Electronics World's renowned news section starts on page 4

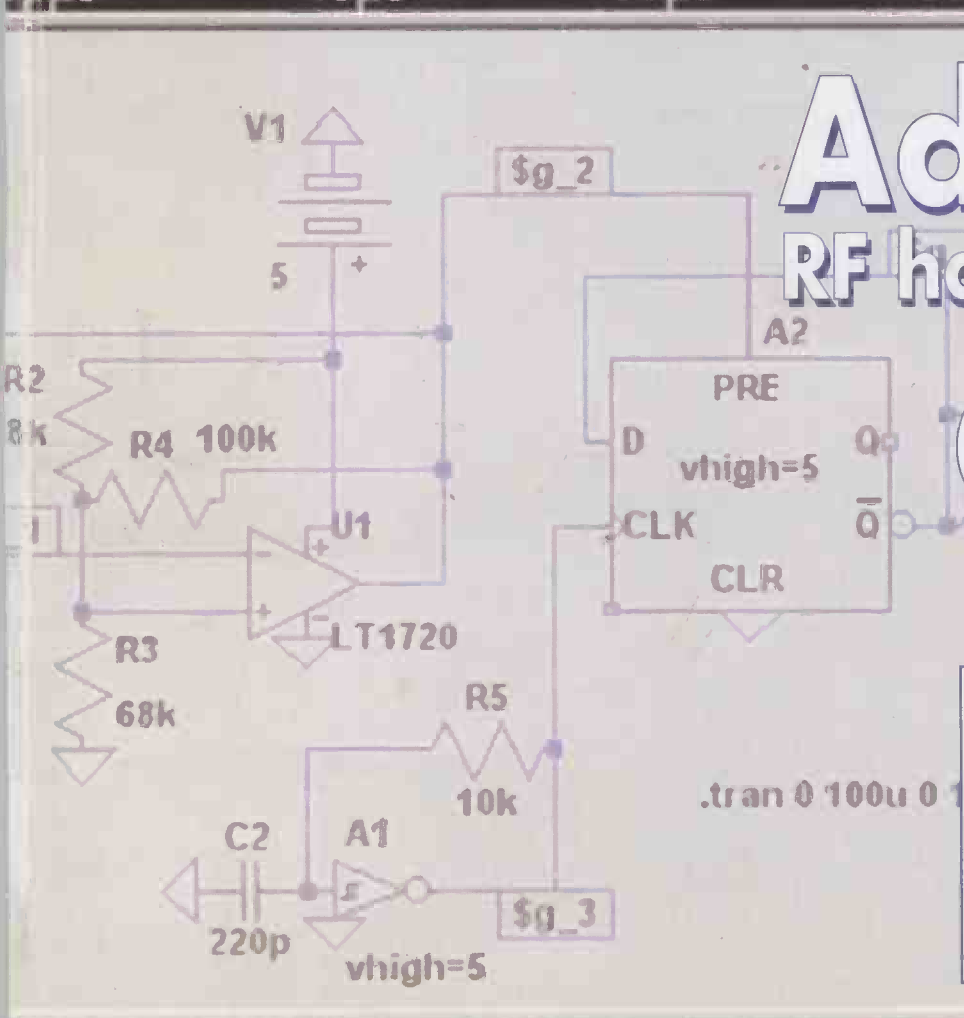
ELECTRONICS WORLD

NOVEMBER 2004 £3.25

the SPICE trial

Advanced RF harmonic theory

Class-A imagineering



Circuit Ideas

- 6 ¼ bit DAC using four output pins
- Long delay timer using one 555
- Audio level and peak metering
- High voltage, simple and fast inverter
- Electronically tuneable active only oscillator

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H.P. 4192A L.F. Imp. Analyser (13MHz)	£4000
Hewlett Packard 4193A Vector Impedance Meter (4-110MHz)	£2900
Hewlett Packard 4278A 1kHz/1MHz Capacitance Meter	£3500
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Marconi 6310 – Prog'ble Sweep gen. (2 to 20GHz) – new	£2500
Marconi 6311 Prog'ble sig. gen. (10MHz to 20GHz)	£2995
Marconi 6313 Prog'ble sig. gen. (10MHz to 26.5GHz)	£3750
R&S SMG (0.1-1GHz) Sig. Generator (opts B1+2)	£2500
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OSCILLOSCOPES

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Agilent (HP) 54600B 100MHz 2 channel digital	£800
Agilent (HP) 54602B 150MHz 4(2+2) channnel digital	£1250
Agilent (HP) 54616B 500MHz 2 channel digital	£1750
Agilent (HP) 54616C 500MHz 2 channel colour	£2750
Agilent (HP) 54645D DSO/Logic Analyser 100MHz 2 channel	£2750
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Hewlett Packard 54600A - 100MHz – 2 channel	£675
Hewlett Packard 54810A 'Infinium' 500MHz 2ch	£2995
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Philips PM3350A - 60MHz >-> 100MS/S 2 channel	£750
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Tektronix TDS 360 200MHz - 2 channel digital	£1200
Tektronix TDS 420A 200MHz - 4 channel digital	£1800
Tektronix TDS 540B 500MHz - 4 channel digital	£2500
Tektronix TDS 640A 500MHz - 4 channel digital	£2700
Tektronix TDS 744A 500MHZ - 4 channel digital	£4250
Tektronix TDS 754C 500MHz - 4 channel digital	£4500

SPECTRUM ANALYSERS

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Agilent (HP) 3562A Dynamic Signal Analyser - Dual Channel	£3500
Agilent (HP) 35665A (opt. 1D1) Dual ch. Dynamic Signal Analyser	£3750
Agilent (HP) 8560A (opt 002 - Tracking Gen.) 50Hz -2.9GHz	£5000
Agilent (HP) 8560E (opt 002 - Tracking Gen) 30Hz - 2.9GHz	£8995
Agilent (HP) 8593E (opt 41/105/130/151/160) 9kHz - 22GHz	£12000
Agilent (HP) 8594E (opt 41/101/105/130) 9kHz - 2.9GHz	£4250
Agilent (HP) 8590A (opt H18) 10kHz - 1.8GHz	£2500
Agilent (HP) 8596E (opts 41/101/105/130) 9kHz - 12.8 GHz	£8000
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Tek 496 (9KHz-1.8GHz)	£2500

Radio Communications Test Sets

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Agilent (HP) E8285A CDMA Mobile Station T/Set	£8500
Anritsu MT8802A (opt 7) Radio Comms Analyser (300kHz-3GHz)	£8500
Hewlett Packard 8920B (opts 1,4,7,11,12)	£6750
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MISCELLANEOUS

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Agilent (HP) 8970B (opt 020) Noise Figure Meter	£3950
Agilent (HP) EPM 441A (opt 2) single ch. Power Meter	£1300
Agilent (HP) 6812A AC Power Source 750VA	£2950
Agilent (HP) 6063B DC Electronic Load 250W (0-10A)	£1000
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EIP 545 Microwave Frequency Counter (18GHz)	£1000
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EIP 575 Source Locking Freq.Counter (18GHz)	£1200
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Hewlett Packard 3784A - Digital Transmission Analyser	£2950
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Hewlett Packard 4274A LCR Meter	£1750
Hewlett Packard 4275A LCR Meter	£2750
Hewlett Packard 4276A LCZ Meter (100MHz-20KHz)	£1400
Hewlett Packard 5342A Microwave Freq.Counter (18GHz)	£850
Hewlett Packard 5385A - 1 GHz Frequency counter	£495
Hewlett Packard 8350B - Sweep Generator Mainframe	£1500
Hewlett Packard 8642A – high performance R/F synthesiser (0.1-1050MHz)	£2500
Hewlett Packard 8901B - Modulation Analyser	£1750
Hewlett Packard 8903A, B and E - Distortion Analyser	from £1000
Hewlett Packard 11729B/C Carrier Noise Test Set	from £2500
Hewlett Packard 85024A High Frequency Probe	£1000
Hewlett Packard 6032A Power Supply (0-60V)-(0-50A)	£2000
Hewlett Packard 5351B Microwave Freq. Counter (26.5GHz)	£2750
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IFR (Marconi) 2051 (opt 1) 10kHz-2.7GHz Sig. Gen.	£5000
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Tektronix AWG 2021 Arbitrary Waveform Gen. (10Hz-250MHz) 2 ch.	£2400
Wayne Kerr 3245 - Precision Inductance Analyser	£1750
Bias unit 3220 and 3225L Cal.Coil available if required.	(P.O.A)
W&G PCM-4 PCM Channel measuring set	£3750
W&G PFJ-8 Error & Jitter Test Set	£6500

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12	Sockets	12LH/MS	@	£39.62 ea. Nett

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12	Sockets	12RH/MS	@	£39.62 ea. Nett

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EDITOR
Phil Reed
p.reed@highburybiz.com

CONSULTANT
Ian Hickman
www.ianhickman.org.uk

CONTRIBUTING EDITOR
Martin Eccles

PRODUCTION EDITOR/DESIGNER
Jane Massey
J.Massey@highburybiz.com

EDITORIAL E-MAILS
EWeditor@highburybiz.com

EDITORIAL ADMINISTRATION
Caroline Fisher
01322 611274
EWadmin@highburybiz.com

GROUP SALES EXECUTIVE
Steve Morley
01322 611254
s.morley@highburybiz.com

PRODUCTION EXECUTIVE
Dean Turner
d.turner@highburybiz.com
01322 611206

CLASSIFIED FAX
01322 616376

MANAGING EDITOR
Bill Evett

PUBLISHING DIRECTOR
Tony Greville

BUSINESS SECTOR MANAGER
Mark Skelton

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SUBSCRIPTION QUERIES

Tel (0) 1353 654431

A late developer

RFID is surely a technology that is about to explode and proliferate into every sphere of modern life. I have been keeping a weather eye on this topic for about fifteen years, since involved, while with Plessey/Siemens, in a study into the feasibility of using Radio Frequency IDentification tags to keep track of children in theme parks. A practical problem - several get lost every day and one or two a year get lost permanently - it is a difficult application for the technology, due to the range required. In retailing, a tag is only required to operate over a short range, a few metres at most.

One of the problems holding back wide scale application of RFID tags has been that of standardisation. It would be a great disadvantage if Europe, the US and Asia went their different ways, with mutually incompatible systems. Fortunately international co-operation is on its way, driven by the needs of industry and commerce. For keeping track of pallets for the transfer of goods, EPCglobal Inc. is developing the proposed the EPCglobalG2 standard. This uses a 96 bit format, and tags using this format are already available, with read and write operating ranges up to a few metres, while operation at UHF permits multiple tags to be read/written per second without collisions.

I don't have details to hand of the allocation of the 96 bits to headers, serial numbers etc. but 2^{96} is a seriously large number, about $7.922816244 \times 10^{28}$! Major concerns, both commercial and governmental, are mandating the use of RFID technology, with Wal-Mart, Tesco, US D.O.D and others requiring its use in their supply chain and logistics systems as early as 2005. And in the EU, new regulations will require the use of tags to keep track of individual animals such as beef herds etc. with the system already proven by pilot trial schemes. Tags complying with ISO 11784/85 were placed subcutaneously in each animal, allowing it to be tracked at various points during its lifetime, and after. Knowing where the beef the housewife buys comes from, should boost her confidence - and sales - in the view of past catastrophes such as scrapie, BSE, foot and mouth disease, etc.

Another burgeoning RFID application is smart tickets, which can be loaded with credit like a phone card, and debited automatically, in accordance with the journeys undertaken by the holder. Card readers on the various forms of transport could simplify matters by enabling a traveller to make a journey involving both buses and trains, without the need for a through-ticket or separate tickets for each stage. Already trialled in Perth, Australia, it is shortly to be tried out on the Seattle mass transit system.

Other applications abound, and Jo(e) Public will have to get used to tags appearing everywhere, whether he or she likes it or not. For instance, tags in clothes may cause the washing machine to inform you of the right programme to use for each, while tags on food packs may enable the smart fridge to warn you when you are running low on butter, or eggs, or milk... In time it will become as difficult to buy a 'dumb' fridge or washing machine as it is now, for those with a 'keep it simple - less to go wrong' philosophy, to buy a car without central locking.

Much more on RFID tag technology and applications can be found in the Philips publication *ON THE MOVE, Identification News*, Volume 6 Issue 3, September 2004. *Philips' World News*, Volume 13 Number 3, September 2004 also covers the topic, with other useful and interesting articles such as Bluetooth/WLAN intercommunication, hands-free car phones, class D amplifiers and electronic ink, among others.

Ian Hickman

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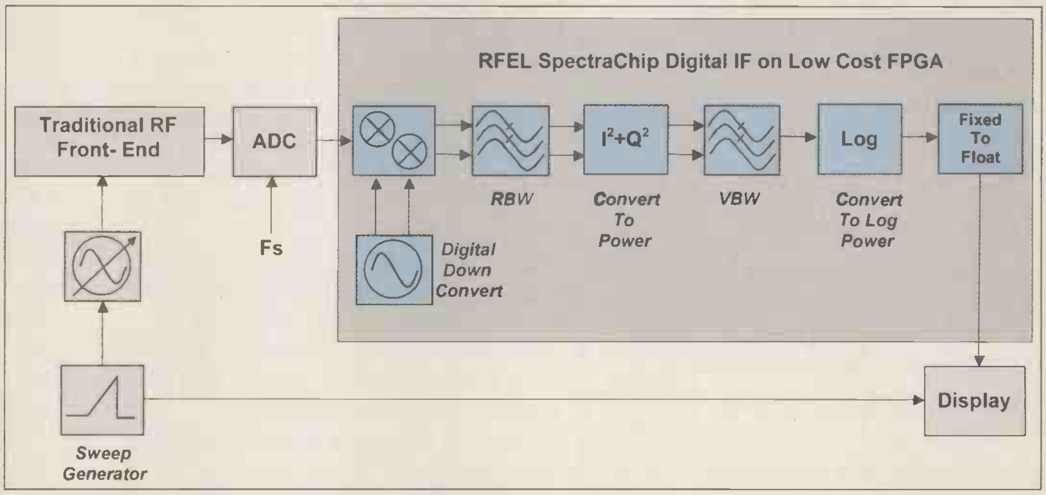
UPDATE

Digital spectrum analyser chip could change test gear



Isle of Wight-based RF Engines has produced a digital intermediate frequency (IF) chip for use in spectrum analysers.

Called 'SpectraChip IF', it "is a digital replacement for analogue IF filtering used in current spectrum analysers" said RFEL, which specialises in DSP firmware for FPGA programmable logic. "It provides a digital implementation of all the standard features, such as resolution bandwidth filtering, video bandwidth filtering and conversion to log power."

Resolution bandwidth is 10Hz to 3MHz in 1-3-10- steps and video bandwidth is 1Hz to 1MHz in 1-10- steps.

Test equipment makers can license SpectraChip IF from RFEL as an off-the-shelf design, although the firm has to customise interfaces for individual customers. "The standard design will fit comfortably into the Xilinx Spartan3 1000 FPGA device costing less than $100," said RFEL. "It consumes a mere 0.5W in operation, or significantly less in the power save mode."

In this form, the FPGA expects a digitised IF signal centred at 21.4MHz with a signal bandwidth of up to 10MHz - although other IF frequencies and signal bandwidths can be accommodated.

In addition to the FPGA, spectrum analyser designers will also have to include a fast analogue to digital converter (ADC). Together "we estimate these will cost 50 per cent of the analogue equivalent; based on the limited information we can get from our customers", said sales executive Simon Underhay.

Dynamic range is restricted to around 85dB by available ADCs. SpectraChip can "match any dynamic range the customer requires", said Underhay.

The product has already been licensed by one test gear firm, with another expected to sign soon, said Underhay.

www.rfel.com

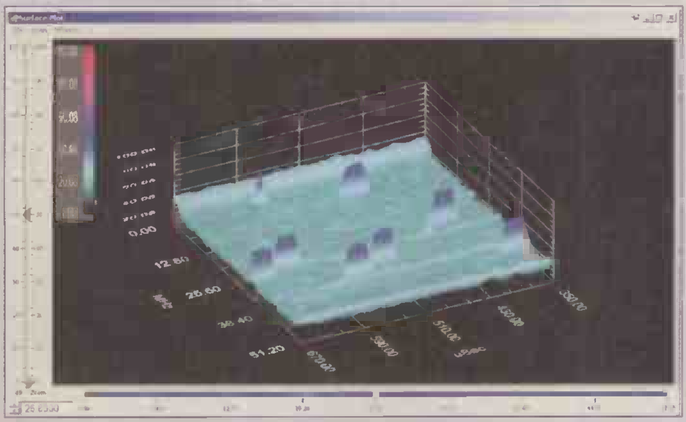
'Free' real-time spectrum analysers

While RF Engine's SpectraChip IF is a replacement for existing spectrum analyser IF strips, adding an ADC and an FPGA raises the prospect of real time spectrum analysis with no extra hardware costs.

Traditional spectrum analysers sweep a filter through the spectrum to develop their amplitude verses frequency display, missing activities at frequencies either side of the instantaneous sweep frequency.

Real-time spectrum analysers continuously pull in the entire spectrum of interest, missing nothing, and are extremely expensive.

According to RFEL, with the right firmware a Xilinx Spartan3 1000 FPGA could take a 5MHz band and measure the instantaneous amplitude in 4,000 separate frequency bands, known as bins, within this.



A 'waterfall' plot of a Bluetooth signal captured in real time by RFEL's SpectraChip MHz' prototype real-time spectrum analyser chip.

RFEL is working on such firmware, for a product to be know as 'SpectraChip MHz'. "Standard Gaussian shape filters can be used to produce a standard spectrum analyser effect, or higher specification, flat top, low transition band filters can be used to achieve power measurement accuracy and good frequency isolation," said RFEL.

As no human can possibly use 4,000 simultaneous channels of data at 5MHz, SpectraChip MHz includes a trigger facility to record waveforms at all frequencies before and after a trigger event at a particular frequency. See the blue diagram for one way of displaying this data. Averaged displays are also possible.

Beyond this, SpectraChip GHz is under development. In a very large fast FPGA, with a 1Gsample/s ADC, this will capture 16,000 bins at 500MHz.

Nanotech gets funding boost

The Government has released the first tranche of its £90m funding for nanotechnology projects in the UK.

The Department of Trade and Industry has made £15m available to 25 projects, spanning anti-corrosion coatings to electronics and printing.

"Nanotechnology is an important and exciting emerging technology, one that has the capacity to improve daily life for us all. It is about designing new products and improving existing ones by making things much smaller, faster, stronger, or more energy efficient," explained DTI minister Nigel Griffiths.

The Government is also making £70m available for nanotechnology through the Research Councils and has set up the UK Micro and Nanotechnology Network to support firms moving into this area. "We want to help organisations turn ideas into reality, helping create jobs and prosperity for companies in the UK," Griffiths added.

Grants awarded under the scheme total up to 50 per cent of a project's total value.

Some high profile projects have won funding under the scheme, including Trikon Technology from South Wales. It has received over £1.5m to help develop an ion beam deposition tool for creating magneto-resistive RAM.

Meanwhile Teraview from Cambridge has over £1m from the fund to continue work on its Terahertz imaging systems. These could revolutionise the detection of cancers.

Two firms – Insight Faraday from Runcorn and Mast Carbon from Guildford, are working on fuel cells.

Qinetiq in Malvern has won funding for molecular gas sensors and for narrow bandgap spintronic devices.

Other projects include improved power electronics contact technology by Dynex Semiconductor; low temperature polymer embossing by Applied Microengineering; and nanotube and nanowire production at Thomas Swan.

A £3m grant has also been given to INEX, a microsystems and nanotechnology facility for industry based in Newcastle.

Facility opens for systems engineers

Loughborough University has opened its £60m Systems Engineering Innovation Centre (SEIC), which combines training and research for systems engineers.

The centre is already committed to training 1,000 engineers from BAE Systems by 2005.

Opening the centre, Minister for Science and Technology Lord Sainsbury said: "Innovation - turning new technologies and ideas into commercially successful products - is the key to business success. Partnership working between academia and industry is vital if we are to capitalise on our world-class science and technology base and pull through knowledge from the lab to the business bottom line."

Joining the University and BAE is the East Midlands Development Agency, making SEIC a unique collaboration between academia, industry and a regional authority.

UK firms working in areas such as aerospace, automotive, food and drink processing, medical technologies and clothing and textiles will benefit, said the centre.

Demand for training in systems engineering is expected to be high. The SEIC highlighted a Department for Education and Skills report which said the UK would need 270,000 engineering professionals by 2010. Up to 45,000 engineers will need to be trained each year to meet that demand.

At the SEIC, some 30 jobs will be created in the first two years, with up to 45 in the longer term.

Support network for Yorkshire firms

Yorkshire electronics firms are getting their own support group that hopes to promote growth and job creation in the Yorkshire and Humber region.

Electronics Yorkshire is a not-for-profit limited company that will establish a network of experts able to provide support to new and existing companies.

"This is a serious opportunity to bring real and tangible benefits to the electronics industry in our region," said Arthur Gillatt, chief executive of Electronics Yorkshire.

The group will also take control of the Training and Technology Centre in Bradford, moving it to Leeds in the process.

"The Network and the new Technology and Training Centre are the first steps in ensuring Yorkshire and the Humber's electronics companies have a brighter future," Gillatt added.

Electronics Yorkshire is funded by Yorkshire Forward and the Learning and Skills Council. Firms wishing to join the group – free of charge – contact Dave Williams on 0113 274 4270.

Computer becomes a music buff

Researchers at the universities of Southampton and Vienna have demonstrated that the complex and individual performance styles of concert pianists can be modelled in unique 'performance alphabets', providing a method of recognising their performance styles by computer.

"Concert pianist Glenn Gould had a unique and instantly recognisable performance style, for which he is rightly renowned," said Southampton. "Indeed, the extent to which pianists such as Gould, Horowitz and Uchida have a discernibly individual style of playing is recognised not just by classical music aficionados, who can hear the differences, but also by computers, which can analyse the differences and model them."

Vienna researchers measured tempo and volume when the same Mozart sonatas were played by Glenn Gould, Daniel Barenboim, Andras Schiff, Mitsuko Uchida, Roland Batik, and Maria Joao Pires.

Professor John Shawe-Taylor of the University of Southampton School of Electronics and Computer Science led the analysis.

"Different players have different ways of building tension or expression in the music," said Shawe-Taylor, "and they [Vienna researchers] represent this raw data for every note and progression of the music as a trajectory, which can be represented visually in tempo-loudness space as a 'performance worm'."

The novel analysis techniques applied to the performance worm data at the University of Southampton were able to distinguish the different performers based on a relatively small sample of their performances.

On a speculative note: "We are seeking common patterns across two different ways of looking at an event," Shawe-Taylor said. "On the one hand we have the musical score, and on the other, its interpretation by an individual concert pianist." If these two could be combined "then we might be able to generate aspects of a Horowitz performance of a piece that he had never actually played".

Lay it on thick

A technique to form layers of ceramic film at room temperature on printed circuit boards has been developed by Fujitsu Laboratories.

The process could be used to produce high quality passive components, particularly capacitors in microwave applications, on standard FR4 PCBs.

Along with Japan's National Institute of Advanced Industrial Science and Technology (AIST), Fujitsu developed two technologies; formation of low temperature ceramic films and multi-layering of those films.

To create the films, the developers used an aerosol deposition, with a gas spraying ceramic powder at high speed onto the board.

Cross-sectional view of a multi-layer capacitor on FR4

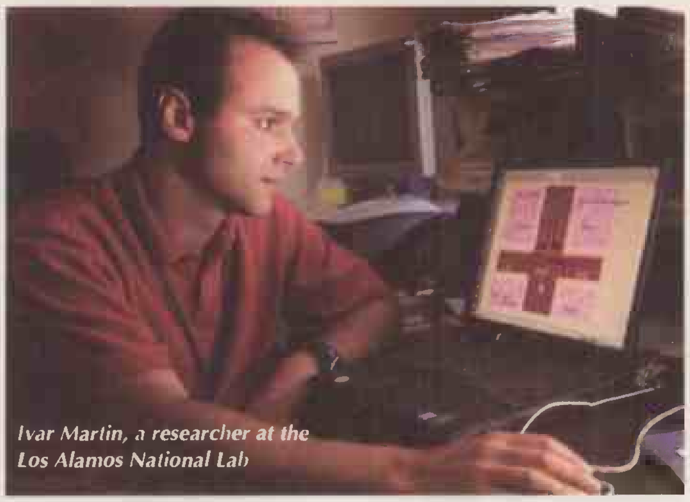
By using ceramics with "unstable surface characteristics", films can be created at room temperature, claimed Fujitsu.

Ceramic films have been created with dielectric constants of 400, some ten times higher than existing techniques which combine ceramic materials with the resin of the board.

Similar results are obtained by sputtering ceramic onto the boards, but this requires high temperatures up to 600°C, impractical for low cost FR4 boards.

Fujitsu said three layer devices made using the films have a capacitance per unit area of 300nF/cm^2.

Electrons get in a spin


Ivar Martin, a researcher at the Los Alamos National Lab

The spin of a single electron in a silicon transistor has been observed by researchers from the University of California.

Working at the Los Alamos National Lab in the US, the team found a way to convert the spin of a single electron in the gate oxide layer of a Fet into a charge. This charge could then be measured by the Fet.

"We believe this is a significant advance in the field of quantum physics. The more that the fields of science and engineering learn about the enigmatic physics of electron spin, the more we will be able to use that knowledge in the future to create nanoscale technologies like spin electronic and quantum computers, that are based on electron spin control," said Los Alamos scientist Ivar Martin.

Future electronic systems may rely less on the statistical maths of thousands of electrons that existing transistors and Fets rely on. Instead single electron memories and transistors may become commercially viable. Studying spin and decoherence is crucial to understanding how electrons behave.

Decoherence is a process undergone by sub-atomic particles such as electrons as they interact with other matter, thereby losing their wavelike properties. Electrons have long decoherence times compared to other particles.

Phone gets a hard drive

Samsung's SPH-V5400 is claimed to be the first mobile phone with an internal hard drive. The drive, which measures 25mm diagonally, can hold 1.5Gbyte of data, compared with a maximum of 100Mbyte in more conventional phones, said the firm. The phone is also equipped with a 56mm QVGA LCD and a mega-pixel camera as well as features including an MP3 player, electronic book and Korean-English/English-Korean dictionaries. Initially, the phone is only being sold in Korea.

Tiny atomic clock size of grain of rice

Scientists at the National Institute of Standards and Technology (NIST) in the US have built an atomic clock just 4.2mm high, the same size as a grain of rice.

The so-called 'chip-scale' atomic clock uses a combination of solid state laser technology, optics and micromachining.

NIST claimed the output frequency of the clock achieves a long-term stability of one part in 10^{10}, better than 10μs per day, while consuming just 75mW.

At the heart of the clock is a vapour cell. Conventional atomic clocks have a cavity defined by the microwaves used to excite the atoms, making the cells in the order of centimetres long.

NIST has used a technique called coherent population trapping. A vertical laser diode is modulated at half the ground state frequency of caesium atoms in the cavity. Modulation produces sidebands that are separated by the caesium resonance frequency.

Optics polarise the light before passing through the caesium-filled cavity. A photodiode detects the output frequency.

A small cell has the added advantage of not requiring much power to maintain a constant temperature of around 80°C in the cell.

CW transmission and detection at terahertz

Teraview, the Cambridge-based terahertz imaging company, has made a solid-state continuous-wave terahertz imaging system.

Previously, the firm has been imaging skin cancer cells using a pulsed lasers pumping a GaAs heterostructure quantum cascade laser (QCL). Output power is 1μW, with returns received by a synchronous detector based around a similar heterostructure.

Now the firm has a continuous wave version (see diagram) with two 800nm infra-red CW laser diodes pumping a modified GaAs QCL. The pumping diodes are tuned so their difference frequency is the terahertz output required.

"We have reduced cost by a factor of ten and size by a factor of eight or nine," said the company's CEO Dr Don Arnone. Power is also down from the pulsed version: "just shy of a microwatt", said Arnone, "but the [synchronous] detector operates in the order of attowatts".

Key to getting the system working has been compensating for the low power of the CW pumps compared with the previous pulsed source.

Final output power depends to a degree on a bias voltage applied to the QCL and re-engineering the hetrostructure to withstand more voltage has allowed sufficient output power to be recovered.

The photographs show a knife and a razor blade, both partially concealed from visible light, viewed by the CW terahertz imaging system.

Chill runs through high energy physics

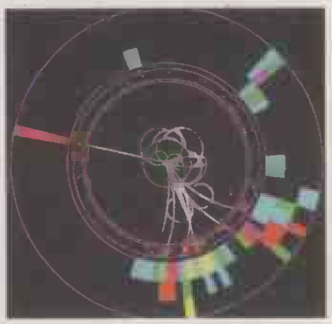
Superconducting electronics cooled to just two degrees above absolute zero will form the basis of the next generation of particle physics accelerator.

With experiments such as the Large Hadron Collider under construction at CERN, an international team of scientists has chosen the next technology, which could see contributions from UK firms.

The electron-positron linear collider will be designed to create impacts with an energy of up to 1TeV (10^{12} electron Volts). It would be able to investigate the Higgs boson (if it exists) and maybe give clues to the nature of dark matter.

To create such huge energies, scientists have chosen to use "cold" technology developed at Deutsches Elektronen-Synchrotron in Germany. (DESY)

"The superconducting technology uses L-band (1.3GHz) radio frequency power for accelerating the electron and positron beams in the two opposing linear accelerators that make up the collider," said George Kalmus, of the Rutherford Appleton Laboratory. By using superconducting niobium cavities in the 20km long colliders, power from the drive klystrons is coupled efficiently.



Simulated response of linear collider detector to supersymmetry displaying the characteristic signature of a lepton, a hadronic jet and missing transverse energy.

Credit: Norman Graf

The LHC at CERN is a circular proton-proton collider. "A linear collider is the logical next step to complement the discoveries that will be made at the LHC," said CERN director general Robert Aymar.

The other technology touted for the linear collider uses X-band structures at room temperature.

Mine hunters meet a robot deminer

Carson the robot, seen lurking in the background, was one of the demining technologies demonstrated to group of humanitarian mine removal organisations by Malvern-based research organisation QinetiQ.

It can be driven remotely up to a mine, place charges to burn out or blow up the mine, then reversing away.

The organisation representatives were in the UK to attend a course at Cranfield University and the QinetiQ visit was part of a de-mining technologies awareness day arranged by Cranfield.

"This is an annual event with Cranfield University. Clearance of mines is a huge concern in many areas of the world and we are always ready to discuss our humanitarian de-mining technologies with those who experience the problem first-hand," said QinetiQ project manager David Lewis.

Together, the organisations attending cover de-mining in Afghanistan, Azerbaijan, Bosnia and Herzegovina, Burundi, Cambodia, Iran, Iraq, Jordan, Kenya, Sri Lanka, Sudan and Vietnam.

Adrio Communications Ltd re-launch radio-electronics.com

Adrio Communications Ltd, owners of the radio-electronics.com website have re-launched the site as resource of free information, data and tutorials for those in the radio and electronics industries.

The site covers a variety of topics which currently includes: cellular telecommunications, private mobile radio (pmr), wireless connectivity, radio receivers, radio propagation, test and measurement techniques, and electronics components.

The front page links to useful articles both on site and elsewhere, and there is a small section that is frequently updated providing interesting quotes, dates, and facts associated with radio and electronics.

To make the site easy to use, time has been spent to ensure that it is easy to navigate, and to further aid any visitors a site search facility is included.

www.radio-electronics.com.

Lasers make mice better

Mouse maker Logitech has turned to lasers to improve tracking performance in optical computer mice.

"We've turned off the red light on these optical mice and replaced it with an invisible laser," said David Henry, Logitech senior v-p of its control devices business unit. "Laser will eventually make the optical mice of today obsolete," he claimed.

An optical mouse tracks by illuminating the surface it slides over, viewing that surface with a detector array through focusing optics. Algorithms in the mouse track marks and blemishes on the surface and infer motion direction and speed.

Smooth plain surfaces, and those with certain patterns, like polished or wood-grain surfaces, can fool the system and prevent the mouse working reliably.

Working with optoelectronic device maker Agilent Technologies, Logitech developed the laser system, which includes a sensor which can capture 5.8Mpixel of detail each second. "The nearly singular wavelength of laser light is capable of revealing much greater surface detail," said Logitech. "In tests conducted at Agilent, the laser mouse was found to have 20 times more sensitivity to surface detail than LED optical mice."

The laser has been built into the MXTM1000 Laser Cordless Mouse, which also includes Logitech's Fast RF 27MHz radio link which "enables the same rate of wireless data transfer as that of corded mice operating through a USB port: up to 125 reports per second, approximately 2.5 times more than other RF-based cordless mice."

Power comes from a lithium ion rechargeable cell which charges in under four hours and "typically lasts 21 days". A ten minute charge will give a day's use. www.logitech.com

Crimewatch in the UK

An image processing project to monitor passengers in public transport systems is being carried out by Thales Research and Technology in Reading.

Part of a European research team, Thales hopes to develop tools that can automatically recognise dangerous or criminal behaviour on systems such as the London Underground.

Charles Attwood from Thales describes the system: "First, a time-base corrector system assigns and aligns the timing on the video feeds. It then digitises the analogue signals, compresses them and sends them to the 'crowd monitor' system

provided by Kingston University in the UK.

"This analyses the video provided and returns alarms based on movements related to crowd behaviour - overcrowding on station platforms, for example."

Motion detection is able to track people as they move through the station. Behaviour recognition software issues alerts if any suspicious activities are spotted.

Called ADVISOR (Annotated Digital Video for Intelligent Surveillance and Optimised Retrieval) the system has been tested by the Barcelona and Brussels metro systems, the latter running 800 cameras.

Alarms are issued when the cameras spot overcrowding, access blocking, the detection of ticket barrier jumping, vandalism, and scenes of violence such as fights and muggings. www-sop.inria.fr/orion/ADVISOR

Many lasers make light work

Washington state-based laser firm Aculight Corporation has produced an external cavity diode laser that combines the outputs from seven diode laser bars, each containing 200 single-mode emitters, to generate 26W of near-diffraction-limited light at 820nm.

"To date, this may be the largest number of diode lasers combined into a single, high-beam-quality light source," said the firm.

Aculight originally developed the 820nm laser for a military customer that needed to project a small spot of light on a faraway target using an efficient light source.

