Electronics World's renowned news section starts on page 4

ELECTRONICS WORLD

DECEMBER 2004 £3.25

What is Tetra?

Simulating power MosFets part III

Mixed spices part II

Simulating ideal transformers using OTAs

Circuit Ideas

- Omni directional ferrite rod receiver
- Lightbulb protector
- Pump monitor
- High efficiency white LED charge pump
- Simple capacitor checker
- Dual rate thermostat
- Two wire flow control
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CONTENTS

DECEMBER 2004 VOLUME 110 NUMBER 1824

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All change!

Welcome to this, my last issue of *Electronics World* as editor.

I have had a great time over the last 27 issues and have got to know a lot of you, both readers and contributors alike (quite often, one in the same thing!) As regular readers will know, my heart lies in the TV programme industry, where I've been practising my engineering skills now for some 34 years, and I've got the opportunity to edit one of the leading trade journals in that industry – so for once, I might actually understand what I'm writing about!

Looking back over the last two-and-bit years, it's quite amazing how the industry has changed, and that is reflected in our readership. Most new readers are coming in from non-English speaking areas such as the Indian sub-continent and the Far East. These areas will surely become the design and manufacturing base for most (consumer) electronics in the near future. And I have noticed that almost all of the electronics I have recently bought have been manufactured in the Far East, in particular China. No longer are products from here 'cheap and nasty'. All the products I purchased, which was computer parts and DVD player are of excellent build quality and design. The DVD machine is so good, I have now completely shelved my idea of using a PC as a 'play anything' machine. This one does it all and is in very neat little 'wife friendly' box!

I will be sad not to have to read your letters, make sense of drawings, check equations and occasionally chat on the phone. But I am leaving this great magazine in the very capable hands of Svetlana Josifovska, who has a superb pedigree in electronics and journalism, having only recently left the post of Editor in Chief of the Institution of Electrical Engineers' portfolio of member publications including the esteemed *IEE Review*. As she knows a lot more about the 'nuts and bolts' of electronics, I am sure you will find her choice of articles more intuitive than mine. And of course, a huge number of you responded to our reader survey and Svetlana will be taking all of your comments on board.

As she will be 'full time' and based in the office in Swanley, (I work mainly from home and am part-time) she will able to give *Electronics World* the attention it deserves. I wish her and the readers and contributors the very best for the future.

Phil Reed

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Angled pits mean ten times more data on a DVD

Physicists at Imperial College London are developing a optical disk storage technique that could hold eight to ten times more data than conventional discs.

"According to our experimental results, we can optimistically estimate that we will be able to store about one terabyte (Tbyte) per disk in total using our method," said research leader Dr Peter Török. That is for a dual-layer, double sided disc which "translates to about 250GByte per layer, ten times the amount that a BluRay disk can hold", he added.

BluRay is Philips' proposed replacement for DVDs.

Along with other DVD-replacement technologies, BluRay uses a 405nm wavelength (blue) laser to read or write a disk and store 25Gbyte/layer - five times more than a DVD.

Blue lasers store more because short wavelengths can be focussed to smaller spots to read smaller surface features on a disc. DVD players use 635 or 650nm (red) lasers, CDs use infra-red at 780nm.

According to Török, optical glass fails to be transparent below 330nm, making it unlikely anyone will bother to develop optical disc players based on lasers shorter than 405nm.

Whereas data is stored as round pits on the surface of an optical disk, with one bit of data stored as the presence or absence of a pit, Török is proposing to use oval pits, or some other shape that can add angular data - calling the scheme MODS - for multiplexed optical data storage.

"We came up with the idea for this disk some years ago," said Török. "But did not have the means to prove whether it worked."

Enter PhD student Peter Munro. Together Török and Munro developed a mathematical model for the reflected light. "We are using a mixture of numerical and analytical techniques that allow us to treat the scattering of light from the disk surface rigorously rather than just having to approximate it," said Török.

332 different angles have been differentiated from experimental pits, made asymmetric with a "sunken step," said Munro. That is over 8bits of data and the team estimates ten bits, or over 1,024 angles, will be possible .

Angular measurement is made using a polarised laser and two optical pick-ups viewing through polarisers set at right angles. The ratio of signal at the two detectors gives the angle.

"That's two photodetectors and a division," said Munro. "There is no difficult signal processing, most of it is done in the optics so data rate can be quite high."

MODS disks would cost approximately the same to manufacture as an ordinary DVD, claims Imperial.

"High density optical data storage comes in handy when manufacturers talk about miniaturisation of the disks," said Török. "In 2002 Philips announced the development of a 3cm diameter disk to store up to 1GByte of data. The future for the mobile device market is likely to require small diameter disks storing much information. This is where a MODS disk could really fill a niche."

Getting serious about gravity

Of the many satellites swarming around the Earth perhaps the most strange is Gravity Probe B, which this summer began a year-long quest to prove some of Einstein's predictions.

The theory of relativity established a century ago by Albert Einstein leads to two predictions: Curved space-time and frame-dragging.

Curved space-time is also called the geodetic effect. The mass of the Earth is said to distort space-time – like the famous rubber sheet analogy. Frame-dragging is a twist put into the Earth's gravity 'well' as the planet spins upon its axis.

The experiment to verify the existence of these properties was first suggested 40 years ago, but the extreme technology required to test the theory has only recently become available.

NASA's probe is aligned onto the star IM Pegasus and four gyroscopes are spun up to 10,000rpm.

If over the course of one year the gyroscopes change their alignment with respect to the probe, then it will be due to curved space-time on one axis and frame-dragging on another.

According to the theory, at an altitude of 640km the former effect should result in a shift off axis of 6.6arcseconds and the latter effect a shift of just 0.042arcsecond.

An arcsecond is covered by a human hair at a distance of around 20 metres.

Gravity Probe B is nothing less than a testament to the art of precise engineering.

The gyroscopes each contain a 38mm diameter quartz ball – the most perfectly spherical and homogenous objects ever fabricated. The balls have a peak-to-valley smoothness of 0.01µm, or less than 40 atomic layers.

They are so smooth that if left spinning they would take 10,000 years to reach 37 per cent of their initial rate. Their drift is better than 10^{-11} degrees/hour.

In order to measure a shift in the satellite's alignment, the gyroscope balls are coated with exactly 1.2µm of superconducting niobium. As the balls spin this creates a magnetic field along the spin axis.

A loop of wire around the gyro allows a superconducting quantum interference device (Squid) to measure changes in axis down to a resolution of 0.001arcsecond.

Because the balls are superconductors, the whole system must be kept at 1.8Kelvin using liquid helium in a 3m long 2,500litre container (called a dewar). This is despite the fact that the probe must pass in and out of direct sunlight twice per day. Evaporation from the container is controlled through one way valves and the excess gas is used to control the satellite's attitude.

Detector finds one nanogram per litre

Optical and electronic techniques have been combined at the University of Southampton to produce a sensitive water pollutant that works simultaneously on 32 chemicals.

Funded by the EU's Environment Programme, the sensor is initially aimed at detecting oestrone - a substance linked with gender change in fish and implicated in falling levels of male fertility.

"Optical sensors have great potential in simultaneous, rapid, high-sensitivity measurement of multiple pollutants in water,' said Professor James Wilkinson of Southampton's Optoelectronics Research Centre, "The biosensor chip enables us to measure a large number of low molecular weight organic pollutants, and we have successfully detected levels at below 1ng/l for oestrone, which is one hundred times better than the original project target."

The detector is based around a glass slide, into the surface of which 3x3μ waveguides are introduced by diffusing potassium ions through a mask. These ions increase local refractive index, so the waveguide acts just like an optical fibre. Beam splitters are also implanted to divide a laser beam and share it equally between 32 implanted rectangles, each 0.1x1mm.

An overall surface coating of low refractive index SiO_2 keeps light in the waveguides, except where windows are etched into it over the rectangles.

Different molecules are printed over the windows, each designed to catch a certain pollutant, previously labelled with fluorescent antibodies.

Through the thickness of the slide, optical fibres carry any florescence through laser-blocking filters to PIN diode photosensors.

Detection range is between 1ng and 1μg/l when detecting oestrone.

The antibodies are selected to fluoresce when illuminated by a 635nm (red) laser - a wavelength chosen because glass does not fluoresce when stimulated with red light.

The equipment, including pumps other glassware, is the size of a PC.

Oscilloscope hits double figures

An oscilloscope with a real-time analogue bandwidth of 13GHz has been introduced by Agilent Technologies.

However, it does not come cheap – priced in the US at $122,500 – and even the active differential probe at over $10,000 costs more than many standard scopes.

The DSO80000 series of scopes has a maximum sample rate of 40Gsample/s at the front end.

The applications for the scopes include measuring the new generation of high-speed serial bus standards such as fibre channel, serial ATA and PCI Express. These are characterised by data rates up to 10Gbit/s and edge rise times of 50picoseconds or less.

Natural gas solid oxide fuel cell

Fuel cell research at St Andrews University has taken a step closer to commercial reality with backing for a spin-out firm.

St Andrews Fuel Cells said it will develop high temperature cells with power outputs between 1kW and 5kW, with prices of $150 per kW. Current prices are around $2,000 per kW, said the firm.

In addition to the low price, the firm said it would increase volumetric efficiency and mass-power density by factors of five.

The technology is called the solid oxide fuel cell (SOFC), a ceramic device running on natural gas.

By using conductors such as scandia-stabilised zirconia and lanthanum gallate-based perovskites the firm hopes to reduce operating temperatures from 1,000°C to perhaps below 750°C, making construction easier.

One problem is avoiding the formation of carbon of the anode.

The firm sees two markets: Short term is the market for low power portable applications such as outdoor pursuits, camping, remote site or military applications. Following this is the combined heat and power (CHP) market, with higher power requirements.

"CHP units will effectively replace gas boilers, currently used for domestic heating and hot water, while also permitting the option of feeding the local electricity generation back into the grid," said the firm.

Bomb detectors capture EMC radiation

A radio receiver detector is under development at the University of Missouri-Rolla's EMC lab. The technique involves capturing electromagnetic radiation leaking from the receiver and slowing it down to make audio signals. Filtering, or pre-processing of the signals, is key to enhancing the sound of a particular receiver.

"There's way too much information to play it back in real time. Right now we're capturing 100ms of data and we're taking 10s to play it back," said Dr Todd Hubing. This research is aimed at detecting remote-controlled road-side bombs in Iraq and these there is no time to develop a sophisticated digital signal processor to analyse returns - hence slowing the signal down so an operator wearing headphones in a truck can pick the wheat from the chaff, much as sonar operators did in the second world war.

According to the university, radio receivers including those found in remote-controlled toys, wireless phones, mobiles phones, and wireless doorbells are used in bombs. The researchers believe that in a year, with funding, they could develop a system for soldiers to identify and locate radio receivers in remote areas.

"Presumably we'd be able to hear anything electronic, particularly if it had a processor in it and there was a lot of electrical activity," Hubing said. "This project started as an effort to identify automobiles based on their radio frequency emissions. This turned out to be much easier than we anticipated."

Chinese Science Park in Swansea

Technium, the Welsh technology incubator, has attracted one of China's largest science parks to Swansea University.

China's Fudan Science Park will establish its first UK base at the Technium, giving the hundreds of companies that it represents access to the UK.

"Wales presents significant and exciting opportunities and many internationally recognised organisations are being attracted to Wales as the place to do business," said Welsh economic development minister Andrew Davies. "The choice of Swansea as the first UK base for Fudan Science Park is clearly a reflection of the favourable business and research environment in Wales."

While the Fudan group wants to help Chinese companies break into the UK, the opposite route is also available, said Professor Yang Yuliang, chairman of Fudan Science Park.

"Our office in Swansea will not only provide a platform for the 200-plus companies in our Park to explore the European market, it is also a window to attract more high-tech companies in Europe to invest in our Park in China," he said.

Pictured during the signing of the deal are Professor Yuliang and Wales' First Minister Rhodri Morgan.

The Welsh Development Agency has also opened an office in Shanghai with the aim of attracting more Chinese firms to the principality.

"Wales offers Chinese companies the ideal platform from which to serve the European market of over 400 million consumers," said Morgan.

When digital really is digital

The digital stills camera, one of the industry's biggest success stories of recent years, may soon finally go truly digital. US firm OmniVision Technologies has developed a five million pixel sensor entirely in CMOS, with the claim that it equals CCD sensors in terms of image quality.

All-digital CMOS is much cheaper to produce than CCD, but the latter has always resisted being ousted from the market due to its higher quality images. CMOS has always had a higher dark current than its analogue CCD cousin and CMOS has struggled to reach the required image resolution.

Omnivision's Jason Lui said the "pixel structure diminishes dark current to unnoticeable levels, a key factor in bringing CMOS image quality to CCD levels".

The OV5610 has a 1/1.8inch optical format, a 2,592x1,944 array and pixels just 2.775μm across. An on-chip 10-bit A/D converter can run at 4frame/s.

The market for digital cameras is over 60 million units this year, rising to around 75 million in 2005. The largest segment next year with 35 per cent of the market will be at five million pixels.

"With increased performance and a lower cost point, CMOS image sensors are expected to grow at roughly seven times the rate of CCD sensors through 2008, enabling CMOS sensors to surpass CCDs for the first time in 2005," said Brian O'Rourke, an analyst at research firm In-Stat/MDR.

C coding gets rules update

The team responsible for MISRA-C, the coding guidelines for automotive engineers, has updated the rules making them applicable to other industries.

MISRA-C was first launched in 1998 comprising 128 rules to make C code safe to use in automotive applications. However, other industries such as aerospace and mining took up the guidelines.

"For the last four years a core team of software engineers has been working with the SAE, JSAE, JAMA, ISO-C panels and many experts in the embedded and software engineering industry, as well as the software tools industry, to produce the second version," said the team.

The developers have improved the rules, making them more targeted and less of the 'blanket' form. Some rules have been removed completely while others, such as those for arithmetic operations have been added.

MISRA-C2 is compatible with the original version, said the team.

0 to the speed of light in 1mm

A group of Imperial College scientists have accelerated a burst of electrons from rest to almost the speed of light in 1mm - something that would take a conventional particle accelerator ten metres, said lead researcher Professor Karl Krushelnick. At this speed each electron carries 75MeV of energy.

The technique used is wake-field acceleration - where a powerful short laser pulse strips electrons out of a puff of gas.

It has been used before, but only to produce electrons at a range of energies between 0 and 75MeV - a 100 per cent energy spread.

Using a 40fs (femtosecond) pulse from the 20TW (terawatt) ASTRA laser at Rutherford Appleton Labs in Oxfordshire, and other improvements, Krushelnick achieved a three per spread to creating a true pulse of high energy electrons. "It's the first time that a real electron beam has been generated by these methods," Krushelnick. "Ultimately our work could lead to the development of an accelerator that scientists could put in a university basement."

Scots lose access to funding

Scotland's Proof of Concept Fund, which has helped many electronics companies, has closed its doors to further applications.

The only money now available is top-up funding for projects already backed through the scheme.

This amounts to £7.4m provided by the European Regional Development Fund.

Almost 60 projects have gained a total of £33m from Proof of Concept, including III-V monolithic mm-wave ICs, RF MEMS, optical switches, silicon sensors for gas detection, and Terahertz Gunn diodes.

Proof of Concept was set up to address the gap between establishing a firm and seed funding. This was deemed to be "restricting the flow of technology from the laboratories to the market place".

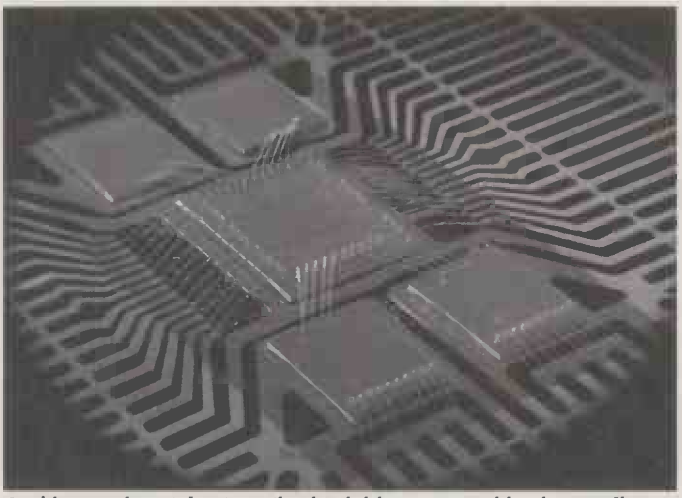
Separate IC boosts mosfet protection

To improve power mosfet protection in chips which drive heavy loads, Philips is developing a concept called 'intelligent power control', particularly for automotive use.

The car electrical environment is a hazardous place for semiconductors, with mosfet load switches likely to have on-chip over-current, -temperature and -voltage protection to increase reliability.

These so-called smart mosfets are robust, but the brute-force protection techniques used are not the best way to minimise power dissipation - important in today's shrinking electronic modules.

Automotive loads are frequently non-linear - a light bulb for instance draws ten times its running current at switch-on - so no simple protection circuit can be optimised to work best under all conditions.



In this experimental automotive load driver a central load controller operates four power mosfets, offering smart protection during faults.

Intelligent power control involves taking all the protection circuitry off the mosfet, except for the current and temperature sensor, and putting it onto a separate chip along with a load controller - which has a high-speed data link to the local microprocessor (MPU).

The control-protection chip sends the MPU current, voltage and temperature information every time the mosfet is switched, and the MPU learns how the mosfet usually responds, shutting it down if anything out of the ordinary happens.

"There will be a dynamic turn-on sequence every time," said Philips application manager Chris Hammerton, "where the microcontroller will do a pre-check, then it will turn the load on, then it will compare the current profile with stored values and what happened last time."

Piezoelectric locks and valves work in harsh environments

Gas valves and door locks constructed around piezoelectric actuators have been developed by UK firm Servocell.

The Harlow-based company says over 20 companies worldwide are evaluating or designing-in the technology, which removes the need for power hungry solenoids or motors.

Servocell developed a long-throw (up to 2mm) piezo actuator as part of a residual current detector (RCD) product. The large displacement is useful, but cuts effective force from kN to around 1N.

Therefore the firm decided to help potential customers by doing its own reference designs. The lock can withstand tonnes of pressure as the actuator does not directly hold the lock closed.

One of the major problems with piezo materials is their sensitivity to humidity, which causes the material to break down. Servocell has issued patents in this area, and believes the locks and valves can work in harsh environments.

The actuators are controlled digitally by a 0-6V input. High voltages for the piezo material are generated internally by a flyback converter.

The firm plans to offer devices that can be powered over an Ethernet link, allowing locks and other controls to be run by existing building networks.

Microscope resolves right down to the atomic scale

A sub-Angstrom resolution image looking straight down on a silicon crystal. Each dumbbell-shaped row of atoms is 0.78Angstrom apart.

Researchers at Oak Ridge National Laboratory have developed a microscope that can take images with resolution below one Angstrom (10^{-10}m).

Using a scanning transmission electron microscope (STEM) running at 300kV, the team was able to resolve down to 0.6Angstrom – right down at the atomic scale.

"Looking down on a silicon crystal, we can see atoms that are only 0.78 Angstroms apart, which is the first unequivocal proof that we're getting sub-Angstrom resolution. The same image shows that we're getting resolution in the 0.6 Angstrom range," said researcher Stephen Pennycook.

Key to the feat is a technology called aberration correction. This corrects errors introduced into the images by imperfections in the electron lenses.

Improved images such as these allow researchers to see individual dopant atoms in a semiconductor, leading to better understanding of the material's properties.

"With aberration correction you can see everything better, basically," Pennycook said. "It's always better to see what's what."

Robot walks up walls and across floors

Flexibot holds out a razor so Professor Mike Topping can shave.

Flexibot is a robot arm that can walk.

Invented at Staffordshire University by Professor Mike Topping, the arm is aimed at assisting disabled people, something which Topping has been designing machines to do for 20 years.

"I have a lot of experience in how robots interact with the disabled," he said. "Flexibot could do a good general purpose job."

Rather than walk free, the robot steps end-over-end between identical 'docking stations', each of which provide firm mechanical support and power.

When it is not walking, the free end of Flexibot extends three claws to form a hand or picks up a custom fitting to perform useful work.

To keep costs down, the arm has no absolute position sensors. Instead, it plugs itself simultaneously into two docking stations with known positions, and from there on measures its own joint angles to estimate location by dead-reckoning - which it can do initially to within 0.1mm.

If the user has a wheelchair, Flexibot can step onto it for mobile use.

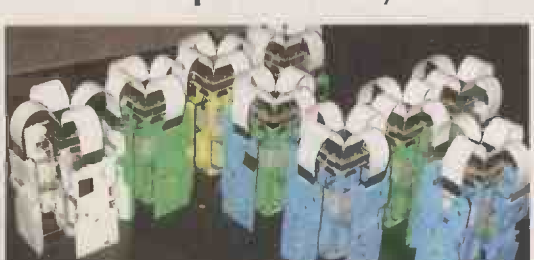
Topping does not design machines to take over from people. "This robot isn't going to force-feed someone," he said. "It brings a spoon to a comfortable position in front of the mouth."

The robot shown, as Topping freely admits, is a little chunky for domestic use. He sees the final version as daintier, but with the same distance between joints.

As a concept, Flexibot has one obvious drawback - it cannot carry anything while walking. "We have an idea for a three-armed robot," said Topping, "It's called Manx."

www.robotic-arm.com

'Transformer' robots are possible if you follow the rules



Self-configuring robots are all very well, but how do you know if a configuration is possible, or if your robot will fall apart as it changes?

Daniela Rus and collegues at Dartmouth College in the US have worked out some simple rules guarantee such self-reconfigurable robots will stick together as they change shape or move across a surface.

Published in the International Journal of Robotics Research: "These latest papers show it is possible to develop self-reconfiguration capabilities in a way that has analytical guarantees," said Rus, who is now working on more complex rules.

Pictured is a two-dimensional self-configuring machine from Dartmouth.

Essex goes high-tech

A major research and teaching facility has been built at the University of Essex.

The Networks Centre is a purpose built five-storey building with over 3,000m^2 of floor space, designed to house the academic and research staff of its departments of electronic systems engineering and computer science.

Within the centre are research laboratories for telecommunications, audio engineering, computer vision and virtual reality, as well as a robotic arena for land-based and flying robots and an intelligent flat, known as iFlat.

iFlat is fitted with computers and sensors which allow it to learn the user's preferences. "It learns, for example, to switch off lights and close blinds to suit the occupant," said the University. "The building will also facilitate inter-departmental research, for example with the cognitive science group in the Department of Psychology."

Professor Huosheng Hu and his football-playing robotic dogs can now play in a high-tech robotic arena within the University of Essex' purpose-built technology building.

photo: Chris Mikami

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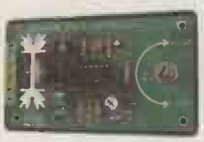

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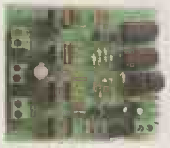

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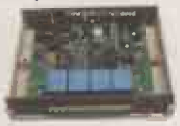

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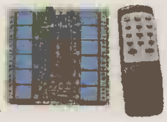

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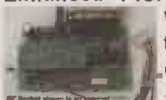

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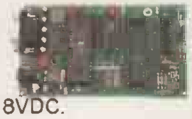

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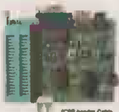

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Simulating power MosFets

In this, the third of a four part series using the Micro-cap6 software, Cyril Bateman introduces a method enabling any Spice user to develop a 'self heating' MosFet model in which the MosFet model's junction temperature and characteristics respond in real time, just like a real MosFet

Spice thermal simulations can take three forms, the simplest constantly monitors the power dissipated within a transistor. By adding a few more components, it is possible to expand this to also monitor the junction temperature, providing a clearer picture as to how much stress is being applied to the device. For this one must have a reasonable model for the transistor's thermal path from its junction to the heatsink. Armed with this additional information and a thermal model for our heatsink, a much improved lateral power MosFet simulation model can be devised, one which not only monitors power dissipation and junction temperature, but uses the calculated junction temperature to continuously update the actual MosFet models characteristics, which like a real transistor now change with change of junction temperature in real time, the subject for this article.

To monitor junction temperature, we add an extra terminal or node to the traditional drawing used to signify a power MosFet and two further terminals one to input a voltage representing the ambient temperature, the second to attach the model for our heatsink, to provide a 'self heating' model. While seeming quite a bit more complex than our traditional power MosFet schematic symbol, our simulations can now closely mimic real components in actual use. Released from the restrictive Spice global simulation temperature, each self heating transistor can now operate using realistic junction temperatures and characteristics. Using such a complete model now becomes simple and straightforward. See **Figure 1**.

Key to this model is to first calculate in real time the instantaneous power dissipated in the transistor, which combined with the

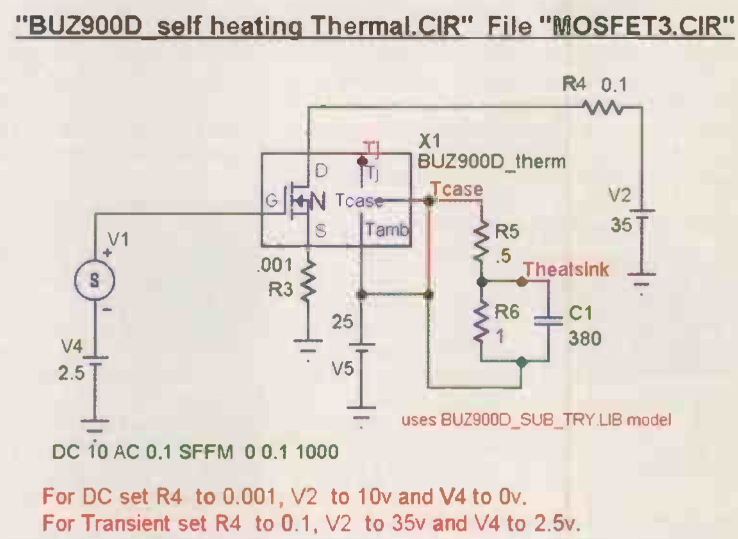

Figure 1: This working schematic illustrates how easily my self heating subcircuit model can be used to accurately model a power MosFets behaviour in end use, fitted with a heatsink. With the link shown in red the circuit emulates the 'infinite heatsink' used to establish datasheet parameters. Removing the red link we model the effect on junction temperature and MosFet behaviour with a heatsink and washer. This X1 symbol represents a netlist some two A4 pages long, the circuit shown in Figure 6.

transistor and heatsink thermal characteristics, results in the transistor junction temperature, calculated in real time. These benefits are not restricted to power MosFets, but could be applied equally well to a BJT transistor.

Power dissipated

The instantaneous power dissipated in any component is obtained by multiplying its through current by the voltage across its terminals. Using Spice2 this instantaneous power can be plotted in real time, converted to RMS or averaged as required.

To facilitate developing our model, we use only those Spice2 devices which can function from within a schematic drawing and a subcircuit model. In last month's model we used a PSpice functional voltage source E_4, to calculate the value of an expression. We now use this device to multiply the voltage across the transistor by its through current, to output a voltage representing instantaneous power dissipation, as 1 volt/watt.

Thermal modelling

We are already familiar with the concept of thermal resistance, where the flow of heat through a medium, whether for double-glazing, roof insulation or a heatsink, can be described in °C per unit of heat. For electrical circuits the unit used is Watts so our heatsink is described as e.g. 1°C per watt, indicating it can dissipate 1 watt by convection, radiation and conduction, for each 1°C increase in its temperature above the local ambient temperature.

All materials possess some mass or

weight so can absorb and store heat, just as a capacitor can accept and store an electric charge. Some materials store more heat per gram weight than others, so are rated for their specific heat. The quantity of heat stored depends on the product of weight and specific heat. Clearly a transistor in a TO3 case weighing some 12 grams stores much less heat, so changes temperature more quickly, than say a 400 gram heatsink.

Over time the quantity of heat stored in any component reduces by radiation, convection or conduction, just like our heatsink so can be assigned a °C per watt rating. For our simulation we use a common analogy where rate of heat flow is controlled by a resistance value and the amount of heat stored is related to capacitance in Farads as shown Table 1.

To model heat flow through a chain of resistors, the voltage calculated to represent the transistor's instantaneous power dissipation must be converted into a current, using a voltage to current converter. This current flowing to ground through our thermal resistor network, develops a voltage across the network, according to the transistor power dissipation and this resistance, representing the transistor junction temperature.

Equivalent circuits

Continuing our analogy, the amount of heat stored and the rate of heat flow is usually modelled using one of two circuits, either an R/C ladder network or a chain of resistors each in parallel with a capacitor. See **Figure 2**.

The thermal transfer of a transistor can be measured by raising its temperature and stabilising at a known value. Having removed the heat source, temperature measurements are taken against time, as the transistor cools to room temperature. From this cooling curve a series network of a chain of parallel resistors and capacitors, the cooling curve thermal model for the transistor can be calculated as shown in **Figure 3**.

Unfortunately while this model provides an accurate representation of the device as tested, if we then need to add or change a heatsink the whole network must be recalculated. If we first convert the original cooling curve network into the equivalent R/C ladder network, this recalculation can be avoided. It is now a simple matter to add another equivalent circuit, for a heatsink etc. using another ladder network, without changing the thermal model for our transistor as shown in **Figure 4**.

The Fairchild FDP038 model[1] used

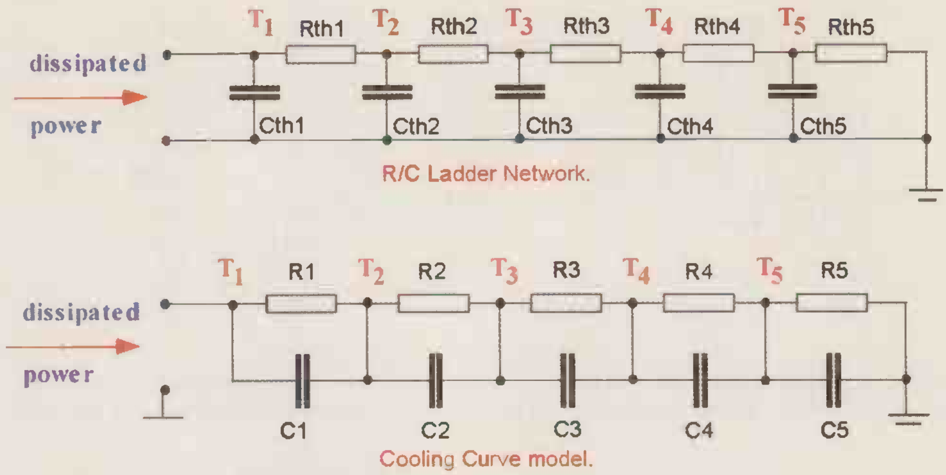

Figure 2: Two common circuits used to model the thermal transfer path in a transistor. Component values for the cooling curve circuit can be derived directly from a cooling curve measurement, favoured by semiconductor makers, unfortunately this circuit cannot easily be fitted with a heatsink. Component values for the R/C ladder network can be used unchanged together with established heatsink models.

Table 1: Common thermal electrical equivalents used for thermal modelling

Thermal equivalent	Units	Electrical equivalent	Units
Resistance	°/watt	Resistance	Volts/amp
Power	Watts	Current	Amps
Temperature	°C or Kelvin	Potential	Volts
Thermal capacity	Watt-seconds/°C	Capacitance	Coulombs/volt

Table 2: Thermal path of TO3 and TO247 transistor packages.

