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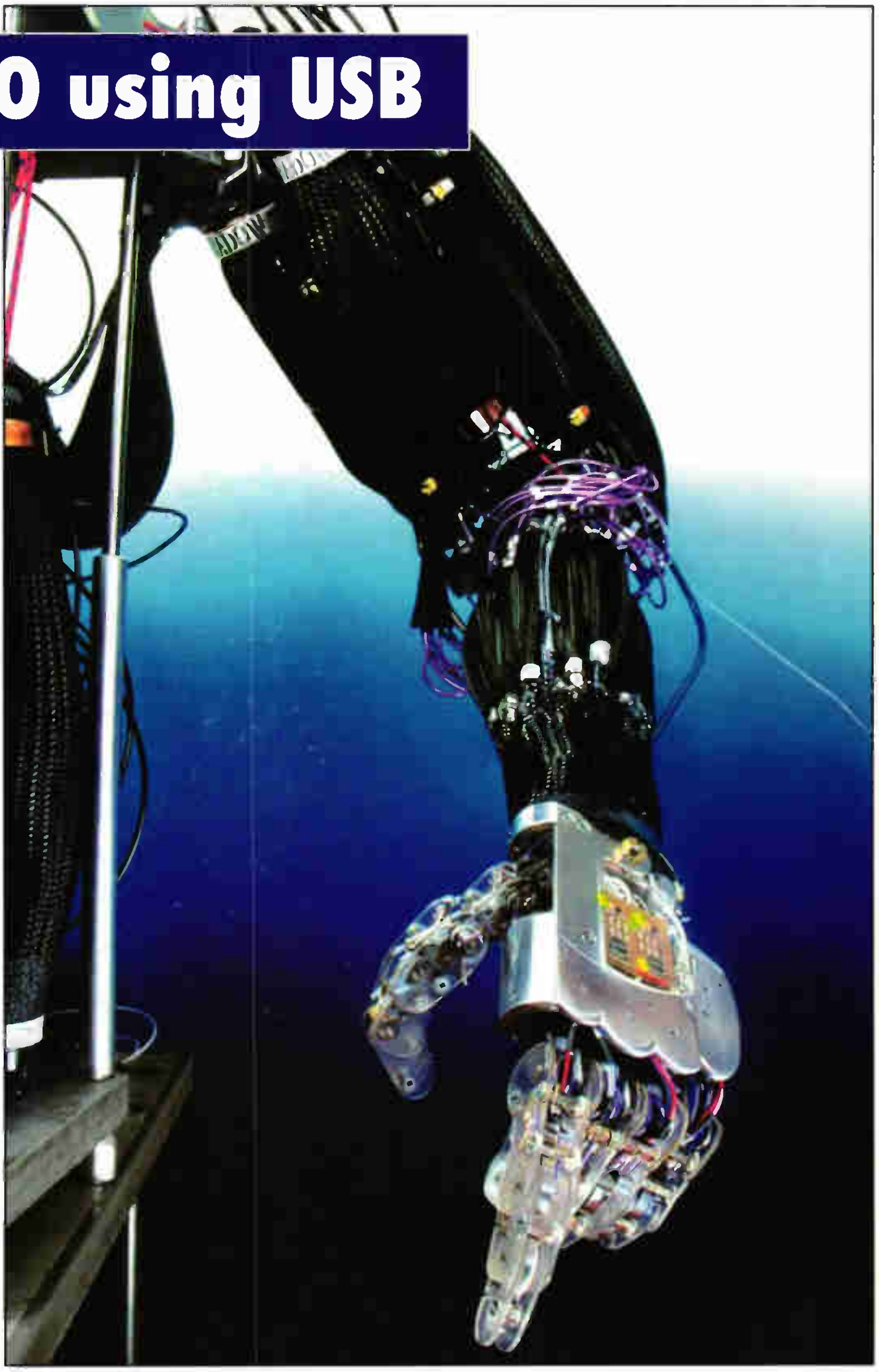
ELECTRONICS WORLD

NOVEMBER 2002 £2.95



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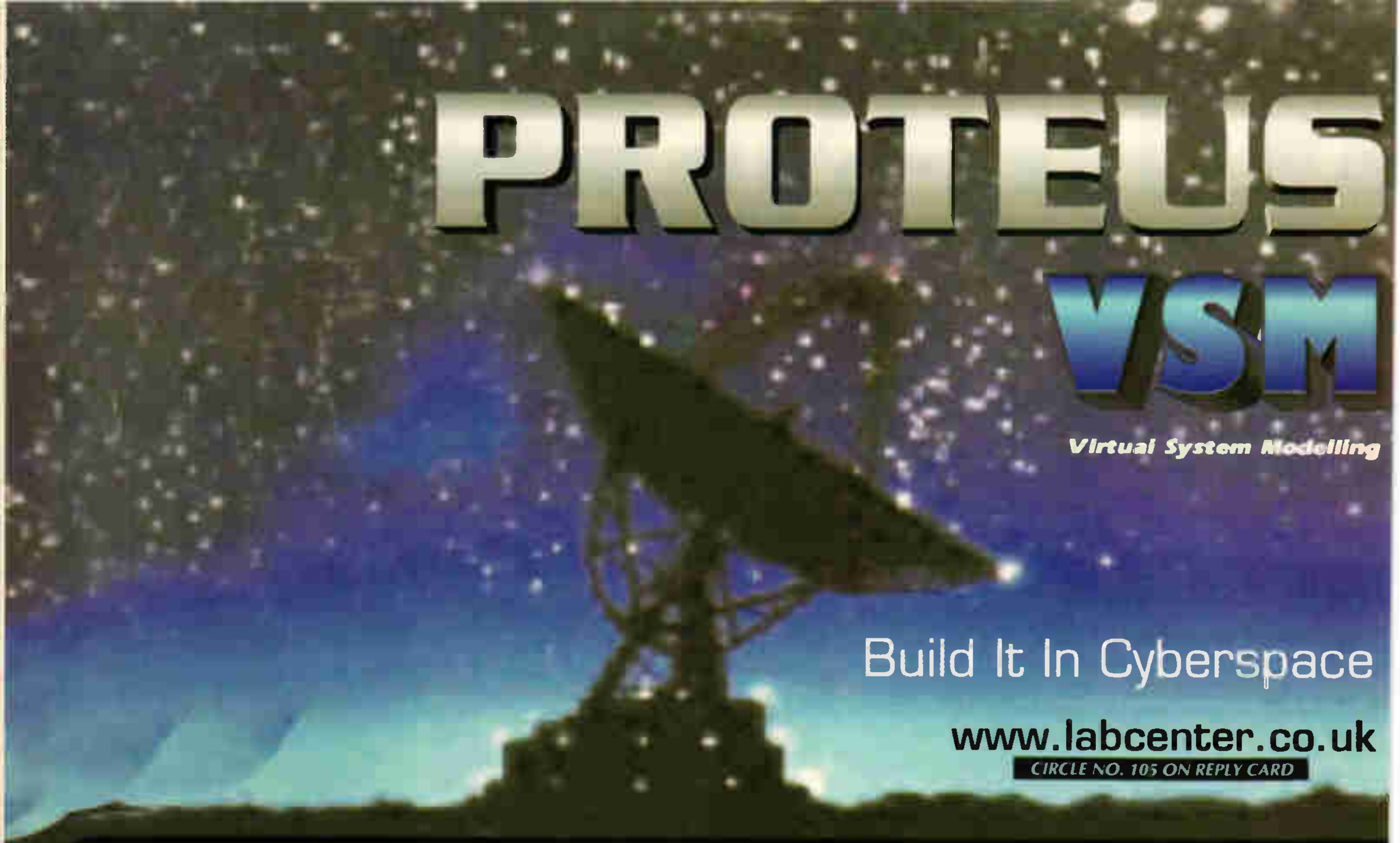
December issue on sale 7 November



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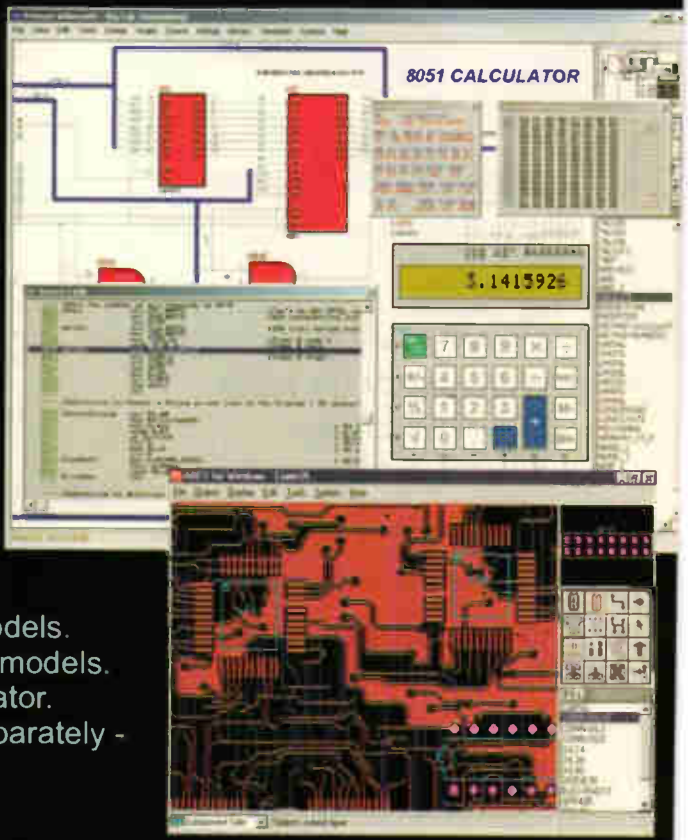

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ISSN 0959-8332
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Recession bites

I have to say that I think the recession and effects of the terrorist attacks on the US are making themselves felt in the electronics industry. Certainly, anybody who is in the entertainment electronics business will be having a hard time.

Why do I say this? I have literally just returned from the International Broadcast Convention in Amsterdam where the mood was one of 'let's hope it can't get any worse'. The lack of spending on the high street affects the companies making the direct sales of home electronics which in turn affects the media companies whose lives depend on manufacturers and suppliers advertising (which fuels TV and radio stations and even magazines like this one.)

The good thing about the exhibition was that the 'time wasters' did not show up and so the quality of visitor was very high. But only things that could save operating costs were being taken seriously at the show, as with all those channels to fill and a shrinking advertising pot – something's got to give.

I think that is true of all manufacturing at the moment, even in non-electronic circles. Ways are having to be found that enable cheaper (and possibly better) products, whether it be a TV programme or a process controller. Slicker manufacturing, lower waste and less labour costs are the vogue.

One of my annoyances at these international industry events (certainly in Europe) is that you never see any government representation. You don't get government ministers (and I'm not just talking about Britain) looking around or even opening a show. There is a creative TV show that happens in Edinburgh every year. For the last two years, the minister responsible for the broadcast industry has snubbed it with the most lame of excuses. What is it about governments that makes them think they can make decisions about industries without (it appears) knowing a whole lot about them?

The equivalent US broadcast show (NAB) at least attracts some government dignitary (I remember Ronal Reagan opening one year) and you see government agencies wandering around the show. Not that that means that good decisions are made. Witness the amazing mess the FCC (broadcast's governing body in the US) has made regarding the introduction of high definition television and digital broadcasting.

Despite my rants above, there was some good news. As I said, processes that save overheads were seriously being looked at, especially if the ROI (return on investment) was short. And some US friends reported that the telecoms industry was picking up again – a bit too late for Marconi – but I expect a lot of you that work in that part of electronics will be breathing a small sigh of relief.

My last rant this month is the imposition by Europe of an Electronic Data Protection Directive that basically forces businesses to retain electronic data for a longer period of time. ISPs, internet companies and telecoms firms will have to keep records of emails and web caches. At the moment it does not extend to the contents – although the UK government can gain access to this in special circumstances. All this is not going to be paid for by governments, but rather indirectly by you and me in our internet charges. This is seriously going to hamper the roll-out of broadband – which just may have been something to hike the electronics industry out of its doldrums. I, for one, do not want people to know what I watch or surf to – if only to keep the target advertisers (and hence lots of junk mail) away from my mail box. And I don't see why I should have to pay for my ISP to store loads of stuff that I don't want it to.

And the UK government wants its citizens to file their tax returns via the internet by 2010 or get fined £3,000.

Phil Reed, Editor.

Electronics World is published monthly. Orders, payments and general correspondence to **Jackie Lowe, Highbury Business Communications, Anne Boleyn House, 9-13 Ewell Road, Cheam, Surrey, SM3 8BZ.** **Newstrade**: Distributed by COMAG, Tavistock Road, West Drayton, Middlesex, UB7 7QE Tel 01895 444055. **Subscriptions**: Wyvern Subscription Services, Link House, 8 Bartholomew's Walk, Ely Cambridge, CB7 4ZD. Telephone 01353 654431. Please notify change of address. Subscription rates 1 year UK £38.00 2 years £61.00 3 years £76.00. Europe/Eu 1 year £54.00 2 years £86.00 3 years £108.00 ROW 1 year £63.00 2 years £101.00 3 years £126 **Overseas advertising agents**: France and Belgium: Pierre Mussard, 18-20 Place de la Madeleine, Paris 75008. United States of America: Ray Barnes, Reed Business Publishing Ltd, 475 Park Avenue South, 2nd Fl New York, NY 10016 Tel; (212) 679 8888 Fax; (212) 679 9455 **USA mailing agents**: Mercury Airfreight International Ltd Inc, 10(b) Englehard Ave, Avenel NJ 07001. Periodicals Postage Paid at Rahway NJ Postmaster. Send address changes to above. **Printed by** Polestar (Colchester) Ltd, **Filmsetting by** G&E 2000 A1 Parkway, Southgate Way, Orton Southgate, Peterborough, PE2 6YN

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Scientists explain metal superconductor

The discovery early last year that magnesium diboride (MgB_2) is a superconductor at 39° kelvin opened up the possibility for cheap superconductors. It also left many scientists puzzled as to the mechanism that allowed the material to superconduct at such a relatively high temperature.

However, researchers at the Lawrence Berkeley National Laboratory and the University of California at Berkeley claim to have figured out MgB_2's properties.

"Structurally, magnesium diboride is almost as simple as pencil lead, graphite," says Professor Steven Louie from Berkeley Labs. "It consists of hexagonal honeycombed planes of boron atoms separated by planes of magnesium atoms, with the magnesiums centred above and below the boron hexagons."

Louie and his team used a theory called Bardeen-Cooper-Schrieffer (BCS) to study the material, showing how electrons combine to form pairs, which leads to superconductivity. To form the pairs, electrons exchange a phonon.

With MgB_2 it seemed that different types of electron were involved in the pairing. Looking at the material's graphite-like planar structure, electrons in the boron plane of the MgB_2 form different pairs to electrons out of the plane of the material.

The differing pairs are nicknamed red and blue electrons. The difference is helping to explain some of the somewhat bizarre properties shown by MgB_2. Both types of pairs are destroyed if the temperature exceeds 39°K, and the material ceases to superconduct.

Bristol bins its batteries

Bristol City Council has begun a scheme to recycle batteries, removing potential toxic chemicals from household waste.

Although not the first recycling proposal for batteries in the UK, it is the first to collect batteries alongside household rubbish, rather than having battery banks scattered around the city.

The scheme will include rechargeable batteries which have outlived their usefulness.

Britain has been slower than other countries to catch on to the idea of separate collections for householders. The Netherlands, Germany, Switzerland and Japan all collect batteries separately from domestic refuse.

It has been calculated that something approaching a billion batteries are sold in the UK every year.

Graphics adds strength to ARM

Coming to a handheld computer or smartphone near you are 3D graphics such as these pictured.

Cambridge processor firm ARM spent the summer showing off the graphics capability it gained by licensing technology from Imagination Technologies.

ARM is using Imaginations's PowerVR MBX core, which uses a tile-based approach to rendering 3D scenes, making it suitable for varying sizes of display.

When combined with an ARM9 processor, the MBX core could easily render images onto screens ranging from 320x240 pixels up to full PC-type displays.

"Our objective with Imagination is to create a standardised graphics platform for a number of different markets," said Noel Hurley, manager for consumer entertainment at ARM. The firm expects the combined technologies to be designed into third generation mobile phones, in-car systems and even set-top boxes.

ARM has developed a range of MBX cores, each targeting different applications. A 3G phone processor would be able to render some one million triangles/s, while a set-top box version could handle 2.5 million, said Hurley. "It's a Sega Dreamcast type of performance," he said.

In order to download games to a mobile handset in reasonable time, ARM has teamed up woth Superscape, another UK firm. It has developed a way of drastically reducing file sizes for 3D gaming.

Images output direct from a test rig combining an ARM processor and 3D graphics core.

Fingerprints get embedded

A Lithuanian firm has developed a fingerprint sensing system aimed at embedded devices.

Neurotechnologija said its Fingercell design is aimed at providing controlled entry to doors, gates and computers. It can be used for both verification of a user or identification, the firm said.

The board is based around a 206MHz StrongARM processor, while the sensor itself is supplied by Authentec.

Software, which takes up some 512kbyte of space on the board, is supplied by the firm in ANSI C format.

"Developing a superior embedded biometric identification system is a long and expensive undertaking. Our FingerCell EDK greatly shortens the process," said Dr Algimantas Malickas, Neurotechnologija's chief executive.

The false rejection rate is claimed to be three per cent, while false acceptance rate is said to be 0.001 per cent. The system can learn and store a new print in under one second.

Evaluation software and a sample fingerprint database can be downloaded from the firm's site at www.neurotechnologija.com

Ben & Jerry's, the ice cream company, is to offset CO_2 emissions at its Vermont factory by supporting the construction of a wind turbine in South Dakota. Through an organisation called NativeEnergy's 'WindBuilders Business Partner' program, the machine will be erected on the Rosebud Sioux Tribe Wind Turbine Project - the first Native American owned and operated large-scale wind turbine in the USA. The turbine is expected to be working by November. www.nativeenergy.com

Linux runs on Xbox

A group of German programmers have managed to boot a version of the Linux operating system on Microsoft's Xbox gaming console.

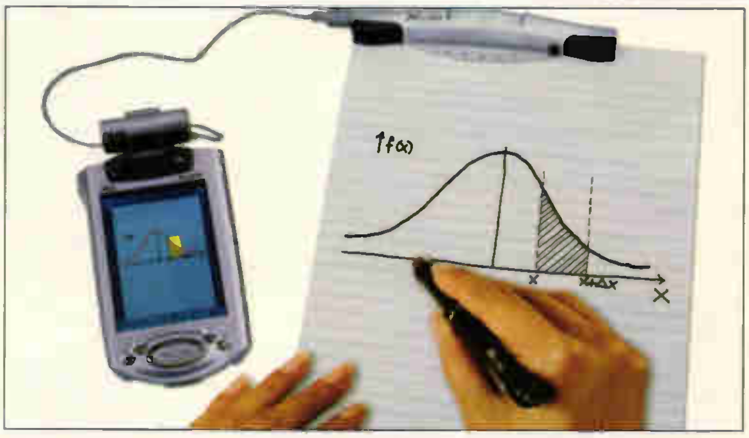
The addition of a $30 chip to allow the machine to run unsigned code, leaves Xbox ready for Linux, the team said. Linux can be booted from flash or, more likely, from a CD inserted in the console.

Since Xbox contains an Intel processor, hard disc and DVD drive, it is perfect for use as a very low cost PC, although not running any Microsoft applications.

Although the operating system is running, the team has yet to add drivers for the full video and audio systems. It plans to do this next, along with drivers for standard keyboard and mouse.

The Xbox could be used as a PC, web server or as a node in a cluster. Being Linux, the operating system can be downloaded and copied freely.

xbox-linux.sourceforge.net



If you think on paper, but work on computer, Seiko could have just the gadget for you.

InkLink, as it is called, is a pocket-sized device designed to transcribe handwritten notes and drawings to a PDAs or computer as they are written with a special pen.

Palm, Pocket PC and Windows are supported and data is transferred through infra-red or USB.

Proprietary software, InkNote Manager, allows users to cut, copy, paste, e-mail and store handwritten notes as .ink vector files, but also allows the files to be exported in .bmp, .png, and .jpg formats.

The unit is 193 x 74 x 143mm and weighs 143g. Price is £99.

www.seikosmart.com

Gentle spray cools better than air or water

As conventional heatsinking approaches its limits, two teams of US researchers are investigating a modified form of spray cooling for over-hot electronics.

While people have been blowing air at, spraying liquid over, and pumping water through hot assemblies for years, these latest efforts micro-spray the semiconductor dies directly.

At University of California, Los Angeles, Professor Elliott Brown found liquid spray-cooling could improve the performance of insulated gate bipolar transistors (IGBTs) in motor drives by as much as 34 per cent. "Significantly greater power than can be achieved with the same chips using conventional cooling," said Brown.

Over at HP Labs, Chandrakant Patel, principal scientist at HP's Thermo-Mechanical Architecture Lab, is cooling chips by spraying them from ink-jet-like nozzles.

According to Patel, existing chip power densities span 40 to $70W/cm^2$, but parts of PC processor chips will reach $200W/cm^2$ within four years. Conventional heatsinking cannot deal with this, he said, and liquid-cooling based on low-boiling point materials runs out at $100W/cm^2$ as bubbles form over hotspots.

Spray cooling is already known, but Patel asked: "Wouldn't it be nice if we could put just enough fluid on the chip so that it goes `psst' and vaporises without formation of any bubbles"?

And this is what he is doing.

Back at UCLA, Brown's research is with power transistors.

Above temperatures of 150°C, said UCLA, transistor life falls and at 200°C they cease to function.

As well as IGBTs, Brown looked at mosfets in 500MHz rf power amplifiers where spraying "was an order of magnitude more effective at removing heat".

In a 60W radio frequency power amplifier, spray-cooling disburses about 20W of heat, he said.

Whereas HP is using special insulating low-boiling point liquids, UCLA researchers found that water is the best at their temperatures. As the water goes straight on the die, its top surface is coated with a conformal dielectric.

The UCLA nozzle is a 4.86x1.53mm block of silicon with a 28x18 hole matrix "to exactly match the layout of the active cells", Brown said.

35µm jets are formed by reactive-ion etching to produce very smooth sidewalls.

Both cooling systems could be neatly packaged as closed systems with the chips they service.
www.hpl.hp.com/news/2002/apr-jun/cooling_demo.html

Cheap computer nears production

Simputer, the computer-for-India, is nearing limited production.

"We have made considerable progress in the development of a new version of the Simputer which has several interesting features that help in product expansion and connectivity," said Shashank Garg, v-p of product development at Encore Software in Bangalore.

Encore is one of the original contributors to the development of the Simputer and also one of the earliest commercial licensees of the Simputer platform.

The computer now has a CompactFlash socket through an optional docking cradle "that could be used for CF-II compliant cards such as a flash card or a Wireless LAN (Wi-Fi) interface. We also offer a choice of colour or monochrome LCD displays," said Garg.

Originally envisaged as a computer for people who cannot read, Simputer uses specially-written Information Mark-up Language (IML). "It could well be called an Illiterate Mark-up Language," said the Simputer Trust.

Standard Simputer evaluation kits

Model	Display	ram Mbyte	flash Mbyte
ESKIT-3216M	mono	32	16
ESKIT-3216C	colour	32	16
ESKIT-6432M	mono	64	32
ESKIT-6432C	colour	64	32

Customised devices will be available for large volume customers and "our current plan calls for launching 200 evaluation kits by the middle of August, and a subsequent launch of 1,000 Simputers, before large scale production starts", said Garg.

Orders for the kits are now being accepted with delivery within eight to ten weeks of receipt, claimed Garg.

www.ncoretech.com
www.simputer.org

Disc drive makers go for Terabytes

UK hard disc technology developer NanoMagnetics has taken another step forward in its work on nanoparticulate magnetic films, by storing 12Gbit of data per square inch.

The latest announcement doubles the density of nanoparticulate systems. "A doubling in areal recording density in just over six weeks highlights the strength of this team and the promise of this technology," said Dr Eric Mayes, the company's co-founder and chief technology officer.

The areal density record for this type of material does not match that achieved today by commercial hard discs, but the materials have fewer theoretical limits on density.

Nanoparticulates, in theory at least, should not be affected by the so-called 'superparamagnetic' limit that will soon be causing problems for conventional magnetic thin films.

The Bristol-based firm believes its technology will lead to discs that store 100 times more data, with drives able to store Terabytes of information.

Not to be outdone, hard drive manufacturer Seagate recently outlined its bid to escape the superparamagnetic limit. It will use heat assisted magnetic recording (HAMR) on future disc drives.

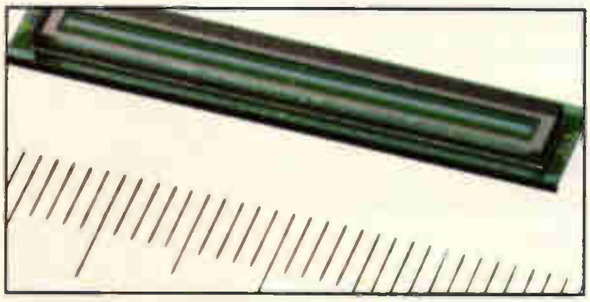
HAMR also uses a particulate structure, with ordered arrays of iron-platinum particles spread across the disc. It heats the area of the disc being written, which makes it more stable, and subsequent cooling is claimed to stabilise the data.

Seagate also expects to reach densities of around 50Gbit/inch^2 with this technique.

Tiny ribbons make light work of displays

Sony is to take forward the grating light valve (GLV) display device invented by California-based Silicon Light Machines, now owned by Cypress Semiconductor.

GLV is a micromachined technology that can achieve a contrast ratio of 3,000:1, claims Sony, making it ideal for projection display products including large-screen TVs.



So far the company does not have hard plans to make finished products.

"We don't have any specific product schedule applying grating light valve display device so far. Currently, Sony is seeking the possibility to introduce this display device as a device itself in next two years," said Sony spokesman Shinji Obana.

The active part of a GLV pixel is six shiny metal ribbons mounted side-by-side facing the same way.

In the relaxed state the ribbons form a highly reflective rectangular mirror.

In the active state, every second ribbon is attracted out of the mirror plane by up to half a wavelength. When fully activated, destructive interference occurs in the light and the pixel reflects hardly any light.

Partial operation, which can be achieved to any degree, produces grey-scales.

Sony's prototype device is 34 x 6 x 2mm, having 1,080 pixels (6,480 ribbons) in one long line. Red, green and blue lasers scan the grating to produce 1,920 x 1,080 high-definition colour images.

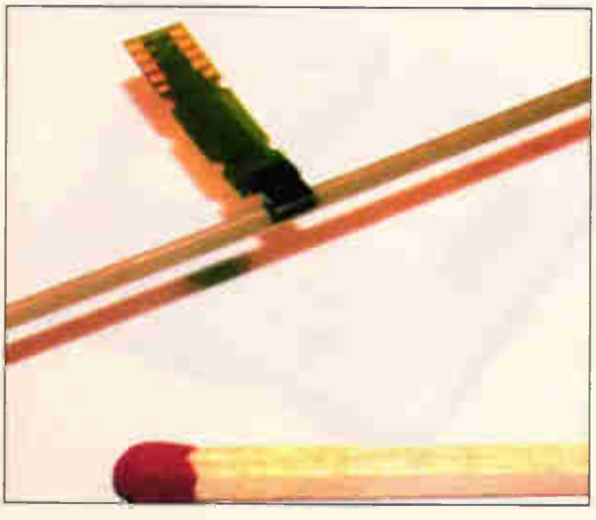
Sony has been working with Silicon Light Machines for a while, with the original aim of developing mass-produced display products.

Since then, in order to make sure the device succeeds, Sony has worked to develop micromachining technology to allow it to make the GLVs as well as developing other components including the laser sources.

Swiss go low for flow

A Swiss firm has developed a liquid flow meter capable of measuring flows as low as 150nanolitre/minute.

Sensirion's device uses the calorimetric principle to measure mass flow. A temperature sensor measures the calorific change from a constant temperature source.



The speed of the liquid through the pipe directly affects the calorific change between the temperature heads. Knowing the cross-sectional area of the pipe allows mass flow to be calculated from the flow rate.

The use of this technique, rather than impeller type systems, means the system can handle bi-directional flows.

The device is also non-invasive to the liquid flow. Sensirion said it can be used to measure viscous materials or those containing contaminants.

At the heart of the device is a CMOS chip which integrates the sensing elements, analogue to digital converter and basic signal processing functions.

Combining all the essential functions on one chip makes the sensor ten times smaller and around 25 times lighter than competing systems, the firm claimed. It is also said to be 100 times more responsive. Response time is down to 20ms, the firm claimed.

Sensirion has already started selling products based around the sensor, the first being the ASL1430. It has either a 10 to 500[micro]l/min range or a 5 to 1,000µl/min range. Accuracy is claimed to be better than 1.3 per cent.

Supply voltage is between 7 and 18V. The sensor will typically draw 20mA at 9V. The data output is RS232 compliant.

Evaluation kits comprising the sensor, PC software, cables and fittings is priced at €3,000.

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Powerful programmer for Atmel 8051 micro controller family. All fuse and lock bits are programmable. Connects to serial port. Can be used with ANY computer and operating system. 4 LEDs indicate programming status. Programs 89C1051, 89C2051, 89C4051, 89C51, 89LV51, 89C52, 89LV52, 89C55, 89LV55, 89S8252, 89LS8252, 89S53 & 89LS53 devices. NO special software needed – uses any terminal emulator program (built into Windows).

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3123KT	ATMEL 89xxx Programmer	£29.95
AS3123	Assembled 3123	£44.95

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Use a PC parrallel port as a real world interface. Unit can be connected to a mixture of analogue and digital inputs from pressure, temperature, movement, sound, light intensity, weight sensors, etc. (not supplied) to sensing switch and relay states. It can then process the input data and use the information to control up to 11 physical devices such as motors, sirens, other relays, servo motors & two-stepper motors.

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All components provided including a plastic case (140mm x 110mm x 35mm) with pre-punched and silk screened front/rear panels to give a professional and attractive finish (see photo). with screen printed front and rear panels supplied. Software utilities & programming examples supplied.

