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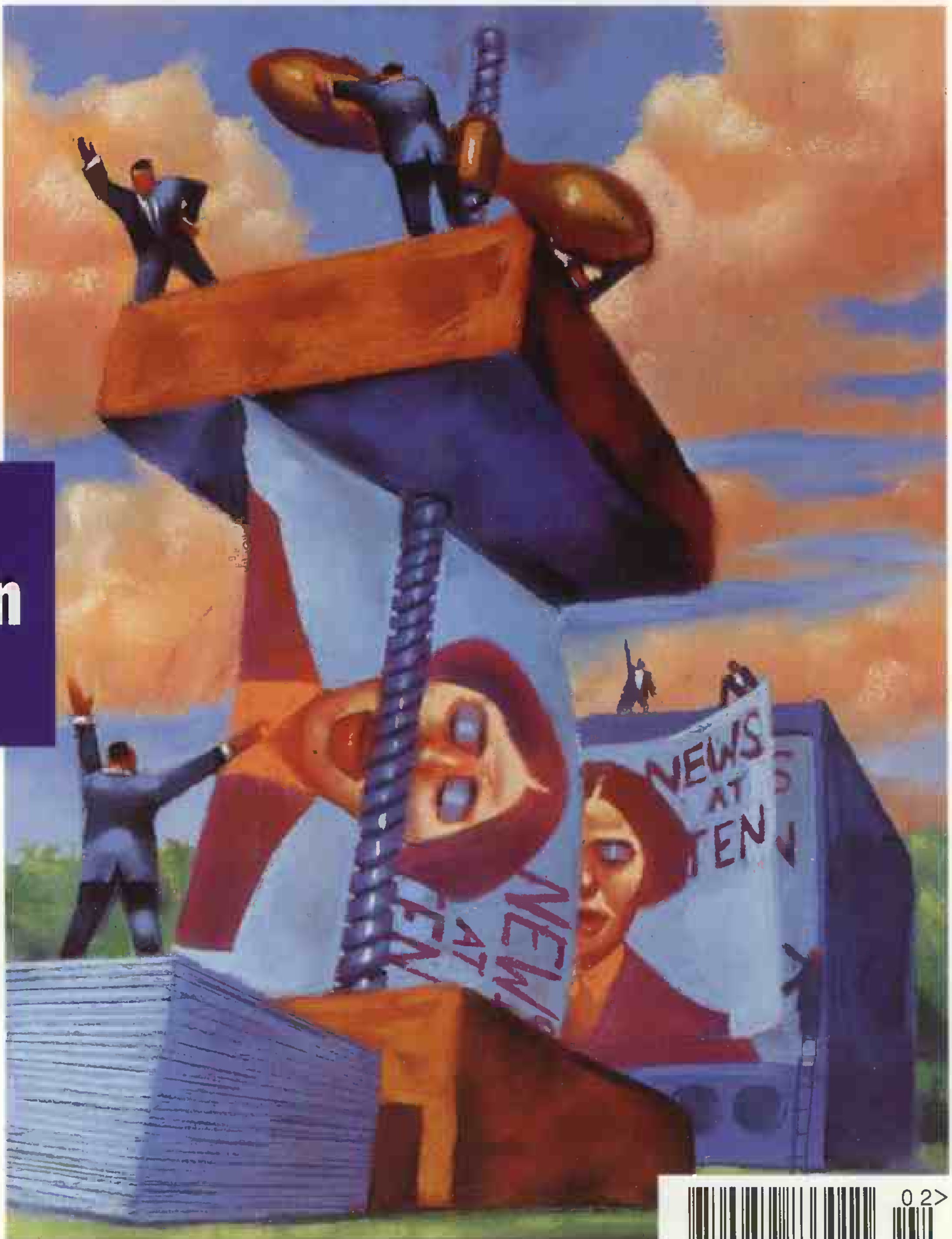


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***Special offer 25% discount on PC interfacing Radio-Code receivers***

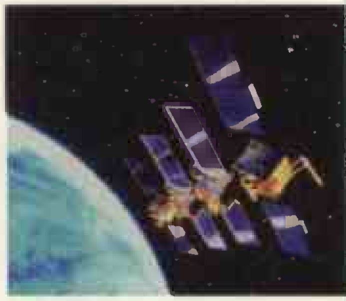
*Read Radio-Code time transmissions into your PC, and control with atomic time precision, page 129.*



# Regulars

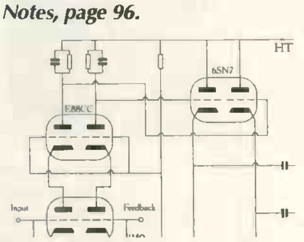


*Cover – Hashim Akib*



*For the first time, satellites have communicated without the intervention of a ground station – Update, page 92.*

*New laser-based sensor promises huge savings for steel makers – Research Notes, page 96.*




*Morgan Jones discusses designing valve audio power amplifiers capable of delivering more than the usual 10W – page 110.*

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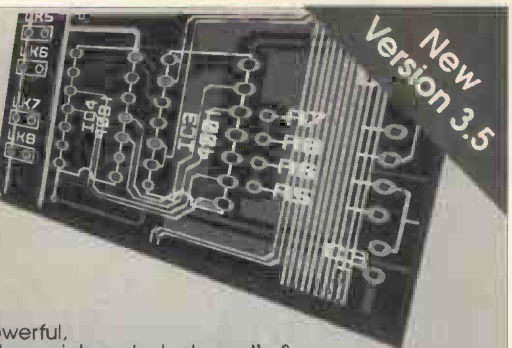


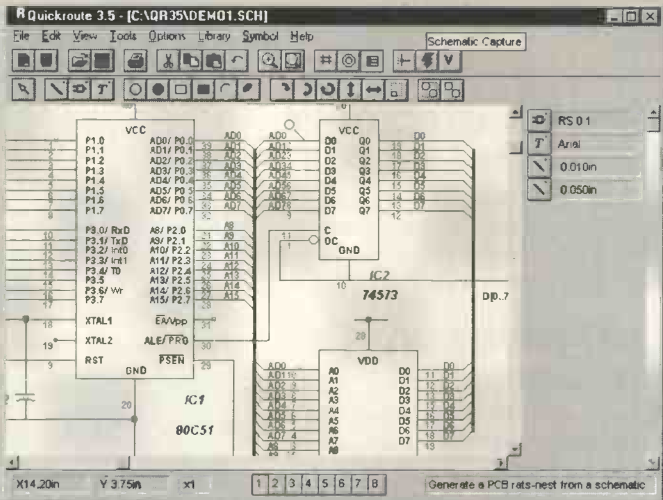


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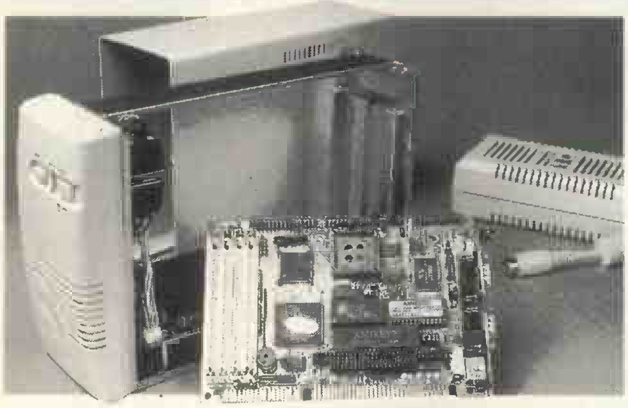

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# Sound reasons for digital tv?

Unveiling the Government's Broadcasting Bill at the end of last year, Heritage Secretary Virginia Bottomley proudly proclaimed that "the rest of the world is watching," because Britain "will be the first major national market to go digital," and "we must look forward to a golden future".

Will this perhaps be a future in which Ms Bottomley's Press Office learns to use new fangled devices like the telephone and fax machine? My invitation to Ms B's briefing arrived by Christmas post – a day after the event. Others in the specialist press never heard anything.

Perhaps it was intentional. We might have spoiled Ms B's day by asking whether the British public will pay money to install the equipment they will need to pay more money to view the new digital services?

When Britain's two satellite services launched around five years ago, both Sky and BSB made the mistake of assuming that people would go out and buy a receiver, and then pay for a dish to be installed. BSB also assumed that viewers would pay a premium of several hundred pounds for MAC instead of PAL.

Rupert Murdoch quickly woke up to reality, branded the brown goods trade "unable to sell a toaster" and employed hit squads to knock on doors, and install Sky systems on free trial. By the time BSB had stopped snoring, and started to organise one-stop installation, the station was collapsing into merger.

For at least a year, David Elstein of Sky has been talking about Sky's plans to transmit more channels, using digital technology. But there are clear signs that his station is now getting cold feet. Even if Sky could afford to give every viewer a free digital receiver – at a starting cost of several hundred pounds – the company would also have to re-employ the hit squads. They would then go up ladders and try to loosen corroded bolts to replace existing LNBs with Universal models that can receive Astra's new high frequency digital transmissions.

The BBC wants to lead the UK – and Europe, and the world – into digital terrestrial television, starting in 1997. At a recent briefing Director General John Birt compared the transition from analogue to digital tv, with the switch from 405 line black and white to 625 line colour. DTTV offers wide screen pictures and digital sound, he revealed to us.

Perhaps someone inside the BBC should tell Boss Birt that his existing analogue tv transmissions already carry Nicam digital stereo sound to 87% of the viewing population, and the other 13% is hoping they will one day get it too.

Birt also seems blissfully ignorant of the debate on wide screen television which has been running since long before the BSB fiasco, and Channel 4's recent decision to cut back on PAL Plus transmission. The British public has been offered wide screen sets for many years and steadfastly refuses to buy them. Indeed Birt admitted that although he has seen a widescreen set, he does not actually own one.

The parallel with the 405/625 line transition is fatuous. It took twenty years, not the fifteen Birt claims, and offered a completely new and dramatic upgrade – colour – at a time when there was very little electronic gadgetry in the shops. Today, consumers are sick to death of new developments and need a good reason to replace existing equipment.

Michael Grade of Channel 4 recently reassured on radio that viewers will not need a new tv set for DTTV, just a set-top digital adaptor. Perhaps one of Grade's staff could ask him how this will give the widescreen pictures which John Birt thinks viewers will see as the reason to go digital.

Presumably someone inside the BBC has a clearer vision than the boss. Hence the BBC's call for the Government to set a firm date for an analogue switch off. This will bully viewers into buying digital equipment whenever they make a new purchase. Based on the fallacious fifteen year 405/625 change-over reference point, the BBC suggests a 15 year gate on analogue services. An extraordinary policy paper, submitted to the Heritage Department by Keith Boyfield and Brian Sturgess, suggests switch-off by 2006, just ten years away. As modern homes are full of tv sets and VCRs that can easily last ten years, this bright idea would succeed mainly in killing sales of all existing analogue equipment.

**"...we must look forward to a golden future"**

Ms Bottomley has settled for a review when half of Britain's households can get digital tv. If this means 50% transmission coverage it is a daft idea. If it means 10 million homes fully kitted out with digital tvs and vcrs, it compares with what happened over 405 shutdown, and makes sense. It's another of those questions the specialist press would have asked Ms B if her staff had learned to use a phone or fax machine.

There is, of course, a hidden agenda in all this. As broadcasters move from analogue to digital transmission, they release frequencies which the government can then sell. Anyone who can rush consumers into buying digital equipment will definitely be in the running for a big thank-you from the Treasury – and probably a nice little knighthood to go with it.

***Barry Fox***


**EDITOR**
Martin Eccles
0181 652 3128

**EDITORIAL ASSISTANT**
Rob Allcock
0181 652 8638

**CONSULTANTS**
Jonathan Campbell
Philip Darrington
Frank Ogden

**DESIGN & PRODUCTION**
Alan Kerr

**EDITORIAL ADMINISTRATION**
Jackie Lowe
0181-652 3614

**E-MAIL ORDERS**
jackie.lowe@rbp.co.uk

**ADVERTISEMENT MANAGER**
Richard Napier
0181-652 3620

**DISPLAY SALES EXECUTIVE**
Malcolm Wells
0181-652 3620

**ADVERTISING PRODUCTION**
Christina Budd
0181-652 8355

**PUBLISHER**
Mick Elliott

**EDITORIAL FAX**
0181-652 8956

**CLASSIFIED FAX**
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**Newstrade:** Distributed by Marketforce (UK) Ltd, 247 Tottenham Court Road London W1P 0AU 0171 261-5108.
**Subscriptions:** Quadrant Subscription Services, Oakfield House, Perrymount Road, Haywards Heath, Sussex RH16 3DH. Telephone 01444 445566. Please notify change of address. Subscription rates 1 year £30 UK 2 years £48.00 3 years £70.00. Surface mail 1 year £35.00 2 years £56.00 3 years £80.00 Air mail Europe/Eu 1 year £43.00 2 years £68.00 ROW 1 year £52.00 2 years £83.00
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**Printed** by BPCC Magazines (Carlisle) Ltd, Newtown Trading Estate, Carlisle. Cumbria, CA2 7NR
Typeset by Wace Publication Imaging 2-4 Powerscroft Road, Sidcup, Kent DA14 5DT

# Brains forced overseas

A lack of vision at the UK Medical Research Council has driven overseas a multi-disciplinary group doing fundamental research on the brain.

The MRC refused to renew the position of the group's computer scientist jeopardising its future.

Late last year, the group moved to the specially established Institute of Neuro-informatics in Zurich, despite it having secured long-term funding and repeatedly expressing its wish to remain in the UK.

Group leader Kevan Martin says he feels no bitterness towards the MRC but questions its decision making.

"Its vision is to the floor rather than the horizon," said Martin. "The MRC unit is directly supported by government; its whole purpose is to support high risk, long term research, and that is exactly what we are doing."

The MRC refused to comment.

The group is developing a hybrid digital-analogue vlsi architecture to model the neocortex – the bulk of the circuitry that makes up the brain.

"The neocortex is a rather fantastic general purpose processor, adaptable to many functions," said Martin. "Understand its workings and you have achieved something fundamental."

The unit's achievements to date include collaborative work with the California Institute of Technology to create a silicon neuron that resembles closely the biological neuro cell.

And the unit has also constructed analogue vlsi devices that incorporate multiple neurons.

The MRC had wanted the group to pursue a more biological slant for the work. This meant there would be no funding for the analogue VLSI work.

**Roy Rubenstein**, *Electronics Weekly*

# Wired for Sea – electronics at the Boat Show

Sailing – that simple, centuries-old communion with wind and tide – is now one of the most high-tech sports in the world, and getting ever higher.

Nowhere is this more apparent than in round-the-world yacht racing, which communications companies seem to regard as an appropriate metaphor for their activities, as well as a proving-ground for their technologies. BT, having provided race communications for the Whitbread races, is now sponsoring one of its own, and took one of the largest stands at this year's Boat Show to proclaim the fact.

The BT Global Challenge involves 14 identical 67ft yachts racing the 'wrong way' – i.e against prevailing winds and currents – around the world. Two of the boats are sponsored by other electronics companies, Motorola and Toshiba, and a third by a consortium of other telecomms companies.

Unluckily for BT, the fact that the hulls are made of steel prevents a repeat of the digitally-compressed tv transmissions which were tried during the last Whitbread. These used Inmarsat A, which needs a much larger antenna – too big to mount away from the deck. This time the fleet will stay in touch via C-Sat, which provides e-mail and Internet, via a conical aerial no bigger than a vacuum flask, and hf ssb radio through Portishead.

Leisure sailors meanwhile are becoming increasingly reliant on electronics. Arguably the most significant launch product at this year's Boat Show was Admiralty charts on CDrom. ARCS – the Admiralty Raster Chart Service – as it is called, provides straightforward digital reproductions of the familiar paper charts, for use with pc-based navigating systems.

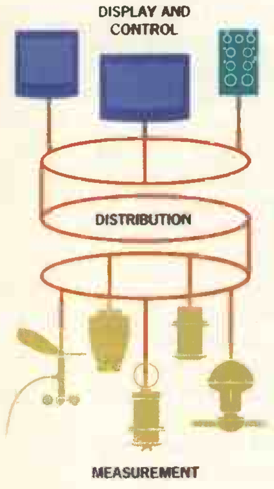
Two forms of display are possible: a 'life-size' close-up of a portion of the chart or a low-resolution overview of the whole thing. As with the paper charts, different scales with differing amounts of detail are available.

Regular updating is available, also on CDroms, issued weekly, but with cumulative information so that just one set of data need be patched in, perhaps just prior to the summer cruise. Charts cost £30 each; update CDroms £10. One chart occupiees about 1Mb of hard-disk space.

ARCS is already supported by PC Maritime's Navmaster Windows-based system, alongside its existing vector-based Livechart cartography.

Elsewhere at the show, integration was this year's great leap forward. Surprisingly, it is only 21 years since the first autopilot was launched. Since then GPS has arrived and wind speed/direction, compass bearing and depth sounder have all been digitised. The result is – or can be – a proliferation of displays and complex harness of wires around the boat. The solution is an integrated system using a single cable with multifunction displays, which is what Autohelm has come up with in in ST80, the latest development of its SeaTalk System. Navico meanwhile has enhanced its own, similar Corus system with an inboard autopilot, Oceanpilot, capable of translating all this information into intelligent coursekeeping.

And if you should fall overboard, make sure you have your PLB7 with you. This tiny device by Sea-Marshall is an electronic beacon, transmitting on the Search and Rescue frequency, which will help your own boat, or a rescue aircraft to find you quickly.




*Above: The four main elements of this on-board information system for navigation at sea are measurement, distribution, display and control. Unlike traditional alternatives, this system keeps the four key elements entirely separate. The benefits: flexible information displays instead of fixed-role instruments, plus choice in the way you put together your boat information system.*


*Right: General overview of the ST80 as installed on the boat.*

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The excellent performance of modern cassette recorders depends totally on the quality of the R/P head.Even the slightest amount of wear can impair the frequency response and distortion levels. Our HC80 is atop quality head from one of the foremost manufacturers in Japan, easily fitted to most standard stereo recorders (except Sony) and will transform the performance over a worn head. Only the fact that we buy these in vast quantities enables us to offer them at the amazing price of only **£11.70** each or 2 for **£17.60.**
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845-008 100g. Reel Special Valve Grade, 20swg .......... £12.90
845-009 100g. Precision PCB Grade, 22swg .......... £14.75
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## CMOS gates switch in 15ps

Toshiba has developed a 0.15μm gate c-mos structure based on a single gate that has a delay of only 15.4ps. This is claimed to be comparable to speeds normally achieved using a 0.1μm c-mos process. Use of a single gate structure was introduced by Toshiba's ULSI Research Laboratory to tackle the cost problem of mass producing the conventional c-mos dual gate structure of N-type and P-type polysilicon for the NMOS and PMOS gate electrode materials.

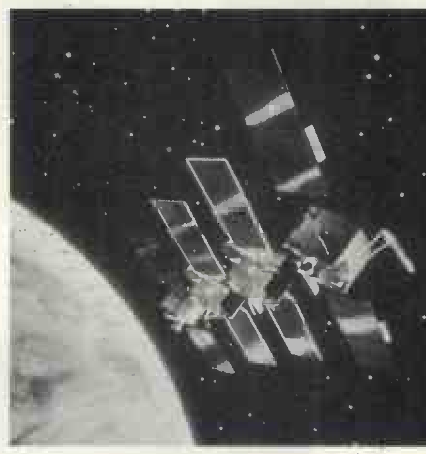
# Satellites communicate for the first time

For the first time, satellites have communicated with each other without the intervention of a ground station. The inaugural message was transmitted between Virginia and Hawaii via two Milstar military communication satellites.

Motorola's satellite mobile phone system Iridium is based on military satellite technology. Iridium satellites are in such low orbits that they can only 'see' a small part of the earth at any one time. To avoid the need for a large number of ground stations, Iridium messages will be relayed between satellites to a satellite with a view of a ground station.

● Japanese and American scientists have used a laser to establish two way communication between a satellite and Earth for the first time.

The US Jet Propulsion Laboratory used a 10W argon laser based at its Table Mountain telescope in California to communicate with the experimental satellite ETV-VI, run by Japan's Communications Research Laboratory (CRL). Results from the test will be used to upgrade the optics on CRL's OICETS satellite due to be launched in 1998. The aim of the research is to let satellites in low earth and geostationary orbits communicate with each other optically.



# Semiconductor makers could face polysilicon shortage

US market research firm Dataquest is predicting a shortage of polysilicon as semiconductor demand continues to grow and production capacity lags demand.

The shortage is predicted to hit semiconductor manufacturers in mid to late 1996 and could last as long as 10 months. But new polysilicon market capacity is reckoned to come online in 1997, which will let it catch up with demand.

"Silicon companies will have high market pricing power throughout the rest of this decade, and particularly in 1996 and 1997," said Clark Fuhs, principal analyst for Dataquest. "For this reason we believe the silicon industry, which has been a historically lower-margin industrial business, will migrate to a business model that more closely resembles the other segments of the semiconductor ecosystem."

Dataquest also predicts a shortage of 200mm wafers as companies prepare new fabs that use the larger format size. The 200mm wafer shortage will begin in 1996 and will continue for much of the rest of the decade, peaking in 1997 and then again in 1999.

The firm estimates that 1995 demand for 200mm wafers worldwide is 1.280m wafers a month. This demand will almost double next year to reach 2.206m wafers and the industry will require 5.213m wafers a month by 2000. However, the industry will only be able to supply 1.956m wafers next year and 3.894m wafers a month by the end of the decade. Dataquest points out that there is a mismatch of supply and demand. This has created a shortage of 100 and 125mm wafer sizes, which should soon be solved.

And it predicts the most serious long term shortages will be in the supply of 150mm wafers.

## Researchers achieve 400Gbit/s down a 100km fibre

Japanese researchers have transmitted digital data at a 400Gbit/s down 100km of optical fibre for the first time. Telecoms developer Nippon Telegraph and Telephone (NTT) has achieved this data throughput by multiplexing four light signals of different wavelengths each carrying a 100Gbit/s data stream. As part of the experiment NTT demonstrated in the laboratory a phase-locked loop timing extraction circuit capable of working at 6.3GHz and a 200Mbit/s optical time division demultiplexer.

The optical receiver must synchronise the incoming data stream and a timing extraction circuit is used to derive the clock signal from the data. NTT used a prescaled PLL which relies on the nonlinearity of a semiconductor laser amplifier to extract a 6.3GHz clock signal which is scaled up to the original 100Gbit/s signal.

In the receiver the optical time division demultiplexer splits the 100Gbit/s signal into 16 separate 6.3Gbit/s streams.

## Pace could loose its set-top lead to Pioneer

Japanese consumer electronics company Pioneer, is poised to snatch the title of 'leading digital set-top maker in Europe' from under Pace Micro Technology's nose. Pioneer plans to mass produce boxes in Europe in early 1997 and does not feel that Pace poses a threat on the European market.

"Pace has been successful in the Asian area and Australia, but its success in Europe has been limited," said Stuart Liddle, business development manager.

However, Pace is not phased by Pioneer's claims, although its current market share of 95 per cent may soon be affected by it.

"We have seen Pioneer's announcements, but we can't comment," said one Pace spokesman.

Pioneer plans to ramp up production to up to 2 million units per year by 1998, with prices close £400. Its manufacturing of set-top boxes in Europe could involve the UK.

"It is very likely our manufacturing will be in Belgium but as far as the UK is concerned, at the moment it is still up in the air," said Liddle.

## BRT to cost Racal £30m pa

Racal is faced with spending hundreds of millions of pounds to modernise British Rail Telecommunications' (BRT) network.

Racal is to pay £132.75m acquiring BRT, which needs around £30m a year for the next three years to keep its existing PDH network up to scratch.

"BRT requires £30m per annum for the next three years, mainly for supporting network resilience and serviceability," said Rupert Hunte, strategic business development director for Racal Network Services, which is to incorporate BRT. ■

Sveltana Josifovska, *Electronics Weekly*

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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.
HP8651A RF Oscillator 22KC/S – 22Mc/s.
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.
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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 £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.
HP8614A signal generator 800Mc/s – 2.4GHz, new colour £400.
HP8616A signal gen 1.8GHz – 4.5GHz, new colour £400.
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HP 3336A or B syn level generator – £500-£600.
HP 3586B or C selective level meter – £750-£1000.
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HP 8683D S/G microwave 2.3 – 13GHz – opt 001 – 003 – £4.5k.
HP 8660 A-B-C syn S/G. AM + FM + 10Kc/s to 110Mc/s PI – 1Mc/s to 1300Mc/s – 1Mc/s to 2600Mc/s – £500-£2000.
HP 8640B S/G AM-FM 512Mc/s or 1024Mc/s. Opt 001 or 002 or 003 – £800-£1250.
HP 86222BX Sweep PI – 01 – 2.4GHz + ATT – £1750.
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HP 86290B Sweep PI – 2 – 18GHz – £1250.
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HP 8620C Mainframe – £250. IEEE – £500.
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HP 8601A Sweep generator .1 – 110Mc/s – £300.
HP 3488A HP – IB switch control unit – £500 + control modules various – £175 each.
HP 8160A 50Mc/s programmable pulse generator – £1000.
HP 853A MF ANZ – £1.5k.
HP 8349A Microwave Amp 2 – 20GHz Solid state – £1500
HP 3585A Analyser 20Hz – 40Mc/s – £4k.
HP 8569B Analyser .01 – 22GHz – £5k.
HP 3580A Analyser 5Hz – 50kHz – £1k.
HP 1980B Oscilloscope measurement system – £600.
HP 3455A Digital voltmeter – £500.
HP 3437A System voltmeter – £300.
HP 3581C Selective voltmeter – £250.
HP 5370A Universal time interval counter – £450.
HP 5335A Universal counter – 200Mc/s – £500.
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HP 3764A Digital transmission analyser – £600.
HP 3770A Amp delay distortion analyser – £400.
HP 3780A Pattern generator detector – £400.
HP 3781A Pattern generator – £400.
HP 3781B Pattern generator (bell) – £300.
HP 3782A Error detector – £400.
HP 3782B Error detector (bell) – £300.
HP 3785A Jitter generator + receiver – £750-£1k.
HP 8006A Word generator – £100-£150.
HP 8016A Word generator – £250.
HP 8170A Logic pattern generator – £500.
HP 59401A Bus system analyser – £350.
HP 59500A Multiprogrammer HP – IB – £300.
Philips PM5390 RF syn – 0.1 – 1GHz – AM + FM – £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 microwave18GHz 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.
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# RESEARCH NOTES

Jonathan Campbell

## Holographic storage gets £32 million fillip

The next generation of data storage systems are likely to be holography-based if a $32Million programme and some of the biggest names in US electronics succeed in their aims.

Holography, where data storage uses lasers to store information as "pages" of electronic patterns within the volume of special optical materials, has looked like an attractive technology for some years. A million or more data bits can be placed on each page and thousands of pages can be stored in material no larger than a small coin. In this way holographic systems offer the possibility of compact devices holding many trillions of bytes of information. Many commercial applications are envisaged, though one of the first users could be the military, looking for a system to help provide its soldiers and command centres with rapid access to the large amounts of information and visual images they expect to need to be successful in the next decade.

However, only recently have some of the essential components and technologies – such as those used in mass-market camcorders and portable computer displays – become available and affordable.

Now a joint university, industry and government consortium has begun to develop the five-year, holographic data-storage system (hdss) programme.

The aim is to develop key components and integrate them into separate write-once and rewritable systems, with a capacity of a trillion bits or more and a data-throughput rate of at least a billion bits a second.

At the same time, a second programme – photo-refractive information storage materials (Prism), with many of the same participants, will work to develop optically-sensitive materials optimised for storing holograms.

The initial goals of the hdss project are to develop several key components for the system, including a high-capacity, high-bandwidth spatial light modulator used for data input; optimised sensor arrays for data output; and a high-power red-light, semiconductor laser. The hdss researchers will also investigate optical systems architecture, such as multiplexing schemes and access modes, data encoding/decoding methods, signal processing techniques, and the requirements of target applications.

Organisations involved in the programmes include, Stanford University, Carnegie-Mellon, IBM, Rockwell and GTE and several others.

The programme's ultimate goal is to integrate all the components into separately optimised systems that will demonstrate write-once and rewritable holographic data storage.

Potential applications for holographic data storage systems include satellite communications, airborne reconnaissance, high-speed digital libraries, rugged storage for tactical vehicles, and image processing for medical, video and military purposes.

*Professor Lamertus Hesselink, Electrical Engineering, Stanford University, Stanford, California 94305-2245, USA.*

## Computing record that is rewriting atomic science

What do you suppose would be the result of setting a computer running continuously for two years to chew over a single problem, using 448 processors, each of which has about the same power as today's fastest PCs.

The answer is a glueball, and despite its rather unattractive name, scientists are delighted.

The point of the research was to calculate the properties of this elusive elementary particle already predicted by theory. In fact the properties were found to match those of a previously unidentified particle detected in several experiments carried out over the last 12 years. So two problems were solved at once – and it only took four hundred million billion arithmetic operations.

The arithmetic was carried out on GF11 – a massively parallel computer designed and built specifically for these type of calculations at the IBM Watson Research Center by Weingarten in collaboration.

The IBM result resolves a long standing puzzle in particle physics. Although glueballs are predicted to exist by quantum chromo dynamics (qcd) – the fundamental theory of nuclear interactions – none had ever before been positively identified in an experiment. It is now clear that glueballs are frequently created in particle accelerators, but have gone unrecognised, because the properties predicted for glueballs by qcd had not been found with sufficient accuracy.

The new qcd calculation provides the first accurate numerical values for the mass of the lightest glueball and for the rate at which it decays into several different combinations of more stable particles. Close agreement between these numbers and the observed properties of a particle named f(1710) make its identification as a glueball practically certain.

So the glueball is now with us for ever.

Over the years we have been used to ever-more exotic labels emanating from the minds of atomic scientists to identify their various building blocks. What a pity that this massive calculation – the largest single numerical feat in the history of computing – and the first instance of a particle's "discovery" by means of a computer, will also be responsible for marking the existence of a particle that sounds more like it has been dissolved out of an atomic scientist's ear.

# Lasers make better steel...

Successful testing of a laser-based sensor, used to track the chemical conversion of iron and carbon into steel, has demonstrated the massive potential of a commercial device, promising huge savings for steelmakers.

The sensor package, developed at Sandia National Laboratories, in Albuquerque, contains two miniature video cameras, filters for controlling both the wavelength and the intensity of the light reaching the cameras, laser diodes, coherent fibre optics for guiding images, single fibre optics for guiding laser light, and circuitry for controlling the filters. All the components are inserted into a compact package near the tip of an oxygen lance that is lowered into the furnace to control the steelmaking.

Sensors monitor temperature at the combustion zone and the bath surface where a blast of oxygen pushes aside thick, foamy slag. The lance instrument package also collects real-time information on the bath height.

By implementing fast, real-time sensors in the steelmaking process, researchers hope to reduce the time for each 'heat', cut oxygen consumption, and improve the efficiency and reliability of the process from heat to heat.

The team is testing several infrared laser methods to measure the temperature, water content, ratio of carbon dioxide to carbon monoxide – indicating how much carbon remains – and the presence of particles above the melt.

Carbon content is normally analysed later, before final metallurgy adjustments, and the final desired composition will vary for automotive sheet metal, steel plate, and other products.

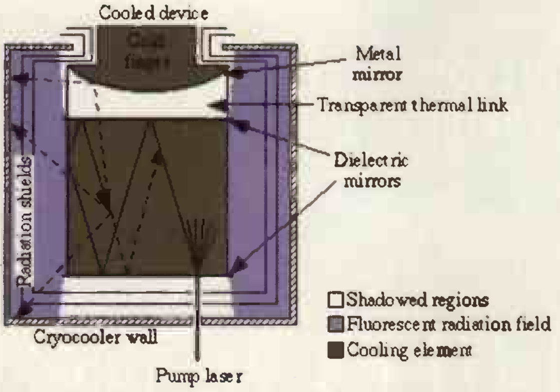
Steel mills currently measure temperatures with single-use, platinum-alloy thermocouples, racking up about $2000 per day for temperature checks.

*Newly developed high-temperature sensing equipment, based on lasers, significantly cuts the cost of steel making.*

*Don Hardesty, Sandia Combustion Research Facility, Sandia National Laboratories, MS 0167 Albuquerque, New Mexico, 87185-0167, USA.*

# ...and take on chilling role

Two per cent efficiency for a process used to cool a solid might not seem anything to get steamed up about. But when that figure is shown to be 10,000 times better than that so far achieved for the much easier problem of cooling a gas, then the extent of the breakthrough made in optical refrigeration – using lasers to cool instead of heat – becomes much clearer.




*Input laser light is shown as the solid lines and fluorescence as the dashed ones. Violet-coloured areas indicate the regions where the fluorescent radiation is most intense. White areas are in the shadow of the dielectric mirrors. These shadowed regions are not completely free of fluorescent radiation, however, due to the band pass of the dielectric mirrors. Hence, a metal mirror and baffles are required to fully eliminate fluorescent-radiation heat loading of the object to be cooled.*

For the first time, scientists are glimpsing the possibilities of constructing a solid-state optical cryo-cooler that could be used in cooling ultrafast computer circuits or for removing heat from electronics in outer space. Researchers are talking optimistically about 'optical refrigerators' being used in satellites to cool infrared cameras or in superconducting relays for cellular telephone calls, within only a few years.

The work is being carried out at the Los Alamos Laboratories, and the latest news is being seen as a major step forward toward the goal of creating a "Los Alamos solid-state optical refrigerator," or Lassor, which would cool electronic devices and scientific instruments to at least liquid nitrogen temperature, 77° above absolute zero – and eventually lower.

When light hits a solid object it usually deposits energy or heat. But under some circumstances, light can absorb energy from the microscopic thermal vibrations in the solid, so decreasing the object's temperature.

If an object excited by radiation at one frequency, can be made to emit radiation at higher frequencies, which carry more energy, the object cools.

In previous experiments, heating has always far exceeded cooling. But by using a tunable laser and modern fibre-optic materials, the researchers have managed to suppress the usual heating and make optical cooling dramatically apparent.

In effect the researchers have discovered how to use laser light to excite an object to special quantum states in which it can trap thermal vibrations but can't create them.

The experiments are actually the first demonstration of a new continuous-solid-state cooling process since the French watchmaker-turned-physicist Jean-Charles-Athanase Peltier discovered thermoelectric cooling in 1834.

In their experiments, the Los Alamos scientists shined a beam of infrared light at a 6mm-long sliver of ultra-pure glass impregnated with ions of the ytterbium.

Ytterbium ions radiate over only a single band of frequencies,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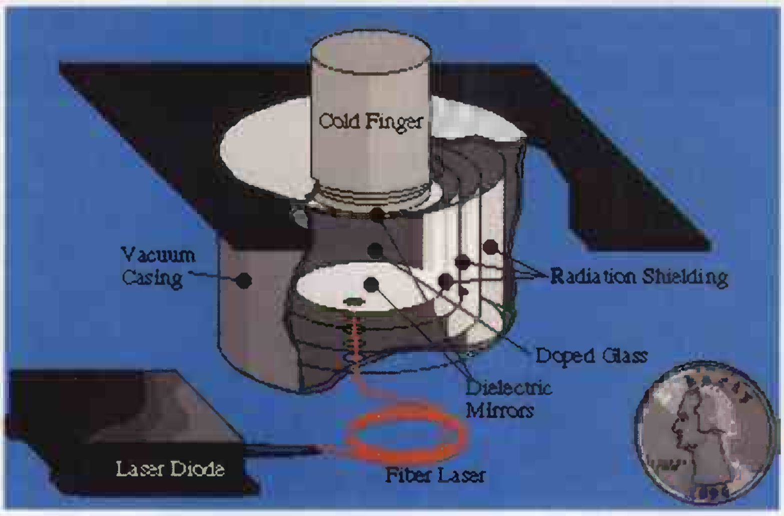



*Design for a fluorescent cryocooler that might be used in space. It is based on using an ytterbium-doped heavy-metal-fluoride glass as the fluorescent cooling element.*

frequencies that are much higher than the possible frequencies at which the glass could vibrate. So when the glass is pumped with laser light at the right frequency, it is unable produce heat.

The Los Alamos team managed to produce a cooling power that was a few percent of the absorbed laser power. That's not a lot in terms of normal cooling devices. But it is quite respectable for cooling high-tech devices to extremely low, or cryogenic, temperatures.

Light for the Lassor would be provided by efficient, compact, high-powered diode lasers in a device with no moving parts and weighing only a 1kg. The technique would be ideal for space, where the Lassors could cool a wide variety of detectors and instruments mounted on satellites.

Since they would be entirely solid-state, the devices would generate no vibrations and could survive for years in the harsh environment of space.

Ultimately, optical refrigerators may find homes in desk-top computers where they could enable superconducting circuits to operate hundreds of times faster than today's conventional electronics.

# Chaotic approach to better electronics

To impose order on chaos – just takes a little more chaos. At least that's the conclusion of three US researchers whose work is forcing scientists to take a new look at the operation and interaction of both natural and artificial non-linear systems. Ultimately it could lead to methods for improving performance of electronic systems by exploiting variations in components.

William Ditto, at the Georgia Institute of Technology and colleagues John Lindner of The College of Wooster and Yuri Braiman of Emory University used computer simulations to study a variety of coupled non-linear systems, including a series of chaotic pendula and a system with a hundred identical oscillators. The systems exhibited chaotic behaviour over both time and space (spatio-temporal chaos), and the activity of each individual element could affect the behaviour of others.

To see what would happen if they increased the disorder and variability of the chaotic systems, the researchers made each pendulum a different length, and programmed each oscillator to respond in a slightly different way.

They expected to see even more disorder and even more turbulent behaviour. But what they got was organised behaviour patterns coming out of the systems. It seems that the diversity or disorder provided a mechanism by which the systems could organise themselves.

How the process works to control chaos isn't fully understood yet, though it looks as if disorder may help move groups of chaotic elements into similar modes of behaviour. Neighbouring elements then begin to lock into the same mode, and "a local domino effect" spreads that behaviour. The result is an organised system of individual elements that repeats its behaviour in a complex but regular way.

The work looks to have direct relevance to electronics – Josephson junctions for instance. A small dc voltage across such junctions, formed by separating two superconductors by a thin insulator, causes an ac current to flow. Because the frequency of the ac current is very sensitive to any ambient magnetic field, such devices (called squids) can be used to measure extremely small magnetic fields. "Our figures of the spatio-temporal evolution of the velocities of arrays of coupled pendula can also represent the evolution of the currents across arrays of coupled Josephson junctions," Lindner told *EW+WW*.

More generally, the work has implications for any system consisting of arrays of identical or near-identical elements, such as vlsi circuits and ccd arrays.

It demonstrates that if the elements are coupled and are non-linear, than the behaviour of arrays of identical elements may be qualitatively different from the behaviour of arrays of slightly disordered elements.

"In fact, small amounts of disorder can literally chaos chaos to order," says Lindner.

The work may turn out to be related to stochastic resonance, a phenomenon in which adding noise to a system actually improves its ability to receive weak signals. Stochastic resonance is already finding applications in electronic systems, and Lindner believes engineers may one day use disorder to enhance performance of electronic systems.

"For certain non-linear systems, maybe you can not only get away with greater variability in your components, but maybe that's what you want," he explains. "A clever engineer may be able to exploit this basic phenomenon to lead to better devices. Surprising as that may sound, having a little inhomogeneity in a system may provide better performance if the elements are nonlinear." ■

*John Lindner is Associate Professor of Physics, The College of Wooster, Wooster, OH 44691, USA. jlindner@chaos.wooster.edu*

# Video COMPRESSION techniques

**John Matthews outlines the techniques being used to cram more image information into less space and bandwidth.**


**Video technology consultancy**
John Matthews is a research and development engineer at Teltec Ireland and is based at Dublin City University. Teltec Ireland is the National Programme for Telecommunications in Ireland and represents a partnership between leading third-level institutions and industry. The Teltec Video Coding Group at DCU has conducted extensive research in the area of low bit-rate video and image coding for telecommunications and multimedia applications. For further information, phone 353-1-7045759, fax 353-1-7045508 or e-mail john.matthews@teltec.dcu.ie.

In recent years, advances in video compression techniques have been at the forefront of the multimedia revolution. The variety of video applications which are becoming available was clearly evident at this year's Telecom 95 show in Geneva, ranging from pc based conferencing systems to high-end video on demand products.

Multimedia is a general term covering the exchange of video, audio and data between people. New applications and products appear almost daily, and there are standards in the multimedia world to ensure that equipment and services from one manufacturer will operate successfully with similar equipment from other companies.

A number of standards are now firmly established, including JPEG, MPEG1, MPEG2 and H.320. Most video-conferencing systems over ISDN now conform to the ITU-T H.320 standard, although there are proprietary systems.

MPEG1 is used for audio/visual storage on cd. JPEG is used for still image transmission, while MPEG-2 is aimed at higher end broadcast systems. Apart from JPEG, these standards define the rules governing the compression of video *and* audio for storage or transmission.

This article looks at the video aspects of these standards. These new standards and compression algorithms have been developed to reduce the bandwidth required for video transmission. In addition, emphasis has been placed on defining methods which can be practically implemented This development has been coupled with progress in the hardware world, where highly integrated single-chip solutions for the algorithms have been brought to the market.

While the video standards have some common features – in particular the fact that they all use discrete cosine transform – they are distinctly different and are geared towards different applications. For example, JPEG could be used for motion video applications, but it would never achieve the same real-time performance as H.261. This is because it does not have inter-frame and motion compensation capability

Also, new standards will emerge, encompassing even more applications. For example, H.263 is a video standard with many similarities to H.261 (the video standard for H.320) which will enable good quality video conferencing over the ordinary telephone line. With MPEG-4, there will be new functions, coupled with improved compression ratios and picture quality.

In this article I outline the video standards already developed and summarise their differences and unique positions. In addition, I also review the up and coming standards.

## Video compression fundamentals

Current standards – JPEG, MPEG and H.261 – have a number of things in common. The most important of these is that they all use the discrete cosine transform, dct, in their algorithms. They each perform quantisation on the resulting dct coefficients, and implement run-length coding on those quantised coefficients, **Fig. 1**.

So what is dct? It is a mathematical transform which translates digital video data from the spatial domain into the frequency domain. Typically, a coding algorithm will divide picture data into blocks of 8-by-8 pixels, where a pixel is a picture element with a value between 0 and 256. It then performs the dct on each block. For each block of 8 by 8 pixels, the dct gives an 8-by-8 block of frequency components.

The rationale behind the dct is as follows: In

## Second generation video coding techniques

Block-based coding methods have disadvantages – in particular blockiness in pictures and the inability to derive information on specific objects in the picture. But is there an alternative?

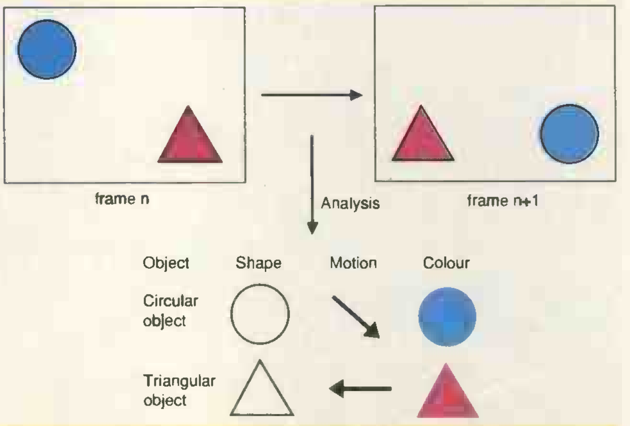
Second generation video coding techniques have been the subject of much research in the nineties, and new techniques such as segmentation, model based coding and object based coding are under intense investigation.

Segmentation is a technique in which the image is described in terms of contours and texture. Contours are abrupt changes in the gray levels of the image, and texture can be thought of as the roughness of the image – or the shade transition across the image. So a segmentation based scheme will attempt to describe an image as textured regions surrounded by contours.

Model based coding techniques have been greatly influenced by progress in the fields of computer vision and computer graphics. The underlying motivation in model based coding is that an image is the projection of an illuminated 3-d object onto a 2-d plane. In object based coding an image is subdivided into objects and each object is described in terms of its shape, motion and colour.

In an object-based coding scheme, a source model provides an abstract means of describing the type of object in the picture. This source model defines the parameter which will be used to identify objects during image analysis. For example, an object might be rigid or flexible.The source model will also describe motion of objects – i.e. fixed direction or arbitrary direction. So a picture is analysed and objects are identified. Each object is then described in terms of shape motion and colour, see illustration.

*Object-based image analysis. New techniques for coding of video signals are being investigated as an alternative to DCT block based techniques. One such scheme is object based coding, where objects are identified in a picture and are coded using their shape motion and colour parameters.*




Of course, certain parts of the picture will not be classified as objects as they will not conform to the source model. These areas are classified as model failure regions, and might be encoded using a more traditional coding scheme. Among the objectives in an object based codec is to have a higher picture quality at comparable or lower bit rates than block based codecs. These objectives are helped by a number of factors.

The process of analysing an image to identify objects is an exhaustive one, but once the objects have been identified, it does not take a lot of information to describe them. Furthermore the only motion vectors which are transmitted are those relating to objects.

Schemes are usually designed so that objects are bigger than the blocks which would exist in corresponding block based coding methods, so there is potential for saving on motion vectors. Thirdly, when an object has been defined and is moving from frame to frame, very little update information may be required for that frame. So even though three parameters are necessary in object based schemes, as against two – motion and colour – for block based schemes, it is still feasible to operate such a scheme on a similar bandwidth.

A further feature of an object based scheme is that it should be able to classify the the objects, so optionally only the most important objects need be transmitted in a low bit rate channel. Editing would also be feasible in such a scheme.

At present, most hardware research in object based coding techniques is carried out using digital signal processors or similar devices. There is still a lot of work to be done both in the areas of algorithm definition and in the implementation of those algorithms in hardware. However, it is an area which is attracting much research interest, and this level of interest has been considerably enhanced due to the interactive features required in MPEG-4.

