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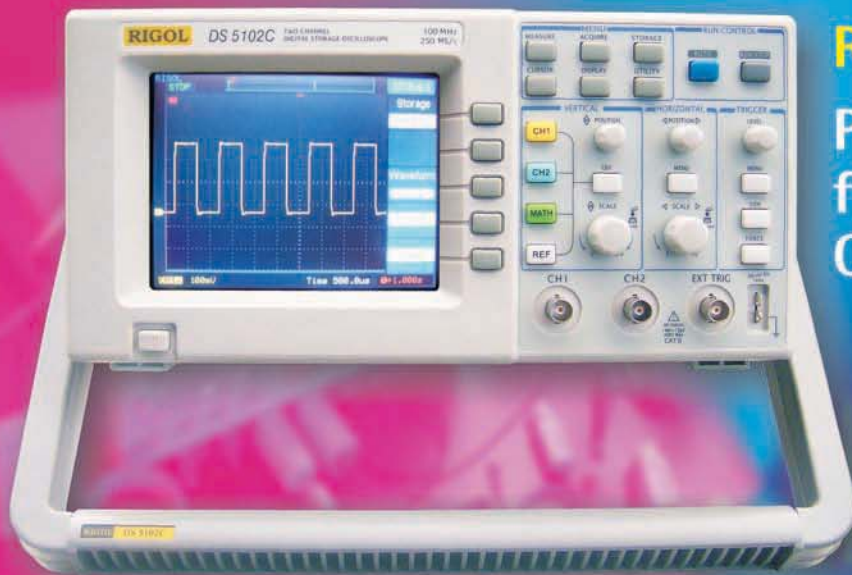
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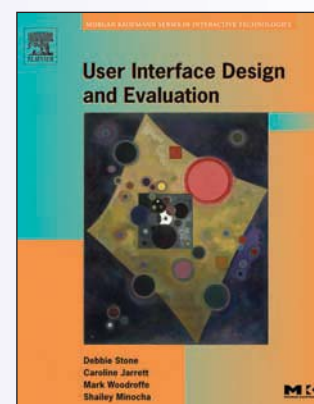
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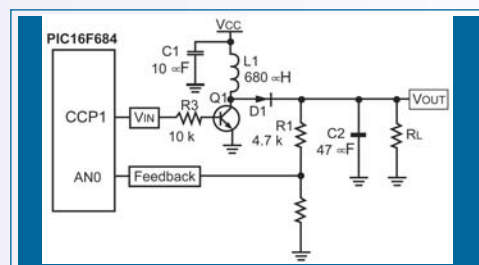
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The colourful world of scopes

Fifteen years ago, mid-range for digital oscilloscopes was in the 300MHz domain. Today, mid-range is an entirely different proposition, covering a market with needs of between 350MHz and 3GHz. Trends nowadays are for new cars, for example, to have a lot of electronics in them, including high-speed serial buses (CAN, LIN, FlexRay etc). We are seeing a lot of high-speed serial buses, such as USB, Gigabit Ethernet and PCI-X among others affecting not only PCs but also consumer electronics. That means that, increasingly, the embedded space is making more use of serial buses. And all of these standards need to be tested. There's a lot of complexity in modern day electronics and developers need to know how all of the different features and functions affect and work with each other in various applications.

With the complexity of systems to be tested, grows the complexity of oscilloscopes with which to monitor and diagnose problems. It's good to know that innovation is rife in this market: we are seeing increasing sample rates, longer record lengths ('deep' memory) and increasingly faster capture of waveforms – and in real time, at that. The newest oscilloscopes offer the highest bandwidth, greatest timing resolution, lowest noise floor and thorough analysis capabilities for current and emerging serial data standards.

As engineers need to handle information on multiple channels – to monitor several signals simultaneously – the capabilities of oscilloscopes need to be expanded too. Capturing anomalies is one important reason why engineers need to capture signals in real time. Checking out the dynamic range of the signal is another.

Today there are two classes of oscilloscopes on the market: DSOs (digital storage oscillo-

scopes) and DPOs (digital phosphor oscilloscopes). Either of these two classes can offer real-time or semi real-time capture, but have different capabilities. The capture rate will vary however, from scope to scope. For example, the DSOs offer a much lower capture rate than DPOs do. DPOs are also about the acquisition rate and not just about improved sample rates, which have already entered the 40GS/s domain. Because of the high acquisition rate, the chance of seeing a rare event in the signal is increased. Oscilloscope makers are increasingly resorting to using more advanced technologies in order to meet the growing needs of engineers. Advanced silicon germanium processes are just one example. Tektronix has worked closely with IBM to ensure that the silicon germanium process can handle its requirements, such as being used in fast analogue to digital converters as well as for triggering purposes. In Tektronix's latest series of scopes – the DPO 7000 – the data that is input in the demux is as fast as 5Gbytes/s.

Such speed will allow users to capture even the most elusive of glitches or unstable conditions.

Fifteen years ago, oscilloscopes would allow users only a limited number of capabilities. Nowadays, bigger displays, different interfaces, different and larger capacity storage mediums and, indeed, interactive configuration control panels, allow the user significantly greater flexibility – all thanks to innovations in this field.

Svetlana Josifovska
Editor

EDITOR: Svetlana Josifovska **E-mail:** svetlana.josifovska@nexusmedia.com **EDITORIAL E-mail:** EWeditor@nexusmedia.com

EDITORIAL ADMINISTRATION: +44 (0) 1322 611274 **E-mail:** EWadmin@nexusmedia.com

PRODUCTION EDITOR/DESIGNER: Jane Massey

SUBSCRIPTIONS: Customer Interface Ltd, Cary Court, Somerton, TA11 7BR **Telephone:** 0870 4287950, **Fax:** 01458 271146

SUBSCRIPTION RATES: 1 year: £45 (UK); €115 (Europe); \$150 US & worldwide

DISPLAY SALES EXECUTIVE: Reuben Gurunlian +44 (0) 1322 611261

PRODUCTION EXECUTIVE: Dean Turner +44 (0) 1322 611206 **E-mail:** dean.turner@nexusmedia.com

PRINTER: William Gibbons Ltd • **ORIGINATION:** Impress Repro A1 Parkway, Southgate Way, Orton Southgate, Peterborough, PE2 6YN

NEWSTRADE: Distributed by Seymour Distribution Ltd, 86 Newman St, London W1T 3EX. • **PUBLISHING DIRECTOR:** Tony Greville

If you are experiencing problems getting copies through your newsagent, please call Debbie Jenner on +44 (0) 1322 611210

Electronics World is published monthly by Nexus Media Communications, Media House, Azalea Drive, Swanley, Kent, BR8 8HU Nexus Media Communications is a trading name of Nexus Holdings Limited. Registered in England. Registered Number 5346404. Registered Office: Hanover House, 14 Hanover Square, London W1S 1HP

ISSN 0959-8332

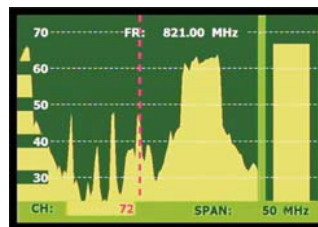
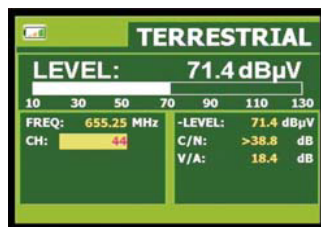
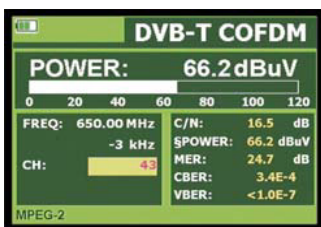
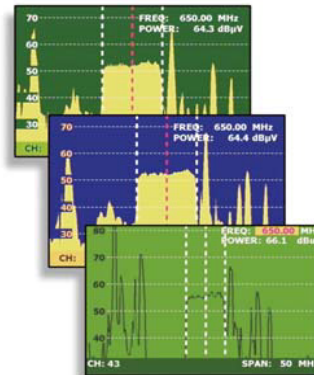
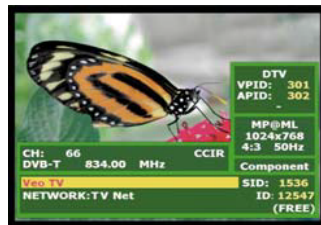
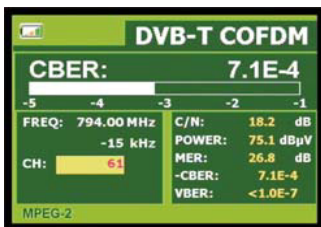


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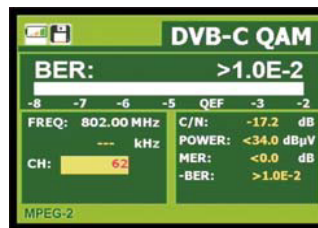
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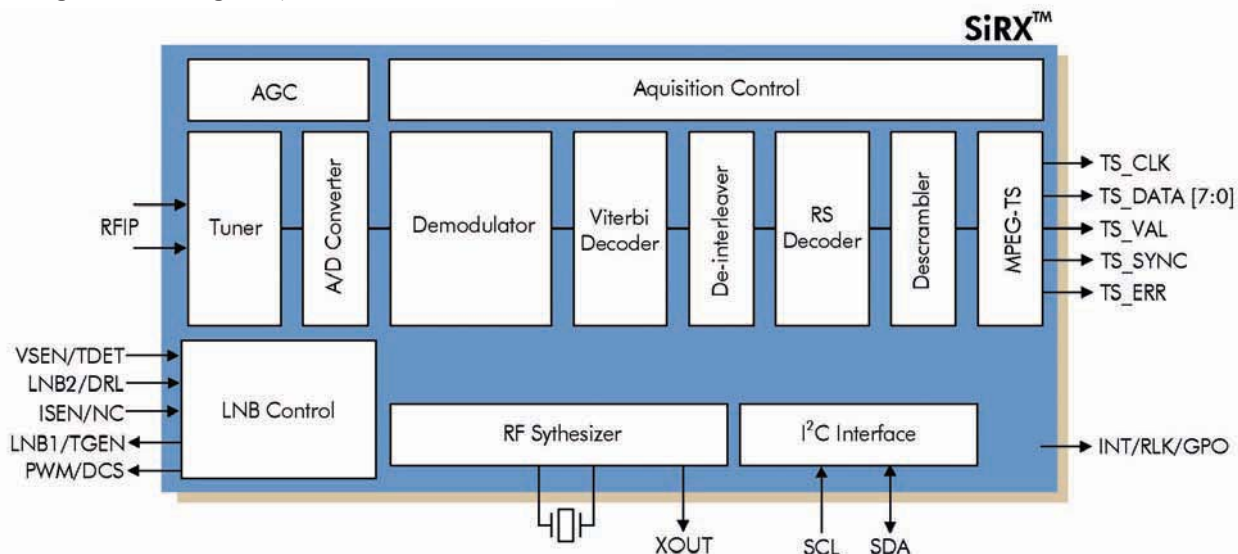
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Block diagram for the single-chip SiRX DBS receiver



Single-chip DBS receiver ready

Silicon Laboratories has last month announced its fully integrated single-chip satellite RF front-end for direct broadcast satellite (DBS) – the SiRX.

Typically, satellite set-top box OEMs buy in the receiver front-end or the network interface module (NIM), but these contain several ICs and many discrete components. They also require functions for sending power and controlling the LNB via the coaxial cable

to be implemented by the OEM outside the NIM. The SiRX integrates all of these functions into a single 6 x 8mm CMOS IC.

The integrated functions, therefore, include a satellite L-band RF tuner, a dual-mode DVB-S/DSS digital demodulator and a power-efficient, step-up supply controller for the LNB. The integration of the LNB supply controller is an industry-first and includes sup-

port for DiSEqC 2.x and legacy tone/voltage LNB signalling.

The SiRX receives input from the LNB, which is part of the satellite dish housing, and outputs a digital bit stream that carries the digital audio and video satellite services for further video/audio decompression and display by the base-band processor.

In addition, the SiRX also contains an on-chip hardware blindscan feature that

improves channel scan time by a factor of 10 compared to many existing solutions. This feature reduces set-top box installation time for FTA applications where channel locations are unknown.

Samples of the SiRX family are available now in a lead-free, RoHS compliant 44-pin QFN package.

Full production is scheduled for the middle of 2006.

IBM says molecular transistors are the future

IBM in Zurich is working on turning single molecules of between 5 and 50 Angstroms into working transistors. "You can use a single molecule as a rectifier, switch, logic or memory, for example," said IBM Zurich's Dr Walter Reiss.

One thing that might stop its quick deployment in commercial applications, however, is the electrode con-

tacts. "The atomic electrodes have to be sharp and the chemical compatibility has to be there for the contact between electrode and transistor (molecule) not to be loose," said Dr Reiss.

One potential solution is to use a metal break junction (MBJ) on a flexible substrate as a contact. However, this is not the only challenge that will

need to be surmounted to turn a molecule into a successful transistor.