"One of the most common lasers around is the diode laser. They're rugged, efficient and a great workhorse for generating optical power," said Dennis Lowenthal, Aculight's v-p for R&D. "But poor beam quality has limited their use in many applications. A technique called spectral beam combining [SBC], however, allowed us to merge the output of several diode laser bars and yet retain the beam quality of a single emitter; the result is a novel high-power, high-quality diode source."

SBC overlays the outputs of many laser emitters into a single, near-diffraction-limited beam. "The SBC technology behind the unique laser could provide an important stepping-stone for the use of diode lasers in many applications that demand high-power, high-beam-quality light," said the firm, which could be used in projection displays, photodynamic therapy and as a pump source for fiber lasers.

www.aculight.com

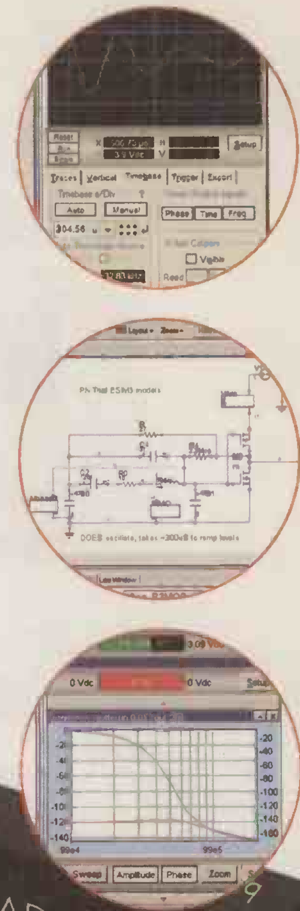
Pilotless plane takes electronics for a ride

Vector P is an unmanned aerial vehicle (UAV) aimed at carrying 9kg of electronics, cameras and sensors.

With a 3m wingspan, the plane can stay aloft for up to five hours and its position and payload information may be monitored from up to 100km via wireless data modems.

The plane can take-off and land autonomously, as well as automatically follow flight plans using GPS. Manual control is also possible.

Built-in safety systems can include auto-parachute deployment in the event of an emergency. www.intellitechmicrosystems.com



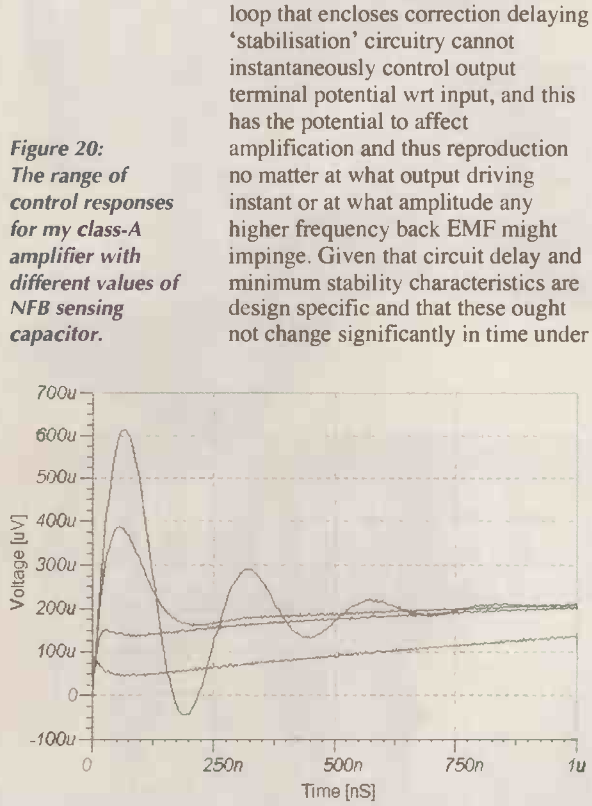
b² Spice AD v5
Professional Mixed Mode Design Software
Analogue & Digital Circuit Design for the PC
RD Research

Class-A imagineering: Part 6

Most high fidelity listeners appreciate the cleanly controlled loudspeaker cone reproduction of class-A amplification. After John Linsley Hood presented his simple 1969 design in *Wireless World* Graham Maynard needed to know why class-A is so much less fatiguing to listen to, because on-going solid-state design efforts are not worthwhile unless the final results actually sound better. Here he concludes his investigation into loudspeaker back-EMF induced amplifier distortion

To date my reverse driven amplifier simulations have been completed under zero input quiescent current conditions that could be said to rarely arise during sound reproduction. However 'passive' loudspeaker crossover frequency separation and dynamic voice coil motion cannot be achieved without some degree of loudspeaker system induced back EMF being generated, and thus any global NFB loop that encloses correction delaying 'stabilisation' circuitry cannot instantaneously control output terminal potential wrt input, and this has the potential to affect amplification and thus reproduction no matter at what output driving instant or at what amplitude any higher frequency back EMF might impinge. Given that circuit delay and minimum stability characteristics are design specific and that these ought not change significantly in time under the effect of any amplitude or phase of loudspeaker generated back EMF, no matter what type of amplification and NFB or output error correction methods are involved, my simulated reverse injection method for amplifier circuit excitation should remain representatively valid.

Figure 20: The range of control responses for my class-A amplifier with different values of NFB sensing capacitor.



Back in my valve amplifier days, I was well aware that varying the value of the pico-Farad capacitor which parallels the output terminal NFB loop sensing resistor could have a notable effect upon reproduction whilst not especially altering the amplifier's measurable steady sinewave audio frequency response and distortion figures. This capacitor affects not only closed loop stability, as may be observed via conventional steady squarewave observation using a real loudspeaker load, but also the 'quality' of sound that a loudspeaker might eventually reproduce;- too high a NFB capacitor can 'dull' or 'suck back' the dynamic detail of reproduction, while too low a value can cause a just perceptible false 'brightness' or 'glassy sheen', and thus the complete omission of this component should be avoided due to the way in which differentially sensed closed NFB loop controlled current correction ends up lagging back EMF modified output terminal voltage error.

Two of the most useful components in my early separate magnetic cartridge pre-amplifier were a dual gang 47kΩ variable in series with 22kΩ, and a 250pF variable capacitor in parallel with short and un-padded pick-up leads, both being finely set to match cartridge loading. As front panel controls both were mostly left in the optimally flat reproduction position, but the capacitor proved very useful in compensating for overly 'dull' or 'bright' pressings. As part of the real world testing for my Part 4 - Figure 10 class-A 25W-8Ω circuit, I actually tried an insulated 100pF air-spaced shortwave variable connected in parallel with its 10kΩ NFB loop sensing resistor to see if there might be any advantage in providing user variability. However amplification did not change significantly because the circuit already had an extremely low propagation delay, even though dynamic response and stability characteristics must have been becoming degraded at maximum and minimum settings respectively. Based upon past experience I had already prototyped using both 15pF and 22pF capacitor values, and was pleased with the neutrality of output when either component was fitted. In view of the long known importance of this component, I more recently wondered how my amplifier might simulate dynamically when the value of its NFB sensing capacitor was varied with 1V.RMS - 10kHz of reverse sinewave drive applied, with the results being observed from t=0; as illustrated in **Figure 20**.

Figure 20 shows four separate response simulations for NFB sensing capacitor values of 1pF, 4.7pF, 22pF and 100pF. The 620μV (-67dB) under-damped oscillation of the 1pF (stray capacitance) controlled response shows why this component should not be omitted; inadequate high frequency NFB loop sensing leads to control overshoot, thus the accuracy and stability of the reverse (loudspeaker back EMF) damping response becomes compromised. The 4.7pF component provides better damping, but at 380μV still does not adequately control back EMF induced overshoot. The 22pF (-80dB) trace corresponds to my earlier Part 5 - Figure 13 simulation, though with different resolution, this value being optimum for the simple two stage amplifier circuit. The 100pF trace appears to establish superior initial error control, but by now it is also beginning to degrade the amplifier's forward measurable phase response and is 'dulling' the first cycle forward amplification (transient) response with its own charging requirement, which causes a gradually settling error offset and increases the equivalent FCD figure I am keen to keep low so that the amplifier will retain top flight and uncompressed dynamic impulse response capabilities.

That the 22pF value subsequently turned out to be correct for quickly returning the simulated 25W-10kHz toneburst signal level zero shift to within 1mV of its biased dc zero, and with Figure 20 separately showing this component's excellent overshoot limiting, this is rather satisfying. Fortuitous perchance? No: the amplifier was already sounding 'good' because it does not appear to introduce audible waveform distortion or characterising qualities or reproduction smear, as more recent simulations have done no more than confirm. It was only after I had finalised the circuit and its write up that later simulator examination revealed the alternating signal zero for a 25W-10kHz sinewave toneburst most quickly returns to genuine zero with a feedback capacitor value of 18pF; also that the 100pF NFB sense capacitor increases the 20kHz forwards simulation measured phase shift to minus 6.7 degrees with a just perceptible alteration of reproduction quality corresponding to an input related equivalent FCD figure of 0.33%.

Squarewave testing?

Sub microsecond oscilloscope observation from t=0 with a suddenly starting sinewave is not easy to arrange on an audio test bench, though it is possible to reverse resistor load drive any amplifier with say a steady 10kHz 1V.p squarewave and monitor its voltage output through each '+ to -' or '- to +' generator induced output stage current crossover. Amplifier deviation and stability disturbance arise much more quickly than with sudden reverse sinewave start-up, and both examinations may well produce findings different to those arising via normal forward continuous squarewave drive with a resistive output termination. The results for reverse simulated steady squarewave drive for my class-A circuit could be shown in a separate illustration, but the nature of the traces do not differ enough from those already shown in figure 20. The amplitude of an initial reverse squarewave driven error spike relates to the natural output impedance of the output stage itself, prior to any global NFB loop or error correction circuitry being able to correct the induced voltage error, and thus initially it remains the same for a given bias setting and topology, irrespective of any NFB loop or error correction arrangements; the time period taken for a return to zero volts at the output terminal relates to the amplifier's separate external impedance, internal parallel capacitance induced propagation delay and global NFB loop characteristics; whilst any oscillation relates to the amplifier's reversely excited stability characteristic.

Now, I do not especially like squarewave testing, nor for that matter 2μF amplifier output terminal loading, because nowhere in real-world audio reproduction is there a similarly sharp response that must be followed or controlled, nor is there a similarly leading current load that must be driven in order to ensure subjective listening satisfaction, besides, the sibilance, musical harmonic and percussive transient responses these tests are most related to are more akin to being a range of abruptly starting harmonics, with the highest and most energetic frequency components being the least sustained. Also, not all squarewave generators have equally fast rise times so that measurements taken in isolation using different equipment, or the results from different simulator programs running with different time and voltage subdivisions might not be objectively satisfactory either. Indeed, simulated squarewave testing can even appear to induce circuit oscillation where this will not arise in real life due to component and wiring impedances not being entered on screen. There is however some real value to be gained from comparing different circuit variations or topologies with an unchanging test bench or simulator set-up, and this is my reason for directly comparing in **Figure 21** the simple Part 4 25W-8Ω class-A circuit with the exampled Part 5 generic class-B circuit.

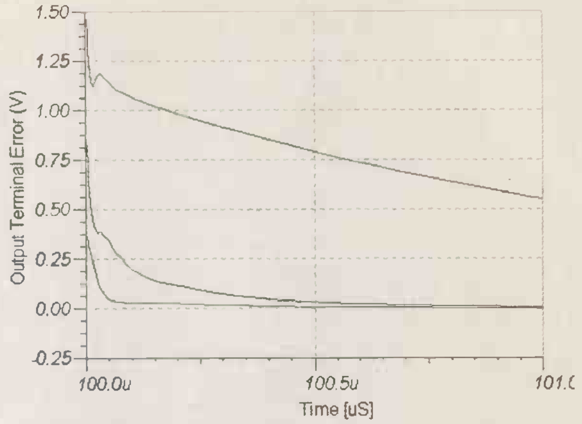


***Figure 21:* Comparing the output terminal responses for the generic class-B and simple class-A.**

Figure 21 illustrates how stored energy within the series output choke has led to the output voltage of the generic class-B amplifier momentarily exceeding the '-' to '+' 1V series resistor driven excitation, with eventual settlement from 1.46V extending off-scale and out to 5μs. Without the series output choke the lowly biased 260nH output impedance generic circuit hits 0.83V, and its C.dom delayed settlement takes 400nS, whilst my simple class-A limits error to 0.36V with a mere 50nS recovery period. The lower the amplitude of the error spike and the faster its stable return to zero, the more competent the amplifier will be at limiting dynamic loudspeaker back EMF induced amplifier-loudspeaker distortion without compressing ongoing dynamic control capabilities and smearing the eventual forward amplified sound-stage image. A non-feedback amplifier would limit error amplitude in relation to its bias and topologically generated output impedance, but would be incapable of subsequently establishing equal but opposite error correction and crossover-loudspeaker damping current flow, but then it could not induce loudspeaker cable, crossover or driver induced 'ringing' against a low output impedance either. There is also always the potential for the reverse squarewave driven error amplitude that is momentarily permitted by a stable class-B amplifier to be greater due to higher natural output impedance with lower standing current, no matter how fast the global or local NFB loop amplified correction, and this is why steady sinewave forward testing procedures alone will not reveal all of the circuit activity that could lead to different amplifiers 'amplifying' and thus 'sounding' differently, especially at those unavoidable instants of music waveform induced overdrive or overload. Unfortunately reverse squarewave simulation can also induce output stage cross-conduction where this is unlikely to arise in a competent real-world finished design due to series load impedances and the amplitude of loudspeaker generated back EMFs reducing with increasing frequency.

Could it be that it has been these momentary load induced and leading w.r.t. circuit delayed NFB loop stabilised amplifier output 'tweeter driving' errors that have been causing many solid-state amplifiers to sound tiringly 'over bright', 'sharp', 'brittle' or 'tizzy', where such effects have come to be formally attributed to minute levels of higher harmonic steady sinewave distortion when evaluated via conventional Fourier analysis with passive and thus non-back EMF generating resistor loads? Also are these 'Fourier' components equally generated when a real cable plus loudspeaker system is the amplifier load instead of a test bench resistor? After all, valve amplifiers have much inferior high frequency harmonic distortion figures at higher output powers without them sounding so dissonant; but then valve amplifiers do not similarly generate an extremely low output impedance against which loudspeaker terminal back EMF can so reactively develop, and I am not aware that amplifier distortion is routinely Fourier analysed when a loudspeaker system is the load anyway. I always found it extremely hard to describe the veiling and mind confusing loudspeaker reproduction of a Part 5 - Figure 14 type of class-B amplifier, for whilst there was no doubt that NFB always firmed up the bass response and reduced single tone distortion, these amplifiers made composite loudspeakers sound as if the musical dynamics had been internally compressed, such that not only was I aware of an anonymising loss of projected image depth and an alteration of detail, especially on leading wavefronts, but that sonic changes led to me becoming disenchanted and disinterested with what had previously been enjoyable listening from known to be good source material. Now that I have simulated these reverse (loudspeaker back EMF) induced leading edge distortions I realise that my ears had been right all along.

Curiously I did observe a similar sonic characteristic on one experimental valve power amplifier where substantial NFB loop sensing had been internally derived directly from the push-pull anodes instead of more conventionally from the loudspeaker output terminal, where the latter leads to a more naturally sounding control of transformer output with a more constant time relationship w.r.t. signal input, though I do feel that valve amplifiers could benefit from using limited amounts of both resistive anode internal and output terminal global types of feedback in a nested configuration, with further additional improvement being achievable by using quiet and directly coupled 50V powered differential solid state first stage circuitry instead of doggedly maintaining an 'all valve' practice with its unavoidable front end thermionic hiss .

Disparate attention required!

In spite of John Linsley-Hood via his *Solid State Audio Power* article in *EW+WW*, Nov 89, p1042, also Ed Cherry as he yet again repeated in his Ironing Out Distortion article *EW*, July 97, p557, as well as many others, all making suggestions that C.doms should not be implemented as a Miller connected capacitor at the VAS stage, my generic class-B circuit is similar to many that are still being written about by some 'established' designer-writers as if THD specifications can prove beyond criticism that these are upper echelon audio amplifier designs. The error waveforms shown in Part 5 -Figures 16, 18&19 demonstrate class-B weaknesses caused by circuit design flaws that regular *EW Letters* readers have been assured as recently as December 2003 are 'solved problems and dead issue' aspects that are 'completely understood', thus amplifier end users and expectant constructors are not getting a fair deal.

Class-B amplifier designers might care to run suddenly starting reverse driven output stage sinewave simulations via a lower value series resistor that will mimic loudspeaker impedance dips, whilst varying output stage, driver and input stage bias currents, because waveform control degradation due to the dynamic invocation of a voltage sensing NFB amplifier's internal closed loop current limitations, are much more significantly consequential for audio reproduction than the normally observed bias controlled minimisation of forward examined steady sinewave crossover distortion with an ideally resistive load alone. Moving the acting end of the C.dom away from the high impedance input stage output node in the way that was so long ago suggested, leaves the VAS collector free to slew at very high speed through the output stage bias range, which usefully minimises VAS stage correction delay and therefore limits loudspeaker back EMF driven conduction crossover commutation as well as conduction crossover induced differential input stage signal waveform modulation. By getting rid of the Miller Effect inducing VAS C.dom, and by minimising the V-ce variable effect of VAS transistor base-collector capacitance by increasing input stage current, or if the connection of a Miller C.dom is essential to prevent instability - by working it against a VAS base-emitter resistor to ensure that its rate of charge w.r.t. input stage current capability at audio frequencies is low enough for flat open loop gain throughout the audio spectrum - loudspeaker induced leading output terminal voltage commutation development at audibly significant frequencies is minimised, and the Part 5 -Figure 19 illustrated zig-zag superimposition plus time elongated crossover distortion anomaly could thus also be minimised, with amplifier-loudspeaker interface induced errors being reduced to extremely low levels.

Fundamental nulling

Over the years I have spent countless non-Fourier hours trimming the amplitude and phase adjusting controls of a fundamental nulling sinewave filter in order to reveal and study output stage conduction crossover plus amplifier generated distortion artefacts, and yet today, this is one method for waveform examination where our modern computer simulation software is much less convenient. We might need to sit back for ten or more separate long running simulations to gradually establish an exact amplitude and phase null for fundamental cancellation at the output voltmeter via a second separately configured voltage generator in series with the measuring instrument. What manually takes thirty seconds with stable test bench gear can easily take ten to twenty minutes on a desktop running with sufficient resolution to fully reveal the amplitude of coincidental non-linearity and any conduction crossover spikes, for these are all still there and it is insufficient resolution with

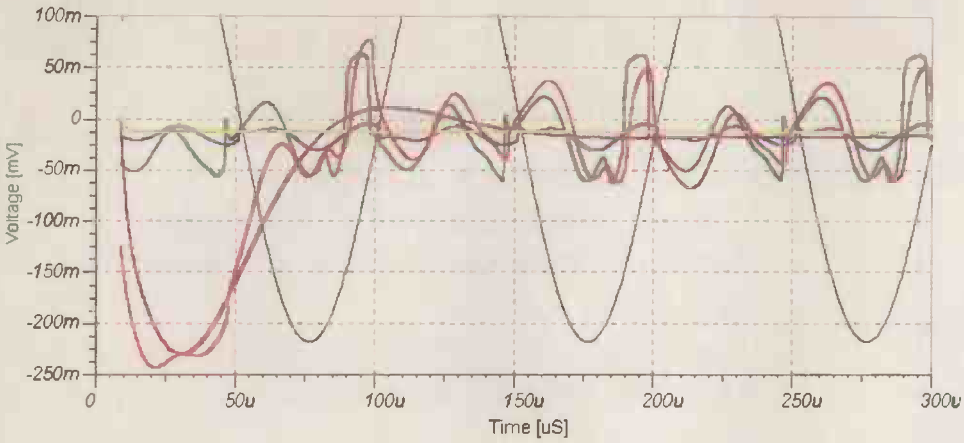


Figure 22: The simulated output terminal residuals for the generic class-B, with and without its Miller connected VAS C.dom and output choke

unrealistically used equipment that leads to them not being revealed. The computer method however is potentially much more useful, for once a specific sinewave null has been established to reveal exactly what the amplified but propagation delayed output terminal waveform is additionally carrying, i.e. the distortion residual that accompanies loudspeaker waveform drive, we can re-start the simulation to reveal and investigate any first cycle distortion that would normally pass undetected before test bench gear is capable of stabilising, and which some simulators might be incapable of analysing.

Now any analogue error arising between t=0 for a suddenly starting and rising sinusoid at amplifier input, and t=0 for the delayed output we eventually hear at amplifier output is not going to be as great as the simplistically 'time shifted' differential sinusoidal waveform subtraction would suggest actually arises during that initial propagation delay period, though as far as on-going music is concerned we cannot fail to hear any errors arising after the t=(propagation delay) period w.r.t. any other coincidentally running waveform already being amplified. This leads me to offer a simulated substantiation for different amplifier-loudspeaker interface distortions being generated by 'loudspeaker' back EMF interacting with different NFB loop controlled output stages in the same manner that my more convenient reverse injection testing has already suggested can arise.

For **Figure 22** I have parallel simulated the steady state responses for five separate generic class-B circuits, each having an identical Part 5 - Figure 14 semiconductor configuration, which is similar to those possessing excellent fundamental open loop gain linearity as investigated by Douglas Self in his 1993-94 *EW+WW* reference series entitled *Distortion in Power Amplifiers*, but which here have the different capacitor-inductor overlays as described below. Each circuit variation is separately fundamental amplitude-phase nulled in order to reveal its individually generated error waveform whilst outputting 15V.RMS at 10kHz into my virtual approximation of the 'Ariel' loudspeaker. The power rails for the generic circuit were increased to ±35V, bias to 30mA, and circuit usage is for illustration purposes only, not to claim any stable real-world reference amplifying capabilities.

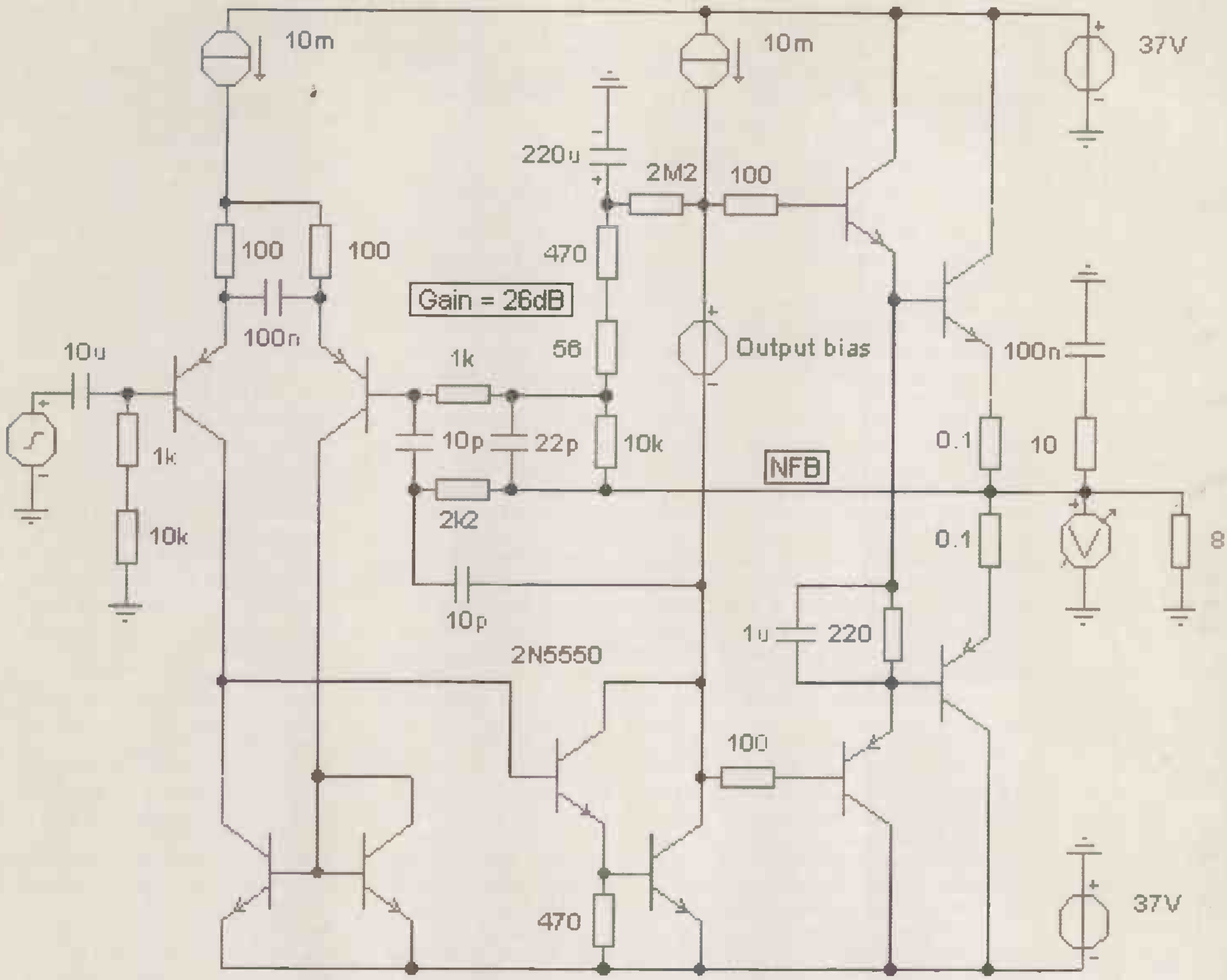
The *black* sinusoidal trace of Figure 22 shows the simulated generic Part 5 - Figure 14 Blue amplifier's output terminal waveform when divided by 100 in order to illustrate a minus 40dB level. Any first cycle amplitude distortion products arising at this level are unlikely to be subjectively inaudible on sharply expressive sound waveform leading edges.

The *blue* trace shows the simulated fundamental-nulled distortion-residual at the loudspeaker terminal for the generic circuit with both its 100pF Miller C.dom and its series output choke fitted, when the amplifier is loaded by a standard 8Ω non-inductive resistor. The amplifier circuit introduces an overall voltage waveform delay of 1.2μs, and the resistor loading ensures that both the output current and output voltage crossovers occur simultaneously.

The *red* trace shows how the amplifier's voltage error, especially during a first amplification cycle, has immediately become seriously compromised by loading it with an approximated virtual 'Ariel' loudspeaker in place of the resistor load. The first cycle error due to series output choke induced voltage error is clear to see, as is the Miller C.dom induced zigzagging back EMF induced reverse commutation which follows every VAS collector polarity change through output bias conduction crossovers, though these internal errors are somewhat smoothed by the series output choke. There is also evidence of prematurely impaired NFB loop current control at higher output voltage amplitude with the reactive loading, even though the virtual loudspeaker should not overload at this frequency. In this simple 10kHz sinewave example 'loudspeaker' loading has increased the voltage waveform delay to 1.8μs, whilst the Miller capacitor charging currents are ±8% of the first transistor's static current after 200μs! The differential input stage transistor current variation becomes ±16% at 20kHz, has highest amplitude and is beginning to 'tear' non-linearly as in Part 5 Figure 19 when the output stage passes through zero conduction crossover, as is clearly illustrated by the red output error trace.

The only causative difference between these blue and red simulations relates to amplifier loading; i.e. 'resistor' and 'loudspeaker'. When it is driven at 10kHz the steady sinewave zero current crossover for this approximated 'Ariel' loudspeaker example leads the zero voltage crossover by 4.8μs, thus the output stage bias is obliged to swing through its current conduction crossover before the voltage cycle ends. The global NFB loop enclosed leading Miller C.dom induced current flow no longer charges about zero output volts, but at approximately ±5V output voltages, and thus there is this global non-linearity, even without any coincidental low frequency drive which would further voltage offset that zero current crossover!

The *green* trace illustrates how removal of the series output choke alone can get rid of its dynamically induced contribution to loudspeaker back EMF-choke generated first cycle transient distortion. However, the unsmoothed Miller C.dom permitted loudspeaker back EMF induced reverse output node commutation, with its significantly higher speed product generation, is now clear to see. Without the series output choke, the amplifier's loudspeaker loaded voltage waveform delay has been reduced to 0.5μs, and thus the zero current class-B crossover for the green trace arises earlier than for the red and mauve traces. (By comparison, the virtual loudspeaker loaded waveform delay with my simple class-A amplifier is just 0.2μs, and there are no output stage induced current crossover discontinuities.)



The *mauve* trace shows how the error residual can be cleaned up by not using a Miller connected C.dom. The much improved high frequency bandwidth response/delay has here been established on the still unchanged class-B base semiconductor foundation by using a low value capacitor two pole HF gain degeneration network between the VAS collector and the differential feedback transistor base, and by adding a 22pF capacitor in parallel with the 10kΩ NFB resistor as used in my own class-A amplifier. See **Figure 23**. Only the Miller Effect inducing C.dom, the differential stage, and NFB loop capacitor arrangements have been modified for this simulation (the VAS driver transistor collector remains 0V connected) the result being a further 40dB reduction in steady state distortion after 200μs. The loudspeaker circuit plus series output choke induced first cycle error still arises, even though this would not be quotable on a specification sheet that merely lists a historically 'established' steady state THD measurement result.

Choke-less and C.com-less

Please note that this Figure 23 circuit has not been tested for stability in the form of a real world construction. It is used here to show how by removing the Miller C.dom alone that it is the internally generated leading differential input stage current flow within the globally closed NFB loop that causes additional class-B conduction crossover generated amplitude error, as opposed to crossover spikes merely being superimposed upon and straddling NFB loop controlled amplitude waveform distortion, as is so often imaged with simplistic passive resistor loading.

Yellow. With the Figure 23 gain degeneration arrangement used in place of the audio frequency gain stultifying Miller C.dom, and with the series output choke removed, we can reduce both input stage distortion and VAS collector delay, leaving the simulated amplifier with the illustrated near straight yellow line error trace. This introduces 1.7μs of voltage waveform delay, with up to 5mV.p-p of crossover distortion spiking, though with a negligible amplitude distortion error. Also note that the first transistor current variation has dropped to less than one hundredth of that with the direct 100pF C.dom connection as it drives the VAS stage buffer; hence the very immediate NFB loop generated output current control in the presence of exactly the same 'loudspeaker circuit' loading as for the green and red traces.

With the '- to +' 1V reverse squarewave Figure 21 test, the initial output stage error for the Figure 23 circuit remains a class-B 0.83V, but

Figure 23:* *The non-'Miller Effect' generating HF gain degeneration circuit used for illustration purposes.

the choke-less and C.dom-less recovery correction takes a mere 100ns, with especially improved performance notable via its -93dB amplitude and 20 degrees of leading 10kHz phase error when examined via a reverse driven steady sinewave investigation as per Part 5 Figure 11 where my class-A measures -64dB and 8 degrees respectively; figures that could question the sensibility of class-A usage.

The only difference between the red and yellow traces relates to NFB and stabilisation control components connected to an *unchanged* Figure 14 transistor layout. The Figure 23 circuit shows that it is possible for steady state distortion and reverse (back EMF) excitation damping levels to remain high at all audio frequencies, when using the earliest of semiconductor devices within bipolar circuitry very similar to that optimised and introduced by Mr Douglas Self as a 'Blameless' design; however unlike the Blameless, real world user stability for the full Figure 23 circuit has not been proven, and I am not presently able to check this out.

Clearly a Miller connected VAS C.dom drawing input stage current at audio frequencies causes a substantial and not subtle increase in dynamic loudspeaker load induced amplifier-loudspeaker interface distortion in the green and red traces when compared to the Blue resistor loaded trace. This confirms that there is a need for reverse injection testing to reveal the leading error caused by components that are invisible to the end user, for Figure 22 shows they can induce additional reproduction affectations when the amplifier is used as part of a real-world sound reproducing system. Figure 23 reveals by removal, how, when a deliberately introduced Miller Effect at an amplifier's three electrode VAS is used to ensure unconditional stability, it can simultaneously degrade global NFB loop control and leave the output stage more free to sharply reverse commutate through a portion of the fixed output stage bias when loudspeaker system back EMF induced current flow momentarily becomes leading phase shifted w.r.t an amplifier's signal input voltage. The leading two pole VAS collector potential derived HF degeneration applied to the NFB sensing input of the differential pair of Figure 23, does not interfere with the balanced voltage to current transductance of the input pairing or the control they exert upon amplifier output w.r.t. input. This same stabilising two pole filter arrangement can be added to other solid state amplifier designs to increase their stability and stabilise load induced amplifier circuit voltage delay.

Bridged amplifiers

It is also worth considering at this point how one amplifier might thus affect its bridge connected partner if the output from a signal driven A'1' is then used to feed a potential divider at the negative input of its A'2' pairing. Integrated circuits and amplifiers that are bridged in this way or via a commoned NFB loop link might bench measure okay when their common load is a passive resistor, yet produce much less convincingly realistic sound reproduction than when they are separately driven by an external phase splitter or via genuine '+' and '-' input reversing. This is due to an A'1's inability to instantly control its output terminal voltage potential under the influence of a portion of loudspeaker plus A'2' delayed error arising at the output of A'1', the error then being momentarily re-amplified by A'2' such that music input generates a range of differently timed voltage errors between A'1' and A'2' outputs. When the capacitor that parallels a NFB loop sensing resistor is omitted within a Miller C.dom stabilised amplifier that might work well on its own, it would not be impossible for leading loudspeaker induced output stage reverse commutation to initiate catastrophic inter-amplifier oscillation. Often, stereo car audio amplifiers run much more cleanly when each channel is separately optimised to drive 'in-phase' just one separate voice coil of a dual voice coil sub-bass driver each, when compared to bridged mode with the voice coils either in inductance increased series, or 4x amplifier current flow parallel. Small and sealed car audio enclosures generate phenomenal back EMF inducing air-spring pressures, so reducing the sound level does not especially improve reproduction quality when d.v.c. drivers are parallel driven by an auto-bridged stereo amplifier, even when power supply rail capabilities have not been exceeded!

Here I have managed to illustrate using a generic circuit similar to some that have been published, constructed and marketed, one mechanism whereby passive crossover circuitry and dynamic loudspeakers can cause many already existing class-B bipolar amplifier output stages to sound increasingly indistinct as rising waveform levels energise complex back EMF induction responses, simply because so many of those amplifiers have been fitted with a series output choke and/or Miller Effect inducing and NFB loop response delaying C.dom!

