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ELECTRONICS WORLD

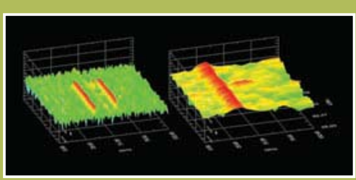
THE ESSENTIAL ELECTRONIC ENGINEERING MAGAZINE

MUTUAL INDUCTANCE
MEASUREMENT

SM ELECTROLYTIC CAPACITORS
TESTING THEIR ROBUSTNESS

FIBRE OPTICS

THE TECHNOLOGIES AND TECHNIQUES CURRENTLY USED



TECHNOLOGY
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Motor Drivers/Controllers

Here are just a few of our controller and driver modules for AC, DC, unipolar/bipolar stepper motors and servo motors. See website for full details.

NEW! PC / Standalone Unipolar Stepper Motor Driver

Drives any 5, 6 or 8-lead unipolar stepper motor rated up to 6 Amps max. Provides speed and direction control. Operates in stand-alone or PC-controlled mode. Up to six 3179 driver boards can be connected to a single parallel port. Supply: 9Vdc. PCB: 80x50mm.
Kit Order Code: 3179KT - **£11.95**
Assembled Order Code: AS3179 - **£19.95**

NEW! Bi-Polar Stepper Motor Driver

Drive any bi-polar stepper motor using externally supplied 5V levels for stepping and direction control. These usually come from software running on a computer.
Supply: 8-30Vdc. PCB: 75x85mm.
Kit Order Code: 3158KT - **£15.95**
Assembled Order Code: AS3158 - **£29.95**

NEW! Bidirectional DC Motor Controller

Controls the speed of most common DC motors (rated up to 16Vdc/5A) in both the forward and reverse direction. The range of control is from fully OFF to fully ON in both directions. The direction and speed are controlled using a single potentiometer. Screw terminal block for connections.
Kit Order Code: 3166KT - **£16.95**
Assembled Order Code: AS3166 - **£25.95**

DC Motor Speed Controller (100V/7.5A)

Control the speed of almost any common DC motor rated up to 100V/7.5A. Pulse width modulation output for maximum motor torque at all speeds. Supply: 5-15Vdc. Box supplied. Dimensions (mm): 60Wx100Lx60H.
Kit Order Code: 3067KT - **£13.95**
Assembled Order Code: AS3067 - **£20.95**

Most items are available in kit form (KT suffix) or assembled and ready for use (AS prefix).

Controllers & Loggers

Here are just a few of the controller and data acquisition and control units we have. See website for full details. Suitable PSU for all units: Order Code PSU445 £8.95

Serial Isolated I/O Relay Module

Computer controlled 8-channel relay board. 5A mains rated relay outputs. 4 isolated digital inputs. Useful in a variety of control and sensing applications. Controlled via serial port for programming (using our new Windows interface, terminal emulator or batch files). Includes plastic case 130x100x30mm. Supply: 12Vdc/500mA.
Kit Order Code: 3108KT - **£54.95**
Assembled Order Code: AS3108 - **£64.95**

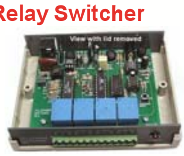
Computer Temperature Data Logger

4-channel temperature logger for serial port. °C or °F. Continuously logs up to 4 separate sensors located 200m+ from board. Wide range of free software applications for storing/using data. PCB just 38x38mm. Powered by PC. Includes one DS1820 sensor and four header cables.
Kit Order Code: 3145KT - **£18.95**
Assembled Order Code: AS3145 - **£25.95**
Additional DS1820 Sensors - **£3.95 each**

Rolling Code 4-Channel UHF Remote

State-of-the-Art. High security. 4 channels. Momentary or latching relay output. Range up to 40m. Up to 15 Tx's can be learnt by one Rx (kit includes one Tx but more available separately). 4 indicator LED 's. Rx: PCB 77x85mm, 12Vdc/6mA (standby). *Two and Ten channel versions also available.*
Kit Order Code: 3180KT - **£44.95**
Assembled Order Code: AS3180 - **£51.95**

NEW! DTMF Telephone Relay Switcher



Call your phone number using a DTMF phone from anywhere in the world and remotely turn on/off any of the 4 relays as desired.
User settable Security Password, Anti-Tamper, Rings to Answer, Auto Hang-up and Lockout. Includes plastic case. Not BT approved. 130x110x30mm. Power: 12Vdc.
Kit Order Code: 3140KT - **£46.95**
Assembled Order Code: AS3140 - **£64.95**

Infrared RC Relay Board

Individually control 12 on-board relays with included infrared remote control unit. Toggle or momentary. 15m+ range. 112x122mm. Supply: 12Vdc/0.5A
Kit Order Code: 3142KT - **£47.95**
Assembled Order Code: AS3142 - **£66.95**

PIC & ATMEL Programmers

We have a wide range of low cost PIC and ATMEL Programmers. Complete range and documentation available from our web site.

Programmer Accessories:
40-pin Wide ZIF socket (ZIF40W) £15.00
18Vdc Power supply (PSU010) £19.95
Leads: Parallel (LDC136) £4.95 / Serial (LDC441) £4.95 / USB (LDC644) £2.95

NEW! USB & Serial Port PIC Programmer

USB/Serial connection. Header cable for ICSP. Free Windows XP software. Wide range of supported PICs - see website for complete listing. ZIF Socket/USB lead not included. Supply: 16-18Vdc.
Kit Order Code: 3149EKT - **£37.95**
Assembled Order Code: AS3149E - **£52.95**

NEW! USB 'All-Flash' PIC Programmer

USB PIC programmer for all 'Flash' devices. No external power supply making it truly portable. Supplied with box and Windows Software. ZIF Socket and USB lead not included.
Assembled Order Code: AS3128 - **£44.95**

"PICALL" PIC Programmer

"PICALL" will program virtually all 8 to 40 pin *serial-mode* AND *parallel-mode* (PIC16C5x family) programmed PIC micro controllers. Free fully functional software. Blank chip auto detect for super fast bulk programming. Parallel port connection. Supply: 16-18Vdc.
Assembled Order Code: AS3117 - **£24.95**

ATMEL 89xxxx Programmer

Uses serial port and any standard terminal comms program. Program/ Read/ Verify Code Data, Write Fuse/Lock Bits, Erase and Blank Check. 4 LED's display the status. ZIF sockets not included. Supply: 16-18Vdc.
Kit Order Code: 3123KT - **£24.95**
Assembled Order Code: AS3123 - **£34.95**

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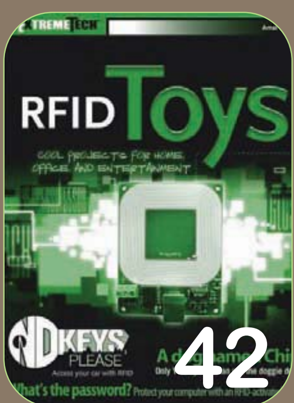

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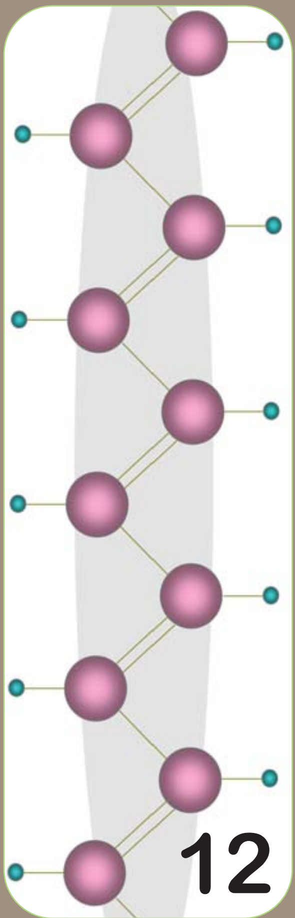

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Building a case for RFID

What's the use of Radio Frequency Identification (RFID)? Not many businesses seem to know the answer to that, it seems. A recent RFID networking event, held at the Olympia exhibition centre in London, highlighted exactly how confusing the whole issue of RFID is for businesses. Not many fully understand the technology and the potential benefits it could bring if harnessed properly.

However, it has not stopped some people dream up ideas, some of which, frankly, too grand to be handled successfully by any technology and not just RFID. For example, at the event there was an automotive think-tank discussing the potential use of RFID to tag all the major parts of a car so that if stolen and then stripped its parts could later be identified. It would be simplest – and cheaper – to give up on the stolen car altogether!

RFID seems to come in useful when locating cars on a large and busy forecourt, or locating parts coming in and going out of a warehouse, but to have a global network of RFID tags which are likely to vary in frequency, power levels, standards, suppliers, part numbers and just about everything else up and down the value chain of a product's life, is almost an impossible task.

Standards could help, but things are not necessarily moving in that direction. For a start, any organisation or retailer (such as M&S and Wal-Mart) that has selected its preferred RFID technology has largely kept it a secret for competitive reasons. This certainly is not conducive to establishing standards.

Each RFID implementation is a large undertaking in its own right; not only because the right type technology should be selected, but, also, supporting infrastructure, including the integration and management software.

At the Olympia event, one RFID integrator said that many of the companies he meets have heard of RFID and fear that they are missing out on something big by not having it installed already. However, his advice is for each firm to analyse its business thoroughly and fit RFID into where it will yield the biggest benefits. Otherwise, companies are bound to end up disappointed if they enter into an expensive installation with unrealistic expectations.

On the other hand, there have been those who have been pleasantly surprised with the benefits RFID has brought to their businesses.

Ultimately, it may boil down to an integrator managing customers' expectations and, for the time being, that seems to be the route forward for RFID.

Svetlana Josifovska
Editor

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A version of the Blackfin DSP backed by Audi

Analog Devices (AD) has launched a version of its Blackfin digital signal processor for the automotive market with backing from car manufacturer Audi.

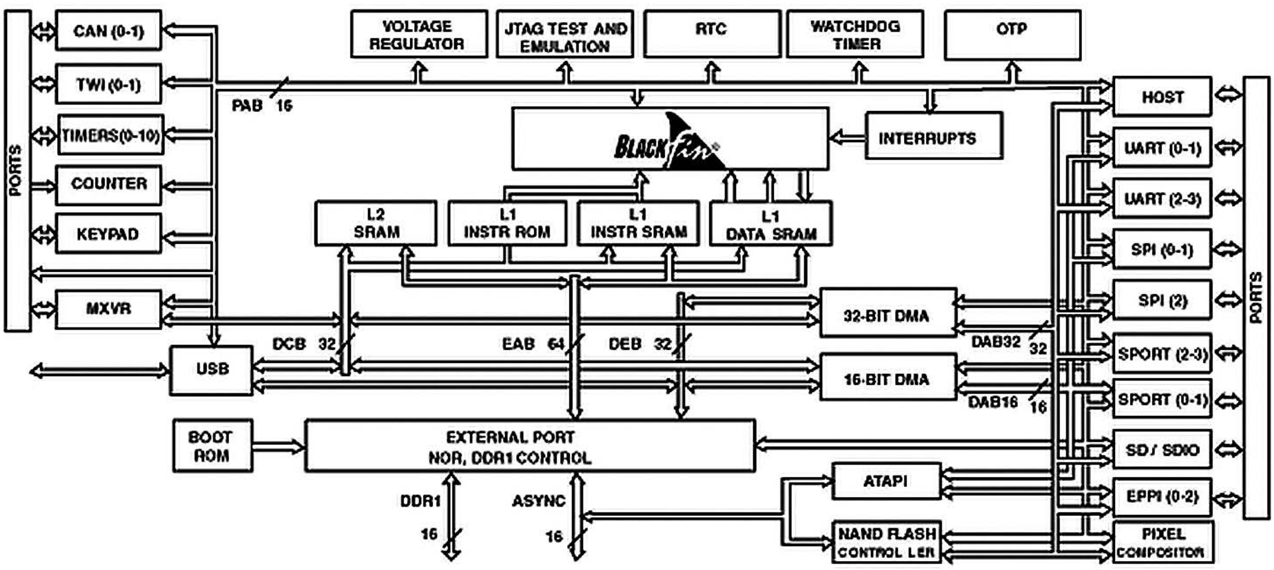
The ADSP-BF54x family has more I/O and memory bandwidth than other families of the core, and integrates both CAN and MOST network interfaces peripherals for multimedia/networked in-vehicle automotive applications, such as audio/video Rear Seat Entertainment systems (RSEs), digital radios, driver assistance systems, navigation head units and hands-free phones.

"System performance has become a crucial enabler for in-vehicle electronics because audio, video and hands-free communication continue to intensify processing demands," said Peter Kohlschmidt, director of Development Infotainment at Audi in Ingolstadt. "The Blackfin processor has proven a good fit for our audio head units."

The family operates at speeds of up to 600MHz and doubles the internal bus bandwidth to 532Mbytes/s with up to 260Kbytes of on-chip memory and up to 152 general-purpose I/O ports. It also adds a high-speed USB On-The-Go (OTG) with integrated PHY, ATAPI controller and NAND flash controller so that PC peripherals could be used.

"The software flexibility is critical for automotive applications, in particular because media formats and communications standards inevitably change in the time it takes applications to go from concept to the highway," said John Croteau, general manager, Convergent Platforms and Services Group at Analog Devices.

Because the software is key to differentiating the system, the family also embeds new Lockbox Secure Technology that protects developers' software code by encrypting any or all of the system. It is configurable so system designers can choose any encryption algorithm they prefer and, also, provides a platform for the digital rights management (DRM) content protection for devices such as media players in the car.



The Blackfin BF549 is aimed at automotive applications with CAN and MOST interfaces

China traipses familiar ground

The DTI has further cut the entry level of R&D investment for the R&D Scoreboard to £19m this year and, yet, no electronics Chinese firms featured in its annual report.

The investment threshold level has already been cut three times over the past three years, from £37m to £22m last year, and now to £19m. According to one of the main authors of the report, Dr Mike Tubbs, this has made the scoreboard wider and allowed entry to many firms from all over the world, but it has not done anything for the Chinese.

"It took Taiwan and South Korea quite some time after doing 'me too' products to become R&D intensive with own products," said Dr Tubbs. "What we are seeing in China is similar: we are seeing [just] manufacturing, then they'll copy [products], then they'll start licensing, and only after that will they decide to go up the value chain [and invest more in R&D]. The whole process takes some time."

South Korea and Taiwan have done well in this year's scoreboard, however. According to the report, South Korea ploughs some 60% of R&D investment into the electronics sector.

Advanced signal analysis system ready for French researchers

FPGA signal analysis experts RF Engines is working with the French Atomic Energy Commission (CEA) Laboratories in Paris to develop a system to automatically select the appropriate resolution bandwidth for required signals out of a wide bandwidth. This is likely to use the simultaneous multi-resolution capability of RF Engines's Pipelined Frequency Transform (PFT).

The company, based in the Isle of Wight, has delivered a high performance real-time, signal analysis system using PFT implemented in an FPGA design. This will be used to acquire and investigate complex signals from across a 25MHz bandwidth in real-time, and uses a combination of novel frequency measurement techniques to acquire the signal from a noisy signal environment and to assist in the characterisation of the modulation type. It also includes data capture for off-line signal analysis.

This work follows on from an earlier evaluation demonstration system by the CEA, which focuses on fundamental research, and a follow-on contract has also been signed to develop the auto select system.

"We have built a very strong relationship with CEA over a number of years, which has resulted in this order for a very advanced signal analysis product for their development work," said John Summers, chief executive of RF Engines. "We are one of the world leaders, if not the leader, in high performance channelisation systems and have been able to apply this expertise to CEA's specific requirement to deliver a solution that exactly meets their needs."

Cirrus combines analogue and digital sound in LCD TVs

Cirrus Logic has launched a highly integrated Class D amplifier for the digital LCD TV market that combines both the analogue and digital audio control for the TV chassis by integrating a stereo analogue-to-digital converter.

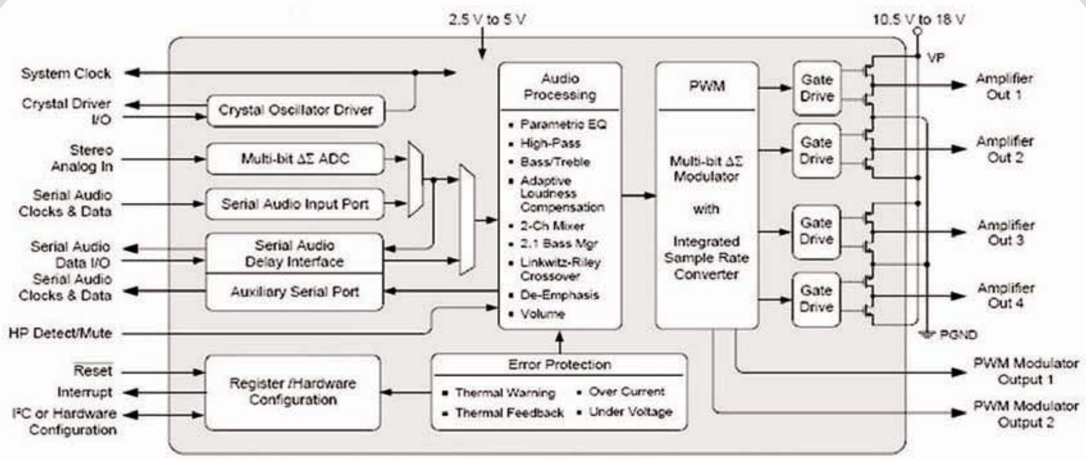
This avoids the need for separate analogue and digital audio systems in the design, but the efficiency of using a Class D amplifier also means no heat sink is needed, reducing the cost and complexity of the system design.

An on-board digital audio processor also handles functions such as parametric equalisation to make sure the sound is at the same level between programmes and advertisements, as well as an auxiliary output for handling lip-sync delay. This can be a major problem for LCD TV displays as it can take a significant time to generate the image and the delay can be hard to introduce in other parts of the signal chain.

"Because of its high level of integration, the CS4525 gives flat-panel DTV manufacturers a superior design solution for managing the audio signal chain, providing a wealth of features and exceptional audio output and quality in a simple-to-use integrated circuit," said Jason Rhode, vice president and general manager, Mixed-Signal Audio Division, Cirrus Logic.

The chip is based on a 24-bit Delta Sigma architecture with a built-in sample rate converter that can manage a variety of incoming audio signals at different frequencies, such as PCM stereo and audio from DVD, while also eliminating clock-jitter effects.

In addition, Cirrus included its patent-pending thermal warning and fold back technology, which scales back output levels automatically if internal chip temperatures produce excessive heat. This is a particularly important feature for flat-panel televisions, whose overall video and system components generate significant amounts of heat within the thin units.



The CS4525 from Cirrus Logic combines a class D amplifier with an ADC to handle both analogue and digital audio through an LCD TV

Freescale and the BMW Group have partnered to bring the industry's first use of FlexRay technology in BMW's new X5 Sports Activity Vehicle (SAV). FlexRay is an in-car communications system designed to deal with the increased high-speed electronic content in cars. One of its features will give drivers access to unprecedented handling and stability control. As such, it is the cornerstone of BMW's AdaptiveDrive capability, which gives drivers a combination of Active Roll Stabilisation and Electronic Damping Control.

* * *

Degussa of Dusseldorf is developing novel materials for lithium-ion batteries that will create a new generation of highly-efficient and lightweight starter batteries for cars. Degussa's new batteries weigh only 2.5kg and are already being used in a Lotus racing car. This compares to conventional starter batteries that typically weigh between 15 and 20kg. At the heart of the battery is Degussa's separator that consists of an ultra-thin ceramic composite material, highly efficient electrodes and electrolyte additives.

* * *

The IET (Institution of Engineering and Technology) announced that it will be actively support F1 in Schools Ltd, a company which promotes engineering to young people around the world through the appeal of Formula One.

A non-for-profit company, F1 in Schools is a multi-disciplinary challenge in which teams of students aged 11-18 deploy CAD/CAM software to design, analyse, manufacture and test miniature CO2-powered balsa wood F1 cars.

This now operates in 22 countries, inspiring students to learn about science and technology, but also other skills such as marketing.

Low power FPGA family

Altera's new line of programmable logic is likely to quell FPGA critics over their typically high power consumption. The new Stratix III family of FPGAs has been specifically designed to consume less power than its predecessors.

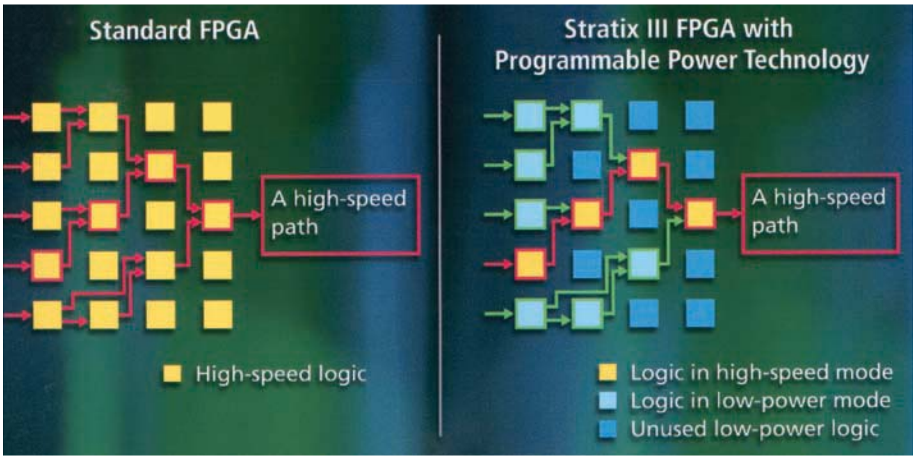
Even though based on 65nm process node, which doubles transistor count and hence power consumption, Altera has applied several innovations to the silicon, design and tools to keep the new family members' power budgets in check. Higher transistor density means higher leakage current too, which leads to greater static power. However, the firm has used strained silicon and triple oxide in the devices' fabrication and optimised the silicon process to reduce power consumption. "For years, FPGAs used to be the glue logic and now they are at the heart of the system. And in most systems today, there's a power budget you cannot exceed. This is the area we think we have most innovation in for the 65nm FPGAs," said Danny Biran, product marketing vice-president.

One of the innovations is applying different voltages to the logic gates in the FPGA. The firm added extra circuitry to deliver this and the penalty is increase in real estate. Another innovation lies in the Quartus-II design software. The tool identifies the design's critical path and automatically adjusts the speed and voltage of each block to minimise power consumption. "We are achieving power savings with programmable power technology. We look at specific blocks – logic, DSPs, memory – and identify the ones where performance is critical, so they can run at higher frequency," said Biran. "The others can run differently and, as such, consume less power. That way, the system can use only the power it needs to run and no more. This is very novel in the industry."

According to Altera's own research, only 20-30% of blocks within a design need to run at higher frequency.

In addition, Stratix III will have a selectable core voltage – 1.1V and 0.9V, which will also help lower static and dynamic power consumption.

The Quartus-II software lowers power in a design by identifying the critical path



EKTN created in the UK

The UK government is investing £3m in a Knowledge Transfer Network for the electronics industry. The Electronics KTN (EKTN) has appointed a chief executive and is set to make the announcement in the next few weeks after contract negotiations.

"The EKTN has a key role to play in acting as a catalyst for the creation of a vibrant, self-confident electronics industry in the UK, in which the sum of the parts is significant, successful and globally competitive," said the chairman, David Kynaston, who is also a vice president of the Intellect trade association. "This will involve all aspects of the sector, operating across domains from academia, research and design through to manufacture, and offering products and services which include hardware, software, tools and methods. It will aim to create an environment in which small and large companies and enterprises can flourish, and to which international ones are drawn."

The role of the EKTN will be decided with the new CEO and the board, said Derek Boyd, CEO of the National Microelectronics Institute (NMI), which was a key part of the bid to set up the network and already runs Knowledge Transfer events for the industry across the UK.

PMN connects public and private mobile networks

Private Mobile Networks (PMN), a wholly owned subsidiary of Yorkshire-based fixed mobile solutions provider TeleWare, launched the PMN SIM card last month, a technology that will enable business users on a private mobile network to move seamlessly between private and public GSM networks without having to make any changes to their mobile phones or be charged large fees.

The network could be deployed as simply as using a laptop and a small, low-power GSM transmitter, or picocell. PMN is using picocells from infrastructure supplier ip.access. When turned on, the phone will automatically search for the preferred network, in this case the PMN network, in order to take advantage of the lowest call charges.

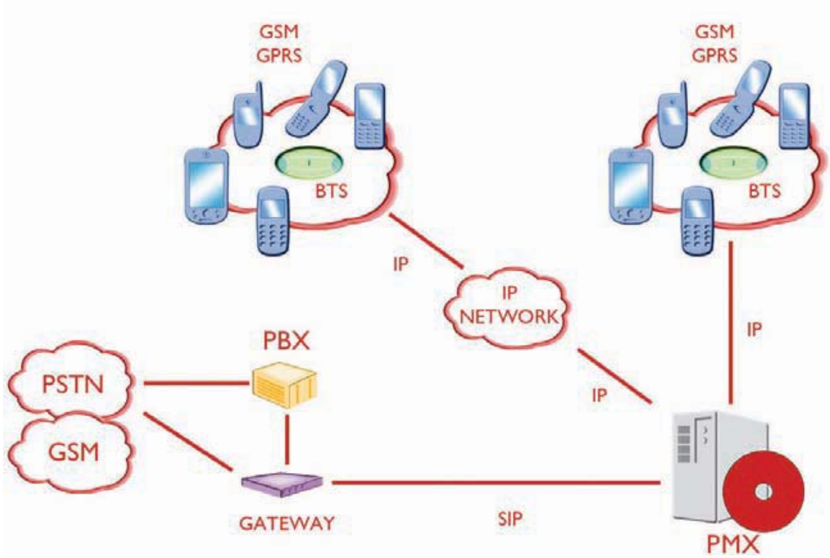
"You have a saving on roaming up to 60% on your usual charges," said Lesley Hanson, marketing director at TeleWare. "And yet, there's no difference in user experience, no new handsets and no loss of capabilities. We believe this is a disruptive technology."

This could be added to PMN's earlier offering – the Private Mobile eXchange (PMX), which allows companies to build a private GSM campus network. Such network is able to carry GSM traffic between mobile phones used within the campus and from mobile phones across intra-site IP networks, without incurring mobile operator call charges.

Private mobile networks enable private, wide-area GSM networks that function within a fixed/mobile convergence environment and use the net for toll bypass. This is achieved by replicating a cellular network in software and interoperating with IP/SIP devices and applications. IP PBXs are the software equivalent to a digital PBX. PMX is the cellular equivalent.

PMN's technology is being used in several pilot schemes, including at the Ministry of Defence in the UK and the Department of Defense in the US.

The future of this technology offers great potential, even for the population at large and not just business users. For example, if a supermarket was to take ownership of such a network, a shopper within the store can use free texts to pinpoint a product's location.



Cheaper call charges are now enabled by PMN's solutions

Innos, the UK-based nanotechnology R&D firm has agreed to provide engineering support and prototyping services for the UK Framework 6 Network of Excellence SiNANO project. The consortium is funded by the EU, with the Engineering and Physical Sciences Research Council (EPSRC) providing additional access to the Innos fabrication facility to the different UK academic partners. The main objective of the project is to explore different technology routes and achieve very high speed, silicon-based nanoscale devices, which can be adopted in the future engineering of ICs.

Each partner in the SiNANO provides different areas of expertise. The research will aim to enhance device performance and integration, in order to meet the ever-increasing demands placed upon components in communications and computing technology.

