Electronics World's renowned news section starts on page 4

ELECTRONICS WORLD

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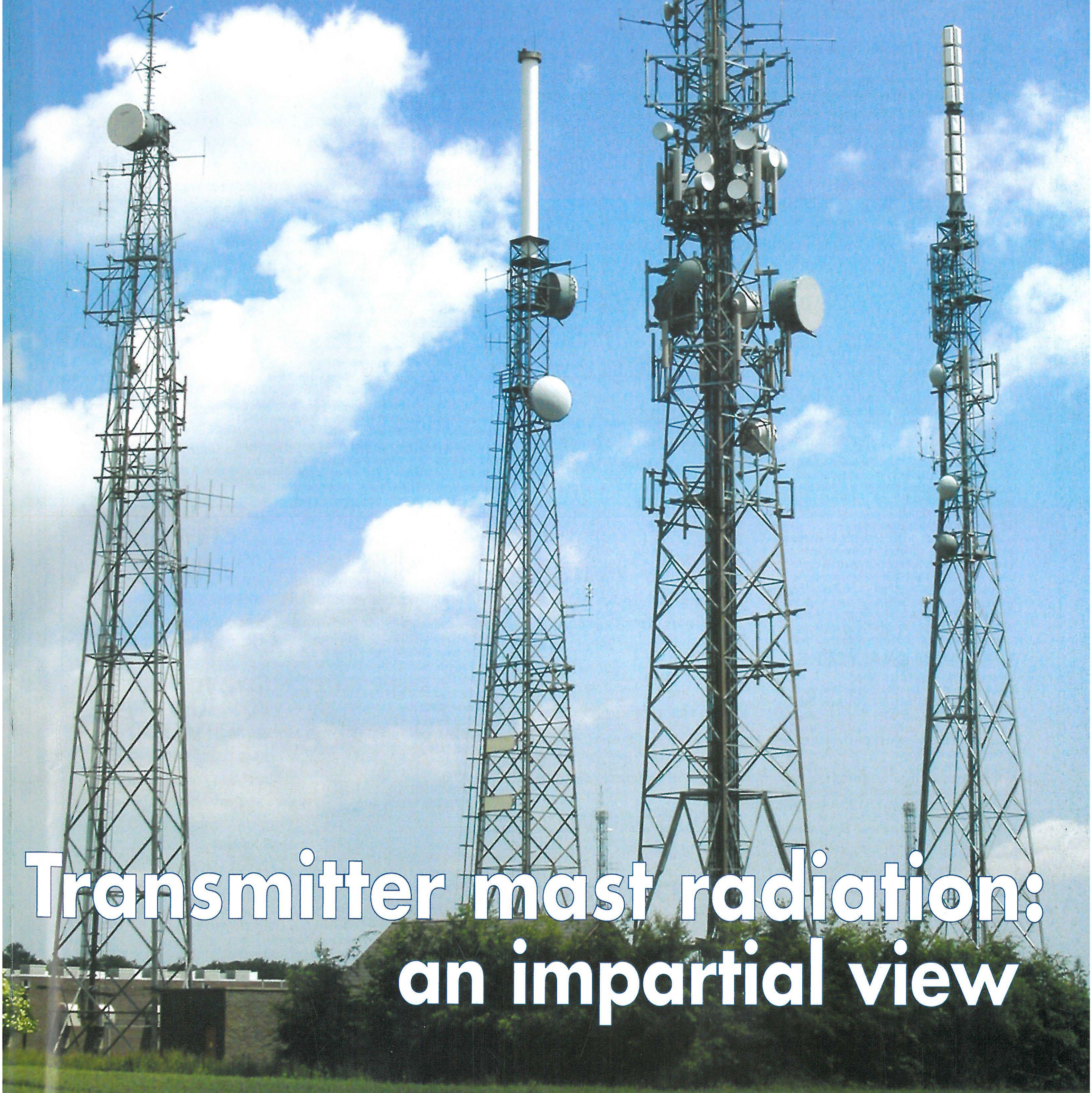

Transmitter mast radiation: an impartial view

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Running out of time!

Many of you must have heard by now that the new CEO of WH Smith seems to think that readers of specialist magazines such as ours are persona non-grata on the scale of her revenue account. Quite wisely I took on subscriptions to my favourite magazines a few years ago. I have also just taken on an electronic subscription to another of my favourite electronic magazines (not *EW*), simply because space is running out on my shelves and in the long run we cannot stop progress. You will gather by now I read profusely and keeping up to date with the industry is crucial for me. But it is still not enough, and I will explain why.

My basic theme is simple; we have got to be smart to survive. For publishing, if not just about anything you do nowadays, is a commercial enterprise of some sort. I can think of many reasons why WH Smith now considers displaying specialist magazines a liability and not an asset, but I don't have time to go into it here.

I started life as an undergraduate in the 1970s and struggled through university at considerable expense. I had good grades (a Bachelor's degree with honours was worth a lot in those days) and I was immediately offered a job with one of the then major British electronic manufacturers whom we shall just call Company X. I stayed for a little over a year and left in spite of the temptation for increased pay offers and promotions as I had solved one of their project's X problems which made millions in later years. I returned to the university (taking a huge loss in salary) because my supervisor offered me a scholarship and he insisted that I was PhD material. Why? The first reason was that the job was really quite boring to my young mind, I could learn nothing new day by day. If it hadn't been for magazines like *WW* now *EW* and many others, I would have been bored to death from intellectual starvation. I need you to survive, I still do. More important to me was that underneath it all, I was not convinced by Company X's performance, in spite of their shareholders naively pumping in millions. Company X went under three years after I left. Lesson no.1: "Better to have an empty pocket than an empty head" (Charley Swayne).

The university system is no better! I went on from being a PhD student to being a faculty member of a world-class university of excellence. Then after nearly two decades service in academia, acquired a long publication record, having brought in collaborative research grants worth millions, produced other PhDs, taught myself to death in over-crowded lecture halls; I was dumped at the age of 45 with compensation that I could barely retire on at that age. That scene is getting worse today. The new mega universities with their pseudo executive CEO Vice-Chancellors/Presidents all the way down to so called line managers (formerly HODs), manufacturing university graduates like a sausage production line, are now calling themselves corporations and some of them are getting away with being paid real world executive salaries and company benefits. Where is the money coming from? Many of these institutions are still governed by numerous dead wood overpaid university dons from the 60s era and are now infiltrated mostly by clones of themselves, with zero fresh ideas to bring forward to this millennium. None have the honesty to even admit that the whole system is in shambles, let alone to stand up against any government using them as a rubbish heap for mostly unemployable and misguided youths. Sadly, these are the real world victims, not me I am afraid, I have my own troubles; remember my battle is to protect my pension funds, no time to fight the cause of the future generation's education.

Lesson number 2 - make sure you are financially independent by age 45, your priority is to make your million by then, because by the time you retire at 65, your million would have the buying power of less than 10% of its initial value - if you are lucky and a wizard at financial management. If you are not awake to it yet, try reading Rich Dad, Poor Dad immediately, but I must warn you, you have to be smarter than even this book's author Robert Kiyosaki. Somehow a world in which you can just walk around profiting through shares, property and other investments, and not ever having to work just doesn't add up. Who will be the doctors, lawyers, carpenters and of course engineers?

After my first redundancy, I went back to work for another major multinational. They trained me to top management, in marketing, finance and business administration, all the things I initially lacked. Guess what? I was made redundant during the September 2001 crash, with a reasonable compensation that still cannot see me through my retirement years as I was then only 55. I am now running my own business, thanks to others who paid for me along the way. With luck I might just be able to retire at 65.

So live smart and prosper for if you are reading this column, I assume you can read, you can add, you know Ohm's law and perhaps even Maxwell's equations. You will be a rare breed in another ten years. Being smart, you and your descendents will control the world economy, have hidden solutions for the food and energy shortage, be able to travel through time and pick the best of the ages to live in – without paying a single penny of tax in this one.

B. H. Schottky

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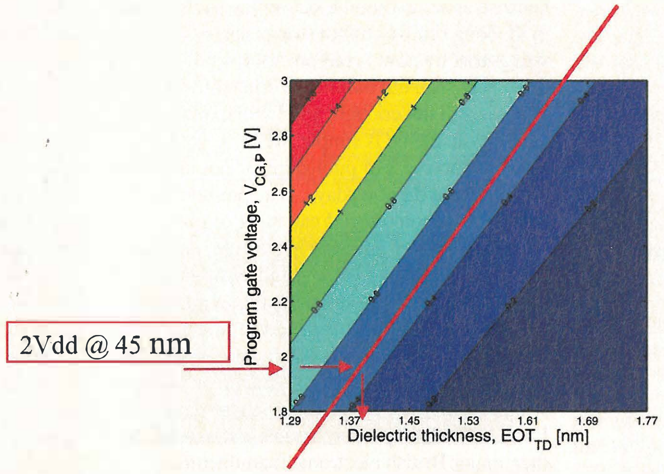
Partners sought to develop novel embedded RAM

IMEC, the Belgian powerhouse for IC research, has begun a project to develop the next generation of memory for production using 45 nanometre process rules.

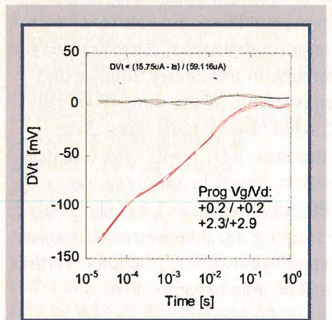
Potential collaborators are invited "to participate in an IMEC industrial affiliation program on embedded RAM concepts for second and higher levels of on-chip cache memory", said IMEC.

The programme will study three concepts: direct tunnelling RAM, ferroelectric fet and floating body cell devices. "The three concepts will be implemented in silicon by year-end to demonstrate their feasibility," said IMEC.

Direct-tunnelling RAM uses a very thin (around 1.5nm) oxide flash memory structure in which charge can be stored on either a floating gate or on a charge-trapping layer.



Modeling results for a direct tunnelling device showing 0.5V of read window for a 10ns write access time based on IMEC's direct tunnelling model.



Retention characteristics of partially depleted silicon-on-insulator (SOI) mosfets when programmed by impact ionization: a residual 150mV Vt window is obtained after 10μs for a non-optimized device. The red curve represents the decay of the charge stored in the bulk of the SOI device due to junction leakage which has to be optimized in order to increase the refresh time of the floating body cell.

In both cases, the use of high-k dielectric materials is being considered as well to lower the write and erase voltages.

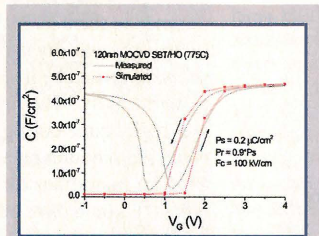
First simulation results of the expected threshold voltage window for different combinations of voltages and tunnel oxide thickness, as obtained from IMEC's tunnelling model, evidence the feasibility of a 10ns programming time at the 45nm node, said IMEC.

The ferroelectric field effect transistor (FeFET) recently regained a lot of attention because of its superior scalability as compared to the capacitor-based ferroelectric RAM. "Also here, the advantage of using high-k materials is substantial since they can be used as a buffer layer between the channel and the ferroelectric in order to lower the write/erase voltages," said IMEC.

Floating body cell is a concept based on the memory effect in silicon-on-insulator (SOI) devices initially developed at IMEC in 1988.

The technology is being adapted for planar as well as FinFET device structures - FinFETs are 3D devices with a gate wrapped around three sides of a channel to increase the gates effect.

Preliminary retention results, obtained on partially depleted SOI mosfets programmed by impact ionisation, said IMEC, show the memory effect in scaled-down SOI technology.



High-frequency C-V curve of a Pt/SBT/HfO2/SiO2/Si metal-ferroelectric-insulator (MFIS) capacitor structure showing ferroelectric hysteresis switching. "Deviation between measured and simulated curve for negative voltages is due to minority carrier supply in the silicon substrate prohibiting deep depletion," said IMEC.

What is the problem with conventional SRAM cache?

IMEC view is: that fast first-level cache memory, which has been and probably will continue to be addressed by static RAMs, are reaching their scaling limits already today due to their drastic increase in relative cell size.

Many high-end microprocessor-based chips will need relatively large amounts of on-chip memory. "Their [the memory] footprint is expected to increase to 80-90 per cent of the chip area in some of these major applications. At the same time, embedded dynamic RAM has never been widely accepted as a mainstream technology option because of limited availability, process complexity and cost issues."

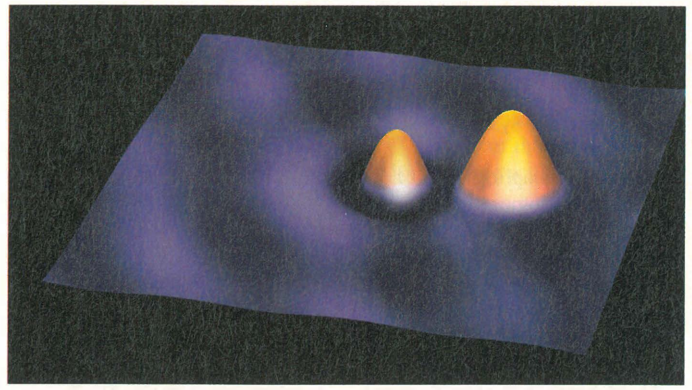
With so much space eaten by first-level cache, new technology is needed to compress the size of second and higher-level embedded volatile memory.

The eRAM project complements IMEC's Flash memory project which started in 2000.

IBM stores charge on individual atoms

Scientists at IBM's Zurich Research Lab and Chalmers University of Technology, Gothenburg have manipulated the charge state of individual atoms.

IBM scientists Jascha Repp and Gerhard Meyer placed and removed a single electron on an individual gold (Au) atom by positioning the tip of a low-temperature scanning tunnelling microscope (STM) above the atom and applying a voltage pulse.



Scanning tunnelling microscope (STM) false-colour three-dimensional image of two gold atoms on an insulating NaCl film surface. The atom on the left-hand side has been intentionally transferred from its neutral state into a negatively charged ion by means of STM manipulation.

Crucial to the charge remaining on the atom is a two atom thick layer of sodium chloride (NaCl) between the gold atoms and the metal substrate.

Charge in the NaCl redistributes as the gold charges, forming a stable local state. In principle, data could be stored long-term on such a structure.

"Both charge states of the atom are stable, that is, an additional electron remains on it until it is removed by a voltage pulse of reversed sign," said IBM.

Owing to the film's large ionic polarisation, in the STM image the new charge state of the gold atom appears as a circular trough around the atom.

"A simple electron transfer with no lasting changes of ion-core positions would not be stable because the electron residing in an excited state on the manipulated Au atom would rapidly tunnel into the metal of the substrate," said Repp.

IC industry hits boom

Worldwide chip sales grew 35 per cent in the first half of 2004, allowing the semiconductor market to reach $102bn.

Figures from market research firm IC Insights show a resugent chip industry, at least in terms of revenue.

The top ten suppliers of ICs are spread globally, with the top five being Intel, Samsung, Texas Instruments, Renesas and Infineon.

Four of the top ten firms – Samsung, TSMC, TI and Infineon – grew at a rate above the average for the top ten. With huge growth in flash memory sales, Samsung grew a staggering 80 per cent when compared with the first half of 2003.

Most analyst firms, including IC Insights, Future Horizons and InStat, believe the chip market will grow at 30 per cent or more for the whole of 2004.

Top ten chip firms

Company	Revenue 1H04	Growth
Intel	$14.9bn	22%
Samsung	$7.4bn	80%
TI	$5.4bn	41%
Renesas	$4.9bn	20%
Infineon	$4.4bn	35%
Toshiba	$4.4bn	28%
ST	$4.2bn	26%
TSMC	$3.7bn	43%
NEC	$3.5bn	30%
Freescale	$2.9bn	26%

MICROMOUSE competitions

On 26 June, Manchester Metropolitan University was delighted to host Micromouse 2004, a national competition for pupils aged 14 to 18 to build and race their self-contained robots – Micromice. The event attracted 40 students and 22 'mice' to compete in the three formula races against the clock and the exciting sudden-death knock-out event, the 'Rat Race'.

Winners: **Formula 4,** King Henry VIII School; **Formula 3,** King Henry VIII School; **Formula 2,** Newcastle-under-Lyme School; **Formula 1 – The Rat Race** Kirkham Grammar School.

There was also a 'grown-up' competition in which our contributor Martin J. Barratt's Micromouse (EW June and July) came third in its class at the Micromouse Competition held at the TIC, Birmingham on the 19th June 2004. In recognition of this it received an award - a silver plated 'brass cheese' (photo).

Funding boost for technology

Science and technology received a £1bn boost from the Government in its latest spending review.

Chancellor Gordon Brown outlined the Government's plans to increase UK spending on research and development to 2.5 per cent of GDP in the next ten years.

This funding plan equates to a nearly six per cent rise in science spending per year until 2008, when public funding for science will reach £5bn, about 0.4 per cent of GDP.

Alongside extra Government spending comes further investment from private sources. The Wellcome Trust has agreed to commit £1.5bn to UK science in he next five years.

The aim is to encourage collaboration between business and universities.

"Only by working with our business and research charity partners will we achieve our goal of R&D in the UK reaching 2.5 per cent of GDP," added Brown.

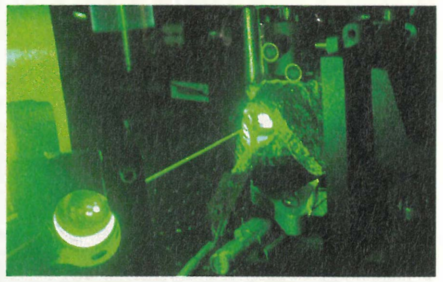
Extra money also goes to science roots in schools. Higher salaries will be given to 'advanced skills teachers' – up to £45,000 in London, teacher training bursaries for science graduates of £7,000, and 'golden hello' payments of £5,000.

Green power laser developed for industry

US laser firm Aculight has produced a 60W 540nm, frequency-doubled, large mode area (LMA) fibre laser.

"We've achieved something very unique with our new fibre laser system," said Roy Mead, Aculight v-p. "We wanted to produce a high-power green laser that was either continuous wave (CW) or quasi-CW. It needed to have excellent beam quality, be highly efficient, lightweight, compact and rugged. Our new system provides 60W of green output at a 10MHz pulsed repetition rate - and it can be even faster than that if we want. We have surpassed all of our goals."

Company engineers initially developed the technology for a military customer. After considering and rejecting modifying a conventional diode-pumped solid-state (DPSS) laser, they chose to develop a fibre system. "There was an opportunity to leapfrog existing technologies to something that had even higher performance in all of the dimensions that our customer wanted," said Mead. "It involved working with LMA fibres."



At the heart of the resulting laser is a ytterbium-doped LMA fibre amplifier capable of achieving high average power, while allowing substantial peak power without the onset of non-linear effects - characteristics essential for efficient frequency doubling, said Aculight.

A seed laser source consisting of a 1,080nm CW fibre oscillator, an amplitude modulator and a preamplifier provides input pulses to the LMA amplifier, which is wound onto a small mandrill to promote good transverse mode quality and polarisation purity. The amplifier is pumped at the output end by free-space coupling optics.

Its output is directed into a pair of frequency-doubling LBO crystals. "The result is a quasi-CW output with excellent beam quality and ten per cent electrical-to-optical efficiency," said Aculight.

"Today's green laser marketplace is dominated by DPSS and argon-ion lasers that produce less than 20W, or function at moderate repetition rates," said Mead. "Our system surpasses those performance characteristics and will likely extend past the 60W that we've achieved so far. And the techniques we've used to wavelength convert can apply to a variety of fibre systems, so will also allow us access to a wide range of wavelengths. This LMA fibre system really fits into a different region of the product space."

Regulation advised for nanotech

Regulation and safety assessments are necessary if the use of nanotechnology is to proceed safely, says a report from the Royal Society and the Royal Academy of Engineering.

In the report, commissioned by the UK Government, a range of benefits in new materials, computing and medicine are identified.

Possibilities include cleaning up contaminated ground, generating solar energy, and targeting drugs to specific parts of the body.

However, risks to people and the environment must be anticipated and managed, says the report. Fortunately, says the report, most forms of nanotechnology pose little or no risk to health. The uncertainty is around small particles on he nanoscale that could get into our environment.

Most areas present no new health or safety risks, but where particles are concerned, size really does matter. Nanoparticles can behave quite differently from larger particles of the same material and this can be exploited in a number of exciting ways. But it is vital that we determine both the positive and negative effects they might have," said Professor Ann Dowling, chair of the working group that produced the report.

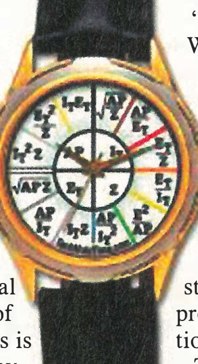
The Royal Society and the Royal Academy of Engineering have called for a Government programme to research possible effects of such particles. Nanoparticles and nanotubes should be treated like other new chemicals, said the report. Approvals for use should be granted by an independent scientific committee. This process has already been undertaken to use titanium dioxide particles in sunscreen.

Nanoparticles, said Dowling, "have different properties from the same chemical in larger form, but currently their production does not trigger additional testing".

Science minister Lord Sainsbury, said: "The Department of Trade and Industry, the Department for Environment, Food and Rural Affairs and Department of Health are already working together with the Health and Safety Executive and the Environment Agency on a number of issues that are raised. The report will help us to focus on the crucial areas and coordinate research across Government."

Time to go Ohm

TechNote Time Watch Company is setting precedence by being the first manufacturer of informative watches geared towards trades-men in the electrical and electronics fields. Their initial product line consists of specialty analogue wrist watches which incorporates Ohm's law, power formulas and a resistor band chart, for AC or DC applications.



It's no secret to electronics and electrical professionals that one of their greatest challenges is the ability to recall many laws, formulas and concepts accurately, or have their reference materials readily available. TechNote Time Watch Company has designed a wristwatch loaded with electrical and electronic information to assist them with just that. Anne Dorsey, Sales and Marketing Manager for TechNote Time said that: "We have knowledge-based products which can simplify many engineer's daily job demands, by making Ohm's Law, power equation formulas and resistor band charts available at all times.

"Our company's 'Magic Wheel' technical watches can save tradesmen valuable time in the field and time saved is money." Anne goes on to say that: "These watches are also a very handy study tool for any student preparing for professional examinations."

TechNote Time Watch Company is a privately held start-up technical watch producer and technical watch distribution company, located in Ocala, Florida. Its strategy is to serve the technical niche market of producing watches specifically for the electronics and electrical professions. www.technotetime.com.

Good prospects for college leavers

Graduate salaries are increasing, with college leavers entering electronics seeing slightly above-average levels of starting pay.

This conclusion comes from a report published by the Association of Graduate Recruiters, which also found that the number of vacancies for graduates is increasing in the UK.

The average starting salary for graduates is now £21,000, up from £20,300 last year, while in electronics starters can expect to earn £21,100.

"The days of escalating starting salaries for graduates appear to be over. Employers are providing graduates with training and development and a remuneration package that is competitive rather than extravagant, covering the cost of living increases" said Carl Gilleard, chief executive of the AGR.

Some 20 per cent of employers offer more than £35,000, while nine per cent offer under £17,000. The most generous firms are, predictably, investment banks who average £35,000.

The biggest rise in starting pay were for scientists and those working in R&D, with a rise of 8.1 per cent.

The AGR surveyed employers across the UK and found that after three years of shrinking opportunity the UK job market has started expanding once more. This year there are 15.5 per cent more positions on offer.

"The findings are good news for the graduate recruitment industry and great news for graduates themselves. Vacancy levels have risen and we expect both salaries and vacancies to continue to remain stable in 2005," said Gilleard.

IT firms had the biggest rise in vacancies, with 52 per cent more jobs available.

Geographically, in should come as no surprise that starting pay is highest in London at £25,000, dropping to £21,500 in the South East. Lowest is Scotland at £18,500 and Northern Ireland at £18,000.

The largest increase in pay was in the Midlands, which grew 8.7 per cent to £20,000.

Jobs application are also improving, said employers, while more are using online psychometric and numeracy tests to screen candidates.

This year's AGR survey was answered by 223 firms who between them filled 14,000 graduate posts this year.

Gunn diode gets new lease of life

A device to detect skin cancer has won researchers from Aberdeen University £200,000 in Government funding.

The money, from Scottish Enterprise's Proof of Concept Fund, will be used to help develop a Gunn diode capable of emitting Terahertz pulses.

Terahertz radiation reacts strongly with human tissue, and can be used to image skin and show up potentially dangerous melanoma. If caught early enough, treatment is usually successful.

"We hope to develop an inexpensive semi-conductor device which would produce Terahertz rays. A small and portable scanner would be a major advance in medical technology, with the potential to reduce skin cancer deaths," said Dr Geoff Dunn, a lecturer at the University's School of Engineering and Physical Sciences.

Existing systems are large, cumbersome and expensive, using lasers to excite gallium arsenide devices to produce the radiation between perhaps 100GHz and 10THz.

Aberdeen's device is based on a Gunn diode, a device with a highly non-linear I-V curve. In fact the devices exhibit a negative resistance through part of their forward conduction curve.

Gunn diodes have always been able to oscillate at reasonably high frequencies of perhaps 100GHz. They are made from doped semiconductor with contacts at each end.

Making them thinner is the key to higher frequency oscillation. Research groups have shown diodes made from gallium nitride that can reach 1THz.

Reaching such frequencies would be the key to making a cheap medical scanner.

"We believe such a scanner could be a vital tool for GPs checking for cancer. But this technology could also be used in a huge range of other commercial sectors which include high resolution Radar systems to wireless office communications," Dunn added.

Scottish Enterprise's Proof of Concept Fund was in 1999 as a three year £1 1m fund - but its success led to this being extended to a £33m fund over six years.

Satellite navigation saved after EU and US sign agreement

Technical and political wrangling that threatened the future of Galileo, Europe's satellite navigation system, have been resolved.

The European and US Governments have signed an agreement that will see GPS, the US system, and Galileo work in harmony.

GPS will be aimed at miltary application, while Galileo will be aimed firmly at civil and commercial use, said the EU.

"This agreement will allow the European project Galileo to become the world standard for civil and commercial use of satellite navigation; it will offer the best possible level of services to all users," said European Commission vice-president Loyola de Palacio.

Talks to make the two systems interoperable have been ongoing for four years. The agreement will boost early sales of Galileo receivers as they will also be able to take signals from the GPS satellites.

The EC has the high expectation that by 2010 the worldwide market for sat-nav will amount to three billion receivers and ¤250bn. Over 100,000 jobs could be created in Europe, claimed the EC.

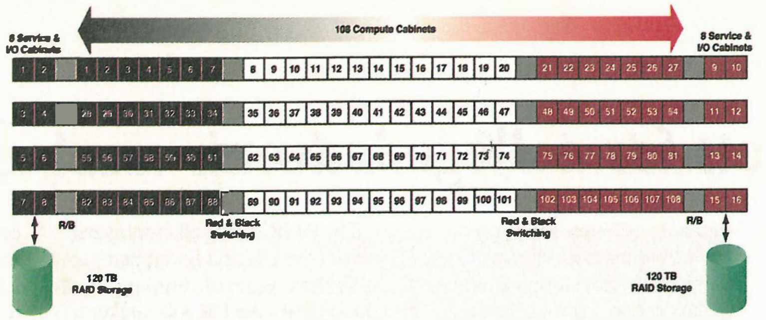
Deployment of the 28-satellite constellation, costing some ¤3.2bn, is due by the end of 2008.

Cats inspire satellite pointing technique

NASA has tested a way of twisting satellites in space which is based on the methods cats use to always land on their feet. To rotate their bodies during a fall, cats push their feet out, twist their bodies, then draw their feet in and twist themselves back to the starting position having rotated themselves. By repeating this, they rotate themselves without needing to push against anything. Satellites usually change the speed of three internal orthogonal flywheels to achieve the same effect or use gas jets - or a third option for satellites orbiting in the Earth's magnetic field: Feed current through coils to develop force. Students at Drury University in Missouri developed the cat-like action to eliminate the need for wheels, thrusters or magnetic reactors from spacecraft with mobile arms - such as future repair robots which could move outside a larger space craft.

"The robot is potentially useful in space exploration. It could enable NASA to turn objects like a satellite or an astronaut without gas jets or spinning gyroscopes," said tutor Gregory Ojakangas, who proposed the technique. The students "did it from scratch, solving many of the problems that inevitably arise in a project. I'm very proud of them."

Supercomputer could point the way for Cray



Red Storm will be a 41.5Tflop computer when it is finished.

Being assembled at Sandia Labs in New Mexico, one quarter of the machine should be up and running in January next year.

Performance testing will begin in early 2005.

By the end of 2005, the machine should be capable of 100Tflops - following an upgrade where each single-processor AMD Opteron chip is replaced with a double processor chip, each running 25 per cent faster than the original.

Developed with supercomputer pioneer Cray, the design may become the centre of the firm's future product line. "From Cray's point of view, the approach we're pioneering here is so powerful they may want their next supercomputers to follow suit," said a Sandia spokesman.

The design is scalable from a single cabinet (96 processors) to approximately 300 cabinets (30,000 processors). In addition, the system was designed to monitor and manage itself.

One of the project aims is to cut power consumption.

Japan's Earth Simulator, currently the world's fastest supercomputer at 35.68Tflop, consumes 8MW peak, said Sandia, compared to a projected 2MW for air-cooled Red Storm. Red Storm will also only take up one third the floor space needed by Earth Simulator.

The machine has four processors to a board and 24 boards to a cabinet. Each processor can have up to 8Gbyte of memory.

Four Cray SeaStar networking chips sit on a daughter board over each processor board. The SeaStars talk to each other "like a Rubik cube with lots of squares on each face," said Sandia. "Cray SeaStars are about a factor of five faster than any current competing capability."

Processor to processor bandwidth is 4.5Gbyte/s bidirectionally.

Europe must invest €6bn

Public/private partnerships need to find ¤6bn per year if Europe is to benefit from research into nano-electronics.

That message came from the European Commission in a report drawn up by CEOs from leading companies and research groups.

"Nanoelectronics is a strategic sector for Europe, with a potential for creating a significant number of highly skilled jobs and boosting growth and competitiveness in most other industrial sectors," said Enterprise and Information Society Commissioner Erkki Liikanen.

Along with electronics firms, the EU has established the European Nano-electronics Initiative Advisory Council (ENIAC), chaired by Pasquale Pistorio, chief executive of STMicroelectronics.

It aims to identify a strategic research agenda for nano-electronics in Europe and implement it.

"Europe cannot afford to miss the next generation of electronic applications that will be for our future economy what oil is for today's economy," said Research Commissioner Philippe Busquin.

Developing nano-electronics will require better co-operation across disciplines and more co-ordinated research.

"Leading the transition to nano-electronics is a challenge that requires our best researchers to work together and our public and private investors to profit from economies of scale," said Busquin.

www.cordis.lu/ist/eniac
www.cordis.lu/nanotechnology

Big magnet – jewel in the crown

Following 13 years of development, the US National High Magnetic Field Laboratory has made a 21.1Tesla superconducting magnet aimed at nuclear magnetic resonance (NMR) research.

The machine is 5m tall, weighs over 13tonnes and holds 40MJ of energy in its field. Samples up to 105mm in diameter can be tested in its 'warm' (non-cryogenic) bore and the machine will remain at full field for five years.

"This very powerful and ultra-wide bore magnet was an extremely challenging system to build, and it represents a significant engineering accomplishment," said director Greg Boebinger. "It is the crown jewel of the laboratory's NMR spectroscopy and imaging program."

The magnet is a concentric assembly of ten superconducting coils connected in series and operated at 1.7K.

Each coil is wound with a monolithic superconductor, composed of either niobium-tin (Nb^3Sn) or niobium-titanium (NbTi) filaments in a copper matrix. To support the magnetic loading, the coils are held with stainless steel over-banding and are vacuum impregnated with cryogenically-tough epoxy. The magnetic field produced is uniform to one part in one billion, claims the lab, in a volume 64 times larger than standard 52mm bore research-grade NMR systems. Fine-tuning of field homogeneity is achieved with a set of superconducting shim coils.

Operating frequency is 900MHz and both NMR spectroscopy and magnetic resonance imaging (MRI) are possible. "Science performed on this resource will range from materials research to macromolecular biological structure determination and non-invasive magnetic resonance imaging of laboratory animals," said the lab.