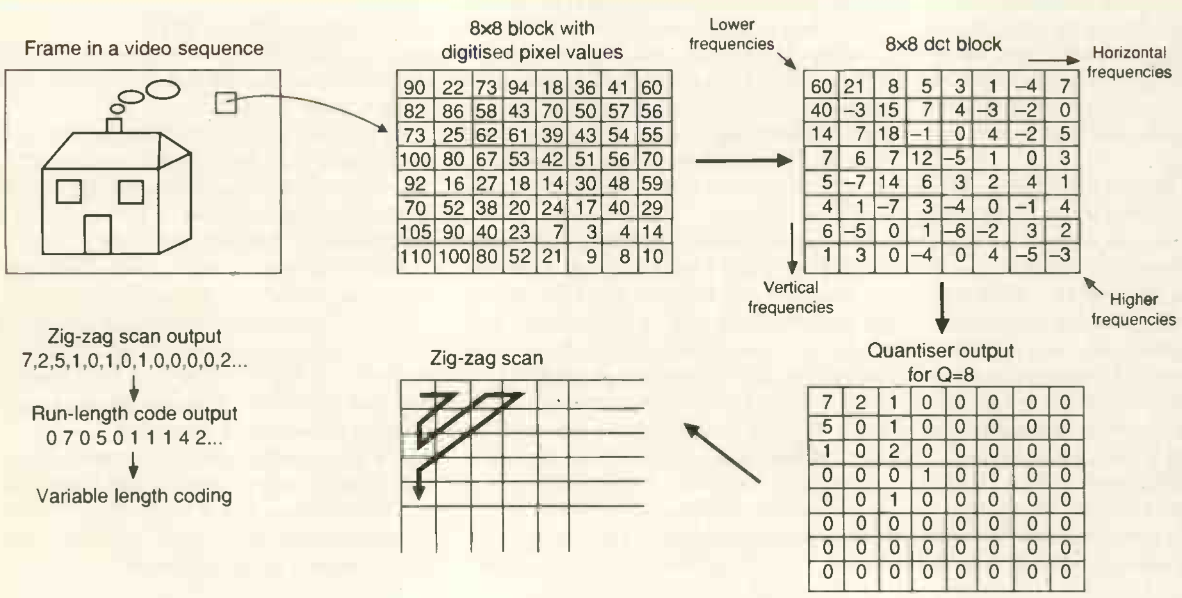



*Fig. 1. DCT-based video coding scheme. The discrete cosine transform (DCT) is at the heart of the coding schemes used in MPEG, JPEG and H.261. This figure shows how an 8 by 8 block of data is extracted from a field of video data. Note the presence of significant low frequencies in the 8-by-8 block of transform coefficients, with most of the higher frequencies going to zero after quantisation.*

*File size of the JPEG original on the left is 1215Kbyte while that of the version on the right, compressed with loss, is only 61Kbyte.*

the spatial domain, picture information is spread thinly over a large number of pixels. In the frequency domain, however, much of the picture information will be contained in the lower frequency components. As a result, it may be possible to discard some of the higher frequency components, without sacrificing too much picture quality.

Performing the dct on the video data typically concentrates much of the picture information into the lower frequency components. The dct itself does not compress data. An 8-by-8 block of pixel data will provide an 8-by-8 block of frequency components. The compression process starts with quantisation and run length encoding.

Quantisation is a process where each of the dct coefficients is divided by an integer and rounded towards zero. In a typical picture, many of the higher frequency components will have low values, so their output after quantisation will be zero. The quantisation integer is user definable, or is adaptable, its value being determined by a control loop. In most cases, users can control how much of the high frequency components they wish to neglect. This is the start of the compression process.

The quantised coefficients are then zig-zag scanned and run length coding, or rlc, is performed. Output of the rlc process will be the values of each non-zero components, preceded by the number of zero valued coefficients before that component.

In most algorithms, further compression is achieved by the use of variable length coding, or vlc. Here commonly occurring strings from the run-length coding process are assigned short code words, while less common strings are assigned longer code words At this point, framing of the data can occur, so compressed data is now ready for transmission or storage.

In motion video, further techniques are used to remove redundancy from the data. The first of these techniques is prediction. Instead of coding and transmitting data for full frames, frame differencing is used. The encoder will code the difference between the current frame and a prediction of what that current frame should be.

The easiest method would be to simply subtract the last frame from the current frame and use the difference. However this method does not work well, because it doesn't take into account the error build up in the transmission channel. So each encoder has an inbuilt model of the decoder in a feedback path. It is therefore a decoded version of the last encoded frame which is used for prediction purposes.

To further eliminate redundancy, motion estimation and compensation is used. The predicted frame is refined to take into account the motion which is estimated to have occurred between it and the current frame. This has the effect of making the predicted frame as similar as possible to the new incoming frame, so the frame difference will be minimised even further.

Most standards do not specify how to do motion estimation and compensation, so a trade off can be made between complexity and performance.

## Still image coding – JPEG

Standard ISO10918, more commonly referred to as JPEG, defines the techniques to be used in the coding of still pictures.

JPEG is the most simple of all the standards under discussion. Because it caters for stills, there is no requirement for frame prediction or motion compensation, so it does not need a feedback loop. The forward process consists of dct, quantisation, zig-zag scan, and run-length and variable-length coding.

However, a number of different options in the JPEG standard, allow users to tailor their systems for different levels of compression and picture quality. Both lossless and lossy coding techniques are referred to.

In lossless coding, the picture can be rebuilt exactly as it was prior to coding. Lossless coding techniques in JPEG are not based on the dct, but are 'prediction' based. Lossless coding is very limited in the amount of compression which it can achieve however. Imaging for medical purposes is an example of a situation where lossless coding is required.

Lossy coding techniques are based on the dct. Among the options available are sequential coding, progressive coding and hierarchical coding.

Sequential coding is the simple dct process already discussed. Progressive coding, as the name implies, allows a progressive build up of picture quality. In this mode, the quantised coefficients are sent in stages. Initially the lower frequency components are stored or transmitted, followed by the higher frequency components.

Another method is to selectively increase the resolution of the coefficients, so that the most-significant bits are first processed, followed by the least-significant bits.

In hierarchical coding, resolution of the picture gradually builds up. The advantage of progressive and hierarchical schemes for JPEG is that they allow the user to select a variable quality level for a still picture. This is important, for example, if a JPEG picture is being transmitted from one location to another.

If a slow transmission channel was used, then sending a very high resolution picture could take a long time. As a result, it might be desirable to send lower quality video. In a 'browse' type application, the receiver could then select specific stills and request those in greater detail.

## Video conferencing and H.261

The most relevant standard today for the video compression part of a conferencing system is the ITU-T standard H.261. Most commercially available systems available today use techniques outlined in H.261. The standard details the syntax for the coded bit stream and specifies how the decoder works.

Implicit in specifying the syntax and decoder will be certain features of the encoder.

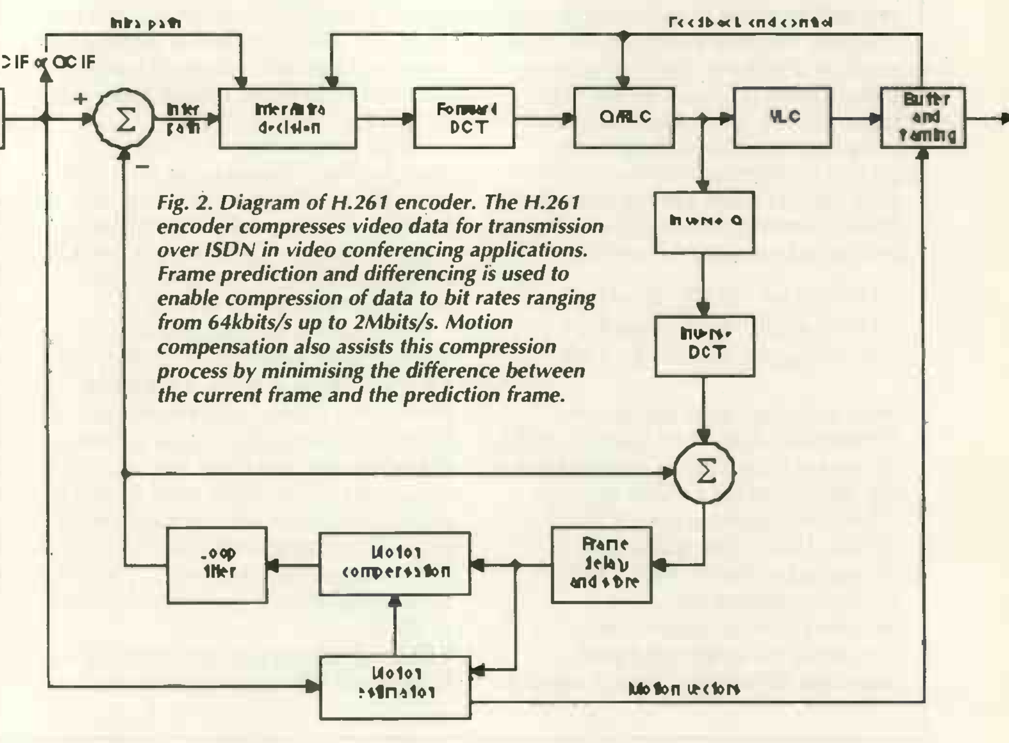


*Fig. 2. Diagram of H.261 encoder. The H.261 encoder compresses video data for transmission over ISDN in video conferencing applications. Frame prediction and differencing is used to enable compression of data to bit rates ranging from 64kbits/s up to 2Mbits/s. Motion compensation also assists this compression process by minimising the difference between the current frame and the prediction frame.*

However many encoder options are left to the user to implement. Examples of such options are the ways in which motion estimation and compensation are performed. **Figure 2** shows a typical H.261 encoder.

The compression algorithm in the H.261 standard operates on YCrCb digital data. There are two resolutions specified for the source data, namely CIF and QCIF, as detailed in the Table. An H.261 compatible decoder must be able to decode a bit stream which had QCIF as its origin, but CIF is optional. Most systems accommodate both resolutions. A system which encodes CIF data to a single ISDN B channel will compress video by approximately 60:1.

H.320 and H.261 are geared towards ISDN, so transmission of coded audio/visual data is normally at a multiple of 64kbit/s – the bandwidth of one ISDN channel.

Typically, since the video and audio must share the channel, there will be some trade-off between audio and video quality – especially at low bit rates. Normally in a single ISDN channel, 16kbit/s would be required for audio, so 48kbit/s would be left for video.

Video-conferencing systems based on pcs typically operate at lower bit rates, usually 64kbit/s. Bigger stand-alone systems usually operate at between 128 and 384kbits/s, and they benefit from the higher resolution of CIF data.

As with JPEG and MPEG, H.261 is a dct based standard. Because it caters for motion video, frame prediction is used with motion compensation in the encoding process.

A reasonably simple motion estimation and compensation scheme is used. The current frame data is divided into 16-by-16 pixel 'macro-blocks', each of which is compared to other 16-by-16 blocks in the last decoded frame, to a displacement of plus or minus 15 pixels in each dimension.

A calculation, such as the sum of the absolute differences between each corresponding pixel in the 16-by-16 macroblocks is made. The nearby 16-by-16 block which gives the minimum overall difference is used to determine the best match. Motion vectors are then calculated based on this match, and these must be sent to the decoder.

The predicted frame is adjusted using the motion vectors, and frame differencing then takes place. In practice the predicted frame will be stored in memory, so the compensation can occur by modifying the memory addresses. Finding the motion vectors, however is a lengthy and computational process.

The variable length codes, produced by the encoder, are fed into the output buffer of the system. This buffer has a variable input rate, but also has a fixed output rate determined by the bandwidth of the transmission channel. The amount of information entering the buffer must be controlled so that the buffer does not overflow. If the buffer overflows picture information is lost, and corruption occurs.

Buffer control schemes can range from being very simple to extremely complex. In simple terms, control is achieved by monitoring the buffer fullness, and adjusting parameters such as the quantisation values to maintain a steady input rate into the buffer. The system also decides whether or not to use frame prediction and motion compensation, and could also decide to drop the frame rate by skipping some frames.

Problems generally occur when there is sudden movement in the picture being encoded. Sudden or rapid movement is almost always a problem for a number of reasons. Firstly, the buffer level increases and can overflow. This happens because there is a larger frame difference due to the high degree of motion between scenes, leading to more quantised coefficients.

More generally the motion compensation can only deal with a limited amount of motion. The effect the viewer sees is a smearing on the screen where the movement is occurring, and jerkiness in the picture is often noticed – especially when frames have been dropped. These are probably two of the most annoying features in a system.

Another annoying feature is poor lip to speech sync, where lip movement and audio output do not exactly coincide. This is caused by different delays for the audio and video data through the system.

## MPEG-1 and video CD

MPEG-1 is an internationally accepted standard for the compression of digital audio and video. The actual standard is an ISO-IEC standard and the term MPEG (Moving Picture Experts Group) comes from the group who started work on the standard. There are three parts to the standard: audio, video and system.

MPEG-1 was originally developed to provide a standard for the storage of audio and video on digital storage media. The standard is optimised for operation at about 1.5Mbits/s. This is significant because it is the data rate for an uncompressed cd and it is also suitable for digital audio tape.

Typically the audio takes about 192kbits/s of this bandwidth, and some bandwidth is also needed for the system data stream. As a result, there will be approximately 1.15Mbit/s available for video.

Certain requirements in accessing stored video and audio have played a large part in the development of this standard. Access to the stored material is important and facilities have been inbuilt for random access, fast forward and reverse, and reverse playback. The synchronisation of audio and video is also very important and error robustness is also inbuilt into the system.

Unlike video conferencing standards, MPEG tends to be quality controlled rather than bit rate controlled. Certain parameters are specified which give a certain quality level, rather than, say, setting the system to operate at a particular bit-rate such as the bandwidth of an ISDN channel.

There are significant differences between the encoding process used for MPEG1 and those used for H.261. In H.261 for example, there are two types of frames, namely intra and inter but in MPEG-1, there are three types.

The intra frame in MPEG, called the I frame, is similar to the intra frame in other standards, and it is encoded without reference to other frames. One difference exists however: in an H.261 system it is desirable to avoid any intra coding. In MPEG however, intra coded frames are needed because they provide the points at which random access can take place in the decoded video.

The inter frame in MPEG is like the one in H.261. Basically a prediction is made of the

## Video colour spaces and resolutions

When people think of colour, the usually think of red, green and blue. These are the primary colours and any other desired colour can be obtained by adding the required amounts of these primary colours together.

Since the advent of television however we have made use of the fact that the eye is much less sensitive to changes in colour than it is to changes in brightness. So television signals are transmitted as a black and white signal plus two colour difference signals providing information on how much the colour varies from white. Because the eye is less sensitive to colour, the colour difference signals are transmitted using a lower bandwidth than the monochrome signal.

The black and white signal is also known as the luminance signal. It is usually referred to as the Y component. The colour difference signals are referred to as the U and V components and they tell how much red and blue exists in a picture. The combination of these signals is known as the YUV colour space. Since the total amount of red green and blue add up to the picture brightness, then the translation between YUV and gamma corrected RGB (referred to as R'G'B') can be calculated using the following matrix.

$$\begin{bmatrix} Y \\ U \\ V \end{bmatrix} = \begin{bmatrix} 0.299 & 0.587 & 0.114 \\ -0.147 & 0.289 & 0.436 \\ 0.615 & -0.515 & -0.1 \end{bmatrix} \begin{bmatrix} R' \\ G' \\ B' \end{bmatrix}$$

Note that even if the signals are transmitted in the YUV space, it will be necessary to be able to convert back to rgb for display on a video monitor.

Another important colour space is YCrCb. This colour space was developed as part of the CCIR Recommendation 601. This recommendation defines the "encoding parameter of digital television for studios", and is a world wide standard for digital component video. Basically, YCrCb represent the same colour space as YUV, but the individual components in YCrCb are scaled and offset versions of the components in the YUV space.

*Table 1. Video resolutions for PAL originated signals*

| Format | Y horiz. | Y vert. | Cr horiz. | Cr vert. | Cb horiz. | Cb vert. |
|---|---|---|---|---|---|---|
| CCIR-601 | 720 | 576 | 360 | 576 | 360 | 576 |
| CIF | 352 | 288 | 176 | 144 | 176 | 144 |
| QCIF | 176 | 144 | 88 | 72 | 88 | 72 |
| SUB QCIF | 128 | 96 | 64 | 48 | 64 | 48 |

CCIR-601 video specifies a 4:2:2 sampling format, and this means that there are only half the number of samples for each of Cr and Cb per line of video as there are Y samples. So it is taking advantage of the fact that the eye is less sensitive to colour. In a line of digital video sampled at 27MHz there will be 720 luminance samples and 360 each of the colour difference samples.

When generating of CIF and QCIF, even further redundancy is built into the colour difference sampling. The format is no longer 4:2:2, but is now 4:1:1. This means that there is one Cr sample and one Cb sample for every four Y samples in the picture.

**Table 1** gives the horizontal and vertical resolutions for CCIR-601, CIF, QCIF and sub-QCIF for. Given that most video coding systems will use CCIR-601 compatible video as their digital source, it is clear that generating CIF or QCIF from CCIR-601 represents significant compression by itself. Converting from CCIR-601 to CIF represents compression of 5:1, and converting to QCIF gives compression of 20:1.

The subsampling of the video signal to generate CIF or QCIF must be implemented quite carefully. It is not sufficient to simply drop the pixels or samples which are not required, and if this is done artifacts can result in the picture due to overlap in the frequency spectrum of the signal. As a result, a process known as decimation is carried out on the video data. A similar process, interpolation, is usually carried out when upsampling CIF or QCIF data up to CCIR-601.

Decimation is a digital filtering process, and in such a process, the data is usually filtered first, and then resampled at the desired rate. In the case of interpolation, where the signal is being upsampled, the data is usually padded with zeros to bring it up to the desired sample rate and is then filtered. This happens in both the horizontal and vertical dimensions.

There is a number of choices for the hardware designer who is required to produce CIF or QCIF. The use of a digital signal processor is an option, but there are chip sets available to resample video. The GEC Plessey VP520 is a dedicated CCIR-601 to CIF/QCIF convertor and it can also convert back from QCIF/CIF to CCIR-601. Other devices for decimating video are manufactured by Philips, Chips & Technologies, Brooktree, Raytheon and Harris – and others.

current frame, and this prediction is motion compensated. The frame difference between this prediction and the actual current frame is taken, resulting in the predicted frame, P.

A third type of frame is used in MPEG-1, known as the bidirectional B frame. B frames are interpolated from earlier and later I and P frames. The way these bidirectional frames are created affects both the ability to random access the video and the quality of that video.

The distance between I frames in a video stream affects the random access capability, while the distance between P frames affects the complexity of the system and the quality of the video. Another interesting thing about MPEG-1 is that frames are not always sent in the order that they are created, so the decoder will need memory to enable the decoder to construct bidirectional frames. See **Fig. 3** for an idea of how this happens.

## Coding for broadcast systems

MPEG-1 was optimised for CD-ROM at bit rates of about 1.5Mbit/s. The international co-operation to develop MPEG worked well, so follow-on work involved addressing broadcast television sample rates using the CCIR 601 recommendation. MPEG-2 was the result.

MPEG-2 is geared towards broadcast technologies, and examples of where it will find applications are catv, digital television, video on demand and DBS. For video-on-demand, trials have been carried out worldwide and ADSL networks were used in these trials.

MPEG-2 has different 'profiles' and 'levels' so its use can be tailored to a particular application. A particular profile will place limitations on the syntax of the encoded video, while a particular level will place limitations on parameters such as frame dimensions or sample rates.

Low level MPEG-2 involves the use of CIF

## Hardware options for image compression

It could be said that design engineers are now spoilt for choice when it comes to the design of video circuitry. New chip sets are constantly entering the marketplace, catering for all of the established standards.

These chip sets are also becoming increasingly more highly integrated. So let us say you have just started working on video, and you wish to design a JPEG, MPEG or H.261 circuit. Whose chips will you look at?

The diagram below shows some of the options your proposed system might comprise. Your first problem is digitising the video. It may need to be preprocessed or scaled. It is then coded and stored or transmitted. After the video has been decoded you may have to do some post processing, ie convert from CIF to CCIR and reconvert the signal to analog PAL for display on a monitor.

Alternatively you may wish to display the video on the pc, so the video signal must be merged with the rest of the graphics going to the screen. I will look briefly at some of the options available for these tasks since an exhaustive survey is beyond the scope of the article.

For digitising PAL signals and converting to CCIR-601, Philips, Brooktree or Raytheon all produce appropriate devices. These companies now offer highly integrated front end solutions, such as the Philips *SAA7110* and the Brooktree *Bt819*. The *SAA7110* needs almost no external components, producing square pixel CCIR compatible data, while the *Bt819* allows scaling of video to one-thirteenth of its original size. Auravision also have a number of interesting devices, such as the *VxP501*, which allows video capture with scaling and also support MPEG/JPEG playback.

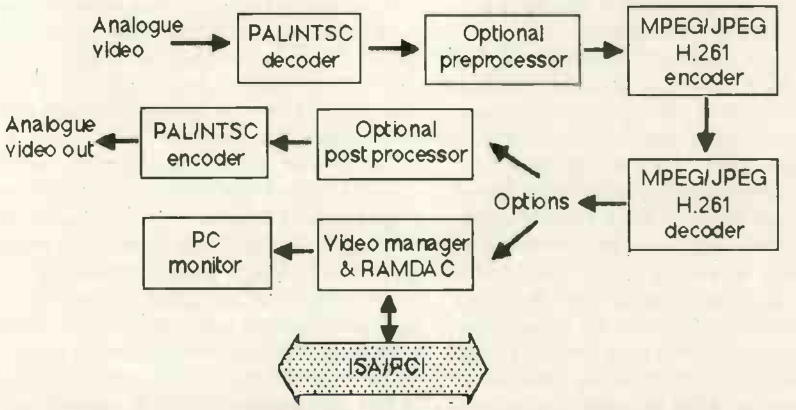
For preprocessing, the GEC-Plessey *VP520* is a dedicated bidirectional CCIR-CIF/QCIF convertor. Other devices are available which are geared to pc video, such as the Philips *SAA7195* video memory controller or the Chips & Technologies VideoPro chip set. Both devices are essentially video memory controllers which allow scaling of video for windowing.

For JPEG coding, options include devices from Zoran, LSI Logic, C-Cube and Winbond. Silicon to support MPEG1 is available from C-Cube, IC Works, SGS-Thomson, Hitachi, IIT and Winbond amongst others. For H.261, packages are available from C-Cube, IIT, GEC Plessey and Array Microsystems. The GEC Plessey chip set is a dedicated H.261 coding/decoding chip set. Array Microsystems produce the Videoflow chip set, based on risc architectures. It is one of a number of chip sets which will support several standards, in this case JPEG, H.261 and MPEG-1, and uses a highly parallel data flow architecture to allow multiple decoding when it accepts several bit streams.

The IIT *VCP* can be used in both H.261 and MPEG-1 codec applications, and will support MPEG-2 decoding. C-Cube has a range of devices which cover all the established standards including the *CLM4200* H.261 codec and the *CL480VCD* MPEG1 system decoder. For MPEG-2 encoding both IBM and C-Cube announced devices this year.

Another option for codec development is that of digital-signal processing Texas Instruments released their *C80* this year. The *C80* contains four parallel processing signal processors with a risc master processor, an integrated video controller and claims 2 BOPs performance.

If a composite PAL signal is required at the output of the system, again Philips, Brooktree and Raytheon should be considered, with each company having a range of encoders suited to different applications. Devices which are becoming more common in the market now are those which will accept video from an MPEG/JPEG decoder and will merge that video with the normal pc graphics for display on the monitor. The Auravision *VxP501* was mentioned, It was designed with interfaces for Zoran and C-Cube JPEG and MPEG devices. Other companies which are strong in this area are Brooktree, Pixel Semiconductors, MCT, Trident and Philips.




***Possible hardware requirements. A typical hardware video coding system will need to digitise analog video, and process that digital video using a JPEG, MPEG or H.261 encoder. The compressed video is then stored or transmitted. In PC based systems the decoded video will be merged with the other PC graphics for display on the monitor. In a stand-alone system, however, the output may be a PAL signal for display on a composite monitor.***

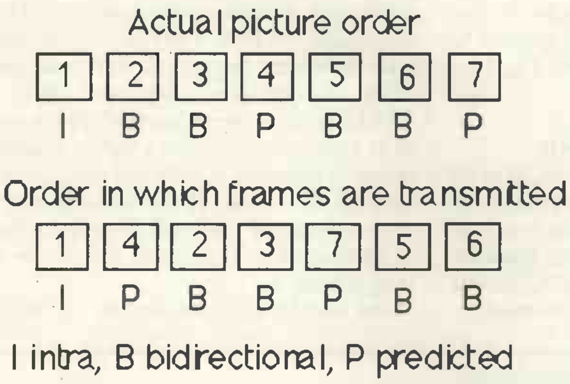



*Fig. 3. Transmission of MPEG 1 frames. MPEG-1 is particularly suitable for video CD, and is optimised for operation at around 1.5Mbits/s. It uses three type of frame, Intra, Prediction and Bidirectional. In order to facilitate the MPEG-1 decoder, these are not transmitted in the same order as they are encoded. Frames 1 and 4 are needed to decode 2 and 3, and frames 4 and 7 are needed to decode 5 and 6.*

resolution with a bit rate for the encoded data up to 4Mbit/s, while main level studio tv applications will require CCIR 601 resolution. Most decoder chips will be capable of operating at main profile, but higher profiles could include features such as spatial scalability.

The syntax in MPEG-2 can be divided into two main categories. There is a scalable syntax and a non-scalable syntax.

The non-scalable syntax is structured as a super-set of MPEG-1, but there are extra tools for handling interlaced video. This is required for studio applications, because PAL and NTSC video sources are interlaced, i.e. two fields are interleaved to form a frame.

Typically MPEG-1 and H.261 use CIF resolution. This is obtained by dropping one field per frame and and further dividing the remaining field. This lowering of video resolution causes a subsequent loss in video quality, so while its acceptable for pc conferencing systems, it is not suitable for studio broadcast.

When handling interlaced video, MPEG-2 allows a frame to be selectively treated as a single picture. Alternatively it allows the two fields to be coded individually. It there is a lot of motion in a picture, it is generally better to code the fields individually.

MPEG-2 incorporates all the functions available in MPEG-1, such as random access, fast forward and reverse and reverse playback. All MPEG-2 decoders will in fact be able to decode an MPEG-1 bit stream.

The scalable syntax in MPEG-2 will find applications in transmission media which do not have constant bit rate. Asynchronous transfer mode, or ATM, is one such application.

There are three types of scalability in MPEG-2, namely signal-to-noise ratio, spatial and temporal. Each of these scalable extensions uses the concept of a base layer and an enhancement layer. The lower base layer is used for a basic video quality, and the higher enhancement layer is used to improve the quality already available from the lower layer. This is useful in applications such as transfer of video over an ATM network, or in general over a channel with a variable bit rate.

An error-robust channel can be used to transmit the base layer video. However, the enhancement layer could be transmitted over a channel which was not so error robust, or was likely to be congested. Since it is for enhancement only, it does not affect the basic video quality if corrupted, or if it fails to reach the decoder.

In temporal scalability, the base layer provides a basic temporal resolution (frames per second), and the higher layer is coded with temporal predictions for further enhancements. The enhancement layer for spatial scalability provides a coded difference signal based on an interpolated prediction of the lower layer. In the case of signal-to-noise ratio scalability error information produced in the encoding process might be used.

## Video over the phone

In March 1995, the ITU-T accepted a new standard known as H.263. The title of this standard is 'Video coding for narrow telecommunications channels at <64kbit/s'. The H.263 standard is geared toward use over the normal telephone line. It will be possible to implement it with a range of options up to and including use of a V.34 modem at 28.8 kbit/s. The umbrella standard for full audio-visual transmission in this way will be H.234

It is expected that chips and products for H.263 will be available sometime next year. These products will probably be for the videophone market, or for pc users who need real time video communications but do not have access to ISDN.

H.263 is similar to, but more complex than, H.261. Much of the work which was done to develop the standard had its origins in the H.261 development. As with H.261, H.263 uses block based methods for compressing and coding video signals.

The discrete cosine transform is used to derive frequency content information from the original spatial information. Quantisation, differential coding between frames, run length and variable length coding techniques then allow compression and coding. As with H.261, motion estimation and compensation can be used in the differential (Inter) coding process.

There are differences between H.261 and H.263 which allow H.263 to operate more effectively at low bit rates. Firstly the picture format for H.263 is QCIF at 176 by 144 elements or sub-QCIF at 128 by 96. Secondly the syntax is different, and improved variable length codes are used. The motion compensation is also different, with H.263 allowing half-pixel accuracy.

A significant difference between the two standards is the use of 'PB' frames in H.263. The idea comes from the use of P and B frames in MPEG. Basically, a PB frame consists of two pictures coded as one.

The P frame is similar to the normal inter coded frame in H.261. The B frame is however derived using bidirectional prediction from the two adjacent P frames, fitting in between these two frames. Experimental work showed that without significant gain in bit rate, a much better picture quality was obtained using these PB frames.

When H.263 products actually arrive to the marketplace, it is possible that there will be a requirement for at least the decoder to be backward compatible with H.261. This may not present too much of a problem however, as H.261 can be seen to be a sub-set of this new standard.

## The future – MPEG-4

The standards already described all use block based coding. In such coding, the image is decomposed into blocks which are encoded independently of each other. There are several problems and limitations attached to this approach. One is 'blockiness' in the picture. Another is the fact that objects in the picture are not defined, so a limited number of functions can be added to the system.

Work has already begun on a new standard, MPEG4, which will support new ways for communicating, accessing and manipulating audio-visual data. With an increasing trend towards wireless communications, and a demand for more interactive communication, many new requirements are arising in audio visual communications which are not catered for by existing standards.

One desirable feature in the new standard will be that of content based accessibility and manipulation. A user should be able to access and manipulate video based on its content. This may only be possible if new schemes are developed for the coding of that video.

Over the past few years, much research has been carried out using advanced coding techniques, such as object based coding, and it is possible that some of these techniques may be used in MPEG-4.

Because the development of this new standard is at such an early stage, no specific method for doing this has been decided. However a desirable coding scheme will be one which can identify objects in a picture and can track the movement of those objects.

It is hoped that MPEG-4 will enable many new uses for multimedia, such as the retrieval of information from on-line libraries. One can imagine being able to browse through a moving picture, clicking on an object in the picture – perhaps a book in a library or an item on a shelf – and being able to download more information on that item.

We are however still a long way away from this scenario – the MPEG-4 standard is not expected to be ratified until 1998. ■

# High power valve audio

**In this extract from his book *Valve Amplifiers*, Morgan Jones discusses ways of designing valve audio amplifiers capable of delivering more than the usual 10W.**

While the amplifier discussed last month was a good design, it has to be admitted that 10W is not a great deal of power, and more may be required. The traditional way of doing this was to use a more powerful valve such as the *EL34*, or even the *KT88*. Another method is to use Class AB; using these techniques, we can obtain 50W from a pair of *EL34* or *KT66*s, or 100W from a pair of *KT88*s. After this, we resort to transmitter valves at enormous cost.

The traditional method has disadvantages:

- Higher powered valves are invariably disproportionately more expensive.
- They use high supply voltages, which makes the smoothing capacitors expensive, and the HT supply a major safety hazard.
- Higher powered valves tend to need higher impedance anode loads, which makes the design of a good transformer difficult.
- Transmitter valves have savage drive requirements, and often need a power valve as a driver.

Fortunately, there are several ways out of this dilemma.

- Lie about output power. In the late 1960s and early 1970s, some quite unpleasant audio amplifiers were made using transistors. Compared to the valve behemoths, these transistor amplifiers were very small and light, but they didn't actually sound any better. In fact, most sounded a lot worse so something was needed to make them sell. The one thing that early transistor amplifiers could do was to provide plenty of power, and as a result the power rating war started.

To make a truly powerful amplifier, a large power supply is needed, but this is expensive. Classical music generally has peaks of only a short duration, and few hi-fi enthusiasts listened to anything else. Amplifiers were designed that could manage higher output powers, but only for a very short time. This allowed power ratings to be increased further, and the 'music power' rating was born.

Music power works like this. Measure the maximum output power at 10% distortion or the onset of clipping, with bursts of 1kHz driving one channel only into a resistive load. By this means, it is perfectly possible to convert a 20W amplifier with a poor power supply into a 50W model. If output is doubled to account for two channels, you can achieve a 100W amplifier. At least four fallacies were used in the previous argument.

- Build more efficient loudspeakers. This is an excellent solution, since inefficient loudspeakers frequently suffer from power compression. This is an effect whereby resistance of the voice coil rises due to temperature. As a result sensitivity is reduced until the coil cools down.

- Drive the loudspeakers more effectively. If drive units are driven by dedicated amplifiers preceded by an active crossover, many benefits result (Colloms, 1985). For the purposes of this discussion, it is sufficient to say that a two way loudspeaker system, driven actively by 10W amplifiers, will go surprisingly loud.

- Parallel output valves. This solution provides many advantages. If multiple pairs of paralleled output valves are used, HT voltage can be kept within safe bounds. This is the case even at 320V if many pairs of *EL84* are used.

With each additional pair of valves, the transformer primary impedance falls, as does the turns ratio. This makes it easier to design a good quality component. Statistically, total anode current per side is better balanced as the number of valves is increased. Deliberate selection will improve this still further.

## Driving higher power output stages

Whether they are composed of paralleled devices or not, higher powered output stages always require more driver circuitry. When the Williamson was investigated – *EW+WW* December 1995 – it was found that it had a dedicated driver stage. However, the large total number of stages made stability a problem. Clearly, a better approach is needed. As before, listing the requirements will help solve the problem.

- A low output impedance to drive the increased input capacitance of the output valves – a cathode follower may be needed.

- Capability for providing a large output voltage with low distortion. This invariably demands some form of a differential pair.

- Wide bandwidth and high gain are also desirable. This is because it would be preferable to use just one set of coupling capacitors, ensuring lf stability. A cascode would be ideal, although a carefully designed pair of dc coupled differential pairs could be even better.

Putting these requirements together results in a cascode differential pair with direct coupled cathode followers. This design is sometimes known as the Hedge circuit after its original inventor, although the original Hedge circuit did not include cathode followers (Hedge, 1956), **Fig. 1**.

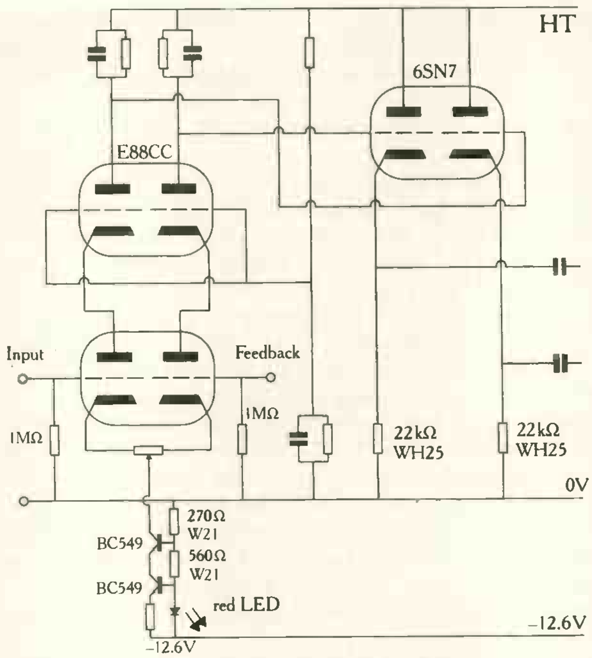



***Fig. 1. Cascade differential pair with direct coupled cathode followers. Sometimes referred to as the Hedge circuit, this design differs because of the inclusion of cathode followers.***

The differential pair is not the ideal phase splitter, so extra care will be taken over this in order to obtain a good result. Anode load resistors should be matched, and generously rated to avoid drift. The constant-current sink should be made to have as high an output resistance as possible. Also stray capacitance to ground from the cathode should be minimised to maintain a high impedance at high frequencies. Matching valves would be useful if possible.

Each pair of valves requires a separate heater supply. Sad, but true. Cathode followers need around 200V superimposed on their heaters. The upper pair of the cascode needs around 100V, and the lower pair 0V. Flirting with this rule will generate problems related to heater cathode insulation breakdown. Emission from the heater to the cathode will be summed with the intentional cathode current. You have been warned.

The place to slug the dominant pole is at the upper anodes. Theoretically, a capacitor between the anodes does the job. In practice however, individual capacitors are necessary to ensure gain roll-off.

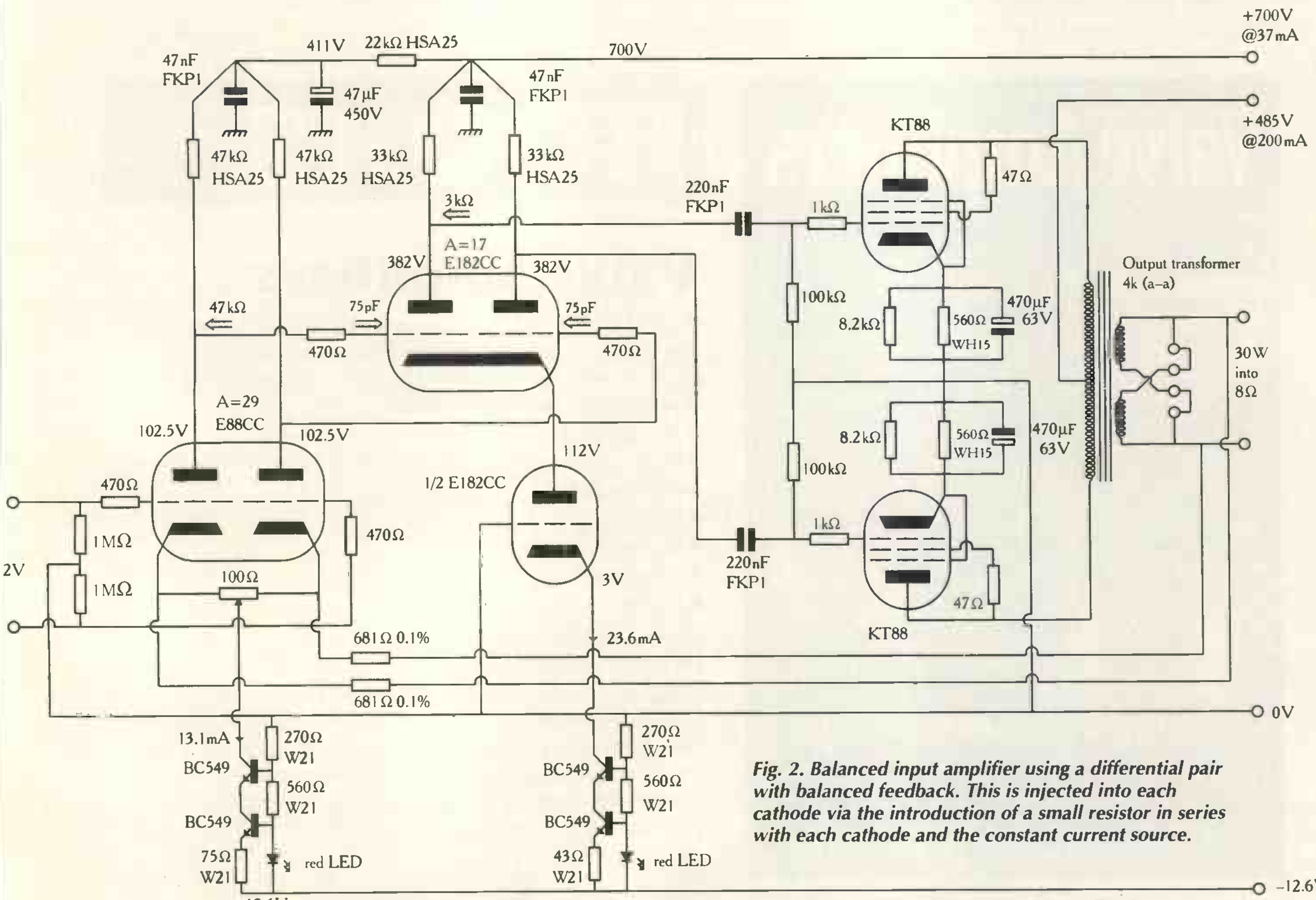



***Fig. 2. Balanced input amplifier using a differential pair with balanced feedback. This is injected into each cathode via the introduction of a small resistor in series with each cathode and the constant current source.***

**Fig. 5.33** *Balanced input amplifier using differential pair and balanced feedback*

As mentioned before, the only really satisfactory valve for use as the lower valve in a cascode is the *E88CC*. Any other type will waste ht.

Cathode voltage on the lower valves is usually quite low, typically around 2.5V. This is insufficient to allow a constant current sink to operate linearly. For this reason, the tail of the sink is usually connected to a subsidiary negative supply.

Feedback from the output can be applied to a grid, which makes calculations of the feedback network much easier.

### Power amps – a balanced alternative

All of the amplifier designs discussed so far accepted an unbalanced input signal presented to the phase splitter. This generates a balanced signal to drive the push-pull output stage. If the pre-amplifier output is already balanced, there would not be a need for a phase splitter. Transmission of the signal from the pre-amplifier to the power amplifier in balanced form would give a great advantage in rejection of induced noise.

The only possible contender for a balanced input stage is the differential pair. But since the input signal is applied to both grids, we need to find a means of implementing global negative feedback. The solution is to add a small resistor in series with each cathode and the constant current source, and inject a balanced feedback signal to each cathode, **Fig. 2**.

Cathode resistors reduce the common mode performance of the differential pair because the cathodes are no longer tied tightly together. Because of this their value should be minimised – a 47Ω resistor is probably the lowest practical value.

Having set the value of $R_k$, resistor $R_F$ can be calculated. Fortunately, this calculation is not nearly as traumatic as that for the 10W amplifier, because the constant current source takes care of the dc conditions. Since each $R_k$ causes such a small amount of local feedback it can be neglected.

If valve balance were perfect, there would be no signal voltage at the top of the constant current source. Therefore this can be treated as being at ground for the purposes of the following calculation. Output of the amplifier is no longer firmly ground referenced, and balances itself about the notional ground of the constant current source. Each leg of the output can therefore be treated as a signal, referred to this notional ground, of half the full output voltage.

The feedback $R_k$, $R_F$ combination is a simple potential divider where the lower arm is loaded by the valve's $r_k$, and the input signal to the divider is half the full output voltage of the amplifier. The required value of $R_F$ can now be found using the normal feedback equation and the potential divider equation, without having to invoke Kirchhoff. However, it is essential that these resistors are accurately matched to avoid unbalancing the circuit. Resistors with a 0.1% tolerance are recommended.

The disadvantage of this arrangement is that the output of the amplifier now has the cathode voltage of the input stage superimposed on both its output terminals. This will not matter to the loudspeaker, even if the cathodes are at slightly different potentials. This is due to the small cathode voltage being heavily attenuated by the potential divider formed by $R_F$ and the loudspeaker voice coil. Also, practical values of $R_F$ would reduce the likely offset to 10mV or less. However, it does mean that neither output of the amplifier may be connected to ground, because this would upset the bias of the input stage. ■

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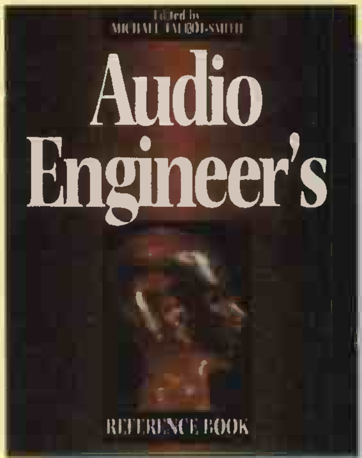
# Reference books to buy

## For Audio Engineers

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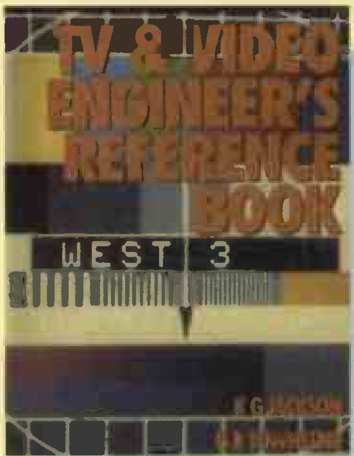


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# Locking onto video sync

**Michael Cox has been designing with a highly integrated solution to sync pulse generation and video genlocking – the *SAA1101*.**

In any video system in which two or more picture sources are to be combined – say, for mixing or special effects – the scanning generators of the picture sources have to be synchronous. This means that the scanning lines must start and finish together both across and down the screen.

If the system is a composite colour one, then the subcarriers in PAL or NTSC must have the same frequency and phase, to within a degree or two. In addition, in a PAL system, the positive and negative line alternation sequences have to be identical.

In older systems, a central sync pulse generator was used, giving timing signals to all sources. More recently, each source carries its own timing generator, which needs to be referenced to other sources in the system. This process is known as genlock.

The design described here produces multi-standard line, field and composite sync pulses, all genlockable to an external composite-video reference.