TO3 package		TO247 package	
R_{th} Ohms	C_{th} Farads	R_{th} Ohms	C_{th} Farads
0.06	0.008	0.058	0.008
0.1	0.04	0.078	0.038
0.28	0.4	0.216	0.391
0.05	4.5	0.049	9.7
Total = 0.5°C/watt		Total = 0.4°C/watt	

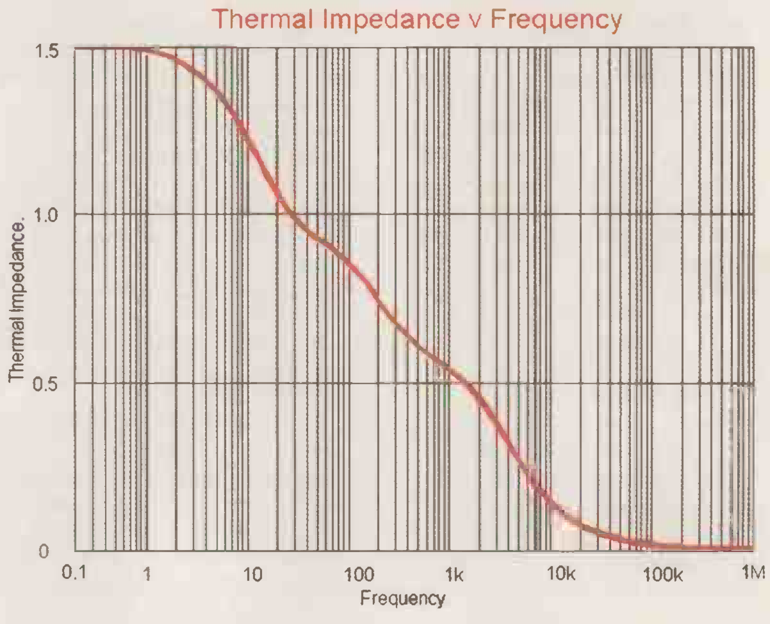

Figure 3: A typical cooling curve plot adapted to calculate the component values for the cooling curve model of Figure 2.

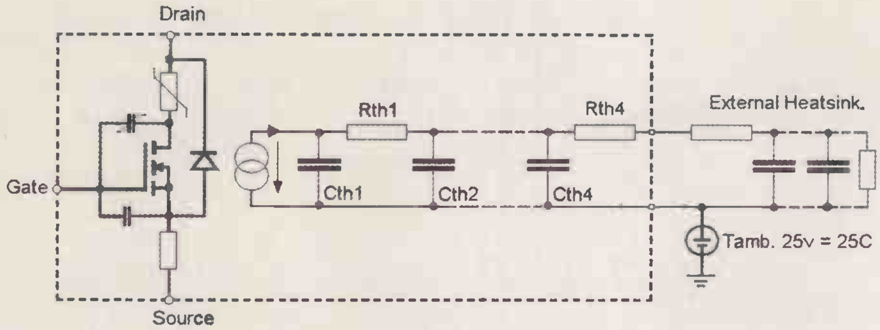

Figure 4: A heatsink model is easily attached to the R/C ladder network thermal model for a transistor. This figure illustrates the basic method used to calculate transistor junction temperature for this article.

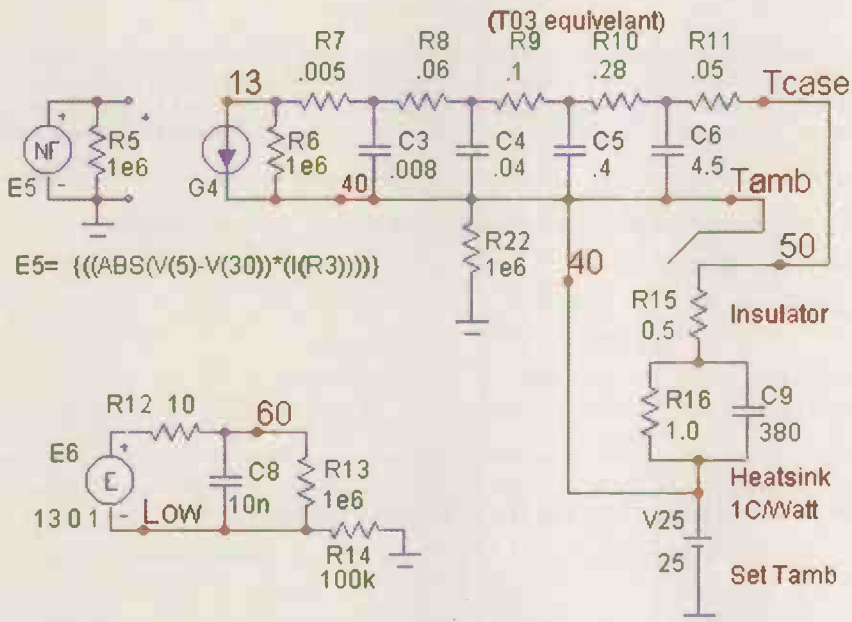

Figure 5: A practical implementation of the figure 4 schematic. E5, left of figure, calculates power dissipated in the MosFet as 1 volt/watt, by multiplying the voltage across the transistor by its through current. G4 is a voltage to current converter; its output current develops a voltage across the thermal path resistors R7 through R11 then to ground. The voltage developed at node 13 represents junction temperature as 1 volt °C. Connecting together nodes Tcase and Tamb, models an infinite heatsink.

to introduce the concepts in my last article, provides only this cooling curve thermal model. The BUZ transistors I wished to model have a 0.5°C/watt thermal rating, but are not provided with any thermal model. After more frantic searching I found a R/C ladder network thermal model for the SKW25N120, a similarly rated power device built into the TO247 case. Amending its final R/C stages to match the thermal mass of the TO3 package produced a realistic thermal model for our TO3 transistor as shown in table 2.

The self heating model

Adding a few more components, this thermal network can be used to calculate and control our MosFet junction temperature and its characteristic behaviour with temperature as in figure 1, resulting in the most realistic Spice2 simulation model possible, one closely aligned to the behaviour of a real MosFet in an actual circuit. In addition we now provide outputs in real time for its junction temperature, as the simulation proceeds. This model clearly provides the most realistic and accurate method possible using Spice2, to calculate distortion in the output stage of an audio power amplifier circuit. See **Figure 5**.

For this self-heating thermal model I used Spice2 devices compatible with my MC6 simulator also PSpice and PSpice equivalent simulators. My choice of behavioural devices was influenced by their ability to output a value from within a schematic, in order to more easily develop the model and also when called from within a subcircuit, as needed for the model's eventual use. Some useful behavioural models can be called either from a schematic, or subcircuit model but cannot be used to output a value from both.

Both Infineon and Fairchild provide thermal models for some of their switching transistor products for use with the latest PSpice or Saber simulators, but to model our audio power MosFets, we must once more dirty our hands, carving a model to suit our needs. As with my last article, I preferred to use the Fairchild modelling method because I could provide the essential data for it from datasheets. Other models that may ultimately prove superior by reducing convergence problems could not be used because they require access to manufacturing data.

A Thermal 'N' MosFet model

For last month's model we used three resistors, two in the drain circuit and one in the source circuit, whose value was forcibly changed according to the Spice2 controlled simulation temperature. For a self heating model, we need to describe the value of similar resistors by temperature using a behavioural resistor model referenced to a controlling voltage node, representing the MosFet's junction temperature as 1volt per °C.

Neither MC6 nor PSspice provide this and Spice2 resistor values can only be directly controlled by the simulation temperature and not by any voltage node. Dynamic temperature control of the TC of the MosFet's resistive elements, cannot be provided without using a controlled behavioural model to replace the three temperature controlled resistors in the last article's model.

Using Ohm's law the function performed by a resistor connected between two circuit nodes, 5 and 7, can be modelled without using an actual resistor: -

I = Voltage (5-7) ÷ (Required Resistance value and temperature coefficients)

In PSpice and equivalent versions of Spice2, the model for a voltage controlled current source can be used with its two voltage controlling nodes connected to its current output nodes. By using the nodes of a current source for its voltage control, sensing the voltage between nodes 5 and 7 this resistor behaviour can be simulated as shown **Figure 6**.

Taking the G1 expression shown in figure 6 as an example, we have the voltage between nodes 5 and 7 expressed as V(5,7) the Spice expression to indicate voltage difference between two nodes, as numerator for the above.

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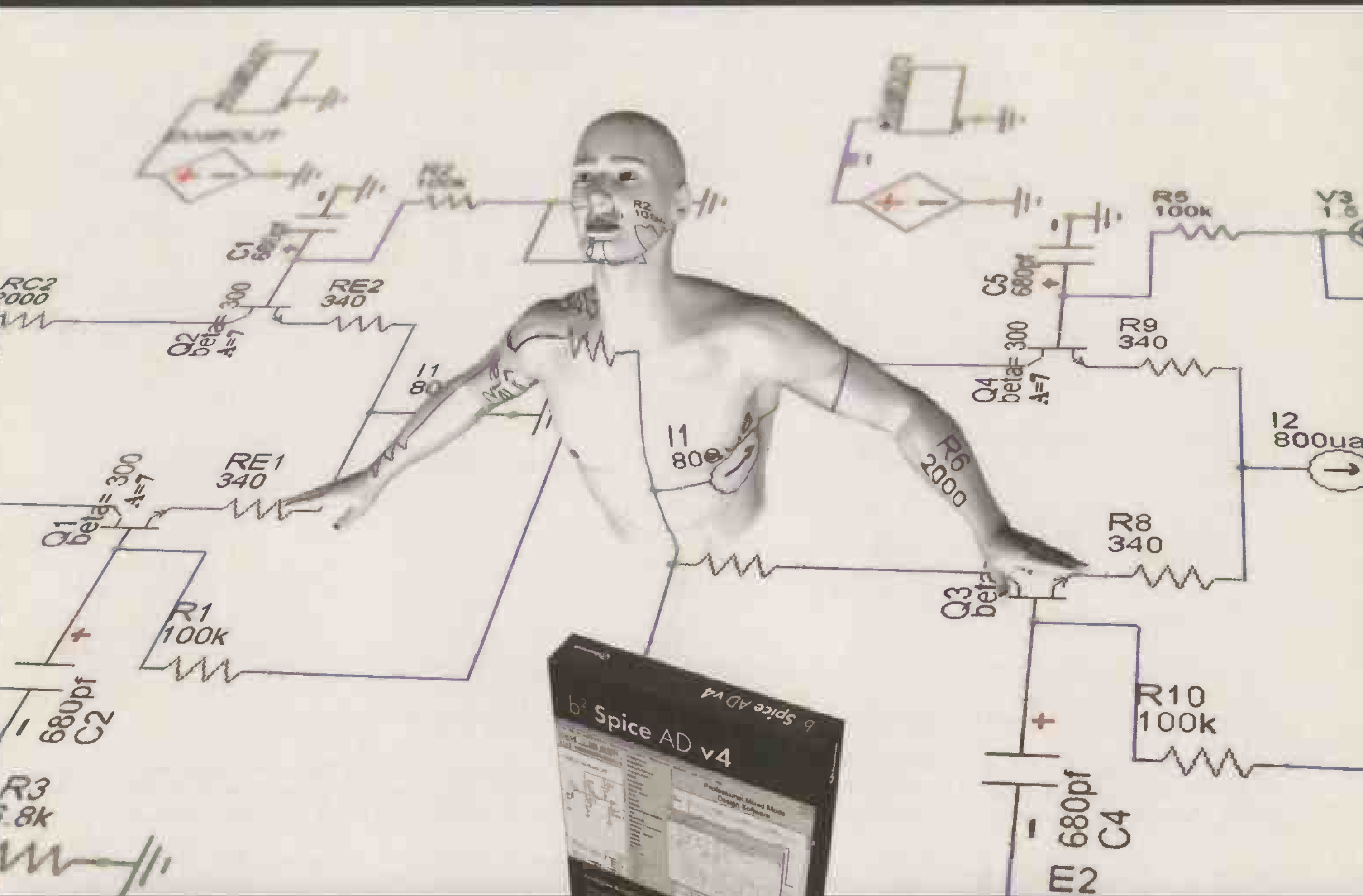

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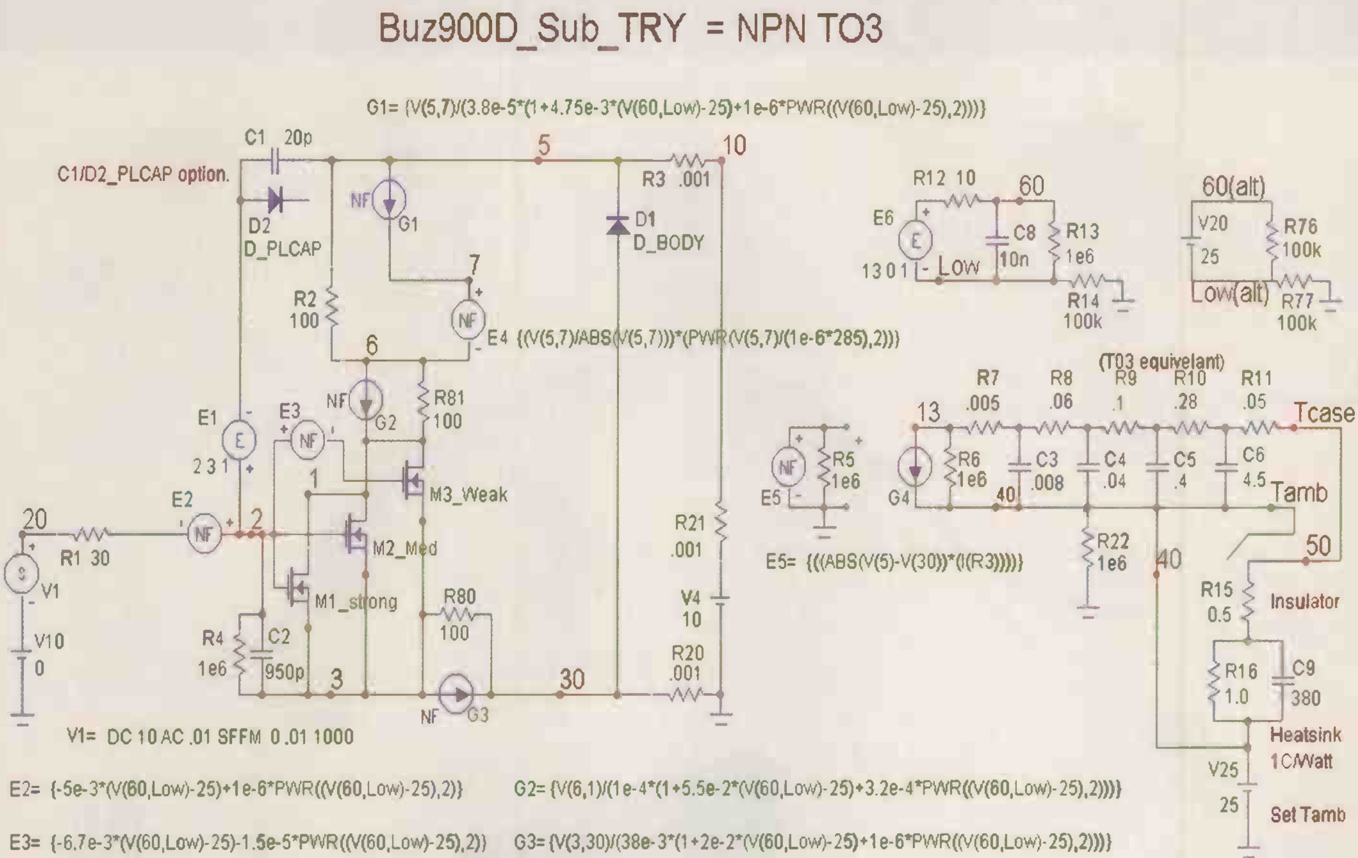


Figure 6 : My finished schematic model for the BUZ900/01D lateral power 'N' MosFet. This circuit works correctly as shown, used either with the C1 or the C1 diode model, chose one only deleting the other. Nodes 60 and Low control the models junction temperature. Move nodes 60 also Low to the alternate positions shown to fix the transistor junction temperature while developing the model. With nodes 60 and Low as shown the model is self-heating.

I = {{V(5,7)÷(3.8E^{-5} × (1+4.75E^{-3} × (V(60,LOW)-25) + 1E^{-6} × PWR ((V(60,LOW) - 25),2)))}}

The denominator for the expression shown above, is a lengthy equation which shows the required resistance value as 3.8E^{-5} Ohms plus a long expression to calculate the value for its two Spice temperature coefficients, related to the increase in transistor junction temperature above 25°C, shown as ((V(60,Low)-25).

In figure 6 of my last article, an almost identical resistance value, 3.91E^{-5} Ohms with TC= 1E^{-3}, 5.05E^{-5} was used for the temperature controlled resistor R3_slc1.

Using a voltage controlled current source and controlling its voltage input nodes provides suitable models to replace the three TC controlled resistors in last month's circuit. These appear in our self-heating thermal circuit as G_1, G_2 and G_3. We also need to replace last month's voltage controlled voltage sources E_2 and E_3 with voltage output function sources, similar to that already used for E_4.

With these exceptions and the thermal modelling/control elements, our self-heating model closely mimics last month's simpler circuit so can use similar values for threshold voltage and KP to those already developed. However it must be remembered your simulator may differ in the way it uses these 'G' voltage controlled current source devices, so do check for this in your manual.

Fundamental to this new model is the thermal modelling and control circuitry shown to the right of last month's now updated version. This is quite different from that used by Fairchild[1] or Infineon[2], so represents my original contribution to this circuit model.

Thermal monitoring circuit

The real time power monitoring circuit was easily arranged using just a simple voltage controlled source, E_5, to calculate the product of voltage drop across the MosFet and its through current. As a result the voltage output from E_5 and developed across resistor R_5, represents the instantaneous power dissipated in the model as 1 volt/watt. To simulate thermal behaviour however we need to model current flow through the thermal path resistors and capacitors, so cannot directly use this voltage.

Converting the voltage output from E_5 into a current using the voltage controlled current source G_4 also allows us to float this current flow circuit above ground, to insert a voltage source to ground representing ambient temperature. The voltage developed by this G_4 output current flowing through the ladder network, R_7 through R_{11} then to earth, represents our TO3 0.5°C watt, thermal path from junction to case. If we connect together the two nodes T_{case} and T_{amb}, to model an infinite heatsink, then apply 25V between T_{amb} and ground as shown, we model the thermal conditions assumed by the transistor maker when drafting the datasheet.

As a result when the transistor is internally dissipating 100 watts, we should find the G_4 current develops 50V across resistors R_7 though R_{11} to the nodes T_{case}/T_{amb} which with 25V to ground for 25°C ambient, results in a voltage between node 13 and ground of 75V, e.g. (100 × 0.5) +25, the expected 75°C transistor junction temperature.

This voltage is used to control the values of the behavioural resistors G_1 through G_3 also the voltage controlled voltage sources E_2, E_3. By these means we have replicated the functions performed in the simpler models of my last article, but instead of relying on the global Spice simulation temperatures to control the modelled transistor characteristics, the model now becomes an independent, self-heating, self-contained component. Calculation of power dissipation combined with the thermal model, is able to calculate the transistor junction temperature in real time. This 'self heating' junction temperature determines the model characteristics, as does a real transistor, in real time.

Thermal models used

The capacitors C_3 to C_6 represent the thermal capacity of the transistor assembled in a TO3 case. By absorbing/storing heat as electrical energy, these capacitors delay the calculated junction temperature rise exactly as does the heat absorbed and stored in the component parts of our transistor and its TO3 case. Without these capacitors the calculated junction temperature would rise and fall unrealistically within each cycle.

Using this current, it is essential we always provide a resistive path to ground, with or without attaching any heatsink, the T_{case} node must never be used unconnected. By shorting T_{case} to T_{amb} we represent an infinite heatsink at 25°C ambient temperature, as used for the datasheet. To represent a transistor insulator or mica washer, having typically 0.5°C/watt thermal resistance, we insert a 0.5Ω resistor between T_{case} and T_{amb} then to represent our 1°C/watt heatsink we insert a 1Ω resistor between this mica washer and the T_{amb} node.

Calculating these capacitor values

Aluminium has a specific heat around 0.9 to 0.95 depending on the grade of aluminium used. The 1°C/watt heatsinks I purchased for these tests (Farnell 150-016) weigh some 400 gms so have a thermal capacity equivalent of (400 × 0.95) or 380 Farads, for our thermal simulation. This 380 Farad capacitor should be connected in parallel with the heatsink thermal resistor as shown. In similar fashion we could assign a capacitor for the insulating washer, but its thermal capacity is negligibly small.

One other point needs explaining, between R_{12} and R_{13} in figure 6 you will see a node '60' marked in bold, also a node 'Low' at the junction of C_8, R_{13}, R_{14}. The voltage between node '60' and 'Low', isolated from ground by R_{14}, represents the MosFet junction temperature. The voltage controlled voltage source E_6 replicates the voltage to ground developed by the thermal resistor path, providing the semi floating voltage difference needed to control our Mosfet models characteristics equations.

Between V_{20} and R_{76} you will see another similar node marked as '60(alt)'. Node 60 is used two ways, first as shown on the drawing it represents the MosFet junction temperature, so controls the transistor's thermal behaviour according to the power dissipated as calculated by E_5, and the current to earth through the transistor thermal path transistors and heatsink. Thus enabling our real time self-heating thermal simulations.

Its second use is if we move node 60 to the 60(alt) position also node Low to the Low(alt) position we can now control the MosFet junction temperature and other characteristics at whatever fixed temperature we chose. Simply adjust the voltage of V_{20} as needed. V_{20} is scaled as 1 volt per °C, hence the 25V shown sets the model to 25°C, exactly as needed when developing/checking the model's performance.

The three nodes shown as 60 for junction temperature, 50 for T_{case} and 40 for T_{amb} are linked out from our final subcircuit model via the schematic shape used, to be accessible in our final schematic simulation circuit as shown in figure 1.

Having described the thermal circuit behaviour, little more about this self-heating model needs

Figure 7: Used in self heating mode with an infinite heatsink at 25∞C ambient, this DC sweep of gate voltage from 0 to 8 volts with 10V drain-source, using the figure 1 circuit, shows how the junction temperature rise lags behind the power dissipated in the transistor, with increasing gate volts. Initially at room temperature, by the end of this sweep, junction temperature has risen to near 92°C.

Table 3 Thermal data common materials used

	ρ[g/cm³]	λth[W/m.C]	c[J/g.C]
Silicon	2.4	140	0.7
Solder (Sn-Pb)	9	60	0.2
Cu	7.6-8.9	310-390	0.38-0.42
Aluminium	2.7	170-230	0.9-0.95
Alumina	3.8	24	0.8
FR4		0.3	
Thermal Paste		0.4-2.6	
Insulating washer		0.9-2.7	
Steel	7.85	79 0.	0.452

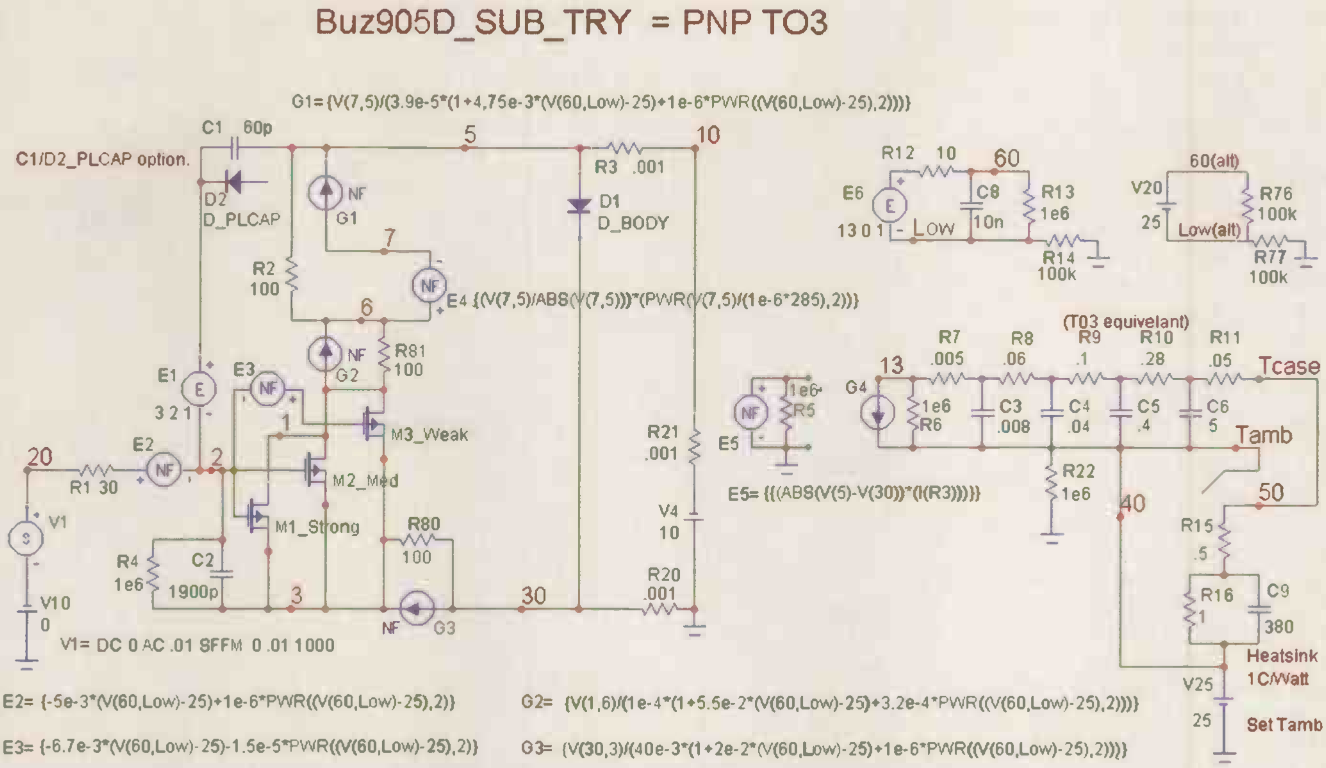


Figure 8: Having completed a model for the 'N' MosFet, the values used in the equations and the MosFet parameters were used to start the 'P' model development. However as can be seen in the figure, the polarity of a few components must be reversed and the three 'N' MosFet devices replaced with 'P' types. No changes are needed for the thermal control network.

explaining. Apart from the G_1, G_2, G_3 behavioural model resistors and E_2, E_3 that now use expressions, the model mimics the thermal model described in my last article. This self heating subcircuit model is developed following the methods already described with its junction temperature controlled by moving nodes 60 and Low to the alternate position, except we now must edit the numbers in a few equations. My MC6 version of Spice2 is particularly fussy about the numbers and types of brackets used for these equations, which do need care when entering into the circuit. The equation values used to model my BUZ900/01D 'N' MosFet[3] or its direct equivalent the EC-20N16/20[4] are as shown in the drawing, Figure 6.

Two final points remain, in this drawing I show a capacitor used to model the drain/gate capacitance, which I believe suffices for audio modelling. I also show the more accurate diode/capacitor model, its polarity must be as shown, which is essential when modelling a fast switching circuit. Use one only and delete the alternative before running the model otherwise Spice will complain about disconnected circuit nodes.

Having developed and checked the model for accuracy, a subcircuit netlist should be exported, edited as needed, then entered into your simulator library as previously described. You can now use the model in your simulations. See **Figure 7**.

Making a 'P' MosFet model

This follows exactly the procedure used for the above except as for last month's 'P' model, the three Level-1 'N' MosFet devices must be replaced by 'P' types and the polarity of some components, as shown in the figure, should be reversed. See **Figure 8**.

It is not necessary to amend any of the thermal path components, these can be used with the 'N' or 'P' type MosFet or indeed almost any similar TO3 device, MosFet or BJT.

When complete, exactly as before export and edit the netlist ready for your subcircuit model, then insert this thermal 'P' model into your simulator library ready for use.

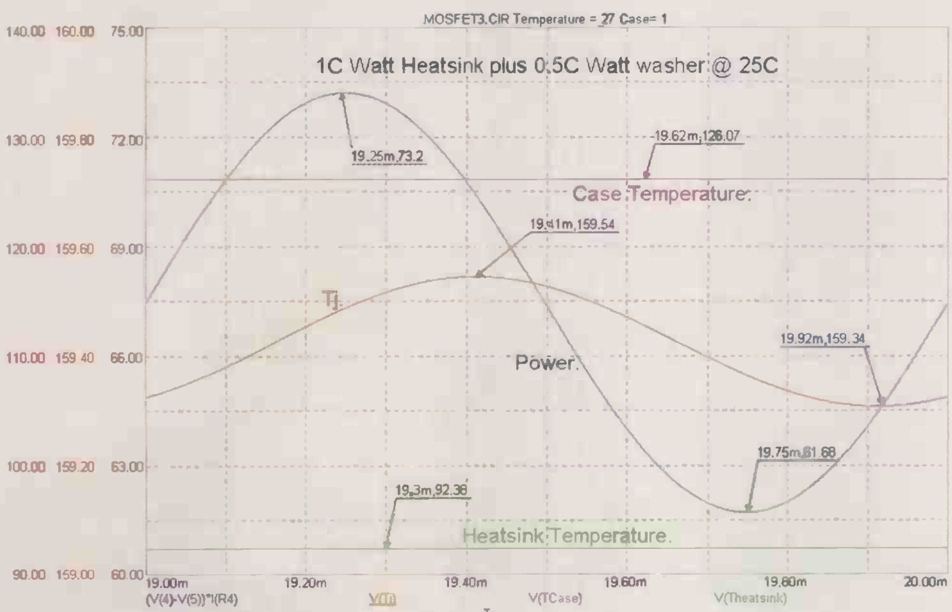
Proving the self-heating model

Having produced a self-heating thermal model, how can we prove it works correctly? The method I used was simple, with a sinewave stimulus it is easy to manually calculate the power dissipated by the subcircuit MosFet model from the product of its voltage drop between the external drain and source terminals shown in the schematic and its through current, when used with an infinite heatsink at a 25°C ambient. The maker's datasheet shows the TO3 thermal resistance as 0.5°C/watt, so simply multiply the power dissipated by 0.5 then adding 25 for ambient temperature to the result we obtain junction temperature. Happily this proved correct, so the model was ready for use. See **Figures 9 & 10**.

Simulating Distortion

To accurately model distortions in a power amplifier with MosFet output devices it is essential our model accurately predicts the device's behaviour for small drain currents while subject to the large drain-source voltages used in a power amplifier. Datasheets mostly concentrate on larger drain currents with much smaller drain-source voltage. For example, the datasheet transfer curves for my chosen devices assumed a 10V drain-source voltage and plotted currents up to their 16A maximum, consequently the curves for currents below 3A were cramped and difficult to read accurately. The drain-source voltage used in a power amplifier would be some 35V or more and a 3A transistor current in a typical output stage is sufficient to produce some 50W of exceptionally low distortion audio. Using increased supply voltages and staying within the device's square law characteristic, more than 100W may be produced.

Figure 9: Proving the models do work. This 20th cycle transient simulation using the figure 1 subcircuit model and infinite heatsink, shows power dissipated peaking at 107.5W and 93.1 minimum or 100.3W mean power, giving a theoretical junction temperature of 75.1°C, close to the models 74.77°C, which temperature has not yet fully stabilised.



10) Removing the red link in figure 1 to now use the 1°C/watt heatsink and a mica washer, we see how dramatically the transistor junction and case temperatures increase. With much lower mean power dissipation of 67.5W, we see the junction temperature now exceeds the 150°C maximum rating for this device. The TO3 case temperature has risen to 126°C and the heatsink to a very hot 92°C.

Refining our subthreshold data

Using the methods described in this and my last article, power MosFet models more suited to modelling audio distortions than supplied with most simulators, can be easily produced using only data extracted from the datasheet. However it is possible you may feel, as I did, the curves for small values of gate voltage and drain current could better match your devices.

After some consideration and realising that used in an audio amplifier the junctions will inevitably be well heated, I decided to mount my devices onto a known heatsink using a known thermal washer. Thus I could measure drain currents using a steady 10V DC drain-source voltage, adjusting the gate voltage as needed to obtain various drain currents up to the 3A maximum which interested me. Then calculate the expected junction temperature rise for each current measurement. By raising the simulation temperature slightly above room temperature as needed to equal the calculated junction temperature for each measurement, I could now tweak the '25°C' subthreshold curve to match these measured values.

I measured six pairs of devices, obtained from three different batches. Each MosFet conducted slightly, around the 0.25mA current suggested by IRF for measuring threshold voltage, even with zero gate-source voltage. Choosing the values for an intermediate device I amended the transfer curves by small adjustments to the threshold and KP values for all three transistor models. Thus providing the best possible model for the subthreshold region, so critical for realistic crossover distortion simulations.

