode which is reflected from a surface onto a position-sensitive detector, subsequent processing turning the detector output into a voltage proportional to displacement. Range is ±5mm to ±20mm and response 9Hz. TwinStar uses triangulation to sense surface variation down to 0.007μm, dual optics allowing measurement of surfaces with sharp variations in height. BeamStar measures the size of objects at distances up to 1m, being a two-piece device in which the size is determined by the portion of the beam reaching the receiver, the smallest detectable object being 0.25mm. Applications include cloth or paper edge detection and particle measurement in fluids. Lucas Control Systems Products. Tel., 01753 537622; fax, 01753 823563.

Reflective sensor. Omron has a very low-profile reflective sensor designed to detect highly reflective surfaces such as paper or polished metal in positions in which there isn't much space; it stands 3mm off the board on a 15 by 4.2mm area. The EE-SY171 is a single unit using an infrared led and a phototransistor on parallel axes, sensing surfaces up to 3.5mm away. Omron Electronics Ltd. Tel., 0181-450 4646; fax, 0181-450 8087.

Alarm systems

Little alarms. Star Micronics MNT-03A and MXT-03A are surface-mounted electromagnetic audio transducers that are 9mm square by 4.5mm high and yet manage to produce 93dB at 10cm on 3.6V. Sound comes out of either the side or the top. Roxburgh Electronics Ltd. Tel., 01724 281770; fax, 01724 281650.

COMPUTER

Computer board-level products

Core controller. Meant for users who need to develop slightly different embedded controllers, Infotec have the K3 Kernel, which is a generic system based on its own real-time operating system software and a hardware core using the NEC V25 cpu. Peripherals such as signal conditioning, clocks and modems are then added to the core to specific requirements. K3 is intended for use in lower-cost, medium-performance control systems such as machinery control and data collection. Infotec Ltd. Tel., 01530 560600; fax, 01530 560111.

Computers

Faster embedded computer. From Ampro, the new Little Board/P5i embedded single-board pc, which gives 25 times the i/o performance of PC/104-based types. It uses a 166MHz Pentium cpu and achieves its i/o throughput by developing the PC/104-Plus bus with a throughput of 132Mbyte/s. Functions include four serial and a parallel i/o port, floppy and enhanced IDE drive interfaces, PCI UltraSCSI, 10BaseT Ethernet and PCI lcd/crt Super VGA display controller. There are also a bootable solid-state disk, embedded-pc bios, a watchdog timer and many other enhancements. Crellon Microsystems. Tel., 01734 776161; fax, 01734 776095.

68EN360-based computer board. BVM's BVME3000 single-board computer family is based on the 68EN360 quad integrated comms controller running at 32MHz. Used as a stand-alone, embedded controller or with a VMEbus interface, it is an alternative to 68040 boards, particularly in communications. Any of the six independent serial ports may use many protocols with external transition modules and in-chip Ethernet is available. BVME3000 takes up to four IndustryPack input/output mezzanine boards with dma access; i/o boards run at both 8MHz and 32MHz. Memory options include up to 16Mbyte of 32-bit dram, 2Mbyte of flash and 512Kbyte of dual-ported sram, an expansion socket taking extra memory. BVM Ltd. Tel., 01489 783589; fax, 01489 780144.

Data communications

Analogue/digital Tx/Rx. Telemetry encoder and decoder units from Wood & Douglas, the ADMR/T handle up to four analogue and 16 digital signals on the same link to form a versatile, general-purpose modem for simple a/d signals. Both units are modular pcb assemblies, interfacing with a 600Ω line or with a standard W&D transmitter and receiver. Serial data is transmitted at 2400baud, analogue i/o having a 12-bit resolution and voltage range of 0-5V. Input and output impedances are 100kΩ and 300Ω. Digital information is in the form of eleven standard and five optional on-off functions. Wood and Douglas Ltd. Tel., 01734 811444; fax, 01734 811567.

Switchable SCSI terminators. Motorola's MCCS142236/7/8/9 SCSI terminators are switchable and therefore remove any need to remove terminations physically, being turned on and off by hardware or software. Both the xx6 and xx8 have eighteen 110Ω terminating resistors with 4pF disconnect capacitance and an on-board 2.85V regulator with current-sinking support, while the x7 has nine, with 3pF capacitance. Motorola. 001 602 732 2397; fax, 001 602 7325020.

Development and evaluation

Faster PIC development. RF Solutions offers a daughter board designed for use with Windows, for its ICEPIC2 in-circuit emulator, which is intended to speed up the development of the Microchip PIC 17C 16-bit microcontroller, providing real-time emulation at up to 25MHz. There is source-level debugging in assembler or C and unlimited hardware trigger breakpoints on any address or range of addresses. The board plugs directly into the ICEPIC2 main pod. RF Solutions Ltd. Tel., 01273 488880; fax, 01273 480661.

C161 starter kit. KiTCON161, a 16-bit starter/evaluation kit for the C161 microcomputer, is based on the Phytec 16MHz C161V microcomputer card and has flash eprom, ram and complete i/o and bus interface. It can be programmed and debugged with the Keil C166 C compiler and HiTOP161/WIN monitor. To help with real applications, the PDP161-T kit has a 161-specific version of the Hitex AX166 romless emulator and C compiler. Hitex (UK) Ltd. Tel., 01203 692066; fax, 01203 692131.