PC scopes get speedy

Pico Technology, the UK firm specialising in PC-based oscilloscopes, has upped the performance of its scopes to reach 200MHz analogue bandwidth. Moreover, using a repetitive sampling technique the devices can achieve a sampling rate of 10Gsample/s.

The PicoScope 3000 series is designed to replace a benchtop oscilloscope with a small unit attached to a PC. The scopes attach via the universal serial bus, and draw all their power from the host computer.

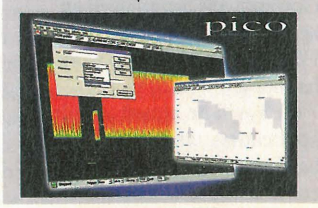
Up to 1Mbyte of memory is available in the scopes to store large sample sizes.

"The combined high sampling rate and large memory buffer of the PicoScope 3206 enable the capture and display of fast-occurring and complex signals," said Alan Tong, technical director at Pico.

The scopes also feature a spectrum analyser mode and two models in the series have built in signal generators.

There are over 30 automated measurements such as frequency, pulse width, rise time, THD and SNR. Display modes include colour traces and analogue intensity (persistance).

Prices for the scopes range from £399 to £799 plus VAT.



Class-A imagineering: Part 5

'High Fidelity' specifications based upon steady sinewave input-output testing procedures with a resistor load have not guaranteed genuinely realistic sounding reproduction, however the simple procedures outlined below should assist in our quest for being able predict at outset which designs will, or will not, drive loudspeakers more accurately. Graham Maynard continues his investigations

After check reading the first four parts of this text it occurred to me that whilst I have claimed my GM 25W-8Ω class-A amplifier will voltage drive real world loudspeakers less erroneously than many other designs arriving with supposedly superior paper specifications, I had still not actually offered any form of demonstrable or test bench measurable proof which might be capable of convincing anyone that I have a basis for my first cycle distortion claims, my reasoning, and my own subjectively based opinions. I also saw that my comments and expressionism relating to the Miller connected VAS C.dom could be deemed as being 'way over the top' by formally trained amplifier designers, yet in so many published articles, circuits, letters and web-pages this is an aspect where there appears to be an on-going lack of understanding, therefore I will further illustrate the waveform distortion this component can introduce; distortion that can become worse when complex loudspeakers are loudly class-B driven.

Back in June 1994 I wrote an open letter to *EW & WW*, p509, suggesting that the performance of an audio amplifier should not only be examined whilst it is driving real loudspeakers, but that it should additionally be bench tested by injecting a signal into its other 'input', i.e. via the loudspeaker terminal and thence the NFB loop controlled output node! My letter was answered in August 1994 *Letters* by Douglas Self, stating that "back EMFs from reactive loads do not cause detectable intermodulation and even if they did, a higher feedback factor would surely reduce rather than increase the effect". Through the years Mr. Self has written so many excellent reference articles, but unfortunately he appeared to not check out my suggestion whilst additionally stating his dislike for my written presentation, which, although being based upon real-world aural observation and intuitive cerebral analysis, was not necessarily wrong just because my brief letter did not offer practical suggestions for any examination methods that could be visualised in a lecture theatre presentation. I was left feeling completely gobsmacked, and wondered why I had bothered to write at all. There really was no point in me doing so again because my suggestion for testing whether reverse impinging loudspeaker generated back EMFs do actually modulate amplifier output - as I, many stage performers and hi-fi aficionados had long experienced in real life - had just received a somewhat scornful, if eminent, burial. I felt that a possible furtherance of widespread understanding via these shared pages, with their additional potential for follow-up and thus overall improvements in sound reproduction for everyone, had been individually undermined by one of our most influential writers.

Ten years have passed, with many recently published designs and present day discussion comments suggesting that precautions for minimising loudspeaker back EMF generated amplifier distortion products are still not being routinely implemented. Also, circuit simulation software can now be used by budding enthusiasts as a first line of design, with programmers suggesting the use of multi-cycle simulations to calculate THD figures in a manner that masks capabilities for investigating other already experienced sound reproduction problems. Multi-cycle simulation cannot fail to assist in establishing the lowest THD figures because this methodology knowingly uses a steady sinewave to optimise waveform averaging about the zero axis (much as does fundamental nulling at the output terminal), and this totally ignores all of the dynamically induced level changes that arise due to the start-up of an initially asymmetric sound waveform which has to initially charge and drive all components and devices in a manner that cannot fail to initiate a momentary shift between alternating voltage and current zeros as the circuit simultaneously amplifies and copes with reactively induced dynamic loudspeaker back EMFs under the controlling influence of propagation delayed (lagging) NFB. If it is possible to hear differences when headphone monitoring amplifier output when switching between resistor and loudspeaker loading, then we need a computer program capable of evaluating loudspeaker induced first cycle changes.

Granted, computers will not burn our fingers and deplete our pocket books if one experimental amplifier construction after another eventually were to end up in the round filing cabinet, and they can save our environment from some of those occasional releases of acrid black smoke which manufacturers like Motorola and Toshiba somehow manage to compress and encapsulate within their devices, but if anyone has yet to choose software, it is better to be aware that some products which rely upon ten or more cycle averaging because they do not fully set up operating conditions before simulating, sometimes cannot be made to complete the Fourier series calculation necessary to produce a distortion figure for any specified single sinewave; and especially for the waveform error that arises within the first cycle of any series, where distortion can be much worse when a virtual crossover inclusive loudspeaker system is the circuit load instead of a passive resistor.

There are few mainstream writers who report upon the behaviour of amplifiers being reverse driven by a separate audio source that mimics loudspeaker back EMFs: see;- Rod Elliott's *Sound Impairment Monitor*, at;- http://sound.westhost.com/sim.htm

Mr. Elliott's work usefully images potential class-B drive inadequacies by illustrating the potential for load induced output stage crossover distortion, but at this time of writing he does not directly mention the amplifier signal path control delay, thus the circuit impedance and first cycle issues I outline here.

Sinewave examination

For decades testers have been determining an amplifier's 'damping factor' via examinations that are equivalent to reverse driving the output stage with a steady sinewave via a nominal load resistor and then dividing the alternating voltage of the driving source by that measured at the output terminal. Some amplifiers have their damping factor specified at different frequencies, though without the reasons for any variation with frequency and its subsequent significance for higher level plus higher frequency dynamic loudspeaker reproduction being explained. I eventually realised that I could use the virtual signal generator of a computer simulation program to reverse inject a test signal into the output terminal of any potential audio amplifier circuit, and thus image not only its steady sinewave responses, but also examine the errors that become introduced as a result of propagation delay within its closed NFB loop allowing the output terminal and interstage responses to become dynamically modified by reverse impinging and high frequency 'reactive loudspeaker back EMF' as the lagging NFB generated output response constantly attempts, but never quite manages, to recover control. Also, computer simulation would allow these observations to be made without any need to 'break' a closed NFB loop, which could otherwise lead to unnecessarily discursive theoretical arguments as to where or how this may or may not be done because of the circuit operation changes that are already so well known to influence findings, and thus provide an opening for objective arguments that could intensify whilst leaving in the shade the originally imperfect fundamental circuit action which is what should really be being considered due to the prior subjective observation of degraded reproduction.

I could simulate that 1994 examination suggestion on any amplifier circuit and study the outline results without needing expensive test equipment; thus I could make additional precautionary checks upon the performance of my own already constructed designs, prior to submitting them for publication.

Figure 11 shows the simulated amplitude and phase plots for swept frequency generator reverse injection of my Part 4-Figure 10, class-A, 25Watt circuit via a nominal 8Ω load resistor in series with its output terminal.

My non-complementary pure class-A amplifier circuit holds the output terminal potential flat at -64dB (<0.1%) for all audio frequencies, with a slight but inconsequential phase lead of 8 degrees by 10kHz. This leading phase is due to the NFB loop's natural inability to make the amplifier respond any more quickly due to internal driver and output stage transistor base related capacitances etc., combined with its entirely natural bias current limitations and the 22pF high frequency NFB response correction capacitor in parallel with the NFB loop sensing resistor. This is a good and most satisfying result for such a simple topology; indeed, maybe this fine performance is due entirely to its circuitry being so basic!

During the early days of 'Hi-Fi' testing, 1kHz was often a reference frequency for measuring specifications, yet examination here will not necessarily reveal all the problems capable of affecting sound reproduction, especially with lowly biased class-B solid state circuitry. I have no interest in 'going easy' on any amplifier in order to present impressive specification figures, and this is why I am much more concerned about how an audio signal amplifying circuit behaves at 10kHz. The Figure 11 swept sinewave simulation for my '25-8' amplifier indicates that its circuitry generates an internal resistance of 5mΩ (-64db) in

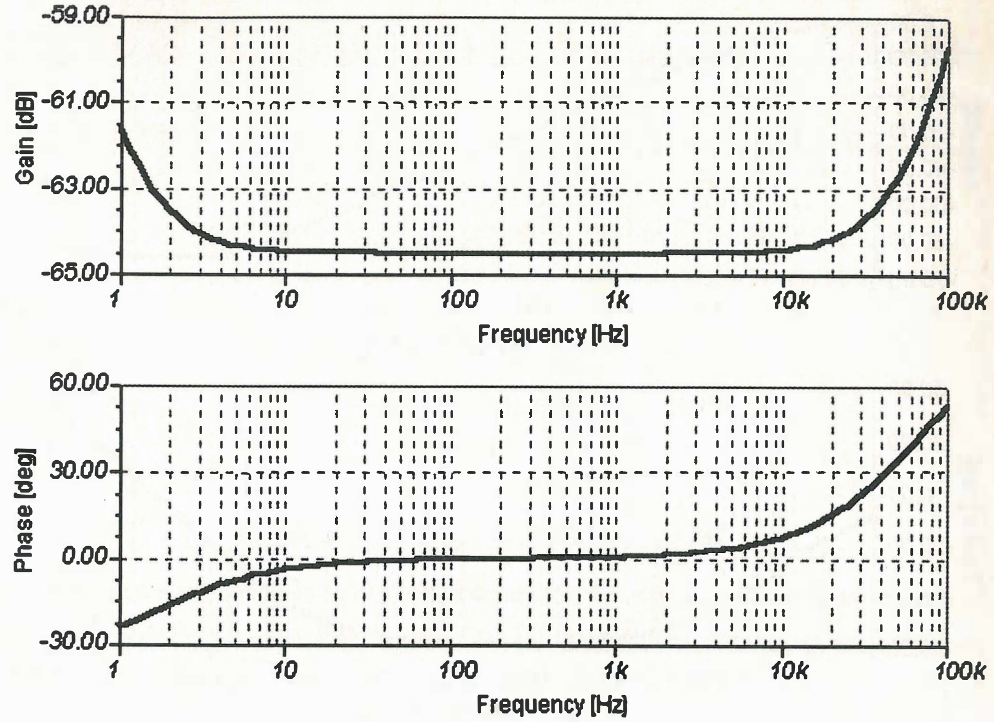


***Figure 11:** The simulated reverse driven amplitude and phase responses for my 25Watt class-A amplifier.*

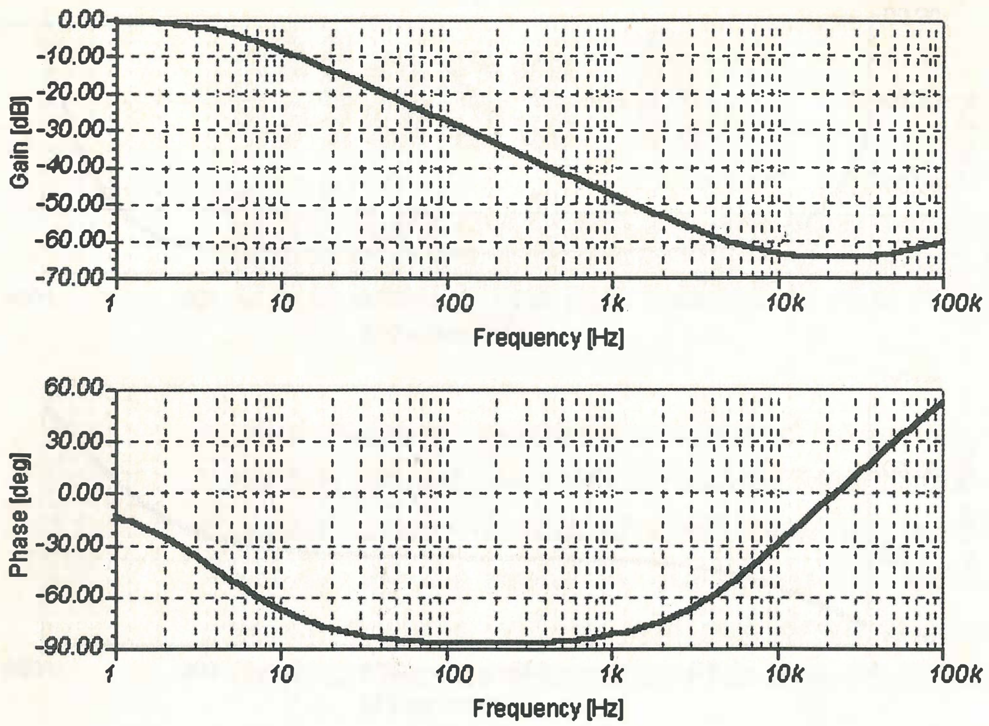
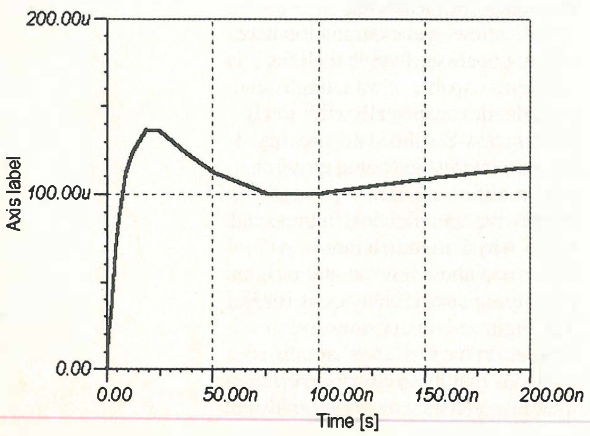


Figure 12: ***The reverse injection characteristics of my Part 4-Figure 10 circuit when its output driving capability is modified by the reactivity of an integral 4m7F series output capacitor.***

Figure 13: ***The 'GM.25-8' class-A amplifier's simulated reverse driven output terminal error at 10kHz due to propagation delay within its closed FB loop.***



series with 10nH (+8 degrees at 10kHz) and 7 Farads (-8 degrees at 4Hz). Note that the 10nH and 7F equivalent values are effectively in series with the internal 5mΩ NFB loop generated output resistance, and are not separately in series with the reverse driving resistor which is externally connected to the NFB output node/amplifier terminal. When we consider loudspeaker driving capabilities this is getting close to the proverbial 'straight wire with gain'; however, the complex characteristics of every component, their leads, and all other interconnects which must physically be used for any real-world fabrication will interpose impedances that cannot fail to alter node current flows and voltage developments, and thus unavoidably degrade the circuit's fundamental control of loudspeaker reproduction. At least there are no additional output stage conduction crossover discontinuities to consider, so this basic circuit is not within itself a significantly limiting factor.

Of course we will be very careful with our interpretation of amplifier simulations, because individual component characteristics (not just their intended values), the physical layout of any topology with regard to the signal-voltage, ground-rail also power-rail impedances, plus circuit isolation and any possible voltage development due to internal and output load generated current flow, can significantly and thus audibly modify the theoretical performance. Also the Figure 11 simulation above is an illustration of circuit response for a swept 'steady sinewave' reverse drive investigation only, as is **Figure 12**, which illustrates the additional phase change that would become apparent at the output terminal of my amplifier due to a back EMF generating 'load' reacting with a series 4.7mF output capacitor inserted between the measuring point (loudspeaker terminal) and the low impedance NFB loop controlled output node of my amplifier. Due to them having a voltage lagging charging characteristic when under reverse sinewave current drive, integral series output capacitors do not suddenly generate additional leading edge (tweeter driving) output potentials, but they do allow loudspeaker generated back EMFs to reactively voltage modulate an amplifier's output w.r.t. the low impedance and NFB loop controlled output node at frequencies above the nominal -3dB roll-off point. This reactivity, and thus the coincidental music driven dynamic shift in the alternating loudspeaker voltage zero level w.r.t. the expected alternating signal waveform voltage, especially after a first half cycle of low frequency drive to a resonant bass loudspeaker, is every bit as audible as was illustrated in my Part 3, Figure 9 simulation.

Dynamic capabilities

There really is nothing wrong with conducting steady sinewave investigations, yet my prime interest has long been studying the way in which individual components and bias levels affect output-driving capabilities when the loudspeaker itself is reactive and is being loudly driven. My concerns include the way in which reverse impinging loudspeaker generated back EMFs affect an amplifier's driving capabilities as it attempts to simultaneously forward amplify complex and constantly changing large amplitude music waveforms. Thus I include **Figure 13**, which simulates the dynamic response of my '25W-8Ω' circuit whilst being reverse driven by a 10kHz sinewave from t=0.

This dynamic response examination reveals the output terminal error potential development due to unavoidably delayed closed NFB loop control behaviour in the presence of a reverse (loudspeaker generated) impingement of 1V RMS (125mA RMS) 10kHz sinewave via its series 'load' resistance of 8Ω, the latter being used in this investigation to preserve nominal closed loop and amplifier output stage operating conditions. The initial, and near instantaneous leading edge, 'load' driven output terminal potential shift is of low level, smoothly controlled and fully settled within 100ns, prior to it subsequently increasing in amplitude to equal that of the steady sinewave -64dB response with its 8 degree shift. The rising edge simulation in Figure 13 with slight overshoot at 20ns illustrates how circuit propagation delay within the NFB loop has denied development of the truly instantaneous dynamic control necessary to counter output node error, whilst the steady sinewave response of Figure 11 merely illustrates the continuously generated damping activity of the amplifier's internal NFB loop reduced output impedance.

Here the significant dynamic response figures are 138μV (-80dB) of initial error at 20ns, which is then followed by the 600μV.RMS (-64dB) of near phase coherent steady sinewave (Figure 11) error. If both the initial/dynamic and full first

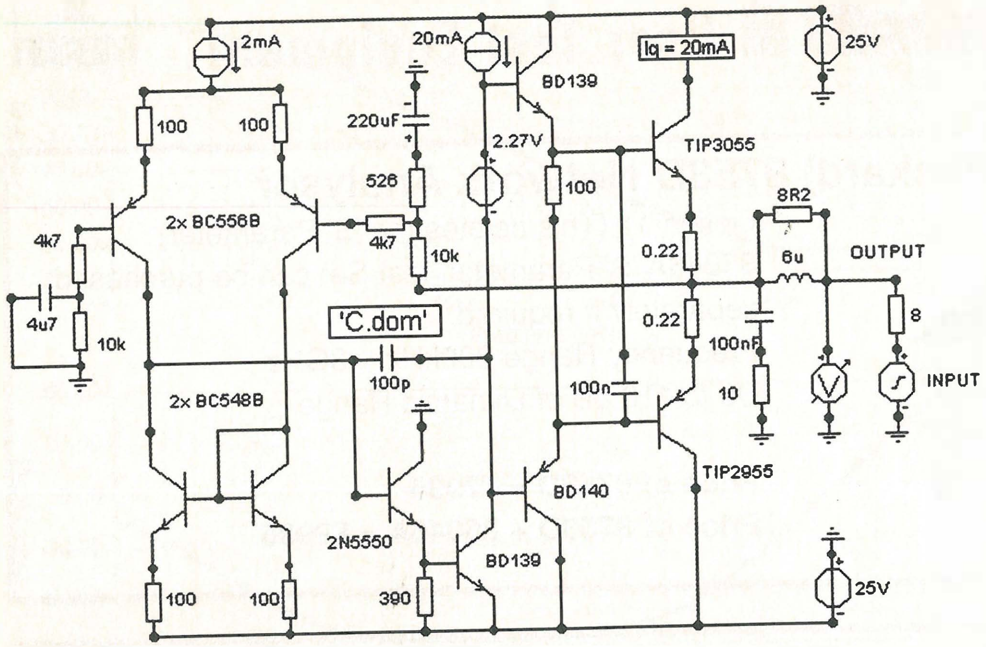


Figure 14: The reverse testing circuit for a generic 20th century 25W-8Ω bipolar class-B amplifier with Miller C.dom and integral series output choke.

cycle/steady sinewave responses had been shown within a single timeframe, then there would be difficulty in resolving the high natural speed of initial class-A amplifier error correction. From a circuit viewpoint there is nothing to worry about here, no matter how much time shift or frequency change might be observable within the back EMF impingement of a dynamically reacting loudspeaker w.r.t prior NFB loop controlled amplifier drive. The competent analogue limitation of this circuit's load driven output terminal potential development has been achieved within the same time period a typical digital amplifier takes to switch once between fully 'on' and fully 'off' conduction states!

At audio frequencies my computer simulated class-A circuit is behaving less inductively than the interconnecting loudspeaker cabling that would become necessary for real world usage. The illustrated 138μV error potential that develops as the output terminal follows back EMF during the amplifier's 20nS propagation delay time period prior to the establishment of NFB loop error sensing and control can be approximately determined from Figure 11, but not necessarily from any 'amplitude only' based damping figure nor a simplistic spot frequency output 'resistance' notation. By multiplying the reverse driving peak sinewave voltage (1.414V) by the -64dB voltage ratio (0.000631) and the 8 degrees Sine ratio (0.1392), all at 10kHz, we get 0.000124V or 124mV. The difference between the former simulated load induced leading error and this the theoretically calculated figure is not due to any software anomaly, but to the reverse driven dynamic overshoot / damping / stability characteristics of the circuit, and these cannot always be determined from steady sinewave established measurements, like those of Figure 11. This error calculation does show however, that the simulation and observation of an unfiltered toneburst from t=0 is as valid as the already acceptable and widely used steady sinewave testing upon which Figure 11 itself is founded.

When a reverse driven amplifier's output terminal (NFB sensing) potential is leading w.r.t. signal input, both the reverse energised dynamic response and stability characteristics are different to those observed via conventional forward testing when the load is purely resistive.

These reverse characteristics are additionally affected by an amplifier's input + VAS + driver + output stage bias current levels, its NFB loop stability w.r.t. input, the value of the series NFB loop sensing capacitor, impedance at the input transistor base w.r.t. ground, and the values of the output Zobel components. What is best for optimising steady sinewave measured forward THD on any particular design is not necessarily best for controlling and thus damping dynamic loudspeaker generated back EMF impulse induction because, and unlike shunt feedback with its single input node, separate 'signal' and 'NFB loop sense' differencing inputs do not always initiate identical output responses. Thus if we were to examine an amplifier by bypassing its input filter and output choke we would alter its fundamental loudspeaker controlling capabilities, and we would then observe falsely enhanced results that can not-will not be achieved in real life.

It is an unavoidable fact that the propagation delay for any amplifier is determined by design specific internal series-parallel connected signal path components plus device and input-output termination impedances and these remain relatively constant prior to the onset of overload induced non-linearities. Thus the 138μV error development at 10kHz arising within the 20ns propagation delay introduced by my circuit becomes 276μV at 20kHz, though with careful wiring and construction, the maximum output terminal error at audio frequencies should still not exceed 0.1% of the equivalent loudspeaker induced reverse potential, even through normal loudspeaker impedance dips.

Unfortunately I would not expect to be able to repeat so quick a response measurement on the test bench, even if I could be sure that any reverse driving source amplifier had not within itself already introduced a first cycle delay such that the sinewave induced error could properly be observed from t=0. Nor would I be able to predict the level of music waveform generated back EMF any loudspeaker is likely to generate, but the simulated Figure 13 reverse dynamic response does at least show that delay within my simple class-A circuit with its high value input capacitor driven by a low impedance source, plus capacitor compensated NFB sensing resistor, does not lead to any significant loudspeaker driven high frequency amplifier products being generated w.r.t. unfiltered and therefore original, first cycle source waveform.

Generic class-B

In view of these findings, and already knowing by hearing that an output choke in series with a low impedance NFB loop controlled output node does cause audible loudspeaker reproduction changes at frequencies below its -3dB hf roll-off point, I wondered how a typical 20th Century class-B bipolar circuit might respond under the same low power reverse sine wave injection testing conditions; hence my examination of the generic **Figure 14** 25W-8Ω circuit.

The simulated amplitude and phase plots for a reverse driven Figure 14 circuit are shown in **Figure 15**. At first sight the generic class-B amplifier appears to have a decent level of steady sinewave damping over most of the audio spectrum, if not so good at higher audio frequencies. However, let us examine the simulated -46dB amplitude response figure at 1kHz. Some designers might separately measure volts and amps at the output terminal to establish and claim that it offers 46dB of loudspeaker damping, or a damping factor of 200, or even specify a 40mΩ output resistance figure, which is a similar value of resistance that would be introduced by a series output fuse. But it is not as simple as this, because the theoretical 40μΩ (yes, micro-Ohms) of simulated output resistance that is internally generated by the NFB loop at low frequencies has been modified by the effect of its integral 6μH output choke, which is outside of and in series with the NFB loop controlled amplifier output.

As I have earlier reasoned, the choke coupled amplifier's output terminal potential development due to reverse impinging loudspeaker back EMF current flow actually develops in leading quadrature w.r.t. the load generated back EMF at audio frequencies, as is now confirmed by the phase plot. This means that the output terminal suddenly develops a -46dB induction potential every time that a loudspeaker develops a 1kHz back EMF w.r.t. the amplifier's choke isolated and NFB loop generated 40μΩ plus series 260nH internal impedance output node. When a series output choke is fitted, the resultant output terminal error potential is due mainly to output choke plus load combination delay and the subsequent leading reaction, not to the closed NFB loop controlled amplifier itself!

Following on from the satisfactory real-life performance of my own class-A amplifier, as additionally confirmed by the reverse dynamic simulation of Figure 13, I wondered how this choke coupled class-B amplifier circuit would simulate whilst its output terminal is being reverse driven by a (dynamically generated reactive loudspeaker system generated back EMF) sinewave via its 8Ω load resistor. Do please note that my reverse drive for this examination is a mere 1kHz-1V.RMS, as illustrated in **Figure 16**.

The Figure 14, 25W-8Ω generic circuit has the same level of NFB as my own class-A design at 1kHz, but its leading series output choke error potential development has allowed the amplifier's output terminal to be driven to 7mV within 7μs. This is the minus 46dB steady sinewave quadrature voltage, and it really is substantial for an amplifier whose output terminal ought to be held very close to zero with a near instantaneously generated equal but opposite 125mA RMS of current flow in order to establish NFB loop controlled back EMF (loudspeaker system) damping.

Oh dear - this is not very inspiring! The start-up of this reverse injection sinewave simulation has alone already illustrated the occurrence of leading choke induced voltage waveform distortion. The choke has interfered with the NFB loop's voltage sensing capability for initiating current correction at the amplifier's real world output terminal, and this ties in with my Part 3-Figure 8 simulation. No *audio* amplifier should be this bad!

The dynamic immediacy of this quadrature induced error potential as it follows a reverse impinging (back EMF) waveform for 7μs prior to a phase shifted sinusoidal voltage development cannot be inaudible, and, because the voltage would rise to 70mV within that same 7μs period at 10kHz, even without any loudspeaker impedance dip, the potential for unpredictable loudspeaker generated intermodulation distortion will be far from inconsequential when compared to the typical 0.01% to 0.001% - non first cycle - THD specifications normally claimed for forward amplification via this type of amplifier when continuous sinewave developments across a passive resistor load are all that are being examined. The choke generated (error) potential suddenly appearing at the Figure 14 amplifier's output terminal is 400 times worse than from my simple 25W class-A circuit. Also, because the choke settlement induced output terminal potentials develop inductively, this becomes like having one after another of suddenly loudspeaker system energised 7μs error voltage generators repeatedly being inserted, with their potentials integrating to generate on-going high frequency induction error between the NFB loop controlled amplifier output node and the sound transducers we actually listen to. This possibility was previously illustrated in Part 2-Figure 6, and shown as differently induced crossover circuit and driver delays affecting loudspeaker voltage in Part 3-Figure 8.

'Choke' - contraindicated

As a direct result of this simulated generic 20th century bipolar class-B

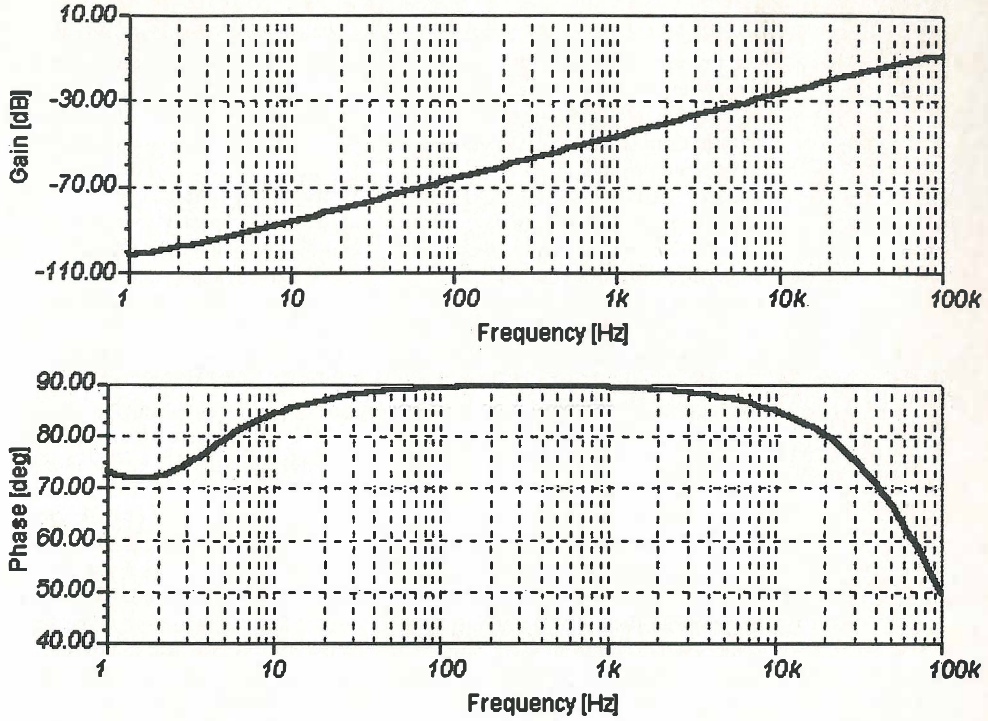


Figure 15: The reverse driven amplitude and phase simulation characteristics for the generic 25W-8Ω class-B amplifier.

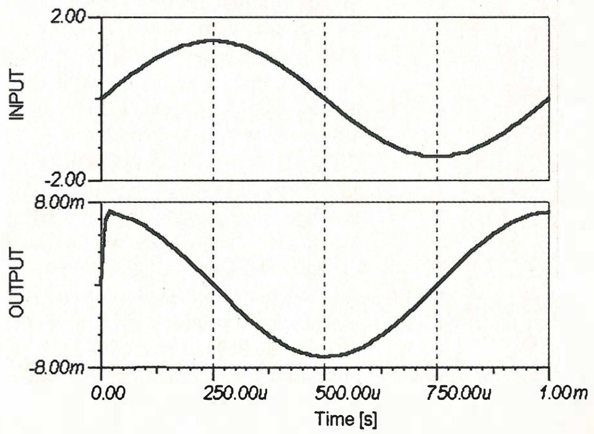


Figure 16: The simulated class-B amplifier's output terminal reverse injection error at 1kHz due to its internal propagation delay plus series output choke impedance.

amplifier's testing 'load' having been rendered representatively 'reactive', the series output choke can be seen allowing the amplifier's output terminal waveform potential to become parasitically modulated in a manner that no amount of NFB can control. This is not a type of distortion that could be measured w.r.t. input in a conventional sinewave driven test situation, because unless the 'back EMF' can be repeatedly pulsed on and off, as it effectively is by musically induced dynamic loudspeaker plus crossover delay-storage-release energy reactions, then the sudden and independently generated step-like potentials and on-going error would occur long before a sinusoidal distortion measuring test set is capable of stabilising. Also, as these errors recur they would not only 'characterise' and smear the sound but shift about average the time and voltage reference points that are essential for any kind of valid steady sinewave distortion measurement to be undertaken and observed in isolation at a Figure 14 like amplifier's output terminal in the first place!