This was easily performed using a 10V power supply, a voltage and current meter and a few resistors to adjust the gate voltage. For this I used my Muirhead decade resistor box, but lacking a similar box a small potentiometer used as a variable resistor would work equally well. See **Figure 11**.

I now have models closely aligned to actual measured values for low drain currents, but to attain this I was forced to accept some divergence from the datasheet at high currents, approaching the saturation region. For my needs that does not matter, since I never expect to use such high drain currents.

When it came to measuring the actual drain currents for the 'P' samples, I found notable differences, compared to the 'N' types. Whereas the 'N' types conducted some 0.25mA at zero gate voltage, increasing steadily with small increases in gate voltage, the 'P' types also conducted some 0.25mA at zero volts but this stayed little changed until almost 0.3V gate-source voltage was reached. This approx. 0.3V difference increased slightly at each voltage step, becoming near 0.65 volts at 2A drain current. Almost as though an additional Schottky diode existed in the source path of the 'P' transistor.

This voltage difference meant the threshold voltage for the weak MosFet had to be moved significantly compared to that for the 'N' type. In some cases the polarity of the controls E_2 or E_3 may need to be reversed to account for this.

Convergence

Convergence is a perennial problem using Spice, except when using the simplest models. This model despite its features is still relatively easy to converge, but is more difficult than last month's simpler model, because controlling the G_1 through G_3 behavioural resistors implies using a closed loop calculation, similar to that needed when calculating a negative feedback loop circuit. However, should convergence problems arise, then adjusting your global defaults away from their default settings but perhaps rather more than suggested for the simpler circuit, should provide a solution.

These circuits worked well and ran quickly in my MC6 simulator, in a quite modest computer, a 750MHz Athlon based system. These three articles represent my many weeks full time experimentation, not so much about the devices themselves as finding the modelling information needed and overcoming many problems I found within the self heating/thermal modelling circuits, using my MC6 simulator.

However, this timescale should not put off any reader interested in creating models for himself.

Following my methods, a couple of days' work should easily produce a pair of working subcircuit models, including the time needed to first work through the FDP circuitry to prove compliance with your chosen simulator and your learning curve. While this may seem a long time compared to obtaining and using the maker's models, if your need is to model circuit distortion with acceptable accuracy, I for one spent considerably longer time, unsuccessfully seeking suitable models.

These same techniques of course could be applied equally well to devising self heating, thermal simulation models for BJT transistors, but now having satisfied my needs and exhausted all my available time, I leave that task to others.

My next article covers using these models, to simulate amplifier distortions, including comparative results from the models 'supplied' in my simulator, my non-thermal and self heating thermal models, comparing the results against measurements on an actual amplifier, my modified Maplin 100 watt[5].
See Figure 12.

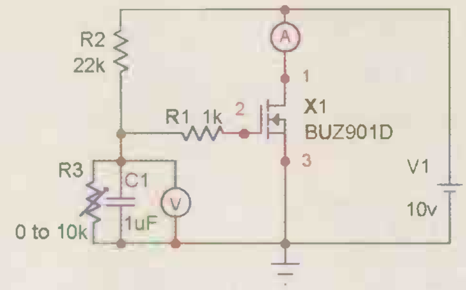


Figure 11: This simple test circuit allows direct measurement of drain current by gate voltage, for the critical subthreshold region that with three overlapping curves in the datasheet, was almost impossible to decipher.

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http//www.fairchildsemi.com/models/PSPICE/Discrete/index.html
2. Infineon Technologies
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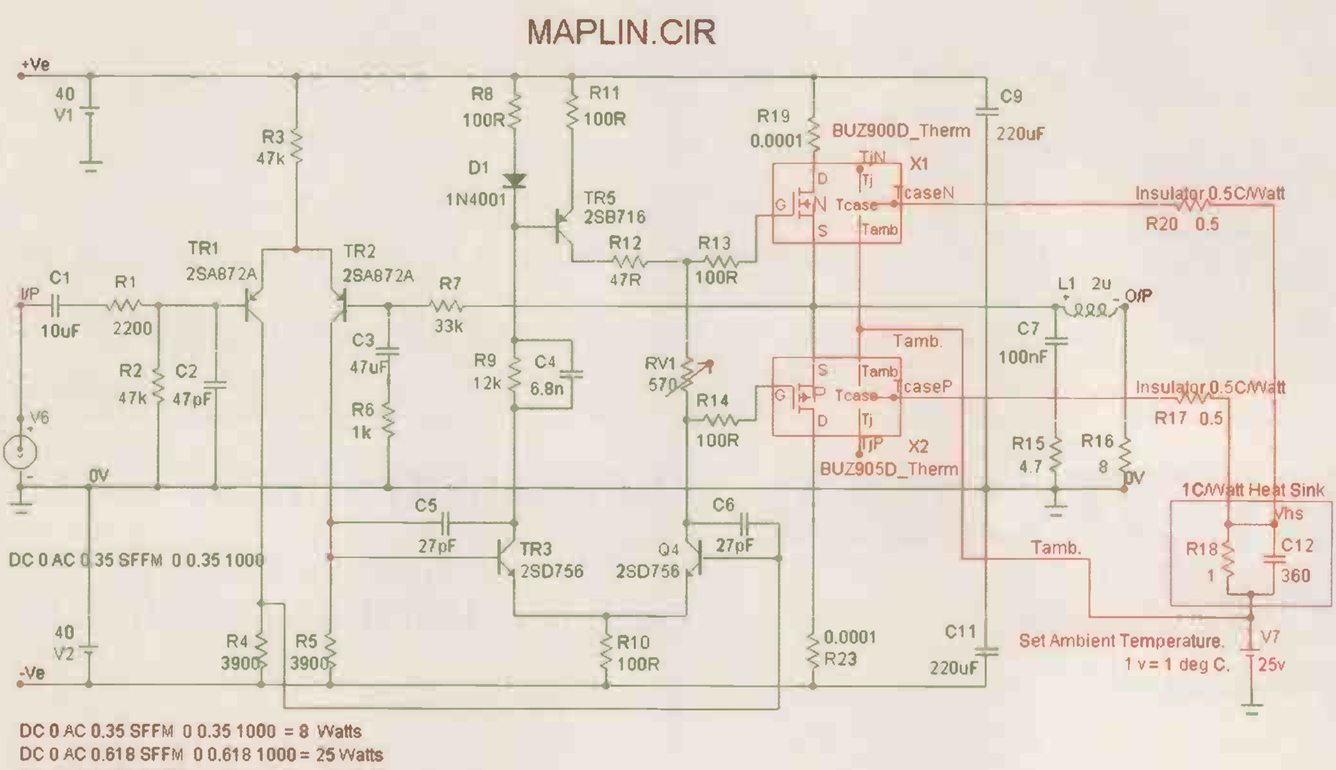


Figure 12: My modified Maplin amplifier modelled using a self heating pair of BUZ900D and 905D lateral power MosFets, together with their associated heatsink and insulating washers. While a little more crowded than using conventional models, this circuit predicted a -91.4dB second harmonic and -91.6dB third harmonic, for 0.00418% THD, slightly lower but satisfying close to the distortions measured on the actual amplifier at 25 watts output power into an 8W load. Justification indeed for my modelling efforts.

Component models used

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

```
BUZ905/06D models used:-
E1 9 2 2 3 1
E2 8 2 VALUE = {{-5E-3*(V(60,LOW)-25)+1E-6*PWR((V(60,LOW)-25),2)}}
E3 4 2 VALUE = {{-6.7E-3*(V(60,LOW)-25)-1.5E-5*PWR((V(60,LOW)-25),2)}}
E4 6 7 VALUE = {{(V(7,5)/ABS(V(7,5)))*(PWR(V(7,5)/(1E-6*285),2))}}
E5 12 0 VALUE = {{(ABS(V(5)-V(30))*(I(R3)))}}
*
G1 7 5 VALUE = {{V(7,5)/(3.9E-5*(1+4.75E-3*(V(60,LOW)-25)+1E-6*PWR((V(60,LOW)-25),2)))}}
G2 1 6 VALUE = {{V(1,6)/(1E-4*(1+5.5E-2*(V(60,LOW)-25)+3.2E-4*PWR((V(60,LOW)-25),2)))}}
G3 30 3 VALUE = {{V(30,3)/(40E-3*(1+2E-2*(V(60,LOW)-25)+1E-6*PWR((V(60,LOW)-25),2)))}}
G4 13 40 12 0 {-1}
*
.MODEL MstroMOD PMOS (LEVEL=1 LAMBDA=0.001 VTO=-2.22 KP=8.75 L=1U W=1U RS=0.3 IS=1E-30 TOX=1
RG=10 N=10 T_ABS=25)
.MODEL  MmedMOD PMOS (LEVEL=1 LAMBDA=0.001 VTO=-0.578 KP=2.75 L=1U W=1U RS=0.395 IS=1E-30 TOX=1
RG=1.36 N=10 T_ABS=25)
.MODEL  MweakMOD PMOS (LEVEL=1 LAMBDA=0.001 VTO=-0.28 KP=0.38 L=1U W=1U RS=2.5 IS=1E-30 TOX=1
RG=25 N=10 T_ABS=25)
*
.MODEL D_BODY D (IS=2.4E-11 N=1.04 TT=1E-9 CJO=4.35E-9 M=0.54 XTI=3.9 T_ABS=25)
.MODEL D_PLCAP D (IS=1e-30 N=10 CJO=300E-12 M=0.47)

Buz900/01D models used:-
E1 2 9 2 3 1
E2 2 8 VALUE = {{-5E-3*(V(60,LOW)-25)+1E-6*PWR((V(60,LOW)-25),2)}}
E3 2 4 VALUE = {{-6.E-3*(V(60,LOW)-25)-1.5E-5*PWR((V(60,LOW)-25),2)}}
E4 7 6 VALUE = {{(V(5,7)/ABS(V(5,7)))*(PWR(V(5,7)/(1E-6*285),2))}}
E5 12 0 VALUE = {ABS{((V(5)-V(30))*(I(R3)))}}
*
G1 5 7 VALUE = {{V(5,7)/(3.95E-5*(1+4.75E-3*(V(60,LOW)-25)+1E-6*PWR((V(60,LOW)-25),2)))}}
G2 6 1 VALUE = {{V(6,1)/(1E-4*(1+5.5E-2*(V(60,LOW)-25)+3.2E-4*PWR((V(60,LOW)-25),2)))}}
G3 3 30 VALUE = {{V(3,30)/(38E-3*(1+2E-2*(V(60,LOW)-25)+1E-6*PWR((V(60,LOW)-25),2)))}}
G4 13 40 12 0 {-1}
*
.MODEL MstroMOD NMOS (LEVEL=1 LAMBDA=0.001 VTO=2.53 KP=7.5 L=1U W=1U RS=0.8 IS=1E-30
+ TOX=1 RG=10 N=10 T_ABS=25)
.MODEL MmedMOD NMOS (LEVEL=1 LAMBDA=0.001 VTO=0.285 KP=3.0 L=1U W=1U RS=0.18 IS=1E-30
+ TOX=1 RG=1.36 N=10 T_ABS=25)
.MODEL  MweakMOD NMOS (LEVEL=1 LAMBDA=0.001 VTO=-0.275 KP=0.2 L=1U W=1U RS=2.9 IS=1E-30 TOX=1
RG=25 N=10 T_ABS=25)
*
.MODEL D_BODY D (IS=2.4E-11 N=1.04 TT=1E-9 CJO=4.35E-9 M=0.54 XTI=3.9 T_ABS=25)
.MODEL D_PLCAP D (IS=1e-30 N=10 CJO=100E-12 M=0.47)
```

Simulating and making simple low-profile WiFi antennae

by Paolo Antoniazzi, IW2ACD and Marco Arecco, IK2WAQ

Proliferation of portable devices that need high-speed connection have created a demand for wireless connectivity[1&2]. Access points and subscriber units traditionally use omnidirectional antennae and are often inadequate to meet the capacity requirements of the network. Directive high-gain antennas are designed to address this market with low-cost solutions.

We have realised a low-profile helix (length=only 4 cm!) to be used in the 2.4GHz amateur band. Thanks to the very small mechanical dimensions, this antenna is particularly interesting also for Wi-Fi (Wireless Lan) applications in today's widely used modes, 802.11b and 802.11g.

It is quite uncommon to use an helix as a WiFi antenna, but it is very simple to build, is a 'no tuning' design and you can connect easily with systems using both vertical and horizontal polarization antennas.

This paper refers to the simulation and measurements of the very low-profile helix. In the past low-pitch helices has been recognized as ineffective radiating elements for a circularly polarised wave. Numerical results using NEC-Win Pro + NEC-Win Synth simulations[3] and field measurements, however, lead to some low-pitch helices with gain comparable to that of a conventionally long helix.

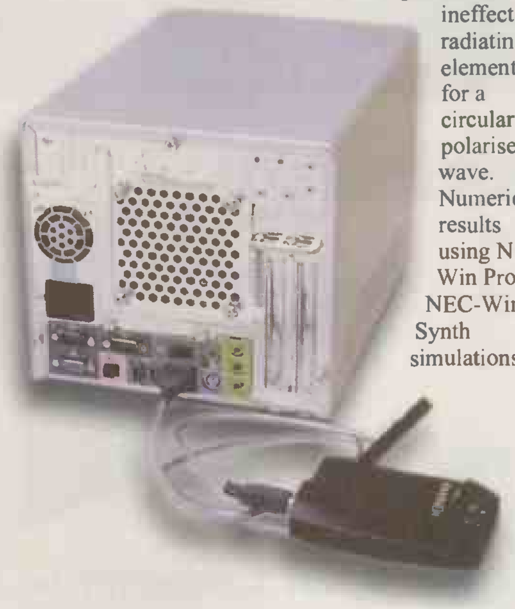


Figure 1: Typical USB access point for WiFi connected to a PC

For circular polarisation applications the axial-mode helix antenna is an interesting candidate, because its good polarization performance is an inherent attribute of the antenna shape without the need for a special feeding arrangement. Polarization properties of the helix have been the subject of several publications[4&5] since the early work of Kraus.

A typical helical antenna operating in the axial mode has a circumference $C = \pi D$ of approximately one wavelength and a pitch spacing S of approximately a quarter wavelength.

Traditionally the pitch angle, an important parameter of the helix, may range from about 12 to 16 degrees; approximately 12 degrees (pitch = 30mm) is typical in most 2.4GHz satellite receiving helices.

The pitch α is the angle that a line tangent to the helix wire makes with the plane perpendicular to the axis of the helix; and it can be found from the relation: $\sin \alpha = S / \pi D$, where S = pitch spacing and D = diameter of helix.

Wave polarisation

The polarisation, orientation and sense of each antenna in a system should be identical in order to optimise signal strength between stations. For example, linearly polarised antennas that are identically oriented (e.g. vertical, horizontal) work best together as do circularly polarised antennas that are using the same sense (RHC, LHC). Even so, circularly polarised antennas are compatible with linearly polarised antennas and vice versa since linearly polarised antennas can 'see' circularly polarised signals in its linear plane.

As you can see in the photo of **Figure 1,** frequently the simple antenna included in common USB Access Points is no vertical-no horizontal, but with casual orientations causing large signal attenuation (see **Table 1**).

When linearly polarised antennas are misaligned by 45 degrees, the signal strength will degrade by 3dB, resulting in up to 50% signal loss. When misaligned by 90 degrees, the signal strength degrades 20dB or more. Likewise, in a circularly polarised system, both antennas must have the same sense or a loss of 20dB or more will be incurred. Combining linearly polarised antennas (TX) with circularly polarised antennas (RX) will incur a loss of 3dB in signal strength between the two formats. A circularly polarised wave radiates energy in both the horizontal and vertical planes as well as every plane in between.

The difference, if any, between the maximum and the minimum peaks as the antenna rotates through all angles, is called the axial ratio or ellipticity and is usually specified in decibels (dB). Normally, if the axial ratio is 0 to 2dB, the antenna is said to be circularly polarised. If the axial ratio is greater than 2dB, the polarization is referred to as elliptical.

Helix antennae

The famous work of J.D. Kraus[6] on helices started in 1946, but only in the 90s was been possible the big simulation study made by D. T. Emerson[7], a very import starting point that cannot be forgotten by every people interested in the simulation and manufacturing of helical axial antennas. Before to start in the simulation phase, using Nec-Win-Pro and NEC-Win Synth, we tried to define the main parameters and the general performances of our antennas (power gain, radiation angle, input SWR and axial ratio).

For the involved people, NEC-Win

Synth is designed to allow users to quickly build complex antenna structures. The structures can be created in multiple ways (47 predefined models are included plus the ability to import NEC, ASCII, and DXF files allows for very creative ways to generate 3D structures). Geometric data is displayed in a spreadsheet with access to 134 predefined functions and constants and 52 user defined variables. Dialog boxes linked to the spreadsheet make it easy to rotate, move or scale individual wires or complete models. As you build and modify your model, the structure is displayed and dynamically updated as edits are made. We used NEC-Win Synth to build the circular reflector for the helix **Figure 2**.

The values used for simulation are:

- Inner radius: 6mm
- Outer radius: 62mm
- Angle 1 (start): 0°
- Angle 2 (end): 360°
- Segment along patch: 36
- Segment across patch: 6

The typical power gain of standard helix antennas can be easy estimated using the graph of **Figure 3** where the performances, at 2.4GHz, of different length antennas are compared: from 2λ to 6.5λ corresponding to a number of turns from 8.3 to 27. The constant parameter of the whole involved helices is the pitch between two contiguous turns that is S=0.24λ (or $\alpha = \sim 12°$, corresponding to 30mm @ 2.4GHz). In this graph the power gain is plotted versus C/λ that means for each antenna length there is an optimum turn diameter that maximise the gain.

The power gain and the directivity are also affected by the dimension and the shape of the ground plane that can be square or circular, but it needs a side or a diameter equal to λ (125mm @ 2.4GHz) to obtain good performance.

With a smaller dimension screen we take the risk of the inversion between the main lobe and the back one of the antenna!

The axial ratio values are included within $1 \div \infty$ and are defined by:

$$AR = |E_\phi| / |E_\theta|$$

Where E_ϕ and E_θ are the electric fields in time-phase quadrature, perpendicular to the axial direction of the helix. The polarisation is as much circular as the AR ratio is near 1 (0dB). The matching of this requirement can be confirmed analysing the radiation patterns generated by the Nec-Win-Pro simulation program.

The following criteria are suggested for the design:

- use of a copper wire, gold or silver plated, having a suitable diameter: 0.024λ (3mm @ 2.4GHz).
- wind the helix in a way to have a cylindrical shape
- divide each turn into 10 segments in order to satisfy the Nec-Win-Pro rule that fixes the minimum ratio between the length of the segment and the wire radius for better simulation accuracy. The use of 20 segments per λ is suggested only for critical regions (complex shapes).
- use of a 6mm stub between the ground plane and the helix, during the simulation phase, to minimise the current induced in the screen by the proximity of the first turn of the helix winding.

Using the above criteria we made a simulation and build two different antennae (see **figure 4**): one with the purpose to receive the AO-40 satellite[8] having 16.7 turns (simulation results: power gain 14.5dB, radiation angle 26°) and another with 5 turns (power gain 12dB, radiation angle 45°) to be used both as a reference antenna and as TX aerial for the directivity measurements described later.

Low-profile helices

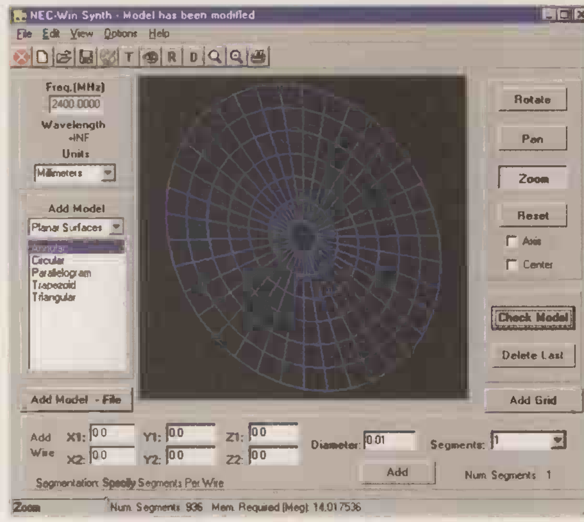
The behaviour of the current versus length of a typical helix shows three different regions:

a) near the feed point where the current decay is exponential
b) near the open end with visible standing wave
c) between the two helix ends where there is a relatively uniform current and small SWR (transmission line)

There are two ways to obtain a good circular polarization helix: 1) tapering the helical turns near the open end, to reduce the reflected current from the arm end, and 2) using only the first helical turns where the decaying current travels from the feed point to the first minimum point.

Starting from these considerations our final low-profile helix uses a pitch S = 0.16λ (20mm @ 2.4GHz) and is both conically wound with a conic 62/41 mm. diameter and very short (only 1.7 turns –as shown in the photo of **Figure 5** during the field tests).

The simulated and measured results are very interesting and the directivity is not so different from that of a conventional multi-turn helix. The equivalent directivity



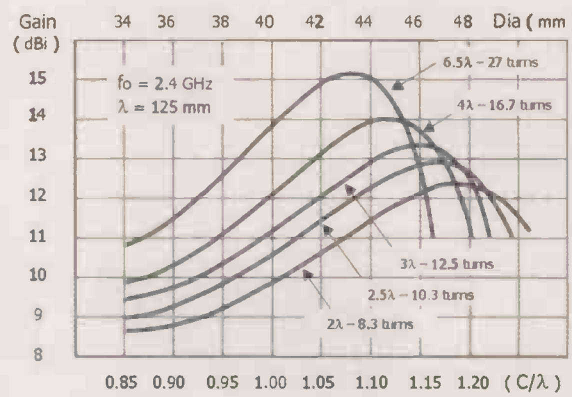


Figure 2: Screen view of the helix reflector design using NEC-Win Synth.

Figure 3: NEC-2 Simulated gain vs helix diameter and C/lambda @ 2.4GHz.

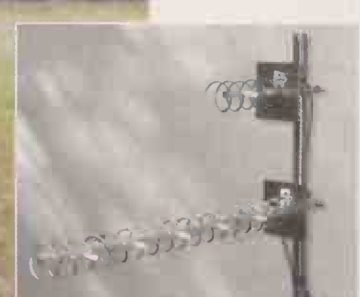


Figure 4: Five and sixteen turn standard helix antennae.

Figure 5: Low-profile helix with 1.7 conic turns during the field tests .

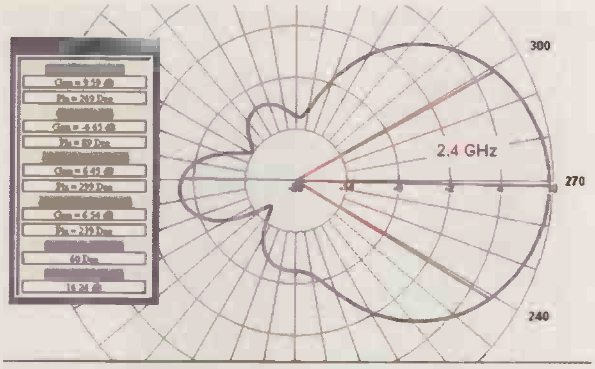


Figure 6: NEC-Win Pro simulated radiation diagram of the 1.7 turn conic 2.4GHz helix(pitch=20mm, conic dia. = 62/41mm)

Figure 7: Input impedance simulation (Smith chart) of the low-profile helix

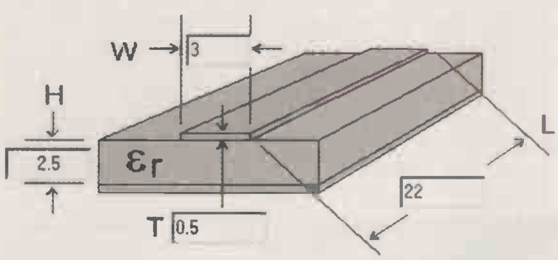


Figure 8: Layout of the λ/4 Teflon transformer calculated using HP-AppCad λ/4 transformation from 130 to 50Ω. The transformer is realised using a Teflon plate (22x30mm) and a copper stripe with W=3mm. Th=0.5mm, at the beginning of the Helix

obtained from radiation angle measurements is about 10dB for the low-profile helix and 11dB for the 150mm long 5 turn helix. The measured radiation angles (-3dB) are respectively 58 and 45 degrees. Thanks to the very small mechanical dimensions, this antenna is particularly interesting for Wi-Fi (Wireless Lan) applications in the 2.4GHz band.

The results obtained with NEC-Win Pro are very stimulating. The radiation diagram and the input impedance (Smith chart) of the low-profile 2.4GHz Helix are shown in **Figures 6 and 7.**

Shown in the **table 2** is a comparison of the simulated and measured values of gain and radiation angle of the 1.7 turn helix of **Figure 9**.

Input impedance matching

A good SWR (Standing Wave Ratio) is guaranteed by the matching between the typical 120-140Ω input impedance of the helix and the 50Ω of the feeding coaxial cable. This is obtained using a λ/4 transformer made by a microstrip realised using industrial Teflon support with h=2.5 mm and line width (W) of 3mm, Z=~81Ω.

The transformer layout is shown in **Figure 8** and was designed using an Agilent tool, the AppCAD (see www.agilent.com). The mechanical realisation is very simple using industrial Teflon strips (see **Figure 9**). As seen in the photo, about 20mm of the beginning part of the helix wire is flattened to about 0.5mm using hammer and vice.

The input connector is an "N" type because the SMA connectors are not mechanically compatible with the helix and matching pad. The final version of the helix for WiFi is shown in **Figure 10** (with a simple plastic radome).

Measurement errors and tricks

It's not very difficult to design and make good helices for the 2.4GHz band. More difficult, are precise measurements. The first critical point is the SWR measurement, because high quality is required of the cables and adapters. The time and money spent on high-quality cables can be wasted if there are large impedance mismatches within the connectors, at the connector-cable interface and with the adapters (only N to SMA, for the 2.4GHz tests). David Slack of Times Microwave Systems[9] writes: "a microwave cable assembly is not 'just a wire'. It is a passive, TEM mode, microwave component and an integral part of a system".

Assuming an high-quality cable is used, the predominant contributor to the SWR of a cable assembly (on a 10-50cm short assembly) is the connector. Improperly compensated geometry changes in low-cost connector interfaces will exhibit very poor SWR characteristics.

Our SWR tests of the helix have been made using an old slotted line . For almost all the tests we used a 2.2 to 2.7GHz generator made with a Minicircuits VCO mod. JTOS-3000 followed by a low-cost wide band amplifier 0535AN2 made by Keps Communication (www.keps.it). Similar amplifiers are also available from Minicircuits (www.minicircuits.com).

The output level from the oscillator is very high (+10dBm), but some attenuation must be included for stability (the wide-band amplifiers oscillate very easy with loads not exactly 50 ohms). To measure the relative gain of the low profile helix, the RF power to the input of a standard 5 turn transmitting helix is also about 10mW followed by a 6dB N-attenuator.

To reduce the measurement errors, the distance between transmitting and receiving antennas has to be considered. To determine this distance, you need to be able to measure the signal level easily with a filtered RF voltmeter having a 20-30dB dynamic range. Also, the wave reaching the receiving antenna should be as planar as possible. The first condition can be easy established starting with the received power and calculating the attenuation experienced by the wave in the open space:

$$A = 32.4 + 20\log(f) + 20\log(d) - G_t - G_r$$

Here, A is the attenuation in decibels, f is the frequency in megahertz, d is the distance in km, Gt is the gain of transmitting antenna indBi and Gr is the gain of receiving antenna, also indBi is obtained by simulation. There is also a simple, easy to remember method[10] of calculating the free space attenuation by considering the distance between the two antennas in terms of wavelengths. When d = λ , A is always 22dB between isotropic antennas.

This equates to 12.5cm at 2400MHz. The attenuation increases by 6dB for each doubling of the path distance. This means that the free space attenuation is 22dB at 0.125 m, 28dB at 0.25m, 34dB at 0.5 m, etc.

Left – Figure 9: A Zoom on the λ/4 impedance matching using a Teflon microstrip transformer

Right – Figure 10: Simple low-loss radome case for the WiFi antenna

Table 1: Signal attenuation versus antenna's polarisation.

Antennas orientation angle (degree)	Linear vs linear polarisation mismatch (dB)	Circular vs linear polarisation mismatch (dB)	Circular vs circular polarisation mismatch (dB)
0	0	3.01	0
15	0.3	3.01	0
30	1.25	3.01	0
45	3.01	3.01	0
60	6.02	3.01	0
75	11.74	3.01	0
90	∞	3.01	0

Table 2: 2.4GHz measurements and simulation of helice's radiation characteristics.

Type	Measured radiation angle °	Equivalent directivity (dB)	Simulated radiation angle °	Simulated directivity (dB)	Notes
16.7 Turns	28	13.5	26	14.5	A040 type
5.0 turns	45	11.0	42	12.0	Reference
1.7 Turns	58	10.0	60	9.6	Low profile

To make the wave reaching the receiving antenna as planar as possible, the capture area in square metres of the receiving antenna is:

$$Ac = Gr\,\lambda^2 / 4\pi$$

This expression is valid for an antenna with no thermal losses and was certainly useful for our experiments. With a circular capture area the minimum distance in meters between the antennas will be:

$$d > nGr\,\lambda / \pi^2$$

A maximum acceptable phase error will also be considered.

For a phase error of 22.5°, which is usually enough, n = 2. If a phase error of only 5° is required, n = 9. In the case where one dimension prevails over the others, the maximum length instead of the capture diameter is used. In this case, the minimum distance in metres becomes[11&12]:

$$d > n\,L / \pi^2$$

where L is the maximum length in metres (50cm for the 16.7 turns helix). The site (10 x 20m) used for our field tests is particularly useful for all helix measurements (typical d = 4 m = 32λ at 2.4GHz).

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Simulating ideal transformers using OTAs

Starting from a resistive ladder network, whose terminal equations are equivalent to those of an ideal transformer, the ideal transformer is simulated using only OTAs and its operation is successfully tested by SPICE software. The turns-ratio of the proposed ideal transformer can be digitally controlled by an external voltage or current source. Yavuz Sari and Abdullah Ferikoglu explain

Inductors and transformers, which are heavy and bulky, are not used in integrated circuits. However, it is possible to realise these component types by using active networks[1]. Operational transconductance amplifiers (OTA)-based networks have been finding a wide range of applications in recent years due the fact that their transconductance can be used as a design parameter. In this study, an electronically adjustable turns-ratio transformer is proposed.

It can be shown that, the resistive ladder network given in **Figure1** is equivalent to an ideal transformer. Simple analysis yields:

$$V_1 = \frac{R_1}{R_2} V_2 \qquad (1)$$

$$I_1 = -\frac{R_2}{R_1} I_2 \qquad (2)$$

Both positive and negative resistors in the network of figure 1 can be implemented using only OTAs[2]. For obtaining an ideal adjustable turns-ratio transformer, transconductance values of OTAs should be chosen as follows,

$$g_{m1} = g_{m2} = g_{m3} = \frac{1}{R_1} \qquad (3)$$

$$g_{m4} = g_{m5} = \frac{1}{R_2 - R_1} \qquad (4)$$

$$I_1 = -\frac{R_2}{R_1} I_2 \qquad (5)$$

where R_2-R_1 must be positive for the condition that the sum of resistors in the middle loop of the network of figure1 must be zero. The resultant adjustable turns-ratio transformer network is shown in **Figure2**.

Results

The obtained adjustable turns-ratio transformer network is tested on a computer by a SPICE simulation software program. The input and output voltage and current waveforms for a load resistor of 1kΩ are used in the SPICE simulation and tranconductances are chosen as, $g_{m1}=g_{m2}=g_{m3}=g_{m4}=g_{m5}=0.2$(mA/V), $g_{m6}=g_{m7}=g_{m8}=0.1$(mA/V).

In the above, we have proved that an original active equivalent of the ideal transformer can be made. It is digital system compatible and can be implemented by integrated circuit technology. Its superiority is that the turns-raito of the transformer can be electronically adjusted by the control voltage or current of the OTAs in the network. It is clear that its operation range is limited by linearity conditions of the OTAs used[3].