"New technology is successful if allows new applications, it is cheaper to manufacture and it offers better performance," said Dr Reiss. "A perfect example of it is the [traditional] semiconductor transistor. It is reliable and cheap and easy to man-

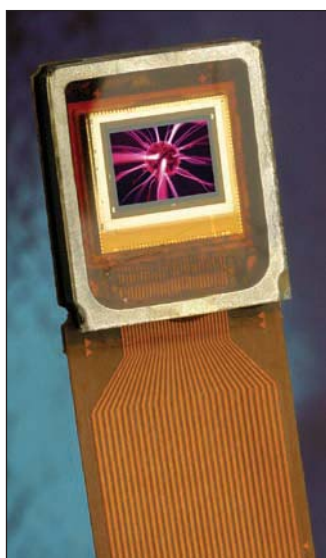
ufacture, as well as it offers better performance. But organic transistors will replace amorphous silicon [in the future] – it's most efficient technology."

"Sometime in the future, we'll probably see molecular atomic devices, but it will depend on how we control them on the nano and microscopic scale," he added.

MED's P-OLEDs are deemed commercial success

Scotland-based polymer organic light emitting diodes (P-OLEDs) developer, MicroEmissive Displays (MED), has started shipping its first displays to a customer in the Far East. They will be used in a 'nightscope' type product, which will allow the consumer to 'see' in the dark. MED co-founder, Ian Underwood, says MED's displays are also suitable for MP3 players, digital still cameras, viewfinders in digital camcorder and as eyewear for games, mobile phones etc.

The plan is to attack the market initially in two directions: for electronics viewfinders (EVF) and wearable displays. "The market size for EVF for digital still cameras will rise from 80-plus million in 2005, to 100-plus million in 2010. For camcorders that number will rise to over 19 million in three years. The EVF share will grow from nine million now to 17 million units in 2010," said Underwood. "There's not much of a market today, but some studies suggest that in the next two years



MicroEmissive Display's P-OLED is one of the smallest displays around

it will start growing thanks to the mobile phone. We are not tightly linked to the mobile phone market, but if the need for wearable displays emerges, we'll be ready," he said.

MED uses a CMOS technology and a polymer to create its displays. The combination offers good image quality, fast moving images and it does not need

polarisers. It also offers low power, small form-factor and light weight, since the CMOS allows high levels of integration and high pixel factor.

However, one big disadvantage is that when the P-OLED image is magnified, the defects that are part of that display are also magnified. "This has challenged and stretched us," said Underwood.

Equally, because of the close vicinity of the pixels, there's lateral conductivity, where the switching of one pixel leads to the switching of the ones close to it. In addition, there's optical coupling between the closely-knit pixels (blue coupling with red etc). "We are a small group and we can't address all of these challenges by ourselves. But, if anybody knows more about these issues then we'd like to hear from them," said Underwood.

MED's P-OLED displays are only 7 x 5mm large. Their usage is currently two hours per day, which Underwood says is sufficient for the operation of EVFs.

Funkwerk Enterprise Communications (FEC) has partnered with Ziehm Imaging to integrate its WLAN technology into Ziehm's line of mobile X-ray products in the medical field. With FEC's technology, X-ray units can communicate wirelessly with the network for image data to be fed directly over standard DICOM (Digital Imaging and Communications in Medicine) to archiving and communication systems, such as PACS (Picture Archiving and Communication System). FEC hardware and software have already been given medical approval as specified by European standard EN60601-1-21.



VaST Systems Technology (VaST), Gaia System Solutions, Toyota Electronics, Toyota Techno Service Corporation and Fujitsu TEN have announced the formation of the V-ECU (virtual electronic control unit) Club, an alliance that invests in virtual platforms for development of integrated electronics for the automotive industry in Japan. Industry research shows a sharp rise in the electronics content in vehicles, with complexity rising proportionately. Today's high-end models have up to 70 electronic control units (ECUs) connected by an average of five buses under the control of millions of lines of software interacting in real time. Recently, BMW reported that up to 70% of the development cost of an ECU is related to software and that some 40% of a vehicle's cost is determined by electronics and software.



Micron and Intel are creating a new company to manufacture NAND Flash memory. The devices are earmarked for use in consumer electronics, removable storage and handheld communications systems. The company will be called IM Flash Technologies. It combines Micron's NAND expertise and Intel's multi-level cell technology. Production will initially take place in manufacturing facilities located in Boise, Manassas and Lehi in the US. The new company will benefit from Micron and Intel's process technology expertise to aggressively convert to advanced 72nm and 50nm nodes. Initial production is expected in early 2006.

Join efforts to solve energy problems



Sir Richard Friend (left), the creator of the first plastic transistors and founder of Plastic Logic, spoke of his concern regarding the world's energy crisis at a recent event at St John's Innovation Centre in Cambridge.

"I am concerned that if we are to keep ourselves warm and fed

in the years to come, we need to create new technologies and markets. What are we going to do about it?" he said.

"There are small examples of how to harness natural resources. The powers of sunlight, wind, tide and biomass are amazing. With a fraction of sunlight harnessed, for example, we could satisfy all of our energy needs," he said.

Sir Richard carried on to say that the Cambridge area, with

its small firms phenomenon, its University and a concentration of VC funding, is good at providing niche solutions that have the potential to turn out "very big". However, he added, to solve the energy problem it will take more than just efforts at universities.

"Now is the time to do something about it and it is not enough to turn to universities to create new industries," he said.

De Montfort University in Leicester, UK, plans to set up a new three-year degree course in Electronics Games Technology. This will be a BSc course, scheduled to start in September 2006. It aims to give students the skills required to work in three main disciplines of the industry – design, electronics and computing software. The course is a collaborative venture between the University's Faculty of Computing Sciences and Engineering and the Faculty of Art and Design. This course follows De Montfort University's launch of the BA (Hons) in Game Art Design degree, in September last year. This course focuses on the artistic elements of creating computer games and works closely with the industry.

Ω

The contamination control and cleanroom industry is in a buoyant mood, says a report from market research company McIlvaine. Some 80% of cleanroom industry professionals state the industry's mood as optimistic or very optimistic, while some 50% have made 'significant' investments in the past 12 months. Around 40% plan to do so in the coming year. The strongest demand is seen in the development of medical devices. Those operating in the semiconductor market have the most advanced procedures. Further growth in this industry is fuelled by nanotechnology applications such as miniaturised sensors and microsurgical instruments. The global cleanroom demand is expected to grow by 45% a year from the current \$11bn.

Ω

Swiss group Huber+Suhner has received the ISO/TS16949 certificate – a stringent automotive industry standard – for its facility in Tzcew/Gdansk, Poland. Here, Huber+Suhner manufactures customised RF components for use in vehicles, as well as radio basestations for mobile telephony. The automotive industry is a big user of RF components for navigation systems, pre-crash sensor systems, distance radar, tyre pressure monitoring, satellite radio and so on. Among the company's clients are Audi, Mercedes, GM, Renault, Volkswagen and specialised first tier suppliers such as Autoliv, Fuba, Harada, Bosch, Delphi and PKC. Huber+Suhner has been producing RF products at its subsidiary company in Tzcew since 2003.

Power electronics can rely on diamonds

Single crystal diamond could replace silicon in power electronic devices, confirmed research carried out at the University of Cambridge. Out of several suitable materials, including silicon, GaAs and GaN, diamond has been found to have the best electric field, better hole and electron mobility and better thermal conductivity than the others. It is also a wide-bandgap semiconductor, which means that its electrons will require much higher energies to change state and, hence, change the structure to a conducting one. Diamond, therefore, retains its electronic control at high voltage and high temperature.



However, costs associated with synthesising diamond are rather high and yields are rather low – only 5mm of high-quality substrate can be made unlike with silicon, where substrate sizes have already reached 300mm.

These are the findings of the University of Cambridge team, led by Professor Gehan Amaratunga, which also works closely with Element Six and Dynex Semiconductor, under the umbrella of the Carbon in Power Electronics (CAPE) project. "You can use diamond in applications for voltages of above 10kV and currents of over 100A, or in amplification, where you'll stack them up in a power system," said Prof Amaratunga. "But, we are far

away from having a diamond transistor."

The team is currently working on increasing the hole mobility in a diamond Schottky device. "If you dope the diamond with boron, you will get a higher hole mobility; it becomes almost metallic and this is a lot better than using SiC."

This diamond structure can be used as a unipolar device for voltages of between 800 and 1700V. "You can make good diamond Schottkys with gold contacts. It is a good device to replace silicon in power electronics systems," added Prof Amaratunga.

Element Six has already created Schottky-barrier diode (SBD) based on a synthesised diamond structure. The SBD is capable of operating at 1700V and a current density of 10A/cm².

No need to type SMS, just dictate it

Dictating SMS and emails and carry out searches on the Yellow Pages, Google or Yahoo by voice from your mobile phone will be available as early as next year in Europe.

Michael Thomson, VP of international marketing of speech recognition firm Nuance Communications, formerly ScanSoft, says that the processing technology, power-saving handsets, voice recognition cores and vocabularies are already here and that operators are asking for such services.

"This area is red hot right now," he said. "Dictation to a mobile phone opens up a lot

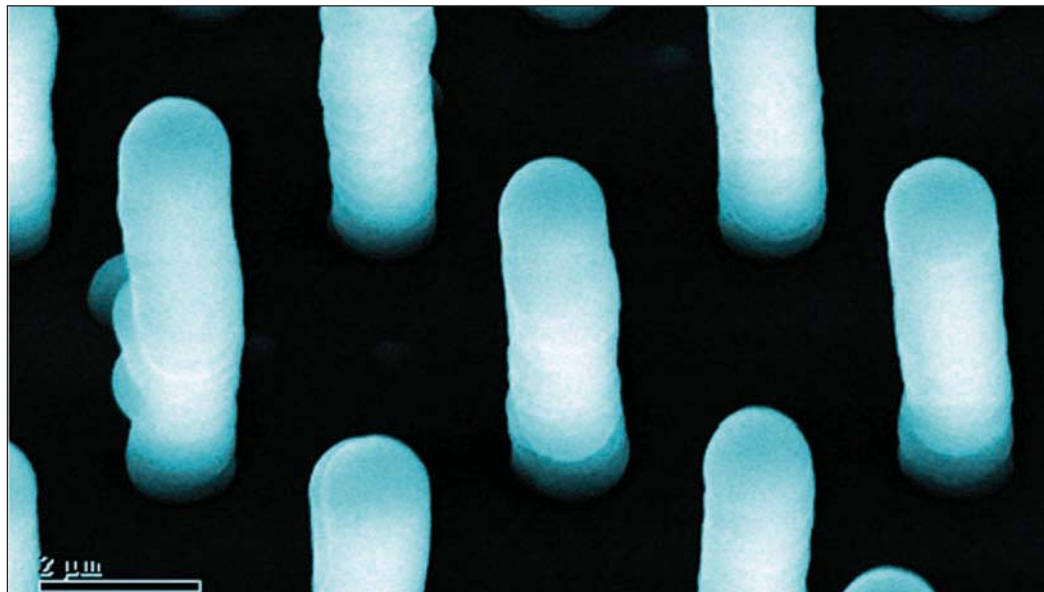
of opportunities for operators and they are asking for this, they are the main driver."

Nuance builds the core technology and also provides the tools and services for such applications. "Only about a year ago, dictation was only available for PCs. Now, embedded dictation is all done in the handset. This is the largest single investment area for us and it requires dozens of engineers," said Thomson.

Europe is expected to see the first dictation-enabled mobile phones during 2006. However, the US is already leading the way with the launch of Samsung's P207

model, available from Cingular Wireless, which contains VoiceSignal's VSpeak full text-to-speech (TTS) solution that can read emails, text messages, web pages or any other text aloud. This offers a complete speech interface solution to look up or dial a contact, dictate a text message or have a text message or other content read aloud. The software's footprint is only 300kB of memory. If combined with a Bluetooth headset, the solution can offer an entirely eyes-free interface. VSpeak is planned for all levels of handsets, including its availability in over 20 languages.

Carbon nanotubes likely to drop out of the IC race



Conductive interconnects created with carbon nanotubes [Source: NASA]

Carbon nanotubes (CNT) are a promising technology but there's a risk that they won't make it into commercial reality for high-density IC integration, says Dr Siegmund Roth of the Max Planck Institute in Germany.

"It is not down to technology but to economy – people get fed up with it [technology] and change. The Kondratiev Cycles imply that each technology starts slowly, then reaches experimental growth and then comes to saturation," said Dr Roth.

Nikolai Dmitriyevich Kondratiev was a Russian economist that lived in the 1920s. He first described economic cycles consisting of four stages – boom, recession, depression and recovery, in their typically 50-year long lifetime. Since the 1700s, four Kondratiev cycles have been observed.

Kondratiev also indicated how technology plays a role in

leading the way out of the contraction stage in the cycle and into the next up-wave.

Dr Roth believes that even though some companies are working with CNT, not many are actually investing in it. "CNT interconnections and vias will come; CNTs transistors for toys and basic research will also come, but the risk is that [transistor] minaturisation will saturate for economic reasons before self-assembly methods for CNT will be ready," said Dr Roth.

"There is still a lot of work to be done to get high IC density with CNT. Research will take another ten years and we are already at 30 years in this [latest] 50-year cycle, so I'm not sure that CNT will be ready by then," he told *Electronics World*.