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ABC Mini 'Hotchip' Board

ABC Starter Pack

Currently learning about microcontrollers? Need to do something more than flash a LED or sound buzzer? The ABC Mini 'Hotchip' Board is based on Atmel's AVR 8535 RISC technology and will interest both the beginner and expert alike. Beginners will find that they can write and test a simple program, using the BASIC programming language, within an hour or two of connecting it up. Experts will like the power and flexibility of the Atmel microntroller. as well as the ease with which the little Hot Chip board can be "designed-in" to a project. The ABC Mini Board 'Starter Pack' includes just about everything you need to get up and experimenting right away. On the hardware side, there's a pre-assembled micro controller PC board with both parallel and serial cables for connection to your PC. Windows software included on CD-ROM features an Assembler, BASIC compiler and in-system programme. The pre-assembled boards only are also available separately.

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ABCMINISP	ABC MINI Starter Pack	£59.95
ABCMINIB	ABC MINI Board Only	£34.95

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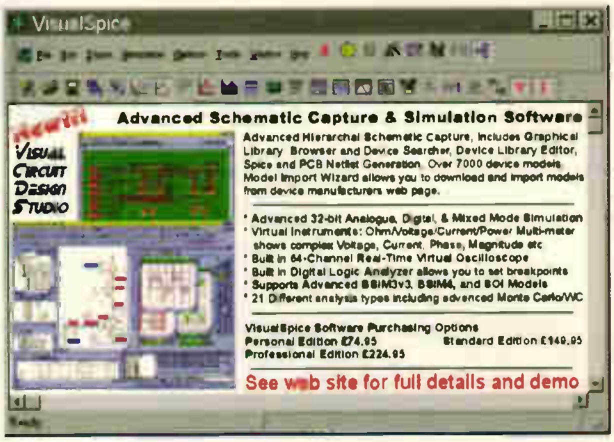

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Order Ref	Description	inc. VAT ea
3108KT	Serial Port Isolated I/O Controller Kit	£54.95
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Full details of these items and over 200 other projects can be found at www.QuasarElectronics.com

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Robot arm uses air muscles

North London robot R&D firm Shadow Robot Company has received £75,000 from lottery-funded NESTA (National Endowment for Science, Technology and the Arts) to develop a robot that can perform tasks for disabled people.

Shadow aims eventually to produce a multi-functional robot that can be guided, trained and programmed to carry out a wide variety of useful tasks for the disabled.

Shadow already has 'air muscles' - tubes which shorten as air is pumped in - which have been used in a fully functional analogue of the human hand.

"Systems based on the hand have the potential to make significant social contributions by enabling and empowering people through the creation of assistive devices," said Shadow.

The device the firm intends to develop will have a hand and arm on a wheeled base, with remote control and camera systems, so a user can guide the robot through simple tasks.

For example, the company envisions a user sending the robot to the bathroom to fetch a drink of water.

Initially the robot will be stupid but "over time, open-source software will be developed to perform more complex tasks automatically".

The Shadow Robot Company is a group of inventors working towards the long-term goal of producing a useful humanoid robot. MD Richard Greenhill, has been working in the robotics field since 1981 and in 1987 he set up the Shadow Robot Project where a group of enthusiasts would meet once a week to build a bipedal robot.

In 1997, the Shadow Robot Company was established to carry out robotic research and prototyping. Since then the company has received a number of contracts to build robots, including the BBC's Tomorrow's World and a Scandinavian toy manufacturer.

www.shadowrobot.co.uk

A soft touch makes sleek products.

Hampshire-based touch switch company Quantum Research Group has announced a six-channel device.

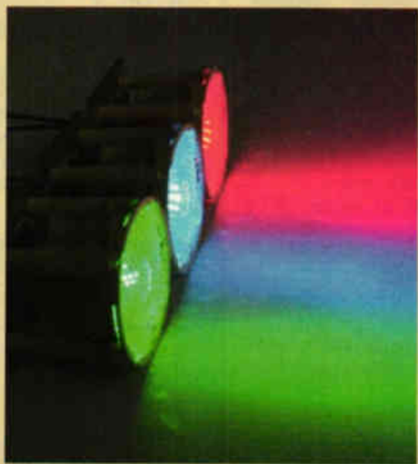
Called QT160, the chip includes a Risc processor which analyses signals to detect touch electrostatically through glass (up to 10cm thick), plastic, stone, ceramic, or even wood. It can also turn conductive objects into touch sensors.

"It not only allows manufacturers to create sleek touch controls, but also offers a practical way to create sealed, IP67 rated keypads," said the company.

The device requires only one capacitor per channel to function. Each of the 6 channels operates independently and sensitivity can be set per channel by changing the capacitor.

Adjacent key suppression blocks response from weaker responding keys and accepts only the dominant key - to solve the problem of large fingers on tightly spaced keys or water films over several keys.

The E160 evaluation board, pictured, demonstrates all these capabilities.



The Italian branch office of US optical company Fraen has developed a family of lenses suitable for Lumileds' Luxeon range of high-intensity LEDs.

The family currently has four lenses - Luxeon calls them 'collimators' - Narrow (6 to 10°), Middle (25 to 30°), Wide (40 to 45°) and Elliptical (10x20 to 12x25°. In each, the actual angle depends on the Luxeon model combined with the lens. Fraen claims they are "available for all the Lumileds Luxeon version: batwing 1W, Lambertian 1W and also for the new 5W Leds". The naked lenses may be combined with Fraen's universal lens holder for convenient mounting and alignment, said the company.

www.fraen.com

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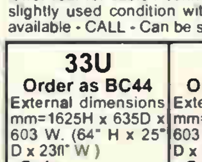

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Low-voltage low-power CMOS integrated architectures for sensor interfaces

Giuseppe Ferri and Nicola C. Guerrini propose some integrated interfaces for capacitive, resistive and temperature sensors designed with portability in mind.

In this work we shall propose some integrated interfaces for capacitive, resistive and temperature sensors. The topologies have been developed in CMOS technology that give the possibly of being included in portable systems where the sensor and the related electronics can be located. In applications like these, it is mandatory that circuits work at low supply voltages, with reduced power consumption and can therefore be supplied by a single-cell battery for all its lifetime (e.g., from 1.5V down to 1.2 V).

We have designed, as a 'general purpose' sensor front-end, a low-voltage low-power CMOS operational amplifier. In particular the amplifier shows low input noise and offset and the following main features: rail-to-rail input voltage range with constant g_m, full output voltage range, good DC voltage gain, high CMRR and PSRR. And special attention has been paid in the parasitic elements, which can affect sensor measurement.

The circuit interfaces have been developed considering the type of sensor and including where possible the cited amplifier. For capacitive sensors, in particular, the capacitive value is often converted to a frequency value through an oscillator configuration. Resistive sensors have interfaces mainly based on bridge circuits.

We propose a topology that recalls the Wheatstone bridge, but the branch resistances have been implemented by MOS transistors. With respect to the traditional passive bridge, this circuit improves sensitivity and resolution. An alternative topology using a phase shifter allows us to avoid the offset and calibration problems belonging to the bridge solution. Other resistive interfaces can be realised by means of the same operational amplifier used for capacitive sensor interfaces, where the sensing element is the resistance instead of the capacitance.

In all the proposed schemes, particular attention has been dedicated to the sensor constraint verification and in particular all the topologies have to show a very low dependence on supply discharges and temperature variations.

Finally, temperature sensors have also been considered. In alternative to traditional topologies, which sense the base-emitter differences in bipolar transistors, a low-voltage low-power bridge structure has been implemented, in CMOS technology, which gives an output voltage proportional to the temperature. A high linearity for a large temperature range variation is guaranteed and resolution is one order of magnitude lower than that of commercial digital thermometers.

Introduction

Modern microelectronic technology is capable of integrating a large amount of transistors on a single chip at reasonable costs. This enables the fabrication of highly complex electronic functions, amongst which are signal processing and data recording, etc. In this sense, it is necessary that the nature of the signal is electrical. Non-electrical information has to be converted through the use of sensors and actuators.

The integrated circuit technology has been continuously pushed towards the fabrication of devices operating with lower and lower operating supply voltages. The reason for

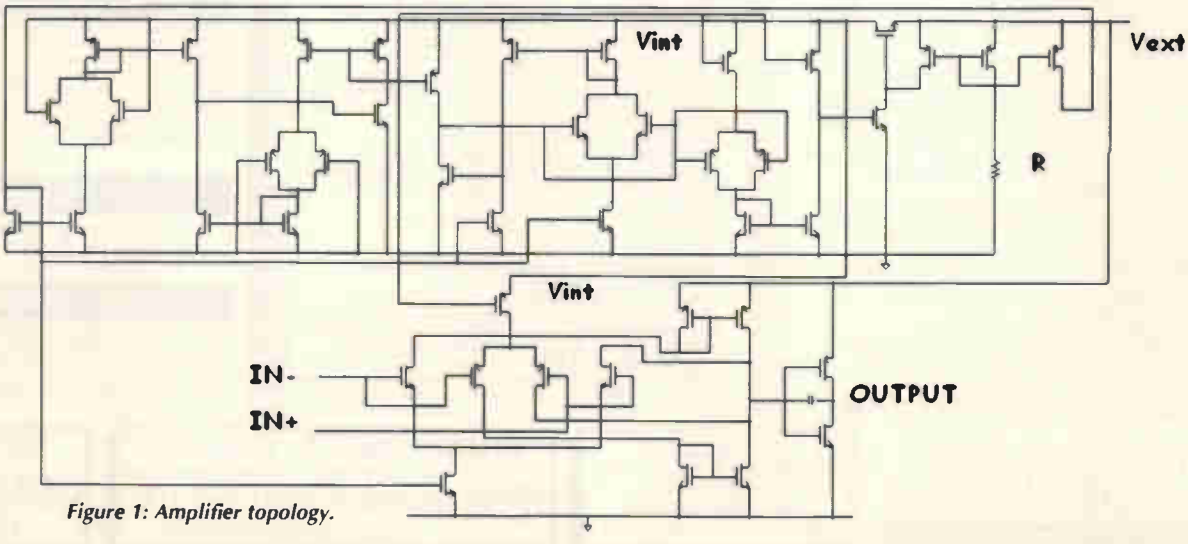


Figure 1: Amplifier topology.

this trend is mainly the increasing use of battery-operated portable electronics and wireless systems, which call for low power consumption, small size and low weight. Especially important is the need to reduce power dissipation in modern analogue and digital systems, which plays an important role in determining this trend. Also, a further push is due to the cultural scientific interest in exploring the technological and physical limits of the integrated devices.

Low-voltage, low-power structures find a natural but fundamental application in sensors and microsystems. In this field, there have not been any particular design constraints in the past, but actually new markets are moving towards the concept of portable and miniaturised products and the low-power solution is mandatory also for sensors and microsystems interfaces.

Hence, mixed analogue and digital electronics are becoming more and more important for sensors, because the chip-scale integration can be utilised for combining, on the same chip, existing standard IC processes, the sensing elements (if of silicon-type) and the processing electronics to fabricate 'smart sensors'. This is exalted by the fact that actually the same materials (silicon, polysilicon, aluminium and dielectrics) are used to fabricate the majority of sensors and electronic circuits.

In this way, CMOS has been proved to be the main sensor technology, because it is able to match the reduction of the technological costs with the design of new low-power interfaces.

Obviously, the first sensor interface has to be analogue, because of the analogue nature of the signal coming from the sensor. Moreover, analogue signal processing offers high functional density and an ability to interface directly the analogue real world of sensors. However, an A/D conversion of the output signal is also possible, so as to improve the quality of data display. In this case, owing to the sensor nature, no particular speed constraints are necessary. Traditional analogue-to-digital converters can be quite good for the purpose.

Sensors

Sensors are physical devices which transfer information from six different energy domains (chemical, optical, mechanical, thermal, magnetic and electrical) into an electrical one, providing a broad variety of electrical signals, which are normally analogue.

With respect to electronic circuits integrated on the same chip, sensors are normally divided into two groups: *active* sensors, which give an output current or voltage, and *passive* sensors, which modify their internal parameters if an external force is applied.

In the first case, resistive bridges or magnetically sensitive transistors can be interfaced to signal processing and conditioning circuitry such as low-noise voltage or current amplifiers. The basic parameters of the passive sensors (the capacitance and the resistance) can be either directly measured or integrated with the sensor and some controlling topologies such as oscillators, bridges, charge amplifiers and switched-capacitor circuits.

In **Table 1,** typical electrical outputs from sensors are shown.

Output signals coming from sensors have the following characteristics: low-level signals, relatively slow sensing parameter variation and the need of initial calibration for long-term drift (it means they generally can be time-variant). For this reason, in order to save the results from measuring errors, the design of low-noise low-offset signal amplifiers with low parasitic transistors is essential. Another important parameter to be considered is the electrical impedance of the sensor, which determines the frequency measurement range. The main sensor characteristics can be summarized in two parameters :

Table 1

type of signal	typical range	type of sensor
voltage	μV	thermopiles, pyro, piezo
current	μA-mA	pyro, magnetic
capacitance	fF-μF	humidity, gas, pressure
charge	pC	piezo
resistance	mΩ-MΩ	pressure, chemical (gas)

Sensitivity = electrical output variation / non-electrical parameter variation

Resolution = minimum detectable non-electrical parameter value in conditions of unitary signal-to-noise ratio.

Sensitivity has to be the highest possible and it has to be evaluated in the typical variation range of the non-electrical parameter. Possibly, it has to be linear and this means that its value does not depend on the working operating point. On the contrary, resolution has to be minimised and is definitively the most important sensor characteristic.

In the following sections, after the description of a low-voltage low-power integrated amplifier, to be used as a general purpose scheme for the sensor interface, examples of capacitive, resistive and temperature sensor interfaces working at low supply voltages with low power consumption will be presented, with the aim of helping the reader towards the concept of standard interface. Unfortunately, the literature suffers from the lack of universal interfaces to be utilised in a broad variety of sensing systems and also commercial integrated circuits are either for very specific applications or very complicated with a heavy digital elaboration part.

Low voltage low power 'general purpose' amplifier

The proposed amplifier[1] can operate with a 1.5 V supply voltage (the minimum being 1.2 V) and the circuit architecture has been designed to obtain low offset, low input noise, rail-to-rail input and output common mode range and low power consumption. The amplifier topology (whose simplified schematic is in **Fig. 1**) has been also optimised for the lowest supply voltage compatible with the technology used (Mietec 0.7µ with threshold voltages of about 0.75V). The circuit has also been redesigned at a 0.5µ technology, and power dissipation has been reduced at less than 200µW.

The input and output full swing characteristic has been obtained in order to have a high signal-to-noise ratio when the amplifier is used in a follower configuration or in front-end circuits, where the signal comes directly from the sensor. This means that the range of the sensor response is not absolutely limited by the amplifier.

The amplifier topology differs from other literature solutions (working at higher supply voltages) for its constant-g_m input stage that ensures a constant gain-bandwidth product over the input common mode voltage and, consequently, simpler frequency compensation. In **Table 2** its main experimental characteristics are shown.

Particular attention has been paid in the design of the amplifier input stage, especially concerning its noise and offset and the input transistors have been designed to operate in weak inversion condition (WI).

In order to have low input noise, a study about the input

Table 2

Amplifier characteristics	Measured values
Input stage swing	rail-to-rail (from V_{ss} to V_{int} + 0.5V)
Output stage swing	rail-to-rail
Equivalent input voltage noise	10 nV/√Hz (@ 1kHz)
Input offset voltage	typ.= 0.2mV ; 3σ value = ± 0.08mV
Input transconductance	constant ($\Delta_{gm\,MAX}$=6 %)
Gain Band Width (GBW)	1.3MHz (PM = 64°)
Low frequency gain	84dB
Power Consumption	0.46mW
Slew Rate	1V/µs
Distortion THD (1Hz,Vpp=60%V_{AL})	–40dB
CMRR	56dB @ 10Hz ; 52dB @100kHz
PSRR+	48dB @ 10Hz ; 26dB @ 100kHz
PSRR–	51dB @ 10Hz ; 32dB @ 100kHz
Chip area	1.2mm²

transistor condition has been done. In particular, the two noise contributions have been considered: thermal (inversely proportional to the input transconductance g_m) and flicker (inversely proportional to MOS sizes and frequency, and consequently dominant at low frequencies):

$$\overline{dv_n^2(f)} = \frac{8kT}{3gm}df + \frac{KF_F}{WLC_{ox}^2}\frac{df}{f}$$

where k is the Boltzmann constant, T the absolute temperature (in K), gm the input transistor.s transconductance, KF_F the Spice flicker noise parameter, C_{ox} the oxide capacitance, and W and L the width and length of the input transistors.

The minimisation of the noise depends on the input transistor condition. In strong inversion (SI) there is a particular relation between W and L. In the case of WI, the previous function is typically minimised using the highest values of W. Unfortunately high values of W give high values of related parasitic capacitances. Of course, the design, in terms of noise minimisation, can be also optimised at the sensor working frequency.

Concerning the offset, an auto zero technique[2] which measures the systematic offset and subtracts it, has been considered. The basic idea of auto zero is the sampling of the unwanted quantity (noise and offset) which is then subtracted from the instantaneous value of the contaminated signal either at the input or the output of the op-amp. This cancellation can also be done at some intermediate node between the input and the output of the op-amp, using an additional input port defined as the nulling input and identified with the letter N in the schematic of **Fig. 2**.

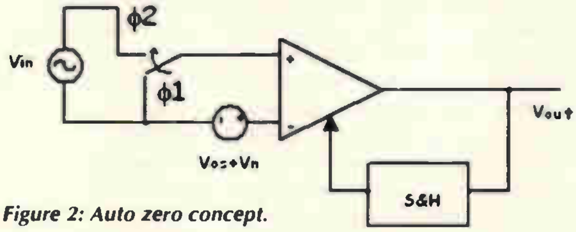


Figure 2: Auto zero concept.

If the noise is constant over time (like a DC offset) it will be cancelled, as is needed in a high precision amplifier or high resolution comparator. If the unwanted disturbance is a low frequency random noise (for example, l/f noise), it will be high-pass filtered and thus strongly reduced at low frequencies, but at the cost of an increased noise floor due to aliasing of the wideband noise inherent to the sampling process.

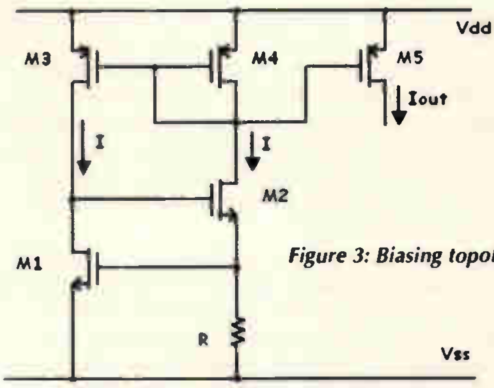


Figure 3: Biasing topology.

The auto zero process requires at least two phases: a sampling phase (ϕ_1) during which the offset voltage V_{OS} and the noise voltage V_N are sampled and stored, and a signal processing phase (ϕ_2) during which the offset free stage is available for operation.

The principle can be used not only to cancel the amplifier offset but also to reduce its low frequency noise It should be noted that the effect of the auto zero is equivalent to subtracting from a time varying noise a recent sample of the same noise. For DC or very low frequency noise this results in a cancellation, so auto zero effectively high-pass filters the noise.

Particular attention has been paid to the layout design (a common centroide scheme) that helps to reduce the random offset coming from the technological spread of integrated transistors.

Finally, another important consideration has to be the biasing of the amplifier and of the all the circuits to be used in sensor interfaces. It is important to implement a biasing current independent of the supply voltage variations[3], so as to avoid performance reduction when the supply battery discharges. **Fig. 3** shows this topology. It is easy to prove that:

$$IR = V_{GS1} = V_{t1} + \sqrt{\frac{2I}{\mu_n C_{ox}\left(\frac{W}{L}\right)_1}}$$

from which, if the transistor is working in WI, we can write :

$$I = \frac{V_t}{R}$$

The current I is independent of the supply voltage and depends very slightly on the temperature. It is mirrored into Iout (see Figure 3) with a unity gain mirror factor.

With commercial products, they typically have no supply voltage operation and so cannot be used in portable applications. For example, the National Semiconductor LM9044, a 'Sensor Interface Amplifier' is fabricated in bipolar technology. The amplifier is absolutely not low-voltage, because the supply voltage is ±60V and the power consumption is 1.3W. Dallas-Maxim Semiconductors seem to be the most suitable among the commercial companies. In particular the MAXIM1452 is a 'sensor signal conditioner', formed by a processor with DAC and operational amplifier incorporating a temperature sensor. The supply voltage is 5V and power consumption is 10mW. Another interesting product is a current-sense amplifier for portable applications, namely the MAXIM4172, whose supply is 3V and power dissipation is 2.4mW. But this product is not specific for sensors. If you want more information, you can consult the web site: http://www.maxim-ic.com.

Low voltage low power capacitive sensor interfaces.

Capacitive sensors have been proved to be good transducers for signal conditioning electronics with reduced current consumption. In fact, they have a high impedance up to reasonably high measurement frequencies and high signal levels.

The capacitive sensors are often interfaced with read-out electronics that performs a capacitance-to-frequency conversion (as with oscillators and phase shifters). The topologies have to respect the following constraints: high dynamic range, good linearity and precision, low input noise and offset, long-term temperature stability, reduced area, low effect of parasitic capacitances and calibration and compensation of the transducer characteristics. These constraints have to be satisfied by interface circuits which, if designed with low-voltage low-power techniques, can be utilised in portable, remote and wireless systems for industrial, biomedical, automotive and consumer applications, where a great need of reliable and miniature sensor systems emerges.

The first two examples here presented will consider the application of the amplifier described in the previous section as a read signal amplifier in capacitive sensors, whose use is rather spread in many sensing systems.

The proposed interface architectures (an oscillator and a phase shifter) have been implemented with low-voltage and low-power constraints. The frequency of the signal at the output of the amplifier allows it to determine the value of the sensor, this frequency being inversely proportional to the capacitance. An automatic storage of the oscillation frequency has also been performed, using a frequency counter controlled by a PC via a GPIB (General Purpose Interface Bus) gate.

Figure. 4 shows the topology of a traditional Schmitt trigger, utilized in an oscillator configuration, for the determination of the capacitance values[4].

These values can be determined by reading the frequency of the oscillator, according to the following relation (if $R_1=R_2$): f_{osc} @ 1 / (2.2R C_{sen}). The circuit has been tested for three decades of capacitance variations (from 1nF to 1μF), which it translates to a frequency span of three decades (from about 20Hz to 20kHz). The frequency values have been stored and elaborated in a personal computer by a GPIB interface. The sensitivity is about 100fF/Hz. Precise values of resistance (in particular of R) have been utilised and non-linear effects (among which the temperature) have been verified to be negligible in the frequency determination.

Figure.5 shows another topology developed to determine the capacitance values[4]. The main block is the traditional phase shifter, realised with the described

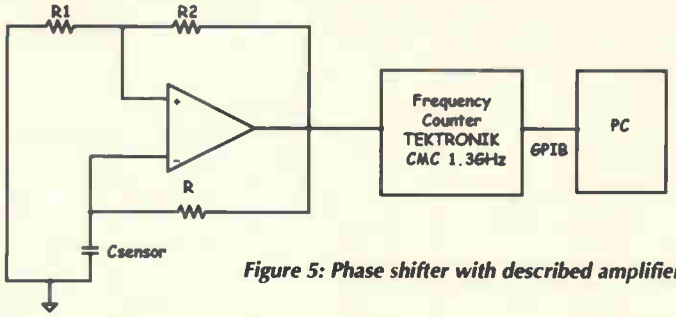


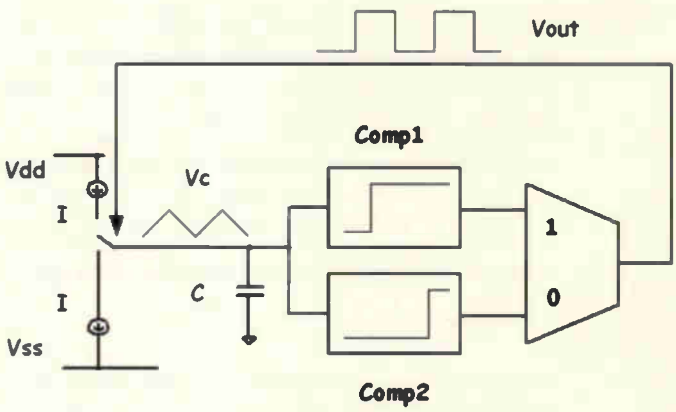
Figure 5: Phase shifter with described amplifier.

amplifier. The input signal is a sinusoidal voltage whose frequency has been set to the initial calibrated value. At the frequency of $f_0=1/(2\pi RC_{sen})$, the output of the amplifier is shifted –90 degrees with respect to the input. Then, input and output curves have been sent to two Schmitt triggers, in order to obtain square-waved voltages. Hence, these signals are directed to a traditional EX-OR gate, supplied at low voltages, and used to automatically detect the frequency f_0. In fact, when the EX-OR gives an output signal at a frequency equal to $2f_0$ with a duty cycle of 50%, the inputs of the EX-OR are in the –90° shifting condition. In this manner, reading f_0, we know the capacitance sensor value. The values of the frequencies at the output of the amplifier and of the EX-OR have been stored in a PC by a GPIB.

The main problem related to the proposed topologies concerns the detection of small capacitance values. In this case, the key aspect of the problem is related to the sensing system, where the sensitivity to parasitic, interconnection wires and noise has to be the lowest. Other applications of the OTA can be easily realized in other sensing systems, as in resistive sensors.

A third capacitive sensor topology has been considered, formed by an oscillator[5]. This topology does not utilise the

Figure 6: Oscillator block diagram.



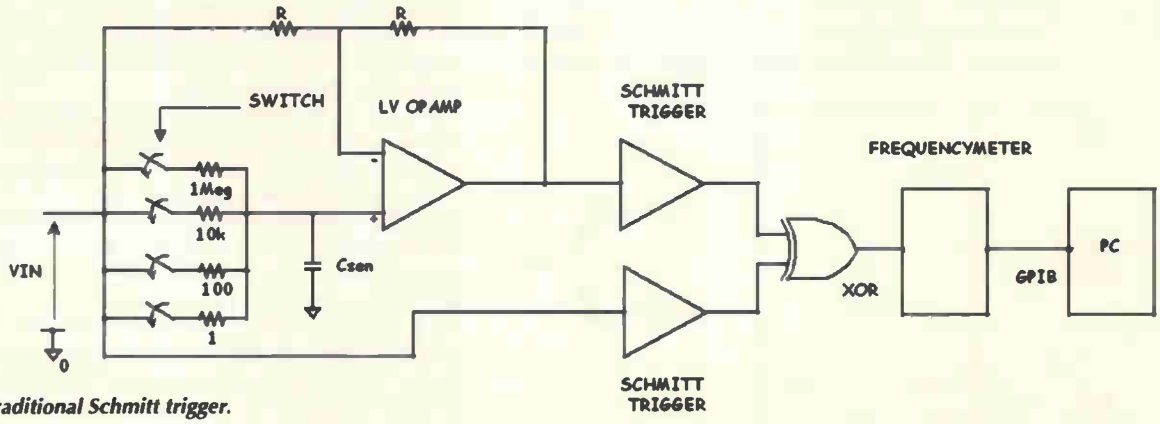


Figure 4: Traditional Schmitt trigger.

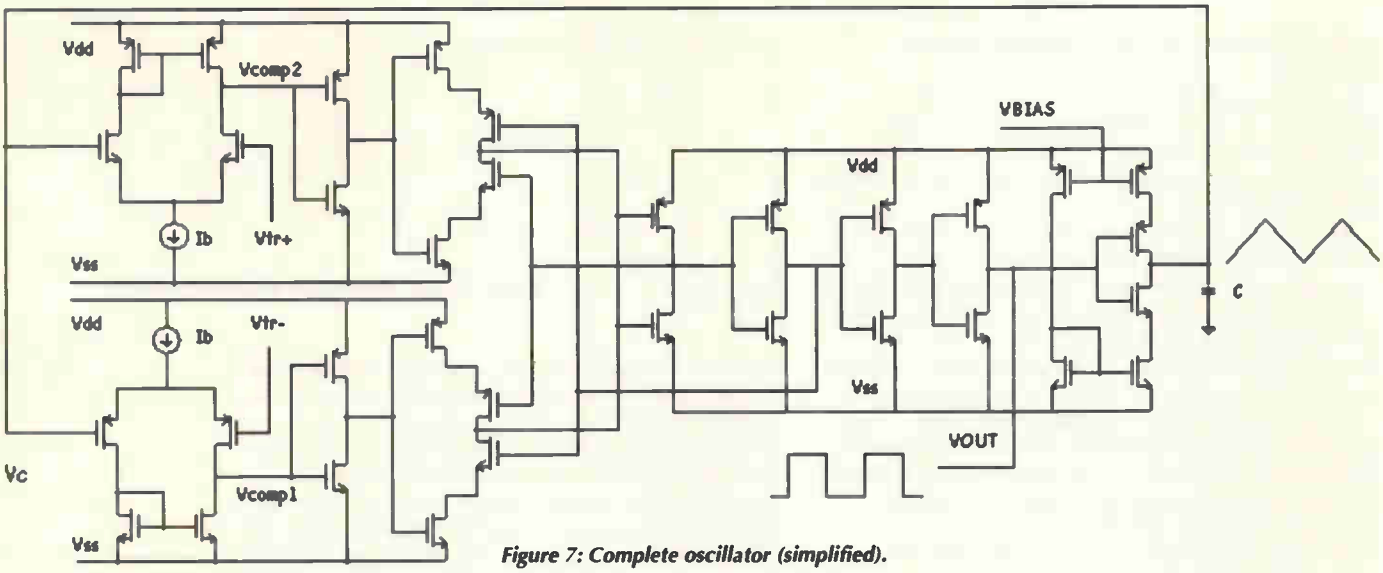


Figure 7: Complete oscillator (simplified).