Computer security

Computer enclosure. Completely proof against dust and water, Intek's stainless steel Armagard Flat Panel Enclosure takes almost any lap-top pc and any flat panel display to form an industrial pc, terminal or remote monitor, no modifications being needed to allow the pc to be housed. A lap-top pc is opened out flat, its display being viewed through a sealed transparent panel and an external membrane keyboard connected to it for data entry. Cable access in by gland plates and the rear door swings out to provide access for servicing. Intek Electronics Ltd. Tel., 01352 810603; fax, 01352 810403.

Computer board level products

PC/104 386 cpu card. PCM-3335 from Impulse Corporation is a full-function 386SX-40 single-board computer that provides support for a flat-panel display or a crt and PC/104 connections. It has an ISA 16-bit data bus and offers 32-bit processing. Memory can be up to 4Mbyte of ram and two IDE hard disk drives and two floppy drives can be accommodated. There are two RS-232 serial ports, a 16C550 uart with 16byte fifo support transfer speeds of up to 115Kb/s. Impulse Corporation Ltd. Tel., 01543 466552; fax, 01543 466553.

Software

Help with CE. For those experiencing mental overload caused by the Low Voltage Directive, Seaward Electronics has the LVD Expert System, a Windows-based software package that is clearly a little like having the bank manager in the broom cupboard. The program includes details of all relevant test standards which show whether a particular piece of equipment comes within the scope of the LVD or, indeed, any other European, US, Japanese or Australian directive. Paraphrasing, hypertext features and diagnostic routines are present to indicate design and engineering practices that will not fall foul of the LVD; check sheets and proforma documents are usable to make Technical Files for declarations. Seaward Electronic Ltd. Tel., 0191 586 3511; fax, 0191 586 0227. ■

Data logging. Three Windows-based packages for use with Datataker 50 and 500/600 data loggers assist with remote communication, logger programming and data acquisition. The suite consists of: DeTerminal, a Windows version of an earlier communications facility, which is now much simpler to use; DeLogger, which allows programming of Datatakers from menus; and DeLogger Pro, which has a fuller version gui than DeLogger with extra facilities such as real-time and historical data in the form of line charts and bar graphs, and analogue and digital meters. The package comes with all new Datatakers and is available for existing users. Data Electronics. Tel., 01462 481291; fax, 01462 481375.

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New rf silicon

Loek Colussi discusses a new bipolar rf transistor technology that rivals GaAs in digital cellular and cordless phone applications.

A new generation of high-performance silicon bipolar rf transistors with transition frequencies in excess of 20GHz has been developed by Philips Semiconductors. These devices are intended for use in low-voltage cordless and cellular telephones.

In addition to small-signal types for use in a phone's rf receiver, these fifth-generation rf wideband transistors include medium-power types. These rival the performance of GaAs devices when used for rf power amplification in a telephone's transmitter. Unlike GaAs devices, however, they operate at high efficiency from a single supply rail – considerably reducing circuit complexity and allowing the design of smaller, lighter, portable phones.

The key to reducing the size and weight of a portable telephone is the use of a smaller battery pack. In order to maximise energy density, this usually means using fewer cells and consequently a lower supply voltage, typically between 3 and 3.6V. Ideally, the entire telephone should then operate from this single supply voltage. Although dc-to-dc converters can be used to create higher supply voltages, they inevitably result in efficiency losses which shorten the telephone's standby and talk times. They also occupy valuable pc board area and increase the telephone's component and assembly costs.

Because the most power-hungry part of a cellular or cordless telephone is its rf power amplifier, it is important that any move to lower supply voltages does not result in a significant loss of efficiency in the power amplifier. The power amplifier should operate at low voltage with high power-added efficiency – i.e. the ratio of rf output power to dc + rf input power. In order to keep component and assembly costs low, it should use as few gain stages as possible and the minimum number of peripheral components. To reduce test time, it should be alignment-free and provide predictable, reliable, performance.

Figure 1 illustrates how Philips Semiconductors' new wideband rf transistors can be used to meet these objectives in an rf power amplifier for DECT telephones. The design operates from a single 3.6V rail and includes bias circuitry for load power adjustment and on/off switching. In addition, it occupies less than 10 by 20 mm of a standard two-layer FR4 laminate pc board.

The amplifier delivers 26dBm output power, achieving a power gain of 29 dB and an overall power added efficiency in excess of 50%.

Amplifying rf power

RF power amplification is achieved using only two of the new wideband devices. Transistor Q_1 – a *BFG425W* – operates in class-A mode at a V_{CE} of 3V and a collector current of 30mA. Under these conditions it provides 18dB of gain and an output power level of

Table 1. Measured source and load impedances of the devices used in the DECT power amplifier.