Carbon is a 'cousin' to silicon; they share the same group number in the periodic table, but with different atomic numbers (6 and 14, respectively). Since the

covalent bonds are a lot stronger in carbon than in metal, this allows electrical conductivity that is 15 times stronger than in copper, for example. The current density in carbon nanotubes is better too, as electrons do not collide with imperfections – loose links in the metal.

Field emitting transistors (FETs) can be made from single nanotubes, but the dimensions of such a device will be a lot larger than in silicon. This means that transistor density in ICs will be limited. Infineon and Samsung have created vertical CNT transistors to save on space, which will lead to better IC integration. However, Infineon has stopped any further investment in this field, for the time being.

"These devices [vertical transistors] exceed the ITRS roadmap for IC integration by 2016, but they require investment," added Dr Roth.



Software Design

- ▶ Begin every step in software design by asking, 'How should this information, this activity, this state, this relationship be represented?'
- ▶ Don't expect to get the representation right first time. Look for opportunities to improve code by re-representation
- ▶ Write no more than necessary
- ▶ Make the first version durable not functional and get it running early
- ▶ Localise information
- ▶ Test to discover information
- ▶ Design software to minimise the damage caused by changes to requirements
- ▶ Steer design by clear criteria
- ▶ Exploit existing solutions
- ▶ Shape your program around these six sufficient concepts: states, events, conditions, actions, objects and types

This month's Top Ten Tips were extracted from the book *'Software Design for Engineers and Scientists'* by John Robinson and published by Newnes, an imprint of Elsevier.

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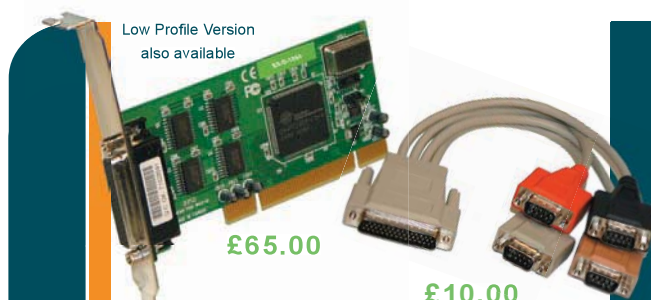
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Test to Benefit

New methodology for testing ASICs provides many benefits to designers says **Gregor Siwinski**

In recent years, hardware emulators have been used for testing ASIC designs at speeds that reach more than 1MHz.

These emulators typically use an array of high capacity FPGAs and allow the ASIC team to run their design at close to real-time speed to verify the design before it goes to production at the foundry.

Such solutions provide many benefits but are very expensive. If the ASIC design needs to be tested in several locations or by several third party companies who will use the final ASIC, then each one needs to receive their copy of the emulator with the prototype design.

Emulators are also inflexible with partitioning of the design. Since the predefined arrays of FPGAs have fixed structures, the design has to be partitioned to fit this structure. This is often one of the biggest problems since the modules cannot be fit into most optimal FPGA sizes and topologies.

Another problem encountered is re-targeting ASIC designs to compile into FPGAs. With different FPGA architectures, the re-targeting process can sometimes take several weeks and require changes in the design to accommodate large memories, multiple clock domains and other FPGA specific limitations.

As a result of these issues, many ASIC teams are adopting a new methodology that includes a custom FPGA prototype for the ASIC design, planned from the beginning of the ASIC project.

At the beginning of the ASIC planning process, a team assumes that all of the code compiles to both ASIC and FPGAs, and that the compilations are tested with each new version of the code. If incompatible, the team can use modified code versions and test the functional equivalence ahead of time. The custom proto-

type can be simulated, based on the FPGA libraries to ensure that the functional and timing results are the same.

When custom hard macros are used in the ASIC, the FPGA prototypes can either include an equivalent IP – core optimised for FPGAs or in some cases, a hard macro can be added to the PCB design in an ASIC form (e.g. ARM processor, memory and others).

Partitioning the design can also be planned based on a module basis. The design team can select the best FPGA to fit each module for optimum cost and performance. In addition to par-

.....
“The biggest difference between FPGA and ASIC design techniques is the use of clock signals”

tioning, the PCB layout for the prototype can be optimised for maximum performance, as well as customised for the required connectivity between FPGAs.

Ideally, the same synthesis tool is used for both ASIC and FPGA compilation, but if it's not possible, a separate FPGA synthesis program in the custom prototype can handle ASIC RTL code. Some ASIC design features can also be re-targeted to FPGAs automatically. The biggest difference between FPGA and ASIC design techniques is the use of clock signals. FPGAs use dedicated clock lines and can typically handle 16 to 32 clocks per device. In ASIC designs, a typical technique is to use multiple clock domains for better power usage. Most ASIC designs use 64 clocks or more. In such cases, automated software can be used to convert ASIC clock management to FPGA clocks.

In addition, high capacity memories cannot be easily accommodated in FPGAs. In such cases, an equivalent external memory can be used on the prototype board with an appropriate logic interface to provide a high capacity memory equivalent. For example, a DDR memory module can be used in the prototype board and configured through a logic wrapper to emulate SRAM or other types of memory.

When the ASIC design is finalised and ready for hardware testing, a prototype board can typically be made within a few days and can include the latest version of the design for high speed testing. If needed, multiple copies of the prototype board can be made for simultaneous testing in several locations.

Connections between FPGAs need to be optimised in tools that can handle high pin count BGA packages. With FPGA packages reaching over 1000 I/Os, the connectivity between FPGAs can be customised to best fit the PCB layout. If needed, the I/O assignment can be customised in FPGA software to assign the signals to the closest pin to connect to the next FPGA device. There are several commercial I/O management tools that will also optimise the FPGA ports to work at higher speeds.

Custom FPGA prototypes promise to offer many benefits to the ASIC designer and complement the traditional ASIC development process with high speed verification solutions at an affordable cost. If project managers plan the FPGA process in the early stage, the verification in hardware can be preformed in a very short time.

Gregor Siwinski is Director of R&D for Aldec



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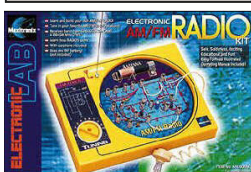


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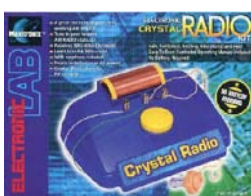
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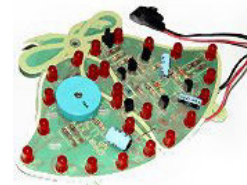


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Are You Tuning into the New Video Buzz?

The video surveillance market, military systems and medical imaging stand to benefit from a new generation of video encoders

By Nick Flaherty

Advanced Video Coding (AVC) is slowly taking over as the dominant standard for digital video, at least in the broadcast world. Also called H.264 and MPEG-4 part 10, encoding in AVC produces a good quality standard definition video stream at 1Mbit/s to 2Mbit/s, compared to 2Mbit/s to 3Mbit/s for today's MPEG-2 technology. Similarly, in high definition (HD), instead of a 19Mbit/s stream in MPEG-2, HD pictures can be broadcast at 10 to 12Mbit/s.

This is great for telephone companies looking to send video over telephone lines using the IP protocol and ADSL2, and for satellite and cable TV operators who want to squeeze more channels into their offerings without having to add new capacity.

As a result, AVC decoders are quickly reaching commodity. STMicroelectronics (ST), Broadcom and Conexant Systems are all shipping single-chip AVC decoders in set-top boxes in the first quarter of 2006, and the march of silicon technology and increasing integration of other devices on the board is bringing the costs down.

Sagem Communication in France is using ST's single-chip decoder, initially in SDTV boxes for Canal+ for receiving digital terrestrial TV, but also plans to launch HDTV boxes before the end of 2005, with volume production in 2006.

However, that doesn't mean the death of MPEG-2 decoders, where the market will continue to grow, says market researcher In-Stat. The worldwide MPEG video chip market revenues for 2004 were \$3.8bn, and MPEG-2 unit shipments are expected to experience growth until 2009, though decreasing prices will cause revenues to decline says the company.

"Buzz is really all there is to MPEG-4 AVC currently," said Michelle Abraham, analyst at In-Stat. "We expect the market will be small in 2005, as many decoder IC suppliers have been conducting interoperability testing with encoder suppliers. We

expect widespread availability of MPEG-4 AVC decoder ICs beginning in 2006."

However, the encoding side of AVC is not so easy. The broadcasters use racks of encoders from suppliers such as Tandberg Television and Harmonic, which use high-end FPGAs from Xilinx and their own ASICs to encode the bitstreams in real time.

Luckily, the latest silicon technology is allowing AVC encoding to fit into significantly cheaper devices.

At the broadcast end, a new company called Telairity has developed a single-chip encoder that will replace up to eight FPGAs in the encoder. "In the 1980s we had RISC and that was very exciting. I think video is the thing now," said Howard Sachs, Telairity's chief executive. Sachs is famous in the industry for developing the ideas behind Very Long Instruction Word (VLIW) technology while at Intergraph and was part of the development team for Intergraph's Clipper VLIW technology.

"The architecture we have is a five-processor architecture, which will give a high definition TV real-time encoder solution in

eight chips and one FPGA, which is the same performance at 24 of TI's chips," he said.

Telairity's T1P2000 single-chip encoder combines five independent vector/scalar cores, a video controller and a DRAM controller supporting an I/O bandwidth up to 5.3Gbit/s. Each vector/scalar core features four vector pipes with independent hardware, an independent scalar unit, 128kbytes of on-chip vector SRAM, a 4kbyte vector SRAM data cache, an 8kbyte scalar scratchpad memory and a 32kbyte instruction cache.

Sachs is talking to the main encoder makers such as Tandberg Television and Harmonic, but he is not just aiming at broadcast video. "We expect to expand our architecture and software expertise into other markets that are very allied to the current



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“In the 1980s we had RISC and that was very exciting. I think video is the thing now”
Howard Sachs, CEO, Telairity

one such as video surveillance, military systems and medical imaging," he said.

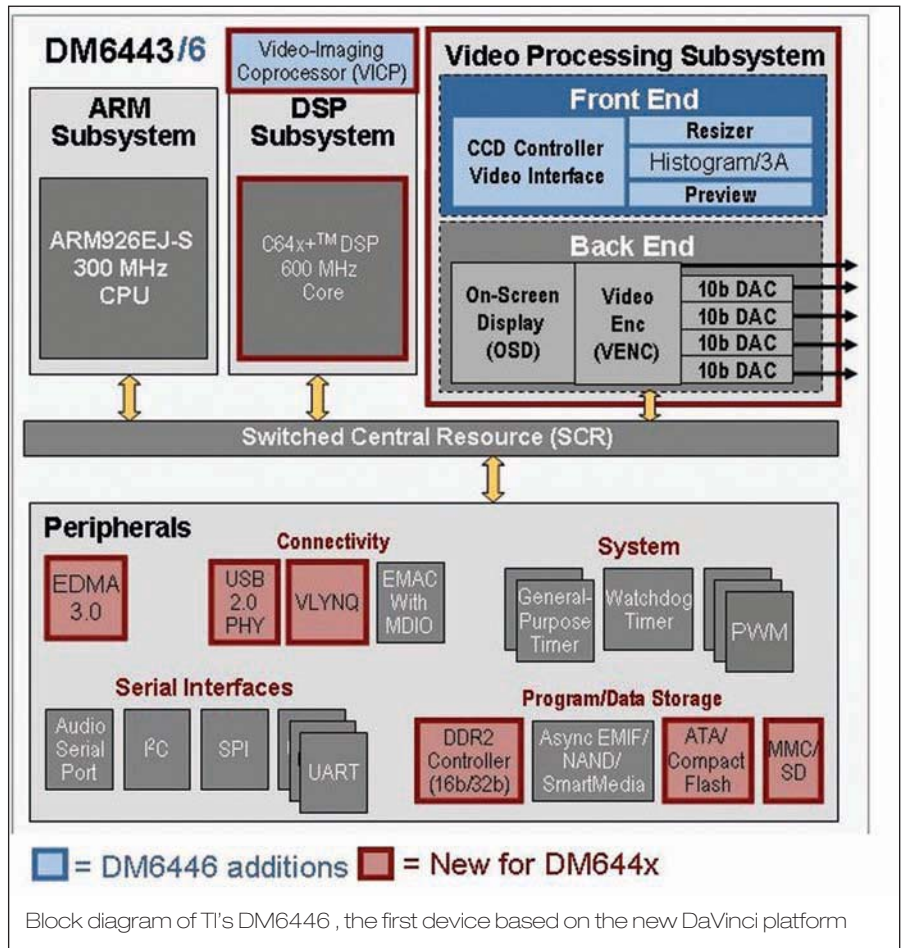
Korean chips designer Intime is also looking at this market, launching a single-chip MPEG-4 encoder and decoder that can handle up to four channels.

The IME6500 provides full standard definition TV resolution performance by using multiple programmable filters, which improve interlacing performance and image quality, whilst reducing noise distortion especially in dark or near dark conditions, which makes the particularly suitable for the surveillance market.