* * *

A smart through-wall radar technology developed by Cambridge Consultants promises to give new tactical advantages to police, special forces and emergency services. Prism 200 is a compact sensing device which will help reduce the high risks involved in tactical entry and siege or hostage situations.

It employs software to evaluate the position and movement of people in rooms and buildings.

The system is the size of a laptop and its operating distance is up to 20m.

* * *

Engineering education programs throughout the world are outdated and do not teach the skills necessary to succeed in an increasingly globalised world, states a recent study initiated by Continental AG. The "*Global Engineering Excellence*" study was conducted at top universities in Germany, Japan, Switzerland, US, Brazil and China. Its findings highlighted that engineering programs are not sufficiently international in terms of student participation and partnership opportunities and that action is required in science, industry and politics to internationalise engineering sciences.

In response, Continental AG has proposed a new joint "*Master of Global Engineering*" study program with participating universities that will encourage study-abroad programs and international scholarships.

BUSINESS CONTINUITY (BC) SOLUTIONS

1 **Offsite web contact database,** courtesy of www.salesforce.com. Sign up for their free single user account and get up to 20MB space free. There are enough free format fields to stretch to your BC plans, listing suppliers, staff and other relevant operational numbers in groups and sub-groups, as required. Salesforce back up the data, giving you a reliable and convenient route to all your essential contact details from anywhere in the world with an Internet connection. Access is via secure 128-bit encrypted https username/password log on, so your data is safe.

2 **Free 0870 numbers** from the likes of www.skycomuk.com. These numbers can be diverted to any underlying landline, or mobile number. Hence, if you find that your entire place of work is unavailable, you can go online and switch the relevant 0870 number to divert straight to your preferred alternative. This gives you a robust solution to key aspects of telecommunications resilience, even in extreme scenarios, where huge sections of staff may need to work from home.

3 **Wireless broadband at home and at work** from any number of suppliers. Many of your staff may have it already, but when catering for flexible operational scenarios within your BC plan, it becomes a high priority. Cost depends on supplier, but make sure key staff have at least 2Mbps access, for as little as £20 a month. We use BT as the largest operator of such networks and seldom have issues. Wireless equipment can be sourced from your local Dixons, Currys, or PC World, or even direct from BT, for less than £50. Spread that expense over the life of the employee and compare it to the value of the employee per hour and, we think you'll agree, it is money well spent. Its value is in the ability to connect to the Internet from multiple places in the home and for visitors to your home to do the same. This makes every household configured and ready for real-time team working over the Internet, including free global voice communication via headsets through www.skype.com.

4 **Laptops with Wi-Fi.** Combined with the wireless broadband solution, your people can take their work with them wherever they need to go. If your BC plan requires physical relocation, which it almost certainly does, then laptops make life for the worker so much easier. Workplace recovery planning can be easier and cheaper by far. The cost difference between a laptop and a desk-bound system is now a few hundred pounds. Security must be addressed through disciplined use of passwords, and encryption may be necessary.

5 **Hosted exchange server services** from the likes of www.fasthosts.com. Starting packages of five fully-functional email boxes can be obtained for as little as £50 per user per year, plus unlimited routing mailboxes for your entire organisation. You gain all the attractive features of MS Exchange, such as shared electronic online diaries, visibility of each other's mailboxes with permissions, shared files, plus web email versions of your standard MS Outlook email, all backed up and hosted for you. When I recently lost my laptop, I switched straight to the online version of my email, using a completely different computer and just kept going. Apple and Lotus Notes offer alternatives here, but as I don't use them I can't recommend them.

6 **Mass backup devices** like the Maxtor external hard drive range. Software provided with such devices often includes a "one-click" complete incremental system backup function. These personal devices will save you and your staff hours of heartache and days of wasted time, if used regularly. Cost depends on size of disk, but look for 100GB of space for around £50-75.

7 **USB memory keys** for essential files, like your BC plan, or your operations procedures manual. If you are canny, you can sometimes get one free from one of your suppliers. Contact www.criticall.co.uk with your personal contact details if you are a budget-holding business continuity professional and we will send you a 1GB memory key in the post, as long as stocks last.

8 **VPN client software** to enable your IT staff, or outsourced service provider to work on your systems remotely. This requires the relevant secure configurations of your firewalls, but will mean that essential IT operations can be run from anywhere in the world with broadband Internet access. We combine this facility with BT's OpenZone offering (www.bt.com), with over 20,000 secure wireless hotspots available globally, via a common username and password. Cost varies with package purchased, but never exceeds £1 an hour.

9 **Fireproof safe.** Some things have to exist on paper. Whether it is copies of insurance certificates, accounts, signed contracts, or copies of your BC plan and operations procedures manual, safes cost a few hundred pounds from the likes of Costco (www.costco.co.uk) and Staples (www.staples.co.uk). If you can't afford a specialist safe, pack files tightly together in metal filing cabinets, or cupboards, as this will help minimise fire damage. If possible, keep them off the floor too, in case of flooding.

10 **Two-way automated notification system** for larger organisations (typically with revenues of over £5m pa and/or over 100 staff). These cost as little as £500 a month from the likes of www.criticall.co.uk, inclusive of several thousand messages. These enable you to broadcast email, pager, SMS and instant email (Blackberry) messages to key recipients. Value for money comes from their versatility. The same system can be used for day to day operations (eg helpdesks, call centres), marketing (eg last-minute offers), team communications (messages to mobile staff), as well as the traditional BC uses. Two-way enables recipients to confirm receipt, including vital messages in response (such as "I'm OK", or "Injured. Please send help"). Such systems are essential when call-cascades just don't work.

Ian Hammond, CEO of Criticall, supplied this month's top ten tips, based on a decade of specialising in emergency notification solutions.

Criticall will be exhibiting at Business Continuity – The Risk Management Expo 2007, held at London's Excel, Docklands from 28-29 March 2007. To register or for more information please visit www.businesscontinuityexpo.co.uk

If you want to send us your top five or ten tips on any engineering and design subject, please write to the Editor at EWeditor@nexusmedia.com.

The future of the MCU Market

Jean Anne Booth, co-founder and Chief Marketing Officer of Luminary Micro, discusses the fragmentation and impending consolidation of the MCU market

The maturation of a marketplace is not based on "divide and conquer" but on "consolidate and prosper". This well-understood stage in a market's development is now happening in the area of microprocessors.

In the early stages of a market, disparate technologies develop, sometimes haphazardly, under the pressure of newly identified needs. As these requirements become more clearly defined, the market becomes more structured and technological innovation gives way to volume growth. However, in these early stages, the market is still very fragmented, with no supplier holding a dominant share.

Eventually, some companies emerge as principal suppliers; in some cases, the few become dominant as an oligopoly, or one may even become the sole monopolistic supplier.

Remember the computer market? As it grew to maturity by the late 1970s, the list of computer suppliers included Honeywell, Sperry, Burroughs, DEC, Data General, IBM's S360, Cray, Amdahl... to name but a few. Each had their own underlying architecture, which ultimately consolidated to one – the x86 – by the mid 90s. The cost of supporting the total ecosystem (software, tools, applications, support etc) was a significant factor in the consolidation to what we now know ubiquitously as the "PC".

The same evolution occurred in the embedded (non-computer) microprocessor market. In the early 1990s, the embedded microprocessor market featured solutions based on x86, i960, 29K, SPARC, MIPS, ARM, 68K, PowerPC and others. This market has consolidated dramatically with ARM market share exceeding 80%. More than 2.3 billion ARM-based embedded processors are shipping each year.

The embedded microcontroller market is a huge market with over $26bn in annual revenue. It is also extremely splintered, with more than 40 suppliers feeding in excess of 50 architectures into the market, yet no single architecture holds as much as 5% share. This market is a good example of one that has grown very big while still remaining fragmented – thus far. With the introduction of the ARM Cortex-M3 core, first brought to market in Luminary Micro's Stellaris MCUs, the stage is set for disruptive forces to consolidate this last silicon processor frontier.

Like the computer market's consolidation, software and ecosystem play a significant role in the consolidation of the microcontroller market. Time-to-market is invariably driven by the software in an embedded design with over 48% of production costs attributed to software (VDC).

Together, Luminary Micro and ARM have the answer to the embedded software developer's dilemma – complete compatibility from $1 to 1GHz, an unequalled range. With entry-level devices at just $1, Luminary Micro's Stellaris MCUs provide ARM-compatibility for the same price as current 8- and 16-bit solutions, while offering significantly more performance.

Cortex-M3 dramatically extends ARM7 capabilities for MCU applications. Stellaris Cortex-M3 MCUs provide a two-fold improve-ment in interrupt performance and a four-fold improvement in MCU control applications. In addition, the compact architecture of the Thumb2 ISA needs only half the flash code space as ARM7 applications.

Stellaris MCUs appeal to the embedded market in other compelling ways: In addition to the many performance, price and functional advantages of Stellaris MCUs, ARM enjoys the broadest knowledge-base of any core in the world. The ARM community has the industry's leading tools and software, together with the most vibrant third-party ecosystem.

Importantly, being part of the ARM community gives developers access to hundreds of third parties, reduced development costs and increased efficiency. The openness of ARM is the catalyst that allows Stellaris MCUs to extend the ARM architecture into the wider 8- and 16-bit market. This is the disruptive force that begins the MCU consolidation. Although this process may take years, during which embedded systems designers and end-users of embedded systems will reap the rewards of increased market competition, the consolidation will happen as it did elsewhere.

Let the revolution begin!

> **"The embedded microcontroller market is a huge market with over $26bn in annual revenue, but it is also extremely splintered, with more than 40 suppliers feeding in excess of 50 architectures"**

FANTASTIC

Solid-state electronics has been with us since the 1950s, when the first transistor-based products appeared on the market. Starting with transistor radios and hearing aids, an explosion in consumer electronics began that shows no signs of abating to this day. On the way, discrete component designs were replaced by integrated circuits that are now ubiquitous in all sorts of electronic appliances. The shift from earlier vacuum tube or valve-based electronics to semiconductor-based electronics was a major development in electrical engineering.

It appears that another significant development in the technology of electronic materials and devices may be just waiting to happen. This time, once again, the underlying idea is quite revolutionary, although it may not have quite the same impact as the development of silicon electronics had, half a century ago. Its beginnings could be traced back to the somewhat accidental discovery of electrical conduction, in a well known polymer called polyacetylene.

FUNCTIONALLY, THESE MATERIALS ARE, IN FACT, SEMI-CONDUCTORS EVEN THOUGH THEY ARE ORGANIC POLYMERS AND ARE NOT NECESSARILY CRYSTALLINE IN STRUCTURE

Plastics, consisting of polymers mixed with small amounts of other additives, are classic non-conductors. When asked about their prominent properties, most people will put the electrical inertness of common plastics at the top of the list. After all, electrical insulation tape, screw driver grips and mains plug bodies are unthinkable without plastics coming to mind.

Consisting of long chains of carbon atoms strongly bonded to each other, with hydrogen atoms connected sideways, plastics have no free electrons to carry an electric current. No wonder then the by chance discovery of a conducting plastic material at the Tokyo Institute of Technology during the 1970s came as a great surprise. It soon became clear that the conducting form of plastics belonged to a special class of polymers called conjugated polymers. These are distinguished by having alternating single and double bonds in the polymer chain. It turned out that adding small amounts of chemical activating agents called 'dopants' could render such polymers conducting.

The basic mechanism of conduction in conducting polymers is not difficult to understand. Graphite is a well-known electrical conductor that has an alternating system of single and double bonds. Such a bond structure contains one unbonded electron per carbon atom. This electron is free and can carry an electric current.

Structurally, graphite consists of sheets of such 'conjugately' bonded carbon atoms, called graphene, stacked one on top of the other. The free electrons exist as a continuous sheet of their own between every two sheets of carbon atoms. Because each sheet extends from one end of a graphite crystal all the way to the opposite end, any sample of graphite exhibits good electrical conductivity.

The peculiar structure of graphite is called a 'layer lattice' and shows much higher conductivity in the plane of the carbon sheets than perpendicular to it. Recently, individual sheets of graphite have been isolated. Such isolated graphene sheets show astounding electronic properties, including superb electrical conductivity that is at par with or better than that of conventional semiconductors.

Bulk samples of conjugately bonded polymers, on the other hand, show extremely small conductivities unless doped by special chemicals. This is because plastics are essentially a disordered assembly of polymer chains, each a few microns long. Unlike graphite, there is no continuous conduction pathway from one end of a sample to the other, resulting in a vanishingly small conductivity.

Addition of suitable dopants creates more electrons by various mechanisms. Some of these extra electrons are capable of jumping from one polymer chain to an adjacent chain thus making inter-chain conduction possible. This then turns any bulk sample of a conjugated polymer conducting.

It is also possible to dope these materials with chemical species that produce electron vacancies or 'holes' that also are capable of carrying current. The value of conductivity is fairly easy to control through the amount of dopant added to the polymer. This now begins to look more like ordinary semiconductors, like silicon, that are routinely doped to produce electron rich (n type) and electron deficient (p type) material.

Functionally, these materials are, in fact, semiconductors, even though they are organic polymers and are not necessarily crystalline in structure. **Figure 1** is a schematic illustration of a short segment of a chain of polyacetylene – one of the first conducting polymers

PLASTIC

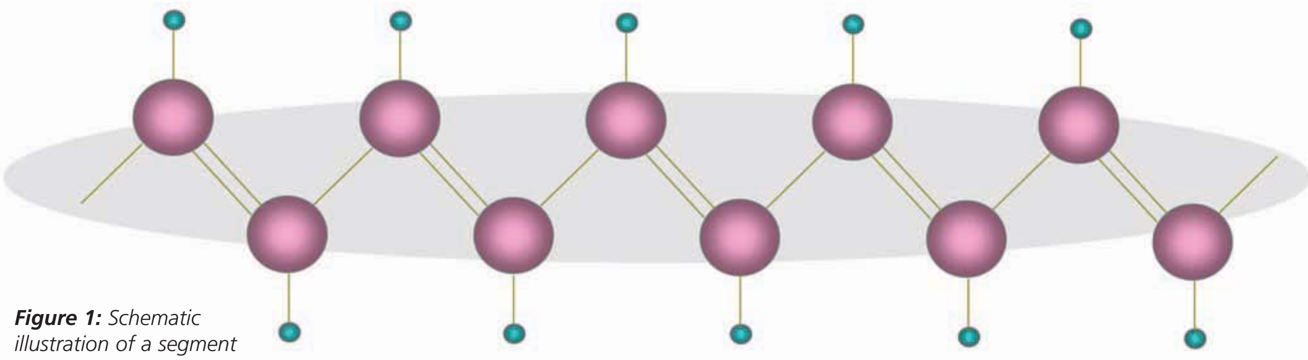


Figure 1: Schematic illustration of a segment of a polyacetylene chain

discovered. The large spheres stand for carbon atoms and the small ones for hydrogen atoms. Alternating single and double bonds between carbon atoms give rise to a concentration of mobile charge carriers; depicted in the image as a grey cloud.

Polymeric semiconductors possess bandgaps and can be doped n and p type just like the more well known inorganic semiconductors. However, on a more fundamental level, the physical origin of their bandgap and the mechanism of doping are very different from that of conventional semiconductors. Recent studies have shown that it is possible to selectively inject either electrons or holes in organic semiconductors using appropriate metallic contacts, without even doping the material.

While work continues to be done to gain a better understanding of the properties of this intriguing class of materials, scientists and engineers are also busy developing useful devices from semiconducting plastics.

Due to their unique properties, for nearly a century, plastics have been an indispensable part of modern life. No other material is quite as adaptable where it comes to creating objects for everyday use. This versatility stems from the myriad ways in which plastics could be processed and the experience has not been lost on those seeking to create novel electronic items from conducting plastics.

Forward-looking companies all over the world have realised the potential inherent in creating devices from a medium that combines the outstanding properties of polymers with the conducting abilities of metals and semiconductors. Researchers in both academic and industrial labs have succeeded in making components like diodes and transistors with semiconducting plastics like polyaniline and polythiophene. The structures of these devices are very similar to that of their counterparts made from conventional semiconductors. What is different, however, is the way devices are constructed with polymers.

Unlike silicon electronics, where discrete and integrated devices are made in elaborate and expensive fabs using high temperature processing techniques and precision patterning equipment, organic microelectronic devices are made using solution processing techniques utilising relatively inexpensive equipment. Polymers are simply dissolved in solvents and then coated by dipping, spinning or spraying on glass or plastic bases to form multi-layer structures.

Unfortunately, photolithography – the technique used for defining conventional device structures, doesn't work with polymers, and alternative methods like stamp embossing have to be used for structure definition. In this technique, polymers are pressed with a 'stamp' made from a hard material, such as diamond or silicon carbide, carrying a replica of the desired pattern in relief. If the right processing conditions are employed, then the design is transferred to the polymer which could be processed further.

Over the past few years, several companies, most prominently Philips, have demonstrated simple integrated circuits made from various polymer combinations. Given the flexibility of plastics in thin layers, this opens up the possibility of flexible electronics if the electronic polymers are deposited on sheets of insulating plastic, used as a non-rigid substrate. Such flexible active integrated circuits have, indeed, been fabricated and these continue to work even when bent through substantial angles.

Perhaps the most talked about application of semiconducting polymers is in various light emitting devices. Since 1996, when the first polymer light emitting diode was demonstrated at Cambridge University, a great deal of effort has been put in the further development of this technology.

It is fairly easy to synthesise photo-emissive polymers that can emit light of any wavelength from infrared to violet. Until a few years ago, a conjugated polymer called poly-

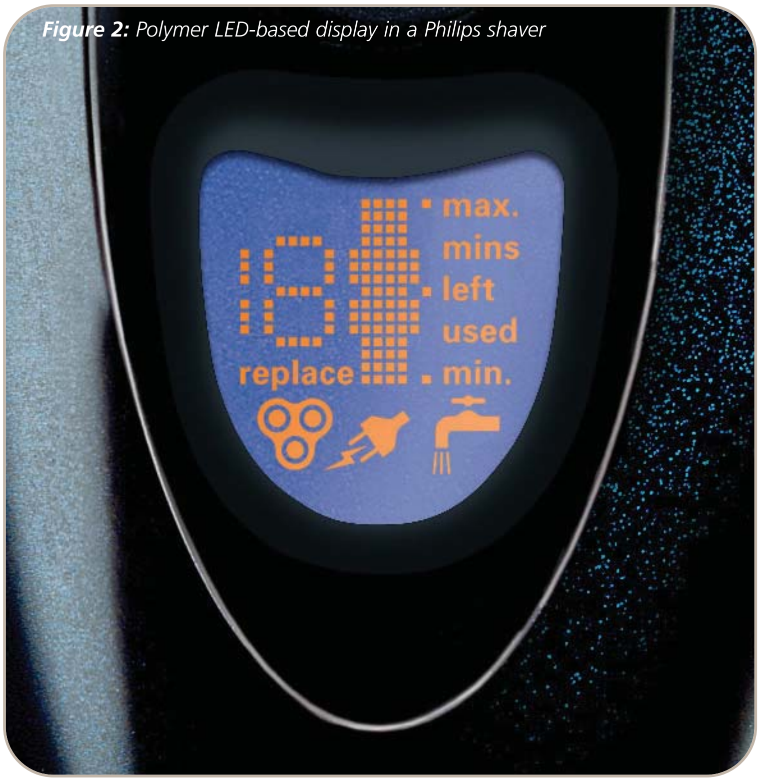


Figure 2: Polymer LED-based display in a Philips shaver

para-vinylene (PPV) was the only material known to emit significant amounts of light when structured as a PN-junction diode, but several other families of conjugated polymers capable of light emission have now been developed. LEDs made from such materials emit beautiful light in vibrant saturated colours.

Initially, these devices exhibited limited lifetimes and suffered from poor efficiency. Intensive research, carried out to understand degradation mechanisms in polymer LEDs, has resulted in much longer lasting devices. Concomitantly, the brightness of these LEDs has also climbed to levels comparable with that of conventional light emitting diodes. These improvements have resulted from the replacement of corrosion-prone calcium cathodes used in older polymer LEDs with more robust lithium-calcium alloys and other materials. Better sealing techniques have also helped and, now, state-of-the-art polymer light emitters boast lifetimes of several thousand hours.

THOUGH THE WORLDWIDE PRODUCTION OF PLASTIC-BASED ELECTRONICS IS RELATIVELY SMALL AT THIS TIME, IT IS CLEARLY SET TO GROW OVER THE COMING YEARS

Dot matrix displays have also been constructed and some have made their way into commercial products like the CXG shaver from Royal Philips Electronics (see **Figure 2**). Similar displays are appearing in other mobile and handheld gadgets like cell phones and personal digital assistants (PDAs).

Compared to longer established displays like backlit monochrome and full-colour liquid crystal displays, the new organic LED-based displays offer sharper colours and all-angle viewing characteristics. The drive circuitry needed for running these displays is virtually identical to that used with their older cousins so that replacing organic LED (OLED) displays for LCD displays is straightforward. The light weight and low power consumption of these devices have made them the favourite of personal immersive display technology companies like eMagin that is already selling impressive 2D and 3D visors for watching films and participating in computer games. The company says that its products are only possible because of the outstanding properties of OLED display panels. Due to their exceptional properties, automakers are also investigating the possibility of incorporating them in vehicle dashboards.

These developments are only a precursor to the emergence of large-sized, full-colour, OLED, flat-panel based displays.

Several companies including Toshiba, Samsung, Kodak, Sanyo and Philips are actively working on bringing such displays to market as computer monitors and flat-panel TVs. An experimental polymer LED panel from Philips can be seen in **Figure 3**.

Active matrix displays like these incorporate one or more Thin Film Transistors (TFT) per pixel, manufactured on the same substrate. Self-luminescent OLED displays, which offer outstanding viewing characteristics, including high contrast, wide viewing angle and fast response times, are widely seen as the leading candidate for the next generation of thin, lightweight displays.

Because the materials used for both LEDs and TFTs are the same and these devices are produced together so OLED TFT displays should eventually be cheaper than other competing variants. These will then provide another alternative to a large and rapidly growing market that is currently based on LCD and plasma display panels. With the race to bring the first OLED-based TVs to the market heating up, all indications are that we are likely to see much increased use of organic displays in consumer electronic products in the coming years.

Displays are by no means the only application for electronics crafted from plastic media. Another area where these materials hold promise is that of solid state lighting. Whereas it is difficult to scale up conventional LEDs to produce more light, organic LEDs naturally lend themselves to manufacturing in the form of large area panels. It is quite easy to coat light emitting polymers from solution onto plastic backing sheets and, thus, produce extended emitting area devices at competitive prices. Flat LEDs like these are being developed with stacked red, green and blue emitting polymers in order to generate white light. Tailoring the exact composition and thickness of polymer layers allows the manufacturer to tune the colour temperature of the emitter so that various desirable shades of white light could be produced.

Another approach is to use phosphors embedded within polymers to convert the polymer's own emission into output light of a different colour. This is considered a more viable route for producing white LEDs. Phosphors made from special nano-scale materials called quantum dots are being investigated for this application. Lighting companies are keen to explore such developments for space lighting applications and at least one company, General Electric in the US, has produced prototype OLED lamps.

Other companies like Osram and Philips are also likely to follow suit. The wide area luminaries provide a fundamentally different lighting experience as compared to inorganic LEDs which are point-like emitters. The possibilities that this offers to illumination designers will be a subject of great interest in the years to come.

While polymer-based devices are good at emitting light, these are also capable of detecting it. Diodes based on organic semiconductors can sense radiation in much the same way as existing photodiodes and phototransistors.

Work on polymer photodetectors has recently picked up with the demonstration of a flexible scanner at Tokyo University. Made entirely on a flexible plastic substrate, the novel scanner has an array of both photodiodes and interface transistors integrated together on a thin backing sheet. The whole scanner is transparent and works by using the ambient light to read any document that is put in contact with it.

Obviously, this kind of a product is only possible with plastic electronics and hints at the unique functionality inherent in electronic polymer-based devices. Yet another area where the use of large-scale polymer photodetectors is being investigated is that of light-to-electricity conversion. Polymer-based solar cells have been shown to work at efficiencies of up to 12% and whereas this figure is low compared to that of the best available mono-crystalline and triple-junction gallium arsenide solar cells (20%-28%), they offer much larger energy-to-cost ratios.
This is because with polymer solar cells, very large area units could be built quite cheaply. These devices are being tested in labs for their ability to stand up to harsh environmental conditions – a key requirement for large-scale deployment.

It is conceivable that over the next few years they will start to appear in commercial solar panel units, providing yet another alternative for clean and renewable generation of electric power.

High-end devices are not the only ones being developed with the new material system. Conducting polymers have already found their way in passive components. Trimmer potentiometers and electrolytic capacitors that employ polymers are commercially available now, of which the former use polymer resistive elements for high reliability and low noise operation. The latter use a polymer electrode as a replacement for the liquid electrolyte electrode used in traditional aluminium electrolytic capacitors.

The new capacitors offer much longer life times and significantly lower Equivalent Series Resistance (ESR), making them especially attractive for filtering and high frequency bypass applications. Other uses include conductive tracks on printed circuit boards, corrosion protection coatings for ferrous metals and spray-on shielding against electromagnetic interference and electrostatic discharge prevention.

Batteries are also a significant area for electronic polymer applications, where polymer electrodes provide improved characteristics over metal electrodes. Lithium polymer batteries are an outstanding example of this kind of usage, providing higher power densities in smaller and lighter units. Fuel cells will also benefit from porous conducting plastics that promise to convert chemical energy into electrical energy in a more efficient way.

This article has looked at a number of application areas for which semiconducting organic polymer-based devices are being developed. The unique properties of this class of materials have generated much interest in recent years with more and more companies getting into plastic-based electronic devices and system.

Though the worldwide production of plastic-based electronics is relatively small at this time, it is clearly set to grow over the coming years and who knows what new products will be enabled by this intriguing system of carbon-based materials in the future.

Figure 3: OLED-based flat panel display undergoing testing at Philips Research Labs

THERE'S SOMETHING they aren't telling you

by Myk Dormer

Take a look in any ISM band module data sheet and, after the predictable text claiming ease of use, long range and low cost (and the sections detailing the pin-out and user interface), you will find an apologetic few paragraphs listing the actual radio specifications.

This is the interesting bit. Radio specifications and the methods used to measure them could occupy a small book, but here I plan to focus on a few important measurements which are all too often omitted. Most of these are receiver parameters. This is because the commonly encountered ISM or 'low power' radio approval specifications, such as EN300-220, AS/NZ 4268:2003 and the American FCC regulations, define the transmitter parameters fairly closely in terms of power output, spurious radiation and frequency accuracy, to guarantee a minimum of interference, both within the ISM bands and with other services. Unfortunately, the receiver performance limits under these approvals are far less stringent.

Let's consider a few important parameters:

- **Sensitivity**: The signal level. This is given in dBm or, sometimes, microvolts. Usually 0.1% or 1% errors at a stated data rate and format, or an equivalent signal to noise or SINAD level.

Always look for a sensitivity measurement which relates to your final application data rate: typical wideband ISM receivers can achieve -115dBm for 12dB SINAD in a 3kHz audio bandwidth, but the same radio at 64kbit/s is unlikely to exceed -105dBm for 1% errors. And it's the sensitivity with your data format that will define your eventual range, assuming none of these following limits are exceeded.

In the case of units with programmable channel widths and data rates, the sensitivity, and, therefore, range claims are frequency given for the slowest data stream, and hence narrowest bandwidth.

- **Adjacent channel**. The ability of the receiver to resist interference from a carrier in the next channel.