*Table 1. Operating modes of the SAA1101 are selectable via three logic levels..*

| X | Y | Z | SYSTEM |
|---|---|---|---|
| 1 | 0 | 1 | PAL B/G |
| 1 | 0 | 0 | SECAM |
| 0 | 1 | 1 | PAL M |
| 0 | 0 | 1 | PAL N |
| 0 | 1 | 0 | NTSC |

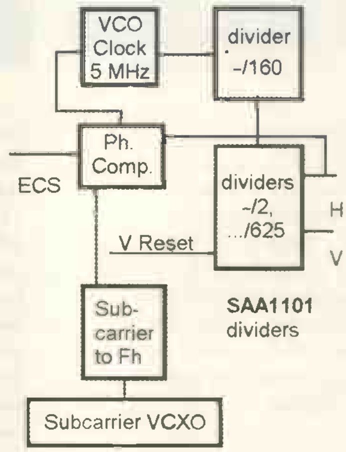



*Fig. 1. Dividers within the SAA1101 allow not only separate and composite sync pulses to be generated, but also allow pulses to be generated for different tv standards.*

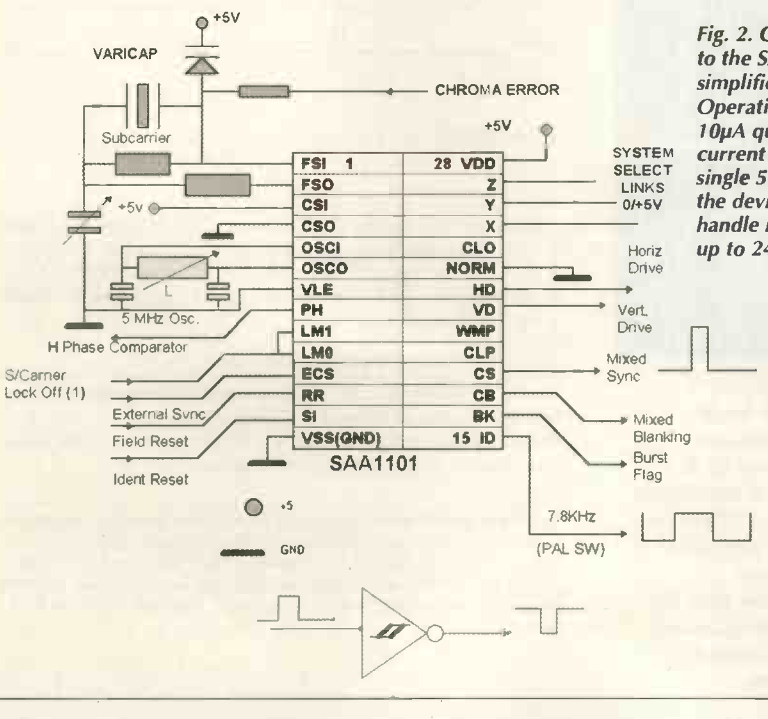



*Fig. 2. Connections to the SAA1101 simplified. Operating with just 10μA quiescent current from a single 5V supply, the device can handle input clocks up to 24MHz.*

## The SAA1101 sync generator

At the heart of the sync generator board is a relatively new chip known as the *SAA1101*, **Fig. 1**. It is a 28-pin device, available in 0.6in spacing dil, or in small-outline sm package.

As **Table 1** shows, by taking pins X, Y and Z to 0V or to 5V, **Fig. 2**, the device can be made to work in PAL, PAL M (Brazil), NTSC and SECAM. In addition the device can operate in 524 or 624 line sequential mode for non standard use, by taking a pin (NORM) to +5V.

It can be set to maintain the frequency relationship between subcarrier and horizontal frequencies, or not. This allows the subcarrier and horizontal oscillators to be locked separately which speeds up the genlock process.

## Genlocking

There is a wide requirement for a genlocking synchronisation generator. Any source, be it camera, video-disc player or test signal generator, may need to be lockable to an external reference signal.

A broadcast synchronisation generator in a studio may need to lock to an external source such as an outside broadcast. The locking process must therefore cause minimal disturbance to viewers.

Normally, horizontal and subcarrier frequencies need to lock almost instantly, while two lines per field are dropped or added until coincidence of the vertical sync block is achieved. This method is rarely used now because of the almost universal use of frame synchronisers at studio inputs.

In a piece of equipment that is fed a reference signal all the time, it is acceptable to reset the vertical counter to achieve rapid field coincidence, **Table 2**. This philosophy is adopted in this design.

## Design details

The starting point for the design is the type of reference signal that is to lock up the synchronisation generator. In this case, it has to be a colour composite signal. In practice, it is usually colour, black, which is 300mV sync, with colour burst but no video.

Refer to main circuit p117. Video input goes to a sync separator IC, the familiar *LM1881*. Note that it is driven by an emitter follower since it has a sync tip restorer that needs to be driven by a low impedance source.

The device also falters if the input, for whatever reason has low level sync, and a high level of burst. It slices the burst as if it is sync and causes chaos downstream of the separator. Hence a simple low pass filter has to be fitted to roll off the chroma before the separator.

Driving impedance requirements dictate that the filter is placed at the emitter follower input, and $R_3/C_{33}$ form the filter. The separator produces a mixed sync output, and a clamp pulse output suitable for colour burst gating.

Vertical output is not so useful as its leading edge corresponds to the first serration in the broad pulse train. This comes too late to reset the field counter in the *SAA1101* synchronisation generator chip.

Schmitt buffer $U_2$, a *74HC14*, inverts the sync signal to drive the ECS input pin of the *SAA1101*. Another section of $U_2$ integrates the sync to produce a field pulse whose leading edge occurs during the first broad pulse. Differentiating then produces a pulse of around 2μs width. These pulses also feed the field reset pin, RR, of the *SAA1101*.

## Internal counters

There are two sets of counters in the *SAA1101*. The first divides the 5MHz clock to horizontal (15.625kHz for PAL) and vertical (50Hz for PAL) pulse trains. The second divides from subcarrier (4.43361875MHz) to $F_h$. This is then compared with the $F_h$ derived from the 5MHz clock to produce an error signal which corrects the 5MHz clock until lock is achieved. **Table 3** shows clock frequencies available by programming *SAA1101* inputs CS1 and CS2.

*Table 2. SAA1101 modes.*

| LM0 | LM1 | Lock Mode |
|---|---|---|
| 0 | 0 | Subcarrier |
| 0 | 1 | slow, ext H ref |
| 1 | 0 | slow, int H ref |
| 1 | 1 | fast, int H ref, V reset |

*Table 3. SAA1101 clock frequencies.*

| CS 0 | CS 1 | Fck | 625 lines | 525 lines |
|---|---|---|---|---|
| 0 | 0 | 160 | 2.500MHz | 2.517842 |
| 0 | 1 | 320 | 5.000 | 5.034964 |
| 1 | 0 | 960 | 15.000 | 15.104893 |
| 1 | 1 | 1440 | 22.500 | 22.657340 |

When the synchronisation generator is free running, overall stability is that of the 4.433MHz crystal oscillator. When it is locked, stability is that of the reference signal used for locking. Two pins on the *SAA1101* chip, $LM_0$, $LM_1$, select the mode of operation, Table 2.

To engage the subcarrier to horizontal lock mode, both $LM_0$ and $LM_1$ need to be taken low. When a reference signal is present, a detector circuit takes $LM_{0,1}$ high, invoking the fast sync lock mode.

Because the slicer input of the separator chip has a sync tip restorer, absence of input makes the output go low. In the ordinary way, this leaves the ECS pin high. This is not a recommended operating mode, as it causes the device to continually slips fields. Accordingly, a resistor and diode, driven from the sync detector, clamp ECS in the low state when there is no locking input to the sync generator.

There are two possibilities for the 5MHz clock oscillator – crystal control or *LC* timing. For speed of pull in, the *LC* type was chosen. The *SAA1101* carries two separate oscillators, both equally suitable for crystal or *LC* stabilisation, Fig. 2.

The internal phase comparator of the *SAA1101* is an edge comparison type, with an output dc of 2.5V.

Because a +5V rail was available, a dc-to-dc converter is used to generate a –5V rail for the error amplifiers and chroma demodulator circuits. Error amplifier outputs can then sit at a nominal 0V, with the varicaps' other electrode taken to ±5V as necessary.

For the 5MHz clock oscillator, an error amplifier is used, to act as a level shifter from the PH output pin to varicap $D_1$. It also provides a convenient point for a phase control. The error signal injects dc into the error amplifier, $U_{8B}$, and shifts the sync edge timing plus or minus a few microseconds relative to the reference input. This is an essential function for integrating a camera or other video source into a practical system.

Smoothing of the error signal is carried out by a *CR* network on the PH pin itself, and by a damping network on the error amplifier feedback loop. Variable inductor $L_5$ is set for correct frequency working, and then finely adjusted to give 0V on the H-Error test point.

## Chroma recovery

In years gone by, various tv receiver chips were available that would serve as subcarrier oscillators, phase-locked loops and identification recovery circuits. Sadly, it is now difficult to find such chips where subcarrier and PAL identification are available to the outside world.

As a result, we have to revert to first principles. The input reference signal is buffered by an emitter follower and drives a very simple band-pass filter, $L_1/C_{10}$. This removes most traces of luminance. Chroma output is then amplified by an *NE592*, $U_5$, and limited by two schottky diodes, $D_{6,7}$, Fig. 3. We are not interested in the chroma during the active line, but a constant amplitude burst is useful.

The limited chroma signal is now demodulated by an *MC1496P*, $U_6$. Its carrier is derived from the oscillator within the *SAA1101*. Recovery amplifier, $U_{8A}$, delivers

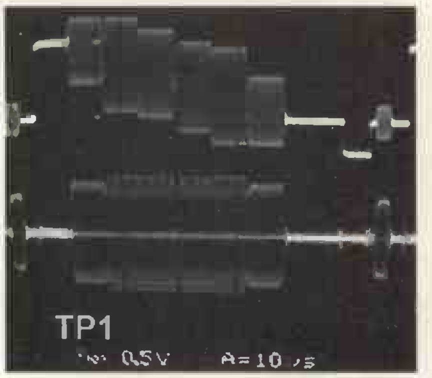



*Fig. 3. Limited chroma. In this design, chroma in the active line is not so important, but a constant-amplitude burst is useful.*

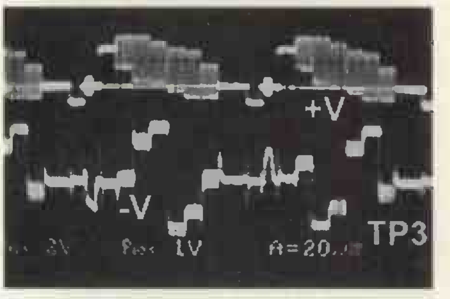



*Fig. 4. Chroma, demodulated using U axis carrier.*

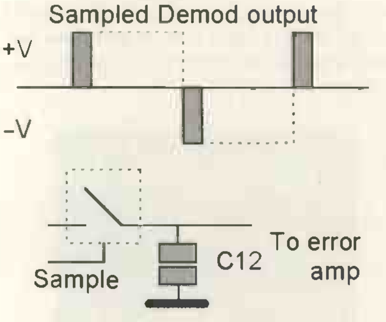



*Fig. 5. PAL switch recovery basics.*

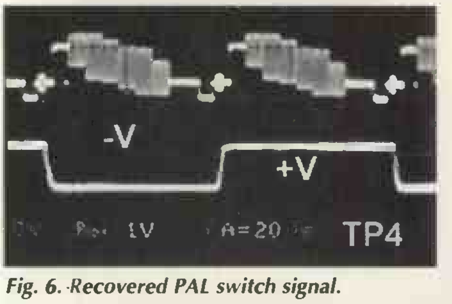



*Fig. 6. Recovered PAL switch signal.*

baseband demodulated chroma to a sample gate, $U_7$, **Fig. 4**. Note that the recovery amplifier and error amplifiers are all sections of a quad BiFet op-amp, in this case, a *TL084*.

The only reason for this choice is that they are cheap, they are reasonably easy to come by – and they work. The sample gate is an *MC4053*.

Output from the demodulator consists of alternate positive and negative burst pulses, together with any active chroma during line time. The gate is turned on during burst time, and storage capacitor, $C_{12}$, has a 7.8Hz square wave on it, **Fig. 5**.

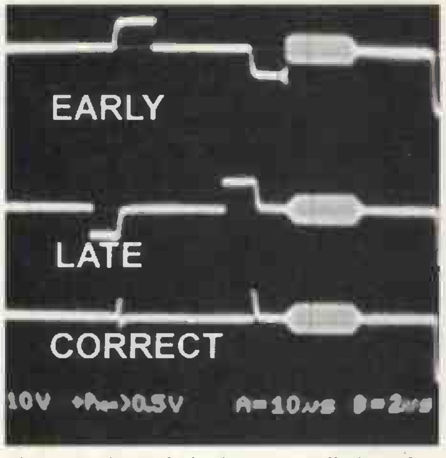



*Fig. 7. Horizontal phasing – cancellation of sync on locking input by sync out from the sync generator, using the A–B display on an oscilloscope.*

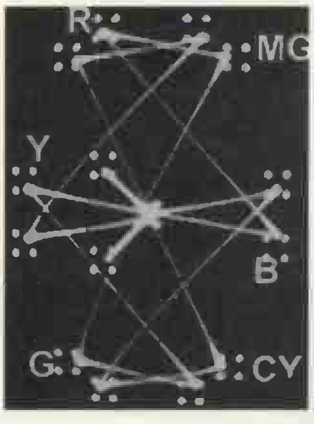



*Fig. 8. Vector scope trace – subcarrier from sync generator is reference, and signal is locking input.*

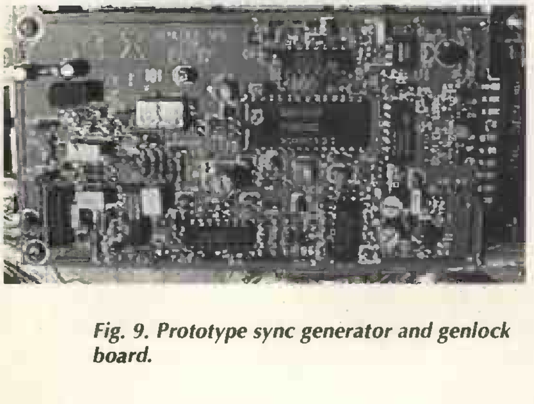


*Fig. 9. Prototype sync generator and genlock board.*

Error amplifier, $U_{8D}$, amplifies this further, and drives the varicap that pulls the subcarrier crystal oscillator into lock with the external reference.

As there is a large amplitude square wave at the chroma-error test point $TP_4$, **Fig. 6**, this is differentiated, sliced by $U_{8C}$, and used to steer the sync generator chip into coincidence with the PAL identification sequence of the reference input.

Subcarrier oscillator output is buffered by a section of another schmitt buffer, $U_{4C}$, and then filtered by an *LC* circuit to give a subcarrier output. Buffer output is also used to drive the chroma demodulator. In this way, the output subcarrier phase is related to that of the reference.

Pulling the *LC* filter circuit either side of resonance provides a simple means of varying subcarrier phase over a range of 180° or so. A similar adjustment to $L_1$ (input chroma filter) although not highly recommended, will allow further adjustment.

## Output stages

My application required negative going pulses at nominal ttl level. As the *SAA1101* outputs are all positive going, a schmitt inverting buffer such as the *74HC14* is ideal.

Because the lowest possible clock frequency for the *SAA1101* is 2.5MHz, waveforms derived from it can only be in multiples of 400ns. The CCIR specification for mixed sync width is 4.7μs, whereas the *SAA1101* gives 4.8μs. As sync is the most critical component – and the most easily measured – it is good practice to try to get it into the middle of the allowed range.

The diode/*CR* network between the mixed sync output of the *SAA1101* and the output buffer delays the leading edge of sync by about 100ns, while doing nothing to the trailing edge. The result is a centre tolerance sync output of 4.7μs.

As already mentioned, the *74HC14* makes a good subcarrier driver. By using capacitor taps on the resonant circuit, subcarrier output is several volts into low impedance.

## Testing the prototype

It is necessary to check the functioning of the locking circuits. Check first that the two oscillators on the chip are running, and at roughly the right frequencies. It helps to have a dual-trace oscilloscope of at least 20MHz bandwidth, and a vectorscope with external subcarrier reference input.

Keep one trace of the 'scope on the locking input, and look at the H error test point with the other. Adjust $L_5$ until the oscillator comes into lock, and then set it finely to give 0V at the test point.

Move to sync output, and set 'scope to A-B. It will be necessary to invert channel 2. The object is to cancel the sync on the locking signal with that from the synchronisation generator.

The H phasing control should be adjusted to bring the two sync signals into time coincidence when one sync cancels the other, **Fig. 7**. Due to differences in rise times between the signals, small 'ears' may be seen at the edges, but when these are symmetrical, coincidence can be assumed.

Return 'scope to normal two channel working. Next, check the 'Demod Chroma' test point. It should bear some resemblance to the waveform shown, Fig. 4.

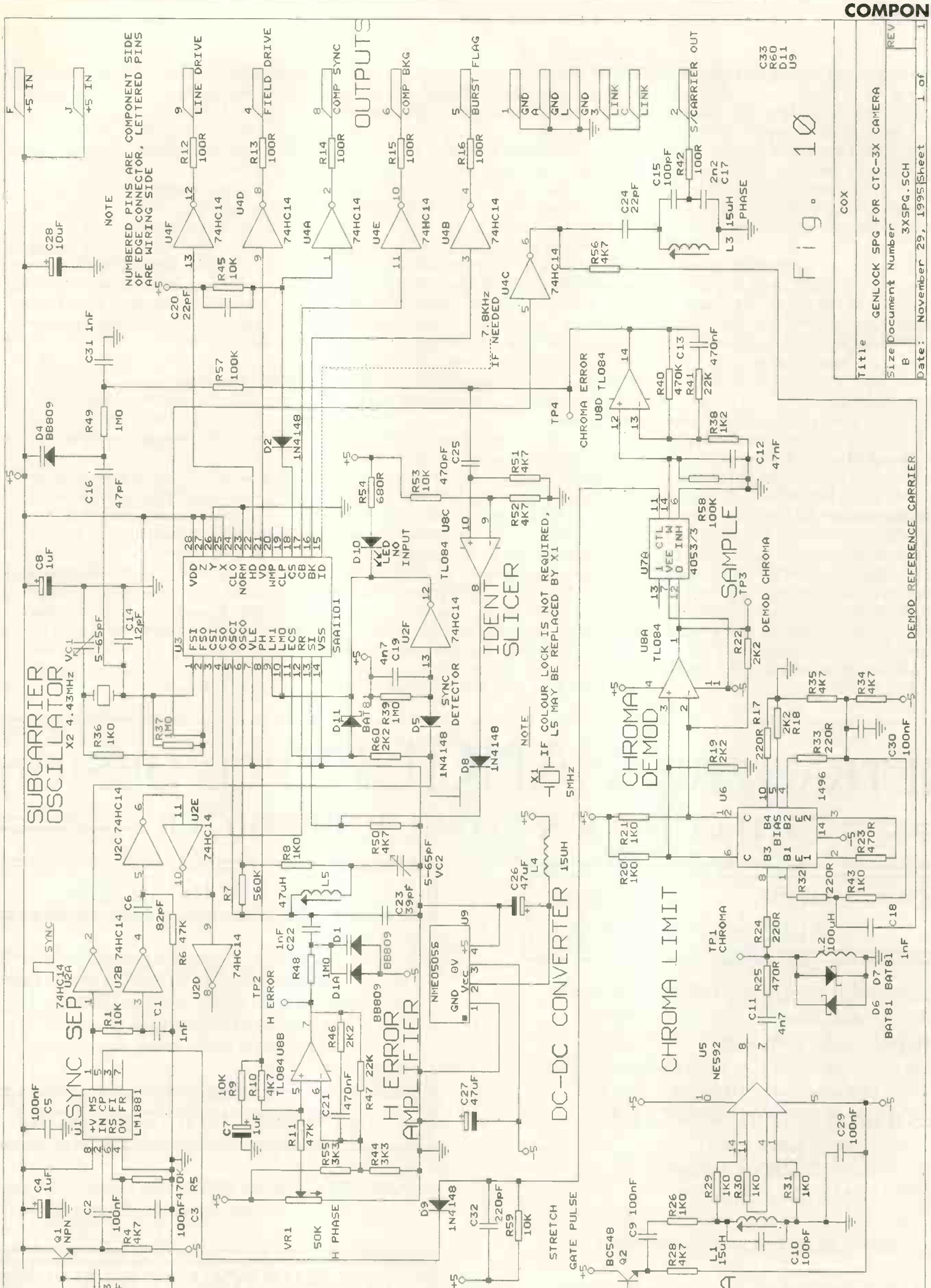
Move to the 'Chroma Error' test point, where a square wave at 7.8kHz should be seen. Set $VC_1$ to make waveform symmetrical about 0V. Connect the subcarrier output to the vectors cope reference input. Loop the sync generator locking signal through the vectors cope channel 1 input.

It helps if colour bars are used for the locking signal, as phase relationships can be seen clearly on the vector display. Check that

***Components for the whole sync generator and genlock circuit.***

| Reference | Part |
|---|---|
| $C_{2,3}$, $C_5$, $C_9$, $C_{29,30}$ | 100nF, ceramic |
| $R_2$ | 75R |
| $R_3$ | 6k8 |
| $R_4$, $R_{10}$, $R_{28}$, $R_{34,35}$, $R_{50\text{-}52}$, $R_{56}$ | 4k7 |
| $C_6$ | 82pF, ceramic |
| $R_7$ | 560k |
| $R_8$, $R_{20,21}$, $R_{26}$, $R_{29\text{-}31}$, $R_{36}$, $R_{43}$ | 1k0 |
| $C_{7,8}$, $C_4$ | 1μF/35V tant |
| $R_9$, $R_1$, $R_{45}$, $R_{53}$, $R_{59}$ | 10k |
| $R_{11}$, $R_6$ | 47k |
| $Q_{1,2}$ | *BC548*, or equivalent |
| $U_1$ | *LM1881*, Nat Sem or *EL4581*, Elantec |
| $U_2$,$U_4$ | *74HC14*, various |
| $U_3$ | *SAA1101*, Philips |
| $R_{12\text{-}16}$, $R_{42}$ | 100R |
| $C_{10}$, $C_{15}$ | 100pF ceramic |
| $R_{18,19}$, $R_{22}$, $R_{46}$, $R_{60}$ | 2k2 |
| $R_{23}$, $R_{25}$ | 470R |
| $R_{24}$, $R_{17}$, $R_{27}$, $R_{32,33}$ | 220R |
| $R_{38}$ | 1k2 |
| $VC_{1,2}$ | 5-65pF, Murata ceramic |
| $R_{39}$, $R_{37}$, $R_{48,49}$ | 1M0 |
| $R_{40}$, $R_5$ | 470k |
| $R_{41}$, $R_{47}$ | 22k |
| $U_5$ | *NE592*, Philips, |
| $U_6$ | *LM1496P*, Nat Sem/Philips |
| $U_7$ | *MC14053B*, Motorola/Harris |
| $U_8$ | *TL084*, Texas, etc |
| $C_{19}$, $C_{11}$ | 4n7 |
| $VR_1$ | 50k, $^3/_8$in sq cermet |
| $D_{6,7}$, $D_{11}$ | *BAT81*, Philips |
| $C_{20}$, $C_{24}$ | 22pF, ceramic |
| $C_{26,27}$ | 47μF, 16V radial |
| $L_2$ | 100μH, axial, |
| $C_{18}$, $C_1$, $C_{22}$, $C_{31}$ | 1nF, ceramic |
| $C_{28}$ | 10μF, 25 v radial |
| $R_{54}$ | 680*R* |
| $L_1$, $L_3$ | 15μH, Toko, 7mm variable |
| $R_{44}$, $R_{55}$ | 3k3 |
| $C_{21}$, $C_{13}$ | 470nF, poly, 0.2in pitch |
| $C_{25}$ | 470pF, ceramic |
| $D_4$, $D_1$, $D_{1A}$ | *BB809*, varicap |
| $C_{32}$ | 220pF, ceramic |
| $R_{57,58}$ | 100k |
| $C_{12}$ | 47nF, poly 0.2in |
| $L_4$ | 15μH, axial |
| $L_5$ | 47μH, Toko, 7mm variable, |
| $X_1$ | 5MHz, 30ppm, 30pF, IQD, EuroQuartz(if $L_5$ not used) |
| $D_{10}$ | led, red, 3mm |
| $C_{14}$ | 12pF, ceramic |
| $C_{33}$ | 10pF, ceramic |
| $C_{17}$ | 2n2, poly, 0.2in pitch |
| $C_{23}$ | 39pF, ceramic |
| $C_{16}$ | 47pF, ceramic |
| $D_9$, $D_2$, $D_5$, $D_8$ | 1N4148 |
| $X_2$ | 4.43MHz, 30ppm, 30pF |

All resistors are 1% $^1/_4$W metal film, (MR25)




*Main circuit of the multi-standard sync generator and genlock. Composite, line and field signal outputs on the right are synchronised with reference composite video, applied at the left-hand side of the drawing.*

adjusting the core of $L_3$ rotates the vectorscope display, Fig. 8.

If a PAL coder is available, use the sync generator to provide all its service signals, and then try to cancel the output waveform with the locking input using the 'scope in A-B mode. If it proves impossible to cancel the burst part of the signal after adjusting the subcarrier phase, the PAL identification sequence is wrong.

It is worth switching the coder on and off a few times to check whether there is a random PAL sequence, or that it is always wrong. A random sequence suggest that the 7.8kHz recovery circuit, $U_{8C}$, is not operating correctly, and needs some attention.

## Implementation

The prototype board was a double-sided board, 3.1in by 5.5in, with a double-row 10-pin 0.156in pitch edge connector.

This defined the size and connections to the replacement. From Fig. 9 you will see that there is adequate space for all the circuitry, on a single pcb, without having to resort to surface mount components.

Lack of a negative rail in the wiring to the sync generator dictated the use of a Newport *NME0505S* dc-to-dc converter to provide a few milliamps at –5V for the op-amps and transistor tails, $U_9$ on the main circuit p117.

The complete circuit diagram of the sync generator card is shown on page 117.

## Developing the generator

For use in a broadcast videotape editing environment, it may be necessary to pay attention to Sc/H phase.

In a PAL signal, subcarrier dot structure repeats over eight fields. In most applications, it is only necessary to respect the four field sequence which is defined by the 7.8kHz PAL identification waveform.

For animation, or where invisible edits are needed, it is necessary to pay attention to the phase of subcarrier with respect to the leading edge of sync on line 1 of field 1. When this relationship is within ±20°, sideways motion on cuts is reduced to a minimum. To achieve this, it is necessary to have a further phase comparison circuit to sample subcarrier phase as stated, and to apply a correction to the H error amplifier.

In some cases, a high degree of subcarrier stability in the free running, i.e. non-genlocked mode may be needed. Most crystals bought off the shelf are cut (AT) with a temperature coefficient that is approximately zero at 25°C.

If the board is in an environment that gets appreciably hotter, then the temperature coefficient gets larger, and the crystal drifts ever further off frequency. Temperature compensated crystal oscillators, tcxo, are available, but usually on extended delivery.

The other solution is to use or make a crystal oven. The crystal has then to be specified to work at 50°C or 75°C, and will be cut to give approximately zero temperature coefficient at that temperature. It will be power hungry – perhaps requiring several watts – and will take some time to reach a stable temperature.

Other outputs available from the *SAA1101*, not used here, are,
- clamp pulse (sync tip)
- 7.8kHz square wave.
- white measurement pulse.

In summary, the *SAA1101* is a significant improvement on its two-chip predecessor, but it is not a real one-chip synchronisation generator, However, it is unlikely that a true one-chip solution would offer the same design flexibility. ■

## Further reading.

Specifications of the world television systems can be found in a CCIR document, published by the International Telecommunications Union (ITU) in Geneva.

Details of the NTSC system can be found in SMPTE Standard 170M, obtainable from, Society of Motion Picture and Television Engineers, 595W. Hartsdale Avenue, White Plains, NY 10607, USA.

Details of the UK PAL I system can be found in "Specification of Television Standards for 625 line System I Transmission in UK", published 1992 by the Radio Communications Agency.

Full details of the *SAA1101* chip will be found in the "Desk Top Video Handbook", published by Philips Semiconductors.

# Modelling cable

**Ben Duncan demonstrates how the behaviour of audio loudspeaker cables can be simulated with increasing realism at audio frequencies.**

Most circuit simulators include a transmission line part. The more sophisticated ones offer several[1]. But for cable lengths below a few hundred metres at 200kHz, and *pro-rata* greater lengths at lower frequencies, cables do not behave like transmission lines[2]. Traditionally, performance has instead been modeled by a lumped network.

## Lumped modelling

**Figure 1** shows the simplest plausible *R-L-C* lumped model, applied in three separate test circuits. Left and right elements are arranged symmetrically. This is done to approximate the fact that substantial cable capacitance is not normally experienced without some series resistance, esr, and inductance of the order of 1μH/m. These parasitics are similar to those found in a real capacitor.

Also, for simplicity, the model is a 'half section' where parasitic values in the return side (lower) have been 'folded-around' into the send side. Overall, the sum of $R_s$ is the total measured values of the send and return resistances in a real cable. This was a 2m length of 2.5mm$^2$ two core PVC insulated flex, as is commonly used for connecting medium to high power amplifiers to full range and bass speakers. But the inductance is defined differently, as the send and return inductances at least partly cancel. Effective series inductance, ESL, is the same as transformer leakage inductance, and the total 2μH is split either side of the capacitance, which is simply the measured value.

In this test setup, the sine source's output impedance is set at 1μΩ, so it can drive the three loads with negligible interaction. On the right, these are a 5.6Ω resistor, which is the resistive portion of a nominal 8Ω speaker; then a 15in drive unit. In this macro, voice coil resistance is external – so it could be stepped. The third load is a macro of a typical two way speaker with passive crossover.

**Figure 2** is a magnified view of the frequency response at each end. The top plot confirms that the response at the source is flat. Below, the other largely flat response is into the resistive load. At hf, it shows how the cable acts as a low pass filter, coming in above 10kHz. This kind of response is often seen in catalogues for shielded cables. The response across the speaker models is more wild, and is a reflection of each speaker's impedance modulus.

With the 2m of cable, the broadest variation is quite subtle at 0.13dB. If the same cable was ten times longer and/or if two cabinets were wired in parallel, or if the resistance was higher,

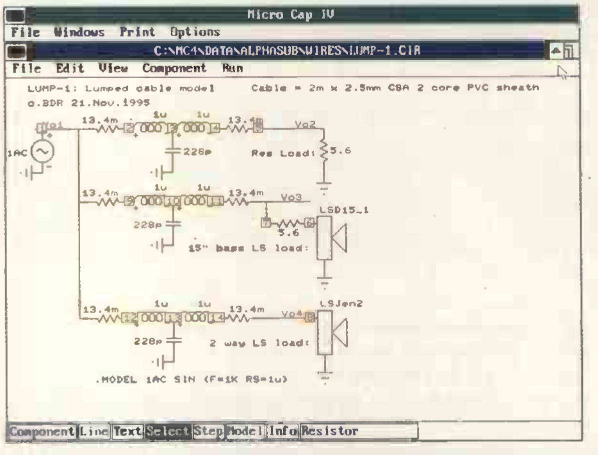



*Fig. 1. Working in MicroCAP IV, the sine source on the left is like a test power amp with infinite current and 1μΩ output source impedance, so it can drive into three different loads down three cable models without its output voltage being significantly modulated.*

*In the centre are three identical lumped speaker cable models, based on 2m of mains cable. These are connected to (top) the resistive part of an 8Ω driver; a model of a 15in bass driver; and (bottom) model of a two way hi-fi speaker. Inside these macros are $R_s$, $L_s$ and $C_s$ that closely simulate the swept impedance characteristics.*

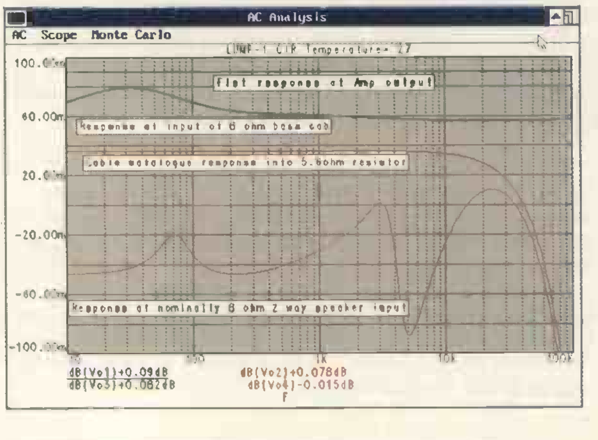



*Fig. 2. Effects predicted by the lumped model as seen on a ±0.1dB scale. On the right, the 8Ω resistance and two-way speaker cause hf roll off, incisive enough to be obvious even on a larger decibel scale. The loudspeaker frequency responses are just evident.*

*You can see that the 15in bass driver resonates at 33Hz, and the two way speaker's bass driver is doing the same at 73Hz. At 4kHz, the lf/hf crossover point is clear from the abrupt phase, followed by a steep inductive rise. Above, the cable's low pass filtration cuts in.*

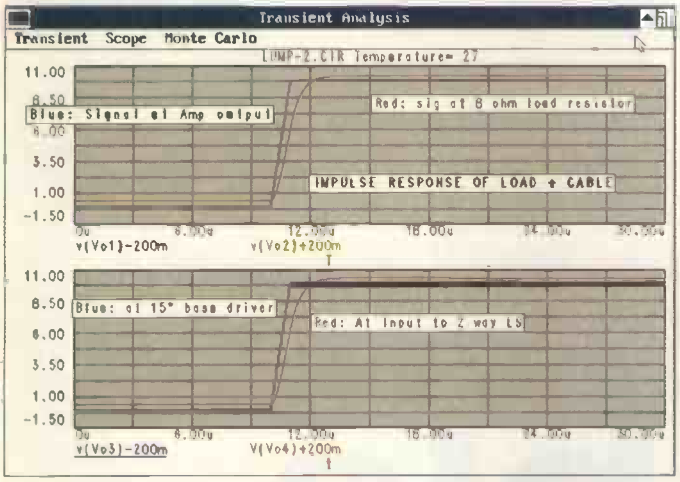



*Fig. 3. Impulse response of the simple lumped cable model driving the three test loads does not tell us much more.*

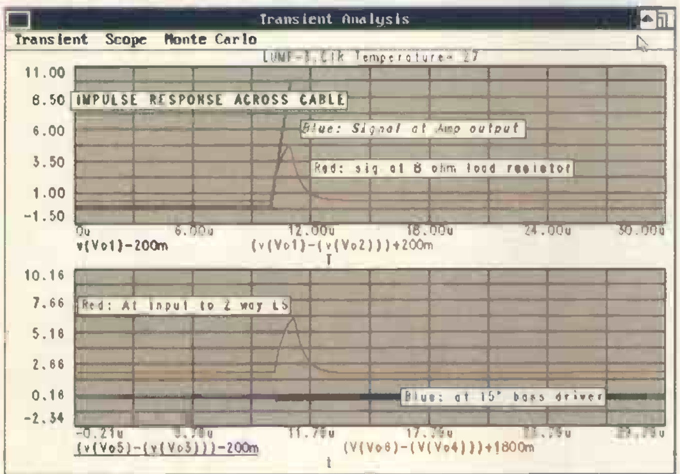



*Fig. 4. Simulation permits differential measurements across lengths of cable that would be difficult or impossible in realspace. Here, the shark fins indicate the leading-edge energy in the pulse that appears across the cable – not the speaker.*

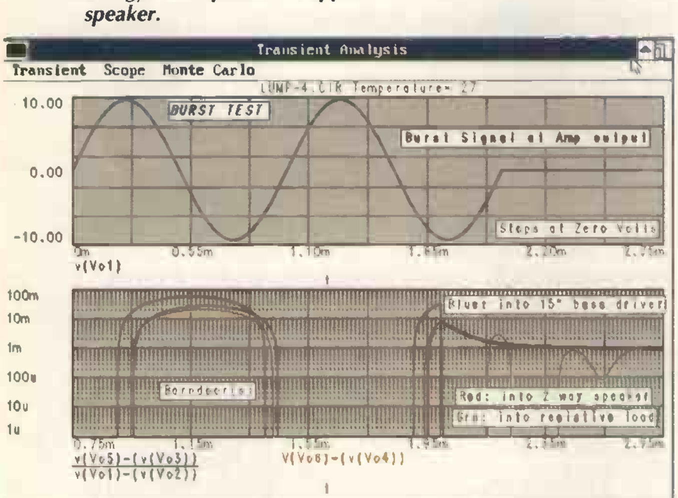



*Fig. 5. The test condition is as Fig. 4, but the signal is a sine wave burst, stopping at 2ms where it is at 0V. Due to phase shift in the cables, the signal at the speaker end at this instant has not reached 0V.*

*The lower graph shows the ensuing losses and damping behaviour across the cable with a log (transient dB) scale. The voltage scale represents current. While smaller than the large signal level (the 'barndoor'), these errors are in a sense infinite, relative to the zero volt stimulus.*

the response variation would be much more prominent. Modelling can come into its own to demonstrate this. This plot highlights the fact that the penalty for inadequately low resistance – and inductance – cabling is not just power loss but also the superimposition of a spurious frequency response deviations.

For **Fig. 3**, the sine source is changed for a +10V pulse with a 1μs risetime, top left waveform, and transient analysis selected. The signal at the load end (offset slightly for clarity) is surprisingly well damped into the two-way speaker model, lower panel. This is because the speaker being modelled has a Zobel network across its input, which helps to make the loading appear resistive at high frequencies. Comparison with the 5.6Ω load resistor's response supports this. The 15in bass speaker has no Zobel network. As a result it is dominantly inductive at the 1μs rise time period (reciprocal of 1MHz), and there is accordingly a small parasitic oscillation, confirmed by smaller timesteps.

## Differential testing

In **Fig. 4**, the test signal is identical, but the measurement condition is now differential *across* the cable. As a result you can see a representation of the current drawn, both into the load and into the cable capacitance, and also the signal that is abstracted by the cable before it reaches the speaker.

Differential testing is physically difficult to perform meaningfully with unshielded, speaker cables of any length. There is a risk of measuring rf contamination. Looping back the cable-under-test for short, noise free sensing connections also requires great care to avoid altering the cable's behaviour. In this simulation, the incoming signal is shown again in the top panel, as a timing reference. Now you can see that the error signal *across* the cable is practically the same for the 8Ω resistor and the two way speaker connections. The error for the bass speaker is much smaller as the inductance draws so little current on a pulse. As before, the oscillation tells us some of the parasitic elements are resonant.

In **Fig. 5**, the test condition is the same except that a two cycle sine wave burst has been substituted (upper panel). In the lower panel, the time scale is a magnified portion. A log scale has been selected so the behaviour can be seen over a 100dB range (five decades). Looking at the 'barn door' centred on 1.15ms, this is the usual appearance of a half a sine wave in logarithmic form.

In the upper panel, 2ms is a zero volt point where the test signal ceases. In the lower panel, the same signal at the speaker ends is not at zero volts due to phase shift. The different behaviours of the currents in each cable can be clearly seen. With the resistive load, the current damps immediately below the 1μV level. The bass driver's current quickly damps to the level represented by 1mV, after which is decays slowly. With the two-way speaker, the current oscillates before settling to the same 1mV pedestal, after nearly 1ms.

## Deeper modelling

The lumped model so far is but a crude first order approximation. It corroborates with little of what critical music listeners hear when different speaker cables are tried. Practical speaker cables are stranded for flexibility. In the real world, copper soon oxidises, or forms other complex ('fractal') molecules on its surface, for example chlorides. Copper oxide is a definite semiconductor. As a result each strand has a longitudinal diodic connection with its neighbour. Below the conduction threshold, the oxide, which can be just a few atoms thick, forms a high value capacitor.

Occasionally and randomly, strands are shorted along the cable length, due to handling, bending and twisting. The only place where the strands are positively ohmically connected is at the connectors, and then only if the wire is soldered or properly crimped. In *MicroCAP IV*, a series of diodes called 'Oxide' were written into the diode library. They have forward thresholds in the tens of millivolt region, and breakdown voltages that are higher but on the same order.

Speaker wires have series inductance and yet must carry substantial peak current, commonly up to 5 or 10A and in some designs, over 100A. Music comprises many sine waves stopping and starting, but series inductance seeks to counter this. Worse, above 3kHz, where transient accuracy is most needed, skin effect intrudes.

This is easy to visualise as highly local eddy current loops within the conductor which subtract from current flow at the interior. In turn, this forces the longitudinal, active current flow increasingly into the outside 'skin' of the wire. The outcome is an additional series impedance that rises at +3dB/octave. This may be viewed as 'the square root of an inductor'. In the green plot in **Fig. 6**, it is modelled using an L-C ladder network. A Laplace function source may be used instead, but only in ac analysis, where Laplaces' violation of causality is not a problem.

Cable insulation acts as a capacitor dielectric. Most affordable cables employ PVC insulation, which is highly polar and suffers high dielectric absorption. For this reason, PVC is not

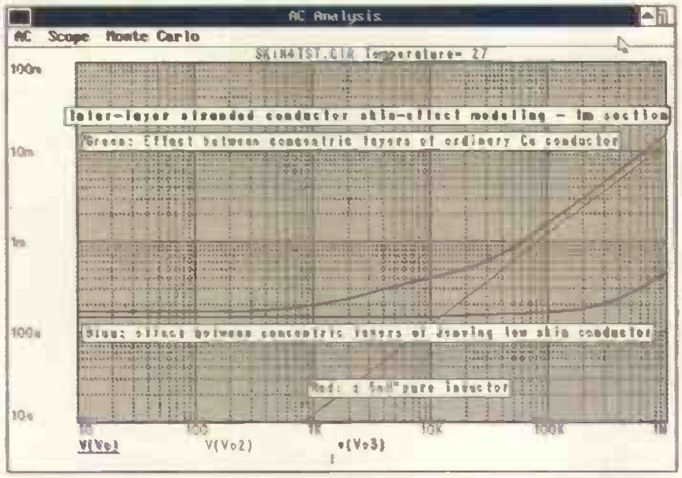



Fig. 6. Two special 'Skin' macros were constructed using L-R ladder networks. Each represents the longitudinal, sectional inductance between strands or bundles of strands, in a stranded conductor. Distances are on the order of 0.3mm section × 1m long.

Here their Z versus frequency is compared to a pure 5nH inductor which slopes at +6dB/octave, exactly twice the rate of skin effect. With ordinary copper, skin effect can be seen coming in below 1kHz, and being swamped by 'real' inductance above 70kHz. With Jenving's patented Supra ply, sectional skin effect is reduced and its onset is displaced over a decade up in frequency.

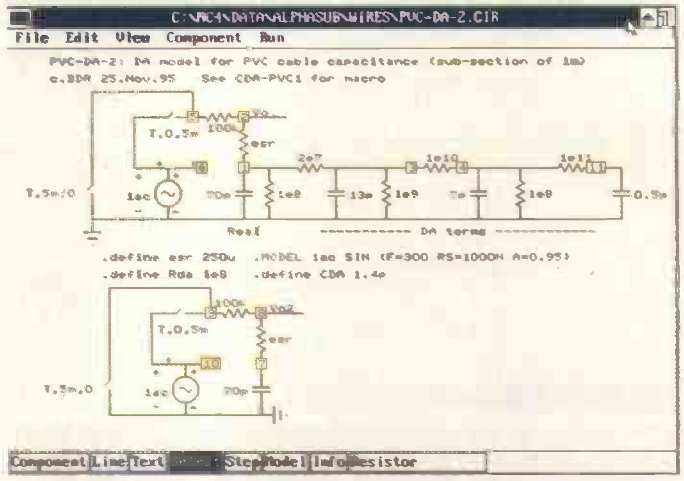



Fig. 7. The RC networks on the right are modeling dielectric absorption in a 0.3mm length of the PVC. On the left, a sine source is connected via a time switch. The point following the switch is shorted after this time. As a result source impedance is zero. In the lower circuit, the stimulus and main capacitor are identical. Only the components simulating dielectric absorption are omitted.

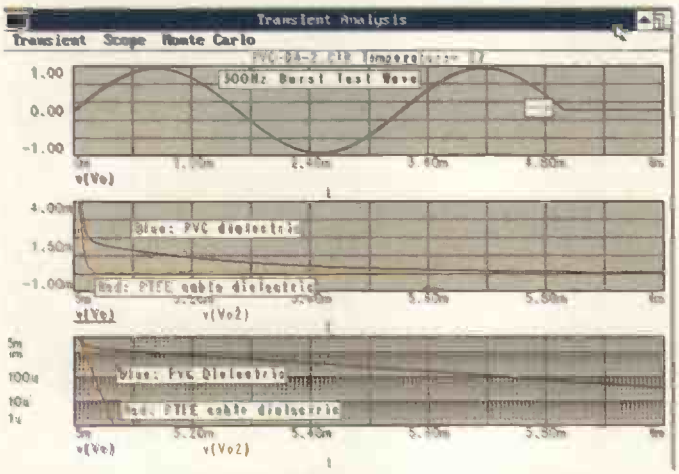



Fig. 8. Effects of dielectric absorption are clear immediately after the burst sinewave signal ceases (upper panel, arrowed). In the centre panel, voltage across the capacitance of an ideal cable falls to 0V within 20µs and at a uniform exponential rate.

However, the discharge of voltage across a PVC dielectric has a dual slope and it takes at least 1ms, or 50 times longer, to come close to 0V. The lowermost panel employs a log scale to show how it has only decayed to 33µV in this time.

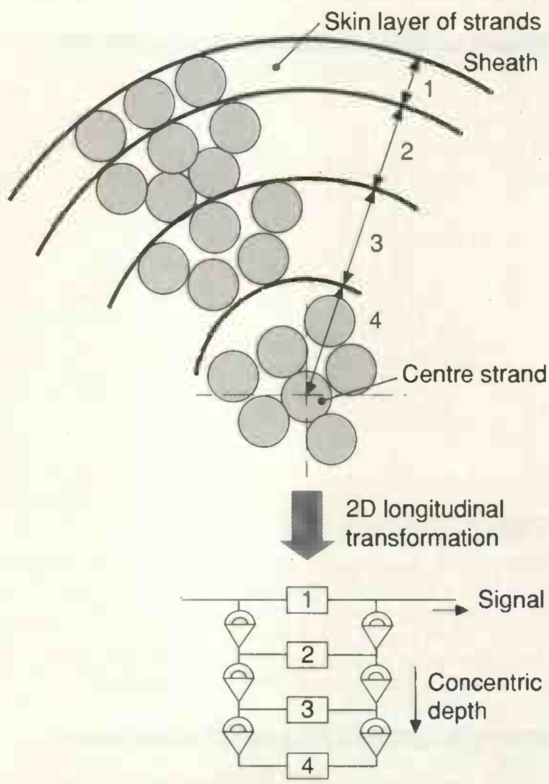



Fig. 9. Top shows end-on view of a round, stranded conductor. Skin effect modelling requires annular sections to be mapped down into two dimensions. For simplicity, annuli of equal longitudinal resistance, hence cross-sectional area are assumed, so the deeper sections are wider.