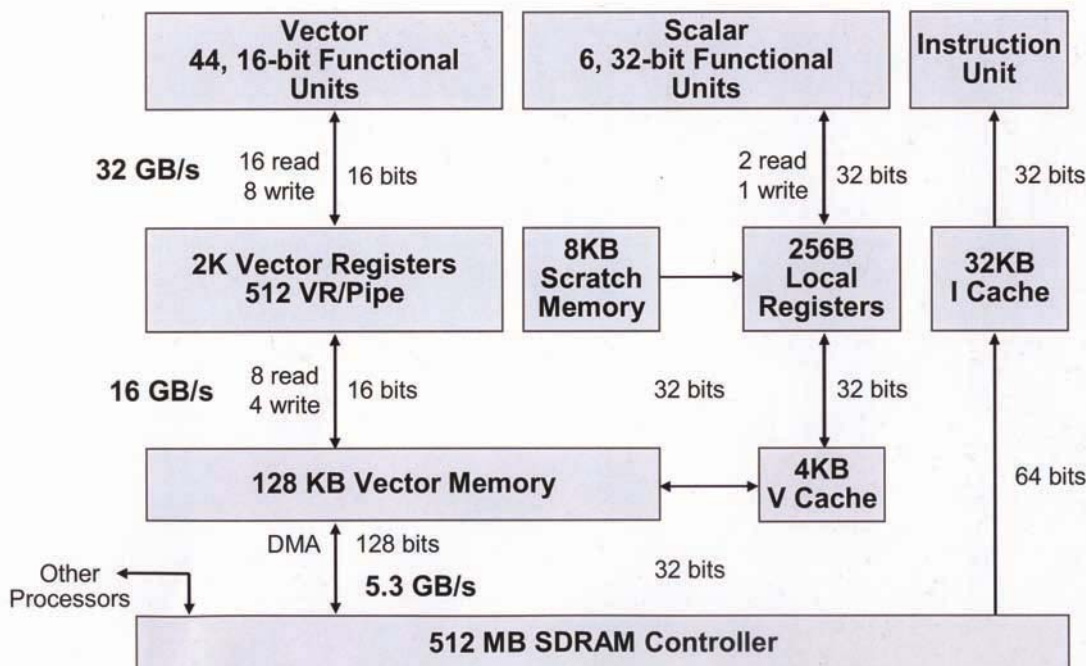
With a PCB footprint of only 17 x 17mm combined with an active power consumption of 300mW, the IME6500 is aimed not only at consumer personal video recorders, but also security, machine vision, medical equipment and CCTV.

The chip is also supported by high-level design software that enables engineers without specific knowledge of DSP or codec applications to implement their design ideas quickly with the IME6500, and there are two reference designs: a 4-channel PCI interface module and a standalone 4-channel digital video recorder.

Meanwhile, a Toronto-based start-up, ViXS Systems, has also developed an MPEG-4 encoder to distribute video between display devices. It is designed for integration in next-generation TVs, PCs, PCTV tuner cards, DVD-



Block diagram of Telairity-1



R machines, set-top boxes and personal video recorders (PVRs). The company is already supplying consumer electronic manufacturers such as Sony, Philips, Funai and Daewoo. Its XCode II encoder is based around a MIPS processor core with hardware accelerators.

Texas Instruments (TI) also sees encoding capability as key. Its DaVinci platform combines an ARM core with the latest digital signal processing core in a wide range of applications, from portable video through surveillance systems to standard definition TV over phone lines and even high definition TV.

The first two part numbers stemming from the DaVinci platform are video devices aimed at IPTV and video surveillance applications, and general-purpose video. "TI wants to accelerate the development of the video market and we want the customers to add their value and not reinvent the wheel each time," said Jean-Marc Charpentier, European DSP business development manager at TI.

The DM6443 and DM6446 both include a 300MHz ARM 926EJ processor core, running MontaVista Linux, alongside a 600MHz C64x DSP core, handling a wide range of software video codecs that TI supplies.

The \$30 DM6443 is aimed at IPTV applications, mainly for standard definition TV but handling up to 720p HDTV for certain

codecs, and includes interfaces to DDR memory and ATA hard disk drives. It consolidates seven other chips, including FPGAs, in a system design, saving at least \$70 of system cost, says Charpentier. But the \$35 DM6446 adds a video encoder for applications such as video surveillance, and the ARM core running Linux makes it easier to use without having to program the DSP directly. "We added the ARM9 because we want to add the Linux community that don't need to go into the details of the DSP," said Charpentier. "The main effort has been done on the software environment."

The key, he says, is having a complete software infrastructure, from low-level operating system drivers to application APIs, making it possible for developers to implement digital video without having to focus resources on writing and optimising codecs or programming a DSP. TI has developed Linux APIs that mask the complex hardware and software details of implementing codecs from developers, enabling them to interchange multimedia codecs without having to modify application code.

AVC is definitely taking off in 2006, and the popularity of AVC in broadcast is driving down the cost of the decoder chips. This, in turn, is driving more cost-effective encoders that can open up video applications in more traditional embedded markets, such as video surveillance and medical imaging.



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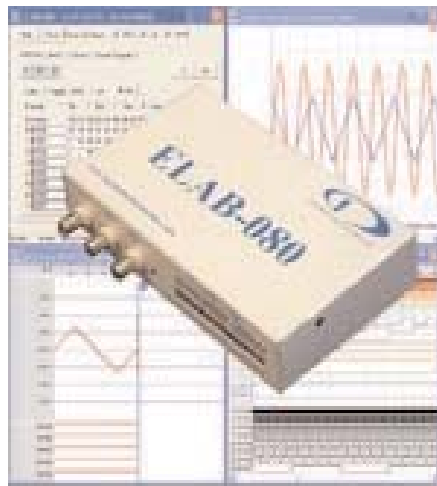
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Circuit Modelling for EMC

Ian Darney identifies the analytical tools involved in deriving values for the L, C and R components from physical data

Back in the January 2005 issue of *Electronics World*, the article "Understanding EMC" on p14 described how to build a couple of transformers, calibrate them and use them for bench tests of interference coupling. One was an injection transformer, used to induce a voltage into a system cable. The other was a current transformer, used to monitor current in the cable. The combination can be used to characterise the configuration under review, carry out conducted susceptibility tests, or perform conducted radiation tests.

It was shown that the results of bench tests can be used to create a circuit model of the equipment under review. Successive approximations to the component values enabled the response of the model to correlate precisely with the measured data. It was shown that a single model can be used to predict conducted emissions and conducted susceptibility.

However, its applicability was restricted to low frequency coupling, where inductance and resistance have a predominant effect. It is a simple matter to build transformers that operate at a much higher frequency than those described, but there is little point in using them if the measured response cannot be simulated.

It is first necessary to establish a systematic method of deriving a circuit model of the configuration under review. The purpose of this article is to identify the analytical tools involved in deriving values for the L, C and R components from physical data. That is, to enable a model to be created before the assembly is built.

Three is a crowd

In general terms, it can be said that only one conductor is needed to transmit electromagnetic fields far and wide, it takes two conductors to transmit a signal efficiently from one location to another, but at least three conductors are involved in the creation of interference.

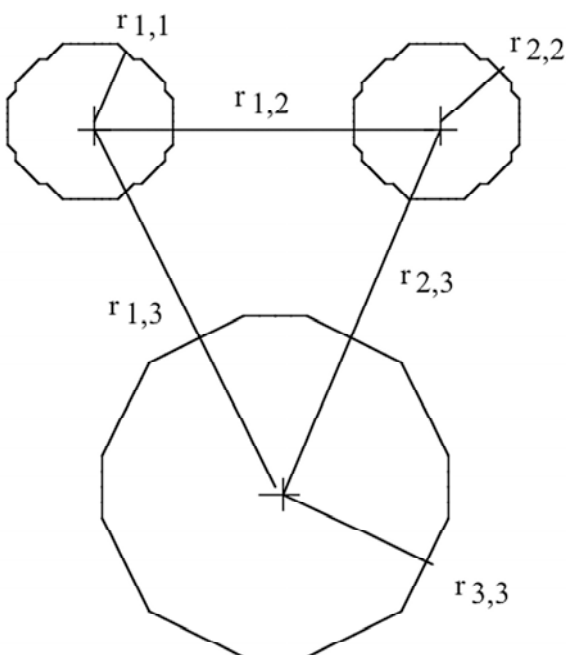
The analysis needs to be able to cater for the interactions between three conductors.

A simple configuration could be a two-conductor cable routed along a third conductor. This third conductor would represent the combined effects of structure and other signal conductors. Such a representation will not provide a perfect simulation, but it is infinitely better than one which assumes that all return conductors form an equipotential surface.

Figure 1 illustrates one way of defining the cross-section of such an assembly. It is assumed that the cross-section is constant and that the length is l . This lends itself to the use of an array of numbers to record all the radial measurements.

$$\mathbf{r} = \begin{pmatrix} r_{1,1} & r_{1,2} & r_{1,3} \\ r_{2,1} & r_{2,2} & r_{2,3} \\ r_{3,1} & r_{3,2} & r_{3,3} \end{pmatrix} \quad (1a)$$

Figure 1: Cross-section of three-conductor assembly



$$r_{i,j} = r_{j,i} \quad (1b)$$

Integers i and j are used to identify the conductors. If they are different, the measurement defines the separation between the axes. If they are the same, the measurement defines the conductor radius.

Calculating the resistance of each conductor is a relatively straightforward task, using:

$$R_i = \frac{\rho \cdot l}{\pi \cdot r_{i,i}^2} \quad (2)$$

where ρ is the conductivity of the conductor. Skin effect is likely to cause the resistance of the conductors to increase with frequency. However, there is no need to invoke this complexity, since inductive and capacitive effects predominate.

Inductance is a parameter used to relate the rate of change of current in the conductor to the voltage along the conductor. If it is assumed that the rate of change of current in an isolated conductor is constant along the length l , then a primitive inductance can be defined as:

$$Lp_{i,i} = \frac{\mu \cdot l}{2 \cdot \pi} \cdot \ln \left(\frac{l}{r_{i,i}} \right) \quad (3a)$$

This parameter can be used to relate the energy level of the conductor to the energy contained in the surrounding magnetic field. Similarly, the energy level in conductor i due to field created by current in conductor j is a function of the inductance:

$$Lp_{i,j} = \frac{\mu \cdot l}{2 \cdot \pi} \cdot \ln \left(\frac{l}{r_{i,j}} \right) \quad (3b)$$

These formulae are a little unusual, in that they contain the length parameter on both sides of the logarithmic term. They highlight the fact that the length of the assembly is of critical importance.

One characteristic of primitive inductors is that $Lp_{i,j}$ is equal to $Lp_{j,i}$, since the radial separations are the same.

An assembly of two parallel conductors is illustrated in **Figure 2**. The voltages Vp_1 and Vp_2 are the absolute voltages along the conductors, since they relate the total energy to the rate of change of current. Since they are associated with primitive inductors, they can be described as primitive voltages. This keeps the terminology systematic. It is assumed that the currents in the two conductors are Ip_1 and Ip_2 , and that the positive direction is from left to right.

If it is assumed that the voltages are sinusoidal functions of time and that the angular frequency is

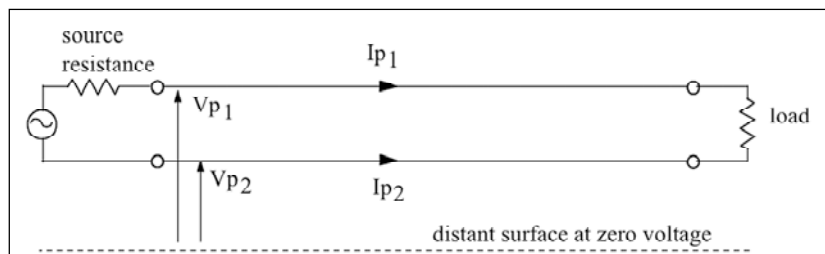


Figure 2: Primitive currents and voltages for a twin conductor

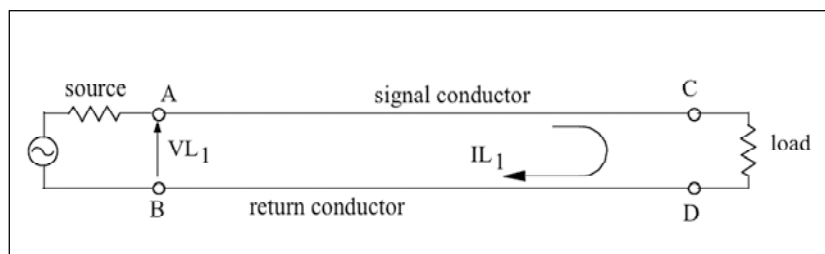


Figure 3: Loop currents and voltages for a twin conductor

ω , then the primitive equations for two parallel conductors are:

$$Vp_1 = j \cdot \omega \cdot (Lp_{1,1} \cdot Ip_1 + Lp_{1,2} \cdot Ip_2) \quad (4a)$$

$$Vp_2 = j \cdot \omega \cdot (Lp_{2,1} \cdot Ip_1 + Lp_{2,2} \cdot Ip_2) \quad (4b)$$

Voltages developed at source and load can be ignored at this stage of the analysis.