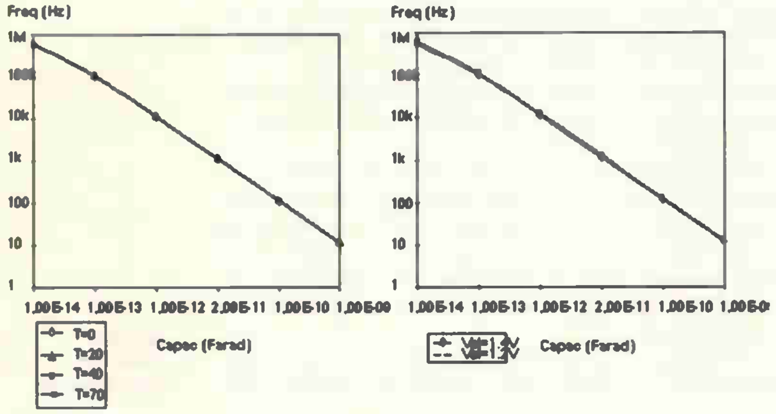


Figure 8: Frequency vs sensor capacitance.

amplifier to overcome the amplifier's imperfections. In order to understand the working principle, let us consider **Fig.6** that shows the oscillator block scheme. The sensor capacitance C is charged and discharged with the constant currents I. The output voltage V_{out} drives the switch. The hysteresis comparator has been designed with two traditional comparators having different threshold voltages V_{tr+} and V_{tr-}. The aim of the comparator is to convert the triangular voltage V_C on the capacitor in a squared-wave, whose duty cycle is set by the choice of the current sources values. The output frequency is the following:

$$f = \frac{I}{2(V_{tr+} - V_{tr\pm})}\frac{1}{C}$$

where the current I has been set to 20nA (by the use of suitable current mirrors) and the quantity in brackets is about 0.9 V, with the utilised technology. This solution can be particularly attractive because it gives an output frequency independent of resistive values. The outputs of each comparator are sent to a two-input multiplexer with the output fed back to the strobe terminal. By doing so, it can be easily shown that for V_C decreasing, the output of the multiplexer is determined by comp1, whereas for V_C increasing, it is determined by comp2.

The simplified schematic of the complete oscillator is in **Fig. 7**. Comparators are formed by simple differential structures. The p-type pair (lower), driven by V_{tr-}, is related to V_{comp1}, while the n-pair (upper), driven by V_{tr+}, generates V_{comp2}. The bias voltage controls the current sources, which charges and discharges C. Indeed, if V_{out} is high, the capacitance is discharging; on the contrary, if V_{out} is low, C is charging.

Figure. 8 shows the output frequency vs. sensor capacitance, at different temperatures (left) and supply voltages (1.4V and 1.2V, right). The circuit shows a very low power consumption (23μW) and a reduced dependence on supply voltage and temperature (0.06%/mV and 0.12%/°C, respectively). The flicker noise contribution has been reduced by the use of high area input transistors.

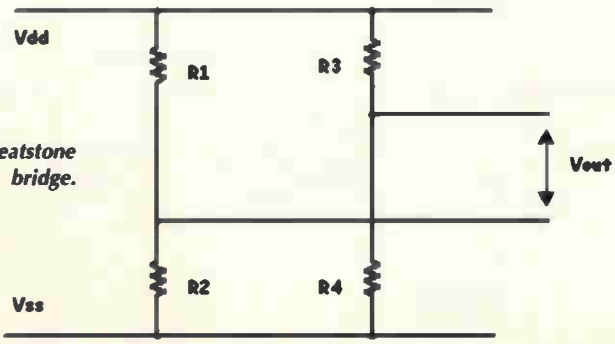


Figure 9: Wheatstone bridge.

Low voltage low power resistive sensor interfaces

Resistance variations are often measured interfacing the sensor element in the Wheatstone bridge configuration **Fig. 9,** where the sensor is one of the four branches of the bridge[6]. If the relative variation of the sensor resistance is less than 5%, an almost linear relation between the voltage output and the sensor variation exists. If V_{cc} is the total supply voltage, in this basic Wheatstone bridge, the sensitivity, defined as the ratio between the differential output voltage and the relative variation of the sensor resistance, is equal to $V_{cc}/4$. It has to be noted that this sensitivity can be increase if a differential stage is cascaded to the bridge output.

A novel circuit topology is presented here, based on the conversion of the passive resistances into active resistances, thus obtaining an improved performance.

The circuit here proposed makes use of CMOS elements in parallel to the resistances in each branch of the bridge. The conceptual scheme is shown in **Fig. 10.** The circuit is symmetrical in order to achieve a high CMRR performance and at the two outputs a common mode feedback circuit (CMFB) has been included to fix the output voltage at the half of the total supply, guaranteeing the maximum output dynamic range. The circuit has been designed to work with a low voltage supply (1.2V) and it also has low power consumption. As an example, the performances of the CMOS bridge have been evaluated in two cases considering the bridge working with sensors of 10KΩ and 1MΩ respectively. **Table 3** summarises the main circuit performances. In order to evaluate the advantages of the proposed topology, the sensitivity and the resolution have been evaluated and compared with those exhibited by the corresponding passive Wheatstone bridge. The sensitivity is 40 times better in the case of a sensor of 1MΩ and 120 times greater in the case of the sensor of 10kΩ, while the resolution has been improved by almost two orders of magnitude.

Another resistive sensor interface is shown in **Fig. 11**. Based on a phase shifter[7], two equal source currents loads and unloads the capacitance, C. The output inverter, formed by transistors MPout and MNout, ensures a square wave at the output. If these transistors have the same β, the output delay T_d will we given by:

$$T_d = \frac{V_{sup} C}{2I}$$

being V_{sup} the total supply voltage. The delay time corresponds both to the capacitive loading time (from the negative supply to $V_{sup}/2$) and to the unloading time (from the positive supply to $V_{sup}/2$).

The proposed circuit performs a maximum delay time of a quarter of the input period (that means a 90° shift), achievable only with ideal current sources. However, the technological spread and the temperature dependence make it impossible to obtain a symmetrical output voltage even if the condition on equal β of transistors Mnout and Mpout is verified. For this reason and with the aim to increase the output shift, two equal phase shifters of Fig. 10 type can be cascaded. Calling with V_{th} is the inverter transition threshold voltage, the first shifter gives a delay of

$$\frac{C}{I}(V_{th} - V_{SS})$$

while the second shifter delay is

$$\frac{C}{I}(V_{DD} - V_{th})$$

This allows a total delay independent on V_{th} whose value is exactly the double of the previous delay. As a consequence, its maximum theoretical value is 180°.

In **Fig.12** we propose a circuit that allows the control of the delay time (and, consequently, of the shift) by the use of a trimmed resistance. Its delay time is proportional to the resistance value, as follows:

$$T_d = \frac{(R1 + R2)C}{R2} Rsensor$$

The measure of the delay time is directly proportional to the value of the sensor resistor. It is also

Table 3

	Sensor Resistance= 10kΩ	Sensor Resistance=1MΩ
Supply voltage	1.2V	1.2V
Power dissipation	254μW	113μW
Sensitivity	1.2mV/Ω	11.8μV/Ω
(corresponding passive bridge)	(0.01mV/Ω)	(0.3μV/Ω)
Resolution	5.45 10^{-5} Ω/(Hz)$^{1/2}$	2.6 10^{-3} Ω/(Hz)$^{1/2}$
(corresponding passive bridge)	(4 10^{-3} Ω/(Hz)$^{1/2}$)	(4.2 10^{-1} Ω/(Hz)$^{1/2}$)

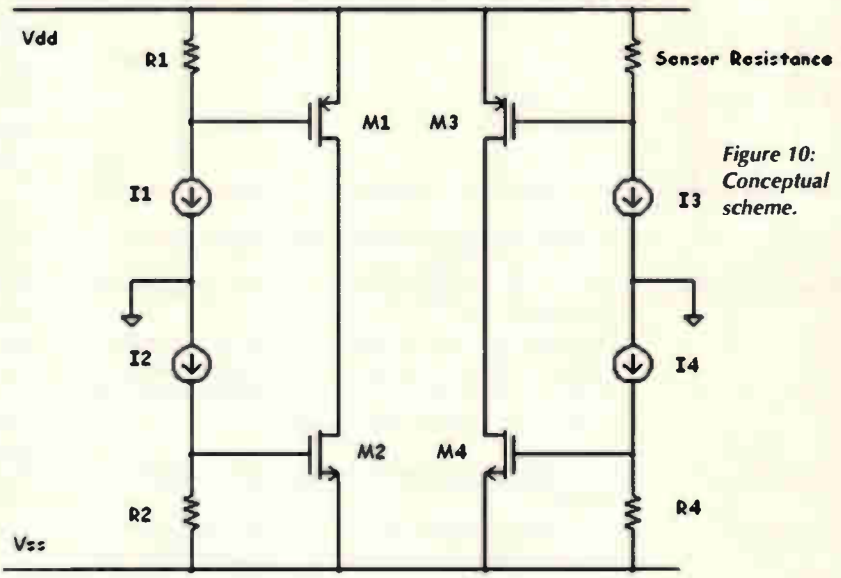


Figure 10: Conceptual scheme.

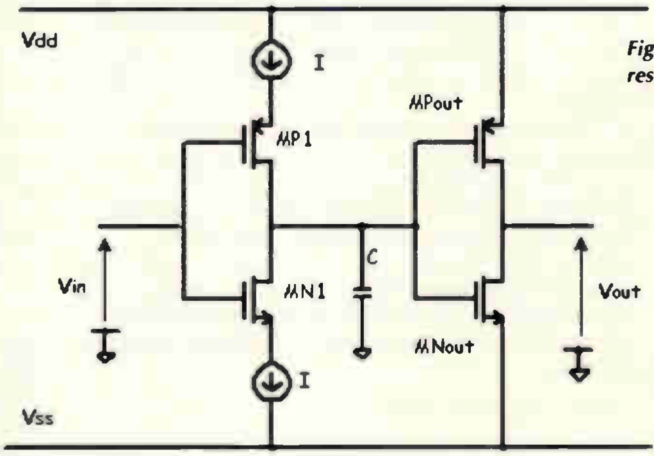


Figure 11: Phase shifter resistive interface.

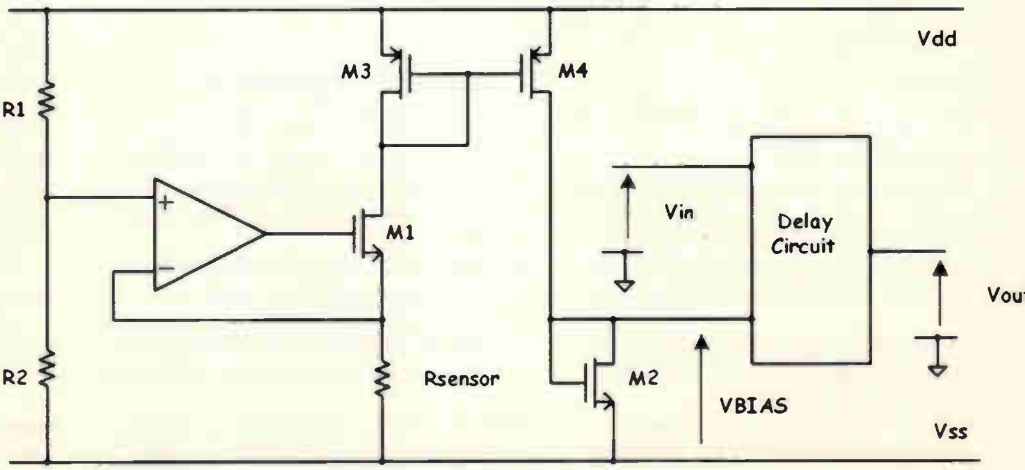


Figure 12: Controlling delay time.

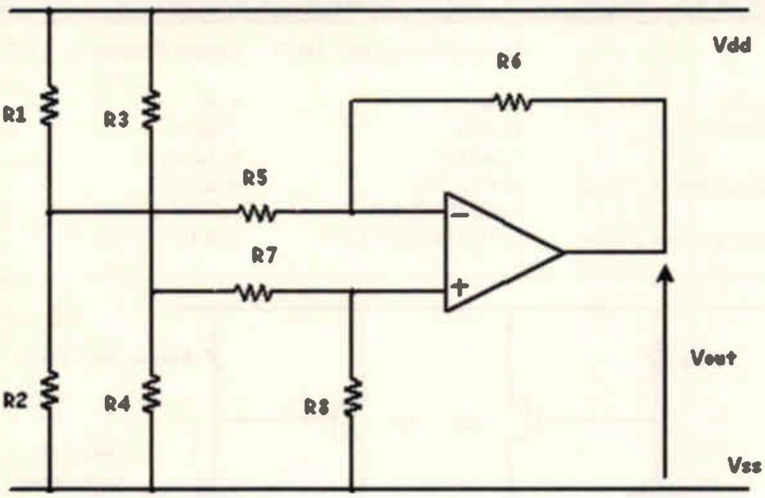


Figure 13: Introducing a differential amplifier.

possible to change the presented topology through the conversion of the shifter into an oscillator. In this case, once more the output frequency is inversely proportional to the sensor value.

Obviously, the first two topologies with the amplifier described in the previous section can be utilised to have a resistive sensor interface where the sensing element is the resistance instead of the capacitance.

Low voltage low power temperature sensor interfaces

Traditional temperature sensors make use of bipolar technology integrated in a chip. They normally sense the difference of two base-emitter voltages, biased by different currents, to detect temperature variation. To improve the sensitivity of this circuit they need a differential amplifier at the output. Unfortunately bipolar technology is very costly, so they actually have to be substituted by cheaper solutions.

A temperature sensor can be designed in a CMOS standard process (Mietec 0.7μ)[8]. It makes use of a temperature sensitive Wheatstone bridge, of Fig. 9 type. The bridge is fabricated using polysilicon resistor layers of positive first order temperature coefficients (a1 = 620ppm/°C) to fabricate opposite resistors R1 and R3 and of negative temperature coefficients (a2 = –2100ppm/°C) to fabricate R2 and R4. In this manner, the output voltage of the bridge will be proportional to temperature variation as follows:

$$Vout = {}^1/_2 Vcc\,(a1 - a2)\,T$$

being Vcc the supply voltage and:

$$R1 = R3 = R0\,(1 + a1\,(T - T0));$$
$$R2 = R4 = R0\,(1 + a2\,(T - T0))$$

The output voltage is independent on the values chosen for the resistances. The circuit has been powered at low voltage supply (1V).The output sensitivity is about 1.3mV/°C. Resolution is 10^{-5} °C, considering a 10kΩ resistance at room temperature. The introduction of a differential amplifier **Fig. 13**, which improves the circuit sensitivity, opens an interesting problem from the noise matching point of view. Using this logic, it is possible to determine the temperature range where the complete system has the best noise performance.

The total output noise depends on both the resistances' noise (thermal) and the amplifier noise (thermal and flicker). Noise calculation is fundamental to determine the system resolution. An accurate evaluation of the noise brings us to the conclusion that noise is generally given from external resistances if the described amplifier is use[9].

A suitable design implies the possibility of having a high linearity for a large temperature range variation. A 0.01°C resolution, one order of magnitude lower that commercial digital thermometers (e.g., National Semiconductor LM 35 and all the other related ICs, whose minimum resolution is 0.5°C at 25°C), can be achieved even if for a confined range of temperature (0-40°C).

Conclusion.

This paper has been written with the aim to give to the *Electronics World* reader some basic concepts about sensor interfacing.

The main aspects can be summarised as follows:

1) Get information about the sensor peculiarities (e.g., non-electrical characteristics, variation range, ...) and if it is resistive, capacitive, etc., and choose an interface topology to work at the best sensor operating frequency, so as to obtain high sensitivity and low resolution.
2) Use a good CMOS technology with low spread of technology parameters.
3) Use a low-voltage low-power interface design: the interface will become of a portable kind.
4) Minimize the effect of the interface in terms of noise, offset, and make the circuit insensitive to supply and temperature variations.

As it is clear from the considerations in the paper, it is very difficult to design a standard sensor interface, but the proposed amplifier can certainly constitute a general-purpose interface to be used in capacitive and resistive sensor applications. ■

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Wide digital I/O from the USB port

Intrigued by Colin Attenborough's recent article on digital i/o using the USB port, but don't like PLDs? Read on, as with this article Colin shows how the PLD can be swapped for four standard CMOS ICs, with increased word width as a bonus.

Some months ago, I showed how a module using the FT8U245AM chip could be used in a simple 16-bit i/o interface connected to the USB port of a PC. While the design used few chips (input/output buffers and a PLD), and, with parallel data handling in the PLD, was fast, it was not easily adaptable to word widths greater than 16 bits. Wider words would demand more PLD pins, with an increase in cost for the PLD.

These shortcomings led me to consider sacrificing speed for the ability to cope with wider words. In a design using a PLD, it is prudent to add buffers to protect the PLD from the outside world. If we use shift registers rather than simple buffers, then serial data to and from the registers reduce pin count on the PLD. Going further, I realised that the control logic between the USB module and the registers could be accommodated in four standard CMOS ICs – the PLD could be discarded. The approach works for any reasonable word widths – just add a register at input and output for each 8-bit increase of word width.

USB module pin functions

As well as the 8 data bi-directional data lines, the USB module has four data control lines. At a high-to-low transition on the NOT_WR input, data are read from the data lines into the transmit FIFO in the module's FT8U245AM chip. The NOT_TXE output goes high while the FIFO is entering the data; while NOT_TXE is high, there must be no more high-to-low transitions on NOT_WR. For data output from the module, each byte written into the module causes a high-to-low transition on the NOT_RXF output. External circuitry must respond by dropping the NOT_READ input for at least 50ns and then raising it again; NOT_RXF will rise in response. While NOT_READ is low, the bi-directional data lines assume a low impedance output state; at all other times, they are high impedance inputs.

Figures 1 and 2 show the circuit of the interface in a 16-bit form. The output registers, 74HC595 (Figure 2), are dual-rank; data are shifted serially along an input register by clock pulses on SCK, and then transferred to the output by a pulse on RCK. The input registers, 74HC589 (Figure 2), are also dual-rank; a positive going edge on RCK transfers input data to a parallel register, a negative-going pulse on *SLOAD transfers it to a serial output register, from where it is shifted out serially by clock pulses on SCK.

Six of the eight data lines of the USB module are used to control data transfer to the interface. (The allocation of function to line is arbitrary.)

D0: Transfer data from parallel to serial section (*SLOAD input of 74HC589s)

D1: Strobe data into output registers (RCK input of 74HC595s)
Load from-world data into parallel registers (RCK input of 74HC589s)
Reset counter for input shift pulses

D2: Serial data to output (SER input of 74HC595)

D3: Clock for output data (SCK input of 74HC595)

D4: Serial data from world (SER_OUT output of 74HC589)

D7: Start input of data

When sending data from the PC to the interface, the NOT_RXF output of the USB module gives a high-to-low transition for each byte sent. A delay circuit, consisting of inverter U1f, C1, R1 and Schmitt NAND gate U2a gives an immediate high-to-low transition on the NOT_RXF input, which lasts for a period defined by C2, R2, (and the threshold voltages of U2a). Inverter U1e disables the output of the input shift register while the outputs of the USB module are active – this avoids contention due to two outputs at the same point possibly trying to pull in opposite directions.

Reading data from the interface to the PC involves

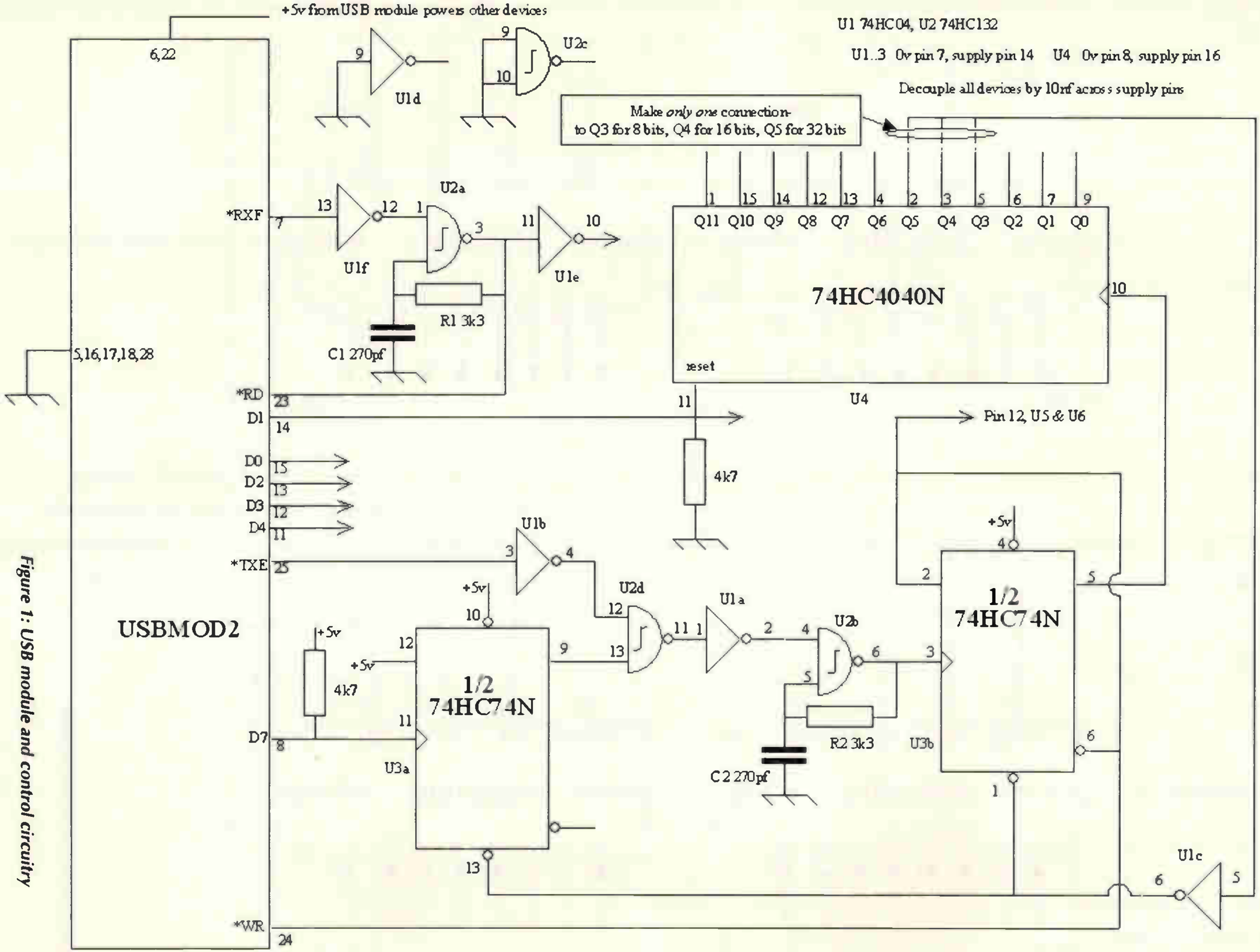

Figure 1: USB module and control circuitry

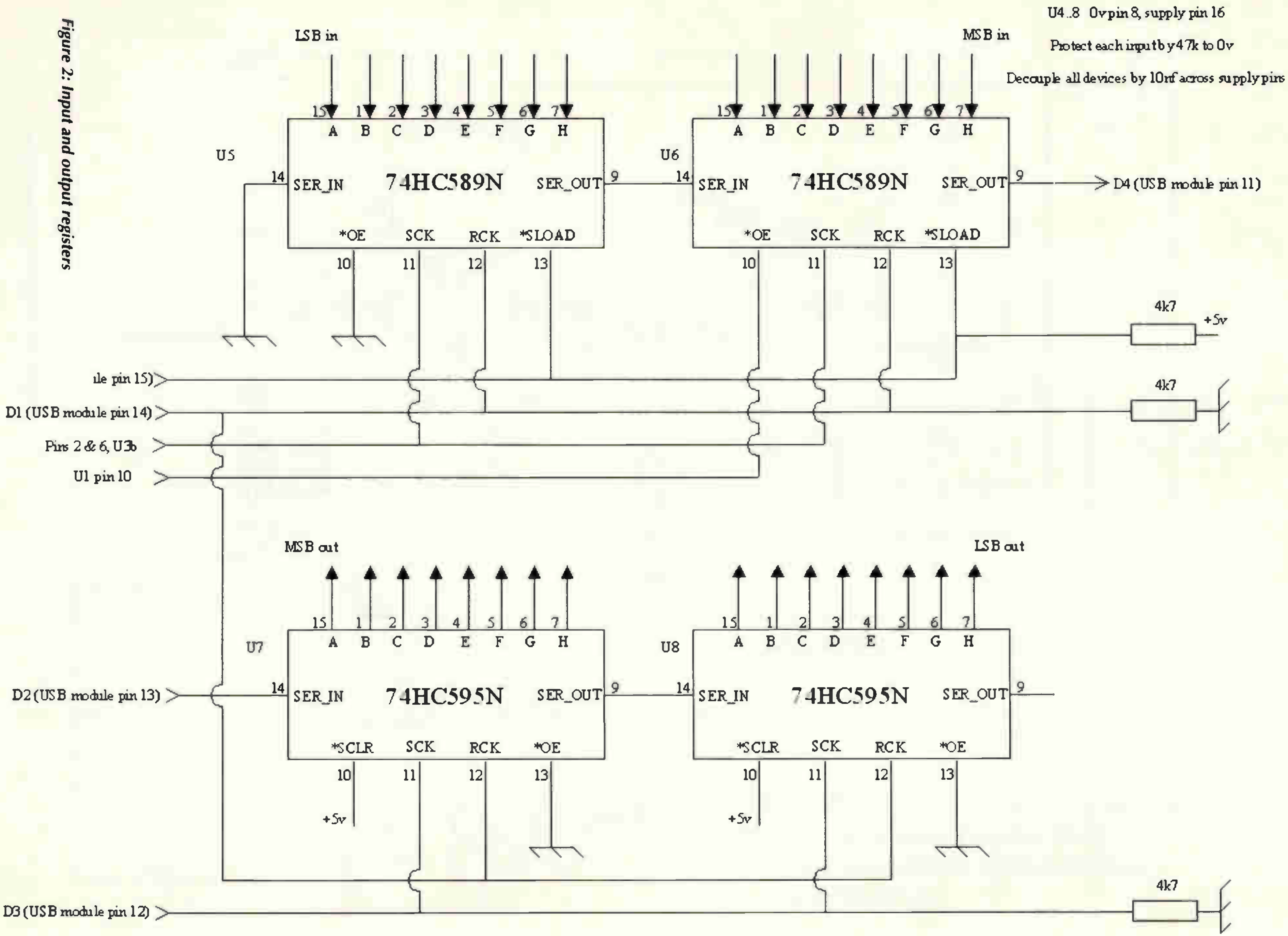

Figure 2: Input and output registers

pulsing the *SLOAD and RCK inputs of the 74HC589 devices to get input data into their shift registers. For an interface with an n-bit wide word, n clock pulses must then be applied to the USB module's NOT_WR input. This process is initiated by a low-to-high transition on the clock input of the D flipflop U3a, putting a logic '1' on the Q output and enabling the oscillator consisting of C2, R2, and Schmitt NAND gate U2b. U1a, U1b and U2d inhibit the oscillator when NOT_TXE is at logic '1'. The D flip-flop, U3b, divides the oscillator frequency by two, providing antiphase drive to the USB module's NOT_WR input, and to the counter U4. When the selected output of the counter goes high, both sections of U3 are reset via U1c, inhibiting the oscillator.

Some of the USB module data lines need pull-up or pull-down resistors. D0 (*SLOAD) and D7 (start input of data) need pull-ups, as the signals they drive are active low; D1, D3 and D4 need pull-downs. D2 must not be pulled in either direction – each data bit for output is put on D2 for two successive writes, and the shift clock is raised for the second of these writes. The system relies on the data line maintaining its state between the two writes – which it does easily, but only if the data line isn't pulled either up or down.

A question of timing

The circuit contains two timings set by analogue components – the delay circuit around U2b for writing to the interface, and the oscillator around U2b for clocking data into the USB module. The period of the delay and the frequency of the oscillator are not closely defined due to wide variations in the threshold levels of the Schmitt trigger gates U2a and U2b. However, the values shown gave periods on the prototype about ten times longer than the minimum values required by the USB module and so should be 'safe'.

Different word widths affect both software and hardware. In the Visual Basic source code, look for the definition of N_BYTES – set it to 1, 2 or 4 for 8, 16 or 32- bit word widths. (If you want, say, 64 bits, then you'll have to provide more check boxes to define the output data and display the input data a different way.) Setting N_BYTES sets up other program parameters, which ensure the correct width of the displayed form, the correct number of check boxes for input data, and the correct length of the strings that communicate data to and from the USB module. (That's why there are three compiled versions for 8, 16 and 32 bits.)

The effects on the hardware of different word widths are twofold. The figures show a 16-bit system; for wider words, insert additional 74HC595s between pin 14 of U7 and pin 13 of the USB module, and additional 74HC589s between pin 9 of U6 and pin 11 of the USB module. The number of clock pulses fed to the NOT_WRITE input of the USB module must also be correctly set by using the appropriate output of the counter U4. Remember that U4's Qn output goes high after 2n pulses have been applied. (Actually, the system will work if, say, a word width of 16 is used but 32 NOT_WRITE pulses are applied; the extra pulses are ignored once the expected word length has been shifted in, but this slows the system avoidably.) If you try to read a 16-bit input word when Q4 has been set up to give eight pulses, you'll get a 'timeout' error in a message box.

The USB module

My previous article used the DLP_USB01 module from DLP Design; this one uses the USBMOD2 from GigaTechnology.com Pty. Both can be obtained from Alpha Micro Components (www.alphamicro.net). My use of the GigaTechnology device here does not imply any criticism of the DLP module. The DLP device contains an EEPROM that stores the USB serial number, product description and other strings. The GigaTechnology device does not include such an EEPROM, but its omission does not affect the operation of the unit described here.

The DLP module is on two rows of 0.1" separated pins, the rows being 1" apart. It can be soldered into a perforated board. The GigaTechnology device is suitable for plugging into a 32-pin 0.6" pitch IC socket. (No, me neither; I used a 40-pin socket and did gentle violence with a very small saw.)

Of course, if you want to use a PLD (or a PIC) then the functions of U1, U2, U3 and U4 can be implemented in a single package. "This is left as an exercise for the student".

Software

The Visual Basic code provides a GUI to allow setting of the outputs and reading of the inputs. Changing a tick box to alter an output line generates a character string reflecting the tick box states; this string is sent to the USB module, providing clock, data and strobe pulses to the output registers. Pressing the "Read" button also sends a string of characters to the USB module, which latch data into the input registers and clock U3a, allowing the U2b oscillator to run and clocking in data from the input registers.