Transistor	Source imp. (Ω)	Load imp. (Ω)	Conditions
BFG425W	(12+0.7j)	(52+102j)	V_{CE}=3.0V; I_C=30mA; f=1.9GHz
BFG21W	(9.1–9.5j)	(9.7–6.4j)	V_{CE}=3.6V; P_o=26dBm; f=1.9GHz

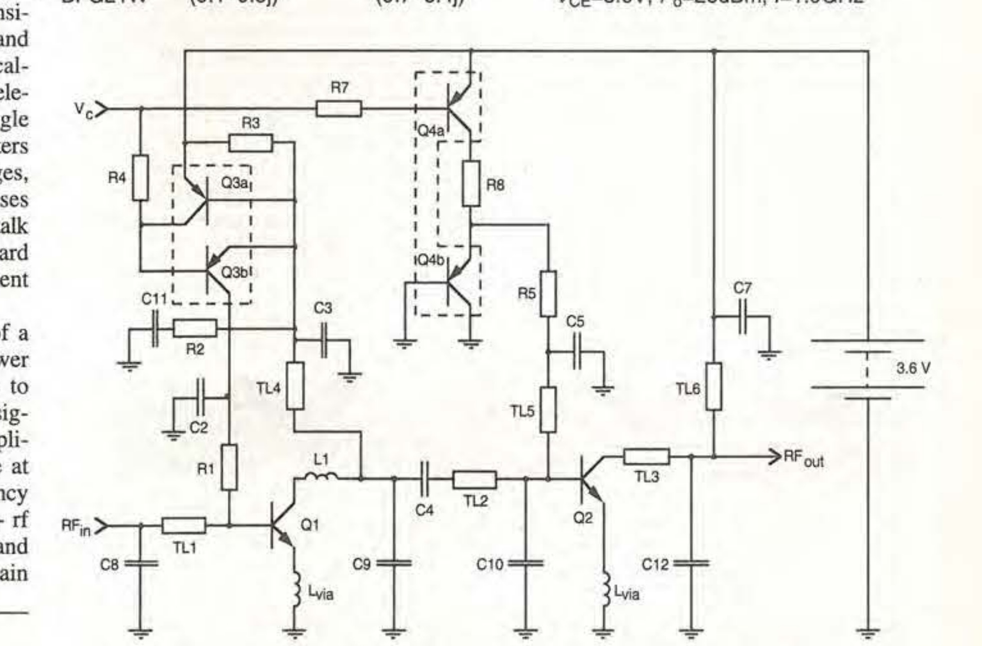


Fig. 1. DECT phone power amplifier using new fifth-generation rf wideband transistors.

Loek Colussi is with Philips Semiconductors' Product Group Transistors and Diodes in The Netherlands.

15dBm for –3 dBm rf input.

Transistor Q_2 is a *BFG21W* medium-power transistor operating in class-AB mode. It drives the telephone's antenna circuit directly. Biased to a base voltage of 0.7V, which results in a quiescent collector current of approximately 1mA, this transistor provides a power gain of 11dB and 26dBm output level.

Under these conditions its collector efficiency is typically 55%.

The measured source and load impedances of the transistors operating under the conditions described above appear in **Table 1**.

Impedance matching networks are therefore required to provide smooth 50Ω matching throughout the amplifier.

Impedance matching

The impedance matching part consists of three separate sections – the input, interstage and output matching networks. Its purpose is to enable the rf transistors to perform optimally with respect to power gain, output power and efficiency.

Fortunately, the inherent impedance levels of the *BFG425W* and *BFG21W* as indicated above are not exceptionally high or low, so they are quite easy to match.

At the input, shunt capacitor C_8 and series microstrip line TL_1 match the 50Ω rf source to the base of Q_1. Base resistor R_1 is used for biasing and has no effect on matching. Between the collector of Q_1 and the base of Q_2, matching is done by series inductor L_1, shunt capacitor C_9 and series transmission line TL_2.

If pcb area is not critical, L_1 can be replaced by a 3.5mm, 50Ω transmission line. Shunt capacitors C_9 and C_{10} partly compensate the influence of bias stubs TL_4 and TL_5, which are both $<\lambda/4$.

At the output side of Q_1, series network R_2/C_{11} is used to increase the k-factor of the first stage to avoid potential instability below 1GHz. The output match is done by series transmission line TL_3 and shunt capacitor C_{12}. Again this capacitor also compensates the influence of bias stub TL_6.

Biasing

The biasing part of the power amplifier incorporates a pair of *PUMT1* dual p-n-p transistors, $Q_{3,4}$. To define the collector current in Q_1, Q_{3a} compares the voltage across R_3 with the forward voltage of its base-emitter junction.

If current in R_3, i.e. Q_1's collector current, increases, Q_{3a} starts to conduct. This reduces the base drive to Q_{3b} which in turn reduces the base drive to Q_1, thereby stabilising Q_1's collector current.

For this circuit to work, control voltage V_C has to be fixed to ground. Voltage on the collector of Q_1 is always 0.6V lower than the supply rail, allowing a 3V collector voltage swing.

The base of class-AB output stage Q_2 is biased by a low impedance voltage source formed by Q_{4b}. The temperature coefficient of Q_{4b}'s base-emitter voltage is roughly the same as that for Q_2, maintaining a quiescent current of approximately 1mA in Q_2's collector – despite ambient temperature changes.

Resistor R_5 prevents thermal runaway of Q_2. Transistor Q_{4a} is driven by control input V_C to cut off the base drive to Q_2 during the interval between rf output pulses.

When used in pulsed mode at nominal supply voltage and output level, the load can be mismatched to a voltage/standing-wave ratio less than or equal to 6:1, in all phases, without damage. The power amplifier can also be operated in CW mode provided that 50Ω output matching can be guaranteed under all conditions.