A more familiar picture of the configuration is shown in **Figure 3**, which defines loop voltages and currents. The relationship between primitive parameters and loop parameters can be defined by comparing Figures 2 and 3.

$$IL_1 = Ip_1 = -Ip_2 \quad (5a)$$

$$VL_1 = Vp_1 - Vp_2 \quad (5b)$$

Substituting primitive currents with loop current and subtracting Equation 4b from 4a leads to the loop equation:

$$VL_1 = j\omega LL_{1,1} \cdot IL_1 \quad (6a)$$

where

$$LL_{1,1} = Lp_{1,1} - Lp_{1,2} - Lp_{2,1} + Lp_{2,2} \quad (6b)$$

Using Equations 3a and 3b to substitute for the primitive inductors relates the loop inductance to the physical parameters:

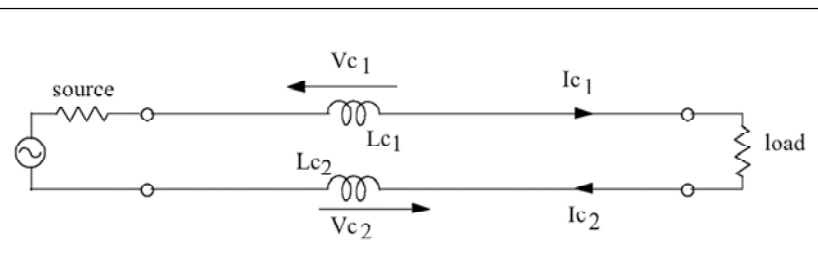


Figure 4: Circuit currents and voltages for a twin conductor

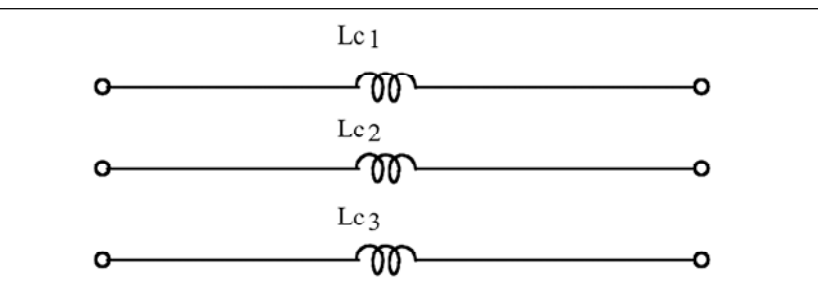


Figure 5: Inductive components of a three-conductor line

$$LL_{1,1} = \frac{\mu \cdot l}{2 \cdot \pi} \cdot \ln \left(\frac{r_{1,2} \cdot r_{1,2}}{r_{1,1} \cdot r_{2,2}} \right) \quad (7)$$

This is the standard equation for the inductance between two parallel wires, in a slightly unusual form. Some important characteristics of this relationship can be highlighted.

First and foremost is the assumption that the current in the return conductor is equal in magnitude to that in the signal conductor. This assumption is not necessarily true. In fact, the more one thinks about it, the less plausible it becomes.

The second point is the fact that two different types of inductance exist. Both types have the dimension of 'henry'. Primitive inductors define the characteristics of the conductor when it is acting as an antenna. Loop inductance is a parameter that can be measured with electronic test equipment.

The third feature of this derivation is of critical importance when EMC of the system is analysed. Loop inductance can only be measured between adjacent terminals. In the case of **Figure 3**, the inductance of the wire pair could be measured by connecting an LCR bridge to terminals A and B, and shorting terminals C and D, or vice-versa. However, when loop inductance is used in a circuit model, vital information is lost. Voltages between A and C, or between B and D, become unknown quantities. Yet, it is precisely these voltages that determine the EMC of the configuration.

Retrieving information

Fortunately, there is an easy way to recover this information. If it is assumed that the inductance of the signal conductor between A and C is L_{C1} , and that between B and D is L_{C2} , then the simple circuit model of **Figure 4** emerges. Here, the circuit voltages are:

$$V_{C1} = j \cdot \omega \cdot L_{C1} \cdot I_{C1} \quad (9a)$$

$$V_{C2} = j \cdot \omega \cdot L_{C2} \cdot I_{C2} \quad (9b)$$

where

$$L_{C1} = L_{p_{1,1}} - L_{p_{1,2}} = \frac{\mu \cdot l}{2 \cdot \pi} \cdot \ln \left(\frac{r_{1,2}}{r_{1,1}} \right) \quad (9a)$$

$$L_{C2} = L_{p_{2,2}} - L_{p_{2,1}} = \frac{\mu \cdot l}{2 \cdot \pi} \cdot \ln \left(\frac{r_{1,2}}{r_{2,2}} \right) \quad (9b)$$

This introduces a third type of inductance – the circuit inductance. Parameters of type 'circuit' are the type that can be used in circuit models. Comparing Equations 9a and 9b with Equation 7 shows that, in this simple configuration, the sum of the circuit inductors is equal to the loop inductance. The process can be summarised:

- > define the primitive parameters
- > set up the primitive equations
- > set up the loop equations
- > define the loop parameters in terms of the primitives
- > set up the circuit equations
- > define the circuit parameters in terms of the primitive parameters.

Applying this process to three parallel conductors results in the circuit model of **Figure 5**. If the radial parameters are as defined in Figure 1, then:

$$L_{C1} = \frac{\mu \cdot l}{2 \cdot \pi} \cdot \ln \left(\frac{r_{1,2} \cdot r_{1,3}}{r_{1,1} \cdot r_{2,3}} \right) \quad (10a)$$

$$L_{C2} = \frac{\mu \cdot l}{2 \cdot \pi} \cdot \ln \left(\frac{r_{1,2} \cdot r_{2,3}}{r_{2,2} \cdot r_{1,3}} \right) \quad (10b)$$

$$L_{C3} = \frac{\mu \cdot l}{2 \cdot \pi} \cdot \ln \left(\frac{r_{1,3} \cdot r_{2,3}}{r_{3,3} \cdot r_{1,2}} \right) \quad (10c)$$

These equations will be familiar to anyone who has dealt with three-conductor power lines.

There exists a capacitive component that is complementary to the primitive inductance:

$$Cp_{i,j} = \frac{2 \cdot \pi \cdot \epsilon \cdot l}{\ln \left(\frac{l}{r_{i,j}} \right)} \quad (11)$$

This capacitance can be used to relate the energy stored in the electric field to the absolute voltage on the conductor. As such, it can be used with the primitive inductance to analyse the performance of antennae. But, it is best to take one step at a time.

Since the logarithmic terms in Equations 11, 3a and 3b are identical, it becomes possible to derive capacitance values directly from inductance values, using:

$$Cp_{i,j} = \frac{\mu \cdot \epsilon \cdot l^2}{Lp_{i,j}} \quad (12)$$

Since this relationship exists with the primitive types, then it remains valid for loop components and circuit components.

Since each conductor possesses the properties of resistance, inductance and capacitance simultaneously, then any circuit model must include all three components, if it is intended to be at all representative. This leads to **Figure 6**, where the inductive and resistive parameters have each been split into two equal components.

Avoid temptation

It is tempting to apply a star-to-delta transformation to obtain values for interconductor capacitance, but this temptation should be avoided. Visibility of the zero-volt node at the junction of the capacitors will be lost. Such visibility is essential if the final, and probably the most important, step in the modelling process is to be achieved.

Electromagnetic theory indicates that currents and voltages vary continuously along a transmission line, and the effect is portrayed as the propagation of an electromagnetic wave in the space bounded by the two conductors. It is equally valid to visualise the propagation of energy as a current that travels along the surface of the conductors at near-light velocity. If this is done, then the three conductors can be treated as three separate transmission lines.

This leads to the model of **Figure 7**, where:

$$ZA_i = Zo_i \cdot \tanh \left(\frac{\theta_i}{2} \right) \quad (13a)$$

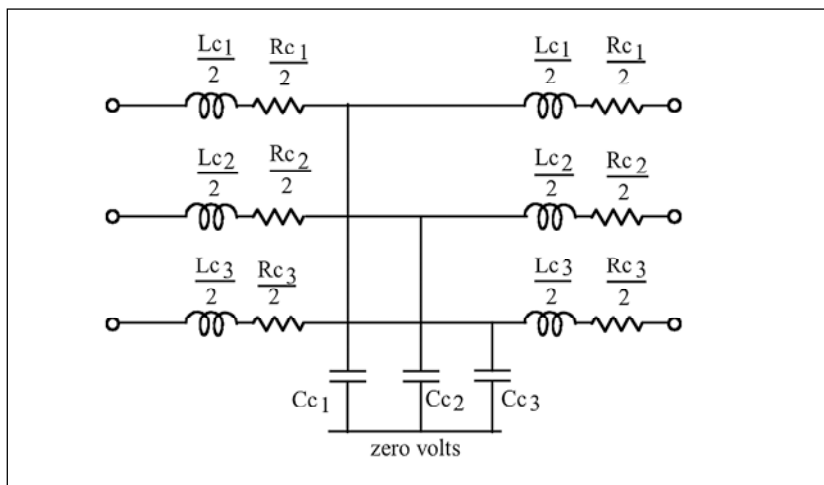


Figure 6: Circuit model of a three-conductor line

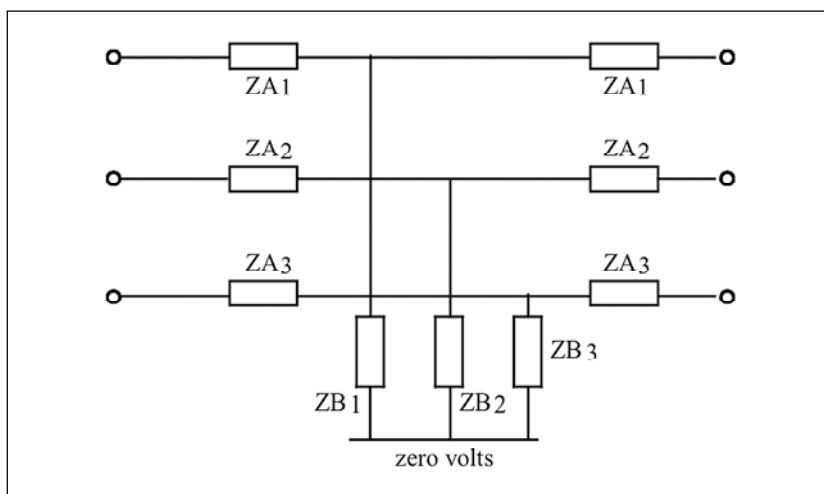


Figure 7: Distributed-parameter model of a three-conductor line

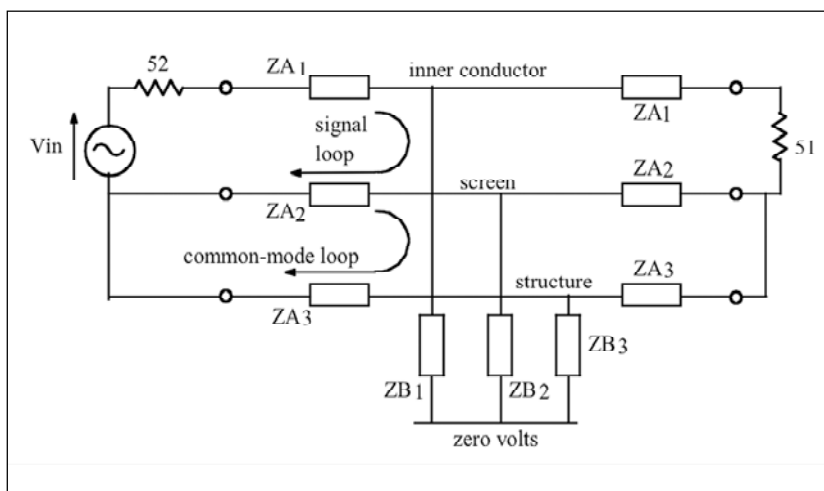


Figure 8: Model of an actual setup

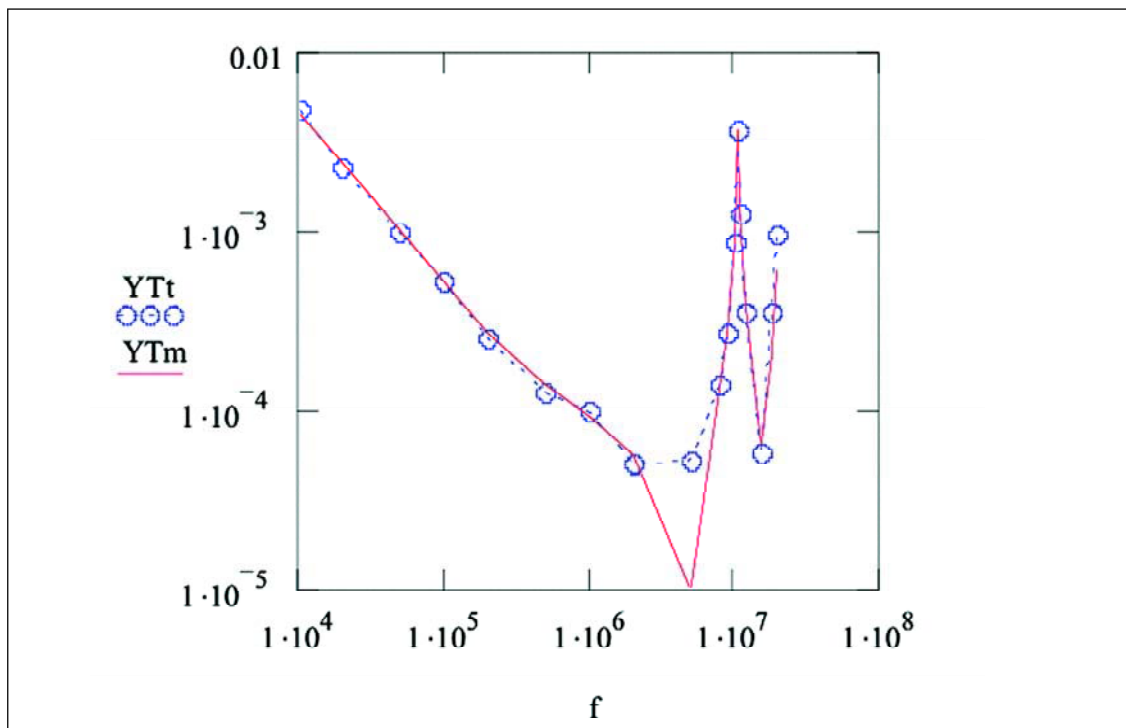


Figure 9: Correlation of circuit model with test results

$$ZB_i = \frac{Z_{o_i}}{\sinh \theta_i} \quad (13b)$$

given that:

$$Z_{o_i} = \sqrt{\frac{Rc_i + j \cdot \omega \cdot Lc_i}{j \cdot \omega \cdot Cc_i}} \quad (14a)$$

and

$$\theta_i = \sqrt{(Rc_i + j \cdot \omega \cdot Lc_i) \cdot (j \cdot \omega \cdot Cc_i)} \quad (14b)$$

Since the ZA and ZB impedances are directly related to the circuit parameters Lc , Cc and Rc , and since these components can be related to physical parameters, it is possible to compute the impedance values at any particular frequency.