The sudden 7mV of transient error for just 1V RMS (125mA RMS) of 1kHz sinusoidal back EMF, separately illustrates a route whereby everyday bass-mid loudspeaker driver and crossover back EMFs could, at ordinary audio frequencies and normal listening levels, cause this generic class-B amplifier's output terminal step like quadrature waveforms to integrate, and thus allow the choke potential to modulate tweeter drive w.r.t. the amplifier's NFB loop controlled output. With mid-frequency cone flexing and an impedance dip to 4Ω at say 3.5kHz, this highly energetic tweeter driving 'amplifier' induction potential could rise to 50mV per equivalent 1V.RMS of back EMF, so what value does a steady sinewave THD figure better than 0.1% really have then? When my approximated virtual Ariel loudspeaker is the driven load for testing a Figure 14 circuit the leading error becomes 57mV/V at 6kHz due to combined loudspeaker driver plus crossover section reactance.

The last three decades of audio development have given us phenomenally low steady sinewave established THD figures without some designers realising that their extra zeros were not actually improving reproduction because their test-bench resistor established measurements ignore effects that are directly related to signal path impedances, propagation delay and the dynamic loudspeaker induction errors that do arise at real-world amplifier-loudspeaker interfaces! Suddenly generated series output choke error potentials which could exceed 10mV/V/kHz through entirely natural dynamic loudspeaker impedance dips and phase change reactions are not going to be insignificant. The greater the number of parallel connected series L plus C crossover sections, the more electro-magnetically efficient the loudspeaker drivers and the more powerfully they are run into voice coil inductance changes, cone, cabinet and air-spring induced back-EMF generation, especially during live stage and studio situations, then the more frequently ongoing and greater these back EMF induced and crossover section 'tuned' errors will become, with the likes of voice, high energy pop music and high power sound reinforcement reproduction being rendered subliminally distracting. Do note however that reactively induced loudspeaker system back EMFs arise at frequencies that are specific to the amplifier-loudspeaker combination itself and thus are different to the waveforms which energised them, the difference arising due to sudden amplifier output changes carrying entirely natural, and asymmetrical wide spectrum amplitude variation, not just a pure single frequency sinewave.

Here is an additional explanation for the cleanliness of bi- and tri-amped audio because even with choke coupled amplifiers mid-bass induced potentials cannot drive a tweeter etc., and this is why correctly designed valve amplifiers with high quality output transformers or a Miller C.dom-less and chokeless class-A can be so satisfying and non-fatiguing due to both types;- not having a basically high pre NFB output impedance, such that correction need only be adjusted between halves rather than be constantly developed and switched between halves via delayed NFB loop control; having a minimal C-bc Miller capacitance cut of their high frequency open loop bandwidth at times when the NFB loop is unable to function linearly; and, normally deriving their NFB loop control potentials directly from the amplifier's loudspeaker connection terminal.

Here too though, is further explanation for the clarity of non-frequency-crossovered multi-small-driver PA loudspeakers and line sources when being driven by choke coupled solid state class-B amplifiers, for these often have less reactively generated mid-band back EMF with which to induce additional high frequency waveform changes, also their fundamental bass resonance occurs at a relatively low frequency such that there is very little choke voltage generation. When large and efficient full range or mid-bass drivers plus tweeters are energised by a choke coupled amplifier however, then unavoidable mid band cone resonance and break-up modes, which are already bad enough in their own right, could lead to the generation of additional fractional tweeter driving parasitics from mid frequency back EMF voltage development across the series inductance. Simple 6dB/octave high and low pass loudspeaker crossovers, or crossover circuitry that betters 12dB/octave but which is lower 'Q' and well damped (less reactive) could induce less (L+C)/output choke 'colouration' error, but then loudspeaker design really ought not be subject to the limitations of inadequate amplifier performance anyway.

Clearly a series output choke is capable of introducing much more dynamic waveform distortion, and thus audible reproduction change, than the base amplifier circuitry itself, such that any low output impedance class-A amplifier which includes one is likely to sound very similar to its class-B counterpart when auditioned through the same loudspeaker systems. The comparative thermal inefficiency of class-A for a similar performance might then be deemed seriously disadvantageous, however, we must not go on to conclude that their sameness of sound justifies a claim that a particular class-B design is as good as class-A designs in general, because unless any series inductance is less than 1uH, its inclusion will have already turned the class-A design into a flawed reference.

As well as class-A, and AB circuits, Mr. John Linsley Hood developed a popular 75W-8Ω bipolar class-B amplifier; however he did not use a resistor damped series output choke! Had he too heard a series output choke degrading the reproduction from his wide-open bandwidth class-A amplifiers in spite of THD measurements suggesting otherwise? As an empirically and theoretically aware designer, also as a widely experienced listener of all amplifier types, Mr. Hood chose to use a 220 milli-Ohm series output resistor which, by its useful degree of load circuit isolation, allowed a sufficient level of unshunted and

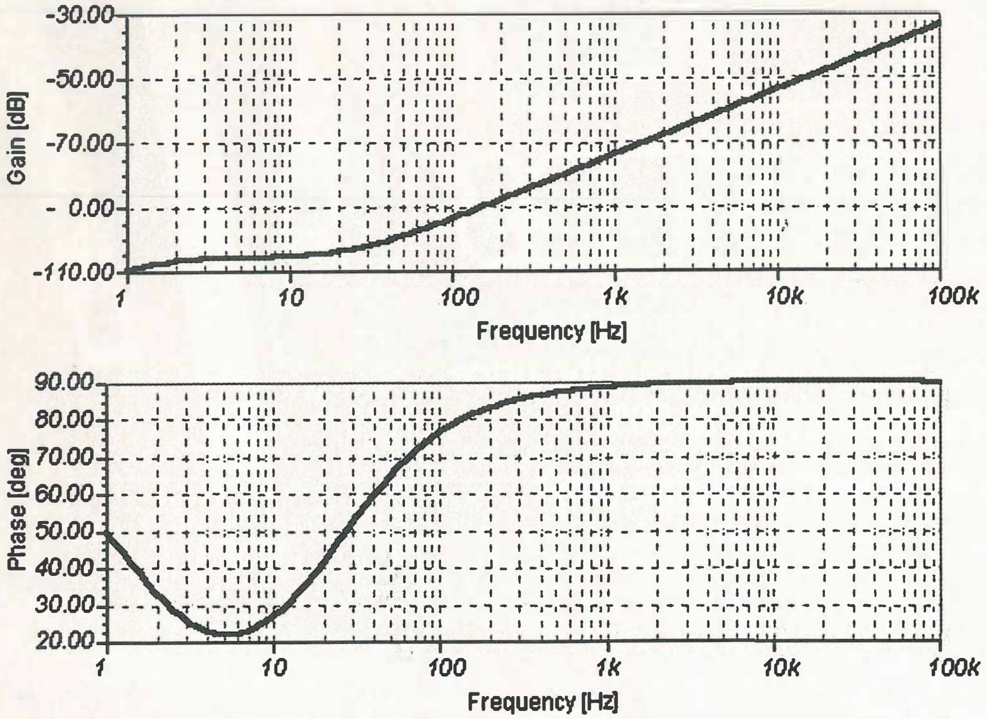
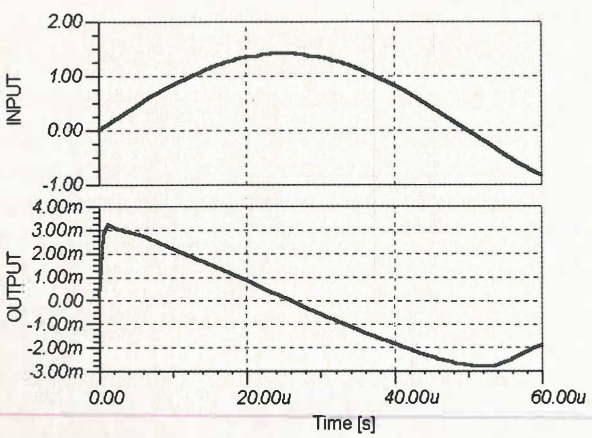


***Figure 17:* The reverse amplitude and phase driven behaviour of the 100pF Miller C.dom stabilised class-B amplifier without its series output choke.**

***Figure 18:* The back EMF generated chokeless class-B output node error due to C.dom induced propagation delay at 10kHz.**



stabilising class-B NFB loop correction development at the same time as providing the lower 'Q' and more valve amplifier like level of loudspeaker damping already achieved by his class-A circuit. The slight loss of output power within that series resistor is an excellent compromise, because Mr Hood's bipolar class-B design cannot generate the series output choke induced type of voltage leading error waveforms w.r.t. an ultra-low impedance NFB node, as has been simulated in this article. The series output resistor (or a series output fuse) portion of output terminal voltage that develops w.r.t. the NFB loop controlled output node, linearly follows any dynamically generated back EMF waveform from zero without any sudden leading choke-loudspeaker induction potential developing between the true NFB loop controlled amplifier output node and the terminals we connect our loudspeakers to. Of course the output of any amplifier which similarly uses a low value series output resistor must itself also be minimally inductive, so that the NFB loop output node potential cannot be leading voltage 'pulled' by reactively modified loudspeaker back EMF w.r.t. amplified signal waveform; also with regard to fuses, I believe that any loudspeaker protection should be at the loudspeaker, in series with, and be appropriate for each individual driver voice coil.

Once again I question myself about the simulation validity of using a 'load' resistor during this reverse injection examination. I conclude that it is acceptable because I am inducing output terminal current flow through the NFB loop controlled output node in much the same way as will any loudspeaker generated back EMF, whether that loudspeaker has a complex impedance or not; a complex impedance that can only further exacerbate any series output terminal inductance generated error affectations due to the composite loudspeaker phase change generating components having the capability for shifting back EMF by more than ninety degrees around crossover and driver resonance frequencies.

Electrostatic loudspeakers do not store and return energy in the same way as do voice-coil energised drivers, thus this reverse testing procedure is not essential for any amplifier that is intended solely for use with series resistor fed electrostatic loudspeakers. Indeed, an old 'pipe and slippers' Hi-Fi amplifier introducing notable propagation delay and having both series capacitive and inductive output characteristics which will lead to it measuring poorly for an equivalent FCD figure and cause it to generate substantial reverse driven amplitude-phase error can still sound good when driving series resistor-transformer fed electrostatic panels, yet it will perform poorly if pressed into service power driving modern-day composite loudspeaker systems, no matter how famous the badge on its front panel. It is not that the magnetted loudspeakers are within themselves especially inferior, just that they are different and have different driving requirements, thus electrostatically energised full range panels cannot be used to prove an audio amplifier's dynamic loudspeaker driving capabilities.

Not just imagineering!

In view of these findings I then wondered how the generic class-B circuit would behave without its series output choke, as has - at long last - become the norm with 21st century design, even though many 'older' circuits remain accessible via the Internet and paper/ROM based publications. This simulation is completed via reverse sinewave driving directly into the NFB loop controlled output node, and Figure 17 traces the Figure 14 amplifier's behaviour after its 6mH choke has been shorted out.

Figure 17 illustrates how removing the series output choke has immediately reduced the high frequency class-B continuous sinewave error voltage amplitude by 28dB, though for the purposes of this illustration stability requirements have been ignored. The chokeless Figure 14 circuit now appears to be more highly reverse damped than when compared to my own '25-8' circuit, however, and once again, the NFB loop controlled output terminal response is still in fully leading quadrature with the reverse injected 'back EMF' for all frequencies above 200Hz, this relating to the 260nH of internal closed NFB loop generated inductance due to the lead charging activity of the 100pF Miller connected VAS capacitor acting directly upon the output node of the differentially sensing input pairing. So, what would this generic circuit's dynamic response be like when it is reverse activated by an externally impinging 'loudspeaker induced' back EMF? To find out I repeated the 10kHz-1V RMS sinewave simulation via load resistance as originally conducted on my own 25W class-A circuit; the resultant output terminal voltage trace being shown in **Figure 18**. After 500ns of Miller connected VAS C.dom induced amplifier propagation delay the 10kHz-1V RMS 8Ω load injected sine wave gives rise to a near instantaneous generation of notably triangulated and 90 degree leading 6mV.p-p output waveform which might still be designer specified as being minus 54dB, or providing a damping factor of 500.

Yet this is somewhat disconcerting too, for it shows that the internal amplifier circuitry itself is not only behaving like the inductor that theory suggests and naturally generating a leading output terminal error potential that increases linearly with frequency, but also that the resulting triangulation shows internal amplifier currents are being badly enough distorted by output stage drive current reversals to indicate that there is indeed a VAS C.dom induced slew rate non-linearity modulation, which was not previously observable on the reverse tested full Figure 14 circuit because its presence was being masked by the voltage lead already introduced by the series output choke! This class-B semiconductor arrangement could have enough NFB to achieve a reasonable conventionally measured steady sinewave THD distortion figure, yet under 10kHz back EMF testing conditions its dynamic (loudspeaker) load induced output error generation has here become exaggerated, even though still not as extreme as it could become with efficient and crossover network driven loudspeakers when high power audio driven back EMFs incite additional output device storage effects, and the Miller connected C.dom leading current flows then cause additional differential input stage non-linearity where speedy and linear feedback ought to be able to counter output stage conduction-crossover distortion.

See the simulation in **Figure 19** illustrating the reverse induction effect when the 8Ω reverse drive is increased to 4V RMS (500mA RMS). Levels greater than about 2V RMS of reverse energisation lead to the Figure 14 output stage conduction crossover spikes become increasingly voltage elongated and extended in time because, as the voltage sensing NFB loop attempts to correct output current errors at supersonic speed, leading C.dom current shunting at the input stage is subtractive and this interferes with the NFB loop controlled correction of load voltage because a greater differentially sensed output terminal error voltage becomes necessary at higher frequencies to initiate NFB control. This is a problem that can lead to premature input stage clipping when differential input transistors are lowly biased and when series emitter resistors are used at input and mirror pairings with an emitter degenerated or a non-Darlington VAS stage. The Miller connected VAS C.dom can cause differential input stage problems when it is not the input stage itself that is the problem, and thus d.c. or steady sinewave single stage optimisations are pointless if, when all circuit elements are subsequently brought together, the combined circuit operation becomes dynamically compromised by interstage circuit charging current insufficiencies that lead to NFB loop delay when attempting back EMF correction, with reverse driven output terminal voltage overshoot then causing an overall amplification response to become non-linear.

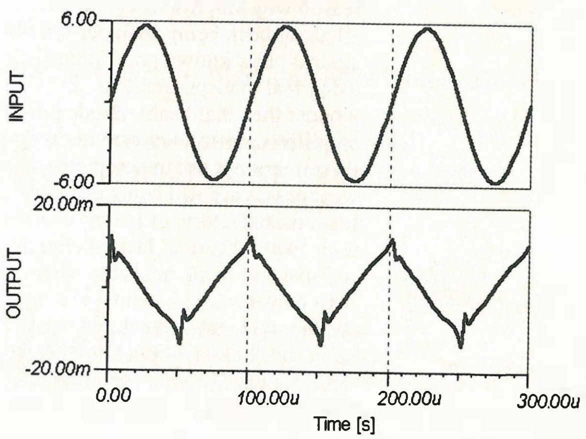
The effect of non-linear and higher frequency leading current flow within circuitry that is encompassed by an output terminal voltage sensing plus internal error correcting closed NFB loop has here been illustrated in a way that cannot be shown up using established steady sinewave testing techniques employing an always 'in phase' passive resistor load. Any NFB loop controlled class-B amplifier is good only for as long as its output and all other stages can drive current linearly without the NFB becoming separately or interactively shunt limited when dynamically attempting voltage sensed output current correction, or without the loaded closed loop encompassed circuitry being rendered slower than the very much higher frequency and differentially sensed 'audio signal versus amplifier non-linearity plus loudspeaker back EMF induction' errors it must correct in order to establish accurate equal but opposite current driven back EMF (damping) control at high audio frequency. Directly coupled pure class-A amplifier circuits already pass high output stage current and thereby retain a higher fundamental level of output terminal damping, thus they will continue to more naturally limit the development of voltage leading (tweeter driving) loudspeaker-system induction at their own output terminals during a momentary degradation of NFB loop control caused by input or load induced overload.

'Miller C.dom' - contraindicated

Yet again I could imagine some designers thinking that these findings are not relevant because I have taken away the output choke that is an essential part of their circuit architecture; but they would be wrong, for voltage error at the NFB loop controlled output node and thus overall amplification linearity itself is affected by load induced current flow, and loudspeaker generated back EMF currents pass whether an output choke has been interposed or not. Thus, when a Miller connected VAS C.dom is being used entirely natural dynamic loudspeaker back EMF generation can cause reverse current driven voltage change at the NFB loop output node and simultaneously render forward signal amplification processes non-linear w.r.t input whether an output choke is used or not (see Part 6); this is additionally illustrated when the conduction crossover induced spikes extend in both time and amplitude beyond the already non linear and triangulated Miller C.dom charging error waveform illustrated! The Miller C.dom slew rate induced NFB limitation problem becomes exacerbated at higher audio frequencies and at higher listening levels with increased loudspeaker efficiency, with output device storage effects that require additional control, and with imbalanced output half characteristics that arise due to the shifted load angles of dynamically and electrically reactive loudspeakers, especially when simultaneously power audio driving close to a bass loudspeaker's resonant frequency.

The dynamically reactive input stage charging of a Miller connected C.dom by a lowly and fixed bias

Figure 19: Miller C.dom delay increases conduction crossover non-linearity with increasing reverse drive.



input transistor pairing can render a class-B VAS collector slow enough at higher audio frequencies to allow the output devices to become loudspeaker back EMF reverse driven through a portion of their crossover bias potential, before the propagation delayed NFB loop can properly minimise the differentially sensed output voltage error; subsequent error correction then becoming exaggerated by overshoot and eventual resettlement. The lower the bias, the more easily that loudspeaker induced back EMF will overcome the output stage prior to establishment of propagation delayed NFB loop control, and it is this predominantly Miller C.dom induced distortion that we can empirically minimise as we adjust the bias of a class-B output stage whilst listening to its loudspeaker output.

It is the ninety degree leading Miller connected C.dom loading within the differential input pairing that prevents a near instantaneous control of the very high speed and sharply discontinuous 'reverse commutating' (conduction crossover) distortion, as is illustrated by the spiking in Figure 19. Leading input transistor current peaks and their associated voltage error developments are anomalies previously reported by John Ellis in my EW March 2003 reference, yet these signal driven conduction crossover anomalies are not in themselves input-output measurable at a resistor monitored output terminal. The current non-linearities independently arising due to imperfect NFB loop corrected output do however lead to a much greater increase in high audio frequency distortion when efficient and loudly driven dynamic loudspeakers generate back EMFs w.r.t. signal amplified output, and this cannot be observable via conventional 'dummy' load test bench resistor measurements, this being the main reason why amplifiers should have all along been being designer tested against other known good amplifiers using real loudspeakers! Little wonder then that beam tetrode power amplifiers are still being made and their minor but not unacceptable weaknesses are still being deemed less unsatisfactory, or maybe even their 'warm hearted' lack of clinical incisiveness being desirable when high power music listening is the requirement, even though modern solid state chassis should already be good enough. Bipolars and Mosfets will never look as impressive as this in the dark though; see: www.turneraudio.com.au/htmlwebpgs02/300monobloc.htm

When a class-B circuit similar to Figure 14 is used to loudly drive complex loudspeaker systems without its series output choke, it can sound much more distorted than standard resistor based test bench error determinations or simulations suggest, also it can be more distorted than when other different stabilisation and NFB arrangements are directly applied to its unchanged base semiconductor circuitry. These Figure 18 and 19 simulations directly illustrate how additional Miller connected C.dom induced current flow through the previously phase coherent differential input sensing transistors can alone cause loudspeaker driven audio frequency deviation due to the resulting inadequate VAS collector voltage slew w.r.t. output stage node induction and this arises whether the input stage is processing any input signal or not. The amplifier's output becomes leading C.dom charge-discharge zig-zag voltage modulated by resultant load driven current variation w.r.t. signal input voltage, and this can sometimes sound like crossover distortion or a voice coil scraping a pole face. And yes of course the distortion can be reduced by increasing NFB, but that means more open loop gain - which means a greater potential for instability within the closed loop - which means more...

It is the signal path, local and global NFB components which are additionally connected to or are missing from any semiconductor topology that lead to a reactive load dynamically generating input or output driven waveform distortions. Applying the very best of transistor parameters to a Figure 14 like circuit makes no significant difference to simulated results, and this mirrors my own real life experience of fitting selected good transistors and not improving the sound through conventional dynamic loudspeakers. I've also observed newer and better output transistors actually degrading electrostatic loudspeaker reproduction after they had been fitted as replacements to a high fidelity amplifier that had been optimised for use with older devices. Often there is nothing to be gained from 'upgrading' an old amplifier by fitting it with the latest transistors, so unless NFB and stabilisation networks are also going to be competently modified, then maybe it is best to do no more than update the R's and C's, check joints, connectors, etc.

Due to the inherent stability of a VAS connected Miller C.dom stabilisation circuit both of the Figure 16 and 18 leading error potentials could have been mathematically determined from the Decibel and sine angle ratios read directly from Figures 15 and 17, as indeed could the capacitor induced first cycle potential shift be calculated from Figure 12, however these steady sinewave based calculations would not necessarily cover the dynamically induced onset of internal signal path capacitance degraded linearity distortion (Figure 18 triangulation) or inadequately controlled output stage conduction crossover distortion at higher output levels (Figure 19 spiking), nor any additional spiky sub-microsecond overshooting due to the reverse driven stability margin on designs which might become dynamically compromised by reactive loading.

Reverse injection simulating with more highly biased input and output pairings or with power Mosfets in place of class-B output bipolars will illustrate the potential for linearising and reducing the triangulated chokeless error waveform between conduction crossovers, thereby recovering a plain leading sinewave phase shift without spiking for the first few back EMF volts. There might then however be temptation for a designer to maintain Mosfet amplifier stability by fitting a single but larger value of Miller connected C.dom in order to overcome possible instability in the 1 to 10MHz region, but then additional capacitive charging currents within the differential input stage could increase the closed NFB loop propagation delay when an input stage mirror is used rather than plain resistors, which could further increase the risks for dynamically induced high frequency back EMF leading voltage (tweeter driving) development, without this showing on a forward steady sinewave and resistor measured nulling display or via a THD specification. If a low angle of reverse driven error cannot be assured at 10kHz, then its dynamically observed NFB loop reverse amplitude response should be extremely fast, stable and of low level for high audio frequency NFB loop product generation to be rendered inaudible when driving real world loudspeakers.

I conclude these reverse driven amplifier investigations in Part 6, by relating findings to simulated fundamental nulling at the output of an input driven but virtual loudspeaker loaded Figure 14 circuit to illustrate amplifier-loudspeaker distortion.

Simulating power MosFets

In this, the first of a planned four part series using the Microcap6 software, Cyril Bateman introduces a hands on approach to Spice circuit simulation and the use of user created power MosFet models, able to accurately mimic actual power MosFet behaviour – a new era in realistic Spice simulations

Memorandum ERL-M520 published in May 1975 by the University of California at Berkeley and still available together with the associated software, describes Laurence Nagel's development of Spice2, a public domain program which commercial developers have enhanced to become the de-facto low frequency circuit simulator.

Spice, short for 'Simulation Program with Integrated Circuit Emphasis', describes both it's intent as well as its weaknesses. At the time it was developed, most design involved discrete transistors only, but an explosion in the variety and complexity of IC designs had begun, mostly targeted to lower frequency analogue circuits, however some low power Mos integrated circuits were also then becoming available. Significantly, development of Spice2 was completed before the power MosFet was produced. As a result, while the core equations built into Spice2 provide for most common components including small signal MosFets, no provision at all was made for modern power MosFets or more recent devices. Capacitors, resistors and inductors were included but only as idealised theoretically perfect components, suited for use in Integrated Circuits, but far removed from almost all discrete, real, components.

Since then, Spice2 itself has undergone various modifications, with newer versions such as the upgrade called 3F4, now included in some of the more recent packaged simulators, including the no-cost WinSpice3. To benefit from these enhancements, these upgraded simulators often require the use of equally enhanced but scarce, component models. For this reason most commercial Spice versions provide backwards compatibility with Spice2G6 and include a library of suitably compatible models. Spice2G6, considered by many the definitive Spice, was the last update of Spice2 written in Fortran, the Spice3 series being similar to Spice2 but re-written in 'C'.

Other developments, made using 'subcircuits' which can be imported instead of using the conventional Spice2 'models', have since provided for components such as the DMOS power MosFets used as switches, as well as MesFets, IGBJTs etc. Some of the most expensive simulators have also been substantially internally upgraded, for example Hspice, a professional version of Spice specialising in the design of complex ICs, now provides for some fifty different MosFet model categories, but not lateral power MosFets. Less costly Spice2 versions are usually limited to just levels one, two and three, making the provision of models suited to calculating power MosFet distortion, unduly difficult, the main topic for this series of articles.

Spice2 origins

The Spice2 development at Berkeley followed the successful completion of two earlier simulation programs. The first developed in 1970/71, called Computer Analysis of Non-linear Circuits Excluding Radiation, abbreviated to Cancer, was initially funded by a grant from the Sprague Electric Company, followed by grants from NSF and ARO which provided for the completion of Cancer and the original Spice program in 1972.

More than one hundred copies of the original Spice program were supplied to universities and electronics companies and used with such success they encouraged the development of the much improved version now known as Spice2, the subject of Laurence Nagels' thesis published in 1975.

The 'Integrated Circuit emphasis' results in part from this funding, Nagel was part of the universities Integrated circuits group, combined with the difficulties then found developing larger analogue integrated circuits. Discrete circuits could be easily prototyped using conventional breadboard methods, but as integrated circuits became more complex, breadboarding was of little use, resulting in increasing dependence on simulation methods.

One commercially successful early development from the Spice2 program was PSpice by Microsim. While other simulators relied on the original Spice2 coding, Microsim significantly overhauled and refined the original codes improving convergence, while adhering to the original UC Berkeley Spice2 standards. PSpice v5.1 for DOS was the first Spice2 based simulator used by this writer.

Component model or subcircuit?

One solution lies in creating a subcircuit, which will be indicated in your schematic as an 'X', rather than a MosFet model indicated as an 'M' in your drawing. While a Level-1 MosFet 'M' model can be simply created using the 'Model' program supplied with MC6 and similar simulators, the resulting model is constrained to use only those parameters provided in Level-1. In contrast, while it cannot be made using this 'Model' program but must be 'hand carved', a subcircuit has freedom to include almost any valid Spice2 'device', permitting use of e.g. a voltage multiplier or a voltage to current converter. However as detailed in your Spice software manual, while some of these Spice2 devices cannot be fully used within a schematic, others may have restrictions used in a subcircuit.

The better Spice2 simulators now provide 'analogue behavioural' modelling tools which can be used within a subcircuit, but here the global standardisation implied by Spice2G6 does not always apply. Specifications for some analogue behavioural model elements unfortunately can vary from one brand of simulator to another, so making some models only able to work in their specified simulator. However quite useful subcircuits can be devised without using complex behavioural elements as shown in **Figure 1**. (*This FDP038 subcircuit model was my first attempt to edit and incorporate a new subcircuit into my MC6 component library. As can be seen, apart from the MosFet being assigned the reference X1, the subcircuit functions within a schematic exactly like any other component. This X1 symbol represents a netlist some two pages long, the circuit shown in figure 2.*

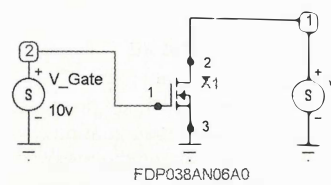


Figure 1: This circuit demonstrates the technique for plotting the IV curves of a MosFet subcircuit model using DC

Audio power MosFet modelling was discussed some years ago in a two part Ian Hegglun article[1]. Now needing to model audio power MosFet distortion for myself, I had optimistically hoped that with this passage of time, Ian's reported problems would have disappeared. Not so, as searches for lateral power MosFet and Spice using Google quickly reveal. Almost no models can be found and those that do exist are mostly simple Level-1 models, not at all well suited to modelling power amplifier distortions. Indeed Google highlights several hundreds of enquiries asking where suitable models might be found. As will be illustrated later, manufacturer's models of power MosFets when used to design audio power amplifiers, can present significant problems for audio designers simulating distortion.

In 1996 the AES pre-print 4298 written by Scott and Parker[2] of the universities of Sydney and Macquarie published a report highlighting this problem together with their modifications of the new Parker Skellen MesFet model, which allowed this model to be used also with power MosFets to produce accurate distortion simulations. Unfortunately[3] few even of the most recent Spice simulators are compatible with this Parker Skellen Mesfet model, even worse I was not able to find a single Parker Skellen MosFet model on the internet other than the one included in this AES pre-print. Some new Spice simulators do provide for the EKV MosFet model[4], but again I was not able to find a single EKV compatible power MosFet model.

Spice2 in practice

The Spice2 simulation software comprises two main components, a simulation engine used to automatically solve circuit conditions according to the Kirchhoff's network laws, and a library of compatible component models.

To allow the software to analyse any circuit, it uses methods similar to those previously used for manual solutions using a slide rule or log tables. At the time Spice2 was developed, portable calculators were rarely available, almost all practising engineers as I well recall, still relied on using their 'guessing stick'. Even as late as 1975, the 'Nelson Jones circuit designers slide rule' which I still on occasion use, was considered an essential briefcase tool.

To provide a fully automatic solution, the software and models used in Spice2 had to be constrained to follow certain rules. As a result the performance characteristics expected, e.g. a silicon junction with temperature and also its voltage curves, are described within the core software using equations. All models in the library are forced to use only combinations of these pre-defined characteristics. Any additional characteristics, such as ideally needed modelling power MosFets could be added by modifying the Spice2 source code, which must then be re-compiled before use. Such enhancements are only practical as part of a major software upgrade and re-issue.

The circuit to be simulated must first be described in a netlist, following a few simple rules. Using a Windows schematic entry version of Spice, this is performed transparently assigning circuit nodes to your schematic drawing which is then automatically converted into a netlist for the simulation, see **Figure 2**. (*Stage 1 in producing the netlist for the subcircuit of figure1. The nodes 10, 20, 30, representing the drain, gate and source connections to the subcircuit model, are translated by the schematic shape editor into the connecting nodes 1, 2 and 3 for the MosFet in figure1. This schematic was used to plot and quickly refine the models response, prior to producing the netlist needed for the library model of figure 1*).

Netlists

The original versions of Spice2 required the circuit be described using a 'Netlist' of components hand written using a text editor, together with a description of the stimulus waveform to be used and details of the required voltage/current outputs. Each circuit node, the connection between two or more components, must be assigned a unique 'node' reference, usually but not necessarily

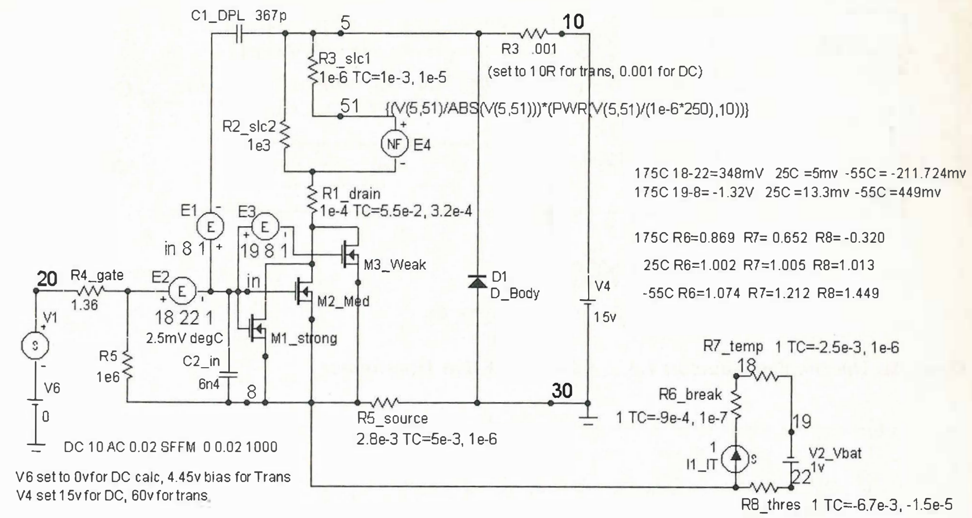


Figure 2: Schematic test version using FDP038 data.