It is possible to increase overall efficiency of the amplifier by a few percent by operating the *BFG425W* in class-AB mode rather than in class-A mode. This also simplifies the biasing circuitry. However, it has the negative effect of reducing the overall power gain, resulting in the need for a higher rf drive level in order to achieve the required 26dBm output power.

If a multi-layer pc board is used, the area occupied by the amplifier can be reduced by burying the transmission lines in the board. Space can be saved by placing the biasing circuitry on the reverse side of the pc board. Inductance of the vias which connect the emitter lead-outs of Q_1 and Q_2 to the ground plane on the reverse of the pc board must be kept lower than 0.1nH in order to maintain rf performance.

PHS-phone applications

With only minor modification this power amplifier is also suitable for use in PHS phones. Although they require a lower rf output power of 21dBm rather than 26dBm, such phones need better linearity performance. To achieve the required linearity, the collector current in the *BFG425W* is reduced to 20mA while the quiescent current through the *BFG21W* is increased to 10mA. In this way, both transistors operate on a more linear part of their gain characteristic.

The *BFG425W*'s collector current can be suitably decreased by increasing the value of R_3 to 22Ω. To increase the quiescent current in the *BFG21W*, the base potential of Q_{4b} is increased by adding a potential divider between the positive supply rail and ground. A divider comprising 330Ω to ground and 18kΩ to V_S works well, although other values can be used to achieve an optimum trade-off between linearity and efficiency. ■

Double-poly transistor technology

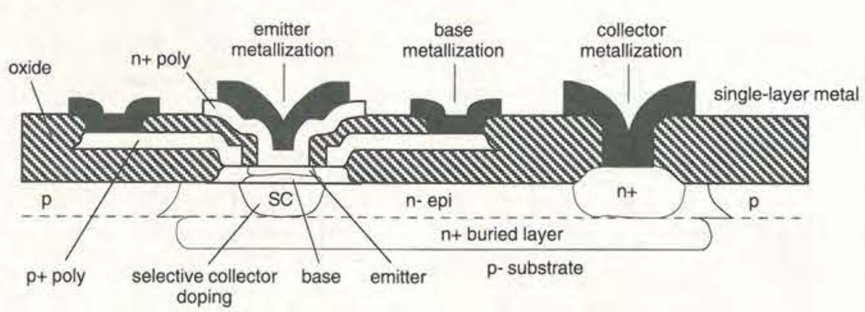
Philips Semiconductors' fifth-generation rf wideband transistors are based on a double-polysilicon buried-layer process that yields bipolar transistors with transition frequencies (f_T) in excess of 20GHz at low V_{CE} voltages. Typical power gains of 11dB at 2GHz allow these transistors to be used in the latest generation of digital cordless and cellular telephones – an application previously dominated by GaAs devices.

To produce bipolar transistors with cut-off frequencies above 20GHz that will operate at V_{CE} voltages of 3V or less, it is necessary to achieve base widths in the order of 100 nm. This is accomplished by using the double-polysilicon transistor structure illustrated below, in which deposited polysilicon is used for both the base and emitter connections.

Very steep doping profiles in the base and emitter regions create the very narrow base widths required for a high cut-off frequency, while sub-micron emitter widths of typically 0.5μm – made possible by the self-aligned nature of the process – ensure a high f_{max}. These sub-micron emitter widths also prevent current crowding effects and help to keep the base resistance low, thereby preventing degradation of power gain. Lateral connections to the base region by p+ polysilicon also help to reduce base resistance and minimise collector-base (Miller) capacitance.

The n+ layer required for the collector is buried within a p- substrate that is connected to the emitter. This enables the transistor die to be bonded directly to the transistor's emitter lead-outs, reducing its emitter inductance and the thermal resistance of the SOT343R plastic surface-mount package. Large area metallisations for the emitter allow the transistors to handle the high emitter current densities required in medium-power types.

Philips Semiconductors' family of fifth-generation rf wideband transistors currently includes five types. The *BFG403W*, *BFG410W* and *BFG425W* are small signal types optimised for maximum f_T at collector currents of 3mA, 10mA and 25mA respectively. They feature gains of over 20dB at 2GHz and noise figures as low as 1.2dB. The two medium power transistors in the range are the *BFG480W* and *BFG21W*, which have their maximum f_T values at collector currents of 80mA and 250mA respectively. They provide power gains at a frequency of 2GHz and a V_{CE} of 3.6V in excess of 14dB and 11dB respectively. Both types achieve typical power added efficiencies greater than 60%.



New wideband bipolar transistors are based on a double-polysilicon buried-layer process yielding transition frequencies (f_T) in excess of 20GHz combined with low V_{CE} voltages.

Components for the DECT power amplifier.