If the conductors are configured as a pair of wires to carry a signal, with a third conductor representing the structure, then the frequency response of the coupling between signal loop and common-mode loop can be simulated. This can be done over any range of frequencies where transmission line analysis is valid. **Figure 8** shows the model of the setup.

At this point, the reader might well be sceptical. However, the proof of the pudding is in the eating. When the tests and analysis described in the article “*Understanding EMC*” were extended to cover a higher range of frequencies, the results were as displayed in **Figure 9**. The parameter YT in the graph is the term used for transfer admittance, in this case the ratio of common-mode current to input voltage V_{in} , in amperes per volt. The 18 blue circles identify the data points derived from test results.

It is possible to create an accurate circuit model of an actual wiring assembly at frequencies well beyond first resonance.

Using this process, anyone can predict the results of conducted susceptibility and conducted emission tests. Further development of the approach allows the prediction of radiated susceptibility and radiated emission.

This article's setups, tests and derivation of equations can be found on

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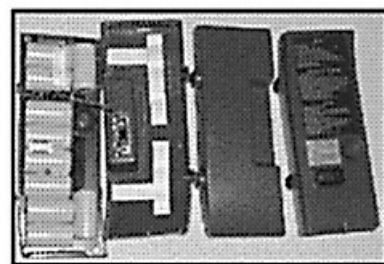
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Suppression Techniques of Over Voltage Transients at Motor Terminals

Chaiwat Choochuan of the Electromagnetic Compatibility and Electrical Machine Drives Systems Laboratory, Shaffner EMC (Thailand), explains the suppression techniques of over voltage transients at motor terminals and conducted EMI in PWM drives systems

The suppression of over voltage transients at motor terminals can be divided into the load reactor and sine-wave filter. One economical solution is the suppression of voltage transients by using a load reactor.

The sine-wave filter is available to improve the inverter output waveform from a PWM type to a near sinusoidal waveform. Using an input EMI filter for the suppression of conducted electromagnetic interference (EMI) is also very effective. The experimental results show the effectiveness of these techniques for over voltage transients and conducted EMI in PWM drives systems.

Adjustable speed drives

Technology to control the speed of AC motors continues to develop at a fast pace. Adjustable-speed drives (ASDs) are a significant part of this development. Energy savings, less wear on the mechanical components and a better process control are the primary factors that push such rapid advancement. The evolution of high-speed switching devices, such as the IGBT has enabled us to increase the carrier frequency of voltage-source PWM inverter.

Since IGBT devices can switch at high carrier frequencies (up to 20kHz), there are several advantages:

- (1) More low speed torque for applications where nominal motor torque is required close to zero speed;
- (2) Quieter motor operation, which reduces audible noise;
- (3) Improved low speed stability, which minimises low speed oscillations.

(3) Improved low speed stability, which minimises low speed oscillations.

Unfortunately, the fast switching IGBT technology also generates high frequency common-mode voltage, which induces high shaft voltages, leads to bearing currents and a very fast voltage rise (dv/dt) in PWM-wave-caused EMI problems, and over voltage at motor terminals when using long cables.

These adverse effects need to be suppressed by some techniques. There are several solutions available to solve this problem, each offering a different degree of protection at a different price. In this paper, the conducted EMI of the PWM inverter, over voltage transients and its mitigation effort are introduced.

Figure 2 shows the simulation waveforms of inverter currents and voltage waveforms for cable lengths of 10m and 50m respectively. It was observed that the transient current spike may be higher than the fundamental current and with a cable length of 50m the motor voltage peak is at 1100V.

Suppression techniques

> Over voltage transient at motor terminals

Figure 1 shows the experimental results of output voltage inverter and motor terminal with long motor cables, high switching frequency and no filter installed. It can be seen that the over voltage (ringing) at motor terminals at each switching instant due to the presence of long cable. Factors affecting motor over-voltage are:

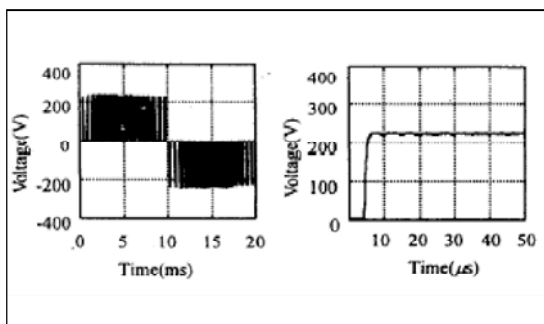
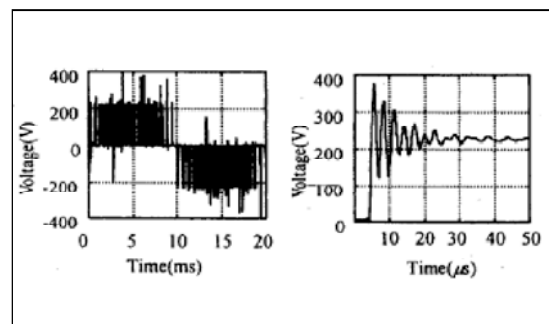


Figure 1: Typical voltage waveforms for an ASD with a long feed – left (a) inverter output voltage right – (b) motor terminal voltage



1. Motor and cable surge impedance
2. Motor load and load reflection coefficient
3. Cable length
4. Magnitude of drives pulse
5. Rise time of drives pulse
6. PWM switching frequency of inverter
7. Modulation strategy (SPWM or Space Vector PWM) and modulation index.

Motor and cable surge impedance mismatch are primarily responsible for the magnitude of peak per unit over-voltage. It has been observed that the amplitude of motor terminal over voltage can exceed 2-3 times the dc-bus voltage due to insufficient dwell time and polarity reversals (Figure 1). The simple output reactor reduces the inverter output, which in turn reduces the over voltage transient of the motor terminal. **Figure 3** shows the proposed output reactor and power circuit.

The design considerations of the simple output reactor for over-voltage suppression are as follows: the effect of core saturation and the flux distribution in the high frequency range. If the core saturates, then performance of the output reactor and the reactor L and R values are changed. Then, the over-voltage suppression of the output reactor is also changed. The values of L and R depend on the flux linkage and the core loss inside the reactor core. In general, the simple output reactor is designed at fundamental frequency, having 1%-5% impedance and the over-voltage suppression is almost proportional to the reactor impedance. However, a higher value of impedance causes a rapid increase in cost and weight.

A higher value of impedance at the fundamental frequency causes a larger voltage drop across the reactor, which reduces the fundamental component of voltage at motor the terminals. A large voltage drop, especially at over load condition, may result in the insufficient torque due to the lack of supply voltage.

One of the methods for overvoltage suppression is the sine-wave filter. The sine-wave filter is an effective means of taming the output voltage waveform and protecting the motor. **Figure 4** shows the sine-wave filter with passive components L,C and R

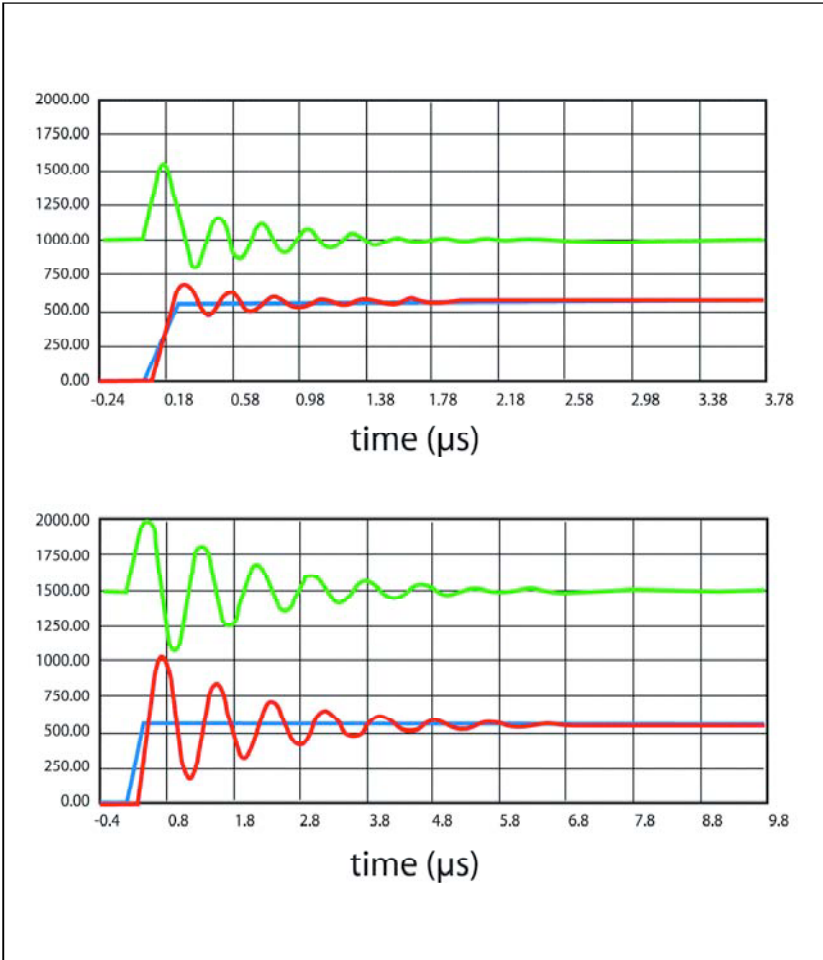


Figure 2: Typical inverter current and voltage waveforms; inverter current (green), inverter voltage (blue), voltage at cable end (red).
Top – 10m
Bottom – 50m

for the inverter output terminals. The structure of RLC filter in 3-phase inverter is shown in Figure 4. The impedance of the inductor of the filter is usually designed as a fraction of the impedance of the motor in fundamental frequency, say less than 5%, thus the filtering characteristic of the filter can be written as (1): (Note: The load is considered as open)

$$H = \frac{j\omega \frac{R}{L} + \frac{1}{LC}}{(j\omega)^2 + j\omega \frac{R}{L} + \frac{1}{LC}} \quad (1)$$

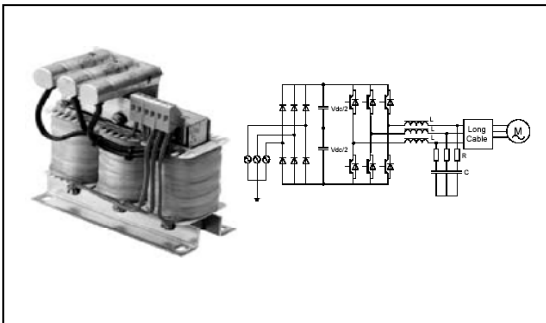
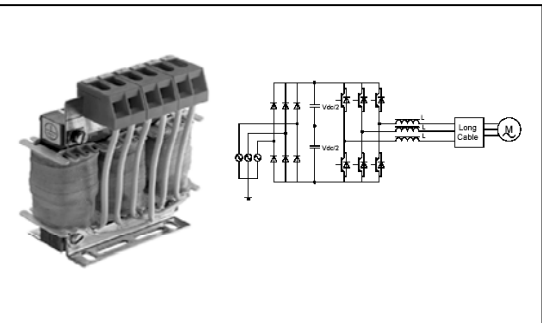


Figure 3 (Left): Photo of proposed output reactor; RWK305-24-KL and power circuit

Figure 4 (Right): Photo of proposed sine-wave filter; FN5010-24-99 and power circuit

$$H = \frac{j\omega 2\xi\omega_o + \omega_o^2}{(j\omega)^2 + j\omega 2\xi\omega_o + \omega_o^2} \quad (2)$$

where

$$\omega_o = \frac{1}{\sqrt{LC}} \quad \text{:resonance frequency}$$

$$\xi = \frac{R}{2\sqrt{\frac{L}{C}}} \quad \text{:damping ratio}$$

When the damping ratio is greater 1, there is constant attenuation in the low frequency, so the damping ratio is usually chosen as smaller than 1. The sine-wave filter can filter the PWM wave to a near sinusoid with THD lower than 5%. In addition, the sine-wave filter has the advantage of high reliability

➤ Electromagnetic conducted emission

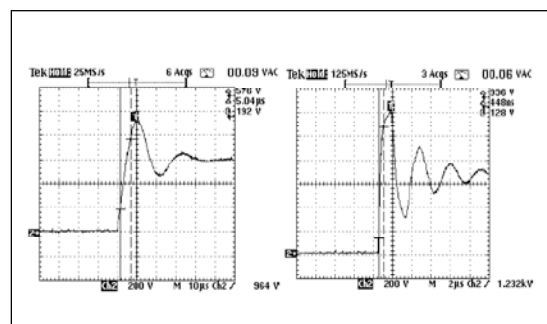
It is well known that conducted EMI noise is a major concern for inverter driven motor drives, especially for inverter drives using high speed switching frequencies. This improves the performance of PWM inverter-fed induction motor systems, such as those that the switching loss and increase the controllability of system. The higher frequency of leakage current by the common-mode voltage of the PWM inverter is the primary concern of the higher conducted and radiated EMI emission. This is because the conducted EMI noise may cause negative interference effects on electronic devices, such AM radio receivers and medical equipment. The power EMI filter is applied to solve these problems.