Any failure to understand the real significance of reactive loudspeaker induced distortion mechanisms which arise due to their frequency dependent dynamic impingements upon either a series output inductor or capacitor coupled output terminal, or a Miller-VAS C.dom HF control reduced amplifier's NFB output node, will devalue the most erudite of amplifier designer's efforts in reducing forward circuit distortion below what once were already excellent 0.1% valve power amplifier levels and especially those that use triodes with their natural load damping anode resistance, or grid aligned beam output tetrodes running in AB1 40% ultra-linear where there is still good damping plus a much lower internal Miller Effect plus lower reverse parasitic electron reflection and re-emission than with a pentode or triode at higher output power. Thus I cannot but wonder what types of amplifiers have been chosen by experienced loudspeaker designers for their own reference work - they might not actually use the better than 0.01% solid state THD specified class-B chassis that some 'experienced' amplifier designers think they should; and, if they did optimise their designs whilst using low impedance but reactive output stages, how would their loudspeakers then sound when driven by other more reverse phase linear plus minimally signal path and thus NFB loop response delayed amplifier types?.

My suggested reverse injection testing procedures, which apply no more than a simple 10kHz-1V.RMS sine wave via an amplifier's nominal load resistance, have been shown to have validity in predicting the likelihood of tweeter voltage modulation from loudspeaker system generated back EMF, due to a lack of coherent output terminal control via a series output choke or correction current delay within a closed voltage sensing global NFB loop solid state class-B amplifier. The procedure is simple, and may be retro-completed on any existing amplifier, old or new; class-A or B; other or dumper; nested feedback or error correction; thermionic or semiconductor; and, analogue or digital! It requires no more than a buffered, or amplified signal generator, a dual trace

oscilloscope, plus a nominal value of non-inductive load resistor. The reverse driving amplifier does not even need to be an ultra low distortion type, and so one channel of a stereo pair could be used to evaluate its partner. Switching to squarewave with an adequately buffered generator could also reveal any NFB loop control instability w.r.t. input, and, if the error amplitude is not near instantly controlled, reveal output stage impedance or a risk for loudspeaker driven pre-NFB controlled output stage reverse commutation through a portion of the output stage biasing potential.

A simple observation of reverse driven output terminal phase shift at 10kHz will indicate whether the chassis under test has a predominantly resistive characteristic that does not momentarily permit the generation of leading error potentials, or, have an inductive characteristic where suddenly generated leading quadrature and 'step like' error potentials could reactively increase with increasing frequency and dynamic loudspeaker impedance dips. Also, in view of the above mentioned momentarily high output stage currents already noted by Mr. D. Self, an increase in the voltage/current level of this reverse examination drive via the 'load' resistor until it dissipates the amplifier's rated power might also reveal any output device conduction crossover induced problems due to signal path induced current insufficiency limitations within the closed NFB loop, which could then become apparent only when driving efficient or crossovered dynamic loudspeakers loudly. Where a capacitive output characteristic is suspected, a second test could be run at 100Hz to check whether the output terminal develops an additional lagging alternating zero-level shift that increases with falling frequency, because this would have a worst effect due to phase shifting around any loudspeaker designer's deliberately intended utilisation of loudspeaker driver-cabinet bass resonance.

Real loudspeaker response

If the output potential of an amplifier being tested remains more than 60dB below that of reverse drive up to maximum rated current flow through the nominal series resistor, then any phase shift at 10kHz might be considered as being inconsequential, but this would be true only if an initial first cycle of reverse (dynamically) generated error is stably controlled by the closed NFB loop, which is why first microsecond and first cycle circuit energisation responses need to be separately simulated from t=0 w.r.t input and t=0 w.r.t output driven signal injection examinations.

Each of the simulated traces illustrated in Figure 22 was derived via individual fundamental amplitude-phase nulling with a differently phase delayed equal-but-opposite 10kHz steady sinewave in order to separately reveal distortion content. This nulling permits voltage error observance at each amplifier's output terminal, due to individual amplifier-loudspeaker interface delay induced reactions with identical loudspeakers and inputs. The traces represent the way in which a 10kHz waveform would be modified by composite loudspeaker reactivity and subsequently erroneously generated amplifier circuit responses. Conventional THD analysis fails to reveal the first cycle distortion due to a composite loudspeaker's impedance reacting with any amplifier's separate, but both integral and internal impedances, as has been is illustrated here. Using an oscilloscope to directly observe the output terminal waveform or differential input base potentials for error during music amplification with loudspeaker loading will not show up this kind of leading current induced crossover distortion because fundamental nulling is necessary to reveal such activity and this cannot be set up for music. Nor can THD measurement and scope monitoring reveal the dynamic activities of Miller C.dom generated 'internal' closed loop inductance or 'external' series inductance waveform modification either. When it is driving a reactive load the generic class-B amplifier does not behave as if the total inductance at its output terminal is a simple summation of both internal plus external attributes, because leading Miller capacitor charging current within the closed NFB loop causes additional 'ringing' through back EMF driven output stage conduction crossovers as it constantly hunts for, but never quite manages to achieve, a steady state correction, as illustrated by the green and red traces of Figure 22, whilst the choke allows dynamic loudspeaker current flow to momentarily modulate the amplifier's output voltage w.r.t. the NFB loop controlled output node, as per the first cycle red and mauve traces.

For those who have the equipment there is a possibility for 24/192 digitally encoding, delaying and decoding any amplifier input signal, and simultaneously comparing it with attenuated amplifier output terminal waveform potential to generate an on-going and live display of 'listening-time' error whilst real loudspeakers are reproducing real music waveforms.

Those amplifiers observed having inductive output characteristics under reverse sinewave examination are likely to be the ones that will be more tiring to listen to because they cannot keep themselves from distorting the leading edges of waveforms due to the influence of complex reactive current flow within real-world dynamic loudspeakers, as has been illustrated by the parasitic tweeter driving wavelets of my Part 3-Figure 8. Thus they cannot keep themselves from fractionally time and voltage smearing detail within the differential stereo sound stage image which the loudspeakers subsequently recreate, whilst amplifiers that pre-correctively overshoot during a cycle of reverse simulation are also likely to sound too 'bright' or 'glassy' when driving reactively complex loudspeakers. Those amplifiers having a capacitive low frequency characteristic might actually sound warmer and louder as the bass response softens with phase shifted NFB current generation at the output terminal because they cannot genuinely damp bass driver voice coil movement; this can lead to kick drums sounding flappy instead of tight.

'Standard' and 'established' amplitude-phase and slew rate measurements observed when amplifiers drive steady sine or square waveforms into a phase linear passive resistor load are so inadequate for determining how an amplifier might actually respond to and therefore 'sound' when it is used to dynamically drive suddenly starting asymmetric sound waveforms into real-world electro-magnetic loudspeakers, because single frequency testing cannot reveal how a family of musically related harmonic components can become differentially phase shifted or intermodulated w.r.t. the original input due to loudspeaker induced back EMFs separately acting against an amplifier's individual dynamic and NFB loop controlled capabilities.

Where the output of a signal generator cannot be gated for first cycle start-up, then an initial leading edge of reverse driven error as per Part 5 - Figure 18 might not be directly observable, especially if the reverse driving source amplifier already introduces its own first cycle waveform distortion. This is where simulator evaluation can be of significant help, because, although fundamental test bench nulling will reveal additional product generation when a real loudspeaker is the load, a real loudspeaker will make much noise and might not survive lengthy voice coil heating; also the test set controls will need endless tweaking as every change impacts upon both the phase and amplitude of null settings - necessary adjustments which prove that internal amplifier behaviour is not as constant as is generally expected. Look at the zero current conduction crossover timings in Figure 22 above, they vary with each of the individual amplifier circuit's internally generated current flows which eventually arise in response to loudspeaker induced back EMF, and thus the zero current crossover timing and amplifier circuit induced distortion will vary with frequency. All five of the Figure 22 amplifiers have identically connected transistors, but their different series-parallel path connected componentature introduced different internal NFB and signal voltage propagation delays, and thus each amplifier generated different loudspeaker back EMF induced soundwave distortions that would (except for the yellow trace example) change and sound different when the loudspeaker is changed!

Continuous display

We might not always be able to discern any amplifier altered delays which arise when a single channel is being tested in isolation, but our ear-brain neurology is capable of perceiving directionally and harmonically induced differential affectations having short time differences that would otherwise relate to 100kHz or more on a cyclic basis. This leads me to suggest another way in which it should be possible to compare different stereo amplifiers by using switch attenuated headphones, or an old series capacitor connected tweeter, or a ground isolated oscilloscope to observe an amplifier's load driving competence whilst it is actually music driving real-world loudspeakers; a method that should also enable investigation for whether a specific phantom sound stage disturbance or undiagnosed manifestation of distortion is actually source or power amplifier related, and which may also be computer simulated, as is illustrated here.

If both channels of a stereo amplifier under test are simultaneously fed together, with headphones, tweeter, oscilloscope or simulated monitoring of the difference between outputs, then nothing should be observable when each channel is driving the same output into equal resistance loads. However, if one resistor is then replaced by a composite loudspeaker system, as is illustrated in **Figure 24**, then any previously level balanced degradation of amplitude linearity or any reactively induced dynamic phase (time) shift in amplifier output will instantly be rendered audibly or visibly detectable. If there is any amplifier non-linearity distortion in the 8Ω output w.r.t. the loudspeaker output, then this too will become part of the audible error or amplitude trace, such that the output from a poor amplifier will appear worse than expected, but this is no bad thing because it is just as important to optimise amplitude linearity as well as minimise dynamically generated loudspeaker back EMF induced amplifier errors. Also, I see the Figure 24 examination arrangement as being a 'live' monitoring set-up because it is capable of automatically 'nominally nulling' the amplified music waveform via an on-going subtraction of a normal propagation delayed and 8Ω resistor loaded zero phase waveform, from that which becomes modified by loudspeaker back EMF generation, such that there will be an on-going 'listening time' indication of loudspeaker induced amplifier distortion whilst the amplifier is real time driving the loudspeaker with complex music signal waveforms.

Figure 25 illustrates the error waveforms as three pairs of amplifiers are simulated together. All of the amplifier circuits have been gain equalised for direct comparison purposes when being simultaneously driven by a single 1V.p - 10kHz signal generator source, to output 10V.RMS each when loaded with an 8Ω resistor. The 'Y' or vertical axis voltage differences displayed here relate to the resistor loaded amplitude distortion w.r.t. the loudspeaker output, plus reactive loudspeaker induced internal impedance error component w.r.t. the initial zero phase output across the resistor load, this being the reason for two crossover spikes appearing on the class-B trace. The initial and any zero voltage level spiking is due to an amplifier's normal output stage current crossover or poor stability response, whilst any second and normally larger amplitude spiking would due to an amplifier's internal stability of NFB loop delayed current control in the presence of loudspeaker generated but phase shifted back EMF.

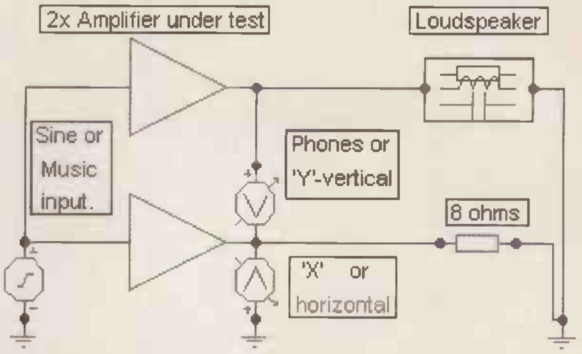


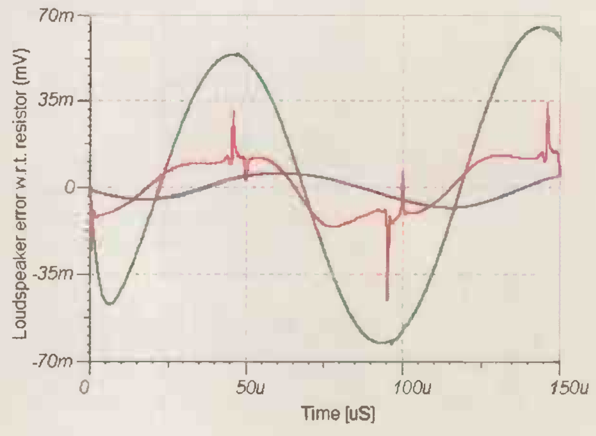
Figure 24: The 'X-Y' amplifier interface distortion monitoring arrangement

The *green* trace in Figure 25 shows the simulated output terminal 'V' difference caused by 8Ω loaded amplifier output nulling of the approximate 'Ariel' loaded amplifier for two Part-5 Figure 14 circuits; please note that the loudspeaker induced loss of 'amplifier' output amplitude shown in the waveform w.r.t. a positively starting common input has been divided by ten, i.e. the simulated difference and quadrature loss in output amplitude between virtual loudspeaker and plain resistor loaded outputs is 1.3V.p-p! The sudden dynamic loss due to the initial amplifier 'permitted' virtual loudspeaker induced series output choke voltage change and the subsequent quadrature phase shift are clear to see.

The *red* trace shows the Miller connected VAS C.dom plus NFB loop generated amplifier inductance of a choke-less Figure 14 'permitting' loudspeaker back EMF induction of initial plus subsequent reverse driven conduction crossover commutation, plus quadrature output terminal voltage development.

The *blue* trace is the resultant from my simple class-A. It has minimal sudden starting error and no crossover distortion, just a smooth and low level of NFB loop permitted

Figure 25: Simulated differential output results



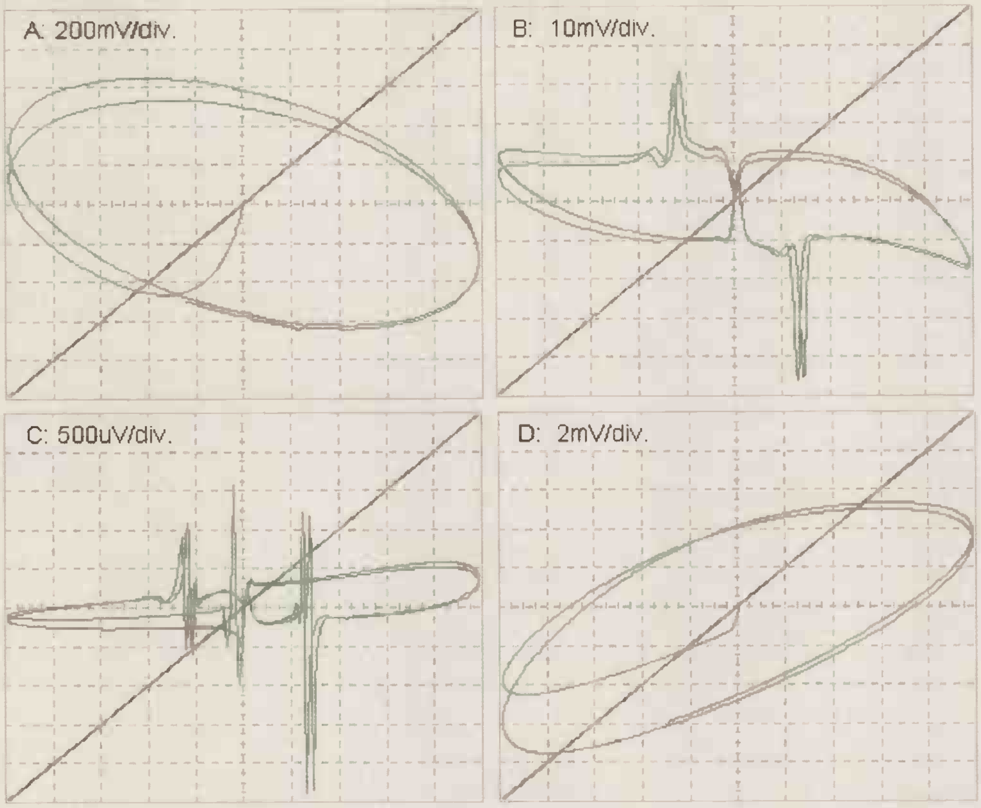


Figure 26: Four two-cycle 10V.RMS-10kHz 'live' interface distortion displays.

phase shift with near resistive loss; i.e. the loudspeaker induced amplifier loss almost mirrors amplifier input. Note that even though this is an excellent sounding amplifier the error (not its output) potential is more shifted than steady sinewave plus resistor examination suggests!

**The most humble of oscilloscopes should be capable of producing this kind of differential display from the 'Y' take off point illustrated in Figure 24. Note however that the oscilloscope input ground is directly connected to the resistor loaded 8Ω amplifier output, so it must be a battery portable type with anything connected to it being ground isolated and insulated, or be ground isolated and powered via a good isolating transformer; also the oscilloscope should be kept well away from the audio frequency signal source and cabling, also the amplifier's input panel. The oscilloscope timebase may be free running, event triggered or externally synchronised to the phase linear 'X' waveform appearing across the 8Ω resistor such that an on-going display of dynamic loudspeaker induced distortion may be viewed during 'live' listening.

'Live' distortion monitoring

My final illustration is Figure 26. It shows the traces from four pairs of amplifiers each being driven at 1V.p - 10kHz and again equally gain matched to output 10V.RMS across an 8Ω resistor. Here the difference between the resistor loaded amplifier under test, and its virtual 'Ariel' loudspeaker loaded stereo partner, goes to the Vertical or 'Y' input of an oscilloscope as set out in Figure 24, whilst the Horizontal or 'X' input is connected to the common ground termination. The oscilloscope ground is again directly connected to the 8Ω loaded amplifier output, so good isolation remains essential.

I well remember using a similar oscilloscope arrangement to monitor transmitter radiation when the BBC first introduced dynamic amplitude modulation control on their AM services. A synchronously demodulated output was fed to both the oscilloscope vertical and brightness inputs, whilst receiver intermediate frequency amplitude was fed to the horizontal input; the result being a flat based triangular display that was not meant to go beyond a 100% modulation pointed top display. Unfortunately it not infrequently did so, though nowadays with their over-enthusiastic automatic dynamic 'optimisation' of studio output being compression limiter adjusted to match the 100% modulation level, supposedly 'necessary' for transmission efficiency and ease of listening at noisy reception sites, our broadcasters are ruining AM reception for everyone due to amplitude demodulators sounding as if 'flat battery' garbled because they become highly non-linear at 90% carrier modulation levels, let alone 100%. This is like audio; what is monitored at an amplifier input or transmitting antenna input, is not the same as what we eventually hear.

I mention this AM radio modulation monitor because it was a display that could be left constantly running and be viewed whilst listening to distorted reception. This is the idea behind my Figure 24 'live' monitoring method for observing music signal induced amplifier interface distortion, because loudspeaker induced amplifier distortion products can be rendered visible and be directly associated with what we are listening to whilst the amplifier under test is music driving real-world loudspeakers. Of course this method of investigation can also be set up using computer simulation software, as here in Figure 26 which shows;-

A; the 1.3V.p-p output vertical input error of the Part 5 - Figure 14 circuit that arises due to series output choke decoupling between the amplifier's loudspeaker loaded output terminal and its true NFB loop controlled output node, w.r.t. the nominally propagation delayed 'X' axis reference output that is more linear and time coherent when the amplifier's load is purely resistive.

B; without a series output choke the Figure 14 circuit error amplitude is limited to between ±10mV, but audio is still going to sound coarse with this degree of conduction crossover spiking at only 10V.RMS output on such a simple waveform. This virtual loudspeaker induced error really is substantial!

C; the very much reduced initial spiking, horizontal start-up, low amplitude distortion and phase shifted crossover spiking of the Miller C.dom-less and series output choke-less Figure 23 circuit is plain to see. Here it has become obvious that not only has one half of the output stage maintained better control than the other by its reduction of vertical input voltage error, but with there being less stray path capacitance charging within the closed global NFB loop, the loop response is more accurately defined, and thus it is more resistant to back EMF induced reverse commutation through a portion of output bias potential. This kind of difference between output halves due to dynamic loading would not be

obvious via other more conventional forms of testing.

D; my simple 25W class-A generates a less vertical (less inductive) first cycle start-up error without any crossover spiking, and a smoothly controlled low level amplitude error. Note however that whilst a pure biased class-A amplifier might perform cleanly it is unable to drive as loudly through loudspeaker impedance dips as can a fixed biased class-A that is free to run into class-AB output operation when dynamic loudspeaker current requirements become momentarily increased; for interest read: www.passdiy.com/pdf/a40.pdf

The vertical axis portion of waveform distortion in C is approximately 25dB lower than in D, but the 'loudspeaker' induced crossover spikes, which in this simple 10V.RMS - 10kHz sinewave example remain sharp, would almost touch the smooth and near resistive better than -60dB trace of D if both were superimposed. Note that both of the chokeless class-B amplifier voltage outputs pass close to the centre zero with slight offset after each half cycle, and that the loudspeaker induced output stage zero current crossovers have become very clearly illustrated due to a degree of sinusoidally generated expansion in the centre region of the display. The illustrated loudspeaker back EMF induced leading current left-up/right-down crossover spikes actually arise between 4 and 4.5μs, whilst the sinewave synchronous graticule width here represents between ±25ms of amplitude display. When musically driven 'loudspeaker' induced errors are displayed it will only be the relative output amplitudes that may be determined, for the frequency and thus the time period of horizontal display width will be inconstant.

In all of these displays the maximum horizontal amplitude was potentiometer adjusted to match the amplitude from the sinusoidal voltage appearing across the 8Ω resistor of each individual amplifier pairing; in real life the sensitivity of one channel on a dual beam scope would need to be most appropriately set for the required front panel switched horizontal deflection. Also note that all simulated second cycle traces are slightly different to the first; this is not especially due to any amplifier weakness, but to composite dynamic loudspeaker characteristics taking longer than one 10kHz waveform time period to stabilise, which can significantly affect amplifiers (especially pure class-A with their naturally limited output current driving capability) that are expected to drive loudspeakers with more dynamically complex music than the medium level simple sinewave comparison illustrated here.

If this kind of dual axis 'live monitoring' display is set up with say a modern digital hand-held oscilloscope, I would expect a good amplifier to maintain a mostly horizontal straight line trace. Those amplifiers generating crossover distortion will produce centre of screen voltage spikes much as illustrated in B and C; those amplifiers which current clip due to dynamic loudspeaker loading with insufficient amplifier bias or output stage current limiting will generate vertical off-screen left-down/right-up error traces closer to the left and right display edges, whilst any semiconductor device current insufficiencies, saturation or cross conduction could become easily identifiable as the differential vertical error responses fall or lift away from what should be a smooth and more horizontal amplitude display. Also where an oscilloscope has a voltage triggerable storage capability it should be possible to study individual error events.

The choke-less and Miller C.dom-less class-A designs discussed in this text are not upset by reverse investigations where the series load is driven up to the amplifier's individual bias limited maximum current capability, but class-B types that use output chokes exhibit notable reactive load current induced error potential generation at their output terminal no matter how accurate the internal NFB arrangements, and, even without an output choke, Part 5 - Figure 14 like circuits can still be reverse driven into additional conduction crossover distortion at higher loudspeaker driving levels due to the Miller connected VAS C.dom capacitor's leading input stage current flow reducing necessary dynamic NFB loop control at the output terminal, and thus failing to prevent additional reactive loudspeaker induced output terminal current flow from reverse driving the output devices through a portion of their VAS controlled forward bias potential, as has now been much more comparatively simulated in Figures 22, 25 and 26. This still occurs at low level in example C for the Figure 23 circuit, because I have deliberately used VAS driver transistor C-bc Miller Effect to degenerate hf gain and thereby improve closed loop global stability, though the error amplitude remains very low.

Figure 21 shows how a fully biased class-A design can be more naturally resistant to initial output terminal leading voltage shift under the influence of dynamically generated loudspeaker back EMF currents. Mosfet class-B amplifier designs tend to be more highly biased and can have naturally lower gate voltage controlled output impedance, but they still need to be checked for additional resistor/collector/drain driven gate-capacitance induced circuit delay creating an additional signal path phase change within what often already are two or more stages of high gain-bandwidth amplification within a global NFB loop, and which often are stabilised through the use of a Miller connected VAS C.dom; the very component that momentarily increases internal high frequency non-linearity as one or other output device drops out of current conduction maybe one to five Mosfet bias volts offset from the alternating signal zero voltage potential. This can show up very clearly via X-Y monitor investigations.

Now I am not stating that all class-B(AB) amplifiers are going to perform badly when driving loudspeaker crossovers and voice coils with suddenly changing asymmetric waveforms, but reverse or differential load testing methods could be a definitive method for 'live' monitoring their capabilities with a view towards topological performance ranking. Also there is no reason why a digitally delayed or this differential examination method could not also be used for checking out different amplifier plus loudspeaker cable characteristics by monitoring loudspeaker terminal potential w.r.t. the resistor loaded reference; maybe those who rely upon steady sinewave voltage measurements will then come to realise that loudspeaker sited 'passive' crossover networks can set up 'dynamically' significant amplifier-cable-loudspeaker induced audio waveform distortion effects.

Conclusions

Quite literally I had all of my 'FCD' (first cycle distortion) reasoning, and both class-A amplifier circuits completed before I thought about simulating these 10kHz reverse driven output response examinations and the parallel-differential distortion monitoring investigations which have further lengthened and thus delayed the originally intended completion of this article. My investigations arose from a personal need to know why simple class-A amplifiers already have a natural advantage, and why they can be so much less cerebrally tiring to listen to, thus I feel that the above findings are worth passing on, even if some readers might dislike my presentation, find related errors that could rightly be stated to affect examination detail but not the nature of this overview, or claim that some of it has been 'old hat'. If anything I have written is wrong then it can be corrected, but please realise that I am unable to cover with a depth I would have ideally preferred to due to real-life limitations and lack of facilities.

Any power amplifier having a low equivalent forward first cycle distortion figure is very likely to have the excellent and phase coherent NFB loop controlled output terminal error correction necessary for dynamically transparent treble and image reproduction via real world loudspeakers, but music cannot be stopped in order for us to make those first cycle or fundamental nulled observations of amplifier performance, and this is where the differential monitoring arrangement, with its continuous and 'live' display capability can give us an instantaneous audible and/or visual indication of real-time amplifier-loudspeaker interface induced amplifier distortion.

Observance of these differential traces has convinced me that good results from the above suggested methods for 10kHz investigation are essential, i.e. a low equivalent FCD output terminal measurement w.r.t. a conventional forward input voltage; a low level of low angle steady sinewave residual w.r.t. a reverse testing load resistance injected input; a low level of initial reverse driven dynamic output terminal voltage development with fast and stable control, and especially, a smooth plus low level of observable differential resistor-loudspeaker load induced dynamic error via the Figure 24 test circuit. Differential testing is especially easy to set up on a good simulation program screen by copy-pasting a resistor loaded amplifier circuit on to the same page and then changing one load into a virtual loudspeaker, linking the inputs to a single source and attaching horizontal and vertical screen oscilloscope inputs as illustrated. Indeed, this method is so instantly revealing of possible reactive loudspeaker induced crossover distortion and loading effects, circuit reactivity and stability, device storage and gain-bandwidth product related differences between stages and output halves, also all signal and NFB loop enclosed impedance - especially parallel, interstage or feed-forward connected capacitors, that it has already become my most used amplifier circuit investigation method. It has already shown up weaknesses on many supposedly 'high-fidelity' amplifier circuits, and thereby offers a practical route for design improvement.

Simulation of the humble JLH-69 with a 22mF output capacitor reveals that its circuit typically manages an equivalent FCD figure of 0.05% with a mere 14 degrees of NFB loop controlled reverse driven phase shift at 10kHz, though with its limited bias and loudspeaker damping control at approximately -30dB that original circuit is not quite as capable of driving complex loudspeaker systems loudly. Any amplifier that can match or better these figures is likely to sound acceptable, though with hindsight, and as with the generic class-B circuit used here for illustration purposes, the original JLH class-As can also be further improved upon without changing the basic transistor arrangement they already use. The Figure 14 circuit has an equivalent first cycle distortion figure of 0.43% and a reverse driven 10kHz phase angle of 85 degrees; the very similar Miller C.dom-less and choke-less Figure 23 an excellent 0.10% FCD and 20 degrees; my simple class-A 0.02% FCD and only 8 degrees of reverse induced phase shift.

An 8Ω resistor measured equivalent 10kHz FCD figure of circa or better than 0.1% within a 100kHz bandwidth should signify a satisfactory standard for dynamic high fidelity reproduction with good pass-bandwidth and fast closed NFB loop generated loudspeaker back EMF control. Under 10kHz reverse driving examinations, an initial error can reveal either series output choke reactivity, enclosed NFB loop propagation delay induced inductance, an inadequate closed loop stability margin, or too high an input (filter) impedance degrading the NFB loop's discrimination capabilities. A low level and low angle of reverse driven steady sinewave error development, especially at higher reverse testing levels will prove not only a good dynamic loudspeaker driving capability, but that all stages within the closed NFB loop have high frequency current phase linearity and are biased with sufficient capability for charging all unavoidable and internal signal path capacitances. Where an input filter has a -3dB turnover below 200kHz its delay could affect and possibly mask subsequent weaknesses in an amplifier's forward tested dynamic response capabilities. Yet an input filter, or a series output choke, must not be bypassed for testing purposes because this will affect reverse loudspeaker back EMF energised closed NFB loop operation and thereby falsely improve the genuinely realisable response speed, which is where these reverse driven and the differential output monitoring investigations come into their own because they allow testers to examine amplifier reactions and look out for those series inductance and minutely delayed but audibly deleterious 'voltage sensed' output-current correction errors that otherwise might have remained unmeasured and thus unstated.

There might indeed always have been more of a Fourier-less 'art' to designing genuinely realistic sounding audio systems than many technical writers have been willing to accept. Properly noted findings of audible change due to single circuit alterations backed up by continuous auditioning, with the development of an ever stronger and empirically refined foundation does have an equally viable basis in its own right because measurements alone cannot fool humans into accepting reproduction errors along the lines of those touched upon in this writing, no matter how superb the quotable THD and slew figures that are still being quoted; that is of course, as long as the loudspeakers, listening room or studio do not overly modify reproduction and the sound levels are not too low or too high when compared to an original performance as to make the learned cerebral perception of human hearing mask out distortion, which does happen more often than is realised. I have often been able to discern amplifier distortion that has been masked at higher output powers by completely closing one ear and lightly closing the other with gentle finger pressure in order to (non-linearly) attenuate high sound pressure levels; and to protect my hearing too! The qualities or weaknesses of high power amplifiers really are best examined in large upholstered theatres, or at a distance in open air and thus quite literally in a free field, also by driving a variety of different loudspeaker types. Empirical designers could find the above differential load examination technique especially useful for it reduces their need to purchase calibrated distortion measuring equipment, and leaves funds for where they can be more productively used.

Time for a cuppa

Through these very lengthy efforts I have come to realise that separate forward and reverse energised first cycle output terminal measurements, backed up by differential resistor-loudspeaker output monitoring studies of an amplifier's response capabilities are essential before anyone can state beyond doubt that any audio frequency amplifying circuit is genuinely capable of being deemed a fully-fledged loudspeaker driving *audio* amplifier, no matter how fantastic its 'conventional' specifications, or how emotional the claims are for one particular type 'sounding' best. Maybe some high-tech amplifiers never did deserve the 'high fidelity' tag that had been claimed for them, and it is little wonder that established retailers have by necessity become so experienced at matching amplifier-loudspeaker combinations.

I cannot claim that these findings are definitive, for I am not especially knowledgeable of theory, but I have sought to test my own beliefs, as well as the understanding of others, in order to ensure that the results of anyone's conscientiously applied theory does not actually conflict with the nature of that which is fundamental and demonstrably observable.

At this point I need a break from the writing, but in so taking I will leave readers to ponder a rather significant question:-

What's inside your amplifier?

As with my previous article, this writing has been long in completion, then freely passed on to *Electronics World* for publication. Sincere thanks are therefore due to the entire *Electronics World* team for their substantial donations paid directly to the Marie Curie Cancer Care Nursing Service.

Thank you very much; Graham Maynard.

Graham hopes to be back soon, examining some of the loudspeaker driving capabilities of ultra-linear beam tetrode output stages, and the challenges they set for competent solid state replacement, which eventually led to his development of a more efficient JLH based class-A power amplifier.

He suggests that anyone interested in class-A amplification visit the excellent web site run by Geoff Moss, and especially Geoff's 'JLH update' page at;- http://www.tcaas.btinternet.co.uk/jlhupdate.htm

Simulating power MosFets

In this, the second of a planned four part series using the Micro-cap6 software, Cyril Bateman introduces a method by which any Spice user can 'hand carve' the power MosFet models needed when designing audio amplifiers

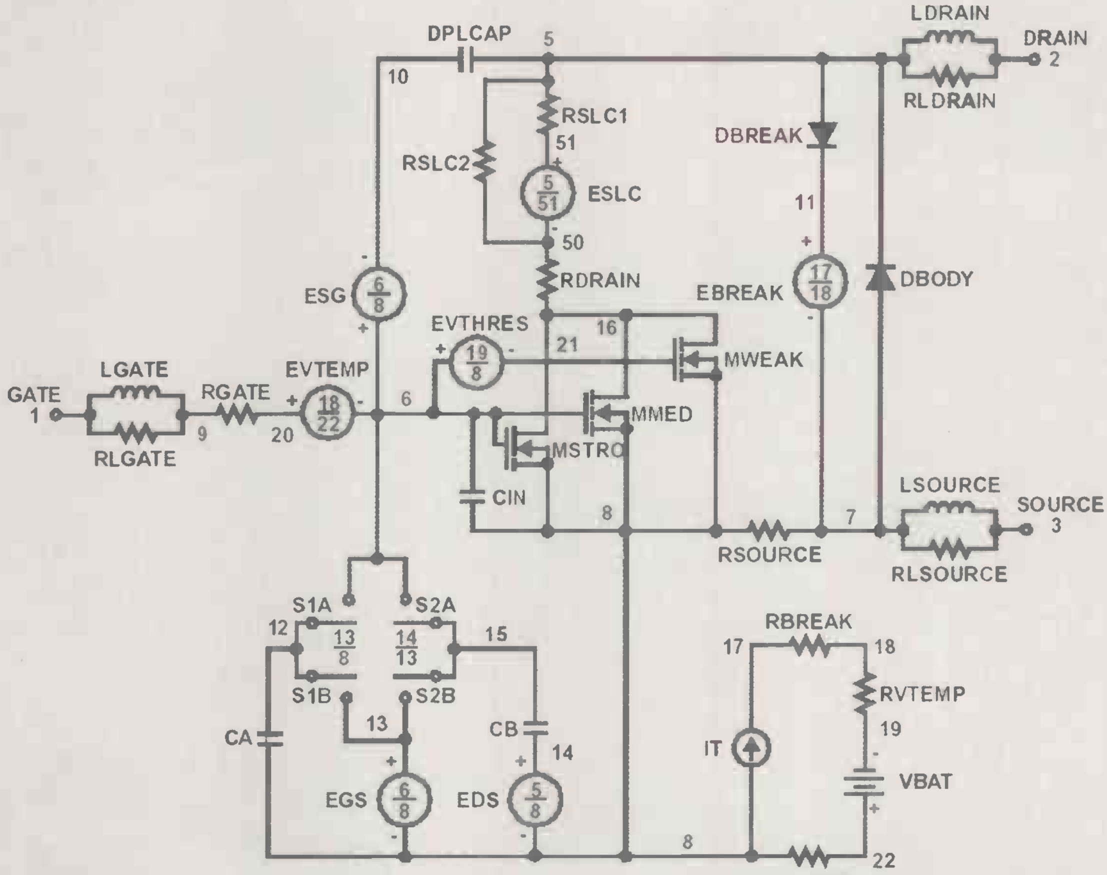


Figure 1: Fairchild's PSpice schematic for their FDP038 subcircuit model. Modelling audio frequency distortion, the circuit voltages used to drive the gate-source capacitance switches should not occur. To minimise convergence problems these switches were deleted from my audio models.