Lower shows the basic form of the two dimensional map. Each layer is isolated by skin effect. The kite-like device is my proposed component symbol designed to symbolise the cable's shrinking effective ohmic diameter under the curved surface. Signal flowing through 2, 3, 4 is in parallel with the skin (1) at dc but increasingly delayed for transients.

used by capacitor makers. In **Fig. 7** PVC's dielectric absorption is modelled by hanging nested *RC* networks from the 'explicit' capacitance, which is a 70pF. Each 1m section has three of these distributed symmetrically. **Figure 8** compares the behaviour of PVC and a perfect dielectric.

## Self-similar meshing

**Figure 9** is an end-on view of wire which shows how a stranded (or solid) conductor can be divided into annuli of equal dc resistivity. Below, the four annuli are transformed into arbitrary, equal longitudinal sections, where coming down the **Y** axis represents depth-towards-centre. Each annulus is simply in parallel at dc, but at hf, they are divided by skin effect.

The circuit fragment in **Fig. 10** shows the beginnings of a higher order cable model. It serves to remind us that analog electronics is ultimately fractal (self-similar) as elements are repeatedly nested within each explicit *R*, *L* and *C* element. At the top left are the test source and control components. The top line is the outside of the conductor, and the first cable section begins with the 1µH inductance on the right of '$V_{in}$' (this initial section omits skin effect for simplicity). Thereafter '$L_{sect}$' is the series or longitudinal sub-inductance, beginning at the wire's surface.

The circuitry below is deeper into the wire, with the skin effect component, as in Fig. 6, between each layer. In series is the inter-strand capacitance $I_{sc}$ and the oxide diodicity (various oxide

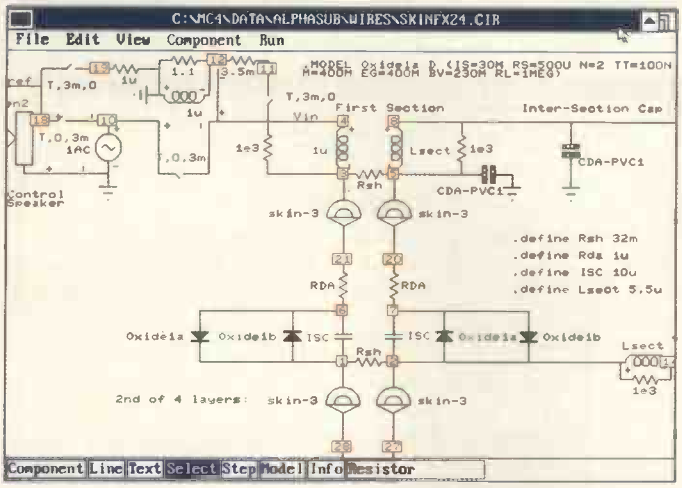



*Fig. 10. Realistic speaker cable modelling begins with the mesh of R,L, PVC cap macros, oxide diodes and skin effect macros seen here. The Y axis is concentric depth, so the lowest layer, not visible hear, represents the innermost strand. At the top left, a second speaker macro is arranged as a control. It receives the same test signal but free of cable.*

*Between the test signal (1AC) and the cable onset is a network through which the cable's input is grounded after the pulse has finished. It comprises a small series resistor and a damped inductor, which simulates the output impedance of a typical power amplifier with high, global negartive feedback. The control speaker is grounded at the same instant.*

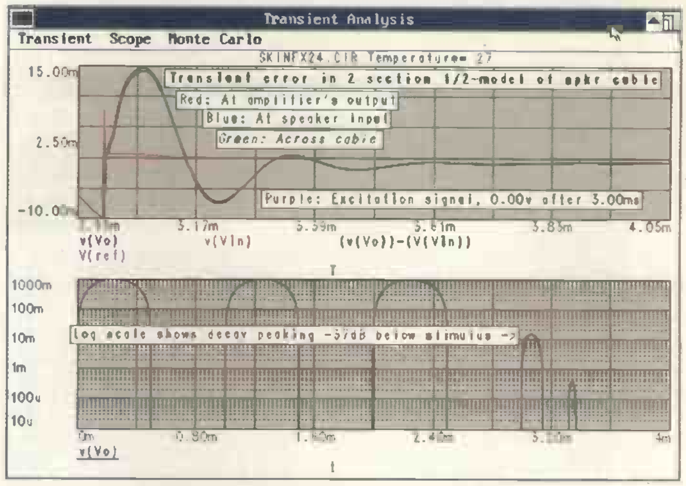



*Fig. 11. Signal residue after the sinewave burst has stopped. The purple plot shows how signal across a speaker fed hypothetically without any cable immediately falls to 0V. The speaker at the end of the cable sections is not so favoured, blue. Initial error is about $^{1}/_{70}$th of the historic peak input signal.*

*This is partly due to finite damping at the amplifier output, red. The green plot looks across the cable to show its dominant contribution. Clearly, with ordinary cable, the amplifier is loosing its grip on controlling the speaker's transient terminal voltage.*

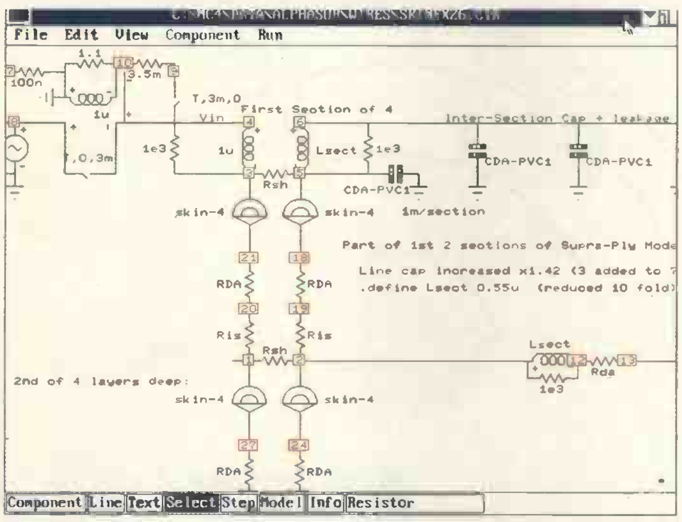



*Fig. 12. In the Jenving cable, simple patented techniques prevent diodicity between strands, so each mesh section is simplified. The Skin-4 macro is the Blue (Jenving) plot in Fig. 6. One extra PVC capacitor covers increased capacitance, while $L_{sect}$ is cut by tenfold.*

*In a full-section model, $L_{sect}$ would be unchanged in effective value, but each would be strongly mutualled to its opposite number by appending 'mutual' statements.*

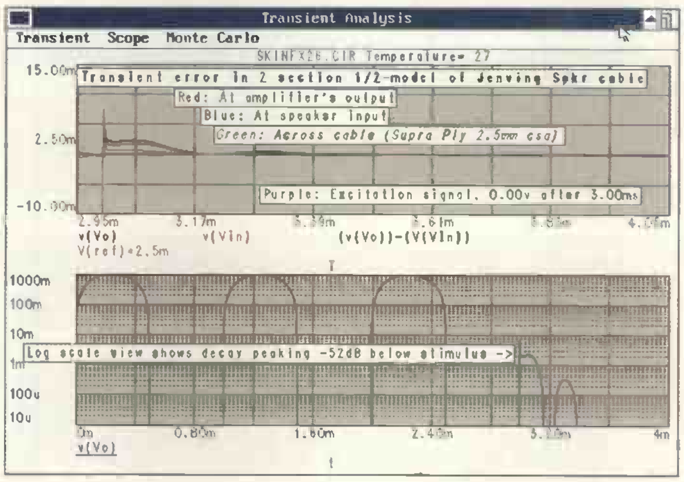



*Fig. 13. Running the Fig. 12 model, showing how the decay error in the Jenving cable construction is much reduced, becoming comparable to the error tail of the power amplifier. This and Fig. 11 corroborate with Audio Precision dsp measurements taken on real cables[4].*

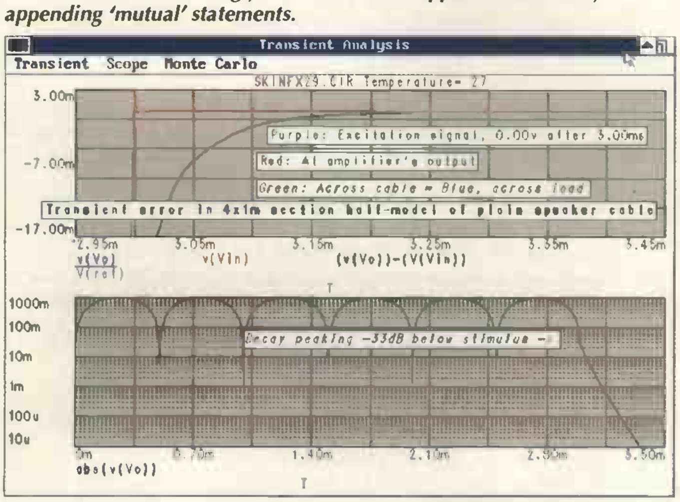



*Fig. 14. Here, the plain 2.5mm cross-sectional cable model (Fig. 10) is driven into a pure 5.6Ω. Peak signal across the cable, green, is surprisingly large. It is about as large as with the speaker load (Fig. 11), but more quickly damped.*

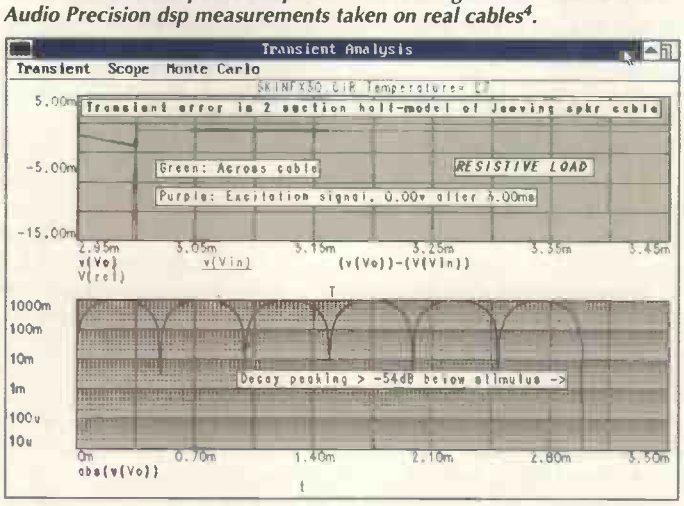



*Fig. 15. Model of Jenving's Supra Ply cable correctly predicts an approximately 20dB lower peak pertubation (–54dB arrowed, lower, where the decay angle changes). The upper plot shows that after 3.0ms, amplifier output error is of the same order or likely dominant.*

diodes 1a, 1b, 1c). Some diodes are 'wild-tied' to other layers, representing a real speaker cable, twisted and crushed after treatment by humans.

Each layer has its own longitudinal sub-inductance, onto the neighbouring section. The shunt resistors ensure the Q of each '$L_{sect}$' is damped below infinity, to avoid unnecessarily prolonging simulation. Along the top line, the cable's shunt capacitance including the dielectric absorption and leakage resistance, is distributed about. Since this, like the lumped one, is a single-ended model for simplicity, the capacitance is to ground. The cable comprises four longitudinal sections, ie. representing 4m. The same Jensen two-way speaker model, as on the left, is connected after section four to ground.

**Figure 11** shows the signal decay after a three cycle sine burst. The model is predicting a pre-referred error across the cable that is 1.5% of 1V, or $^1/_{66}$th of the peak signal level that did exist. Strictly, the error is very much higher, approaching infinity, for at least 1ms, and this is what sensitive listeners notice with music, when the slurring happens repeatedly.

In **Fig. 12**, the mesh values and skin macro have been adjusted to simulate an advanced cable construction, patented worldwide by Jenving of Sweden[4]. It has very high mutual inductance hence low loop inductance ('$L_{sect}$'). Also, it is almost free from skin effect and has no diodicity. Hence it has negligible capacitance between strands. These parasitics are replaced by small resistances. Shunt capacitance is slightly higher but this has no ill effect on transmitting current pulses into speakers; it only concerns poorly designed power amplifiers.

**Figure 13** shows how – excepting the sharp 'inductor spiking' for a few microseconds after 3.00ms, which is a problem for the amplifier's negative feedback – the Jenving construction is 'smears less'. That is, it allows a given, typical loudspeaker load to damp much more quickly and tidily. Events in the 0-30ms period after a given musical attack, can be highly audible, being in the 'early arrival' window before masking room reflections arrive.

**Figure 14** and **15** complete the picture, confirming that ordinary cable can store enough energy to be the cause of much of the initial peaking in Fig 11. Verdict: Mr Nalty may have been right all along[5]. ■

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# Notes on free phasing

**Needing a combination of simplicity, cheapness and very high stability, oscillators for free-phasing electronic organ notes make an interesting design challenge. Ian Hickman discusses his new solution to the problem.**

Practical analogue circuit design is fraught with snags, compromises and difficulties. These are well illustrated by the subject of this article – keyed tone generators – such as might be used in the two tone alarm generator of an hf radio telephone or a hundred other applications.

One such application is tone sources in an electronic organ. There are two main varieties of electronic organ, namely divider organs, and free phase organs. Divider types use a digital 'top octave generator' to produce the twelve semitones of the equal tempered scale. All the intervals are, if not exact, at least very close, and of course 'set in concrete'.

Each semitone output is applied to a binary divider such as the seven stage *CD4024* to provide the lower octaves. Advantages of this approach include cheapness and simplicity. It also produces an organ which is always in tune, but there are a number of snags as well.

With all twelve semitones of seven or more octaves available all the time, each individual note has to be passed when the corresponding key is pressed, or else blocked, by its own keying circuit. It is difficult to obtain sufficient attenuation when notes are not supposed to be sounding, leading to a residual background noise aptly described by the term 'beehive'. Also, square waves contain no even harmonics, so some combining of different octave outputs for each note is necessary if a convincing variety of pipe-like sounds is to be achieved, adding to the complexity. This is especially so where open diapasons are concerned.

However, for anyone wanting at least anything like the richness of sound provided by a real pipe organ, a major snag is the use of dividers to provide the various octave pitches. For example, if while sounding middle C an octave coupler is activated, then C' – the C one octave above – will also start to sound. But since C was obtained by dividing C' by two in the first place, the two notes are locked together and the octave is too perfect.

In fact, all you have done is to change the harmonic content of C: if you didn't hear the

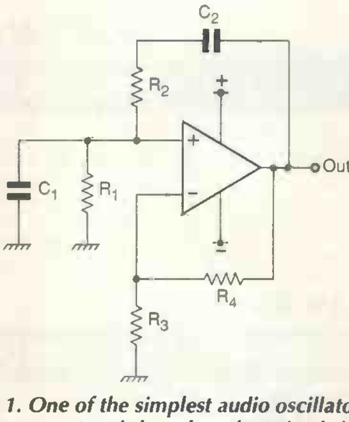



*Fig. 1. One of the simplest audio oscillators or tone generators is based on the Wien bridge.*

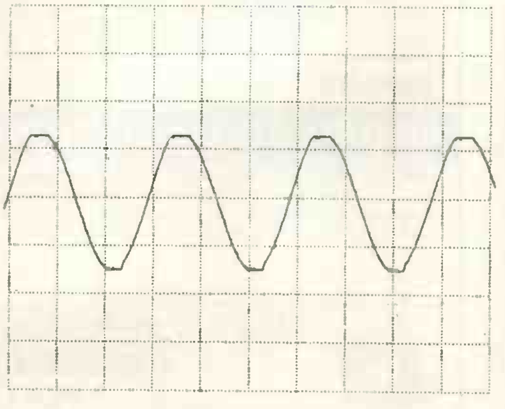


*Fig. 2. Output of an oscillator based on Fig. 1 shows little distortion.*

two notes starting to sound at different times, you would never know that there were supposed to be two separate notes sounding. For this reason more than any other there is still a lively interest in 'free phase' designs, despite the availability of palliatives such as phase modulated delay lines which try to 'unlock' the various octaves.

## An oscillator for free phase designs

A true free phase organ needs a separate oscillator for each note of the rank – or for half that number using an ingenious scheme for sharing one oscillator for each adjacent pair of semitones. This is on the premise that normal music does not require both to sound at once[1]. For example, a flute stop would have 61 generators. The usual arrangement is C,, – two octaves below middle C – to C''' three octaves above.

On an 'eight-foot rank' – so called because eight foot is the length of the lowest pitch open flue pipe of the five octaves – middle C sounds at that pitch. On the other hand on a four-foot rank, middle C would sound the note C', and on a 16 foot rank, the note C,. To simulate the richness of a pipe organ, several ranks of generators are needed, corresponding to the different stops on a real organ. Clearly, economy is a prime consideration in choosing an oscillator design, but equally important is stability. With 61 individual independent generators per rank, retuning would otherwise be an endless chore.

## Oscillator options

In the past, many electronic organ builders have used *LC* oscillators, the inductor using a gapped laminated core. This type of oscillator has the advantage of not needing a separate keying circuit; it performs its own keying function by switching the supply to the maintaining transistor.

Output is taken from a point in the circuit where there is no change in dc level between the on and off states. This avoids keying thumps, while the smooth build-up and decay of the amplitude avoids the slightest suggestion of 'key clicks'. (These clicks plague many other designs of keyed oscillators and keying circuits). Many such ranks are still in use, but the size and cost of using *LC* oscillators provides a strong incentive to seek alternative designs.

I decided to design a cheap simple keyed oscillator needing no separate keying circuit. Instead it only requires a single-pole normally-open switch for each key contact. Some published designs require, at each key, one changeover contact plus two normally open contacts. A single-pole normally-open contact is preferred to a normally closed option since the worst that dust can then do is to prevent a note from sounding when played. A normally closed contact can cause a note to be 'stuck on'.

One of the simplest possible oscillators consists of a Wien bridge and an op-amp, **Fig. 1**. Attenuation from the op-amp output to its non-inverting input via $R_1$, $R_2$, $C_1$, and $C_2$ is infinite at 0Hz and infinite frequency, and a minimum of a factor of three at the frequency given by $f=1/(2\pi RC)$, if $R_1=R_2=R$ and $C_1=C_2=C$. This forms the narrow band positive feedback path.

If attenuation in the broadband negative feedback branch via $R_3$ and $R_4$ is less than 3:1 the circuit will not oscillate. But if it is equal to (or due to the finite gain of the op-amp, slightly greater than) 3:1, then the circuit will oscillate. With no special amplitude stabilising measures, amplitude of the oscillation will build up until limited by the output hitting the supply rails. This causes little distortion if the positive feedback signal at the non-inverting input barely exceeds the negative feedback at the inverting input, **Fig. 2**.

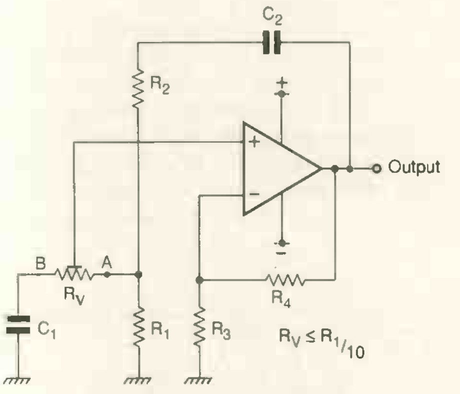



*Fig. 3. Adding trimmer potentiometer $R_V$ permits tuning of the oscillator without changing the attenuation via the Wien network – provided the resistance of the pot track does not exceed a tenth of the reactance of $C_1$. $R_1=R_2=100k$.*

Surprisingly, using the circuit shown, with an *LM324* op-amp, there is no audible change in pitch as the supply rails are varied from ±3V to ±15V.

To make a practical organ tone generator, some means of tuning is required, and this is by no means straightforward. Varying any one of $R_1$, $R_2$, $C_1$ or $C_2$ will change the frequency, but will also change the attenuation in the positive feedback path. Depending on which way the attenuation changes, this can cause oscillation to stop. Alternatively it results in limiting so hard that the signal verges on a squarewave.

A two-gang resistor will do the job, but this

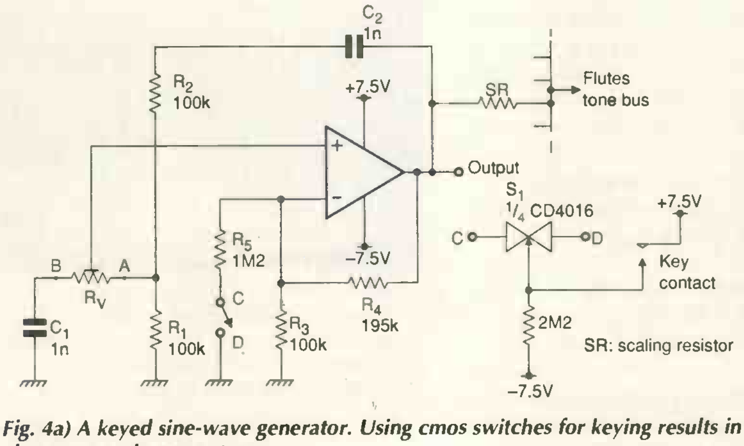
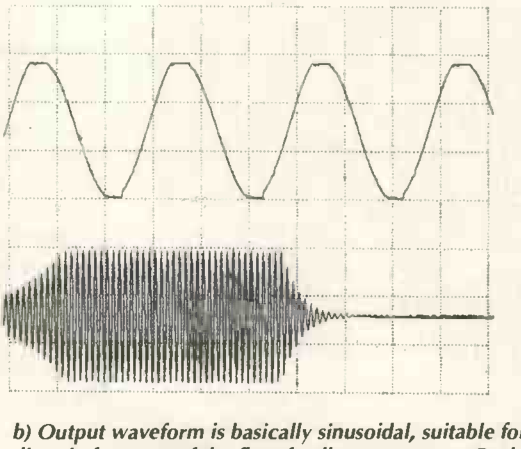



*Fig. 4a) A keyed sine-wave generator. Using cmos switches for keying results in cheapness and compactness.*



*b) Output waveform is basically sinusoidal, suitable for use directly for stops of the flute family, upper trace. Begin and end transients are smooth and free from any incidental dc shift, lower trace.*

is hardly practicable on a one-per-note basis. Fortunately, as is so often the case in analogue circuit design where only a small parameter change is required, a little ingenuity can provide the solution, **Fig.** 3.

If reactance of the capacitor at the operating frequency is ten times the track resistance of the potentiometer, the voltage at B will be only 0.5% smaller than at A. Since these voltages are in quadrature, the voltage across the resistor will be a tenth of that across the capacitor. However, as the wiper of the pot is moved from A towards B, additional phase lag is introduced onto the signal fed to the op-amp's non-inverting terminal.

To compensate for this, maintaining zero phase shift from the op-amp's output to its non-inverting input, the frequency must fall. Due to the low Q of the *RC* network (its Q = 1/3), a small change in phase shift causes a much larger compensating change in frequency than would be the case with an *LC* circuit.

At the operating frequency, the reactance of $C_1$ equals $R_1$. So in Fig. 3, track resistance of the potentiometer should not exceed 10kΩ. This provides almost three semitones tuning range, while a 4.7kΩ pot provides over one semitone.

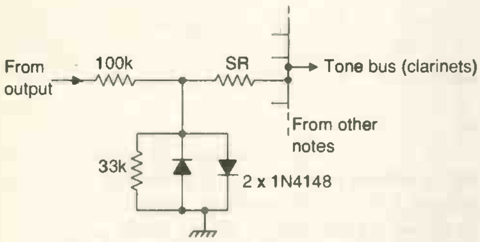
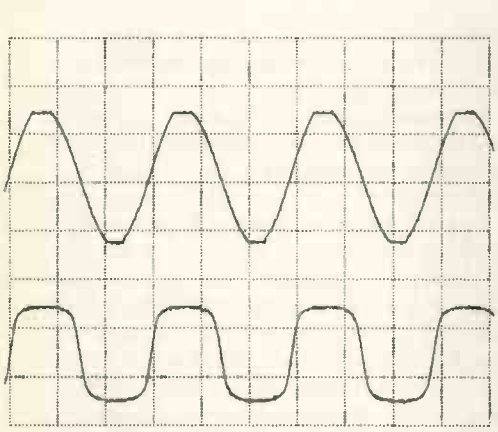
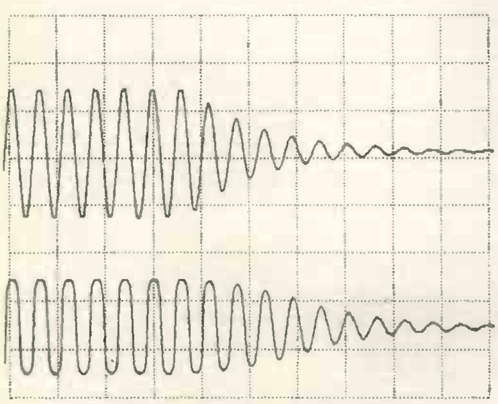



***Fig. 5a) Simple clipper circuit provides an approximation to a square-wave for simulating, say, a clarinet.***



***b) Comparing the 'square-wave', lower trace, with the input sine-wave.***



***c) Due to the limiting action of the diodes, the ending transient of the square-wave output is extended compared to that of the sine-wave.***

## A stable, tuneable oscillator

From my records I found that I developed this circuit in 1982, but I know that it has been independently derived by others[2]. It has a further advantage in that the wiper of the potentiometer feeds an op-amp input, ie a high impedance. Except in the case of wire-wound types, the resistance from one end of a potentiometer to the wiper plus that from the wiper to the other end, exceeds the end-to-end track resistance, due to wiper contact resistance.

Contact resistance is relatively less stable than the track resistance, so tuning by making part of $R_1$ or $R_2$ a potentiometer would be impracticable on stability grounds, quite apart from the incidental change in loop gain. As it is, $C_1$, $C_2$ can be polystyrene types, available in E12 values at 1% or more cheaply 2.5% selection tolerance. Resistors should all be metal film types. Using polystyrene capacitors and metal-film resistors, long term stability of the oscillators should be adequate to ensure that only occasional retuning is necessary.

Over the temperature range 20°C to 60°C, the breadboard circuit exhibited a temperature coefficient of –0.02%/°C, using polycarbonate capacitors. Frequency shift with change of ambient temperature can be expected to be – for all practical purposes – the same for all notes, provided of course that the capacitors used all have the same type of dielectric.

## Designing a keyed oscillator

Having arrived at a stable, tuneable oscillator, it remained to add a keying facility, which can be achieved by altering the ratio of $R_3$ and $R_4$. This has to be effected by the key contact, but the latter cannot be used to modify the component values directly, if – as is likely – it is required to add octave and suboctave couplers. These, when activated, sound the note an octave above, and/or an octave below each note played.

Richness of sound is increased and, because of the inevitable slight departure from exact octaves when using individual generators, creates a desirable chorus effect just as in a pipe organ. As a result, key switches should simply key a dc control signal, instructing the generator to sound when the corresponding key is depressed. The circuit itself will be controlled by an electronic switch. Cmos switches are cheap and readily available and, like the *LM324* op-amp, come four to a pack, for example the *CD4016*.

**Figure 4a)** shows such a keyed oscillator while Fig. **4b)**, upper trace, shows the output waveform, which is basically sinusoidal. Being so, it is suitable for use directly as the

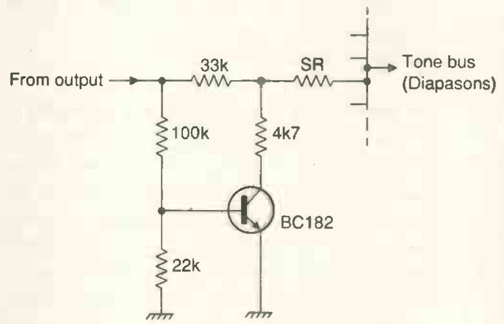
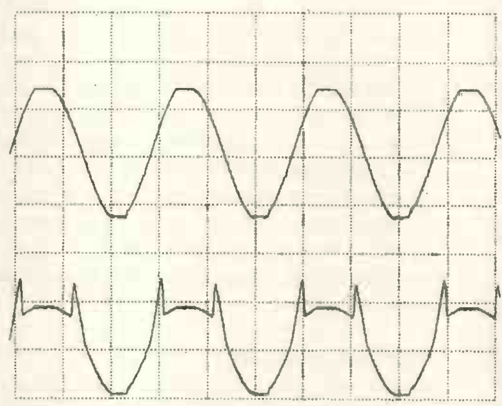
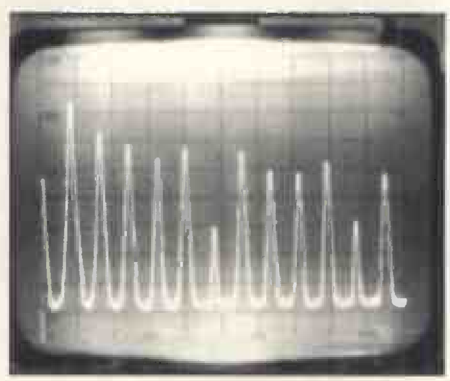



***Fig. 6a) Circuit for adding second – and other – harmonics to the sine-wave.***



***b) Output of the above circuit, lower trace, compared with the sine-wave input, upper trace.***



***c) Showing the fundamental at about 1.7kHz, the second harmonic about 10dB down – about right for an open diapason – and many other harmonics. (10dB/division vertical, 2kHz/division horizontal, span 0-20kHz.)***

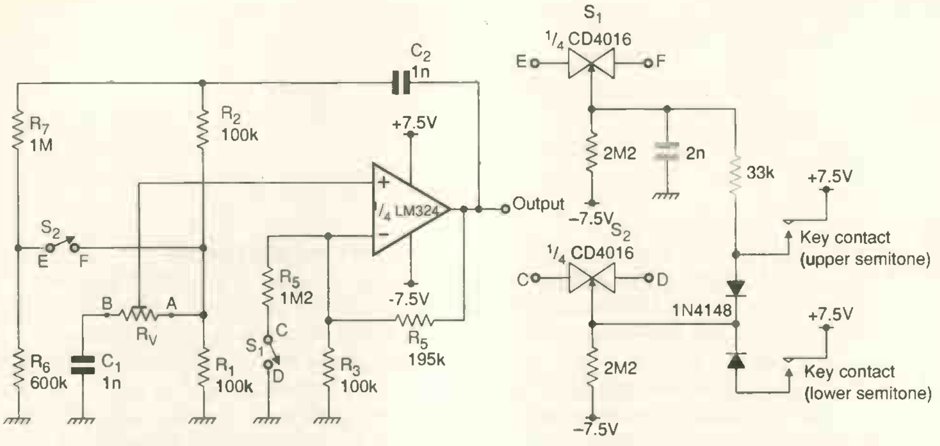



*Fig. 7. Circuit Fig. 6 modified to sound either of two adjacent semitones, according to which key is pressed. Addition of both $R_6$ and $R_7$ keeps the loop gain the same when $S_2$ is closed, leaving the amount of clipping at the rails the same for either semitone (see Fig. 8a).*

basis of stops of the flute family. Figure 4b), lower trace, shows the starting and ending transients. These are clean and smooth. Having no associated dc level shifts, they give complete freedom from key clicks and thumps respectively.

The note sounds when $R_5$ is grounded via $S_1$, one section of a *CD4016*. In view of the supply voltage rating of this device, the circuit is run on ±7V rails instead of the more usual ±12 or ±15V. The 2.2MΩ resistor of Fig. 4a) normally holds the control pin of $S_1$ at –7V, the key contact raising this to +7V to sound the note.

The rate of build-up of the tone depends on how much greater than 3:1 is the attenuation from the op-amp's output back to its inverting input when the key is depressed. The rate of decay is set by how much attenuation is less than 3:1 when the key is released. As a result, by suitable selection of $R_3$, $R_4$ and $R_5$, attack and decay times can be separately adjusted.

Although Fig. 4a) behaves like a high Q tuned circuit, this is only because the feedback is just too much or too little to allow it to oscillate. Where the frequency determining network has a high Q in its own right, for example an *LC* oscillator, the build-up transient will generally be as fast as the decay – or faster if the maintaining circuit is heavily overcoupled.

## Creating other tone colours

While a near sine-wave is fine for flute type stops, waveforms with higher harmonic content are needed to simulate many other pipe sounds. A near square-wave, with its absence of even order harmonics, is ideal for stops of the clarinet family. **Figure 5a)** shows a simple add-on circuit to provide it. One per note is required.

Figure **5b)**, lower trace, shows the 'square-wave', compared with the input sine-wave driving it, upper trace. Due to its rather smooth shape, the harmonics – especially the very high ones – roll off rather faster than a true square-wave, but it sounds very acceptable. Figure **5c)** shows the ending transient, which – due to the limiting action of the diodes – is extended compared with the sine-wave. In practice, this is of no consequence, provided it is smooth, well controlled and free from clicks or thumps. The ear is much less sensitive to the end of a note than it is to its beginning.

For other types of sound, some second harmonic is essential, for example open diapasons. Being a quarter of a wavelength long, stopped diapason pipes are an exception, but even these, if of large square cross section tend to show some second harmonic.

**Figure 6a)** shows an interesting shaper circuit, originally published in an American magazine, and modified here with suitable component values for the available drive voltage. Figure **6b)** shows the output voltage, lower trace, compared with the input sine-wave, upper trace.

Experimentation with the relative values of the four resistors enables a wide variety of waveshapes, and hence of harmonic contents, to be achieved. However, when even harmonics are introduced, the circuit reduces the area under positive-going half cycles more than under the negative-going ones. This means that it introduces a small dc component, which results in an offset at the keyed output relative to ground when sounding.

The result is a slight tendency to produce keying thump, mitigated somewhat by the fact that the driving sine-wave builds up and dies away gradually. This effect is found in nearly all schemes for introducing second harmonic, and the thump can be largely suppressed by passing the output through a high pass filter. The filter need not be provided on a one-per-note basis, but on the other hand one per rank cannot be effective over the whole keyboard.

Figure 6b) type tone generator outputs can therefore be combined on an octave basis, passed through an appropriate high-pass filter and the five filter outputs combined for feeding to further voicing and tone-forming filters. If passed through a high-pass circuit providing attenuation of the fundamental relative to the harmonics, a sound like a really fiery reed stop results.

By these means, three different stop types can be derived from a single rank of generators, but of course in no way does this make it equivalent to three independent ranks. Drawing two of the three stops together simply changes the harmonic content of a note. It therefore contributes nothing to the chorus effect, whereas with two different speaking stops drawn on a pipe organ, two different pipes sound for each note.

Nevertheless, it is convenient to have three different tone colours available, even if drawing them in different combinations merely provides further different tone colours. In particular, one output can be voiced as a very loud stop and another as a quiet one. If the loud one were drawn, the quiet one would not be heard anyway – even on a real pipe organ.

## Cutting cost and complexity

However simple the tone generator, the requirement for one per note per rank means a lot of circuitry is needed.

The scheme of reference 1 sharing a generator between two adjacent semitones is therefore very attractive, but that used a relaxation oscillator. Changing the pitch of a Wien bridge oscillator is not so simple however, as pulling the frequency of a relaxation oscillator. This is because, as noted earlier, while changing either $R_1$ or $R_2$ alone will change the frequency, it will also change the required ratio of $R_3$ and $R_4$.

What is needed is a way of simultaneously changing both $R_1$ and $R_2$, using – for economy – just a single pole switch, such as a single section of a *CD4016*. Here again, as the parameter change required is a small percentage, one equal tempered semitone represents a 5.9% change in frequency – a little ingenuity can supply the answer, **Fig.** 7.

While the two additional resistors connected to switch $S_2$ will marginally increase the frequency of oscillation when $S_2$ is open, values can be found which will cause a further increase of exactly a semitone in pitch when it is closed. This occurs without changing the positive feedback level. As a result the degree of clipping is unchanged – compare the two semitone outputs in **Fig. 8a)** – leaving the harmonic content virtually unchanged, Fig **8b)**.

In 8b), the semitone frequency separation of the two fundamentals is only just visible. However the separation becomes two semitones or about 12% at the second harmonic, and so on in proportion to the order of the harmonic. The starting and ending transients of the upper semitone are also unchanged, due to circuit arrangement maintaining the same

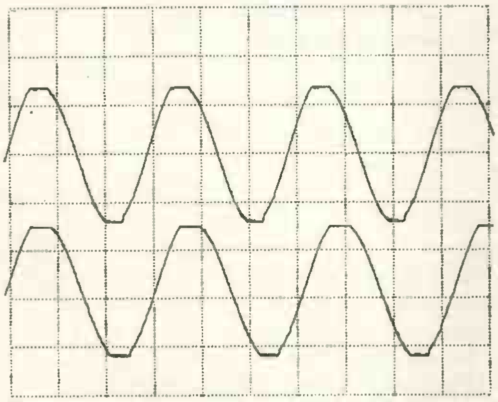
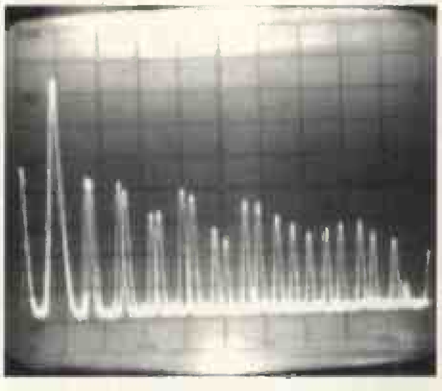
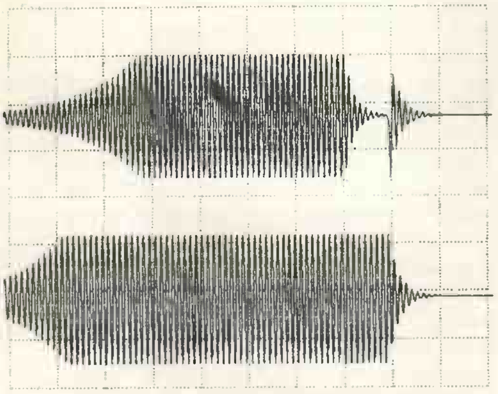


*Fig. 8a) Two sine-wave outputs, a semitone apart.*



*b) As a consequence, amplitude and harmonic content of the circuit's sine-wave output is virtually the same for both semitones. (10dB/div vertical, 2kHz/div horizontal, span 0-20kHz.)*



*c) Delaying the removal of the semitone pitch change control signal to avoid chirp on end transient of the square-wave output when sounding the upper tone causes a hiccup in the ending transient of the upper tone sine-wave output, audible as a slight key click.*

degree of clipping for both semitones.

For experimentation purposes the actual frequencies were regarded as unimportant, the semitone shift being the essence of the exercise. But the two notes – in the region of 1700Hz – correspond roughly to A" and A" flat. There is a small effect on the accuracy of the semitone change, depending on the setting of the tuning potentiometer. This amounts to a few cents more or less than a semitone with the tuning potentiometer at one extreme end of its range or the other, where one cent represents one hundredth of a semitone.

The two diodes in Fig. 7 are arranged so that either of the two adjacent semitone keys will close $S_1$. This causes the note to sound, but only when the key for the upper note is pressed will $S_2$ be closed, giving the higher of the two pitches. If both keys are pressed at once, the upper semitone sounds. In some shared note schemes accidentally pressing both keys together causes a totally different, unrelated note to sound.

With the optional capacitor (2nF) absent, the pitch will revert to the lower semitone immediately the upper semitone key is released. Consequently, this causes the tail of the note to be at the lower semitone frequency. Strangely, this results in but the barest trace of key click on the sine-wave output, presumably because of the rapid decay of the tone, Fig. 4b). However, the decay of the square-wave output is much slower, due to the limiting action of the diodes, and this is clearly visible in Fig. 5c).

On the square-wave output, the pitch change during the ending transient of the upper semitone gives a much more obtrusive key click. The 2nF capacitor suppresses this by delaying the return to the lower pitch when the key is released. The optional resistor (33k) is necessary to control the capacitor charging current, otherwise a key click appears at the beginning of the upper semitone square-wave output.

Unfortunately, while the optional components suppress any key click on either semitone on the square-wave output, they create a very audible key click on release of the upper semitone sine-wave output. This is caused by charge injection in the switch circuit $S_2$, from the control input to that section of the *CD4016*.

With the capacitor delaying the opening of the switch, it now occurs when the sine-wave has all but died away. As the switch is connected directly to the op-amp's non-inverting input, it shock excites the oscillator into ringing – visible on the upper trace (upper semitone) in Fig. 8c). By comparison, the lower semitone sine-wave output is of course unaffected, lower trace.

## Further enhancements

Charge injection in electronic switches is a well known phenomenon, and in later designs of switch ICs it has been greatly reduced, but these would be too expensive in the numbers required for this application.

Clearly there is scope for further development here. For example, the capacitor at the control input of $S_2$ could be grounded not directly, but via another section of the *CD4016*. This additional section would be switched on when square-wave was selected, but not for sine-wave. All additional switch sections would have their control inputs connected together and controlled by the stop switches, being on for clarinet (square-wave) type stops but off for flutes (sine waves).

Having concentrated on the basic one-per-note, or one-per pair of notes tone generator, a word on controlling the generators from the keyboard will not go amiss. For a very simple organ of just one rank, key switches can control $S_1$ for each note directly, and $S_2$ – if using the shared generator scheme – via diodes as in Fig. 7a).

If it is desired to incorporate octave and sub-octave couplers, this can be achieved by adding diodes and resistors. However the complexity increases alarmingly, especially with the shared generator scheme. It increases further if it is desired to have two or more ranks of generators with the option of sounding these at different pitches. As a result, for all but the least ambitious designs some other scheme is called for.

A microcontroller can be used to scan the keyboard and set or clear latches controlling $S_1$, and $S_2$ if used, in accordance with the stops drawn. But a simpler approach is to employ one of the variations on the multiplex scheme, which has been described many times in the literature, for example Refs 3 and 4. ■

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# Fractional-*N* synthesis

**Cosmo Little reveals the benefits of a little-known frequency synthesis technique that is cheaper than direct synthesis – and requires much less power.**

A fractional-*N* synthesiser is essentially a single-loop digital synthesiser where the loop divider has been modified to divide by an integer plus a fraction – instead of by an integer alone. In this way, extended frequency resolution is obtained.

Because division by a fraction is not possible, the divider approximates to the required fractional division ratio over a period of time.

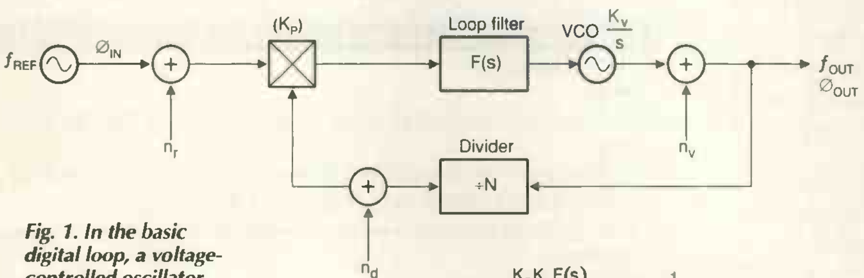



*Fig. 1. In the basic digital loop, a voltage-controlled oscillator running the output frequency is divided by integer* N.

$$G(s) = \frac{K_p K_v F(s)}{s} \qquad H = \frac{1}{N}$$

Transfer function $\dfrac{\phi_{OUT}}{n_r} = \dfrac{G(s)}{1+G(s)H}$

Reduction of $n_v$ or $n_d$ $\dfrac{\phi_{OUT}}{n_v, n_d} = \dfrac{1}{1+G(s)H}$

$n_r$, $n_d$, and $n_v$ are additive phase noise sources

Time to acquire phase lock (zero initial frequency error)

$$\approx \frac{1}{2\pi f_n} \approx \frac{1}{\omega_n}$$

Time to pull in from frequency offset $\Delta f$

$$\approx \frac{(2\pi\Delta f)^2}{2\xi\omega_n^{\,3}}$$

(2nd order loop with phase detector) Note: this formula does not apply to phase/frequency detector

In this first article I review the operation of the single loop digital synthesiser, and will consider the basic type of fractional-*N* synthesiser. In the second article, we will conclude the review of fractional-*N* techniques, and will look at practical implementations of the method. Both articles will be illustrated by *MathCad* simulations which will be made available to readers as the original *MathCad* documents. This will enable readers to experiment with the design of fractional-*N* synthesisers to suit specific applications.

## Benefits of fractional-*N*

The fractional-*N* synthesiser is believed to have been invented by Hewlett Packard in the early eighties. It was used in a number of synthesisers – including the *3325A*, which achieved micro-hertz resolution with only one phase-lock loop.

The same concept has been used in other instruments, such as the Marconi *2022* signal generator, and can be used as the high resolution loop in a multi-loop synthesiser which increases the possible frequency range into the gigahertz region.

Modern digital large-scale integration chips have generated considerable interest in the direct digital synthesiser, which has largely eclipsed the fractional-*N* technique. nevertheless, fractional-*N* has many advantages, and is worth serious consideration by any designer faced with the task of producing a medium performance frequency synthesiser with small frequency resolution.

Fractional-*N* can be implemented much more cheaply than a direct digital synthesiser, or dds. It also uses much less power as it does not require fast logic or a fast video d-to-a converter for the signal reconstruction.

This article reviews the different evolutions of the fractional-$N$ synthesiser from the most basic idea of non constant division ratios in a digital phase-lock loop, pll, to the development of the two accumulator fractional-$N$ loop with its division by a choice of four different integers and analogue correction.