The EMI line filter is usually located between the AC input terminal and the diode rectifier, and is expected to provide the attenuation of common and difference-mode high frequency current components.

Figure 6 shows the standard three phase EMI filter schematic and photo of EMI filter.

Measurement results

The suggested mitigation techniques for over voltage phenomena include load reactor, sine-wave filter and EMI input filter. The test results are



obtained on a 5.5kW 400V IGBT drive and an induction motor. A 50m shielded cable between the inverter and induction motor was employed. A line-impedance stabilising network (LISN) is connected between the AC power source and the inverter system to provide a stable source impedance to the inverter systems in frequency range of 150kHz-30MHz.

Figure 7 shows the motor terminal voltage without and with the output reactor. **Figures 8a** and **b** show the motor current with and without output reactor connected respectively.

As can be seen, the charging current or transient current spike can cause inverter over-current faults or poor operation, due to inaccurate current feedback measurements such as flux vector or sensorless speed control.

Figure 5: Photo of proposed sine-wave filter with return to the DC link; FN530-16-99 and power circuit

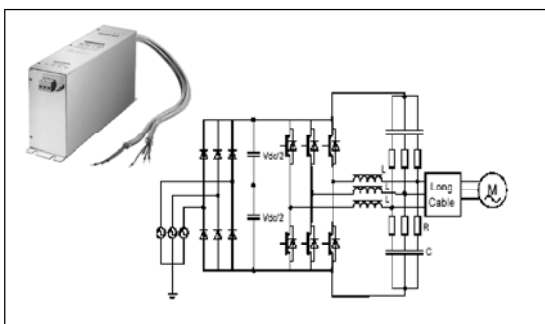


Figure 6: Standard three phase EMI filter schematic and photo of EMI filter; FN3258-55-52

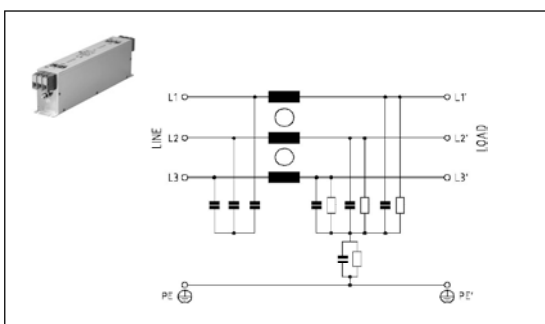


Figure 7 (Far right): The experimental results motor terminal voltage with long cable and high switching frequency (50m and carrier frequency 18kHz) Left – shows the motor terminal voltage without output reactor. Right – with proposed output reactor RWK305-24-KL

and large capacity, so it is widely used in applications where large power capacity is needed but sine-wave filters for this purpose cost approximately 2-3 times the cost of a simple output reactor.

Another sine-wave filter circuit proposes a return to the DC-link, with the purpose of connecting the LC filter star point to a point in the DC-link, as shown in **Figure 5**. The solution of this topology has an important advantage over the classical sine-wave filter: it reduces both common-mode (CM) and differential mode components (DM).

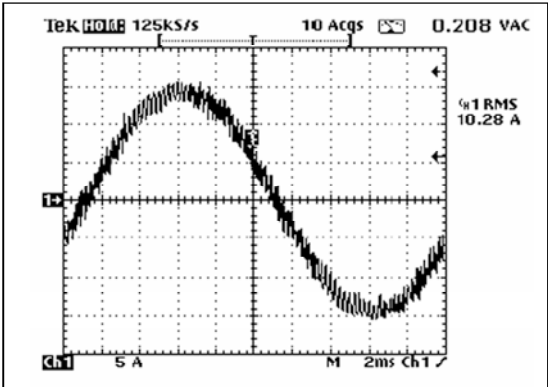
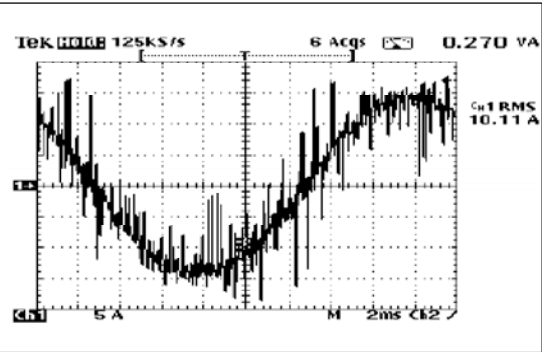


Figure 8: The experimental results of motor current with long cable a) without output reactor (b) with proposed output reactor: RWK305-24-KL

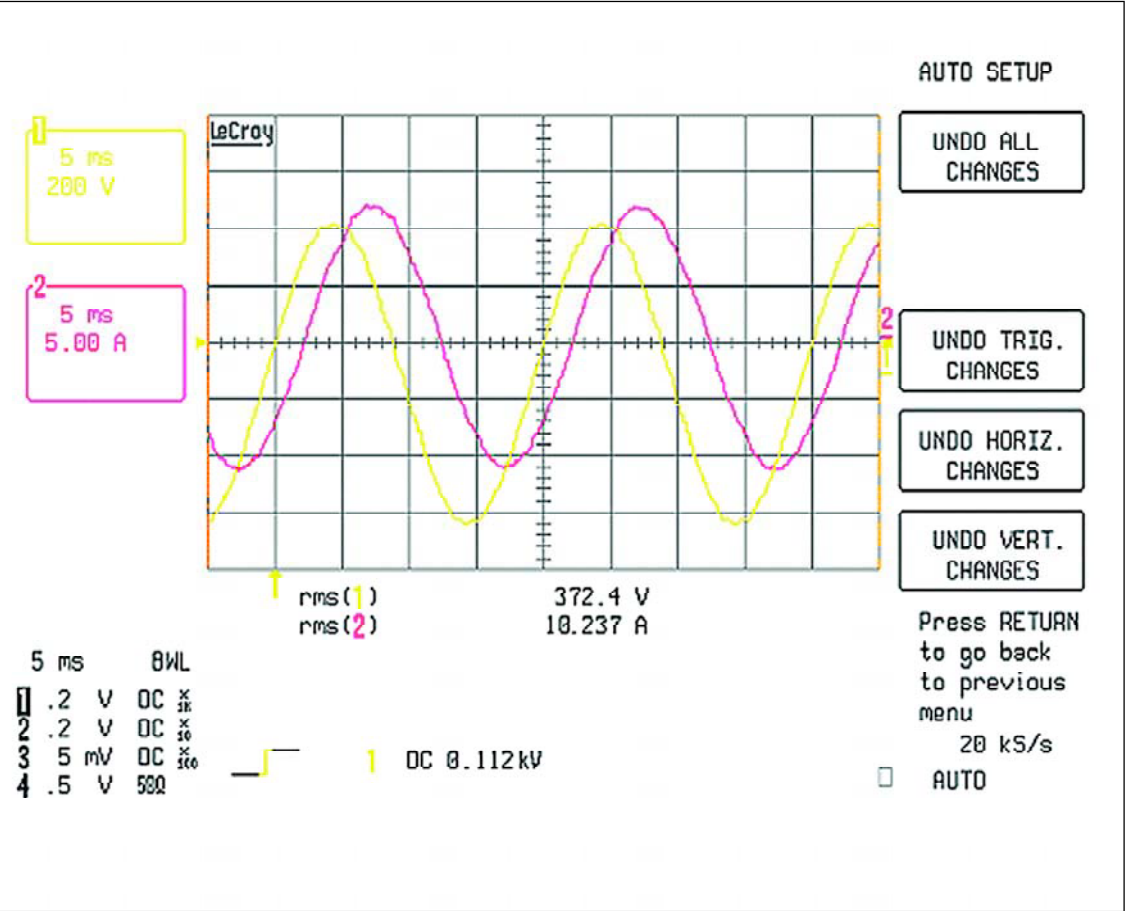


Figure 9: Waveform at the motor terminal line-to-line voltage and current

Figure 9 shows the experimental results of motor terminal line-to-line voltage and current. It is shown that the sine-wave filter can eliminate the peak voltage at the motor terminals. Note that the application and design of inverter output filter is dependent on the cable length, and the losses of filter are dependent on the motor drives system kilovoltamperes.

Conclusion

In this paper, some application issues for PWM drives have been investigated. The effect of high rate

of voltage rise in inverter output (dv/dt) generates the excessive over voltage in the motor terminals and conducted EMI of the PWM inverter have been addressed and explained. Suppression techniques have been shown to significantly reduce the transient over voltages at motor terminals and reduce the common-mode EMI using passive filter. The reduction contributes to improved power quality and reliability of PWM induction motor drives systems.

The Ground Plane: Lord of the Board

Tim Williams, a consultant with Elmac Services and author of *'EMC for Product Designers'* emphasises the importance of giving due care to the ground plane in all designs

The way that a circuit is laid out on a PC board is critical to its EMC performance. At high frequency all tracks must be regarded as components in their own right, since the coupling of high frequency signals from one part of the circuit to another is mediated by their stray capacitance and inductance. The schematic rarely – if ever – tells this side of the story; PC tracks are nothing more than lines on the diagram joining pins of lumped components such as ICs, resistors and capacitors.

Every track on a board exhibits its own partial self-inductance, and mutual inductance and capacitance to other nearby areas of copper. If these are uncontrolled then the performance of the board at high frequency – and that means the EMC performance of the circuit – is also uncontrolled. The simplest and most effective way of fixing the characteristics of all such tracks is to implement a ground plane on the board.

The workings of the ground plane

First of all, forget any notion of the ground plane being used for screening: it may do, but that is not its primary function. What you expect from it is to provide a high-frequency return path for every signal and power rail on the board.

Running a signal and its ground return close together – so that neighbouring currents flow in opposite directions – reduces the inductive impedance of the ground path by a factor equivalent to the mutual inductance between the two:

Effective inductance of signal loop = $2(L - M)$
where L is the self-inductance of each half of the loop, M is the mutual inductance between the halves.

This effect is critical to understanding PCB track routing at radio frequencies: you need to control the 0V current return path. Keeping the signal and 0V paths adjacent along their length – and therefore

maximising their mutual inductance – will ensure not only minimum coupling with the magnetic fields around the PCB, but also minimum impedance of the return path and, therefore, minimum noise voltage developed along it. This will keep down the “noisiness” of the whole board, which is mostly responsible for common mode emissions. It will also improve its immunity to external interference, which is determined by the conversion of incoming common mode disturbances to differential mode within the circuit.

This is entirely possible to do with assigning a 0V return track to each signal and running the two next to each other, or on adjacent layers, throughout the board. This is, of course, somewhat tiresome and awkward in a board with hundreds of such pairs and it isn't necessary. If you provide a 0V (ground) plane as a continuous layer adjacent to each track layer, then the job is done for you. The return current for each high frequency signal track automatically finds its own preferential path through the plane, which will be that path that is directly underneath the signal. The geometry ensures that this path has the least enclosed loop area and, therefore, the least total path inductance. All you have to do is make sure that the component 0V connections, which carry the return path through, have a short and direct connection to the 0V plane – which is easy, and that the 0V plane remains unbroken along the length of any of these paths, which, despite also being easy on a multi-layer board, still occasionally suffers from poor implementation.

Ground planes in the xy direction

The effect of a slot in the direction of the current flow is to increase the 0V return path inductance. There are two problems caused by extra inductance here:

- Any high frequency current passing through this inductance will develop a voltage across it. This

voltage then appears across the two parts of the plane, each of which has a high capacitance to the environment due to its surface area. Even on its own, this structure forms an efficient radiating dipole, which is only enhanced when cables are attached to the board;

- The inductance itself represents increased magnetic coupling to the environment through the enclosed loop area.