These 'rejection' specifications are expressed as a ratio (always in dB) between the level of the interferer on the adjacent channel and the wanted signal, at the point when the wanted signal has been jammed. It is usual to conduct these tests at 3dB above the quoted limiting sensitivity level.

- **Spurious rejection**. The resistance to interference on certain, specified frequencies.

These should be points where the receiver is particularly susceptible, such as 'image' (carrier frequency ± intermediate frequency x 2), 'half IF' (carrier ± intermediate frequency x 0.5) and for dual conversion designs 'second image' (carrier ± second intermediate frequency x 2).

A responsible designer will usually test for spurious responses at multiples of the reference and processor clocks too, as well as looking for intermediate frequency filter related stopband response defects or 'humps'.

- **Blocking**. The resistance to interference at a series of arbitrarily defined frequencies (normally ±1, 2, 5 and 10MHz from the carrier).

The final catch-all specification, this relates more to the device's large signal handling ability and the quality of the filter stop-bands, rather than known 'weaknesses' in the design.

- **Intermodulation**. The resistance to multiple interferers, at frequencies which can result in a mixing product on-channel. Normally tested with a pair of interferers at ±n and ±2n from the wanted carrier. ('n' must be equal to or greater than the unit channel spacing. Typically, this test is conducted with n = 2).

This is an exacting test of receiver large signal handling capability, particularly in the RF amplifier and mixer stages, and is usually where simple, low current designs fail. Unfortunately, in a crowded band, it is intermodulation that eventually limits overall performance, to the point that frequency allocation plans for multiple channel systems are normally arranged to minimise possible intermodulation effects. In fact, there is a whole market for software which automates this 'intermodulation planning' process.

But why does it matter?

The receiver sensitivity defines range (with a given power transmitter, at a given channel bandwidth, with given aerials) but only in the absence of other signals. In

> "The receiver sensitivity defines range but only in the absence of other signals. In the real world, it is the receiver 'rejection' parameters that will limit performance"

Table 1:

	PMR	'class1'	'class 2 '	'single chip'
adjacent	70dB	60dB	(no spec)	30-45 dB
spurious	70dB	60dB	(no spec)	40dB
intermod.	65dB	(no spec)	(no spec)	30dB
blocking	90dB	84dB	30-60dB	50-60dB

the real world, it is the receiver 'rejection' parameters (adjacent, the various 'spurious' and blocking) that will limit performance. A certain degree of planning can minimise the effects of poor intermodulation and, maybe, adjacent channel rejection, but deficient spurious, and especially blocking, performance cannot be ignored.

In these cases other transmitters on other frequencies will compromise your system. There is no use in having good enough sensitivity to reach out ten miles if a garage door opener in the next street can block you. If your supplier isn't quoting a particular parameter, then it's usually because it's embarrassingly poor.

Lastly, here are a few example figures. In **Table 1** are the basic receiver specifications for radios in the PMR (radio telephone) bands under EN300-086, for safety critical class 1 and class 2 ISM band radios approved to EN300-220, and one of the better 'single chip' receiver designs.

Note 1: The term 'dBm' refers to 'power in dB relative to 1mW'. The impedance should be quoted also, but this is usually 50Ω.
Power (dBm) = 10 x log (power in milliwatts), or
Power (dBm) = 10 x log (20 x (rms voltage across 50Ω)^2)
0dBm = 1mW + 27dBm = 500mW – 118dBm = 0.28µV

Note 2: SINAD is a commonly used measurement applied to receiver audio. It is the ratio (expressed in dB) between (signal + noise + distortion) and (noise + distortion). This is measured by applying a carrier, modulated with a 1kHz sinewave, to the receiver. The recovered audio is filtered, which defines the audio bandwidth, often 300-3400Hz, or sometimes a narrower 'psophometric' filter with a particularly shaped passband, and then fed through a narrow 1kHz notch filter. The total power levels before and after the notch filter are compared. 12dB SINAD is about the noise limit for comfortably intelligible speech.

There are many measurement instruments (dedicated 'sinadders', communication test sets) that automate this measurement.

Myk Dormer is Senior RF Design Engineer at Radiometrix Ltd
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....PB-FREE REFLOW EFFECT ON...

N. BLATTAU AND **C. HILLMAN** FROM DFR SOLUTIONS, **C. WIEST** FROM THE UNIVERSITY OF MARYLAND, AND **J. WRIGHT** AND **R. SCHATZ** FROM NIC COMPONENTS IN NEW YORK PRESENT THEIR INVESTIGATION OF THE SUSCEPTIBILITY OF V-CHIP CAPACITORS TO CASE DEFORMATION WHEN SUBJECTED TO PB-FREE REFLOW CONDITIONS

In addition to a change in materials, the movement to Pb-free will also result in significantly higher temperatures during surface mount assembly. Initial attempts at surface mount reflow have noted the initiation of case deformation aluminum liquid electrolytic capacitors in a surface mount configuration (V-chip). Of special concern is not only the visually observable change in component dimensions, but the possibility of damaged capacitors escaping into the field and inducing widespread field returns.

To assess this issue, a wide range of V-chip capacitors were subjected to two series of experiments. In the first set, capacitors of various size and electrolyte formulation were exposed to a modified reflow condition, where the peak temperature was maintained until deformation of the aluminum housing was observed. Based on time to deformation determined in this experiment and the reflow parameters defined in J-STD-020C, a more limited population of V-chip capacitors were exposed to simulated reflow conditions and then subjected to highly accelerated life tests. Results suggest that the strongest predictor of deformation is the volume of the capacitor, with the smallest and largest case sizes having the potential to deform during reasonable Pb-free reflow conditions.

When exposed to elevated temperature conditions designed to accelerate electrolyte evaporation, V-chip capacitors showed limited differentiation in time to failure as a function of reflow conditions or the presence or absence of case deformation.

Figure 1: Aluminum liquid electrolytic capacitor, on the left-hand side, that has experienced case distortion after an extended time at Pb-free reflow temperatures

Key Concerns

The transition to Pb-free manufacturing, to ensure compliance with RoHS legislation, has resulted in substantial concern over the possibility of unknown reliability issues in product released to customers. One particular area identified has been the observation of bulged or deformed surface mount aluminum liquid electrolyte capacitors (aka V-chip) subjected to temperatures recommended for Pb-free reflow (see **Figure 1**). Initial reports have simply been limited to observation, with little to no quantitative information available on process guidelines or degradation in capacitor performance.

The purpose of this investigation was to provide the industry with an initial accounting of the susceptibility of V-chip capacitors to case deformation and assess the potential for potential reliability issues after exposure to Pb-free reflow conditions.

Ranges of electrolytic capacitor part types were selected to investigate the influence of electrolyte formulation, capacitor dimensions and rated voltage. A listing of the various capacitors used in this investigation is provided in **Table 1**. All capacitors were subjected to an initial inspection to identify any potential anomalies or defects. No anomalies were identified.

Reflow Sensitivity Level (RSL) describes the potential

...ELECTROLYTIC CAPACITORS

for surface mount aluminum liquid electrolytic capacitors to experience deformation or degradation when exposed to elevated reflow temperatures for extended periods of time. To assess reflow sensitivity, the capacitors were subjected to a modified Pb-free reflow profile.

Series	Capacitance (F)	Tolerance	Voltage (VDC)	Size (h x d) (mm)	Rated Lifetime
NACETP[1]PT	10	20%	16	3X5.5	2000 hrs at 85ºC
NACE	1	20%	50	4X5.5	2000 hrs at 85ºC
NACE	10	20%	50	6.3X5.5	2000 hrs at 85ºC
NACEWTP[2]PT	100	20%	16	6.3X5.5	1000 hrs at 105ºC
NACEW	1000	20%	6.3	6.3X5.5	1000 hrs at 105ºC
NACHLTP[3]PT	33	20%	25	6.3X6.1	5000 hrs at 105ºC
NACE	22	20%	63	6.3X8	2000 hrs at 85ºC
NACE	220	20%	35	8X10.5	2000 hrs at 85ºC
NACTTP[4]PT	47	20%	35	8X10.5	1500 hrs at 125ºC
NACT	220	20%	25	10X10.5	1500 hrs at 125ºC
NACE	330	20%	50	12.5X14	2000 hrs at 85ºC
NACE	3300	20%	16	16X17	2000 hrs at 85ºC

Table 1: *Surface mount electrolytic capacitors subjected to reflow sensitivity analysis*

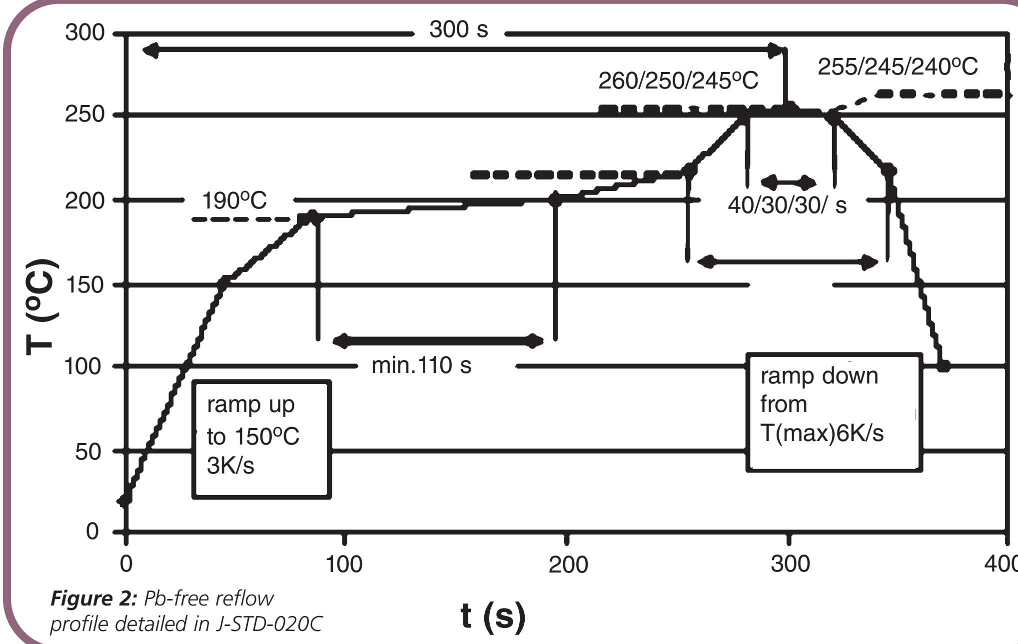


Figure 2: *Pb-free reflow profile detailed in J-STD-020C*

Experimental Procedure

Three capacitors of the same part type were placed in a reflow simulation chamber. The temperature in the chamber was regulated using a K-type thermocouple, which was attached with Kapton tape to one of the capacitors on the top of its aluminum housing.

The time/temperature behaviour was adjusted to ensure a preheat and ramp rate that was representative of a Pb-free reflow profile. The specific parameters chosen were based upon IPC/JEDEC J-STD-020C Moisture/Reflow Sensitivity Classification for Non-hermetic Solid State Surface Mount Devices (see **Figure 2**). Capacitors

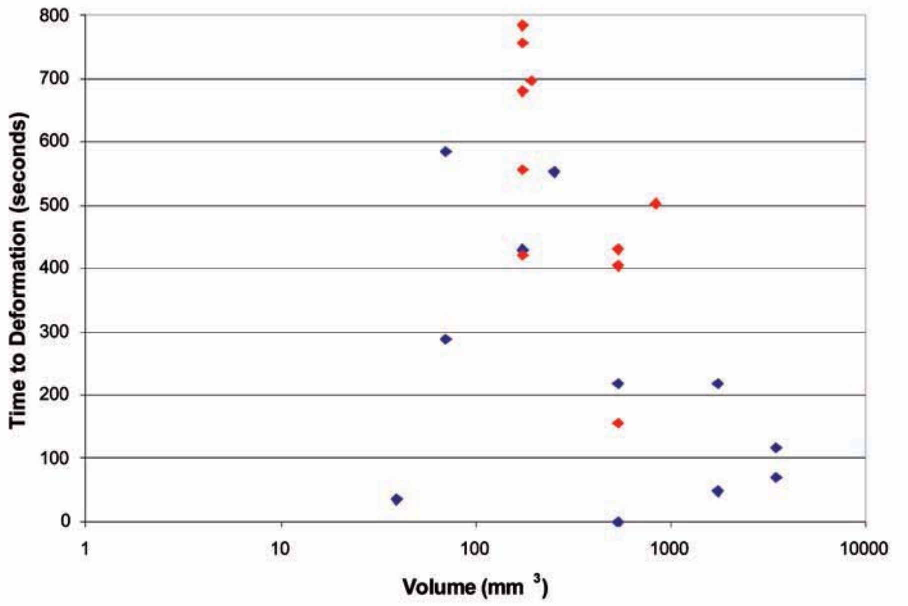


Figure 3: *Time to deformation as a function of capacitor volume at 235P^{oP}C peak reflow temperature. The diamonds in red are indicative of extended temperature or extended life capacitors.*

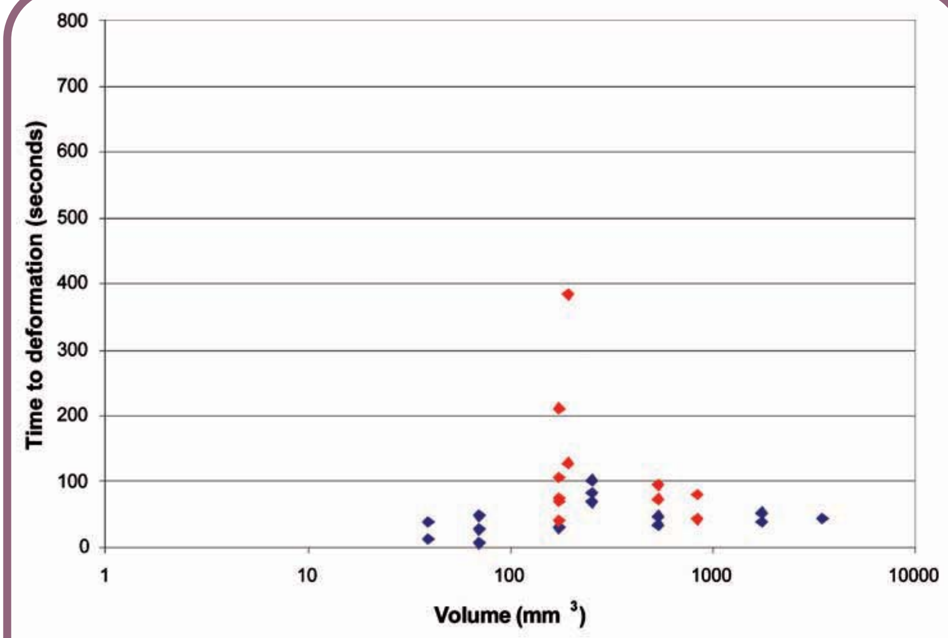


Figure 4: *Time to deformation as a function of capacitor volume at 260P^{oP}C peak reflow temperature. The diamonds in red are indicative of extended temperature or extended life capacitors.*

were ramped up to a preheat temperature of 190P^{oP}C at a rate of 3P^{oP}C/sec. This temperature was maintained for 120 seconds and increased to the peak temperature at a rate of 3P^{oP}C/s. The capacitors were then held at the peak temperature until physical deformation of the alumi-nium housing was observed.

The peak temperatures chosen were 235P^{oP}C and 260P^{oP}C. These temperatures were based upon the minimum recommended reflow temperature for tin/silver/copper (SAC) solder alloys and the maximum expected peak temperature detailed by J-STD-020C.

The results for time to deformation at 235P^{oP}C and 260P^{oP}C peak temperature are displayed in **Figure 3** and **Figure 4**, respectively. One capacitor experienced deformation during the ramp-up to the 235P^{oP}C peak temperature and is, therefore, not displayed in Figure 3.

Discussion

Several trends were identified upon review of the preliminary results displayed in Figures 3 and 4. The time to deformation seemed to be strongly dependent upon the volume of the capacitor, with the maximum time to deformation observed to occur at moderate volumes (100-500mmP^{3P}). The smallest and largest capacitors seemed to be most prone to deformation, with the smallest capacitors at 235P^{oP}C and several more capacitors at 260P^{oP}C experiencing deformation before the 40-second hold time defined in J-STD-020C. This may suggest that the smallest and largest surface mount aluminum liquid electrolytic capacitors could experience deformation in more severe Pb-free reflow conditions.

The dependency on volume is expected based upon the steps involved in case deformation. The first step is introducing a sufficient amount of energy into the system to raise the temperature beyond the boiling point of the electrolyte. The temperature on the outside of the capacitor, where the thermocouple was placed, can be considered the energy flowing into the system. The heat capacity of the electrolyte, which relates energy introduced into the system to a ΔT, is dependent upon the moles of electrolyte present in the system. Therefore, for a given outside temperature, the smaller capacitors will equilibrate with the ambient conditions more rapidly then larger capacitors.

As the boiling point is reached, the vapour pressure will increase rapidly. The larger the volume of liquid present, the larger the pressure is within the cylinder

Sample Set	Part Number	Reflow Conditions		
		Benign	As per J-STD-020C	Severe
Volume < 350 mm^3	NACE1R0M50V4X5 NACE220M63V6.3X8	235°C / 30 sec	250°C / 30 sec	260°C 30 sec
			250°C / 40 sec	
Volume > 350 mm^3	NACE331M50V12.5X14	235°C / 30 sec	245°C / 30 sec	260°C 30 sec
			245°C / 40 sec	
Extended Lifetime	NACHL330M25V6.3X6.1	N/A	N/A	260°C 20 sec

Table 2: *Experimental design selected for assessing long-term degradation*

for a given temperature. The case resisting the pressure, the surface area, scales with dimensional unit squared. Since the pressure on the can inducing deformation scales with dimensional unit cubed, it can be seen how a more benign time/temperature environment within the larger electrolytic capacitors would induce earlier deformation.

In general, extended temperature or extended lifetime capacitors displayed a less severe sensitivity to reflow conditions, with all extended temperature/lifetime capacitors having a time to deformation exceeding 40s.

While deformation behaviour as a function of reflow temperature and capacitor volume was demonstrated, of greater concern is the possibility that capacitors that did not experience deformation may have experienced some degree of unobservable damage or degradation that would result in a limited lifetime.

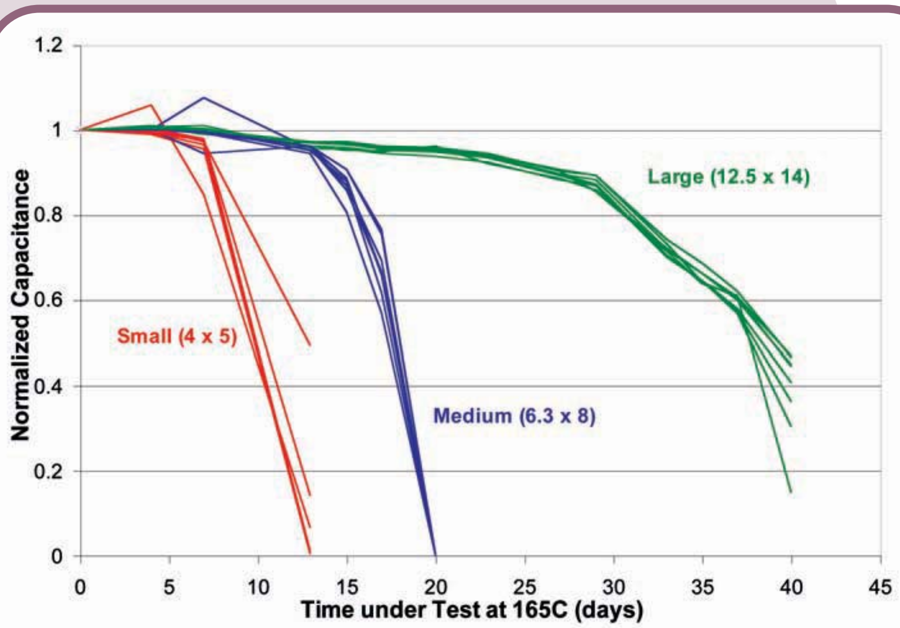


Figure 5: *Change in normalised capacitance as a function of time-under-test. Changes in degradation behaviour by volume (small, medium, large) can be observed*

Long-Term Degradation

To assess the potential of degradation during exposure to Pb-free reflow conditions, a limited sample set of capacitors listed in Table 1 were subjected to a range of peak reflow temperatures and hold times. Due to the behaviours observed during the first set of experiments, the sample set was primarily based on volume, with an extended lifetime part also selected for comparison purposes. The experimental design is detailed in **Table 2**. Three capacitors from each part number were exposed to each reflow condition.

After exposure to the reflow conditions detailed in Table 2, the capacitors were subjected to accelerated test conditions. The accelerated test conditions used to assess degradation behaviour were based on previous experiments performed on electrolytic capacitors, which determined that testing to industry specifications would be insufficient to assess long-term reliability. The test temperature was selected to ensure capacitor failure within a reasonable time period, without inducing inappropriate failure modes. In addition, since damage to the seal during reflow was the primary concern, a high temperature was desirable to ensure that evaporation of the electrolyte was the mechanism that induced capacitor failure.

Given these requirements, the test conditions were set at 25VDC at 165P^{oP}C. The applied voltage helped ensure maintenance of the dielectric without inducing dielectric breakdown. The elevated temperature, while above industry and company specifications, was significantly below the boiling point of the liquid electrolyte (180P^{oP}C to 200P^{oP}C).

Capacitors were periodically pulled from the environmental chamber for capacitance and Equivalent Series Resistance (ESR) measurements. The results are displayed in **Figures 5** and **6**.

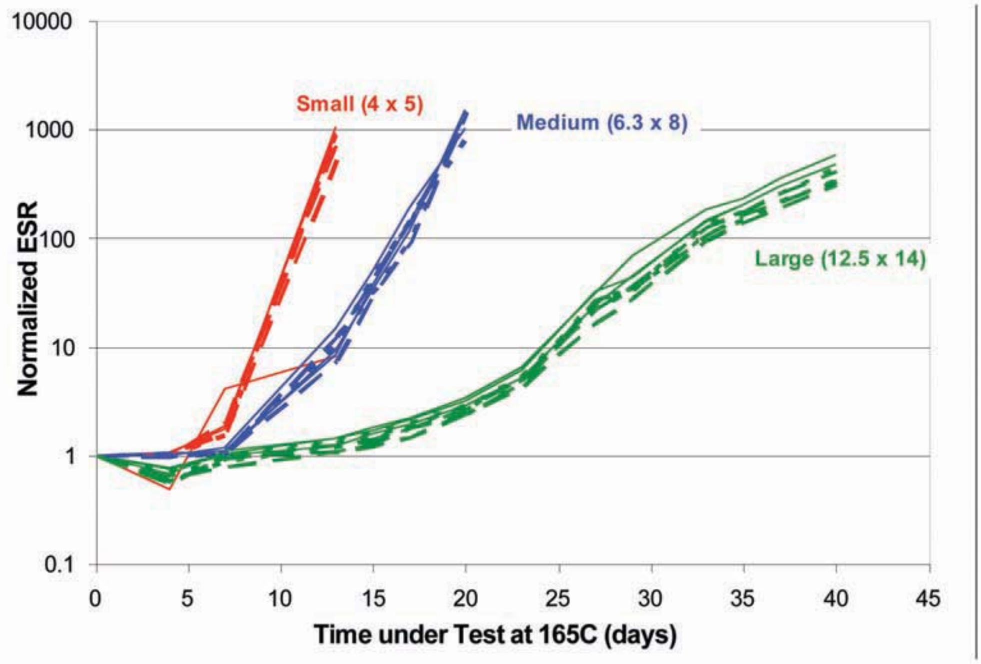


Figure 6: *Change in normalised Equivalent Series Resistance (ESR) as a function of time-under-test. Changes in degradation behaviour by volume (small, medium, large) can be observed*

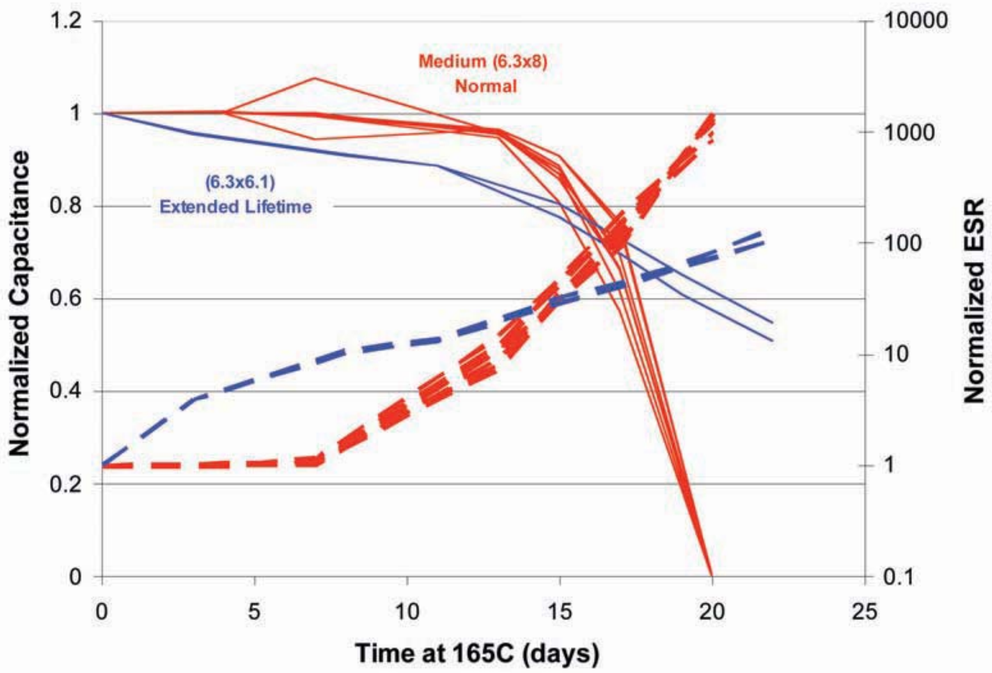


Figure 7: *Comparison of degradation behaviour of general purpose and extended lifetime capacitors*

Long-Term Degradation?

The results show that within normal variations, the reflow exposure conditions had no influence on long-term degradation behaviour. This observation was found to be true even when the capacitor was observed to have experienced case deformation after Pb-free reflow simulation. The primary influence on time to failure under accelerated test conditions was determined to be capacitor volume. Extended lifetime capacitors were found to have similar degradation behaviour to the general-purpose capacitors, as shown in **Figure 7**.

Size Dependence

The potential for case distortion in V-chip capacitors during Pb-free reflow was found to primarily be dependent upon the volume of the capacitor, with small, less than 100mmP3P, and large, greater than 1000mmP3P, showing the greatest degree of susceptibility. Capacitors with extended temperature or extended lifetime capacitors in general showed more robust behaviour.

Exposure to a range of Pb-free reflow conditions and the occurrence of case deformation seemed to have minimal influence of the reliability of the V-chip capacitors in environments designed to accelerate electrolyte evaporation, the most common root-cause for electrolytic capacitor failure in the field.

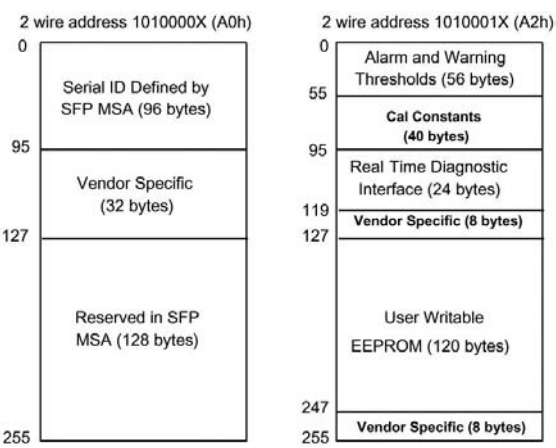
Integrated Diagnostics for SFP OPTICAL TRANSCEIVERS

Fibre optic transceivers are used to provide electrical to optical (E/O) conversion required for common long-distance transmission media. With each new generation of optical transceiver come additional feature sets defined to enable increasingly sophisticated link management. These now include capabilities to measure component parametric and diagnostic information externally through a link management interface.