Performance of the fractional-$N$ loop is well suited to simulation on a general purpose maths package, as its performance limitations can be investigated independently of performance limitations caused by the hardware implementation. In this article, analyses are supported by *MathCad* documents.

Initially, I will review the basic digital loop from a performance point of view, and discuss the basic trade offs between resolution, reference frequency suppression, phase noise, and tuning time. These relationships are not derived formally, as this has been done many times in existing literature.

## Basic digital pll

**Figure 1** is a block diagram of a single loop synthesiser. This will no doubt be fairly familiar to most readers. A voltage-controlled oscillator running at the desired output frequency, $f_o$, is divided by an integer, $N$.

Phase of the output frequency from the divider is then compared with the phase of a reference frequency, and the phase error is converted to a voltage error by the phase detector. This voltage error is processed by the loop filter, which is usually an integrator and zero, with one or more higher frequency poles.

Output from the loop filter is now used to correct the vco frequency to hold the phase error at the phase detector to near zero. This loop can be formally analysed to predict such behaviour such as loop stability, forward transfer function – which describes how closely the phase fluctuations of the vco follow those of the reference – behaviour of the loop to phase errors added to the vco signal, ie suppression of phase noise of the vco, and many other performance characteristics. I do not analyse the digital loop in detail here, but I have included references to the details.

This is a summary of the most important performance trade offs:

**Frequency resolution.** This equals the reference frequency. Output frequency is $f_{ref} \times N$

**Natural loop frequency**. This is defined as the –3dB point of the closed loop gain. It is an important parameter that affects several other performance measures. For generality it is best considered as a fraction of $f_{ref}$.

Note that the digital loop is actually a sampled data system. Phase comparisons are only available at the rate of the reference frequency. As a result, in theory, the loop frequency cannot be higher than $f_{ref}/2$.

In practice, the loop frequency will need to be $f_{ref}/10$, or even lower. The main reason for this will be to suppress leakage of unwanted high frequency signals from the phase detector. These occur at multiples of the reference frequency, and will phase modulate the vco, producing spurious sidebands.

Some phase detectors are better than others in this respect. The sample and hold detector is generally considered to have the lowest reference frequency feed through. However, the digital phase/frequency comparator is still used in a great number of single chip synthesisers. The choice of phase detector becomes important for another reason concerned with the fractional-$N$ mechanism, so I will return to the discussion of phase detectors later.

## Designing the loop

The natural loop frequency – and damping factor – may be calculated from the component values used in the type 2, second-order loop filter. The circuit and equations are given in **Fig. 2**. Real implementations, however, always have additional poles, and the simplest way to find the natural loop frequency, and also the phase and gain margins, is to plot the closed loop gain and phase. This is easy with tools such as *MathCad.*

The op-amp is modelled as an amplifier with finite gain and one low frequency pole. Component values in Fig. 2 for the fifth-order loop are entered. The loop divider ($N$), phase detector constant (Kp), and vco gain constant (Kv) are also entered. The document then plots graphs of closed loop gain and phase, forward gain, reduction of vco phase noise, and loop filter response. This enables estimates of natural loop frequency, loop stability, reference frequency suppression, and vco phase noise to be made.

The natural loop frequency affects a number of performance tradeoffs of the digital loop. Those that are most important to synthesiser designers are phase noise, spurious sidebands, and tuning time.

Phase noise of the vco can be divided into two regions, frequency offsets well within the loop natural frequency, and frequency offsets well outside it. At frequencies well within the loop frequency, the vco is controlled by the reference frequency phase. Any phase fluctuation of the reference will appear on the vco output, multiplied by the loop division factor, $N$. In terms of sideband level, the increase in level from the reference to the vco will be $20 \times \log(N)$.

For example, if the reference has 100Hz sidebands at a level of –80dBc (80dB below the carrier) due to phase modulation , and $N$ is 1000, then the 100Hz sidebands will appear on the vco at a level of –20dBc.

Exactly the same calculation applies to noise sidebands. In the example a reference phase noise level of –80dBc/Hz will be transferred to the vco at a level of –20dBc/Hz. If $N$ is very large, the vco may well end up with worse phase noise than if it was free running.

This in fact was the case in an early version of a low-cost signal generator made by a famous manufacturer. As far as I can remember, the vco operated at 160MHz, with a comparison frequency of 100Hz. In this case $N$ would be 1.6 million, and the phase-noise multiplication factor 124dB.

Within the loop bandwidth, which was only about 10Hz, output of the signal generator had so much residual fm that it produced an audible wobble when used to test ssb receivers.

### Alternative frequency synthesis techniques

Methods of frequency synthesis can be divided into a number of distinct types.

Direct frequency synthesisers use only harmonic multipliers, mixers, and dividers to generate their output. No phase-locked loops are used. This type of synthesiser switches in a few μs, but is extremely complicated, with a vast quantity of analogue circuitry. They are essentially obsolete. Interested readers are referred to the block diagram of the *HP5105* in reference 3.

Digital synthesisers use one or more digital phase lock loops, ie a phase-locked loop with a digital divider between the vco and the phase comparator. More than one loop is commonly used, with mixers inside and outside the loops. A huge variety of different configurations have evolved to provide small frequency increments with fast lock-up time and good phase noise.

A large number of frequency synthesisers in everyday use are single-loop types. This includes all domestic radios and tvs, and the great majority of commercial communications equipment operating in the vhf, uhf and low microwave bands. The use of multi-loop mixing types is mainly in laboratory instrumentation such as signal generators and spectrum analysers, and in hf transceivers.

The third type of synthesiser is the direct synthesisers. These synthesise the output sinewave by outputting a sampled approximation of the waveform at a rate determined by a high-frequency clock. As all the circuitry is digital, it is suitable for large scale integration. This type is becoming more popular with the astonishing increase in density, and reduction in price, of LSI digital chips. A direct synthesiser may be used in a single loop with mixer configuration to provide small frequency increments with reasonable simplicity.

This made measurement of SINAD near impossible.

The manufacturer soon replaced this signal generator with a version that used a fractional-$N$ loop to avoid this problem.

## Calculating phase noise

This digression all helps to illustrate the problems with digital loops with high loop division factors.

Note that reference noise must include divider noise and phase detector noise. If this is not done, it is possible to make a large error in the calculation of phase noise within the loop bandwidth. Reference oscillator phase noise will be reduced by the reference divider chain by the same factor $20\times\log(R)$, where $R$ is the total divider ratio.

For example, a crystal oscillator at 10MHz may have phase noise of –100dBc/Hz at a 100Hz offset. If this was then divided down to 100Hz, you might assume that the phase noise at the divider output would be –200dBc/Hz from the above formula. But this would be incorrect, as the divider noise floor is unlikely to be below –160dBc/Hz. For examples of divider and phase detector noise, see reference 1, p86.

Outside the loop bandwidth the vco is uncontrolled by the loop – except for the unwanted reference frequency components which generate discrete sidebands. As a result phase noise reverts to that of the free running vco.

A loop with low reference frequency is used to obtain high resolution, and a low loop natural frequency suppresses reference frequency sidebands. But with such a combination, the phase noise may be too high to meet the designer's initial specification. Nothing can be done about this except to build a better vco.

## Phase shift considerations

Spurious discrete sidebands have already been mentioned. These are generally caused by reference frequency leakage from the phase detector.

In an attempt to push the loop natural frequency higher, designers have come up with elaborate filtering schemes after the loop filter/integrator. These often take the form of twin-T notch filters tuned to the reference frequency and its second and even third harmonic.

These can be useful, but it is essential to consider the effect of the phase shift of these filters on the loop stability. This could be done using the *MathCad* document for the pll, by adding the $s$-plane transfer function of a twin-T notch filter to loop filter function $F(s)$.

Tuning time, which is the time to acquire phase lock after a new frequency has been programmed, is a very complicated parameter to analyse. This is because it involves non linear operation of the phase detector, with the loop initially slipping cycles during acquisition.

Phase detectors of the digital phase/frequency tristate comparator type will always acquire lock. Even if there is a large initial frequency difference, they always provide an output to pull the vco in the correct direction. They are almost always used in commercial single-chip digital synthesisers, although often an alternative phase detector is provided for use once the loop has locked.

Tuning time is always inversely proportional to the loop natural frequency. Some approximate formulae are given in Fig. 1.

## Multi-loop synthesisers

To summarise, a single-loop synthesiser is simple and economical, but has serious performance problems when fine frequency resolution is required.

Many transceiver applications have fixed channel spacings of 12.5 or 25kHz, and in these cases it is possible to design single loop synthesisers with vco frequencies up to 200MHz with adequate performance.

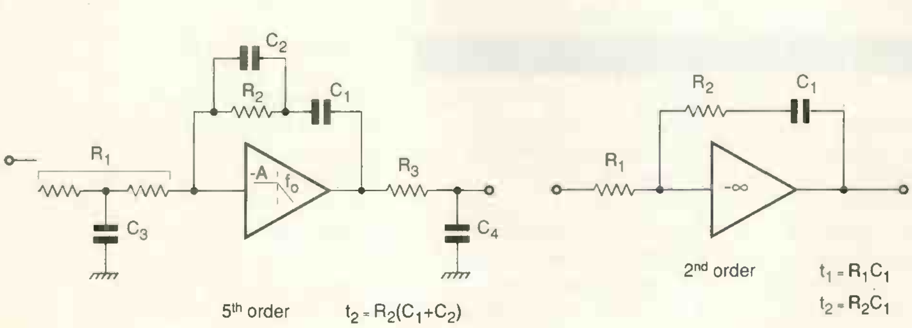
The same applies to domestic radios where a channel spacing of 9kHz is used, in Europe, on the amplitude modulation broadcast bands, and a channel spacing of 100kHz on the fm broadcast band. But what about a general coverage short wave receiver?

In this case the receiver will probably use up-conversion to a first intermediate frequency of 45MHz, with a vco range of 45 to 75MHz. At least 50Hz resolution is required for ssb reception. A single-loop synthesiser will be terrible, as you can appreciate from the formulae provided.

The traditional solution is multi-loop synthesisers – synthesisers with mixers in the loop. These can be incredibly complicated. A circuit of an early Yaesu amateur transceiver is typical of Japanese transceiver design in the eighties. It uses four internal vcos – not including the output vcos – with four single-chip programmable divider/ phase detectors, five mixers, six fixed divider chips, and a mass of loop filters, low-pass, and tuned filters to try and control all the unwanted mixer products. To be fair to the Yaesu designers, current amateur transceivers use much simpler schemes, usually based on a direct digital synthesiser – dds – as part of a dual loop.

Direct-digital synthesis is a method of directly generating a sine wave output by accumulation of phase, table look up of the sine function, and reconstruction of the output with a d-to-a converter. Analysis of the dds would require another article, but it is not the universal solution to synthesiser design that some manufacturers of the chips might claim.

Due to the fast logic required – clock rates of at least four times the output frequency are required – and the fast d-to-a converter, prices of the chips are very high, around £20 to £40 for the dds chip, and about the same for the d-to-a converter. Spectral purity is a problem, especially if the output frequency is multiplied in a digital loop.




$t_2 = R_2(C_1+C_2)$

$t_3 = C_2R_2$

$t_4 = \frac{R_1C_3}{4}$

$t_5 = R_3C_4$

$f_{POLE} = \frac{1}{2\pi t_4}$

$f_{mod} = \frac{1}{2\pi t_5}$

$f_o$ = op amp low frequency pole

A = op amp dc gain

$t_1 = R_1C_1$

$t_2 = R_2C_1$

$f_n = \frac{1}{2\pi}\sqrt{\frac{K_vK_p}{Nt_1}}$

$\xi = \pi f n t_2$

*Fig. 2. Loop filter options in outline. Real implementations always have additional poles.*

## Fractional-$N$ in practice

Figure 1 provides some insight into how the fractional-$N$ technique works.

If the loop divider is changed at the end of each reference cycle, ie when the $N$ counter overflows and reloads from its latch, the average value of $N$ taken over many reference cycles may be made a non-integer value.

Frequency of the vco is still be given by $f_{vco}=N\times f_{ref}$, but now $N$ is no longer integer, giving us fractional frequency resolution.

As an example, if $N$ is 100, and $f_{ref}$ is 1kHz, output frequency will be 100kHz. If, every 10 reference cycles, $N$ is changed to 101 for 1 reference cycle, the long term average of the loop divider will be (9×100+101)/10=100.1 and output frequency will be 100.1kHz.

You may already have noticed the weakness of this idea. Assuming that a very-low loop natural frequency is used, the loop can be considered as if it were open. The vco is considered to be on exactly the correct frequency, ie 100.1kHz.

Now, after the first reference cycle we will

Phase lock loop comparison frequency $f_{ref} := 10^5$. Integer part of loop divider $N := 10$

Fractional part of loop divider $M := 7$. Accumulator modulus $Modulo := 2^8$.

Correction DAC modulus $DAC := 2^8$

$i := 0..(Modulo \cdot 8 - 1)$ $\quad T_{error_0} := 0$ $\quad A_0 := 0$ $\quad N_{D_0} := N + 1$

$f_{VCO} := f_{ref}\left(N + \frac{M}{Modulo}\right)$ $\quad T_{ref} := \frac{1}{f_{ref}}$ $\quad f_{VCO} = 1002734.375$

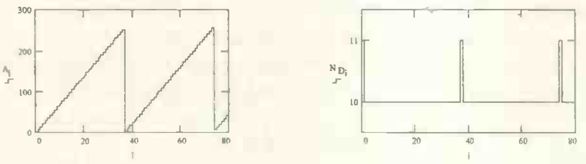
The next statement forms the vector of values in the accumulator. $\quad A_{i+1} := \mathrm{mod}(A_i + M, Modulo)$

Special vector function definition that returns 1 if next entry in vector is less than current entry. Serves to mark accumulator overflows. If index is <0, returns 0. $\quad OVFL(V,j) := \mathrm{if}(j<0, 0, \mathrm{if}(V_{j+1}<V_j, 1, 0))$

Logic section to create vector of divisors based on accumulator overflow. Order of these statements is important

$N_{D_{i+1}} := N$ $\quad N_{D_{i+1}} := \mathrm{if}(OVFL(A,i)=1, N_{D_{i+1}}+1, N_{D_{i+1}})$ $\quad \mathrm{mean}(N_D) = 10.027818$

Graphs of accumulator content and loop divisor



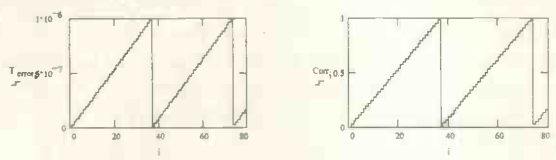

Create vector of time errors due to non exact division. $\quad T_{error_{i+1}} := T_{error_i} + T_{ref} - \frac{N_{D_{i+1}}}{f_{VCO}}$

Create correction voltage from accumulator A. Accumulator is truncated to no of DAC bits $\quad Corr_i := \frac{1}{DAC} \cdot \mathrm{floor}\left(\frac{A_i}{Modulo} \cdot DAC\right) \cdot \left(\frac{f_{ref}}{f_{VCO}}\right) \cdot 2 \cdot \pi \cdot 1.59$

Create error voltage vector at output of phase detector. Phase detector constant assumed equal to 1.59. Note that correction voltage is assumed to be accurate to 5% $\quad V_{error_i} := 2 \cdot \pi \cdot \frac{T_{error_i}}{T_{ref}} \cdot 1.59 - 0.95 \cdot Corr_i$

Remove DC component from error voltage $\quad MV := \mathrm{mean}(V_{error})$ $\quad V_{error} := \overrightarrow{(V_{error} - MV)}$

Graphs of time error, correction, and voltage error

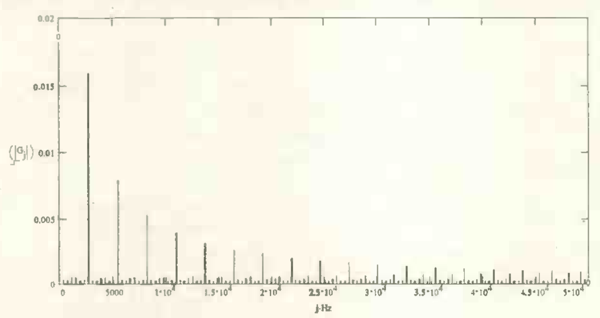



Fourier transform of error voltage. Bin spacing = f.REF/no of samples

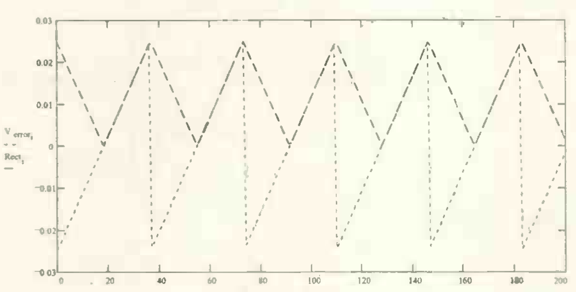
$G := 2 \cdot \mathrm{FFT}(V_{error})$ $\quad \mathrm{length}(V_{error}) = 2.048 \cdot 10^3$ $\quad \mathrm{length}(G) = 1.025 \cdot 10^3$ $\quad Hz := \frac{f_{ref}}{\mathrm{length}(V_{error})}$

$j := 0..\mathrm{length}(G) - 1$ $\quad \mathrm{WRITEPRN}(freq1) := G$ $\quad \mathrm{WRITE}(freqinc1) := Hz$

spectrum of error voltage(y axis in volts eqiv to peak voltage of a transformed sine wave)




Automatic adjustment of error voltage correction by synchronous rectification of V.error $\quad Rect_i := \mathrm{if}\left(\mathrm{floor}\left(\frac{2 \cdot A_i}{Modulo}\right), V_{error_i}, -V_{error_i}\right)$




$\mathrm{mean}(Rect) = 0.01245$

loop fixed parameters:

op amp low frequency pole and open loop gain: $f_0 := 10$ $\quad A_0 := 100000$

loop filter component values: $R1 := 4700$ $\quad R2 := 10000$ $\quad C1 := 10 \cdot 10^{-9}$ $\quad C2 := .5 \cdot 10^{-9}$

Phase detector gain constant ( volts per radian), $Kp := 1.59$

VCO gain constant (radians per volt second), $Kv := 50 \cdot 10^3$

VCO modulation pole(Hz), $f_{mod} := 30 \cdot 10^3$

Loop division ratio, $N := 10$

Additional low pass pole(Hz), $f_{pole} := 20000$

Frequency sweep:

$i := 1..num$ $\quad num := 100$ $\quad f_{start} := 100$ $\quad f_{stop} := 1 \cdot 10^6$

$f_i := f_{start} \cdot \left(\frac{f_{stop}}{f_{start}}\right)^{\frac{i}{num}}$

calculated values:

$t_0 := \frac{1}{2 \cdot \pi \cdot f_0}$ $\quad t_1 := C1 \cdot R1$ $\quad t_2 := R2 \cdot (C1 + C2)$ $\quad t_3 := C2 \cdot R2$ $\quad t_{mod} := \frac{1}{2 \cdot \pi \cdot f_{mod}}$ $\quad t_{pole} := \frac{1}{2 \cdot \pi \cdot f_{pole}}$

$s_i := 2 \cdot \pi \cdot j \cdot f_i$

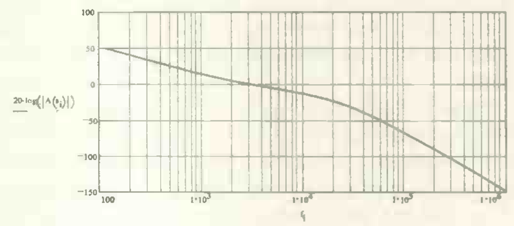
$$F(s) := \frac{A_0 \cdot (1 + s \cdot t_2)}{s^3 \cdot t_0 \cdot t_1 \cdot t_3 + s^2 \cdot (A_0 \cdot t_1 \cdot t_3 + t_1 \cdot t_3 + t_0 \cdot t_1 + t_0 \cdot t_2) + s \cdot (A_0 \cdot t_1 + t_1 + t_2 + t_0) + 1} \cdot \left(\frac{1}{1 + s \cdot t_{mod}} \cdot \frac{1}{1 + s \cdot t_{pole}}\right)$$

$G(s) := \frac{Kp \cdot Kv \cdot F(s)}{s}$ $\quad H := \frac{1}{N}$ $\quad t_{mod} = 5.30516 \cdot 10^{-6}$ $\quad t_1 = 4.7 \cdot 10^{-5}$

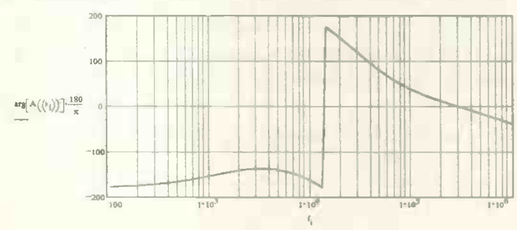
$A(s) := G(s) \cdot H$ $\quad B(s) := \frac{G(s)}{1 + G(s) \cdot H}$ $\quad t_{pole} = 7.95775 \cdot 10^{-6}$ $\quad t_0 = 0.01592$ $\quad t_2 = 1.05 \cdot 10^{-4}$ $\quad t_3 = 5 \cdot 10^{-6}$

Performance graphs:

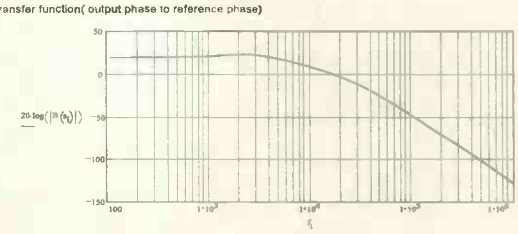
Modulus of closed loop gain



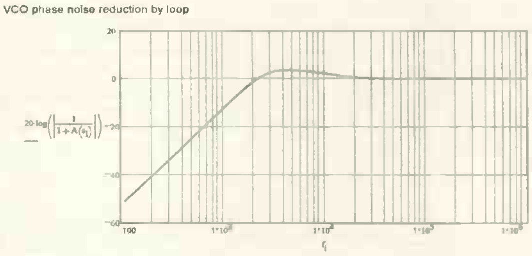

Phase of closed loop gain



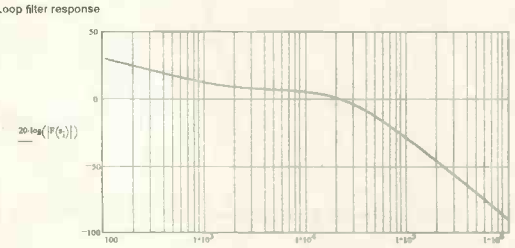

Transfer function( output phase to reference phase)




VCO phase noise reduction by loop




Loop filter response




***Left, simulation of the fractional-N loop with analogue correction. Right, a simple way to display the natural loop frequency. Component values in Fig. 2 for the fifth loop are entered as well as values for the loop divider (N), phase detector constant ($K_p$), and vco gain constant ($K_v$). The MathCad document shows plotted graphs of closed loop gain and phase, forward gain, reduction of vco phase noise and loop filter response.***

have divided by 100. The time required for the counter to overflow, and deliver the output pulse to the phase detector, will be 100×1/100100 or 999μs. The reference pulse or edge has a period of 1000μs. As a result, the first reference cycle has resulted in a time error of 1μs at the phase detector, the divider output appearing earlier. This corresponds to a phase error of 6.28mrad.

The next reference cycle produces another incremental error of 1μs, giving a total error of 2μs. Errors build up until the tenth reference cycle, at which point a divide by 101 is carried out. Incremental error due to this division will be –9μs, the divider pulse appearing later.

The total error is now zero. The error waveform is a sawtooth, with a peak-to-peak amplitude of just less than $1/f_{vco}$. In the example above, the range was from 0 to 9μs. In a real loop this would produce a net input to the loop integrator, and the phase of the vco would shift to ensure that the error waveform had no dc component.

The error waveform is a sawtooth with a period of 100ms – 10 reference cycles – or a fundamental frequency of 100Hz. This is equal to the gain in resolution of the loop. In order to filter out the error voltage, which would otherwise phase modulate the vco, it is necessary to reduce the loop bandwidth. As a result little is gained over using a loop with a reference frequency of 100Hz in the first place.

Both cases are not quite the same, as in the fractional loop phase comparisons are still at a 1kHz rate. Also, the error voltage to be filtered out does not depend on the characteristics of the phase detector, as it does with reference frequency feed through.

Amplitude of the time error waveform is equal to the reciprocal of the vco frequency, and is proportional to $1/N$. However, the loop will multiply the error voltage by its forward transfer function, which at low frequencies is equal to $N$. As a result the vco spurious sideband level will tend to be independent of $N$.

## Cancelling error voltage

In order to make the fractional loop work properly, initially, some means of controlling the loop divider is needed to generate any required frequency increment. Secondly some means of cancelling out the error voltage in the loop is necessary.

Control of the divisor can be arranged by using an accumulator, with a phase increment added into it at the end of each reference cycle. When the accumulator overflows, the loop divisor is changed from $N$ to $N$+1. Either a decimal or a binary modulus may be used. The value of the phase increment is called the fractional divider, or $M$. Output frequency is now $(N+M/modulus)\times f_{ref}$.

This relationship can be verified by trying a few examples. If $M$ is set to 1 less than the modulus, the accumulator will overflow on every reference cycle except 1. The loop will divide by $N$+1 for (*modulus*–1) cycles and then by $N$ for 1 cycle.

Another situation that is of interest is when $M$ is not a factor of the modulus. For example $M$=7 and modulus is equal to 100. Now the accumulator will overflow on the fifteenth reference cycle, but will not contain zero. The time error waveform will be the sum of two sawtooths, one with a period of about fifteen reference cycles, and the other with a period of 100 reference cycles. The lowest frequency component will be $f_{ref}/modulus$. If $M$ is a factor of *modulus*, the lowest frequency will be $f_{ref}\times M/modulus$.

Cancellation of the error voltage from the phase detector can be achieved easily since the contents of the accumulator can be considered as a binary – or decimal – integer exactly following the error waveform. Parallel output of the accumulator is fed to a d-to-a converter, its voltage is scaled appropriately, and it is added to the output from the phase detector. This exactly cancels the error voltage.

Now all the elements for the fractional-$N$ synthesiser exist. **Figure 3** gives a block diagram of one type of practical implementation based on discrete logic. Other implementations will be considered later.

## Simulating analogue correction

Rather than analyse this circuit in detail in the text, it is easier to introduce the first of the two *MathCad* documents shown. This is a simulation of a fractional-$N$ loop with analogue

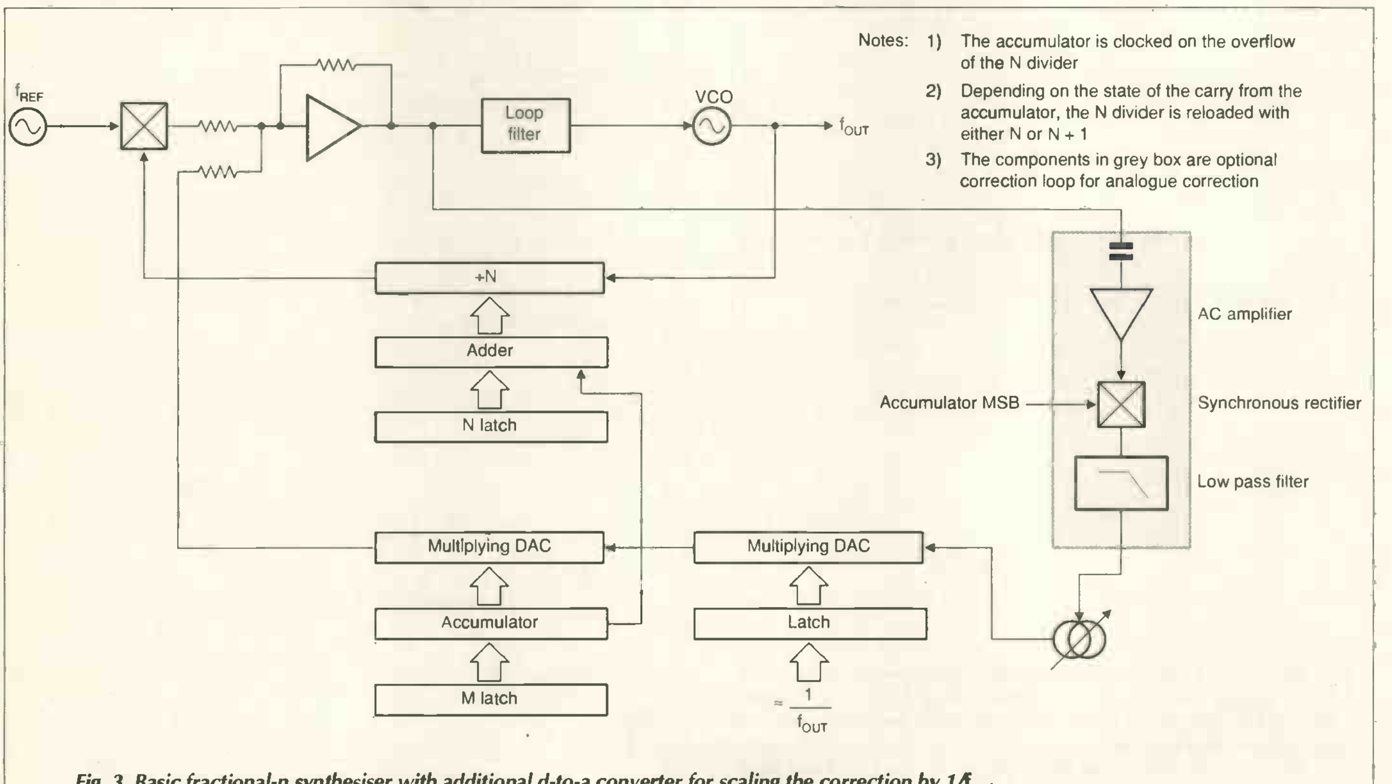



*Fig. 3. Basic fractional-n synthesiser with additional d-to-a converter for scaling the correction by $1/f_{out}$.*

correction. It should be fairly easy to understand to users of *MathCad*.

It is not possible to simulate a program loop with *MathCad*. Instead, the simulation must be made in a linear manner, proceeding from the start to the end of the document. The length of the simulation is set by the range variable '$i$', which must be a power of 2 so that the Fourier transform will work. The accumulator modulus is set to 256, and the length of the simulation to eight complete cycles (2048).

Various vectors are generated, such as the vector of the accumulator contents and the vector of the loop divider. Most of these are illustrated by graphs. The effect of using a d-to-a converter to generate the correction voltage with a modulus less than that of the accumulator may be simulated. This is quite likely to be the case in a practical implementation of the synthesiser, when the accumulator modulus may be much greater than 256.

The addition of the correction voltage to the phase detector output voltage is deliberately made in error by 5%. This is in order to make the rest of the document more interesting. Those of you with *MathCad* can verify that the correction does in fact exactly cancel the error by changing the factor 0.95 to 1.

The large graph is the Fourier transform of the error voltage as it is at the input to the loop filter. This graph does not directly predict the spectrum of spurious sidebands of the vco, as the error voltage will be modified by the loop forward transfer response.

The fractional frequency generated in this simulation is 2734.375Hz. The major component of the error voltage is at this frequency and its harmonics. Smaller components are visible at spacing of 390.625Hz which is $f_{ref}/256$.

As the adjustment of the correction voltage is very important for the elimination of spurious sidebands, it would be useful if there was a method of automatically balancing the error voltage to zero. This is in theory possible if we consider that the polarity of the error voltage will change if the correction is changed from too large to too small. As it stands the error voltage will always be balanced about 0V to give the requirement that the input to the loop integrator has a mean of zero.

However if you synchronously rectify the ac error voltage using the most-significant bit of the accumulator, and then filter the resulting waveform, you get a dc value proportional to the unbalance of the correction. This may be used in a long time-constant feedback loop to correct the magnitude of the correction voltage. This idea is shown in Fig. 3, and is simulated in the last part of the document.

As I mentioned earlier, the magnitude of the required correction must be scaled by $1/f_{out}$ in order to keep the adjustment of the correction correct. This can be done by means of an additional multiplying d-to-a converter, Fig. 3.

This concludes the discussion of the fractional-$N$ loop operation. In a further article I will examine an important modification of the single accumulator fractional-$N$ loop which goes a long way to reducing spurious outputs – even without analogue correction. ■

### References


1. Ulrich L., 'Digital PLL Frequency Synthesisers', Rhode, Prentice Hall.
2. William F. Edan, 'Frequency Synthesis by Phase Lock', Wiley.
3. Vadim Manassewitsch, 'Frequency Synthesisers, Theory and Design', Wiley.
4. Alain Blanchard, 'Phase Locked Loops. Application to coherent receiver design'.

# Designing cascade RC oscillators

**Cascade *RC* oscillators are simple, stable, and offer very fast start up. David Griffiths explains how this long-established circuit configuration benefits from modern op-amp technology.**

The new breed of dual op-amps with rail-to-rail output swing allows a very simple implementation of the *RC* cascade oscillator.

In addition to the timing elements shown in **Fig. 1**, the cascade oscillator needs little more than three resistors and a single eight-pin IC, provided a stable supply voltage is available. Sadly, distortion performance is only around the 2-3% level – mainly third harmonic. However, those concerned with sensor instrumentation at fixed frequencies should welcome the following:

- Excellent stability of output amplitude, which is also highly predictable and has a very low temperature coefficient.
- Very stable timing relation, with ageing and temperature, between sine and square wave outputs, which is a boon in phase sensitive detection schemes.
- Almost instant start and settling characteristic, which allows burst operation of sensors to minimise battery power consumption.

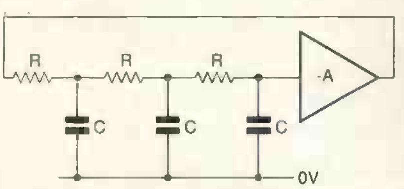



*Fig. 1. In principle, the RC phase-retard cascade oscillator is similar to its CR phase-advancing counterpart.*

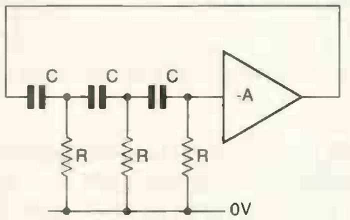



*Fig. 2. The common CR-network based cascade oscillator relies on phase advance.*

## The *RC* cascade alternative

The above principles were expounded in *Wireless World* by L Nelson-Jones[1]. He concentrated not on the common high-pass *CR* cascade, but on the less well known low-pass *RC* scheme. As a re-cap, Nelson-Jones' introduction went along the following lines.

**Figure 2** shows the more common cascade *CR* phase advance oscillator. If the inverting gain of the amplifier exactly equals the attenuation of the cascade when it is giving 180° phase shift, then oscillations at that frequency will be sustained by this arrangement.

This still leaves the awkward problem of precisely controlling the gain to maintain stable oscillation amplitude. This could entail thermistor control, but then ambient temperature changes have a big effect on the amplitude, and it is difficult to avoid prolonged amplitude ringing at switch-on.

In my opinion, a Wien bridge oscillator with thermistor gain control beats this approach hands-down – especially if the amplification is arranged as virtual earth amplifiers giving ultra-low distortion as in the elegant scheme due to John Linsley Hood[2].

Alternatively, gain control in Fig. 2 can be exercised with oscillation amplitude sensing or rectification. The ensuing voltage is then applied to vary the channel resistance of a fet used in the feedback path controlling amplifier gain. In my experience, this approach always seems to give more amplitude temperature coefficient than initially expected, as well as increasing distortion.

The necessary 180° phase shift in the cascade can be equally well achieved via phase retardation generated by interchanging the positions of *R* and *C* to give Fig. 1. Again, the same less than ideal schemes could be used to try to control the gain at the critical value to maintain oscillation amplitude stability. However, Nelson-Jones' insight was that, if you primarily want amplitude stability and can bear some distortion, then it is much better to run the maintaining amplifier as near as possible as an ideal limiter.

## Cascade transfer characteristics

Assuming that the maintaining amplifier is not clipping and has infinite input impedance combined with zero output impedance, the analysis is as shown in the equations panel. This gives mesh equations for both types of cascade.

In the phase retard case, if the cascade is driven at angular frequency $\omega$ at amplitude $V_1$, then the output amplitude $V_2$ is related by,

$$V_1 = V_2.\{1-5\alpha^2 - j(\alpha^3 - 6\alpha)\} \qquad (1)$$

where $\alpha=\omega RC$ and $j=\sqrt{-1}$. When $\alpha=\sqrt{6}$, the imaginary term is zero and $V_1=-29\times V_2$, showing the phase shift is 180°. Angular oscillation frequency is therefore $\omega=\alpha/RC$. That is, the phase retard *RC* cascade oscillator will ideally oscillate at a frequency '*f*' given by,

$$f = \frac{\sqrt{6}}{2\pi RC} \qquad (2)$$

As an aside, in the case of the phase advance cascade, the expression for the frequency at which the imaginary term is zero has the $\sqrt{6}$ on the *bottom* line. Despite this six-fold difference in frequency for the two configurations using the same *R* and *C* values, attenuation at the 180° phase shift condition is again –29, as expected from symmetry considerations.

Substituting 2α, 3α, 4α and 5α into equation (1) allows the attenuation to be calculated for the second, third, fourth and fifth harmonics of the oscillation frequency. This yields attenuation factors of 148.1, 443.6, 1004 and 1916 respectively, and shows a misprint for the third harmonic figure in the original article. This gives an attenuation of 5.1, 15.3, 34.6 and 66.1 respectively compared to the attenuation of the fundamental at which oscillation can occur. These figures agree with those calculated by Nelson-Jones.

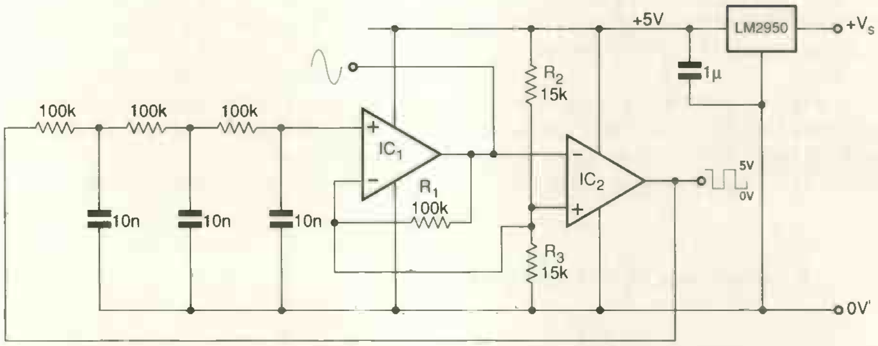



*Fig. 3. With modern rail-to-rail-swing op-amps, this amplitude-stabilised RC oscillator achieves a sine-wave output of around 3.2V pk-pk from a 5V supply. Square-wave output is in antiphase relative to the sine wave.*

## The ideal case

It is now appropriate to consider the ideal case where the inverting amplifier of Fig. 1 is given infinite gain and behaves as an ideal limiter with its output voltage only at one or other of two levels. The phase-retard *RC* cascade is as a result driven with a square wave and the question arises as to the waveform that emerges from it. Using the harmonic attenuation factors calculated above and expressing the driving waveform as a Fourier series of harmonically related sinusoids, the output waveform can be calculated as follows.

Orthogonality of sines and cosines when integrated over a complete cycle means that a square wave of amplitude 'A' can be represented by the Fourier series,

$$\frac{4}{\pi}.A.\left[\sin\theta+\frac{1}{3}.\sin3\theta+\frac{1}{5}.\sin5\theta+...\right]$$

This means that if the amplifier in Fig. 1 were ideally limiting between 0V and +5V supply voltage, then the amplitude of the fundamental harmonic contained 'in' the square wave is $4\times5/\pi=6.37$V. The fact that it should be greater than the swing permitted by the available power-supply voltage is intriguing.

An unforeseen time constant in an amplifying chain can knock the corners off fast rising edges. Knowing this, you might imagine that the corners just get progressively rounded off as the limited waveform progresses down the cascade. However, it does not happen like this because the repetition rate of the square wave is fast compared to the cascade time constants.

At the first capacitor down the chain you can see only a spiky sequence of exponential rises and falls, without any 'flat bits' left from the square wave. The waveform on the second capacitor is remarkably close to being triangular. As a result, it is initially surprising that an oscilloscope shows a presentable sine wave at the end of the cascade.

The Fourier series shows that in the ideal case the third harmonic is the lowest distortion component that would be present and would have a third of the magnitude of the fundamental harmonic. We have already established that this third harmonic is subject to 15.3 times more attenuation in the cascade than the fundamental. As a result the third harmonic at the output only contributes 1 part in 45.9 to the signal, ie a distortion level of 2.18%. Calculation gives a fifth harmonic level of 0.3%, with relatively negligible amounts of higher harmonics.

## Implementating the design

Because of the attenuation in the cascade and the need for it to be lightly loaded, a buffer amplifier is needed as well as the limiter. These could both be driven in parallel by the

### RC Cascade oscillator performance

Measurements on a number of units oscillating at about 400Hz and using the *LMC660* and *AD822* op-amps have shown an ac amplitude temperature coefficient around the 10-20ppm/°C level. This occurred over excursions of 25°C above and below ambient temperature, in addition to that imposed by the tempco of the supply voltage. As a result, for many measurement situations, there is no need to use ratiometric techniques to compensate for drifts in the carrier amplitude. This allows you to base an adequately stable calibration on simply measuring the output voltage from a sensor.

As might be expected from the toggled nature of the driving waveform and passive shaping, the circuit start-up and settling are very rapid indeed. At switch-on, the uncharged timing capacitors hold the output of $IC_1$ low, ensuring that the limiter output goes high.

After a delay of about one and a half oscillation periods, output of $IC_1$ is high enough to toggle the limiter and the oscillation cycle starts. Observations on an analogue oscilloscope suggest that after two further cycles the amplitude has settled to within 1% of its final value. This gives a total start-up time of around three cycles.

At frequencies up to about 1kHz, this circuit shows remarkably good constancy of phase relation between the sine and square wave outputs without having to use a comparator for $IC_2$. As a result a space saving dual op-amp can be used.

An *AD822AN* based oscillator, again running at about 400Hz, was carefully examined in this respect. It showed less than 0.2° shift change between the sine and square waves for a 30°C rise in circuit temperature above ambient. Similarly good long-term stability can be expected, making this oscillator useful for applications with phase sensitive detection.

Since expected distortion is not good, examination of this aspect has been limited to using a passive Tee to filter out the fundamental and check that amplitude of the residue is consistent with the theoretical predictions.

cascade output, but it seems better to drive the limiter from the buffer amplifier. This increases the overdrive to speed up the transitions, **Fig. 3**.

A single supply voltage rather than dual is shown because positive line regulators or references tend to have better constancy of output than their negative line counterparts. This is important in maintaining the ac amplitude stability which is a prime feature of this oscillator.

Op-amp $IC_2$ performs the limiting function and must have an output which can swing to within a few millivolts of its supply voltages. Resistors $R_{2,3}$ are of equal value and as a result fairly closely establish the mid-point voltage between these output limits. If these resistors are 1% tolerance, then the tolerance on the mid-point voltage is also 1% – as the resistors are equi-valued. This is more than good enough in view of other imperfections in the circuit.

My first reaction was to decouple the mid-point of $R_{2,3}$ to 0V with a large capacitor to provide a low impedance reference voltage. But the resistance of $R_1$ in parallel with $R_2$ can be used to define the appropriate gain needed in the buffer amplifier – as well as generating the correct reference voltage for the limiter. This also ensures dc coupling round the oscillator loop, enabling the circuit to be used down to low frequencies limited only by practical values for timing resistors and capacitors.

As we have already calculated that the amplitude of the fundamental component driving the cascade is 6.37V, and know that it is attenuated by a factor of 29 in the network, the buffer amplifier is presented with a signal of $0.22V_{pk\text{-}pk}$. With $R$ at 100kΩ and $R_1$ and $R_2$ at 15kΩ, gain of the $IC_1$ stage is 14.3, giving a sinewave output of $3.2V_{pk\text{-}pk}$.

Because of the gain of the $IC_1$ stage and the comparatively high resistor values used, care needs to be taken to reduce unintentional capacitive coupling from the square wave to the input of $IC_1$. If this is not done, the sinewave output will be adorned with spikes where the limiter toggles. A sensible layout, with a grounded pcb trace to guard this fed back voltage, is all that is needed.

If output from the cascade is $0.22V_{pk\text{-}pk}$, then the junction of $R_2$ and $R_3$ must be going up and down by the same amount. This would seem at first sight to be a hopeless reference voltage for the limiter. However, when output of the cascade is in mid-excursion it must ideally be at +2.5V, since it is fed by a 1:1 squarewave of amplitude 5V. Under these conditions, no current will flow in $R_1$ since the junction of $R_2$ and $R_3$ is already at +2.5V by virtue of the current flowing through these resistors from the supply line. This is exactly the correct condition to toggle the limiter.

## Op-amp criteria

Since load on $IC_2$ is greater than 100kΩ, it is likely that any op-amp with rail-to-rail output capability will limit its output within some 10mV of 0V and +5V under these conditions. As a result, it would be satisfactory in this respect. However, $IC_2$ also needs to come out of saturation quickly from either limit and to slew rapidly in either direction at the same rate.

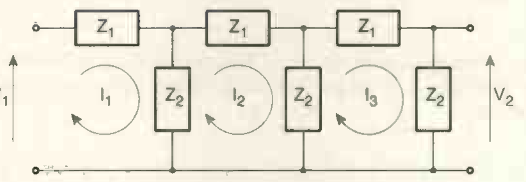
The new (dual) *AD822AN* seems a good choice in these respects, but the older (quad) *LMC660C* has worked well in a production run. As for the +5V regulator, there is much to be said for the *LP2950ACZ*-5.0. Housed in a TO-92 package, this inexpensive low-drop out type offers 20ppm/°C typical temperature coefficient. It is however essential to remember to decouple its output as detailed in the data sheet.