R_1	560
R_2	10
R_3	18
R_4	100k
R_5	10
R_6	not required
R_7	10k
R_8	180
L_1	1n8
C_1	not required
C_2	10n
C_3	8p2
C_4	8p2
C_5	8p2
C_6	not required
C_7	8p2
C_8	1p8
C_9	1p8
C_{10}	2p7
C_{11}	10n
C_{12}	2p7
Q_1	BFG425W
Q_2	BFG21W
Q_3	PUMT1
Q_4	PUMT1
TL_1	Length 6.5mm; Width 0.5mm
TL_2	Length 3.0mm; Width 1.2mm
TL_3	Length 4.5mm; Width 0.5mm
TL_4	Length 7.5mm; Width 0.2mm
TL_5	Length 7.5mm; Width 0.2mm
TL_6	Length 6.5mm; Width 0.2mm

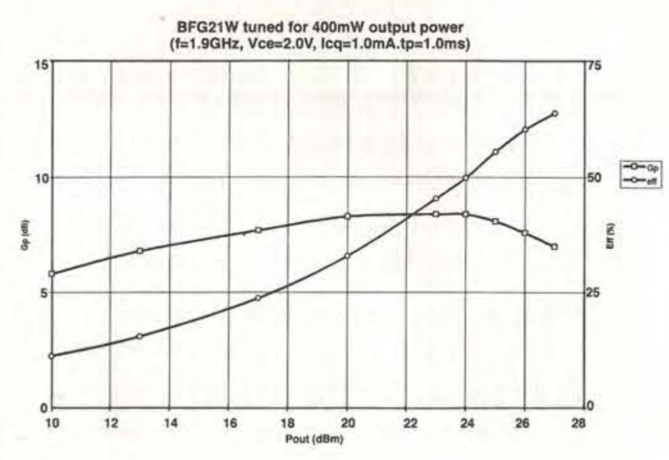
Power gain – even at 2V

Experiments using a scaled-up version of Philips Semiconductors' *BFG21W* double polysilicon rf transistor indicate that this technology can be used to design DECT power amplifiers that operate from 2.4V battery packs.

As indicated in **a)** below, the transistors tested provide a power gain, G_p, of 7dB at the required 27dBm DECT output power level when operating with a V_{CE} of only 2.0 V. Equally impressive, their power added efficiency at this output level is almost 64%, allowing 2.4V DECT telephones to achieve long standby and talk times.

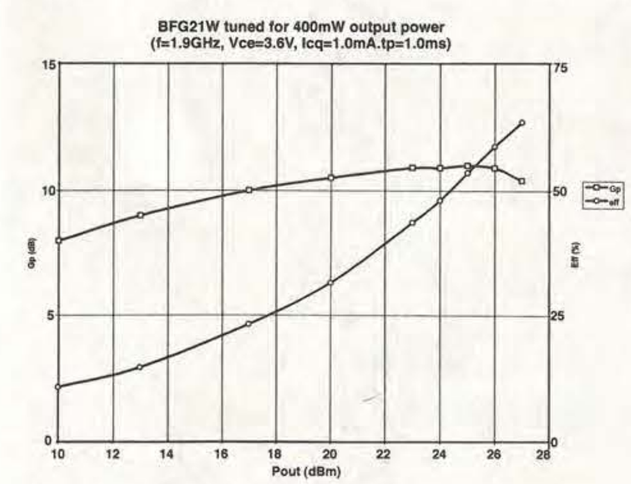
To compensate for the lower power gain of the output stage at 2.0V (7dB compared to the 10.5dB shown in **b)** for a V_{CE} of 3.6V), power amplifiers that use these double-polysilicon transistors would require three gain stages rather than the two described in the main article above. However, the driver and pre-driver stages would each need to provide only 10 to 15dB of power gain in order for the amplifier to be driven at –5dBm.

Detailed test results on the the transistors used in this evaluation can be obtained from Philips Semiconductors' Transistors and Diodes Product Group in Nijmegen, The Netherlands.



Gain and Efficiency vs. P_{out}
f=1.9GHz; Vs=2.0V; Icq=1mA; tp=1.0ms;
tuned for 26dBm output power

P_{out}(dBm)	P_{in} (dBm)	I_c (mA)	G_p (dB)	Eff (%)
10	4.2	45	5.8	11.11111
13	6.2	65	6.8	15.34817
17	9.3	106	7.7	23.64091
20	11.7	152	8.3	32.89474
23	14.6	220	8.4	45.34687
24	15.6	252	8.4	49.83902
25	16.9	285	8.1	55.47856
26	18.4	330	7.6	60.31927
26.5	19.5	350	7	63.81194



Gain and Efficiency vs. P_{out}
f=1.9GHz; Vs=3.6V; Icq=1mA; tp=1.0ms;
tuned for 26dBm output power

P_{out}(dBm)	P_{in} (dBm)	I_c (mA)	G_p (dB)	Eff (%)
10	2	26	8	10.684
13	4	38	9	14.585
17	7	60	10	23.203
20	9.5	88	10.5	31.566
23	12.1	127	10.9	43.641
24	13.1	145	10.9	48.12
25	14	164	11	53.562
26	15.1	188	10.9	58.822
27	16.6	219	10.4	63.57

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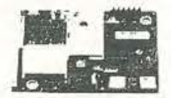

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Alternative inverter drive

Linear power output stages are at their most efficient when driving a rail-to-rail square wave. Conventional motors on the other hand prefer a sine-wave drive. Irving Gottleib describes how to get the best of both worlds in an unconventional way.

For some applications, a sine-wave is preferable to the square-wave output delivered by most dc-to-ac inverters. Among other things, square waves can roughen the torque characteristics of motors and they increase hysteresis and eddy-current losses. Also, the harmonic content of the square-wave format tends to agitate electromagnetic and radio-frequency interference problems.