Diagnostic functions in fibre optic transceivers provide cost-effective performance monitoring – a capability that is becoming widely used to assure reliable, high-performance optical interfaces for mission-critical Fibre-Channel, Ethernet and SONET/SDH applications.

The ability to observe optical transmit and receive parameters is not new. High reliability telecommunication applications have used discrete optical transmitters and receivers with various monitoring features since the 1980s, primarily to assess laser reliability. Similar capabilities began to appear on integrated optical transceivers in the early 1990s.

Figure 1: *SFF-8472 digital diagnostic memory map*



An early example of parametric monitoring can be found on the January 1998 SFF (Small Form Factor) 2x10 "through-hole" MSA (not affiliated with the SFF Committee), which allocates five pins for direct sensing of laser bias current, laser optical power and received optical power. These pins provide an output voltage or current proportional to each device parameter. Since the pins provide an analogue output, the host/system designer had to provide for digitisation and interpretation of sensed values.

A new generation of extended capabilities has been agreed upon for SFP (Small Form Pluggable) transceivers. The Small Form Factor Committee (SFF) has voted upon and accepted document SFF-8472 Digital Diagnostic Monitoring Interface for Optical Transceivers, which defines more user-friendly monitoring and control features for real-time optical power levels, laser bias current and temperature and supply voltage.

New Capabilities And Their Uses

In August 2002, the SFF-8472 multi-source document was completed, which merged capabilities of 2x10 SFF and standard SFP MSAs. The agreement adds to the standard SFP provisions for on-board sensing and digitisation of parametric data, scaling and calibration of the result, and provides a common memory structure to store the information for access over the existing two-wire serial interface. Details of SFF-8472 can be found on the SFF Committee's public website at **www.sffcommittee.org**.

RANDY CLARK OF APPLICATIONS DEVELOPMENT IN FIBRE OPTIC COMPONENTS AT AVAGO TECHNOLOGIES DISCUSSES MONITORING TECHNOLOGY WIDELY USED TO ASSURE RELIABLE, HIGH-PERFORMANCE OPTICAL INTERFACES FOR MISSION CRITICAL FIBRE-CHANNEL, ETHERNET AND SONET/SDH APPLICATIONS

SFF-8472 retains the legacy SFP/GBIC memory map at address A0h, with added provisions for identifying the presence of SFF-8472 feature sets. A second 256-byte memory address (**Figure 1**) was added at address A2h which, in addition to providing parametric sense infor-mation, defines flags for alarm or warning conditions, mirroring of hard pin status, limited digital control capability and user-writable non-volatile cells.

The purpose of SFF-8472 is to aid in predicting transceiver lifetime, isolating system failure and verifying component compliance in field installations. The predictive failure feature allows a host to identify potential link problems before system performance is impacted. Prior identification of link problems enables a host to service an application via "fail over" to a redundant link, or replace a suspect device, maintaining system uptime in the process.

For applications where ultra-high system uptime is required (such as "five nines" or 99.99999%), a digital SFP provides a means to monitor two real-time laser metrics associated with observing laser degradation and predicting failure: average laser bias current (Tx_Bias) and average laser optical power (Tx_Power).

SFF-8472 devices provide real-time access to transceiver internal supply voltage and temperature, allowing a host to identify potential component compliance issues. Received optical power is also available to assess compliance of a cable plant and remote transmitter.

The fault isolation feature allows a host to quickly pinpoint the location of a link, minimising system downtime. For optical links, the ability to identify a fault at a local device, remote device or cable plant is crucial to speeding service of an installation. SFF-8472 real-time monitors of Tx_Bias, Tx_Power, Vcc, Temp and Rx OMA (optical modulation amplitude) can be used to assess local transceiver current operating conditions. In addition, status flags Tx Fault, Tx Disable and Rx LOS (loss of signal) are mirrored in memory and available via the two-wire serial interface.

Component evaluation is a more casual use of the SFF-8472 real-time monitors of Tx_Bias, Tx_Power, Vcc, Temp and Rx_OMA. Potential uses are as debugging aids for system installation and design, and transceiver parametric evaluation for factory or field qualification. For example, temperature per module can be observed in high-density applications to facilitate thermal evaluation of blades and systems.

Identification Of SFF-8472 Compliant SFPs

To determine if an SFP transceiver is SFF-8472 compliant, the user must access the base memory (**Figure 2**) looking to see if digital diagnostics capability

Figure 2: *SFF-8472 address A0h memory contents*

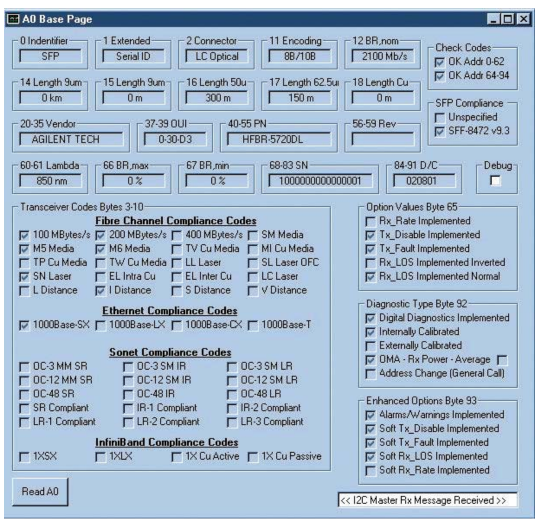
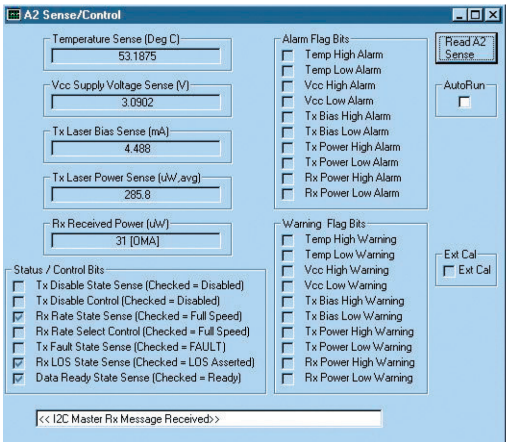


Figure 3: *SFF-8472 address A2h sense related memory contents*



is indicated. Once known compliant, the user identifies which host-based maths algorithm is required to properly scale the real-time monitoring results. Most devices require a simple scaling of each SFP real-time result by a constant defined in SFF-8472. Several additional bits identify which optional “soft” control and mirror func-tions are enabled on the SFP. Each bit will indicate if a feature is present: examples being “soft” assertion of Tx_Disable, mirroring of Tx_Fault and Rx_LOS pin status, and implementation of alarm and warning flags for real-time sense parameters.

Parametric Monitoring Details

SFP applications tend to be high density due to their small external dimensions and availability at 1-2.5Gbps (intermediate) data rates. In these applications, the ease of use associated with an SFP receptacle cage is sometimes offset by the cage’s restriction of air flow required to cool each device. This makes transceiver internal temperature (**Figure 3**) a useful metric for assessing in-situ compliance to electro-optical specifications and long-term reliability expectations. Historically, transceiver temperature was measured in system qualification using thermocouples and could not be monitored in field installations. Temperature will commonly be measured on SFPs using sense circuitry, mounted directly on the internal PCB. These measured values can be used to observe drifts in thermal operating point, be empirically correlated to case in a given host application or used to detect extreme fluctuations due to failure or fault in a system thermal environment.

Pluggable applications add additional uncertainty to the transceiver internal supply voltage due to potential resistive drops across pluggable contacts. This, added to voltage lost across a host filter network, can drop the transceiver internal supply voltage over 100mV from the host supply, creating a potential specification compliance issue. Supply voltage is measured on SFPs using sense circuitry, mounted on the internal printed circuit board. Transmit or receive supply voltages may be monitored for this read-back. These measured values can be used to observe drifts in supply voltage operating point, be empirically correlated to SFP pins in a given host application or used to detect extreme supply voltage fluctuations (that is droop) due to failure or fault in a system power supply environment.

Historically, the most useful transceiver parameter for in-situ reliability prediction has been laser bias current. Common integrated optical transceivers are operated in constant optical power mode, modifying laser bias current with a closed loop feedback network to maintain a fixed launched power. Because laser efficiency and threshold vary with temperature, most SFPs will need to change laser bias current to maintain constant optical power. Drifts in laser bias current not associa-ted with routine variations in temperature or supply voltage can be an indication of pending laser reliability problems. Differentiating between normal bias variations and those due to reliability drifts requires some analysis. At quiescent thermal and voltage conditions no significant drift in bias current should be seen.

Each manufacturer can provide information on absolute maximum and minimum laser bias currents for the particular device technology. System software can compare the value of bias current (I_{Bias}) against these limits, looking for catastrophic laser problems. For finer reliability prediction, comparison of bias current values must take SFP operating conditions into account. At equivalent thermal conditions, increases in I_{Bias} may be indicative of laser degradation. It is recommended that system designers looking for fine reliability prediction consult SFP suppliers for more information.

For observing average transmit optical output power, most SFP transceivers use a low speed PIN diode mounted adjacent to the transmitter’s laser to sample the laser’s optical output power. The resulting PIN current is used by laser drivers to maintain constant optical power and also to facilitate real-time monitoring of average transmitted optical power. Typical SFPs are operated in constant power mode and will rarely exhibit variations in Tx power. Should significant variations in this parameter be seen, it indicates that the SFP may be no longer functioning properly.

For received optical power, the digital SFP specifi-cation provides flexibility for manufacturers to monitor modulated or average optical power. Each has advantages for providing system designers and installers with status information on incoming optical signals. Fibre optic standards themselves are split on the use of OMA (optical modulation amplitude) or average power for compliance specifications. Fibre Channel FC-PI uses OMA for all input optical specifi-cations at 1Gbps and 2Gbps. Ethernet 802.3 at 1Gbps and SONET/ATM OC-48 use average input optical power for their compliance specifications. OMA gives the system information on peak-to-peak magnitude of incoming optical signals. Since system installers can measure average optical power directly off a cable with a handheld meter, supplying modulated power information provides installers and debuggers with an additional tool to verify compliance of a link (cable plant and remote optical transmitter).

Status Indicator And Control Bits

To supplement the traditional SFP hard pin functions, SFF-8472 provides "soft" monitor and control of pinned functions over the two-wire serial interface. Because of increased latency using soft indicators, it is recommended that hard pins continue to be used for fault and LOS interrupts. Many systems hard wire in a logical OR such alarm indicators from transceivers and physical layer ICs. For these applications, soft indicators can be used by host controllers to provide additional fault isolation information. Soft control of Tx_disable and Rate_select can be useful in latency-tolerant applications.

SFF-8472 provides memory locations for optional alarm and warning flags associated with each real-time sensor. Decision making is handled by the transceiver, using factory set thresholds for high alarms, low alarms, high warnings and low warnings. Alarms indicate conditions associated with link fault or significant functionality issues. Warnings indicate potential compliance violations or non-ideal operating conditions. Using these indicators, many system controllers can gain the benefits of real-time monitoring capabilities without having to routinely poll and analyse parametric information. Should an alarm or warning indicator be found, a host can also read real-time parametric information as part of fault isolation.

Two-wire Serial Interface Bus: Atmel Protocol Emulation

As defined in legacy SFP and GBIC agreements, the SFF-8472 serial interface uses the two-wire serial CMOS EEPROM protocol defined for the Atmel AT24C01A/02/04 family of serial EEPROMs. However, SFF-8472 represents the first generation of devices that will generally not use this type of component to directly manage the interface. As a result, there are protocol emulation behaviours that may be important to system designers not explicitly defined in these MSA documents. It is recommended that attention be paid to bus issues as part of accommodating SFF-8472 transceivers in host to transceiver communication links.

SFP transceivers based on SFF-8472 are the first multisourced digital diagnostic capable optical transceivers available for data and telecommunication applications. They provide real-time monitoring information on key electro-optical parameters and important digital control and status features. Emerging 10Gbps form-factors are incorporating related capabilities, ensuring that advanced control and monitoring functionality plays a key role in future generation optical components.

FULL-BAND TUNABILITY

Figure 1: Output modulated power from module; inset shows back-back eye

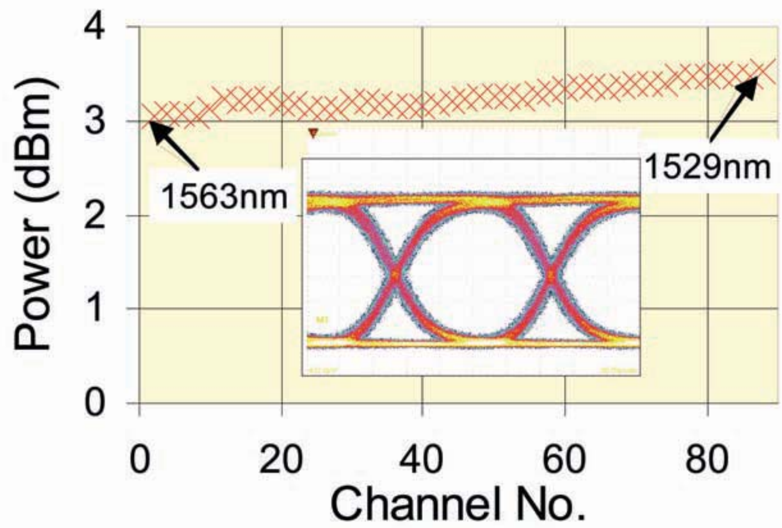
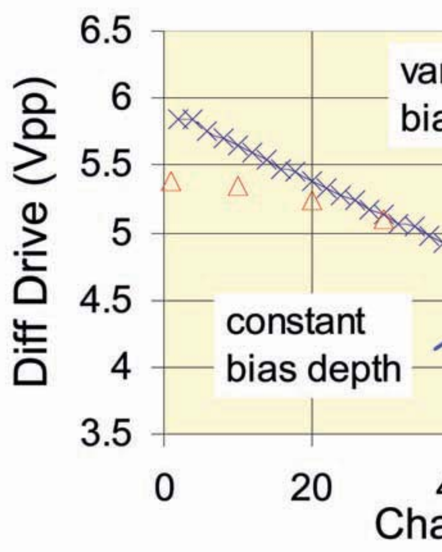


Figure 2: Measured drive amplitude over C-ba



R. A. Griffin, B. Pugh, J. Fraser, I. B. Betty, K. Anderson, G. Busico, C. Edge and T. Simmons **present a compact, high power, MQW InP Mach-Zehnder, full-band tunable transmitter. Variation in modulation characteristics with wavelength intrinsic to MQW InP MZMs is shown to be readily managed**

High performance, small size and low voltage requirements make InP Mach-Zehnder Modulators (MZMs) very attractive for demanding transponder and transceiver applications at 10Gb/s. To satisfy the growing requirement for inventory reduction and automatic wavelength provisioning, however, devices that can operate over multiple DWDM channels while maintaining compact size are required. Here, we report the combination of a full-band tunable DS-DBR laser with MQW InP MZM in a co-packaged transmitter module. We demonstrate high output power and low dispersion penalty over the C-band and show that wavelength variation can be readily managed while retaining a very compact module footprint.

Full-Band Module

The full-band tunable laser source we employ is a Digital-Supermode Distributed Bragg Reflector (DS-DBR) design, a monolithic device fabricated using conventional InP processing. These surface ridge lasers have an MQW active region and bulk phase tuning section. The rear section has an e-beam written phase grating reflector that provides a sharp and flat comb reflectance response. The front section is a linearly chirped Bragg grating that is coarsely tuned to 'digitally' select one of the rear supermodes. Tuning is achieved through carrier injection for complete coverage of C- or L-band, with simplified calibration compared to alternative multi-section DBRs. An integrated semiconductor amplifier (SOA) at the laser output provides high power and allows for straightforward power levelling over channels.

The optical couplers performing the split and combine-function for the MQW InP MZM are configured to give an output null with zero applied volts. The split ratio of the couplers is carefully tailored to produce a controlled DC extinction ratio (ER), providing enhanced transmission performance.

The MQW modulator core consists of twenty InGaAsP quantum wells and twenty-one Q = 1.1μm InGaAsP barriers, producing a steep absorption edge around 1400nm. Large absorption changes with applied electrode field due to the QCSE are converted to refractive index changes in the C-band through the Kramers-Kronig relation, providing an optical phase shift to the propagating mode.

for 10Gb/s DWDM

and for constant and variable bias depths

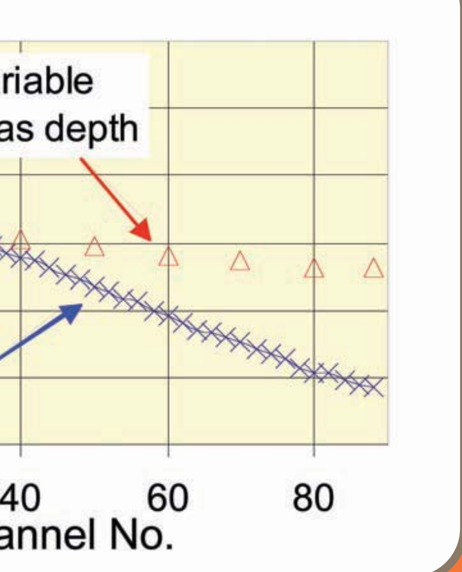
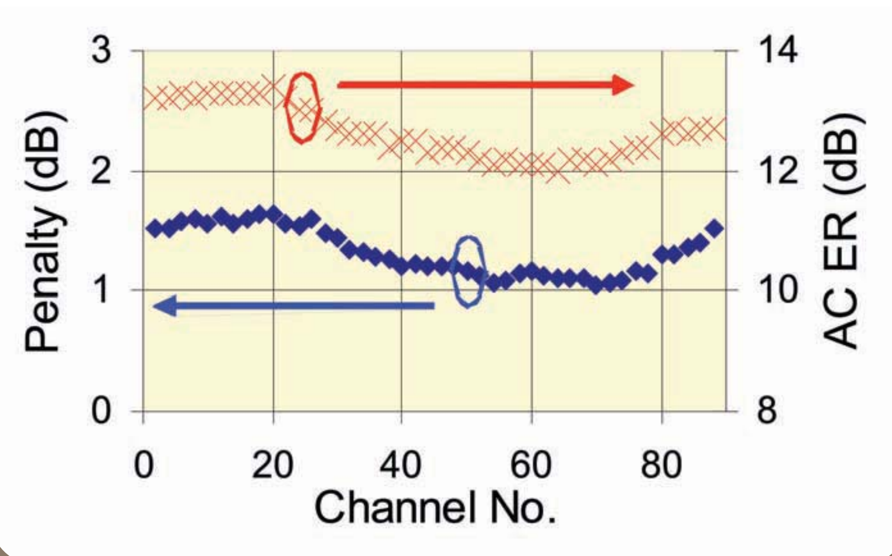


Figure 3: *Measured dispersion penalty and dynamic ER*



Utilising the QCSE enables an extremely compact design with high-bandwidth, lumped-element electrodes only 600μm long, ~ 50 times shorter than conventional lithium niobate MZMs.

Co-packaging of the DS-DBR laser and MQW InP MZM was achieved in a compact module with industry-standard 14-pin footprint (30mm x 12.7mm). Coplanar feed-throughs were used for DC laser and modulator connections, and a coplanar GSGSG arrangement was used for the RF modulation path. The InP MZ was mounted on an Al2O3 carrier with terminated 50Ω coplanar transmission lines that allow independent drive signals to each modulator arm. Differential modulation is compatible with low-cost SiGe or GaAs driver ICs. The small dimensions of the MZM allowed the modulator to be mounted together with the laser on a common Thermo-Electric Cooler (TEC), ensuring high stability. An isolator was included between laser and modulator; anti-reflection coating of the InP modulator chip ensured low back-reflection for the module.

Fine Tuning

A power levelling routine was employed to adjust the module output power. For each of 88 50GHz-spaced channels spanning 1529-1563nm, the MZM bias was adjusted for maximum CW transmission; SOA current was iteratively adjusted for a target power, while the DBR currents were fine-tuned to maintain required wavelength.

Accuracy better than 2GHz was achieved for all channels, with SMSR > 40 dB. Measured optical power under modulation for 100GHz channels is shown in **Figure 1**, demonstrating > +3dBm modulated power for all channels. The slight tilt in power is a result of different bias conditions between levelling and modulated operation. The inset to Figure 1 shows the back-back eye recorded for 10.7Gb/s operation. A consistently clean eye with low jitter and high ER was observed over full-band tuning, with no observable interaction between modulator and laser.

Since the detuning between absorption edge and operating wavelength changes as the source is tuned through the C-band, the modulation efficiency and absorption of the MZ electrodes vary with wavelength. Phase modulation efficiency monotonically decreases with longer wavelengths, as does the residual voltage-induced absorption. While variation may be reduced by increasing the detuning from the absorption band, this approach requires longer electrodes and potentially velocity matching to maintain suitable bandwidth. Our approach is to use a relatively small detuning for high efficiency with short lumped-element electrodes, easing manufacturing tolerances and building on many years of production experience with fixed wavelength MQW InP MZs.

To set the DC bias on the modulator arms, V_{left} and V_{right}, we adopted two alternative strategies. For the simplest approach, we kept the bias depth constant. For each channel, differential bias was adjusted to set 50% crossing point for the output optical eye using a standard control arrangement – applying a low frequency pilot tone to the modulator driver and monitoring the optical output with a low frequency photodiode. Differential drive amplitude applied to the MZM was adjusted for maximum dynamic ER. As shown by the data points in **Figure 2**, for constant bias depth the required modulation amplitude varies in a linear fashion with channel number, with a range from 5.8Vpp to 3.9Vpp. This variation is within the capabilities of widely available low-cost differential driver ICs.

A second strategy was to vary the bias depth linearly with laser channel number to compensate for wavelength-dependent effects. With this approach, the range of modulation amplitude was significantly reduced, with only 0.6V variation over the band.

The non-linear phase-voltage characteristic of the InP MZMs devices generates negative chirp in a differential drive configuration; at quadrature, the left arm is biased more deeply than the right and, hence, the dynamic phase modulation is higher, producing negative chirp and improving transmission performance for standard SMF.

Limiting the ER in a controlled fashion also provides an additional contribution to negative chirp, improving transmission performance relative to high DC ER operation. Dispersion penalty of 44 100GHz-spaced channels was measured for 10.7Gb/s transmission over 1600ps/nm with a BER of 10^{-9}. Applied data was a 2^{31}-1 PRBS; a standard PIN-based receiver was employed, together with a CDR chip-set. As shown in **Figure 3**, dispersion penalty < 2dB was achieved for all channels, while maintaining ER > 12dB.

As for single-wavelength InP MZs, dispersion penalty of each channel was strongly correlated with the DC ER. Any wavelength-dependent trend in penalty was weak (< 0.2dB), confirming that simple DC screening is sufficient to ensure transmission performance over band. A DC ER in the range 17-24dB provides low dispersion penalty while maintaining high dynamic ER.

Promising Approach

Co-packaging a MQW InP MZM together with a DS-DBR laser has realised a full-band tunable transmitter, exhibiting excellent performance in a compact footprint. Modulated output power > +3dBm for all channels was achieved, together with low dispersion penalty for transmission over 1600ps/nm at 10.7Gb/s.

It was shown that variation in modulation characteristics with wavelength can be readily managed, with modulation amplitude variation as low as ±6% across the band. Providing full tunability in a compact footprint while preserving the high waveform quality associated with Mach-Zehnder transmitters is a promising approach to satisfy growing industry demands in metro and long-haul applications.

DESIGN and ANALYSIS of DS-DBR LASERS for C and L-BAND

Wideband tuneability based on Digital Supermode Distributed Bragg Reflector (DS-DBR) laser technology for complete C-Band coverage has been reported before. The DS-DBR laser is a monolithic InP semiconductor laser, which has all the benefits of DFB technology: high yield, low cost, simple packaging with no moving parts or reliability hazards. This paper will present equivalent L-Band performance based on the DS DBR technology platform. The lasers are physically indistinguishable, fabricated using the same process and mask sets, and offer complete and overlapping coverage of both bands, with a combined tuning range in excess of 85nm. Since the C and L-Band device are identical, the same control electronics and packaging can be used for both.

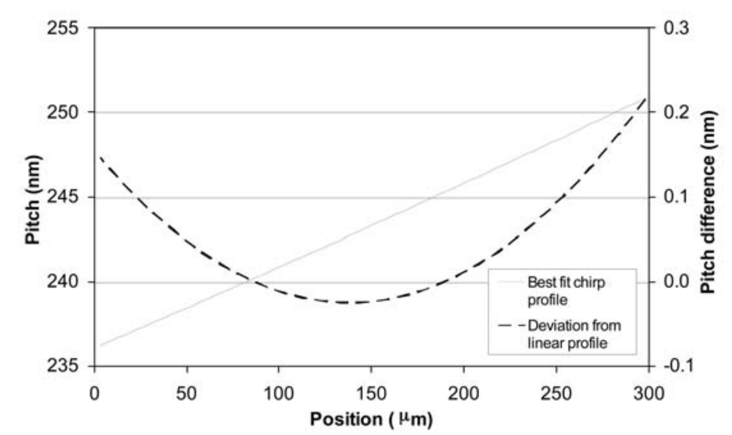


Figure 1: *Front grating chirp rate.* ***Inset:*** *rear phase grating comb reflector*

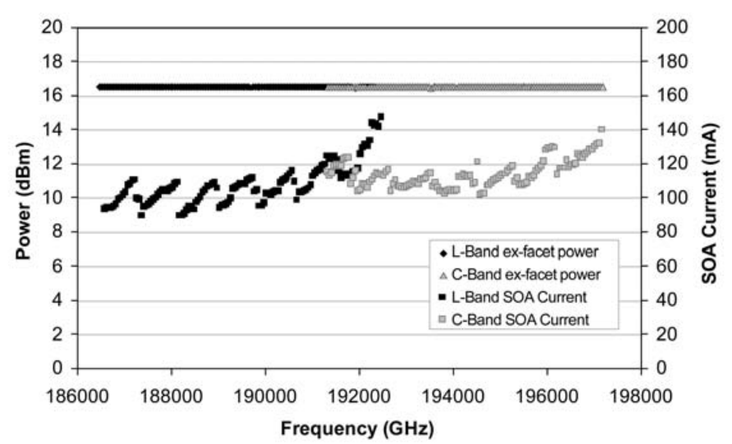


Figure 2: *Power levelled output and SOA current versus channel frequency, for C and L-Band*

Laser Design And Fabrication

Both the C and L-Band DS-DBR devices have been fabricated using the same InP process route. It is a surface ridge waveguide device, with a front chirped grating, rear phase grating, phase section, gain section and monolithically integrated Semiconductor Optical Amplifier (SOA) to boost output power.

The challenge in creating an L-Band version of the DS-DBR is two-fold. First, subtle changes to both the active (gain) and passive (grating) material compositions are required to support a lower frequency of operation and avoid any fall-off in performance. Second, the grating design itself must be translated to lower frequencies. This requires a good understanding of the wavelength dependence of the refractive index of the passive material, which was obtained from analysis of a number of 3-section DBR lasers fabricated to operate across the C and L Bands.