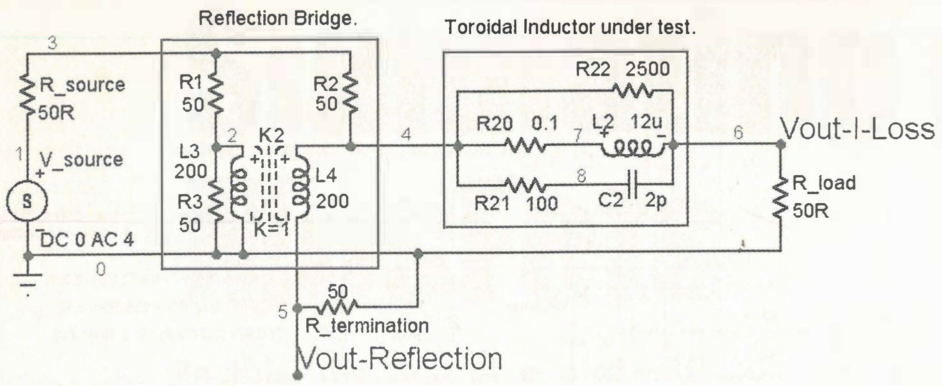


Figure 3a: The coupled inductors L.3, L.4 for a 1:1 Balun Transformer

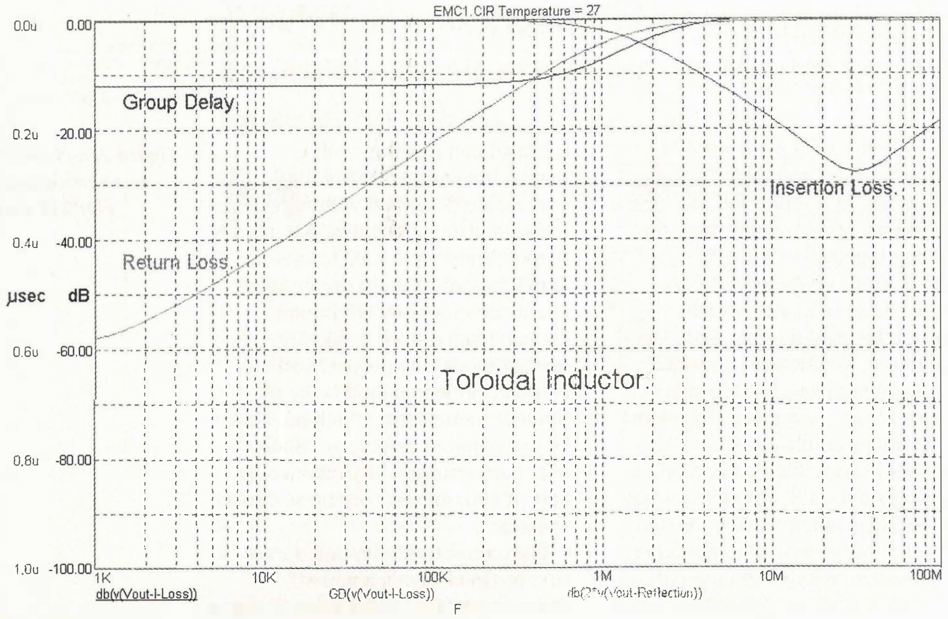


Figure 3b: These plots for insertion loss, return loss and group delay were made using the figure3A circuit. They can be quickly produced using the most basic Spice2 simulator.

Figure 3C: Example of a Spice2 Netlist exported from Figure3A

```
C:\MC6\DATA\EMC1.CIR
Multiple Analysis
 * Represents a network
analysis Reflection
Bridge.
 *6dB Directional Bridge
- see HP 8721A data.
 K2 L3 L4 1
 L3 2 0 200
 L4 4 5 200
 R1 2 3 50
 R2 4 3 50
 R3 0 2 50
 R_TERMINATION 5 0 50
 *Inductor under test
 C2 8 6 2P
 L2 7 6 12U
 R20 4 7 0.1
 R21 4 8 100
 R22 4 6 2500
 *V_source and R_load
 R_LOAD 0 6 50R
 R_SOURCE 1 3 50R
 V_SOURCE 1 0 DC 0 AC 4
 .END
```

using a number, except for ground or earth which must be numbered as 0, a reserved number meaning ground reference, node zero. This node 0 ground is a mandatory requirement and must exist in every circuit. All other circuit nodes must have two or more connections, also a DC current path to this ground node, both are essential requirements of the Spice2 algorithms.

Today the Windows schematic versions still use a similar netlist, but it has been made transparent for most circuit design needs, except when creating a new device model subcircuit to add to your library of components. As has already been seen, in the absence of a satisfactory manufacturer's model, we must create our new subcircuit component model ourselves.

This new component must first be described as a netlist, following the simple rules needed to allow the simulator to recognise the new subcircuit model. When finished, for use with a schematic circuit, it must then be associated with a schematic circuit drawing 'shape' before inserting into your component library.

Using a Windows Spice2 simulator, the simplest way to produce a new subcircuit model is to draw it initially as a schematic circuit, complete with a stimulus, voltage sources and loads, much as it will eventually be used, when called from within your schematic. This allows your model to be quickly tested by running simulations, adjusting its components as necessary by amending values in the 'model' schematic.

When satisfied, simply output and save to a file as a Spice2 textfile netlist, as shown in **Figure 3**, which can be opened, modified as needed then saved, using a simple text editor which does not insert control codes. For this I prefer to use Windows Notepad. *(Figure 3a: Illustrating how easily Spice2 can be used to model insertion loss, the reflected signal indicating return loss as well as group delay of the toroidal inductor. The box shown as 'Reflection Bridge' represents the classic reflection bridge part of a network analyser. As can be seen to realistically model an inductor at frequency, requires a number of additional components representing its losses.*

Spice2 – warts and all

Equipped with a suitable Spice2 software package and a component model library, are our simulations accurate, can we rely on the results? Provided our models are correctly designed, suited to our task and used within their frequency limit the answer has to be a qualified 'yes'.

The only way to determine whether a Spice2 model has been correctly designed, is to perform a simulation, then compare its modelled results against actual measured results, alternately against the datasheet values. The best simulator used with an inadequate component model, can never provide accurate answers.

One example can be found in op-amp models, mostly these are relatively simple, small, 'macromodels' produced to approximate the op-amp behaviour in many applications. Reading the small print attached to each model should indicate whether it is suited to your needs or whether you should obtain a more complete model. The most complete op-amp models use large numbers of transistors effectively replicating the full schematic drawing. While these transistor level models produce the most accurate simulations, they are considerably more complex so run more slowly than when using a macromodel.

One good thing in Spice2 is that it takes no account of whether any component is being overstressed so depending on your design, it may produce seemingly impossible outputs, if that is what the Spice2 interpretation of the model used and your design implies. For this reason the most complete subcircuit models do include components to restrict say the voltage output from an op-amp to realistic levels. A few of the most recent switching power MosFet models are also this complete, so can indicate when exceeding their breakdown voltages. However as a bonus, simulation using Spice2 does not break overstressed devices, allowing you to interrogate the circuit to ascertain potentially damaging conditions, without risking expensive breadboard components, or destroying a power amplifier.

Two other points should be noted. When performing an AC (frequency domain) analysis Spice2 cannot take account of any voltage level dependant non-linearities, but may account for frequency dependant parameters. Spice2 effectively performs only a 'small signal' analysis, ignoring any voltage dependencies, even though you may have specified a very large stimulus voltage.

When performing a transient (time domain) analysis, Spice2 cannot take account of any frequency dependant variables. This is particularly important with capacitors, which can exhibit both frequency and voltage dependant parameters. If such a capacitor has for example its ESR specified to change by frequency for AC analysis, when performing a transient analysis Spice2 defaults its frequency parameter (FREQ) to zero, making the capacitor ESR appear as infinite, not at all a useful value.

The Spice2 simulator calculates all results using a process called 'convergence', typically using the modified Newton-Raphson method and trapezoidal integration. Essentially guessing a result which it then uses to re-calculate the circuit, noting the amount and direction of any errors. It then refines this guess and repeats the process, in order to converge towards its final answer. Occasionally this convergence process goes wrong and instead of each re-iteration approaching closer to the final answer, it diverges until stopped by the inbuilt error limits or exceeds the floating-point calculation limits.

This process is rather like skimming pebbles across a pond. Mostly the pebble approaches the water surface at a suitable angle, bouncing across the surface several times, just like the actions needed for a circuit to converge. Occasionally however, a pebble approaches at the wrong angle to disappear instantly beneath the surface, like Spice2 failing to converge.

Each Spice2 simulator includes a table of global parameters, essentially simulation error tolerances, which determine when the simulator has converged sufficiently towards a solution. These standard global parameters were established during the Spice2 development so suit the simulations made many years ago. Modern circuit needs can be more difficult to simulate so Spice2 fails to converge, the calculation exceeds the permitted floating-point limits, displaying error messages. Even worse on occasion the program shuts down, so it is wise to save the most recent design, before running a simulation.

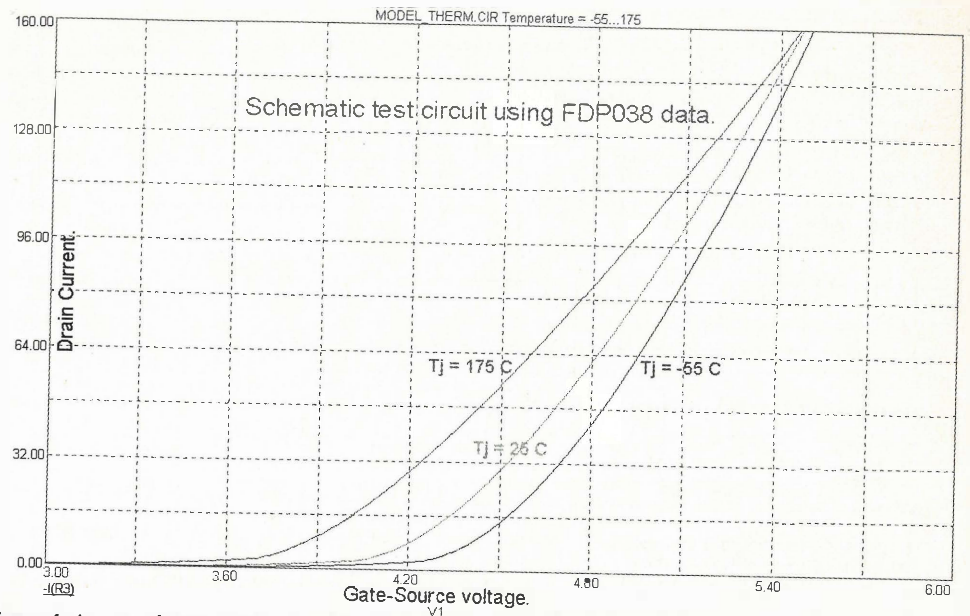


***Figure 4:** Accurately replicating the datasheet curves, these simulation plots illustrate how change in junction temperature affects drain current with gate-source voltage for a power MosFet. Note especially the large shift in threshold voltage, which affects the crossover region in a power amplifier.*

Other simulators

Prior to obtaining PSpice v5.1 for my PC, I had written a number of dedicated simulation programs running on Hewlett Packard technical workstations, used to design also automatically production test, capacitors and EMC filters. Subsequently I wrote for my personal use a PC based version using visual Basic, this program, called EMCfiltr, was used by several companies for their own design needs, also to illustrate my EMCfilter article and recent letters page.

While Spice2 is targeted to transient or time domain simulation, many other simulators target the frequency domain, simply because many components, especially capacitors, inductors and certain semiconductor's characteristics vary more with frequency than with amplitude. Some also vary in both domains, which presents additional difficulties leading to the use of an 'harmonic balance' type simulator.

The original de facto frequency domain simulator was called Touchstone and was marketed by Hewlett Packard. It was superseded by the Hewlett Packard developed MDS or Microwave Design System. I used MDS successfully some twelve years ago designing high frequency components. At that time it was the only commercial simulator that accepted dielectric constants up to some 5000, as needed for capacitor and EMC filter RF simulations. Today MDS has been updated to the Hewlett Packard ADS or 'analogue design system'.

One university devised simulator I particularly admire comes from Finland, not the US. APLAC has many advantages over other software, in that it provides for time domain, frequency domain and harmonic balance simulations, with or without thermal and electromagnetic simulations. A most complete simulator, with the added benefit that a size limited, no cost, student evaluation version is readily available.

You should also be aware of another expensive but excellent simulator, called 'Saber', simply because some semiconductor makers now offer simulation models for Saber as well as Spice2. Saber models are incompatible with any Spice based simulator.

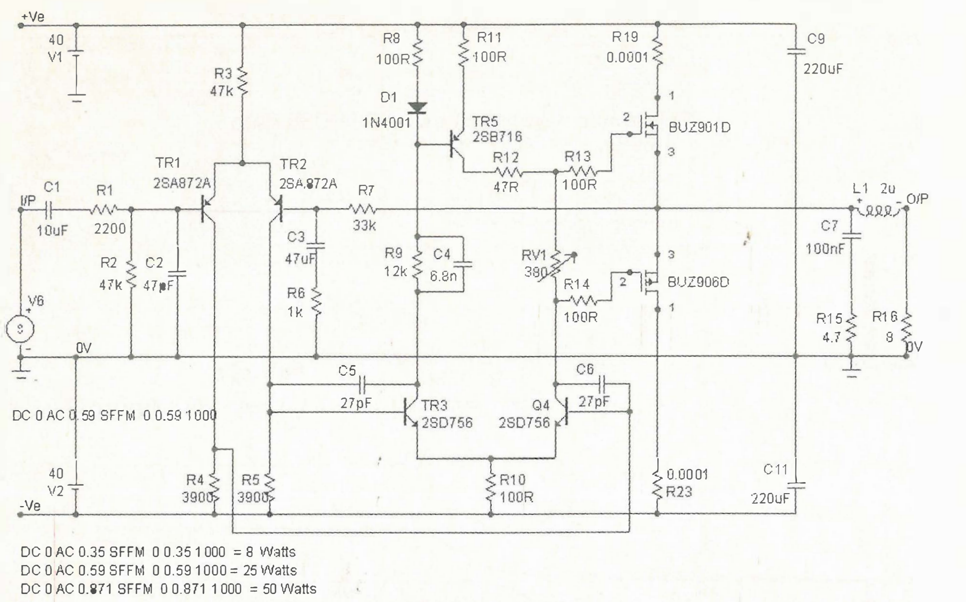


Figure 5: The circuit of my revised Maplin amplifier featured in my November 2003 article. In order to minimise the Level-1 MosFet model errors when modelling its distortions, I used my revised BUZ901/906 Level-1 models for this example, however these models cannot correctly reproduce the subthreshold region curves as does the FDP038 subcircuit.

Useful reference

When I first started using a PSpice simulator, I found the book ***Spice - a guide to circuit simulation and analysis using PSpice*** by Paul Tuinenga, written while he was with the Microsim Corporation, especially useful. At the time it was written, it predated the Windows software so assumes using netlists throughout. However it is easily read and understood so several years later and now using the MC6 Windows schematic software with its two manuals totalling 900 pages and 'on screen' help files, I still find the Tuinenga book helps, by quickly answering many questions. That's not to say that using a Windows Spice2 simulator is difficult, because it isn't, but rather that from time to time I find myself involved in trying to perform a more difficult or non-standard task than usual, or perhaps its because I'm getting old. I commend this book to any reader wanting to start experimenting with Spice2 software since with it much useful work can be performed using one of the many no cost, size limited, student Spice2 versions.

These problems can often be removed by adjusting the global parameter convergence limits, the maximum permitted time step for transient simulations or even slight adjustments to the circuits supply or stimulus voltages. A particular limitation when modelling distortion for power MosFets is that Spice2 models every component using a common temperature control, as indicated by the note about temperature, which heads every output plot. For many components instructing Spice2 to model using say a 55°C junction or ambient temperature may well be correct but not for the power stage which may experience junction temperatures way in excess of 100°C. As study of a datasheet reveals, power MosFet characteristics do vary significantly with junction temperature, as shown in **Figure 4**.

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 schematic '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 includes the primitive components needed to model a simple op-amp within its core software.