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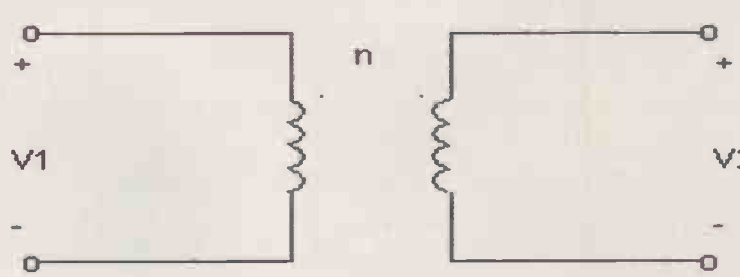


Figure 1:
The circuit symbol (a), and resistive equivalent of the ideal transformer with turns-ratio of R1/R2 (b)

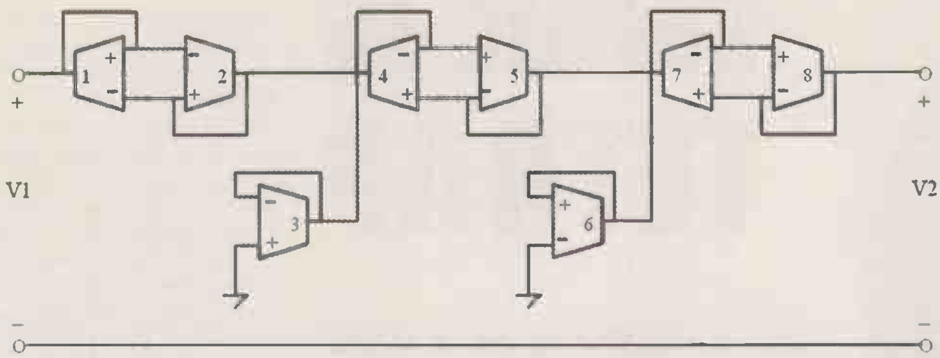
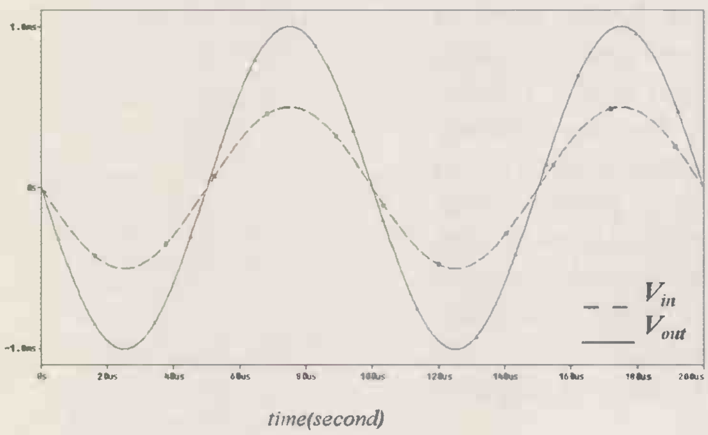
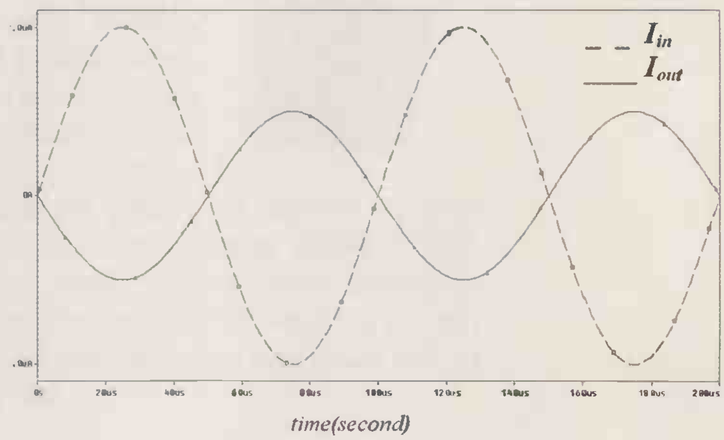


Figure 2: An OTA realisation of the ideal transformer in figure 1

Below left – Figure 3: The input and output voltage waveforms of the realised ideal transformer with n=1/2

Below right – Figure 4: The input and output current waveforms of the realised ideal transformer with n=1/2





What is TETRA?

Nokia THR850 WAP enabled TETRA phone
(Image courtesy Nokia)

Communication is an essential feature of today's world. Cellular phones have seen a remarkable level growth since their introduction in the 1980s, demonstrating the importance of mobile communications in today's world. For many business or professional applications other forms of communication are available that may be more suitable than cellular phones. These PMR services have been available to the business user for many years.

Ian Poole investigates

The letters PMR stand for Professional (sometimes Private) Mobile Radio. Originally systems consisted of simple radio transceivers mounted in vehicles that could communicate with a fixed base station. These basic systems were first introduced many years ago well before the days when mobile phones were available and they were ideal for use by many mobile services including taxis, utility services and the like as well as the emergency services where frequent communication was required with a base station. Spectrum efficiency was very poor because only intermittent use was made of the allocated frequencies, even if several users shared the same frequency in one location.

To enhance the services, improve spectrum efficiency, and provide a far more effective service new systems were developed. Trunked services where mobile users could be linked through to their office via a series of linked base stations enabled mobile stations to use the service over a much wider area. Also by enabling a far greater number of users access to the system on a shared basis, the utilisation of each channel was increased, thereby considerably improving the spectrum efficiency. A number of systems were developed, but the one that has by far the greatest acceptance around the world is specified under MPT1327.

MPT1327 provided many advances, but it was essentially an analogue system. With developments in digital technology and the widespread use of mobile phones which also used digital technology it was clear that further improvements could be made, and as a result TETRA was born.

Evolution

The name TETRA stands for TErrestrial Trunked RAdio. Aimed at a variety of users including the police, ambulance and fire services, it is equally applicable for utilities, public access, fleet management, transport services, and many other users. It offers the advantages of digital radio whilst still maintaining the advantages of a PMR system.

The TETRA system was developed with the support of the European Commission and ETSI (European Telecommunications Standards Institute) members. Work started in 1990 with the first standards being published in 1995. Representatives from all interested parties including manufacturers, users, operators and other experts were included in developing these standards. Although it is aimed at being used by a wide variety of professional users, there has been an emphasis on ensuring the special needs of the emergency services were met. To achieve this many of the lessons learned in the development and deployment of the digital cellular telecommunications networks were taken into account. This ensured a successful development cycle.

Features

TETRA offers many new and valuable features. These include a fast call set-up time, which is a particularly important requirement for the emergency services. It also has excellent group communication support, direct mode operation between individual radios, packet data and circuit data transfer services, better economy of frequency spectrum use than the previous systems and advanced security features. The system also supports a number of other features including call hold, call barring, call diversion, and ambience listening.

The system uses Time Division Multiple Access (TDMA) technology with four user channels on one radio carrier and 25kHz spacing between carriers. This makes it inherently efficient in the way that it uses the frequency spectrum. Data can be transmitted at 7.2kbits per second for a single channel. This can be increased four fold to 28.8kbits per second when multislot operation is employed.

For emergency systems in Europe the frequency bands 380-383MHz and 390-393MHz have been allocated. Additionally, whole or appropriate parts of the bands 383-395MHz and 393-395MHz can be utilised should the bandwidth be required. For civil systems in Europe the frequency bands 410-430MHz, 870-876MHz / 915-921MHz, 450-470MHz, 385-390MHz / 395-399.9MHz, have been allocated.

The TETRA trunking facility provides a pooling of all radio channels that are then allocated on demand to individual users, in both voice and data modes. By the provision of national and multi-national networks, national and international roaming can be supported, the user being in constant communication. TETRA supports point-to-point, and point-to-multipoint communications both by the use of the TETRA infrastructure and by the use of Direct Mode without infrastructure.

Operation

There are three different modes in which TETRA can be run. They are voice plus data (V+D), Direct Mode Operation (DMO), and Packet Data Optimised (PDO).

The most commonly used mode is V+D. This mode allows switching between speech and data transmissions, and can even carry

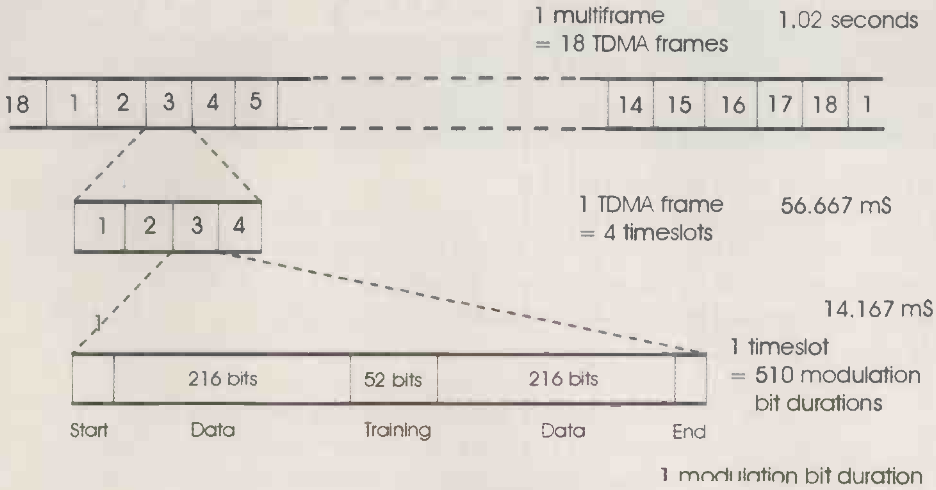


Figure 2: TETRA data structure.

both by using different slots in the same channel. Full duplex is supported with base station and mobile radio unit's frequencies normally being offset by about 10MHz to enable interference levels between the transmitter and receiver in the station to be reduced to an acceptable level.

DMO is used for direct communication between two mobile units and supports both voice and data, however full duplex is not supported in this mode. Only simplex is used. This is particularly useful as it allows the mobile stations to communicate with each other even when they are outside the range of the base station.

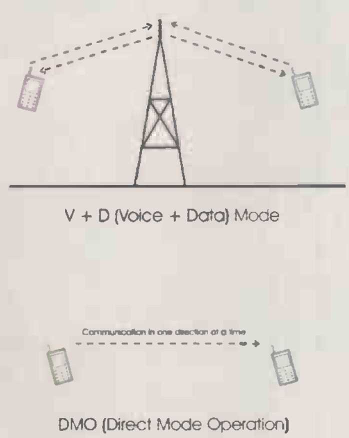


Figure 1: TETRA modes of operation

Finally the PDO mode is optimised for data only transmissions. It has been devised with the idea that much higher volumes of data will be needed in the future and it is anticipated that further developments will be built upon this standard.

Data structures

TETRA uses TDMA enabling efficient use of the spectrum by allowing several users to share a single frequency. As the speech is digitised, both voice and data are transmitted digitally and multiplexed into the four slots on each channel. Digitisation of the speech is accomplished using a system that enables the data to be transmitted at a rate of only 4.567 kbits/second. This low data rate can be achieved because the process that is used takes into account the fact that the waveform is human speech rather than any varying waveform. The digitisation process also has the advantage that it renders the transmission secure from casual listeners. For greater levels of security that might be required by the police or other similar organisations it is possible to encrypt the data. This would be achieved by using an additional security or encryption module.

The data transmitted by the base station has to allow room for the control data. This is achieved by splitting what is termed a multiframe lasting 1.02 seconds into 18 frames and allowing the control data to be transmitted every 18th frame. Each frame is then split into four time slots. A frame lasts 56.667ms. Each time slot then takes up 14.167ms. Of the 14.167ms only 14 milliseconds is used. The remaining time is required for the transmitter to ramp up and down. The data structure has a length of 255 symbols or 510 modulation bits. It consists of a start sequence that is followed by 216 bits of scrambled data, a sequence of 52 bits of what is termed a training sequence. A further 216 bits of scrambled data follows and then the stream is completed by a stop sequence. The training sequence in the middle of the data is required to allow the receiver to adjust its equaliser for optimum reception of the whole message.

The data is modulated onto the carrier using differential quaternary phase shift keying. This modulation method shifts the phase of the RF carrier in steps of $\pm\pi/4$ or $\pm3\pi/4$ depending upon the data to be transmitted. Once generated the RF signal is filtered to remove any sidebands that extend out beyond the allotted bandwidth. These are generated by the sharp transitions in the digital data. A form of filter with a root raised cosine response and a roll off factor of 0.35 is used. Similarly the incoming signal is filtered in the same way to aid recovery of the data.

Beyond this, TETRA uses error tolerant modulation and encoding formats. The data is prepared with redundant information that can be used to provide error detection and correction. The transmitter of each mobile station is only active during the time slot that the system assigns it to use. As a result the data is transmitted in bursts. The fact that the transmitter is only active for part of the time has the advantage that the drain on the battery of the mobile station is not as great as if the transmitter was radiating a signal continuously. The base station however normally radiates continuously as it has many mobile stations to service.

One important feature of TETRA is that the call set up time is short. It occurs in less than 300ms and can be as little as 150ms when operating in DMO. This is much shorter than the time it takes for a standard cellular telecommunications system to connect. This is very important for the emergency services where time delays can be very critical.

Summary

Although the take-up has not been as swift as that for cellular phones, and in the UK there have been some business setbacks, TETRA is now widely established with more than 325 contracts in 55 countries. Certificates of equipment interoperability have been issued to ten manufacturers for equipment with more under test.

For the future further work is under way to enhance the performance of TETRA. Higher data rates are being planned to keep the system in line with modern requirements. Standardisation of speech codecs is also being investigated to ensure complete compatibility whilst also providing the optimum coding, and further improvements are being included to enhance the features available as well as optimising the spectrum efficiency, network capacity and quality of service.

'Mixed Spices' Part II

Alistair Macfarlane of Electric Fields continues his investigation of Spice simulation programs, and shows how they compared with real life.

First up this month is **Orcad/Pspice** which is a Spice simulation product from Cadence; seemingly a fusion of Orcad's PCB design and Cadence's Pspice. I had thought of Pspice as the grand-daddy of Spice programs, but quickly ran into problems. My first attempt was to try to use the Pspice 'Student' version, which I downloaded. But when I tried to create my circuit and hit the usual trouble, I contacted the agents, Parallel Systems Ltd., who suggested that I use their 'Orcadlite' demo program instead. Still hitting problems with missing models for my circuit, I was offered a time-limited full version of the program. This arrived with 5 CDs (for the full Orcad suite) and a complicated licensing procedure which took the best part of a day to get to work. (I currently use Windows 98SE due to warnings about stability on the newer platforms, and the latest version of Pspice does not support 98SE, so I had to use the older version of the program.)

Rather like Icaps4, this is all rather ponderous with apparently great power but not intuitively easy to use. Due to the installation delays I had to rush to try to get my circuit to work, but initially failed miserably. Part of the problem was the ease of picking up parts which seemed right but were from the capture library only and didn't have simulation models. And finding a humble 1N4002 diode proved impossible until I discovered that their search engine only found diodes if the number was prefixed with the letter 'D'! Library size is a claimed 14,000+ models.

After sending the non-working circuit file off to the agents and receiving it back, I still couldn't get it to converge, so sent it off again. This time it came back and (without any reported changes) the transient analysis actually ran. However it took an whopping 2 minutes plus for a graph to be produced. On all the other programs I used a maximum time step of 10nS, both to get good resolution and to give a measurable simulation time. The maximum timestep in Pspice was set to 100nS as it failed to converge at 10nS (or 1μS).

Unfortunately something was wrong and the results were not at all what I expected. The frequency of the digital clock oscillator was around 45MHz (!), against a calculated value of some 455kHz (f=1/2.2RC), and the output of the D-type was a soft sawtooth which soft switched the transistor. This high frequency would explain the slow run time. There was a message about the D/A interface failing to converge and a large number of 'serious' warnings popped up, so perhaps the reason was buried in these error messages.

I again asked the agents about the problem, but failed to receive a reply before my timed license ran out. The agents then told me that the fault was due to the need to individually set the inverter Initial Conditions, and sent me a plot showing that there was a ringing on the edges of the inverter waveforms (see screenshot below).

I am aware that this can sometimes happen with 74AC buffered gates, but interestingly (and worryingly) it was the only other Spice program apart from MC7 to pick this up (and MC7 only showed it with a 74AC part). Perhaps that's why it is the most expensive product tested. From the plot they sent me showing the edge ring, the frequency was still

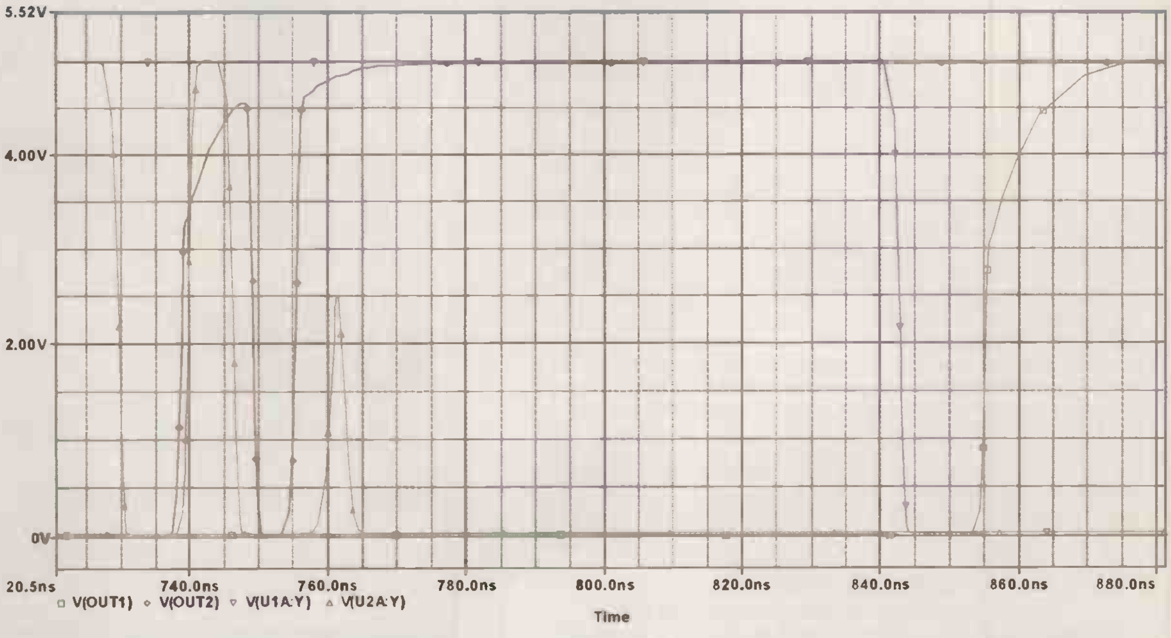


Figure 1

very high at an estimated 4.2MHz.

As shown in the screen grab in **Figure 1**, the simulation does have a form of marching graphs, but these appear to download the data in lumps every few seconds rather than printing each point individually as the others seem to do. Monte Carlo/sensitivity analysis is offered with an optimiser as you would expect in an expensive program like this. A neat idea is their 'Smoke' alarm, which warns about the possibility of a stressed component by comparing simulation results with the component's ratings. It even claims to check for transistor secondary breakdown.

Pspice A-D costs a hefty £3995, which includes the Orcad capture/PCB suite (see **Figures 2 & 3**). To buy Pspice on its own would cost £6500 plus another £900 for a capture front end. For more info go to: www.parallel-systems.co.uk). There is also a user group through www.orcadpcb.com. Again, with Spice programs you often need to know what to expect before believing what you get!

Spiceage/Spicycle is a UK programme from Those Engineers. It really fell at the first fence, as I could not install it using the 'PC Install' installation program provided, which is apparently no longer supported. But I attempted to persevere and contacted the company, and was then sent a version without the installer. I copied over the relevant directories but although the capture interface was quite intuitive, I quickly found it could not run my circuit sensibly. Again I made contact with Charles Clarke of Those Engineers, who modified the digital models so that the simulation ran for him. (He was also concerned that I might be using Windows XP). But he had added and changed quite a few parts so when the results were still not accurate I began to think that this program was perhaps past its sell-by date. Part of it is reported to still be 16-bit legacy code and whatever that implies the screen shot (**Figure 4**) shows up its failings all too clearly. According to Mr. Clarke, the simulation took 90 seconds to run, but examining the traces show multiple switching on the D-type edges and erratic transistor operation. I went back to try some of the demos provided, but probably due to the faulty install procedure I could not get any of them to run. I decided not to pursue this testing further. If you can get it all to work, there are quite a few interesting circuits to run, however, such as a Phase-locked loop. Spiceage and Spice cycle are the least expensive costing only £45 points per module; they can be found at www.spiceage.com.

Superspice is another UK offering, this time from Anasoft, run by Kevin Aylward, an analogue design engineer. I had high hopes for this program, since from the capture screen with what seemed like dozens of icons (see **Figure 5**), it looked powerful. However I immediately ran into trouble with frequent computer hangs and general buggy behaviour. I contacted Kevin, who voiced the opinion that the problems were due to my use of 98SE, and that I should be using XP! However the program is advertised as being suitable for everything from Win95 up. In fact other Spice program makers had specifically named XP as being a source of problems, so I'm not sure about this assertion.

I endeavoured to create the circuit in capture, but ran into problem after problem. Even the available demos were buggy, and after running a simulation, graph windows would come and go, appear as thumbnails, and usually end up hanging up, so in the end I didn't get a screenshot of a simulation. Which is a pity as there were some interesting circuits included such as switched capacitor filters, CCFL drivers (with a negative resistance characteristic) and so on. I think this could potentially be a good tool, however again I feel that getting the 'hangs' out of a program should be a first priority before it is unleashed on a paying customer and there is still work to be done. Superspice comes in a student version at £40 up to the Professional Gold version at £220; free to approved educators. (www.anasoft.co.uk).

Simetrix/Simplis is yet another UK designed Spice program, from Catena Software Ltd. It comes in various versions, including a Linux based one (It has a Unix origin). It is described as 'affordable' simulation software, and at around £1590 for the basic Simetrix AD Plus 4.5 is in the middle of the range. But adding the Simplis extension takes the price to £2690.

The initial impression of the program was of a reasonably intuitive capture program with a component toolbar for easy selection, (see **Figure 6**) but no context based 'Help' on any but the intro screen. I couldn't find a slow recovery diode in the limited model lists, so had to use a fast one. Trying to move component text was a pain; in fact the first minor bug showed up when I tried to move the text for C3 and found it moved at 180 degrees to

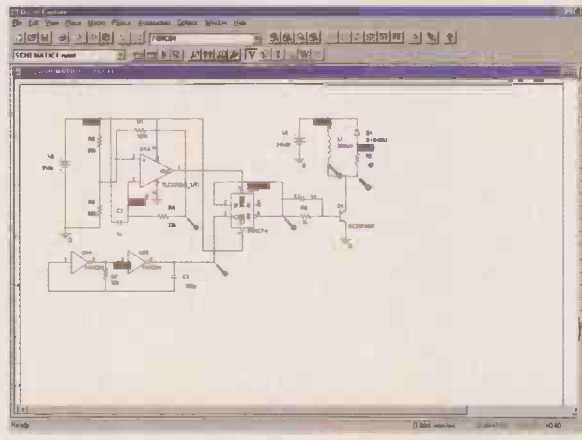


Figure 2

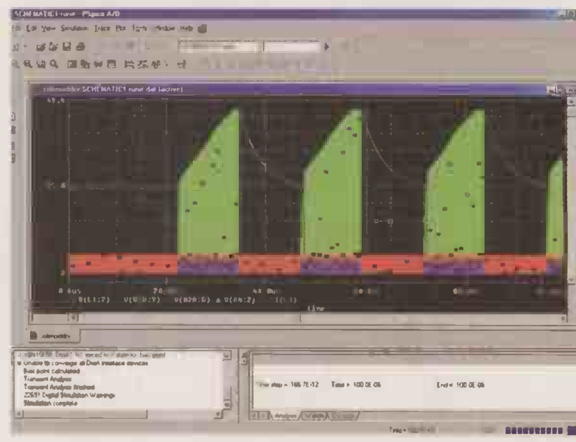


Figure 3

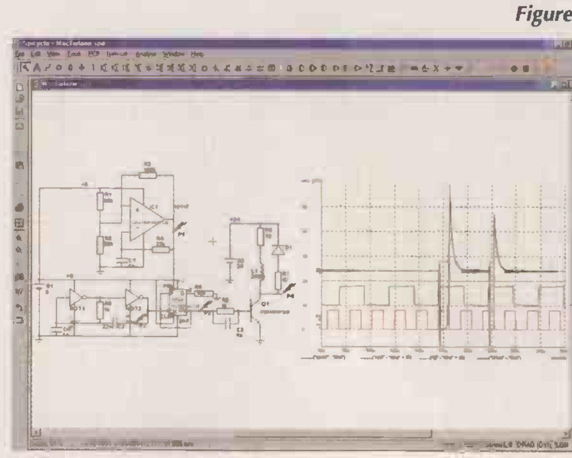


Figure 4

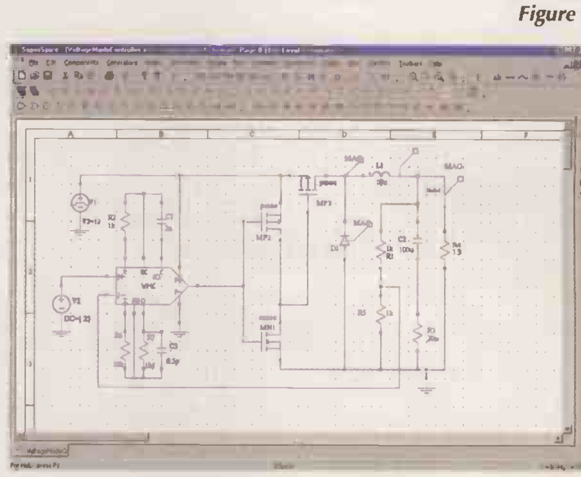


Figure 5

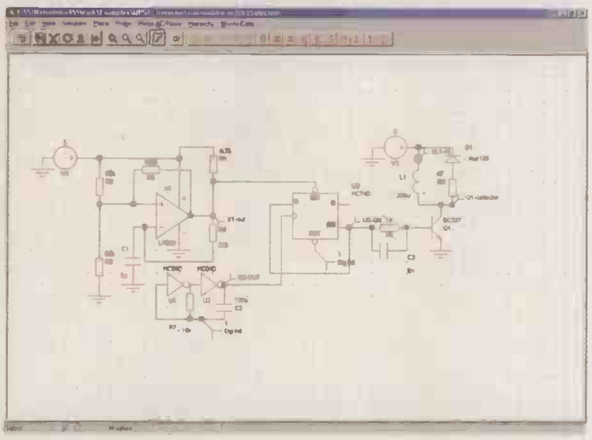


Figure 6

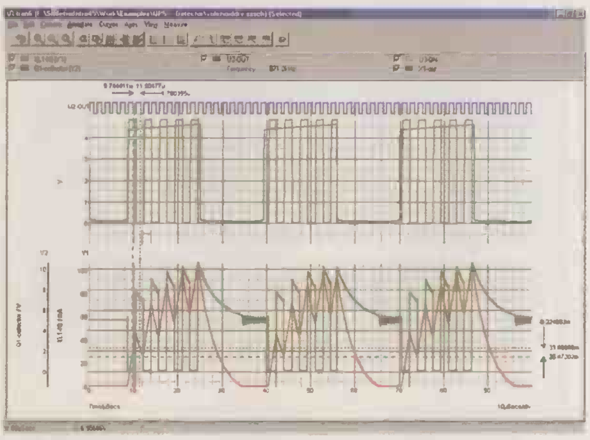


Figure 7

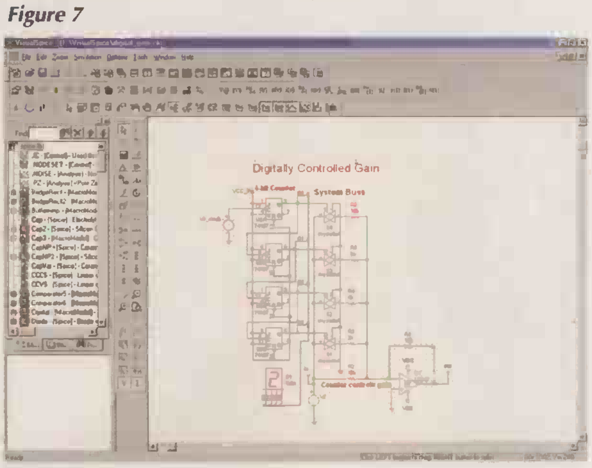


Figure 8

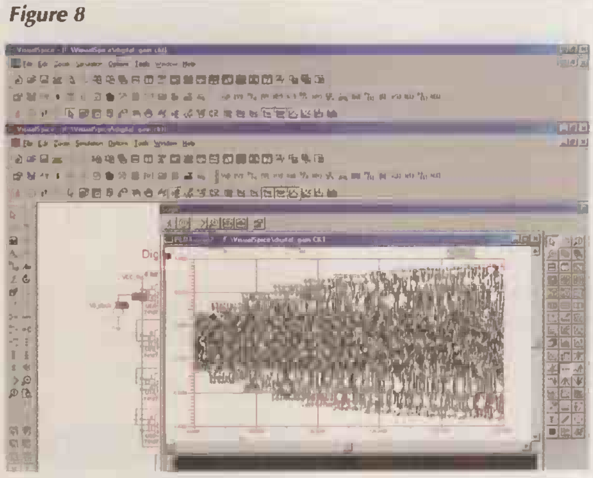


Figure 9

where I set it. To move a component meant first detaching it, and the wiring pen had to be toggled on and off (what's wrong with the Esc key?) I contacted their support as the simulation (of course) wouldn't run first time, and got a helpful pointer to setting the initial conditions to get the two oscillators to run. The simulation then ran flawlessly, taking 7 seconds. The graphs were clear and there were two cursors, which allow a delta to be measured as shown in **Figure 7**. And an FFT could be run on any of the chosen waveforms. The frequency from the digital oscillator was high at 571kHz compared with the calculated value of 455kHz; this may be due to some model differences. Monte Carlo analysis is also provided.

Simplis is an extension to the basic Spice program to run switching circuits a claimed 10 to 50 times faster, using a piece-wise-linear method to cut down the amount of data needed in a Spice simulation. Trying one of the demo SMPS programs showed that this seems indeed to be the case. (I couldn't figure out if it was possible to use my 'reference' circuit with Simplis.) It ran 20ms of a 50kHz boost converter in around three seconds, rather impressive. (But perhaps some switching data may be lost in the process in comparison to Spice?) It also allowed a Bode frequency and phase plot to be made of the switching converter loop in an equally fast time, a very useful feature for designing control loop optimisation. Unfortunately after a few runs the program decided I'd exceeded some limit and refused to run beyond an 8ms time; kicking in an annoying DOS window which had to be removed at each attempt. But all in all, the best of the British.

Visual Spice is a US designed program supplied by Quasar Electronics in the UK.

From the capture screen, **Figure 8**, it looked powerful, having plenty of icons, but trying to use it in anger was to prove impossible. The demo will not allow simulation or saving of a circuit other than those provided, and even then will not run even an unintentionally modified schematic, or changed simulation set-up. In fact it wouldn't run much at all with any sincerity; whilst a few of the animations ran, graph traces were unintelligible (see **Figure 9**) and eventually the whole program got weirder and weirder until I decided enough was enough and reinstalled the whole thing. But although I did manage to capture a small part of my circuit with relative ease, deleting a component involved clicking on an icon which changed the cursor to a lightning bolt, which then had to be clicked on the item to be removed. (What's wrong with the 'delete' key?) The help file was all I had to go on and that was pretty hard to follow. I contacted Quasar, but assistance was slow in coming and there was no offer of a more advanced version or better help. At that point I capitulated. Visual Spice is however inexpensive; there are four versions costing from £74.95 for the simplest, to £179.95 for the Advanced. More info at: (www.quasarelectronics.com/visual_spice).