## Considerations for higher frequencies

With performance at frequencies where the propagation delay in $IC_2$ becomes significant compared to the cycle time, it should be noted that this delay is equivalent to a phase lag. As a result the cascade does not need to produce a full 180° phase shift to satisfy the Barkhausen criterion for oscillation that there be 360° phase change round the complete feedback loop. This means oscillations will always occur at frequencies somewhat less than that predicted by equation (2).

Accordingly, attenuation of the fundamental through the cascade will necessarily be somewhat less than the value of 29 implied by

## Cascade oscillator equations




The generalised cascade gives the following mesh equations:

$$V_1 = I_1(Z_1+Z_2) - I_2Z_2 \qquad (3)$$

$$0 = -I_1Z_2 + I_2(Z_1+2Z_2) - I_3Z_2 \qquad (4)$$

$$0 = I_2Z_2 - I_3(Z_1+2Z_2) \qquad (5)$$

Adding equ.(3) x $Z_2$ to equ.(4) x $(Z_1+Z_2)$ eliminates $I_1$, yielding,

$$V_1Z_2 = I_2\{(Z_1+2Z_2).(Z_1+Z_2) - Z_2{}^2\} - I_3.Z_2(Z_1+Z_2) \qquad (6)$$

Substituting $I_2$ from equ.(5) into equ.(6), and putting,

$$I_3 = \frac{V_2}{Z_2}$$

eventually yields :

$$V_1 = V_2\left\{\left(\frac{Z_1}{Z_2}+5\right)\left(\frac{Z_1}{Z_2}\right)^2 + 6\frac{Z_1}{Z_2} + 1\right\} \qquad (7)$$

In the CR phase advance cascade, Fig. 1, $Z_1$ is given by $C$ at angular frequency ω and $Z_2$ by $R$, yielding,

$$V_1 = V_2\left\{1 - \frac{5}{\omega^2C^2R^2} + \frac{j}{\omega CR}\left(\frac{1}{\omega^2C^2R^2} - 6\right)\right\} \qquad (8)$$

The imaginary term is zero when $\omega = \dfrac{1}{RC\sqrt{6}}$, giving $V_1 = -29V_2$.

In the RC phase retard cascade of Fig. 1,

$$V_1 = V_2\{1 - 5\omega^2R^2C^2 - j\omega RC(\omega^2R^2C^2 - 6)\} \qquad (9)$$

Writing $\alpha = \omega RC$, the phase shift θ given by,

$$\tan^{-1}\theta = -\left(\frac{\alpha^3 - 6\alpha}{1 - 5\alpha^2}\right)$$

indicating 180° phase shift when,

$$\omega = \frac{\sqrt{6}}{RC}.$$

equation (1) and give more output swing from the buffer amplifier than might otherwise be expected.

With an *AD822AN* at frequencies up to about 500Hz, discrepancies from the ideal due to circuit delays are not large. At 400Hz the sinewave output amplitude is about 5% larger than given by simple theory with about a 5° shift between the sine and square waves. However, as noted earlier, the temperature coefficient of the amplitude and phase shift is still admirably low.

With timing element *R* at 30kΩ and *C* at 1nF, oscillation occurs at about 10kHz. However, the phase shift between sine and square waves is now such that in a phase sensitive detection scheme you would have to use a separate comparator to generate a good switching reference from the sinewave.

Time delay around the oscillation loop produces an equivalent phase shift such that oscillation occurs at only 75% of the frequency given by equation (2) and the reduced attenuation through the cascade gives some 75% greater output than at low frequency, requiring that $R_1$ be reduced to about 68kΩ. While these figures are consistent with the (awkward) equations, it is much easier to do a bit of trial and error under these conditions rather than try to calculate the degree of trimming needed.

With this degree of non-ideality, you might expect the amplitude temperature coefficient of the output to be very poor. But while it is worse than at low frequencies it is still acceptable for many applications. Output of a trial unit with an *AD822AN* increased by just 1% for a temperature rise of 30°C above ambient, ie about 0.03%/°C. At 25kHz this only increased to about 0.06%/°C.

Because of the dc coupling in the amplitude maintaining and stabilising loop, operation down to very low frequency is possible. With timing element *R* at 10MΩ and *C* at 2.2μF, the resulting oscillation period of 60s is probably near the practical limit, as electrolytic capacitors can not be used because of their leakage currents.

Start-up and oscillation at this frequency, **Fig. 4**, was captured with a logging dvm and can be seen to be as described for audio frequency operation. Sinewave amplitude and frequency were well within the limits set by the 1% resistor tolerances and the precision of measuring and selecting equi-value timing capacitors. This is no surprise since the op-amps behave close to the ideal under these conditions.

If the oscillator output must be symmetrical about 0V, this is easily achieved by using dual ±5V supplies, albeit with a reduction in the amplitude temperature cofficient, as discussed earlier. Increased swing now required from the limiter takes longer with the slew rate of a given op-amp. This somewhat increases delay around the maintaining loop and, as a result, the errors arising from this mechanism. ■

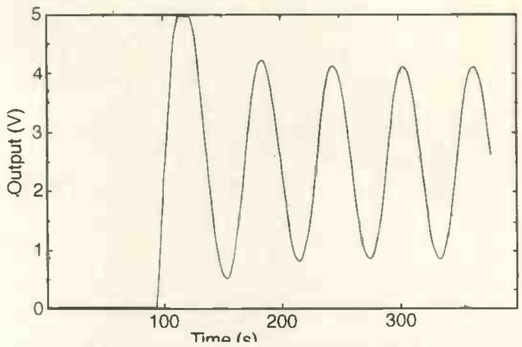



***Fig. 4. Start-up characteristic of a 60s period RC cascade oscillator demonstrates extremely fast stabilisation.***

Acknowledgements to the directors of Chelsea Instruments for their permission to publish this work.

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# Audio power with a new loop

**In Marcel van de Gevel's high-performance power amplifier, quiescent current is stabilised via feedback, using a non-linear common-mode loop.**

In a conventional class (A)B amplifier, quiescent current through the output devices depends on temperature differences between the output devices and a temperature sensor. The temperature sensor is usually mounted on a heat sink.

Due to thermal resistance from the output devices to the heat sink, and thermal capacitance of the heat sink, there are large differences in temperature after a sudden change in power dissipation. Hence, quiescent current of an audio amplifier changes every time the music volume changes. Non-optimal quiescent current causes extra distortion known as transient crossover distortion[1].

Designers of operational amplifiers sometimes use a different technique for class (A)B biasing known as non-linear common-mode loop, or class-AB control loop[2]. However, circuits described in reference 2, and in many other articles, are not suitable for a discrete amplifier. This is because they use IC techniques like emitter-area scaled transistors.

With a different kind of non-linear network, however, it becomes possible to use non-linear common-mode loops in an almost wholly discrete audio power amplifier. With this technique, thermal problems like thermal runaway and transient crossover distortion can be avoided as can quiescent current adjustment.

Further, the technique can be applied to common-emitter or mixed common-emitter/common collector output stages, or their mos equivalents.

## Designing the output stage

A simplified schematic of the output stage is depicted in **Fig. 1**. Voltage source $V_1$ produces about 10V above the output voltage, so that the gate of $Tr_{13}$ can swing above the positive supply rail.

The circuit may seem rather asymmetrical, having a source follower for the positive side of the signal and a common source stage for the negative side. However, both mosfets are current driven. Under current drive, source followers and common source stages behave almost identically, **Fig. 2**.

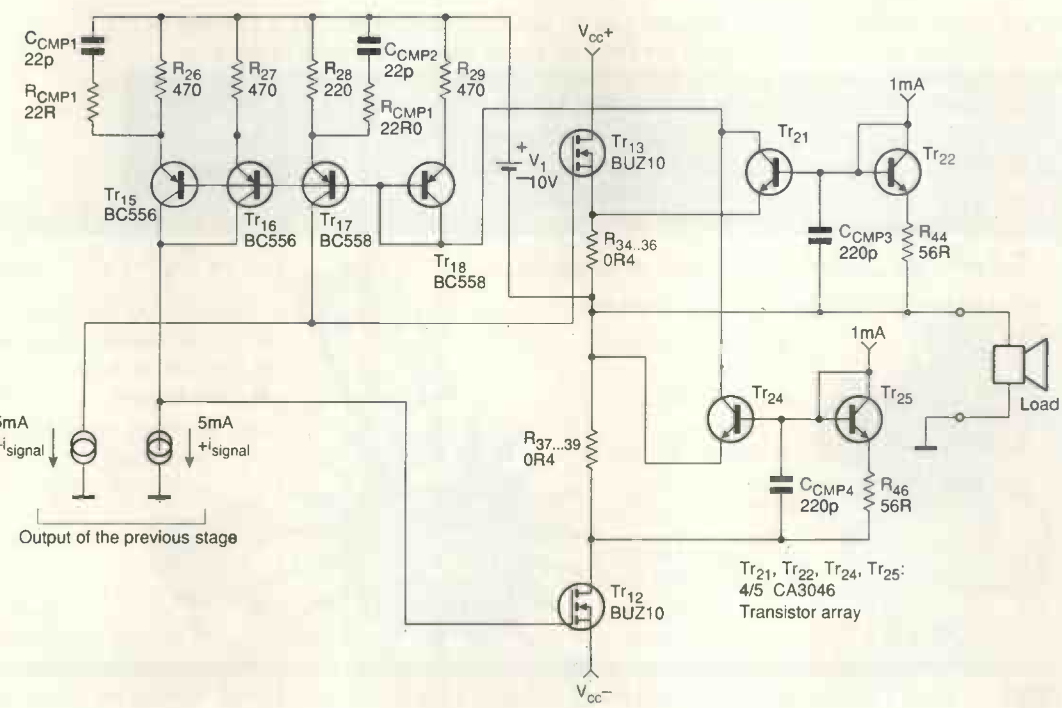


*Fig. 1. Simplified output stage of a power amplifier incorporating a non-linear common-mode loop. The circuit may look asymmetrical, but the source follower and common-source stages behave similarly when current driven.*

The class-AB control loop – a non-linear common-mode loop – consists of a non-linear network around $Tr_{21,22}$, and $Tr_{24,25}$, a current mirror with two outputs $Tr_{15-18}$ and output mosfets $Tr_{12,13}$.

When current through $Tr_{13}$ or $Tr_{12}$ becomes too small, the current through $Tr_{21}$ or $Tr_{24}$ increases respectively. Gates of both mosfets are charged by the current mirror until both currents are again large enough. When the currents through both mosfets become large simultaneously, currents through $Tr_{21}$ and $Tr_{24}$ become small.

Now, the bias currents of the previous stage are larger than the currents through the current mirror and the gates of both mosfets are discharged until the smallest current becomes small enough. When currents through the mosfets are of the same order of magnitude, they both have an influence on the non-linear network and a smooth transition from one output device to the other is realised.

Quiescent current is about 100mA. When the current through one output device is large, current through the other device is still about 45mA, indicating that my amplifier is clearly a non-switching type.

The influence of the class-AB control loop on the difference between the drain currents of $Tr_{12,13}$, and thus on output current, is kept as small as possible by always driving both mosfets equally. When the current through one mosfet is at its minimum, and the normal differential-mode loop tries to discharge its gate further, the common-mode loop responds by increasing the currents through the current mirror.

The gate of the other mosfet now charges quicker than it would have had current through the mirror not increased. In other words, the drive current of one output mosfet is passed on to the other when this is necessary. In this way, sudden changes in gain – similar to the $g_m$-doubling effect – are prevented.

Components $C_{cmp1-4}$ and $R_{cmp1-2}$ improve stability of the non-linear common-mode loop.

## Amplifier circuitry

One channel of the amplifier is shown in **Fig. 3**, and the protection circuit and power supply in **Fig. 4**. I built a stereo amplifier, consisting of two channels with a common protection circuit and power supply. Output mosfets of both channels are mounted with electrical insulation on a 1K/W heat sink.

In the protection circuit, $Tr_{37}$ has its own star-shaped heat sink, 60K/W or less. The other transistors do not require heat sinking. A modified bootstrap network comprising $R_{25}$, $D_{14,15}$ and $C_{10}$ corresponds to $V_1$ in Fig. 1.

In order to keep the current controlled during clipping, recovery from clipping, slewing, switch-on and switch-off, the gates are always discharged when currents through the current mirror are small. For this reason, $R_{12,14}$, $D_8$ and $R_{33}$ and $R_{24}$ have been incorporated into the circuit.

An anti-saturation circuit comprising $Tr_{14}$ and $D_{18,19}$ prevents saturation of the current mirror. Without this circuit, a 5A current spike flows through $Tr_{13}$ and $Tr_{12}$ for a few microseconds during recovery from clipping.

The normal differential-mode feedback loop consists of the differential pair $Tr_{2,3}$, phase splitter $Tr_9$, common base stages $Tr_{10,11}$, output mosfets $Tr_{12,13}$ and the feedback network comprising $R_{4,5}$ and $C_5$.

## Major poles in the loop

Without frequency compensation, the normal feedback loop has three major poles, ie poles that have an important influence on high-frequency behaviour. Circuits can have several major poles, sometimes referred to as dominant poles. However, this can cause confusion as 'dominant pole' is often used to describe the very lowest pole[5].

Capacitances between the bases and emitters of $Tr_{2,3}$ and $Tr_9$ contribute two major poles. In the output stage, the high-frequency transfer is mainly determined by the gate to drain capacitance – the third major pole. This capacitance acts as a Miller capacitor, causing open-loop output impedance to drop to a few ohms. It also makes the high-frequency transfer of the last stage less sensitive to the widely varying transconductance factor of the output mosfets.

In theory, the three major poles could be moved into their desired positions – for example, three equal negative real poles or third-order Butterworth positions – with two compensation networks[6]. Unfortunately, simulations with a pole-zero extraction program and a root locus program show that the influence of all the non-major poles and zeros together is too large. As a result, a less subtle approach is necessary.

Heavy pole-zero compensation in the second stage comprising $L_1$, $R_{13}$ and $C_7$ reduces the number of major poles to two. This is

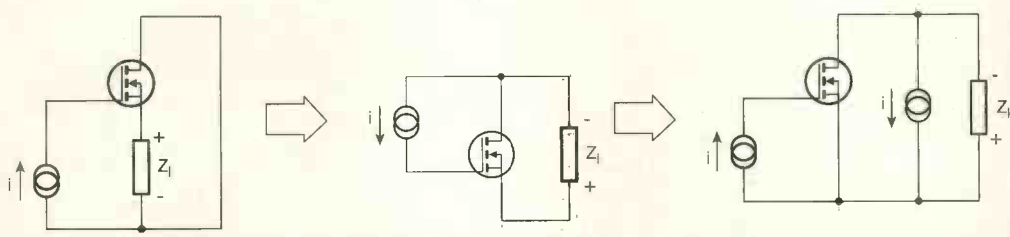


*Fig. 2. By redrawing the circuit and applying a Blakesley transformation, the source follower on the left is shown to be equivalent to the common source stage with the extra current source on the right. As long as the current gain factor of the mosfet is much larger than unity, the influence of the extra current source is negligible.*

### Mosfets versus bipolar transistors

Mosfets are used because of their typically large current-gain factor and good high-frequency behaviour. Readers who have read Self's article on mosfets and bipolar transistors[3] might think it foolish to choose mosfets, so I shall explain why his arguments do not apply in this case.

As Self correctly points out, mosfets have a lower transconductance than bipolar transistors – a major disadvantage in some circuits. However, the transconductance does not matter much when the output devices are current driven, as they are in my circuit. Current gain on the other hand has a direct influence on loop gain, and should be large.

Self also says that the sudden transition from a zero to a quadratic voltage to current transfer gives nasty corners in class-B gain plots. However, whether the voltage-to-current transfer of the output devices starts in a smooth way or not does not matter in my circuit. Output devices do not determine how the current is divided between the two sides of the output stage.

Further, the voltage-to-current transfer of a real mosfet starts in a much smoother way than the voltage to current transfer of a *PSpice* power mosfet. Designers of low-power mos circuitry know that mosfets do not switch abruptly from an 'off-state' to the quadratic, strong inversion, region.

At low gate-source voltages, mosfets are in the so-called weak inversion or sub-threshold region, where gate-source voltage to drain current transfer is exponential[4]. As gate-source voltage rises, the mosfet's transfer gradually changes from exponential to quadratic. The range where the transfer is neither exponential nor quadratic is known as the moderate inversion region.

For example, measurements of a *BUZ71A* power mosfet show that the mosfet is in weak inversion from below 1μA up to about 10mA. The mosfet is in moderate inversion from about ten to a few hundred milliamps. *PSpice* simulation of an *IRF240* mosfet using a *PSpice* library file shows no weak or moderate inversion regions at all, which is physically impossible. Unfortunately, there is no *BUZ71A* model in this library.

achieved by covering the pole of the first stage and lowering the bandwidth. Capacitor $C_5$ generates a zero in the feedback network which pulls the root locus well into the left half plane. In the closed-loop response the two major poles almost end up on the negative real axis.

Output filter $L_2$, $R_{47}$ and $C_{13}$ [7] decreases the influence of strange load impedances and suppresses spurious high-frequency signals picked up by the loudspeaker leads. The input filter formed using $R_{1,2}$ and $C_{1,3,4}$, suppresses high-frequency signals and prevents slew-rate limiting when the amplifier is subjected to unrealistic test signals like square waves. The response of this filter is approximately a second-order Butterworth one, with a cut-off frequency around 140kHz.

Experimentation has shown that the amplifier is stable with an 8Ω load, a loudspeaker load and an unrealistic but often used 8Ω//2μF test load.

The circuit around the *TL071* is a dc bias servo loop, which gives a second-order Butterworth high-pass response with $C_2$ and $R_3$ – at least when output impedance of the preamplifier is zero. Damping of the response increases when output impedance is not zero.

## Protection network

The circuit around $Tr_{19,20}$ is a simple protection network. When output current is greater than about 10A, this thyristor-like structure triggers and turns off the output stage. It remains turned off until the amplifier is switched off for around thirty seconds and then switched back on again.

The collector of $Tr_2$ connects to another protection circuit, **Fig. 4**, which responds if current through $Tr_2$ is less than a quarter or greater than three quarters of the tail current. This will only occur if the amplifier is clipping or if the amplifier has broken down, otherwise feedback keeps the signal levels small.

Transistors $Tr_{27\text{-}32}$ comprise a dual current window comparator – dual, because both channels of a stereo amplifier have one common protection circuit. If current through $Tr_2$ has a too large or small a value for longer than about 20ms, $C_{15,16}$ discharge, the output relay switches off and three-

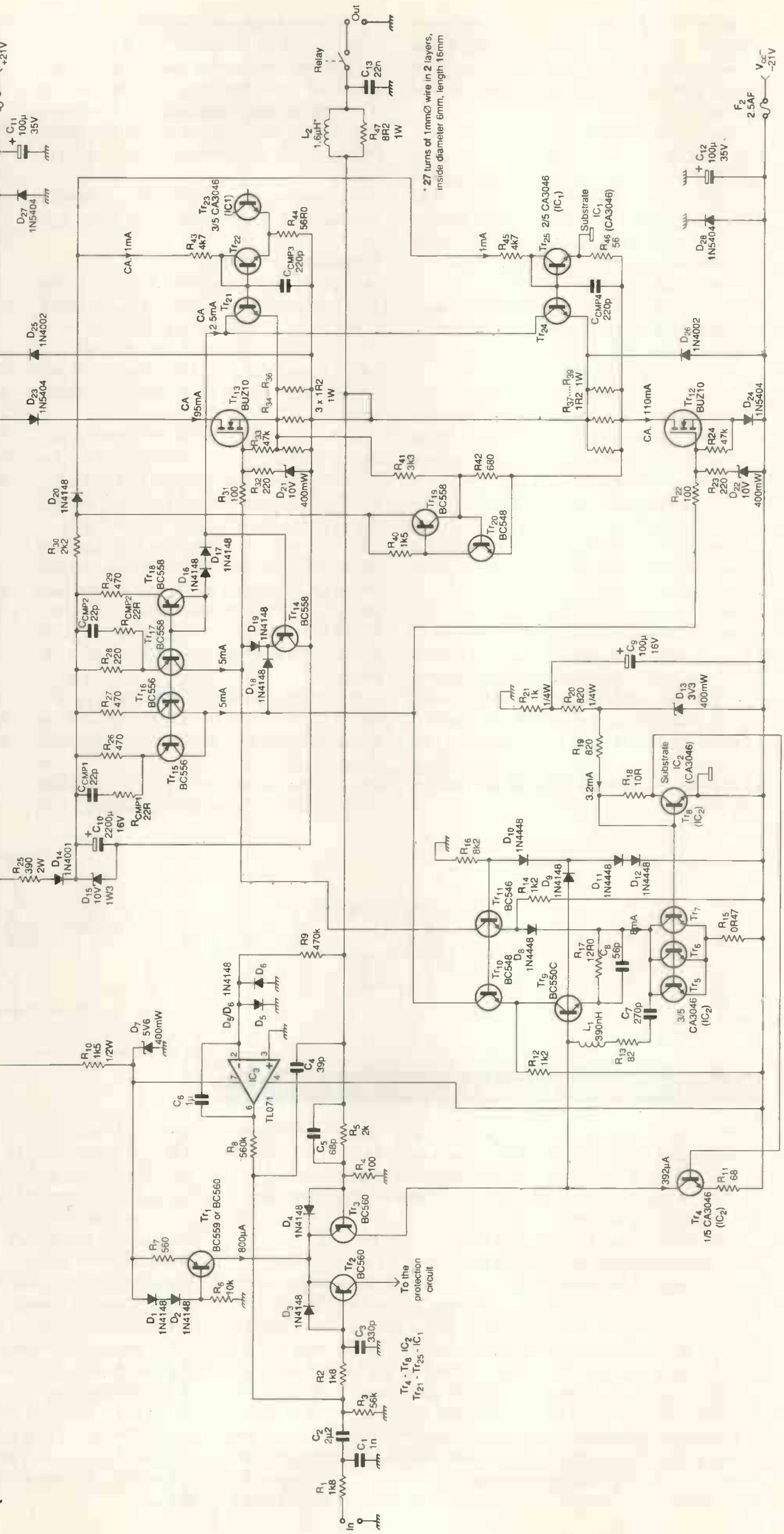

*Fig. 3. Entire schematic of one channel of the non-linear common-mode loop power amplifier.*

colour led $D_{36}$ turns red. This process protects the loudspeakers and provides a warning.

When the amplifier clips, the three-colour led emits yellow/orange light for about a second. If it clips severely for prolonged periods, the output relay turns off and the led turns red. This is done because the harmonics generated by the clipping amplifier are not good for the tweeters. A few seconds after the volume of the signal is reduced, the output relay turns on again and the led turns green.

As usual, protection circuitry prevents switch-on and switch-off plops from reaching the loudspeaker. The time needed to charge $C_{15,16}$ determines the switch-on delay with $R_{68}$ and $C_{14}$ the switch-off delay.

## Measurement results

Total harmonic distortion at 10kHz and 16W into 8Ω – about 80% of maximum power – was measured to be about 0.006%, dropping to 0.0025% at 10W. The –3 dB points of the frequency response were about 1.1Hz and 143kHz.

A simplified Quad/Baxandall-like subtractive test[8,9] showed that the distortion on a real music signal was much smaller than 0.1%. With these methods, the desired signal can be attenuated without attenuating noise and distortion. With about 60dB suppression of the signal, the residue still sounded like music rather than distortion so the distortion level must be well below –60dB or 0.1%. ■

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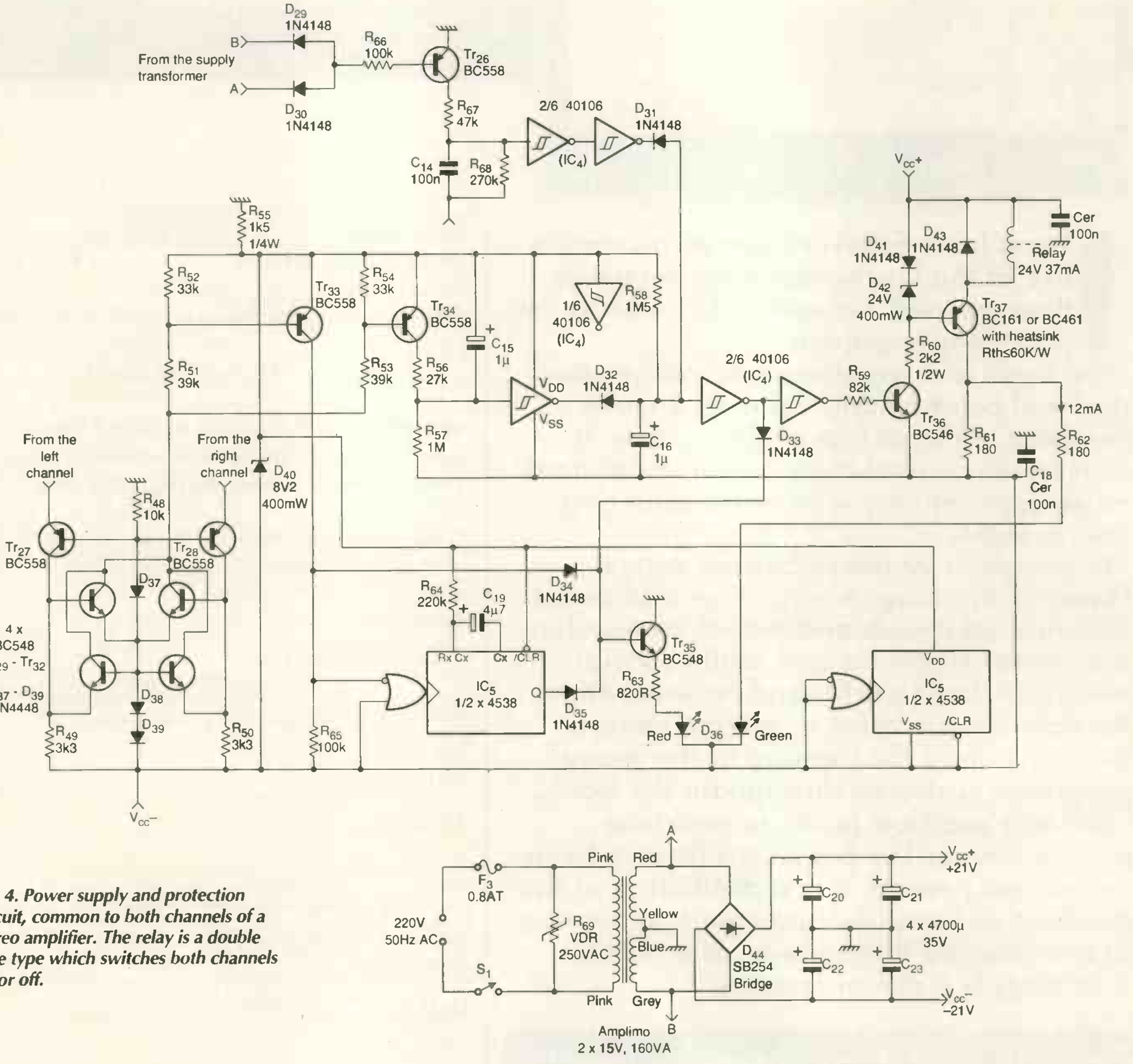
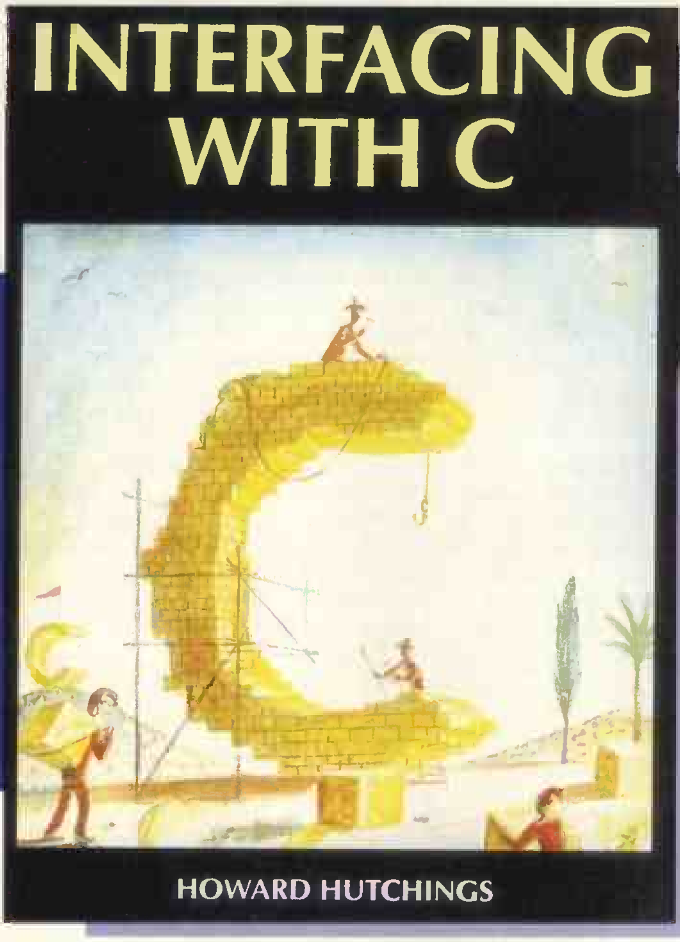



*Fig. 4. Power supply and protection circuit, common to both channels of a stereo amplifier. The relay is a double pole type which switches both channels on or off.*

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Again, almost any dual-gate mosfet will work in place of the *BRF84* as the mixer. Since the receiver used has a narrow-band front end, mixer output needed no tuning and was successful in spite of much 7MHz activity in the evenings. If necessary, either an 8MHz low-pass filter or a 4-6MHz tuned circuit should be suitable.

***Peter Parker (VK1PK)***
*Garran, ACT*
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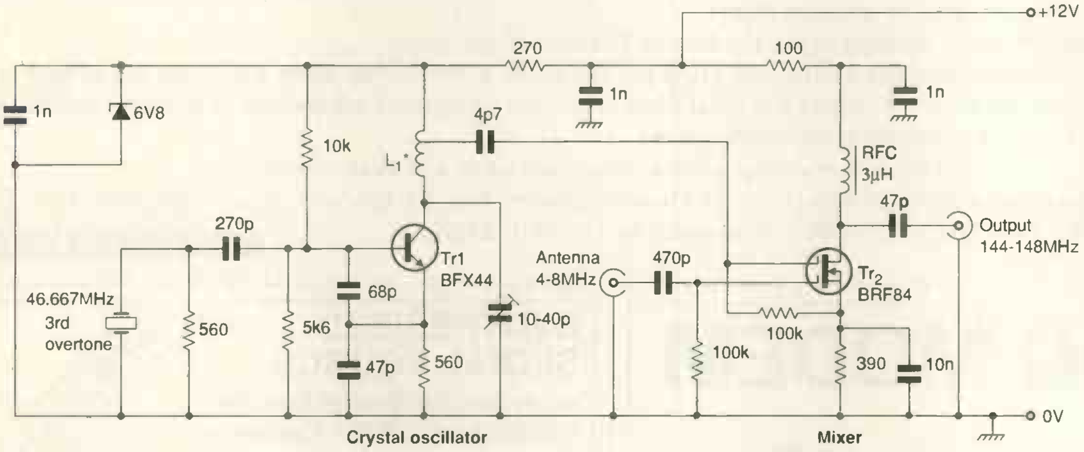


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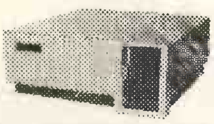

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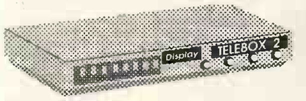

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| | |
|---|---|
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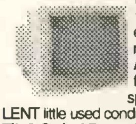

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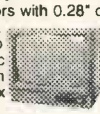

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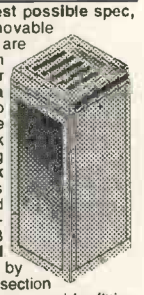

Made by **Eurocraft Enclosures** Ltd to the **highest possible** spec, rack features **all steel construction** with removable side, front and back doors. Front and back doors are **hinged for easy access and all are lockable** with five **secure 5 lever barrel locks.** The front door is constructed of double walled steel with a **'designer style'** smoked acrylic front panel to enable status indicators to be seen through the panel, yet remain unobtrusive. Internally the rack features fully slotted reinforced vertical fixing members to **take the heaviest of 19" rack equipment.** The two movable vertical fixing struts (extras available) are pre punched for standard 'cage nuts'. A **mains distribution panel** internally mounted to the bottom rear, provides 8 x IEC 3 pin Euro sockets and 1 x 13 amp 3 pin **switched utility socket.** Overall ventilation is provided by fully louvered back door and double skinned top section with top and side louvres. The top panel may be removed for fitting of **integral fans** to the sub plate etc. Other features include: fitted castors and floor levelers, prepunched utility panel at lower rear for cable / connector access etc. Supplied in **excellent**, slightly used condition with keys. Colour **Royal blue.** External dimensions mm=1625H x 635D x 603 W. ( 64" H x 25" D x 23¾" W )

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# Flasher for dogs

This circuit uses a single, momentary contact, push button to turn an led flasher on and off.

$IC_{1a}$ and $IC_{1b}$ with $R_1$ form a bistable. Capacitor $C_1$ charges, via $R_2$, to the opposite logical state of the input of $IC_{1a}$. When the switch is closed, $C_1$ forces $IC_{1a}$ to change state. Positive feedback through the two gates then stores the new state.

Oscillation of $IC_{1a}$ and $IC_{1b}$, when the switch is closed, is prevented by keeping resistor $R_1$ significantly lower in value than resistor $R_2$.

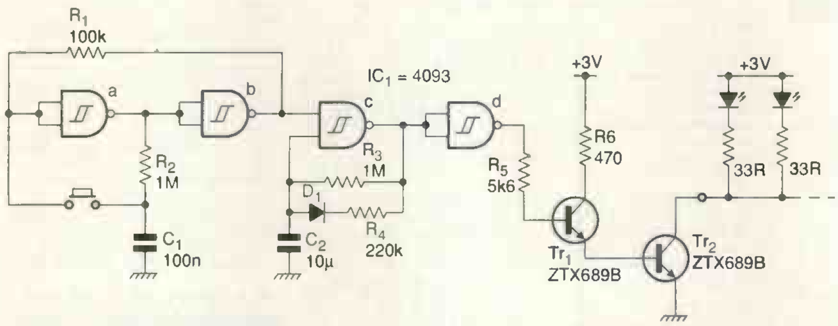
$IC_{1c}$ forms a gated oscillator and is the only reason for using two input nand Schmitts in the circuit. $D_1$ and $R_4$ make the oscillator low output period shorter than the high period.

$IC_{1d}$ corrects the polarity of the output and the two transistors act as a buffer. Transistor $Tr_2$ will drive up to a couple of dozen leds providing the battery will stand it. Transistor $Tr_1$ can be omitted if only one led or two leds are used, but $R_5$ may need to be reduced.

Quiescent current was under one microamp in the prototype. To maintain this, and allow $IC_{1c}$ to operate correctly, $C_2$ must be a low leakage type.

The original design was powered by two AAA cells and drives six leds on a collar so that I can locate my errant dog during night-time woodland walkies!

***Steve Bush***
*Epsom*




# Frequency comparator with hysteresis

A phase-locked loop IC, the *PC74HCT4046*, forms the core of the comparator and provides hysteresis.

Potentiometer $VR_1$ sets the voltage on the IC's voltage-controlled oscillator, $C_1$ and $R_1$ being the timing components. Vco output goes to one input of the phase/frequency comparator and input $f_i$ to the other. Output of the comparator goes to the circuit output via a low-pass filter and by way of the feedback resistor $R_f$ to the bottom of $R_1$.

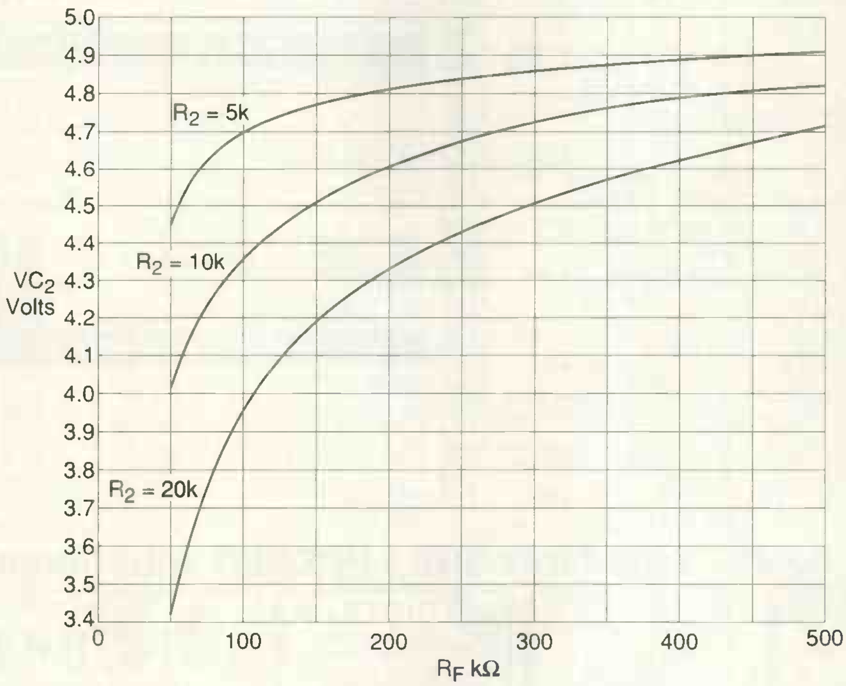
When the input is at a lower frequency than the vco output $FH$, the comparator output is low; when it exceeds the vco output, the reverse applies and the voltage at the junction of $R_1$ and $R_v$ increases, decreasing the current through $R_1$, since the voltage at pin 11 equals that at pin 9.

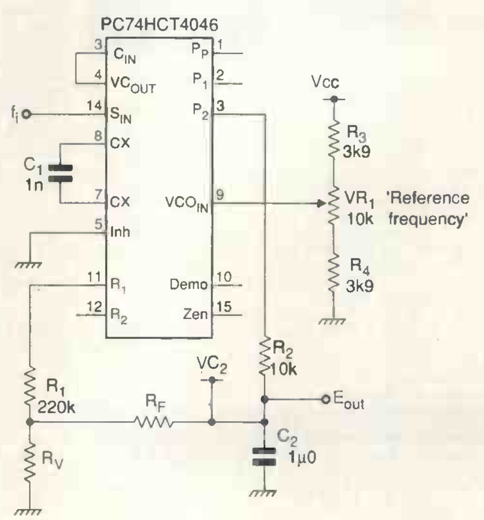
This decreasing current lowers the vco frequency $FL$, so that $FH$-$FL$ is the hysteresis.

$FL/FH = 1 - V_{C2}/(VCO_{in}(1+R_f/R_v))$

where $V_{C2}$ is a function of $R_2$ and $R_f$.

***W Dijkstra***
*Waalre*
*The Netherlands*




***Frequency comparator provides hysteresis adjustable by resistor values.***

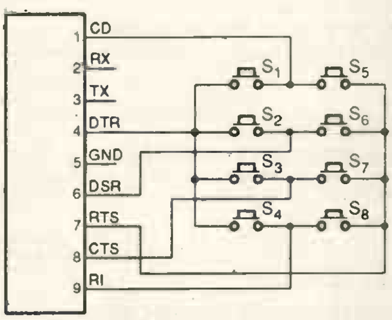
# Control software flow through pc's serial port

Many pc application programs use special defined keys (for example Alt+R) to select different procedures. In the manufacturing environment, sometimes it is inconvenient to access the keyboard directly.

For instance, an application program may define 'push any key to repeat test'. It makes more sense to use just a regular push button rather than a keyboard. The diagram below shows how to use the pc's serial port to connect eight push buttons. Pushing any one of those buttons will lead the program to a specified application procedure.

This power-less approach is controlled via a C program – right.

***Yongping Xia***
*Torrance*
*USA*




***This circuit reads eight buttons through the pc's serial port.***

```
#include <dos.h>
#include <conio.h>
#include <stdio.h>

#define MCR 4 /* modem control register */
#define MSR 6 /* modem status register */

void app_1(void)/* your 1st application */
{
     printf("button 1 is pushed");
}
void app_2(void)/* your 2nd application */
{
     printf("button 2 is pushed");
}
void app_3(void)/* your 3rd application */
{
     print("button 3 is pushed");
}
void app_4(void)/* your 4th application */
{
     printf("button 4 is pushed");
}
void app_5(void)/* your 5th application */
{
     printf("button 5 is pushed");
}
void app_6(void)/* your 6th application */
{
     printf("button 6 is pushed");
}
void app_7(void)/* your 7th application */
{
     printf("button 7 is pushed");
}
void app_8(void)/* your 8th application */
{
     printf("button 8 is pushed");
}
void main(void)
{
     int base_address1=0x3f8;/* COM1 address
*/
     int base_address2=0x2f8;/* COM2 address
*/
     int data1, data2;
     do{
        clrscr();
        outportb(base_address2+MCR, 0x01);
               /* set COM2's DTR high and RTS
low */
        delay(1);
        data1=inportb(base_address2+MSR)/16;
        delay(10);
        data2=inportb(base_address2+MSR)/16;
        if (data1==data2)
         {
         switch(data2){
          case 0x08:app_1();
               break;
          case 0x02:app_2();
               break;
          case 0x01:app_3();
               break;
          case 0x04:app_4();
          }
         }
        outportb(base_address2+MCR, 0x02);
               /* set COM2's DTR low and RTS
high */
        delay(1);
        data1=inportb(base_address2+MSR)/16;
        delay(10);
        data2=inportb(base_address2+MSR)/16;
        if (data1==data2)
         {
         switch(data2){
          case 0x08:app_5();
               break;
          case 0x02:app_6();
               break;
          case 0x01:app_7();
               break;
          case 0x04:app_8();
          }
         }
        delay(200);
     }while(!kbhit());/* hit any key to quit
*/
}
```

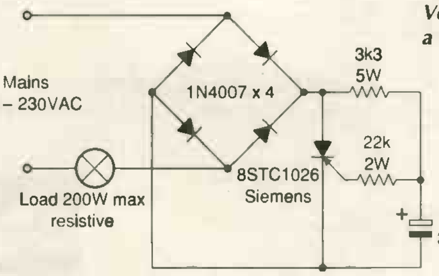
# Simple power flasher

For a simple task, a simple circuit. This is a flasher circuit for a mains-powered lamp.

Initially, the capacitor is discharged and starts to charge through the 3.3kΩ resistor, triggering the scr. The lamp therefore comes on and stays on while current from the capacitor sustains the scr. As the capacitor discharges, the lamp goes off.

Using a 60W bulb, frequency is around 1.1Hz at nominal mains voltage and temperature, decreasing slightly with a 25W lamp; duty cycle ratio is 1:1. Capacitance largely determines frequency and it may be found that a different gate resistor is needed for other scrs.

***D Di Mario***
*Milan*
*Italy*



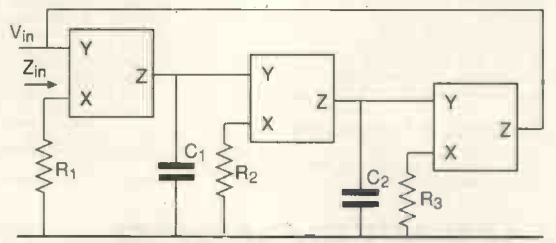

***Very simple circuit to flash a 60W lamp at about 1Hz.***

# Frequency-dependent negative conductance

Current conveyors do not suffer from the bandwidth restriction with feedback seen in common op-amps, so that this fdnc is able to work at much higher frequencies. Further benefits of using cc amplifiers include stability with inductive and capacitive loads.

Input impedance is $-\omega^2 B$, where $B = R_1R_2R_3C_1C_2$.

***K L Sunil Kumar***
*Visakhapatnam*
*India*




***Using current conveyors instead of op-amps in this fdnc increases its bandwidth and improves stability with reactive loads.***

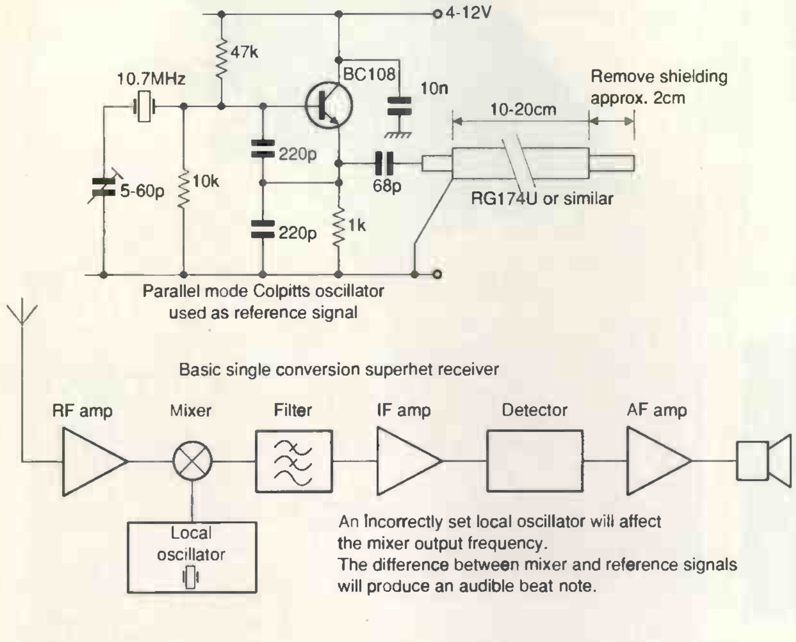
# Retuning local-oscillator crystals

In a receiver using a crystal-tuned local oscillator, changing channel frequency requires the crystal to be fine tuned to produce the same IF. This circuit assists in the process.