When creating a subcircuit for our power MosFet, then the subcircuit nodes representing its drain, gate and source electrodes must be assigned in the first line of the 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
```

Using the conventional, mostly Level-1 models supplied with my simulator, I had no choice but to accept a global simulation temperature. In 1991 a paper by W. Hepp and C. Wheatley, referenced in the FDP038 datasheet, described how this problem could be overcome to model the power stages at working temperatures, with the remaining components simulating at their realistic ambient temperature. Unfortunately again I was not able to find any audio power MosFet models which used this feature.

When modelling distortion of a conventional power MosFet audio amplifier, the models' behaviour as one device is turning off and the other commencing to conduct, in the MosFet low current subthreshold region, is crucial.

The Spice2 Level-1 MosFet model is quite basic so cannot properly model this subthreshold region. As will be seen later, simulations using a Level-1 MosFet model do not model bias currents or harmonic distortion at all accurately. This applies also to some subcircuit models, mostly simple extensions of a Level-1 model using external capacitances, to better model a switching transistor. These simple subcircuit models also do not accurately model the subthreshold region, so produce unacceptable results when modelling an audio amplifier.

Examining a model using a text editor quickly reveals the models status.

User defined model

I decided I had little choice but try to produce a more acceptable model myself. My copy of MC6 was provided with a model module intended to facilitate creating one's own semiconductor simulation models, something I had not previously needed to use, despite using this simulator for more than four years. Unfortunately while the MC6 simulator supports Levels 1, 2, 3, 4, 5 and 8 MosFet models, its accompanying model creation program only produces MosFet models at Level-1 and cannot be used to make a subcircuit.

My capacitor and inductor model needs however had been quite different, frequently I was forced to hand craft more acceptable models

for these, since a capacitor and inductor model development facility is not provided by the MC6 semiconductor modelling aid.

Not having access to a curve tracer I tried following the MosFet maker's published data, but with little success. My interests were mostly for drain currents of three amps and smaller, down to just a few milliamps for the crossover region. The maker's data which effectively concentrated on drain currents from three amps to the sixteen amps maximum of the chosen device, did not help. Unfortunately I had already selected and obtained these particular transistors, mostly because the datasheets illustrated good matching between N and P types, not apparently available with other lesser-rated devices.

After many trials I obtained a pair of "Level-1" MosFet models which appeared to provide a closer fit to the datasheet at low drain currents than did the models supplied with my simulator. So I commenced my experimental simulations as shown in **Figure 5**. My original task had looked quite simple, I had hoped to produce simulated distortions in line with the practical measured results made on my modified Maplin amplifier. My new Level-1 MosFet models worked in that I could reproduce the input and output waveforms both open loop and closed loop. The high frequency response curves however looked particularly suspect, clear indication of badly modelled MosFet capacitances, when compared with actual measurements on the amplifier. Worse still, using the same output stage bias currents as used for the measurements when simulating distortion, although the distortion simulated using these new models was smaller than when using the models provided with the simulator, all simulated distortions were many times greater than those measured on the actual amplifier, clearly illustrating the problem reported by Ian Hegglun, shown in **Figure 6**.

Originally I started this exercise as a useful way to recuperate following my cataract operation the last week in October 2003. The hospital having forbidden me any exertions for 6-8 weeks, to allow my eye to heal. This exercise had now subtly changed from an interesting topic into a challenge which I was now determined to solve. As an added incentive, three notable audio designers had recently mailed me asking questions about this very topic, which I had been unable to answer.

Having studied a great many

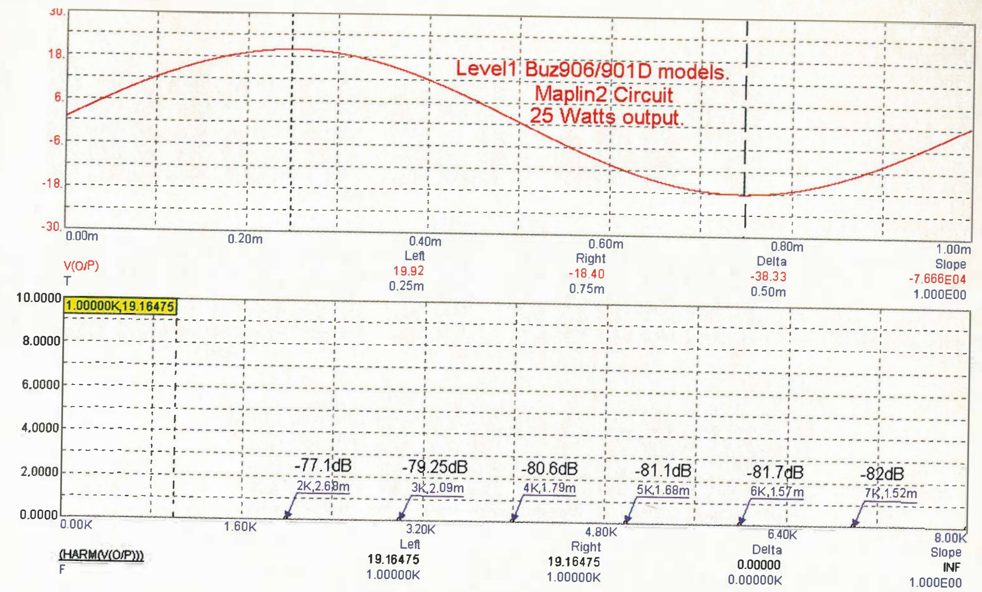


Figure 6: While these BUZ901/906 models produced less distortion than other Level-1 models tried, distortions in this simulation far exceed those measured on the actual amplifier. Second harmonic of -77.1dB and third of -79.25dB are gross errors compared with the -92.1dB and -94.3 dB actually measured on my working amplifier.

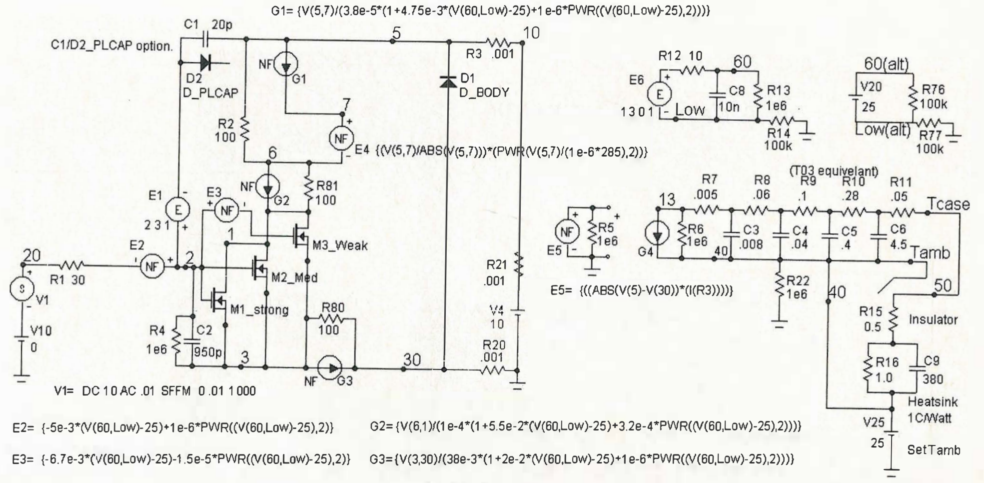


Figure 7: Self heating models accurately simulate distortion

reference documents downloaded from internet, I slowly realised much of the information I presently lacked, could become available, enabling me to devise thermal subcircuit models more suited to simulating audio distortion. Eventually after several weeks work, I managed to produce self contained, self heating models for my BUZ900/05D lateral power MosFets, overcoming the Spice2 global temperature restrictions, with models which closely replicate the devices subthreshold conduction.

These self heating models can be used together with heatsink models, to continuously monitor the transistor junction temperature, automatically adjusting the MosFet characteristics with each change in junction temperature just like a real transistor, introducing a new era in realistic Spice simulation for the user **Figure 7**.

Creating a working power MosFet subcircuit thermal model from a datasheet, forms the topic of my next article.

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2. A Mosfet model for the simulation of amplifier non-linearity. J Scott, A Parker. http://www.elec.mq.au/cnerf/papers/AES4298.pdf
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PIC crystal tester and frequency counter

Ever rummaged around in the junk box and found unidentified crystals? Or even wanted to verify the frequency when fault finding? Then Hamid Mustafa's little box is for you

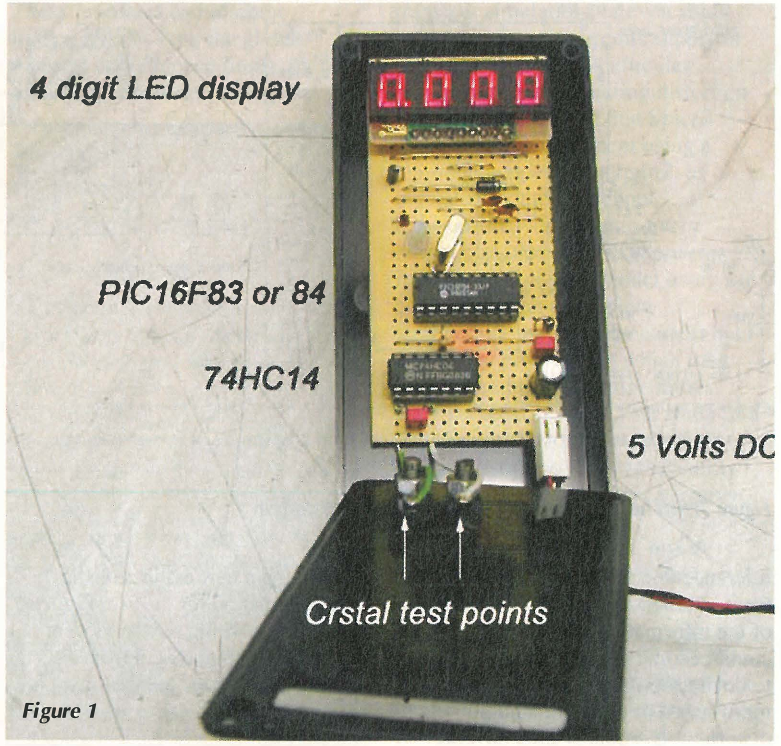


Figure 1

This crystal tester was designed for testing the quartz crystal elements used in thin film coating machines. In the machine, a thin piece of quartz is used, attached to an oscillator circuit and as the parts get coated, so does the crystal, which lowers its resonant frequency. The frequency change is a measure of coating thickness in microns. This tester can also be used for testing crystals used in colour TVs and video recorders. Colour decoder TV crystals usually resonate at 4.43MHz (3.58MHz for NTSC).

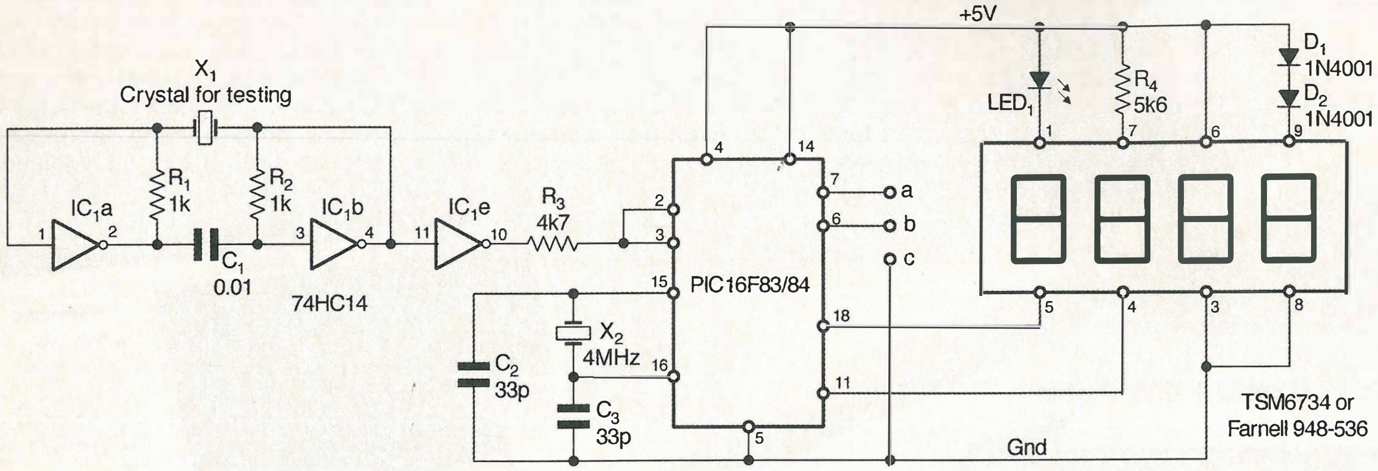
The heart of the tester is a PIC16F83/4 programmed to function as a frequency counter driving a 4 digit LED display, TSM4000, TSM6734 or Farnell 948-536. In this type of display, data is serially written and only two port pins are needed from the PIC. The PIC itself operates with a 4MHz crystal clock as shown in **Figure 1**. An LED flashes at 5Hz to indicate that the crystal being tested is OK and its frequency is within the range of the tester.

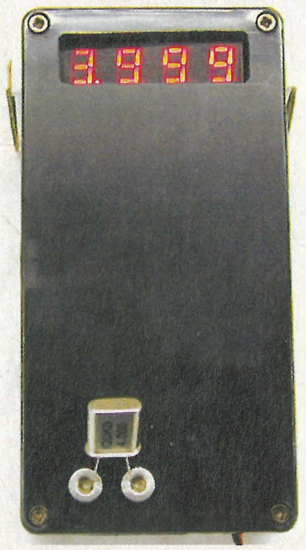
Because the display has 4 digits, the measured frequency range is 1KHz to 9.999MHz. However the 4 digits are intended to show MHz, and 1KHz will be displayed as 0.001. Although this tester was not intended to be a frequency counter, by adding a second four digit display or a 20 digit LCD display and modifying the software, an 8 digit frequency counter can be made with a range of 1 Hz to 50MHz. If used as a frequency counter, remove the 74HC14 from its socket and feed the signal to be measured to pin 10 of the socket.

The hardware

Figure 1 shows the simplicity of the hardware and the wiring. A 74HC14 hex inverter is used to oscillate the crystal being tested. The pulses from the oscillator are fed to the counter input of the PIC through a 4k7 resistor and are counted. The counter input pin of the PIC (pin 3) is connected to an I/O pin, pin 2 which is configured as an input (high impedance) while counting. When the counting time is up, pin 2 is configured to be an output, and is switched to LOW.

This shorts the counter input to ground preventing further pulses from reaching the counter while the PIC program reads the counter and prescaler values to determine the frequency. The 4-digit display is driven from two port pins of the PIC, pins 10 and 11. A PCB was not made





for this project due to the small number of components and wiring. It can be built on a small piece of Veroboard 8cm x 4cm.

If a more precise reading of the measured frequency is desirable, replace the 33pF capacitor on pin 16 of the PIC with a 15pF capacitor in parallel with a 50pF variable capacitor and adjust while testing a known crystal.

The author built three prototypes using three different 4MHz CPU crystals and compared frequency measurements with an 8-digit frequency counter and found the accuracy more than adequate.

The 4 digit LED display

The TSM 4000 or TSM 6734 has nine pins. Only two pins are connected to the PIC, the others are for brightness, power, enable etc. If a TSM6734 cannot be obtained, a Farnell 948-536 can be used instead. When a TSM6734 is used, link pins 5 and 6 of the PIC. For the Farnell display link pins 6 and 7 of the PIC. Also, for the Farnell 948-536 type display, diodes Dl and D2 and the resistor R4 are not needed. For the Farnell display, leave pins 7 and 9 of the display unconnected. The Farnell 948-536 has a link on either side of the 9-pin connector, these must be jumpered.

How the software works

The PIC program was written in C language and includes lots of comments and explanations. It should be a good starting point for anyone wanting to learn to program PICs in C language.

The entry point of the program is the function main( ). The first thing the program does is configure port B pins as output except bit 0, and port A pins as all input. Next, the display is initialised by taking the data pin low and clocking the clock pin 40 times. This blanks the display by ensuring its internal registers are zeroed. Next, the option register is written with 11100111. This configures TMR0 as a counter with a prescaler of 256. This means that every time the prescaler reaches a count of 256 it will overflow, reset to 0 and increment TMR0. Note that the prescaler can count at 50MHz.

The software then waits for exactly 1ms and when the 1ms is up, it shorts the counter pin to ground by making pin 2 an output and making it LOW. The number of pulses received in 1ms are now safely in TMR0 and the prescaler.

Because we count pulses in 1ms instead of pulses per second, the frequency displayed will be in KHz. With 4 digits, 4433618 Hz will be displayed as 4.433. The value of TMR0 can be read directly, but the prescaler cannot. To get the count in the prescaler, we clock the counter's input pin (pin 3) using pin 2 (remember, pin2 and 3 are linked). The number of pulses needed to overflow the prescaler and increment TMR0 gives us the prescaler count.

For example, if the prescaler was holding a count of 200, clocking it 56 times would overflow it. Therefore the prescaler count = 256 - 56 = 200. The only thing left to do is to multiply the value in TMR0 by 256 and add to it the prescaler's count. We then convert this total count to 4 digits using printf( ) and display the 4 digits. After displaying the frequency, we delay 0.2 seconds before measuring again. This prevents possible fast flicker of the last digit.

A kit containg a pre-programmed PIC may be available from EW if there is enough demand. The target price will be €20, $20 US or £15, which will include postage, a floppy disk with the C program, assembly code, hex code for programming your own, Veroboard assembly instructions and photos of the author's prototype. For more details and an order form, please contact Caroline Fisher, details on page 3.

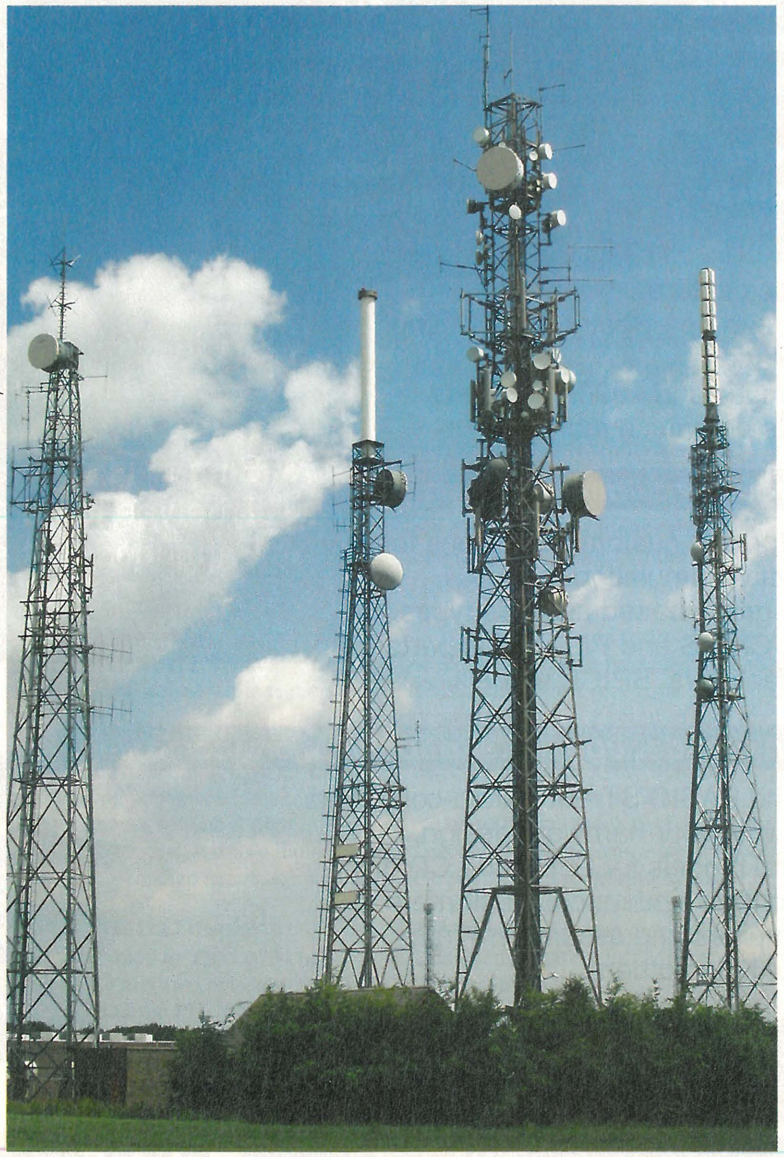
On the floppy you will also find instructions on how to make your own 4 digit serially driven led display which will connect directly to the crystal tester. This uses an M5450 and four HDSP-7501 common anode LED displays.

Readers requiring just the code files, please email Caroline Fisher with 'Frq counter' as the subject and she'll mail them right back to you.

RADHAZ:

The unmentionable hazard?

Every few weeks the popular press features a new article relating to potential health dangers of mobile phones or base stations. Large sums of money are being spent on research in laboratories all over the world. In this article Brian Collins reviews the sources of non-ionizing radiation to which we are exposed, and our present state of knowledge on their effects



Multi-user transmitting site

It is widely understood that some kinds of radiation are harmful to human health, although we all live in a world full of natural radiation of every frequency – from noise at the low-frequency end of the radio spectrum to hard ultra-violet (UV) radiation. Without it we would not be here.

Fortunately the earth is surrounded by ionized particles that absorb most of the more harmful radiation at frequencies from UV upwards. Almost all the radiation that reaches the ground is non-ionizing radiation: the energy carried is not sufficient to ionise atoms by knocking out charged particles. At UV and visible light frequencies there is still enough energy to break the bonds that hold molecules together. We are all familiar with plastics which discolour in the sunshine and dyes that fade in the light. Infra-red (IR) radiation is mainly associated with the transfer of heat. When food is placed in a grill it cooks because the IR radiation is absorbed in the surface of the food, imparting its energy to the molecules from which the food is made.

As the frequency falls further we enter the radio frequency (RF) spectrum. Electromagnetic waves at radio frequency ('radio waves') carry energy which is transferred into any object which absorbs them. We all know that if we irradiate food with a few hundred watts of microwave energy it will get hot and cook. Today's world is full of sources of RF energy with powers from a small fraction of a watt up to a million watts (1MW) and more; we need to understand the effects they can have on our bodies. Clearly we must avoid cooking ourselves, but are there other effects we need to guard against? This apparently simple question is at the heart of the current debate.

Proving negatives

It's clear that there are no major short-term risks in normal levels of exposure to RF fields. If there were we would have seen an epidemic of disease or early death among exposed populations or groups of workers. What is less certain is that continuing exposure to fields at lower intensities causes no long-term ill effects.

The methods of science are not well suited to proving that a suggested cause has no long term effect. There are two reasons for this. Firstly, we don't know for certain what effect we are looking for, and secondly we can only assign some degree of probability to our conclusions. The result of thorough and well-conducted investigations costing millions of pounds/euro/dollars can only be expressed in terms that sound to many laymen like prevarication. Any experimental campaign will conclude at best that in the conditions of the experiment (at the frequency, exposure intensity, duration, modulation scheme &c) the effects searched for (mortality, soft tissue cancers of a particular type, nightmares &c) in the population (or animal or cell type) studied are (within a stated confidence limit) less than (a stated incidence). This inevitable state of affairs means that while investigators can use their best imagination and experience, they can never say that something is absolutely 'safe', and those who assert that an effect has been missed (or worse that something is being hidden) can always make any statement relating to safety look insecure by challenging the conditions and subject of the experiment, whether the right effects were looked for, and the sensitivity of the outcome. Worse, different groups are seen as having different interests in 'proving' that some situation is 'safe', so the credentials of the investigators and the source of their funds is seen to colour the result.

A well known technique is that of epidemiology. Take a sample population (say a million mobile phone users) and compare them for the incidence of some possible effect (say increased hair growth) with a control population (say a million people without mobile phones). It sounds easy, yet there are many problems. Is phone use the only difference between the populations? Obviously not; in particular we may find that the mean income of the non phone-users is lower than that of the phone users. Anyhow, what do we mean by a phone user? Is my wife, who only turns her phone on when she needs to make a call a 'phone user'? These confounding factors, working like noise in a communications system, reduce the sensitivity of the experiment and make the results fuzzy and imprecise.

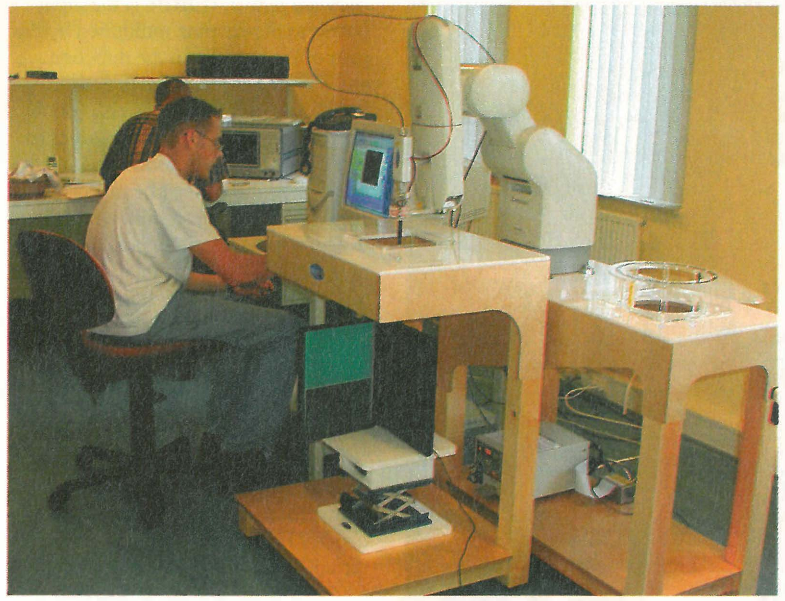


Figure 1: IndexSar measurement equipment

Even the existence of a study often creates concern in the popular press. The headline will always read *'Scientists seek (or fear) connection between mobile phones and x'*, and as we have seen, negative results will always look less than categorical.

As far as mobile phone usage is concerned, I believe that my probability of death from the effects of RF radiation from my phone are less than the probability that my phone will at some point save my life (and/or someone else's life).

Individual sensitivity

There's a huge difference between individuals' ability to sense exposure to RF fields. Some individuals are reported to be able to detect when a laboratory signal generator is turned on and off at power levels of only a few milliwatts, even when connected only to a coaxial cable with an open connector placed in front of the test subject. I was recently within 10m of an antenna radiating 600kW at about 1MHz and could sense it was on only because my feet became hot where I stood on a ground conductor. It's natural that people with enhanced sensitivity will be more concerned about any effects of RF fields, and their enhanced sensitivity to RF is often paralleled by enhanced sensitivity to other environmental stimuli. (Try Googling for 'Electromagnetic hypersensitivity' if you are interested.)

Specifying and measuring limits of exposure

There are various ways of measuring the level of exposure of individuals to RF fields. For waves propagating in free space the field is characterised by the power density (W/m^2). Close to an antenna, consideration must be given to separate criteria for the E-field (V/m) and H-field (A/m). If we wish to investigate the absorption of fields into the body we describe the level of absorption in watts per kilogram (W/kg) of body tissue; this quantity is referred to as the specific absorption rate (SAR). We may be interested in whole-body exposure, or more often the power density in each separate 10g or even 1g of tissue.

The body responsible for internationally specified limits is ICNIRP (International Committee on Non-Ionising Radiation Protection). National guidelines are laid down in many countries, often by health and safety administrations; these often coincide with ICNIRP, but sometimes differ – often in the direction of being less stringent. The competent authority in the UK is the National Radiological Protection Board (NRPB), while in the US the FCC, OSHA and IEEE are responsible for various aspects of regulation. The ICNIRP limits have been set with a large factor of safety below any known ill effects.

Calibrated hazard meters are now readily available and are routinely worn by riggers and engineering staff required to climb antenna structures. They usually operate over a wide frequency range and have an audible alarm that sounds if the safety limit is exceeded.

Measuring the absorption of energy in human tissue is difficult, especially as high spatial resolution is needed to make sure there are no small hot-spots which might suffer damage. Measurements are usually made by placing a phone close to a bath of liquid electrolyte with similar

Typical streetworks base station site

electrical properties to a human head and probing inside the electrolyte with a remotely-controlled probe. A typical measurement system with probe scanning inside a liquid-filled 'phantom head' is shown in **Figure 1**. It is now routine practice to measure the SAR distribution in this type of apparatus for every new model of mobile handset.

Effective radiated power

Most antennas concentrate the power they radiate into a beam which may be directional in the azimuth or elevation plane, or both. The *effective isotropic radiated power*, (EIRP), is the input power to the antenna multiplied by the gain of the antenna in the direction we are interested in, relative to an isotropic radiator [or in decibels, eirp (dBW) = Pin(dBW) + G(dBi)]. The larger an antenna, measured in wavelengths, the more directional it can be made. The concept of eirp is not valid when the distance from the antenna to the point of interest is small (certainly when it is less than the longest dimension of the antenna). Fields and power flows close to an antenna are best calculated using an electromagnetic simulation program or by measurement with a small probe.

The power density at a distance from an antenna is easily calculated. If an isotropic antenna radiates p watts, then at a distance r the power density is $p/(4\pi r^2)$ w/m^2, so for an eirp of P watts the power density is $P/(4\pi r^2)$ w/m^2 and the associated field strength $E(V/m) = \sqrt{(120\pi P)}$. These simple relationships apply to the far field of the antenna, which starts about 10m away from a typical base station antenna.

The effects of exposure to high fields

These are well documented. At very high levels of continuous exposure the body is heated and death will be caused by heat stroke when the body's cooling system is no longer able to cope. Pulsed fields can give rise to sensations such as pinging in the ears. Some tissues, notably the retina and the testes can be damaged at sub-lethal exposures, resulting in blindness or sterility. Eyes are particularly at risk from radiation in the upper microwave bands where most energy is deposited in surface tissues (such as the cornea) which have relatively low blood flow, and consequently are not efficiently cooled.

Exposure limits are set with the object of limiting the rise in body temperature to a small value within the normal limits of variation during light exercise. To give some sense of scale, the total power radiated by a GSM mobile phone typically has a peak power of around 1 watt, transmitted in short bursts for only 1/8 of the time for which the transmitter is active – a mean power less than 1/8 watt. Summer sunshine in the UK has a typical power flux of around 1000W/m^2, so when sitting on a beach on a sunny day we may be irradiated by 500W of electromagnetic energy, much of it deposited in the surface tissues in the form of heat.

Most standards recognise acceptable occupational exposure limits and set lower limits for the general public, a population which may include potentially more vulnerable individuals – for example children, old people with poor health, pregnant women – and whose exposure may be for much longer time periods than those of a typical worker.

RF exposure in the 21st century

Mobile phones

The currently most publicised source of exposure is the mobile phone and its associated base stations. The mobile phone is the first portable transmitter to be carried and used by a huge number of individuals, so it is natural that its safety should be of great concern. Unlike occupational exposure – generally experienced by healthy adults, mobile phones are used by every section of the community including groups who may be more sensitive to low levels of energy absorption, children, pregnant women, old people, and by people with circulatory problems, cancer and other pre-existing medical conditions. The peak SAR associated with a typical mobile phone is often between 50% and 90% of the ICNIRP limit when operating at maximum power. The actual power radiated by a mobile phone is automatically reduced to the minimum necessary to establish communication; the largest powers are required when the user is close to the limit of coverage of a base station.

Microwave ovens

The leakage of energy from a microwave oven is required to be less than 5mW/cm^2 at a distance of 5cm from any surface of the oven (BS 5175:1976). Even so the total power escaping may exceed 0.5 watt – say 1/1000th of the total RF power generated – representing a screening effectiveness of only -30dB. It's not surprising that there are stories of people putting mobile phones inside a microwave oven and successfully making calls to them! (You should only be able to do this if you live fairly close to a base station – and if you want to try, first make sure the oven is OFF.) The door seal will wear and become dirty through use, so the screening effectiveness is likely to fall as the oven ages. Even so the HSE circular to local authorities states 'The incidence rate … for injuries arising at microwave ovens … appears minimal and derives almost entirely from reports published in the USA.'

Mobile phone base stations

A base station is equipped with several transceivers, each generating up to about 40W. Allowing for the loss of signal through filters and cables the total power reaching the antennas is typically up to about 200W for each of the three sectors served by the base station. Some stations use two antennas for each sector and each will transmit up to 100W; where dual polar antennas are used a single antenna is used for each sector and transmits up to 200W. The associated power densities are the same. In order to provide the best possible coverage, the base antennas are built as a vertical array of radiating elements. This focuses the radiated energy into a narrow beam directed just below the horizontal; typically the beam is 6° wide and is directed 2° below the horizontal. On a typical low structure (10m high) the beam centre reaches head level 285m from the mast base.

If we do the calculation, we find that the power density in the example will be 0.013W/m^2, compared with an ICNIRP limit of 10W/m^2. Even allowing for sidelobes that lie below the main beam but are of lower gain, there is a substantial margin below the recommended limit. This is a pessimistic example as we have used the shortest structure combined with a power near the high limit of what would be encountered. When four networks are co-sited on a 20m structure we can see that the power densities at head height will still be well below recommended limits. Sites on building roofs can only create a hazard for people walking in front of them and close to their beam centre line – typically window cleaners and service personnel; to alert people to this hazard suitable fences and warning notices are usually provided.

So-called 'streetworks' base stations are mounted on structures like lamp posts, often only 8 – 10m high. These typically transmit 4 x 2W carriers, with an eirp of 16dBW. The gain of a typical base station antenna in the direction of the ground below the antenna is typically 30dB less than that in the maximum direction so there is no brightly lit area below the structure.

To put the potential hazard provided by a base station into perspective, I conducted the following experiment. A standard RF hazard meter with an alarm level set at one half of the ICNIRP limit was moved towards a standard GSM handset while a call was in progress. The hazard meter beeped when it was about 1cm from the antenna of the handset (the user's ear would normally be closer than this). The same hazard meter was then moved towards the front of a typical 12-element 1800MHz base station antenna fed with 2 x 20W signals; the meter beeped when it was 30cm (1ft) from the front of the antenna. The simple distribution of the total power over the physical area occupied by the antenna reduces the power density, even at the antenna itself, and as the power leaves the antenna it is progressively distributed over a larger area, typically the frontal area of a beam 60° wide in azimuth and 5° wide in the elevation plane.

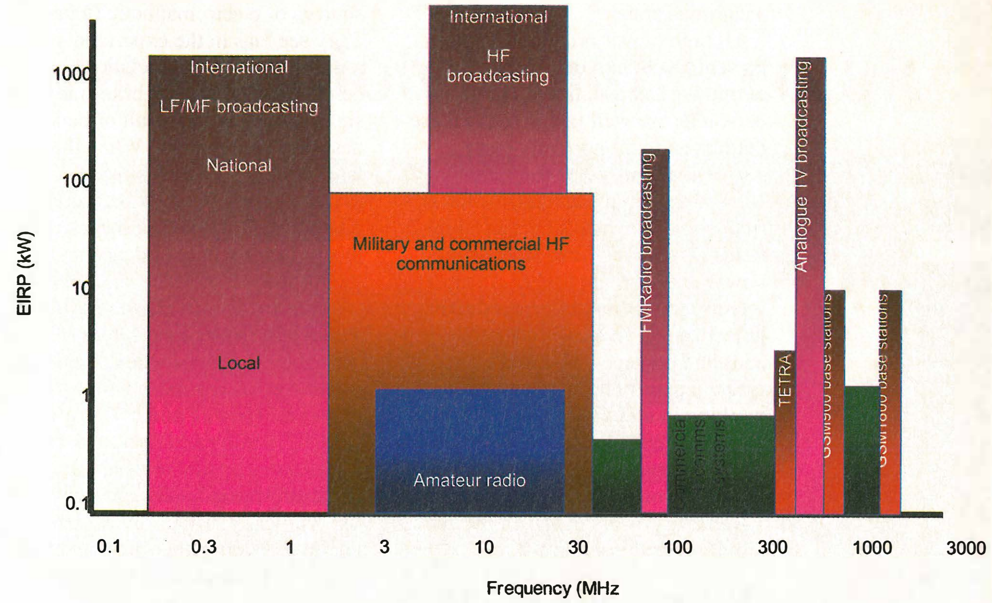


Figure 2: Simplified diagram showing typical range of EIRP radiated by major spectrum users

TETRA