Microcap 7 is designed in the USA by Spectrum Software, started by Andy Thomson around 1980 and now sold by Rainbow Software in the UK., I have used each new version since the mid-eighties and have (and regularly use) the full V7, therefore this might be considered to have something of an unfair advantage. But without doubt I have kept the best until last. I was able to build and run the standard circuit in less than five minutes; it ran first time and gave close to the correct results in that the digital clock frequency was 496kHz, 9% higher than calculated. It did not show any edge rings on the 74HC04, but did pick up occasional nanosecond ringing when I used a 74AC04 (as Pspice showed with the 74HC part) and then only when using trapezoidal integration and not in gear. See **Figure 10**. Correctly the D-type ignored them as too fast. Perhaps there are differences in the models for these parts. Microcap offers two different models for the 1N4002 diode. One has an almost zero recovery time (unrealistic) and the 1N4002GP has a normal slow recovery. The latter produced the same ringing artefacts that B2Spice had demonstrated, unlike all the others. It would seem they must use the 'unrealistic' diode model. MC7's simulation was also the fastest of all, doing much more in just under two seconds. It correctly showed the reverse diode recovery and failure to completely saturate the transistor. Transistor power dissipation showed as 0.15W

Everything about MC7 seems to have been thoughtfully designed for easy, quick and accurate usage, from the schematic capture to analysing the (marching) graphs. For the beginner, there are a number of animations found under the help menu, which cover most of the basic and some more advanced features. Capture is a breeze, once you realise that rotating a component is just a matter of holding the left mouse button down and clicking the right

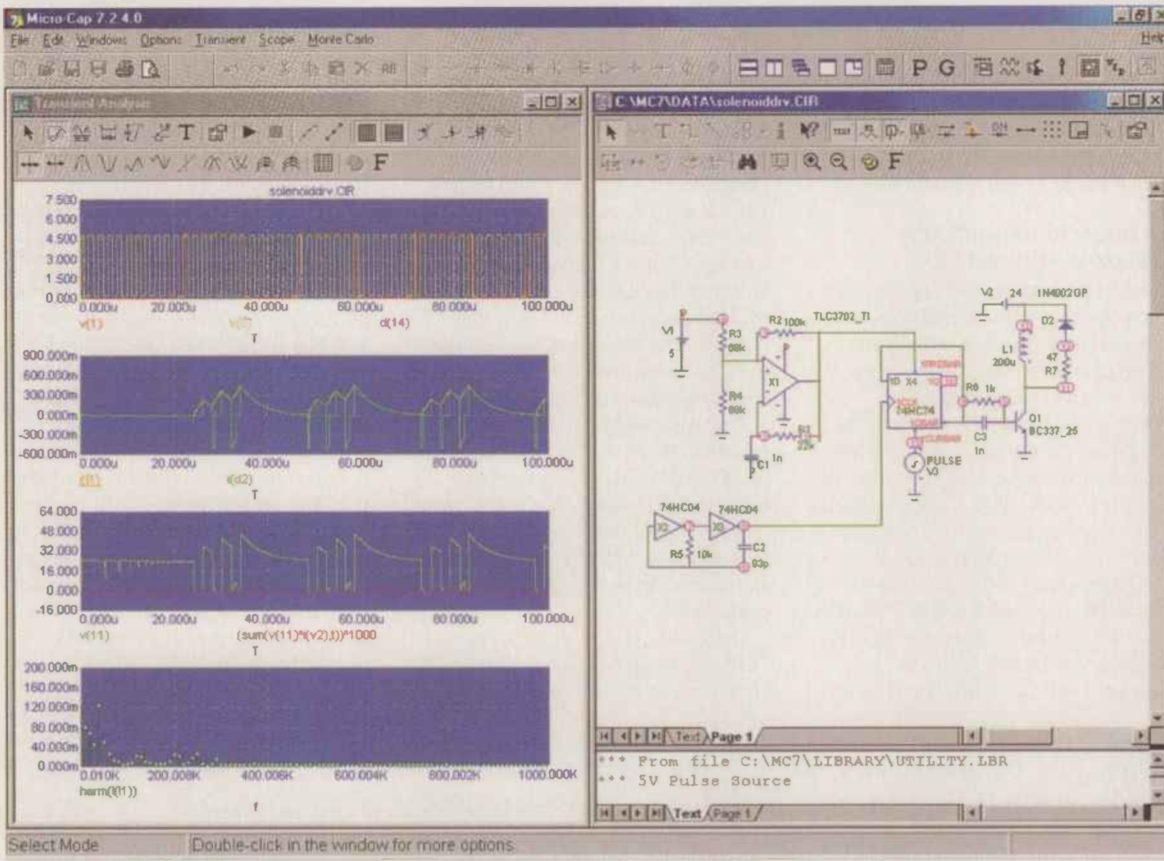
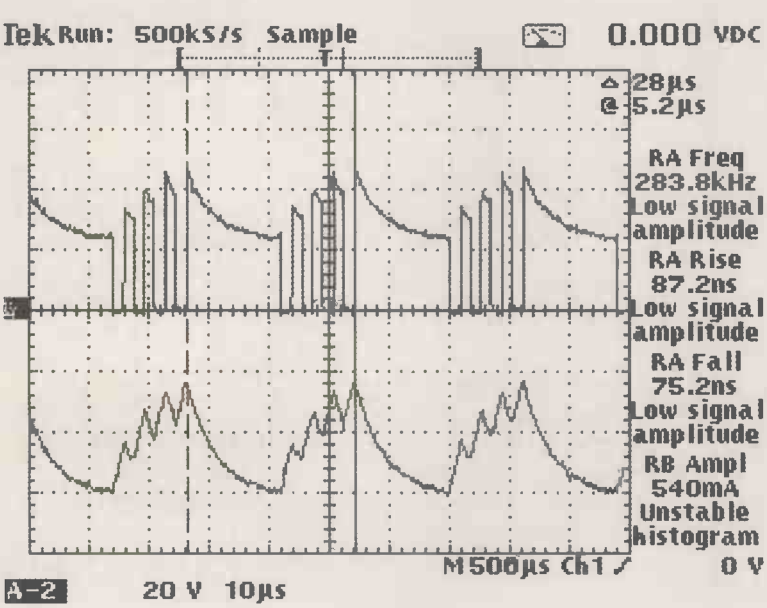


Figure 10

one to go through each of the eight degrees of freedom. Intuitive selection of basic components is from a small toolbar and more advanced parts such as ICs, sub-circuits, sources etc. from a manageably sized drop down series of alphanumeric menus. A lesson to others!

Running the simulation first throws up a window which allows selection of what to plot. You can either define this by node numbers or text name, the latter having the advantage that if you change the circuit, the text names stay put. Or you can define a current or voltage across a device (e.g. I(R1)), digital node or include an equation such as device power, RMS/AVG values or many other more exotic math functions. You can find lists of these by right clicking on the trace boxes. The traces can be in the same axis, or be tiled in as many windows as will fit the screen if you wish. And the X-axis can be time or frequency for FFT, harmonics etc. You can set fixed or auto-scaling, temperature, operating point, zero node voltages or leave them to run the simulation on from the previous finish point, all in that one window; very convenient.

The marching plots, selected by an almost unlimited number of colours and backgrounds, are updated as the points are calculated and the simulation can be stopped by Esc key or pause button. The plot window defaults to zoom mode, so that it's instantly possible to zoom into a section of any waveform (and go back as there is unlimited 'undo'). One more click selects the measurement mode, with two cursors, which can lock on to high/low points, or waveform inflections by another button click, reading the values (and both deltas and slopes) below the trace in the same colour, making it fast and easy to analyse the waveforms. The performance function allows an instant measurement of waveform properties such as frequency, rise time and so on. And Microcap have an ingenious 3-D plot which can be used for temperature or component sweeping. Monte-Carlo, sensitivity analysis and optimisation are all provided and there is an active and passive filter design feature. Averaging models of SMPS parts are available and Microcap recently re-wrote the Jiles-Atherton models for magnetic cores, which enables much faster simulation of devices using



Left: Figure 11

ferrite cores etc. They also have a user group www.micro-cap-subscribe@yahoogroups.com and publish quarterly newsletters, freely downloadable. MC7 costs £2450.86 from www.micro-cap.co.uk . (MC8 has just been announced.)

So how did they actually compare with real life?

I constructed the circuit using the same components as in the simulations, and powered it up. The digital oscillator showed no edge ringing whatsoever, even with a 74AC04, and ran at 560kHz. See **Figure 11**. However the 100pF cap measured 7% low so a more realistic frequency of 520kHz would be expected with a 'perfect' component. The supply voltage was 3% high and gate threshold tolerances could account for the frequency difference between measured and simulated. (MC7 was the closest at less than 5% difference). The analogue comparator oscillator measured 36kHz but here the 1nF cap was 15% high so this equates to a design frequency of 41kHz, exactly what most predicted. However I was somewhat bemused when the output transistor fried itself to a short unexpectedly. Investigations showed that the speed up capacitor across the base drive resistor was too small to switch off the transistor fast enough, and needed to be increased to at least 2.2nF. (This would seem to indicate a Spice model shortcoming in Miller capacitance and/or base stored charge). Also the current rose to higher than the design level, which is explained in part by the lower frequency due to the 1.15nF cap, and by a progressive saturation of the small ferrite inductor I had available. Transistor dissipation calculated from the 30°C rise I measured would indicate around 0.17W, versus around 0.15W from MC7's simulation. Again the higher current explains this.

There was no sign of the transistor failing to saturate, but here the advantage of having simulated the circuit would have allowed a careful designer to allow for more base drive, a higher gain transistor and most importantly a fast recovery diode!

So there you have it, Spice is an excellent way of checking out the general operation of a circuit and tolerancing it, but you always need to take care with what you read into it. Component models in Spice are usually based on an attempt to be the perfect mathematical equivalent and allowances must be made for real-life parasitics and approximations. It can be so very powerful, when you can just try something out in minutes which might take days of calculations to solve, or be very frustrating when it refuses to converge.

Selection of the best program for your application has to be a personal choice based on cost, user-friendliness and/or sheer mathematical 'grunt'. I hope this article has helped show where some of the pitfalls may lie.

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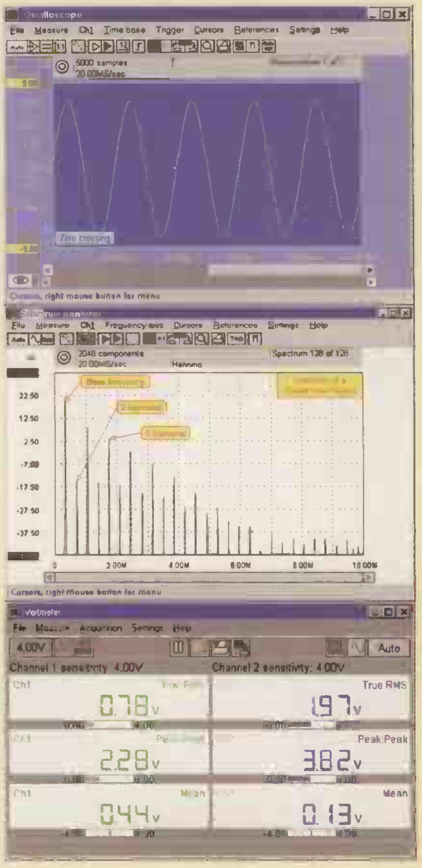

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Omni directional ferrite rod receiver

With a radio receiver using a ferrite rod aerial, the accepted wisdom is that there will always be a pair of strong nulls in the polar response, meaning that the receiver must be positioned in an often inconvenient position, and that it will not be suitable for use in a moving vehicle. The usual solution of using a telescopic rod aerial is a poor one, because an aerial of this type is at its least satisfactory at low frequencies, particularly on LW.

Attempts to get around this by, for instance, putting half the winding on one rod and winding the other half on a rod at right angles results in a combination which still works well, but with a pair of nulls which are just as strong.

If, however the RF phase of the signal from one rod can be delayed by 90 degrees so that the signal is in quadrature with the signal from the other rod before it is combined and amplified, then there is no condition where the signal from one direction can cancel the signal from the other, and the result is a ferrite rod radio with no nulls. Adjusted properly, the signal will vary by at most 2 to 1, easily coped with by even a simple AGC circuit, let alone the amplified one shown here. The circuit shown is for a single station TRF LW receiver intended for Radio 4 UK on 198kHz.

Each ferrite rod (in this case about 5 inches, but not critical) has its own tuned winding easily set up by sliding one rod in and out of the winding with the other disconnected and then fixing. Using such a large winding with a small tuning capacitor results in both a larger signal and a broader bandwidth than is normally obtained on a long wave receiver. Those used to the 'afterthought' LW facility on most modern radios will be quite surprised by the bright relatively noise-free result possible.

The necessary phase shifts are obtained by series resistors around gate 2 which allow the internal capacitance of the gate to cause extra phase delay, and a pair of simple cascaded T filters to give a little phase advance after gate 1 from the other aerial. Gates 1 & 2 give sufficient isolation between the two rods to allow the signals here to be simply connected together and fed to the input of gate 3 without interaction.

One advantage of using the unbuffered 4069 instead of more accepted RF devices is that the gain characteristic of each stage allows a very powerful AGC circuit to be easily set up. Pulling a stage away from the centre of its characteristic where the gain is at maximum (the quiescent state give by the 1M resistors) results in a rapid reduction in gain, provided here when the BC548 is biased 'on' by a signal of more than about 0.6V from the detector diodes.

This circuit, with modification to aerial windings and coupling capacitors, (chosen here to keep AF gain of the RF stages to a minimum) can be easily be adapted for omni-directional reception of 60kHz atomic clock signals.

Andrew Ziemacki
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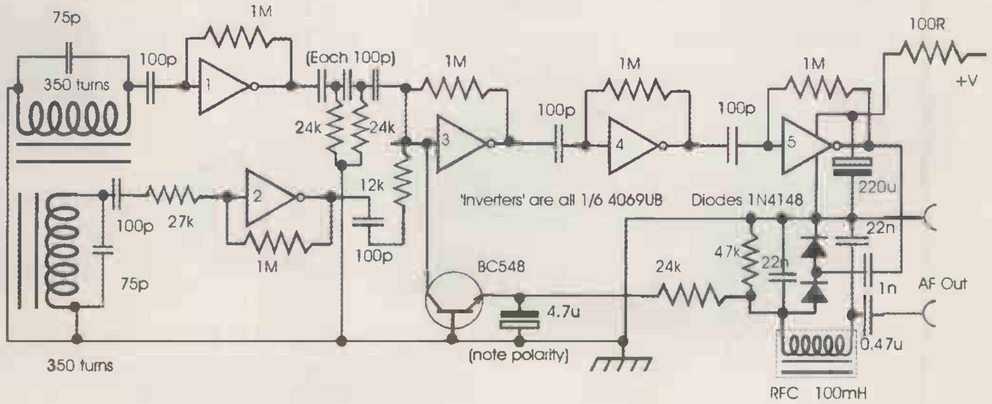

Multi-station irrigation control using mains timer

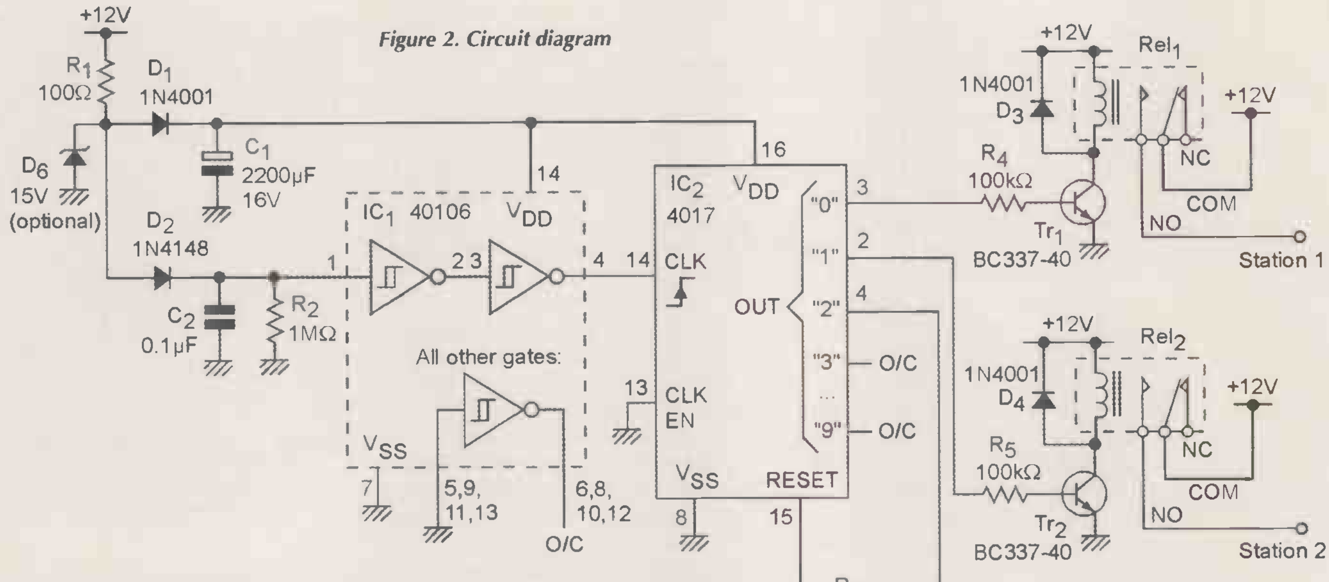

Figure 2. Circuit diagram

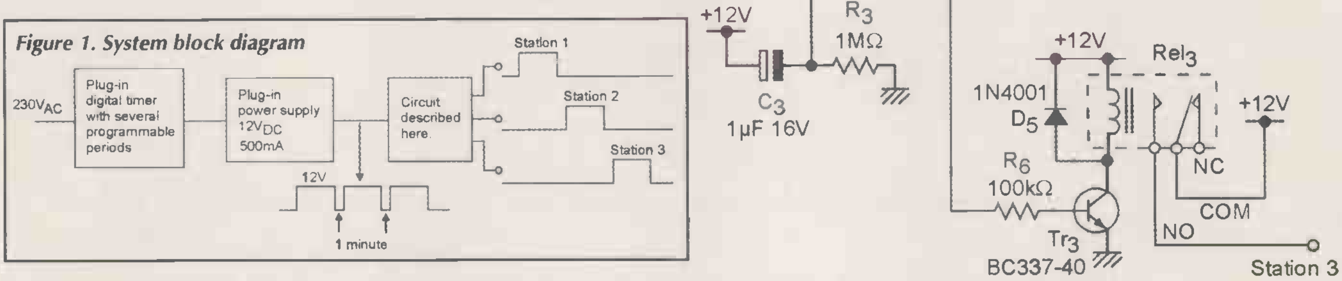

Figure 1. System block diagram

This simple circuit turns an ordinary mains timer into a multi-station controller for a garden sprinkler system. I had a spare timer and thought there must be a way to use it for automatic sprinkler control of more than one zone. This circuit might in general be a lower cost alternative to a commercial irrigation controller, since the latter costs a lot more than the common household timer.

The principle of operation is explained in **Figure 1**. The timer is programmed to turn on for a series of periods, and off for 1 minute in between. Each time the timer turns on, the selector circuit described here switches the power to a different sprinkler zone. The duration of the on periods determine for how long the sprinklers will water the garden.

A 12V DC power supply is plugged into the 230V AC socket of the timer, to drive the circuit described here, as well as the solenoid valves of the sprinkler system. These solenoids should be rated 12V DC, although the common 24V AC solenoids appear to work well enough on 12V DC. Alternatively, a separate supply can be used to drive the solenoid valves, in place of the 12V on the relay contacts.

A 24VAC solenoid will operate on a lower DC voltage. The DC current is larger, because it is not affected by coil inductance. The advantage of AC drive is that the coil inductance increases when the solenoid closes, thereby drawing less current in hold mode. The inrush current for 24V 50Hz AC is typically 0.4A and the hold current 0.2A.

When controlled by DC, the voltage should be lower than 24V to avoid potential overheating. One way to limit the DC hold current is to use a transformer with a high-voltage low-current secondary and a large capacitor (e.g. 2200μF) after the diode rectifier bridge. The capacitor will provide a large initial voltage to close the solenoid, and then the transformer drops to lower voltage under load.

The circuit diagram is shown in **Figure 2**. The core of the circuit is the 4017 decade counter/divider. It switches a '1' from output 0 to output 9 each time the input is clocked. The status of IC_2 must be maintained for the minute the timer turns off and the supply is interrupted. This is the function of C_1. It is charged up via R_1 and D_1. R_1 limits the charge current. Zener diode D_6 can be added to protect the CMOS ICs from voltage spikes.

Most 4017 counters need a short clock rise and fall time. The supply doesn't go up fast enough for this and would cause false triggering. Schmitt trigger gates IC_1 are added to decrease the clock transition times. An exception would be ST's HCF4017, which has unlimited clock rise and fall times. If you use this device, you could leave out IC_1, although the Schmitt trigger input does provide protection against spurious triggering.

D_2, C_2 and R_2 are added to prevent false triggering, particularly by the drop in the supply voltage when a solenoid valve is turned on and a large current drawn. R_3 and C_3 prevent clocking of IC_2 the first time the power is turned on, so that output '0' is high for the first zone.

During the 1 minute intervals that the power is turned off, C_1 is partially discharged by base resistors R_4 and R_5. When the timer switches of permanently at the end of the zone 3 period, C_1 is discharged fully by R_6. This takes about 10 minutes. Actually, there is plenty of time – C_1 only needs to discharge in time for the next garden watering session, which is typically the next day.

The transistors must be the high gain version of the particular transistor type, e.g. BC337-40 or BC109C, with a current transfer ratio h_{FE} of at least 250. The relays must be a low current type, with a coil resistance of at least 400Ω.

If you have a mechanical timer which can only be set in increments of 15 minutes, the circuit needs to be modified. Since the off intervals will be 15 minutes long, C_1 must keep its charge for longer. Either C_1 must be increased to at least 22000µF, or the transistors replaced with Darlingtons and the base resistors increased to at least 1MΩ.

The number of zones could of course be increased by adding more relay circuits to the output of IC_2. The number of periods programmed into the timer must not be more than the number of relay circuits implemented. If the high output of IC_2 goes beyond the relay circuits, e.g. output '3' in this case, there is no resistor to discharge C_1.

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via email
South Africa

Lightbulb protector

The circuit, **Figure 1**, is basically a standard phase controlled triac switching circuit, which is what a lamp dimmer consists of, except we have replaced the control knob with a light dependant resistor, LDR_1.

LDR_1, together with R_1, is used to control the conduction angle of SCR_1 (TRIAC). The value of R_1 is chosen so that when LDR_1 is shielded from ambient light with a piece of black tubing that is clamped at both ends, the lamp just begins to glow. Under normal operating conditions LDR_1 is shielded with a piece of black tubing that is open at one end. This end faces the lamp so that LDR_1 can sense the brightness of the lamp **Figure 2**.

When power is initially applied to the circuit, LDR_1 has a high resistance. R_1 causes SCR_1 to conduct sufficiently to light up the lamp dimly. The (low) light from the lamp causes the value of LDR_1 to drop, which causes SCR_1 to conduct longer. This in turn causes the lamp to glow brighter which leads to the value of LDR_1 dropping even further. A complete chain reaction is started which leads to LDR_1 reaching its lowest value and the lamp reaching maximum brightness in less than a quarter of a second. R_4 and C_3 have been included in the circuit to suppress any noise generated by SCR_1.

The circuit is fairly simple and straight forward. It would find many practical uses where one would not relish the thought of having to replace bulbs very often. It works exceptionally well when used to supply a set of lamps such as a large chandelier. LDR_1 is simply focussed on any lamp in the chain. Remember if the load is going to be excessive, a suitable heatsink should be attached to SCR_1.

Nick Moodley
KWA-Zulu Natal
South Africa

Parts List:

R_1	180kΩ
R_2	3,3kΩ
R_3	18kΩ
R_4	10Ω 1W
C_1	100nF 250VAC
C_2	47nF 100VAC
C_3	100nF 250VAC
DIAC	DB3
SCR	TIC 206 D
LDR_1	4mm light dependant resistor

Figure 1

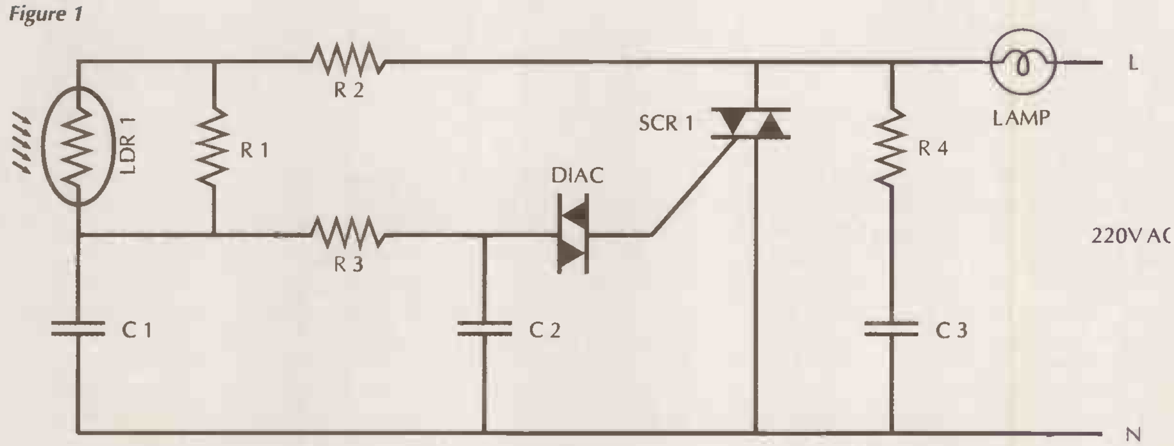
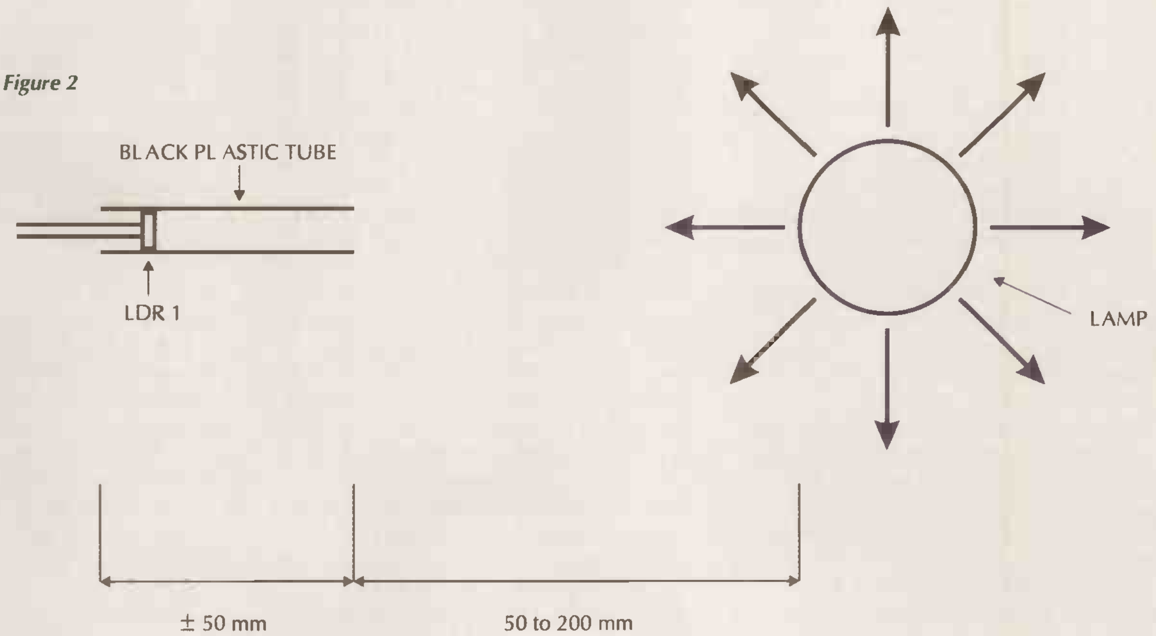


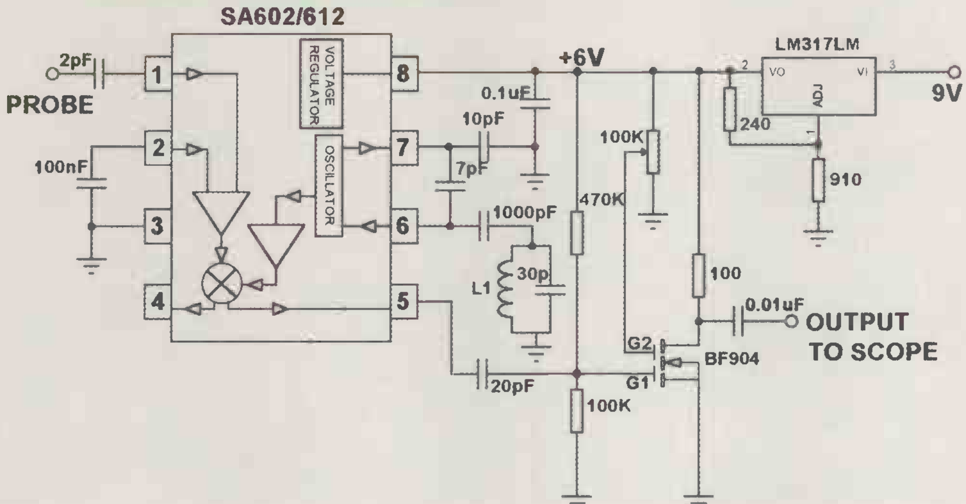
Figure 2



125RF Probe for 20MHz oscilloscopes

This uses the double balanced mixer/oscillator in the Philips SA612. The tuned components (L1 wound on a ¼ inch former; 6 turns) are set up for the 100MHz area, but can hopefully be scaled up to 200MHz. The 30pF tank capacitor can probably be replaced with a small trimmer if required, to extend the measurement range. No anti-image filter is used on the input, as it is assumed that the signal frequency components will be roughly known, and you just want to know if a signal is present or not.

The dual gate mosfet amplifies the now IF signal. It will also amplify the sum signal, which the low frequency scope will not see and is usually filtered out. Gate 2 of the mosfet can be used for calibration with a known level signal. Sometimes this is the agc control input, but here, it sets the stage gain at a fixed level. If calibration can't be done because you don't even have temporary access to a 100MHz scope, then the probe can still be used to measure relative levels of signals. I now know which of my FM transmitters puts out the largest signal without having to walk half way to the shops.



B Teleki
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Pump monitor

In times of drought the level of water in boreholes may become critical and limit the amount of water that can be withdrawn before the level drops below the pump. As the pump relies on the water for heat removal, power applied with no water to pump can cause damage.

The arrangement shown employs a moving iron meter to monitor pump current remotely which for the case in question was 4A reducing to 3A for no water flow. The pump is intermittent so that an acoustic sounder is incorporated to draw attention to the pump operation. The voltage drop across the moving iron (about 0.4V) is insufficient to activate a sounder so a low voltage transformer is connected in reverse to increase the AC volts to 15V whilst current is flowing.

A standard red LED is connected in series with the acoustic sounder to pass only positive going half cycles and simultaneously function as an indicator. Note that the same arrangement would serve to alert to excessive supply current for many motor applications.