Power MosFet transistors are said to be easier to manufacture than the equivalent junction transistor, but vastly more difficult to model accurately. For this reason most Spice2 users rely on using models provided by the transistor maker. In my last article I outlined how more accurate models, better suited to modelling distortions, might be produced by a user 'hand carving' a subcircuit model using data from the datasheet, supplemented with a few simple low current DC measurements.

In this second article, we get our hands dirty producing our first subcircuit model, one which does not need complex Spice2 Analog Behavioral devices. Using three Spice2 Level-1 MosFets together with selected Spice2 devices, this model should be usable within most versions of Spice2. My next article describes a most complete model using more complex behavioral devices which do require a PSpice compatible simulator.

To accurately model distortions in a power amplifier with MosFet output devices, our model must accurately predict MosFet behaviour for small drain currents while subject to the drain-source voltages used in power amplifiers. Datasheets mostly concentrate on larger drain currents at a small drain-source voltage. For example, the datasheet transfer curves for my chosen devices used a 10V

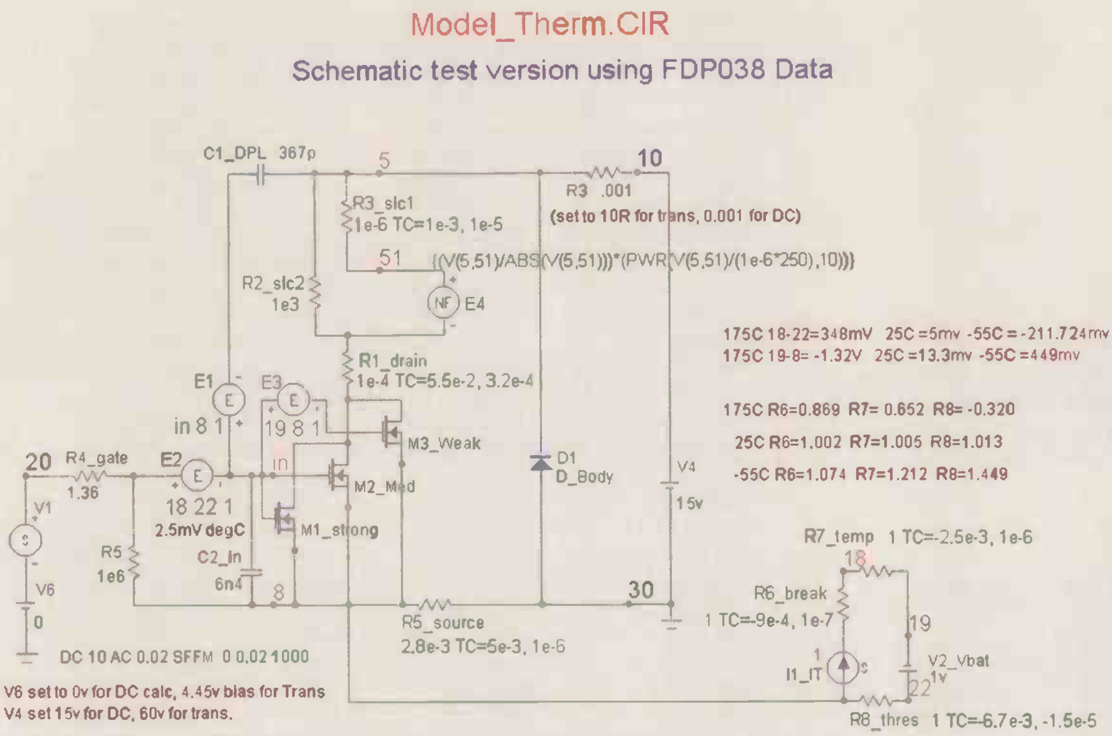


Figure 2: Nodes 10, 20, 30, representing drain, gate and source connections into the subcircuit model, will be translated by the MC6 schematic shape editor into the connecting nodes 1, 2 and 3 for the final subcircuit model. This schematic was used to plot and quickly refine the model, the first step towards the netlist needed for our subcircuit model.

drain-source voltage and plotted currents up to their 16A maximum. The curves for currents below 3A were cramped and difficult to read accurately.

The drain-source voltage used in a power amplifier might be 35 volts or more and 3A transistor current in a typical output stage is sufficient to produce some 50 watts of audio. Using increased supply voltages and staying within the MosFet square law characteristic, more than 100 watts may be produced.

Initial exploration

When I commenced this work, my knowledge of power MosFets or their problems was limited, mostly gleaned from private correspondence with Ian Hegglun during his lengthy investigations and from reading his two part article, May/July 1999. So followed many hours searching my literature files and internet using Google, as well as visiting and searching most power MosFet makers sites and downloading every interesting file, in total some 200Mbytes.

After several weeks work I attained my aim, usable Spice2 power MosFet models better able to predict audio distortions than conventional models. Not by devising much original work but rather by selecting and combining the most appropriate, existing, non-propriety modelling methods, able to work from a datasheet without needing details of the transistor construction or any specialised measurements.

Claiming to provide both electrical and thermal models, Fairchild's FDP038AN06A0 datasheet and two application notes AN7532/7533 caught my attention[1]. On examination the thermal model claim did not really stack up, but the electrical model could be used as a knowledge base. It would be possible to produce similar models with information extracted from a datasheet and using only relatively common Spice2 devices. Whenever I undertake a new computing venture, I like to start either with a simple, easily verified task, or if more complex, less well understood, one which can be separately verified. For this the FDP038 model looked ideal.

A complete model requires those characteristics which change with temperature, current or voltage in use, be described. Subjected either to reversed gate-source voltages of more than two or three volts or a very small drain-source voltage, power MosFet input capacitances change abruptly. Change of Cgd is modelled by a diode, DPLCAP and the change of Cgs is modelled using the combinations of switches and capacitors CA and CB. Such voltages are unlikely to occur in audio circuits and switches can increase convergence problems so were omitted from my models, as were D_Break and E_break, which model the transistor breakdown voltage, all easily reinstated if required, see **Figure 1**

Component models used

With the exception of the models used for the three Level-1 MosFets and the diodes, all other model values and equations needed are input in the schematics and visible in the figures. Schematic drawings and full subcircuit netlists are available from the editorial offices (contact Caroline Fisher, details page 3), but to save space in the issue have reduced the netlists to those device models needed, but not detailed on the schematic drawings, to implement the circuits:-

```
BUZ900/01D semiconductor models only:-
.MODEL MstroMOD NMOS (LEVEL=1 LAMBDA=0.001
VTO=2.53 KP=7.5 L=1U W=1U RS=0.8 IS=1E-30 TOX=1
RG=10 N=10 T_ABS=25)
*
.MODEL MmedMOD NMOS (LEVEL=1 LAMBDA=0.001
VTO=0.285 KP=3.0 L=1U W=1U RS=0.18 IS=1E-30 TOX=1
RG=1.36 N=10 T_ABS=25)
*
.MODEL MweakMOD NMOS (LEVEL=1 LAMBDA=0.001
VTO=0.275 KP=0.2 L=1U W=1U RS=2.9 IS=1E-30 TOX=1
RG=25 N=10 T_ABS=25)
*
.MODEL D_BODY D (IS=2.4E-11 N=1.04 TT=1E-9
CJO=1.6E-9 M=0.54 XTI=3.9 T_ABS=25)
*
.MODEL D_PLCAP D (IS=1e-30 N=10 CJO=100E-12
M=0.47)

BUZ905/06D semiconductor models only:-
.MODEL MstroMOD PMOS (LEVEL=1 LAMBDA=0.001
VTO=-2.22 KP=8.75 L=1U W=1U RS=0.3 IS=1E-30 TOX=1
RG=10 N=10 T_ABS=25)
*
.MODEL MmedMOD PMOS (LEVEL=1 LAMBDA=0.001
VTO=-0.578 KP=2.75 L=1U W=1U RS=0.395 IS=1E-30
TOX=1 RG=1.36 N=10 T_ABS=25)
*
.MODEL MweakMOD PMOS (LEVEL=1 LAMBDA=0.001
VTO=-0.28 KP=0.38 L=1U W=1U RS=2.5 IS=1E-30 TOX=1
RG=25 N=10 T_ABS=25)
*
.MODEL D_BODY D (IS=2.4E-11 N=1.04 TT=1E-9 CJO=2E-
9 M=0.54 XTI=3.9 T_ABS=25)
*
.MODEL D_PLCAP D (IS=1e-30 N=10 CJO=300E-12
M=0.47)
```

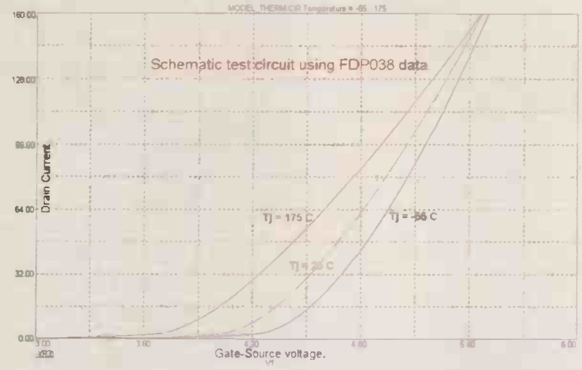


Figure 3: These DC analysis plots from the figure 2 schematic illustrate how changes in junction temperature affects drain current with gate-source voltage for this power MosFet. The crossover region in a power amplifier is affected by changes in threshold voltage.

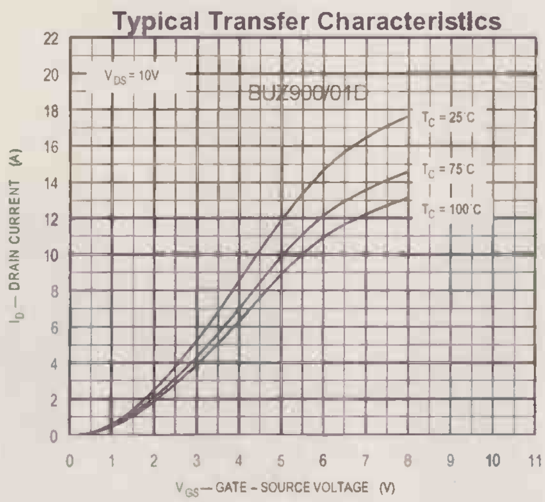


Figure 4: The datasheet curves for the BUZ900/01D MosFet are small, so difficult to read especially for small currents. These transfer curves were printed as large as possible, then annotated with actual spot values to facilitate developing my model.

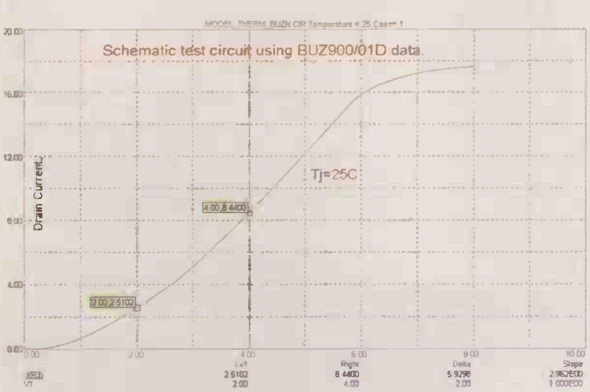


Figure 5: Attaining this smooth merging of the sub threshold conduction - square law regions, important when modelling distortion, completes stage 1 for our model. Our 25°C curve shows good agreement with the datasheet.

MosFet input capacitances Cds and Cgd also change with drain-source voltage, but both curves flatten with increasing voltage. These capacitance changes could cause distortion and affect the transistor's apparent gain with frequency. Using AC simulations to model frequency response Spice2 performs a small signal analysis, so cannot replicate these capacitance changes with drain-source voltage. Modelling distortion using transient simulations, the subthreshold/crossover distortion region is critical, however drain-source voltage will then be large, minimising these capacitance changes.

Obtaining model data

I found the Fairchild[1] and Infineon[2] sites the most useful. Both companies produce a large selection of switching power MosFets for which they list models and a good selection of application notes. I already had a copy of the Harris CD ROM from 1998 which provided many early application notes. One of the earliest attempts to provide an adequate model is described in AN9210 dated February 1992. Now part of Fairchild, several of these Harris application notes have been updated and re-issued, explaining in some detail the Fairchild PSpice models.

Two meanings seem to be inferred by the description thermal model, some makers use it to describe the transistor's internal heat flow from junction to case, others provide a transistor model which 'self heats' according to the power dissipated. For clarity I shall use the term self heating for such devices.

Exploring the FDP038 PSpice model

Having downloaded the datasheet, their PSpice model file and AN7532/33 application notes, I redrew the model as a conventional schematic, simplified as above but complete with stimulus, supply voltage and loads, enabling the model to be run and tested. This schematic, **Figure 2**,must provide the model descriptions in the FDP038 PSpice electrical model netlist.

Three Level-1 MosFets are used connected in parallel to model the gain block for the full current range, from the sub-threshold region to the devices limit. The M1_strong MosFet provides the curve for the square law and saturation regions, M1_Med fills in from the sub-threshold to the point when M1 takes over. M3_Weak provides the small currents required near zero volts and initiates the sub-threshold curve.

The Level-1 data set applied to these MosFets is severely restricted, temperature is permanently fixed at 25°C, both W and L are set to 1 micron, reducing the Level-1 g_{fs} equation to KP. Using essentially just two parameters, threshold voltage and KP for each MosFet, facilitates attaining the desired transfer curve in the subthreshold and square law regions. A small value of source resistance can be used to flatten the curve from individual MosFets as gate voltage increases, smoothing the transition from one MosFet to another. The settings used for TOX, IS and N are chosen to minimise convergence difficulties.

Having restricted the MosFet characteristics to provide a gain curve locked to 25°C, we must separately provide the variations in threshold voltage, source and drain resistance with temperature and voltage. Three Spice2 voltage controlled voltage sources E1, E2, E3 and one PSpice function source E4 are used, so must be compatible with your simulator.

The small network of components bottom right in the schematic uses three resistors each having specific temperature coefficients. Used with E2 and E3 they adjust the threshold voltage with temperature by adding or subtracting a small correction voltage to the gate voltages. At 25°C these resistors remain as 1Ω, the network is balanced so E2 and E3 make no adjustments.

With the circuit still set to simulate at 25°C, the resistors in the source and drain leads can be adjusted to match the 25°C transfer characteristic curve. Resistor R5_source, acts to rapidly flatten the transfer curve with increasing gate voltage. Two resistors and E4 in the drain circuit are used to control the transfer curve at high drain currents. The resistor R1-drain reduces current in the linear region.

The multiplier X in the E4 equation ((1E-6 × X) also the exponent '10' of the power statement) models the space charge limiting effect. Reducing X or the exponent '10' rounds off the curve at high currents. The modelling needed to replicate the MosFet's 25°C transfer curve is now complete.

When simulations are run at temperatures other than 25°C, the three resistor network, bottom right in the schematic, becomes unbalanced according to the TC of each resistor and the simulation temperature. E2 and E3 now adjust the threshold voltages by adding or subtracting a small voltage to the gate voltages. To aid understanding, I ran a few simulations to ascertain the level of these applied

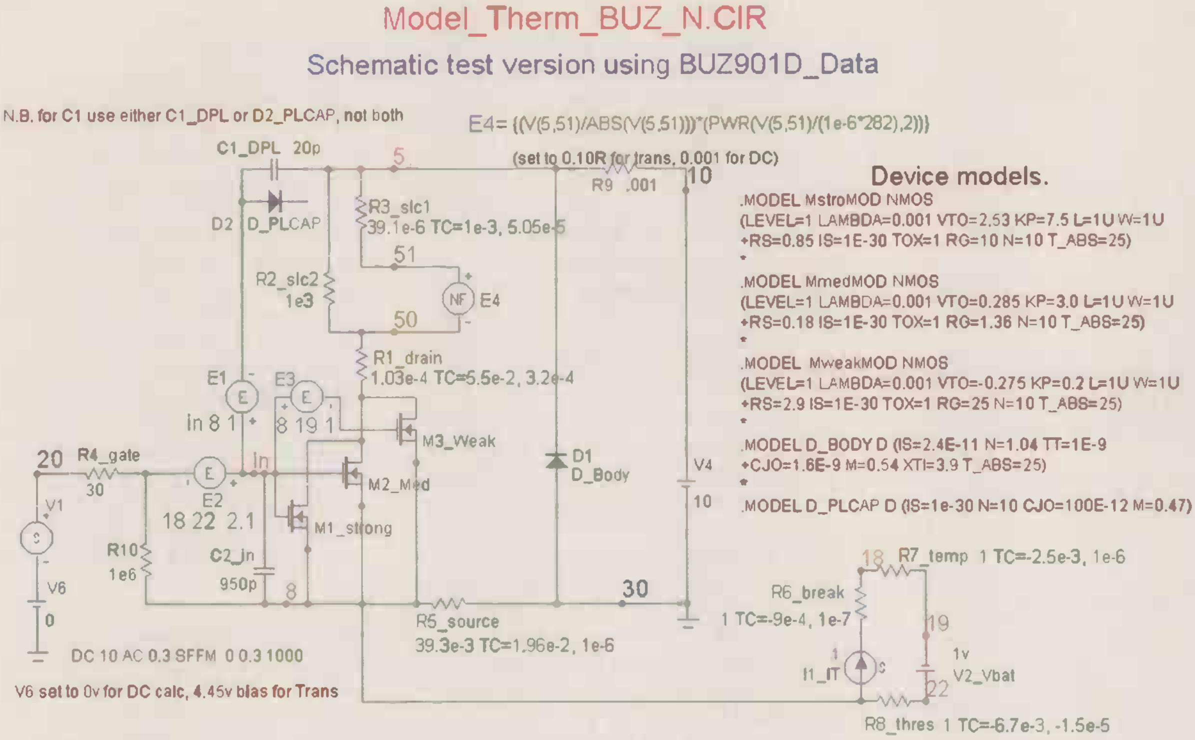


***Figure 6:** Final schematic circuit model for a BUZ900/01D power MosFet showing the values used to model the R_Source also the resistors in the drain circuit. The gain setting for E2 was increased to 2.1 times to match the change in threshold voltages with temperature. Also shows the polarity needed for D2 if used to replace C1 capacitor.*

corrections by temperature. The results are listed in figure 2, directly above the resistor network.

Final proof that this schematic version of the model works correctly can be seen in the attached plot of temperature/transfer curves, which comply particularly well with those shown in the datasheet, especially in the subthreshold region, see **Figure 3**.

The simplest way to craft our desired lateral 'N' MosFet subcircuit model is to save and rename a copy of this FDP038 working schematic, using its settings as starting values for the new model. I decided to model the BUZ900/01D transistor[3] or its direct equivalent, the Exicon EC-20N16/20[4].

A BUZ900/01D non-self heating model

The MosFet labelled M3_weak is adjusted first, setting its threshold voltage just above or below zero volts as needed with a small value, 1 or less for KP. Paying attention only to the lowest current section of the datasheet transfer curve, running DC simulations at 25°C and plotting the result as adjustments proceed, as shown in **Figure 4**.

MosFet M2_med is then adjusted, starting with a threshold voltage around 0.5 volts and KP around two to three. Adjust these values to match the transfer curve from some 0.5 volts to perhaps two volts, simulating and plotting as before.

Finally adjust MosFet M1_strong, starting with a threshold voltage around two to three volts and KP perhaps five to ten. Adjust these values to follow the transfer curve from some 2 volts up to some 5 or 6 volts gate-source, simulating and plotting as before.

Should you find any of the above curves increasing too quickly at voltages above their threshold but cannot reduce the value of KP and still match the curve at lower voltages, simply add a small amount of source resistance in the relevant MosFet model, to flatten its curve with gate voltage. Any source resistance used within these Level-1 MosFets remains fixed and cannot change with temperature, so use the minimum values needed.

My chosen devices are 'double die' types packaged in a single TO-3 case, so exhibit increased transconductance compared to many transistors. When modelling other types, you may need to adjust downwards my suggested starting KP values.

Having attained a good match with the transfer curves up to say 5 to 6 volts, slightly more perhaps with single die lateral MosFets, you will probably find the curve continues as a nearly straight line, indicating excessive drain currents for gate voltages of 7 and above.

This is easily corrected by adjusting the temperature dependant source resistance R5_source, a major contributor to the MosFet $R_{ds(on)}$. This should significantly flatten the curve but will not produce the rounding off needed as the MosFet approaches its saturation region. That is controlled by the two resistors R3-SLC1, R1_drain and the PSpice function source E4.

(V(5,51)÷ABS(V(5,51))) × (PWR(V(5,51) ÷(1e-6×220),10))}

This E4 function is controlled according to the voltage drop between nodes 5 and 51, the voltage drop across R3_SLC1, using the expression shown above as used in the figure 2 schematic. In Spice2 terminology the voltage between two nodes is described as e.g. V(5,51) signifying the voltage drop across the resistor R3_slc1. Both produce a large, dramatic rounding off effect. Initially I find it best to vary R3_SLC1 in small amounts then adjust the value in brackets (1e-6×220),10). Small changes in the '220' value produces modest flattening and rather more dramatically when adjusting the power '10'. With some models this power originally shown in figure 2 as a '10' factor for the FDP038, may also have to be reduced to avoid excessive 'floating point range exceeded' errors.

To expedite this process, I enlarged and printed the datasheet graph as large as possible, then using ruler and calculator noted the relevant drain currents by voltage on my printout. When satisfied your 25°C plot replicates the data sheet, plot the curves for higher and lower temperatures, by running simulations with the temperature reset as needed. See **Figure 5**.

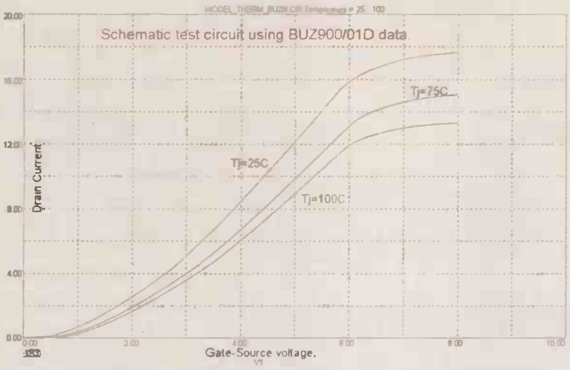


Figure 7: Using the values shown in figure 6, these temperature plots display particularly good agreement with the datasheet for drain currents up to some 12A.

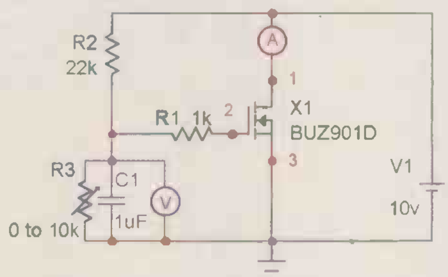


Figure 8: The simple test circuit used to measure the subthreshold region transfer curves, providing additional data points for this critical crossover region. Measuring a 'P' MosFet the supply polarity should be reversed.

Figure 9: Using the final values for the 'N' device schematic model as starting values for the 'P' device schematic quickly produced these curves, closely replicating the datasheet subthreshold conduction, to complete our pair of schematic models.

Before explaining this sequence, it is now appropriate to further explain the circuit, I1_IT, R6_break, R7_temp, R8_thres, and V2_bat, bottom right in the schematic.

The three resistors each have a value of 1Ω and an associated **TC=** xxx instruction. The first number following the TC= indicates a linear temperature coefficient, the second a quadratic coefficient, translating as <value> RΩx(1+TC1(T-Tnom)+TC2$(T\text{-}Tnom)^2$). When simulating at 25°C these TC coefficients have little effect, resulting in a near 1Ω value for each resistor, so E2 and E3 have little effect on the threshold voltages and the gate source voltage controls E1. With higher or lower temperatures, these resistor values increase or reduce in line with their stated coefficients.

To help my understanding I plotted the effects of temperature on this network, which are tabled in the figure 2 drawing. Increasing or reducing temperature produces the voltages used to control E2, and E3, which now replicate the effect change in temperature has on the threshold voltage. Used with a suitable capacitive diode model for C1_DPL, E1 results in the dramatic change in the gate drain capacitance with gate voltage already discussed. Used with a capacitor as shown, it has no effect and capacitance stays constant.

Run and plot simulations for the upper and lower datasheet transfer curves. In all probability these will now approximate the datasheet curves. If not small adjustments to the TC values for R1_drain, R3_SLC1 and R5_source, should suffice. You may also need to modify the 'gain' values for E2 also E3. Changing these or the resistor TC values should have little or no effect on the 25°C curve. See **Figure 6**.

As a final check plot the transfer curves by temperature in sequence as one plot simply by 'stepping' the plot temperature using a temperature list, e.g. 25, 75, 100, see **Figure 7**.

Improving subthreshold data

It is possible you may feel, as I did, the curves for small values of gate voltage and drain current should better match your devices. Most makers datasheets are produced assuming 25°C, using extremely short test pulses in an attempt to avoid transistor self heating and retain the junction at this temperature. Lacking a suitable pulse measuring method, pulse testing was not an option.

After some consideration I decided to mount my devices onto a known heatsink using a known thermal washer. I would measure drain currents using a steady 10V DC drain-source voltage, adjusting the gate voltage as needed to obtain various currents, from less than 1mA to the 3A maximum which interested me, then calculate the expected junction temperature for each measurement. By increasing the simulation temperature slightly above room temperature to match each measurement temperature, I could now tweak the '25°C' subthreshold curve for each measurement, to match my measured drain current values.

I measured six pairs of devices, obtained from three different batches. I found each MosFet did conduct slightly, even when the gate was directly connected to the source, i.e. zero gate-source voltage. Choosing the values from an intermediate device I amended the transfer curves by small adjustments to the threshold and KP for all three transistor models. Ensuring the best possible model in the subthreshold region for realistic simulations of crossover distortion.See **Figure 8**.

With models now closely aligned to actual measured values for low drain currents, I was forced to accept some divergence at currents approaching the saturation region. For my needs that matters little, since I don't expect to use such high drain currents.

'P' Types

Following the above process but with the polarity of some components reversed in the schematic, I produced a model for my BUZ905/6D 'P' MosFets. Having completed my initial learning curve and using the values developed for the 'N' model as a starting point, this 'P' model was developed rather more quickly and easily. See **Figure 9**.

However, when I measured the drain currents for my 'P' samples, I found notable differences compared to the 'N' types. The 'N' types conducted some 0.25mA at zero gate voltage and current increased steadily with each small increase in gate voltage. The 'P' types also conducted some 0.25mA at zero volts but their current did not increase until some 0.3V gate-source voltage was applied. This approx. 0.3V difference increased slightly with each voltage step, becoming nearly 0.65 volts for 2A drain current. As if an additional Schottky diode existed in the 'P' device source path.

This voltage difference meant the threshold voltage for the weak MosFet had to be moved significantly to model the 'P' type, compared to that used for the 'N' type.

Netlist

The next stage in producing our final subcircuit model is to export this schematic netlist into a simple text editor such as Windows notepad. One which does not insert control codes.

First we should assign the node numbers we wish to use for our subcircuit, to a printout of the schematic. As already explained the node numbers connected to the gate, drain and source electrodes must match those specified in the first subcircuit line also the schematic shape we intend to use.

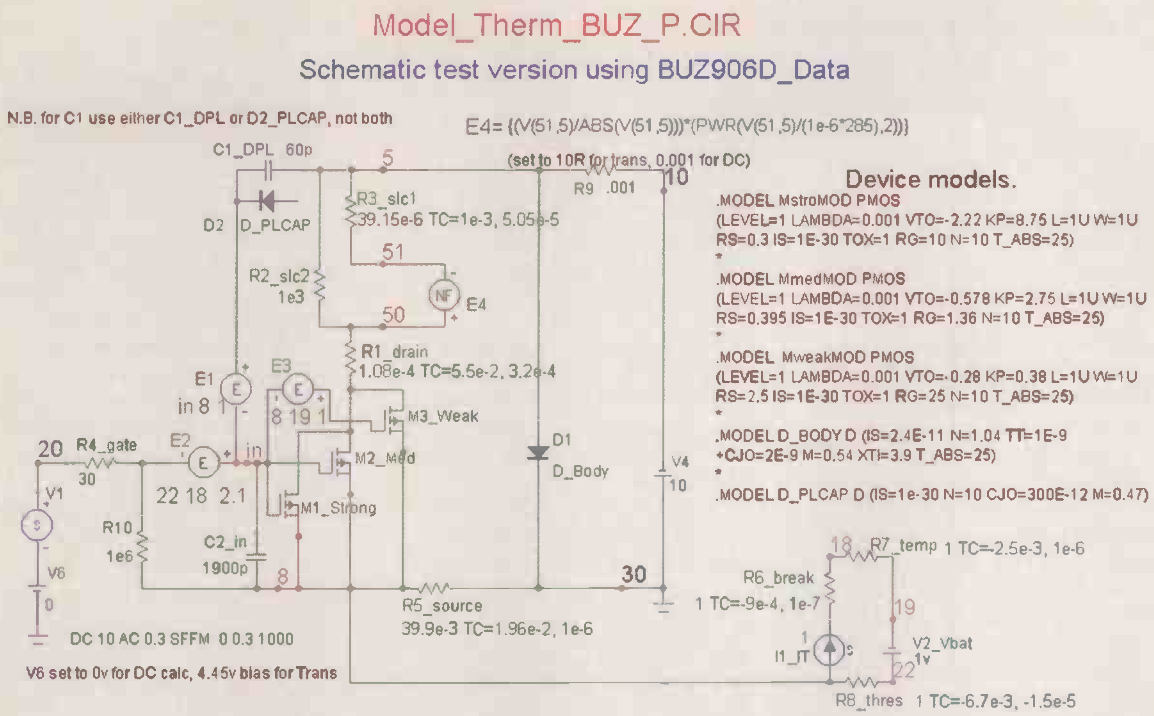


Figure 10: Assigning 'fixed' reference nodes to our 'P' MosFet schematic, facilitates converting from this schematic format to the netlist required for the final subcircuit model.

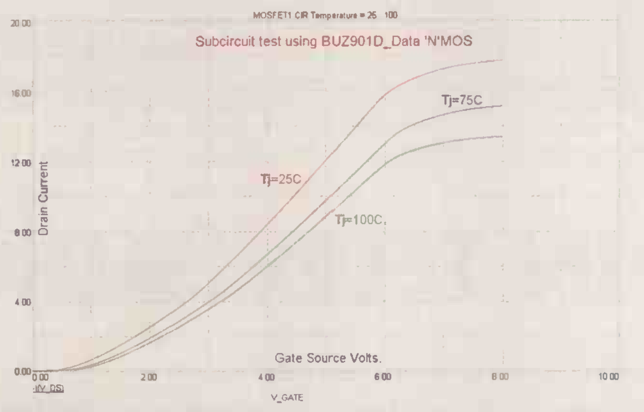


Figure 11: Having completed the netlist needed for the subcircuit models, we can at last test the finished subcircuit models. Using MC6 I found some final tweaking was needed.

In all probability your Spice2 schematic software will already have assigned these nodes, which is why my drawing shows apparently odd nodes, in larger fonts, e.g. the '5', '51', '50', '8', etc on my schematic. MC6 and possibly your own simulator automatically assigns circuit nodes, any circuit change results in the nodes automatically renumbering. Hence where needed for calculations or to plot results, to avoid confusion or errors I assign these named nodes which do remain constant. See **Figure 10**.

Your exported netlist will include some items not needed for the subcircuit model, also the automatically assigned nodes will not agree with the nodes used in the formulae or the schematic shape nodal requirements. Manually edit the nodes in the netlist and formulae as needed, insert the opening statement and .ENDS etc needed for a subcircuit to complete our subcircuit model, then insert into your library.

Having added the subcircuit into the library as required by your particular simulator and explained in my last article, you assign the model to a schematic 'shape' using the shape routines supplied by your simulator. You can now import the model into a drawing, just like any other component. Assign stimulus and voltage sources together with any required test load and test run your model. A small amount of model tweaking, by amending and resaving the subcircuit netlist, resulted in my final models. See **Figure 11**.

Convergence

As stated in my last article, convergence is a perennial problem using Spice2 other than with the simplest models. This model, despite it's features, is relatively easy to converge, however should problems arise then adjusting your Spice2 defaults away from their default settings should provide a solution. **See table 1 on page 32.**

Subcircuit rules.

A subcircuit can easily be called from a net-list:- Simply by entering its name and nodes,
X<name> {nodes} <subcircuit name>
e.g. X12 10 20 30 BUF905D

Calling a subcircuit from a Windows schematic is even simpler. All subcircuits in your library will have been assigned a representative component 'shape'. Any subcircuit included in your library is simply selected and its 'shape' inserted into position in your schematic exactly like any conventional component, resistor, capacitor, transistor, op-amp etc. This schematic shape will initiate the subcircuit call for you.