Using this information, the chirped front grating can be designed to correctly produce reflection peaks with the necessary frequency spacing. The optimum chirp rate is shown in **Figure 1** and shows slight, but important deviations from simple linearity. Second, the rear phase grating must be redesigned to produce the comb of seven peaks with the correct peak spacing at lower frequency. Merely translating

J P DUCK, L PONNAMPALAM, A J WARD, D J ROBBINS, G BUSICO, N D WHITBREAD AND **D C J REID** FROM BOOKHAM TECHNOLOGY HERE PRESENT MONOLITHIC, L-BAND, DS-DBR LASERS WITH PERFORMANCE EQUIVALENT TO THE C BAND DEVICE. LASER PAIRS CAN COVER OVER 200 CONSECUTIVE 50GHZ CHANNELS WITH SMSR >45DB AND EX-FACET POWER LEVELLED AT 16.5DBM

the C-Band design would produce a comb with the wrong peak spacing, leading to problems in achieving optimal overlap between supermodes.

The completed designs were fabricated using a Leica VB6 e-beam writer for grating exposure, leading to very consistent and repeatable grating behaviour from batch to batch.

C And L-Band Performance Comparison

Figure 2 shows the power levelled performance of both a C and L-Band DS-DBR device. Both devices comfortably achieve the target 16.5dBm ex-facet power across the full operating frequency range.

Also shown is the underlying SOA current, used to achieve the power levelling. The fine structure of the SOA current is indicative of the underlying tuning mechanism, where a high SOA current correlates with a high, rear grating, tuning current. Towards the high frequency extreme of the L-Band device operation (~192000GHz), there is a further increase in SOA current to compensate for the underlying roll-off of the gain band. The same behaviour is observed for the high frequency extreme of the C-Band device (~197000GHz). The respective gain bands could be optimised to limit this effect, although care must be taken to maintain an operating frequency overlap.

The pair of devices achieve 213 consecutive 50GHz spaced channels, from 186550GHz to 197150GHz, spanning the C and L-Bands.

Side Mode Suppression Ratio (SMSR) is another key parameter and all 213 channels are above the 40dB target. In fact, the current devices comfortably exceed this target for all but the extremes of the wavelength operating range, see **Figure 3**. Again, by further optimising the respective gain bands, the SMSR can be improved. RIN has also been measured (see **Figure 4**) with a worst case resonance value better than -150dB/Hz for both, the C and L-Band devices.

It is expected that an L-band laser will produce a lower output power than the equivalent C-Band device. As **Figure 5** and **Figure 6** show the power of both devices saturate as the gain and SOA currents are increased, but the maximum saturated power of the C-Band device (95-100mW) is higher than the L Band device (85mW).

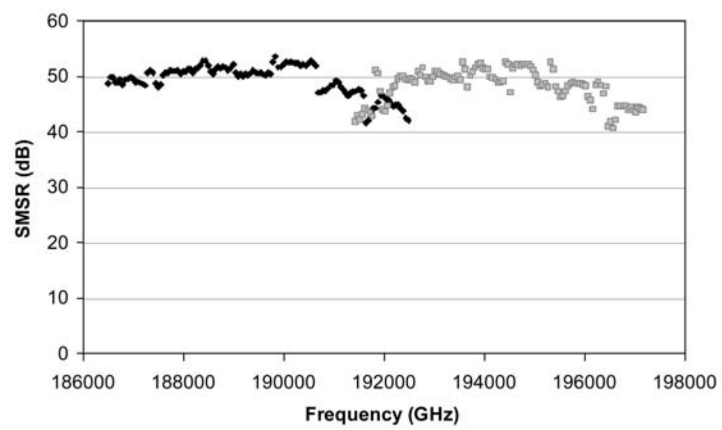


Figure 3: *SMSR versus channel frequency, across the C and L-Band*

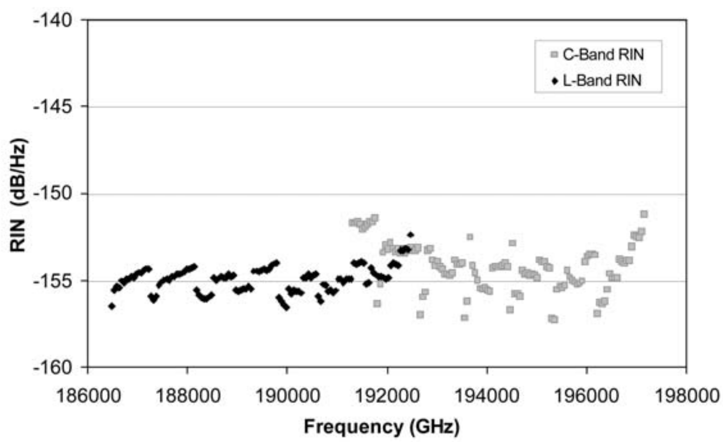


Figure 4: *Resonance peak value of RIN*

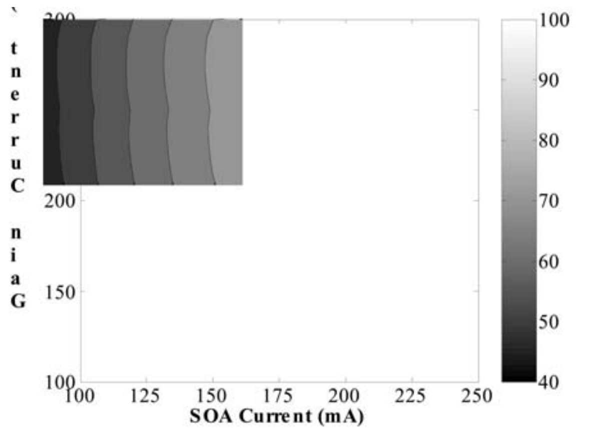


Figure 5: *C-Band free-running power contours*

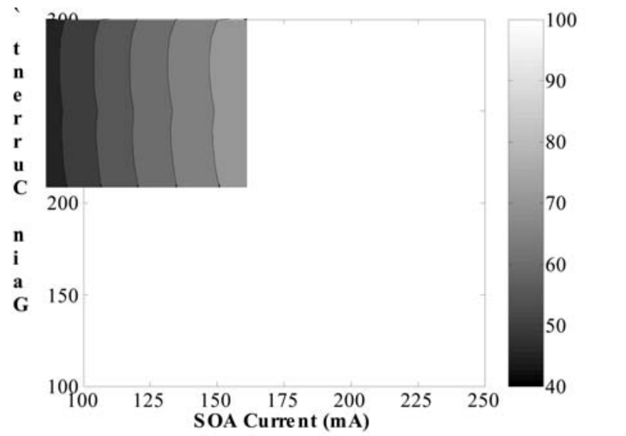


Figure 6: *L-Band free-running power contours*

Moreover, by looking at the photo-current induced in the SOA, it is possible to show that the intrinsic power produced by the laser cavity is greater in the C-Band device than the L Band one. The reason that both devices can be power-levelled at the same level is that, as the figures below show, the target ex-facet power of 16.5dBm (45mW) is well within the saturated output power of both devices.

Translated With Greater Performance

It has been shown that the DS-DBR technology platform can be translated to L-Band whilst maintaining performance equivalent to C-Band. Over 200 consecutive 50GHz channels have been demonstrated using two DS-DBR lasers. The L-Band laser has matched the high output power and SMSR of the C-Band device. RIN is unaffected.

Future work will continue to be based on extending the DS-DBR technology platform. Higher output power, high temperature operation, broader tuning range and monolithic integrated laser-modulators are all of interest.

REMOTE MONITORING FOR fibre-to-the-home

FIBRE-TO-THE-HOME (FTTH) OPERATOR'S ABILITY TO IDENTIFY AND LOCATE FAULTS REMOTELY AND SUCCESSFULLY REQUIRES THE CORRECT EQUIPMENT. **SIMON RICHARDSON** PRESENTS A SOLUTION THAT BRINGS THE ADDITIONAL BENEFIT OF BRINGING MAJOR COST SAVINGS TOO

Equipment pricing continues to be a major consideration in the telecommunications industry and optical component vendors face fierce competition to remain cost competitive. Nowhere is this more important than in the access network where fibre-to-the-home (FTTH) networks are starting to be deployed with tight budgets. However, from an operator's perspective, the initial equipment outlay is only part of the story and the ongoing ownership costs are equally important for longer term profitability.

Analysing The Costs

A major cost-factor here is the operator's ability to identify and locate faults in its network so that it can send the correct engineering resources to the fault location quickly with the necessary equipment to fix the problem with minimum system downtime.

Electronic fault diagnostics is built into the active equipment directly. Optical transceivers, for example, are designed to have several layers of warning alarms to highlight problems such as low output power or laser failure, but these features do not help to identify the location of faults in the optical infrastructure.

So, when problems occur in the optical domain, engineers equipped with an Optical Time-Domain Reflectometre (OTDR) are needed to locate the problem. But, good OTDRs are not particularly cheap, and cheap OTDRs are not necessarily good enough to locate faults in some circumstances. In any case, they require trained users to analyse the information that they provide. It is, therefore, an expensive solution to have a team of trained engineers, each equipped with OTDRs for field diagnostics.

Obviously, it's not just the equipment cost which needs to be factored in here, but also the cost of maintaining an engineering team to use and interpret the diagnostics equipment.

The cost doesn't end there either; once the fault is detected, operators need to dispatch engineers with the necessary equipment to fix the fault. The remedy could require splicing, which means each engineer requires a splicer, even if it is not being intensively used (another piece of capital outlay), or the problem could be fixed by simply cleaning a connector, which requires less expensive tools. Overall, this is an inefficient use of capital equipment.

Once the problem is solved, some operators may carry out another diagnostics check to confirm that the fault has been repaired correctly, which essentially necessitates at least three engineering visits to fix one problem. This method of working also has the disadvantage of considerable system downtime, as well as the difficulty of getting access points to attach the OTDR.

Fibre Monitoring

An alternative solution is the use of a static 'Fibre Monitoring' system (FiMo), which is permanently situated in the central office along with the transmission equipment. Japanese equipment manufacturer Fujikura has been supplying static fibre monitoring systems into Japanese Metro markets for over ten years now, more recently, it has developed this system for use in optical access networks.

Fibre-to-the-home continues to grow in Europe replacing copper networks. This migration of fibre into the access network actually amplifies the commercial need for a static diagnostics system. The infrastructure is more accessible to the public and other utilities providers, so there is greater potential for damage caused by third parties.

In addition, the dispersed fibre infrastructure means that engineering teams could spend more time travelling to the fault location rather than fixing it. PON networks are strong contenders for large FTTH deployments in Europe, but this topology introduces additional technical challenges for static monitoring because of the additional complications caused by monitoring OTDR traces through optical splitters.

Figure 1: FiMo system overview

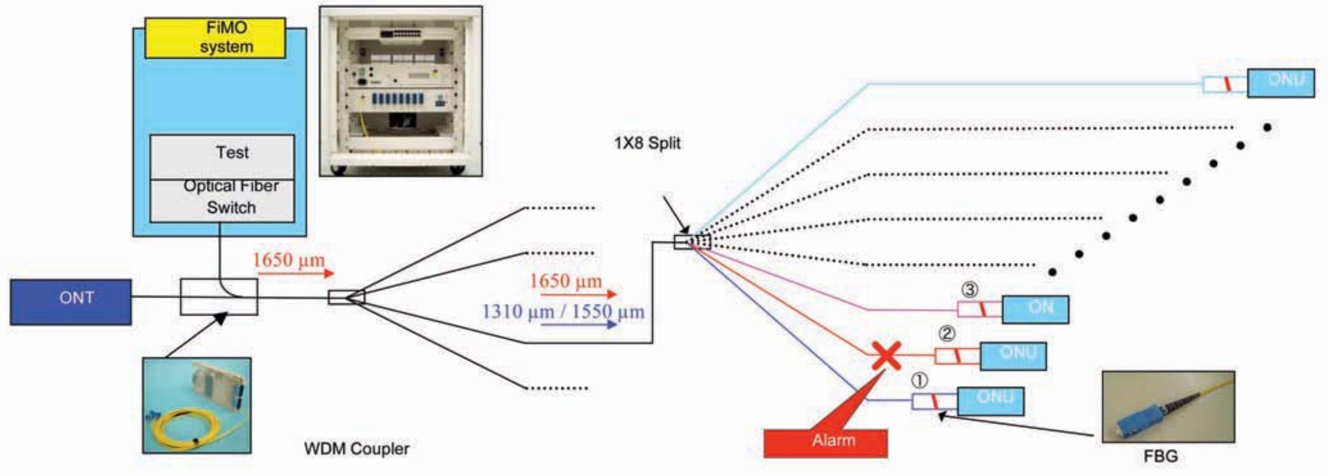


Figure 1 shows a schematic of the FiMo system. The static fibre monitoring concept itself is very simple. In live networks we obtain an OTDR trace by sending a wavelength outside of the transmission windows through an optical switch coupled to the network. (Dark fibre can be monitored at 1550nm). The switch can serve up to 2000 data-lines from a single OTDR input to optimise the cost efficiency of the unit. The trade-off is that as the number of lines monitored increases, which lowers the cost per customer, there is a corresponding decrease in the cycle rate as more lines need monitoring.

Each output from the optical switch is coupled to the data lines using WDM couplers, which are specifically designed to have a very low insertion loss at the signal wavelengths.

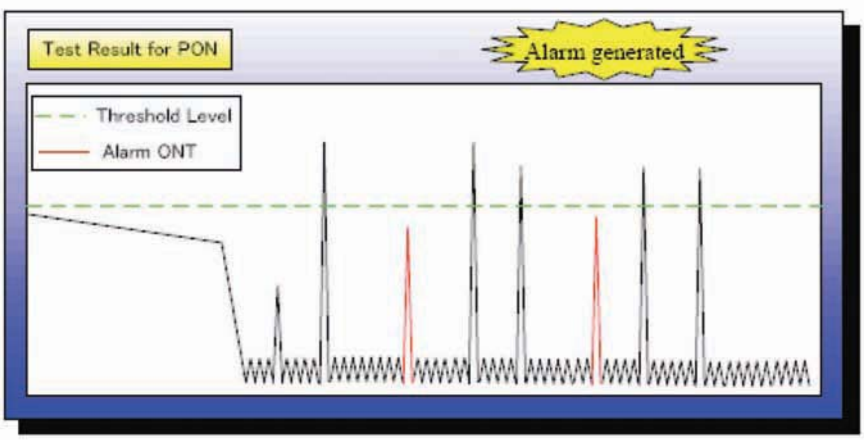
So, in point-to-point systems, the fibre monitoring system located in the central office can then simply cycle through each data-line, in turn monitoring the optical network for problematic losses. Any errors that occur then trigger alarms to the engineering team, highlighting that there is a problem and indicating the location. In PON systems the picture is more complicated because the optical splitters make the trace more difficult to interpret.

PON Network Differences

In Japanese and American PON networks, the initial data line is split either inside the central office or closer to the customers using Planer Light Circuit (PLC) splitters. Within Europe, some operators have looked at PON networks using double star topologies with two layers of split located close to the end users. Typically this would utilise a 1 x 4 splitter followed by four 1 x 8s. So the OTDR trace now contains up to 32 branches, which is difficult to analyse in a meaningful way.

Economically, the PON infrastructure also has the potential to decrease the cost of monitoring, since 64,000 users can potentially be monitored continuously from a single OTDR, but the challenge is to make sense of the OTDR traces. It takes some user

Figure 2: Alarm generation



Alarm content displayed after alarm generated

State	Date	LV	Unit	Port	No.	Distance	Type	Value	Message
Alarm	2006/9/18 11:00	LV2	Unit1	Port1	1	7.00km	Ref	5.0dB	At 7.00km point. Level down of Return loss is detected.(5.0dB)
Alarm	2006/9/18 11:00	LV2	Unit1	Port1	1	7.04km	Ref	4.5dB	At 7.04km point. level down of Return loss is detected.(4.5dB)

skill to interpret traces even on a single line, but when you incorporate splitters, particularly in double star systems, the trace becomes very complicated.

A solution to this problem is to incorporate Fibre Bragg Gratings (FBG) close to the actual users, that provide a reflection peaks which can be identified by the OTDR. These FBGs have minimal impact on the insertion loss of the data wavelengths because they are narrow band devices that can be designed to operate at the OTDR wavelength outside of the transmission window.

The reflection peaks serve as markers which identify the end of the optical network. This also has additional benefits in de-marking the passive infrastructure from the active. This has important implications for operators if local loop unbundling is required. The incumbent operator owns the optical infrastructure, whilst the active equipment is potentially under the control of the other service provider. Monitoring the optical infrastructure can, therefore, minimise disputes between the two companies regarding where the responsibility of the fault lies.

Tracing Traces

So, in a PON network each OTDR trace from each switch now contains visible features, such as the optical splitters and up to 32 reflection peaks from the FBGs. The initial trace is stored in the FiMo computer and this can be automatically compared to subsequent traces as the OTDR is cycled through the optical switch. Any infrastructure problems can now be visibly seen as a departure from the original OTDR trace. The system will then automatically send alarms to the engineering team, including the magnitude of the loss and the location of the fault.

The magnitude of the losses that trigger alarms can be adjusted by presetting threshold values to account for minor fluctuations in the network (small bend losses, for example). The location of the fault is identified by making use of the FBG reflection peaks (see **Figure 2**).

It is not trivial to visually relate the OTDR trace on star topologies to the physical position of the fault, and this is where the FBG reflections help. If an individual or group of peaks drops below a threshold value, you can identify where the fault is generated. For example in a double star network, if eight FBG peaks go below threshold, then it is highly likely that the fault lies between the primary and secondary splitter. If one goes down then you know the fault is either electronic (in which case the transceivers should also record a problem) or located in the final customer drop. The fibre monitoring system can use these indicators to locate the fault using bespoke software.

One obstacle is that this system needs to have each feature registered as the first trace is recorded. It would be very difficult to do this manually, because it relies heavily on the network operator keeping exact records of the fibre lengths installed in its networks and, of course, there will be uncertainties in these lengths which could be problematic. It would also be very time-consuming to enter each feature manually and, of course, this would also open possibilities for human error.

However, Fujikura has found a solution to this problem. In PON networks the downstream signal is broadcast to all 32 users, but the upstream traffic is Time Division Multiplexed with each transceiver at the user side being allocated time slots in which to send signals upstream. To enable this multiplexing, the transceivers at central office – called the Optical Line Terminal (OLT), and the user side transceivers – called the optical network units (ONU), communicate with each other to establish the optical path length between them. This is important because differences in the optical path length equate to time differences in the data transit time, which would result in jumbled data being received by the OLT. This process is called ranging and is integral to the PON concept. It is also useful for the fibre monitoring system.

The ranging information along with the ONU registration name is usually stored on a system database. Therefore, the RFTS (Remote Fibre Test System) can interrogate this database using its Network Database Management System (NDMS) to compare the features on the OTDR trace with the registration information obtained through ranging. The FiMo system can then identify which ONU corresponds to each reflection peak on the OTDR trace and each peak can then be labelled with the ONU registration number. This process is carried out when the FiMo system records its first trace, eliminating the need for operator input or accurate (manual) record keeping.

Additional Factors To Consider

There are some additional factors to consider which effect access networks more than Metro networks. Access networks are subjected to a higher rate of physical changes than Metro networks, for example it is unlikely that all of the end users will subscribe to the FTTH when it is initially deployed. This is perhaps not such an issue in new build sites, but would certainly be important for existing residences. It is, therefore, important that the FiMo system has the ability to adapt to intentional network changes without triggering false alarms.

To do this, any changes to the network are high-lighted to the operator who then has the ability to override the alarms if the change is an intentional one. If the change is intentional then the new configu-ration is taken as the reference trace and the monito-ring can continue.

Although each local exchange needs at least one FiMo system to monitor the local network, several fibre monitoring systems can also be connected to a single remote server. This enables a single location to monitor the access network of an entire country.

The FiMo system is proven technology. Japan started its PON deployments over ten years ago. There was initially regulatory pressure to provide live system diagnostics in the access market to maintain service standards. This was the initial motive for Fujikura to start developing this system in Japan. As the technology matured, the cost saving element clearly became an unexpected additional feature of the system. The system now continues to be deployed through choice rather than regulation, simply because it is showing its cost effectiveness through field use. It is the ongoing cost saving element that is currently the attractive feature for price sensitive access networks which is sparking interest in these systems for operators outside of Japan.

Measuring MUTUAL using the

In this article, Dogan Ibrahim describes a method for finding the mutual inductance between a pair of coils by using the Atlas LCR passive analyser

The Atlas LCR passive component analyser from Peak Electronics is used to measure all types of passive components, including resistors, capacitors and inductors.

As shown in **Figure 1**, the analyser is a small plastic moulded unit with a 2-line by 16-character LCD display. The unit comes with a pair of black and red clip-on probes. The usage of the device is very easy. The component to be measured is connected to the probes and the left-hand on-test button is pressed. After a five second countdown the component is analysed and then its value measured and displayed.

Pressing the test button during countdown cancels the countdown and starts the analysis instead. Once the component value is displayed, pressing the right-hand button scrolls the display through additional test data, such as the test frequency used during the measurement. The measurement can be repeated any time by pressing the left-hand test button.

Figure 1: The Atlas passive LCR analyser

The Atlas LCR analyser measures the capacitors in the range 1pF to 10,000μF, inductors from 1μH to 10H and resistors from 1Ω to 2MΩ. The basic accuracy of the unit is specified as 1%. I have tested a large number of capacitors and resistors and they were all found to be within their specified tolerances. The unit was also used to test inductances of known values. Several RF coils were tested ranging from a few μH to several mH and they all gave satisfactory results.

Although the Atlas LCR analyser can measure the inductance of a coil, in many RF applications we need to know the mutual inductance between two coupled coils and, hence, the voltage generated in a coupled coil when the voltage in the other coil is varied. Before looking at the mutual induc-tance test procedure, it will be useful to review the basic theory of inductance and the mutual inductance.

INDUCTANCE

There are many formulae to calculate the inductance of a coil of any given shape. Basically, the inductance is directly related to the square of the number of turns, the cross sectional area of the coil, and is inversely proportional to the length of the coil. One of the most popular formulae used to calculate the inductance of a single-layer air-core coil is by Wheller:

$$L = \frac{n^2 a^2}{9a + 10l} \quad (1)$$

where:

L = the inductance (μH)
n = number of turns
a = coil radius (inches)
l = coil length (inches)

Equation 1 is usually used to calculate the number of turns for a required inductance, where:

$$n = \sqrt{\frac{L(9a + 10l)}{a^2}} \quad (2)$$

The above formula is accurate when the wire diameter is less than 10% of the coil diameter, when the coil is closely wound on a single layer and when the coil length is 0.4 to three times the diameter.

The inductance of a multi-layer air-core coil can be calculated by using the following Wheller formula:

INDUCTANCE

ATLAS LCR ANALYSER

$$L = \frac{0.8n^2a^2}{6a + 9b + 10c} \qquad (3)$$

where:

L = inductance (μH)
n = number of turns
a = coil radius (inches)
b = coil length (inches)
c = difference between the outer and inner radii of the coil (inches)

An easy way to determine the inductance of a particular coil is to find its resonant frequency (f) with a known capacitor (C). The value of the inductance can then be calculated as:

$$L = \frac{1}{4\pi^2 f^2 C} \qquad (4)$$

Measuring the value of an inductor using the Atlas LCR analyser is very easy. Connect the inductor to the probes and press the on test button. The value of the inductance will be displayed on the LCD after a few seconds. Pressing the scroll-off button will show the measurement frequency and the DC resistance.

MUTUAL INDUCTANCE

When a pair of coils is close to each other, the magnetic field produced by the current in one coil produces a voltage in the other coil as a result of the inductive coupling of the two coils. This inductive coupling, known as the mutual inductance is very important in coupled RF inductors and transformers. For the circuit of **Figure 2**, the following equations can be written:

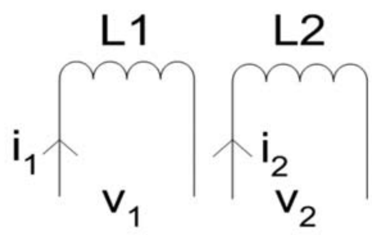
$$V_1 = L_1 \frac{di_1}{dt} + M \frac{di_2}{dt} \qquad (5)$$

$$V_2 = L_2 \frac{di_2}{dt} + M \frac{di_1}{dt} \qquad (6)$$

where:

L1 = inductance of coil 1
L2 = inductance of coil 2
M = mutual inductance between the coils
V_1 = voltage across coil 1
V_2 = voltage across coil 2
i_1 = current through coil 1
i_2 = current through coil 2

Figure 2: *Two inductively coupled coils*



Thus, V_1 is produced by i_1 reacting on itself and partially by i_2 reacting on i_1. Similarly, V_2 is produced by i_2 reacting on itself and i_1 reacting on i_2.

Mutual inductance depends entirely on the geometry and the orientation of the two coils with respect to each other. The sign of the mutual inductance can be positive or negative, depending on the relative direction of the two coils.

If the two coils are wound in the same direction and the current flows in both coils in the same direction then M is considered to be positive, otherwise M is taken as negative. Assuming that all of the flux lines from one coil embraces the other coil, the mutual inductance between the coils can be calculated approximately by the following formula:

$$M = \frac{\mu_0 n_1 n_2 A}{l} \qquad (7)$$

where μ_o is the permeability of free air ($4x10^{-7}$), n_1 and n_2 are the number of turns of each coil, A is the cross sectional area, and l is the length of the coils.

When two coils are normally connected in series, the total inductance (L_T) is equal to the sum of the inductances of each coil. i. e. $L_T = L_1 + L_2$. But if the two coils are coupled to each other inductively, then the total inductance is given by:

$$L_T = L_1 + L_2 \pm 2M \qquad (8)$$

where the sign of M is taken as described earlier. The degree of coupling in inductively coupled coils is measured by the coupling coefficient k, where:

$$M = k\sqrt{L_1 L_2} \qquad (9)$$

or:

$$k = \frac{M}{\sqrt{L_1 L_2}} \quad (10)$$

If the two coils are coupled perfectly where all the lines of magnetic flux from one coil embraces the other coil, then the coefficient of coupling becomes $k = 1$. **Equation 8** is very important in finding the mutual inductance of a pair of inductively coupled coils and it forms the basis of our procedure to find the mutual inductance. **Equation 10** is also important, since it is an indication of the degree of coupling between the two coils.

MEASURING THE MUTUAL INDUCTANCE

An easy and reliable method for measuring the mutual inductance between a pair of coils is first to measure the inductances of each coil independently and then to measure the total inductance when the coils are connected in series. Then, we can use Equation 8 to calculate the mutual inductance. The procedure using the Atlas LCR analyser is as follows:

- Measure the inductance of the first coil, call this L_1;
- Measure the inductance of the second coil, call this L_2;
- Connected the two coils in series and measure the total inductance, call this L_T;
- Using Equation 8 calculate the mutual inductance between the coils as:

$$M = \frac{L_T - (L_1 + L_2)}{2} \quad (11)$$

- Calculate the coefficient of coupling using Equation 10.

An example calculation is given below.

Example 1:

In this example, the mutual inductance between a pair of coils wound on a ferrite core is calculated. First of all, the inductance of each coil was measured using the Atlas LCR analyser and the results were found to be $L_1 = 775\mu H$ and $L_2 = 1.193mH$.