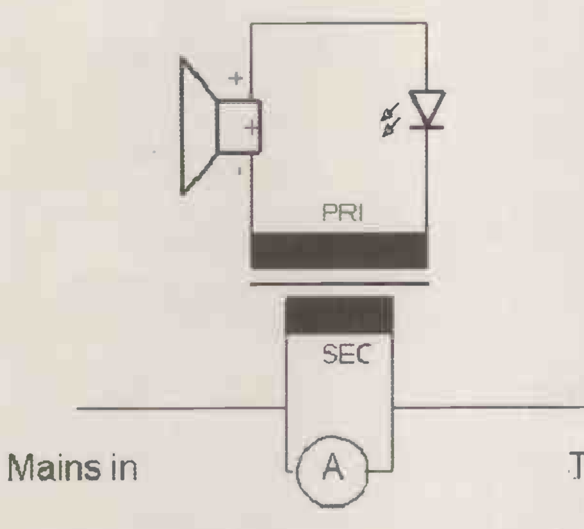
The circuit employed:

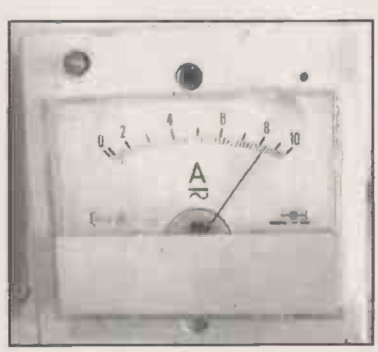
5A moving iron meter: Farnell part number 143-513

Sounder: Farnell part number 413-6380

Red LED: Farnell 322-544 (the piv could be exceeded so that a further series diode may be prudent)

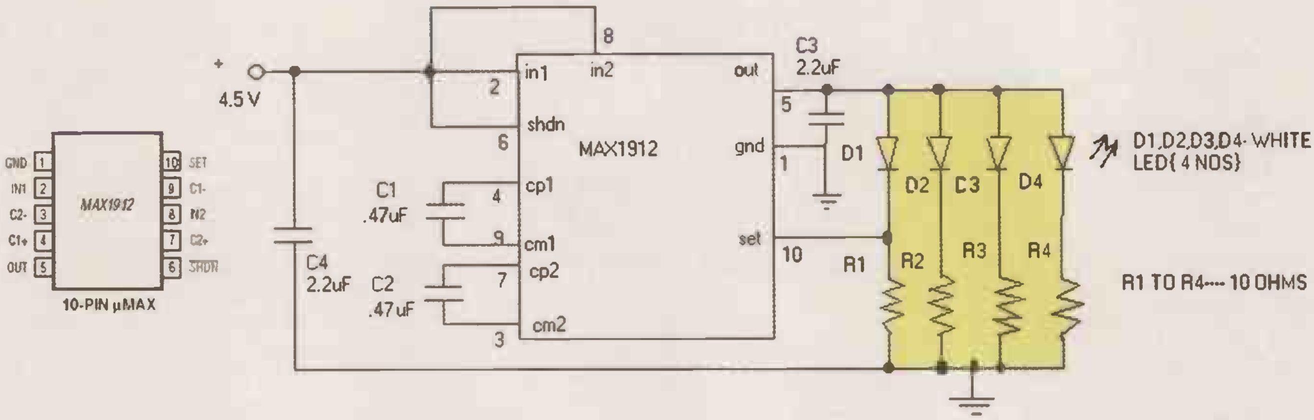
Transformer was removed from a 9V power plug.





Douglas Dwyer
Okehampton
Devon
UK

High efficiency white LED charge pump



In the past whenever something had to be illuminated, we had to use yellow LEDs as a substitute for white light. But as white LEDs are now available anywhere they will replace incandescent bulbs in traffic lights and other applications, drawing less power and lasting over ten times longer. White LEDs are twice as bright as incandescent bulbs, and will help preserve natural resources due to the use of non-toxic materials and efficiency.

A typical 60-watt light bulb puts out a lot of electromagnetic energy in the infrared part of the spectrum which can't be seen but is felt as heat. Replacing incandescent light bulbs with white LEDs would reduce the energy needed to power them.

The recent wireless communication revolution has brought colour LCD displays to cellular phones and PDAs. White LEDs provide the perfect backlight solution for this application. However, a single-cell Li+ battery delivers 3.6V nominal and 4.2V maximum, not enough to drive white LEDs which have a forward voltage of 3.5V typical and 4.0V maximum at I = 20mA to 25mA. Cellular phone and PDA manufacturers are seeking an economic and efficient boost solution for white LED backlighting.

The given circuit design is a simple one, which drives white LEDs with a regulated output voltage or current (up to 60mA) from an unregulated input supply (2.7V to 5.3V). It is a DC-DC converter requiring only four small ceramic capacitors and no inductors. We know the charge-pump solution is the most economic because it does not require an inductor. Input ripple is minimised by a unique regulation scheme that maintains a fixed 750kHz switching frequency over a wide load range. Also included are logic-level shutdown and soft-start to reduce input current surges at start-up. It employs a 750kHz fixed-frequency 50% duty-cycle clock.

The IC_1 (MAX1912) includes soft-start circuitry to limit inrush current at turn-on. When starting up with the output voltage at zero, the output capacitor is charged through a ramped current source, directly from the input with no charge-pump action until the output voltage is near the input voltage. If the output is shorted to ground, the part remains in this mode without damage until the short is removed.

Once the output capacitor is charged to the input voltage, the charge-pumping action begins. Start-up surge current is minimised by ramping up charge on the transfer capacitors. As soon as regulation is reached, soft-start ends and the circuit operates normally. If the SET voltage reaches regulation within 2048 clock cycles (typically 2.7ms), the circuit begins to run in normal mode. If the SET voltage is not reached by 2048 cycles, the soft-start sequence is repeated. The devices will continue to repeat the soft-start sequence until the SET voltage reaches the regulation point. The IC1 shuts down when the die temperature reaches +160°C. Normal operation continues after the die cools by 15°C. This prevents damage if an excessive load is applied or the output is shorted to ground.

Due to the high switching frequency and large transient currents produced by IC1, careful board layout necessary. A true ground plane is a must. To minimise high frequency input noise ripple, it is especially important that the filter capacitor be placed with the shortest distance to IC_1 (⅛ inch or less).

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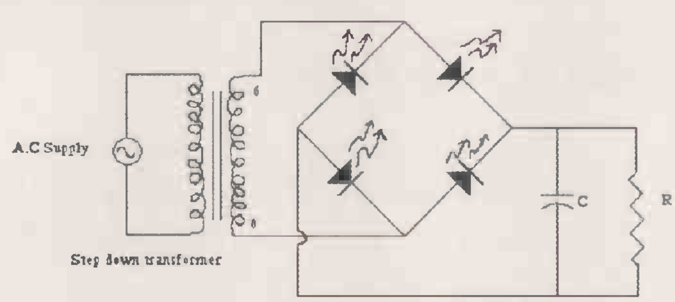
Full wave bridge rectifier using LEDs

Could LEDs be used to assemble a bridge rectifier? Of course it is possible, but what are the technical and practical difficulties. LEDs are made of Gallium Arsenide,GaAsP or GaP. With reference to GaP LEDs they have a reverse voltage of 5V. They can support a continuous forward current of 50mA and a peak forward current of 1A.With forward drop across each LED to be 2V, the maximum drop can only be 3V DC. To get higher voltages more than one diode has to be connected in series and to obtain higher current output more than one LED has to be connected in parallel.



In the circuit a bridge circuit made of LEDs is connected to the output of a stepdown transformer and a load resistor is connected across the outputs. Rectification occurs as in an ordinary bridge but the ripple factor is found to be greater. In a bridge rectifier using ordinary silicon diodes the ripple factor

$$\text{Ripple factor} = \frac{\text{RMS value of a.c component}}{\text{Effective value of D.C}}$$

is found to be 0.485 but in case of LEDs the ripple factor is higher of 0.787, i.e. LEDs are used in place of silicon diodes. If a capacitor filter is used, the ripple factor can be further decreased. By using a resistor R=270 and a capacitor C = 22μF, the ripple factor was found to be 0.287. So if silicon diodes are unavailable GaAs LEDs can be used for small voltages and small currents. However the cut in voltage is greater. If higher voltage is required to be rectified, LEDs should be connected in series such that the reverse breakdown voltage across each diode should not be greater than 5V.

Shery Joseph Gregory
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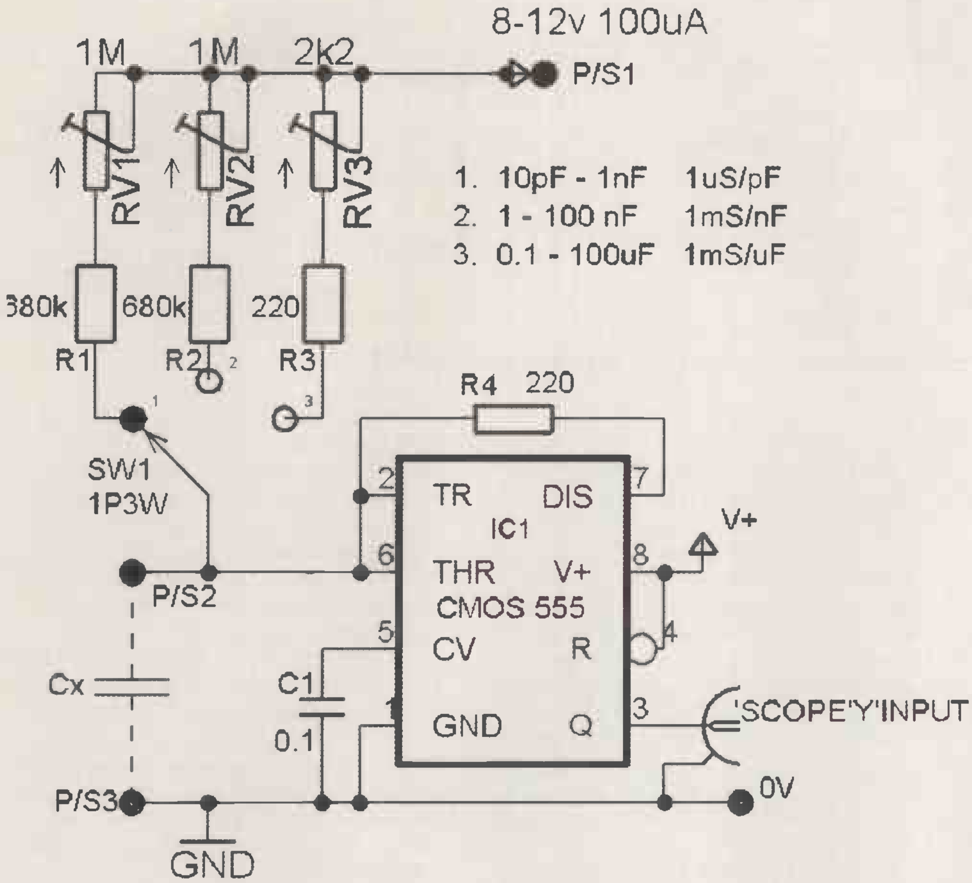
Simple capacitor checker



When developing a prototype, I have always subscribed to the principle that time spent checking the value of a component before soldering is more productive that time spent bug fixing when the circuit does not perform as predicted. A good quality LCR bridge is indispensable for properly testing components but a simpler method has its attractions on the workbench.

Resistor values can be 'checked' on a multimeter (incorrectly coded resistors are seldom encountered, but do exist, however the colours are not always easy to discriminate), capacitor values on the other hand are not readily checked.

This circuit of a capacitor checker is small, inexpensive, simple, and can be connected to the 'Y' input of any oscilloscope.IC_1, CMOS 555 timer, functions in the normal astable mode with a large mark-space ratio, the short duration -ve pulse being determined by C_x and R_4 whilst the +ve pulse is determined by C_x and the selected resistors RV_1,RV_2 and RV_3 with their associated fixed resistors.

Initial calibration is straightforward. On each range, using the closest tolerance capacitors available, preset $RV_{1..3}$ is adjusted indicate the correct value on the appropriate time base of the oscilloscope.

Here is an example – calibrating range 2. Assuming a 10x10 graticule and a 1nF reference capacitor, the oscilloscope should be set to 100μs/division and RV_2 adjusted for a +ve pulse of exactly 10 divisions – hence 1nF/division on the 1ms/division range).

Whilst this device is not a substitute for the facilities of an LCR bridge, it does provide a quick value check in the range 10pF – 100μF (1pF – 1000μF can be achieved). When checking small values of capacitor, there is some inaccuracy due to self capacitance of the input circuit, this is accommodated by range 3 which can be calibrated at 100pF.

The checker can be battery powered and connected by the usual oscilloscope lead, however I have assembled it on a 1" x 1.5" p.c.b. mounted in a small box which carries the switch, access to the presets, the connectors for Cx and a chassis mounted BNC for direct connection to the Y input. A flying lead is connected to an (extra) front panel socket which derives the 100μA from the oscilloscope power unit

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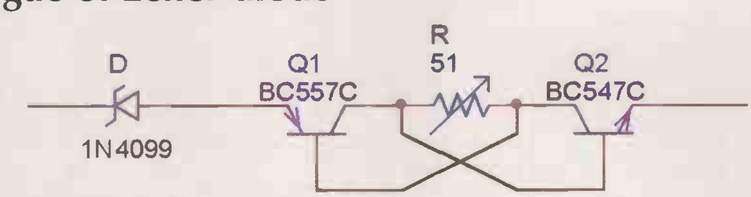
A zero-resistance analogue of zener diode



The dynamic resistance is an important parameter of zener diodes because zener diodes better stabilise the voltage when they have small dynamic resistance. The dynamic resistance of the proposed analogue of zener diode is equal to zero.

The circuit consists of a usual zener diode D and a negative resistor that consists of the resistor R and transistors Q1 and Q2. The total dynamic resistance of the circuit is

$$R_\Sigma = R_z + 2\frac{k(273+T°C)}{eI} - R$$

where R_z is the dynamic resistance of the zener diode, Boltzmann's constant K=1.38·10^{-23} J/K, electron charge, e=1.6·10^{-19} *coulomb*, I is the current of the circuit (i.e. the current of the zener diode), $T°C$ is the temperature in Celsius. The second addend of the expression is the dynamic resistance of the transistors Q_1 and Q_2 when R=0.

Let us assume that $I = 0.005A$, $T = 25°C$ and $R = 3\Omega$. In that case $R_\Sigma = 16.7 - R$. Therefore the dynamic resistance of the circuit is equal to zero when the resistance of the trimming resistor is equal to 16.7Ω.

The stabilisation voltage of the circuit $V_\Sigma \approx V_Z + 1.2V$, where V_Z is the own voltage of the zener diode D. Any stabilisation voltage can be obtained by means of the appropriate choice of the zener diode D.

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Moldova

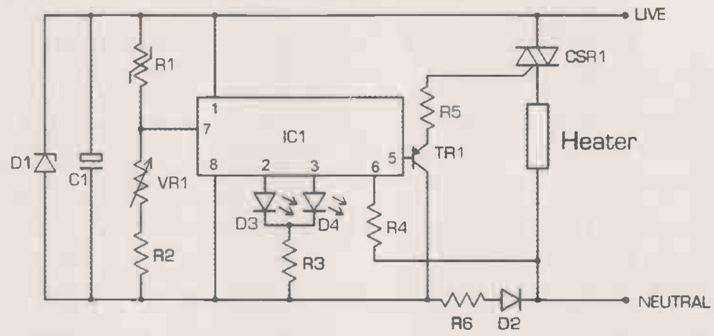
Dual rate thermostat

This circuit controls a domestic hot water immersion heater, for installations where electricity is available at a lower price overnight (for example UK's Economy 7). If the heater is connected only to the Economy 7 supply, then use of too much hot water during the day will result in there being no hot water until midnight. Alternatively, if the heater is connected to the normal supply, then no advantage is taken of the cheaper overnight electricity.

This circuit sets the thermostat to two different temperatures, approximately 20°C higher when cheaper electricity is available, so that the heater is only switched on during daytime rates if the water temperature drops by more than 20°C below the overnight temperature setting.

The program uses a pseudorandom number generator to implement proportional control, and counts mains cycles to determine when to change thermostat temperatures. The software reads the voltage at the a/d pin, and uses this value to look up in a table the percentage power required. Two different tables are used, one for each temperature setting. The power required is compared to a pseudorandom number. If it is higher than the pseudorandom number then the triac is turned on for a full cycle, by producing two pulses - one on the positive zero crossing and one on the negative zero crossing. This ensures that there is no net DC supplied to the load which would contravene EN61000-3-2 (Harmonic currents). The a/d is read each mains cycle, and a new pseudorandom number is generated.

Second-rate Thermostat



D_2, R_6, D_1 and C_1 produce a low current 5V supply referenced to mains live. The thermistor probe is connected to the PIC's a/d converter. R_4 is connected to neutral by a high value resistor to produce a 50Hz clock. TR_1 amplifies the output of IC_1 to produce 50mA gate pulses for the triac. As the load is entirely resistive, a 100μs pulse is used. This ensures that the load current through the triac has exceeded the holding current value, but keeps the average supply current to 500μA. A tri-colour LED is used to indicate whether the circuit is in night or day mode, and flashes yellow in the case of an open- or short-circuited probe.

It should be noted that the probe is directly connected to mains live, and so should have insulation to withstand 2.5kV (as required by the low-voltage directive). This could be done by employing one of the thermally conductive sleeves which are sold for mounting live power transistors to earthed heatsinks. I used EPCOS part number B57020-M2502-A17 because it is supplied in an insulating sleeve.

The circuit is synchronised to the Economy 7 timer simply by switching it ON at the end of the cheap rate period. It was done this way because I am more likely to be awake and active at 8am than at 1am. Nocturnal readers should adjust the software to reverse the switching! Any reader who needs the software, please contact Caroline Fisher (details page 3) quoting CI 224 as the reference.

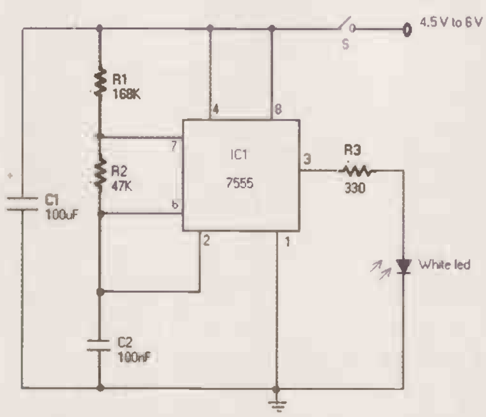
__Ian Benton__
Ilkeston
Derbyshire
UK

R_1 = NTC Thermistor 5k 25°C
R_2 = 3.9k ¼W
R_3 = 2.2k ¼W
R_4 = 1M ¼W 375V
R_5 = 47 ¼W
R_6 = 15k 2W
C_1 = 220μF 16V
D_1 = BZX55C 5V1
D_2 = 1N4004
D_3/D_4 = red/green LED
TR_1 = BC640
CSR_1 = BTA16-600BW
IC_1 = PIC12C671
VR_1 = 1k

LED torch

A common problem with small torches is the short life-span both of the batteries and the bulb. The average incandescent torch, for instance, consumes around 2 Watts. The circuit design described here is a simple torch light using a white led in place of incandescent bulb.

A white LED is, in reality, a blue LED (light-emitting diode) surrounded by a phosphorescent dye that glows white when it is struck by blue light. This is a similar process to that in fluorescent lamps, where the coating glows white when it is irradiated by the ultraviolet light that the tube generates internally. A white LED has a continuous spectrum similar to daylight, i.e. slightly blue. White LEDs are twice as bright as incandescent bulbs and will help preserve natural resources due to the use of non-toxic materials. While the forward voltage drop of



a traditional green LED is between 1.8V and 2.7V, a white LED has a higher forward voltage drop that lies between 3.1V and 4.0V depending upon the manufacturer. This means that whereas a green LED can be powered directly from the commonly used Li-Ion battery with a linear regulator and a ballast resistor, a white LED requires the battery voltage to be boosted.

The LED torch consumes between 20 to 40mW, giving it more than 50 times longer service from 4 AA alkaline batteries. This torch is based on a 7555 timer (IC_1) running in astable mode. A white LED (D_1) produces 400 mcd light output and when focussed, can illuminate objects at 20 metres.

A convex lens with short focal length is placed in front of the LED to focus the beam. If banding occurs at the beam's perimeter, use another very short focal length lens directly in front of the LED to smooth the beam.

__D. Prabakaran__
Tamilnadu
India

Precision A-weighting filter

Most good value designs of A-weighting filters for consumer audio which correspond with the NAB/ANSI S1.4-1986 standard (not the CCIR 468 standard, which is mainly used in studio environments) try to solve the filter transfer function with passive components only - followed by an OPA to create the required gain at 1kHz (0dB). The disadvantage of these approaches is the fact that the original filter requires 2-pole solutions at both ends of the filter's frequency range which cannot be resolved by passive components on their own. There must be at least one active device somewhere between the filter creating components which form the two 2-pole solutions.

To overcome this problem, a satisfying approach could be the adaptation of an A-weighting filter design proposed by Mr. W. Adam in Figure 6 of his very interesting 1989 article[1] **Figure 1**. A PSpice comparison between the ideal A-filter curve **Figures 2 & 4** and figure 1's transfer curve shows significant differences (+3.2 / -3.7dB, **Figure 5** curve X). Although these figures fall within the tolerance range set by the standard, it is not sufficent when talking about precision. Only a handful of additional components will easily produce far better 'good value' results (+0.25 / -0.12dB, **Figure 5** curve Y).

The OPAs shown in the improved W. Adam design **Figure 3** can be OP27, TL071 or similar devices. Resistors are 1% E96 types, capacitors are measured within 1%.

Burkhard Vogel
Stuttgart
Germany

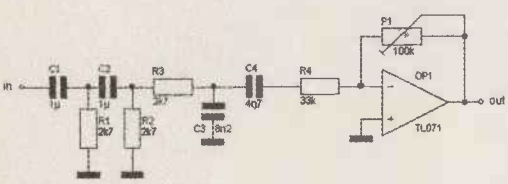


Figure 1: Original W Adam design (1989)

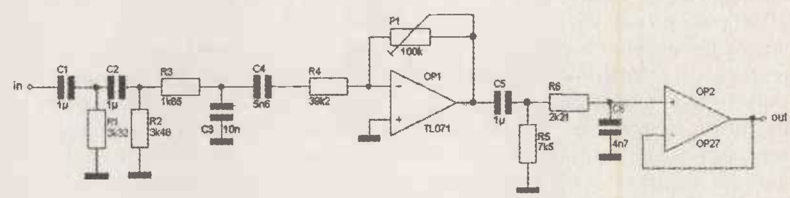


Figure 3: W Adam design improved

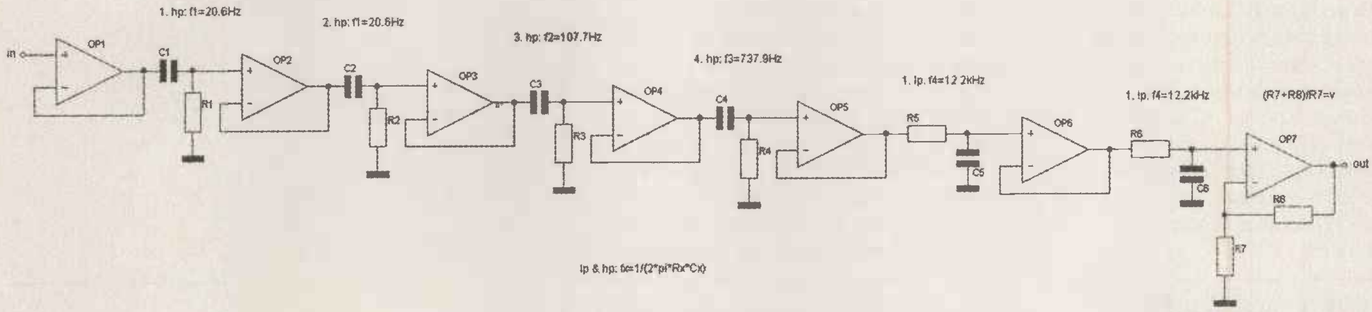


Figure 2: Ideal A-weighting filter (ANSI & NAB)

Figure 4

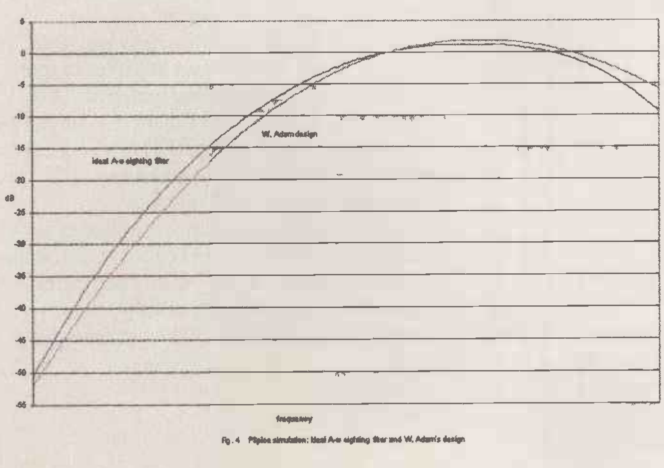
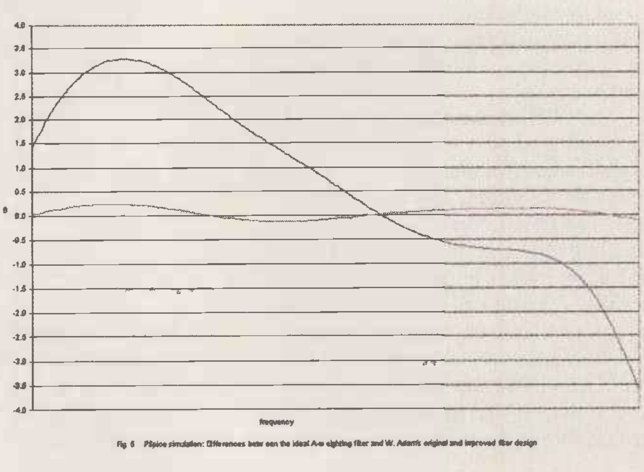


Figure 5



Appendix

1. A-weighting filter poles (ANSI)[2]:

2 at f1=20.6Hz
1 at f2=107.7Hz
1 at f3=737.9Hz
2 at f4=12200Hz

2. A-weighting filter transfer function A(f) (figure 4)

$$A(f) = v * \frac{1}{\sqrt{1+(\frac{1}{f*\frac{1}{f1}})^2}} * \frac{1}{\sqrt{1+(\frac{1}{f*\frac{1}{f1}})^2}} * \frac{1}{\sqrt{1+(\frac{1}{f*\frac{1}{f2}})^2}} * \frac{1}{\sqrt{1+(\frac{1}{f*\frac{1}{f3}})^2}} * \frac{1}{\sqrt{1+(f*\frac{1}{f4})^2}} * \frac{1}{\sqrt{1+(f*\frac{1}{f4})^2}}$$

'v' is the amplification of the A-filter to create 0dB overall gain at 1kHz:

$$G(1kHz) = \frac{1}{\sqrt{1+(\frac{20.6}{1000})^2}} * \frac{1}{\sqrt{1+(\frac{20.6}{1000})^2}} * \frac{1}{\sqrt{1+(\frac{107.7}{1000})^2}} * \frac{1}{\sqrt{1+(\frac{737.9}{1000})^2}} * \frac{1}{\sqrt{1+(\frac{1000}{12200})^2}} * \frac{1}{\sqrt{1+(\frac{1000}{12200})^2}}$$

$$v = \frac{1}{G(1kHz)} = 1.25889663 \int 1.9998014\ dB$$

3. Results:

A f Hz	B A(f) fig. 2 ideal dB	C A(f) fig.3 measured dB	D Δ B - C dB
31,5	-39,53	-39,37	-0,16
200	-10,85	-10,70	-0,15
1.000	0,00	0,00	0,00
2.500	1,27	1,19	0,08
6.300	-0,11	-0,22	0,11
12.500	-4,25	-4,25	0,00
16.000	-6,70	-6,68	-0,02
20.000	-9,34	-9,16	-0,18

4. Originally A(f) was created for noise measurement purposes only. A-filtering will 'improve' any noise figure in a given audio bandwidth B (eg. 20Hz - 20kHz) by a factor of a(B):

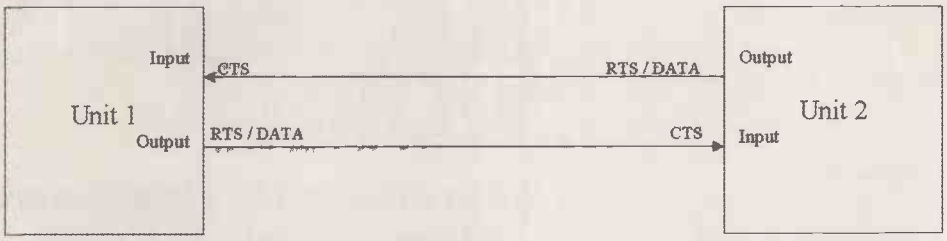
$$a(B) = \frac{1}{B} * \int_{20}^{20000} A(f)df = 0,74 \int -2,62dB$$

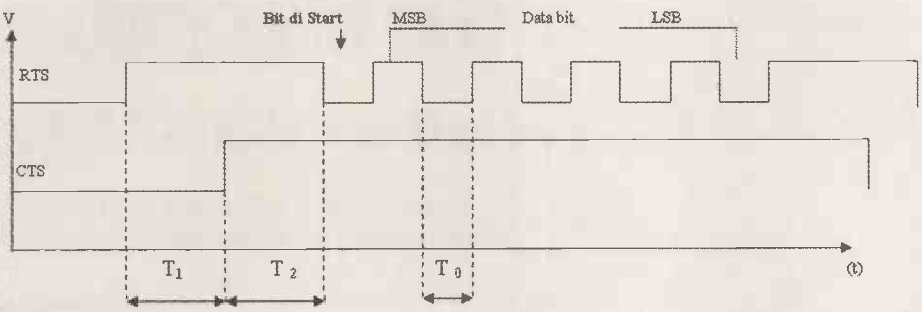
My own CLIO40 16Bit measuring system could verify this factor as well.

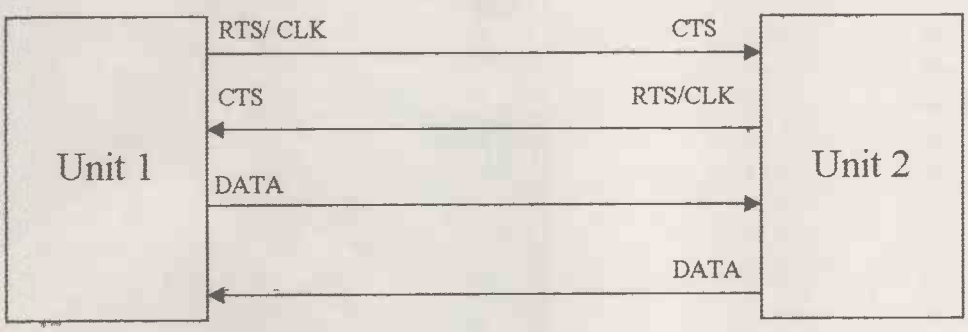
References:

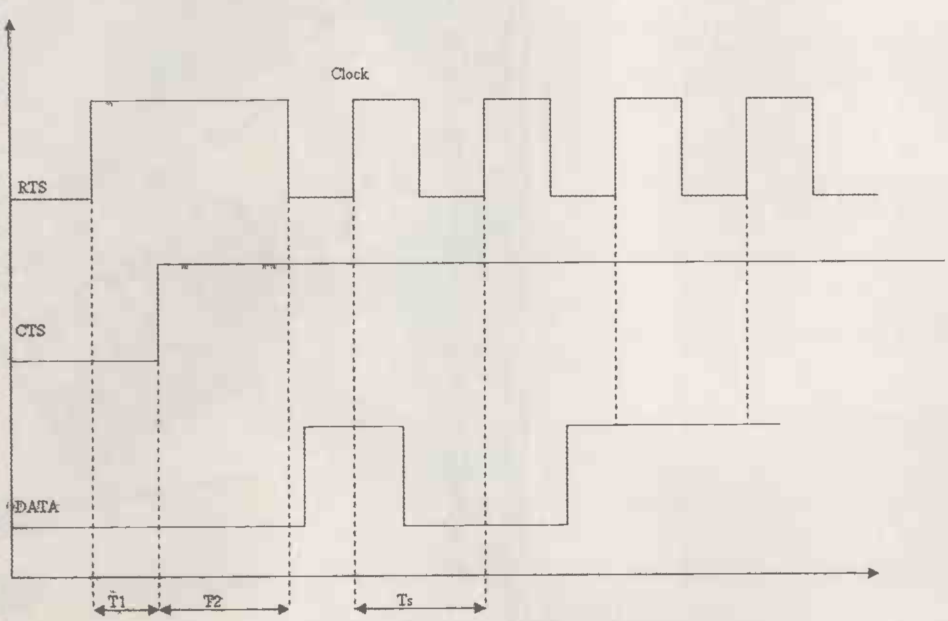
1. Wilfried Adam: Designing low-noise audio amplifiers *E&WW* June 1989, p. 628ff
2. Product Technology Partners: Noise measurement briefing, **www.ptpart.co.uk/noise.htm**

Two wire flow control









T1 and T2 are the timeout and Ts is the sample time.