Using a subcircuit model to circumvent the limitations inbuilt in Spice2 is not new, if you look you will see that Op-amps are modelled as relatively large subcircuits even though Spice2 does include the primitive components needed to model a simple op-amp within its core software.

When creating a subcircuit for our power MosFet, the subcircuit nodes representing its drain, gate and source electrodes must be assigned in the first line of its netlist. The last line must be terminated with .ENDS and not .END as is usual for a netlist:-

```
e.g. .SUBCKT BUF905D 10 20 30
* DRAIN, GATE, SOURCE
* any required explanation e.g. modified 12/10/03
* an asterisk at the beginning of a line indicates a remark statement.
C1 9 5 50p
D1 3 30 D_BODY
M1 1 2 3 3 MstroMOD ;then follows the rest of the subcircuit netlist lines
*
```

Global Defaults, amended settings as used.

Name	Default	Revised	Simulation controlled
ABSTOL=	1pA	1µA	maximum current accuracy
CHGTOL=	0.01pC	1pC	maximum charge accuracy
ITL1=	100	1000	Permitted iterations for DC analysis
ITL2=	50	1000	Permitted iterations for DC analysis
ITL4=	10	1000	Permitted iterations for transient analysis
RELTOL=	0.001	0.01	Relative accuracy of voltages - currents

Thermal modelling

If you obtained the datasheet for the FDP038 from Fairchild, you will notice they claim it provides both electrical and thermal models. The electrical model is certainly fine and runs extremely well producing unusually good agreement with the datasheet curves. However the claim to also provide a thermal model in my view is optimistic. It is true they do supply a model of the thermal path from the transistor junction to the outside world, but the format used makes attaching a heatsink difficult.

I did many experiments but could find no way this thermal model could be made to control the junction temperature. Without additional components it could not even be used to monitor the junction temperature.

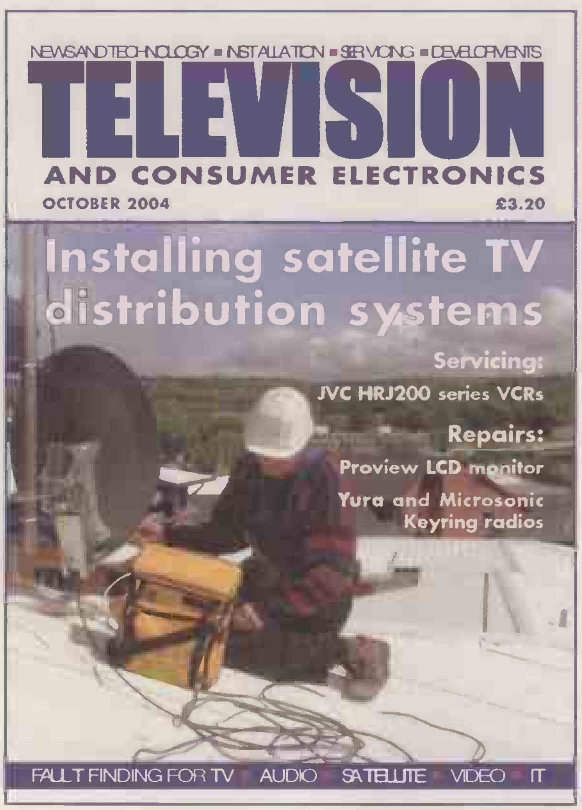
Obviously while the subcircuit model described in this article can be used to model distortion at various junction temperatures, the most desirable model of all would be one which 'self heated'. Fitted with an electro-thermal representation of a heatsink and insulating washer, not only could transistor junction temperature be measured, the MosFet characteristics also would change in real time with change of junction temperature, even during one cycle at 1kHz, just as in a real transistor.

These improvements are possible and form the subject for my next article. I have already produced working self heating thermal models for my pair of BUZ transistors. As will be seen, these thermal models do provide distortion simulations satisfying close to those measured on my modified Maplin amplifier. However these thermal models must use behavioural modelling elements which may differ from or even not be available in your Spice2 simulator.

References

1. FDP038AN06A0 spice model. **http//www.fairchildsemi.com/models/PSPICE/Discrete/index.html**
2. Infineon Technologies **http://www.infineon.com/simulate**
3. BUZ900/01D lateral power MosFet. **http//www.magnetec-UK.com**
4. EC-20N16/20 lateral power MosFet **http//www.exicon.com**

NEXT MONTH IN TELEVISION



TEST REPORT: **the DigiFusion FVRT100 DVR**

The FVRT100 is a hard-disk recorder for use with DTT reception. Roger Thomas bought one to replace his elderly VCR and gives it a thorough check out.

PROJECT: **an electronic stethoscope**

This simple but efficient design uses an electret microphone for sound pickup, a TL431 adjustable regulator (comparator) for amplification and a MOSFET as the output stage. There are all sorts of uses for stethoscopes. In particular they can be helpful for activity detection when fault-finding.

THE LUNEBERG LENS AERIAL

J. LeJeune describes an unusual aerial for satellite signal reception.

VINTAGE REPAIRS: **the Collaro Conquest autochanger**

Michael Maurice describes the operation of this record changer, which dates from the Fifties/Sixties, and what's involved when repair is necessary.

Available October 20

'Mixed Spices' Part I

Spice (Simulation Program with Integrated Circuit Emphasis) was originally developed by the University of Berkeley in California around 1968. First used mostly by academic types to simulate and optimise the behaviour of semiconductor devices, gradually it became less dependant on the original arcane text-based language as it was integrated with schematic capture and graphing capabilities, and available in ever more complex PC/Windows based programs. Recently the number of programs offering Spice simulations has increased until there are nearly as many flavours as on your supermarket shelves. Alistair Macfarlane of Electric Fields Design Consultancy, a long term Spice user, reports

Analysing many different Spice programs is a somewhat daunting undertaking. Each and every program designer has their own idea of how the schematic capture, layout, help, graphing and modelling should be done, never mind the maths behind it all. This involves solving a large number of differential equations using matrix algebra, and running various iterative processes until the (potentially many) nodes of the circuit all settle down ('converge') to prescribed limits.

From a long web surfing session I identified 12 programs which looked promising to examine and if possible, test. Some of the other more complex (and expensive) workstation based or IC design types were not tested, due to both time constraints and a suspicion that these would be of less immediate interest to readers.

To learn even the basics of any new Spice program thoroughly requires at least a day or two of effort, and the features in some of these programs can be so complex that even then it's only possible to scratch the surface of what they can theoretically do. Spice provides many different types of analyses, such as DC, AC (e.g. Bode plot, phase shift) Transient analysis, frequency domain, noise, distortion, digital, complex maths such as RMS values, Nyquist control loop stability, Smith charts for RF and so on. Sensitivity /MonteCarlo/worst case/optimisation analysis is sometimes provided to help with tolerancing a design. Most have the choice of integration methods from Trapezoidal, Gear and sometimes Backward Euler. I have been working with Spice since the mid-80's in my capacity as a circuit designer and latterly a consultant, and still find features I have not used - yet.

With so many programs to examine, ease of learning was considered to be an important part of the process. I quickly found that many programs have so many layers of menus, options and choices that it is easy to miss the function you might want because it is hidden several layers down behind some button which perhaps does not appear immediately relevant. Personally, in a design business rather than academic application, I want to use Spice both to design and to investigate the operation of a given circuit quickly, easily and reliably, and then go on to build and test a prototype (if possible), and then PCB design. A good simulation can find and take out the question marks and perhaps save a design iteration or two. And circuit elements in Spice do not burst into flames or explode when overloaded, causing damage to either the circuit or designer!

With Surface Mount components becoming smaller by the year, the days of prototyping boards and hanging a scope probe on a convenient wire-ended resistor are gone; it's much easier if you get it right first time! (Most of the programs provide a netlist output for PCB design.) With that kind of schedule I may not go back to using Spice for some time, so when the next design pops up I don't want to have to relearn some quirky program from scratch again. Nor as a circuit designer do I wish to have to become a software guru, or have to have to call on an IT department (or the program's tech support department) for help. I dare say there are many other applications where Spice is to be used more constantly and familiarisation is not so much of a problem. To assist with learning, a few programs provide animated runs to show how some of the functions can be carried out. Others rely on a short tutorial and some solely on plodding through a help file, hoping that the relevant terms can be found in the index – not always the case!

Most of the Spice providers offer demo programs with circuits which they have created, and it would be easy just to run some of these and treat the results as a guide to performance. However being a cynical engineer I of course wondered whether these demo circuits could have been massaged so the bugs have been removed (actually, not every time!) and there is even the possibility of using already prepared files which appear to run. Often the makers provide these demos with limited time, size or model database; a few offered full versions for the purpose of my test.

To try to make sure I was comparing as realistically as possible, I decided to try to create a simple mixed-signal circuit to pulse-width modulate a coil, which hopefully would be within the limits of the demos available (albeit – theoretically - almost insultingly easy for some of the programs!). I could then see how easy - or difficult - it might be to capture this to schematic, run the simulations and examine the data produced using a transient analysis, which to me is the more interesting - and more challenging to the software. My experience of Spice has shown that oscillators in particular can be problematic due to the strong feedback implied, and indeed this proved to be the case. (The circuit has two, one analogue, one digital, feeding a D-type Flip-flop divide by 2, which in turn drives an npn transistor controlling the inductor current). The flywheel diode used was a deliberately 'bad' choice being a slow recovery type.

The Spice programs to be tested were 5Spice, B2Spice, Edwin, Icaps 4, LTSpice/SwitcherCAD, Microcap 7, Multisim 7, Orcad/Pspice, Simetrix, Superspice, Spiceage/Spicycle and Visual Spice. Not tested were programs like HSpice, Accusim, Smart Spice, Dolphin Spice, Spectre, nVisage etc. as I consider these are more for the big companies.

5Spice

5Spice is a collaborative effort between an Englishman, Mike Smith and an American, Richard Andresen, 30 years an analogue design engineer. Mike developed the WinSpice engine and Richard the schematic capture front end. The program demo is a little minimalist and I suspect 5Spice may be at quite an early stage of development. It has only limited if any mixed signal/digital capability and no obvious FFT and advanced features like Monte Carlo (although the registered version claims to have noise and distortion, which could imply some sort of FFT is available). To me, one of the more useful features during a Spice transient analysis is to have the program graph the results at the same time as the simulation is running. (These are known in some programs as 'Marching Waves'.) 5Spice does not appear have this feature. During a long simulation, it is important to know what is happening, so that if things are going hopelessly wrong it can be aborted.

Unfortunately I was unable to get even the simplest part of my circuit (the opamp relaxation oscillator) to run properly. A 5V supply, 4 resistors, a capacitor and comparator managed to generate –1.6kV in a 100μS run. (Useful perhaps as a source of renewable energy, but unfortunately not exactly accurate!) And to make my case for marching waves, on one attempted run, 5Spice told me the simulation was running but after waiting a goodly time looking at this message, I found the computer had hung! And hanging up was something which happened all to often with various programs to be discussed later. Life is too short to have to watch your computer reboot and Scandisk run through multi-GB drives; I quickly got bored with making yet another cup of coffee. I feel that such basic faults should have been ironed out well before the programs concerned were put on the market. So I had to be content with running their astable demo.
The screen grabs are shown in **Figures 1 & 2.**

In fairness to 5Spice, the front end is fairly intuitive to use and construct a circuit, with a component toolbar down the right side of the screen. But there is no auto-numbering of component references; each has to be put in manually. And junctions of more than two wires need a junction point to ensure connection; all too easy to forget. As well as this minor irritation, opamp shapes come either as three wire (no supplies) or an 7-wire version which has only 4 pins identified. 'Dangling pins' like this need a no-connection symbol. There is no facility for expanding graphs other than one cursor. All rather basic, but then 5Spice costs just $169. They can be found at www.5spice.com.

B2Spice

B2Spice is created by Beige Bag Software in the USA. I wondered where the name came from; it turns out that they originally wanted to call it 'Brown Bag' to signify good value for money but as that was already taken settled for the present name! I first tried to run the demo but was getting nowhere so the UK distributor arranged for a full version 4 'Pro' to be made available to me. At one stage I was asked to try a beta of their latest version 5, but that proved much too buggy, so after several crashes I reverted to V4. This is a much more advanced program than 5Spice, but does not support FFT and other DSP functions; that is still to come in V5, along (I am told) with many new features. Digital is limited, and the digital models proved to be somewhat eccentric, having a dead band between 0.5V and 4.5V where it was considered an illegal level and everything stopped. However the boys at Beige Bag bent

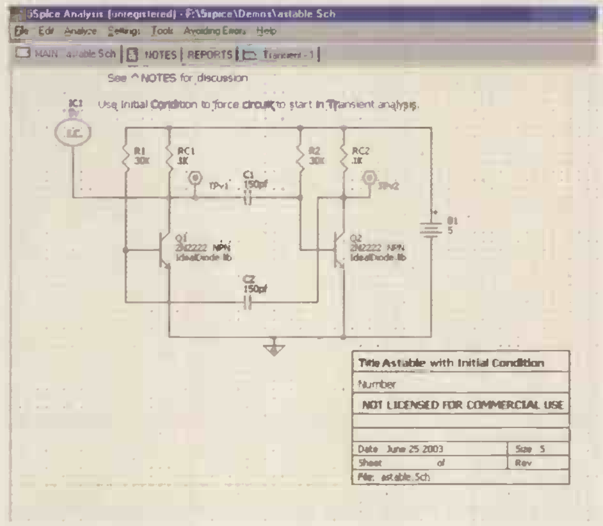


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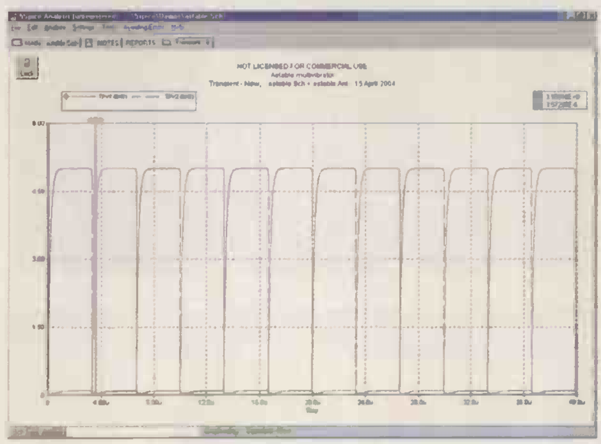
Figure 2: below



over backwards to help me and with some tweaking got my circuit to run. But even then not all was well; curious artefacts appeared, namely dips in the transistor collector voltage trace without any base drive (see screen dump). These are probably due to ringing between the slow diode and 'perfect' inductor, but this goes to show one of the maxims of using Spice; often you need to know pretty much what the answer should be before you can rely on your results! The simulation took a slow 16 seconds to run!

Creating the schematic was reasonably intuitive; there was no component toolbar but the components were listed in categories under the Devices menu. Unfortunately sometimes this meant a long scroll down through many components to find the desired part. And placing a part and changing your mind was a pain; the 'Esc' key didn't get you out as I'd expect, and the only way I could find was to continue to place it then delete! And rotating required Ctrl-R, a rather clunky

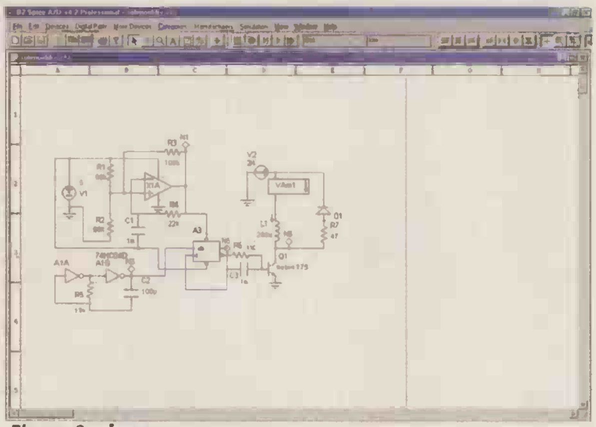


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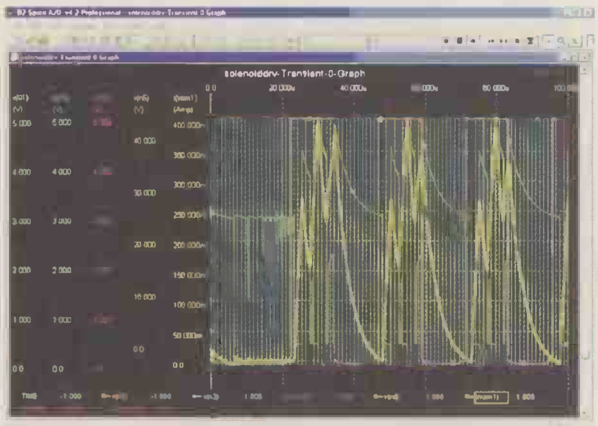
Figure 4: below



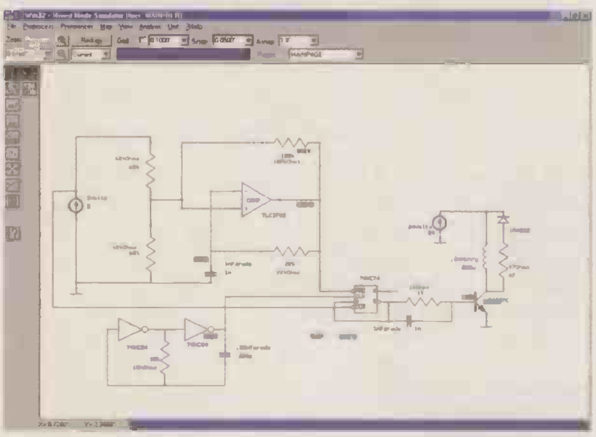
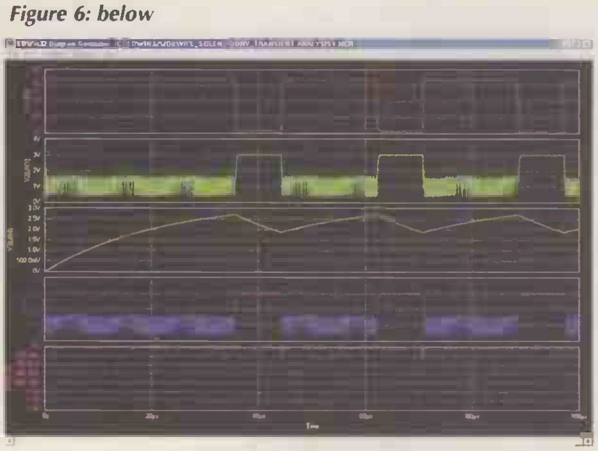


Figure 5: above

Figure 6: below



method. I found that after I had enabled the component names (such as the 74HC04) for some reason all but that name disappeared from the schematic, and nothing I have been able to do since has persuaded them to return! At least it didn't hang.

Graphing produced a nice set of (marching) traces but rather limited in that the left y-axes are all 'sort-off' commoned (but not quite) and the autoscaling doesn't leave any headroom. It's hard to see, but the first dip in the collector trace actually goes negative. And you are limited to just one cursor. In my experience two are desirable so that the delta can be measured. Apparently it is possible to tile the traces, but 'a bit involved'. But otherwise the trace expansion worked well and I could look into the switching edges easily. The failure of the transistor to stay saturated with the extra current due to the diode's slow recovery time was clearly seen in the collector graph.

B2Spice is available from RDResearch (www.spice-software.com) costing £139 for the standard and £239 for the Pro versions. The screen grabs are shown in **Figures 3 & 4**.

Edspice

Edspice is part of the Edwin suite of EDA programmes from Visionics, originally a Swedish company. It is bundled with schematic capture, mixed signal simulation, thermal analyser, electromagnetic analyser, signal integrity simulation and PCB layout. I purchased the non-commercial version out of interest but had never explored the Spice capability. There are a number of demo circuits which run OK, so I tried to build my 'standard' circuit using the capture module. It is not an easy program to learn due to numerous similar looking icons and quirky menus. Also the net and wiring are apparently separate. When wiring up the net and wires coincide, but deleting the wire does not delete the net; this has to be deleted separately.

With all these features, the program gives the impression of being potentially quite powerful, but is not intuitive or easy to understand from the long-winded help files. The first part of the circuit - when eventually created - did apparently run but without producing a graph. I contacted Visionics who sent me a quick reply (from India) attaching two files containing sections of my proposed circuit. Unfortunately these were for Windows XP, and I use 98SE, so that didn't work. I created my full circuit and sent it off again to Visionics to see if they could get it to work. The circuit arrived back after a few days and I was able to run it in an encouraging 7 seconds. But the results were completely wrong. The digital oscillator waveform was weird (there seems to be some interaction between the two oscillators) and the voltages too low to switch the transistor. So 'nil pointes' for Edwin, I'm afraid. The screen grabs are shown in **Figures 5 & 6**.

Icap 4

Icap 4 is a somewhat scarily complicated Spice program suite from Intusoft, a US company. After the (by now usual) unsuccessful attempt to use their demo, followed by several e-mails for help to their representatives in the UK, Technology Sources Ltd., I was able to create my circuit and run it. In fact I was limited to 20 components in the demo, but the IC models used up an unknown number of sub-circuits internally, so the representatives created special models to stay within the limit. There are a number of quirks in the schematic capture, which no doubt regular use could make easier. Dangling pins – even an output - are a no-no; you must attach a wire to the pin, even if it doesn't go anywhere! Selecting a component or text for moving or rotating involves the slightly annoying delay of holding the cursor clicked on the part for a second or so, otherwise it doesn't move. Commands and functions are hidden under layers of menus, but for the power available it could be worth the intellectual effort (and patience) to find them. Unfortunately this complexity does not make it very intuitive; running a simulation throws up a bewildering array of three Windows on top of the first one, including a set of thumbnail marching graphs. In fact Intusoft are one of the vendors who produce a set of animations of the use of their program, and in one of these they mention that 'this problem is being addressed'. It needs to be.

Buried underneath it all is the menu to find Intuscope, the 'real' graph program. This is also quite powerful but not very intuitive; at least in the demo, traces have to be selected each time, but the default is to tile them, which makes for easy viewing.

Icap can have the thumbnail probe waveforms included in the schematic, as shown on the next page. The graph window below is not easy to find but once expanded to full size has a number of useful functions such as expanding the traces, changing scales, math functions and so on. The screen grabs are shown in **Figures 7 & 8**.

The simulation took a rather mediocre 9 seconds to run. (At one stage I changed the Set and Reset connections to the D-type and on the next run the waveforms went soft edged. But after saving and re-running the correct shapes returned.) After that the waveforms from Intuscope can be post-processed using various scripts using the ICL language, again very powerful but not to my mind user-friendly. Many templates for these scripts are provided, however. The database is claimed to hold over 18,000 models including such esoteric devices as Hall-effect. Intusoft also appear to specialise in SMPS and magnetics design, and have a number of 'averaging' models which can assist in SMPS control loop optimisation without the pain of waiting while every high-frequency clock cycle is simulated. A neat feature of their SMPS design is the ability to superimpose the statutory EMI limits on the FFT graphs. And a filter design tool option is also provided. Monte Carlo and optimisation etc are available to those savvy (or brave) enough to delve into the scripting language. Another neat idea is the provision of a description of the component when looking to select a part, for example a Mosfet has the voltage, current and Rdson specified, speeding the selection process without having to pick up a databook. There are far too many other options available in Icaps4 to be covered in the scope of this article. Intusoft produce quarterly newsletters with lots of useful hints. Icap 4 comes in numerous different versions, from the entry level Icap/Rx at £530 to the professional at £2860. Deals are available for educators (www.technology-sources.com).

LTSpice/SwitcherCadIII

LTSpice/SwitcherCadIII is a freeware program produced by Linear Technology. It was the surprise of my testing; a well featured Spice program covering AC, DC, Noise, Transient analysis and even FFT for the just cost of a download! Intended to be used for designing with Linear Technology SMPS parts, Switched capacitor filters, and other analogue components, this little program runs very well. Especially since Linear's designers have in some way optimised the switching models to run very fast so that the much slower control loop time constants can be established without having to wait a lifetime. But even without these special models, the program ran the solenoid simulation (with marching waves) in just 2½ seconds! And Linear's models of capacitors etc. can be 'real' with the parasitics such as ESR and ESL included. There is however no Monte Carlo analysis and another limitation is that there seems to be no provision for cursors on the graphs. Unfortunately the only diode models available were fast recovery types so I couldn't see what it did with a slow diode.

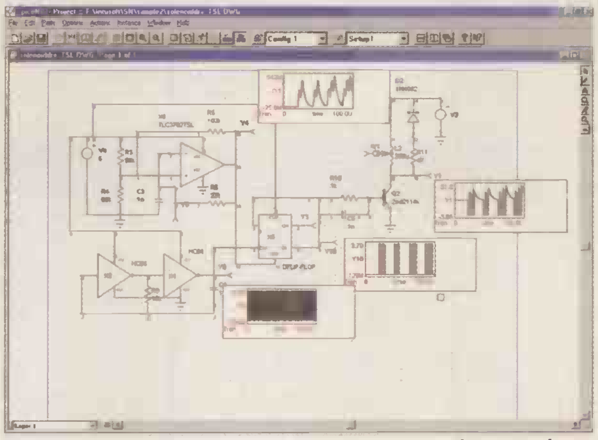
There was a difference in the frequency of the Schmidt oscillator from what was expected using a 'real' 74HC14 (2.2MHz vs. 485kHz) but this can be attributed to too small a value of hysteresis in the model. I couldn't get the two-gate oscillator to run; perhaps that's a challenge for someone with more time? In fact its 'Digital' modelling is a little limited but there are adequate models for playing with and the possibility of importing models, which could expand its use considerably. LTSPice has an active user group (http://groups.yahoo.com/group/Ltspice) and can even export netlists to various PCB design programs. The screen grab is shown in **Figure 9**.

All things being considered, an excellent entry level program for anyone interested in getting into Spice. And did I mention - it's free!

Multisim 7

Multisim 7 is the Spice simulation part of the Electronics Workbench EDA suite of programs; it claims to have 16,000 models and to cope with VHDL and Verilog as well. I struggled with this from the start. Its demo does not allow a save, so that each time I had to start again meant rebuilding the circuit from scratch. It hung the computer on too many occasions (even twice in the middle of schematic capture, with a 'Run Time' error), which of course meant every time starting again after a reboot; this quickly got more than just boring and eventually I gave up. Capture is slightly quirky and erratic, in that you cannot connect more than one wire to a component pin, and I found wires came and went rather arbitrarily. Component values are selected from a table of 'standard' values, and this was limited to 1 Ohm to 22 Megohms for resistors. I never did find how to use (say) a 0.001 ohm resistor for current sensing!

Since I couldn't get my circuit to run, even after a discussion with the technical support (who in this case was concerned that I might be using XP!), I looked at the demo circuits provided. I chose the AC-DC converter to try which is basically a rectifier followed by a LP filter. It ran OK but seemed to ignore the stop time I'd set. The only way to stop it was to press the pause button. And the graphs appear as an oscilloscope trace. The resolution was coarse and the sinewave distorted, in spite of setting a 10μs time step. The only way to view the traces for analysis was to use the view graph command, although at least there were a pair of



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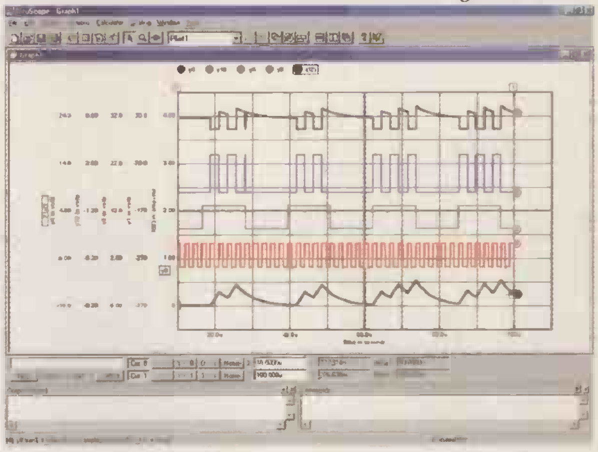

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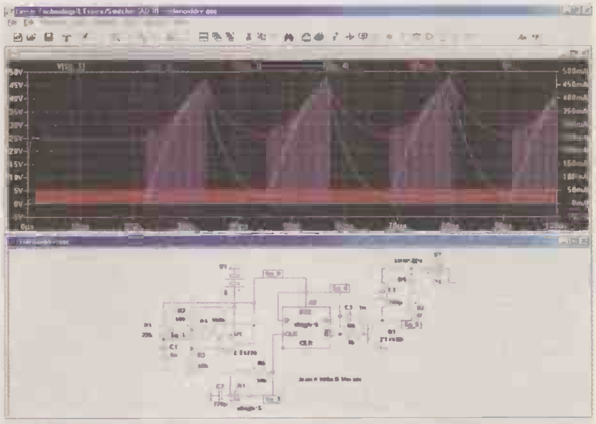
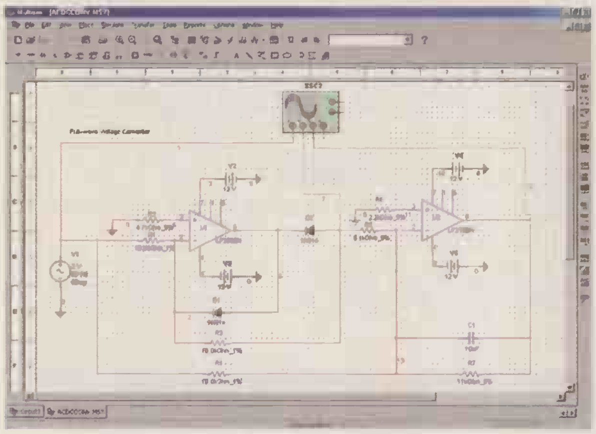


cursors provided. Personally I dislike the use of 'Virtual Instruments' although I can see their value for educational purposes. They tend to clutter up the schematic especially if you need to measure a lot of different voltages or currents. The screen grabs are shown in **Figures 10 & 11**.



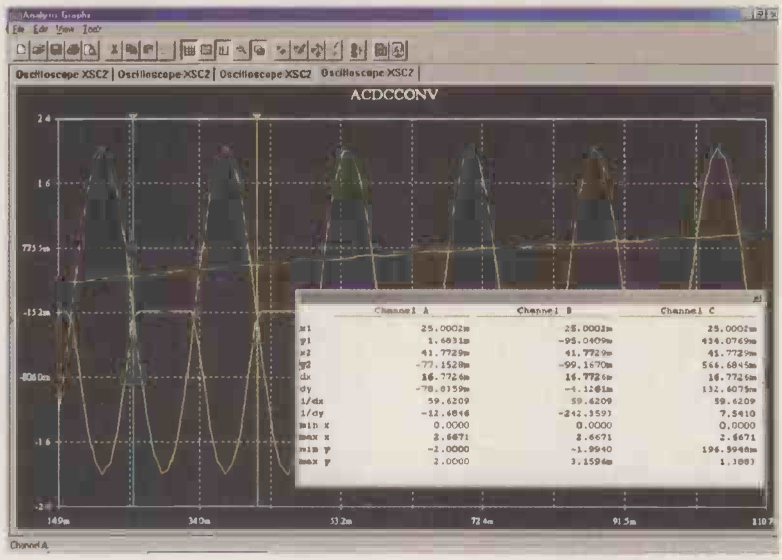


Figure 10: above

Figure 11: right

Readers may be interested in the following further reading, available from the EW book service operated for us by Boffin books at www.boffinbooks.com

SPICE: A GUIDE TO CIRCUIT SIMILATION AND ANALYSIS USING PSPICE: Paul W Tuinenga, Prentice Hall 1992 & 1995. Limited Availability as out of print item - price on application.

ANALYSIS AND DESIGN LINEAR CIRCUITS 3E WITH PSPICE FOR LINEAR CIRCUITS (USES PSPICE VERSION 9.2) SET, Thomas, John Wiley & Sons Inc, 2001, hardback £97.50 0-471-20929-5

BASIC ENGINEERING CIRCUITS ANALYSIS 7E WITH PSPICE FOR LINEAR CIRCUITS (USES PSPICE VERSION 9.2) SET, Irwin, John Wiley & Sons Inc, 2001, hardback £93.50 0-471-20924-4

CIRCUITS 5E WITH ETA & CD WITH FIRST LAB IN CIRCUITS & ELECTRONICS AND PSPICE FOR LINEAR CIRCUITS SET, Dorf, John Wiley & Sons Inc, 2001 hardback £122.00 0-471-21997-5

CIRCUITS 5E WITH HAND ON PSPICE SET, Dorf., John Wiley and Sons (WIE), 2000 hardback £115.00 0-471-43969-X

CIRCUITS ANALYSIS 2E WITH PSPICE FOR LINEAR CIRCUI TS (USES PSPICE VERSION 9. 2) SET, Cunningham, John Wiley & Sons Inc, 2001 hardback £98.50 0-471-20933-3

ELECTRIC CIRCUITS ANALYSIS 5E WITH PSPICE FOR LINEARCIRCUITS (USES PSPICE VERSION 9.2) SET, Johnson, John Wiley & Sons Inc, 2001 hardback £89.50 0-471-20932-5

ELECTRIC CIRCUITS FUNDAMENTALS: PSPICE MANUAL, Sergio Franco; James S. Kang (California State Polytechnic University, Pomona, USA), Saunders College Publishing/Harcourt Brace, 1995, paperback £21.99 0-03-003534-1

ELECTRIC CIRCUITS REVISED AND PSPICE SUPPLEMENT PACKAGE, Nilsson, Prentice Hall, 2002 hardback £72.99 0-13-077079-5

ELECTRONIC DEVICES AND CIRCUITS: EWB/PSPICE DATA DISK, Hassul. US Imports & PHIPEs, floppy disk £9.99 (inc.VAT) 0-13-505009-X

FUNDAMENTALS OF ELECTRIC CIRCUITS WITH PSPICE FOR LINEAR CIRCUITS (USES PSPICE VERSION 9.2) SET, Paul. John Wiley & Sons Inc, 2001 hardback £76.95 0-471-20928-7

INTRODUCTION TO ELECTRIC CIRCUITS 6E WITH PSPICE F OR LINEAR CIRCUITS SET, Dorf, John Wiley & Sons Inc, 2003 hardback £101.00 0-471-48730-9

INTRODUCTION TO PSPICE FOR ELECTRIC CIRCUITS, Nilsson, Prentice Hall, 2002 hardback £15.98 (inc.VAT) 0-13-009470-6

MICROSIM PSPICE FOR WINDOWS: OPERATIONAL AMPLIFIERS AND DIGITAL CIRCUITS: V. 2, Roy Goody (Mission College, USA). 2nd ed, US Imports & PHIPEs, 1997, paperback £49.99 (inc.VAT) 0-13-655804-6

ORCAD PSPICE FOR WINDOWS: DEVICES, CIRCUITS, AND OPERATIONAL AMPLIFIERS: VOLUME II, Roy W. Goody. US Imports & PHIPEs, 2001, paperback £25.99 0-13-015797-X

ORCAD PSPICE FOR WINDOWS VOLUME 1:DC AND AC CIRCUITS: DC AND AC CIRCUITS: DC AND AC CIRCUITS, Roy W. Goody, US Imports & PHIPEs, 2000 paperback £31.99 0-13-015796-1

PSPICE FOR LINEAR CIRCUITS (USES PSPICE VERSION 9.2), Svoboda, John Wiley & Sons Inc, 2003, paperback 17.50 0-471-20194-4

PSPICE FOR SIMULATION OF POWER ELECTRONIC CIRCUITS, R.S. Ramshaw; D. Schuurman (both of University of Waterloo, Canada). Kluwer Academic Publishers, 1996, paperback £72.00 0-412-75140-2

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SPICE FOR CIRCUITS AND ELECTRONICS USING PSPICE SCHEMATICS, Rashid. Prentice Hall, 2003 paperback £34.99 0-13-101988-0

Advanced RF harmonic theory ... in practice

Opamp and ADC technology have progressed so far that they have outreached high-resolution harmonic distortion measurements. A fresh look is needed at the practical aspects of the measurement process, allowing designers to debug their latest creations.