A dynamic link library provides the functions that interface Visual Basic to the USB driver; the DLL is in the same directory as the other source files.

What's on the disc

The disc contains three directories:

- Vb Source- the source code for readers with Visual Basic, who can examine and modify the program.
- Compiled- containing three subdirectories, Install8, Install16 and Install32, for readers without Visual Basic.
- Driver- containing a zipped version of the driver that must be installed before the USB module can be driven.

The D10202 file in the Driver directory should be unzipped and stored in a directory somewhere. (You'll find it also contains the dynamic link library- this is already embedded in the compiled versions of the code. It's also included among the source code files.) When you plug in the USB module for the first time, the system will search for a driver; browse your way to where the driver has been unzipped and it will be installed.

The driver file was downloaded from the FTDI website www.ftdichip.com; it works entirely satisfactorily, but readers may wish to check for an updated version. ■

Acknowledgements

The author is grateful to his employers, Cambridge Consultants Limited, for permission to publish this article.

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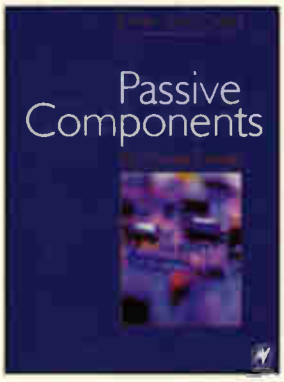

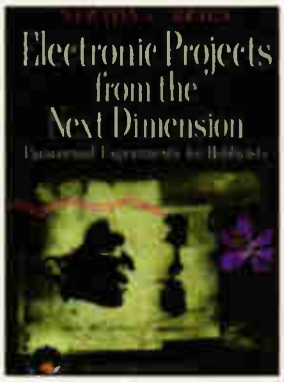

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Pulsar Electronics is offering a 10Gbit/s physical layer (PHY) device from US firm Mysticom, which the supplier claims features the industry's lowest power and best bit error rate (BER). Based on a patent-pending architecture that incorporates what the firm called an all-digital receiver, the Mysticom MY3004 reduces BER by two orders of magnitude compared with conventional transceivers. Adaptive line-conditioning software is supported which means the device automatically adjusts signal parameters whenever system configurations change. In addition, the device supports a variety of frame formats that allow it to be used as an Ethernet, InfiniBand, or Fibre Channel Transceiver. Packaged in a 17x17mm ball grid array the device is a four-channel 3.125 Gbit/s serialiser/deserialiser (SerDes) device with an XAUI/XGMII HSTL 1.5 volt compliant interface. It features a power consumption of 1.1W equating to 200mW per channel and selectable 8B/10B encoding/decoding with timing options which provide hook up to upstream Asic, FPGA or other devices. It supports source-centred and flexible source-synchronous timing.

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Siroyan signs up to ewb-based SoC evaluation

IP company Siroyan has added its scalable DSP technology to SOCworks, the web-based SoC evaluation and simulation environment from Sonics. Under the terms of the agreement, SoC designers will get free access to Siroyan's recently launched SRA328 DSP core to assess performance and interoperability. SOCworks is an online environment that is designed to offer services for the locating and evaluating commercially available semiconductor IP cores by enabling their remote assembly into models of targeted SoC applications for full chip data flow simulations. Siroyan's OneDSP architecture uses instruction-level parallelism to provide up to 32 execution-unit clusters to be implemented by the licensee with either 32 or 64-bit data paths. At 200MHz, OneDSP will scale from 400 MMACs, with just the master cluster deployed, through to 25,600 MMACs, with a full complement of 32 clusters.

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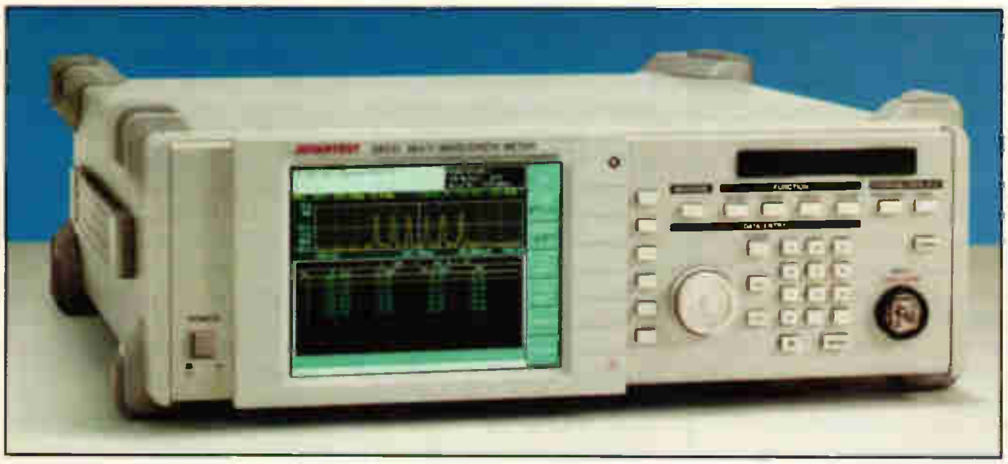

PCB prototyping tool supports IP re-use

Zuken has upgraded its Hot-Stage virtual prototyping tool to enable electrical constraints to be defined and reused within a schematic capture environment. Version 4.03 enables definition of high-level meta-constraints for complete signal paths within the schematic, each constraint being capable of application to hierarchy levels. The tools have always allowed users to establish electrical and physical constraints from a central database of information, but only logical attributes could be defined within the schematic. Effective IP re-use requires the ability to capture not only logical net constraints, but also the constraints associated with electrical nets and complete timing paths. At the schematic level, engineers have always been able to specify delays for logical nets N1 and N2, but have not been able to specify overall path delays. Such low-level attribute definition leads to over-constrained routing, particularly in high-speed designs.

Zuken
Tel: +44(0) 1454 207800
www.zuken.com

75A Current sensor

Allegro Microsystems has developed a current sensor based on a precision linear Hall-effect integrated circuit coupled with a

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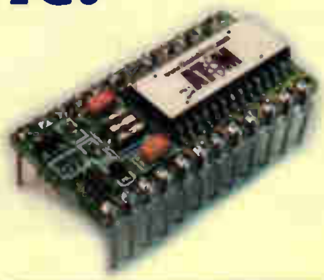

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www.milinst.com

Milford Instruments Limited, Tel 01977 683665, Fax 01977 681465, sales@milinst.com

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magnetic circuit optimised to maximise device sensitivity. The entire assembly is housed in a package configuration with automotive or industrial applications in mind. The first device based on the design is the ACS750, which will measure currents of up to ±75A with an accuracy better than 1% and linearity better than 0.5% said the supplier. The sensor is designed for through-hole PCB mounting, with the power leads electrically isolated from the sensor leads. The power leadframe used for current sensing is designed for low power loss, and has an internal resistance of less than 1mΩ. The sensor exhibits a low thermal drift of offset voltage (±0.008%) and of sensitivity (±0.01%) and has a response time of 23μs.
Allegro Microsystems
Tel: 0033 450 512359
www.allegromicro.com

SHDSL modem puts 2.3Mbit/s on unshielded cable

The Microtel Crocus SHDSL modem uses single-pair high-bit-rate digital subscriber line technology to provide full duplex transmission at up to 2.3Mbit/s over a single two-wire unconditioned unshielded twisted-pair cable. Operation is

based on the new G.SHDSL standard for higher speeds and longer loop ranges, and the variable line transmission rate can be increased to 4.6Mbit/s using a two-pair version of the modem. The use of TC-PAM (trellis-coded pulse-amplitude modulation) is designed to ensure spectral compatibility with other transmissions in the same bundle, so that the risks of crosstalk and intermodulation effects are minimised. It is manageable under HP OpenView, and is supplied with advanced free maintenance software.
Microtel
Tel: +44(0) 1322 552020
www.microtel.co.uk

RF screening of radiated emissions

Microponents has patented its RF screening system to provide designs with radiated emissions compliance. The system, which is aimed at small-scale electronics manufacturers, allows engineers to build screening enclosures of various sizes, shapes and heights. A design feature is the way in which a profile can be formed whilst maintaining the 2.54mm spacing to fit on development boards. Other features are multi-fence heights, assembly without tools and solderability.
Microponents
Tel: +44(0) 121 3800100
www.microponents.com

Development kit with a view

Impact Memec is offering the LabVIEW development kit from Xicor for evaluating and developing systems incorporating the firm's programmable mixed-signal devices. The development environment uses low-level device drivers to allow programmability of specific features such as voltage gains, comparator threshold levels and other key analogue parameters. The kit includes schematics for the hardware interface that is connected via a PC parallel port. The software operates with any PC running Microsoft Windows.
Impact Memec
Tel: +44(0) 1296 336100
www.impact.uk.memec.com

Waveform analyser has gigaword of memory

The DL750 ScopeCorder from Yokogawa is a portable waveform measuring instrument that combines recorder and oscilloscope technologies. The instrument's exceptional memory length of up to one gigaword coupled with 10Msample/s sampling makes it suitable to the capture and analysis of high-speed signals over long time periods of up to ten days. It can act as an oscilloscope for capturing instant events and as a data recorder for prolonged trend measurement. A dual-capture function allows the instrument to capture slow trend data and high-speed events simultaneously, while a high-speed processing facility called Giga-zoom can process the entire 1 gigaword of data stored in the memory and display it. The instrument can capture data on up to 16 analogue channels plus one 16-bit logic input. Data is sampled at up to 10Msamples/s with 12 bits A/D resolution via a 2-channel isolated input module, or at 1Msample/s at 16 bits using a high-sensitivity (1mV/div) input module.
Yokogawa
Tel: +44(0) 1494 459200
www.matron.co.uk

Audio DAC works off 3V

Wolfson Microelectronics has introduced a family of audio DACs that can run at supply voltages down to 3V. The first device, the WM8726, is a 24-bit, 192kHz stereo DAC which is targeted at applications such as digital TV and digital set top boxes, home theatre systems, MP3, CD and DVD players. The WM8726 is a pin compatible upgrade to the Scottish firm's WM8725, featuring on-chip digital volume control, improved linearity and a data interface that supports industry-standard DSP data formats. The device has a signal-to-noise ratio (SNR) of 100dB and Total Harmonic Distortion (THD) at –95dB. It is available in a 14-pin SOIC package.
Motorola
Tel: +44(0) 1355 565000
www.mototrola.com

Low profile PCI cards have 2 or 4 ports

Brain Boxes are offering two low-profile PCI communications cards. The two port and four port cards are less than 85mm in height. The cards follow a one port version released to coincide with the launch of low profile PCI as a standard in February 2000. Low profile PCI is similar

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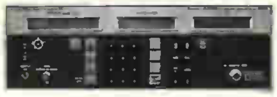

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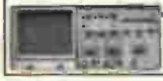

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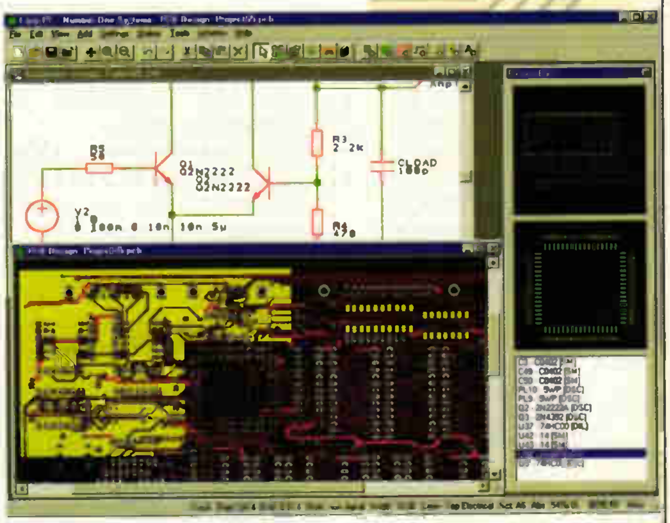

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to PCI in terms of electricals, signals, and drivers, but possesses a more compact form factor with a shorter raw card and new mounting bracket. The two new cards are backward compatible with standard PCI brackets. Both new cards are offered with a customer support package including a three year warranty.
Brain Boxes
Tel: +44(0) 151 220 2500
www.brainboxes.com

Small package buck voltage regulator

A buck regulator in a package claimed to be 36% smaller than the commonly used TO-263 (D2PAK) is available from Micrel Semiconductor. The MIC4685, a member of SuperSwitcher family, is a 200kHz, PWM buck regulator house in a 7-lead SPAK package. The device has an input voltage range of 14V to 34V, and a 3A output current. Output voltage is adjustable down to 1.25V. Features include a cycle-by-cycle current limit, frequency foldback short circuit protection, and thermal shutdown. Samples are available from stock, and production quantities are on a 12 week lead time. Efficiency is over 85% and a typical shutdown current is 150µA.
Micrel Semiconductors
www.micrel.com

1.5Gbit/s serial connectors

Honda is offering a family of 1.5Gbit/s serial ATA connectors. The company also expects to release 3Gbit/s and 6Gbit/s versions by 2004 and 2007 respectively which it said is in line with the projected rollout of the serial data interface standard. They are designed to support the bandwidths which will be required in RAID arrays and multimedia PCs for transferring data between high capacity HDDs and to CD-ROM/DVD drives with read/write capabilities. Features include a passive locking system and a polarisation feature, as well as hold downs to improve retention to the PCB. The serial ATA interface is specified for cable runs of up to 100cm. It has been implemented on a cable-mount signal connector, a combined PCB mount connector and a 7-pin PCB mount signal connector.
Honda
Tel: +44(0) 1793 523388
www.hondaconnectors.co.uk

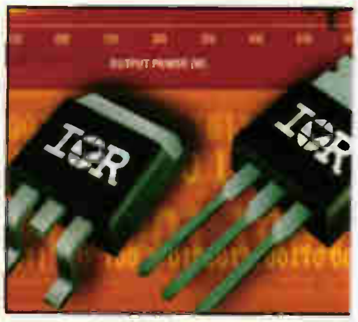
EMC horn antenna for 18GHz

A double-ridge horn antenna has been added to the range offered by Schaffner. The BHA9118 is suitable for applications such as RF and EMC measurements to MIL, SAE, IEEE, IEC and FCC standards. The 1GHz to 18GHz antenna is a linearly polarised broadband unit is capable of

250V/m, at one metre. The design is rugged and corrosion-resistant, and can be used to receive and transmit. A precision N-type connector is used to provide low VSWR and high power handling capability. Each antenna is individually calibrated, following ANSI 063.5 traceable to national standards. The antanna is supplied with a calibration certificate that contains the measured antenna factor data certifying precise antenna performance.
Schaffner
Tel: +44(0) 1306 710205
www.schaffner.com

Power Mosfets cut on-resistance by 40 per cent

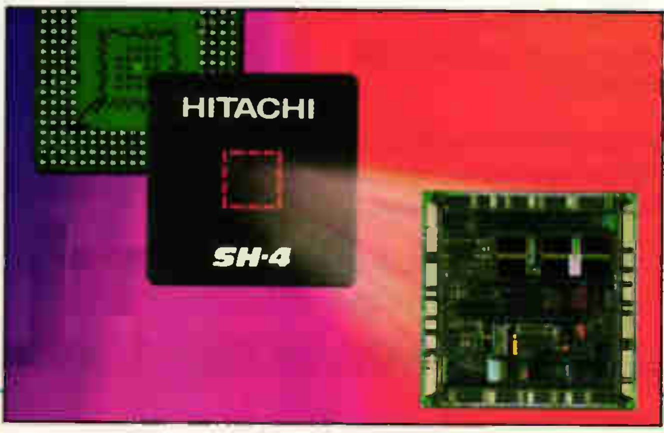
International Rectifier has introduced the 100V, IRFB4710 and IRFS4710 HEXFET power Mosfets that it claimed increase power density in 48V input, half- or full-bridge topologies used in the top-end DC-DC converters required for telecoms and datacomm systems. The IRFB4710 device is rated at 75A in the TO-220 package, or 30% higher current than previous generation devices. The Mosfets have a 40% lower on-resistance, or RDS(on), than previously available devices. According to



the supplier, in-circuit testing shows that four IRFB4710s run 20°C cooler at 350W compared to industry-standard devices, The IRFS4710 MOSFET, in a D2Pak package is optimised for 100 to 300W, board mounted power systems. The IRFB4710, in a TO-220 package is designed for 3kW to 5kW shoe-box type power supplies used in wireless base stations. ■
International Rectifier
Tel: +44(0) 20 8645 8003
www.irf.com

32-bit superscalar embedded processor

Hitachi has announced two SH-4 32-bit superscalar Risc embedded microprocessors that offer higher frequency and a new cache architecture for multimedia applications, car information systems and networking applications. The devices include a superscalar archtecture, full floating point engine, SIMD (Single Instruction Multiple Data) acceleration for 3D and DSP, and 16-bit fixed instruction length for low memory footprint. The SH7751R offers an integrated PCI controller while the SH7750R has a 64-bit SDRAM interface for increased memory bandwidth. Both offer increased performance (430MIPS/1.7Gflops) and higher frequency (240MHz) than existing SH-4 devices, with 64-bit bandwidth to provide a memory throughput up to 120MHz SDRAM access at a peak of 960Mbyte/s. The new cache architecture from direct map to 2-way set associatve reduces the cache miss and increases the real time capabilities. Operating at 1.5V, they have four additional DMA channels (eight in total), support a temperature range from –40 to +85°C, and provide 16kbytes instruction and 32kbytes data cache memory. The devices support the complete multimedia software stack which includes operating systems, JavaVM, browser, e-mail, OSGI, VoIP, H.323 suitable for Java-based embedded applications, Internet appliances and Multimedia interactive TV.
Hitachi
Tel: +44(0) 1628 585161
www.hitachi-eu.com



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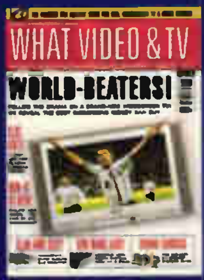

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Clear hand-written notes on paper are a minimum requirement: disks with separate drawing and text files in a popular form are best – but please label the disk clearly. Where software or files are available from us, please email Jackie Lowe with the circuit idea name as the subject.

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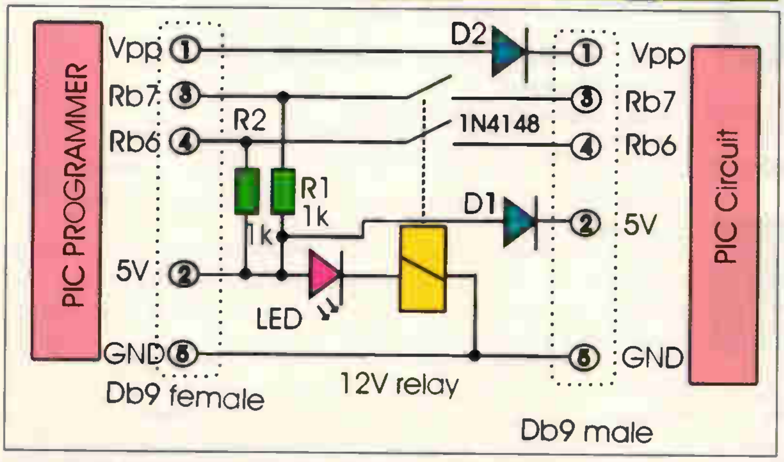
An on line programming adapter solves PIC programming pins problem

The most important secret in embedded processors is on line programming. It is done either with a ROM emulator or with a flash programming. These methods allow programming and testing in real devices. Sometimes all the programming pins of the processor are used by an external load. Then on line programming does not work. Also an on line programmer prevents operation of some processor pins. I faced this problem in my last project, namely a 12" LED display for horse races. ULN2803 in RB7 and RB6 prevented the on line programming. After I had added some booster resistors the programmer worked, but it prevented normal operation of the circuit.



The solution is to automatically switch the on line programmer on and off. This circuit works on any PIC programmer. The circuit uses a standard 12 volt 2 pole DIP relay to switch programming pins RB7 and RB6 on and off with 5V from the programmer. Diode D1 prevents circuit power to switch the relay on with normal power supply. D2 separates the reset circuit from programmer. I used a LED in series with the relay to get a visual indication and to reduce 5V loading. I found that a 12V relay works with 3V and 18mA! The R1 and R2 add some boost to the programmer for external loads. The resistors are automatically switched off after programming. Using PC software I can switch the programmer 5V on. I do not need any other power supply for testing. Which is very convenient for field programming! I used a small utility box with two DB9 connectors to build this very useful programmer adapter. When needed, I simply put this adapter in series of my normal PIC programmer.

Pekka Ritamaki
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Two-wire PIR enhancement

What follows is a low cost and novel means of removing the need for the six-wire bundle currently used to connect to PIR and other detectors in domestic and commercial alarm applications. The bundle is replaced with a more easily routeable two-wire solution. This circuit also offers certain anti-tamper enhancements over the existing six-wire designs.

Driven by the need to route wiring to a new PIR for my home alarm system, this circuit allows the use of single twisted pair to provide the full PIR functionality, which usually requires a three-pair solution.

In normal operating conditions, standard wired PIR detectors exist in one of the following three different current states:

- Idle – requiring about 15mA for the detector circuitry.
- Alarmed – about 70mA. Alarm relay closed, 50mA flows in alarm loop, plus idle current.
- Tampered, where wire has been cut – 0mA.

These values are based on an ageing Wickes system.

A simple current window detector is used in the circuit described here to distinguish between these states. Selection of R_1 and R_2 determines the window of normal operation. Nominally, a current of 15mA is required by my PIR in its non-triggered state, but this will vary according to the detector type used.

In this design, upper and lower thresholds have been set to give roughly an 18mA window of normal operation (8- 26mA).

Resistor R_1 and the V_{be} of Tr_1 determine the upper threshold while the lower threshold is set by a combination of Tr_2, R_1 and the R_5/R_6 divider.

In the idle state, Tr_1 and Tr_2 are not conducting and a current of about 15mA flows through R_1. When triggered, the PIR alarm relay (relay contacts in series with 200Ω) switches in and a total current of more than 70mA flows. The resulting voltage across R_1 switches Tr_1 into conduction, driving Tr_3 which draws current via R_9 (200Ω in this case to simulate the PIR alarm relay) triggering the alarm circuit of the original three-wire circuit.

If there's no common ground between the power and alarm circuits, a relay could be used in place of Tr_3.

In the event that the two-wire loop is cut, or the PIR case is opened causing the anti tamper switch to break the circuit, no current will flow and Tr_2 conducts. This causes the normally-closed relay RL_1 to become open circuit.

Resistor R_4 provides a little hysteresis and additional base drive to Tr_2. An open circuit at RL_1 triggers the anti-tamper circuit of the original alarm circuit. For a number of reasons, including absence of a common ground with the power circuit, a simple relay contact in the existing loop was used to convey the tampering signal.

In the event that an attempt is made to tamper with the PIR or its connecting wires by bypassing them, cutting them, or other means, current drawn will fall outside its operating window and Tr_1 or Tr_3 will conduct, stimulating an alarm. This scheme offers potentially greater tamper resistance than the original design.

Diode D_1, together with R_1, helps reduce voltage drop across the circuit to ensure continued operation of the PIR power circuits under alarm or other high current conditions.

Values shown give a normal operating window of between about 8mA and 26mA, which gives good immunity to false alarms. The debouncing feature of the parent alarm system itself and C_1 and C_2 complement this. As current in the alarm or tamper states are considerably outside the normal window of operation, switching is quick and reliable.

Maximum current delivered by the combined alarm circuit and power circuit of the parent system is understood to be no more than 100mA per port. Resistors $R_{1,2}$ should have enough dissipation handling capability to cater for a long-term short-circuit tamper.

Careful selection of R_5/R_6 could reduce current taken by the circuit to a few milliamps in current-sensitive applications.

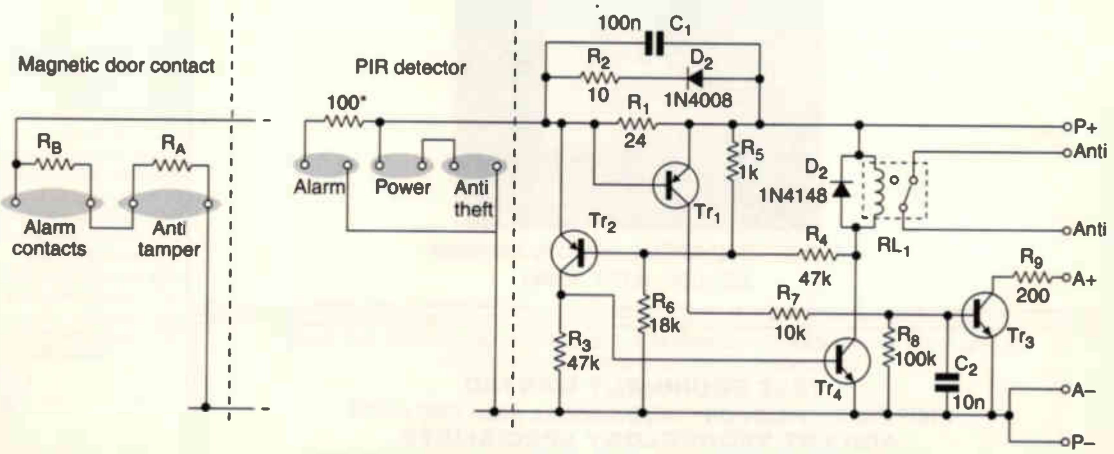
Rewiring of PIR detectors

In order to use the circuit described, it is necessary to carry out some simple rewiring of the detectors as shown in the diagram. This rewiring can be carried out simply at the screw terminal connectors of the PIR.

The anti tamper switch on the case of the sensor should be connected in series with the parallel combination of PIR power circuit and the alarm circuit.

In order to reduce maximum current drawn via the alarm relay of the PIR and hence through R_1, D_1 and R_2, an additional resistor has been added in series with the alarm contacts of the PIR. A value of 200Ω has worked well.

The alarm contact of the PIR used, remains in conduction for about three



This interface reduces the number of wires needed to connect a PIR sensor from six to two. Transistors $Tr_{1,2}$ are BC212L while $Tr_{3,4}$ are BCW34. Line P+ is the positive alarm system supply connection and P– is the negative line. The alarm connections are A+ and A– and the two conductors marked 'Anti' are the anti-tamper circuit connections.

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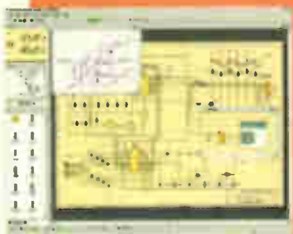

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QUICKROUTE SYSTEMS

seconds following each triggering, after which the PIR is inactive for about three seconds before being rearmed.

Door/window magnetic reed contacts

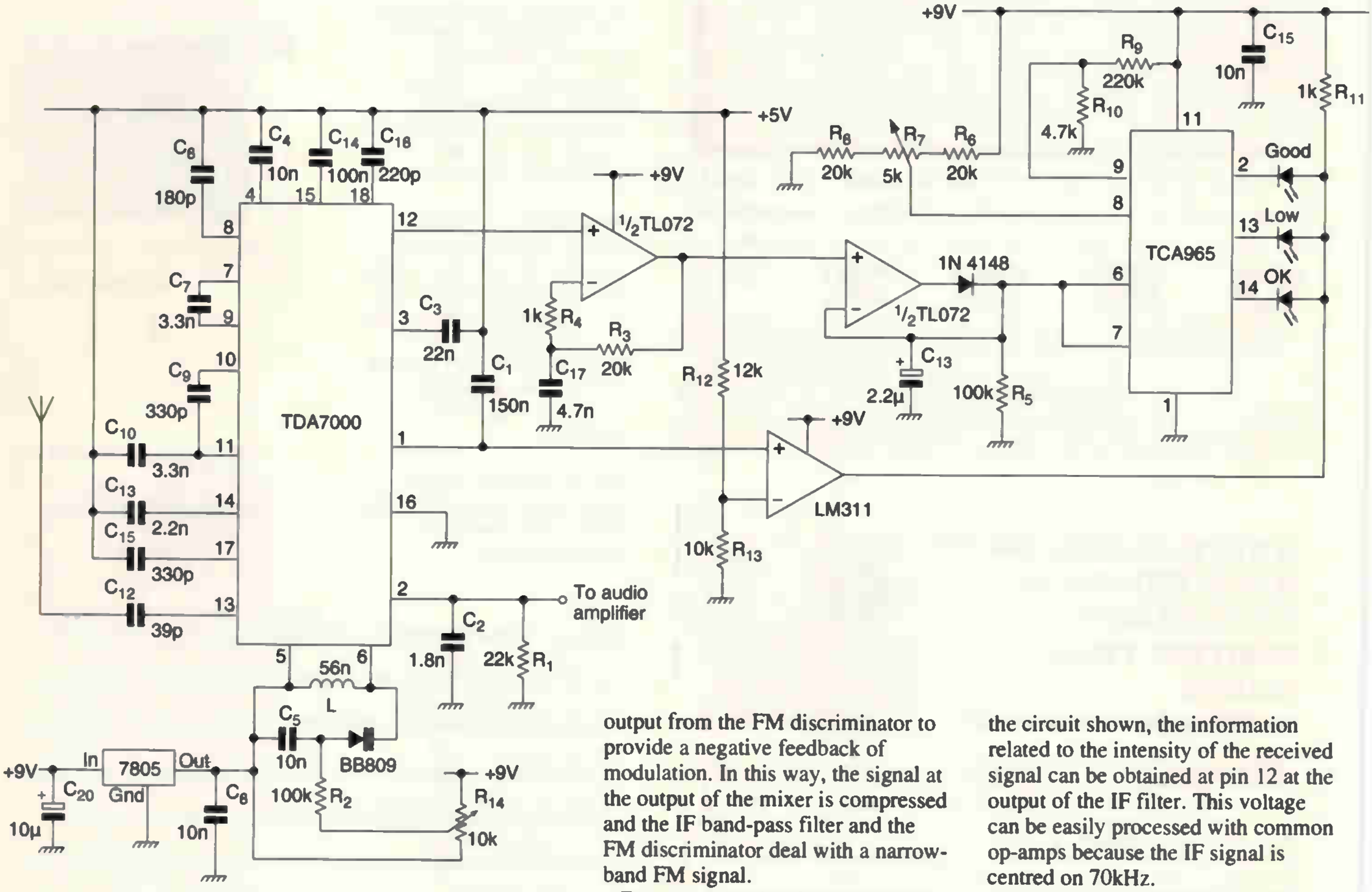
Re-wiring of door contacts is also necessary in order to provide a three-state system and allow full operation of the current-window circuit. Fortunately this is easily achieved at the screw terminal connectors by adding two resistors in the R_A, R_B positions shown in the diagram.