A reference frequency of 10.7MHz, generated by a simple crystal oscillator, is coupled capacitively to the mixer by simply putting the end of its output coaxial cable, near the filter.

If an unmodulated signal is now injected at the receiver input, a nominal 10.7MHz IF is produced which beats with the 10.7MHz reference oscillator output, the beat being heard at the receiver output. Adjusting the local oscillator for zero beat gives the correct lo frequency. Use the circuit for either single or double conversion superhets.

***Glyn Roberts***
*Walsall*
*West Midlands*



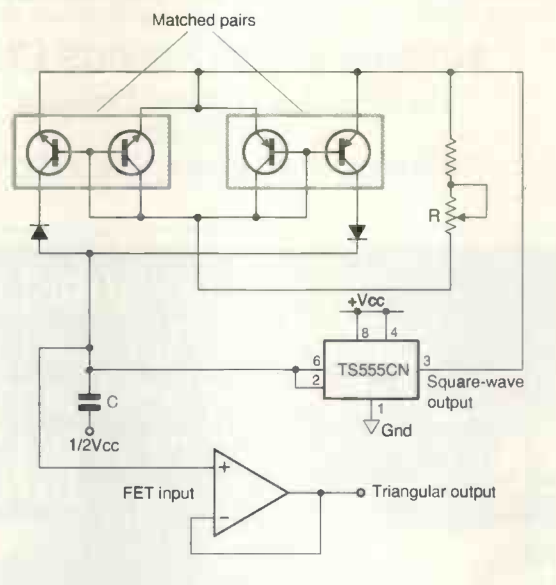

***Beat oscillator helps to retune crystal-based receivers to different channels.***

# Linear square and triangle generator

Constant-current charging and discharging linearises the triangular output from a *555*-based function generator.

When the timer output is at $V_{cc}$, the timing capacitor charges through the p-n-p transistor current mirror. As the ramp reaches $2V_{cc}/3$, the *555* output goes to ground and, since the capacitor voltage is now higher, the capacitor discharges through the n-p-n current mirror until the ramp reaches $V_{cc}/3$. Charging current is adjusted by *R*. Output frequency is variable up to 2.1MHz.

***Lee Szymanski***
*Stamford*
*Lincolnshire*



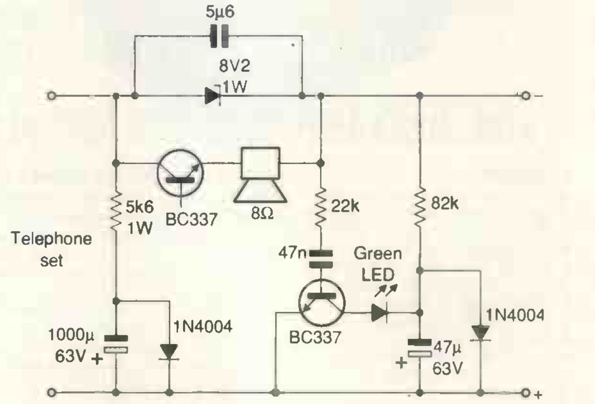

# Telephone line monitor

Exploiting the tendency exhibited by some n-p-n transistors to oscillate when connected in reverse, this circuit uses one to monitor a telephone line and give warning of untoward activity. In normal operation, the circuit has no effect on telephone calls, taking abuse such as reverse voltages, spikes and wrong connections in its stride.

Normal conditions show as a rapidly flashing green led, which stops flashing when a call is made or received. A high-pitched sound shows that reverse polarity is applied on the line side and the connection of another telephone in parallel or a short circuit on the line side sound fades and the led stops flashing. Disconnection or a blown fuse gives a low-frequency tone for nearly two minutes.

Conversation triggers the led every few minutes and ringing or dialling causes it to flash at a different frequency. Normally, the speaker is virtually out of circuit to provide privacy.

***D Di Mario***
*Milan*
*Italy*




***Telephone monitor gives an indication of fault conditions or unusual activity on the line.***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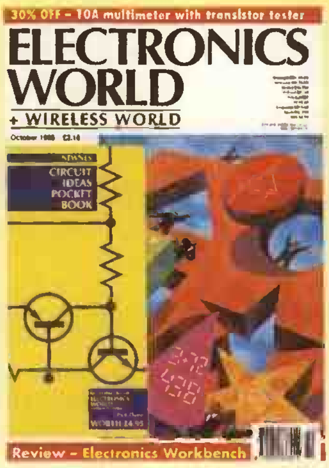


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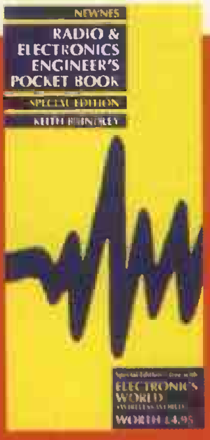


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# Edge detector/doubler

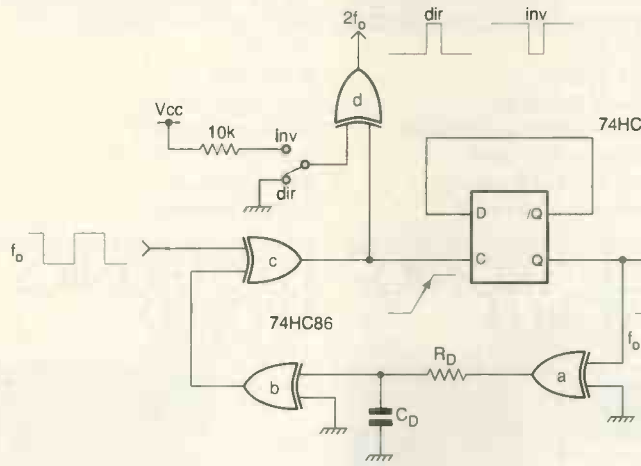
In a similar manner to that of the circuit by Mike McGlinchy (Aug.1994 Circuit Ideas), this circuit responds to both transitions of square input, but this one is self-clocked and a little simpler, although it does have one extra IC.

Delay $R_DC_D$ gives control of output pulse width $t_w$ and may be split between sections a and b of the *74HC86* to equalise propagation delay, allow smaller components or increase pulse width.

It is essential to use cmos logic because of the source-current limitation to $R_D$ when positive-going and the logic zero threshold when negative-going. Values in the table are for a 10V supply.

***John A Haase***
*Colorado State University*

*Edge detector producing output pulses on both transitions of the input at fin/2 and 2fin, positive or negative.*




| | | | | | |
|---|---|---|---|---|---|
| $t_w$, ns | 50 | 70 | 150 | 230 | 400 |
| $C_D$, pF | 0 | 0 | 24 | 50 | 100 |
| $R_D$, kΩ | 0 | 4.7 | 4.7 | 4.7 | 4.7 |

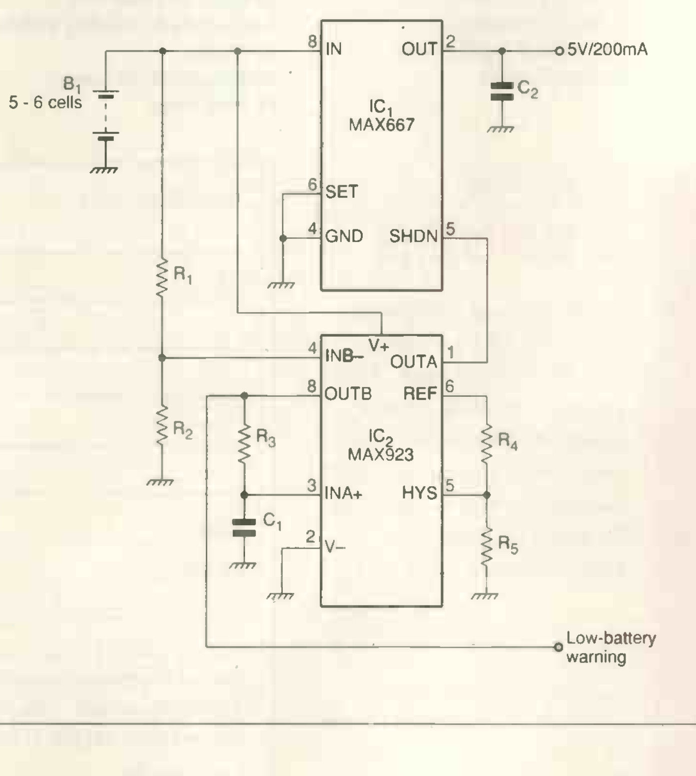
# Low-battery monitor shuts down gracefully

After detecting a low-battery condition, this circuit allows a definite time for emergency housekeeping tasks before shutting down a controlling processor, rather than waiting until battery voltage decreases further. Current drawn while quiescent is a few microamps, so that discharged cells are protected. Accurate voltage monitoring, achieved by the close tolerance of a comparator threshold, allows the battery-low warning to be positioned exactly on the knee of the NiCad discharge characteristic.

Low-dropout linear regulator $IC_1$ supplies 250mA to the output power line, dropping only 350mV at 200mA; $IC_2$ is a dual comparator/±1% voltage reference. If the fraction of the battery voltage at the junction of $R_1$ and $R_2$ falls below the internal 1.182V reference voltage, OUTB goes high, serving as a warning and also charging $C_1$ through $R_3$. When the voltage at pin 3 rises to the reference voltage, OUTA shuts $IC_1$ down. To obtain ±25mV of hysteresis, make $R_4$ 49.9kΩ and $R_5$ 2.4MΩ.

If, for example, $R_3$ is 1MΩ, $C_3$ is calculated by $V_{th}=V_{OUTB}(1-e^{-t/\tau})$, where $V_{OUTB}$ is 4.9V and $\tau$ is $R_3C_1$. For a 1s delay, $\tau$ is 3.6s and $C_1$ is 3.6μF. Alternatively, a standard 3.9μF gives a delay of around 1s; use a low-leakage type. During shutdown, $C_1$ becomes charged and needs about 6s to discharge when operation resumes.

***Craig Falkenham and Larry Suppan***
*Maxim Integrated Products Ltd*
*Theale*
*Berkshire*




*On detecting low battery voltage, this very frugal circuit issues a warning signal and shuts a power line down after a precise time interval, rather than after the usual, somewhat indefinite time taken for the battery voltage to decrease even further.*

# LETTERS

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

## Reflections on optical links

In his article 'Reflections on Optoelectronics' in the November issue, Ian Hickman says that increasing the gain of the photodiode amplifier by 40 dB extends the range of the optical link by a factor of 100, for example from 2m to 200m. I don't agree with this assertion.

Assume, for the moment, in terms of linear units rather than decibels and consider Fig. 12a) of the article. Amplifier $A_2$ has a voltage gain of 100, so if we transfer the oscilloscope probe from the output of $A_1$ to the output of $A_2$ and we want to observe the same signal amplitude, than the current produced by the photodiode has to be reduced by a factor of 100.

Current produced by the photodiode is proportional to the incident light power, which therefore has to be reduced by the same factor.

A ratio of 1:100 in incident light power corresponds to a ratio of 10:1 in distance, so Fig. 12c) displays the simulated output of $A_2$ at a range of 20m rather than 200m.

If we want to argue in terms of the sometimes deceiving decibel, we have to point out that the photodiode is a non linear transducer which can be thought as made up by an electrically linear receiving antenna followed by a quadratic detector.

The antenna establishes a linear relation between the alternating electric field associated to incident light and the voltage produced; the detector establishes a linear relation between the dc current produced by itself and the mean square of the alternative voltage produced by the antenna.

In fact, as the incident light power is proportional to the square of the relevant alternative electric field, we can verify the linearity of the relation between the dc current produced by the photodiode and the incident light power.

So a 20 dB loss in the incident light power (relevant to a 10:1 distance ratio) causes a 40dB loss in the signal current, just recovered by amplifier $A_2$.

To recover the signal loss due to a 100:1 ratio in distance, the amplifier $A_2$ should have to gain 80dB; unfortunately at a range of 200m the output signal of $A_2$ would appear as totally buried by noise since the signal to noise ratio was just enough at a range of 20m, as displayed by Fig. 12c).

***Carlo Carli***
*Ferrara*
*Italy*

## Clarified linear modulation

In his letter of Dec 1995 concerning my article 'Modulating Linearly' – July 1995 – Mr West makes the point that "intermodulation products are usually generated at the power amplifier final stage".

I entirely agree, indeed, the second paragraph of the article runs "In an hf ssb transmitter, it is likely to be the transmitter power amplifier output stage that is principally responsible for... intermodulation products." It goes on to point out that as clean a test signal as possible is desirable for test and measurement purposes.

Nevertheless, it is true that hf ssb power amplifiers only produce the amount of intermodulation products commonly observed, because they are permitted to do so by current regulations 25dB below either tone for R3E, J2E and H3E without privacy device; 35dB with privacy device and for A3E, B8E, R7B, B7B and B7W, per CCIR Recommendation 326.

There is no incentive for manufacturers to produce 'cleaner' power amplifiers, bearing in mind that this would involve extra costs.

However, if this were necessary, the required techniques are already to hand. The Polar Loop technique[1,2] was intended to permit the use of ssb with 5kHz channel spacing at vhf, should this standard ever be introduced. Power amplifier intermodulation products of 55dB below either tone were demonstrated, and the principle

## But EMC testing is *not* required

In Letters, Dec. '95, Chris Bore makes an interesting point regarding emc emissions. He appears, however to be misinformed on the emc directive requirements. Equipment does *not* have to be tested – contrary to what many test houses would have you believe.

Conformance can be shown by submitting a Technical Construction (TCF) to notified body, or you can self-certify if you are confident that it meets the standards.

Secondly, common sense shows that 2W is not sensible limit for input power. The point of emc is to stop interference to radio communications. A device radiating more than a few tens of milliwatts on a broadcast or communications band would clearly see interference.

However, it is still reasonable to assume that devices using sufficiently low power could not emit over the test limits – digital watches for example.

The same would apply to items using low clock speeds and slow logic, as radiated emissions are only measured above 30MHz for most products. If you can show calculations to justify this, you could use the TCF route or self-certify on this basis – there would be no need for a change in the regulations.

Still on the subject of emc, I've found a very cheap way of assessing emissions. From a ham radio shop, for less than £300, you can get a scanning receiver covering 1MHz to 1GHz with a signal strength meter. This is obviously not very accurate, but it does tell you everything needed for development work – ie are there emissions, if so where, and if I make a change does the level go up or down? If you cannot 'hear' your product on this, it is reasonable to assume it won't interfere with anyone else's.

***Mike Harrison***
*Loughton*
*Essex*

## EMC critique deserved

Chris Bore's letter regarding emc and low power circuits has already received some comment in another trade journal, namely *Electronics Weekly*.

Criticism of his thesis has given examples such as oscillators or a photographic flash gun which could produce interference at spot frequencies. And one way of checking an IR remote control is to listen for the interference it can induce into an am radio at zero range.

However, his main point deserves sympathetic consideration for any low power item whose only interference output (if any) would be white or pink noise. After all, the measurement band is up to 1GHz.

As Mr Bore said, let someone competent do the calculations, and come up with some useful figures, including allowance for power delivered to load. The 2W must be power available to cause interference and represent device inefficiency.

***R J Higginson***
*Edgbaston*
*Birmingham*

## Arguments on EMC partly right

I think that Chris Bore's arguments are sound when cw or quasi-cw rf signals are being generated from conventional power supplies. Forty years ago, a young graduate showed, to his joy, how to generate short pulses of 10kV and 200kA using a modest 1kW for less than one minute from the mains. It soon became necessary to measure these pulses, and in a short time conducted interference pulses of 1000V and 100A ($10^5$W) were easily but inadvertently produced.

Pulsed power has moved into everyday use – for example high power lasers, and electric traction. As the rise time approaches 1ns or less, the possibility of radiating an appreciable part of the pulse power from leads only a few centimetres long becomes likely in bad designs. Switched mode power supplies are using ever faster and shorter pulses and have internal pulse powers exceeding their average input and output powers.

I must admit to having enjoyed this subject, especially the challenge of separating a $10^{12}$W source from a diagnostic of $10^{-6}$W (180dB). But now that emc regulations have moved into the third of the 'Three Cultures'[1], I am thankful at not having to justify it in terms of the European regulations.

***E. Thornton***
*Gloucester*

**Reference**


1. Maurice Ramsey, 'Three Cultures', *Physics World*, p72, Aug 1994. (UK Institute of Physics).

would be directly applicable at hf.

The same sort of reduction in odd order intermodulation levels could be achieved by the related Cartesian Loop system, which I believe was also developed at Bath University.

These techniques require the resolution of a sample of the transmitter output into its real and imaginary components at intermediate frequency in order to close the loop. This naturally requires a fair amount of kit, so in the absence of mandatory regulations requiring that sort of performance, it is not surprising that extensive use has not been made of these schemes.

In the mid 'eighties, I developed a simpler arrangement, which was applied to a 150W broadband hf power amplifier module. The latter was designed to be multicoupled up in stages, to provide various powers up to 1kW PEP[3]. This also reduced the third order intermodulation products to 60dB or more below PEP.

Even with this degree of transmitter output stage linearity, it remains true that the main use for an ultralinear modulator is in test and measurement. For any out of band intermodulation products produced by the modulator will usually be adequately suppressed by the following sideband filter. The exception is where modulation is performed without a sideband filter, for example by quadrature modulation.

However, I cannot agree that odd order intermodulation products add to the intelligibility of speech *per se*. Indeed they degrade it. The improvement in hf communications in difficult conditions arises from the fact that, with a very poor signal to noise ratio, the degradation due to intermodulations is more than offset by the improved signal to noise ratio brought about by the increased average radiated power. The instantaneous compression provided by intermediate-frequency clipping followed by a second sideband filter – compared with the slower compression provided by a VOGAD – is very effective at emphasising the quieter components of speech such as unvoiced consonants – particularly sibilants – which otherwise get lost in the noise.

**References**


1. Petrovic V and Gosling W, 'Polar Loop Transmitter', *Electronics Letters*, 1979, 15,(10),pp. 286-8.
2. V Petrovic and C N Smith, 'The Design of VHF SSB Transmitters' IEE Conference on Communications Equipment and Systems, April 1982, pp. 150-5.
3. 'Single Sideband Transmitters', UK Patent 2209639B granted 1 May 1991, assigned to Siemens Plessey Electronic Systems Ltd.


***Ian Hickman***
*Waterlooville*
*Hampshire*

## Complete loony

I am fairly sure that loonier minds than mine will point out that Morgan Jones' valve power amplifier design shown on page 27, January is flawed.

The "concertina" phase splitter concept works adequately at low frequencies but falls over in the higher ranges: the splitter anode presents a much higher drive impedance than that of the cathode mirror. While signal levels look the same at low frequencies, the Miller effect present with high gain, high capacitance audio power pentodes knocks off the sharp signal edges in the top half of the circuit. It is much better to use a double triode virtual earth phase split circuit which produces a symmetrical output in every respect.

But it is all academic really. Who but a complete loony suffering from terminal nostalgia would seriously consider valves for anything? Which is probably the reason that Mr Jones didn't consider dynamic output impedance... But then it takes one to know one.

***Frank Ogden***
*Consulting Editor*

## Fields and health

In his response to Roger Coghill's letter – *EW+WW* May, 1995 – your correspondent Colin Davidson offers to ensure that the IEE will give the widest possible publicity to any evidence concerning possible relationships between low-level, low-frequency electromagnetic fields and health – *EW+WW* Dec, 1995.

There is indeed, abundant evidence of the existence of a direct relationship between all forms of electromagnetic field and living tissue. The demonstration of this relationship relies on concepts of coherent polyphase computation rather than on crude measures of field strength.

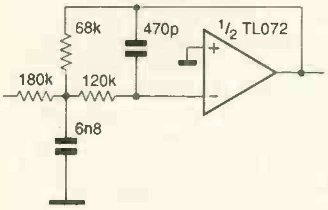
It is principally governed by a bi-linear modulo-four arithmetic as an expression of phase conjugate quadrature rather than by decimal mathematics. The IEE would be well advised to study the subject of bioelectronics with close attention, since the ultimate expression of these effects is that of heuristic electromagnetic resonance as the driving force of evolution.

The Institution should therefore contact the Department of Health or Sir John Maddo, editor of Nature, to whom much relevant information has already been supplied.

***BEP Clement***
*Clement Neuronic Systems*
*Powys*

## Sallen & Key misread?

Before accusing me of departing from the truth, or at least of drawing wrong conclusions, Mr. Skirrow (Letters, December) would have done well to read my November letter more carefully, even to repeat my simple experiment. The distortion figures I tabulated were measured with no capacitors at all in circuit. I did not try the *5532* op-amp, but anyone interested could easily do so. It just seemed odd to me that after all these years the S&K circuit continues to appear in data-sheets without a health warning.




I also wrote to one of the American op-amp manufacturers, and have since heard that the explanation is thought to be the variation of input bias current with common-mode input level. I have also heard from one or two interested readers who have detected audible distortion in S&K circuit.

Mr Skirrow criticises my use of 110k resistors, saying that 3k3 would be optimal, but not why. No doubt 3k3 would reduce distortion, because of the very effect I reported, and probably make it negligible for many applications. But why put up with this constraint – and the cost of larger capacitors – when a better circuit is available?

The replacement Rauch 1kHz circuit referred to in my previous letter is shown here, nominal Q=1.47, and $F_0$ gain unity. It adds no detectable second harmonic (<2ppm) and uses capacitors easily available in polystyrene.

***A. D. Ryder***
*Bolton*
*Lancashire*

## Foster-Seeley related?

Your recent articles on valve amplifiers and Richard Brice's article on the Seeley-Foster discriminator, Dec 1995, have awakened interest and nostalgia. I built a Williamson amplifier while still a student at university, and a valve fm tuner a few years later. Though at the time grasping their essential operating principles, my main memories were of not really understanding the operation of the fm discriminator. Time may not have fully solved the latter problem.

Richard Brice's equivalent circuit for the loosely coupled IF transformer bears a remarkable resemblance to a disastrously designed pulse transformer in which the leakage inductance L(1+k) greatly exceeds the primary inductance –kL.

I found it academically interesting to try to explain the 90° phase shift discussed in his article without resorting to his more rigorous analysis. Clearly, when closely (perfectly) coupled the leakage inductance is zero and AB in his figure represents the primary inductance and the identical reflected secondary inductance of the perfect transformer. Input and output voltages will be in phase.

When loosely coupled, AB (–kL) represents the low valued coupling inductance of the intermediate-frequency transformer, the voltage across which drives the right-hand side of the equivalent circuit. This latter consists of a series resonant combination of capacitance *C* and inductance L(1+k) whose overall series impedance is zero.

There is however a loss term, the right-hand parallel resistance *R*, which I would transform into the time honoured series equivalent series resistance *r*, in series with *L* and *C*.

Voltage generator AB therefore sees to its right-hand side a pure resistance r, and a current will flow in phase with $V_{AB}$. This means that the current through *L* and *C* at resonance is in phase with $V_{AB}$. I assume that *r*, though small, is larger than $j\omega kL$, and does not significantly load the coupling inductance. The voltage across *L* (or *C*) will be 90° out of phase with the current through them so there will be a 90° phase shift between points A and C.

Coupling inductance k*L* is in series with the primary inductance *L* on the left-hand side and their voltages will therefore be in phase. Hence, providing the coupling inductance is not appreciably loaded, $V_{in}$ and $V_{out}$, will be 90° out of phase.

Nostalgically, after about 10 years, the transistor arrived and the mono Williamson was replaced by a stereo pair of Tobey and Dinsdales. The phase-locked loop took over the fm discriminator after a further couple of decades. ■

***E. Thornton.***
*Gloucester*

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## ACTIVE

### A-to-D and D-to-A converters

**12-bit a-to-d**. Analog's *AD9042* analogue-to-digital converter samples at 41Msample/s minimum while giving true 12-bit performance with a spurious-free dynamic range of 80dB at half the sampling rate; intermodulation distortion is 90dB at the Nyquist frequency. Power consumption is 575mW from a single 5V supply. The device is made in Analog's XFCB technique, which is extremely fast complementary bipolar. Analog Devices Ltd. Tel., 01932 266000; fax, 01932 247401.

### Discrete active devices

**6.4GHz power fets**. Hewlett-Packard has the *IM5964-xL* series of internal matched gallium arsenide power fets (Imfets) to operate in the 5.9-6.4GHz comms band. Typical output is over 4W and the devices contain internal matching networks. Hewlett-Packard Ltd. Tel., 01344 366666; fax, 01344 362269.

**Fast diodes**. *BAS* family diodes from ITT provide extremely fast switching and are extremely versatile. These extreme diodes have forward voltages of 1V at 100mA and 1.25V at 200mA, with reverse breakdown at 120V-250V. Power dissipation is 200mW at 25°C. ITT Semiconductors. Tel., 01932 336116; fax, 01932 33148.

### Linear integrated circuits

**Multiplexers**. Analog's *ADG608/609* are eight-channel and differential four-channel multiplexers for 3V, 5V and ±5V supplies. Economy is served by the provision of an on resistance of under 30Ω and less than 0.5nA leakage current, so making them usable in precise circuitry that often requires several more components. Power consumption lies between 0.1μW and 1.5μW, depending on supply, and operation is free from latch-up to 100mA or more. There are also *ADG438F/439F*, with eight single-ended channels and four differential inputs. Both devices are fault-protected, but without the usual loss of performance. Analog Devices Ltd. Tel., 01932 266000; fax, 01932 247401.

### Memory chips

**Flash proms**. EDI has a range of flash simm and plcc modules based on a standard JEDEC 80-pin simm or 68-pin plcc arrangement, with a 32-bit data bus in capacities up to 32Mb. The proms work on 5V and give both more board space and increased memory up to 256Mb, with access times of 100-150ns. EDI (UK). Tel., 01276 472637; fax, 01276 473748.

### Microwave components

**Coaxial attenuators**. *Atlantic Microwave's* new range of attenuators comprises SMA, 2.92mm and Type N units for frequencies from zero to 40GHz. Attenuation is from 0.5dB to 60dB, with accuracy at 30dB of ±0.75dB. Rating is 2W at 25°C and versions with cooling fins to handle up to 50W. They are normally male/female in-line, but can be male/male or female/female and there is a hexagonal body section, if required, to allow accurate setting of coupling torque. Atlantic Microwave Ltd. Tel., 01376 550220; fax, 01376 552145.

**Travelling-wave-tube amplifier**. Thorn Microwave Devices announces the *PTX7439* amplifier, which uses a 9-10.5GHz (other frequencies to order) travelling-wave tube matched to an encapsulated, switched-mode power supply. The amplifier is designed for battery-powered and airborne use, offering an efficiency of 30% minimum at a power output of 45dBm. Built-in circuitry allows monitoring of correct operation. Thorn Microwave Devices Ltd. Tel., 0181 5735555; fax, 0181 5691839.

**Broad-band klystrons**. TMD has two new klystrons for future airborne systems, featuring small size, high efficiency and phase stability in vibration and shock. *PT5184* produces 50kW at 5% bandwidth in X band with 1.5kW of average power, periodic permanent-magnet focussed, although a solenoidally focussed version is being developed that will produce 100kW peak and 5kW average output. *PT6470* gives 1-3kW peak at 2% bandwidth in X band at a very high duty ratio. Thorn Microwave Devices Ltd. Tel., 0181 5735555; fax, 0181 5691839.

### Optical devices

**Wafer inspection**. For the inspection of silicon wafers at high magnification, Vision Engineering has introduced a wafer loader for its *5E Dynascope* optical inspection system. It uses the WED Semiconductor Wafer Handler for rapid operation and precise positioning of wafers before cutting and encapsulation. Vision Engineering Ltd. Tel., 01483 223417; fax, 01483 223297.

Synchronous dram. Designed to match 64-bit memory bus bandwidth, two synchronous drams from Smart can be used to make main memories to run faster than any previously available. They are organised as 1M by 64 and 2M by 64 and offer a throughput of 900Mbyte/s, both being 168-pin unbuffered dimms with serial presence detection. Smart Modular Technology Tel., 01908 234030 fax, 01908 234191

## PASSIVE

### Connectors and cabling

**Printer connectors**. Fujitsu's *FCN-240R* series of parallel PC-to-printer connectors conform to IEEE P1284-C standards, European CE Mark and the CG standard and are a direct replacement for Centronics connectors. They are protected against emi and esd. There are right-angled sockets for pcb mounting and plugs come with light plastic shells or die-cast covers. Rating is 1Adc/240Vac with a contact resistance of 30mΩ; insulation resistance is 1GΩ. Inelco Ltd. Tel., 01734 810799; fax, 01734 810844.

**Comms cable**. Montrose/CDT of Massachusetts has a new line of pairs-in-metal-foil cable designed to comply with new crosstalk standards. Each pair is individually shielded by the foil and the whole braided overall with a pvc or zero-halogen jacket. *EMAXX 300* shows a –55dB performance at 300MHz, while *EMAXX 400* gives the same attenuation at 400MHz. The cables are smaller than comparable types. Montrose/CDT. Tel., 001 (508) 791 3161; fax, 001 (508) 793 9862.

**Mains/phone/data services**. *Rendar* offers an integrated electrical services system to enable 'Plug-and-go' installation of the three types of service in an office complex, based on two, four and six gang UK and IEC socket styles with variable orientation and a choice of switching and fuse options. Rcd, mcd and filter modules are available for protection against earth leakage, overcurrent and transients and the system is easily adapted in the light of future needs. Rendar Ltd. Tel., 01243 866741; fax, 01243 841486.

### Displays

**Colour lcd controller**. For colour liquid-crystal displays, a controller card from Inelco allows the display of 22981 colours. It has 512K of dram on board and comes with utility software. Since the video bios is user-programmable, driving parameters such as refresh rate and resolution can be varied to suit. A *SmartMap* facility allows the intelligent conversion of colour to a grey scale in text mode to solve the problem of applications being optimised for colour crts. The card fits any 16-bit, ISA-compatible slot. Inelco Ltd. Tel., 01734 810799; fax, 01734 810844.

### Filters

**Relay filters**. Solid-state filters contribute enough 150-400kHz noise to come between the domestic EN50081-2 and the industrial EN50081-2 standards. Relay maker *Crydom* has introduced relays with filters to suppress this noise in single and three-phase applications, simply being connected across incoming line or phases. One filter suffices for several relays at currents over 15A. Noise reduction at 50A is from 70dBμV to 35dBμV at 150kHz and to

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40dBμV at 250kHz at 30A. Crydom Europe. Tel., 0181 763 0550; fax, 0181 763 0499.

## Hardware

**PC card cover.** Molex's *Snapper* is a one-piece, stainless-steel cover for PCMCIA PC cards that needs no epoxy or tape to fit the card to the frame. It is usable with Types I and II cards In such applications as modems needing shielding. No heat or pressure is needed; a small arbor press closes and secures the cover, which is complete with 68-circuit standard interface connector, grounding clips and i/o connectors, also available with solder tails. A range of cable assemblies and I/o connectors is available and other kits for different PC cards. Molex Electronics Ltd. Tel., 01420 477070; fax, 01420 478185.

**Shielded touch screens.** Lucas *Duralith* resistive touch screens provide electrostatic and magnetic Interference shielding. They comply fully with emc legislation, now mandatory, and will protect the screen and other components against discharge from an external object – a finger – charged to several kilovolts. The emi shielding not only protects the screens but reduces emissions. Anders Electronics plc. Tel., 0171 3887171; fax, 0171 3872951.

**EMC filters for 3-phase.** *FN 258* is a filter for three-phase industrial frequency inverters with a universal voltage rating of 480V by Schaffner, meeting EN 133200 and the American UL 1283. The filters are in nine variants for currents from 1A to 180A and, since its temperature rating Is 10°C higher than usual, it can be used in most conditions without de-rating. Two-stage filtering provides for output cables up to 75m in length. Schaffner EMC Ltd. Tel., 01734 770070; fax, 01734 792969.

**Embedded PC chassis.** IMS announces the *MBPC-641 Microbox* chassis of about the size of a shoe box to contain an embedded industrial PC controller safely and in a small space. It has a four-slot PC/AT bus backplane, a 65W power supply with fans and a number of connectors for i/o. There is a range of plug-in 286/386/486 cards with solid-state disks to fit the chassis, with three slots spare. Integrated Measurement Systems Ltd. Tel., 01703 771143; fax, 01703 704301.

## Test and measurement

**Surge testing.** A surge generator from Seaward, the *THOR* tests electrical and electronic equipment for surge immunity by replicating large surges of the type caused by power bursts, lightning and some other sources. Outputs are selectable from 0.5kV to 4kV at up to 16A and an oscilloscope connection is provided. The unit is controlled by a microprocessor to give easy operation and a library of test routines. Seaward Electronic Ltd. Tel., 0191 586 3511; fax, 0191 586 0227.

**Non-contact profile measurement.** UBM offers the *UBR200* non-contact measurement system, which can be used with a *Microfocus* optical sensor to replace the stylus in a profilometer, so reducing the risk of damage and increasing speed of measurement. No modifications are necessary and the system works exactly as before. *Microfocus* automatically controls lens position to maintain focus on the object's surface and a second system monitors lens position, which is a replica of the surface profile. Two ranges are produced: ±500μm and ±50μm, the laser power being selected to match. Advanced Products and Technologies Ltd. Tel., 01865 724863; fax, 01865 725831.

**Laser fault finder.** *ME301* by the Spanish company Molher Electronica is a visual fault locator using a laser to find several types of fault In optical fibres, emitting a red glow to indicate a point of high loss in the fibre caused by tight bends or crimps, bad connections, poor splices or breaks, and will identify fibres. Output is selected for cw, low-power cw and a 2Hz pulsed signal, the selection being indicated by a led. A timer can be selected to allow five minutes of operation. Molher Electronica sa. Tel., 0034 1 462 25 62; fax, 0034 1 465 53 82.

**Gas monitoring.** The cost of area gas monitoring for health and building applications is reduced by CBISS's *Intelligent Sampling System Mk 2*, which collects samples of up to four gases in eight areas. Since that is equivalent to 32 monitoring points, the cost is brought down to £140 per point. Gas comes in through ptfe or nylon tube to the central analyser, which allows alarms and data logging to be carried out centrally without the use of additional modules. Auto-zero and calibration are incorporated and detector faults are shown and their results ignored. Windows software for data acquisition is supplied. A wide range of gases can be detected. CBISS Ltd. Tel., 0151 3431543; fax, 0151 3431847.

**Analogue/digital dmm.** From Di-loG, the *DL-295* digital multimeter, which measures voltage, current, resistance, frequency to 200kHz and temperature in the –40°C to 1370°C. In addition to the 3.75-digit readout, there is also a 43-segment analogue bar-graph display with a 'zoom' facility for precise readings. Other facilities are a comparison function between reference and measured values, lowest and highest readings over time and a hold facility. Di-loG Ltd. Tel., 01707 375550; fax, 01707 393277.

**Function generators.** Yokogawa *FG200/300* are synthesised function generators with touch-screen control. The *FG200* series has sweep and modulation on two independent channels at 0.000001Hz-15MHz and up to ±10V, the *FG300* types also providing arbitrary sweep and waveform definition. All waveform parameters can be set independently for the two channels. An extra on the *FG300* instruments is a floppy drive for waveform output or input. Martron Instruments Ltd. Tel., 01494 459200; fax, 01494 535002.

**PC-to-storage-oscilloscope module.** Converting a PC into a 22kHz digital storage oscilloscope, the Allison *O-Scope* takes the form of a module to connect to the printer port and software. Features include a spectrum analyser mode and the normal facilities offered by the PC for saving traces and printing can be used. Allison envisages the *O-Scope* being used in audio, data logging, car electronics and the like. Allison Technology Corp. Tel., 001 800 980 9806; fax, 001 713 777 4746.

**Am/fm signal generator.** The *Topward 8015* 2MHz frequency generator offers both amplitude and frequency modulation at a cost of £159. Its features include 5mV-20Vpk output on sine, square, triangular, ramp and pulse waveforms, a polarity inversion switch, variable dc offset and variable duty ratio on ramps and pulses. There is an internal 400Hz modulation oscillator, with provision for an external source. Tandem Technology Ltd. Tel., 01243 576121; fax, 01243 576119.

**Keypad for lefties.** If you are bored with the number pad on the right of your keyboard, try the Cherry *G80-3700* – one that can go anywhere. It is a separate pad to link directly with ordinary IBM-compatible keyboards by way of a PC pass-through port. It has the usual number pad keys, plus four programmable function keys. It is claimed to be of benefit for left-handed operators. Cherry Electrical Products Ltd. Tel., 01582 763100; fax, 01582 768883.

## Literature

**Test gear.** TTi has a new catalogue of other people's instruments – a range of equipment from the world's makers, including oscilloscopes, power supplies, audio analysers, video test equipment and mains analysers. Manufacturers include Tektronix, Stanford and Hitachi. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

**Power components.** A new 150-page catalogue is produced by *Pico*, containing details of surface-mounted and plug-in transformers and inductors, dc-to-dc converters and power supplies. The transformer section has a wide selection of audio and ultra-miniature components. Ginsbury (UK) Ltd. Tel., 01634 290903; fax, 01634 290904.

**Panel meters.** Europa Components has produced a 20-page catalogue of Crompton Greaves DIN standard panel meters, which meet the *DIN57411* sheet 1/VDE0411 pt1, proving that not only are specification names becoming longer, but that these meters are suitable for very harsh surroundings. Main product is the *S100* range, which is adaptable to most requirements, being of moving-iron and moving-coil types, conforming to all manner of other specifications, all contained in glass-filled polycarbonate cases with a black bezel. Europa Components & Equipment plc. Tel., 0181-953 2379; fax, 0181-207 6646.

**Modems and GPS.** Rockwell offers a 50-page handbook and guide to its range of modems and Global Positioning System devices, boards and evaluation products. It contains a glossary of terms in the telecomms, datacomms and navigation fields. Telecom Design Communications Ltd. Tel., 01256 332800; fax, 01256 332810.

**Telecomms.** A short catalogue from Stanford Telecom gives details of products in the areas of demodulation and spread-spectrum asics and boards, frequency synthesiser boards and subsystems, forward error correction at up to 45Mb/s and digital communications. BFI IBEXSA Electronics Ltd. Tel., 01622 882467; fax, 01622 882469.

**Computer-based instruments.** National has a catalogue of software and hardware to form instrumentation and industrial automation based on the use of computers. In over 600 pages, the catalogue is in six sections: software, data acquisition, GPIB, VXI/MXI, industrial communications and customer education. National Instruments UK. Tel., 01635 572400; fax, 01635 523154.

**Three books on VMEbus.** Three reference books and an 'edited highlights' from a market report on the VMEbus are available from Vita Europe. There is a product directory to about 3000 products from around the world, a 300-page handbook for engineers and programmers and the VME64 specification for 8 to 64 bit parallel-bus architectures. All the books cost around £35 and the report is free. VITA Europe Ltd. Tel., 01329 841272; fax, 01329 846166.


**Slotted opto-sensor.** Omron believes its *EESX 1101* range of transmissive, slotted opto-sensors to be the smallest available. The whole thing is only 4.3 by 4 by 5mm and led and phototransistor each fit into a width of 1.15mm, leaving a 2mm wide slot. A Fresnel lens focuses the led output on the detector to give a high current. Devices are of two kinds: a standard phototransistor or a photo-ic output for improved switching speed, the latter having an amplifier and regulator, with a Schmitt. Omron Electronics Ltd. Tel., 0181 450 4646; fax, 0181 450 8087.

**PIC guide.** Polar Electronics has updated the *Beginners Guide to the Microchip PIC*, but it is still at the old price of £19.95 from catalogue companies such as Maplin, Farnell, RS and Rapid. It is now spirally bound to lay flat and contains information on hardware and software design, assembly and debugging. The price includes a disk of useful design software. Most of the new information presented is on the 16Cxx family. For volume purchases, speak to Ian Ewin or Peter Greenslade at Resource, Polar Group on 01525 858200, fax: 01525 858101.

**Wire, cable and tubes.** *Alpha Wire* introduces the Master Catalogue, which takes the form of a guide to the selection and specification of wire, cable and tubing for the electronics industry. In 400 pages, the publication is in ten sections and includes shrinkable tubing, coaxial cables, lan cables, and all kinds of more familiar products such as hook-up wire. There are also 38 pages of technical data, including information on cable shielding. Alpha Wire Ltd. Tel., 01932 772422; fax, 01932 772433.

## Materials

**Tamper detection.** Electrolube offers *Bloc'Lube*, which gives the game away when someone who shouldn't attacks a screw or tuned coil. It is applied to the component, remaining tacky for a while until final adjustment is made and drying brittle, so that any further attempt at adjustment cracks it and shows. You can get most of it off with a screwdriver and any bits left with a solvent. Electrolube Ltd. Tel., 01734 403014/031; fax, 01734 403084.

**Solder paste.** Loctite's new solder pastes are particularly formulated for use in precise, miniature circuitry. The *3824*, *3825* and *3828* are of natural rosin, mildly activated and exhibit 0.2mm 'slump' on 0.7mm pads after an hour at room temperature. No cleaning is needed after reflow soldering and the wet strength is retained for long periods. Loctite UK Ltd. Tel., 01707 821000; fax, 01707 821200.

## Printers and controllers

**Thermal printer.** A 42-column version of Epson's M-TS60 thermal printer, the *M-TS63*, is now available. It operates in both directions to print two lines at the same time at a rate of 2.4lines/s. In spite of the width and high speed, the unit's price is only 70% that of comparable printers. The unit measures less than 10cm wide and comes with a serial RS-232C board to provide a paper feed switch, self test, 5V power supervisor and head-jam and motor-stall protection. Able Systems Ltd. Tel., 01606 48621; fax, 01606 44903.

**Another thermal printer.** Fujitsu claims the world paper speed record for its *FTP600 series* thermal line dot printers at 100mm/s, running at 40% of the power needed by earlier types. The printers are available in 2in, 3in and 4in versions, producing 480, 640 and 832dots/line. *FTP621/631/641* are produced as mechanisms only or complete with interface or as a set with microcontroller and gate array. Paper cutting mechanisms are offered. Inelco Ltd. Tel., 01734 810799; fax, 01734 810844.

## Production equipment

**Hot jet handpiece.** The *Royel 100 hot jet handpiece* is suitable for reflow soldering or solder paste reflow in surface-mounting application, shrinking tubing and component test. It can be operated from a variable air flow and digital feedback-controlled power unit such as the HJ1000 or one of the *Royel* workstations. An N-type thermocouple gives continuous readout of air or gas temperature. Production Equipment Sales Ltd. Tel., 01323 811694; fax, 01323 811528.

## Power supplies

**Micropower dc-to-dc converter.** *SC1652CS* from Semtech is a converter designed for lcd bias contrast application, having an 87% efficiency. It is an inverting type driving an external switch to generate programmable negative voltages, output being scaled to –40V or more by two resistors. Quiescent current is 80μA or 0.7μA when shut down. Semtech Ltd. Tel., 01592 773520; fax, 01592 774781.

**100W dc-to-dc converter.** Abbott's *SMH50* and *SMH100* modules complete the family of 200kHz single and dual output converters of 50, 100 and 200/280W ratings, with complementary emc filters, transient filters and ac/dc front end modules. Standard input is 18-36V and outputs cover the 2-48V range. There is no current de-rating up to 100°C. Abbott Electronics Ltd. Tel., 01233 623404; fax, 01233 641777.

**Low-loss regulators.** Input/output difference voltage of Sanken's *SI-3001N* ic voltage regulators is 1V and power dissipation 1.5W with no heat sink; output current is 1.5A. Protection against over current, too high an input voltage and overheating. The series is suitable for use when a constant output voltage is needed without on/off control. Allegro MicroSystems Inc. Tel., 01932 253355; fax, 01932 246622.

Three-phase harmonic analysis. Fluke's *41B Power Harmonics Analyser* is a hand-held instrument measuring true rms voltage and current, frequency and power factor and displaying three-phase power. Display is as a waveform, as a bar-graph showing the level of harmonics present or as a numeric value, data being downloaded if required. PIDA. Tel. (and fax), 0756 799737.

**Lower-cost switchers.** Calex says its new 100W switched-mode power supply, while retaining all the advantages possessed by that type of design, now costs about the same as an unregulated transformer type. its output is 24V, 5A and is proof against shorts; ripple less than 50mVpk-pk; regulation less than 1% overall; and mains buffering greater than 20ms at 5A. The supply conforms to the relevant interference, emission and safety standards. Calex Electronics Ltd. Tel., 01525 373178; fax, 01525 851319.

**Lamp transformers.** Meant to drive the cold-cathode fluorescent lamps used in flat-panel displays, transformers by *Coiltronics* come in power ratings of 2.5, 4, 6 and 14W in a variety of mounting styles. They are usable with floating or tied-secondary designs and give up to 30mA at 40-80kHz. Output is sinusoidal and emi/rfi consequently low. Microelectronics Technology Ltd. Tel., 01844 278781; fax, 01844 278746.

**DC-to-DC converters.** Melcher's *IMR* family of converters offers ranges of 3, 6 and 15W and is for use in reasonable environments. Single or double outputs are 5V, 12V or 15Vdc

Please quote *"Electronics World"* when seeking further information

and input range is 2:1. The smallest unit comes in a 24-pin dil package, the other two being in 2in square modules, both types having standard pinouts. Open and short circuit protection is provided and i/o isolation is 500Vdc. Melcher Ltd. Tel., 01425 474752; fax, 01425 474768.

## Switches and relays

**Solenoid.** BLP's new *PED Series 66* is a low-cost device suited to uses in which a high force:stroke ratio is needed, but where there isn't much space. It measures 30 by 16 by 14mm, has a 3W continuous rating and pulls a load of 0.1kgf at 10mm stroke; 1.1kgf at 1mm. Coils are for 5, 6, 12, 24 and 48V and the solenoids are available in push and pull versions. BLP Components Ltd. Tel., 01638 665161; fax, 01638 660718.