Leslie Green CEng MIEE investigates

Audio engineers use 24+ bit acquisition systems to measure harmonic distortion 120dB below the fundamental. Microwave engineers use spectrum analysers to measure harmonic distortion 85dB below the fundamental out into the far gigahertz region. In between the extremes of the audio and microwave regions lies the vast bulk of modern designers. We want to design for low harmonic distortion in the low megahertz range, but neither of the aforementioned disciplines is of much use.

Spectrum analysers

Modern spectrum analysers are enhanced with digital technology, reducing the noise floor and hence increasing the dynamic range. It is therefore quite usual for spectrum analysers to quote dynamic ranges in excess of 120dB. They also have extended low frequency operation so that some work down to the kilohertz region. You would therefore be forgiven for assuming that such a device would be ideal for measuring harmonic distortion in the low megahertz region.

Of course 'microwave and RF' engineers have their own terminology. Instead of measuring relative to the fundamental, they measure relative to the 'carrier'. A second harmonic 80dB lower than the fundamental is then simply –80dBc. HD2 is a convenient shorthand notation for second harmonic distortion, and will be used henceforth.

Although a spectrum analyser has a huge *non-harmonic* dynamic range, the HD2 specification seems quite poor. Consider the HD2 specification of the Agilent E4402 9kHz to 3GHz spectrum analyser:

10MHz to 500MHz:< –65dBc
500MHz to 1.5GHz:< –75dBc
1.5GHz to 2.0GHz:< –85dBc
>2GHz : < –100dBc

Observe that HD2 gets better as the frequency is increased. Also notice that there is no HD2 specification below 10MHz, despite the fact that the analyser is otherwise specified down to 9kHz operation. The bottom line is that spectrum analysers specified for –100dBc at 1MHz just do not exist.

The key problem with a spectrum analyser for measuring harmonic distortion at ultra low levels is that the displayed distortion is critically dependant on the size of the signal presented to the input stage, 'the mixer'. This effect is shown quite clearly in the measured results of **Figure 1**.

The best HD2 measurement on a spectrum analyser is achieved by reducing the signal at the mixer until the second harmonic is just sitting on the noise floor. Whilst this can be done by using the input attenuator built into the spectrum analyser, this attenuator usually has 10dB steps. Such a coarse attenuator does not allow the best possible measurements. It is for this reason that the Anritsu MS2681A, for example, uses an input attenuator with 2dB steps. One can of course use an external attenuator, but this is not ideal from a usability point of view.

In order to minimise the noise floor of the spectrum analyser, the 'local' measurement bandwidth, the *resolution bandwidth*, has to be made as low as possible. Minimising the resolution bandwidth slows down the sweep speed and makes it impractical to see the fundamental and second harmonic at the same time.

The optimum measurement is achieved by centring the display on the fundamental, taking a reading, then re-centring on the harmonic and taking a reading. All very inconvenient and time consuming; not something that even a proficient user should be expected to realise.

When I wrote out the table of HD2 specifications for the Agilent E4402, I missed out a crucial piece of information, the signal level at the input to the mixer. In order to estimate the best HD2 reading that can be made, you are expected to realise that HD2 improves at a rate of 1dB for every dB that you reduce the input signal until the second harmonic reaches the noise floor of the spectrum analyser. This 1dB/dB slope is seen quite clearly in **Figure 1**, provided the signal is initially low enough.

Equipment

If you look at the data sheet for the Analog Devices opamp AD8021, for example, you will see typical distortion specs at 1MHz of HD2 = –93dBc and HD3 = –108dBc. As a user you might reasonably ask how this high level of performance can be verified in your design.

In the case of an opamp, it can be wired into a unity gain inverting configuration. If the input is supplied with a clean sinewave then the output should be an anti-phase sinewave plus whatever harmonics the opamp itself produces. If the input and output are summed using a pair of equal resistors, the fundamentals will largely cancel. The remaining signal will therefore contain a much higher proportion of the opamp's harmonics.

This technique has effectively boosted the resolution of the measuring system. In fact, any technique which suppresses the fundamental will have the same resolution enhancing benefit.

Unfortunately, because of the delay through the opamp, the input and output fundamentals will not be exactly in-phase. They therefore will not null particularly well, even if a fine gain control is included in the summing network. A phase shift of θ degrees will prevent the fundamental from being nulled by more than $-20\times\log_{10}[\sin(\theta)]$. Nulling the fundamental by more than 40dB requires the input-to-output phase shift to be less than 0.57°.

Introducing such a small phase shift is apparently easy, just an RC filter. But capacitors and inductors are notorious for *creating* harmonic distortions!

Obviously if you used a ceramic capacitor you would only use an NP0 dielectric, knowing full well that Z5U and Y5V types are useless for analogue work.

However, even metallised film capacitors are known to create harmonic distortions, so it is unsafe to assume that the filter you have made will be completely free from harmonic distortion. The only safe answer is to use a low-loss delay line to produce the phase shift, preferably a delay line having an air dielectric.

Having nulled the pure sinewave you can now ... but wait a minute, where did the pure sinewave come from? A typical RF signal generator at 1MHz and above will have HD2 and HD3 levels of around –40dBc to –50dBc, depending on the output level. You cannot buy a signal generator that produces less than –90dBc harmonics above 1MHz; and the best I have found only gives –60dBc.

In the case of the opamp scheme, you might feel that the input harmonics from the signal generator would also be nulled by summation with their counterparts after passing through the opamp. The trouble is that the phase shift network now has to vary with frequency appropriately to correctly phase-up the fundamental, second and third harmonics, and the gain also needs to be constant over this 3:1 frequency range. Realistically then, it is still important to use a very pure sinewave source.

Standards

Let's suppose that you make yourself a fixed frequency ultra-pure sinewave generator. Having made it, you don't know how well it performs so you want to calibrate it. I checked with the UK National Physical Laboratory (NPL) and they are unable to calibrate such a generator. I also made the same enquiry with the US National Institute of Standards and Technology (NIST, formerly NBS) but they did not respond.

Of the five calibration laboratories in the UK accredited for distortion measuring equipment, not one was able to calibrate anything above 100kHz. Thus whatever confirmation you can get from other sources, the measurement is not going to be traceable to national standards. This being the case, you need a much better understanding of fundamental metrology than usual; there is no measurement establishment to corroborate and backup your readings.

Text books suggest using a notch filter to kill off the fundamental, but do not then explain how to prove 'beyond reasonable doubt' that the notch filter is linear. It is very easy to make a notch filter that is non-linear at the ultra-low distortion levels we are talking about.

The solution to our signal source problem is to use either a notch-pass filter or a low-pass filter; either way we get rid of the output harmonics as required. The solution to the measurement problem is either a notch filter or a high-pass filter; either way we need to remove at least 60dB of the fundamental in order to boost the system measurement capability by at least 60dB. The problem is that neither filter is testable on its own; they can only be tested as a pair.

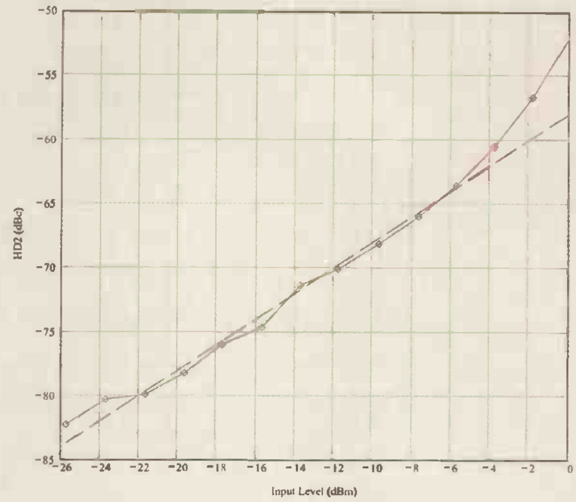


***Figure 1:* Measured second harmonic distortion characteristic of an Advantest R3261C spectrum analyser at 1.1MHz.**

Theory

The way a spectrum analyser works, the only place harmonic distortion can be created is in the mixer. The input signal is then converted to a fixed intermediate frequency so that no further frequency related information is processed. You always need to test for distortion in the mixer by using extra attenuation at the spectrum analyser input. If an additional 2dB attenuation reduces the fundamental by 2dB, but the harmonic by 4dB or more, then the harmonic is being generated by the mixer and is not present in the signal.

The theory behind the 1dB per dB reduction of internal second harmonic distortion with input attenuation is very simple. The spectrum analyser's input mixer can be characterised by a power series expansion:

$$V_O = A_0 + A_1 V_1 + A_2 V_1^2 + A_3 V_1^3 + \ldots$$

Using a cosine input signal gives:

$$V_O = A_0 + A_1 V\cos(\omega t) + A_2 V^2\cos^2(\omega t) + A_3 V^3\cos^3(\omega t) + \ldots$$

The cos squared term can be converted to a double frequency term using standard trigonometric identities and it is then clear that the resulting double frequency term is multiplied by the same V^2 factor that was originally in front of the cos squared term. Hence when the input voltage V is increased by a certain number of decibels, the second harmonic is increased by double that amount.

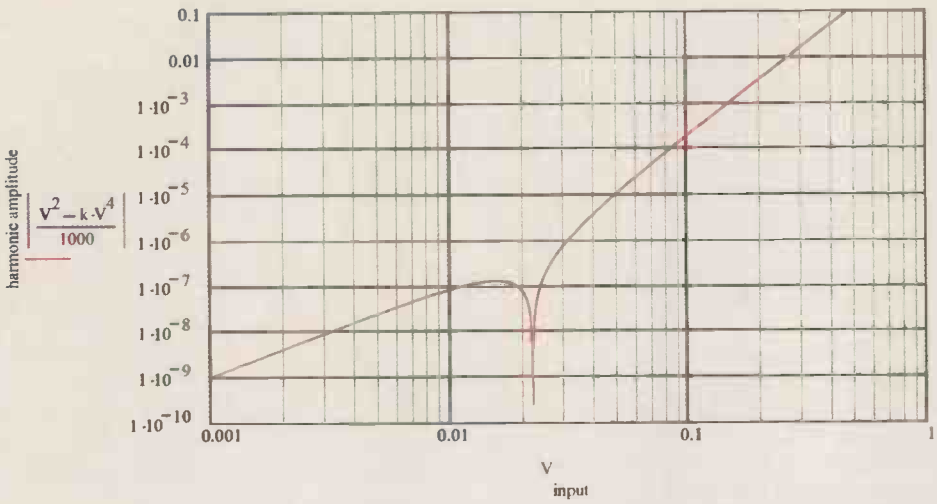


Figure 2: Theoretical second harmonic distortion due only to second and forth order terms in the power series representation.

The more complete answer is that all even-order powers of the input signal, created by the even-order powers in the power series, result in some second harmonic contribution. (Likewise, all odd-order powers of the input signal contribute to the third harmonic distortion.) However the higher order even powers will contribute more second harmonic distortion as the input signal is increased by virtue of their higher index of the V term in the power series expansion. What therefore happens is that the V^4 term creates second harmonic distortion that increases at four times the dB rate of the input signal. A graph of this situation shows the increased slope at increased signal levels (**Figure 2**).

From the graph it is seen that the initial slope of the curve is two decades per decade (= 2dB/dB). At the far right hand end of the curve, the slope is 4dB/dB due to the fourth order distortion. The dip in the middle is due to the fourth order term being of opposite sign to the second order term. Hence it is not *guaranteed* that harmonic distortion monotonically decreases as the input signal is decreased.

The measurements of **Figure 3** validate the theory, showing that higher order terms in the power series give rise to a great many local minima in HD2. The signal generator used for this test was a Krohn-Hite 4402B, which typically generates very pure sinewaves. Although a 12 bit system has 1 part in 4096 resolution (72dB), more resolution is achieved when you FFT the time domain data. The FFT spreads out the quantisation noise, giving more resolution.

The measurement was performed using a 4096 point FFT and the resulting FFT was averaged to minimise the noise. Furthermore, additional resolution was achieved by putting the Accura into a dithered oversampling mode; the resulting –110dBFS noise floor is quite impressive. It seems likely that these high-order power series terms are due to the integral non-linearity errors of the ADC used, an Analog Devices AD9432BST–105.

Since the Krohn-Hite is not specified up to a distortion level which eliminates its effects from the test, another test helps to show that it is the acquisition system giving rise to these wiggles in the HD2 reading. By using a fixed 4 div ptp signal and shifting it up and down the screen, changes in the signal source are eliminated. **Figure 4** shows that there is still interesting structure in the HD2 readings.

The point to be stressed here is that simple devices, like the diodes used in a spectrum analyser mixer, will tend to have a monotonic increasing transfer function. Such a transfer function will have a power series representation where the terms all have the same sign.

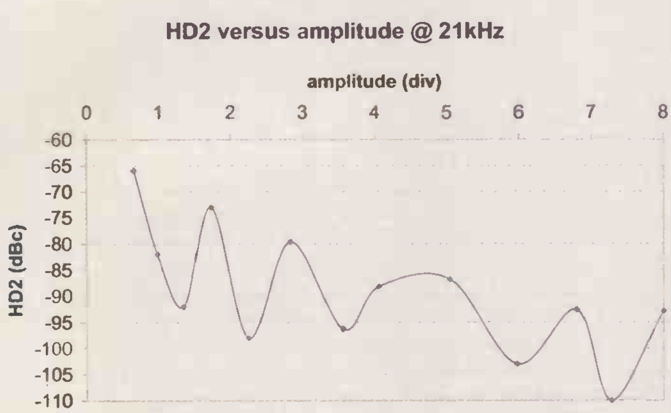
In a complex linearity-corrected device such as an ADC, the wiggles in the transfer function created by the linearisation process amount to dominant higher order terms which are not all of the same sign.

Consider just second harmonic distortion. For any system, if you reduce the input signal sufficiently the second harmonic distortion created by the second order term will eventually dominate, unless you hit the system noise floor first. The reason is that the second harmonic distortion due to the higher order terms will have been reducing much faster than that produced by the lower order terms as the signal amplitude was reduced. For an ADC, however, the signal level required for the second order term to be dominant may not be found because, in practice, the signal is seldom reduced below one tenth of full scale.

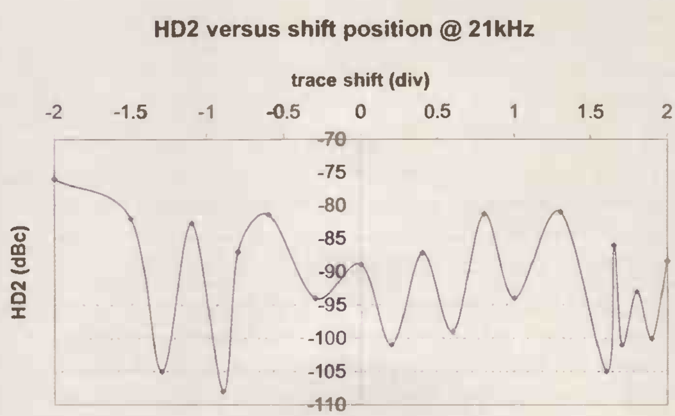
Signal Source

The first step in any testing process is to generate a pure sinusoidal signal, which, as mentioned earlier, is achieved using a filter. I bought a custom-made notch-pass filter from Allen Avionics for several hundred pounds. A centre frequency of 1.1MHz was chosen, rather than 1MHz, to minimise aliasing problems.

This filter was tested to –95dBc insertion loss at the second harmonic by the manufacturer. That is a very easy figure to verify using a signal generator at 2.2MHz feeding into a spectrum analyser. There is no dynamic range consideration because



Left – Figure 3: Second harmonic distortion in a Nicolet Accura 100 12-bit digital storage oscilloscope as a function of the peak to peak input signal.



Right – Figure 4: Second harmonic distortion in an Accura with a 4 div ptp signal moved to different screen positions.

you can use a nice big signal at the input to the filter.

In normal use, with this much attenuation at the second harmonic, 'none' of the second harmonic present in the signal generator will get through to the output; –45dBc HD2 at the input, for example, would give rise to –140dBc at the output! However it is quite possible for internal distortion within the filter to create its own second harmonic distortion.

What is clear is that I cannot 'turn down' the signal from the signal generator, look at the signal on the wave analyser and see if the distortion changes. (By 'wave analyser' I mean any one of a spectrum analyser, an oscilloscope with FFT capability, or a dedicated FFT analyser.)

In fact it is not even possible to say that reducing the output level of an RF signal generator is guaranteed to reduce its harmonic content! If the generator consists of an oscillator, a variable attenuator and an output stage, the oscillator harmonics will be fairly constant, whereas the output stage harmonics will be changing with amplitude. It is therefore quite possible for the oscillator harmonics to null with the output stage harmonics at one or more amplitude levels, thereby making the harmonic content a non-monotonic function of the output amplitude.

One test I could do is to initially use a 3dB pad at the input to the filter. (RF people use the simple term 'pad' instead of 'inline 50Ω attenuator'.) If the 3dB pad is first used at the filter input, then moved to the filter output, the overall harmonic distortion should be substantially the same in both cases. If it is not the same, the filter was causing a significant amount of distortion. **Figure 5** makes this test clearer.

A slight caution is in order at this point. The 3dB pad between the signal generator and the filter will improve the matching between the two devices. Therefore, when the pad is removed, the system may not give exactly the same performance due to the mismatch. This problem can be minimised by always using additional pads at the signal generator output and at the wave analyser input. Mismatch changes will then be considerably lessened when the moveable pad of figure 5 is used.

Now of course the 3dB pad which is being moved may introduce distortion of its own. In this case it might cause a difference between the readings in the test of figure 5. Thus the test is most useful when no change is detected.

Then again, the change in distortion in the filter may just happen to cancel the additional distortion due to the 3dB pad, causing no overall change! In a situation like this, you need another independent method to verify the test. If the test also worked with a 6dB pad then the confidence level would be increased.

Absolute harmonic filter

Ideally a filter is needed which can remove at least 10dB of the harmonics, but which is guaranteed not to introduce harmonic distortions of its own. Clearly no LCR filter can perform this function because any inductors and capacitors in such a filter would be under suspicion and cause an uncertain result. Air-wound inductors should give reasonably distortion free results, as should air-dielectric capacitors, but a piece of wire should give the best possible results.

What we are talking about is a quarter wave transformer used as a notch filter. For those who are rusty on transmission line theory, an open-circuit length of loss-free coaxial cable will appear as a short-circuit when the signal frequency is equal to the reciprocal of four times the delay time of the cable. A 1MHz fundamental requires the 2MHz

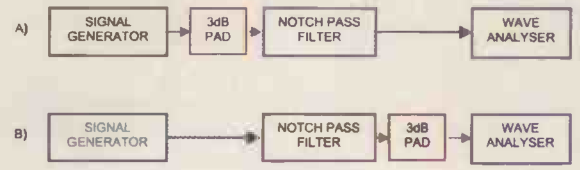


Figure 5: The pad is moved from the input to the output of the notch-pass filter. If the reading on the wave analyser changes, the filter was creating distortion.

second harmonic to be removed and this requires a cable that is 125ns long. RG58 coax has a delay of around 5ns/metre, so that means around 25 metres of cable.

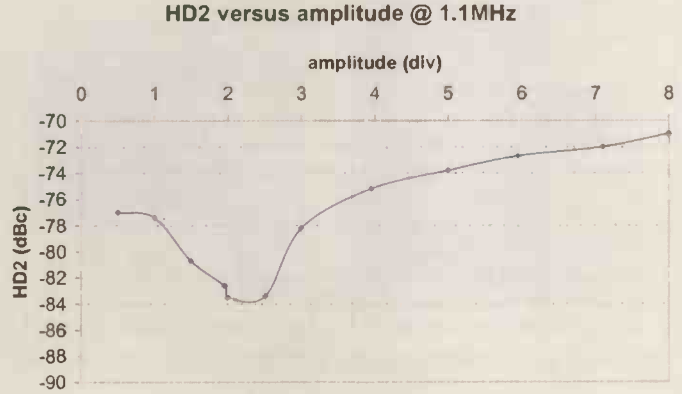
The loss of the cable determines the depth of the attenuation notch, so don't use skinny coax such as RG174. In my case, the second harmonic reduction of the (chunky) RG58 absolute filter was 20dB. Please note that propagation delays are dependant on dielectric constant. You should always cut the cable too long in the first instance, measure the notch frequency and iteratively tune the notch by successive measurements, calculations, and snips.

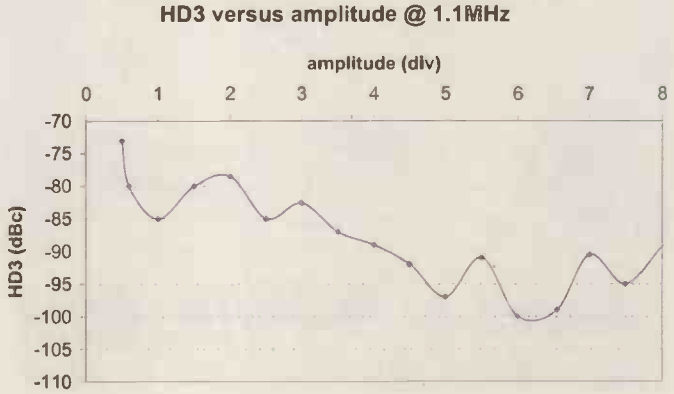
A simple T-piece junction can be used at the input to the wave analyser, the open circuit coax line being attached in parallel with the input signal. A switched attenuator will be needed just before the input to the wave analyser/quarter wave line to cope with the loss at the fundamental frequency when the quarter wave line is attached. Obviously you switch in some additional attenuation when the quarter wave line is not in circuit, thereby maintaining the amplitude of the fundamental applied to the wave analyser between the two tests.

This absolute harmonic filter method should give a very high confidence that the notch-pass filter is, or is not, putting out sufficient harmonics to be measured in your system. Having determined that the notch pass filter was not producing significant harmonics by using an absolute harmonic filter, I was confident that the readings taken

Left – Figure 6: Second harmonic distortion in the Accura driven from a pure source.

Right – Figure 7: Third harmonic distortion in the Accura driven from a pure source.





were due to the oscilloscope and not the signal generator or the filter. **(Figure 6, Figure 7)**.

Using the absolute harmonic filter method, the source/filter combination can be qualified for use at any arbitrary level of resolution. Therefore the next step is to increase the wave analyser resolution by using either a high-pass filter or notch filter, thereby lessening the fundamental content.

The harmonics visible with this configuration should be minimal if the high-pass filter is linear. The absolute filter is used at the input side of the high-pass filter to see if there is a change in the harmonic level. If there is no change, and the harmonic is too big relative to the enhanced measurement being made, then the high-pass filter itself needs to be improved. If, on the other hand, the harmonic drops, then the harmonic you were previously seeing was a valid enhanced view of the harmonic coming through the notch-pass filter. One should still do a test with a 3dB pad moved from input to output of the high-pass filter, however.

Another thing to watch out for is an increase of harmonic distortion when the absolute filter is used. This would mean that harmonic signals from the generator/low-pass filter combination were nulling with distortion created by the high-pass filter.

In conclusion

Higher frequencies and higher loads make low harmonic distortion more difficult to achieve. However, absolute filters using transmission lines get shorter, smaller and more effective at higher frequencies. Unless you can measure harmonic distortion to a sufficient resolution, and with confidence, you cannot easily evaluate optimum components for your precision application.

At the moment, accurate measurements of ultra-low levels of harmonic distortion in the low megahertz region are laborious and can only be done at spot frequencies. However, it will only be a matter of a few years before time domain equipment such as FFT analysers and oscilloscopes totally outperform frequency domain equipment such as spectrum analysers in this measurement region. Specialist filters will be needed in the first instance to calibrate these FFT analysers, and absolute harmonic filters will be needed to calibrate the specialist filters.

RF signal generators in the 1MHz to 10MHz region are generally very poor on harmonic distortion and it is to be hoped that they will also improve to support the measurement efforts; at least a factor of ten improvement would be appropriate.

It is also important for national metrology institutes and accredited test laboratories to upgrade their capabilities to provide the necessary traceability of harmonic distortion measurements at these frequency and distortion values.

I should finally admit that even cables and connectors are known to be non-linear! The non-linearity manifests itself as passive intermodulation distortion (PIM). PIM is important in RF transmitter cables, with figures in the region of –140dBc to –160dBc being quoted. Thus well-made cables should be acceptable at the level of –110dBc which is required for the present purposes.

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Bandwidth	DC to 10MHz
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Input capacitance	40pF+oscilloscope capacitance
Working voltage	600V DC or pk-pk AC
Switch position 2	
Bandwidth	DC to 150MHz
Rise time	2.4ns
Input resistance	10MΩ ±1% if oscilloscope i/p is 1MΩ
Input capacitance	12pF if oscilloscope i/p is 20pF
Compensation range	10-60pF
Working voltage	600V DC or pk-pk AC
Switch position 'Ref'	
Probe tip grounded via 9MΩ, scope i/p grounded	

6¼ bit DAC requires only four output pins

All microcontrollers with the exception of a few primordial 8051s have the ability to set each pin as either an output driven high, an output driven low or a high impedance input. Using all three states allows a DAC with 81 output levels to be produced using just four output pins, whereas a standard binary DAC using 4 pins would produce only16 levels.

R_5 and R_6 bias the non-inverting input of an op-amp to half the supply (2.5V). The circuit operates in exactly the same way as the binary-weighted-resistors DAC, with the op-amp operating as a virtual earth amplifier, summing the currents through resistors R_1 to R_4, except that the resistors are ternary weighted. With all microcontroller outputs in their high impedance state, no current flows in R_1-R_4 and the output is at 2.5V. With R_4 taken to 0V, the output is at one LSB above 2.5V, with R4 taken to +5V, the output is at one LSB below 2.5V. With R_3 taken to 0V, the output is at three LSB above 2.5V, and with R_3 high the output is three LSB below 2.5V.

To produce an output of two LSB above or below 2.5V, R_3 is taken high and R_4 low, or vice versa. Unlike the usual weighted resistors DAC, this circuit can operate off a +5V only supply, If the output is only required from 0V to +5V, use an op-amp with rail-to-rail outputs (doesn't need rail-to-rail inputs) and R_{oh} and R_{ol} are not required.

To bias the output to 0V R_{oh} should be the same as R_7. R_{ol} can similarly be used to bias the output to the centre of a +12V supply. C_1 is filters out the glitches caused as the inputs change state. This circuit has been tried with PIC and AVR microcontrollers. PIC software has to access a bank-select register twice to tri-state the outputs, so the AVR is a better choice if any output speed is required. An example of the software is shown below. Table 1 shows the data to be written to the Port output and Data Direction registers (TRIS registers in the case of the PIC).

Note that the data is for the AVR microcontroller which uses a zero to select high-impedance state on the output, whereas the PIC uses a one.

The circuit can be extended to more bits, it will give the equivalent of 58% (log3/log2) more bits over a binary system: there is a sequence of E24 resistors that almost fits the ternary weighting values:

1.1 - 3.3 - 10 - 30 - 91 - 270 - 820

The fit is about 1% so for more than 100 output levels, better tolerance resistors approximating more closely to a ternary scale would be require if monotonicity is not to suffer. The table below shows the port values necessary for each output voltage. The first column refers to a single-ended output, the second to a bipolar output centred on 2.5V.

The following is an example of the AVR code:

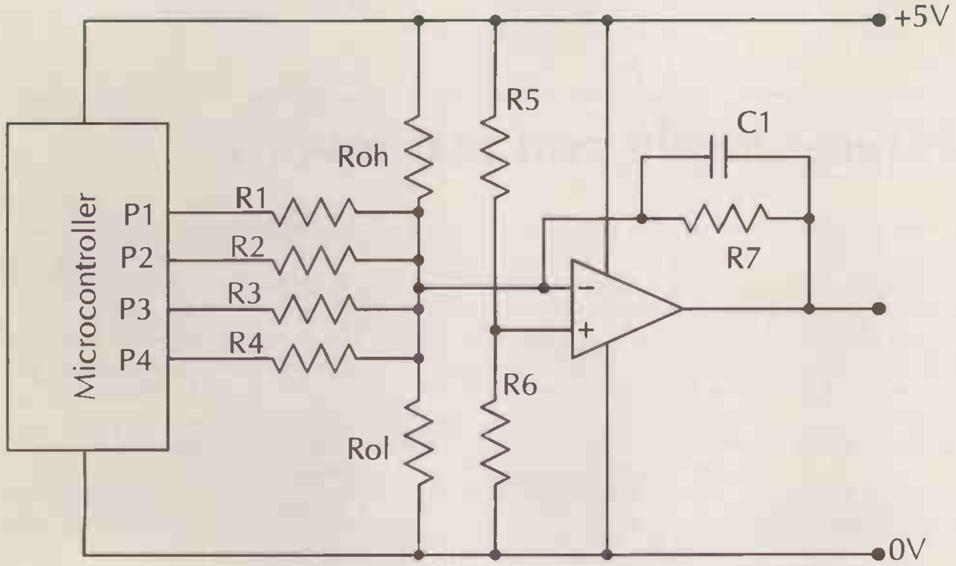
```
LPM ;LOAD PORT DATA FROM
TABLE
MOV R1,R0
INC ZL
LPM ;LOAD TRISTATE DATA
FROM TABLE
LDI YL,0 ;TRISTATE ALL
OUTPUTS
OUT DDRB,YL
OUT PORTB,R1 ;OUTPUT DATA
TO PORT
OUT DDRB,R0 ;OUTPUT DATA TO
TRISTATE REGISTER
INC ZL
RJMP LOOP
```

The Z register points to a table of values to be output, each 16-bit word containing the port data (least significant byte) and the tristate data (most significant byte). It only takes two clock cycles to update the output (which is 250ns for an AVR running at 8MHz).

Perhaps some reader could inform me as to whether an R-3R ladder circuit exists!

Ian Benton
Ilkeston
Derbyshire
UK

	Port Value				DDR Values					Port Value				DDR Values			
	P1	P2	P3	P4	P1	P2	P3	P4		P1	P2	P3	P4	P1	P2	P3	P4
0 -40	0	0	0	0	1	1	1	1	41 1	0	0	0	1	0	0	0	1
1 -39	0	0	0	0	1	1	1	0	42 2	0	0	1	0	0	0	1	1
2 -38	0	0	0	1	1	1	1	1	43 3	0	0	1	0	0	0	1	0
3 -37	0	0	0	0	1	1	0	1	44 4	0	0	1	1	0	0	1	1
4 -36	0	0	0	0	1	1	0	0	45 5	0	1	0	0	0	1	1	1
5 -35	0	0	0	1	1	1	0	1	46 6	0	1	0	0	0	1	1	0
6 -34	0	0	1	0	1	1	1	1	47 7	0	1	0	1	0	1	1	1
7 -33	0	0	1	0	1	1	1	0	48 8	0	1	0	0	0	1	0	1
8 -32	0	0	1	1	1	1	1	1	49 9	0	1	0	0	0	1	0	0
9 -31	0	0	0	0	1	0	1	1	50 10	0	1	0	1	0	1	0	1
10 -30	0	0	0	0	1	0	1	0	51 11	0	1	1	0	0	1	1	1
11 -29	0	0	0	1	1	0	1	1	52 12	0	1	1	0	0	1	1	0
12 -28	0	0	0	0	1	0	0	1	53 13	0	1	1	1	0	1	1	1
13 -27	0	0	0	0	1	0	0	0	54 14	1	0	0	0	1	1	1	1
14 -26	0	0	0	1	1	0	0	1	55 15	1	0	0	0	1	1	1	0
15 -25	0	0	1	0	1	0	1	1	56 16	1	0	0	1	1	1	1	1
16 -24	0	0	1	0	1	0	1	0	57 17	1	0	0	0	1	1	0	1
17 -23	0	0	1	1	1	0	1	1	58 18	1	0	0	0	1	1	0	0
18 -22	0	1	0	0	1	1	1	1	59 19	1	0	0	1	1	1	0	1
19 -21	0	1	0	0	1	1	1	0	60 20	1	0	1	0	1	1	1	1
20 -20	0	1	0	1	1	1	1	1	61 21	1	0	1	0	1	1	1	0
21 -19	0	1	0	0	1	1	0	1	62 22	1	0	1	1	1	1	1	1
22 -18	0	1	0	0	1	1	0	0	63 23	1	0	0	0	1	0	1	1
23 -17	0	1	0	1	1	1	0	1	64 24	1	0	0	0	1	0	1	0
24 -16	0	1	1	0	1	1	1	1	65 25	1	0	0	1	1	0	1	1
25 -15	0	1	1	0	1	1	1	0	66 26	1	0	0	0	1	0	0	1
26 -14	0	1	1	1	1	1	1	1	67 27	1	0	0	0	1	0	0	0
27 -13	0	0	0	0	0	1	1	1	68 28	1	0	0	1	1	0	0	1
28 -12	0	0	0	0	0	1	1	0	69 29	1	0	1	0	1	0	1	1
29 -11	0	0	0	1	0	1	1	1	70 30	1	0	1	0	1	0	1	0
30 -10	0	0	0	0	0	1	0	1	71 31	1	0	1	1	1	0	1	1
31 -9	0	0	0	0	0	1	0	0	72 32	1	1	0	0	1	1	1	1
32 -8	0	0	0	1	0	1	0	1	73 33	1	1	0	0	1	1	1	0
33 -7	0	0	1	0	0	1	1	1	74 34	1	1	0	1	1	1	1	1
34 -6	0	0	1	0	0	1	1	0	75 35	1	1	0	0	1	1	0	1
35 -5	0	0	1	1	0	1	1	1	76 36	1	1	0	0	1	1	0	0
36 -4	0	0	0	0	0	0	1	1	77 37	1	1	0	1	1	1	0	1
37 -3	0	0	0	0	0	0	1	0	78 38	1	1	1	0	1	1	1	1
38 -2	0	0	0	1	0	0	1	1	79 39	1	1	1	0	1	1	1	0
39 -1	0	0	0	0	0	0	0	1	80 40	1	1	1	1	1	1	1	1
40 0	0	0	0	0	0	0	0	0									



R1 = 10k
R2 = 30k
R3 = 91k
R4 = 270k
R5 = R6 = 22k
R7 = 6.8k
C1 = see text

Audio level and peak metering

Engaged in a recent audio project, I had left the Vu and peak-metering circuit to last thinking that it would involve a simple rectifier, possibly enhanced by an op-amp driving a capacitor. Once involved in the detail, I soon found that this was not a simple task.