For optimum system performance, the resistors chosen should ensure that no more than a few milliamps of current flows. Resistors R_1, R_5 and R_6 should be scaled accordingly.

This arrangement also provides improved anti tamper resistance when compared to the original system for installations using reed type contacts as this circuit will trip the alarm when a short circuit is used to bypass anti tamper circuits.

If necessary, the circuit can also be locally powered to allow expansion of an alarm system without placing additional demand on the central system's power circuits.

Ian Jennings
Caldicot
Monmouthshire
UK



Improve the TDA7000 receiver with a signal-strength display

Philips' TDA7000 integrates a monophonic FM radio receiver all the way from the antenna to the audio output. Only one tuneable *LC* circuit for the local oscillator, a few capacitors, two resistors and a potentiometer to control the variable-capacitance diode tuning need to be added to the IC to make a complete receiver.

The 7000 has a frequency-locked-loop structure. The basic idea is simple. Local oscillator output is frequency-modulated by filtered output from the FM discriminator to provide a negative feedback of modulation. In this way, the signal at the output of the mixer is compressed and the IF band-pass filter and the FM discriminator deal with a narrow-band FM signal.

For a compression factor K=3,the original FM bandwidth is reduced to 180/3=60kHz. The IF can be as low as 70kHz so neither ceramic filters nor complex LC tanks are necessary to realise the IF filter. A simple active filter with operational amplifiers fulfils the task.

The IC incorporates a correlation muting system that suppresses interstation noise and spurious responses due to detuning. The muting circuit is built around a second mixer. Its output is available externally at pin 1 and can also be used to drive a tuning indicator.

If your application demands a high performance FM radio you can add a signal-strength display to the 7000. In the circuit shown, the information related to the intensity of the received signal can be obtained at pin 12 at the output of the IF filter. This voltage can be easily processed with common op-amps because the IF signal is centred on 70kHz.

Voltage at pin 12 is DC coupled to an amplifier. Next, an envelope detector built around the TL082 gives a DC voltage analogue to the received signal strength. This voltage is compared in the Siemens TCA965 window discriminator with a voltage derived from R_6, R_7 and R_8 – the window centre – and R_9, R_{10} for the window half width.

Three LEDs show the result of the comparison – 'low', 'OK', or good – but only when the tuning is correct. When it is, the voltage at pin 1 reaches its maximum value and the comparator LM311 enables the TCA965.

José Ma Miguel-Lopez
Barcelona, Spain

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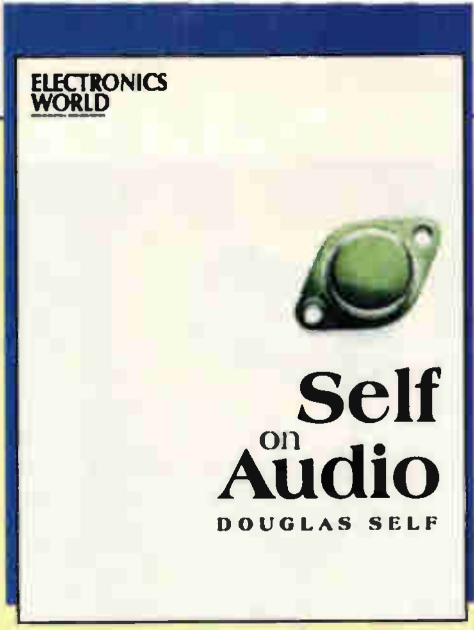

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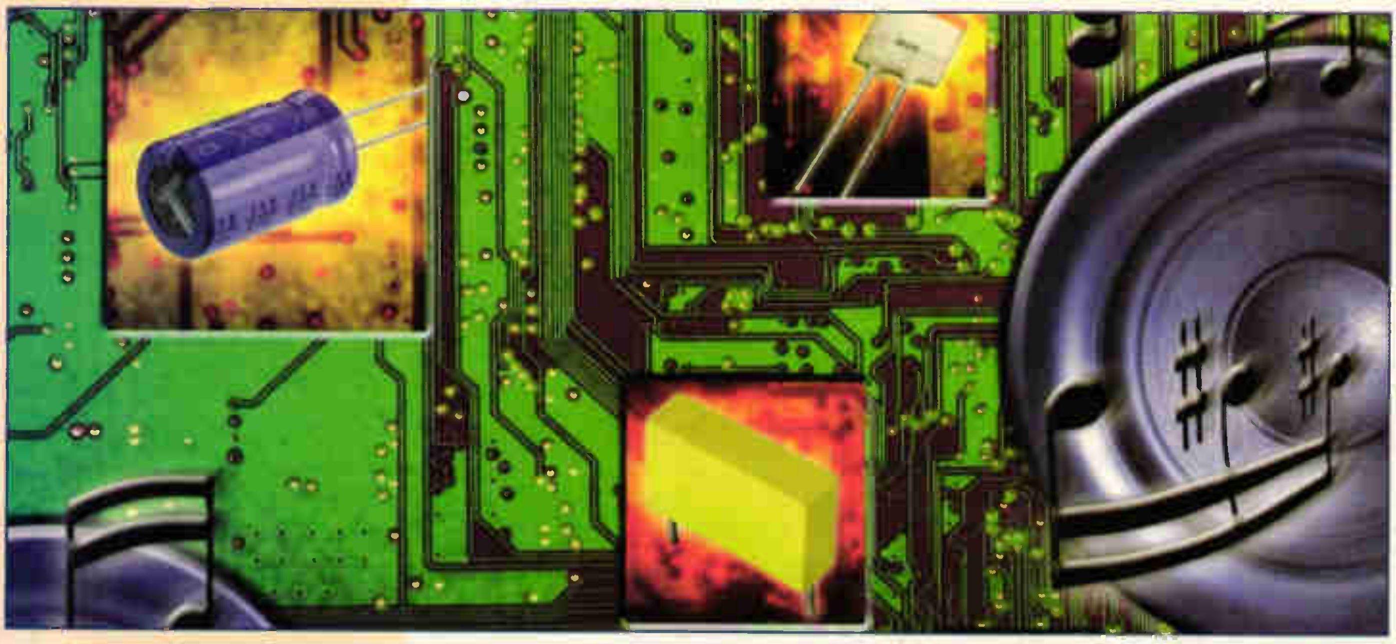

Capacitor sound 4

This month, Cyril concentrates on the difficult area of 100nF to 1μF, which usually for size and cost reasons means using metallised PET products

Readers of my previous articles will have seen that many capacitors do introduce distortions onto a pure sine wave test signal. In some instances distortion results from the loading the capacitor imposes onto its driver. In others, the capacitor generates the distortion within itself.[1]

Capacitors are not categorised for distortion in manufacture, so a distorting capacitor would not be accepted as reject by its maker. Using my easily replicated test method, capacitor distortions can now be measured, surpassing speculation. Equipment designers can now select capacitors for each circuit requirement.

For capacitances of 10nF and smaller, the safe solution is to use C0G ceramic or extended foil/film capacitors. Made with Polystyrene or Polypropylene dielectrics and with leadwires soldered or welded directly to the extended foil electrodes. Avoiding altogether capacitors made with metallised film dielectrics.

These idealised choices minimise all measurable distortion products. While this presents a counsel of perfection, as an engineer I believe prior knowledge of the best and worst extremes should form part of any compromise.

Such near ideal capacitors are not easily

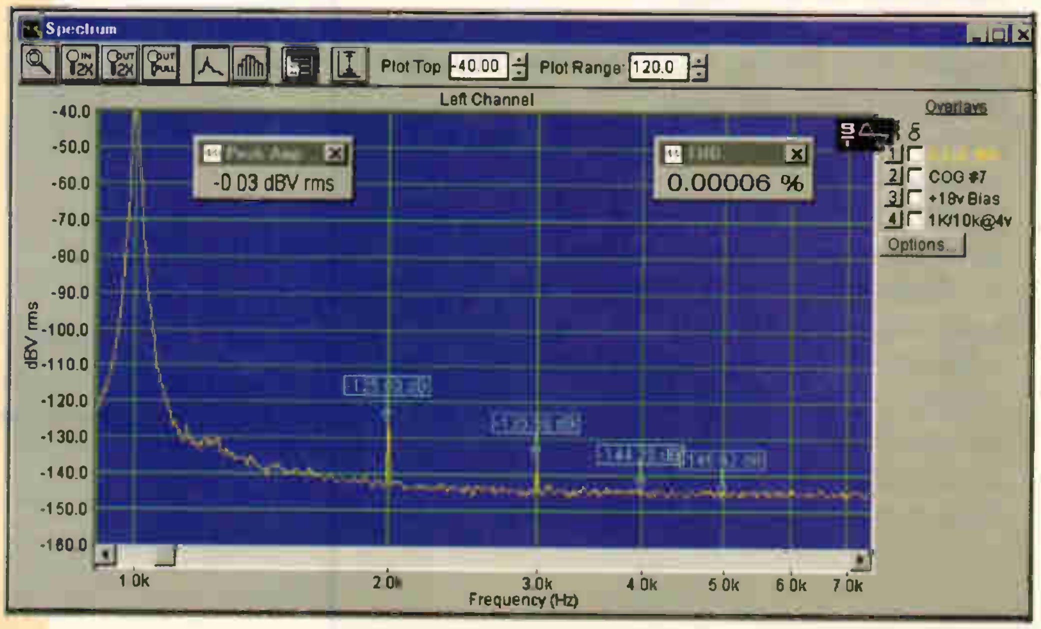

Figure 1: Distortion measurement of a 100nF 50 volt C0G ceramic, using 100Hz and 1kHz signals at 4 volts with 18 volt DC bias. With no bias this multilayer capacitor measured just 0.00004%. Second harmonic was -131.7dB, other harmonics remained as shown.

available in acceptable sizes or costs for higher capacitance values. Finding suitable low distortion 0.1μF and 1μF capacitors proved almost impossible. High 'k' BX, X7R, W5R and Z5U capacitors produce far too much distortion for our needs.[1]

Multi-layer ceramics of 100nF 50 volt manufactured in C0G, produce little distortion, with and without DC bias, but are not easily available in small quantities. **Fig. 1**

The worst capacitor?

A 100nF ceramic disc capacitor is still available. Having the thinnest possible high-k dielectric it provides the worst possible distortion. Despite this, a number of papers found on the internet choose to use this style on which to base their ceramic capacitor measurements and opinions.[2]

Originally called a 'transcap', it pre-dated all low cost 0.1μF film capacitors by many years. It was developed as the smallest, lowest possible cost capacitor, used in pocket transistor AM radios.

A conventional high 'k' ceramic, re-sintered in a reducing atmosphere, becomes a semi-conducting disc measuring a few Ohms resistance. The outer few surface molecules are re-oxidised when the electrode silver is fired in air, to become the dielectric. If sectioned, you will find a black disc, apparently made from charcoal. Using a high power microscope, you may just see an extremely thin, much lighter coloured dielectric layer covering the outer surfaces.[3]

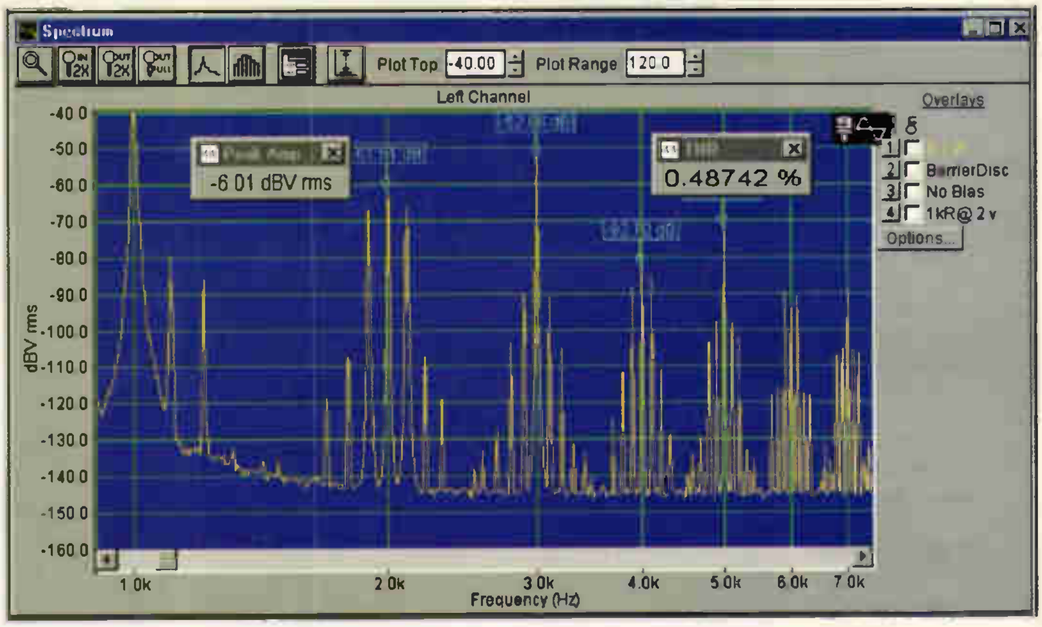

Figure 2: The worst distortion of more than 2000 capacitor measurements. The test voltage had to be reduced to two volts AC with no DC bias, to avoid harmonics overloading my soundcard.

Such devices have no place in any audio system. So take care if offered a small ceramic disc, having significantly greater capacitance than found in conventional disc capacitors.[4] **Fig. 2**

Electrolytics

Tantalum or aluminium electrolytic capacitors are available in these values and form the subject of my next article. Meanwhile we will investigate the options available in film capacitors.

Very low distortion foil and film, Polypropylene (PP) and Polyethylene terephthalate (PET) capacitors are available but are large and usually expensive. The lowest cost, smallest size capacitors, are made with metallised PET.

Metallised PET capacitors

In the drive, some thirty years ago, to size and cost reduce the 0.1μF capacitor, two problems had to be addressed:

1) First was to produce satisfactory quality, extremely thin metallised PET. In 1978, the Dupont 'Mylar'® capacitor film became available at a thickness of 1.5 microns, some 20 times thinner than human hair.

2) The second was to develop low labour cost methods to wind small capacitor elements. For capacitor makers this was difficult because of the high cost and large numbers of

Metallised film dielectrics

All common film capacitor dielectrics, other than Polystyrene, can be metallised, to produce a negligibly thick electrode. This metallised coating, usually aluminium, is produced by evaporating metal ingots inside a vacuum chamber. The film is stretched taut and passed through the chamber at controlled speed. To prevent overheating, the film passes over refrigerated rollers.

The metallised coating is so thin, it is transparent. Thickness is monitored by measuring resistance, typically a few Ohms per square, of the metallised surface.

PET and PPS films are easily metallised and provide good adhesion to an evaporated aluminium coating. Untreated Polypropylene has a smooth, waxy surface, which inhibits adhesion.

Various pre-treatments have been applied to PP to improve electrode adhesion. These include mechanical roughening and exposure to high voltage ionisation fields. However, a metallised electrode is often applied to a higher resistance value, i.e. thinner, onto PP than other films.

Contact to the metallised electrode is made by spraying minute metal particles, evaporated inside a high temperature spray gun, onto each end of the capacitor winding. This is known as a 'Schoop' connection. The volume of air needed to propel the metal particles ensures the film surface is only exposed to relatively cool metal and so does not melt.

This 'schoop' metal spray end connection is also used to manufacture some makes of foil and film capacitors and those with double-sided metallised carrier film electrodes. The conductive end spray, short circuits together all turns of a wound capacitor, ensuring minimal self-inductance.

When sufficient 'end spray' thickness has been applied, the capacitor leadwires are attached, usually by soldering or electrical resistance 'welding.' Properly applied this 'schoop' end spray provides a good connection to the electrodes, able to carry significant current. The extremely thin metallised film electrodes obviously cannot handle high currents. When overloaded, visible electrode 'edge burning' occurs, ultimately leading to an open circuit capacitor.

The resistance of the metallised electrode combined with aluminium's temperature coefficient of 0.0039, results in a non-linear resistance. This may at least partially explain some of the larger third harmonic distortions. One simple indicator of the current carrying ability of the 'schoop' end connection into the electrodes used, can be seen in the peak current ratings claimed for the capacitor. For example a 10nF metallised PET capacitor might be rated for 30v/μsec, foil and PET has a much higher current carrying ability, being rated as high as 1000v/μsec.

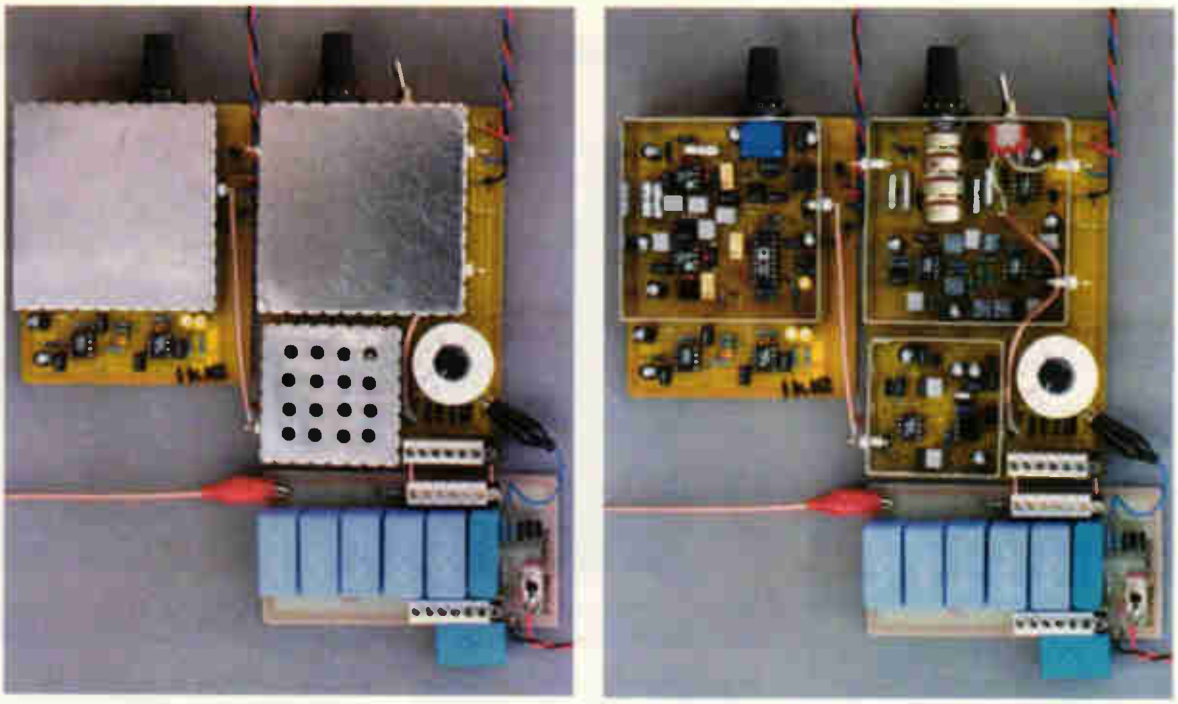

Figure 3: Finalised measurement system using two test signals, 100Hz and 1kHz, to measure capacitor intermodulation and harmonic distortion, with and without DC bias voltage. The capacitor under test is mounted directly onto the DC bias buffer network. The red crock clip and screened cable supply the 100Hz signal. All screening case lids must be fitted while measuring distortion.

automatic winding machines needed to produce capacitors in volume.

The major German capacitor makers were leading these developments. Wima with others, worked to develop intricate machines capable of automatically winding individual small capacitors. The Siemens company, now Epcos, sought a different solution, their so-called 'stacked' capacitor.

Despite their name, stacked film capacitors are first wound onto a large diameter wheel, to make a 'mother' capacitor. When all possible processing stages are complete, this 'mother' is sawn into short lengths, each a discrete capacitor element.[5].

During my initial distortion measurements on metallised PET capacitors, I was curious as to whether these two processes would result in different distortion characteristics.

Concentrating my measurements on known wound BC Components type 470 and known stacked Epcos capacitors, I did find differences. The stacked film capacitors usually exhibited an increased third harmonic, compared to this wound type. My initial stocks were too small to be statistically valid, so more capacitors were purchased.

Wound v Stacked metallised PET

At this time I measured distortion using only a single pure 1kHz tone and no DC bias. With 4 volts dropped across the capacitor, my equipment noise floor was below -140dB. Loaded with a 0.5% metal film resistor, distortion measured 0.00005%. Similarly the best capacitors typically measured 0.00006%, with second harmonic better than -125dB, third and higher harmonics better than -130dB.

Measuring 25 type 470 capacitors I found three having more than ten times higher distortion. Even harmonics were little changed, but third harmonic increased to -100dB, fifth to -115dB. Measuring another 25 capacitors I found two with high distortion. I set a good/bad limit at -120dB, any harmonic exceeding this level being viewed as bad.

The next step was measuring 25 stacked capacitors and I found most measured as bad. Distortions varied from 0.00034% to 0.0018% and many displayed -90dB third harmonics.

Was this difference genuine or was my sample still not statistically significant? Measuring more capacitors, I found some also having increased second harmonic distortions. I had anticipated finding third harmonic variations but did not understand these second harmonic problems,

PET of course has significant

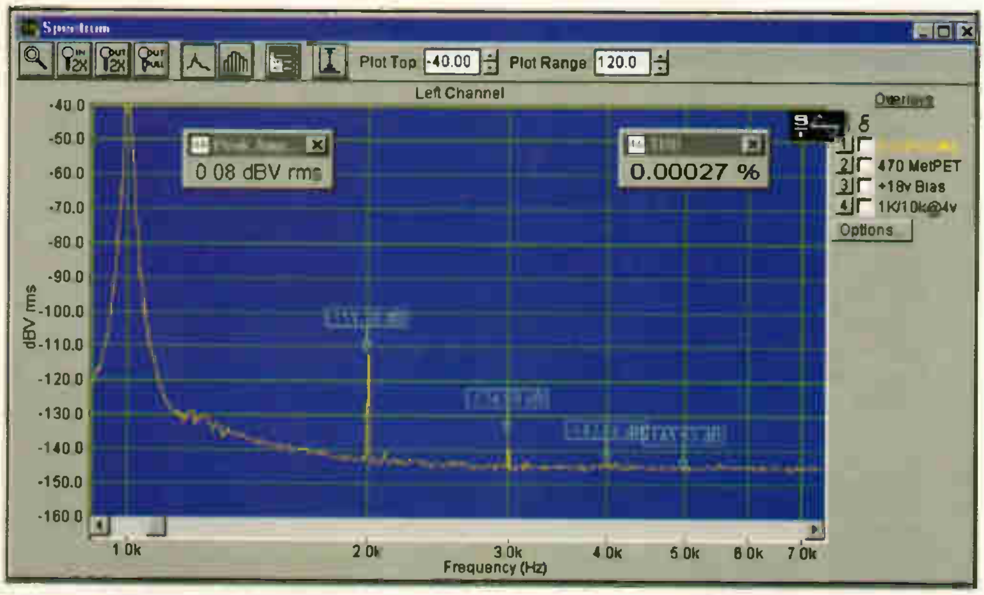

***Figure 4:** With no bias, this exceptionally good 0.1µF 63 volt type 470 metallised PET capacitor from BC Components, with magnetic leadwires, measured just 0.00004% distortion. With 18 volts DC bias, the second harmonic increased 22dB from -133.3dB to -111.4dB (distortion increased six times.) Intermodulation products are just visible, either side of 2kHz.*

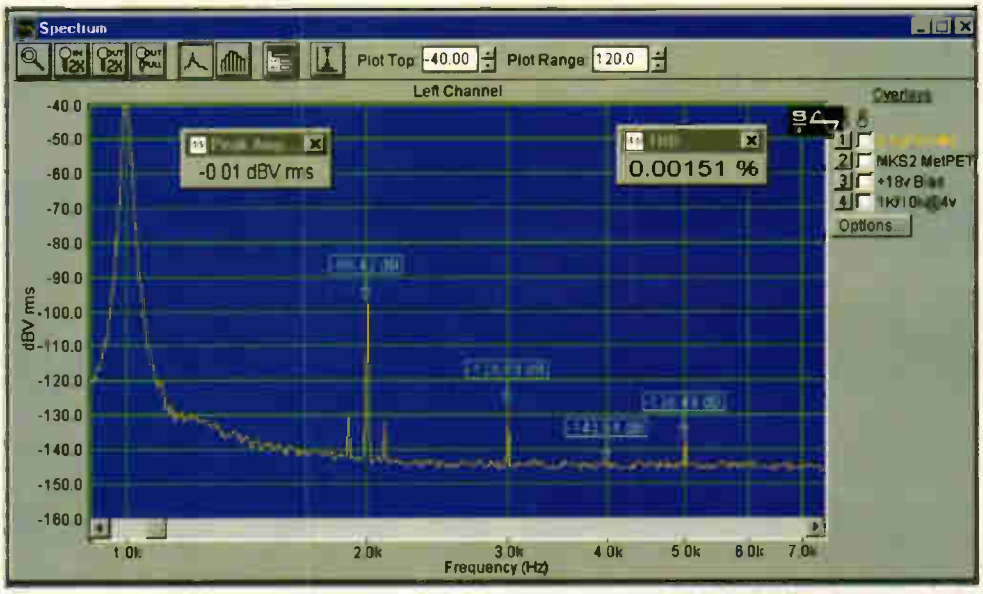

Figure 5: Distortion measurement of a typical MKS2, with no DC bias measured just 0.00007%. With 18 volts DC bias the second harmonic increased 31dB from -128.3dB to -96.4dB. Intermodulation products and other harmonic levels did not change.

dielectric absorption, typically 0.5%.[6] Several capacitors, pre-selected as good and very bad distortion, were accurately measured for capacitance and tanδ at 1kHz using my precision bridge, initially unbiased then with 30 volts DC bias. The biggest capacitance change found was less than 0.01% and with tanδ values remaining constant regardless of bias voltage, seemed to rule out any dielectric absorption effects.

Somewhat puzzled, I decided to expand my distortion measurements, changing the measurement stimulus in small steps and varying one test parameter only at a time. I would also look for intermodulation using two test frequencies and explore the affects of change of DC bias voltage. I had no choice but start again, repeating almost 1000 single frequency distortion measurements already saved to disk, both of film and electrolytic capacitors.

Revised measurements

To prove my DC bias buffer contributed no distortion, I measured my near perfect 1μF KP capacitor. Using 6 volt test signals at 100Hz and 1kHz and 50 volts DC bias, its distortion measured 0.00006%. This DC bias buffer was then used for all measurements. **Fig. 3.**

A 'good' 0.1μF 63v type 470 wound capacitor, tanδ 0.00337, measured similar distortion when tested with no DC bias. Intermodulation was just visible either side of the second harmonic. With 18 volt DC bias, second harmonic increased by 22dB and distortion to 0.00027%. **Fig. 4.**

A batch of Wima MKS2 wound capacitors consistently show increased intermodulation products and third and fifth harmonics. Typical no bias distortions measured around 0.0001%. With 18 volt DC bias the

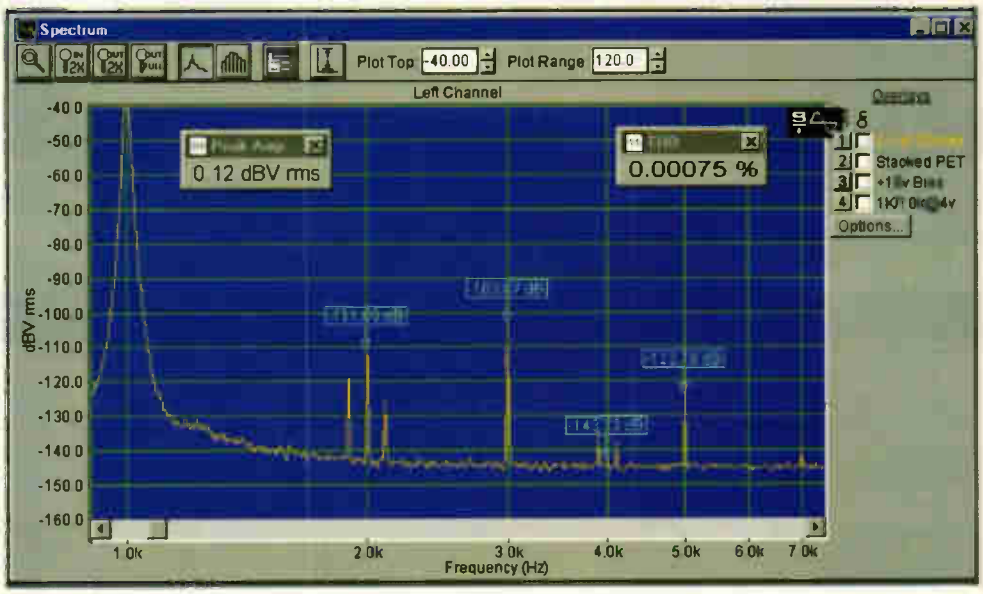

Figure 6: A 0.1μF 100 volt stacked metallised PET, with magnetic leadwires, displays increased odd harmonics and intermodulation components. The second harmonic of this much larger capacitor made with thicker PET, increased less with DC bias, compared to figures 4 and 5.