## Television components

**Character generator.** From Philips, the *PCA8516* stand-alone on-screen display generator, which allows the display of up to 256 high-resolution characters from a font containing 253 custom characters. On a 12 by 18 dot-matrix area, the device displays Japanese and Chinese writing systems and adjacent cells can be combined vertically and horizontally to make icons and semi-graphics. It is programmable to suit all common television scan formats. Philips Semiconductors (Eindhoven). Tel., 00 31 40 722091; fax, 00 31 40 724825.

## Transducers and sensors

**Bending actuator.** For large displacements at low voltages, Philips offers the *CMA* (ceramic multilayer actuator), which is a piezoceramic strip with electrodes to cause it to bend under applied voltage and actuate external equipment such as valves. Since the piezo layers are very thin, drive voltages from 10V to 60V produce electric field strengths up to 3kV/mm; conventional types of actuator would need up to 1kV to give the same field strength. Speed is higher and the device is smaller than is usual, requiring only voltage drive and therefore much less power. Philips Components. Tel., 00 31 40 722790; fax, 00 31 40 724547.

**Linear actuator.** *Electro-Thrust* is a linear actuator combining the simplicity of a pneumatic type with the precision of a mechanical cylinder. It is programmable, repeatable to within ±0.013mm and comes in stroke lengths in the 50-1000mm range. Velocity can be controlled to speeds up to 1250mm/s at thrusts of up to 7200N. The actuators are available in frame sizes of 32, 50 and 80mm in nine metric ISO mounting styles. The range of leadscrew pitches and drive ratios makes for easy matching to an application, as do the four rod end choices. Parker Hannifin plc, Digiplan Division. Tel., 01202 699000; fax, 01202 695750.

## Vision systems

**Stereo vision.** Sundance has a dual digital video interface module, the *SMT318*, to provide stereo vision in embedded systems. It provides a digital data-capture node for TMS320C4X TIM-based systems and can be used as the interface to two digital cameras. The two interfaces give a peak acquisition bandwidth of 60Mbyte/s and linking them gives a single 16-bit interface. Independent control of each channel is provided and there are 10 general-purpose, programmable i/o lines. The interface can be used with Pulnix, Dalsa and EEV cameras. Sundance Multiprocessor Technology Ltd. Tel., 01494 431203; fax, 01494 726363.

# COMPUTER

## Computer board-level products

**Pentium motherboards.** *Cosworth*, from Apricot, is an ISA/PCI motherboard for the full Pentium Pro family, the relevant sockets, clock and bus speeds up to 66MHz being provided. There is on-board memory of up to 1Gbyte in 3.3V dimms and a Cirrus Logic Alpine GD543x or 544x chip set copes with graphics. Dual-mode PCI IDE ports with two sockets are provided for hard disks or Atapi CD-rom drives. Apricot Computers Ltd. Tel., 0121 717 7171; fax, 0121 717 3692.

## Data acquisition

**Plug-and-play.** A new multi-function data acquisition board from National, the 1Msample/s *AT-MIO-16E-1* is plug-and-play ISA-compatible and uses the company's E Series architecture to eliminate jumpers, switches or potentiometers. There are 16 single-ended inputs, 16 pseudo-differential inputs with a shared common or eight 12-bit full differential inputs; two analogue outputs have 12-bit resolution, eight digital i/o lines sink 24mA on each line and there are two 24-hour counter/timer channels. National Instruments UK. Tel., 01635 572400; fax, 01635 523154.

## Data communications

**Wireless lan chipset.** Harris introduces a four-member chipset, *Prism*, for 2.4GHz direct-sequence, spread-spectrum, wireless lan systems, to be used as the core of transceivers for Type II PCMCIA and ISA network adaptor cards for 4Mb/s working. Features offered include a resistor-programmable filter, programmed timers and threshold levels in the baseband processor. Harris Semiconductor UK. Tel., 01276 686886; fax, 01276 682323.

## Development and evaluation

**Rom emulator.** Nexus's *ROMbox* is a rom emulator designed to assist code and hardware development by allowing rom image downloading at 200kb/s from a PC, and also by providing aids to debugging of both hardware and software. Trace functions are included for address matching and a bidirectional link obviates the need for an extra serial port for development. Up to 16 ROMboxes can be daisy-chained on a single cable to support 128b buses. The unit can be plugged into a 32-pin or 28-pin rom socket and adaptors are available for plcc sockets, the connection to the PC parallel port being by a single cable. Two versions have memory of 1Mb and 4Mb. Nexus Electronics Ltd. Tel., 01223 576100; fax, 01223 576619.

**Device drivers.** A Windows tool from IAR, *DriveWay-51*, is launched as "the fastest way to generate device drivers for the 8051 family". It is produced in association with the Israeli company AiSys and allows integration of on-chip peripherals into designs, automatically producing documented and tested driver C source code for each 8051 peripheral. Test functions and documentation are generated automatically, as is an on-line data sheet on the chip's peripherals, modes, registers and pins. You will need 5Mbyte of hard disk space and 4Mb of memory. IAR Systems Ltd. Tel., 0171 9243334; fax, 0171 9245341.

## Computer peripherals

**RS-485 as GPIB.** National's *GPIB-485CT-A* is an external box that makes a computer with an RS-485 port behave as a full-function IEEE 488.2 controller, an RS-485 device appearing as a GPIB device, the system being effective over a distance of up to 4000ft. National also announces the *PCI-GPIB*, a plug-in GPIB instrument control board for computers with PCI bus running Windows 95. It handles data transfer rates to 1.5Mbyte/s and implements the HS488 GPIB protocol for programmed i/o transfers at 3.7Mbyte/s or more. With this board, a computer is able to monitor and control several thousand pieces of equipment. National Instruments UK. Tel., 01635 572400; fax, 01635 523154.

Software

**Waveform generator.** Thurlby Thandar introduces a Windows-based package, *WaveForm DSP*, to support arbitrary waveform generation on the Model TG1010 function generator, which is a 10MHz direct digital synthesiser. The package creates, analyses and edits waveforms, which have been made by any combination of drawing, mathematical expression, output from a dso, or taken from a library or pasted from other applications; waveforms may be drawn by mouse and smoothed with curve-fitting algorithms. It switches easily from the time to the frequency domain, both possessing editing, drawing and library features. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

## Computer security

**Data protection.** Jetico, Inc., of Finland, produces the *BestCrypt+* data protection system for PC compatibles, which is said to provide the most secure storage and yet offer easy access control to the encrypted data on disk, where it behaves transparently to an authorised user from any application program. The hardware is an add-on board and the software contains Control Panel for MS-DOS and Windows. Encryption is by means of the Russian Federal standard GOST 38147-89, which is, apparently, well known as an uncrackable algorithm, being $10^{50}$ times more uncrackable than the American standard. And it isn't any good trying to guess a password, because they have thought of that, too, and made it impossible to get it even by looking over someone's shoulder. Jetico Inc. Tel., 00358-31-316-5215; fax, 00358-31-316-5901. e-mail jetico@scl.fi ■

# Load dump generators for emc testing

Automotive generators that cope with the new ISO 7637 standard and the more stringent emc test requirements in a modern car are the subject of Schaffner application note INS0002.

Entitled *Discussion on load-dump generator designs and suitability for automotive EMC testing,* the note describes how *NSG500x* emc test generators are of use in evaluating automotive emc.

Transient and burst generators typically need to generate very fast rise times and short pulse widths. A typical transient generator is shown in **Fig. 1**.

Pulse shaping components are typically passive components – capacitors, inductors, resistors – because current solid state technology cannot generally provide the pulse shape control with the rise times and energies involved. As a result of this output resistance $R_i$ is typically specified as a variable value to allow for some control over the energy delivered to the equipment under test – $R_L$ in Fig. 1. This mechanism however provides very limited energy control on the output. If you take the case where $R_L$=$R_i$ it is clear that half the pulse energy is dissipated internally in the generator rather than the equipment under test, or eut. In addition pulse amplitude at the eut would be half the value programmed.

In the case of the load dump generator for the *NSG5000,* because rise times and pulse width requirements are in the millisecond region, this pulse generator is based entirely on solid state technology. This means pulse shape and energy content, delivered to the eut, are fully programmable. As a result, test pulses delivered to the load can be much more predictable. This provides the user with test modes previously not possible with traditional passive circuit designs.

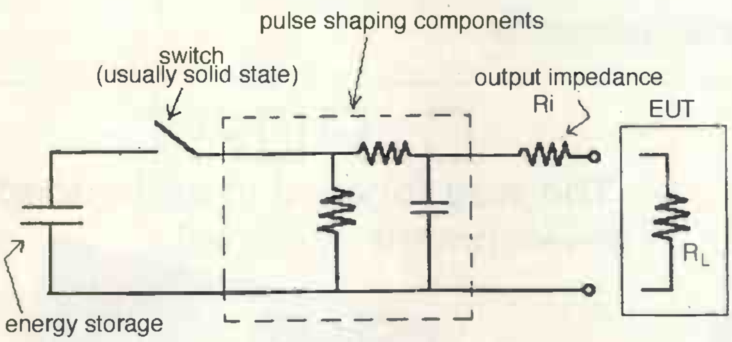



***Fig. 1.** Simplified block diagram of a transient generator. Both transient and burst generators typically have requirements for generating very fast rise times and short pulse widths.*

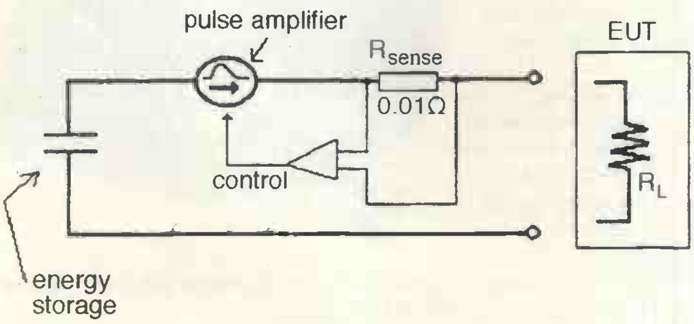



***Fig. 2.** Simplified schematic of the NSG 5005 operating in current mode. An in-line sense resistor develops a feedback voltage to control the pulse amplifier.*

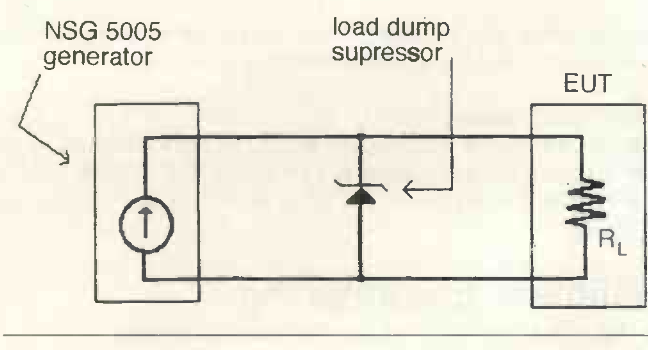



***Fig. 3.** Centralised load dump suppressor in the test harness to the EUT helps simulate the type of environment that the EUT might meet in an automobile.*

## Operating modes

The instrument is intended for three basic modes of operation, one of which is current mode output.

This mode of operation could be considered a new departure in terms of the type of pulses and pulse specifications that are presently specified in the ISO 7637 standard. However, outputting a current pulse shape is a much more accurate representation of what actually happens with a load dump from an alternator in a modern automobile.

An in-line sense resistor is used to develop a feedback voltage to control the pulse amplifier, **Fig. 2**. The value of this resistor is very small so it does not dissipate a significant amount of energy.

In this mode the pulse shape would normally be defined into a short circuit. Once the eut internal resistance is low enough to allow the maximum programmed peak current to flow then the pulse shape programmed will be replicated exactly at the output. Voltage developed at the output terminals, ie the equipment under test, is determined by load impedance $R_L$.

Maximum output voltage available from the *NSG5005* is 200V so the unit will clamp around this level if no other limiting device is present in the test circuit. A more common situation would be to include a centralised load-dump suppressor in the test harness to the eut, **Fig. 3**. Again this also helps to represent the environment that the eut might meet in the automobile more accurately.

If the eut load represents a high inductance or capacitance then the current shape may be distorted from that programmed.

The note also describes the normal voltage mode output and the voltage mode with external $R_i$ resistance unit. Maximum performance limits of the *NSG5005* being a maximum pulse amplitude of 200V, pulse width of 500ms and a maximum current of 200A.

An appendix describes how pulse shape characteristics depend on the eut load and how quickly it drains energy from the generator.

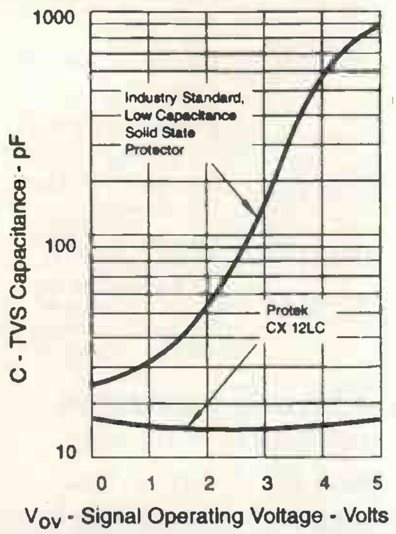
***Schaffner EMC Ltd.** Ashville way, Molly Millar's lane, Wokingham, Berkshire. Tel, 01734 770070, fax, 01734 792 969.*

# Surge protection solutions

Transient voltage protection products are the subject of a data and applications manual from Protek Devices. Dedicated to engineering solutions for the transient environment, the manual is split into five sections covering tvs diodes, Discrete TVS diodes, power tvs assemblies and high power surge suppressor modules. The fifth section covers application notes.

An interesting device designed to protect interfacing equipment from induced lightning or switching transients is detailed in section 4. Called the *CX 12LC* module, the device is a two stage, hybrid surge protector with a low clamping voltage, high energy handling capabilities and an operating data range up to 100Mbit/s. It is designed for high data rate applications over the operating voltage range of a computer.

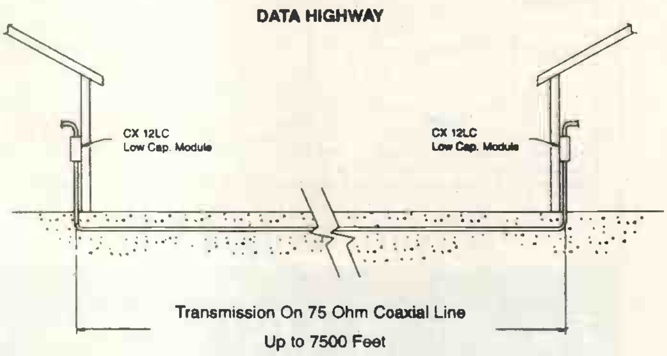
Coupling capacitors inserted across the data line must be very low to prevent signal distortion or loss of data on the LAN network. The diagram (left) shows a comparison of two protection products inserted in the data line for transient voltage protection. The bottom line is for the CX 12LC, and the top line is for a more standard device. Due to the drastic change in capacitance of the product, top line, signal distortion, loss of data or even access to the computer may be a problem.




***Capacitance in protection devices can be a source of significant signal attenuation.***

***Protek Devices**, 2929 Fair Lane, Tempe, Arizona 85282, USA, Tel, 602-431-8101, fax, 602-431-2288.*

***In a typical data line application, as shown, video and serial data transmission lines are susceptible to lightning strikes and surges from ac power lines. The CX 12LC is designed to protect against such surges.***




# 'Uncrackable' electronic lock

Claimed to be the world's first unpickable electronic lock, the dynamic key alarm micro – *DKA1* – from Electronic Research and design Ltd lends itself to many security applications.

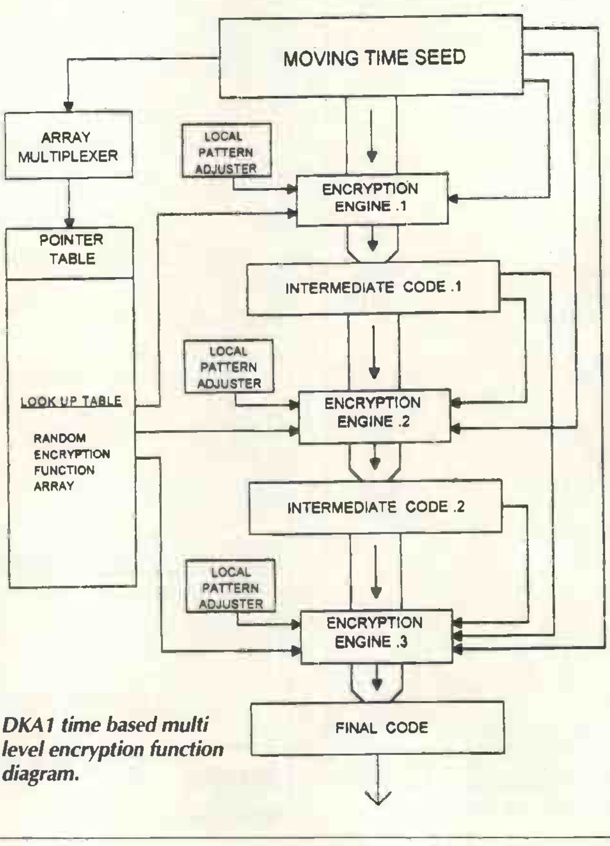
The device is particularly applicable to keyless lock designs relying on radio transmission. Such lock systems have traditionally been unsuitable for high-security applications.

Detailed in its technical data brief, the *DKA1* is said to incorporate levels of security that have only recently become technically feasible. The system includes generation III military type encryption techniques based on the manufacturer's time-based multi-level encryption technology. This is said to be uncrackable – even by the most advanced 'grabber predictors'.

When originally programmed, a base time seed code is generated. This is different for every key and forms the start time-seed. This seed is then clocked in real time and a proportion of the encryption mechanism is 'weighted' by the current time seed variables. These variables are constantly changing. This means that the encryption engine parameters are changing all the time. Together with the third level encryption on encryption coding techniques (multi level), this combines to make it impossible to calculate/predict consecutive codes. Even with knowledge of the multi-level encryption algorithm, it would still be impossible to decode the current time based multi-level encryption code without the knowledge of the current time seed. This can only be known by that particular key's run-time file in the decoding to which it was initially synchronised.

Conversely, because the decoder has an identical run-time file and the same moving time seed, it is able to de-encrypt the incoming code and observe any comparison, see diagram.

The brief also details the device's specifications, alarm features and diagnostics.




***DKA1 time based multi level encryption function diagram.***

***Electronic Research and Development**, K&K House, Station Approach, Rickmansworth Road, Watford, Herts, WD1 7LU. Tel. 01923 240525, fax 220011.*

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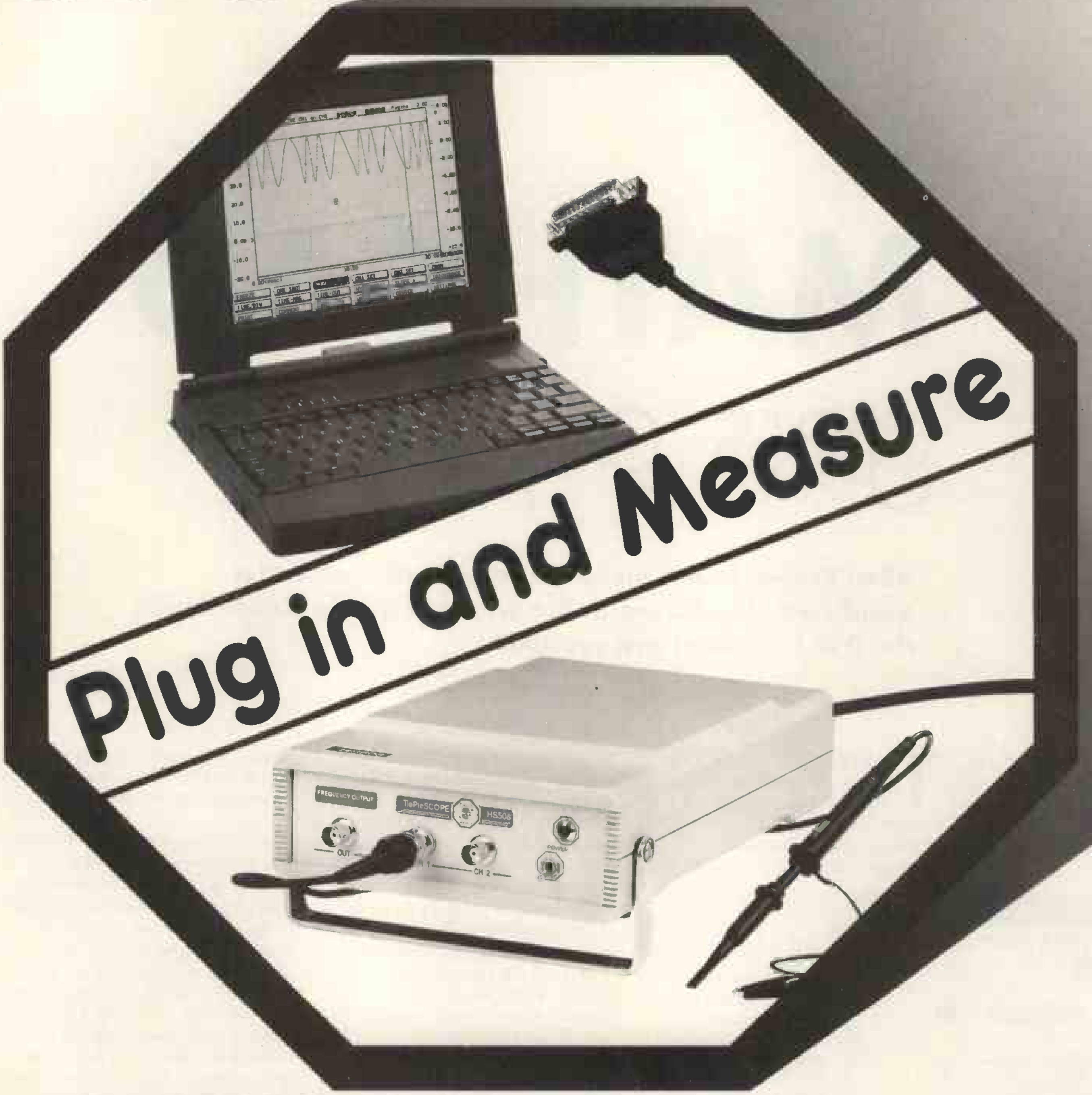


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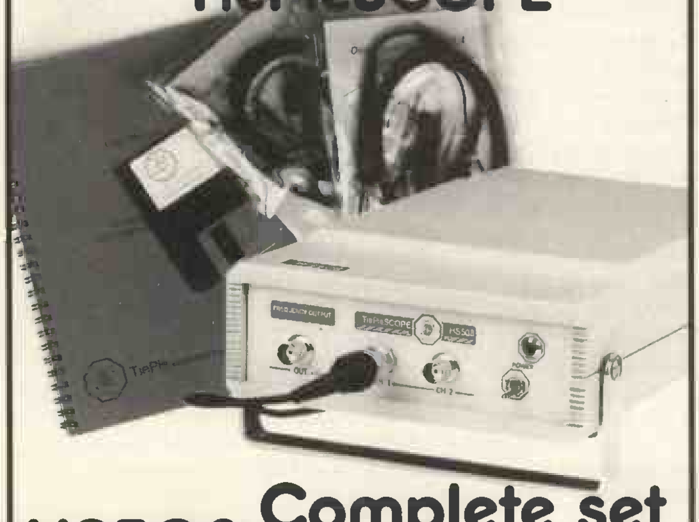

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# MATHCAD gets a plus

**Allen Brown found the latest release of *Mathcad* a significant step forward. But was he equally pleased with the DSP Function-Pack soption?**

The well known maths software package *Mathcad* has recently been upgraded to version 6, and renamed *Mathcad Plus 6*. Version 6 is an enhancement of version 5 with some very interesting developments – not least the ability to execute program loops. In addition, *Mathcad* is one of the few 32-bit programs that runs under Windows 3.1. It also runs under Windows 95, albeit with a patch program.

The new version retains the basic initial design of using the pc screen as a scratch pad where the user is able to express and solve equations, draw graphs and construct tables. The package makes most of these tasks relatively easy. It is also able to perform analytical operations via the Maple Library. Although this feature was introduced a few version ago, it still remains as a powerful and useful feature of the package.

Newcomers to the software will probably find the immense array of functions and options bewildering. It is not a software package that the new user will learn overnight. It will probably take you several weeks to master it. However there are a number of design features, in addition to the normal help menus, that will help you gain an insight into the package's potential.

Because of its very general nature, *Mathcad* can be used by anyone who has a need to manipulate numbers or perform modelling tasks. To add to its appeal, the package can be bought with a variety of 'function packs' that contain functions developed for specific applications. In this review I will be looking at the signal-processing function pack.

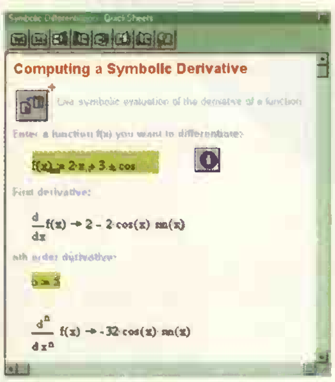



*Fig. 1. A new feature of Mathcad Plus 6 is the inclusion of Quicksheets. Each one gives an example of how a Mathcad function can be used. This example shows how the first derivative can be obtained.*

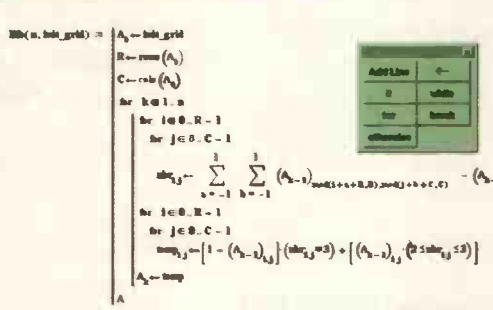



*Fig. 2. Program loops and structures can be included in a Mathcad object; they have similar constructs as found in traditional programming languages.*

## QuickSheets for faster learning

One innovation introduced into *Mathcad Plus 6* is the concept of the QuickSheet. It is well known that one of the most effective methods of learning is through the use of examples and this is the principle employed in the QuickSheet facility. The user is provided with a table of contents, each entry leads to a further menu and eventually to an example of how a function is used. An example is shown in **Fig. 1** illustrating how the derivative function is operated.

Examples, showing how the various functions and operations work within *Mathcad*, can be viewed. More importantly, many of the examples are easy to follow.

## Programming methods

One of the problems with previous versions of this package was its inability to perform programs with conditional loops. In version 6, this deficiency is remedied by the introduction of a set of programming functions. These include 'for', 'while', 'if', 'break' and 'otherwise'. They are evoked from the programming constructs palette and an example of how the program is

## Signal processing function pack

The makers of *Mathcad* provide many special purpose function packs. One of particular interest to electronics engineers is dedicated to signal processing.

Each pack is a collection of special functions compiled into a dynamic linked library (DLL) that is accessed by the software. Although the SIGNAL-PROCESSING PACK comes with a small booklet which lists the functions, when installed a Signal Processing Electronic Book is loaded on the PC.

All information regarding the operation of the pack is accessible by entering the 'Electronic Book'. This provides numerous examples of how each function in the pack can be used. Having an electronic book dispenses with the need for yet another paper manual. If you want copies of specific pages from the electronic book, you merely print them as required. The only irritating aspect of the *Mathcad*'s electronics books is the pale blue font which makes them difficult to read.

Sixty four functions in the signal-processing pack fall into the following categories,

- Transforms
- Spectral analysis
- Time series analysis
- Spectral analysis
- Digital filtering.

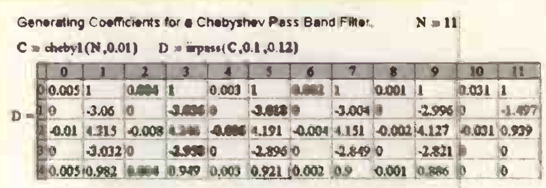
Fast Fourier transforms and inverse FFTs already form part of the *Mathcad* package and the signal-processing pack have several spectral analysis functions that complement these. These include cepstrum, for finding harmonics in spectra, costr, sintr and cosine and sine Fourier transforms. There is also a discrete Hartly transform which is similar to the Fourier Transform except it does not use complex maths.

A number of window functions are available for shaping input data before any spectral analysis is performed on it. These are important for resolving small spectral peaks lying in the shallows of much larger ones.

Although many of the functions are useful one has the feeling that they could be easier to use and this is particularly true of the digital filter functions. These operate in a rather strange manner. Normally the user will known what the stop band attenuation is, the pass band ripple and the transition frequency width between the bands. Filter design software should then provide the number of coefficients (number of filter taps) and value of the coefficients.

In the signal-processing pack, the user is expected to provide the number of coefficients. Not only that but once the coefficients have been generated there is no direct, easy method for displaying the transfer function of the filter. The user needs this to determine whether the filter satisfies the specifications. To add to the confusion, the coefficients generated are larger than unity. As a result, they do not lend themselves to easy quantisation for implementing on fixed point DSP chips (see Fig. 7). This makes me wonder whether an engineer was consulted when the signal-processing pack was designed.

To make matters worse, there are no functions for designing elliptical filters. This is a nuisance since this type of filter is used frequently for sharp cut-off filters with minimal tap count.



Generating Coefficients for a Chebyshev Pass Band Filter. N = 11

C = cheby1(N,0.01) D = iirpass(C,0.1,0.12)

D =

| | 0 | 1 | 2 | 3 | 4 | 5 | 6 | 7 | 8 | 9 | 10 | 11 |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 0 | 0.005 | 1 | [illegible] | 1 | 0.003 | 1 | [illegible] | 1 | 0.001 | 1 | 0.031 | 1 |
| 1 | 0 | -3.06 | 0 | [illegible] | 0 | [illegible] | 0 | -3.004 | 0 | -2.996 | 0 | -1.497 |
| 2 | -0.01 | 4.315 | -0.008 | [illegible] | [illegible] | 4.191 | -0.004 | 4.151 | -0.002 | 4.127 | -0.031 | 0.939 |
| 3 | 0 | -3.032 | 0 | [illegible] | 0 | -2.896 | 0 | -2.849 | 0 | -2.821 | [illegible] | 0 |
| 4 | 0.005 | 0.982 | [illegible] | 0.949 | 0.003 | 0.921 | 0.002 | 0.9 | 0.001 | 0.886 | 0 | 0 |

*Fig. 7. Mathcad's signal-processing function pack has functions for designing digital filters. However when the coefficients of an infinite impulse response (IIR) filter are displayed, the format is not immediately useful to a digital filter designer.*

constructed in shown in **Fig. 2**.

Also shown, as an insert, is the programming palette. The while statement is very useful for iterating a loop until a condition is met at which point the program stops executing. As an extra safeguard the break function can be included to ensure that the loop terminates in the event of the normal exit condition not been satisfied.

As with other programming languages, loops can be nested within other loops to give several variable dependencies. It does however take a while to get used to the programming technique. As with all programming languages, it is necessary to think of your problem in terms of the language constructs. But first you must be very familiar with the operation of the constructs.

### Handling non-linear differential equations

One of the exciting aspects of *Mathcad Plus 6* is the facility for solving nonlinear differential equations numerically. Although *Mathcad* has been a very effective tool for modelling linear processes, most real world problems are nonlinear. Man has a history of trying to impose linearity upon nature which is intrinsically nonlinear.

The principal function for tackling nonlinear differential equations is 'rkfixed'. This function evokes the fourth-order Runge-Kutta algorithm which can be used to solve any order, and even systems, of nonlinear differential equations.

**Figure 3** shows how the function models relaxation oscillations in a semiconductor laser. There are other algorithms available in the package and their application depends on the nature of the nonlinear differential equations to solve, and on the accuracy of the required solutions. In general, systems of nonlinear differential equations can be classified as 'smooth', 'slowly varying' and 'stiff'. The function for tackling smooth systems is bulstoer, after Bulirsch-Stoer, and for slowly varying is Rkadapt, a modified version of the Runge-Kutta algorithm. Stiff systems can be solved by using 'stiffb', again after Bulirsch-Stoer, or 'stiffr' after Rosenbrock.

Solving nonlinear differential equations can be achieved provided the user has knowledge of the initial conditions. If this information is not available, it may be possible to use the 'sbvla' or the 'bvalfit' functions which employ 'boundary value' techniques.

Given partial knowledge of a

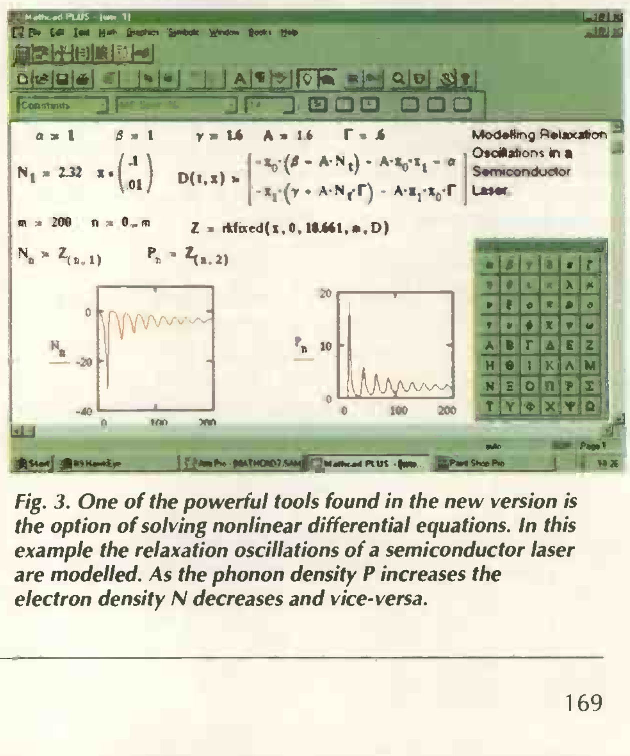



*Fig. 3. One of the powerful tools found in the new version is the option of solving nonlinear differential equations. In this example the relaxation oscillations of a semiconductor laser are modelled. As the phonon density P increases the electron density N decreases and vice-versa.*

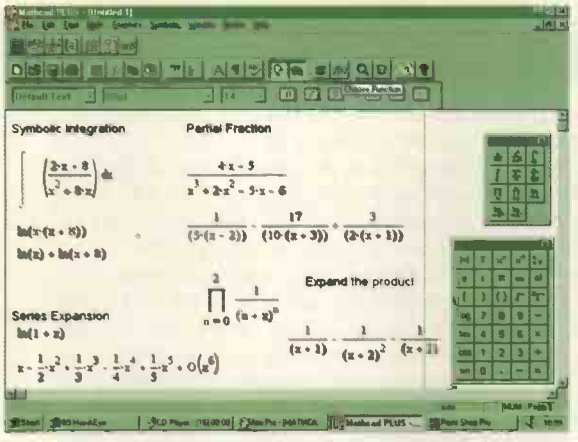



*Fig. 4. Mathcad uses a symbolic calculator to perform analytical evaluations of equations. This figure shows a few examples of what can be done with the calculator.*

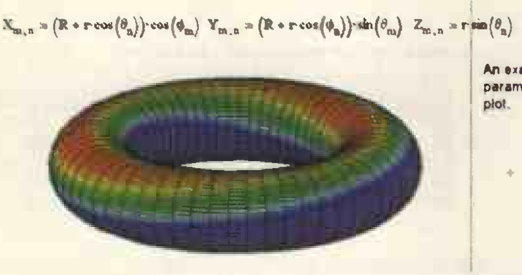



*Fig. 5. Customary 3D plots can be generated with Mathcad, but their ease of construction could be improved. Not only that but object rotation to obtain a different perspective is not at all convenient.*

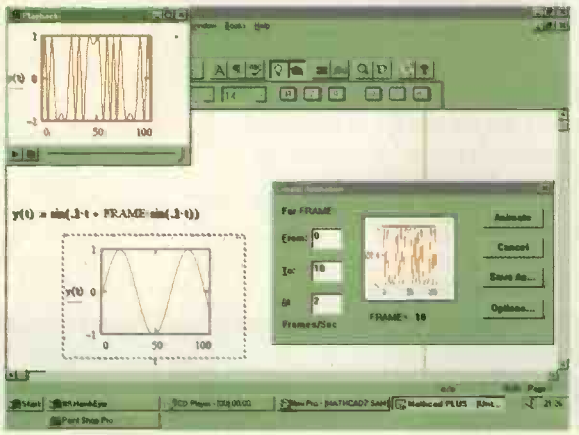


*Fig. 6. For imaginative users, Mathcad now has the provision for generating video clips which can show how a plotted function changes as one of the variables is allowed to change from frame to frame.*

problem these functions allow you to determine appropriate initial conditions that can be used in the application of functions just mentioned. Collectively these functions form quite an impressive arsenal with which to solve differential equations.

## Symbolic calculations

*Mathcad 6 Plus* retains the ability to solve equations symbolically. It uses the Maple engine from the Canadian company Waterloo. Its performance is usually quite impressive. When the pc is configured for 32-bit disk access, the speed of execution of the 'symbolic calculator' is surprisingly fast. Its popularity in education goes without saying. Judging by the falling standards in mathematical skills by students entering universities, this aspect of the package will be seized on with great enthusiasm.

The software is able to perform a whole array of symbolic processing operations including integration, polynomial expansion, simplification, partial fraction expansions and many other general symbolic algebraic functions. However, sometimes the symbolic expansions do not perform even on simple expressions.

I had to conclude that it mostly provides an answer but not always. **Figure 4** illustrated a number of symbolic operations that were successfully performed by *Mathcad*.

## Graphing and plotting

Graphing options offered by *Mathcad* are not too dissimilar to those offered in general purpose graphics packages. These days, high quality data plotting software is freely available. As a result, the package has a lot in common with most graphics packages – contour plots, polar plots, three dimensional solid modelling with rotation and bar charts.

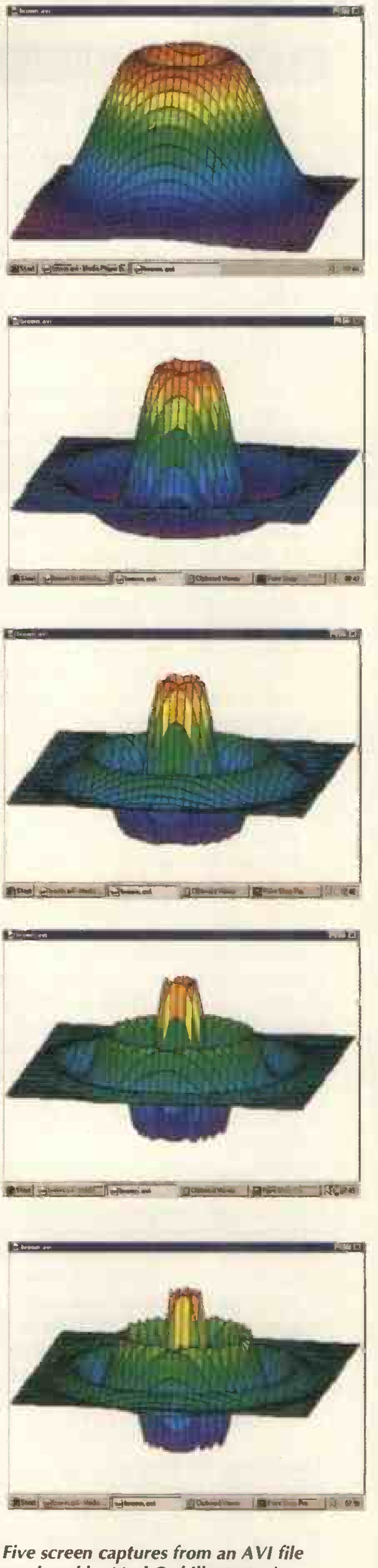
However, it should be mentioned that producing a surface plot is not as easy as I would like. It is still necessary to construct a matrix beforehand, which is not the most intuitive method of generating a three-dimensional plot.

On the whole, facilities for generating and manipulating three-dimensional plots are rather clumsy. An example of a colour coded parametric plot is shown in **Fig. 5**. There is also a provision for importing images, however this facility is by no means free of problems as it failed to function properly. Not all bit-mapped images are recognised.

An interesting addition in the new version is the option to generate animation plots. This feature employs the Microsoft multimedia facilities for video clips, .AVI files.

**Figure 6** shows how easy it is to create an animation. A function is defined with a variable called FRAME. Once the animation has been evoked a dialogue box appears, on the right of Fig. 6, where the range of FRAME is defined. The user also chooses the number of frames per second.

Once the plotting area has been selected, the animation process begins.



*Five screen captures from an AVI file produced by MathCad illustrate the package's animation capability.*

The frames are then compiled into .AVI video, Fig. 6, top left, and played at will. This could be useful for dynamic solid modelling to illustrate vibration modes in something like an optical fibre.

The software is quite impressive when solving simultaneous equations using the lsolve function. Solving simultaneous equations boils down to solving the matrix equation,

$$x=A^{-1}v$$

For example, if this represents 500 simultaneous equations with 500 unknowns, $v$, and 500 solutions, $x$, *Mathcad* takes just over 14s to solve it when running on a 120MHz *Pentium*, representing 28ms for each unknown element in $x$. Bearing in mind that the matrix $A$ contains a quarter of a million floating point elements, it can be said that the pc has come of age for computational operations.

Solving simultaneous equations is a frequent requirement for statistical work. In the new version, there is also a substantial range of statistical functions. Should the need arise to import or export data to and from Windows applications, the package has a facility for creating dynamic data exchange, DDE, interfaces. This is a useful component if the user is interested in using *Mathcad* to process data direct from a data acquisition card in real-time.

## Summing up

Looking at all the features that *Mathcad Plus 6* has to offer, one wonders what the next version will have in store. There could be improved 3D plotting. It still remains a very well designed software product that will continue to find appeal amongst many different kinds of users. Whether you want to perform simple calculations or quite involved nonlinear modelling the package is a very approachable product.

The user guide has passed through several iterations and the volume for the new version is an invaluable document for any user – experienced or otherwise. The 694 page guide contains many examples and is very readable. It certainly complements the help file.

My enthusiasm for the signal-processing function pack however is not so forthcoming. Although it has many useful functions, many lack functionality and are basically unwieldy and difficult to use. These days there are many software packages for designing and analysing digital filters. Unfortunately the signal-processing function pack falls far short of most of them. ■

## Availability

Adept Scientific Micro Systems Ltd, 6 Business Centre West, Avenue One, Letchworth SG6 2HB, Tel. 01462-480055, fax 01462-480213. Price – *Mathcad Plus 6*, £395 excluding VAT, *Signal Processing Function Pack*, £195 excluding VAT. Educational discounts are available in some cases.

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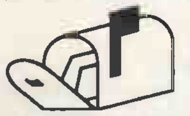

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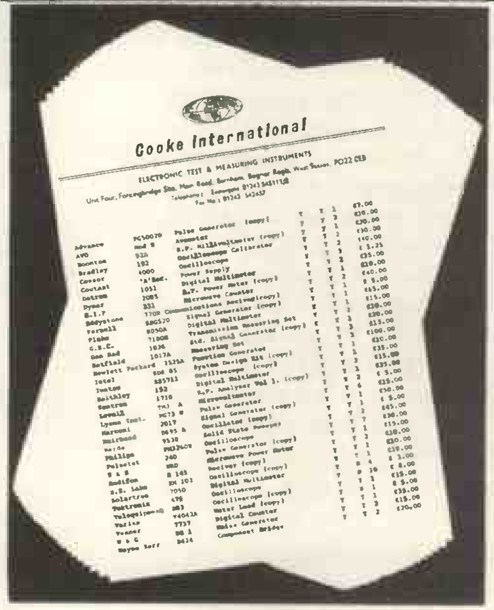


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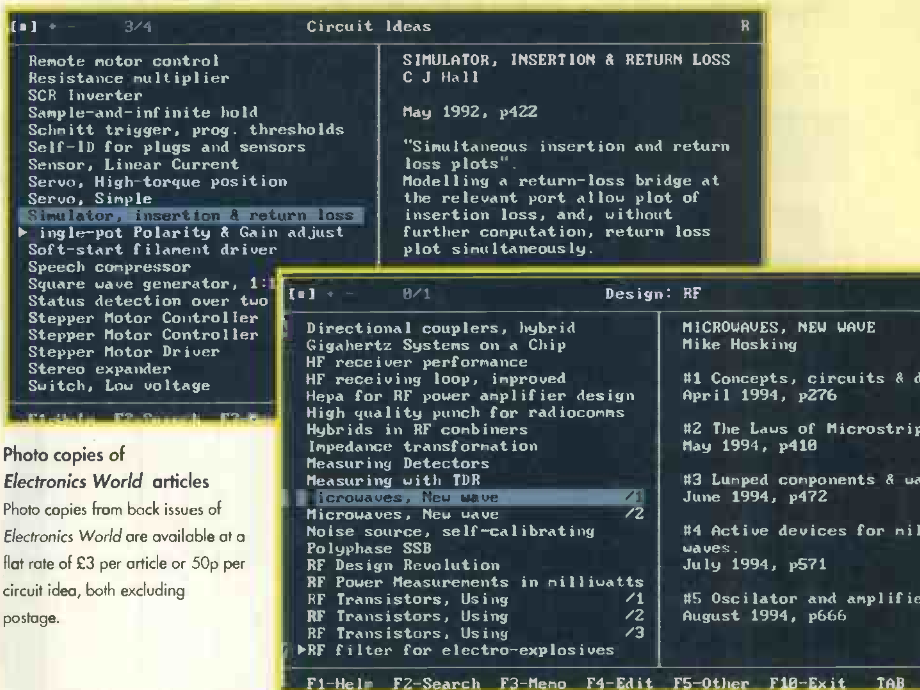


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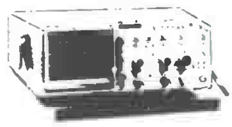

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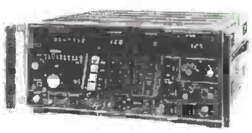

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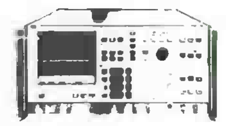

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