To develop a data transfer system, one often needs a data flow control between the transmitter and the receiver. The popular RS232 protocol is a simple mode to transfer data that has the RTS/CTS (Request To Send/Clear To Send) protocol as data flow control. This protocol needs 4 wires to link the two units. Here, I show that it is not generally necessary to have 4 wires, and that two wires can suffice to develop a data flow control. This is an advantage in the microcontroller system design.

The equipment, labelled as Unit 1 and Unit 2, must have an output pin to send the data and an input pin to receive the data, the diagram in **Figure 1** shows a system for an Asynchronous Transfer Mode.

For simplicity, we assume that the 'no-data-to-transfer' state is OFF. When a Unit has to transfer data, it sets the output pin to ON, and waits for the other Unit to set the other line (its input pin) to the ON position. Similar to the RTS/CTS protocol, the Unit will send the data after it has verified the ON state of the input pin.

In this mode the output pin is used as Tx and RTS signal, the input pin is used as Rx and CTS signal. **Figure 2** shows the signal timings.

T_1 is defined as the transmitter timeout; T_2 is defined as the receiver timeout, and T_o is the bit rate. These three values must be chosen by the designer for each specific application. Similarly, it is possible to develop a system for a Synchronous Transfer Mode, in this case both units must have an Output pin to send the data, a second Output pin to generate the RTS and Clock signal and an input pin to receive the data. **Figure 3** shows the schematic and the **Figure 4** shows the timing for a Synchronous Transfer Mode.

T_1 and T_2 are the timeout and T_s is the sample time.

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Popoli
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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'.

Preamps please

Please – in *EW* you have published lots of amplifiers over the years, but never any preamps - especially with a wide range of inputs for MC, MM, Phono, DC, Tape etc. Please could you amend this large gap!

R Phillips
Bournemouth
Hampshire
UK

Any takers? – Ed

Talcum powder rejuvenation

First, thank you for taking your job in a serious, enthusiastic and clearly-minded way. I enjoy reading *EW* or *WW* as much today as I did more than forty years ago. I have filled in the reader survey and will send it tomorrow.

Second, may I suggest a tip which can save much time, labour and anger, though it's admittedly very low-tech.

Ever heard an old floppy disk squeak grimly in the drive as you try to reach the files you want back after some years of quiet oblivion? The sound generally anticipates what the screen message confirms a moment later: you won't get your data.

If this happens, don't try again! You'll easily warp the disk surface, which has lost its anti-friction properties probably for absorbed humidity. You might as well damage the drive's heads or impair their alignment.

Here is a solution which has worked in 80-90% of my cases, and is so rough and simple that surely many will take it as a joke:

Take the diskette, reverse it and let some talcum powder, of the standard after-bathing type, in its central opening and shake well to distribute the powder evenly inside the diskette casing; then blow the excess powder away (you can do it more technically with low-pressure compressed air), pushing the metal slide aside to open the head accesses.

Unless poured in kilograms, talcum powder won't clog the drive's heads nor damage its internals in any way: it is too soft to be abrasive. In fact it will just absorb humidity inside the diskette casing and reduce friction dramatically.

Before taking the floppy disk back into the drive, test the evenness of its rotation by hand: if you feel some point of friction repeat the process. When it 'feels' OK, go ahead and you'll be pleased to hear no more squeaking and, if neither the disk nor the drive haven't been damaged before, to access your data normally.

This trick worked perfectly with over forty old floppies of mine plus some of my friends', which seemed lost forever.

The fluid-absorbing and friction-reducing capacity of talcum powder also helped putting an old, good quality cassette deck back into service. I found that the lubricant in the transport mechanism had hardened to the point that some sliding parts were stuck to each other, moving with difficulty. I tried to take the old lubricant away with a solvent, then gave a drop of fresh, fluid oil: but to no avail, as the intricacy of the mechanism did not let the solvent nor the lubricant reach the key points.

Before proceeding to dismount the (rather complex) mechanism, I decided to try the talcum powder trick. I threw in a little and then activated the commands repeatedly: some parts began moving almost immediately, showing that it did work.

For the most badly stuck parts I had to repeat the operation three times, but finally (after two days) the talcum powder absorbed all old and new lubricant residues, then came off quietly, allowing free movement - and final light re-lubrication. The transport now works normally.

I use standard commercial talcum powder sold in any supermarket, but would suggest to avoid the scented types, as any aether oil even in minimal quantities might possibly damage plastic parts in the long term.

Usual disclaimer: no liability implied or accepted, try at your own risk, don't steal talcum powder from your wives.

Eugenio Costa
Sanremo
Italy

'Leakage'

Physicists attended the 17th International conference on relativity in Dublin recently where the hot topic of conversation was Stephen Hawking's announcement that he had solved the 'information paradox'. 'Leakage' seems to be at the root of the solution, but Roger Penrose speaking to BBCi News, said "I think he's going the wrong way" Young readers who may be a bit confused about what's going on might find Jacob Bekenstein's article *Information in the Holographic universe (Scientific American, August, 2003, pp48-55)* helpful in explaining the consensus view of most of the physicists working in this area.

'Leakage' however, is well supported. It was invoked to explain phosphorescence which we can see in LEDs, CRT traces etc.; and sensory pathologies associated with myelin around axoas, such as motor-neurone disease (ALS) and adenoleukodystrophy suggested that, although erussic acid (Lorenzo's Oil) will not cure ADL, it will prevent it in about 90% of siblings genetically pre-disposed to ADL. Ramachandran also used 'leakage' as an explanation of several cognitive pathologies although, in this fold, excessive leakage was implied; for normal development, especially speech, some leakage is essential. An interesting finding has just been published which shows that 'sign-exposed' babies of deaf parents produce two distinct types of hand waving at 1 and 2.5Hz compared to speech exposed babies who only who only show the high frequency pattern (rhythm). Laura-Ann Petitto believes this lower frequency (probably 0.8Hz) is a "unique rhythm characteristic of natural language" ..."passed on genetically". Rama's work supports the former conclusion, but we disagree with the genetic mechanism, unless we equate genetics to evolution. What is more likely explained is that babies' brains resonate with the Earth-Moon resonance's (longitudinal), at 0.8 and

0.4Hz as we suggested[1] and these frequencies are controlled by the bifurcation constant quotient 1.87. In this sense we can see, if vaguely, the possibility of the relation between the gravitational sensor Gun and electro magnetic sensor Tun which Einstein was convinced of up to the end. In the 5 D-anti-de sitter universe described by Bekstein, ϑ and λ are both attractive but attracted towards different attractors; suns, plants and moons for ϑ black holes for λ!

Tony Callegari
Much Hadham
Hertfordshire
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[1]New Scientist 17/7/24, p.8

Patent spoof?

I notice in the New Products section of the September edition of *Electronics World* that a colo(u)rimeter using a beam from an LED has been patented. It is difficult to see what can reasonably be said to be sufficiently novel to be worthy of a patent. Certainly not measuring haemoglobin concentration by means of colo(u)rimetry. Nor is the use of an LED to produce a substantially monochromatic beam or even using two with different wavelengths, one wavelength being strongly absorbed by haemoglobin and used to measure concentration. The other wavelength being used to correct for absorption by the solution in which the red blood cells are suspended and by the cuvette.

It is of course entirely possible that the reason for opting for a patent rather than a registered design is in fact altruistic. By preventing a company taking out a patent on such a device it becomes possible to ensure that it is not priced out of reach of medical teams in poorer countries. But in either case it is a very poor reflection on the patent system.

In biology we are already seeing the problems of patents with much too wide a scope being granted. A cancer researcher who creates a trans-genic strain of animal cannot make use of individuals in that strain without the agreement of an American company which bought up a patent granted to cover the creation of any trans-genic animal.

We seem to be moving into an era of 'unlimited patentability' where the main criterion is not novelty but whether something is capable of commercial exploitation. Thus we have the Nippon Electric Company trying to obtain a software patent on the principle of an optimising compiler, (EP0646864). By no stretch of the imagination can this be said to be 'novel'.

It has been claimed that the motivation of the European Patent Office in granting patents which make sweeping claims on essentially non-novel ideas is that they make money out of it. I'll leave that for the Euro - philes and -phobes to argue about.

Dr. Les May
Rochdale
Lancashire
UK

Magnoflux nonsense

I was very disappointed by the Magnoflux Aether article. His entire theory falls apart, he has negative neutrons that he recognises the field of physics disagrees with but proceeds with the "in reality they are negative" where is he deriving this information from, perhaps the Nobel council should be informed? His justification is that the e/m ratio must be balanced as it is a constant for all matter. Wrong - it is a constant for all electrons, or protons, etc. they have very different masses and hence very different e/m ratios and that's not even taking relativity into account.

The irony is that you ran this asinine article written by a chemical engineer on physics and you ran it in the same magazine with an article on pseudo-science.

Even if he is on to some new starting theory, shouldn't it be sent to Nature or Physical Review for peer review, hardly appropriate for an electronics magazine.

"Science is like sex, it has its practical purposes, but that's not why we do it." -Richard Feynman (Theoretical Physicist).

Daniel Lemieux
By email

Cyril's conundrum

I offer this in response to Ian Cuthbert's request for any 'Any takers?' (*Letters* page 49 September 2004).

Remove the resistor along the edge across which the total resistance is to be determined. In this case the resistor connected between nodes A and B. (This resistor will be included again in the last stage.) The network is then rearranged in two dimensional form. **Figure 1**

Two ways in which analysis can proceed from this are given here.

A) The circuit symmetry and all the resistors having the same value allows branch currents to be assigned as shown.

As

$V_{AB} - V_{AC} - V_{CD} - V_{DB} = 0$

then

$V_{AB} - I_1R - (I_1 - I_2)R - I_1R = 0$

and

$V_{AB} - 3I_1R + I_2R = 0$ (1)

From loop CDFE

$V_{CE} + V_{DC} + V_{FD} + V_{EF} = 0$

then

$I_2R - (I_1 - I_2)R + I_2R + 2I_2R = 0$

giving

$I_2 = I_1/5$ (2)

Using (2) to substitute for I_2 in (1) gives

$V_{AB} = (14/5)I_1R$ (3)

and because

$R_{AB} = V_{AB}/2I_1$ (4)

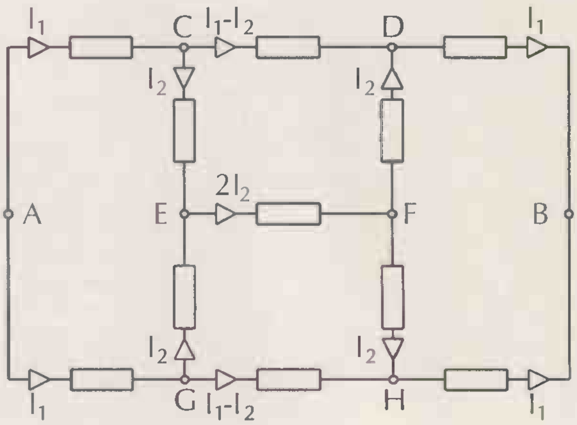
then by using (3) to substitute for V_{AB} in (4) produces $R_{AB} = (7/5)R$

Placing the original resistor back in the circuit, R in parallel with (7/5)R, to find the total resistance which produces (7/12)R

B) Star to delta transformations can be applied to the stars centered on nodes C, G and F. This will produce three sets of two resistors in parallel. Combining these will produce a star embedded in a delta. Transforming this star to a delta will allow further simplification and ultimately yields the same result.

Kerry Bodman
By email

Figure 1



Further to Cyril's Cube Puzzle

I take up Ian Cuthbert's challenge (*EW* Sept. 2004) to show that the perceived resistance of the cube when measured between opposite ends of one edge is 7R/12.

Figure 1 shows the cube constructed from resistors of identical value, R. We wish to calculate the resistance between corners A and B, symbolised as WAB. Using the same kind of symmetry arguments presented by Cyril (*EW* Dec. 2003), we see that corners H and F have identical potentials and can therefore be connected together, as can corners C and G. The resultant simplified

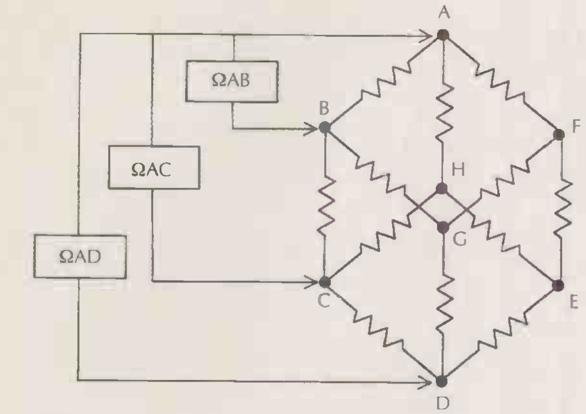


Figure 1

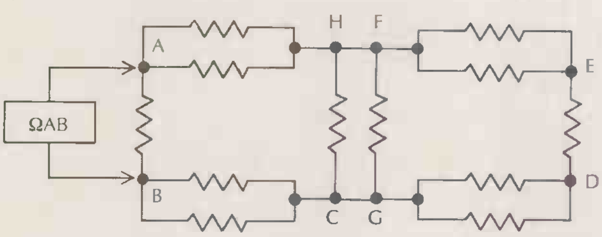


Figure 2

network is shown in **Figure 2**, which calculates out as 7R/12.

For completeness, I identify a further resistance measurement that can be made, WAC, which is between the opposite corners of one side of the cube. With a little thought it can be seen that corners B, H, G and E are all at the same (mid) potential. These corners can therefore all be joined together, which has the effect of short-circuiting the resistor connected between B and G, and also that between H and E. Alternatively, since no current flows through these two resistors, we could simply omit them.

Both of these simplifications are represented in the simplified network shown in **Figure 3**, where the dotted line can be treated as either a short-circuit or an open-circuit (I think it easiest to assume the latter). The resultant resistance is then easily calculated to be 3R/4.

The final resistance measurement,

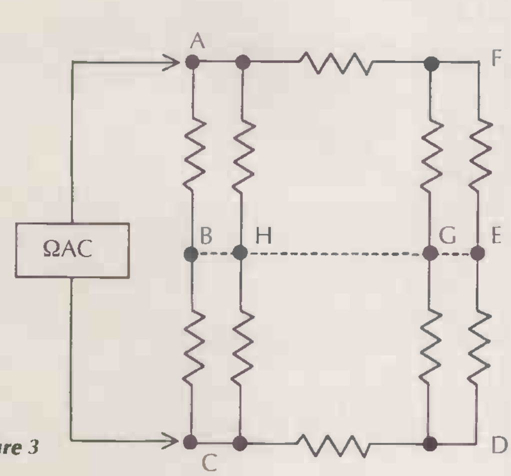


Figure 3

ΩAD, is that which has already been described by Cyril, and has the value 5R/6.

There are no other resistance measurements possible for a cube comprising identical resistors.

I note that the ratio of the three resistances ΩAB:ΩAC:ΩAD is 7:9:10. Any significance in this, anyone?

Steve Hughes
Waltham Chase
Southampton
UK

Powers that be

In response to the letter from Mr. Skeggs in September, from the indices rules, indices are added in order to multiply.

Thus: 10^1 can be shown as:
$10^1 = 10^{0.5} \text{ X } 10^{0.5}$

Anything to the power 0.5 gives the square root. 10^1 = Sq root 10 X Sq root 10

Thus: $10^1 = 10$. Also 10^0 can be shown as: $10^0 = 10^1 \text{ X } 10^{-1}$

As anything to a negative power is a reciprocal and from the previous example, then, 10^0 = 10 X 1/10

Thus: $10^0 = 1$.

Also $10^{1.5}$ again from the first example, can be shown as:
$10^{1.5} = 10^1 \text{ X } 10^{0.5}$
= 10 X Sq root 10

Thus: $10^{1.5}$ = 10 X 3.16227 = 31.6227 approx.

I had an enlightened maths teacher some time in the 1950s, I guess about the same time as Mr Skeggs was struggling, who did not leave any loose ends.

May I suggest a beautiful book called *The Nothing that is* by Robert Kaplan that tells how the ancients thought through some of these mathematical quirks.

Robin Duffin.
Northwood
Middsex
UK

Powers that be II

Mr. Skeggs should have been given a better explanation of why any number to the power 0 is 1, and about fractional and negative powers.

Mathematicians like consistency. It's good to have a notation that remains logical for all possible cases. Consider, for example, 32 ÷ 4 = 8. This is exactly $2^5 \div 2^2 = 2^3$.

This indicates a general rule: $X^a \div X^b = X^{(a-b)}$, and if you try it you find it always works. Mathematicians have a more rigorous proof than 'it always works'. Now consider 4 ÷ 32 = ⅛. Applying the above rule get $2^2 \div 2^5 = 2^{(-3)}$. So that indicates that $2^{(-3)} = 1 \div 2^3$ = ⅛.

Once again, if you try examples, it always works and there is a better but much more abstruse proof that it works. Now consider 4 ÷ 4 = 1. We get $2^2 \div 2^2 = 2^{(2-2)} = 2^0 = 1$. Again, it always works.

What could $2^{(1/2)}$ mean? Well, $2^{(1/2)} \text{ x } 2^{(1/2)} = 2^{(1/2 + 1/2)} = 2^1 = 2$. So $2^{(1/2)}$ is the square root of 2.

$2^{(3/2)} = 2^1 \text{ x } 2^{(1/2)}$ = 2 x sqrt(2) = sqrt(2^3) = sqrt(8). And $2^{(-3/2)}$ is the square root of ⅛.

On another subject – I understand that almost all of Turing's work during WW2 was classified until at least 30 years after 1945, so I doubt that it was discussed even in academia before then.

I also have comments on *Class A Imagineering*. Mr. Ward (*Letters*, September) has been working things out for himself, which should be encouraged, but has picked up some popular misconceptions on the way. I suspect he will soon realise their falsity, even if he doesn't immediately believe me now.

Asymmetrical non-linearity does indeed generate second harmonics, but also a succession of higher even-order harmonics, depending on how sharp the non-linearity is. Equally, symmetrical non-linearity produces third and a succession of higher odd-order harmonics.

Even-order harmonics sound 'better' because all of them except even multiples of 7, 11, 13 ... are consonant, whereas dissonance starts with the 7th in the series of odd-order harmonics.

Mr. Ward is quite right about loudspeaker impedance at 200Hz being close to the nominal (strictly 'rated') value. For small loudspeakers, the frequency is often a bit higher. But beyond that the popular errors set in. Loudspeakers give a flat(tish!) frequency response with a constant VOLTAGE input, not constant POWER. Provided the impedance doesn't go LOW, so that the amplifier runs out of current, it will quite happily provide the necessary constant voltage to 8 ohms and to 80 ohms. At least, it will if it has a low output source impedance (e.g. 0.4 ohms or less in the case in question).

Transformers (properly designed) are not 'inductive'. The impedance on one side appears on the other side multiplied or divided by the square of the turns ratio. The transformer's magnetising inductance (effectively in parallel with the primary) should be a high impedance at even the lowest frequency, and the leakage inductance (effectively in series with the load impedance) should be a low impedance at even the highest frequency. (More complex models can be made, but that is good enough for most purposes.)

You do NOT want the amplifier /loudspeaker interface to be a 'match' in the sense of 'equal impedances'. Loudspeakers want constant voltage drive, which means that the amplifier output source impedance needs to be much lower than the loudspeaker's impedance (but not ridiculously lower: less than 1/20th is quite good enough). The transformer should be as nearly 'transparent' as possible, i.e. its own impedances (magnetising and leakage inductance and the winding resistances, plus winding and interwinding capacitances) should be negligible at all frequencies. It IS possible, but it requires careful design and often a generous budget.

Some valve amplifier don't have such a low output impedance. In that case, more volts are delivered to the loudspeaker at frequencies where its impedance is higher. This occurs around the main low-frequency resonance and at high frequencies where the voice-coil inductance (1 to 10mH, typically) has significant impedance compared with the voice coil resistance. The result - 'good bass response and a reasonable top end' to quote Mr. Ward. But now you know why!

Why isn't the bass 'so deep' as with transistor amplifiers? Because the valve output transformer isn't ideal. Below some low frequency its performance drops off, whereas a transistor amplifier may go down to nearly DC. If you make the transformer bigger, heavier and more costly, you can get lower frequencies through it. Measuring the output voltage of a valve amplifier with the loudspeaker as a load IS valid, but only for that loudspeaker; a different loudspeaker will give a different answer, valid for that loudspeaker alone.

Mr Ward is quite right to say that the impedance of a loudspeaker is not a pure resistance at most frequencies. Normally, it is resistive at only two; the exact main low-frequency resonance, where it may be 40 to 80 ohms, and at that mid-frequency point that Mr. Ward mentions, where the voice-coil inductance and the motional capacitance are series resonant, giving a minimum in the impedance/frequency curve, which may be 6.4 ohms. At all other frequencies, it is more or less reactive, so that the voltage and current are not in phase and the actual power absorbed is anybody's guess. 95% to 99% only goes to heat up the voice coil, anyway.

All this means that amplifier 'power' doesn't matter (but few will believe that!); it's the voltage and current capability, considered separately, that matter. The voltage capability is responsible for the sound pressure, while the current capability ensures that you get the required voltage even if the loudspeaker impedance at the signal frequency is low (an '8 ohm' loudspeaker can go down to 6.4 ohms and still be legitimate, but some go much lower, and crossover networks can present very low impedances to impulsive signals).

John Woodgate
By email

Class A imagineering I

I tried to study the article mentioned in the header, but I do not understand why you publish an article that is so far below your normal standards. For one the style of writing is miserable. A sentence lasting for 33 lines is very difficult to read, not only for me as a non-UK reader, but for everybody. Could you please make sure that a normal writing style is used in future?

A second objection is about the content. Somebody who needs a whole column to explain that the group delay of an RC circuit is only constant up to (approx.) the -3dB point is clearly not understanding what is really going on. Furthermore, the example taken (10k ohms, 1nF) has a 3dB point at 16kHz. Certainly no decent amplifier design will be using such values. What Mr. Maynard is trying to explain is that if a signal contains frequencies that are outside the bandwidth of an RC circuit, some frequencies will be attenuated. I really do not agree with him calling this distortion.

A little further on in the article, he refers to the output choke and calls the (linear) distortion it causes the reason that loudspeakers can sound differently when connected to different amplifiers. He does not give any proof of this statement, which I think not acceptable. We are interested in facts which are obtained from measuring or from listening, but in science there is no place for vague and not substantiated suggestions.

Michel Nieuwenhuizen
By email

Class A imagineering II

I started reading Graham Maynard's articles (*Class-A Imagineering*) with no prospect of acquiring any new knowledge or wisdom. I found it funny at first, although I don't think that was intended, but Graham's turgid and obscure style soon became tedious. There seemed little or nothing of value from a technical perspective, and the recounting of his experiences along with shonky interpretations could hold my attention for only one or two pages at a time. One needs to be clear when writing technical articles, and when the writer introduces new terms or concepts it is incumbent on him or her to explain them. Just what does "phasily 'tonal' and 'pass-bandy'" mean? "Class-A Imaginings" would be a more accurate title. I don't doubt Graham's enthusiasm nor that his belief is genuine, but unfortunately his experience setting up discos has left him with some egregious misconceptions.

The failing of Graham's philosophy – and it is a philosophy, not a technology – is that a listener can identify failings in an audio system better than measuring instruments. This has never been demonstrated as far as I am aware, and is in complete contradiction to scientific knowledge established decades ago. Although extremely unlikely, Graham may posses the hearing of a dog or a bat, but while that might be a cause for scientific curiosity it is not a rational reason for changing proven engineering techniques based on science.

The difficulty with listening tests is that they rely on the reporting of the test subjects (i.e. people), which is notoriously unreliable. What Graham must do before he can be afforded any credibility is to furnish proof that he or others actually heard the audible effects they claim. To do that he must detail what processes were used to avoid pitfalls such as the Placebo Effect and the Expectation Effect. This requires quite extensive and rigorous efforts, and simple lounge room auditions (for example) are useless. Double blind and AB testing are designed to avoid these traps. He must also demonstrate that if a statistically significant effect was demonstrated there was no measurable difference.

Furthermore it is unacceptable for a group of people listen and discuss together what they perceive. It is easy to demonstrate (and has been) that in such circumstances the members of the group soon start to agree about what they perceive even when those perceptions are false.

It is essential that test subjects do not discuss their perceptions before the data are collected, and it is also essential that the test subjects are unaware of what specific hardware they were listening to when they record their impressions (blind testing). It might be acceptable for subjects to know what equipment is being used and the devices are under test, but it is essential that they do not know which particular device they are listening to when they make their assessment. When these basic requirements have been satisfied subjectivists invariably show the same insensitivity to measurable effects as normal people with less than golden ears. Sometimes the golden eared report audible effects that could not exist. Indeed, it was reported (http://www.verber.com/mark/cables.html) that some subjectivists claimed to be able to detect a difference between cables when unbeknownst to them only one cable was used.

Over the decades that I have followed the subjectivist vs. rationalist (more appropriate than "objectivist") debate I have been continually impressed by the lack of technical expertise displayed by subjectivists. For example, they complain that sine wave testing does not reveal the complete behaviour of audio equipment. That is true to the extent that it reveals only linear distortion, so a full set of specifications must include data such as output power, THD, slew rate (full power bandwidth), noise, etc. The curious thing is that the human ear breaks the incoming sound into a set of frequency bands (that is, essentially a set of sine waves), and the brain recombines these into the sound we perceive. Sine wave testing would seem to emulate the behaviour of the human ear pretty well, but furthermore anyone who has had any experience with Fourier analysis or Laplace Transforms would know that the frequency response infers the transient response. Frequency response and transient response are two sides of the one coin. It is sometimes difficult to calculate one from the other, but in most circumstances relating to audio equipment that is not the case.

Graham demonstrably fails to address the fundamental issue when in his second article he rehashes the issue about the response of a first order low pass filter. The 3dB point (63kHz) is way above the limit of human hearing, but he seems to claim that it produces audible phase shift. He shows that the filter does not behave exactly like a pure time delay, but this is neither new nor remarkable. Nobody claimed that it did (although I was stupidly accused of holding that view). The point at issue, which Graham has avoided, is whether the phase shift is audible. The range of frequencies where the phase delay is significant and not like a simple delay (above 20kHz) is outside human perception, and over the range of frequencies where humans are sensitive (slightly) to phase (below 10kHz) the difference between the filter's response and a pure delay is inconsequential. I didn't even claim that inserting the filter produced no audible effect, simply that the phase response is not a rational explanation. The filter could have increased the noise in the audio band for example, or it may have reduced slew induced distortion by limiting the rate of rise of the incoming signal. Perhaps it blocked some RF interference that created audible distortion. Graham does not tell if these possible effects were studied and discounted, but they are plausible while the phase shift theory is not.

Tests have been performed to establish if certain filters produce audible differences (http://www.pcavtech.com/abx/abx_f4.htm). It was reported that a fourth order filter with a 3dB cutoff at 17kHz did not produce any verifiable difference to the perceived sound, although filters with slightly lower 3dB frequencies did. If the 17kHz filter with its savage phase response (about 180 degrees at 17kHz) has no demonstrable effect, then it is risible to suggest that a first order filter with a mere 45 degrees at 63kHz (and nothing significant below 20kHz) will. Furthermore, the 17kHz filter will introduce significant phase shift at much lower frequencies (about 60 degrees at 6.8kHz), but that had no demonstrably audible effect. Clearly human hearing is not very sensitive to excess phase shift. Graham needs to have his hearing tested.

Trying to have a sensible debate with subjectivists is like chasing the soap around the bottom of the shower. They deny or lack knowledge of established science, and cannot or will not produce the slightest shred of credible evidence for their claims. People claim to have seen Elvis but we don't take them seriously. We don't grant them space in technical journals to justify their wild claims, and we certainly shouldn't pay them to expound their views.

Which raises the question, why does this subjectivist mumbo-jumbo get so much space in *EW*? Is there such a paucity of technical expertise across the editorial staff that they cannot discern valid science and engineering from this tripe? Has the editorial staff been stacked with acolytes of this religion? Why should *EW* waste readers' money and misinform many when there are plenty of other magazines devoted to faith based audio? (I hesitate to use the term hi-fi there because frequently subjectivist equipment is not.) I don't think you appreciate how silly, or maybe cynical, it makes *EW* appear to informed readers, and you do a considerable disservice to those who wish to learn the truth. Those proffering subjectivist articles for *EW* should be turned away at the door.

Science and engineering are not democratic disciplines, if you cannot back up your claims with credible evidence then your theory is suspect and your opinions pretty worthless. One cannot simply choose which parts of science to accept and which parts to deny as false. Furthermore if one's philosophy is discounted by the bulk of long established scientific knowledge then you are almost certainly wrong, and if you are wrong you are wrong, end of story. We don't expect Satanists to be given equal time in pulpits, and we would be rightly alarmed if the Astronomer Royal were to believe in a flat earth. Publishing this subjectivist rubbish might fill pages, but it is not really possible to have a sensible debate between subjectivists and rational engineers in my experience, so I don't believe you are providing any positive public benefit by wasting our time and money on subjectivists' ravings.

At the head of his fine piece titled Audio amplifier distortion is not a mystery (*Wireless World*, Nov 1977) the great Peter Baxandall had this quote from Bertrand Russell: "Some things are believed because people feel as if they must be true, and in such cases an immense weight of evidence is necessary to dispel that belief". One can only imagine how dismayed PB would feel to see how the standards of a once fine and reliable journal have slipped. By publishing articles such as Graham Maynard's you only make the task of dispelling these preposterous philosophies much harder. I call on you to cease and desist forthwith. And I recommend to Graham that he study audiology, and not bother us again until he can demonstrate some understanding of the field.

I thank you for your time.

Phil Denniss
Sydney
Australia

I'm afraid that both myself and the bulk of readers commenting on Graham's articles disagree with you. I have for many years been in the entertainment industry and can confirm that there are those with both 'golden ears' and 'golden eyes' who seem to be able to spot things that mere test equipment can't. But I do agree that a blind test would be a fine idea. – Ed.

Imagineering and Catt

I returned home from a lengthy visit to the UK (where I did not read any *EW* issues) to find that during my absence John Linsley Hood had passed away.

JLH made outstanding contributions to the science and, dare I say it, the art of amplifier design. We readers will miss his articles in *EW*.

Permit me to comment on Mr. Maynard's articles in the June, July, August, and September, 2004 issues of *EW*.

A jury is instructed to wait until ALL the evidence is in before beginning deliberations, but I am unable to restrain myself any longer.

It appears that Mr. Maynard is advancing the magnitude of First Cycle Distortion (FCD) as the reason for the differences he experiences in listening to various types of amplifiers. He further suggests that the 10kHz FCD should be kept well below 1%.

Although I have not carried out any listening tests myself, this is not relevant to a serious question I would like to raise. Let us apply the 10kHz FCD test signal to the terminals of a loudspeaker. (Any reasonably wide band amplifier of reasonably low THD but exhibiting 'significant' FCD can output a reasonable approximation to the theoretical 10kHz test signal merely by carefully pre-distorting the input to the amplifier).

Now consider the chain of sound reproduction beginning at the loudspeaker terminals and ending up at a presumed ideal measuring microphone several feet away. In this path we have a large number of components such as crossover networks, voice coils, diaphragms, cones, and the air path between the cone and the measuring microphone. The air path characteristics are determined by its density, its compressibility and its viscosity. As a result, it alone probably contributes a substantial amount of phase dispersion (i.e. a delay that varies with frequency) to the situation.