Much publicity has been given to health concerns related to TETRA, the TErrestrial TRunked RAdio system being introduced for use by the police and other public services. The main source of concern relate to the relatively high power in use at the handsets, and the low frequency of the bursts they transmit. A report published in the 1980s suggested that the diffusion of sodium from brain cells was accelerated by exposure to RF signals with a pulse frequency around 16Hz. A large amount of more recent work has failed to confirm this effect and no health-related effects have been identified for exposures at or below the ICNIRP limits. Future developments in which handsets transmit in more than one burst in each frame (TETRA uses a frame with four bursts) might lead to users experiencing SARs at or above the ICNIRP limit, and this development may require additional precautions to meet the limit. TETRA base stations transmit a continuous signal and the associated eirps are generally lower than those used for GSM, so there are no special reasons for concern.

Radio and TV transmitters

The highest power densities are radiated by the main UHF television transmitters. A total eirp of 6MW is radiated by several TV stations in the UK, but the antennas are even more directional than those used for mobile radio, and the structures are typically 300m high, so the beam maximum is directed at the ground more than 8km from the structure (if the structure is on flat ground) and the ground level power density is very small. A classic epidemiological study was conducted a few years ago around the UHF TV station at Sutton Coldfield (UK).

VHF radio transmitters operate at rather lower eirps at frequencies around 100MHz. Although the ground level power densities are modest, this frequency band corresponds to a wavelength of 3m, so body resonances may increase the effect of exposure; this possibility is included in the ICNIRP limit which is correspondingly reduced.

Terrestrial digital radio and TV transmitters operate at much lower mean eirps, although their peak/mean ratio is relatively high.

High power radio systems at HF and lower frequencies create very high local field strengths at ground level and antenna systems are usually fenced to prevent access to hazardous zones.

All high power broadcast systems present hazards to those working within the radio stations. These personnel are well instructed in these matters and will normally carry personal alarms when they need to work near active antennas. At lower frequencies the heating effect of the fields is relatively low, so permitted power densities are higher. Typical effects are that any large conducting object carries induced currents/voltages and contact with a hand can create a painful burn; even when working in areas within safe limits personnel will usually wear insulating gloves while handling wire ropes and steelwork.

Relative powers of the all the above are illustrated in **Figure 2**.

The present state of knowledge

The Stuart Report (UK) is one of the most comprehensive and objective reviews of the possible effects of mobile radio systems. Some comprehensive epidemiological studies are beginning to report results and there are many on-going studies that will report over the next few years. Work on non-thermal effects is examining matters like the change in action potential in cells exposed to RF fields, and looking for effects on memory and perception. While various effects are reported it is not clear whether any has negative health significance.

The UK Independent Expert Group on Mobile Phones (IEGMP), chaired by Sir Richard Doll has reported even more recently. If you are seriously interested in the subject it is well worth reading. The research work reported spans investigations into the incidence of cancers (whether induced by RF fields operating alone or in combination with a number of other factors). The overwhelming impression on reading the document is of the wide scope and diligence of the work reported.

Are the published results biased?

Some people see much of the available information as being the result of a conspiracy not to tell 'the truth' about the perceived dangers of sources of electromagnetic fields. They see bias in the experts' dismissal of positive associations as being unrepeatable, of little statistical significance, or the result of badly designed experiments. When the language of the conclusions of an experimental campaign is careful, guarded and precise the sceptic sees this as lacking in confidence and hiding 'the truth'.

There are several reasons why the actual bias in results is likely to be in the conservative direction. Results reporting positive associations between exposure and health effects are newsworthy and bring publicity to the investigators responsible; grant allocating bodies are more likely to back further investigation of positive results rather than negative ('no effect') results. A newspaper will headline a positive result, but there are no headlines for solid research producing a conclusion of no adverse effects. This view is confirmed if you dip into the research literature: you may well recognise many of the positive findings, while the bulk of the negative reports are unfamiliar.

Reading the reports of many investigations suggests that there has been too little involvement by engineers with detailed knowledge of radio systems. There is little sign that the medics are being led by the nose by technical experts from the mobile radio industry. The strong impression is that many experiments would have been better designed – and more reliable results would have been obtained – if more advice had been taken from engineers before the investigations had begun.

The good news is that the more of the published literature you study, the more you realise just how much investigative work has been done and is in progress. While everyone is properly guarded about the possible emergence of long-term effects, the strong consensus is that no effects on health occur within the ICNIRP exposure limits. The 2003 report of the Independent Advisory Group on Non-Ionizing Radiation (Documents of NRBP Vol 14 No 2) concludes: "The weight of evidence now available does not suggest that there are adverse health effects from exposures to RF fields below guideline levels, but the published research on RF exposures and health has limitations, and mobile phones have only been in widespread use for a relatively short time. The possibility therefore remains open that there could be health effects from exposure to RF fields below guideline levels: hence continued research is needed."

The perception paradox

The 2002 report on the possible health effects of mobile phone systems to the French Senate comments on the fact that although most public protest relates to the siting of base stations, most people who oppose their construction use mobile phones and allow their children to use mobile phones, failing to admit that if a risk does exist it is in respect of handsets, which create far higher levels of personal exposure. As we have observed, mobiles used at the coverage limit of a cell radiate much more power than one near a base station, so the successful protesters enjoy a much higher level of RF exposure than those who live across the road from a typical base station.

Further information

There is a vast and fast-growing literature on the limits of RF exposure and the study of possible health effects. The author's website at www/bscassociates.co.uk/links.html carries links to all the documents mentioned in this article as well as to a number of other authoritative websites and sources of further references. The reports of the Independent Experts Group are written in language accessible to anyone with an interested lay person's knowledge of both radio and physiology and anyone interested in the subject will find they are well worth reading.

Anyone working with radio transmitters needs to be aware of the current regulations and to make sure staff working with live antennas are aware of potential hazards and carry hazard warning devices.

The author

Brian Collins has spent more than 40 years designing and building antennas for every application.

He currently runs his own consultancy business while retaining part time employment as Technical Director at CSA Ltd (base station antennas) and Chief Applications Engineer at Antenova Ltd (handset and other consumer antennas).

The Catt anomaly

Correspondence on the Catt anomaly has erupted again recently, and Ian Hickman was prompted to take a closer look. Here are some of his thoughts on the subject, a topic of which he has been vaguely aware for many years

I seem to remember reading an article by Ivor Catt in these pages, many years ago, considering what would happen when a sinewave signal travelled along a feeder, and encountered a termination at the far end, consisting of a perfect diode. This, of course, is quite a complicated scenario, since waveform distortion will occur, leading to the generation of harmonic frequencies. However, just recently, as a result of the reappearance of Ivor in these pages, I paid a visit to his website. You can find an animation demonstrating the Catt Anomaly (he prefers to call it the Catt Question or the E-M Question) at http://www.electromagnetism.demon.co.uk/catanoi.htm and it is in fact a very much simpler scenario than that mentioned above. It set me thinking, and some of these thoughts are set out below. As usual, my approach has been in terms of the fundamental properties of the circuit, rather than a mathematical description of the phenomena. I have always been surprised that so many engineers, including many of my colleagues when I worked in industry, seemed happy simply to turn the handle on the textbook maths and arrive at a working circuit, with no real understanding - or even curiosity - as to the details of what was actually going on, and just *how* the circuit in fact functioned. (Maths does not *explain* anything, it simply provides you with the tools to calculate the quantitative relationships between the variables. The maths itself is just arranged to mirror the facts, once they have been understood and explained.)

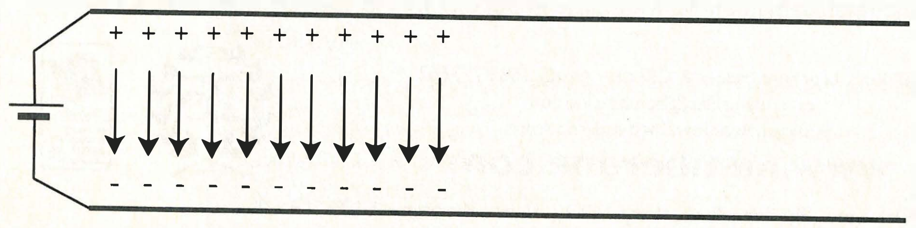


Figure 1: Showing a voltage step function propagating along a two-wire line.

Vector diagrams, Argand diagrams, Nyquist diagrams, Bode plots and the like can provide an insight into the functioning of a circuit, which an equation does not, at least not to the ordinary mortal. An example is the functioning of a second order phase lock loop: the treatment in my Radio Frequency Handbook[1] uses graphical methods to elucidate the workings, arriving at the result without resort to any complicated mathematics. In the same way, the analysis below is based firmly on the physical properties, the relationships between voltage, current, charge, capacitance, inductance, power, energy etc. in the circuit.

The Catt Anomaly

Figure 1 shows a battery connected to a two wire transmission line of finite length, and we may assume, from the position that the TEM wavefront has reached, that the battery was connected a (very!) short while previously. On the part of the line which has already been charged by the source, positive charge appears on the top conductor, and negative charge on the bottom conductor. Electric field lines are indicated, originating on the positive charges and terminating on the negative charges. The electric field strength, in volts per metre, will be equal to the potential difference between the upper and lower conductors, divided by the distance between them. At least, this is the case for the field lines shown, in this two dimensional representation. There will also be longer such lines, starting on the upper conductor and terminating on the lower, in the same plane as the lines shown, but bulging out in front of the paper, and likewise behind. There are even notional lines, of negligible importance, heading off vertically from the top conductor towards infinity, and likewise downwards from the lower, assuming the two conductors are located somewhere in free space. The electric force results in an electric flux, of strength depending upon the permitivity of the medium in which the line is immersed. This is analogous to the way magnetic field strength H results in a magnetic flux density B, depending upon the permeability of the medium. For simplicity, let's assume the line is in air, or even in vacuo.

The Catt Anomaly, as I understand it, lies in the question as to where the charge at any point on the surface of the lower conductor comes from, bearing in mind that its greatest concentration will be where the lines of electric force landing on it are most concentrated, i.e. directly underneath the upper conductor. Ivor points out that it cannot be delivered by the displacement current flowing between the lines, as displacement current is not a flow of real current[2].

Ivor says on his website that he has asked well known academics where the negative charge on the lower conductor comes from. He divides those who could be induced to give any answer at all, into "Westerners" and "Southerners". The former say the charge comes from the source at the left-hand end of the line, and the latter, from within the line itself at that point. But, he maintains, it cannot come from the left-hand end as that would suppose that it travels at the speed of light[3]. But the negative charge consists of electrons, each of which has a small but finite mass

(9.109×10^{-31}kg), and therefore cannot travel at the speed of light.

The Southerners' explanation entails charge moving from the inside of the material of the line, to the surface i.e. at right-angles to the direction of propagation, which he also he also finds problematic. On the face of it, these two positions seem to be irreconcilable, and perhaps for this reason, Ivor assumes they are both wrong. On the contrary, I maintain, they are in fact both right.

From the outset, I was puzzled by Ivor's fixation on the negative charge on the lower conductor. Surely any complete explanation must account equally for the positive charge on the upper conductor. So that became my point of departure.

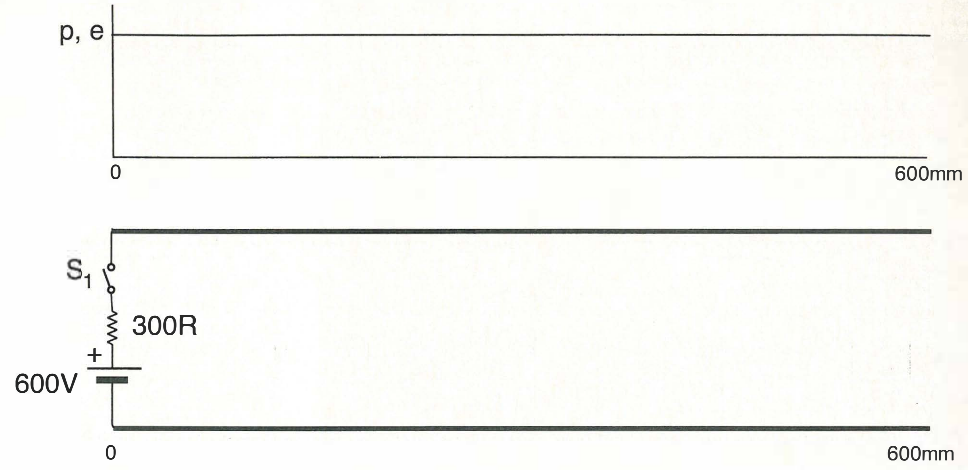


Figure 2: A 300Ω open-circuit line, about to be connected to a matched source.

A different view

Consider the line before the battery is connected. The metal - let's assume it is 100% pure copper - consists of atoms which are in "fixed" positions, the material being a solid. Although fixed, they are in fact each vibrating about some mean position, assuming the temperature is not absolute zero. Nevertheless, for the purpose of this exercise, that is immaterial; they cannot actually wander around, and they will therefore be regarded as stationary. This applies to the protons and neutrons in the nucleus and also to most of the electrons, in each atom.

But there are also "free electrons" forming a sort of "conducting gas" within the conductor. These may be on a somewhat freer rein, but they too, for the purposes of this argument, can be considered as stationary. Thus at every point along the line there are equal numbers of protons and electrons per unit length. This has been plotted in a graph at the top of **Figure 2**, where the x axis is length, the same as the line itself. Before the switch is closed, the equal number of protons and electrons per unit length is indicated by the black line labelled "p, e". Thus there is no net charge at any point on either conductor, and the potential difference between them is everywhere zero.

An atom is not proprietorial about its free electron: if the latter departs to the left as part of an electric current, the positive charge on the nucleus is happy to let it go, provided that another electron comes along from the right to take its place, which will always be the case in a complete circuit.

Now consider the case when the switch is closed, **Figure 3**. Conventional current flows into the top conductor of the line, but in fact a conventional current flowing to the right actually consists purely of electrons moving to the left (at least, in a metallic conductor, which - unlike a semiconductor - has no mobile positive charges or "holes"). Assuming the velocity of the wavefront on the line is the same as the speed of light (for an open air-line, it won't be that much less), then after half a nanosecond it will have reached the point indicated by the line A A', and after a further half nanosecond it will have reached B B', as indicated in **Figure 3**. The instant the switch is closed, the leftmost electron in the upper conductor (the "first" electron) will start to move, eventually passing via the switch and into the matched source, the internal resistance and emf of which have been shown separately. The second electron now finds the third electron closer to it than the first. As like charges repel each other the more, the closer they are together, the second electron experiences a net force pushing it to the left and starts to move leftwards also. Now, the third electron finds the fourth electron closer to it than the second and also starts to move - and so on. The electrons may be moving at a snail's pace relative to the speed of light, but the *disturbance* just described propagates along the line at the speed of light, reaching B B' in just a nanosecond.

To the right of B B' there are still equal numbers of protons and electrons per unit length of line, so there is no net charge and the voltage on the line there is as yet zero. To the left of B B' a line of electrons is moving to the left at a constant speed, the spacing between the moving electrons being everywhere equal, $x + \delta$ say; slightly greater than to the right of B B'. Therefore there is everywhere a slight deficit of electrons per unit length of the upper conductor, their number being represented by the green line "e" in **Figure 3** (not to scale). The resultant constant net positive charge per unit length is responsible for the constant positive potential on the line, relative to the lower conductor, indicated in **Figure 3** by the red line labelled "V". With a line of electrons all moving at the same speed and with a constant spacing between them, the current in the line to the left of B B' is everywhere constant, indicated by the blue line labelled "I" in **Figure 3**. The ratio of V to I gives the "characteristic impedance" or "surge impedance" of the line, which I will assume to be 300Ω, a typical value for a two-wire air-line, though it could be anywhere from about a third to five times that value, depending upon the thickness of the conductors, their spacing and the dielectric separating them.

Exactly the same mechanism which has been used to account for the appearance of a positive charge on the upper conductor, accounts equally well for the appearance of a negative charge on the lower conductor. Only now, as conventional current flows from the source into the upper conductor, it returns from the lower conductor, into the negative pole of the source. This implies that the negative pole forces electrons into the left-hand end of the lower line. What was the first electron there now finds one closer to it on its left than the second electron on its right and - I won't bore you with a blow by blow account again, but to the left of B B' the electrons are slightly closer together ($x - \delta$) than to the right, all equally spaced and moving rightwards at a constant speed. They thus constitute a conventional negative current flow to the left, indicated by the blue line below the baseline in the lower graph in **Figure 3**. This is basically the Westerner view advocated by Dr. Neil MacEwan and similar to the

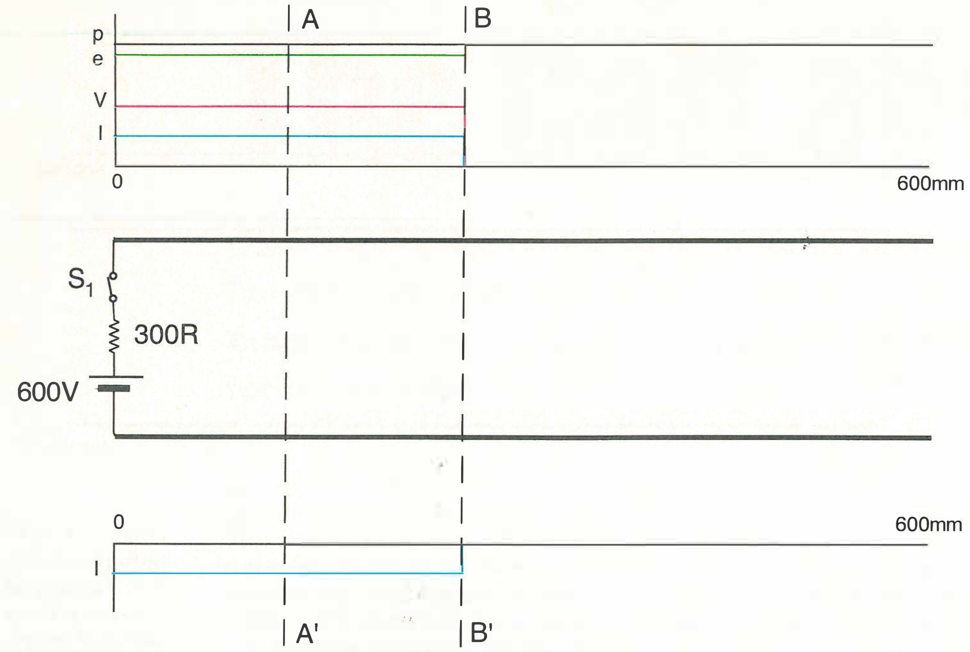


Figure 3: The position 1ns after switch closure.

analogy put forward by Dr. J. W. Mink of the IEEE. He points to the analogy of a droplet of water entering one end of a pipe, promptly forcing a drop out of the other end, but not of course the same drop.

Although I have described events on the upper and lower conductors separately, the boundary between the charged and uncharged sections of the line proceeds at the same rate on both conductors, and indeed this common boundary *is* the wavefront. Also, the foregoing might seem to imply a neat single line of free electrons, rather like peas in a peashooter. But the current of 1A flowing in the line consists of the passage past any point of a charge of one Coulomb (1C) per second. The charge on an electron is 1.602×10^{-19}C, so there are 6.242×10^{18} electrons entering the lower and leaving the upper conductor per second. The speed with which the electrons are moving depends on how many of them are involved in carrying the current. In a line composed of very thin conductors, the current will presumably be carried by fewer electrons, travelling faster, than in one with very thick conductors. In either case, the velocity of the electrons is very much less than that of light. Assuming the same conductor spacing, these two lines would clearly have very different characteristic impedances, so the applied voltage needed to cause a current of 1A to flow would differ greatly. Nevertheless, the mechanism of propagation of the wavefront on the line is as described; but many electrons are involved rather than just the "first", "second" etc. I can see no flaw in the Westerners' explanation of how the negative charge appears on the lower conductor, resulting from the slight bunching up of the free electrons due to those entering the line. Professor Pepper's contribution is to point out that the negative charge at any point on the lower conductor, past which the wavefront has travelled, will distribute itself over the circumference of that conductor, in proportion to the density of lines of electric flux terminating at each point around the circumference. MacEwan explains how the charge got there in the first place; Pepper explains, given that it is there, how it distributes itself over the surface of the conductor, as a result of the electric field between them. We must assume that the deficit of electrons in the upper conductor (the positive charge, due to nuclei short of one attendant electron) distributes itself on the *surface*, around the circumference, in a mirror image of the distribution of the additional electrons on the lower conductor, there being no deficit of electrons *within* the upper conductor.

During the two nanoseconds the wavefront takes to traverse the line depicted in **Figure 3**, the source supplies 300nJ/ns of energy to the line - delivered by a current of 1A - and "thinks" it is connected to a 300Ω resistor. If the length of the ideal loss-free line is infinite, the source will continue to "see" a 300Ω resistor indefinitely, and similarly if the end of the line in **Figure 3** is terminated in a real 300Ω resistor. But the 300Wdissipation in the resistor will not commence until 2ns after the switch is closed. During that time, 600nJ of energy is stored in the electric and magnetic fields of the line, and 300W will continue to be dissipated in the resistor for 2ns after the switch is subsequently opened.

It is worth considering in detail what happens at 2ns, when the wavefront reaches the open circuit end of the line in **Figure 3**. The conventional "explanation" is that an open circuit reflects the voltage in-phase and the current out of phase, while a short circuit reflects the voltage out of phase and the current in-phase. A more complete analysis follows. At time t = 2ns the source has delivered 600nJoules of energy to the line (up to this instant the termination, if any, is immaterial).

So the source has supplied 2nCoulombs of charge to the line.

There is a uniform PD of 300 Volts along the line.

Q = CV, therefore 2nCoulombs = C × 300 where C is the capacitance between the conductors. (It's just unfortunate that C stands both for capacitance and Coulomb.)

Therefore capacitance of line = $2 \times 10^{-9} / 300 = 6.667 \times 10^{-12}$ Farads.

Energy stored = $\frac{1}{2}CV^2 = \frac{1}{2} \times 6.667 \times 10^{-12} \times 300^2$ = 300n Joules.

This is half the energy put in, so the rest must be in the magnetic field.

Energy stored = $\frac{1}{2}LI^2$.

Therefore $\frac{1}{2}L \times 1^2$ = 300n Joules.

Therefore L= 600n Henries.

It may seem strange to talk of the inductance of a 600mm line which is open circuit, but it is quite logical if attention is restricted purely to the instant t = 2ns, when there is current flowing in the whole length of each conductor.

Now for a lossless line, as assumed,

Line impedance = $(L/C)^{0.5} = (600 \times 10^{-9} / 6.667 \times 10^{-12})^{0.5} = 300\Omega$

which all seems to tie up. Half of the total energy supplied is stored in the electric field, and half in the magnetic.

In the open circuit case, *after* t = 2ns, what actually happens is this:-

The electron at the extreme right-hand end of the lower conductor (the "last" electron) cannot move substantially to the right; it is at an open circuit. But the potential between the conductors at the input is still 300V, so the source continues to feed electrons into the left-hand end of the line. The "train" of electrons thus continues on its route, forcing the "second but last" electron, closer to the stationary last electron. When the spacing between them falls from $(x - \delta)$ to $(x - 2\delta)$, the force of repulsion between them becomes so

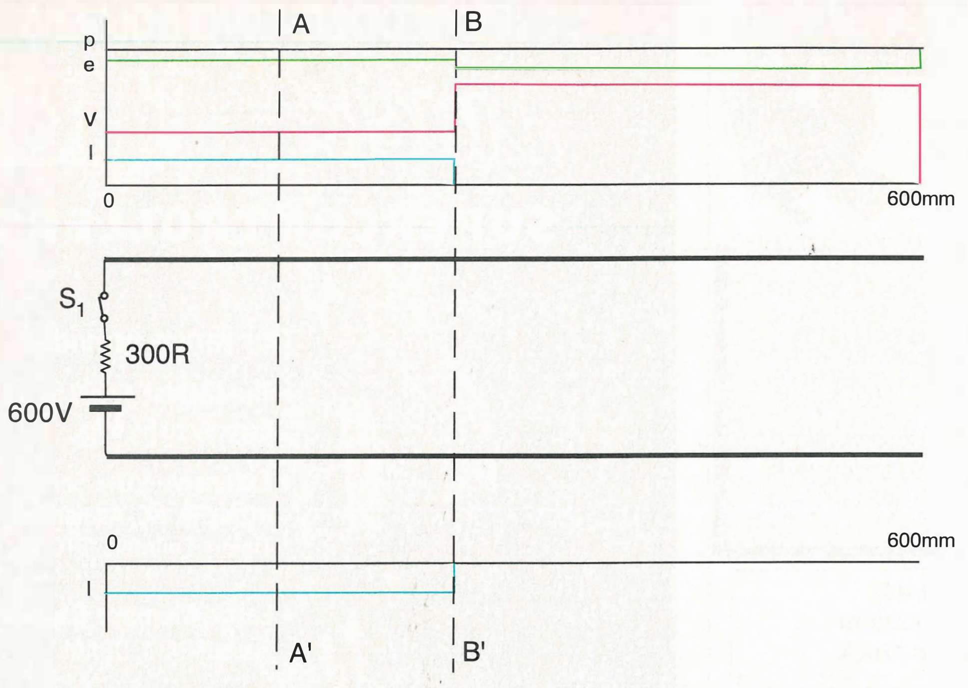


Figure 4: The position 3ns after switch closure.

great as to force the second but last electron to a halt. Successively, the third but last and other electrons also grind to a halt, so the current at the right-hand end of the line is zero. The boundary between the moving and stationary electrons propagates to the left at the speed of light, even though the speed of movement of individual electrons is comparatively very slow.

Simultaneously, when the right-most electron in the upper conductor starts to move to the left, becoming the last "truck" in a "goods train" of electrons steaming to the left, the nucleus of the right-most atom finds no electron arriving from the right to replace it. It is therefore unwilling to relinquish its electron entirely. The electron truck is therefore promptly uncoupled again from the train. The next atom in from the left is similarly unwilling to relinquish its electron, since no replacement is available from the right. The second from right electron is thus also uncoupled from the train, but the spacing between these two now stationary electrons is $(x + 2\delta)$. The train continues to the left, surrendering 1nC/ns of charge to the source, but getting steadily shorter as more and more electron trucks are abandoned on the rails, all at a spacing of $(x + 2\delta)$, until after three nanoseconds, the situation is as depicted in **Figure 4**. Whereas electron spacing of $(x \pm \delta)$ on the conductors equated to a potential difference between them of 300V, $(x \pm 2\delta)$ equates to a potential difference 600V.

To the right of B B', the deficit of electrons per unit length of the upper conductor is twice as great as to the left of B B', indicated by the green line labelled e on the graph.

Therefore the potential on the upper conductor, relative to the lower, is now 600V, as per the red line labelled V, while the current on the line to the right of B B' is zero, since the electrons there are stationary. At four nanoseconds after switch S1 was closed, the length of the "goods train" has reduced to zero, all the "trucks" having been abandoned at a spacing of $(x + 2\delta)$, the electrons on the lower line are all at $(x - 2\delta)$ and the potential on the line is everywhere 600V.

There is thus no potential difference between the two ends of the 300Ω source resistor, so the flow of conventional current being supplied by the source EMF ceases abruptly, leaving as many extra electrons per unit length on the lower conductor as the deficit of electrons per unit length on the upper conductor.

During the period t = 2ns to t = 4ns, the length of line in which current is flowing reduces steadily from 600mm to zero, and thus the *effective* inductance (storing energy in a magnetic field) reduces similarly from 600nH to zero, converting the 300nJ of energy stored in the magnetic field to energy stored in the electric field. Together with the 600nJ supplied by the source during this period, this brings the total energy stored in the line to 1200nJ, finally all stored in the electric field.

The apparent flow of conventional current from the upper to the lower conductor is what is meant by displacement current, a concept introduced by Maxwell to retain consistency with Kirchoff's First Law, in cases where there is no obvious circuit. Note that the notional displacement current only flows where the electric field strength (voltage gradient) is changing. In **Figure 3**, at one nanosecond after the switch S1 is closed, the voltage on the line to the left of B B' is everywhere constant at 300V, and to the right of B B' is everywhere zero. So the displacement current of 1A flows only at the point where the wavefront is. This implies that if the voltage step is instantaneous, the length of conductors between which the displacement current flows is infinitesimal, and the displacement current density therefore infinite. This is a major difficulty, and one reason why the concept of displacement current can be consigned to history.