The coils were then connected in series and the total inductance was measured to be $L_T = 3.816mH$. Using **Equation 11**, the mutual inductance is calculated as:

$$M = \frac{3.816 - (1.193 + 0.775)}{2} = 924\mu H$$

Also, the coefficient of coupling is found from Equation 10 as:

$$k = \frac{0.924}{\sqrt{1.193x0.775}} = 0.96$$

VERIFYING THE RESULTS

It is always a good practice to verify that a measured component is actually within the expected limits. The value of the measured mutual inductance can be verified by many methods. One method is to apply a sinusoidal voltage to one of the coils and then to measure the amplitude and the phase of the voltage across the other coil. Knowing the component values we can easily verify the value of the measured mutual inductance. Another method is to apply a sinusoidal current to one of the coils and then to measure the voltage across the other coil.

Another commonly used simple method is to apply a step voltage to one of the coils and then to plot the time response of the voltage across the other coil. From the step response we can verify the value of the mutual inductance. All three methods are described here, although the step response technique is used by the author to verify the mutual inductance measured in Example 1.

Appling Voltage

As shown in **Figure 3**, the coils have resistances R_1 and R_2 and a sinusoidal voltage is applied to one of the coils. We can write the following equations for this circuit:

$$\begin{bmatrix} V_1 \\ 0 \end{bmatrix} = \begin{bmatrix} R_1 + jwL_1 & -jwM \\ -jwM & R_2 + jwL_2 + R_L \end{bmatrix} \begin{bmatrix} i_1 \\ i_2 \end{bmatrix} \quad (12)$$

The output voltage is calculated as:

$$V_2 = i_2 R_L = \frac{R_L \begin{vmatrix} R_1 + jwL_1 & V_1 \\ -jwM & 0 \end{vmatrix}}{\begin{vmatrix} R_1 + jwL_1 & -jwM \\ -jwM & R_2 + jwL_2 + R_L \end{vmatrix}}$$

$$= \frac{jwR_L M V_1}{(R_1 R_2 + R_1 R_L - w^2 L_1 L_2 + w^2 M^2) + jwL_1(R_2 + R_L)}$$

or:

$$V_2 = \frac{jwR_L M V_1}{R_1(R_2 + R_L) + w^2 L_1 L_2 (k^2 - 1) + jwL_1(R_2 + R_L)} \quad (13)$$

The amplitude and the phase of V_2 are:

$$|V_2| = \frac{wR_L M V_1}{\sqrt{\left[R_1(R_2 + R_L) + w^2 L_1 L_2 (k^2 - 1)\right] + w^2 L_1{}^2 (R_2 + R_L)^2}} \quad (14)$$

and:

$$\angle V_2 = -90 - \tan^{-1} \left[\frac{wL_1(R_2 + R_L)}{R_1(R_2 + R_L) + w^2 L_1 L_2 (k^2 - 1)}\right] \quad (15)$$

Knowing the values of the components, we can calculate the amplitude and the phase of the output voltage and compare it with the experimental value. Note that the resistances of the coils can be measured with the Atlas LCR analyser.

Applying Current

As shown in **Figure 4**, a sinusoidal current is applied to coil 1. The voltage across coil 2 is given by:

$$V_2 = M \frac{di_1}{dt} \quad (16)$$

For example, if the applied current is:

$$i_1 = 2\sin(300t)$$

then the voltage across the other coil will be:

$$V_2 = M \frac{d(2\sin(300t))}{dt} = 600M\cos(300t)$$

Using the measured value of M and by plotting or observing V_2, we can verify the value of M.

Step Response

Considering Figure 3, a step voltage is applied to coil 1 and the voltage across coil 2 is plotted. We can write the following equations for the coils:

$$V_1 = R_1 i_1 + L_1 \frac{di_1}{dt} + M \frac{di_2}{dt} \quad (17)$$

and:

$$V_2 = R_2 i_2 + L_2 \frac{di_2}{dt} + M \frac{di_1}{dt} \quad (18)$$

Suppose that we now suddenly connect a battery to coil 1 at time t=0. The other coil is assumed to be open circuit (or connected to a voltmeter with a high internal impedance) and thus $i_2 = 0$. From **Equation 17**:

$$V_1 = R_1 i_1 + L_1 \frac{di_1}{dt} \quad (19)$$

The step response of **Equation 19** is the well known:

$$i_1 = \frac{V_1}{R_1}(1 - e^{(-tR_1/L_1)}) \quad (20)$$

Using **Equation 20** in **18** and noting that $i_2 = 0$, we get:

or:

$$V_2 = V_1 \frac{M}{L_1} e^{(-tR_1/L_1)} \quad (21)$$

Plotting **Equation 21** we get an exponentially decaying graph as shown in **Figure 5**. From this graph at t=0 the value of V_2 is equal to:

$$V_2 = V_1 \frac{M}{L_1} \quad (22)$$

Thus, we can verify the value of M easily by comparing it with the experimental result obtained from the step response.

In order to verify the mutual inductance of the coils used in Example 1, a step input of V_1=0.4V was applied to coil 1 and the step response, shown in **Figure 6**, was obtained using the Pico ADC-200 PC-based oscilloscope. From this figure, at t=0 the voltage across coil 2 is measured to be V_2 = 0.48V. From Equation 22, V_2 = 0.4 x 0.924/0.775 = 0.476V, which is very close to the measured value, verifying that the mutual inductance between the coils is M = 924μH.

A Method For Finding Mutual Inductance

This paper has described a method for finding the mutual inductance between a pair of coils. The Atlas LCR passive analyser is used to measure the self-inductance of the two coils, and then a simple method is described for calculating and verifying the mutual inductance.

The overall measurement and calculation process takes no more than 30 seconds using the Atlas LCR analyser. The optional verification process simply requires the step response to be plotted either using a data logger or a suitable oscilloscope (eg a PC-based oscilloscope).

Figure 3: *A sinusoidal voltage is applied to one of the coils*

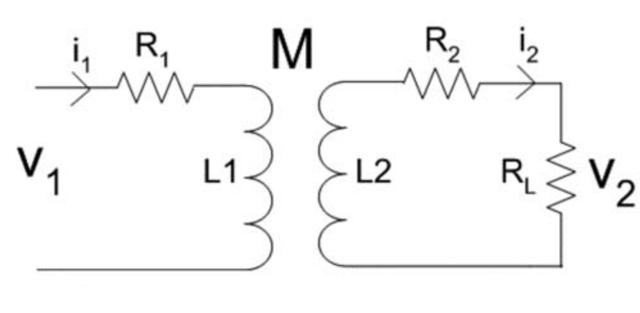


Figure 4: *A current is applied to one of the coils*

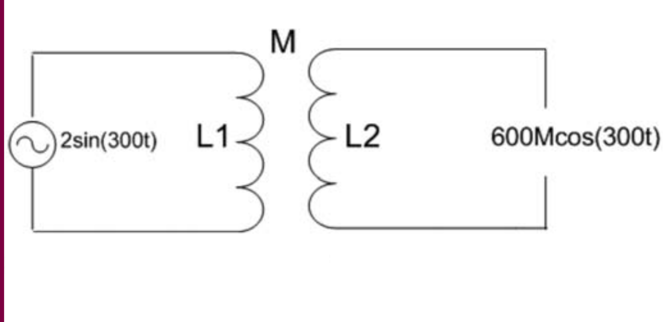


Figure 5: *Step response of coil 2*

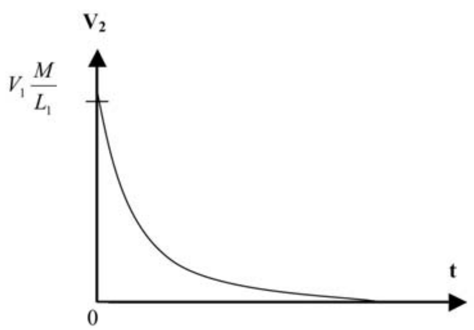
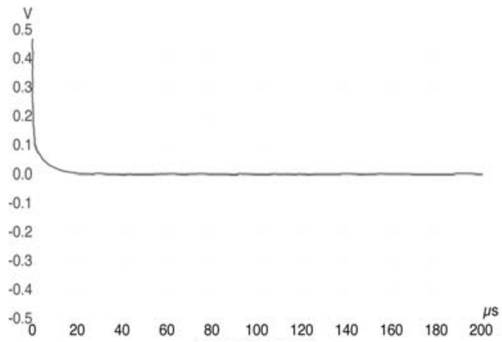


Figure 6: *Step response for the Example 1*



Book Review

RFID TOYS – 11 Cool Projects for Home, Office and Entertainment

Amal Graafstra

ExtremeTech

We've all heard of "tagging prisoners" and we've seen those fancy plastic dongles that the supermarkets clip on to their bottles of more expensive plonk – and now we've got "smart" wheelie-bins. Somewhere inside all these, there's a transmitter and an aerial, and I've always been curious about the technology.

More recently, there's been a plague of sticky labels on ordinary products, like plugs and sockets, or shirts and socks; these also contain a simple Radio Frequency Identification (RFID) device which causes a squawk if you try to run off without paying – or if the shop assistant forgets to defuse it. These in particular have piqued my interest, simply because of the implication that the tags cost almost nothing to make.

This book not only offers an insight into how these things work, but also provides a cook-book for some pretty good home-made products based on RFID technology. I must admit that, when I opened the book, I didn't expect it to do more than satisfy my curiosity, but I've one project under design already, so it's pretty convincing.

> ***This is an interesting and stimulating book. The author's intention is not that it be used as a blueprint, but that the various projects in it should show what can be done, and spark off one's own ideas***

The book covers both passive and active RFID systems. Passive tags simply respond with their identification code to a prompt from the sensor and, therefore, have very short range, typically only a few inches. Active tags contain a battery and, thus, come more expensive and with a limited service life, but their range is up to several hundred feet, and writable tags can provide more complex bespoke coding.

Chapter 1 offers a very brief introduction to the uses of RFID, describes the types of tag available and lists the frequency bands allocated for their use. The 125-148kHz band is for passive tags, often used for animal tracking and access control, but has its limitations because a fair amount of copper antenna has to be crammed into the tag for it to work at all. Next is 13.56MHz, again passive, but with a more compact style because the antenna can be printed on to a paper substrate. This is used for book control in libraries and for access monitoring, but suffers serious interference from metals, so design needs to take this into account. Then we have 915MHz, with very low cost tags, ideal for supply chain product tracking because anti-collision capabilities at this frequency allow simultaneous reads and, therefore, high data rates. The interference problems here are from liquids and the human body.

Finally, the most complex systems use active tags at 433MHz and 2.5GHz, for road tolls, vehicle management and asset tracking, with very expensive tags but far greater range.

The rest of the book offers a range of projects well within the capabilities of the interested amateur. They are tackled at three levels: relay operation, BASIC Stamp microprocessor interface, or a USB connection into a PC, so there's a lot of scope. The author, ever mindful of our over-protective legislation, warns of the dangers of cutting oneself or burning the house down, but throughout the text he gives very practical advice on all the mechanical aspects of the work. There's a great deal of sawing and drilling in this book, in which everything from the front door to a commercially-built circuit is fairly brutally hacked to pieces and rebuilt, all explained in detail in a sort of "Surgery for Dummies" style.

Internet access is essential, because all the software to run these projects is downloaded – most of it entirely free – from either the hardware suppliers or from the author himself. And that's not something to be sneezed at, because the author offers a free upgrade and maintenance service, simply because he's an enthusiast and encourages participation.

Chapter 2 details a keyless access project for the home, modifying an electronic deadbolt to work from a USB RFID tag reader, to log who comes in the front door and when. The instructions are extremely clear, with 37 photographs in this section alone, but it would still take a brave man to carve up the door in quite the way the author does.

Chapter 3 provides keyless access to the car, though, obviously, without the computer – just a relay connected to the central locking mechanism. The constraints of working in the vehicle environment, surrounded by metal and plagued by damp, means you end up with a pretty ugly package, but it looks robust.

Chapter 4 goes further into the computer connection, detailing a very nifty way of logging in to Windows XP by waving a tag over the keyboard rather than typing in a password. It really is pretty clever, although you need to be confident about butchering the keyboard and altering Windows registry to finish the project off.

Chapter 5 shows how to make an RFID-enabled safe. This is an excellent tutorial on the use of a BASIC Stamp microprocessor to process the tag information, rather than being limited to a relay, or having to go into Windows.

Chapter 6 moves up from the kHz range of tags to 13.56MHz, tackling resonant loop antennas and anti-collision systems. This level allows unique product identification, down to individual items, and is what some big stores use for inventory control. The author details a "Smart Shelf", linked to a PC, which can register any tagged item placed on it.

Chapter 7 takes the 13.56MHz technology further, with a pet-door that underlines the need to be aware of the relative orientation of tag and reader for accurate operation. This again uses the BASIC Stamp-2 rather than a PC, so is very useful as a template for other ideas, but I have my doubts about electronic pet doors, as I have seen a terrified cat completely demolish one by barrelling straight through it without waiting for the pick-up signal to operate.

Chapter 8 moves up again to active RFID on 433MHz, with an employee time-and-tracking system. Obviously, this is a large and complex project, the hardware is more expensive and the software for this one isn't free. The author isn't particularly happy with the proprietary attitude of his supplier, but admits that it's the best option for the project, which is, like the others, very fully described. It would be ideal for my local comprehensive school, which has a dreadful problem keeping children in the right place.

Chapter 9 extends active tags to asset monitoring, with some useful DIY breadboard-style circuitry for 433MHz antenna-switching, controlled by another BASIC Stamp device. It's all good skill-boosting stuff and well worth getting into.

Chapter 10 introduces writable tags, touching on data encoding to save precious memory space, before showing how to program a BASIC Stamp to write a tag's identity and subsequently read it. There's a lot of useful stuff in here. But it's in Chapter 11, "Extreme RFID", that the author introduces his coup de théatre, when he reveals that he has had a subcutaneous tag (a 2mm x 12mm glass tag at 134kHz) installed in his hand to save carrying one in his pocket, and he discusses the potential uses of such implants – typically, for keyless entry and logging on to networks, as per the projects in earlier chapters.

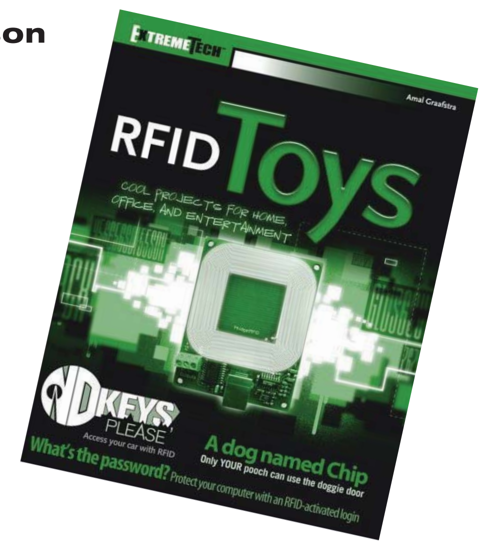
This is an interesting and stimulating book. The author's intention is not that it be used as a blueprint, but that the various projects in it should show what can be done, and spark off one's own ideas. Parts lists and supplier details are provided throughout, although they are all American, which raises costs significantly unless one can find domestic sources. The UK isn't as far forward as it might be – Google gives over 4.1 million hits for "RFID" worldwide, but only 47,000 on a UK search.

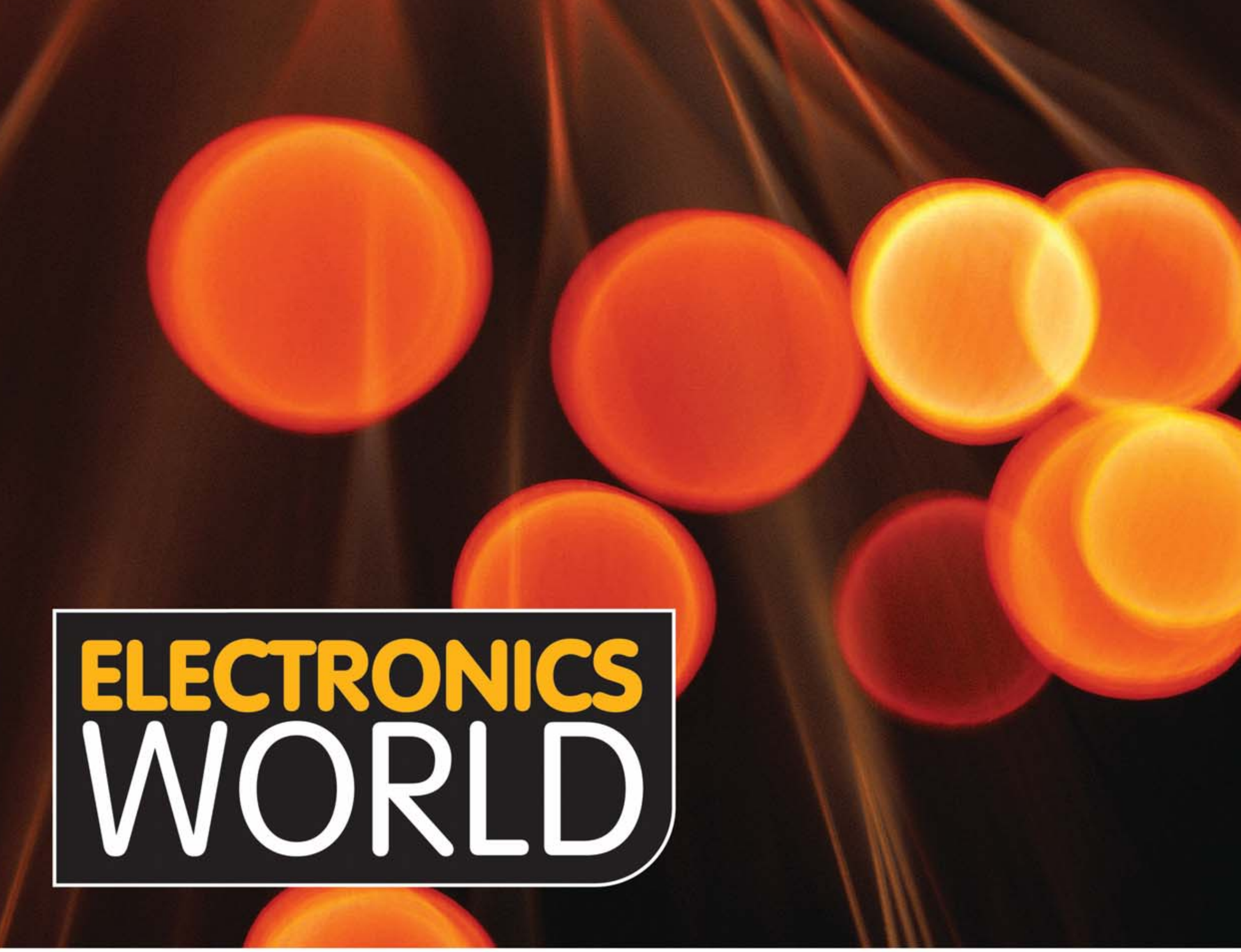
Buying into a typical multi-tag prototyping kit here can cost upwards of £600, which is well beyond the pocket of the home experimenter that the author is aiming at in this book. The best bet in the UK seems to be eBay, where you can pick up a complete door-lock kit from Hong Kong, with two fobs and a programmer, for less than £20. But if you do this, you're on your own when it comes to the brain surgery the author reveals, to get more out of the system and his software will be useless.

There's a good summary with technical specs in the appendix, and there are lots of websites to consult, many of them with hardware for sale and free software to download. If you want to sample what's on offer, try the author's **www.rfidtoys.net**, although it does tend to be more of a user forum and, so, isn't as helpful as it might be.

All in all, this is an excellent book, and the author's very approachable and personable style, on top of his obvious technical competence, does much to commend it – an ideal introduction to RFID for the amateur.

Hedley Richardson





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Voltage-Mode Multiplier Implementation Employing Current Conveyors

Analogue multipliers have received great attention due to the fact that they possess a wide range of applications in many areas such as signal processing, telecommunication, control and instrumentation.

In the literature, analogue multipliers are designed using different techniques. MOS four-quadrant analogue multiplier is one of the techniques. Another technique for designing analogue multipliers is to use operational amplifiers (OAs) and/or current conveyors (CCs). In the literature, second-generation current conveyors (CCIIs), differential difference current conveyors (DDCCs) and their applications are reported. Compared to the voltage-mode counterparts such as OAs, CCs have several advantageous such as greater linearity, wider bandwidth and better dynamic range.

In this paper, a novel voltage-mode multiplier circuit employing only one CCII+, one DDCC+, one grounded resistor and identical four NMOS transistors is proposed. The circuit presented here performs the multiplication of two analogue voltage-mode signals with an adjustable multiplier constant depending on the resistor value of the proposed network and the gain factor of the identical NMOS transistors. The DC voltage at the output of the proposed multiplier can be adjusted arbitrarily. The proposed circuit employs no capacitors. The SPICE simulation results are also included to demonstrate the performance of the proposed analogue multiplier circuit.

Description of the multiplier

A CCII is a three-port versatile active device and has the following voltage and current relationships as given in:

$$I_Y = 0 \qquad V_X = \beta V_Y \qquad I_Z = \pm\alpha I_X \tag{1}$$

Also, a five-port DDCC is represented by the following:

$$I_{Y1} = I_{Y2} = I_{Y3} = 0 \qquad V_X = \beta_1 V_{Y1} - \beta_2 V_{Y2} + \beta_3 V_{Y3} \qquad I_Z = \pm\alpha I_X \tag{2}$$

Current convention is such that all currents flow into the CCs, CCII+ and DDCC. The sign of α in Equations 1 and 2 denotes the type of the CCs. The frequency dependent non-ideal current gain α and voltage gain β in Equations 1 and 2 are ideally equal to unity. Each of the bulks of the identical NMOS transistors in **Figure 1** must be connected to its source. Applying routine analysis for the circuit in Figure 1, the current-voltage relationships are obtained as in:

$$\begin{aligned} I_{D1} &= \frac{k_n}{2}(V_{in1}(t) - V_{Tn} - V_{ss})^2 \\ I_{D2} &= \frac{k_n}{2}(V_{in2}(t) - V_{Tn} - V_{ss})^2 \\ I_{D3} &= \frac{k_n}{2}(V_{BB} - V_{Tn} - V_{ss})^2 \\ I_{D4} &= \frac{k_n}{2}(V_{in1}(t) + V_{in2}(t) - V_{Tn} - V_{ss})^2 \end{aligned} \tag{3}$$

where:
I_{Di} (i = 1, 2, 3, 4) is as seen in Figure 1.
In Equation 3, k_n is the gain factor and V_{Tn} is the threshold voltage. As a result, k_n is defined as in:

$$k_{n\,j} = \left(\frac{W}{L}\right)_j \mu_n C_{ox} \qquad (j = 1, 2, 3, 4) \tag{4}$$

where μ_n and C_{ox} are called the surface mobility and the gate capacitance per unit area, respectively.

As a result, from Equation 3, the following inequality conditions that are necessary for the identical NMOS transistors to be on are obtained in:

$$\begin{aligned} &(V_{in1}(t) - V_{Tn} - V_{ss}) \geq 0 \\ &(V_{in2}(t) - V_{Tn} - V_{ss}) \geq 0 \\ &(V_{BB} - V_{Tn} - V_{ss}) \geq 0 \\ &(V_{in1}(t) + V_{in2}(t) - V_{Tn} - V_{ss}) \geq 0 \end{aligned} \tag{5}$$

Note that the four inequality conditions in Equation 5 must be satisfied simultaneously. The conditions given by the set of equations 5 can be rewritten as:

$$V_{ss} \leq V_{in1}(t) + V_{in2}(t) - V_{Tn} \tag{6}$$

On the other hand, an input signal $V_j(t)$ can be expressed as follows:

$$V_{in\,j}(t) = V_{in\,j}(t)_+ + V_{in\,j}(t)_- \tag{7}$$

In Equation 7, $V_{in\,j}(t)_+$ (j = 1, 2) and $V_{in\,j}(t)$ can be expressed as:

$$V_{in\,j}(t)_+ = \begin{cases} V_{in\,j}(t), & \text{for } V_{in\,j}(t) \geq 0 \\ 0, & \text{otherwise} \end{cases}$$

and

$$V_{in\,j}(t)_- = \begin{cases} V_{in\,j}(t), & \text{for } V_{in\,j}(t) < 0 \\ 0, & \text{otherwise} \end{cases}$$

According to the input signals defined in Equation 7, the non-equality in Equation 6 turns the form in:

$$V_{ss} \le V_{in1}(t)_- + V_{in2}(t)_- - V_{Tn} \quad (8)$$

All of the transistors M_1, M_2 and M_4 are diode-connected and operate in the saturation region. However, in order to operate the transistor M_3 in the saturation region, the following condition must be satisfied in:

$$V_{in1}(t)_- + V_{in2}(t)_- \ge V_{BB} - V_{Tn} \quad (9)$$

Otherwise, the third NMOS goes into the linear region and extra DC term occurs at the output. Then, $V_{out}(t)$ is found to be in:

$$V_{out}(t) = R_L\,(I_{D4} + I_{D3} - I_{D2} - I_{D1}) \quad (10)$$

where I_{Di} (i = 1, 2, 3, 4) is given in Equation 3.

Therefore, $V_{out}(t)$ in Equation 10 is calculated as:

$$V_{out}(t) = R_L\,k_n\,V_{in1}(t)\,V_{in2}(t) + \frac{R_L\,k_n}{2}(V_{BB}^2 - 2V_{BB}(V_{Tn} + V_{ss})) \quad (11)$$

Note that the additional DC term $R_L\,k_n\,(V^2_{BB} - 2V_{BB}(V_{Tn} + V_{SS}))/2$ can be adjusted to any value by changing the value of V_{BB}. Therefore, by using Equation 5, the DC term with respect to V_{BB} is expressed as follows:

For $V_{Tn} + V_{SS} < V_{BB} < 0$, the DC term is negative.
For $V_{BB} < 0$, the DC term is positive.
For $V_{BB} = 0$, the DC term is zero.

Note that the inequality $V_{Tn} + V_{SS} < 0$ is always chosen.

If minus-type CCII and DDCC are employed for the network in Figure 1, $V_{out}(t)$ with $V_{BB} = 0$ is obtained with 180° phase shift in:

$$V_{out}(t) = -R_L\,k_n\,V_{in1}(t)\,V_{in2}(t) \quad (12)$$

If non-ideal current gain α and voltage gain β are taken into account, the multiplication with $R_L\,k_n = 1$ and $V_{BB} = 0$ in Equation 11 becomes:

$$V_{out} = V_{in1}(t)\,V_{in2}(t)\alpha_2\beta_{11}\beta_{13} + a_2V_{in1}^2(t) + b_2V_{in2}^2(t) + a_1V_{in1}(t) + b_1V_{in2}(t) + a_0 \quad (13)$$

Consequently, the undesired terms due to non-idealities, a_2, b_2, a_1, b_1, $a_0 = 0$ are computed as in the set of equations 14.

$$\begin{aligned} a_2 &= \frac{\alpha_2\beta_{21}^2 - \alpha_1}{2} \\ b_2 &= \frac{\alpha_2\beta_{23}^2 - \alpha_1}{2} \\ a_1 &= (V_{Tn} + V_{ss}\beta_1)\alpha_1 - (V_{Tn} + V_{ss})\alpha_2\beta_{21} \\ b_1 &= (V_{Tn} + V_{ss}\beta_1)\alpha_1 - (V_{Tn} + V_{ss})\alpha_2\beta_{23} \\ a_0 &= (V_{Tn} + V_{ss})^2\alpha_2 - (V_{Tn} + V_{ss}\beta_1)^2\alpha_1 \end{aligned} \quad (14)$$

Here, a_1 and b_1 are the corresponding current and voltage non-idealities of the first CC, respectively. Also a_2 and $b_{2\,k}$ (k = 1, 2, 3) are the corresponding current and voltage non-idealities of the second CC, respectively.