On the other hand, a switching circuit generating square wave power is noted for high efficiency, since it allows the switching transistors to operate with minimal thermal stress. Obviously it would be nice to retain the square-wave switcher, but at the same time obtain sinusoidal output.

In **Fig. 1a)** is a basic saturable-core oscillator. This particular circuit makes use of an auto-transformer winding, and the switching transistors operate in the common-collector mode. Any of the other saturable-core oscillator circuits would be equally satisfactory for our purposes. In Fig. **1b)**, a band-pass filter is associated with the output winding to produce a sine-wave. Sometimes, a simpler low-pass filter is similarly used, but it is then more difficult to get a good quality sine-wave.

A further technique is depicted in Fig. **1c)**. The inclusion of the large inductor, *L*, enables the output winding to be resonated. Although the transistors still operate as a square-wave switching circuit, the desired sine-wave output is obtained. Noted that it would not be feasible to tune the output winding of the basic inverter circuit of Fig. **1a**.

The driven inverter of Fig. **1d** is a class-B amplifier. This has fairly-good possibilities, but you should be prepared to cope with crossover distortion and with higher transistor dissipation than in the self-excited switching circuits.

A different approach

Yet another approach to the problem makes use of parametric phenomena in magnetic cores. Briefly stated, voltage can be induced in the secondary of a transformer via variation in inductance, as well as variation in flux linkage.

You won't find much mention of this in traditional engineering texts though. This is because it is usually assumed that transformers are designed and operated to function over the essentially linear region of their magnetisation curves. Such operation minimises hysteresis loss and maximises efficiency.

You know, however, that violent non-linearity is to be found in saturable-core inverter transformers. As magnetic saturation approaches in these cores, permeability rapidly decreases, as does the inductance of associated windings.

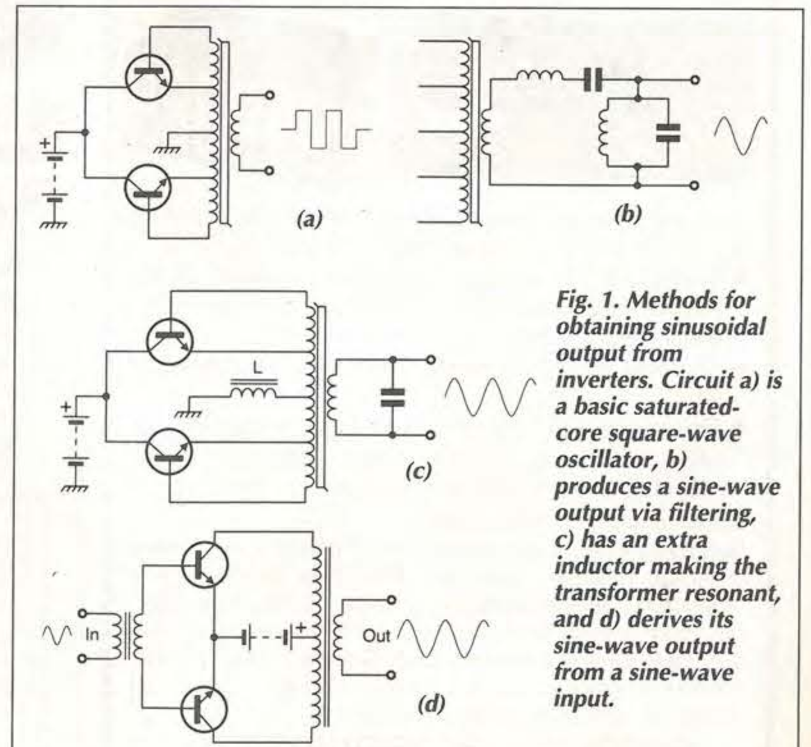

Fig. 1. Methods for obtaining sinusoidal output from inverters. Circuit a) is a basic saturated-core square-wave oscillator, b) produces a sine-wave output via filtering, c) has an extra inductor making the transformer resonant, and d) derives its sine-wave output from a sine-wave input.

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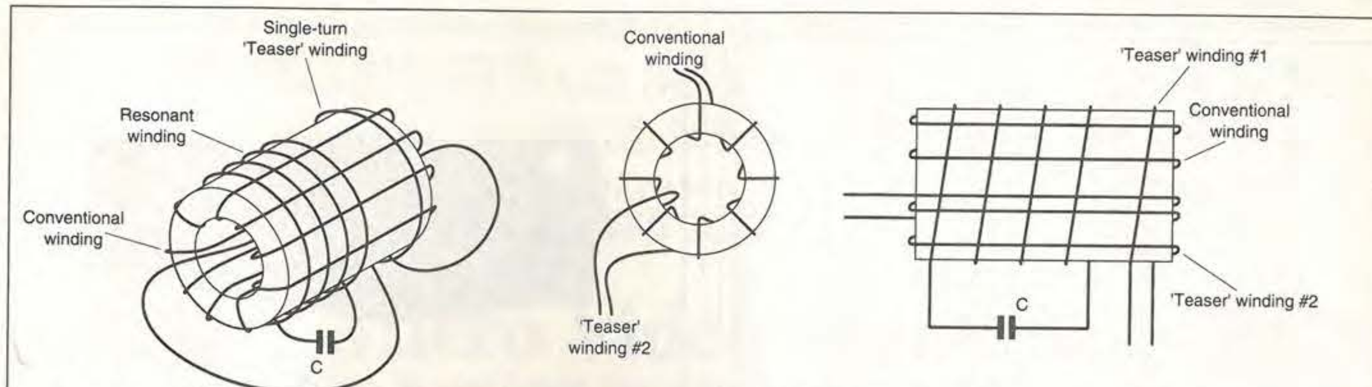