My requirements were for a circuit, which would capture the peak potential of an audio signal to within 5% between 20Hz and 20kHz and store the potential (slowly decaying) for observation by the user. The RMS value should also be available.

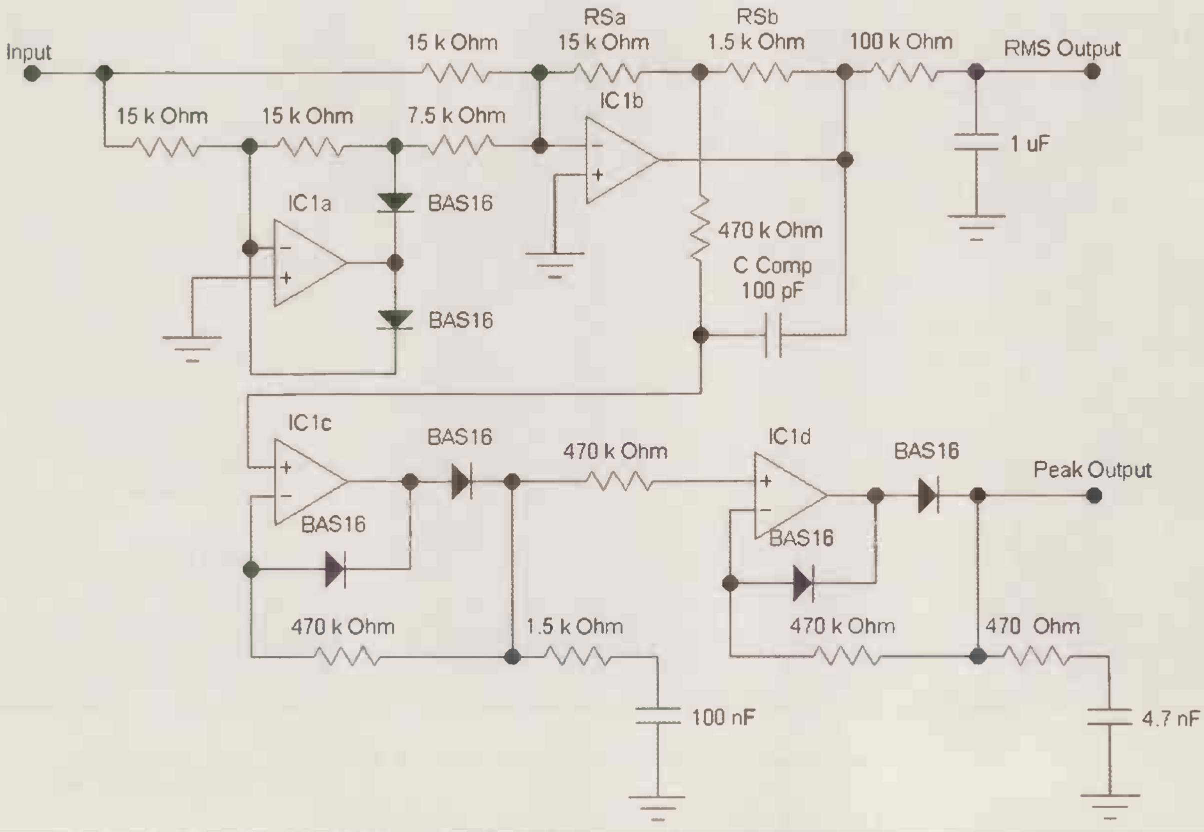
Charging a large capacitor at high frequencies without causing phase lag and thereby inaccuracy is the route of the problem, as the frequency rises, the peak value tends toward the average instead of the peak due to the limited current output capabilities of an op-amp.

The answer was to 'pass on' the acquired result and any errors from a frequency limited low frequency stage to another stage capable of dealing with the higher frequencies, but less capable of storing the value for long time periods.

IC1a&b form a standard precision rectifier that can easily be scaled by multiplying RSa & RSb by the required ratio. RSb, which factors the average value to RMS, should ideally be 11% of RSa. C Comp goes someway to compensating for the speed limitations, which are mainly due to op-amp slew rates & input capacitance, and diode capacitance.

With the values shown, the output ripple from the peak detector is +0, -7% at 20Hz, and is 98% accurate at 20kHz, and has a decay time of approximately 50 milliseconds to 30%. The RMS output ripple is ±2% at 20Hz and has a decay time of approximately 125ms.

John Charlesworth
Wombleton
York



High voltage, simple and fast inverter

Figure 1 shows a very simple high voltage MOSFET inverter, where the problem of driving the high side MOSFET is solved using a low voltage transistor Q_1 and a special arrangement through D_6. This is an extremely fast inverter, much faster than those driven through optocouplers, so dead time problems are minimised.

The inverter has the usual blocking diodes D_4 and D_6, and the parallel diodes D_5 and D_8.

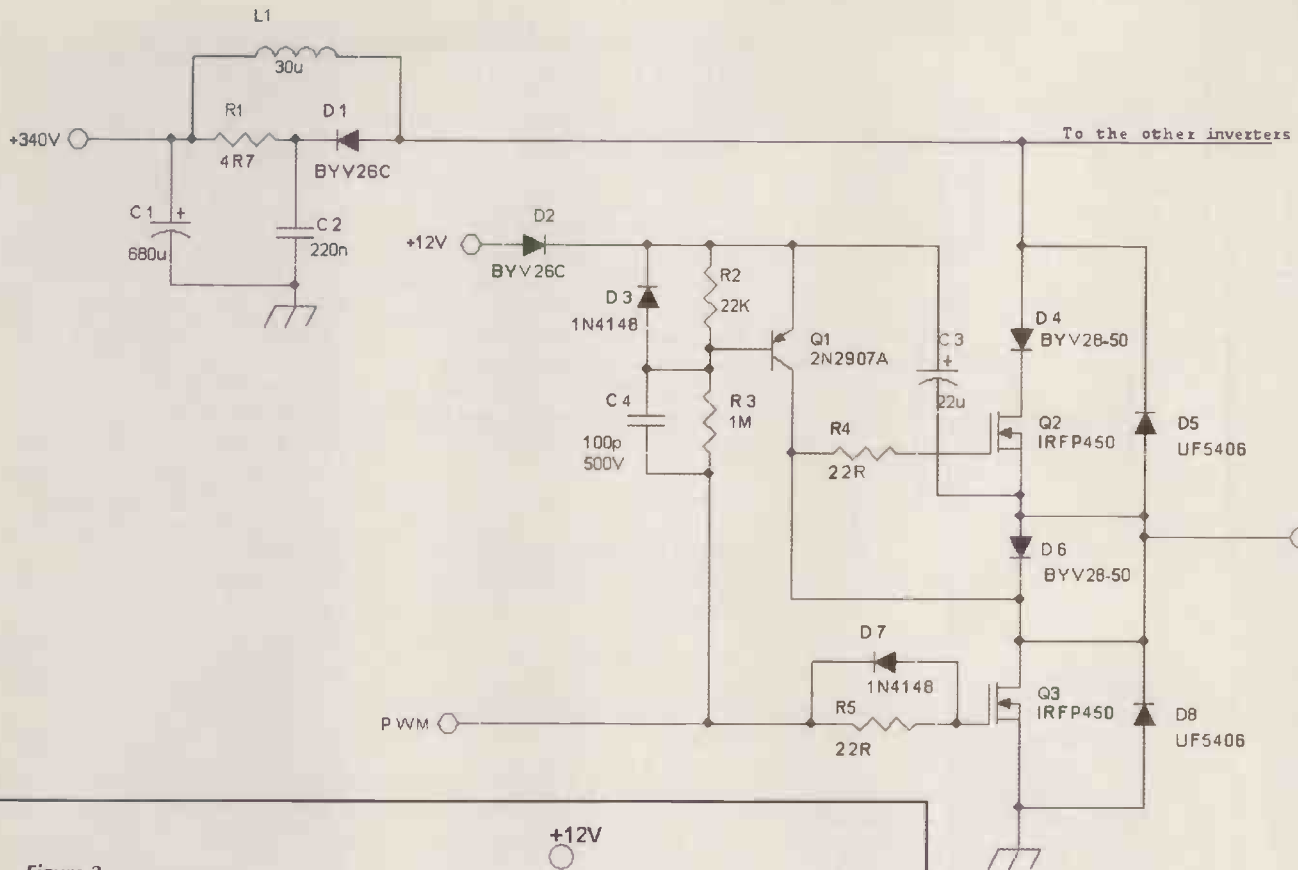
Turn off of Q_2 is performed by Q_3: when Q_3 turns on, Q_2 gate is short-circuited to ground through R_4 (which limits current and dampens oscillations). Q_2 gate is discharged very fast, discharging time being only limited by the value of R_4. Q_1 stays off thanks to R_2 and C_3 is charged to +12V through D_2. The gate pulse creates a current through C_4 and D_3 protects de base-emitter junction of Q_1.

Turn on of Q_2: when the control input (PWM in the diagram) goes low, Q_3 turns off fast, thanks to D_7. A displacement current ($C4 \cdot dV/dt$) flows through C_4 to the base of Q_1. Q_1 charges the output capacitance of Q_3 and the gate capacitance of Q_2, which turns on. C supplies the collector current.

If the period is long, Q_1 stays conducting and compensating the leakage of Q_3. If D_6 were a Schottky diode, which is very leaky, R_1 should be reduced.

High voltage, simple and fast inverter (*continued*)

Figure 1



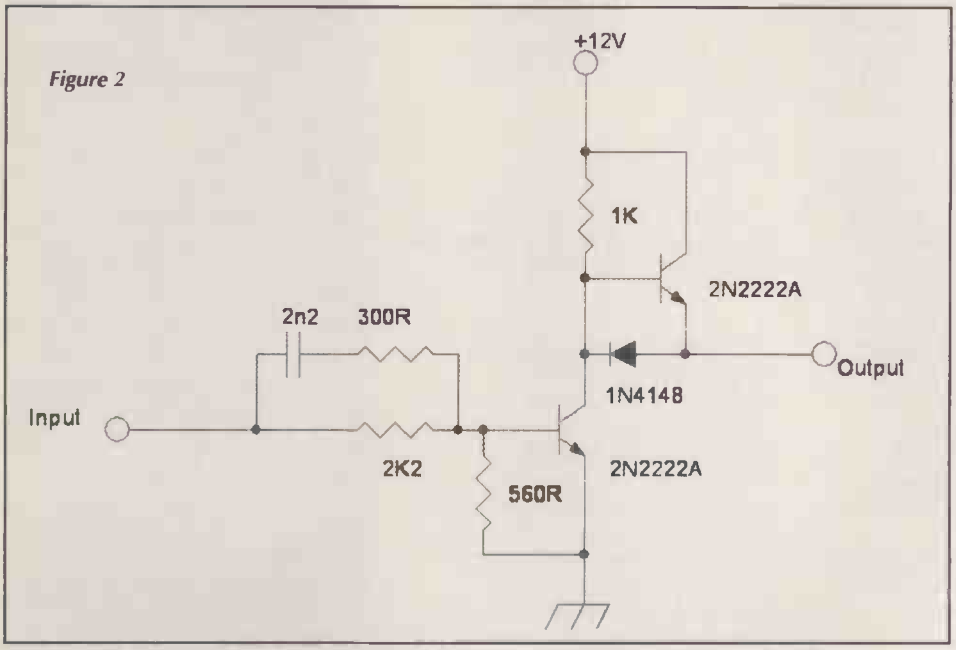


Figure 2

There is a very short cross conduction between both MOSFETs, more apparent when Q_3 turns off and Q_2 on. A small inductor in series with the main supply limits the current spikes. The inductor (L_1) needs a snubber made by D_1, R_1, C_2. (the inductor value shown is very conservative and can be smaller).

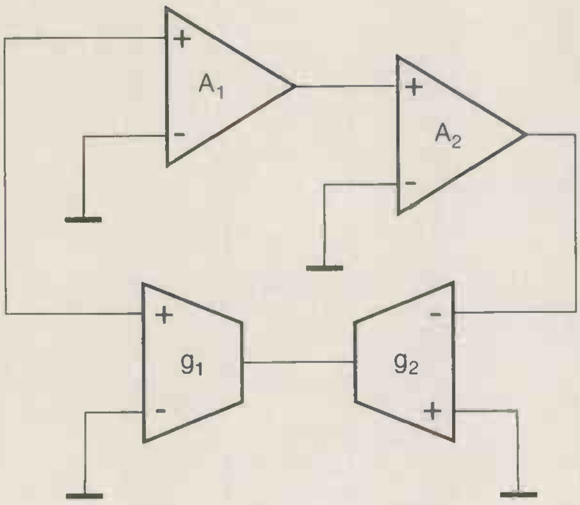
The values shown are for a 370W three-phase inverter with 150% overload capacity. If the MOSFET is changed, the value of C_4 has to change according to the total gate charge plus the output capacitance of Q_3 (which is much lower and an can be neglected). The capacitor current is amplified by Q_1, so $C_4 > Qg_2 \times h_{FE1}$. Do not make C4 value higher than necessary as the base current in Q_1 would be too high.

To get all speed advantages of the circuit, the PWM signal should be able to drive Q_3 very fast. If necessary, a buffer circuit as shown in **Figure** 2 can be used. It can be driven by a single CMOS gate.

This is probably the simplest high voltage inverter one can design. It has been used in thousands of three-phase motor drives from 0.37 to 0.75kW. It was protected under the patents US4802075 and EP0274336, but now these patents are free to use.

***Francesco Casanellas** CEng, MIEE, SMIEE*
Aiguafreda
Spain

Electronically tuneable active only oscillator



Presented here is a novel current and/or voltage controlled active-only sinusoidal oscillator based on and employing two operational transconductance amplifiers (OTAs) and as many internally compensated operational amplifiers (ICOAs). The architecture is thus fully integrable with programmability features as it is devoid of external passive components. The circuit is canonical as it employs the absolute minimum number of components. The circuit offers considerable economy in the chip area, as it is intrinsically free from oscillation. The oscillation frequency besides being temperature insensitive is linearly adjustable through the bias voltage of ICOAs1, lending electronic control over it, which well suits IC design techniques. The frequency of oscillation can also be adjusted non-linearly through the bias current of the OTAs.

A routine analysis of the proposed oscillator circuit depicted in the figure yields the following characteristic equation:

$$A_1A_2g_2 + g_1 = 0 \quad (1)$$

Since no complex term is present in the characteristic equation, the circuit as such is free from the preset condition of oscillation. Due to this several IC related advantages have accrued viz. reduction in chip area, power consumption and parasitic effects. Moreover, the absence of the preset condition of oscillation eliminates the possibility of the circuit ceasing to oscillate or producing a distorted signal. Circuits that have a preset condition of oscillation are constrained with a critical design problem of ceasing to oscillate or they produce a signal with a fair amount of distortion in the event of a marginal change in the value of components involved in the condition2-12. This is quite possible when the circuit is allowed to operate in varied environmental conditions. As a result, design procedure becomes involved and the use of costly precision components is inevitable. Assuming the open loop gain of OAs of the form

$$Ai=Bi/s \quad (2)$$

where Bi is the gain bandwidth product of ith OA, equation (1) reduces to

$$s^2g_2+B_1B_2g_2=0 \quad (3)$$

The frequency of oscillation is given as

$$\omega 0=\sqrt{}B_1B_2g_2/g_1 \quad (4)$$

The frequency of oscillation in terms of bias current of OTAs is given as

$$\omega 0=\div B_1B_2I_{02}/I_{01} \quad (5)$$

where transconductance gain $g_i=I_{02}/2V_{Ti}$, I_{01} is the bias current and VTi is the thermal voltage of ith OTA.

An inspection of equation (5) reveals that the frequency of oscillation can be tuned either non-linearly through the bias current of OTAs or linearly through bias supply of ICOAs. The transconductance gain of the OTA is susceptible to change with variation in ambient temperature, but since frequency of oscillation is the function of ratio of transconductance gains as well, resulting in its stabilization as the ratio of trasconductance, gains remain temperature invariant and as such can work well under changing environmental conditions. This feature will have the positive fallout of further improving the functional performance of the circuit. The active sensitivities of w0 are very low and are less than unity.

N. A. Shah and S. Z. Iqbal
The University of Kashmir
Srinagar
India

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Long delay timer using only one 555 chip

One application for this circuit would be a personal radio sleep timer add-on unit. The example sleep timer uses a transistor as a switch but other applications could use it to drive a relay or isolated triac for heavier loads.

My first attempt to build a simple one chip (7555) battery saving 30 minute sleep timer for a personal radio failed due the electrolytic capacitor's leakage current preventing the monostable from timing out. I then used the standard long timer circuit of a 555 astable and digital counter IC, which worked as desired but I felt there must be someway of making a single 555 do the job. The solution seemed rather simple and the circuit is shown in **Figure 1.**

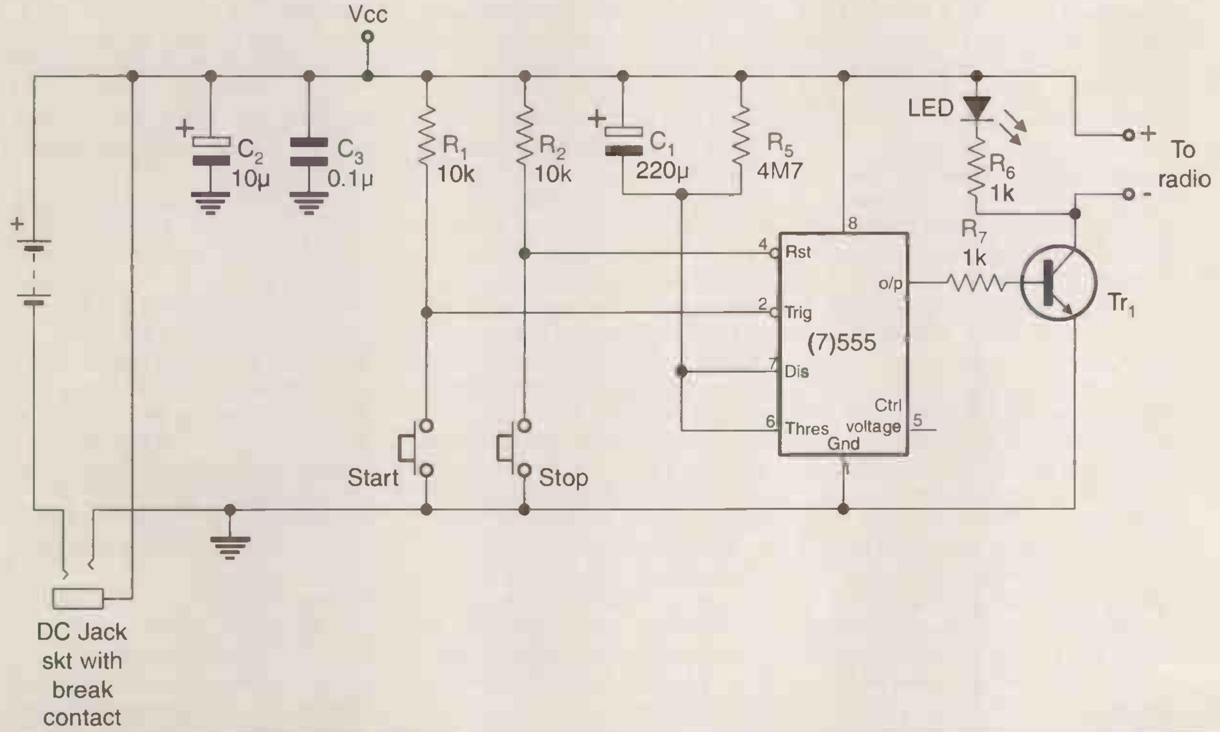
The main circuit uses a CMOS 7555CN (for the low quiescent current, 60μA) to turn the radio on for up to 25 minutes using TR_1 as a switch. TR_1 has a low 1k base resistor to ensure it is saturated. The 'reset' button turns the radio off, and the 'start' button turns it on (thus starting the timeout period), or restarts the timeout period. I also added a DC jack with a make break contact so that I could plug in a solar panel/mains battery eliminator to further save batteries.

The timing components C_1 and R_5 have been rearranged so that the discharge pin actually charges C_1 and so the timing period is set by the C_1-R_5 discharge time, which means that any leakage current will simply slightly shorten the delay by discharging C_1 faster. By referencing C_1 to V_{cc} rather than ground, the 555 threshold pin sees the voltage (V_{cc} - VoltageAcrossC_1), which makes it appear as if C_1 was charging as expected. Thus the delay remains $1.1C_1R_5$ seconds.

I wasn't sure if any diode clamps were needed at the threshold input, so just decided to rely on those present in the IC as part of its static protection circuitry. The power indicator LED_1 only uses 1mA.

I housed the timer PCB and two AA cells in a case of a similar size to the radio, joined the two with stick-on Velcro pads and connected power to the radio using wires ending in dummy AA cells (Maplin YX92A).

Alan Bradley
Belfast
UK



LETTERS
to the editor

Letters to "Electronics World" Highbury Business, Media House, Azalea Drive, Swanley, Kent, BR8 8HU
e-mail EWletters@highburybiz.com using subject heading 'Letters'.

Nutters

I am moved to commend *Electronics World* upon its extraordinarily equability, giving voice to non-mainstream contributors. Perhaps though, a line should be drawn at harmless nutters (non-pejorative). Deciphering the text of *An electric universe* (Aug 04), if I grasp correctly, the cornerstone is that astronomical bodies are highly charged. An old notion easily discounted by observation and calculation. Of course bodies do tend towards a degree of charge but it cannot usually be significant. It's not necessary to elucidate because information is readily available and can be derived from first principals anyway. Nor does your contributor impart authority by invoking mythology, but in his terms he might reflect upon Matthew 7:26 (a foolish man which built his house upon the sand) before constructing his philosophical edifice.

Intelligence, to which I bow, is apparent, but rational appears to be subsumed by a prior agenda. This leads seamlessly to Mr. Catt whose words I have been aware of for nigh on thirty years. The same words juxtaposed with no discernible order. *The Catt question* and other railing have been rejoined ad nauseam but all that results is the same discredited tautologies at higher volume. Failure to comprehend explanation does not render the explanation false. So what is the point of continued indulgence?

Understanding of all things has been forged by brave souls who often suffered for their heresy. But demonstrably flawed random noise merely exposes intellectual caesura and encourages the dissolution of enlightenment. Pandered to, the proponents learn nothing and go to their graves ignorant, arrogant and sad. I would suggest that it is a disservice to them and the scientific community at large to be party to promulgation of their self deceptions.

Andrew S Robertson
Ayrshire
Scotland
UK

Curious result

Bravo to Lee C.F. Sallows for his article *A Curious New Result in Switching Theory* (May '04).

As a one-time designer of state logic systems, I shared his obvious excitement of the chase, and rejoice in his discovery and incisive clear thinking.

'Impractical' I'm sure, but neither 'futile' nor 'redundant' Lee, and perhaps one of the most significant circuit boards ever constructed. Puts me in mind of Alan Turing's Machine – useless but beautiful.

This was my introduction to Moore's Circuit, but I recognised the basic decode/re-code as the method I used in a state logic sequence controller without memory for the first wine bag filling machines, minimising IC packages rather than inverters.

Which made me think; some of the outputs drove air cylinders and some inputs came from switches sensing cylinder positions, forming a 'memory' for the current state of the machine. I would be interested in Mr Sallows' view on how this external memory relates to Moore's Circuit, by definition without memory.

My intuition tells me that the concept of a memory-less system may only be a useful abstraction, and that all 'non-trivial' machines will have state memory in some form, albeit obscure.

And 'I dips me lid' to *EW* for again 'going deeper' (Thomas Scarborough, May letters) and drawing me to think deeper.

What makes *EW* different? The depth that comes from Self, Hickman, Catt, Sallows, JLH, JPB, Scroggie, et al, writing up the mistakes they (and others) made, and how they applied theory and practice to sort them out, gaining insight and passing it on.

The bits that don't fit are the truly interesting ones.

Roly Roper
Ivanhoe
Melbourne
Australia

Hybrid amp

With reference to Mr. Jeff Macaulay's reply in the August *EW;* I'm afraid he has not got the drift of my letter or that of Mr. Aylward. The objection was not to valves in general, or even to valves in hybrid circuits. The objection was to using a valve in such a manner that does not make any significant contribution to the performance of the circuit, while unnecessarily increasing the complexity and cost of the circuit. The fact that a valve is present in the circuit, without affecting the signal, is not sufficient to term the unit a hybrid amplifier, any more than the presence of a neon bulb as a power indicator would be.

Valves are fine, when they are used appropriately. I am yet to find a solid-state domestic AM tuner which can stand comparison with any 50-year-old valve tuner with a triode-hexode oscillator/mixer and a pentode IF stage, in its selectivity, lack of oscillator pulling, or its overload handling ability. I'm not so convinced about the superior sound of valve Hi-Fi, however, and I generally agree with Mr. Macaulay's comments about output impedance (and the vagaries of output transformers) being responsible for the "valve sound."

The point I was trying to make was that the valve in Mr. Macaulay's amplifier is operating purely as a current source, which could have been implemented much more simply using an FET; without compromising performance.

The reasoning that a solid-state current source would cause a large switch-on current surge through the speaker is, at best, dubious. Mr. Macaulay's circuit inherently has a slow turn-on due to the charging of capacitor C1 (through the signal source), with a time constant of the order of 0.68 second. Since the terminal voltage of the current source only rises at the same rate, the surge, with a capacitor coupled output, and enhancement mode FETs, is not likely to be anything of the order of the 10 Amps that Mr. Macaulay mentions in his letter.

I feel quite certain that Mr. Aylward did not recommend the removal of the valve without substituting another current source in its place, as Mr. Macaulay seems to believe. I hope Mr. Macaulay does not think *EW* readers are that ignorant.

Using a valve involves the addition of a filament supply and an increase in heat dissipation. Further, the lifetime of a valve is limited. While it may be fun & trendy to use valves, engineering common sense dictates that they be used only where they make a significant difference.

By the way, what was Mr. Clive Steven's August article all about? I sincerely hope that was a glorious tug at the lower extremities!

Joseph Carri
By email

Cottaging

I'm scratching my head as to how I said that D. Self was running a cottage industry, I was speaking hypothetically. According to SEB my critic's logic, all nil-distortion amplifiers sound the same, therefore all makers of nil-distortion amplifiers should pack it in and D. Self hypothetically should be the sole manufacturer world wide, I was giving D. Self a backhanded compliment, can he not see that? I could easily have chosen a number of Americans, some Japanese or Europeans. I chose him.

I'll repeat what I said in previous letters D. Self deserves world wide acclaim, for his nil-distortion power amps built using cheap parts. I am a follower of J.L.H's philosophy on power amps using MOSFETS from the late seventies onward John's main aim was to design a MOSFET amp, I agree, but this is the only difference I have worth mentioning with D. Self. It was D. Self that put forward power amp circuits without an input filter being accepted as the way to go, I copied that.

I will go further and say and say he richly deserves an OBE for his services to the British Hi-fi industry. There is only one problem with this - to achieve this he needs to be put forward by his peers and after 12 years of slaughtering them in pages of EW this might not be forthcoming , it's a shame really for technically he deserves it and Doug, I'm praising you not slagging you off.

I agree with Jeff McCauley, the perfect amp should have no sound, but what percentage of power amps are like that to more than 10% world wide, why? It's all down to business practice. What amp maker is going to sell his top of the range amp at entry level prices, they would need to be mad and who could afford them, only those earning big bucks who are approximately 10% of a country. The only way around that is to build your own.

A point I should have made previously which brings to public notice, I said previously, in my first letter a year ago, that I listened to music on my Stax Lamba Electrostatics, these required a quality audio interface plugged into the power amp, output (12 Watts RMS) after working with my amp using an 8 Ohm resistor at the output I then thought it better to use a real world load i.e., the interface unit. To my horror the 1KHz square wave on the scope screen looked as if it had been hit with a sledge hammer a third of the way up the leading edge vertical, it veered off to the right at an angle of 40 degrees. This turned out to be because of one of the wire wounds around the output (I had already removed the source resistors) was inductive, therefore two inductors in series, loss of HF, Scroggie Radio Handbook, 6th edition. I removed the resistor and peace was restored, a small adjustment in the feedback removed the slight over-shoot.

It has to be asked how many people in this country as abroad used the Stax's with the transformer interface, with a commercial amp, only self build , those with a scope could try bypassing the output components and fitting headphone only sockets to try it electrostatic only, it does exactly as Graham Maynard's so far.

My critic, SEB, made out in a round about way that I should be reading the *Beano* and *Dandy* comics not *EW*. Well no, I don't have letters after my name, but I was among the first post-war children to be given an intelligence test (now IQ test). I remember too, I finished so quickly that the boy next to me copied my answers, the teacher caught him. I passed my 'qualifying' with an 'A' the top mark, I was three years at High School, when high school meant high school but had to leave as I came from a poor family and had to go out to work to bring money into the house.

SEB mentions that my writings are the same as those on DIY/Hi-Fi websites, sorry to disappoint SEB, but I don't own or have had a PC. So if my thoughts are the same as those on web sites, it is coincidental, although judging by the emails from that source on John's demise they sound very intelligent so I'll take it as a compliment.

Although not relevant here, for the past 20 years I have been a member of the theosophical society of Britain, from which I was given a diploma for intellectual and spiritual reasoning in philosophy, religion and reincarnation, etc.

This takes me onto Clive Stevens' article, very interesting, on a practical level, it looks like a diode valve, Sun/Anode, - Earth/Cathode. Even DoWH to deposits on the earth/cathode. On time he's 'hit the nail on the head'. On a scientific, and for that matter also spiritually, you can't argue against time. No Time = No distance = No human body spirit, only the end as life as we know it, humanly. In spirit, you think of those you love and they are with you, no time, no distance. You think therefore it is. What Clive says regarding the end of time is also said to be the Bible, but in 2000 year old terms = the sun and moon shall give light – the stars shall fall from the sky. The heavens shall roll up like a document. There are now several cosmologists who now believe there is a God. I realise this is a 'hot potato', arguments on both sides but it's what I believe in and won't be changing my mind.

Lastly, if I was in a pub drinking the local ale and looking for some interesting conversation and had to choose between men in grey suits and Ivor Catt, Ivor would win every time, I certainly like his style.

D. Lucas
Anstruther
Fife
Scotland
UK

Catt's litter

Ian Hickman's article *The Catt anomaly, Electronics World* October 2004, p38, compresses history. In 1982 Catt suggested that there was an anomaly in Classical Electromagnetic Theory. However, after some decades of suppression (of this suggestion and also of his own theories), Catt decided to concede that the reigning Electromagnetic Theory of 1910 was perfect, as many experts have assured us. Catt then asked 'The Catt Question', which humbly asks for detail on the perfect theory which has ruled for a century. 'The Catt Question' should be minimal, merely asking where the negative electric charge, which all agree appears on the bottom conductor, comes from. It asks nothing about how and why it reaches its necessary position.

A decade ago, it took four years to force two luminaries, Pepper FRS and McEwan, to comment. They contradicted each other. All luminaries then went silent.

The £2,000 letter, published by me in the August issue of *EW*, p57, offers money to any student who prevails on his accredited expert to write anything on the subject.

'The Catt Question' is only a question, and makes no assumptions. Ian Hickman is wrong to write; "Ivor assumes they are both wrong. On the contrary, I maintain, they are in fact both right."

Hickman knows, following Professor Ziman's repeated statement; "The aim of science is to achieve consensus," that it is necessary for all salaried luminaries to sing from the same hymn sheet, so as not to frighten the horses (students). If there are two conflicting theories, Westerner McEwan's and Southerner Pepper's, then students must be warned. Otherwise, in confusion and despair, the number of students studying physics will drop even more rapidly.

If, as Hickman asserts, both Westerners and Southerners are right, then it is necessary that such luminaries, for instance McEwan, Reader in Electromagnetics, and Pepper FRS, not Hickman, say so. In fact, McEwan and Pepper both say the other is wrong; ".... I am prepared to take slight issue with Prof Pepper - again in a completely friendly way I hope - about the main component of the velocity ....". Southerner Pepper says; ".... charge supplied from [the west] outside the system would have to travel at light velocity as well, which is clearly impossible." It is no solution for Hickman contradict them both, and to write; ".... they are in fact both right." They, not Hickman, control the content of university courses. Hickman merely provides obfuscatory waffle to give them cover. Try to keep it simple. Once I came to accept that nothing new in electromagnetic theory is allowed, I spent decades honing this simplest possible question on the old.