Simulations of the proposed multiplier

The presented circuit in Figure 1 is simulated using a SPICE program, thus the input signals $V_{in1}(t)$ and $V_{in2}(t)$ are chosen at 100kHz with amplitudes 200mV peak and 300mV peak, respectively. The corresponding ideal output signal $V_{out}(t)$ is demonstrated in **Figure 2** with a magnitude of 600mV. Here, k_nR_L = 10 is chosen.

Conclusion

In this study, a circuit performing the multiplication of two analogue voltage-mode signals is proposed. The presented circuit employs two CCs, four identical NMOS transistors and one resistor. The proposed multiplier needs no passive element matching conditions and/or cancellation constraints, and employs only one grounded resistor as a passive element thus it is easy to implement the introduced topology in fully integrated circuit technology.

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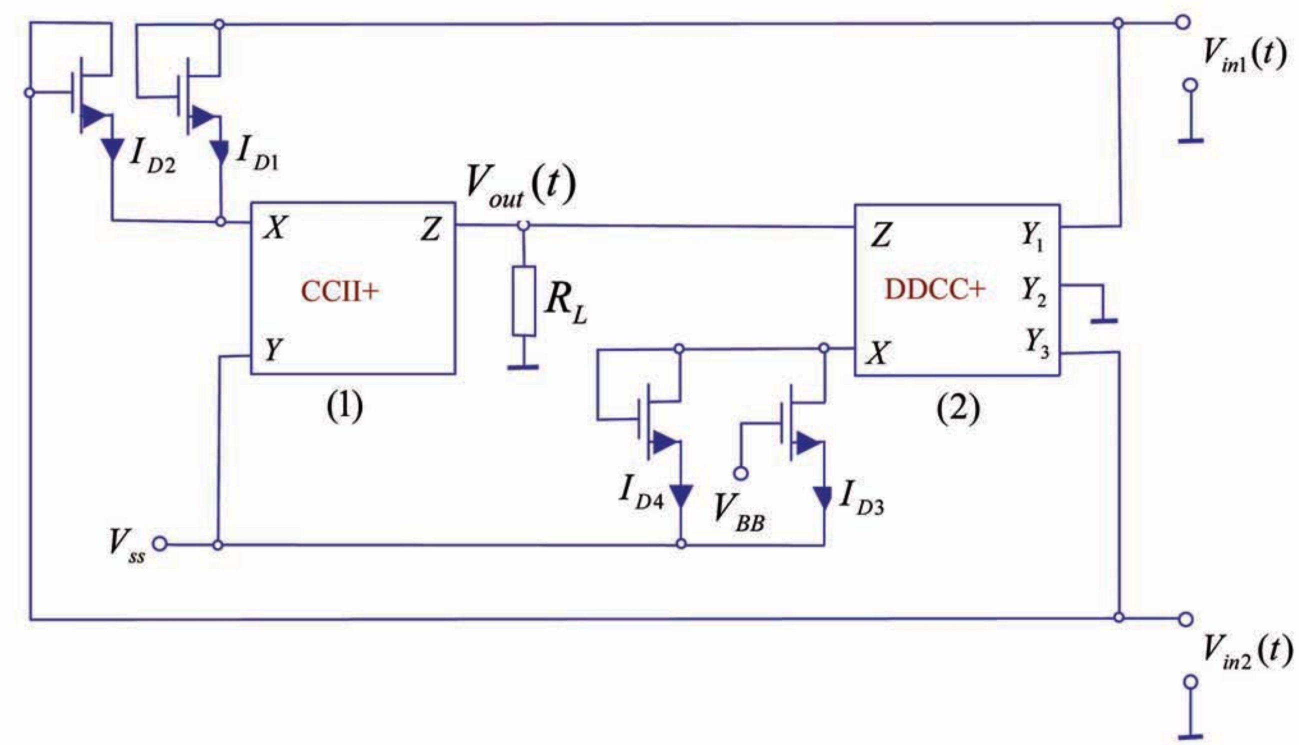

Figure 1: *The proposed analogue voltage-mode multiplier*

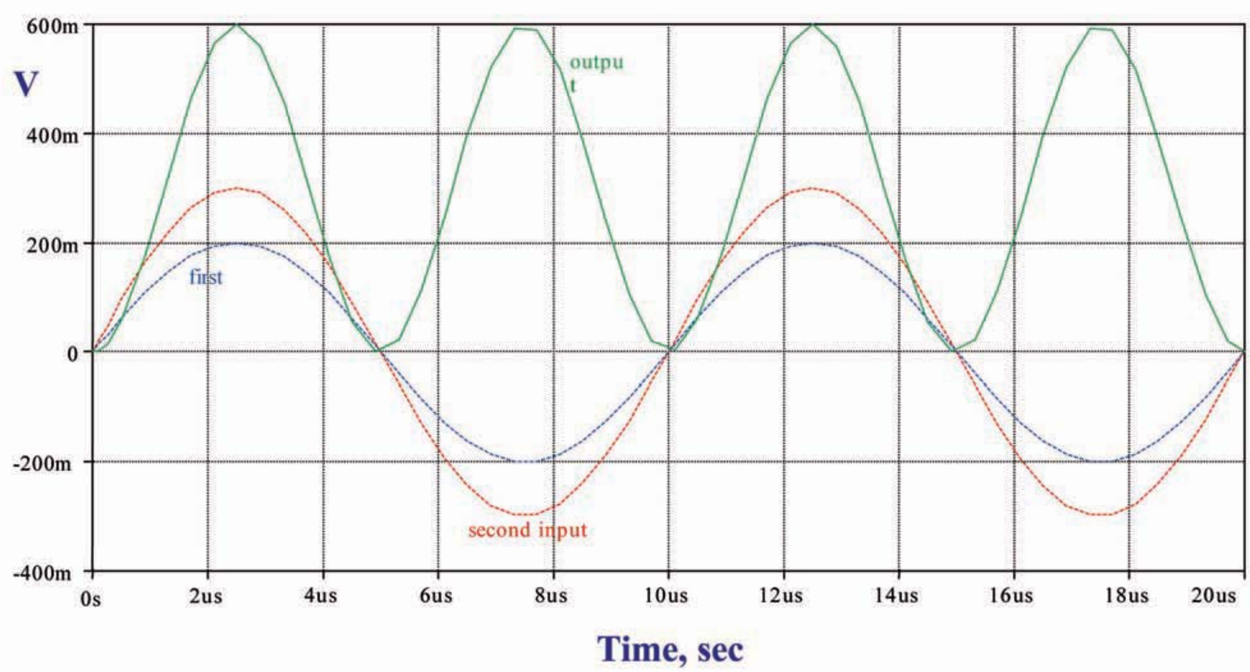

Figure 2: *Time domain responses of the proposed multiplier for the two input signals at 100kHz*

Circuit And Fold-Back Current For Negative Rail

Many electronic systems require both positive and negative supplies to operate properly. In these applications, the common and inexpensive voltage regulator µA723 from Texas Instruments can be used as a positive or negative regulator, where the handled output current can be extended by adding an external pass transistor.

As stated in Reference 1, in order to protect the pass transistor from overheating damage when a short-circuit or an overcurrent occurs, a fold-back protection, which reduces or folds back the output current during fault, is desired. This diminishes the power-handling requirements for the pass transistor.

The design and implementation of this output current protection is well-described in the manufacturer's datasheet, i.e. the µA723 from Texas Instruments or the MC1723C from ON Semiconductor, for a positive configuration, while no information is given for the negative configuration. But, normally, same kind of limitation is required for both polarities.

The circuit in **Figure 1** implements this protection for a negative output regulator, by adding few components (resistors R_{3a} and R_{3b} and the feedback resistor R_4), and using the internal current limiter transistor of the µA723 regulator. For this circuit, you can calculate the maximum foldover and short-circuit currents, I_{knee} and I_{sc}, respectively, considering absolute values, as follows:

$$\left|I_{knee}\right| = \frac{0.7}{R_s} + \frac{R_3}{R_s \cdot R_4} \cdot \left(-\left|V_{in}\right| + \left|V_{out}\right| + 0.7\right) \qquad (1)$$

$$\left|I_{sc}\right| = \frac{0.7}{R_s} + \frac{R_3}{R_s \cdot R_4} \cdot \left(-\left|V_{in}\right| + 0.7\right) \qquad (2)$$

As explained in Reference 2, the Zener diode D1 must be designed for the required input voltage tolerance ($V_{Z1} \geq |V_{in}| - V_{BE(Q1)} - 3$ and $V_{Z1} \geq |V_{in}| - |V_{out}| - V_{BE(Q1)} + 6$, which limits the input voltage range). V_{cc-} (pin 7 of IC1) is connected to the Zener diode D_2 instead to V_{out} (as opposed to Reference 2), which is fixed to 15V and permits to supply properly the regulator also in an output fault condition.

In a practical design, you select values for I_{knee}, I_{sc} and equal R_{3a} and R_{3b}, and then use **Equations 1** and **2** to calculate the resistors R_s and R_4.

In Figure 1, the maximum and short-circuit currents are

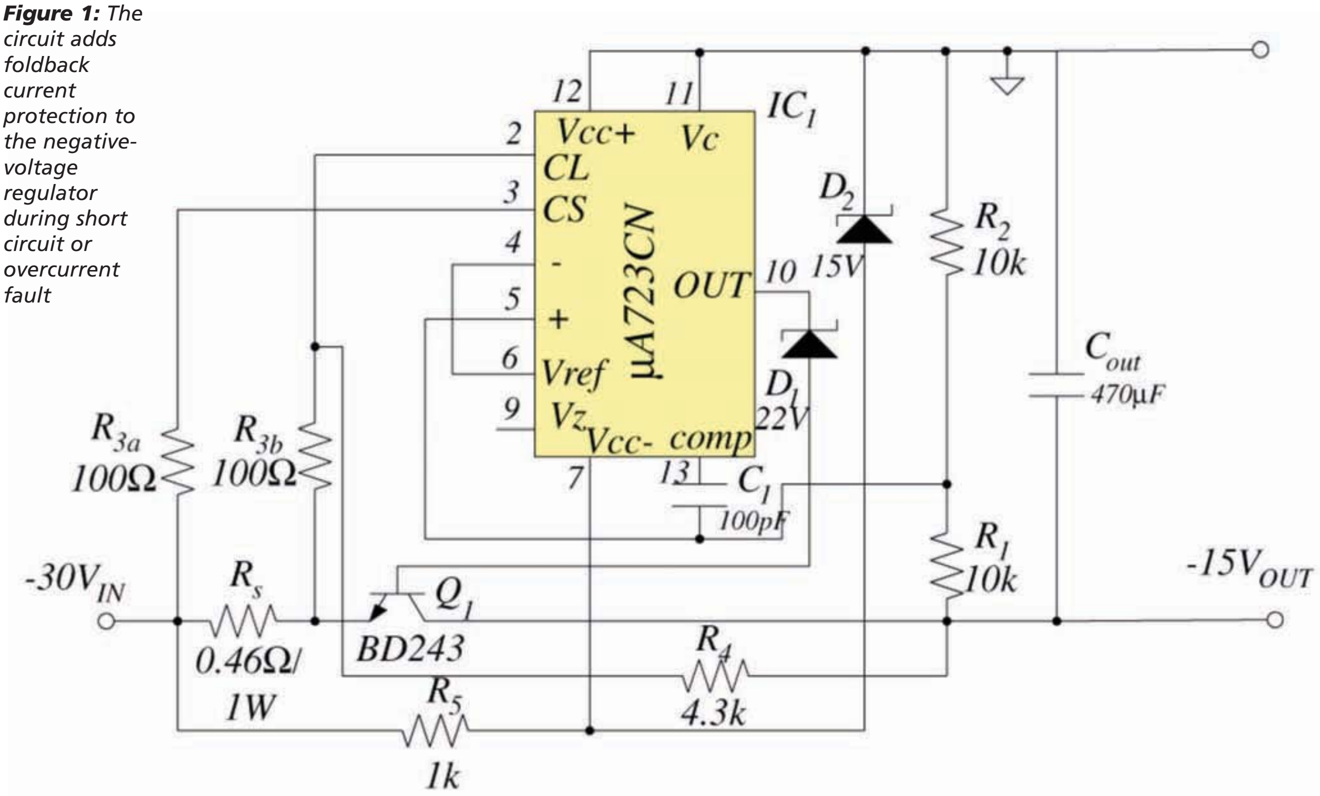

Figure 1: *The circuit adds foldback current protection to the negative-voltage regulator during short circuit or overcurrent fault*

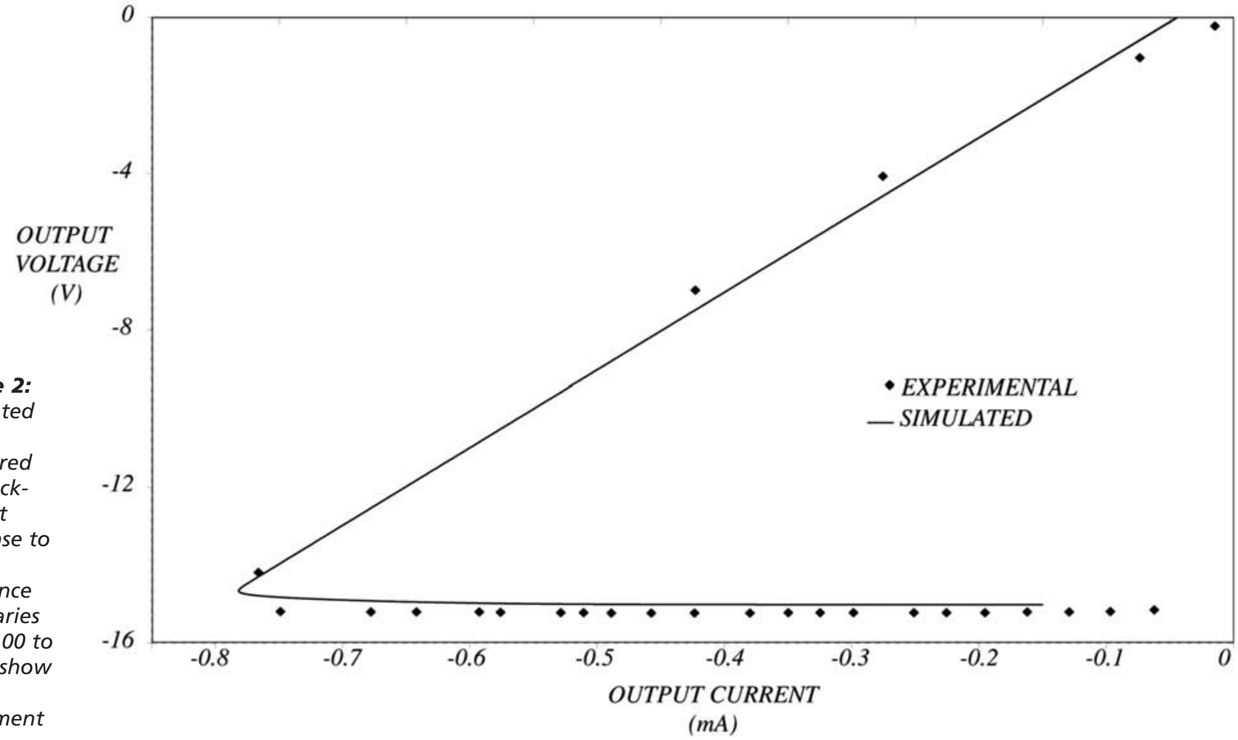

Figure 2: *Simulated and measured foldback-current response to a load resistance that varies from 100 to 0.01Ω show close agreement*

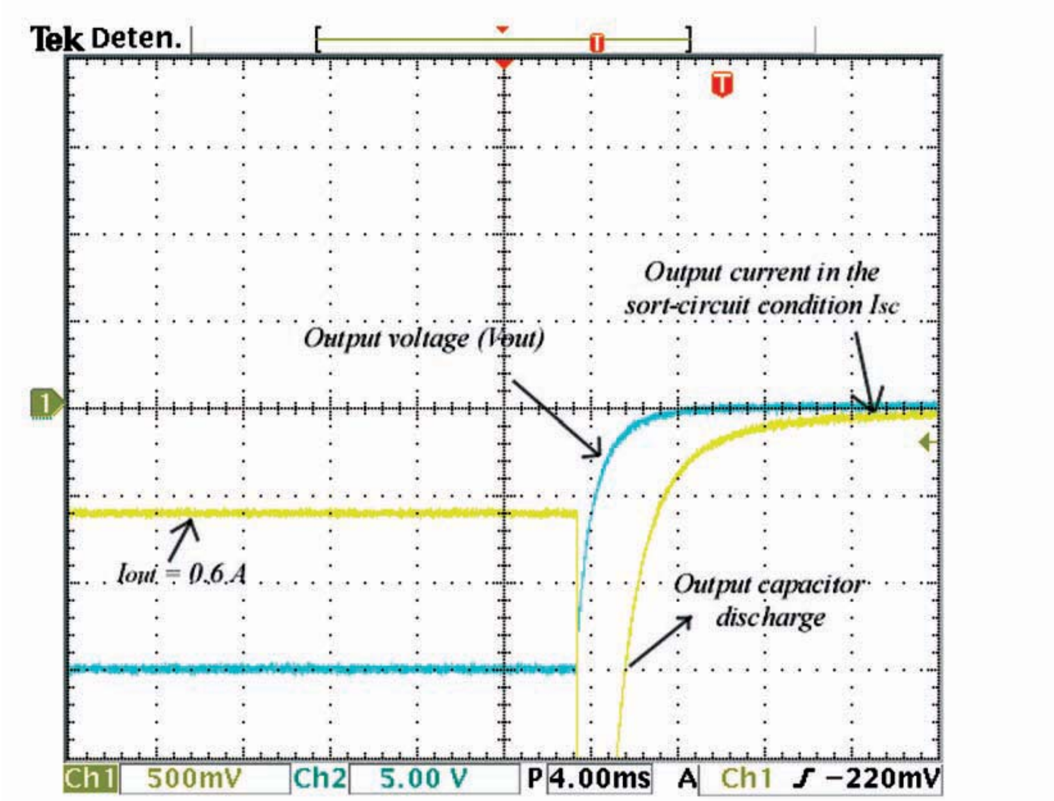

Figure 3: *Output current and voltage in the circuit of Figure 1 in a short-circuit. Current is reduced finally to* I_{sc}

fixed to 0.8A and 0.05A, respectively. With R_{3a} and R_{3b} set to 100Ω, solving the equations results in R_s=0.46Ω and R_4 = 4.3kΩ.

You can demonstrate the circuit's performance by applying a variable load from 100Ω to short circuit. As **Figure 2** shows, the output's simulated and measured voltage-versus-current characteristic, V_{out} and I_{out}, are in close agreement.

Finally, **Figure 3** shows an oscilloscope photo of the output current and voltage when a short circuit is suddenly applied at the output.

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What's wrong with DAB?

Given the big, brash, "What's wrong with DAB?" on the cover of the December 2006 issue (p18), I expected an article with DAB in the title; moreover, I anticipated a serious debate on the whole issue of the poor thinking behind the introduction of DAB, the increase in costs and complexity, the increased power consumption, etc, etc, whereas what we got was just one weakness in the technology, linked to demodulation.

Given the government's wholesale stitch-up on the 'sale of bandwidth' issue and industries' compliance in forcing the population to make obsolete a small fortune's worth of perfectly good equipment well before its natural lifespan (and the environmental concerns associated with that), and particularly given that the same government wants to tax us wholesale over global warming, has it not occurred to anyone else in our line of work that these two issues are linked?

Why are we continuing to sell analogue radios and TVs, when the transition to digital has been on the cards for years? Given that the digital TV switchover is now scheduled for 2012 and that the life of a good TV is around 15 years, we should have stopped selling analogue TVs in 1997.

Similarly, I bet the market will still be full of AM/FM only radios when DAB is rolled out nationally and the FM transmitters turned off.

The disposal of millions of radios and TVs well before their natural 'end of life' is environmentally criminal and it's this government that has to answer for it. If they are really concerned about global warming, then it's about time they entered into 'joined-up politics' and thought about the effects their policies are having. Digital TV and DAB have nothing to do with improved quality of reception, they have been biased towards greater station availability at the cost of quality, and all because this government anticipates making more money out of doing so.

These are the real issues and it's about time we owned up to them.

We have this big environmental debate going on at present, global warming, carbon-neutral technologies, green generation, etc, etc – it's all good stuff but pointless unless we all reduce our power consumption. It's something we all have to do, but it will involve some serious self-criticism: Should I use the car less, stop flying everywhere, buy local goods, etc? However, all this is pointless if we think that replacing our old 4:3 CRT TV with a 16:9 plasma screen is somehow going to save energy (because it's a more modern bit of kit).

We need to consider the entire life-cycle of the product: manufacture, distribution, use, disposal… If we are replacing perfectly good working equipment early, just to get the latest gadget, the whole life-cycle needs to be recalculated.

Like washing machines and refrigerators, our TVs, Hi-Fis and radios also need to be energy rated. If the new DAB radio set is going to consume power at four times that of your old AM/FM one, and it's on eight hours a day – every day, that's some serious extra consumption.

Don't get me wrong, I'm as much a 'gadget man' as the next guy, but I do think we need to honest about what we are selling.

If we are having a public debate, as we appear to be, let's get 'all the cards on the table'.

If Britain is going to lead the world on green issues (as Tony Blair is suggesting), then we have to get our own house in order, before preaching to the rest. Perhaps only then we can convince our US allies to curb their gross profligacy of energy consumption?

Even better, perhaps we could sell them the 'green' technology. Now that would be a new 'Industrial Revolution'! Well, a man can dream.

Graham Field
UK

Not bright enough

Regarding the UKDL article on page 57 of the December issue of Electronics World, focusing on flat panel displays, if you care to work out the contrast ratio (CR) for the dimly lit room case given, you will find that it is 16.9:1 and not 450:1 as stated. To have a CR of 450:1 one would need a reflected light of 0.79 lux, which would be somewhat dark. The second calculation is correct at 3.66:1 so my calculations seem OK.

As the figure of 450 is used a bit later on in the text, it can not be a typo error, so I, and many others I am sure, would like to know what is intended, since the article is most interesting otherwise.

John Freeman
UK

Chris Williams, Network Director of UKDL replies:

Thanks very much to John Freeman for showing up the error I made in my calculations. The "dark room" contrast ratio of 450 would, of course, require a background reflection from the screen of just 0.79Cd/m^2 and not 25Cd/m^2 as I wrote in my column.

Train information display at Waterloo railway station in London with poor CR

I can easily achieve a level of 0.70Cd/m^2 when watching TV by closing the curtains and turning the room lights down to an absolute minimum but, as John suggests, a higher value of reflectance is probably more realistic and would represent most people's typical viewing habits.

John's comments do further back my assertion that TV manufacturers' quotation of a contrast ratio measured as hundreds (or even thousands) to one is totally meaningless and downright wrong for most actual "real world" installations.

8-pin FLASH PICmicro Microcontrollers

The Flash-based PICmicro microcontrollers (MCUs) are used in a wide range of everyday products from smoke detectors to industrial, automotive and medical products. The PIC12F/16F family of devices with on-chip voltage comparators merge all the advantages of the PICmicro MCU architecture and the flexibility of Flash program memory with the mixed signal nature of a voltage comparator. Together they form a low-cost hybrid digital/analogue building block with the power and flexibility to work in an analogue world. The flexibility of Flash and the development tool suite that includes a low cost In-Circuit Debugger, In-Circuit Serial Programming (ICSP) and MPLAB ICE 2000 emulation, make these devices ideal for just about any embedded control application.

The following series of Tips 'n Tricks can be applied to a variety of applications to help make the most of discrete voltage comparators or microcontrollers with on-chip voltage comparators.

TIP 1: MULTI-VIBRATOR (SQUARE WAVE OUTPUT)

A multi-vibrator is an oscillator designed around a voltage comparator or operational amplifier (see **Figure 1**). Resistors R1 through R3 form a hysteresis feedback path from the output to the non-inverting input. Resistor RT and capacitor CT form a time delay network between the output and the inverting input.

At the start of the cycle, CT is discharged holding the non-inverting input at ground, forcing the output high. A high output forces the non-inverting input to the high threshold voltage and charges CT through RT. When the voltage across CT reaches the high threshold voltage, the output is forced low. A low output drops the non-inverting input to the low threshold voltage and discharges CT through RT. When the voltage across CT reaches the low threshold voltage, the output is forced high and the cycle starts over.

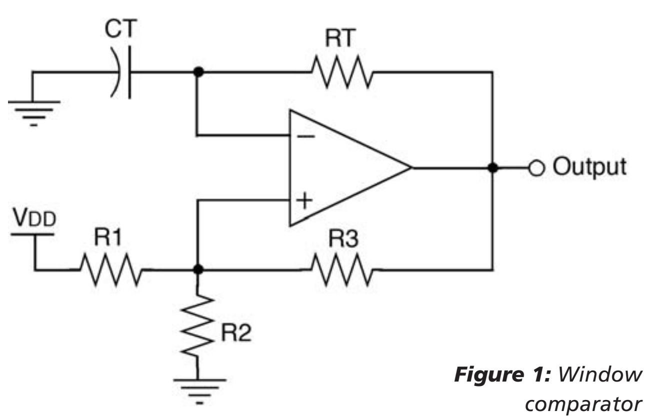


Figure 1: *Window comparator*

To design a multi-vibrator, first design the hysteresis feedback path using the procedure in TIP #3 of the December issue of *Electronics World*. Be careful to choose threshold voltages (VTH and VTL) that are evenly spaced within the common mode range of the comparator and centred on VDD/2. Then use VTH and VTL to calculate values for RT and CT that will result in the desired oscillation frequency FOSC. **Equation 1** defines the relationship between RT, CT, VTH, VTL and FOSC.

$$F_{OSC} = \frac{1}{2 * RT * CT * \ln(V_{TH}/V_{TL})}$$

Example:

- VDD = 5V, VTH = 3.333V, VTL = 1.666V
- R1 to R2 to R3 = 10k
- RT = 15kHz, CT = .1μF for FOSC = 480Hz

TIP 2: MULTI-VIBRATOR (RAMP WAVE OUTPUT)

A multi-vibrator (ramp wave output) is an oscillator designed around a voltage comparator or operational amplifier that produces an asymmetrical output waveform (see **Figure 2**). Resistors R1 through R3 form a hysteresis feedback path from the output to the non-inverting input. Resistor RT, diode D1 and capacitor CT form a time delay network between the output and the inverting input.

At the start of the cycle, CT is discharged holding the non-inverting input at ground, forcing the output high. A high output forces the non-inverting input to the high threshold voltage (see TIP #3 in the December issue of *Electronics World*) and charges CT through RT. When the voltage across CT reaches the high threshold voltage, the output is forced low. A low output drops the non-inverting input to the low threshold voltage and discharges CT through D1. Because the dynamic on resistance of the diode is significantly lower than RT, the discharge of CT is small when compared to the charge time, and the resulting waveform across CT is a pseudo ramp function with a ramping charge phase and a short, sharp discharge phase.