Fig. 2. Modified toroidal transformer for producing sine-waves. You start with a conventional winding, as used in saturable-core oscillators, then add windings as shown. a) is a perspective view, b) is the front view showing how one lead of the 'teaser' winding goes through the solenoid and c) is the side view.

Can this relationship be put to practical use?

In the sketches of **Fig. 2**, the salient feature of the modified toroidal transformer is the introduction of a resonant secondary winding. Note that this winding is placed *over* the outer rim of the toroid.

The spatial relationship is such that no ordinary mutual flux linkage exists between the conventional primary winding and this unconventional secondary winding. Rather, the new winding senses the changing inductance of the core. This results in the parametrically induced emf. You can also think of this resonant winding as a shock-excited oscillator.

As you can see, the modification involves a bit more than just the resonant winding. Additionally, two single-turn links are used to couple the primary and secondaries by ordinary electromagnetic means. This enhances energy transfer. Overall then, the modified transformer uses both flux-cutting and inductance change to transfer energy. The links are the so-called 'teaser' windings. A single pass through the hole of the toroid comprises a single turn.

Does this idea worry you?

The unorthodox configuration of the modified transformer could, understandably, upset those of you used to more conventional formats.

The schematic diagram of **Fig. 3** should help clarify matters. Here the X between the conventionally wound primary winding and the added resonant winding symbolises the lack of ordinary electromagnetic coupling between these windings.

As I pointed out, the absence of such flux-cutting energy transfer is brought about by the spatial orientation of these two windings. This brings us to the single-turn teaser windings which are geometrically arranged so as to promote some coupling via ordinary mutual induction. Thus, input and output windings are also link-coupled.

To many practitioners, an interesting aspect of this scheme is that it calls for a bit of experimentation. Clearly, some kind of average value of inductance must be involved in the tuned output circuit. And although the Q of this resonant tank must necessarily impact both energy transfer and wave purity, it is not easy to quantify things for general applications.

I conducted investigations with a nominally 20W inverter at several tens of kilohertz; I obtained a very good sinusoidal output and I felt that the use of appropriate scaling factors should enable operation at other power levels and at other frequencies.

It may be wise first to get the feel of this unusual circuitry and then proceed empirically in tailoring the resonant winding and the *L/C* ratio to conform to your specific needs. Also, if you already have an operational inverter using a saturating toroidal output transformer, much time and effort can be saved by placing the new winding(s) on this toroid.

At first attempt, about the same number of turns should be used for the resonant winding as the total number of turns on the primary winding. Then, one or two decade capacitor boxes will facilitate search for resonance. An oscilloscope is particularly useful in as much as one can observe both magnitude and waveshape.

Several things should be born in mind in interpreting results. You may encounter sub-multiple resonances, but none of these will compare in magnitude and wave-purity with the true resonance of the fundamental oscillation frequency.

To a considerable extent, energy transfer will improve with the Q of the resonant windings. This in turn corresponds to a high ratio of C to L. Resonant impedance of a parallel-resonant LC tank is given in ohms by $\sqrt{(L/C)}$ with L expressed in henries and C in farads; high Q implies low impedance.

A load resistor connected across the resonant output winding of such value that its presence reduces the amplitude of the sine-wave to half its unloaded value establishes the output impedance. Initially, at least, you should aim for an output impedance of about 250Ω.

To secure the output voltage you need, experimental flexibility is well served by employing either taps or an auto-transformer addition of a few turns. Clearly, Q, energy transfer, voltage, and output impedance are all interrelated and that optimisation for the requirements of a particular application can be an experimenter's delight. At the same time, the basic operation is readily forthcoming, being neither elusive nor critical.

Once optimised, this scheme is likely to compel selection over the other techniques in matters of cost, board surface area, and wave-purity. ■

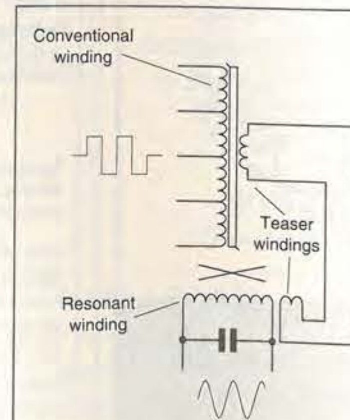

Fig. 3. Schematic of the modified toroidal transformer. The resonant winding is wound around the outer rim of the toroid. One 'teaser' winding is a single-turn link adjacent to, or over, the resonant winding. The other teaser winding is a single turn link formed by a single pass through the hole of the toroid. The conventional winding is the commonly used toroidal winding arranged for operation with push-pull switching transistors.