With this in mind, I would wager that the 10kHz FCD of this path is close to 100% for any loudspeaker on the market today. If I'm right, what possible benefit is there in reducing the 10kHz FCD of an amplifier to a fraction of a percent as proposed by Mr. Maynard?

Mr. Maynard also makes much of the unrealistic nature of a continuous sine wave. I believe that the 10kHz FCD test signal proposed by Mr. Maynard is also unrealistic. Its first derivative is not continuous at t=0. This is not possible in a signal derived from instruments playing in air, whether they be string, cymbal, triangle, glockenspiel, or drum. All these instruments playing in air involve lossy energy storage in the instruments themselves and lossy energy storage in the transmission medium (air). Given the large number of energy storage elements involved, no signal impinging on a microphone from such instruments can exhibit any discontinuous low order derivatives, let alone a discontinuous first order derivative.

It would be nice to capture a large sample of sound segments from various instruments playing alone and together. Then, a sensible test suite of realistic 'stress' signals could be determined.

On a different subject, in August, one of your correspondents (unsigned) asked for "Less Catt Please" (page 57, August, 2004). I am thankful that at least the request was not to banish Catt.

First, I must declare my interest. In 1962 and 1963, my wife and I were fortunate to live in the apartment directly below that of Ivor and his wife, Freda (in Culver City, California). We formed a good friendship, and our respective children played with each other.

It was always a pleasure to debate technical and non-technical issues with Ivor, although he was very difficult to pin down at times. To a degree, he personified the uncertainty principle – you could, perhaps, pin down his position, but not at the same time the speed with which he was changing his position. I can tell you from personal experience that he has not a malicious – but also not a diplomatic – bone in his body.

My ambition was to make a lot of money (then unfashionable in socialist Britain, let alone in semi-communist Wales whence I hail); his, I think, was to make a mark. Despite such divergent ambitions, we got along well. In 1964 or thereabouts, Ivor went back to the UK while we stayed in California where we still reside. I believe he has achieved his ambition.

Ivor could fairly be described as an iconoclast's iconoclast. Thank heaven there are such folks. For us personally, he justifies a solecism: "most unique character we've met".

I'm pleased that you rejected the request for "less Catt", but I'm disappointed that you said that you went along with the majority vote. I would hope that you would have rejected the request irrespective of the vote. On the other hand, I'm grateful that a majority of your correspondents are indeed supporting the concept of more Catt.

While reflecting on Ivor, I recalled another friend of mine, Dr. Peter Williams of Paisley University. He and I cut our electronic teeth together at Porth County Grammar School in Wales the early 50s. We both read *Wireless World* at our local library, and fortunately our parents tolerated our radio experiments. Peter, apart from being a generous contributor of his time to less developed countries, was a regular contributor to *Wireless World* in the mid 70s (see, e.g. March 1974, page 45, part of the *Circards* series). His articles were lucid and educational; his circuits were elegant. Perhaps he could be persuaded to contribute once again.

Keep up the good work.

Martyn A. Lewis
Pacific Palisades
California
USA

Foreign language

Leafing through the September issue I found my attention drawn to the letter from Robert Baines, in which he asks whether it would be possible for the article 'An Electric Universe' in the August issue to be translated into English.

Having subsequently studied the article in question with great care, I believe I can now answer Robert Baines' question without ambiguity: No, it would be quite impossible to translate this article into English.

John Eades
By email

Powers that be III

Regarding Mr. Skeggs letter in *EW* September, I too was once confused with this and many other areas of Math's and electronics. A college of mine explained it to me and it all slotted into place. The trouble is that when people write down numbers and formulae, they assume the reader takes a lot for granted like not writing the '2' in front of the square root symbol but cubed root has to have the number '3' in front as to distinguish between square root and cubed root.

On the power front you have to remember that every number is raised to the power of 1 and is also divided by 1, but you never see this written down when expressing numbers.

So when you take 10 raised to the power of 2 you have to remember that it is already raised to the power of 1. Similarly, if you raise 10 to the power of zero you again have to remember it is already raised to the power of 1 and as you remove this 1 and make it zero you are dividing that number by its self hence the result 1.

Adam Rouse
Cornwall
UK

NEW PRODUCTS

Please quote *Electronics World* when seeking further information

Fuel sensor goes live at Electronica 2004

Position and movement sensor solutions company Hamlin is launching a new range of non-contact automotive fuel level sensors at the Electronica 2004 exhibition in Munich from the 9th to 12th November. Although the sensor has been designed for the hostile environment of an automotive tank in mind, the product is very well suited to other applications where low cost, accurate and reliable fluid level measurement is required. The component is of modular design with a standard body and an easy to customise float and float arm. Accurate reliable level sensing is achieved using an annular magnet and programmable Hall effect sensing element.

www.hamlin.com

Osprey Metals becomes Sandvik Osprey

To reinforce its commitment to activities in the field of controlled expansion alloys (CE Alloys), Sandvik Materials Technology has recently changed the name of its subsidiary Osprey Metals to Sandvik Osprey. CE Alloys are currently being used for optical packages in imaging systems; guide bars on printed circuit boards in embedded computer products; carrier plates for advanced sensor devices; backing plates for cladding to conventional circuit board materials and assembly fixtures for use in fabrication of micro-processors. Sandvik Osprey sells its range of ultra-lightweight, silicon aluminium, controlled expansion alloys in the form of machined and electroplated carriers, housings, packages and structural components.

www.sandvik.com

New partnership targets returns

A new service will be offered by Aon Network Services and Amethyst Group to major electrical and consumer electronics retailers. 'No fault found' returns, can account for as much as 50% of all products returned. Aon, a risk management organisation will provide a call centre of trained agents skilled in guiding customers through the basic faults and operating procedures of electrical and consumer electronics products. If that fails to resolve the problem, a network of field-based engineers will be deployed to fix the problem in the customer's home. Only products with a major fault will be considered for return to the retailer.

Amethyst, customer management and supply chain solutions provider, will provide outbound retail and on-line logistics management and support by consolidating distribution of products from multiple suppliers through a single warehouse. It will also manage all returns via its centrally located distribution centre. The team will inspect and classify each product and determine whether it should be repaired, returned to manufacturer under warranty or scrapped.

www.amethystgroup-uk.com

Battery authentication for portable applications

Following recent news of 'exploding' mobile phones caused by counterfeit batteries, Microchip announced that its Keelog secure algorithm can now be used for battery authentication in portable applications. The technology allows an application to simply and securely differentiate between genuine and counterfeit batteries. Using counterfeit batteries can lead to a potentially dangerous situation for the end user. Using the concept of IFF – Identify Friend or Foe – where 'Friend' is a genuine battery and 'Foe' is counterfeit, Keelog with its proprietary encryption and decryption algorithms provides a high level of security without adding excessive complexity to the system.

www.microchip.com/keeloq

Smallest LMOS logic IC series grows

Toshiba Electronics Europe recently launched a series of low power consumption logic devices for portable systems such as mobile phones, PDAs and notebook PCs. The logic MOS (LMOS) IC family – the LVP – features an operating voltage range of between 0.9V and 3.6V, a choice of 12 basic logic gates and a selection of three package options: fSV that measures 1.0mm x 1.0mm, ESV that measures 1.6mm x 1.6mm and USV, measuring 2.0mm x 2.1mm.

Propagation delay t_{pd} is 2.5ns (typ) at V_{cc} of 3.3V. Operating temperature is from -40°C to 85°C.

www.toshiba-components.com

Microchip expands its MCU family

US-based microcontroller (MCU) supplier Microchip has expanded its 28 and 40/44-pin PIC flash portfolio with eight new devices. They are aimed at mid-range applications that require program memory of up to 32kB.

The new family offers standard flash and enhanced flash memory with endurance of up to 100,000 erase/write cycles and data retention of up to 40 years. The core performance is 40MHz (10MIPS), the voltage range is 2V to 5.5V and the temperature ranges from -40°C to 125°C. In addition, they comprise a 32kHz to 32MHz software-configurable internal oscillator, a 10-bit 100k sample-per-second analogue-to-digital converter offering up to 13 channels, as well as two analogue comparators.

The new MCUs are supported by Microchip's development systems including the MPLAB Integrated Development Environment (IDE), MPLAB C18 compiler, MPLAB ICD 2 in-circuit debugger and MPLAB ICE 2000 in-circuit emulator.

www.microchip.com

Please quote *Electronics World* when seeking further information

More power from Power-One

Power-One of the US has launched its latest AC-DC power supply – the MPU200-1048 – that provides a 4.2A 48VDC output. The 200W product has a field MTBF in excess of 1,000,000 hours, based on data compiled from an installed base of over 500,000 similar MPU-series products, making it suitable for telecom, datacom, medical and industrial applications. Among its standard features are: power factor correction (PFC), 85 to 264VAC input range and active current sharing with N+1 redundancy. Interface signals include remote sense, output good, input power fail warning and global inhibit. Internal protections are provided for overtemperature, overcurrent and overvoltage.

On-board EMI filtering provides Class B compliance to FCC CFR Title 47, Part 15, Sub-Part B - conducted; and EN55022/CISPR 22 conducted. Regulatory agency certifications include UL recognition to UL60950/CSA 22.2 No. 60950-00 and TUV approval to EN60950-1.

The MPU200-1048 comes in a 203.2mm x 106.7mm x 38.1mm package that fits inside a 1U chassis. A DC-input version is also available in this same form factor. Prices start from $185 for 250-unit quantities and are typically available in six to ten weeks' time.

www.power-one.com

Swinging signal analyser

The SR785, now available in the UK from TTi (Thurlby Thandar Instruments), is a dual-channel dynamic signal analyzer, suitable for analysing both electrical and mechanical systems.
The SR785 uses a 32-bit floating-point DSP that delivers a 102.4kHz real-time bandwidth on both channels simultaneously. Bandwidth is not sacrificed for the number of channels utilised.
Two precision 16-bit analogue-to-digital convertors provide 90dB dynamic range in FFT mode and 145dB in swept-sine mode. With up to 800 lines of spectral resolution, the SR785 allows the user to zoom in on any portion of the 476mHz to 102.4kHz range.
TTi has developed a measurement architecture that allows each input channel to function as a separate analyser with its own span, centre frequency, resolution and averaging modes. This allows the user to view a wideband display and at the same time zoom in on specific spectra. This architecture also provides simultaneous storage of all measurements and averaging modes. Vector-averaged, RMS-averaged and unaveraged data are all available without the need to start the measurement again.
The SR785 costs £9680 (plus VAT).

www.tti-test.com

Sockets for large cables

A new PYF-14 series have been recently added to Omron's portfolio of relays and sockets. The sockets are suitable for use with Omron's MY relays and H3Y series timers and are available with two connection configurations for greater flexibility.
The PYF-14 ESN is a conventional design, whilst the PYF-14 ESS places the input terminals and output terminals on separate sides of the socket, allowing better safety and easier wiring. Both feature rising-up terminals for ease of connection and accommodate more cables and larger cable diameters. Further options include a metal spring clip facilitating secure relay installation or a plastic holding clip for quick component ejection.
The sockets provide both screw mounting and DIN-rail mounting facilities, featuring a footprint of 27mm x 82mm.
Conforming to all relevant international standards, the sockets are rated for currents up to 12A at 300V and offer an insulation voltage greater than 3kV. Suitable for use in the most demanding environments, the sockets are rated for operating temperatures between -400C and +850C.

www.europe.omron.com

NEC Electronics sings a new tune

Three new mobile-phone sound chips have been launched by NEC Electronics, which claims offer a "superior sound". The series consists of the microPD9993 IC that offers 64 polyphonic tones and is the first LSI device to support MP3 and advanced audio coding (AAC) playback; the microPD9996 for low-end monaural (single-speaker) mobile phones, and the smallest in class monaural microPD9995.

The microPD9993 has a DSP core for MP3 and AAC decoding. The device also uses surround sound from DiMAGIC's Adaptive Surround Technology, which processes two audio channels to produce five channels of stereo-wide surround sound. These features are expected to significantly enhance the sound quality in mobile phones, allowing users to listen through headphones to enjoy music stored in their phones at CD quality.

The microPD9996 chip is optimised for cost-effective solutions as it eliminates high-end functions such as surround sound and external digital input/output (I/O) whilst offering monaural sound.

The microPD9995 is the smallest in class with a package of 4.38mm x 4.38mm. Pin count has been reduced to 48.

www.ee.nec.de

B2Spice version goes up a notch

Version 5 of the B2Spice package is now available from RD Research. After a development of two years, the company has made extensive enhancements to the software's simulation capabilities, which now include a "scenario editor" that will allow users to sweep any parameter for any component. In addition, the user interface has been redesigned for easier and quicker design. There's a new parts browser, which allows users to navigate through a pop-up menu tree structure with great ease.

Amongst v5's feature rich enhancements is a "live circuit" feature that will allow users to modify components while a simulation is running and see the effects immediately. Parameter sweeping of any circuit/program/model/device parameter is available for every test, as is Monte Carlo analysis. A parts bin to store most frequently used parts is provided along with interactive "live" components such as switches, buttons, LEDs and others.

A new "circuit wizard" feature will auto-generate many circuit designs either as a new circuit or as a sub-circuit part that can plug into an existing circuit.

RD Research is offering it on a full 30 day evaluation basis.

www.spice-software.com

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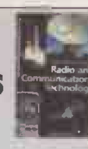
NEWNES GUIDE TO RADIO AND COMMUNICATIONS TECHNOLOGY

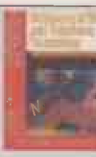


Ian Poole

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

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

DICTIONARY OF VIDEO AND TELEVISION TECHNOLOGY

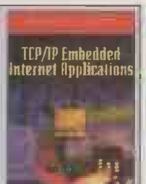


Jack Tsatsoulin

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

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

TCP/IP EMBEDDED INTERNET APPLICATIONS



Edward Insam

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

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

ELECTRONIC CLASSICS: COLLECTING, RESTORATION AND REPAIR

Andrew Emmerson

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

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

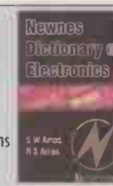
INTRODUCTION TO DIGITAL SYSTEMS

John Crisp

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

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

NEWNES DICTIONARY OF ELECTRONICS

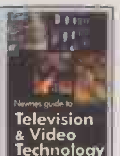


S W Amos; R S Amos

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

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

NEWNES GUIDE TO TELEVISION & VIDEO TECHNOLOGY



Eugene Trundle

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

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

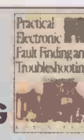
NEWNES GUIDE TO DIGITAL TV

Richard Brice

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

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

PRACTICAL ELECTRONIC FAULT FINDING AND TROUBLESHOOTING

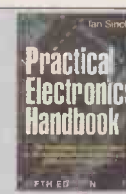


Robin Pain (Design Engineer, Cotag International Ltd)

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

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

PRACTICAL ELECTRONICS HANDBOOK



Ian Sinclair

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

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

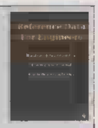
RSGB RADIO & ELECTRONICS COOKBOOK

Radio Society of Great Britain

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

Nov 2000 ▲ 336 pages ▲ PB ▲ Illustrations
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
HB ▲ Published in UK
Code 0-7506-7291-9 **£90.00**

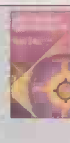
FUZZY LOGIC FOR EMBEDDED SYSTEMS APPLICATIONS

Dr Ahmad Ibrahim

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

Oct 2003 ▲ 312 pages ▲ PB
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
Code 0-7506-7339-7 **£19.99**

HANDBOOK OF RF AND WIRELESS TECHNOLOGIES

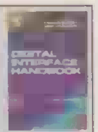
Farid Dowla (Lawrence Livermore National Laboratory, University of California-Davis, USA)

Subjects covered include third-generation wireless, wireless sensor networks, RF power amplifiers, spread spectrum modulation, signal propagation and antennaes. 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 231 line illustrations ▲ PB

Code 0-2405-1909-4 **£29.99**

TV FAULT-FINDING GUIDE

Edited by Peter Marlow

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

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

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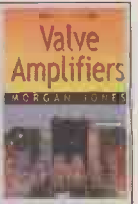
VALVE AMPLIFIERS



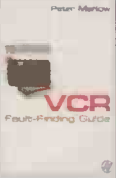
Morgan Jones

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

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

Code 0-7506-5694-8 **£29.99**

VCR FAULT-FINDING GUIDE



Edited by Peter Marlow

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

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

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DESIGNING AUTONOMOUS MOBILE ROBOTS

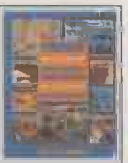
John Holland

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

Code 0-7506-7683-3 **£35.00**

VIDEO AND CAMCORDER SERVICING AND TECHNOLOGY



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

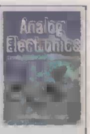
Keith Jack

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 Published in UK

Code 1-878707-56-6 **£50.00**

ANALOG ELECTRONICS: CIRCUITS, SYSTEMS AND SIGNAL PROCESSING



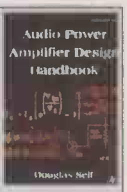
D Crecraft (formerly Open University); S Gergely (formerly University of Coventry)

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

Code 0-7506-5095-8 **£19.99**

AUDIO POWER AMPLIFIER DESIGN HANDBOOK



Douglas Self

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

3rd edition ▲ May 2002 ▲ 448 pages 62 line illustrations ▲ Index ▲ PB Published in UK

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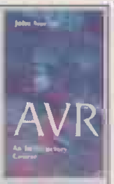

John Morton (University of Cambridge)

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 Glossary ▲ Further reading ▲ Index ▲ PB Published in UK

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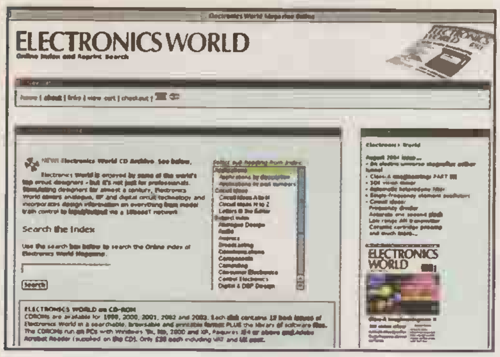

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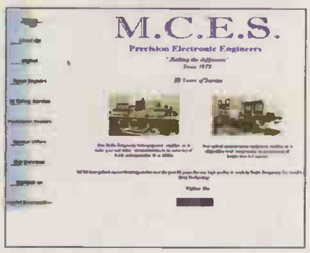

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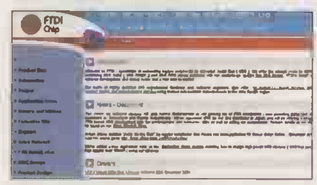

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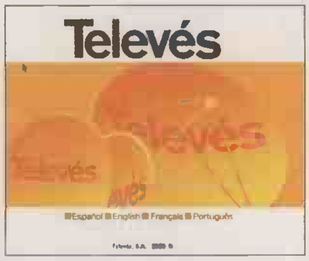

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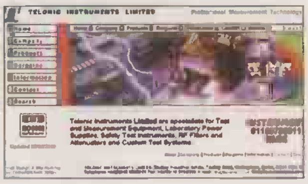

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AMPLIFIERS	Sale (GBP)	Rent (GBP)
AT/HP 8349A/001 20GHz 15dB 21dBm Amplifier	2250	96
AT/HP 8449B 26.5GHz 26dB +7dBm Pre-amplifier	3500	146
Amplifier Research 10W1000B 1GHz 10W RF Amplifier	2350	72
Amplifier Research 1W1000 1GHz 1W RF Amplifier	950	48
EG&G 5204 0.5Hz-100kHz Lock In Amplifier	1350	55
Kalmus KMS737LC 25W 10kHz-1GHz Amplifier	4500	136
DATACOMMS		
Fluke DSP4000 Cat 5e/6 LAN Cable Tester	3450	144
Fluke E3 P1 Handheld E3/ATM Network Analyser	1750	71
Microtest PENTA SCANNER+ Cat 5 Cable Tester	975	50
Tek 1502C/04 High Resolution Metallic TDR	3650	149
Tek 1503C Metallic TDR	2650	114
Wavetek LT8600 Cat 5e/6 LAN Cable Tester	2750	135
EMC		
AT/HP 11945A/E51 Close Field Probe Set With Preamp	3150	156
AT/HP 8542E 9kHz-2.9GHz EMI Receiver	23250	968
Schaffner NSG1025 Fast Transient/Burst Generator	1950	59
FREQUENCY COUNTERS		
AT/HP 53131A 225MHz 10 Digit Universal Counter	950	39
AT/HP 53132A 225MHz 12 Digit Frequency Counter	1300	62
AT/HP 5348A 26.5GHz Counter/Power Meter	4250	170
AT/HP 5350B 20GHz Frequency Counter	1675	68
AT/HP 5350B/010 20GHz Frequency Counter	1995	72
AT/HP 5370A 100MHz Universal Time Interval Counter	1250	50
AT/HP 5372A 500MHz Frequency/Time Interval Analyser	1995	82
Racal 1992 1.3GHz Frequency Counter	950	30
FUNCTION GENERATORS		
AT/HP 3314A 20MHz Function Generator	1250	50
AT/HP 3324A 21MHz Function Generator	1350	54
AT/HP 3325B 21MHz Function Generator	2050	62
AT/HP 3335A 81MHz Function Generator	1850	74
AT/HP 8111A 20MHz Function Generator	1150	46
AT/HP 8116A 50MHz Function Generator	1895	76
AT/HP 8165A 50MHz Function Generator	1350	54
AT/HP 8904A/001/002/003/004 600kHz Function Generator	2950	91
Tek AWG610 2.6 Gs/s Arbitary Waveform Generator	16950	698
LOGIC ANALYSERS		
AT/HP 1652B 80 Channel Logic Analyser	2150	108
AT/HP 1660A 500MHz Timing 100MHz State 136 Ch Log An	2950	89
AT/HP 1662A 500MHz Timing 100MHz State 68 Ch Log An	2550	77
AT/HP 1670G 500MHz Timing 150MHz State 136 Ch Log An	6550	328
AT/HP E2423A SCSI Bus Preprocessor	100	10
NETWORK ANALYSERS		
AT/HP 3575A Gain/Phase Meter	1350	56
AT/HP 3577A 5Hz-200MHz Vector Network Analyser	4750	190
AT/HP 41952A 500MHz Transmission/Reflection Test Set	1950	58
AT/HP 4195A 500MHz Vector Network/Spectrum Analyser	6950	209
AT/HP 8720B 20GHz Vector Network Analyser	20950	825
AT/HP 8752C 1.3GHz Network Analyser	7250	261
AT/HP 8753C 3GHz Vector Network Analyser	6950	285
AT/HP 8753D/1D5 3GHz Vector Network Ana c/w S Param	10250	308
AT/HP 8753E/010 3GHz Vector Network Ana c/w S Param	14000	504
AT/HP 8753ES/002/006/010 6GHz Vector Network Analyser	19500	585
AT/HP 89441A-Various option sets available - prices from	11950	486
Anritsu MS4624B 9GHz Vector Network Analyser	18450	743
Anritsu S331A 3.3GHz Sitemaster Scalar Network Analyser	3950	170
Anritsu S331B 3.3GHz Sitemaster Scalar Network Analyser	4950	213
Anritsu S400A/05 4GHz Sitemaster Scalar Network Analyser	5250	189

OSCILLOSCOPES	Sale (GBP)	Rent (GBP)
AT/HP 54111D 2 Channel 500MHz 2GS/s Digitising Scope	1975	80
AT/HP 54112D 4 Channel 100MHz 100MS/s Digitising Scope	2450	103
AT/HP 54502A 2 Channel 400MHz 400MS/s Digitising Scope	1450	58
AT/HP 54600B 2 Channel 100MHz 20MS/s Digitising Scope	1250	40
AT/HP 54602B 4 Channel 150MHz 20MS/s Digital Scope	1850	81
AT/HP 54622D 2 + 16 Ch 100MHz 200MS/s Digi Scope	2750	115
AT/HP 54650A HPIB Interface Module	155	16
AT/HP 54825A 4 Channel 500MHz 2GS/s Digitising Scope	6250	189
AT/HP 54845A 4 Channel 1.5GHz 8GS/s Infinium Scope	9950	399
Lecroy 9374L 4 Channel 1GHz 2GS/s Digitising Scope	4350	195
Lecroy 9384AL/WP1/2/FD/GP/DDM/PRML/MC1/4 4 Ch 1GHz	4750	199
Tek TDS210 2 Channel 60MHz 1GS/s Digitising Scope	600	30
Tek TDS320/14 2 Channel 100MHz 500MS/s Digitising Scope	1500	45
Tek TDS350 2 Channel 200MHz 1GS/s Digitising Scope	1950	80
Tek TDS420A 4 Channel 200MHz 100MS/s Digitising Scope	3600	155
Tek TDS724A/1M 2 Channel 500MHz 1GS/s Digitising Scope	3800	152
POWER METERS		
AT/HP 11722A 2.6GHz Power Sensor Module	1150	35
AT/HP 437B RF Power Meter	1250	55
AT/HP 438A Dual Channel RF Power Meter	1550	47
AT/HP 8481A 10MHz-18GHz 100mW Power Sensor	500	25
AT/HP 8481H 10MHz-18GHz 3W Power Sensor	650	33
Anritsu ML2437A RF Power Meter (Single Channel)	1395	60
Anritsu MA2442A 18GHz Power Sensor (High Accuracy)	695	29
Anritsu MA2444A 10MHz-40GHz Power Sensor	825	42
Anritsu MA2473A 10MHz-32GHz Power Sensor	1200	50
Anritsu ML2438A RF Power Meter (Dual Channel)	1995	80
Gigatronics 80401A 10MHz-18GHz 200mW Mod Pwr Sensor	800	40
Gigatronics 80601A 10MHz-18GHz 200mW Mod Pwr Sensor	950	50
Gigatronics 8542C Dual Channel Power Meter	2350	97
Gigatronics 8652A Dual Channel RF Power Meter	3750	150
POWER SUPPLIES		
Wide Range of AT/HP Programmable DC Supplies .....from	600	30
Keithley 248 5000V 5mA High Voltage Supply	1850	74
Kikusui PLZ-300W 300W Electronic Load	995	50
PULSE GENERATORS		
AT/HP 8011A 20MHz Pulse Generator	995	50
AT/HP 8012A 50MHz Pulse Generator	995	50
AT/HP 8130A 300MHz Pulse Generator	5650	230
AT/HP 8133A 33-3000MHz Pulse Generator	12750	537
AT/HP 8160A 50MHz Pulse Generator	2950	119
RF SWEEP GENERATORS		
AT/HP 8340B 10MHz-26.5GHz Synthesised Sweep Generator	9950	399
AT/HP 83640A/001/004/008 10MHz-40GHz Synth Sweeper	19950	790
AT/HP 83650B 50GHz Synthesised Sweeper	38650	1495
Anritsu 68177B/2C 10MHz-50GHz Synthesised Sweeper	19500	759
Gigatronics 8911 10MHz-20GHz Swept Signal Generator	4950	199

Check out our new Winter 2004/05 Product Guide !!
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SIGNAL & SPECTRUM ANALYSERS	Sale (GBP)	Rent (GBP)
Advantest R3261C 9kHz-2.6GHz Spectrum Analyser	4500	185
Advantest R3365A 100Hz-8GHz Spectrum Analyser c/w TG	9850	391
Advantest R3371A 100Hz-26.5GHz Spectrum Analyser c/w TG	13950	549
Advantest R4131B 3.5GHz Spectrum Analyser	3250	130
AT/HP 3561A 100kHz Dynamic Signal Analyser	2550	108
AT/HP 3585A 40MHz Spectrum Analyser	3500	106
AT/HP 53310A 200MHz Modulation Domain Analyser	2950	90
AT/HP 8560A/002 2.9GHz Spectrum Analyser	6250	250
AT/HP 8562A 22GHz Spectrum Analyser	10950	329
AT/HP 8563A/103/104/H09 22GHz Spectrum Analyser	7950	240
AT/HP 8563E 9kHz-26.5GHz Spectrum Analyser	17500	685
AT/HP 8591A/010/021 1.8GHz Spectrum Analyser With TG	3950	119
AT/HP 8594E 2.9GHz Spectrum Analyser	4650	149
AT/HP 8901B 1.3GHz Modulation Analyser	1950	82
AT/HP 8903B/001/013/051 20Hz-100kHz Audio Analyser	1850	56
Anritsu MS2663C/1/2/4/6 9kHz-8GHz Spectrum Analyser	8350	338
Anritsu MS2665C 21.2GHz Spectrum Analyser	11750	470
Anritsu MS2667C 9kHz-30GHz Spectrum Analyser	14500	596
Anritsu MS2711A 3GHz Handheld Spectrum Analyser	3600	144
IFR 2392 9kHz-2.9GHz Spectrum Analyser	4650	140
R&S FSP7 9kHz-7GHz Spectrum Analyser	11750	423
SIGNAL GENERATORS		
AT/HP 8642A 1GHz Synthesised Signal Generator	1950	78
AT/HP 8643A 1GHz Signal Generator	6975	279
AT/HP 8648A Synthesised Signal Generator	2500	100
AT/HP 8657B/001 2GHz Synthesised Signal Generator	2350	71
AT/HP E4421B 250kHz-3GHz Synthesised Signal Generator	5500	220
AT/HP E4432A/1E5 3GHz Synthesised Signal Generator	6950	278
AT/HP E4433A/1E5 250kHz-4GHz Synthesised Signal Gen	7950	239
Anritsu 68047C 10MHz-20GHz Synthesised Signal Generator	7950	318
Marconi 2031/002 2.7GHz Synthesised Signal Generator	4500	135
Marconi 2041/001 2.7GHz Low Noise Signal Generator	6950	278
TELECOMS		
AT/HP 37722A 2MBPS Digital Telecom Analyser	2650	106
Trend AURORA DUET Basic & Primary Rate ISDN Tester	1950	75
Trend AURORA DUET Basic Rate ISDN Tester	995	50
Trend AURORA PLUS Basic Rate ISDN Tester	350	28
TTC 147 2MBPS Digital Communications Analyser	3500	106
TTC Fireberd Interfaces - many in stock from	395	12
TTC Fireberd 6000A Communication Analyser	3650	110
TTC Fireberd 6000A/5 Communications Analyser	3950	119
TTC Fireberd PR-45 Printer For Fireberd 6000	350	15
TTC ISU 6000-4 Interface Switching Unit For 4 Modules	1650	85
TTC TIMS-45 TIMS Test Set For Fireberd 4000/6000	750	23
W&G PFA-35 2MB/s Digital Communications Analyser	3950	158
TV & VIDEO		
Minolta CA-100 CRT Colour Analyser	2000	80
Philips PM5515T/RGB TV Pattern Gen c/w Teletext + RGB	1750	55
R&S SFQ TV Test Transmitter Various Option sets avail from	17950	646
WIRELESS		
AT/HP 11759C RF Channel Simulator	4750	143
AT/HP 83220E/010 GSM/PCS/DCS1800 (1710-1900) Test Set	1950	59
AT/HP 8920B/1/4/7/13/14 1GHz Radio Comms Test Set	3950	119
AT/HP 8922M/001/006/010 1GHz GSM MS Test Set	3950	158
IFR 2935 GSM 900/1800/1900 Test Head	4950	198
IFR 2967 Radio Comms Test Set with GSM	5950	245
Marconi 2955B 1GHz Radio Comms Test Set	3500	126
Racal 6103/001/002/014 Digital Mobile Radio Test Set	3950	119

The only company with the confidence to offer 12 month warranty as standard, with 24 month available!

FREE
Palm Zire shipped with every order over £5K

Prices shown are for guidance in £UK GBP, exclusive of VAT and Ex-Works. All items subject to prior sale. Rental prices are per week for a rental period of 4 weeks. Free carriage to UK mainland addresses on sale items. Rental or non UK deliveries will be charged at cost. This is just a selection of equipment we have available - if you don't see what you want, please call. All items are supplied fully tested and refurbished. All manuals and accessories required for normal operation included. Certificate of Conformance supplied as standard. Certificate of Calibration available at additional cost. Test Equipment Solutions Ltd Terms and Conditions apply. All E&OE.

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