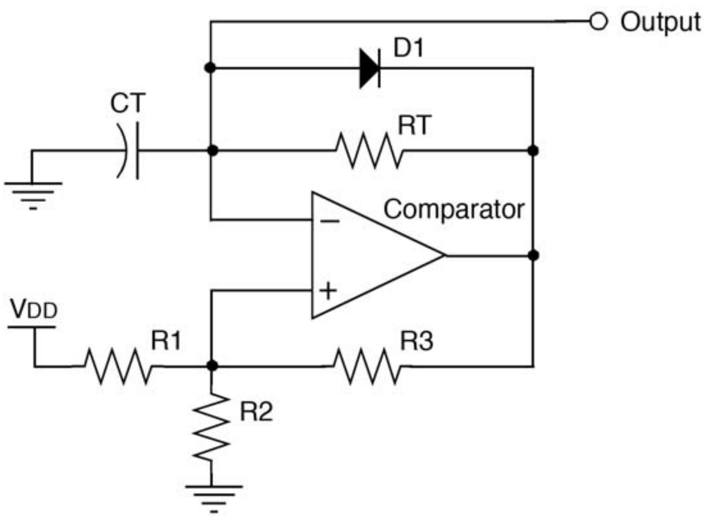


Figure 2: *Ramp waveform multi-vibrator*

To design this multi-vibrator, first design the hysteresis feedback path using the procedure in TIP #3 of December issue of *Electronics World*. Remember that the peak-to-peak amplitude of the ramp wave will be determined by the hysteresis limits.

Also, be careful to choose threshold voltages (VTH and VTL) that are evenly spaced within the common mode range of the comparator. Then use VTH and VTL to calculate values for RT and CT that will result in the desired oscillation frequency FOSC. **Equation 2** defines the relationship between RT, CT, VTH, VTL and FOSC.

$$F_{OSC} = \frac{1}{R_T * C_T * \ln(V_{TH}/V_{TL})}$$

This assumes that the dynamic on resistance of D1 is much less than RT.

Example:

- VDD = 5V, VTL = 1.666V and VTH = 3.333V
- R1, R2 and R3 = 10k
- RT = 15k, CT = .1µF for a FOSC = 906Hz

*Note: Replacing RT with a current limiting diode will significantly improve the linearity of the ramp waveform. Using the example shown above, a CCL1000 (1mA Central Semiconductor CLD), will produce a very linear 6kHz output (see **Equation 3**).*

$$F_{OSC} = \frac{I_{CLD}}{C\,(V_{TH} - V_{TL})}$$

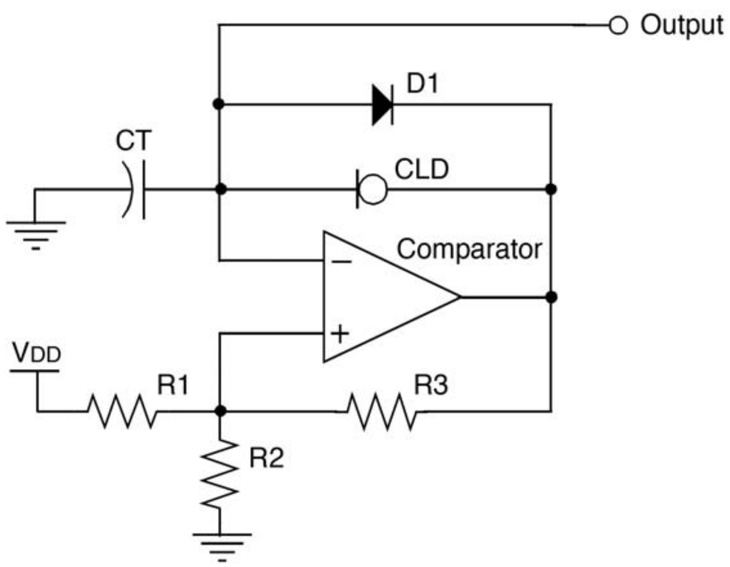


Figure 3: *Alternate ramp waveform multi-vibrator using a CLD*

LCD PICmicro Microcontrollers

TIP 3: CAPACITIVE VOLTAGE DOUBLER

This tip takes the multi-vibrator described in TIP #1 and builds a capacitive voltage doubler around it (see **Figure 4**). The circuit works by alternately charging capacitor C1 through diode D1 and, then, charge balancing the energy in C1 with C2 through diode D2.

At the start of the cycle, the output of the multi-vibrator is low and charge current flows from VDD through D1 and into C1. When the output of the multi-vibrator goes high, D1 is reverse-biased and the charge current stops. The voltage across C1 is added to the output voltage of the multi-vibrator, creating a voltage at the positive terminal of C1 which is 2 x VDD. This voltage forward biases D2 and the charge in C1 is shared with C2. When the output of the multi-vibrator goes low again, the cycle starts over.

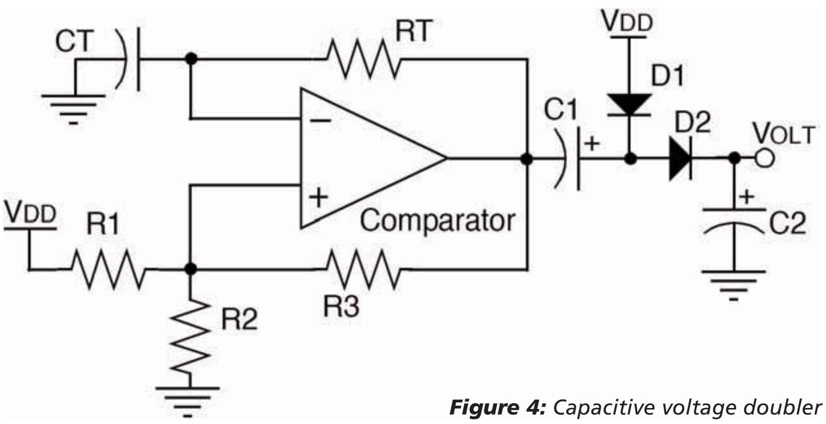


Figure 4: *Capacitive voltage doubler*

Note: The output voltage of a capacitive double is unregulated and will sag with increasing load current. Typically, the output is modelled as a voltage source with a series resistance (see ***Figure 5****).*

To design a voltage doubler, first determine the maximum tolerable output resistance, based on the required output current and the minimum tolerable output voltage. Remember that the output current will be limited to one half of the output capability of the comparator. Then choose a transfer capacitance and switching frequency using **Equation 4**.

$$R_{OUT} = \frac{1}{F_{SWITCH} * C1}$$

Note: ROUT will be slightly higher due to the dynamic resistance of the diodes.

Once the switching frequency is determined, design a square-wave multi-vibrator as described in TIP #1. Finally, select diodes D1 and D2 for their current rating and set C2 equal to C1.

Example:
From TIP #1, the values are modified for a FOSC of 4.8kHz.
- C1 and C2 = 10µF
- ROUT = 21Ω

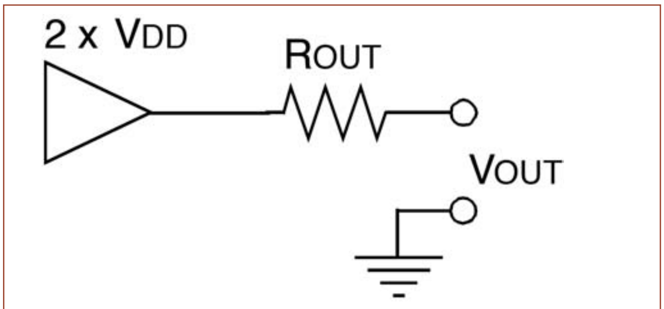


Figure 5: *Equivalent output model*

TIP 4: PWM GENERATOR

This tip shows how the multi-vibrator (ramp wave) can be used to generate a voltage-controlled PWM signal. The ramp wave multi-vibrator operates as described in TIP #2, generating a positive-going ramp wave. A second comparator compares the instantaneous voltage of the ramp wave with the incoming voltage to generate the PWM output (see **Figure 6**).

When the ramp starts, it is below the input voltage, and the output of the second comparator is pulled high, starting the PWM pulse. The output remains high until the ramp wave voltage exceeds the input, then the output of the second comparator goes low ending the PWM pulse. The output of the second comparator remains low for the remainder of the ramp waveform. When the ramp waveform returns to zero at the start of the next cycle, the second comparator output goes high again and the cycle starts over.

To design a PWM generator, start with the design of a

Win a Microchip PICDEM.net Lite Internet/Ethernet Demonstration Board

Electronics World is offering its readers the chance to win a Microchip PICDEM.net Lite Demo Board and MPLAB ICD 2 Debugger.

The PICDEM.net Lite demonstration board is an Internet/Ethernet demonstration board using the PIC18F452 microcontroller and TCP/IP firmware. The board supports any 40-pin DIP device that conforms to the standard pin-out used by the PIC16F877 or PIC18F452.

The PICDEM.net board is used to experiment with Microchip's various TCP/IP solutions. The user has immediate network access after the initial set up of the IP address. The Flash microcontroller allows modifications to the demonstration program to add application software.

The breadboard area includes a regulated 5V power supply for the addition of sensors or custom circuits for testing. Other standard or custom stack control software can be loaded for evaluation.

The board now uses the free Microchip TCP/IP stack, which is available in Application Note AN833 (DS00833). Please refer to this document for code samples.

The Microchip TCP/IP stack is a suite of programs that can either provide services to standard TCP/IP-based applications (HTTP server, Mail Client, etc) or be used in a custom TCP/IP-based application. Potential users do not need to know all of the intricacies of the TCP/IP specifications to use it, and those interested only in the accompanying HTTP server application need not have specific knowledge of TCP/IP.

The TCP/IP stack is implemented in a modular fashion, with all of its services creating highly abstracted layers, each layer accessing services from one or more layers directly below it. The stack is written in the C program-ming language, intended for both Microchip C18 and HI-TECH PICC 18 compilers, and is designed to run on Microchip's PIC18 family of microcontrollers only.

Although, this particular implementation is specifically targeted to run on Microchip's PICDEM.net Internet/ Ethernet demo board, it can be easily retargeted to any hardware equipped with a PIC18 microcontroller. The PICDEM.net supports Ethernet and RS-232 interfaces. With a standard web browser such as Microsoft Explorer, HTML web pages generated by the PICmicro MCU can be viewed. The initial board configuration is performed via the RS-232 port using a standard terminal program to configure the IP, Ethernet, addresses etc for the board. The demo board is also equipped with a 6-pin modular connector to interface directly with the MPLAB ICD 2 In-Circuit Debugger. With MPLAB ICD 2, the developer can now modify or re-program the onboard Flash-based PICmicro device to meet the specific needs.

A generous breadboarding area is also available to add special circuits for experimentation. The area is large enough to add an embedded modem to provide for dial-up capability. Several status indicators and user interface devices are provided, including a 16 x 2 LCD indicator and LEDs.

For the chance to win a these development kits, please log onto **www.microchip-comp.com/ew-picdem**

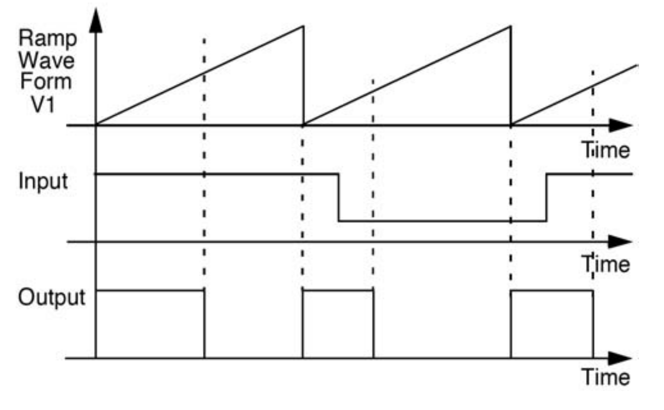


Figure 6: *PWM wave forms*

ramp wave multi-vibrator using the design procedure from TIP #2. Choose high and low threshold voltages for the multi-vibrators hysteresis feedback that are slightly above and below the desired PWM control voltages. Using the example values from TIP #2 will result in a minimum pulse width at an input voltage of 1.7V and a maximum at an input of 3.2V.

Note: *The PWM control voltage will produce a 0% duty cycle for inputs below the low threshold of the multi-vibrator. Control voltage greater than the high threshold voltage will produce a 100% duty cycle output.*

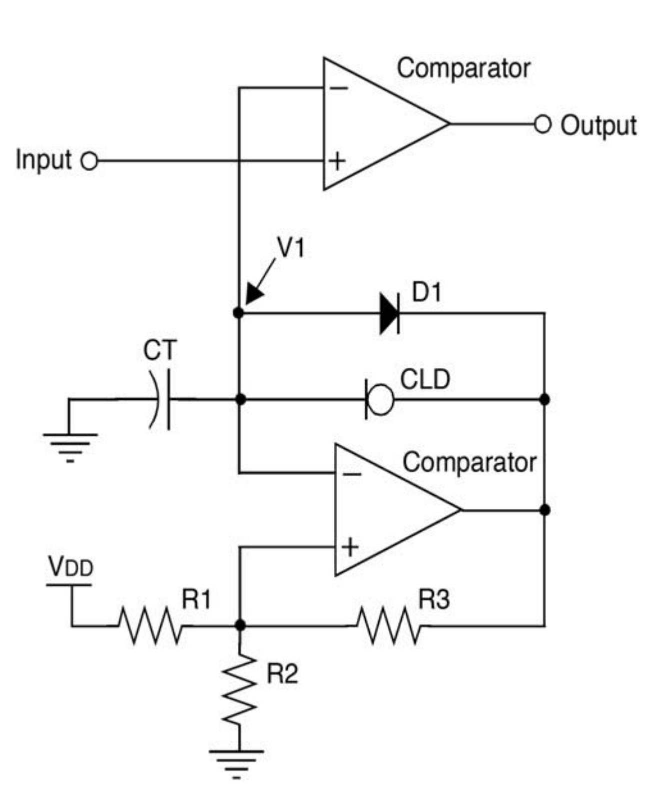


Figure 7: *PWM circuit*

Living with RoHS – the big questions

Now that RoHS is law (as of 1st of July this year), there are more questions than ever about how to cope with it. Recent research showed that the UK wasn't prepared for the deadline. Only 12% of design engineers, buyers and MRO engineers were fully compliant in readiness, ahead of RoHS officially coming into force. Whilst 37% of respondents revealed that they were "close to becoming compliant", a further 28% confessed that they had only just "started to become compliant".

There's still clearly a lot that needs to be done by the design engineering community but the main thing for engineers to realise is that they aren't alone in their quest to become compliant. Wide ranges of support services exist to help people along the way, such as those on offer at www.rohs.info. The fact that the deadline has passed means that it is even more important to access the help that exists.

The research – which was conducted amongst 263 UK design engineers, buyers and MRO engineers – shows that distributors are playing a vital role in ensuring compliance is achieved. Around 46% of those surveyed had chosen to approach a distributor for reliable RoHS support, followed by 22% who preferred to directly approach the manufacturer. Interestingly, only 9% have been relying on the government for RoHS support.

By its nature, online support is the fastest way to find out about the latest RoHS complaint products. Signing up to automatic email notification or online 'Bill of Materials' conversion services are effective ways to get new part numbers for old non-compliant components and upgrade to the latest RoHS offerings. But being able to speak to experts is also proving key for engineers who have achieved compliance.

Whilst 53% of respondents from the research considered online technical help and support services to be either "extremely" or "very" important, 39% also considered telephone technical help and support services to be "extremely/very important". There are still many grey areas around the new legislation that people are unsure about – exemptions and due diligence are just two of the 'hot potatoes'. Being able to access expert opinion over the coming months on these issues will be hugely important as the real effects of RoHS start to take place.

There are still many questions that need answering about the scope of the legislation and it will be essential to keep on top of the products that are under review for exemption. A recent example of this is semiconductor evaluation boards. Distributors and manufacturers alike believed these to be out of scope but the National Weights and Measures Laboratory, the body responsible for policing RoHS, has decided they're in.

It often isn't clear if a product is within the scope of RoHS or not. The situation for many types of industrial product will depend on how they are used. Equipment that is not dependent on electricity is also excluded such as gas boilers and petrol lawnmowers.

Q: In terms of WEEE, please can you clarify who exactly the producer would be and what their responsibilities are?

A: WEEE is a producer responsibility directive but also impacts retailers, importers, local authorities and councils, and end users (with partial responsibility in the B2B sector). Producers include every company in the UK that manufacturers, imports or re-brands electrical and electronic equipment. Please see **www.farnellinone.co.uk** for more information.

Q: Is RoHS going to China?

A: China will adopt a new legislation called the '*Administrative Measure on the control of Pollution Caused by Electronics Information Products*' in March 2007. This will affect most types of electrical equipment that is sold in China. Although it is based on the EU's RoHS directive, it has several significant differences that will become clear as its legislation is rolled out – watch this space!

Q: Can you help define what WEEE involves?

A: The obligations for WEEE are wide ranging. A 'producer' will have to pay for the collection, treatment, recovery and recycling of WEEE. This will include:

- Meeting recovery and recycling targets set by the category
- Marking WEEE products from 1st April 2007
- Providing information on new EEE placed on the market
- Placing a unique producer number on invoices after 1st June 2007
- Reporting on weights and numbers placed on the market in the UK by category
- Reporting on weights and numbers placed across member states and record obligations met.

Q: We sell into China and are concerned about how they will define 'on the market'. Can you shed some light on this please?

A: Unlike the EU RoHS, put on the market in China is defined as when the product is sold. Products listed in a catalogue must comply and it will be necessary to have products pre-tested by one of the 200 licensed analysis labs in China. It's worth noting that China RoHS comes into force on 1st March 2007 and so all EIPs that are sold in China must be labelled from this date.

Q: How do I show I'm WEEE compliant?

A: Evidence of treatment should be supplied by an authorised treatment facility and evidence that you use an accredited recycler or exporter. This evidence will be used to justify issuing certificates of compliance at year-end. It's worth noting that evidence may be brought and sold to meet targets.

Q: Dates for WEEE legislation in the UK keep changing. What are the major deadlines?

A: This varies depending on the 'role' you play in WEEE compliance. But here are a few key dates that you should note:

- Compliance schemes apply to agencies between 4th January and 31st January 2007
- Sign up of producers to compliance schemes should be completed by 15th March 2007
- By 1st April 2007, compliance schemes should register their members to agencies; all non-household and household obligations commence
- Remaining household obligations commence on 1st July 2007.

Gary Nevison is chairman of the AFDEC RoHS team, board director at Electronics Yorkshire and head of product market strategy at Farnell InOne. As such he is our industry expert who will try and answer any questions that you might have relating to the issues of RoHS and WEEE. Your questions will be published together with Gary's answers in the following issues of Electronics World. Please email your questions to EWeditor@nexusmedia.com, marking them as RoHS or WEEE.

How organic is your TRANSISTOR?

By Chris Williams, UKDL

At the core of the nascent Plastic Electronics revolution is the fact that diodes and transistors can now be printed on to flexible plastic substrates using inks that are composed of semiconductive organic polymers. This process can take place at low temperatures (including room temperature) and the inks can be printed using ink-jet printers, not entirely dissimilar to the one you have at home.

So, can organic polymers – i.e. long chain molecules based on carbon – fully replace silicon? Let's look at what is involved.

Many know the tortuous route that silicon goes through to create a transistor circuit: first a single crystal structure ingot is grown from molten silicon. The ingot is then sliced into multiple individual wafers. These wafers are then subject to a series of high-temperature sequential deposition and patterning processes to create the epitaxial layer which contains separate areas of p+ and n- doped materials, in order to create the required transistor junctions. The wafers will then be scribed and diced into individual circuits – the "silicon chip" – each of which may contain single transistors, or hundreds, thousands or millions of interconnected transistors to perform the necessary circuit functions for which it is designed.

The system benefit is that electrons can travel at high speed through the crystalline semiconductor material with little obstruction. This has allowed the development of low voltage, low power and high-efficiency devices.

The benefit of using silicon reduces dramatically when moving away from crystalline silicon (device grown on a single crystal wafer) to the different types of silicon that are widely used in the displays industry. Here, we want to utilise the switching properties of silicon transistors, but can't create devices using crystalline silicon. Imagine that the LCD display used in your laptop computer has at least one transistor at each pixel position in the display. If we wanted to use crystalline transistors, we would need to grow silicon wafers of the same size as the display itself. Can you imagine the difficulty in making a silicon wafer that would be 15 inches in diagonal? Or if we want TVs, how about wafers up to 40 inches or more in diameter? It can't be done – at least not for the price we want to pay for a computer screen or a TV.

The answer has been to use amorphous silicon to create transistors across the whole surface of the display screen. Invented in the UK at the University of Dundee, this process has become the *de facto* standard in display factories around the world in Active Matrix Thin Film Transistor LCDs (AM TFT LCDs) and is based on high temperature sequential deposition and patterning processes of silicon and its required dopants directly onto glass substrates. This method has the ability to create transistors at any point on very large substrates – developments in Japan are leading to the deposition of amorphous silicon on glass substrates up to 3 x 2.5m in a proposed Generation 10 LCD plant. That is pretty impressive and the equipment to do this is extremely expensive. The complete LCD fabrication plant will cost up to $10bn by the time it has been completed.

Amorphous silicon as a process that can be scaled for use with very large substrates, but the cost is huge and the transistor performance is very poor compared to crystalline silicon – amorphous silicon has a semiconductor mobility that is some 600 times slower than crystalline silicon. This has a huge impact – it means that an amorphous silicon transistor cannot pass as many electrons in a given time as a crystalline silicon transistor, so everything it does is slower in comparison. The reason for this is that the amorphous silicon is literally deposited and patterned as multitudes of independent domains of silicon, so when electrons want to pass through the system as part of a circuit function, they have a reasonable speed through each silicon domain, but when they arrive at a domain boundary, it takes energy and time for the electrons to jump across into the next silicon domain.

Even with this fundamental limitation, amorphous silicon has become the *de facto* standard method for creating active matrix backplanes for displays. It works; it is reasonably reliable and repeatable, as all the transistors in a display system will need to behave pretty much identically. If I apply a voltage for a specific time to a transistor I expect it to pass a certain amount of charge based on the number of electrons that pass through it whilst it is switched "on". If I apply the same voltage for the same time to a number of transistors I need them all to pass the same amount of charge. Why? The number of electrons they pass sets the charge on a capacitor which determines how much each pixel or sub-pixel is turned on. I want all of the pixels to respond in a similar fashion as it is no good if the transistors all have different characteristics and allow different amounts of charge to pass through as this will result in totally random shades of colour to be presented on the display screen.

So, going back to the original question: Can organic (carbon-based polymer) transistors replace silicon? The answer is "yes" and "no". Organic transistors are slower than crystalline silicon and, today, most are even slower than amorphous silicon transistors. But, the production cost of making organic transistors will be less, probably much less, than their amorphous silicon counterparts. For this reason alone, the effort will be put in at companies around the world to solve the current problems of scale (depositing printed transistors on bigger substrates), improving device performance and maintaining uniformity of characteristics to allow printed organic transistors to become a real competitor to amorphous silicon. Expect great advances over the next year.

Chris Williams is Network Director at UK Displays & Lighting KTN (Knowledge Transfer Network)

Cherry Switches To Tiny Sizes

Cherry Electrical Products's range of small-size switches now includes the new DH ultra-miniature switch, measuring just 8.2 x 2.7 x 6.2mm.

These minuscule devices offer a maximum switching voltage of 30VDC and a maximum switched current between 300 and 500mADC, depending on model. With a mechanical life of over 50,000 operations, parts operate between –25°C and +70°C.

Especially designed for low switching currents and voltages, these switches are available with optional integrated auxiliary actuators and can be soldered or used lying/standing on a circuit board.

Cherry switches and control assemblies allow automotive manufacturers to offer their customers products that include door, seat and console controls, intelligent door latch and lock systems, advanced steering wheel and instrument panel controls and leading-edge seat belt sensors.

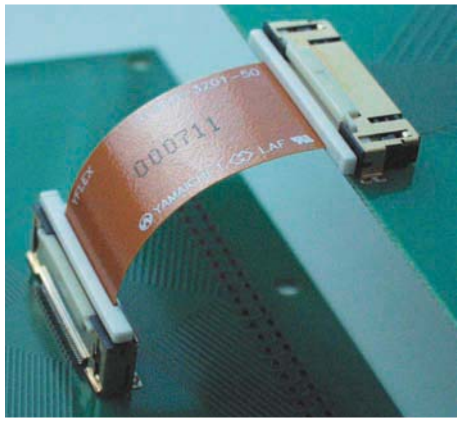
From a heritage as a leading manufacturer of snap action switches, Cherry has expanded its line of products to sensors and controls. Cherry's affordable, high-performance, speed and proximity sensors, electronic oven controls and patented cook top sensors bring greater functionality, efficiency and safety to a variety of everyday products.

www.cherry.de

Flexible Interconnect System For 12.5Gbps Applications

A sophisticated high-performance flexible interconnect system is now available from ACAL Radiatron offering design engineers and system builders a compact alternative to traditional options.

Offering significant advantages over polyamide, the Y-Flex system manufactured by Yamaichi features superior performance characteristics derived from its specialised construction based on liquid crystal polymers, delivering lower thermal expansion and less absorption of humidity. The flexible 100Ω impedance system is seen as ideal for present-day and next-generation high-speed data rates up to 12.5Gbps. It is shielded to ensure high immunity to crosstalk.



Y-Flex is used in conjunction with the HF503 connector, which is polarised and can be vertical or right-angle for PCB mounting. These fine-pitch surface-mount connectors feature an integral 'stiffener' for mechanical strength and a quick-lock retaining latch with an audible click. They require a low insertion force on mounting and can be disconnected easily via a simple release button for maintenance or development needs.

A small bend radius allows the interconnects to be installed in the most compact spaces. They are available in a range of standard lengths, from 70mm to 430mm, or as customised lengths to suit individual applications. Connector options are from 20-way to 60-way.

www.acal-radiatron.com

Link For Tasking At The Mathworks

The MathWorks today announced Link for Tasking, a new product that builds, validates and verifies automatically generated code using MATLAB and Simulink with Tasking, Altium's compiler tool chain for embedded software development.

Tightly integrated with The MathWorks products for Model-Based Design, Link for Tasking enables engineers to generate algorithm code, libraries, makefiles and test harnesses from a Simulink model, and automatically load them into the Tasking Integrated Development Environment (IDE) for execution plus analysis.

This improved workflow minimises the need for hand coding, which decreases errors and shortens the time typically needed for on-target rapid prototyping and production code deployment, two factors that are critical to successful delivery of sophisticated electronic products in the automotive industry.

Link for Tasking provides Processor-In-the-Loop (PIL) testing features that enable embedded object code to be co-simulated with Simulink. Once the code is loaded into the Tasking IDE or deployed into the embedded hardware, it can be tested with the same executable specification used for code generation. This capability automates code verification and validation, without requiring users to switch between tools.

Unlike other PIL solutions that only support testing for one type of microprocessor, Link for Tasking enables PIL testing for many popular microprocessors supported by Tasking, including Infineon TriCore.

www.mathworks.com/products/tasking

Website Directory

To reserve your website space phone Matthew Dawe 01322 611261 or email matthew.dawe@nexusmedia.com

Alpha Wire

www.xtraguard.com

Leading supplier of advance solutions and service in wire, cabling and tubing, Alpha Wire International, has announced the launch of a new website dedicated solely to its range of hazard-matched cables, XTRA•GUARD that provide extreme performance for extreme environments. The new website www.xtraguard.com represents an industry first for a website dedicated solely to a single brand of cable. The site offers an intuitive selection aid to help engineers correctly specify cable for their application's extreme environment.

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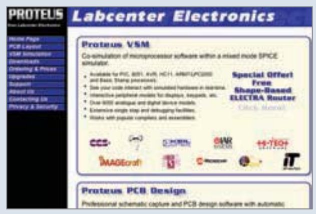

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