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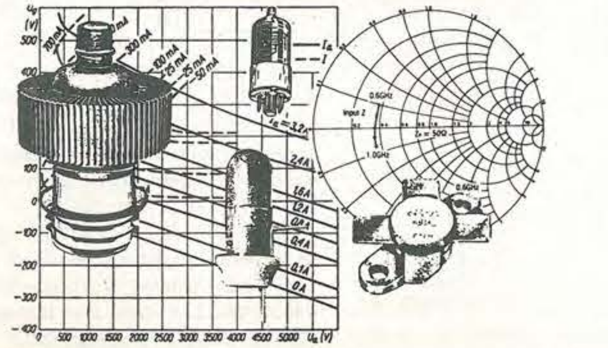

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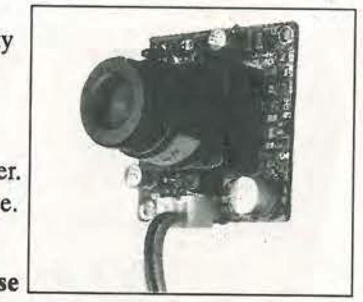

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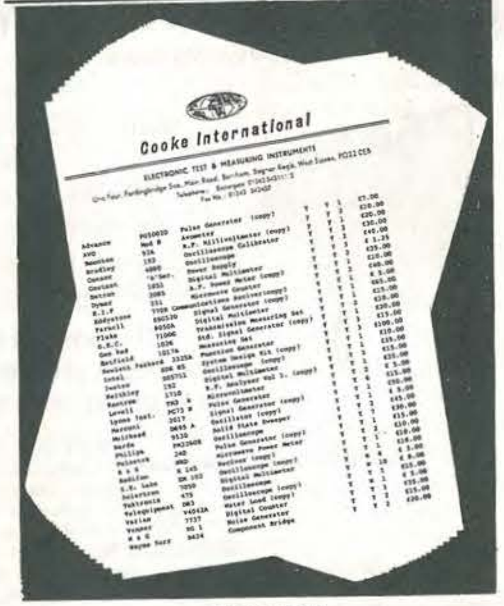

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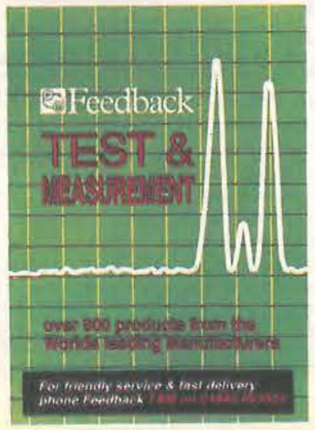

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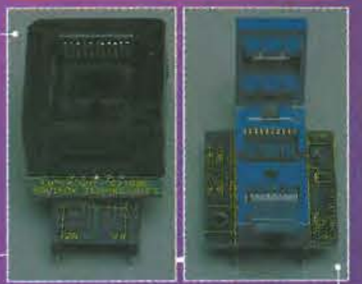

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RAM (bytes)	128	128	256	256	256	256	128	64
EEPROM	-	-	-	-	-	2K	-	-
In-system re-programmable	-	-	-	-	-	YES	-	-
I/O Pins	32	32	32	32	32	32	15	15
16-bit Timer/Counters	2	2	3	3	3	3	2	1
Watchdog timer	-	-	-	-	-	YES	-	-
Interrupt sources	6	6	8	8	8	9	6	3
Serial UART (full duplex)	YES	YES	YES	YES	YES	YES	YES	-
SPI Interface	-	-	-	-	-	YES	-	-
Analogue comparator	-	-	-	-	-	-	YES	YES
Data pointers	1	1	1	1	1	2	1	1
Package Pins (DIL)	40	40	40	40	40	40	20	20

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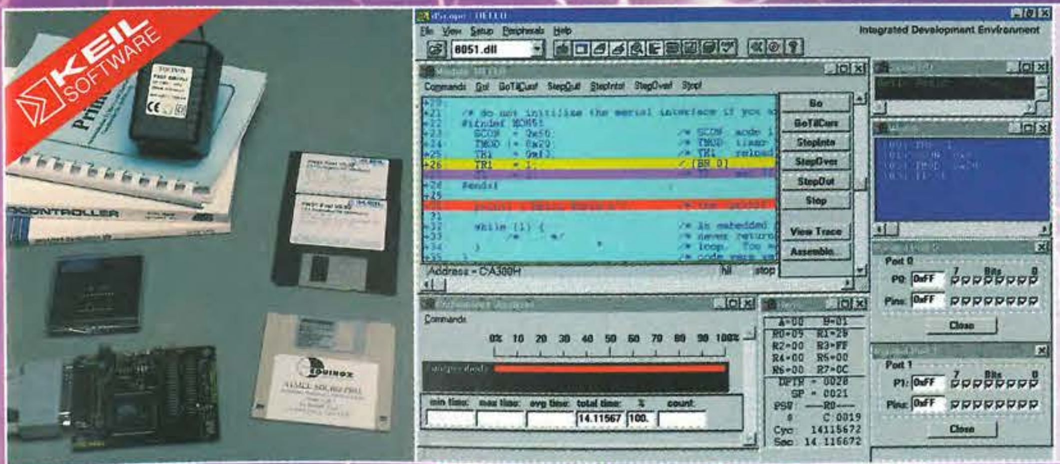

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