Is there a crucial difference between the 300Ω characteristic impedance of the line, and a capacitor, or a 300Ω resistor? At twice the voltage, a capacitor stores *four* times as much energy, and a resistor dissipates energy at *four* times the rate. In the case of the open circuit line, at 2ns, the energy stored at 300V is 600nJ, and after 4ns, at twice the voltage, the stored energy only *doubles* to 1200nJ. But after 2ns, only 300nJ is stored in the capacitance of the line, the other half of the energy supplied being stored in its inductance. Thus the energy stored in the line's *capacitance* quadruples when the voltage doubles to 600V. During the four nanoseconds, 1200nJ has been dissipated in the internal resistance of the source, so the total energy supplied is 2400nJ.

At four nanoseconds, the energy stored in the line is $\frac{1}{2}CV^2 = \frac{1}{2} \times 6.667 \times 10^{-12} \times 600^2 = 1200nJ$, exactly the same as for a discrete capacitor of the same capacitance as the line. The crucial difference is that if the discrete capacitor is connected to a 300Ω resistor, it will deliver an initial 2A, decaying exponentially whereas the charged line will deliver a constant 1A for 4ns, ceasing abruptly thereafter. This is why a delay line rather than a capacitor is used to supply anode power to a magnetron, to form a pulse flat-top pulse in a radar transmitter.

Conclusions

An explanation of the appearance of negative charge on the lower conductor of a line, on which a Transverse Electric Magnetic wavefront from a matched voltage source, is propagating, has been presented. This explanation has been extended to the case of an open circuit line of finite length. An explanation, in similar terms, of what happens when a TEM wave from a matched ideal current source propagates along an ideal lossless short-circuited line, can also be developed. However, the situation looks, at first sight, a little more complicated, and I am still thinking about it. So, in common with all the best textbooks, I can only say that this question "is left as an exercise for the reader". The explanation of the appearance of charge on the lower conductor here given, has been presented not, indeed, entirely without mathematics, but with no more than simple arithmetic, and appeal to universally known relationships in the science of electricity.

The discussion above applies in the case of an ideal lossless line, but applies in broad principle to practical lines. However, there are two unwanted characteristics of real lines which have been ignored: attenuation and dispersion. These are interesting, important and well worth studying and becoming familiar with. But there is enough meat in this article for now, so I hope to look at these topics later. For the present then, here endeth the first lesson.

Acknowledgment: The author gratefully acknowledges helpful discussion and comment in the preparation of this article from his old colleague and longtime friend, M. H. Gross, C.Eng., MIEE.

References:

1 *Newnes Practical RF Handbook*, 3rd Edition 2002, ISBN 0 7506 5369 8, Butterworth-Heinemann

2 Displacement current, and Maxwell's Equations generally, seem to have fallen into disfavour since I was at college: at one company, where I was working as a contract engineer on the development of epirbs (emergency position-indicating radio beacons), the manager to whom I reported (younger than me) stoutly maintained that he had never heard of displacement current, and didn't believe such a thing existed. In a sense he is right of course; it certainly does not account for the negative charge on the lower conductor. Displacement current is a notional flow which is used to account for an apparent flow of current where there is no conducting circuit (e.g. "through" a capacitor when the voltage gradient in the dielectric changes). Displacement current is proportional to the rate of change of the electric field strength D in volts/m:- displacement current = k dD/dt in the one dimensional case, it's rather more complicated in the general three dimensional case - see Maxwell's Equations.

3 (or very nearly at the speed of light, in an open air-line. In a coax the speed is only about two thirds that of light, in a delay cable used to feed the Y plates of an oscilloscope, it is much lower still, while in a loaded telephone cable it is only about one twentieth of the speed of light).

Fact: most circuit ideas sent to *Electronics World* get published

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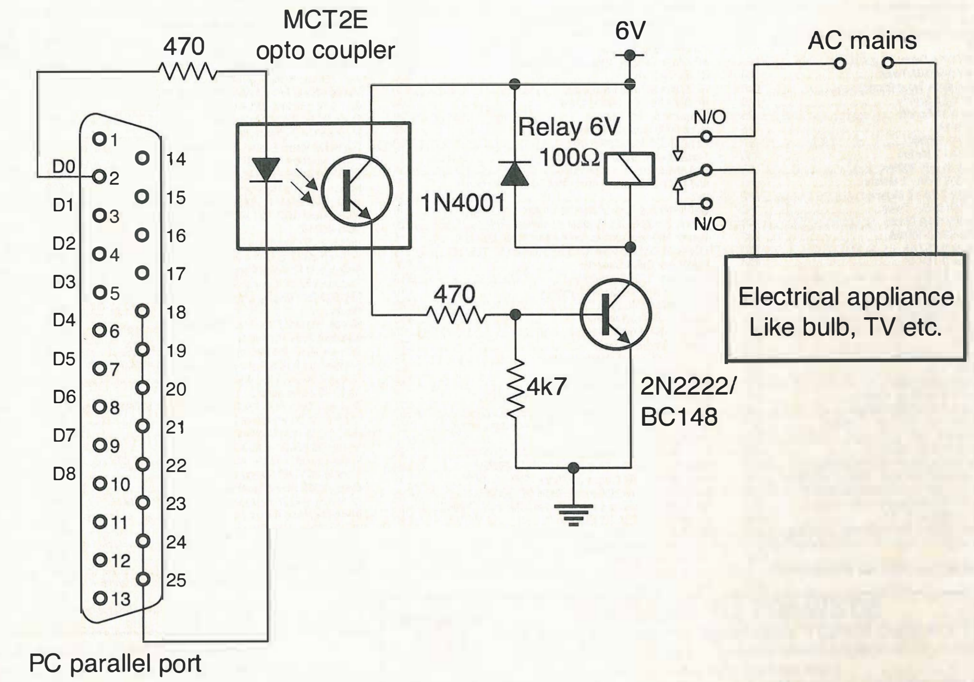
Controlling electrical appliances using PC

Here is a circuit for using the printer port of a PC, for control applications using software and some interface hardware. The interface circuit along with the given software can be used with the printer port of any PC for controlling up to eight pieces of equipment.

The interface circuit shown in the figure is drawn for only one device, being controlled by D0 bit at pin 2 of the 25-pin parallel port. Identical circuits for the remaining data bits D1 through D7 (available at pins 3 through 9) have to be similarly wired. The use of an optocoupler ensures complete isolation of the PC from the relay driver circuitry.

Lots of ways to control the hardware can be implemented using software. In C/C++ one can use the outportb (portno, value) function where portno is the parallel port address (usually 378 hex for LPT1) and 'value' is the data that is to be sent to the port. For a value=0 all the outputs (D0-D7) are off. For value=1 D0 is ON, value=2 D1 is ON, value=4, D2 is ON and son on e.g. If value =29 (decimal)=00011101(binary)> D0, D2,D3, D4 are ON and the rest are OFF.

Raj K. Gorkhali
Kathmandu
Nepal



Current-mode oscillator

Sinusoidal Oscillators play a very important role in many areas including instrumentation, measurements, communication, control systems etc. A great deal of research work has been carried out on the OTA – based sinusoidal oscillators and as a result of these efforts a good number of oscillators have been reported. Almost all the oscillators are voltage mode and it seems that very little work has been carried out on current mode oscillators, which enjoy certain advantageous features over their Voltage-Mode counterparts. Although these oscillators have several salient features yet they face problems like presence of condition of oscillation to be satisfied, constraints to be imposed on the values of transconductances and passive components, besides having cumbersome design procedure. Some of the oscillator circuits apart from employing excessive number of components use floating capacitors, which are not economically fabricated in IC construction. Moreover from the most of the circuits reported earlier, quadrature type oscillators cannot be obtained through use of additional components or inducing change in the circuit topology or nature of components.

This communication is concerned with the introduction of oscillator circuit that is suitable for implementation in IC form. The main feature of proposed oscillator circuit is that it is free from the preset condition of oscillation besides its frequency of oscillation is electronically tuneable and its variation linear over a wide range. The use of two OTAs is essentially required to achieve linearity in the frequency of oscillation. The circuit has been made to generate high frequency sinusoids by using single pole model of Operational Amplifier, the use of which has eliminated the requirement of second capacitor that is a minimum to be employed in a second order oscillator function. With the introduction of two CFs the current signals are obtainable at high impedance levels and are also out of phase by 90°.

A routine analysis of the proposed circuit of **Figure 1**, assuming that OA is characterised by its single pole model at higher operating frequency yields the following equation

$$s^2 + \frac{(RBg_1g_2)}{C} = 0 \qquad (1)$$

where B= A0w0 is the gain band width product of OA.

From (1) it is clear that the oscillator is free from the preset condition of oscillation. The frequency of oscillation is given by

$$f_0 = \frac{1}{2\pi}\sqrt{\frac{RBg_1g_2}{C}} \qquad (2)$$

For gl = g2 = g fo is given by

$$f_0 = \frac{g}{2\pi}\sqrt{\frac{RB}{C}} \qquad (3)$$

Since g = IB/2VT the frequency of oscillation is given by

$$f_o = \frac{I_B}{4\pi VT}\sqrt{\frac{RB}{C}} \qquad (4)$$

As is evident from (4) that the frequency of oscillation is linearly tuneable through bias current of OTAs.

The quadrature sinusoidal output signals can be obtained through use of two CFs as shown in Fig. 2. We can sense out the current I(R) in the resistor R, which is in quadrature with the current I(C) through capacitor C.

SENSITIVITY:- The active and passive sensitivity of this oscillator are,

$$S_R^{\omega_o} = S_B^{\omega_o} = S_{g_1}^{\omega_o} = S_{g_2}^{\omega_o} = -S_C^{\omega_o} = 0.5$$

which are very low.

N. A. Shah and S. Z. Iqbal
Srinagar
India

Figure 1: Proposed oscillator configuration.

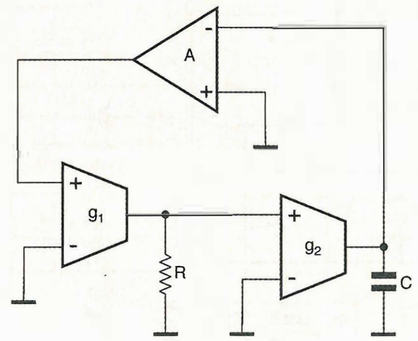
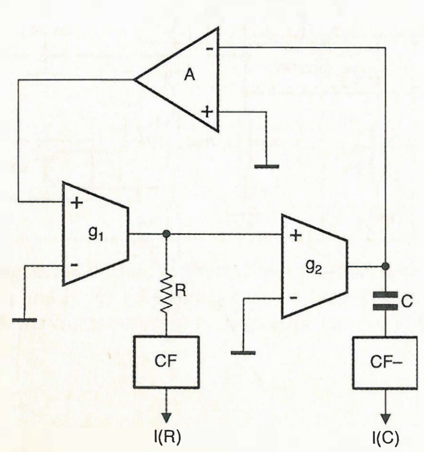


Figure 2: Current-Mode oscillator configuration.



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4. C.C. CHEN, H.W. TSAO and S.I.LIU, Current- Mode Oscillator using single Current Follower, *ELECTRONICS LETTERS Vol. 27*. pp 2056-2059 October 1991.
5. M.T. Abuelma'atti and H.A. Al- Zaher, Current- Mode Oscillator Using Single FTFN, *IEEE Transaction on CAS-II Vol. 46*. pp. 69-77. January 1999.

A simple and low cost 1000V high voltage driver

High voltage drivers play an important role for driving Piezo devices, electro-optic components, PMT and spectrometry, etc. A simple and low cost 1000V high voltage driver is shown below in Figure 1. We adopted off-line current mode control schemes and a flyback switching power supply design. The UC3844 is a major control part and the operating frequency is set at 100kHz. The UC3844 provides frequency modulation to reduce the switching frequency in the light load and no load conditions.

The feedback voltage, which is derived from the output of error amplifier, is taken as the indicator for load conditions. Once, the feedback voltage is lower than the green-mode threshold voltage, the switching frequency will start to reduce. Indeed, all of the power losses are in direct proportion to the switching frequency, such as the switching loss of the transistor, the core loss of the transformer and inductors, the power loss of the snubber etc.

The frequency modulation in the PWM controller can reduce the power consumption of the power supply in light load and no load conditions. The PWM operation in normal load and high load conditions are normal and not affected by the frequency modulation. Pin 8 (FB) of the UC3844 sums the current sense signal, the output voltage feedback signal and any added slope compensation.

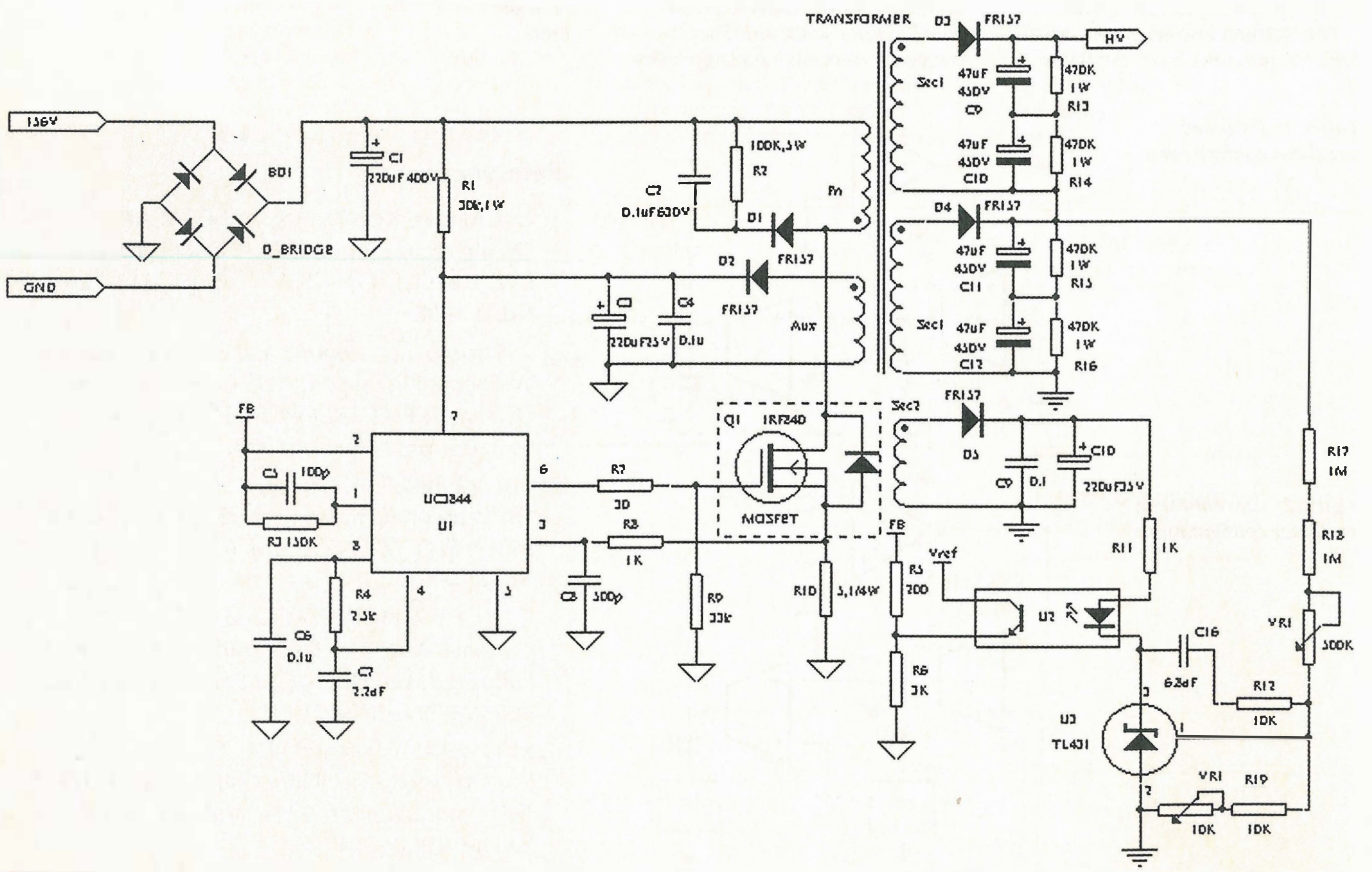
Referring to the feedback control circuit, we used a TL431 adjustable shunt voltage regulator to detect the output signal pass through PC817opto-isolator to feedback control FB pin of UC3844. TL431 acts as an open-loop error amplifier with a 2.5V temperature compensation reference. The output voltage may be adjusted to any value between Vref and 36 volts. When the voltage output is lower than the required level, it will automatically adjust the voltage apply to the UC3844 to compensate the pulse width modulation of the output triggering signal.

Ceramic capacitor bypass capacitors (0.1mf) from Vcc and Vref to ground will provide low-impedance paths for high frequency transients. We used the Tomita made transformer having an EI25-2E6 core set. To prevent core saturation, the gap is about 1mm. The primary winding is 70 turns and the wire gauge no. 28. Both the secondary windings are 105 turns and the wire guage is no.34.

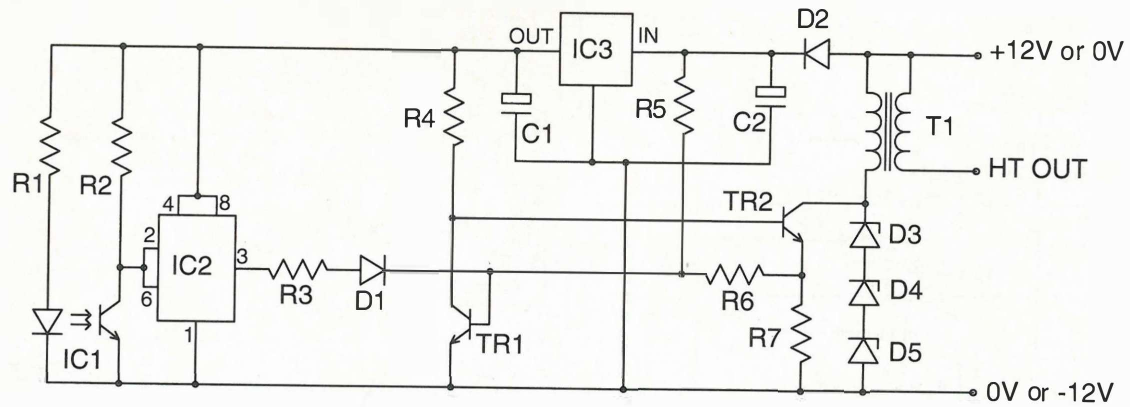
Both primary. and secondly auxiliary windings are five turns and six turns, respectively. These are also wire guage 34.

The DC output voltage of **Figure 1** is fixed at 1000V and the output current is 10mA. The input voltage is 110VAC. The DC output voltage could be set in a 50V range by adjusting VR1. The line regulation is less than 1%. The load regulation is also less than 1%. The power efficiency is better than 80% at full load.

Prasit Champa, Prathumthani Thailand and Tai-Shan Liao Hsinchu
Taiwan
Republic of China



Contactless electronic ignition



R1 = 150 ¼W; **R2** = 10k ¼W; **R3** = 1k ¼W; **R4** = 10 1W; **R5** = 15k ¼W; **R6** = 220 ¼W; **R7** = 0.1 5W; **D1** = 1N4148; **D2** = 1N4004; **D3,4,5** = 1N5388B; **TR1** = ZTX618; **TR2** = BU2722; **C1** = 470µF 25V; **C2** = 10µF 16V; **IC1** = EE-SG3 (Omron); **IC2** = ICM7555 **IC3** = LM7805; **T1** = #11220 (Intermotor).

Not so long ago the pages of electronics magazines were full of automotive circuits, but now there is so much electronics already under the bonnet that there is little left to add.

This circuit would be of interest to anyone interested in older cars fitted with the standard coil and contact breaker 'Kettering' ignition circuit . Such cars are often left in the garage for weeks on end and a circuit that makes them easier to start can be well worth having. An advantage of this circuit is that it can be used with both positive and negative earth vehicles without any changes. A little ingenuity is required to construct the optical sensor. The Omron part was chosen as it has fixing holes arranged so that the slot is horizontal. A rotating vane is attached to the shaft of the distributor which interrupts the light passing through IC_1 (I cheated and bought one from the Luminition web site), IC_1 is connected to the rest of the circuit using extra-flexible wire - the base plate of most distributors moves with the vacuum advance.

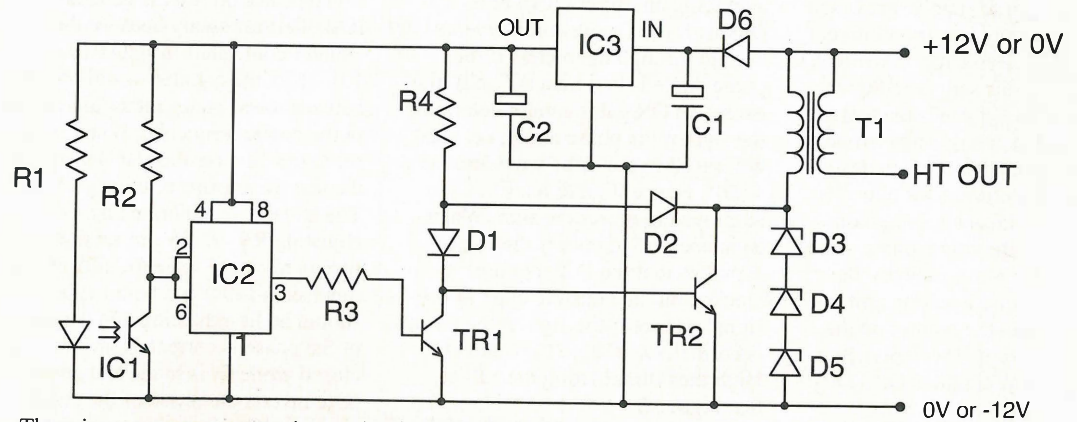
The circuit is a basic 'flyback' power supply: with the vane blocking the beam in IC1, base current is supplied to TR_2 by R_4. It is well known that both high-speed transistors and high voltage transistors have low gain. TR_2 is a close relative of the well-known BU208 horizontal deflection transistor and is both high speed and high voltage, and its H_{fe} is truly miserable, worst case being as low as eight, so it needs half an amp of base current to get 4A through the primary of T_1. Current is sensed by R_7, and TR_1 removes TR_2's base current to keep T_1's primary current at 4A.

When the vane is removed from the beam of IC_1, TR_1 is turned fully on, removing the base drive from TR_2 and any stored charge. The magnetic field in T_1 begins to reduce, and in doing so produces a voltage of opposite polarity. The secondary voltage keeps rising until it is high enough to arc across the sparking-plug gap. D_3,D_4 and D_5 clamp the primary flyback voltage to 600V, which allows a maximum secondary voltage of 39kV.

The capacitor (condenser) normally connected across the points should *not* be left connected across TR_2.

An advantage of inductive energy storage over capacitive is that should the arc extinguish whilst there is still energy stored in the coil, the output voltage will increase again until the arc re-strikes. At low engine speeds the vane moves across the light source quite slowly, and IC_2 a CMOS 555 timer, used in its most simple form as a Schmitt trigger to sharpen up the edge.

R_5 is deliberately returned to the unregulated supply. This increases the coil current (and thus the energy stored) at low supply voltages - in other words whilst the starter motor is running. T_1 is the type of coil that is normally used with a ballast resistor (it is not required in this circuit) - it has a primary resistance of less than 1Ω, and a primary inductance of about 10mH. If it is wished to retain an existing 'self-ballasted' coil, which has a primary resistance of 3-4Ω, then the following circuit should be used.



R1 = 150 ¼W; **R2** = 10k ¼W; **R3** = 1k ¼W; **R4** = 10 1W; **C1** = 470µF 25V; **C2** = 10µF 16V; **TR1** = ZTX618; **TR2** = BU2722; **D1,D2** = UF4007; **D3,D4,D5** = 1N5388B; **D6** = 1N4004; **T1** = Ignition coil; **C1** = EE-SG3 (Omron); **IC2** = ICM7555; **C3** = LM7805

There is now no current sensing as the current is limited by the series resistance and it is necessary to add an anti-saturation clamp consisting of D_1 and D_2 to TR_2 to make sure it turns off quickly. Unfortunately, spark energy is now lower with the starter running, as maximum current is simply the battery voltage divided by the coil primary resistance.

Ian Benton
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Three phase meter

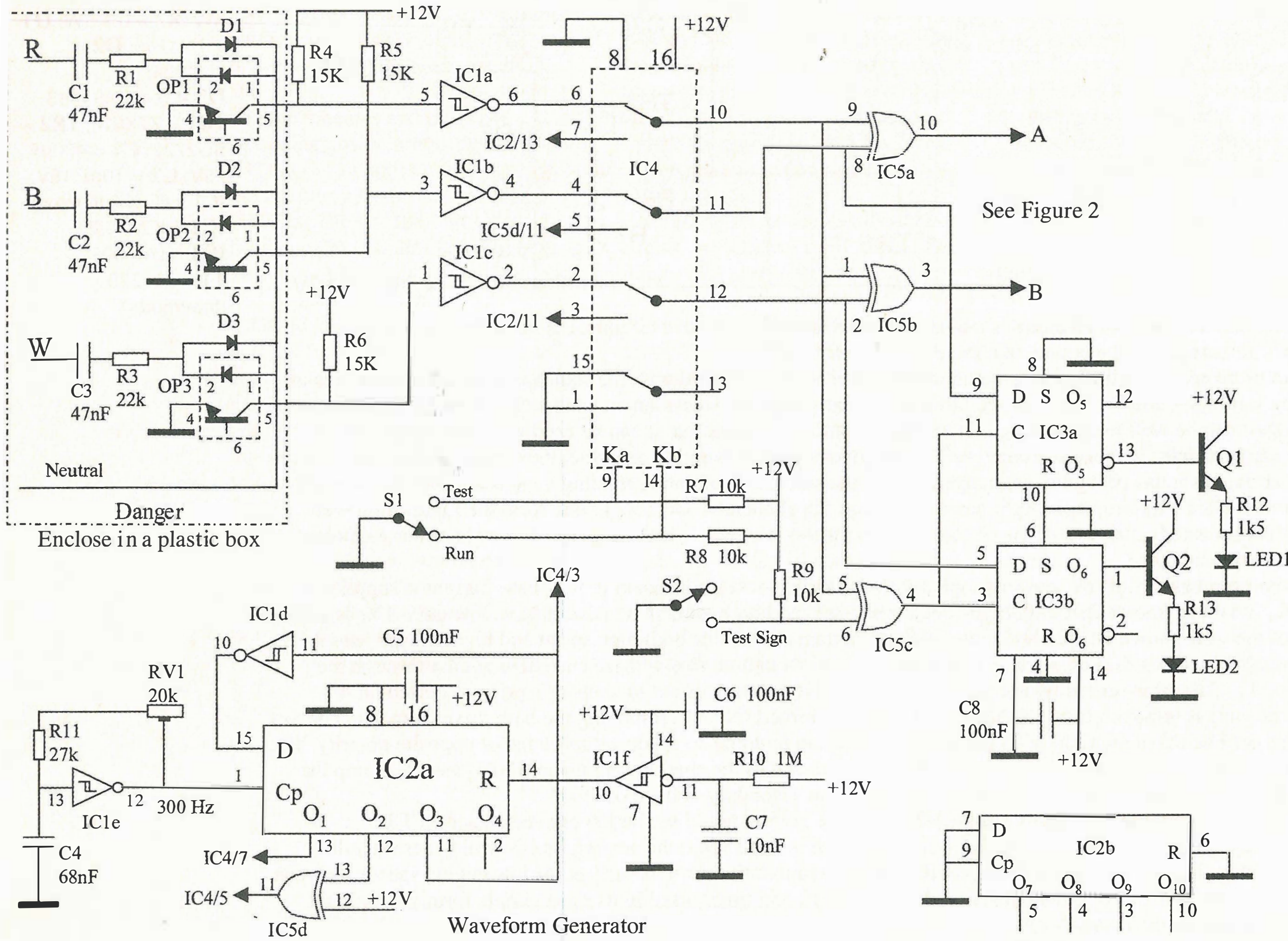


Figure 1: Digital section

A friend has a lathe with a 3-phase motor and lives in a country area with only a 1-phase supply. Provision of a 3-phase supply was prohibitive. So he bought a device that converts a single-phase mains supply to three phases. However, it needs manual adjustment of the phase angle when under load. So he asked me to design and build a phase meter for him.

The 3 phase mains are even more dangerous than the single phase since there is 380-415 Volts between the phases. Thus a safe isolation unit is required between the mains and the low voltage section. This isolation is provided by opto-couplers OP_1, OP_2 and OP_3.

The 3 phase signals from the opto-couplers are squared by the Schmitt triggers IC1a, b & c and then applied to the exclusive OR gates IC_5a & b via IC_4. The output of the exclusive OR gates is a square wave signal whose mark to space ratio is proportional to the phase difference. These signals are applied to the averaging amplifiers IC_6a & b. Diodes D_4 & D_5 prevent a reverse current through the meters if the phase angle is less than 90°. NB. the exclusive OR gates cannot determine the sign of the phase angle, i.e. -120° will result in the same waveform as +120°. Hence IC_3a & b are employed to detect the sign. When S_2 is open, IC_5c inverts the $IC_{4/10}$ signal so that the D Types are clocked on the negative edge of this signal. Hence if the sign of the phase is correct, O_5 will go Low and O_6 High thus illuminating the LEDs. See **Figure 5**.

In order to calibrate the meters, IC_2a is configured as 3 stage Twisted Ring Counter which generates a 3 phase test waveform. Refer to Figure 4. Note that the phase difference between O_1, O_2 and O_3 is 60° but since O_2 is inverted by IC_5d, the correct phase differences (120°) are obtained. See **Figure 5**.

The power on reset is generated by IC_1f. This necessary because the counter could start in state 010 or 101. If so, the counter would oscillate between these states rather than count in the correct sequence. When S_1 is set to the Test position, IC4 switches the test waveforms to IC_5a, $_5$b & $_5$c. The meters are calibrated by adjusting RV_2 & RV_3 to set the meters to centre scale (0.5mA which represents 120°) and both LEDs should be lit indicating that the sign of the phase is correct. If switch S_2 is closed while S_1 is in the test position, IC_5c inverts the phase of the clock signal to IC_3a & $_3$b thus extinguishing the LEDs.

Len Cox
Forest Hill
Vic
Australia

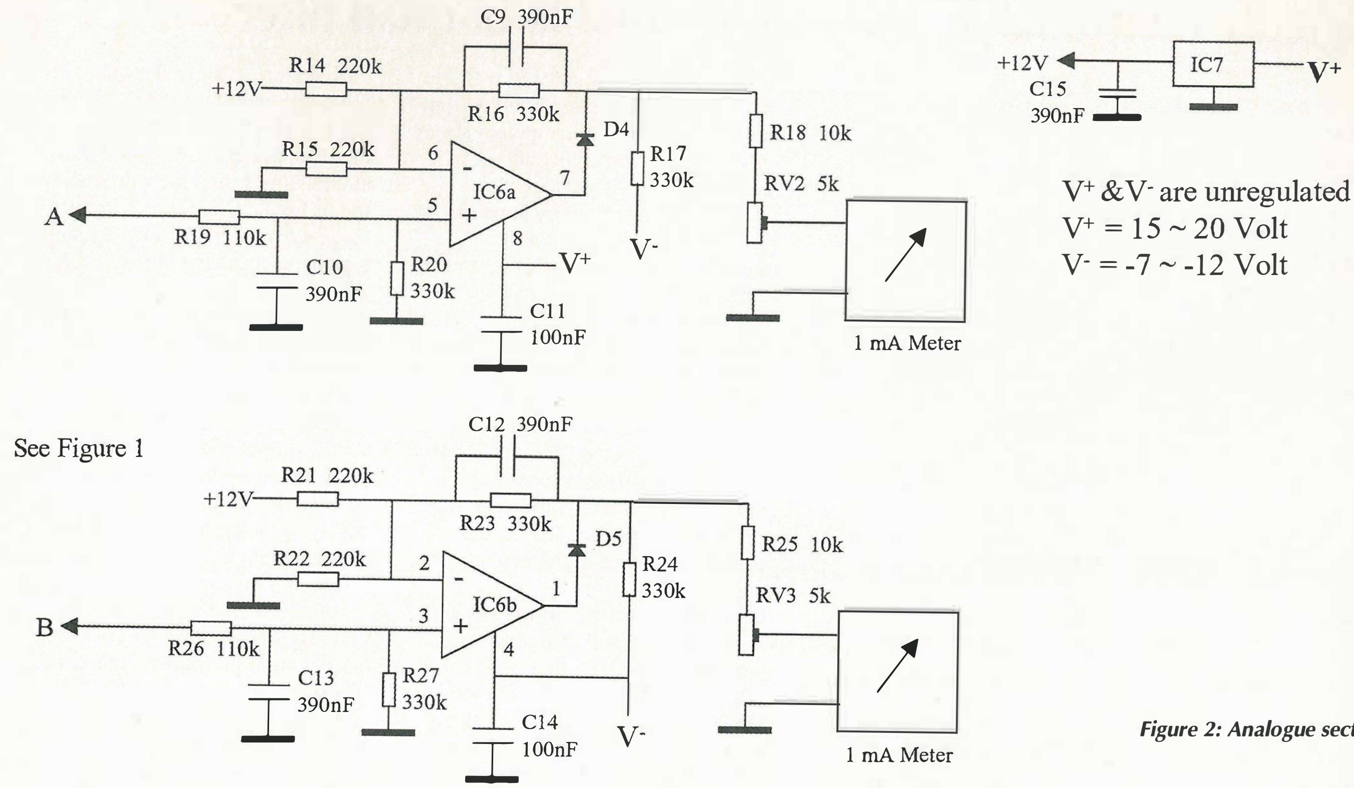


Figure 2: Analogue section

Figure 3: Count sequence

O_1	O_2	O_3
0	0	0
1	0	0
1	1	0
1	1	1
0	1	1
0	0	1

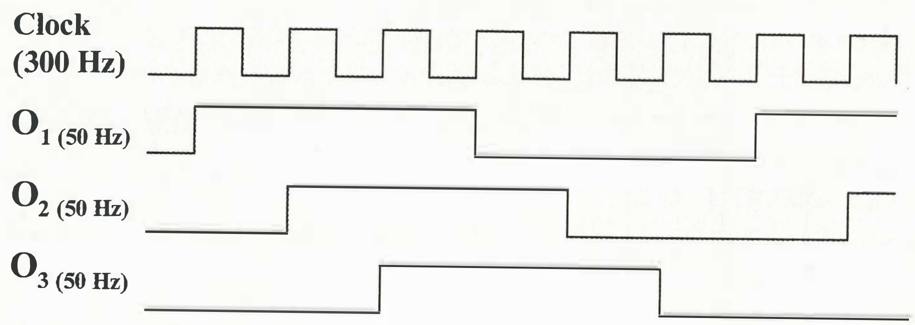


Figure 4: Waveforms of twisted ring counter

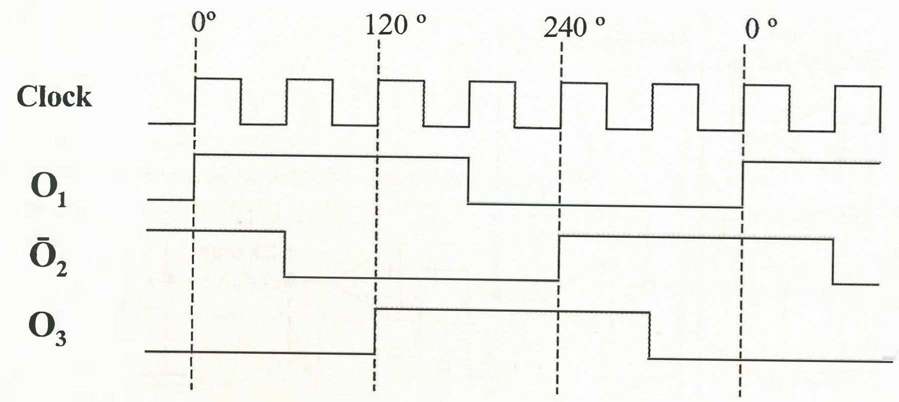
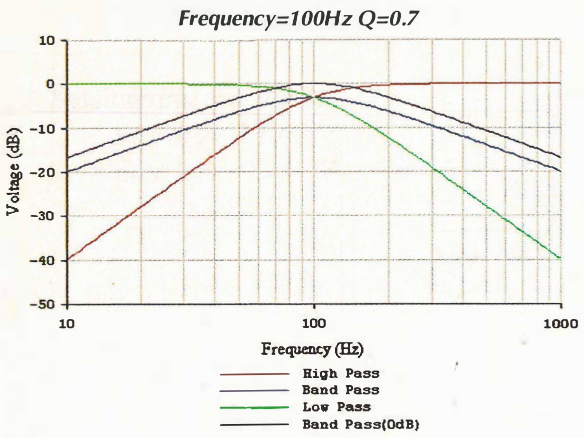


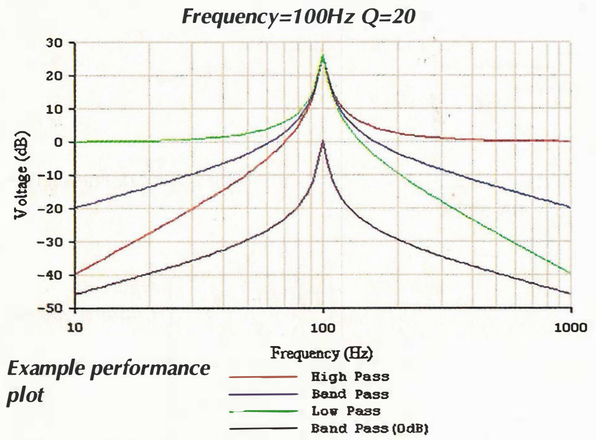
Figure 5: Waveformss showing O2 inverted & phase angle

Parts List

$OP_{1,\,2,\,3}$ 4N25 opto-coupler
IC_1 40106 Hex Schmitt Trigger
IC_2 4015 Dual 4 Bit Shift Register
IC_3 4013 Dual D Type Flip Flop
IC_4 4019 Quad 2 input Multiplexer
IC_5 4030 Quad Exclusive OR gate
IC_6 LF353 Dual Op Amp
IC_7 78L12 Voltage Regulator
$C_{1,\,2,\,3}$ 47nF Capacitor 250V Mains rated
C_4 68nF 50 Volt
$C_{5,\,6,\,7,\,8}$ 100nF 50 Volt ceramic
$C_{9,\,10,\,12,\,13}$ 390nF 50 Volt ceramic
$C_{11,\,14}$ 100nF 50 Volt ceramic
$R_{1,\,2,\,3}$ 22k ½Watt 5%
$R_{14,\,15,\,21,\,22}$ 220k ¼ Watt, 1%
$R_{16,\,20,\,23,\,27}$ 330k ¼ Watt, 1%
$R_{19,\,26}$ 110k ¼ Watt, 1%
All other resistors are ¼ Watt, 5%
$D_{1,\,2,\,3,\,4,\,5}$ 1N4148
Q_1, Q_2 PN100
RV_1 20k Trim pot
$RV_{2,\,3}$ 5k Trim pot

Digitally controllable, truly state variable bi-quad filter





Example performance plot

Controlling the frequencies of filters by using potentiometers would seem simple, however, although the output ratio of potentiometers both mechanical and digital are very accurate, the absolute track resistance is usually specified by the manufacturer as having a tolerance of ±20% which cannot provide for repeatable results.

By dividing the potential across a resistor terminated into a virtual earth point, the conductance can be reduced thereby increasing the effective resistance. This can be used to control RC time constants independent of the end-to-end track resistance of the pot. To further isolate the physical properties of the pot, the output from the wiper is directed to a voltage follower reducing the wiper impedance to near zero.

In circuit shown, the R's in the Bi-Quad Filter are multiplied by the division ratio of the potentiometers 'F'. Using the values shown, the full scale frequency is set to 256Hz such that if a 256 tap digital pot is used, each step will change the frequency by 1Hz meaning that the count sent to the pot will be the centre frequency e.g. 200 counts = 200Hz.

In order to achieve the same 'rational' control of the Q, the pot 'Q' is arranged to control the gain of the feedback loop instead of the attenuation. Using the values shown, the full scale Q is set to 2.56 such that each step in a 256 tap digital pot represents a change in Q of 0.01 e.g. 100 counts = a Q of 1.00. Qs of 25 or more can be reliably and accurately achieved by scaling the resistor values.

By arranging the feedback in this fashion there is also an added benefit; At the output of the gain stage, there is available a band pass signal whose level at the centre frequency is always exactly 0dB, independent of the value of Q.

Though the circuit was primarily designed to be digital controlled, mechanical potentiometers are just as effective giving the same linear scale.

A small word of caution; Do not use the pots around 0%, as most filters do not operate correctly at zero Hz or with zero Q! This one is no exception.

John Charlesworth
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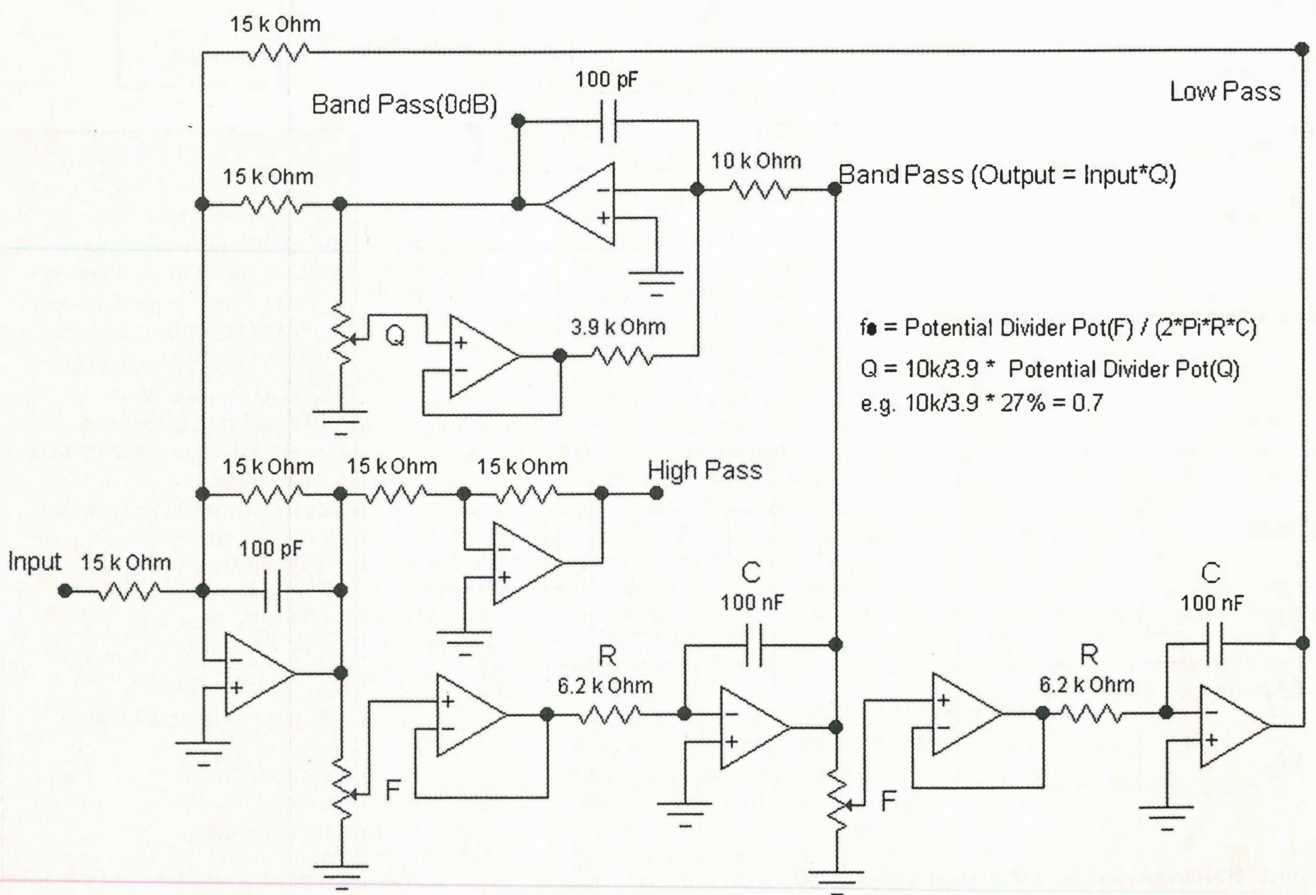
UK



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Specifications

Switch position 1	
Bandwidth	DC to 10MHz
Input resistance	1MΩ – i.e. oscilloscope i/p
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	

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'.

Class A Imagineering

Someone needs to answer Mr Maynard's comments on capacitive load coupling before some poor sod goes out to buy 22mF caps! The 'error' voltage shown in his simulation is simply the inevitable dc shift that starts during the first half cycle of an applied signal. This phenomenon occurs everywhere capacitive coupling is used regardless of the technology used in the electronics, indeed valve cathode decoupling caps are prime offenders in this regard.

In his haste to rubbish capacitive coupling in output stages he has overlooked the fact that loudspeaker systems are mechanical high pass 2nd order and above filters. The gain of high pass filters to dc signals is zero, therefore these 'errors' cannot produce an audible output!

Capacitive output coupling does deliver the goods down to the calculated -3dB point.

What's not so easy to defend is transformer coupling. The only way to produce a secondary voltage that is related to the primary voltage is to feed from a zero impedance source. This is never done although it could be with some output stage redesign. Ironically it is easier to produce good results in the bass with transistor drive than with valves, when using output transformers.

If you were to substitute 22mF caps in most single power supply amps you would probably run into problems. Providing the extra surge current didn't either fry your rectifier or the output devices you wouldn't attain any benefit. Believe it or not engineers have been designing amplifiers for a long time now. Many examples of their work has survived unscathed for decades and still gives good service. The problems that their designers coped with are also unchanged with the passage of time.

A good solution then is a good solution now.

Jeff Macauley
Sussex
UK

Graham Maynard replies: *I thank Mr Macaulay for his letter and agree with him that a potential shift develops during the first half cycle of sinewave current flow via a capacitor, also I feel that his view of resonant loudspeaker systems being like high pass mechanical filters already suggests that any series driving impedance ought not lead to any capacitively reactive interaction.*

Around bass driver loudspeaker resonance frequencies, voice coil currents are rarely directly in phase with an amplifier's signal driven output voltage, and it is the resultant loudspeaker current that generates the capacitor 'error' potential, which then becomes inserted between the low distortion NFB loop controlled amplifier's voltage output node and the loudspeaker system we listen to. The series capacitor error potential is not predictable, for it varies in relation to a music driven loudspeaker's characteristics and as illustrated, it can be substantial; i.e. it is not fixed, nor in direct relation to the amplifier's low distortion voltage output. Loudspeaker induced capacitor error potential becomes superimposed upon genuinely amplified and low distortion music output voltage, and this was my reason for having already included a separate 4m7F amplitude plus phase plot in the yet to be printed Part 5.

I have not rubbished capacitor coupling, merely suggested that a 22mF fitment will reduce the dynamically and reactively generated series output capacitor induced loudspeaker reproduction distortion that can be audible at frequencies above a steady sinewave -3dB roll-off point. Nor have I been hasty, for as I have already stated, I ran comparative listening tests with different capacitor values over thirty years ago, a time when what is now termed as 'sub-bass' frequency output was much less prodigious. An entirely natural change in bass driver control is clearly audible for those who might care to repeat my comparisons; though not all loudspeakers are equally capable of inducing and/or revealing that bass/sub-bass change.

Mr Macauley suggests that transformer coupling is less easy to defend, and I have said valves would not be my future aim. However, most valve amplifiers take their NFB directly from the loudspeaker connecting side of an output transformer such that the transformer inclusive amplifier effectively presents a lower driving impedance at its output terminal. Yet to maintain stable low frequency solid-state amplifier driving capabilities a series output capacitor must normally be fully outside of its amplifier's NFB loop, and thus any series capacitor 'generated' error potential cannot be corrected. In the past I have attempted to minimise capacitor error by splitting the output capacitor and including part within the NFB loop, and by enclosing it within the NFB loop with compensatory input/gain/output tapering, but these methods present more problems than the results are worth, especially when a single larger value component is such a simple remedy.

Yes many examples of decades old amplifiers have survived unscathed and are serviceable, but there are precious few of the older solid-staters that we would actively seek out as being our first choice for driving modern, high resolution, dynamic loudspeakers. Actually I would be more inclined to feel sorry for the 'poor sods' whose amplifiers are already fitted with 2m2F or 4m7F series output capacitors, and yet, because they cannot miss what they have never had, they might already be entirely satisfied and thus see no need to alter an existing system.

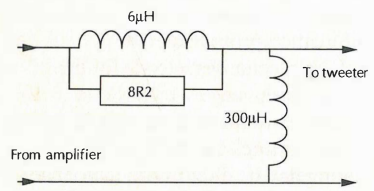
Valve/transformer design did not remain stationary either;- for general interest see http://www.audioresearch.com/ref600.html

Graham Maynard

Class A Imagineering

I am afraid that Mr Maynard has made some fundamental errors in his series on audio amplifiers and loudspeakers. I would like to refer in particular to Part 3 where he seems to be trying to prove that a small series inductor in series with the output of an audio amplifier can have serious effects on the output fidelity.

Firstly, referring to the "error" signals that he quotes in Figure 8. Here he is comparing the drive signals to the first node of the tweeter crossover network with and without a 6µH inductor that has a parallel 8R2 resistor. Sure, there will be a difference - it would be expected. At frequencies above the crossover frequency then we can largely ignore the 4µ7 capacitor and we are then left with the circuit shown in my Figure 1. This is an inductive potential divider with a ratio of 1 to 0.98. Sure the upper arm has a resistor across it, but at say 20kHz, the reactance of the 6µH inductor is about 3/4 Ohm so the parallel resistor of 8R2 is largely irrelevant. A potential divider ratio of 0.98 gives an error of only 0.02% - i.e. 0.17dB.