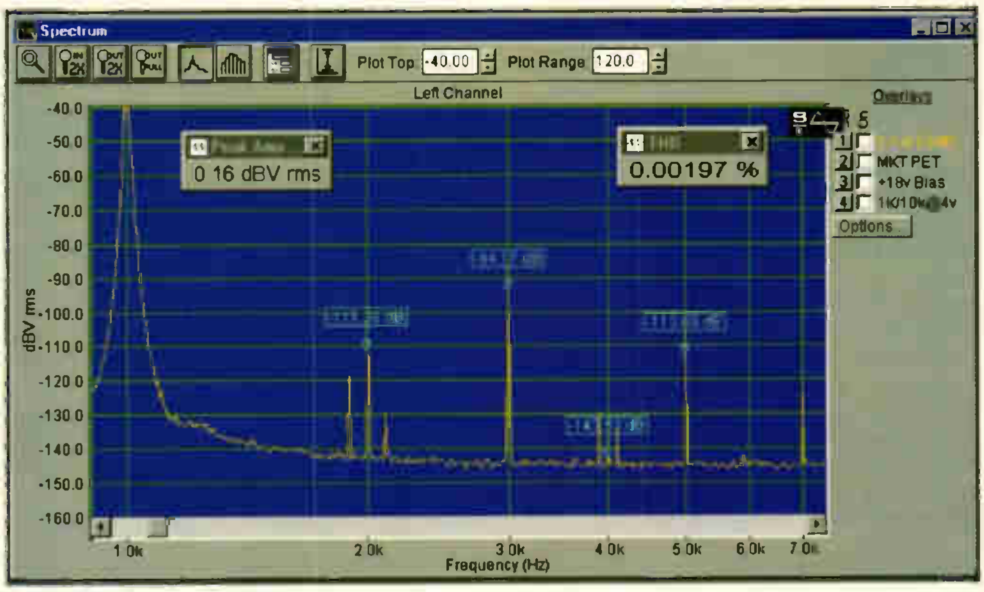

Figure 7: A different maker's very much smaller, 63 volt stacked metallised PET capacitor with copper lead wires exhibits worse distortions than those shown in figure 6. Notice however a family likeness of distortion components, similar to figure 6 but quite different from figure 4.

second harmonic increased by 32dB and distortion measured 0.00151%. With a tanδ of 0.00272, this capacitor was dismantled to confirm it was wound construction. **Fig. 5**

A much bigger, 100 volt rated, uncased stacked capacitor with tanδ 0.00352, shows a very high third harmonic level and increased intermodulation products, typical of the construction. Made using thicker dielectric, its second harmonic increased by 16dB when biased with 18 volts. Due to its third harmonic, high distortions were measured with and without bias. **Fig. 6**

Third and odd harmonics vary with AC test signal, but DC bias from 0 volts to 30 volts, has almost no affect. These enormous changes in second harmonic, tested with and without bias, clearly result from the DC bias, dielectric thickness and dielectric absorption. (see box Dielectric Absorption.)

Uncertain of their construction, I ordered just ten MKT capacitors (Farnell 814-192) and all behaved similarly. Exceptionally high distortion with and without bias, dominated by the near -90dB third and -113dB fifth harmonics. With tanδ 0.00371, this capacitor was dismantled to confirm it was stacked construction. **Fig. 7**

With such large variations in harmonic distortion, it seemed all small metallised PET capacitors should be distortion tested, to avoid building obviously 'bad' capacitors into the signal paths of audio equipment

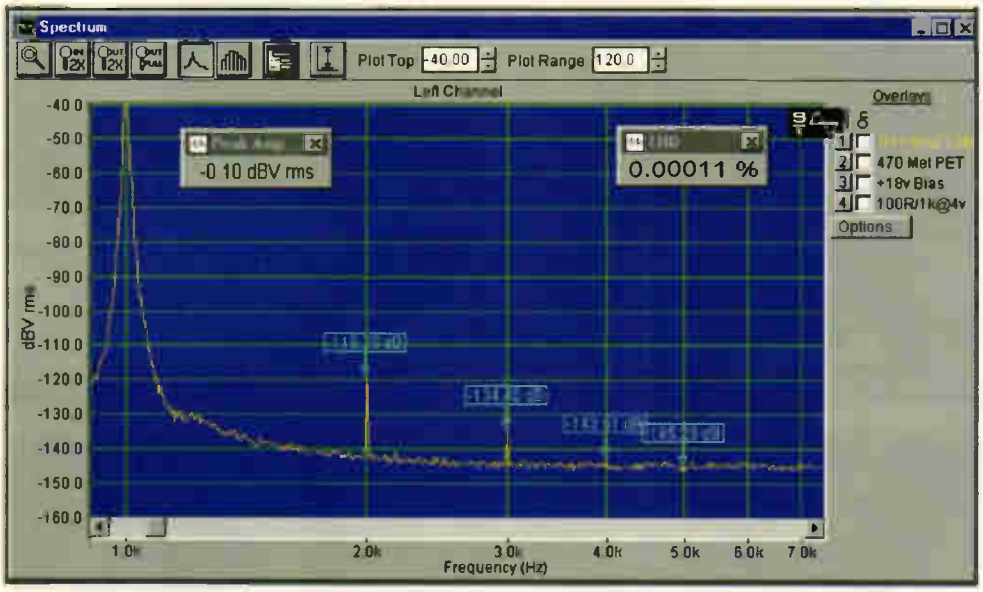

Figure 8: The first of two plots which explore the effect an increase in metallised PET film thickness might have on distortions. With no bias, distortion of this 100 volt capacitor measured 0.00006%, second harmonic -126.2dB. With DC bias, second harmonic increased by 7dB and distortion to 0.00011%

Dielectric Absorption

Two major dielectric characteristics exist, polar and non-polar. By polar I am not referring to an electrolytic capacitor, but to the way the dielectric responds when subject to voltage stress. This stress relates to the voltage gradient across the dielectric and not just the applied voltage. In other words it is stress in volts per micron, which matters.

Non-polar dielectrics, for example vacuum and air, are little affected by voltage stress. Solid dielectrics which behave in a similar fashion are termed 'non-polar'. Most solid dielectrics and insulators however are affected to some extent, increasing roughly in line with their dielectric constant or 'k' value. This 'k' value is the increase in measured capacitance when the chosen dielectric is used to replace a vacuum or more usually, air.

When a dielectric is subject to voltage stress, electrons are attracted towards the positive electrode. The electron spin orbits become distorted creating mechanical stress and a so-called 'space charge' within the dielectric. This mechanical stress produces some heat rise in the dielectric and a power loss, called dielectric loss. Non-polar dielectrics exhibit very small power or dielectric losses. Polar dielectrics are much more lossy. Having been charged to a voltage, it takes much longer for the electron spin orbits in a polar dielectric to return to their original uncharged state. Polar dielectrics produce easily measured 'dielectric absorption' effects.

Dielectric behaviour with voltage depends on the voltage gradient, in terms of volts/micron as well as on the characteristics of the dielectric. Its effects are more readily apparent with very thin dielectric. The lowest voltage, 50 and 63 volt rated metallised PET film capacitors are often made using 1 micron or thinner film. As will be seen in my next article, the dielectric used to make small low voltage electrolytics is perhaps one hundred times thinner. Consequently we should anticipate increased effects from dielectric absorption.

Foil and film capacitors cannot 'self heal' so must be made using relatively thick dielectric films. As a consequence we find that foil and film PET capacitors can provide low distortion, even when subject to DC bias voltages.

Dielectric absorption is usually measured by fully charging the capacitor for several minutes to a DC voltage, followed by a rapid discharge into a low value resistor for a few seconds. The capacitor is then left to rest for some time after which any 'recovered' DC voltage is measured. The ratio of recovered voltage to charge voltage is called dielectric absorption.

So how might dielectric absorption affect the distortion produced by a capacitor? Many fanciful descriptions can be found, describing smearing, time delays and compression. My AC capacitance and distortion measurements, simply do not support these claims. The main characteristic I have found, which clearly relates to dielectric absorption, is the magnitude of the second harmonic. This does increase with applied AC or DC voltage stress and especially so with thin materials, having known higher dielectric absorption. For example, the PET (Polyethylene Terephthalate)and PEN (Polyethylene Naphthalate) dielectric films have almost identical characteristics except for dielectric absorption. Comparative distortion measurements with and without DC bias, made on metallised PEN and metallised PET capacitors, show that PEN capacitors do produce much larger second harmonics. The PEN material at 1.2%, has almost three times greater dielectric absorption than PET[6]

The 1μF problem

To approach our idealised capacitor we need the small size provided by metallised PET, the low distortions found using Polypropylene and low cost. These qualities could be approached using metallised Polycarbonate, but Polycarbonate capacitors have become extremely expensive. With the production of Bayer Makrofol Polycarbonate film having ceased, metallised Polycarbonate capacitors may disappear.

A great many 0.1μF metallised PET capacitors had been measured, without finding clear reasons for their widely differing distortions. Would measurements at 1μF help?

I decided to measure the same make and style, rated at both 63 volt and 100 volt, to explore the D. Self comment that 63 volt capacitors exhibit ten times more distortion than 100 volt.[7] Provided the maximum capacitance possible at these voltages in both case sizes is obtained, dielectric absorption effects related in volts per micron to the differing film thickness used should be observed. It seemed probable that the 63 volt capacitor would exhibit increased second harmonic compared to the 100 volt version.

I choose to measure the 470 style capacitors, because 0.47μF at 100 volt and 1μF at 63 volt, were the maximum capacitances available in the case size. I soldered together several pairs of 0.47μF to produce near 1μF 100 volt capacitors.

Measured within a few minutes of each other, with no bias voltage, the 63 volt and 100 volt capacitors measured almost identically, with distortion at 0.00007% and 0.00006% respectively.

Re-measured with 18 volt DC bias, the third and higher harmonics were unchanged but second harmonic levels increased for both voltage ratings. Second harmonic for the 63 volt capacitors increased by +12.5dB, the 100 volt capacitors by +7dB, giving measured distortions of 0.00024% and 0.00011% respectively. **Fig. 8** These figures equate well with the expected differences in film

Technical Support

Interested readers are free to build a system for personal use or educational use in schools and colleges. Commercial users and replicators should first contact the author.

A professionally produced set of three FR4 printed circuit boards, with solder resist and legends, for the 1kHz signal generator, the output buffer amplifier/notch filter/pre-amplifier and the DC bias buffer network, comprising a 'with DC bias, single frequency, distortion test system'. Complete with component parts lists and assembly notes, the set of three boards costs £32.50.

Post/packing to UK address £2.50. Post/packing to EU address £3.50, rest of world £5.50.

As a service to Non-UK readers, but only if ordered together with the above PCBs, I can now supply one four gang potentiometer with each set of boards, re-tinned and tested, for an additional £5.00 inclusive of postage.

Falcon Electronics (EW September) has these potentiometers in stock.

Postal Orders or Cheques, for pounds sterling only, to C. Bateman. 'Nimrod' New Road. ACLE. Norfolk NR13 3BD England.

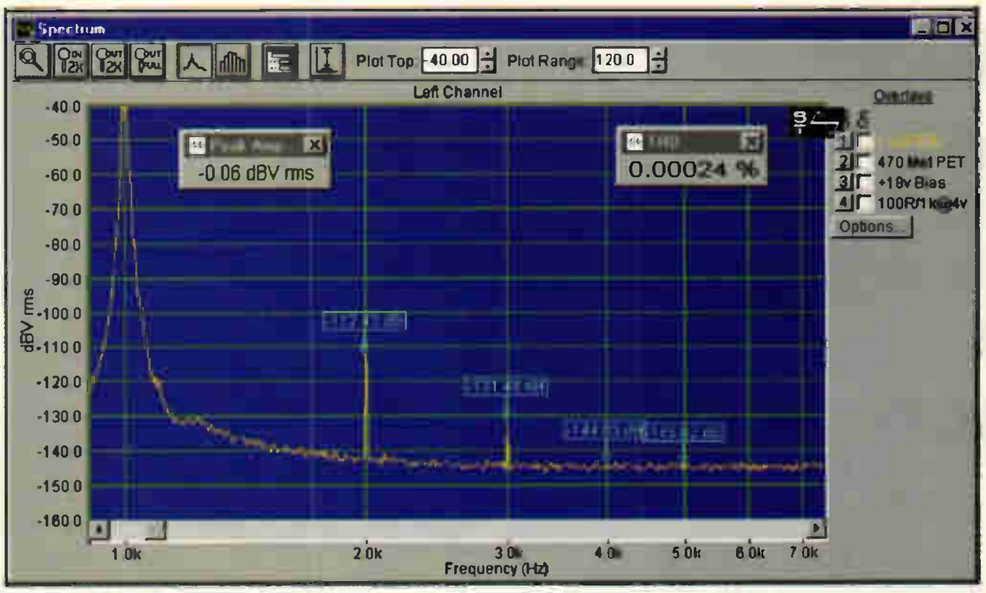

Figure 9: Distortion of the 63 volt capacitor, same make comparison with figure 8. With no bias, distortion measured just 0.00007% with second harmonic at -124.9dB. With DC bias, second harmonic increased by 12.5dB. At 0.00024%, distortion is double that of the 100 volt capacitors.

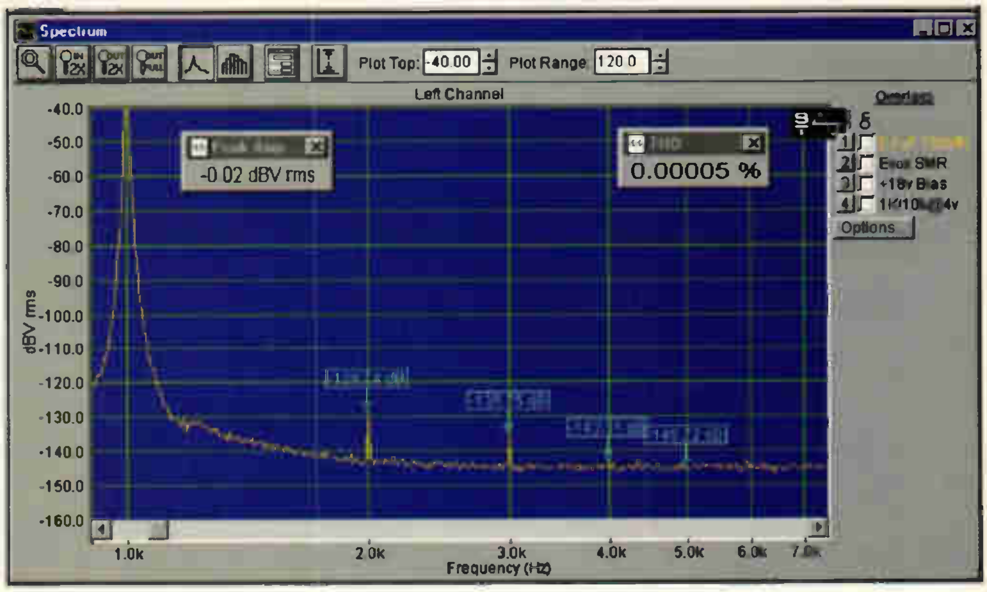

Figure 10: This 0.1μF 100 volt Evox Rifa SMR capacitor, provides a superb low distortion performance with or without DC bias, even as high as 30 volts and no intermodulation products were seen.

thickness and confirmed the effect dielectric absorption has on second harmonic distortion. (see box Dielectric Absorption.) Some factor other than rated voltage, must account for Douglas's reported observation. **Fig. 9**

Further measurements on 1μF metallised PET capacitors, using 25 pieces of the wound type 470, and a similar quantity of stacked film capacitors, revealed nothing new. Distortion patterns established by the smaller capacitors were being repeated. I also had 10 pieces of wound capacitors type 370, dated 1995. These produced harmonic levels with and without bias remarkably similar to those measured on the MKS2 types.

Possible mechanisms

These tests clearly illustrate how audible problems can exist using metallised PET capacitors in low distortion audio. I now sympathise with listeners who complain about amplifier sounds, when using metallised PET capacitors.

Lacking the facility to assemble test capacitors using known differences in materials and processes, I can only speculate as to possible reasons for the different third harmonic distortion levels I found. These may result from differences in manufacture of the basic film or the vacuum deposition of the metallised electrodes, processes that vary from maker to maker.

It might even be as simple as the electrode metallisation thickness used. Perhaps thickness gives the wrong impression, this aluminium coating is so thin, like mirror sunglass lenses, it is quite transparent. Its thickness is measured in Ohms/square, typically some 2 to 4 Ohms.

One convenient explanation for these differences might be the use of copper versus magnetic leadwires. Not so, the lowest distortion, type 470 metallised PET capacitors tested use magnetic leads whilst the worst distortion stacked types used copper.

More likely are differences in the metal compositions and spray application methods used, to produce the 'schoop' end connections.[5] Aluminium metallised electrode has an electro-chemical potential of +1.66 volt, magnetic leads +0.44 volt, copper wires -0.337 volt. For the 'schoop' connection, a variety of other metals are used, having intermediate, mostly positive potentials. Possible 'Seebeck' effects should not be ignored. (see box Metallised film dielectrics.)

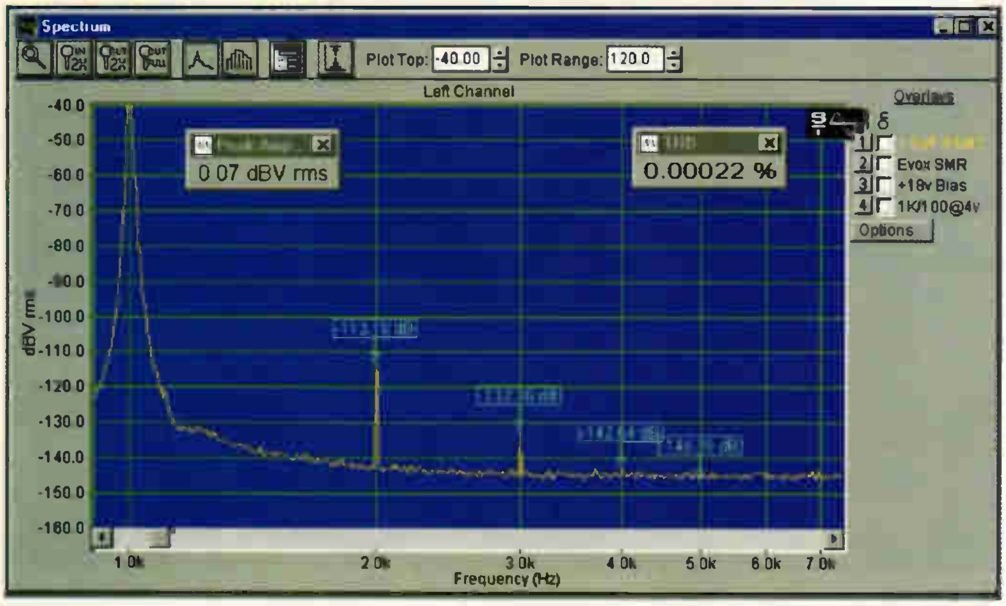

Figure 11: Little larger than their 0.1μF, this 1.0μF 63 volt SMR capacitor from Evox Rifa behaves impeccably with or without DC bias up to 10 volts. With increased DC bias its second harmonic does increase but no intermodulation products were seen.

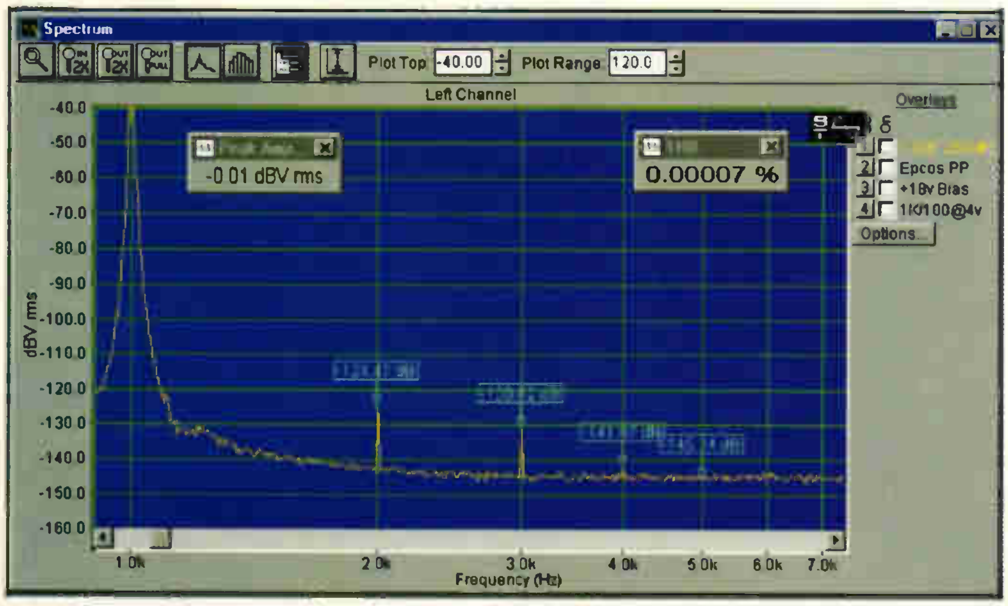

Figure 12: If you have room for a capacitor with 22mm lead spacing, this 1.0μF 250 volt Epcos with DC bias voltage, distorts less than most capacitors with no bias.

Intermodulation distortion

From many measurements using AC voltages from 0.5 to 6 volts, intermodulation products are produced in metallised PET capacitors according to the level of third harmonic the capacitor produces. For example a 'bad' capacitor exhibits intermodulation when subject to much less than 1 volt AC. A capacitor developing smaller third harmonic, shows no visible intermodulation until its AC voltage exceeds 3 volts. **Fig. 4**

The best metallised PET capacitors produced almost no distortion with no DC bias, but when used to block DC, second harmonic distortion increased rapidly with increasing DC bias voltage. Depending on circuit arrangements, many capacitors could produce audible distortions. Perhaps this should not surprise us. Audiophiles have claimed to be able to 'hear' PET capacitors for many years.

I believe that for 0.1μF to 1μF values, metallised PET capacitors should first be distortion tested. Because of their rapid increase in second harmonic with DC bias, they should not be used with significant DC bias, relative to their rated voltage, in high quality audio equipment. Having so far failed to find a physically small, economic, low distortion solution, is one possible?

Polyphenylene Sulphide

A much better but little used, slightly more expensive dielectric has been available for many years.[8] It is available metallised down to 1.2 microns and with a 'k' of 3, it provides capacitors slightly larger than metallised PET.[6] It has many other benefits. Usable to 125°C, it provides a near flat temperature coefficient and tanδ slightly higher than metallised Polypropylene. It has a small dielectric absorption of 0.05%, better than Polycarbonate and ten times better than PET.

Like Polycarbonate, Polystyrene and C0G ceramic, it provides superb long-term capacitance stability, changing only 0.3% maximum in 2 years. It seems Polyphenylene Sulphide (PPS) should provide acceptable size, low distortion capacitors.

I used 0.1μF 50 volt, 5mm centres Evox Rifa SMR metallised PPS capacitors, in my tanδ meter assemblies. Measurements of 25 pieces displayed extremely low distortion. This stock was purchased from RS, who has dropped the product from its catalogues, so I sought another stockist. The Farnell web site recently listed a small selection of Evox Rifa Polyphenylene Sulphide capacitors. Maximum stock value in 5mm lead spacing is 10nF, with up to 1μF at 63 volt in 10mm centres and at 100 volt in 15mm. The largest value, 3.3μF at 63 volt, has 15mm centres.

The 0.1μF 100 volt SMR produced superb results with and without DC bias voltage. **Fig. 10** The 1μF 63 volt produced superb results if biased to less than 10 volts but with increasing bias, second harmonic distortion increases. The larger 1μF 100 volt should be less sensitive. **Fig. 11** Both SMR types tested have small case size and 10mm lead spacing.

Bigger the better?

Another new Farnell line is Polypropylene capacitors from Epcos (Siemens). The second harmonic of the 1μF 5% 250v B32653, with 22mm centres, changes little with DC bias up to 30 volts, distortion is then 0.00008%, a superb performance. **Fig. 12** The 0.1μF 5% 400v 15mm centres B32652, measured 0.00005% with 30 volts DC bias. **Fig. 13** Distortions from these 0.1μF and 1μF Epcos capacitors were not bettered by any similar sized capacitor I tested. With double the PCB footprint of the SMR types they may not fit your space. No doubt these new lines will appear in the Farnell catalogue.

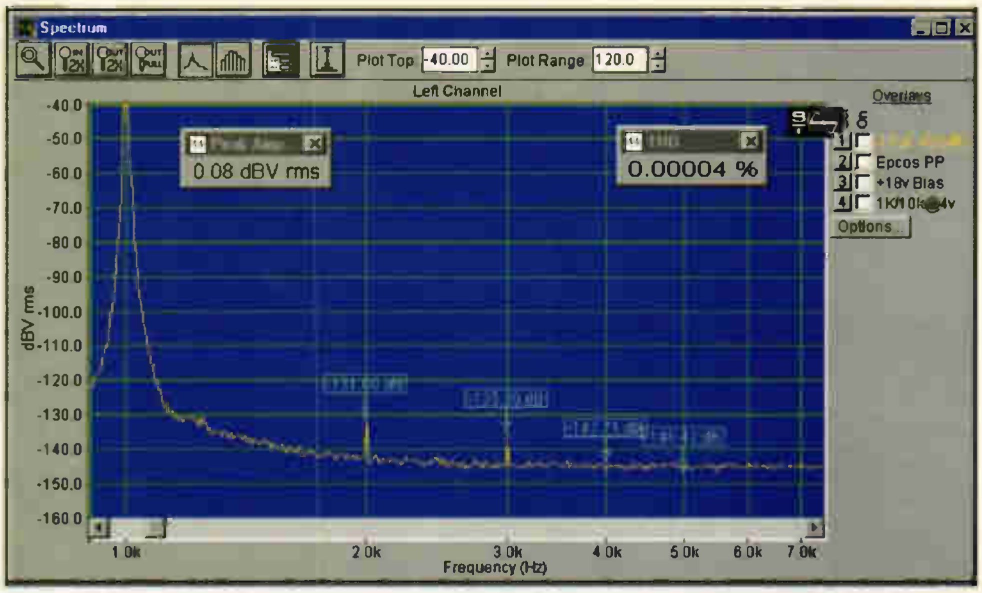

Figure 13: As good as Polystyrene? Distortions from this Epcos 15mm lead spacing 0.1μF 400 volt capacitor barely change even with 30 volts DC bias.

Maintaining designed performance

Having measured several hundred metallised PET capacitors, I found many with extremely low distortions if measured without DC bias. I also found far too many showing very bad distortions, both DC biased and unbiased, yet metallised PET capacitors continue to be used in the signal paths of high quality audio amplifier designs.

To ensure the claimed performance of a published audio circuit can be repeated, the designer should declare the make, model and rated voltages of the capacitors. Simply stating ceramic, film etc. is totally unacceptable. These tests illustrate how a capacitor with an acceptable single frequency distortion test, can produce significant intermodulation on audio when presented with multiple frequencies.

Many years ago Ivor Brown presented the case that amplifier tests should comprise three test signals. This seems to have been completely ignored, at least in EW amplifier design articles.[9] Single tone 1kHz amplifier harmonic distortion tests ignore distortions caused by the rising impedance of capacitors at low frequencies. It is now clear that large amplitude bass notes and drum beats in music can result in peculiar intermodulation distortions, in an otherwise apparently good amplifier.

For my part I shall disregard any published audio designs which do not report low frequency intermodulation distortion claims or low frequency harmonic distortion results, especially if the capacitors used are not properly chosen and adequately defined.

In my next article we introduce that most complex of capacitors, the electrolytic, then explore which produces the least distortion at 1μF, a metallised film or an electrolytic capacitor. ■

Some of the illustrations in last month's article were accidentally reduced too much, resulting in some of the text bordering on being illegible. Those diagrams are repeated over the next pages in larger form. Apologies.

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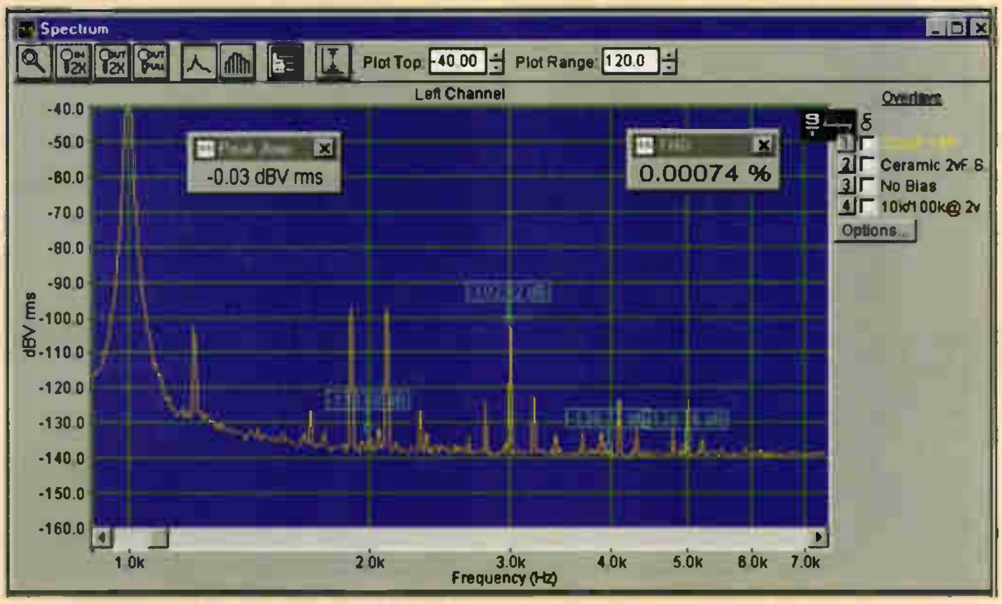

Figure 1: Y5P is a medium 'k' class 2 ceramic. Tested with two signals, 100Hz and 1kHz at 2 volts amplitude, with no bias network, it produces many new intermodulation distortion frequencies.