With Harold Hillman, Reader in Biophysics, and others, I have found worldwide cases in science and academia where all of today's experts refuse to define their ruling theories. One lethal example is AIDS.

Historically, it is unprecedented for all text book writers and salaried expert teachers to be exposed for refusing to define the rudiments of their craft. Biophysics lecturer Dr Luca Turin, UCL, comments on The Catt Question; "It belongs in Chapter One of all the textbooks." The implications go far beyond an abstruse technical question. Further information is at www.ivorcatt.co.uk or www.ivorcatt.com/44.htm

Ivor Catt,
St. Albans
Hertfordshire
UK.

Pseudo science

In the August 2004 issue, page 52 *Engineering versus pseudo-science* Mr. Green makes a fundamental error about Newton's second law. The second paragraph is trying to illustrate the second law (F=ma) and at the end of the paragraph gives an incorrect example.

"In fact modern car manufacturers agree with Aristotle, using bigger engines in cars when they want higher top speeds." is correct by itself, but it has nothing to do with F=ma. A larger engine usually brings a higher power-to-weight ratio and thus a higher ACCELERATION, which would be relevant to the paragraph. A "higher top speed" has to do with the resistance of motion (the air resistance, rolling resistance and transmission losses) balancing with the output force of the tyres on the road, so a larger engine would allow a car to go faster.

I don't think this point is pedantic, but a fundamental piece of understanding of forces. It is rather amusing that the first sentence thus applies to the writer of the article: "*Men are deplorably ignorant to natural things...*" in this particular situation. Thanks for a great magazine

Peter Rolfe
By email

Not me guv

Like Mr Andy Holt, I too am amused by the articles of Mr. Ivor Catt.

May I fill in a gap he left in his May article, according to the letter of Mr. Penny, by adding some additional information?

The influence of residual magnetism in mains transformers is underestimated and often unknown. Compared with the old fashion laminated silicon steel transformers, the use of a modern ferroxcube ring-core transformer is more problematic.

As a transformer is switched off, the core can have residual magnetism. If the transformer is switched on, such that the mains current creates a magnetic field in the same direction as the magnetic remanence, the core will saturate. While the change of the magnetic flux drops almost to zero, the primary induction disappears and the current is only limited by the low DC resistance of the primary coil. To cope with this high surge current, there is a need for a high value primary fuse. But this is, in case of a failure, in conflict with the safety requirements when a much lower fuse value is needed to avoid red glowing components on the PCB board.

I add some measurements to show the problem.

With a modern digital oscilloscope and some means of switching on and off the mains at a defined point in the voltage cycle, it is possible to measure the behaviour of transformers.

The setup to measure the hysteresis loop was picked out of

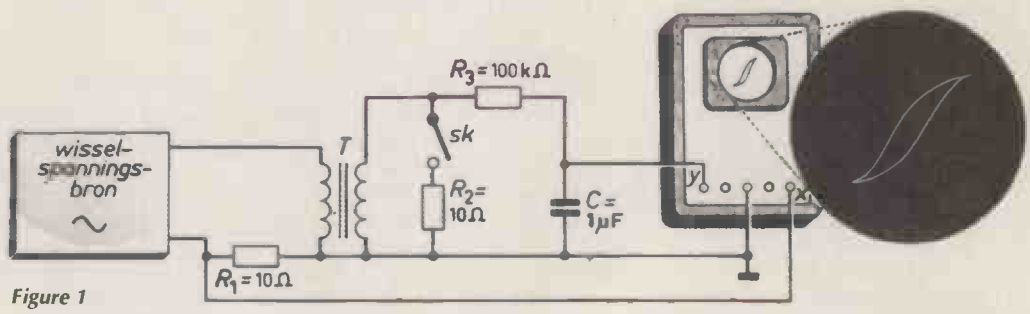
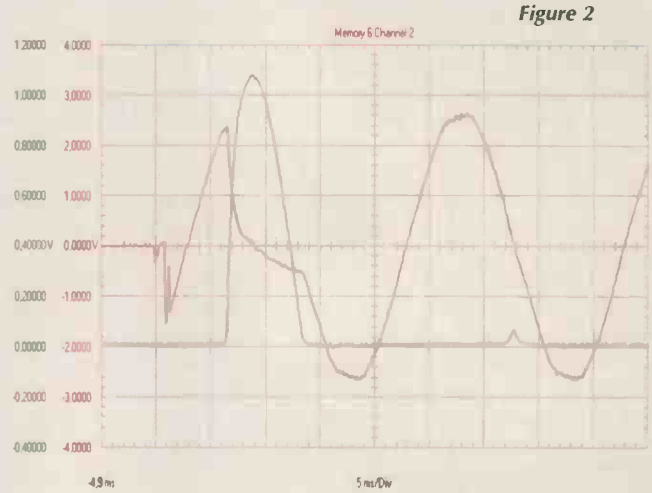


Figure 1

Figure 2



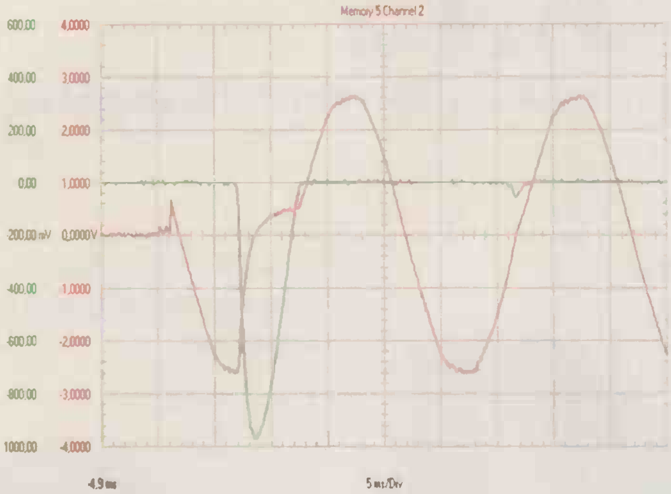


Figure 3

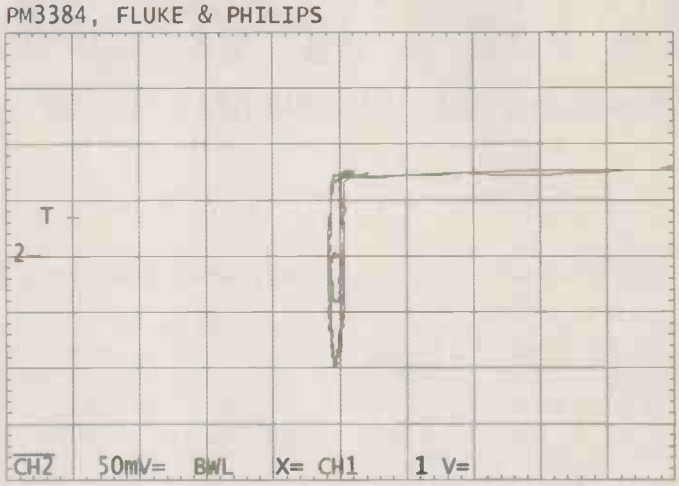


Figure 4

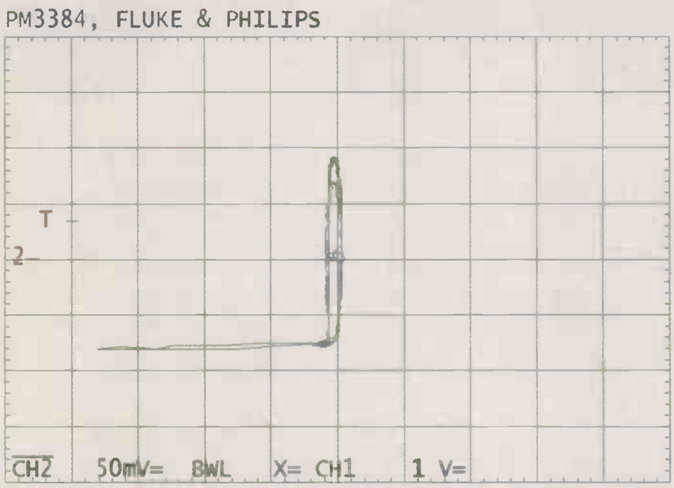


Figure 5

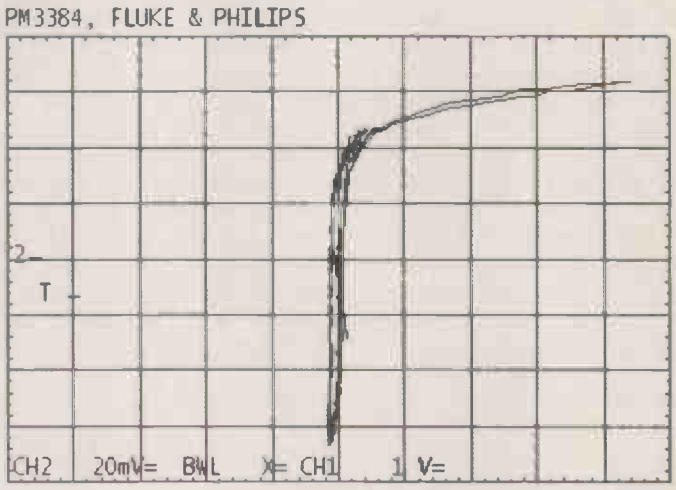
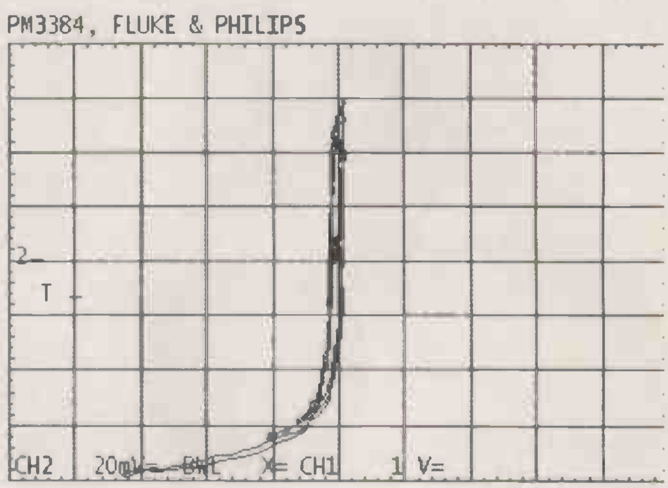


Figure 6

Figure 7



the book *101 Proeven met de oscilloscoop*, NV Centrex, Eindhoven 1966, see **Figure 1**. In that time this Dutch publication was also available in English, French and German. The resistor value R_1 was changed to 0.05 ohm for the current measurements. R_3 with C serve as a 90-degree phase shift. R_2 is the transformer load resistor. The oscilloscope is set in the X-Y mode.

WARNING: Note that the oscilloscope is connected to the mains. Take the necessary safety precautions.

A few years ago my attention was caught while a 1A fuse, 20x5mm IEC127 slow, blew up when I switched on an unloaded ring-core transformer. It became even stranger when I did measure a high inrush current, above 22A peak, while at the same time the secondary voltage dropped to zero. See **Figure 2** and **Figure 3**. The single spike is the primary current plot. It can occur in the positive or the negative phase of the supply voltage. It depends on the magnetisation prehistory. The other line in the plot is the secondary output voltage.

The manufacturer of this ferroxcube transformer was unknown, but by it size it was estimated as a 100VA-120VA transformer with a primary DC resistance of 13 ohms.

Later on I did compare the results with a similar old silicon steel E-I core transformer, Monacor TR 134, bought in 1980, and specified as primary 220V, secondary max 24V, 5A. The measured primary DC resistance was 11 ohms.

I used different load conditions with both transformers, but I only include the measurements with a 22 ohm load. It shows best the hysteresis curve.

Figure 4 and **Figure** 5 show the curve of the ferroxcube ring-core transformer and the large current overshoot in fig. 4. The X-axis is related to current. The Y-axis is the phase shifted secondary voltage across the capacitor in fig.1.

Figure 6 and **Figure 7** show that even with silicon steel a slight saturation is possible, but extreme current spikes did not occur. The saturation bend is slightly less sharp as can be expected with this kind of material. Note that the sensitivity of the Y channel is 20mV instead of 50mV.

E. Vanderfeesten
Belgium

NEW PRODUCTS

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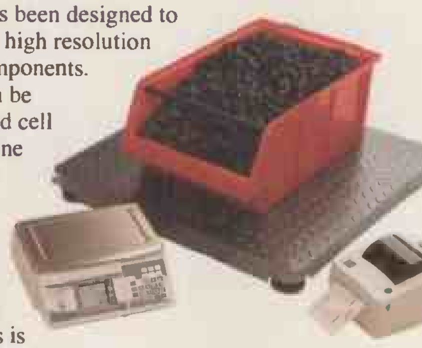
RPR solution

An IEEE 802.17 standard compliant Resilient Packet Ring (RPR) solution for the Intel® IXP2XXX product line of network processors is available from Wipro Technologies. It includes integrated data plane and control plane components implemented on Intel® IXP2XXX network processors and allows for more efficient use of SONET/SDH ring bandwidth while transporting Data traffic. Common applications for this solution are in the Metro Optical and Ethernet products to enable efficient transport of Ethernet over existing infrastructure. RPR offers a sub 50ms restoration time after a failure and up to 8 times over-subscription of the network under certain conditions.

www.wipro.com

New scale you can count on

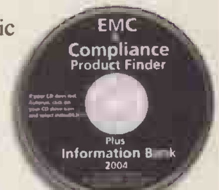


A new weighing scale has been designed to provide fast and accurate high resolution counting of parts and components.

The bench top scale can be connected to any four-load cell platform, with up to 6 tonne capacity. The G235 from Avery Weigh-Tronix is easy to operate with a range of counting methods provided to suit the product. A choice of flexible sampling routines is available including reverse and negative sampling which eliminates unnecessary handling of delicate components. An accumulator function allows the operator to break a large quantity into smaller batches which will be totalled at the end of the routine. RS232 output is supplied as standard for connection to a printer or PC. The software includes 8 print format options offering a variety of layouts. A re-chargable battery is optional, which provides 200 hours of continuous operation, or it can be powered from six dry cell batteries for 800 hours local base use only.

Enquiries: info@awtxglobal.com
0870 90 50061

Free CD...everything you wanted to know about EMC



EMCIA is the UK trade association for companies involved in electromagnetic compatibility. Formed in 2002 and comprising 34 members the association is also an accredited trade organisation with UK Trade and Investment. Available now is the EMC & Compliance CDROM which contains a relevant directives with lists of appropriate standards, a product finder and directory of companies. Included are articles covering design techniques for EMC, EMC for systems and installations and EMC testing.

www.compliance-club.com

The changing face of watches

Fossil Inc., designers of contemporary, quality fashion watches and accessories, has teamed up with the flexible display technology company, Pelikon, to produce the world's first wristwatch incorporating complex segmented electroluminescent (SEL) display technology.

Pelikon, which developes the application of SELs and InterfaceDisplays™, has been working with Fossil to produce a series of eye-catching, innovative watch face displays that feature both animation and illumination.

Fossil designers have worked closely with Pelikon to develop the concept, taking their current range of animation watches to the next level with the addition of light-up features. With Pelikon's ability to use SEL technology to deliver new watch faces incorporating features that customers will have never experienced before.

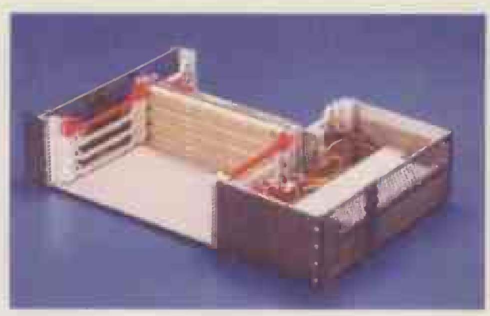
The SEL display at the heart of the watch is the thinnest self-illuminating display commercially available; ten times thinner than the animated LCD it replaces. Sitting behind the traditional analogue hands, the SEL display, a microprocessor controlled animation, fires up at the brief touch of a button. Hold that button down, and the whole face lights at once, backlighting the hands for night-time time telling.

Available in the US now, and worldwide in Spring 2005.

www.fossil.com
jholliday@fossil.com

Enquiries regarding SEL displays contact sales@pelikon.com

High density horizontal solutions for CompactPCI and VME



Rittal has extended its range of Slim-Box products that reflect the trend for smaller and more efficient systems. The high-density electronic packaging solutions feature 2 slots per 1U of height. The new models include VME as well as increased range of CompactPCI (CPCI) systems and "neutral" versions for custom backplanes, which can include switched fabric.

All models in the horizontal mount range are equipped for 6U front plug-in boards and include fans for left to right side cooling. The options, available as standard configurations, include provision for rear I/O transitions, plug-in hot swap or A TX power supplies, and "neutral" systems without a backplane for greater customer choice.

The systems that can be configured as 482.6mm rack mounting units or, without mounting flanges, as desk or bench top systems are built from steel, with mounting flanges that may be fitted either at the front or up to 100mm towards the rear if preferred. Inside, each 1U of height is equipped with 4 codable guide rails to create 2 standard (4HP) slots, with additional models also having 80mm rear I/O (CPCI) or 160mm rear I/O (VME).

Rittal Electronics Systems
www.rittal.co.uk

Please quote *Electronics World* when seeking further information

Seeing the way forward

A monitoring and display breakthrough from Cambridge Consultants could give athletes the means to optimise their training and performance without the need for the conventional armoury of sensors and wrist displays. Athletes of tomorrow will need only to wear a pair of glasses that incorporate everything needed to monitor and display key biomedical parameters in real time.

Performance improvement depends increasingly on access to detailed biomedical data. Technospecs incorporate both data capture and information delivery. Sophisticated sensors measure factors such as heart rate and blood oxygen levels, and that information is delivered in real-time in the form of a head-up display originally developed for jet fighter pilots.

Behind Technospecs lies work on a number of technologies that will transform the concept into a commercial product. One further intriguing implementation possibility is the addition of a wireless communications link to the Technospecs chip, so that an athlete could transfer performance data to a PC for detailed analysis, or to the trainer.

The compact head-up display technology also presents opportunities in automotive sport where the ability to display sensor readings in a helmet or viewfinder could eliminate the need to look down at controls. Monitoring of real time physical data also has applications beyond sport, such as in the armed forces or emergency services.

Cambridge Consultants
www.cambridgeconsultants.com

Envirowise and Intellect announce WEEE and RoHS supply chain initiative

Environmental support and advisory service, Envirowise, has launched a free service to help UK electronics and computer manufacturers set up product supply chain partnerships aimed at improving environmental performance, achieving cost savings and complying with WEEE and RoHS legislation on waste electrical and electronics equipment and hazardous substances.

Envirowise will provide free support and advice on design issues, waste minimisation, compliance with WEEE and RoHS legislation, and fulfilment of ISO 14001 commitments to help the host company manage its supply chain environmental performance, coordinate promotion and awareness, organise events and workshops, and offer free on-site design and environmental consultant visits and counselling.

Companies requiring further information on electronics industry supply chain partnerships should contact Philip Price at Envirowise on 0870 190 6355, fax him on 0870 190 6361, or e-mail: philip.price@envirowise.gov.uk.

Organisations who require general advice and information on WEEE and RoHS or companies who wish to take advantage of free Designtrack and FastTrack one-day site visits can obtain further information through the Envirowise website (www.envirowise.gov.uk) or via the Environment and Energy Helpline – Telephone 0800 585794.

New EMC antenna

Schaffner Ltd has introduced the BHA 9220 horn antenna for EMC immunity and emission testing. A linearly polarised broadband antenna for EMC and RF measurements over the frequency range 200MHz to 2GHz, it may be used both as a transmitter and receiver. It complements the existing BHA 9118 model, which covers the frequency range 1 to 18GHz.

All Schaffner antennas are individually calibrated at their in-house UKAS accredited facility, which also offers on-going re-calibration services. A full range of accessories is also available including, masts, tripods, adapters and mountings.

Full details of the Schaffner Antenna range can be found in the new EMC Antennas catalogue, please contact Andrew Kotas on +44 (0) 118 977 0070 for a copy or email AKotas@schaffner.com

www.schaffner.com

New 3-in-1PXI FlexDMM instrument

The National Instruments PXI-4072 FlexDMM gives engineers the functionality of three common instruments in a single-slot 3U PXI module – a 6½-digit multimeter, an LCR meter, and a 1.8 MS/s isolated digitizer.

The NI PXI-4072 FlexDMM brings digital multimeters one step closer to providing universal measurement capability by offering engineers the 20 most common ATE measurements, including voltage, current, capacitance, inductance, temperature and resistance. By integrating these measurements into one PXI module, engineers reduce test system size and cost, increase throughput and shorten test development time.

www.ni.com/uk

Queen's Award – nominations wanted

Outstanding businesses from the electronics sector have just weeks left in which to apply for a Queen's Award for Enterprise – the UK's most prestigious accolade for business success. Applications can be made in three categories – International Trade, Innovation and Sustainable Development. The Awards are open to any sized business, from two-person partnerships to large multinationals.

This year for the first time, nominations are also being invited from the electronics sector for individuals felt to be worthy of a new royal Award for contributions to encouraging a UK enterprise culture.

The Queen's Award for Enterprise Promotion is for outstanding individuals who promote enterprise by teaching other people enterprise skills or who, as mentors or role models, inspire others to become successful entrepreneurs. Nominations are sought from colleagues, associates and beneficiaries of their work.

The deadline for all applications and nominations is 31 October 2004.

For more information, visit www.queensawards.org.uk

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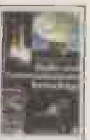
NEWNES GUIDE TO RADIO AND COMMUNICATIONS TECHNOLOGY

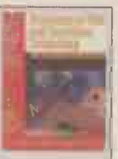


Ian Poole

This is a guide to the technology and applications of modern radio and communications equipment. The author's approach provides a useful foundation for college students and technicians seeking an update on the latest technology.

Jul 2003 ▲ 352 pages ▲ Index ▲ PB
Published in UK
Code 0-7506-5612-3 **£16.99**

DICTIONARY OF VIDEO AND TELEVISION TECHNOLOGY

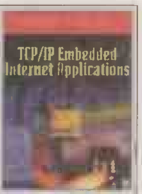


Jack Tsatsoulin

This work provides comprehensive and contemporary information on the essential concepts and terms in video and television, including coverage of test and measurement procedures. The CD accompanying the text includes an electronic version of the book.

Sept 2002 ▲ 365 pages & CD-Rom
Published in UK
Code 1-878707-99-X **£29.99**

TCP/IP EMBEDDED INTERNET APPLICATIONS

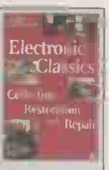


Edward Insam

This text highlights an engineer's approach to Internet protocols and applications, reducing R&D time for engineers. The book aims to be the main design guide for the cutting edge of Internet-enabled electronic products and systems.

Aug 2003 ▲ 384 pages ▲ Index ▲ Appendix
PB ▲ Published in UK
Code 0-7506-5735-9 **£32.50**

ELECTRONIC CLASSICS: COLLECTING, RESTORATION AND REPAIR



Andrew Emmerson

This text encompasses all aspects of buying, collecting, restoring, repairing, sourcing parts, professional services, clubs and societies. The first part covers technical aspects of restoration and details where components can be found. The second part presents useful information for collectors.

Aug 1998 ▲ 256 pages ▲ Index
10 halftones ▲ 50 line illustrations ▲ PB
Published in UK
Code 0-7506-3788-9 **£21.99**

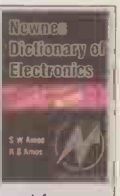
INTRODUCTION TO DIGITAL SYSTEMS

John Crisp

This self-study text introduces digital electronics from first principles, before going on to cover all the main areas of knowledge and expertise. It covers the practicalities of designing and building circuits, including fault-finding and the use of test equipment.

Feb 2000 ▲ 302 pages ▲ Glossary ▲ Index
PB ▲ Published in UK
Code 0-7506-4583-0 **£18.99**

NEWNES DICTIONARY OF ELECTRONICS



S W Amos; R S Amos

Aimed at engineers, technicians and students working in the field of electronics, this dictionary provides clear and concise definitions, including TV, radio and computing terms, with illustrations and circuit diagrams.

4th edition ▲ Mar 2002 ▲ 394 pages
100 illustrations ▲ PB ▲ Published in UK
Code 0-7506-5642-5 **£12.99**

NEWNES GUIDE TO TELEVISION & VIDEO TECHNOLOGY

Eugene Trundle

An exploration of television and video technology. It covers the fundamentals of digital television (satellite, cable and terrestrial) and digital video, as well as providing a grounding in analogue systems.

3rd edition ▲ Feb 2001 ▲ 432 pages ▲ Index
PB ▲ Published in UK
Code 0-7506-4810-4 **£17.99**

NEWNES GUIDE TO DIGITAL TV

Richard Brice

Covering all aspects of digital television, this text encompasses the electronics of the equipment, data compression, television production, servicing and the different transition methods - terrestrial, satellite and cable. The text has been updated with developments since the 2000 edition.

2nd edition ▲ Oct 2002 ▲ 304 pages ▲ Index
45 illustrations ▲ 15 photographs ▲ HB
Published in UK
Code 0-7506-5721-9 **£24.99**

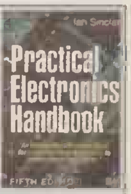
PRACTICAL ELECTRONIC FAULT FINDING AND TROUBLESHOOTING

Robin Pain (Design Engineer, Cotag International Ltd)

A text using simple circuit examples to illustrate principles and concepts fundamental to the process of analog and digital fault finding. It aims to help the reader tackle any job, from fixing a TV to improving the sound of a hi-fi. A digital multimeter and oscilloscope are needed for these jobs.

Apr 1996 ▲ 284 pages ▲ Index
50 line illustrations ▲ PB ▲ Published in UK
Code 0-7506-2461-2 **£21.99**

PRACTICAL ELECTRONICS HANDBOOK



Ian Sinclair

A collection of all the key data, facts, practical guidance and circuit design basics needed by a spectrum of students, electronics enthusiasts, technicians and circuit designers. It provides explanations and practical guidance, and includes new sections on SHF techniques and intruder alarms.

5th edition ▲ Feb 2000 ▲ 571 pages
Illustrations ▲ PB ▲ Published in UK
Code 0-7506-4585-7 **£16.99**

RSGB RADIO & ELECTRONICS COOKBOOK

Radio Society of Great Britain

Only a basic knowledge of electronics is assumed for this collection of electronics projects, and it is ideal for all electronics and DIY enthusiasts and experimenters. Designed by the RSGB, the UK radio amateurs federation, the projects are clearly explained step by step.

Nov 2000 ▲ 336 pages ▲ PB ▲ Illustrations
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Code 0-7506-5214-4 **£17.99**

REFERENCE DATA FOR ENGINEERS: RADIO, ELECTRONICS, COMPUTERS AND COMMUNICATIONS

Mac E Van Valkenburg; Edited by Wendy Middleton

Written by professionals for professionals, this is a complete reference for engineers, covering a broad range of topics. As well as addressing radio technology data, it covers digital electronics, computers and communications.

9th edition ▲ Aug 2001
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FUZZY LOGIC FOR EMBEDDED SYSTEMS APPLICATIONS

Dr Ahmad Ibrahim

Written by a recognised expert in the field, this covers all the basic theory relevant to electronics design, with particular emphasis on embedded systems. It shows how the techniques can be applied to shorten design cycles and handle logic problems that are tough to solve using conventional linear techniques. All the latest advances in the field are discussed, and practical circuit design examples presented.

Oct 2003 ▲ 312 pages ▲ PB
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Stuart Ball

Provides hard-to-find information on interfacing analog devices and technologies to the purely digital world of embedded microprocessors. Gives the reader the insight and perspective of a real embedded systems design engineer, including tips that only a hands-on professional would know. Covers important considerations for both hardware and software systems when linking analog and digital devices.

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Farid Dowla (Lawrence Livermore National Laboratory, University of California-Davis, USA)

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November 2003 ▲ 720 pages ▲ HB

Code 0-7506-7695-7 **£59.99**

DIGITAL INTERFACE HANDBOOK

John Watkinson, Francis Rumsey

A detailed manual for those who need to get to grips with digital audio and video systems. It sheds light on the differences between audio interfaces and shows how to make devices "talk to each other" in the digital domain. Also includes detailed coverage of all the regularly used digital video interfaces. New information included in this third edition: dedicated audio interfaces, audio over computer network interfaces and revised material on practical audio interfacing and synchronisation.

3rd edition ▲ September 2003 ▲ 392 pages
231 line illustrations ▲ PB

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TV FAULT-FINDING GUIDE

Edited by Peter Marlow

A distillation of the most-used fault reports from 11 years of *Television* magazine. Arranged by make and model, it features over 200 reports on over 300 models of television, including diagnosis and repair advice.

Mar 2000 ▲ 387 pages ▲ Illustrations
PB ▲ Published in UK

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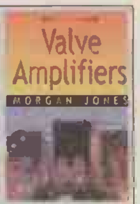
VALVE AMPLIFIERS



Morgan Jones

The author's straightforward approach, using as little maths as possible, should be of use to those with only a limited knowledge of the field as well as being the standard reference for experts in valve audio. Design principles and construction techniques are also provided.

3rd edition ▲ Aug 2003 ▲ 624 pages ▲ Index
PB ▲ Published in UK

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VCR FAULT-FINDING GUIDE

Edited by Peter Marlow

A distillation of the most-used fault reports from 11 years of *Television* magazine. Arranged by make and model, it features over 2000 reports on over 200 models of VCR, including diagnosis and repair advice.

Mar 2000 ▲ 447 pages ▲ Illustrations ▲ PB
Published in UK

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John Holland

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January 2004 ▲ PB & CD-Rom ▲ 352 pages

Code 0-7506-7683-3 **£35.00**

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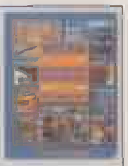

Steve Beeching

A comprehensive guide to domestic VCR technology and repair techniques. This edition brings the information fully-up-to-date, with expanded coverage of camcorders, sections on DVD equipment and the latest VCR technology.

5th edition ▲ Apr 2001 ▲ 323 pages
Illustrations ▲ PB ▲ Published in UK

Code 0-7506-5039-7 **£20.99**

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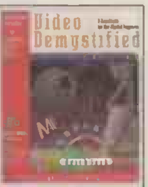

Keith Jack

This edition has been updated to include information on digital television, datacasting, interactive video, digital camcorders and VCRs, and video interfacing. Coverage is international, including European, Asian and North/South American video standards, methods and techniques.

3rd edition ▲ Jul 2001 ▲ 784 pages & CD-Rom
References ▲ Glossary ▲ Index ▲ PB
Published in UK

Code 1-878707-56-6 **£50.00**

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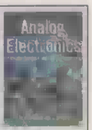


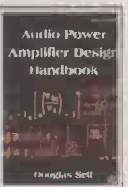
D Crecraft (formerly Open University); S Gergely (formerly University of Coventry)

Designed to complement the digital systems modules and develop the essential skills needed for RF circuit design, this book demystifies the art of analogue circuit design and analysis. The content meets the requirements of first and second year electronics engineering courses.

May 2002 ▲ 435 pages ▲ 150 line illustrations
PB ▲ Published in UK

Code 0-7506-5095-8 **£19.99**

AUDIO POWER AMPLIFIER DESIGN HANDBOOK



Douglas Self

This text on the design of audio amplifiers is based on Douglas Self's work in *Electronics World* magazine and a career at the cutting edge of audio design. This handbook aims to provide a practical guide to the state of the art, and includes detailed design and construction information.

3rd edition ▲ May 2002 ▲ 448 pages
62 line illustrations ▲ Index ▲ PB
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John Morton (University of Cambridge)

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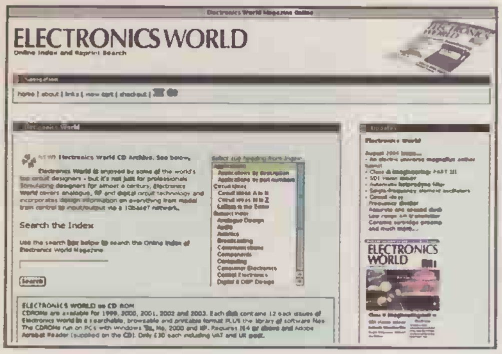

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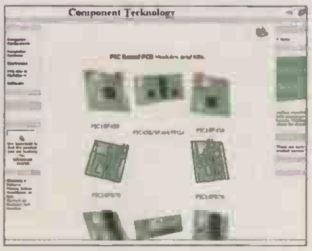

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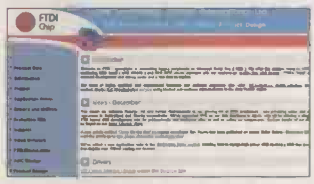

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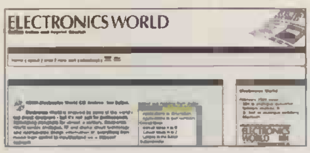

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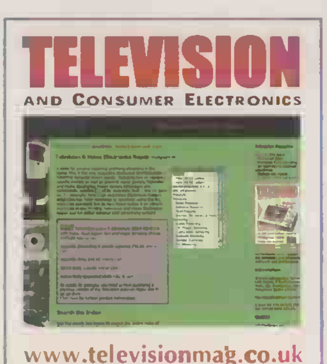

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TTC Fireberd PR-45 Printer For Fireberd 6000	350	15
TV & VIDEO		
Calan 3010R Sweep / Ingress Analyser	2950	124
Minolta CA-100 CRT Colour Analyser	2000	60
Philips PM5515I+RG8 TV Pattern Generator with RG8	1650	50
Philips PM5515IT+RG8 TV Pattern Gen with Teletext+RG8	1750	55
WIRELESS		
AT/HP 3708A Noise & Interference Test Set	4950	199
AT/HP 3708A/001 Noise And Interference Test Set	5750	222
IFR 2967 Radio Comms Test Set with GSM	5950	245
IFR 54421-003J RF Directional Power Head	250	20
Marconi 2945/05 Radio Comms Test Set	5950	179
Marconi 2955A/2957A 1GHz Radio Comms Tester With AMPS	2500	90
Marconi 2955B 1GHz Radio Comms Test Set	3500	126
Marconi 2955R 1GHz Radio Comms Test Set	2950	107
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01753 596000
Fax: 01753 59 6001
www.TestEquipmentHQ.com
email: info@TestEquipmentHQ.com

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