So all the output inductor does is to give a reduced tweeter sensitivity over its active range of some 0.17dB. Hardly a distortion and being so small it would not be expected to be audible. Heating up the tweeter voice coil by about 6∞C would have a very similar effect.

But what of the spiky error results? That surely is different? Yes, the errors Mr. Maynard quotes will exist - but only for the test conditions he specifies.

The fundamental error is using tone-burst waveforms without filtering them. It is a similar problem to the mains transformer problem referred to by Mr. Catt. In practice all audio signals will have been through bandwidth limiting devices such as microphones, microphone transformers, line amplifiers, recording gear, etc. etc. None of these go down to D.C. nor to much above 30kHz or so. If we feed a tone-burst through a simulated audio bandpass filter, then we will find that D.C. type transients are minimal, as are the sudden sine wave starts from zero that Mr. Maynard assumes. Mr. Maynard's test signal has appreciable out-of-audio-band components and it is these that are causing his spike errors.

So basically series output chokes merely drop tweeter sensitivity very slightly and seem to have no other untoward effect.

Arthur R. Bailey Ph.D. M.Sc. F.I.E.E.
Ilkley
West Yorkshire
UK

Graham Maynard replies: *I sincerely thank Mr Bailey for his letter and this opportunity for me to cover his comments. I fondly remember reading Mr Bailey's pages back in the 1960s; they were instrumental in founding my Hi-Fi knowledge.*

Mr Bailey raises several points, however he omits the 4.7µF capacitor and then goes on to approximate my illustrated tweeter crossover section as if no more than two inductors in series as a potential divider. However it is the capacitor that has the greatest series impedance, and this component is able to fractionally resonate with the exampled 6µH series output choke between an amplifier's low impedance output node and the rest of that bulked series/parallel tweeter section components and driver. This gives rise to a tuned error peak (not a spike) arising at the amplifier's output terminal after the momentary and reactive series inductance initiated voltage charging delay. Look at the intersection of 6µH and 4.7µF on a resonance chart - this brings us right back into audio territory.

Mr. Bailey suggests that my errors will exist only for the test conditions specified, alluding that I really ought not be suddenly starting a sine wave from zero - i.e. a toneburst - without first filtering it. Sorry ?

Ever listened to music via a SSB filter?

Ever listened to music via a CW filter?

Are we listening to sound waves or not? (Thank goodness Mr Catt did not set in the universal mould!)

Simulator programs are not limited by human thought processes, they calculate, and if properly informed provide accurate results. Due to time and facility limitations I cannot satisfy even my own wish to be fully accurate with these simulations, but I would ask all readers to wait until Part 6 when I pull this section of text together. Then, if the doubters believe my tests are fundamentally flawed I look forward to hearing their explanation as to how I have wrongly managed to remove simulated distortion without changing the testing conditions themselves.

Graham Maynard

Errata – Hybrid amp

It has been found that the ECC82 can be directly substituted for the ECC88 in the circuit so you can use whatever device can be easier obtained for V1. Note however that the '82 heater wiring is different, consult a tube manual for details. The 2SC2547E used for Q1 is hard to substitute for. A ZTX600B can be used instead if R2's value is increased to 20kΩ. R1's value has been set at 56kΩ. Some may find that the circuit lacks sensitivity and on reflection a value of 27kΩ is recommended for this component. The gain is defined by the ratio of R3 to R1 and the value of R1 can be varied over a large range if a different input sensitivity is required. For experimenters, R1 = 40S where S is the RMS input voltage required for full output in volts and R1 is expressed in kΩ.

The layout diagram on page 51 has some errors on it as follows:-

Q4 and Q3 are shown with the top pin of the devices being the gates, the centre pins are the drains and the lower pins are the sources. Looking at Q4 the unmarked gate connected resistor is R8. It needs to be soldered directly to the gate pin. The drain of Q4, centre pin should connect to the common earth, CE point not to the + terminal of C4 as shown. The source pins of both Q3 and Q45 are correctly shown connected together. In addition the + terminal of C4 needs to be connected to Q4's source.

Finally the connection shown between V1's cathode (K), and the gate of Q3 should be removed. Make a new connection between the collector of Q2 (COL) and V1 grid, pin (G). Thankfully none of the above listed errors is device threatening! The circuit will now be ready to set up as described in the article.

Apologies for the above.

Jeff Macauley
Sussex
UK

Erratum

There is an error in my article about noise and moving-magnet cartridges in *Electronics World* October 2003. In figure 5, the lower half of the double pole switch is drawn in the wrong position.

In the 40dB mode, the 243Ω resistor should be connected to the 33µF bipolar electrolytic capacitor and the left-hand side of the 232kΩ resistor should be connected to ground.

In the 30dB mode, the 243Ω resistor should be disconnected from the 33µF, and the 232kohm should be connected to the 226kΩ.

Marcel van de Gevel
By email

Alkaline battery failures 2004

I wonder if any other reader has experienced the sort of problems I have had with alkaline batteries, mainly this year. I would recommend checking the're rarely used or backup battery equipment soonest!

In the last few months I have found several cases of cells either leaking, splitting their case, or bulging at the ends. In several instances this has required dismantling and cleaning of the unit. These have not been either excessively old or over discharged at heavy load.

For example TV and VCR remote controls lying unused for six months, quartz clock, cells just loose 'on the shelf', cells in a flashgun which was working before and was definitely not left on.

I know they can burst if shorted, but in each burst case where they were in equipment the unit worked OK afterwards and there was no evidence of shorting.

Cells were mostly AA, AAA, C size, though 1 PP3 also failed. The manufacturers involved were all reputable brands, such as. Duracell, Ever Ready, Panasonic, Ray-O-Vac etc. and all within the marked use by date.

Peter Hague AMIEE
By email

More JLH

I read Ian Hickman's excellent tribute to John Linsley Hood in the May Edition, and like him I share the sadness at the passing of a great contributor to the magazine.

I met John only once at a Symposium at Reading University at which he made a presentation. I was able to have a short while to talk with him on that occasion. He was a very quiet man with a wealth of experience especially in Audio matters, as is well demonstrated in his many articles.

In the May Article Ian Hickman says that at least one of the contributors mentioned along with John is 'still with us' – I believe I may be the person referred to. Unlike John my past articles ranged over a wide range of subjects. Although now retired I still get involved in electronics generally, although I have not done much writing in the past 20 years, as for quite a while I ran my own small business in electronic design work, and much of what I was doing could not be written up due to it being for a couple of local companies, and commercial security prevented publication.

I still read *Electronics World* every month, a habit which goes back to the early 1940s when as a young teenager I often had to wait for long hours for the bus home from High Wycombe after school, and I would go into the Library Reading Room where they had the *Wireless World and Electronic Engineering*, and also if memory serves me right the *Wireless Engineer* (now long gone). It was there that I began to learn my electronics.

Laurence Nelson-Jones
Bornemouth
Dorset
UK

CD-R failure

In response to Simon Wright who in *EW* July 2004 wrote a letter about CD-R failure, there is a short article on the following internet address that may be of interest: http://www.images.dk/earkiv.nsf/doc/permanence?OpenDocument

This is an article on the permanence of CD-R media in English by Jacob Trock. Another short article can be found at:http://www.medialinenews.com/issues/2001/news/0314/0314.1.shtml by the same author. The articles conclude with the same finding as that of Simon Wright.

Jacob Trock has studied CD-R's permanence under the Royal Danish Academy of Fine Arts, School of Conservation.

Henning Tousted
By email

Mystery of magnetic lines of force

Wilf James asked for some enlightenment on the question of the existence of magnetic lines (*Letters*, August '04) which is what Jacob Beckenstein wrote about in his article in *Scientific American*, (August, 2003 pp48-55) entitled 'Information in the Holographic Universe'. Summarising the article, under 'Augurs of a Revolution', Beckenstein wrote "although the holographic way of thinking is not fully understood, it seems to be here to stay. And with it comes a realisation that the fundamental belief, prevalent for 50 years, that field theory is the ultimate language of physics must give way." The fundamental theory, "a final theory, must be concerned not with fields, not even with space-time, but rather with information exchange among physical processes."

Consciousness and language will be part of final theory and in this respect, 'a global network of 60 sensors* is being built to police the Nuclear Test Ban Treaty [which] will also provide useful data about: tornadoes, seismic resonance, animal communication, shivers down the spine (or up, Kundalini?) and haunted houses inter alia – e.g. the origin of language; the article following Beckenstein's (ibid, pp 57-63) is perhaps a taster for what's to come.

In the article, entitled *Questioning the DelphicOracle*, the authors explain that Apollo's priestess sat over a chasm, from which ethylene issued, to induce a higher state of consciousness which enabled her to prophesise the future. Her name, Puthia (Pythia in English) is inappropriate for a sweet smelling gas. So is Apollo, it's not derived from 'polloi' (many), according to Curtius, who used 'Apollon', in his introduction, to emphasise mistakes made by early scholars in assigning provenance, including Plato – no less – who took particular interest in people's names (vide 'Cratylus') following Socrates, who contended that 'true' names, whose sounds capture the entity's essence, were assigned by ancient 'legislators' who had special insights. The root 'apo' means 'turn around' or 'land across the sea', but Curtius suspected, due to vowel change and consonant doubling, that it could relate to Apolio or Apullia, towns in Italy, which is where Puthia comes in, i.e. A pull-, whose gematria is 1.8655 – the ratio of bifurcation constants: F1, F2. Or to turn the argument round, as Oliver Heaviside, Jacobi and many others suggested: if at first you don't succeed – invent.' In other words: bifurcation processes in the holographic universe code for the origin of language which Rama Chandran re-iterated recently in the Reith Lectures for 2003, although stating that he didn't want to go that far!

Support for some of these conjectures, in the form of ELF, may also come from the Apollo Project, due to come on-stream this year, utilising mirrors left on the moon by the Apollo mission, the global network of 60 sensors is specifically designed to look for frequencies below (less than 4 peddars) 20Hz. At the other extreme, Casimir effect researchers are looking for frequency cut-offs at around 1600 GHz! – and what we can learn from that? Not a lot, I suspect, but – there you go: 'You pays your money and you takes your choice.' One of the reasons for these wide divergences (base 10 expo123 in the worst scenario) may be found in Number theory. Roger Penrose in his 'Large, Small and Human Mind', Appendix 1, 'Goodstein's Theorem and Mathematical Thinking', shows that any number like the very large just previous and say 137, the comical constant (λ) can be shown to equal zero. It's not a new idea; and perhaps, we should leave it till later.

When the implications of the data (e.g. Tiller's) becomes quite unavoidable! I hope this might provide some of the illumination Wilf asked for.

Tony Callegari
Much Hadham
Hertfordshire
UK

**New Scientist*, 26 June 04, page 51. 'The word-Infrasound'.

The cat's anomaly

Watching my cat play this morning, a profound thought flickered through my brain. Once I had tracked it down, I found that I had, apparently, an idea that was worth pursuing. It goes like this:

She was playing with a couple of toy mice. Now they came in a packet with several others, and all look the same. She obviously preferred one particular one, thus she can tell the difference, but, I would suspect that every one is countable as just 'one' to her? The "natural", or non-human way of looking at the Universe.

In mathematics in human terms, I would state the most basic mathematical postulate as:

Call an 'entity' by the sign **1** – put another identical entity beside the first, and I can designate the group by the sign **2**

That is, symbolically add **1** to **1** and the answer is **2** – on that basis, the whole of mathematics is built (call it the fundamental equation).

Now, the anomaly that my cat pointed out is that, in fact, NO entity is identical to another (many better people than I have already pointed this out, for example, Heisenberg) and so the fundamental equation breaks down. Thus I concluded that a better form of mathematics is needed, or maybe the whole Universe cannot be described on the basis of current mathematical theory? Or maybe it doesn't exist?

Can anyone explain away this anomaly for me?

Tony Batchelor
Denmark

Injustice to Newton

I have just looked up what Leslie Green wrote in the August issue of *EW* p52.... "Consider what is known today as Newton's second law: F=ma". To my disappointment I find this version of the law is backed up by several reference books.

But it is not what Isaac Newton wrote when he published "Principia Mathematica Philosophiae Naturalis" in 1686/7. It was in Latin, but we can write the second law in Mathematical notation. Ideally it would use his elegant fluxations, - in which a dot over a variable indicates rate of change with time. I can't type the dot so I have to use Leibnitz. The product of mass and velocity was called motion (=u), later momentum.

Newton's second law ....
F= du/dt where u= mv.
The product leads to
F= m(dv/dt) + v(dm/dt).

In words.... Force equals (mass times acceleration) plus (velocity times rate of change of mass). Newton took direction into account, it could be expressed with vectors.

This proper form of Newton's law copes with the speed of light limitation by allowing mass change and Mr Green's page 52 constant force does not necessarily generate unending acceleration.

C.W. Rimell
Spalding
Lincolnshire
UK

Drawing standards

I bought my first copy of *Wireless World* in 1980, the same year as Clive Sinclair brought out his personal computer. Printers were 9-pin dot-matrix, and graphics were nonexistent, yet the circuit diagrams in Wireless World were beautifully drawn in a consistent style and perfectly legible. Since then, computing power has increased enormously, graphics packages can draw in thousands of different fonts, printers can print over a thousand dots to the inch, but what has happened to *Electronics World's* circuit diagrams? They seem to have gone in quite the opposite direction. Nowadays, some of them are barely legible, some look like a five-year-old has drawn them on etch-a-sketch, there is no consistency in the symbols used. It was nice to have my circuit idea printed, but why are the symbols in my circuit twice the size of those on the facing page, and how come the diagonal lines have become pixellated? It wasn't even sent to you in a bitmap format!

On the subject of Circuit diagrams, compare the circuit diagram of the Linsley-Hood amplifier reprinted the June 2004 edition, with Graham Maynard's amplifier circuit diagram from the August 2004 edition, and ask an impartial observer to guess which came from the days where the best we had was a 9-pin dot-matrix printer, and which came from the era where professional-quality publishing software is available on every desktop.

Ian Benton
By email

You are correct. But you are forgetting the economic factor. Back in the 70s, Wireless World had a much higher circulation, made more profit and could afford more staff. My policy on circuit diagrams is simple. Circuit Ideas get little or no 'tarting' unless the drawing is really bad.

Feature articles get more time (and money) spent on them – but again – if the drawing is good, although not in the house style – I tend to let it through. This is because I would rather spend my budget on things that make a difference – the fact that the drawings may be different article to article is not that important, compared to having the article there in the first place. – Ed

Catt's 377Ω space predicts G to within 1.7%

My Electronic Universe article, Apr. 2003 *EW*, proves $G = 3H^2/(4\pi r)$. Here H is Hubble constant and r is the density of universe responsible for causing gravity by reaction of Catt's 377Ω space to the big bang. Considering the density, it is highest at early times and thus density increases in the observable space-time trajectory, as we look further into the past with increasing distance.

But the increasing spread of matter with increasing distance partly offsets this increase, as proven when we put the observed Hubble equation ($v = Hr$) into the mass continuity equation and solve it. For spherical symmetry, $dx = dy = dz = dr$. Mass continuity implies: $d\rho/dt = -\nabla.(\rho v) = -\nabla.(\rho Hr) = 3d(\rho Hr)/dr = -3\rho H$. Solving $dr/dt = -3rH$ by rearranging, integrating, then using exponentials to get rid of the natural logarithms (resulting from the integration) gives the increased density to be ρe^{3Ht}, where e is Euler's constant (2.718 ...). In the absence of gravitational retardation (i.e. with the cause of gravity as inward reaction of space to the outward big bang), $H = 1/t$ when $H = v/r = c/$(radius of universe) $= 1/t$, where t is the age of the universe, so $e^{3Ht} = e^3$ and observed $G = 3H^2/(4\pi e^3\rho)$.

Nugent, *Physical Review Letters* (v75 p394), cites decay of nickel-63 from supernovae, obtaining H = 50 km/sec/Mps (where 1 Mps = 3.086×10^{22}m). The density of visible matter at our local time has long been known to be 4x10-28 kg/m3. However, White and Fabian in the March 1995 *Monthly Notices of the Royal Astronomical Society*, using the Einstein Observatory satellite data, estimate that invisible gas increases this density by 15%.

Using these data, $G = 3H^2/(4\pi e^3\rho) = 6.783 \times 10^{-11}$ Nm^{2}kg^{-2}, 1.65% higher than the physical measurement for G of 6.673×10^{-11} Nm^{2}kg^{-2}. So current data predicts acceleration of just under 10ms^{-2} at the Earth's surface, compared to the observed value of about 9.8ms-^{2}. This proves Catt's insistence on the reality of the 377Ω fabric of space beyond any reasonable doubt.

Nigel Cook
Colchester
Essex
UK

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The Kölnarena, Germany's largest ultra modern, multi-functional events forum has installed a set of the art digital announcement system manufactured by Interalia, designers and manufacturers of digital voice announcement solutions. Specified and installed by Laauser & Vohl, the TROUNCER system is used as a public address system providing visitors with information and jingles. TROUNCER can be installed in a variety of environments and has been designed to provide high-quality audio messages and control visual information displays. Audio is saved as .WAV at any available sampling rate including full CD quality 44.1k. Interalia's headquarters are in Canada with subsidiary locations in the USA and UK.

Interalia
www.interaliac.com

Changes to hazardous waste

A new guide published by Envirowise helps management teams to deal with new EU hazardous waste rules which came into force in July 2004.

The publication, Hazardous Waste Management - essential information for business (GG469), alerts senior managers to the opportunities and risks posed by the Landfill Directive and the forthcoming implementation of the European Hazardous Waste List.

To order a copy visit www.envirowise.gov.uk/hazwaste

Component Kits

A range of component kits is available from new supplier Fast Component. "Electronic hobbyists are having a rough time, leaded components are becoming more difficult to find, and suppliers are imposing 'small order' or high shipping charges". We are offering a fresh and more economic way to buy electronic components" says co-founder Thomas Arundel.

"As former product designers we were frustrated with time and effort it took to build up a decent stock of bench-top components. Once we started investigating a supply chain, e realised that we could sell its of commonly required components at a third of the price they would cost separately."

Fast Components has opened their range with kits containing, 1000 1% metal film resistors spread over 93 values, 240 50V ceramic capacitors spread over 24 values, 93 radial electrolytic capacitors spread over 12 values, and 100 miniature polyester capacitors spread over 9 values. The resistor kit retails at £9.99, and the capacitor kits at £4.99. Customers can order direct from www.fastcomponents.co.uk, where they can pay securely with credit cards, PayPal, or Cheque / Postal Order. Full Adobe PDF datasheets of all kits are available on the website.

"We've started with a useful collection of kits to get hobbyists (especially newcomers) started" says Thomas. "We're planning to follow up with more passives, surface mount kits, mechanical hardware kits and semiconductors through the autumn."

Postage & packaging is at cost with no minimum order charge.

Fast Components Ltd
www.fastcomponents.co.uk

Gear-tooth sensor

A Hall-effect gear-tooth sensor module ideal for obtaining speed and duty-cycle information in automotive applications such as transmission speed sensing has been introduced by Allegro MicroSystems Europe.

The new device, ATS643LSH, combines a Hall-effect sensor IC and magnet in an optimised configuration to provide a user-friendly solution for true zero-speed digital gear-tooth sensing in two-wire applications. The integrated circuit incorporates a dual-element Hall-effect sensor and signal-processing circuitry that switches in response to differential magnetic signals created by ferrous targets. The device contains a sophisticated compensating circuit to eliminate magnet and system offsets.

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Allegro MicroSystems Europe Ltd
www.allegromicro.com

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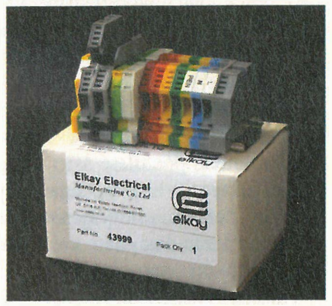

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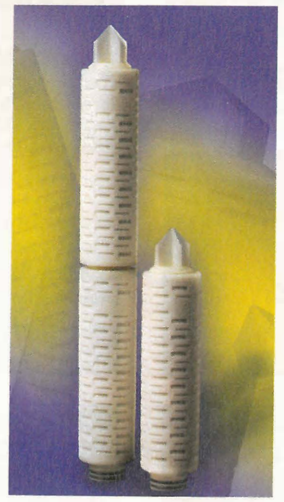

The new CUNO PhotoSHIELD™ Nylon filter cartridge is a highly retentive membrane filter element designed to provide the highest level of gel and particle removal from photoresist and ancillary chemicals while also saving processing time and reducing costs. Allowing high flow rates while maintaining full efficiency, these filters achieve the required particle specifications in less time, which results in lower filtration costs by reducing energy consumption, pump wear and labour costs.

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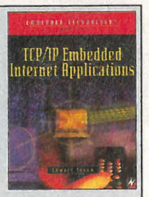

Edward Insam

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**

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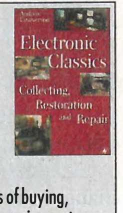
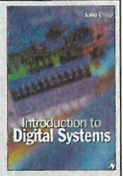


Andrew Emmerson

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

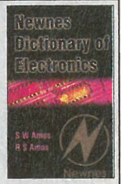


John Crisp

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**

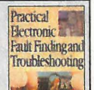
NEWNES GUIDE TO DIGITAL TV

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

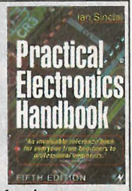


Robin Pain (Design Engineer, Cotag International Ltd)

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**

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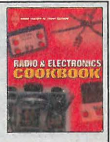

Radio Society of Great Britain

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
Published in UK
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
1568 pages & CD-Rom ▲ 1385 line illustrations
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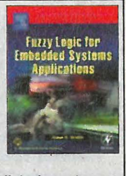

Dr Ahmad Ibrahim

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
Code 0-7506-7605-1 **£35.00**

ANALOG INTERFACING TO EMBEDDED MICROPROCESSORS

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.

March 2001 ▲ 288 pages ▲ PB
100 line illustrations
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Farid Dowla (Lawrence Livermore National Laboratory, University of California-Davis, USA)

Subjects covered include third-generation wireless, wireless sensor networks, RF power amplifiers, spread spectrum modulation, signal propagation and antennas. This is far more than just a tutorial or reference guide — it is a "guided tour" through the world of cutting-edge RF and wireless design, combining theory, applications and philosophies behind the RF/wireless design process.

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
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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
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3rd edition ▲ Aug 2003 ▲ 624 pages ▲ Index
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Mar 2000 ▲ 447 pages ▲ Illustrations ▲ PB
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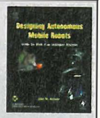

John Holland

Covering topics such as advanced sensor fusion, control systems for a wide array of application sensors and instrumentation, and fuzzy logic applications, this volume is essential reading for engineers undertaking robotics projects as well as for undergraduate and graduate students studying robotic engineering, artificial intelligence and cognitive science. Its state-of-the-art treatment of core concepts in mobile robotics helps and challenges readers in exploring new avenues in an exciting field.

January 2004 ▲ PB & CD-Rom ▲ 352 pages
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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**

VIDEO DEMYSTIFIED

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
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Code 1-878707-56-6 **£50.00**

ANALOG ELECTRONICS: CIRCUITS, SYSTEMS AND SIGNAL PROCESSING

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
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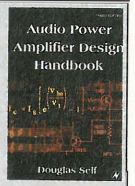

Douglas Self

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.

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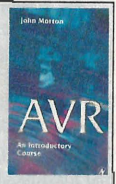

John Morton (University of Cambridge)

AVRs have certain strengths compared to other microcontrollers but are less well-known or understood. The author wrote this guide to introduce AVRs to those who have little or no microcontroller background and to encourage them to start using these useful devices with easy projects to try.

Sep 2002 ▲ 256 pages ▲ 150 illustrations
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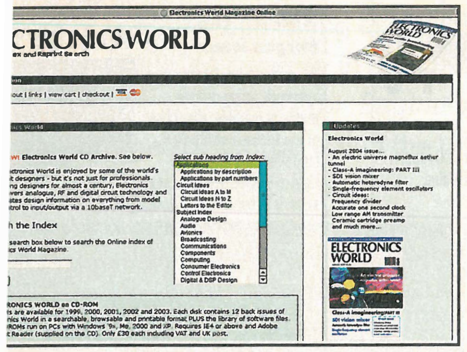

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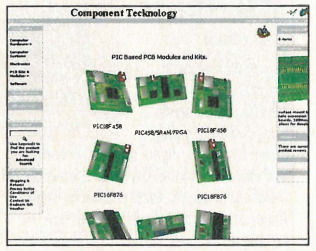

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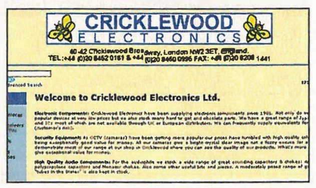

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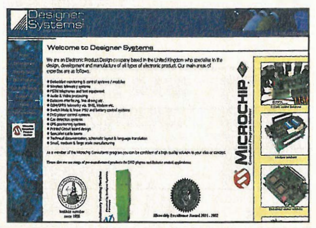

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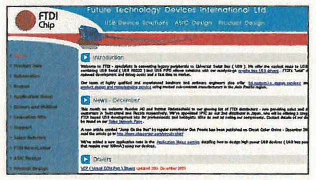

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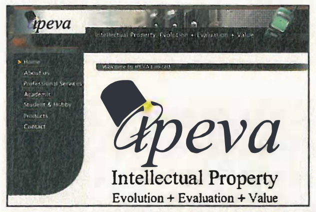

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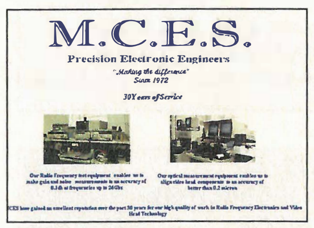

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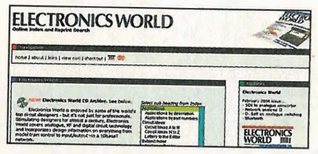

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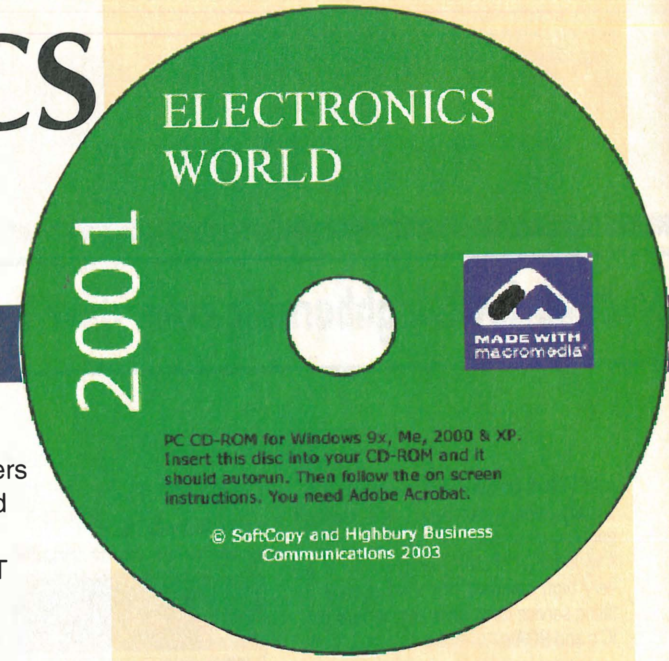

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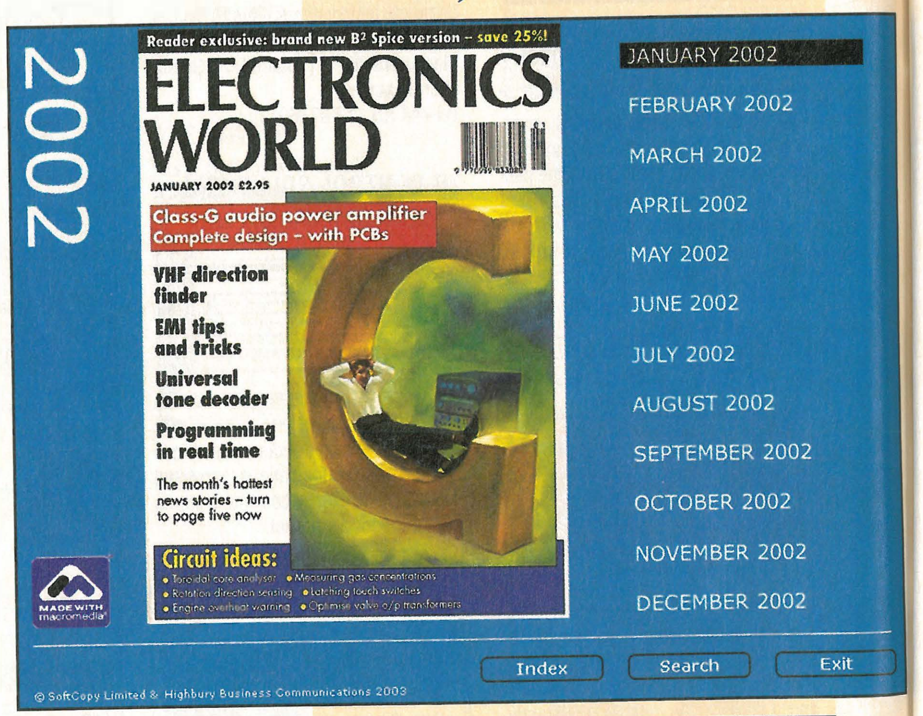

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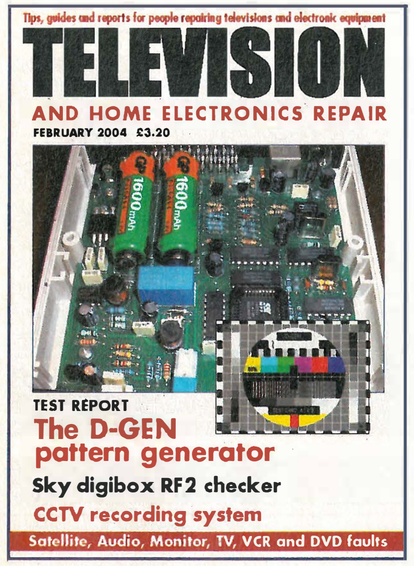

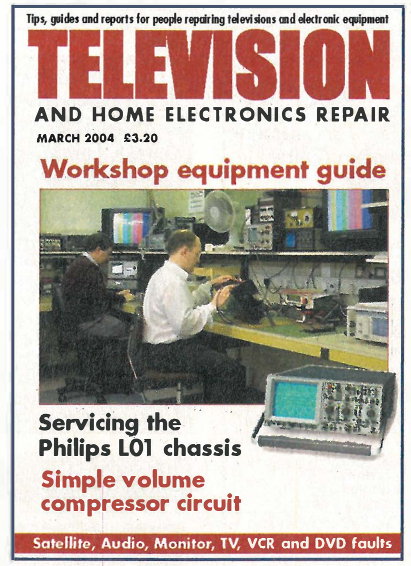

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