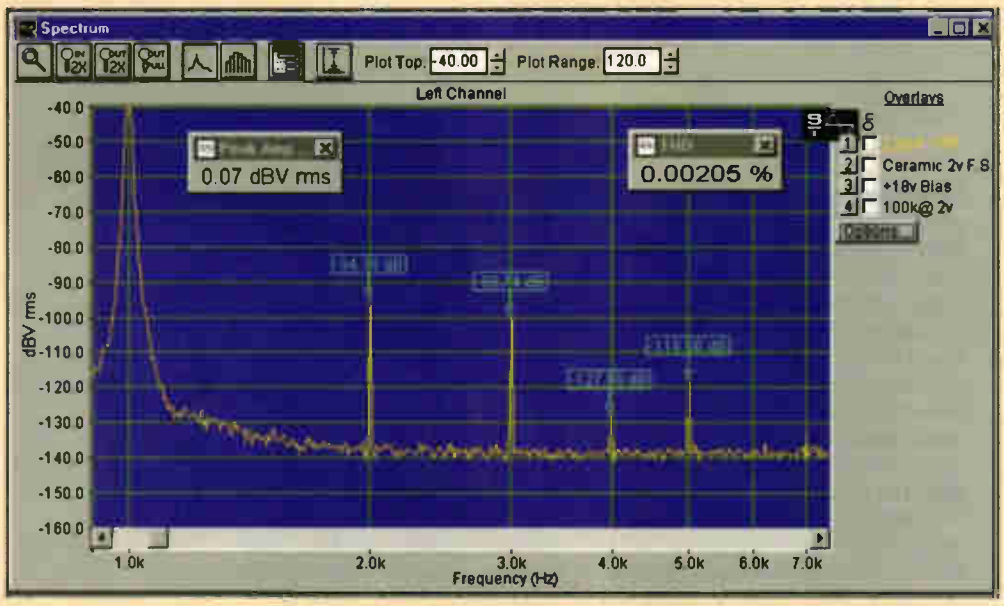

Figure 2: The figure 1 capacitor tested using 1kHz only with 18 volt DC bias. Compared to its 0 volt bias test, second harmonic has increased 23dB, a 14 times distortion increase.

Figure 5: Distortion measurement of a Class 1 ceramic using 100Hz and 1kHz signals at 4 volts and 18 volt DC bias. With no bias this tiny 10nF 50 volt C0G multilayer capacitor measured just 0.00006%. Second harmonic was -128.5dB, the other levels remained as shown.

Figure 6: A Class 2 X7R 10nF capacitor from the same maker as figure 5 and tested the same. This test dramatically shows the impact an increase in both tanδ and dielectric absorption have on capacitor distortions.

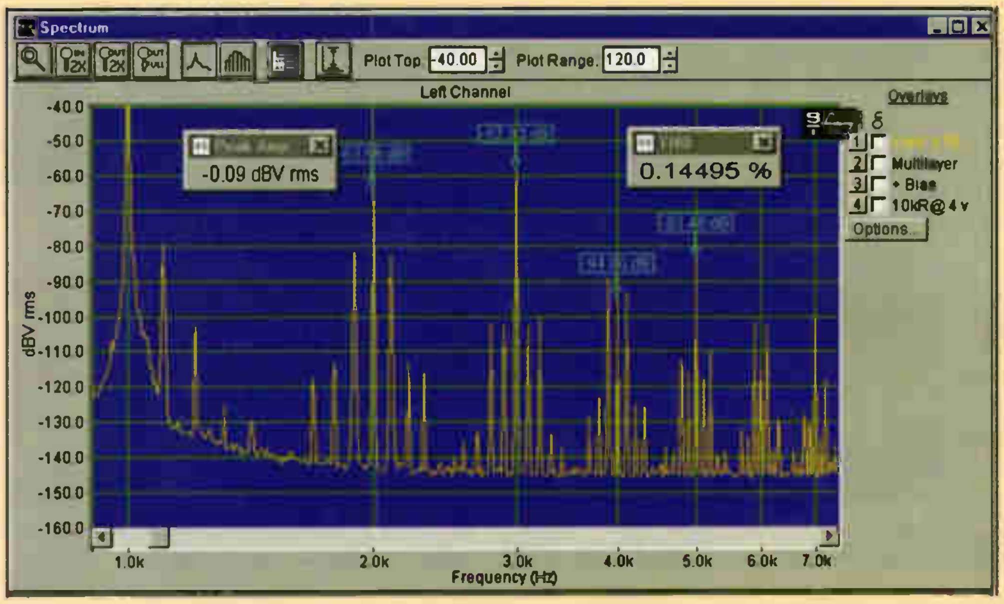

Figure 7: This now discontinued Philips extended foil/Polystyrene 1% axial lead capacitor, with 4 volt signals and 18 volt DC bias, shows negligible distortion. With test signals increased to 6 volt and DC bias to 30 volt second harmonic increased less than 4dB and distortion to 0.00007%. There was no visible intermodulation.

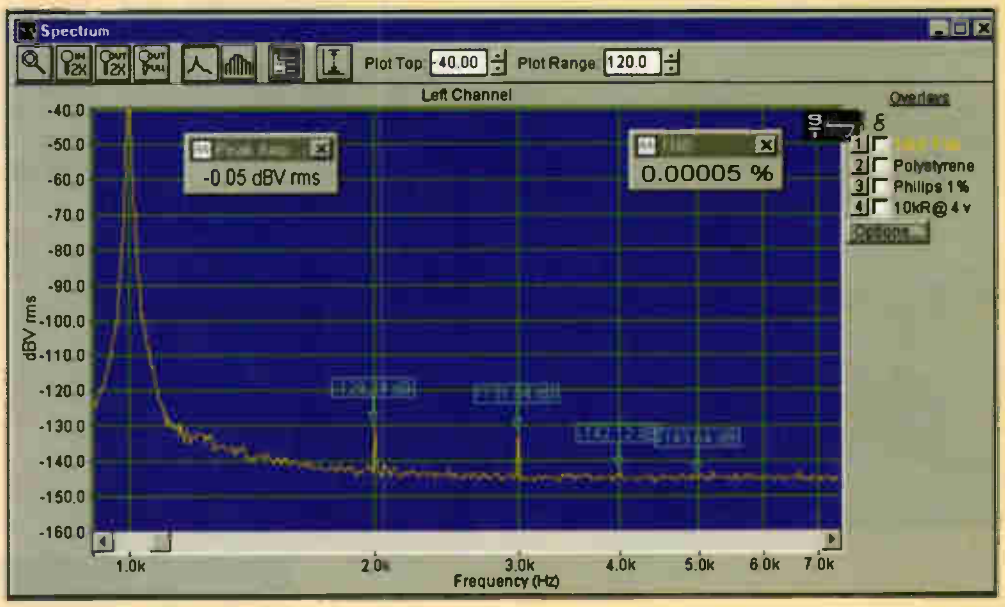

Figure 8: The makers replacement extended foil/Polypropylene shows the same 0.00005% distortion but second harmonic is 1dB worse. With test signals increased to 6 volts and DC bias to 30 volts second harmonic increased just over 5dB, distortion to 0.00008%. Again, no visible intermodulation.

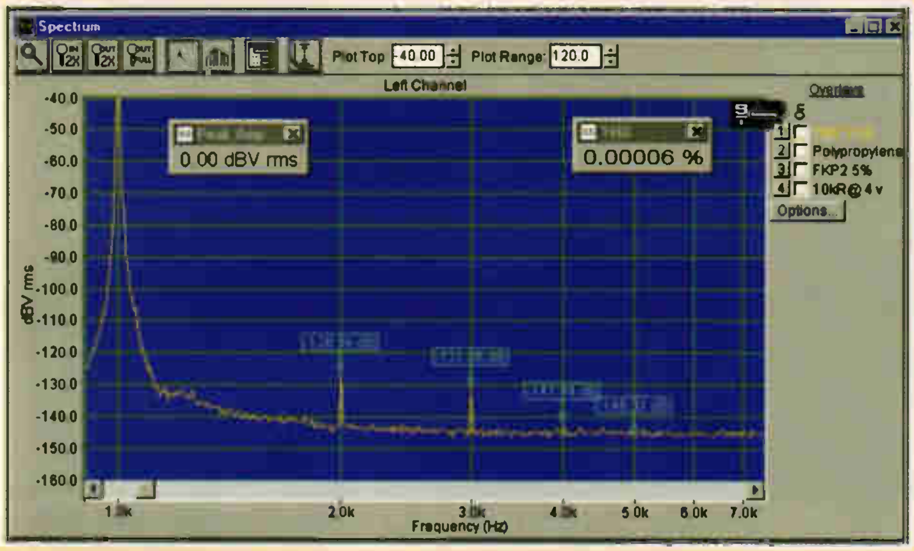

Figure 9: The small Wima FKP2 foil/Polypropylene capacitor shows similar performance except for 2dB increased second harmonic. Distortion just 0.00008% with 6 volts stimulus and 30 volts DC bias.

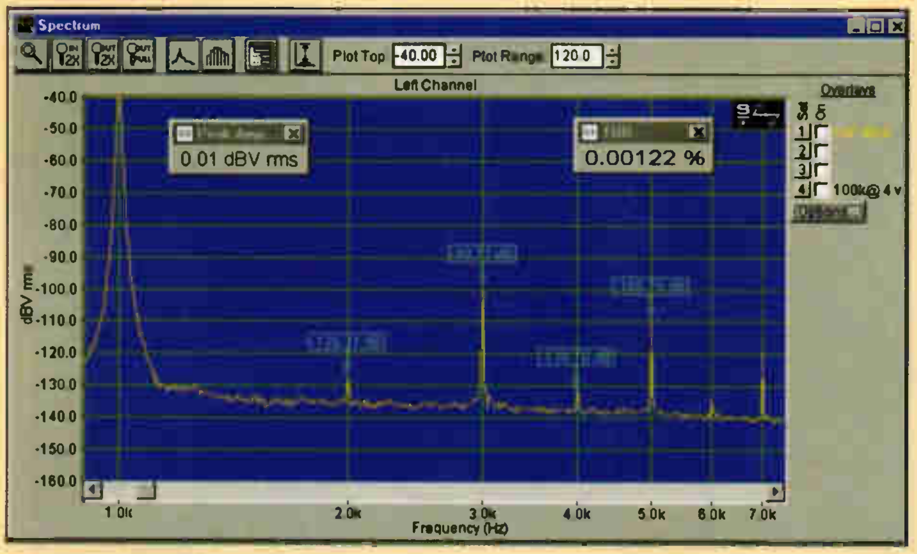

Figure 10: Despite cleaning and re-tinning its oxidised lead out wires, this 1nF Mica capacitor, tested using 1kHz only at 4 volts and no bias, clearly has an internal non-ohmic connection problem.

Figure 11: Tested with no bias, this 0.1µF MKS2 metallised PET capacitor measured 0.00016% with clearly visible intermodulation products. With 18 volts DC bias, the second harmonic increased dramatically, from -119.0dB to -92.9dB and harmonic distortion to 0.00225%.

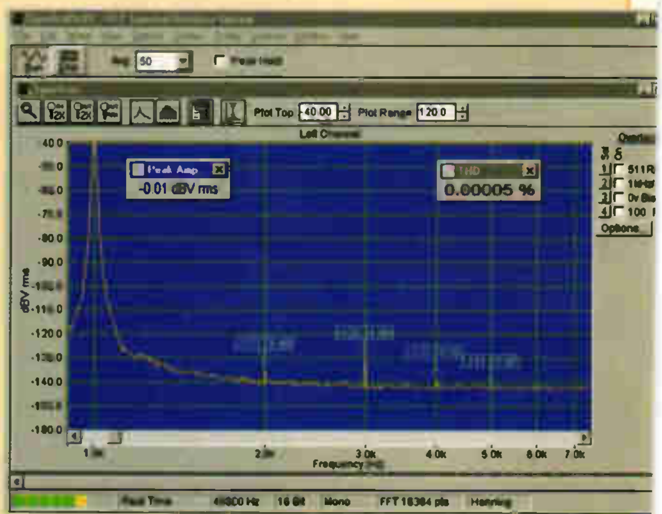

Figure 12: The Plus232 software shows a green then yellow signal strength meter, bottom left, changing dramatically to red at the soundcard overload level. My 'standard' measurement settings can be seen. Loaded with a 511Ω resistor, all harmonics are well below 0.5 ppm distortion.

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Part 1

RF Auto-transformers - Transmission Line Devices modelled using SPICE

Nic Hamilton (G4TXG) proposes an improved model for the small-signal RF auto-transformer with a ferrite core and illustrates it by building a SPICE model for a typical small-signal RF transformer.

The usual model for RF transformers is the same as the model used for low frequencies. Although the model is fine for 50 and 60Hz, it provides inaccurate results at RF and offers no enlightenment on the way the transformer works. I'll start by putting the transmission line into the transmission line transformer and modelling the turns.

I have a PC based on a P2 233MHz processor and this is now slow by modern standards, yet it can run a SPICE simulation of the frequency response of a circuit with hundreds of components in a reasonable time. I am also interested in RF. There are plenty of models around for surface mounted transistors and most passive components, but many circuits use transformers, and when I take a twin-hole 'Balun' core (also known as pig-nose, dual-aperture, double-aperture or binocular) and thread through a few turns of wire, how do I model it, and how do I use the model to optimise the transformer design? This is not just an idle curiosity: RF transformers appear in baluns, hybrid couplers, mixers, low-noise amplifiers, wide-band data circuits and switched-mode power supplies. Measurements show that Balun cores give a wider bandwidth than similar transformers wound on toroids and that 'monofilar' RF auto-transformers give a wider bandwidth than similar transformers using multi-filar windings with defined impedances[1].

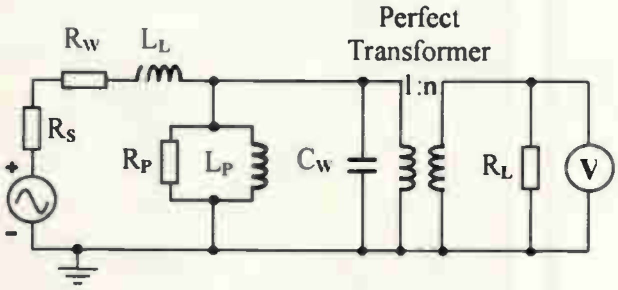

Fig. 1. Lumped equivalent circuit of an impedance matching transformer

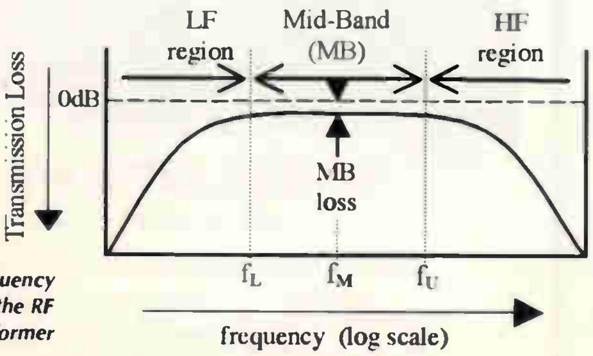

Fig. 2. Frequency response of the RF transformer

Why use an auto-transformer?

RF auto-transformers are usually used as impedance matching devices. Impedance matching can also be achieved using networks containing capacitors and inductors, transmission lines, resistors, or a mixture of these. The transformer's main advantage is that it works over a wider bandwidth with less loss than the other options. Impedance matching has two effects. It minimises the transmission loss (can give insertion gain), and it maximises the return loss (minimises VSWR).

The transmission loss is usually the more important of the two, but consider the following illustration. Connecting 50Ω coaxial cable to 75Ω cable in an otherwise matched system will cause a 1.5:1 VSWR and a transmission loss. Ask most non-specialist engineers in the street for an estimate of this loss and the most frequent answer is 3dB, but in fact the loss is only 0.177dB. Therefore, using a matching transformer in this circuit could give an insertion gain of 0.177dB. However, the loss of the transformer is likely to be 0.2dB anyway, so there would be little point in using a transformer to minimise insertion loss.

But, if the 50Ω to 75Ω cable join occurs in a long run that connects to a TV, and if the source and load impedances are not perfectly matched to the cables, ghosting could occur on the picture due to the signal reflected from the join. A pair of resistors in an L configuration could be used to achieve a perfect broadband match, but the additional loss of the resistor network would be almost 6dB. In this case, the use of an auto-transformer would be justified.

What is the 'usual' auto-transformer model?

Figure 1 shows the lumped element equivalent circuit of an impedance matching transformer. In this model, the source impedance R_S and load impedance R_L are matched by a perfect transformer with an appropriate turns ratio. Imperfections associated with this transformer are present in both the primary and the secondary windings, but for convenience, are collected on the primary side of the transformer. L_P is the inductance of the primary winding, and R_P is the associated core loss. L_L is the leakage inductance representing the imperfect coupling of the primary and secondary windings, and R_W is the winding resistance. C_W is the winding capacitance.

Figure 2 shows the frequency response that results from this model. The L_F droop below f_L is due to the low reactance of L_P compared with R_S, the MB attenuation is due to R_P or R_W, and the HF droop above f_U is attributed to the high reactance of L_L compared to R_S, or to the low reactance of C_W.

For a given frequency response, f_L and f_U are defined by the transformer's maximum allowable VSWR or insertion loss. Where insertion loss is the prime concern, flatness of the frequency response is important, and f_L and f_U may be set at the frequency where the insertion loss rises by 1, 2 or 3dB in excess of the Mid Band loss. But remember that, ultimately, most matching transformers must have an insertion gain to be useful.

What's wrong with the 'usual' model?

The primary winding inductance is represented by L_P, a single lumped inductor. An assumption inherent in all lumped elements is that the current flow into a component is equal to the current flow from it. For L_P, this is true at DC, and is a reasonable approximation up to f_M. But at frequencies higher than f_M, the electrical length of the winding is no longer small compared to a wavelength, and the current flow in the primary winding is no longer uniform in magnitude and phase along the winding's length. This defect, with others, gives the 'usual' model several shortcomings:

a) The values of the components of the model change with frequency. For example, the winding resistance depends on the skin effect, which is a function of frequency.

b) Above f_U, the model's HF attenuation increases continuously as frequency is increased. Real transformers have additional HF pass-bands with a transmission loss only slightly greater than the MB loss.

c) The 'usual' model's phase response is hopelessly wrong at the high end of the MF region. This can be very important, for example, in modelling amplifiers with transformer feedback.

d) f_U depends on L_L, the leakage inductance. This element, which is so vital to the transformer's performance, is not predictable for the RF transformer.

e) Impedance matching transformers tend to suffer from increased VSWR near f_U. A recognised way of curing this is by placing a low value capacitor in parallel with the transformer's low impedance winding. The 'usual' model predicts that a lower value capacitor could be equally well connected in parallel with the high impedance winding. In fact, placing a capacitor there degrades the transformer's performance.

In spite of this, the 'usual' model does have several useful features, which will be retained in the development of the SPICE model.

Transmission lines: the heart of a better transformer model?

The 'usual' model fails because it uses a lumped inductor to simulate a winding that is not electrically short compared to a wavelength. The way forward must be to use transmission line theory. Consider an RF transformer with a 2:1 turns ratio consisting of just two turns wound on a large twin-hole core. The winding of this transformer will be two wires lying side by side. The obvious deduction is that a useful starting point for a model would be a parallel wire transmission line of length equal to one turn.

Similarly, a transformer with three turns could be based on three transmission-lines, each representing the characteristic impedance between each possible pair of turns. A four-turn transformer would have six transmission lines, and an n-turn transformer would have n(n-1)/2 transmission lines. So a model of a transformer with 8 turns could have 28 transmission lines. I tried this model[2]. It was relatively successful, but as the number of turns is increased, the number of transmission lines in the model gets rapidly out of hand. A transformer with 12 turns would have 66 transmission lines. And the 'characteristic impedances' of these transmission lines will be far from equal, and impossible to calculate if the windings are not laid perfectly.

The basis of the new model

Instead of modelling each transmission line between each turn, the model I now suggest uses a transmission line to represent the characteristic impedance Z_0 between one turn and the generality of the other turns. These other turns could be considered to be similar in effect to the outer conductor of a coaxial cable as shown in **Fig. 3**.

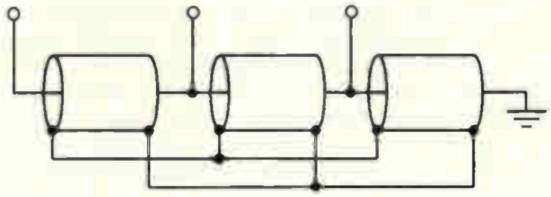

Fig. 3. Transmission line model of a three turn transformer

In this model, one transmission line represents one turn. However each turn has two coupling mechanisms. Firstly, each turn is a transmission line causing extra time delay. This forces the current in the individual turns out of phase with each other. Secondly, each turn is coupled to the rest by the traditional 'Faraday' transformer action of the core. This tends to force the phases of the currents in the individual turns together.

There are two ways of modelling transmission lines in SPICE.

a) Use the SPICE transmission line model. I used SPICE 2G.6, 3F4 and 3F5. While the latter were OK, the implementation of SPICE 2G.6 that I used had a faulty model. The transmission line could be connected to give power gain[2].

b) Use a lumped element low pass filter model. This uses inductors and capacitors of sufficiently small value that the HF cut-off frequency is above the highest frequency of interest. This gives models with hundreds of components that take too long to run.

SPICE transmission lines work as a transformer

The SPICE transmission line model simulates only the energy flowing in either direction between the conductors of the line - it does not simulate energy flow external to the line. **Fig. 4** shows an equivalent circuit of the half of the SPICE transmission line model that propagates signals from left to right. The other half of the equivalent circuit is a superimposed mirror image of the circuit that allows signals to propagate from right to left. Notice that the input is isolated from the output. This is a form of transformer action.

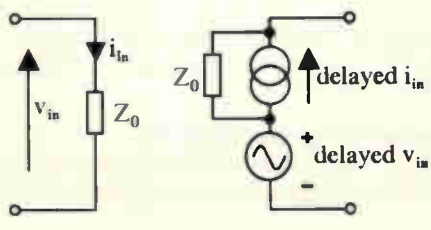

Fig. 4. The forward transmission element of the SPICE transmission line model.

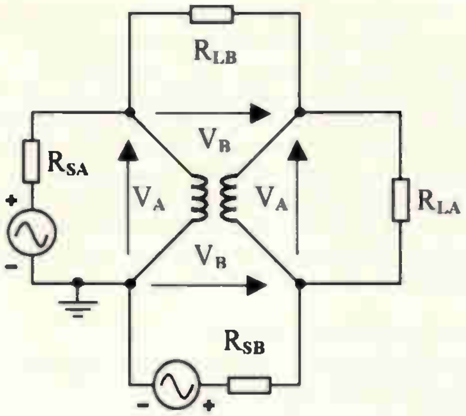

Fig. 5. The perfect transformer in the centre can be rotated by 90° leaving the RF performance of the circuit almost unchanged.

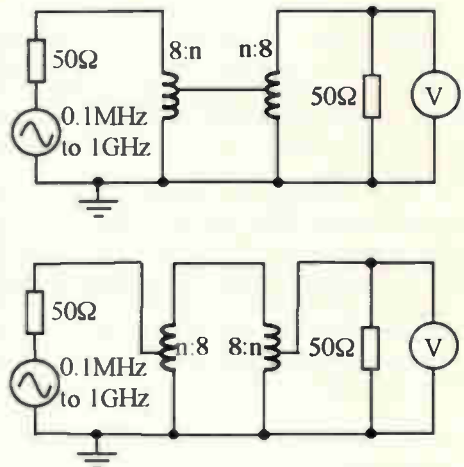

Fig.7. Measurement of Insertion Loss spectrum of two transformers connected back-to-back a) Step-down circuit with low centre impedance b) Step-up circuit with high centre impedance

A perfect 1:1 transformer is shown at the centre of **Fig. 5**. The normal 'Faraday' transformer action propagates energy from left to right. However energy will also propagate from the bottom to the top of Fig. 5 by what Sevick[3] has called the Guanella mode. A little thought shows that the transformer could be rotated by 90°, leaving the circuit's performance unchanged except in the matter of DC connections, and a slight difference in phase delay. This is a very useful technique for visualising and analysing RF transformers. Take Fig. 3, treat the transmission lines as transformers and apply a 90° twist.

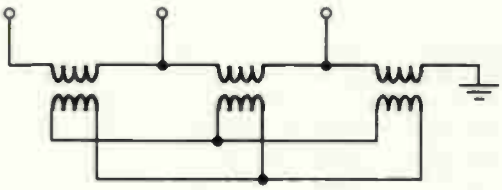

Fig.6. The three turn transmission line transformer with lines represented by 1:1 transformers with a 90° twist.

The result is **Fig. 6**, which illustrates the way the transmission line transformer works; the three 'secondary windings' are wired in parallel. This ensures that each turn or 'primary winding' is coupled to the next.

Frequency response of the transmission line model

Examine Fig. 3, and assume that there are no secondary connections, so that the whole transformer operates as a choke. At some frequency, the electrical length of the three turns added together comes to a quarter wavelength. At this frequency, the transformer will present a short-circuit. Furthermore, as this frequency is approached, the transformer becomes increasingly inductive. This presents a practical limit to f_U, unless the higher frequency pass-bands are to be exploited.

To model the turns, the winding length must be known, and this depends on the shape and size of the magnetic core. Core shapes are derived from the fundamental relationship between the current flow and the lines of magnetic flux. In the transformer, both the winding length and the magnetic path length must be minimised. This can be achieved in one of two basic ways:

a) Toroidal flux lines and a radial winding, as in the toroidal core.

b) Toroidal winding with radial flux lines, as in the pot core.

The double-aperture core can be regarded as a pot core with the sides removed for access, and it is ideal for use as a transformer core.

Description of the transformers

I took Siemens-Matsushita B62152A8X30 cores (available from RS, etc.). These double-aperture ferrite cores are small: 3.6x2.5x2.1mm, and the holes have a diameter of just 0.8mm. I threaded 8 turns on each core. The average DC resistance of the transformers' 8 turns was 0.049Ω, and by comparing this with the resistance of 1 metre of wire, I deduced that the mean winding length was 71mm. Each connection wire was about 6mm long. I wound 10 such auto-transformers, two each with a turns ratio of 8:1, 8:2, 8:3, 8:4 and an 8-turn choke with no tap point.

I connected each pair of transformers at the low impedance side, as shown in **Fig. 7a**, and measured the insertion loss spectrum. The 8:7, 8:6 and 8:5 transformers were the 8:1, 8:2 and 8:3 transformers with the end connections reversed. I halved the insertion loss of the circuit to give the transmission loss of one transformer. The results are shown in **Fig. 8a**. I repeated the process, but connecting the high impedance side as shown in **Fig. 7b** giving the results in **Fig. 8b**. These are the graphs that the model must reproduce. However, I found later that the high frequency attenuation peak in Fig. 8a was to be due to a spurious resonance of the test jig at 474MHz; it should be ignored.

Discussion of results

Note that the high frequency cut-off f_U of the transformer depends on the transformer's turns ratio. This is a characteristic of all auto-transformers, and is similar to LC matching circuits, where the greater the impedance ratio to be achieved, the narrower the bandwidth of the resulting impedance match.

The SPICE transmission line is specified by two fundamental parameters: the time delay t_d and the characteristic impedance Z_0 of the line. SPICE works in capitals, so these are entered as T_D and Z_0. However, when the transmission line under consideration is short compared to a wavelength, an odd effect occurs, and the effects of t_d and Z_0 become related as shown in **Fig. 9**. Conduct the following thought experiment to demonstrate the truth of this. Consider a section of transmission line with t_d=1ps and Z_0=5Ω in series with a 50Ω circuit. At low frequencies, the insertion loss of the 5Ω line will be negligible, but the insertion phase shift and group delay will be measurable. Because of the 10:1 VSWR at each end of the 5Ω line, the signal is reflected repeatedly between the ends, and a small amount of signal leaks from the ends at each reflection. The result is that the insertion group delay of the 5Ω 1ps line is about 5ps. Halving Z_0 to 2.5Ω halves the signal leakage, and so doubles the insertion group delay to about 10ps. So, providing the line is very short compared to a wavelength, the insertion group delay is the only important parameter. This delay can be achieved by an infinite number of different values of td and Z_0. However, of the two parameters, t_d is much the easier to estimate in a transformer, and the correct high frequency response will depend on selection of the correct value of Z_0.

Setting the delay t_d for the lines

The time delay t_d of the line is determined by its electrical length. The physical length of the 8-turn winding in each transformer is about 0.072m. Assuming that the signal travels at the speed of light, the time delay is $0.072/3x10^8 \approx 240ps$, and the delay of each turns is 240/8(30ps. However, the SPICE transmission line has an alternative method of defining the time delay. This is achieved by setting two sub-parameters: F and N_L. F is the frequency in Hz and N_L is the normalised electrical length of the transmission line with respect to wavelength in the line at frequency F. It all sounds very complex. But, by using an alternative definition where F is the propagation velocity in metres/second, then N_L is the physical length of the line in metres. As N_L is a parameter that must be entered into the model elsewhere, it is attractive to use this definition. So the problem is to find the propagation velocity. The signal travels down the transmission line at the speed of light only if the transmission line is:

a) suspended in a vacuum, and the relative permittivity $\varepsilon_r=1$.

b) has no magnetic materials nearby, and the relative permeability $\mu_r=1$.

(εr is also known as the relative dielectric constant. This is a poor name because if $\varepsilon r>1$, εr is not a constant, it is a function of frequency). If the dielectric is other than air, the signal travels at less than the speed of light by a factor equal to the square root of ε_r. The more slowly the signal travels, the greater the electrical length of the cable compared to its physical length. Take an example. A coax cable has a polythene dielectric, $\varepsilon_r=2.3$. In this cable, the electrical length will exceed the physical length by a factor of $\sqrt{2.3}=1.52$. This assumes that $\mu_r=1$, which, in the case of the RF transformer, it does not. Most of the dielectric between the conductors is air with $\varepsilon_r = 1$. The other dielectric is the thin 'enamel' coating which insulates the wire, which has $\varepsilon_r \approx 5$. Let's assume that the overall $\varepsilon_r \approx 1.44$ and $\mu_r=1$, and therefore the velocity ratio is 1.2. So $F=3x10^8/1.2=2.5x10^8$. N_L, the length of each turn, is 0.072/8=0.009 metres. So t_d, had it been entered in that format, would have been $0.009/2.5x10^8=36ps$.

Setting Z_0 for the lines

Assume for a moment that Z_0 is set purely by the geometry of the windings. There is an approximation for finding the Z_0 of a transmission line with two parallel enamelled wires in contact[4]. This indicates that Z_0 will be greater than 50Ω. There is also an approximation to find Z_0 for a transmission line consisting of a conductor in the centre of a square of four similar conductors[5] carrying an equal share of the return current. This approximation indicates that Z_0 should be less than 150Ω.

At this point, I have a confession to make. Given a

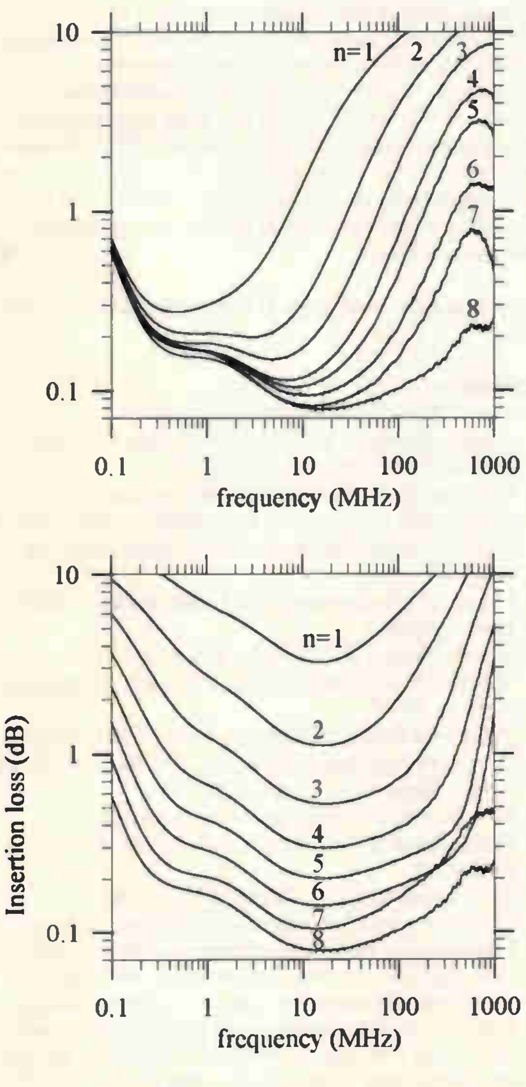

Fig.8. Insertion Loss spectrum of one transformer measured in back-to-back circuits of Fig 7
a) Loss of a step down transformer
b) Loss of a step up transformer

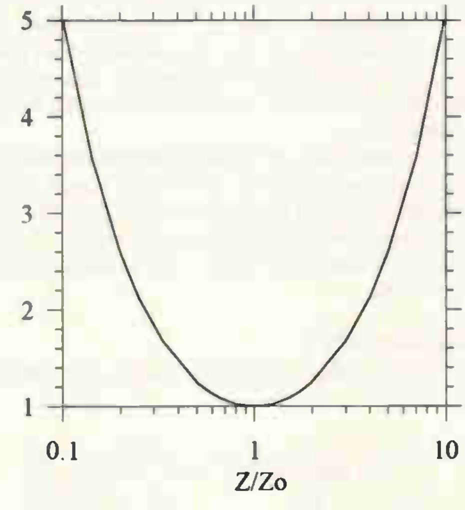

Fig. 9. The additional delay factor of a short transmission line characterised by Z & td, inserted in a line with characteristic impedance Zo

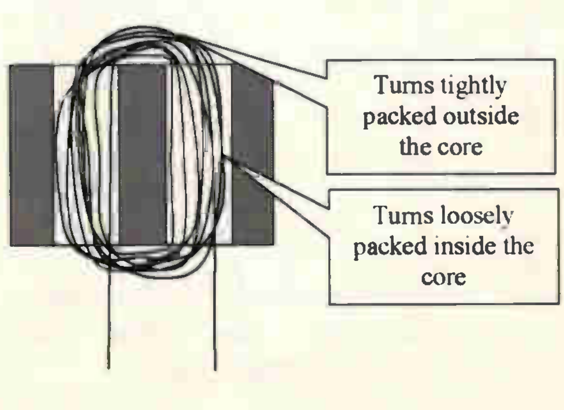

Fig. 10. Eight turns hand wound on a small core. What is Z0?

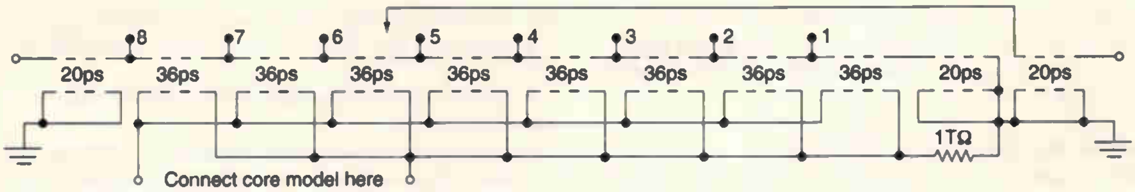

Fig. 11. SPICE model of a loss-free auto-transformer. Turns have F= 2.5E8 NL=0.009 Z0=100 and connection wires have F= 3.0E8 NL=0.006 Z0=150

transformer, I cannot think of a way of determining a value for Z_0 for use in the transformer's SPICE model. There are many difficulties:

a) The simple formulae for impedance of multi-wire lines are based on one signal conductor and several return conductors. These return conductors all carry the same voltage and current. This condition does not apply inside a transformer.

b) **Fig. 10.** shows that Z_0 is not constant over one turn. Even if it were constant, the average impedance would be impossible to calculate.

c) We have had to assume that $\varepsilon r \approx 1.44$. The choice of ε_r also affects Z_0.

d) For the turns on the outside of the windings, the core must have a significant effect, as ferrite's ε_r and μ_r are both very large.

e) In the model, the transmission line represents both the phase delay of the turn and the mutual coupling of the turn with the transformer core. It is not clear to me how the magnetic coupling should be reflected in the impedance chosen for the transmission lines.

f) Z_0 is a simple real number independent of frequency only if the line has no loss. Although the line has loss, the assumption is one that is frequently made (e.g. 50Ω cable).

However, experiment with different values of Z_0 in the model shows that Z_0 does not have a great effect on the loss predicted by the transformer model. The predicted loss increases as Z_0 increases, and this is more noticeable near 2:1 turns ratio than at very large or very small ratios. Z_0=100Ω gives a reasonable fit to the data. This is a good enough value to start from, but does not take into account the winding losses.

The connection wires are modelled as transmission lines. They are 6mm long and in air, so $F=3\times10^8$ and NL=0.006. The equivalent value of t_d would be 20ps. The impedance of the connection wires is high because they are a long way from the ground-plane relative to the diameter of the wire. Say 150Ω.

Results from SPICE

Fig. 11. shows a simple model of a transformer. The core

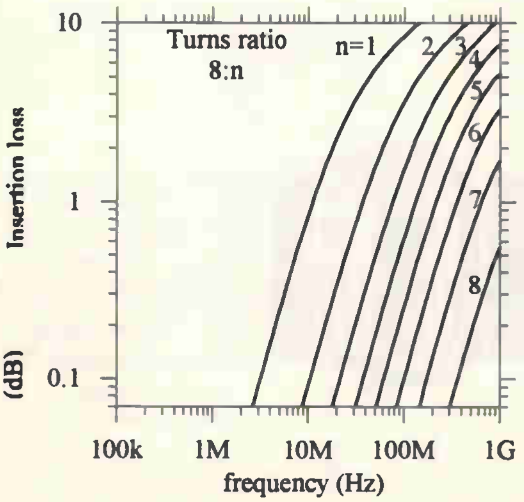

Fig 12. Simulated insertion loss spectrum of the loss-free step-down transformer in Fig 11

model will be discussed later, but an ideal core with no loss and an infinite inductance per turn is modelled by leaving the core model connections open-circuit. If two of such transformer models are connected back to back as in Fig 7, the simulator gives **Fig. 12**, which has a high-frequency performance similar to the measured values shown in Fig 8a.

The simulation results in Fig. 12 are for transformers with no resistive elements, and so represents a theoretical limit to auto-transformer performance. If you want the transformer to work at higher frequencies than Fig. 12, changing the core material is fruitless. You must either shorten the winding length, use filter matching techniques[6], or abandon the simple auto-transformer and use something more complex with multiple transformers. Examples include a chain of auto-transformers with a smaller impedance ratio, diplexers to route the signal to multiple transformers with different frequency responses, or transformers wound with defined transmission line impedances[7,8]. Sevick[3] has named these types Guanella transformers. Guanella transformers have their optimum performance when the transmission-lines are a quarter of a wavelength long, and this performance is extended to lower frequencies by ferrite loading of the secondary transmission line. ■

In the next part, I will show how to model the ferrite core loss.

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LETTERS to the editor

Letters to "Electronics World" Highbury Business Communications, Anne Boleyn House, 9-13 Ewell Road, Cheam Road, Surrey SM3 8BZ e-mail j.lowe@highburybiz.com using subject heading 'Letters'.

MFB (or not?)

Jeff Macaulay's article (Sept. 2002) offers an interesting way of improving the performance of a traditional woofer and I would not seek to belittle that. However, I do not accept that the technique described can be classified as motional feedback, because all of the necessary steps are not present.

At no point is there an analogue of the back emf of the drive unit. At no point is this subtracted from the input signal to produce a cone velocity error. There is no high loop gain applied to a cone velocity error and no compensation to obtain stability. As there is no motional feedback, there is no extension of the frequency response. This is evident from the fact that external equalisation is needed.

Negative output impedance is one way of increasing damping, but I can (and do) increase the damping of woofers by fitting stronger magnets. I doubt that I would be allowed to claim this as motional feedback.

John Watkinson
Reading, UK.

Expect to hear more of John's ideas in the future - Ed

No conspiracy

I believe the debate about radiation from mobile phones (and overhead power lines) has been compromised by the silly ideas put forward as possible modes of action. Let us be clear that the radiation quanta are of far too low energy to cause any chemical reaction (you need light for that, which is why we see in the band we do); there is no chance of building up cyclotron resonance in hydroxyl ions (for example) as their motion is being continually randomised by thermal agitation; and the tissue temperature rise is negligible compared to that produced by a hot bath or even a hot cup of tea. It is not therefore surprising that studies looking for physical changes such as cancer produce at best indeterminate and statistically dubious results, nor that the village of Wychbold (slap bang underneath the 500kW Droitwich transmitter) does not seem to be a cancer hotspot. I think it entirely possible, however, that just as mobile phone interference causes my CD player's processor to crash, it may equally well disrupt the tiny electrical signals in the brain and effectively cause part of the brain to crash. Psychological and psychosomatic effects are therefore much more obvious. It is also the case that the brain's response to 50Hz fields is considerably enhanced by stimulants such as amphetamine and its relatives, popular among the 'rave culture' but probably not among 'bio-EMC' researchers. It might prove profitable to correlate amphetamine-induced psychoses with living under power lines.

Pigeon
By email

Measuring low distortion

While I applaud the attempts to use low cost measurements whenever they are adequate for the task in hand, I fear this is not so with the Richard Black attempts to measure speaker cable distortions.

Before attempting any serious distortion measurements, especially those planned for publication, one should first attempt to quantify the level of distortion that might be expected. Then add a safety margin, typically 10dB, as an absolute minimum should be provided.

In like fashion any signal source used must also be of lesser distortion, since any attempt to quantify small distortion differences, swamped by much larger distortions, is unrealistic.

Researching back publications, in measurements reported by G. Millard (BBC engineering), carbon-film resistors typically exhibit THD around -120dB -130dB, mostly third harmonic and intermodulation around -100 to -120dB.

Using my equipment I too have measured similar values with carbon film but even with a dynamic range of -140dB cannot satisfactorily measure differences in metal film resistors. For that much lower distortion equipment, or a different measurement, is required.

Surely one must assume that all

Shock hazard

I welcome the full debate my letter on the shock hazard of 1mA has brought. I expected (and hoped) another reader would find a standard (the IEC 60479) bringing real authority to the debate. My top the of head and practical observation guesses were in the right ballpark, if a bit high on the Ventricular Fibrillation threshold.

Obviously individuals should always be protected form the chance of electric shock by suitable barriers and by limiting powers and energy to as low a level as practically possible.

In my work we follow EN60204 "Safety of Machinery, Electrical Equipment of Machines" which generally requires a double barrier between the user and live parts. Covers require tools to remove them and interlocks cut the power to exposed parts or secondary insulation (such as terminal block covers) prevents in inadvertent contact. Many items in the domestic environment fall well short of this goal. On the subject of safety, the dangers of UV light are perhaps under stated (Tiaraju Vasconcellos "Wager's Ozoniser" (August 2002) and Jose A.Senna "Another UV source for EPROM erasing" (September 2002)). The short wavelength UV from an unshielded mercury vapour lamp quartz arc tube is readily damaging to the eyes, like viewing a welding arc. The outer filter glass bulb should never be removed. Such lamps always have outer glass bulbs or are operated in enclosures with further glass filters. Today UV concerns extend to quartz halogen lamps also and the popular 20 and 50W dichroic low voltage spotlights now have glass filters.

***Paul Bennett** MIEE CEng, UK*

copper based cables will produce less third harmonic distortion than the best resistor?

I can find no statement as to exactly which soundcard or even test voltages he used. Most modern cards are 16 bit only but a select few are 20 bit or better. The dynamic range of an unaided 16 bit card is not sufficient, which is why I had to spend time and money developing a notch filter/preamplifier.

For my initial experiments, in August 2001, I did try using my Soundblaster 1024 Live card with the signal generators in 'Cool Edit', to measure capacitors. These signals are far from distortion free and all attempts failed. Noise and distortion from the generator far exceeded and clouded those I was attempting to measure. I also tried measuring intermodulation distortion differences, using these signals. To say the results I found were utterly confusing would be an understatement.

From my experiences I suspect that Black's attempts to measure cable distortions using his system were doomed to failure from the beginning. Perhaps he also was making the wrong measurements.

Even using the much-vaunted AP test set would not, I believe, be sufficient to measure distortions in speaker cables.

Failure to measure differences using a test method may prove comforting but it does not prove that cable distortions or differences will not be found in more realistic tests.

To paraphrase a sentence from my first Capacitor sounds article "truly audible differences must be both understandable and measurable. Understanding in terms of construction. Measurements may however require a change in measurement technique."

From my earlier work on speaker cables I know that audible distortion differences certainly do occur with change in cable construction. Measuring them is quite another matter.

I also have some comments on last month's letters. John Jardine makes the point he would like to see all schematics redrawn; Martin Eccles frequently did get schematics redrawn. As a contributor I have suffered many hours trying to identify and remedy errors in redrawn schematics, often in barely legible Faxed proof copies. I accept the present arrangement of posting hardcopy proofs is easier but even so I for one shudder each time I receive a redrawn schematic for corrections, so would certainly not welcome such a move as routine.

Having regularly submitted many exceptionally well developed constructional and measurement projects, all complete with PCB layouts, I would hope the quality of the schematics I submit as HPGL plot files should be acceptable. If not and the decision is to redraw all with no increase in page rate, then for me no more schematics unless very small and simple. End of project design publication in EW for me.

As to the Michael Edinger letter, I suspect he wrote that quickly with little thought as to practicality. While I cannot speak for Mr. Black, I suspect that like me he is freelance and not an EW employee. In which case all expenditure in time, materials and travel must be recouped from the modest EW page rate.

From experience, especially of my recent projects, which so many readers have admired, that quality of article cannot be cost effective for one person, yet alone two sharing the page rate and incurring extra expenses. While suitable recompense can overcome many problems that also would seem unlikely.

As to his suggestion of downloading PCB drawing BMP files, no way, even compressed the file sizes would be enormous. He may have a broadband access, most do not. As to hardware, even a modest PCB would require using screen resolutions of 1800 x 1200 as a minimum. I do not have such capabilty and to reduce eyestrain always prefer to work at 800 x 600 resolution.

I could output 'Gerber' but how many readers can accept Gerber, I suspect none. It might be feasible to output a compressed PCL file, but again how many can read and print PCL. As to PDF, from PCB software that requires a number of file conversions again with the creation of errors.

I can speak on this with some authority, since of late I have spent considerable time trying to find how best to output such files from a Windows machine to CD Rom. If any reader has a proven practical method that does not cost an arm and a leg, I would like to be so advised.

He like many others also seems over enamoured of common simulators. I wrote my first dedicated simulator way back in 1983. Since then I have used Touchstone and Hewlett Packard's Microwave Design system running under Unix, also many variations of Spice both DOS and Windows as well as a Microwave version of Spice from Compact Software. For most of my published projects I used Microcap6.

While usable models for transistors and ICs are freely available and can provide reasonably accurate results, as soon as a design involves capacitors larger than a couple of hundred pF or inductors more than a couple of hundred μH, goodbye accuracy, especially calculating distortion and frequency response, even at audio frequencies.

Perhaps if he too purchased a copy

CPU architecture

A fast, agile computer is essential for the function of modern defensive weapons such as an anti-missile. A simple homing warhead in an anti-missile might be foxed by an incoming target that employs counter-measures. Counter-counter-measures are difficult.

In my work for spastics, I have examined numerous computer architectures in the course of my study. The biggest problem that I have encountered is in the realm of input-output. It is in the pursuit of an improved connection between input-output and a computer's software core that I wish to suggest the 'differentiated multi-processor'. A classical multi-processor consists of a handful of similar cores tightly connected using shared memory. It seems to me that the performance of this system can be improved by individualising each core according to function. That is the basic idea.

To give an example of how such a computer might be implemented, take the M68000 microprocessor family. The 32-bit 68020 member of the family is a particularly interesting microprocessor core because of its exceptional support for high-level language. I am particularly referring to its firmware support for modules, the CALLM and RETNI instructions. Using dual-port RAM memory, which is now freely available, it does seem to be feasible to connect the high-speed computation-oriented 68020 directly to other M68000 family members as a tightly-coupled multi-processor. Hardware 8-bit data bus communication, to simplify the wiring, would be feasible because the 68020 offers dynamic data bus sizing, while there are still M68000 family members, akin to the old 68008, that can be hard wired for 8-bit, operation. But so what? It is plain to me that here is a way round that perennial old problem with microprocessors, package pin-out restriction. Enough said!

Allan Campbell
Newcastle, UK

of the SPICE2 document ERL-M250 from Berkeley as I did many years ago, he might find it illuminating.
Cyril Bateman
UK

We often have to re-draw circuit diagrams, not because the author has made a mess of it, but usually because the original design software will not export to something other than a bitmap (usually via MS Paint). This normally produces 'jagged' resistors, gaps in the lines connecting components and the text is far too small to be read if printed. Complex circuit diagrams that go through a manual re-draw process have to be checked, which of course takes time. Each author is different, which can make proofing a time consuming process. In an ideal world, the circuit software would output a standard DXF file, which could then be just tweaked, MS Equation Editor would allow you to export its funny little text box pictures into something normal and above all the Mac (and Quark layout software) would recognise a whole lot more than it does at present. And I could just email a PDF file to the author to print, correct and re-scan to be sent back to me. - Ed

Spectrum pricing

Dr Rudd is correct in his observation that auctions are normally designed to maximise revenue. But that is not the same as realising the full economic value of the spectrum. When the objectives of the 3G auction were announced to the House of Commons on May 18th 1998, the words used were: 'realise the full economic value to consumers, industry and the taxpayer..' What the taxpayer has gained has been lost by industry and/or consumers. (Even the taxpayers' gains are short term. To properly take into account tax revenues over the 20 year licence period would lead to a very different auction design.) Since the most profitable bidders will win, an auction will normally maximise industry profit, but intervention is needed if you want to include consumer benefit in the maximisation (unless you have perfect competition!) As this was not done it could be argued that the auction was the wrong tool for the stated objective. Neither did it comply with the EU Licensing Directive, which requires 'due weight to be given to maximise benefits to users' and 'ensure optimal use'. I covered this in more detail in an article in the June issue of Land Mobile.

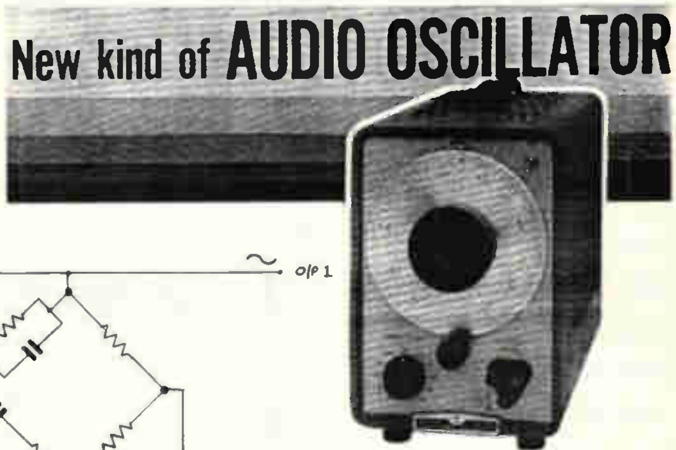
Dr Rudd also comments on the practice of reserving bands for specific services. While it is true that this can raise prices in certain bands, it will reduce others and increase choice. Overall, intelligent band planning will have a net positive effect on consumer benefits. ■

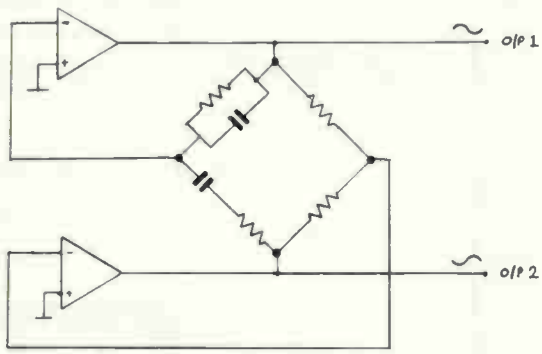
James Page
Manager, Radio Regulations, Nokia UK Ltd.

Wein revisited

I read with some interest Cyril Bateman's article about Capacitor Sound in the July issue. In particular, his modification of the Wien-bridge oscillator described by John Linsley Hood caught my eye.

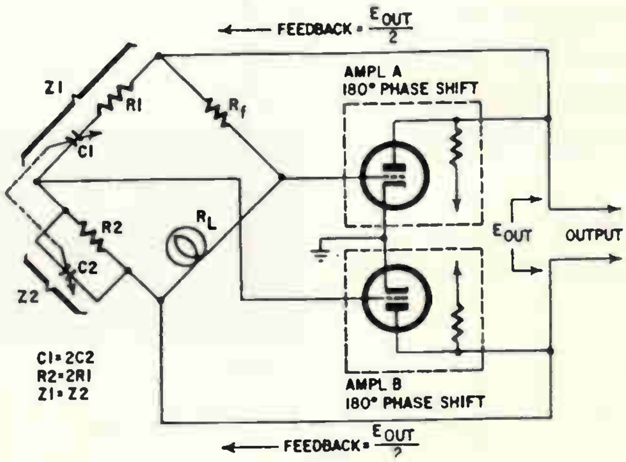
The two amplifier version of the wien-bridge oscillator seems to have a rather chequered history and to have been independently discovered several times. The occasions that I know (and I don't pretend to be an expert on the matter) are 1959 by Hewlett Packard[1], 1960 by W. Woodman of Nash and Thompson [2], an unknown date probably in the 1960's by Peter Baxandall [3] and 1981 by John Linsley Hood[4]. Intriguingly, the earliest version had equal output voltages as utilised by Cyril.





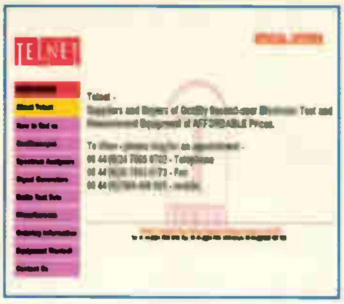
One advantage of the circuit is that the outputs from the two amplifiers are in antiphase. By making the two voltages equal, a balanced output can be obtained. In the days of valves, this was then an ideal way to drive a push pull output stage. I believe this was why HP adopted it on their 1959 oscillator. The block diagram in fig.1, with the bridge redrawn to show its balanced nature, illustrates their circuit. (Actually their bridge circuit was rather more complicated as complex RC circuits were used to enable 20Hz to 20kHz to be covered without band switching.)

P.F. Gascoyne
Wantage, Oxon.UK



References:

1. New kind of Audio Oscillator. Robert F. Scott. Radio Electronics Jan 1959 p58.
2. Letter to the Editor. W.Woodman. Wireless World Dec 1960 p610.
3. Private communication from J. Linsley Hood.
4. Wien-bridge oscillator with low harmonic distortion. J L Linsley Hood. Wireless World May 1981 p51.



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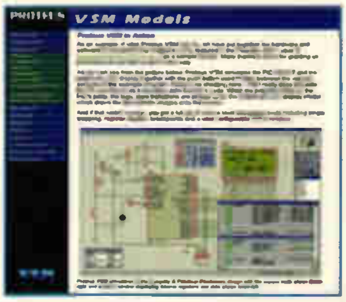

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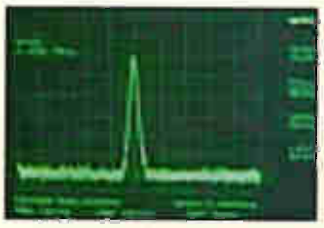

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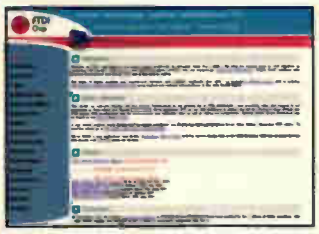

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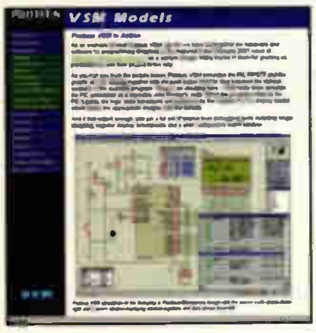

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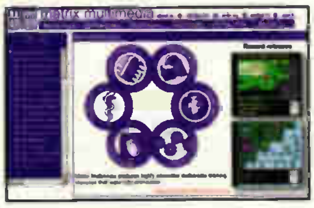

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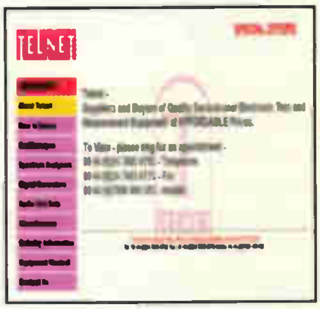

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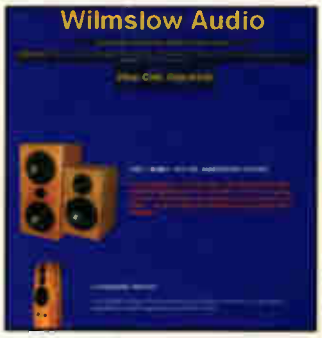

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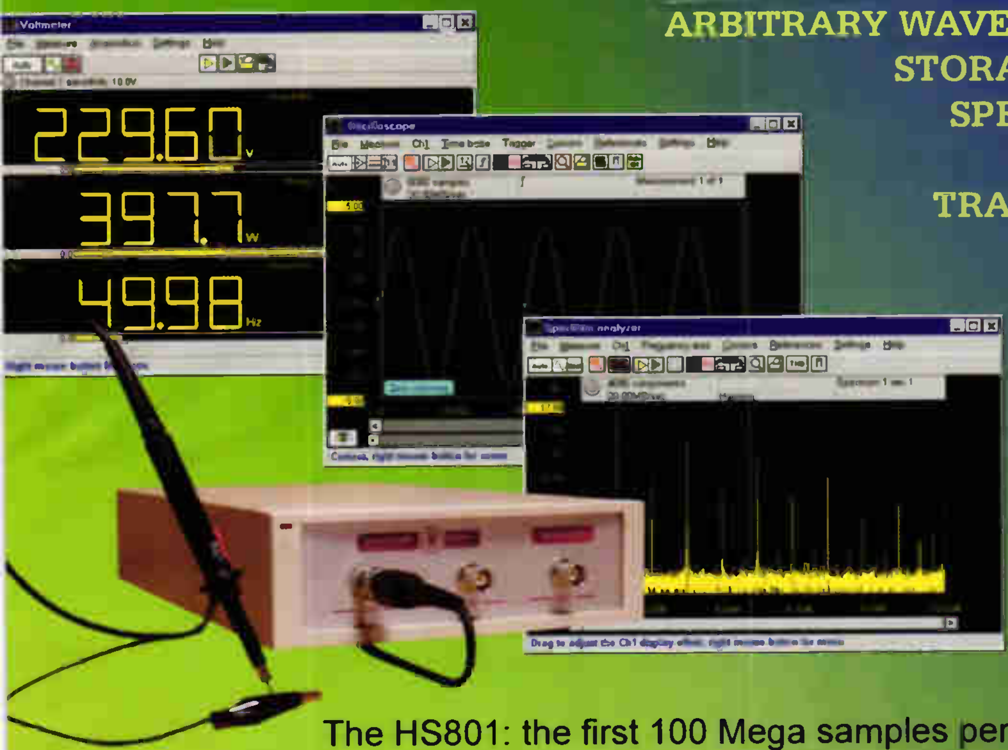

Reliability

The HS801: the first 100 Mega samples per second measuring instrument that consists of a MOST (Multimeter, Oscilloscope, Spectrum analyzer and Transient recorder) and an AWG (Arbitrary Waveform Generator). This new MOST portable and compact measuring instrument can solve almost every measurement problem. With the integrated AWG you can generate every signal you want.

- The versatile software has a user-defined toolbar with which over 50 instrument settings quick and easy can be accessed. An intelligent auto setup allows the inexperienced user to perform measurements immediately. Through the use of a setting file, the user has the possibility to save an instrument setup and recall it at a later moment. The setup time of the instrument is hereby reduced to a minimum.

- When a quick indication of the input signal is required, a simple click on the auto setup button will immediately give a good overview of the signal. The auto setup function ensures a proper setup of the time base, the trigger levels and the input sensitivities.

- The sophisticated cursor read outs have 21 possible read outs. Besides the usual read outs, like voltage and time, also quantities like rise time and frequency are displayed.

- Measured signals and instrument settings can be saved on disk. This enables the creation of a library of measured signals. Text balloons can be added to a signal, for special comments.

- The (colour) print outs can be supplied with three common text lines (e.g. company info) and three lines with measurement specific information.

- The HS801 has an 8 bit resolution and a maximum sampling speed of 100 MHz. The input range is 0.1 volt full scale to 80 volt full scale. The record length is 32K/64K samples. The AWG has a 10 bit resolution and a sample speed of 25 MHz. The HS801 is connected to the parallel printer port of a computer.

- The minimum system requirement is a PC with a 486 processor and 8 Mbyte RAM available. The software runs in Windows 3.xx / 95 / 98 or Windows NT / 2000 / XP and DOS 3.3 or higher.

- TiePie engineering (UK), 28 Stephenson Road, Industrial Estate, St. Ives, Cambridgeshire, PE17 3WJ, UK
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