Therefore, either there should be no split within the plane to add this extra inductance, or high frequency currents should be prevented from flowing between the two separated segments. This can be achieved through careful circuit design for emissions control, but it doesn't address the problem of immunity to incoming RF fields. The magnetic and electric coupling is still present for these fields and so they will develop voltages between the two ground segments, which are likely to create circuit susceptibilities.

There is a convention, encouraged by many device application notes, to create different 0V planes for different parts of the circuit. For instance, analogue 0V and digital 0V, which are then linked at one point on the board, typically at an A-D converter. The purpose of this trick is to prevent digital noise currents from flowing in the analogue circuit and corrupting it, that is, it relates to internal EMC. Unfortunately, it creates problems for external EMC and introduces difficult questions that could be avoided by a single system-wide 0V plane:

- Where should you join the planes when there are multiple analogue-digital interfaces?
- What about power returns that are common to both analogue and digital?
- Where do you run signals that must cross the break between the analogue and digital sections? Any of these signals will be exposed to incoming interference that will appear between the planes, or will develop such interference themselves.
- If you are going to bond the 0V planes at RF to the chassis of the product, which is always a recommended practice to minimise common mode coupling into or out of the PCB, how do you choose which of the planes should be bonded?

For all these reasons, a preferable way to proceed is to use a single, system-wide 0V plane and carefully lay out the board so that digital noise is effectively segregated from low-level analogue circuits. Avoid as far as possible any discontinuities in this plane and make sure that you do not run any critical (noisy or low-level) tracks across any discontinuities that remain, such as slots formed by close-packed vias, or near to the edges of the plane. There is no merit in putting "moats" in this 0V plane.

In the centre of a plane inductive effects are minimal, but towards the edges fringing magnetic fields are created and the inductive impedance of the plane

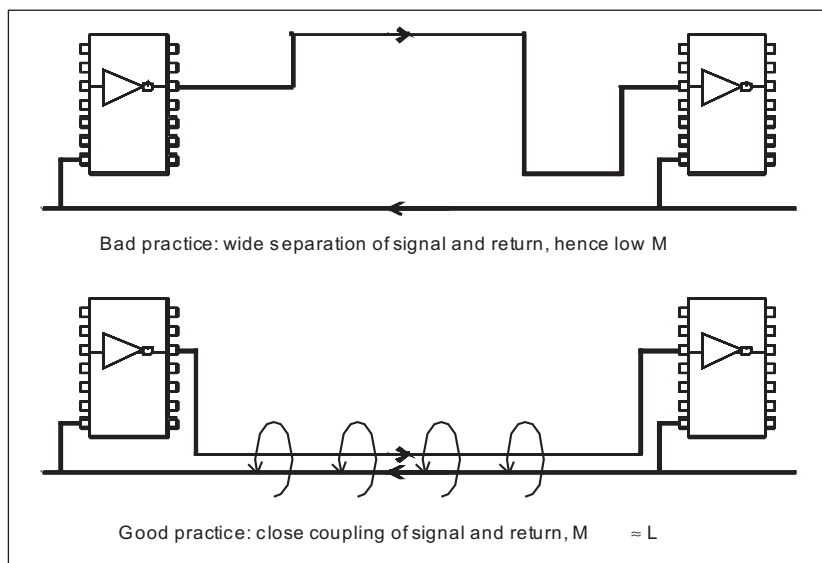
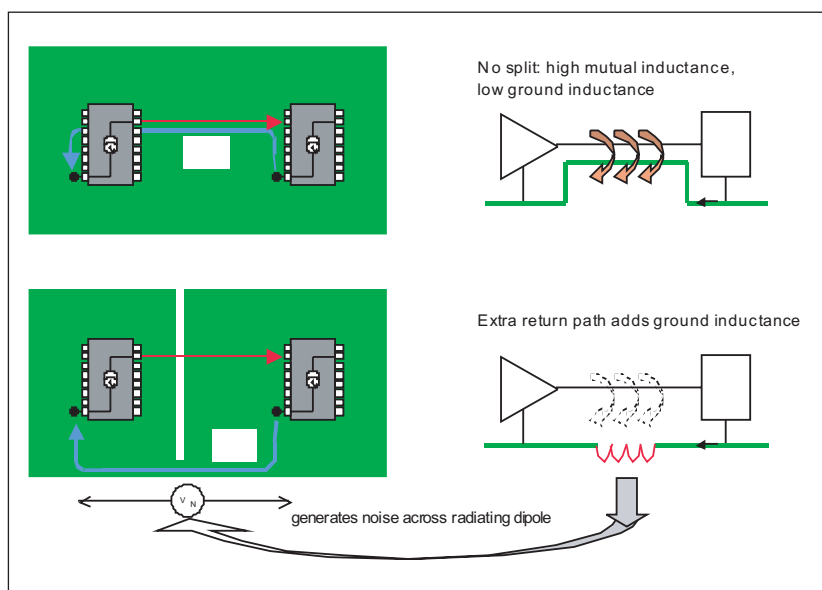


Figure 1: Separation of signal and return

rises. A good rule of thumb is to keep critical tracks further away than $10h$ from the edge of the plane or the edge of any discontinuity, where h is the separation distance in the z -direction between the layer that carries the track and the layer that carries the plane.

This can also apply to power plane layers placed against a 0V plane. If the power and 0V planes were to have exactly co-located edges, the transmission line formed by the two planes would be terminated in a sharp open circuit and the problem of transmission line resonances would be potentially severe, leading to a high degree of coupling to and from the PCB at the resonant frequencies. Staggering the edges of the planes softens the impedance mismatch and reduces the Q of the resonance. In fact, in contrast to the recommendation not to split the 0V plane, it is a good idea to split power planes into several smaller planes each of which supplies a par-

Figure 2: The effect of a discontinuity on return current



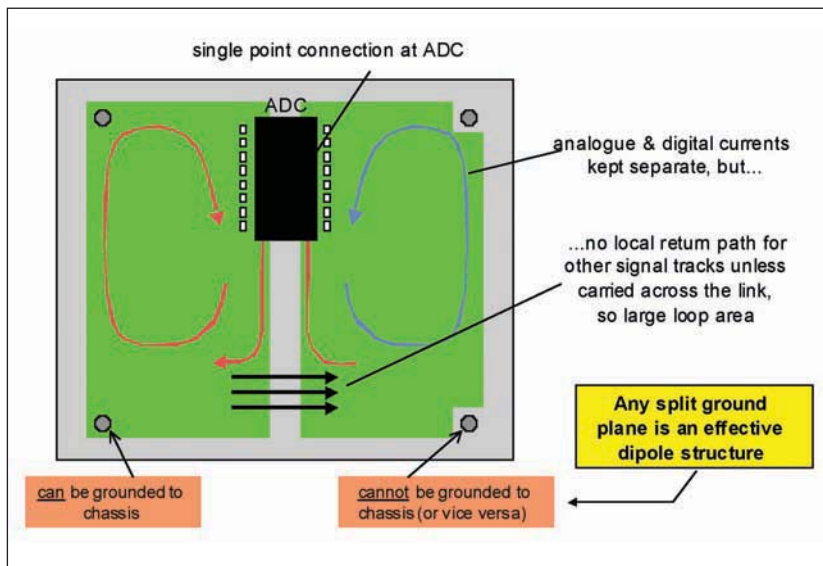


Figure 3: The problems of split ground planes

ticular application, and to decouple these from each other with small ferrite chip inductors.

Ground planes in the z direction

Successful implementation of a 0V plane on a PCB requires careful attention to the layer stack-up, that is, the order in which the different tracking purposes are assigned to different layers. In multilayer (four layers or more) configurations, the most important aspect is that every signal layer should be adjacent to a 0V or power plane layer. Also, power and 0V planes should be on adjacent layers, possibly with deliberately thin separation, to take advantage of the interlayer capacitance for high frequency decoupling. Critical tracks, for instance those carrying high di/dt signals such as clocks, should be routed adjacent to 0V rather than power planes. Such tracks should also not jump through vias from one 0V reference layer to another, unless the 0V layers are tied together with vias near that point.

For boards with more than six layers, multiple 0V plane layers are needed so that tracks and power plane layers can always be located next to a 0V plane, but it is also essential to tie these together with grounding vias at small intervals – intended connections to 0V will provide this automatically, but in areas of the board where this doesn't happen, a 1cm spacing of such "stitching" vias will assure good performance of the planes up to about 2GHz.

One consequence of the dependence on layer separation is that the detail of the vertical sandwich construction of the PCB can be as important for its EMC as is the layout pattern. The separation is determined by the thickness of the pre-preg and core materials used in the build of the bare board. Very often, this is left to the PCB manufacturer and is not specified by the circuit designer, or anyone else. As a result, you end up with the default thick-

nesses used by a particular board supplier, which might be perfectly adequate for the EMC performance of the product and so is never questioned. But, if another board supplier is chosen during the product life cycle, it is entirely possible that a different set of thicknesses could be used which result in changed performance. To guard against this, make sure that you specify layer thicknesses in the PCB drawing, if necessary checking with your preferred supplier what their defaults will be.

Constant impedance tracks

It is becoming commonplace to find that some tracks, because they carry very high frequencies or are routed long distances, have to be treated as transmission lines so that reflections along their length are minimised. This contributes both, good external EMC in the form of lower emissions and good signal integrity, improving the reliability of the circuit. This means that the tracks have to avoid discontinuities in their linear and cross-sectional geometry. This in turn requires that:

They should only run in one layer, with no vias except at their ends

This layer should be adjacent to one (microstrip) or two (stripline) 0V planes

The track cross-section dimensions and dielectric constant of the layer should be defined and controlled throughout the layer to give the correct constant transmission line impedance

Changes in direction of these tracks should be gradual rather than abrupt.

This then requires that at least one layer on a PCB is devoted to these "constant impedance" tracks, and at least one adjacent ground plane layer (completely unbroken) is essential to maintain their characteristics.

Devoting a layer to ground

The ground plane has enough functions on any electronic circuit board to ensure that at least one layer needs to be devoted to it alone. For all multilayer boards, it can be incorporated as a matter of course. Even for low cost double sided boards, with care, a ground plane can be implemented on one side, provided that the rules about avoiding breaks are understood. When such a plane is implemented properly, the external EMC of the board is automatically improved. Even simple analogue circuits are susceptible to exposure from radiated RF fields in the environment and any designer who consciously avoids putting in a ground plane, on the grounds of cost, will find that satisfactory EMC performance will need greater cost and effort than if the plane had been implemented from the start.

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Input capacitance	40pF+oscilloscope capacitance
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Switch position 2

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Input capacitance	12pF if oscilloscope i/p is 20pF
Compensation range	10-60pF
Working voltage	600V DC or pk-pk AC

Switch position 'Ref'

Probe tip grounded via 9M Ω , scope i/p grounded

Power Conversion Topology and EMI

Philip Lioio, Senior Field Applications Engineer at Vicor Corporation, in Andover, Massachusetts, analyses the noise for different DC-DC converter topologies

Electronic systems throughout the world are powered by direct current (usually after AC rectification), which in most cases is converted by DC-DC converters to the voltages needed by each part of the system.

Today, power conversion is usually accomplished by high-density DC-DC converter components based on high-frequency switching technologies.

Efficient high-frequency operation has long been recognised as the key to achieving high-power density and improved performance in switch-mode converters. High-frequency operation translates into smaller magnetics and capacitors, faster response times, lower noise levels and smaller filters.

Unfortunately, however, all DC-DC converters generate electromagnetic interference (EMI) or noise. This noise – common mode, differential mode and radiated noise – can vary widely among DC-DC converters from supplier to supplier and topology to topology. The reasons for this are diverse, but they include fundamental differences in the way noise is produced.

Knowing your noise

Although many different topologies are used, none is superior in every respect. Some applications have requirements that are best satisfied by a specific topology. Although the number of designs, or topologies, of DC-DC converter components certainly number in the hundreds, two are dominant: pulse-width modulation (PWM) and quasi-resonant zero-current switching (ZCS).

Full consideration of the large number of topologies available could be a daunting task, so it is helpful to consider the noise performance of the two main topological classes. Specifically, a comparison is made between DC-DC converters using fixed frequency PWM and variable frequency quasi-resonant ZCS.

PWM converters are power density limited because they inherently trade off efficiency against operating frequency. The problem is the switching losses – heat is generated in the switching element each time it discontinuously makes and breaks inductive current flow during its brief turn-on and turn-off transitions.

Power dissipation, due to switching losses,

increases directly with operating frequency in PWM converters until it becomes a dominant loss factor. At that point, efficiency declines rapidly and the thermal and electrical stresses on the switch element become unmanageable. Losses attributable to non-ZCS result in a “frequency barrier” which limits achievable power density in conventional converters.

Quasi-resonant ZCS converters overcome the frequency barrier by implementing a forward converter switching at zero current. Each switch cycle delivers a quantised “packet” of energy to the converter output, with switch turn-on and turn-off occurring at zero current. ZCS results in an essentially lossless switch.

ZCS converters can operate at frequencies in excess of 1MHz. By eliminating the fast current discontinuities, characteristic of conventional topologies, zero-current switching results in a virtually lossless transfer of energy from input to output, with dramatically reduced levels of conducted and radiated noise.

The noise generated by the switch is a major difference between PWM and ZCS converters. **Figure 1** shows a comparison of conducted input noise generated by ZCS and PWM DC-DC converters.

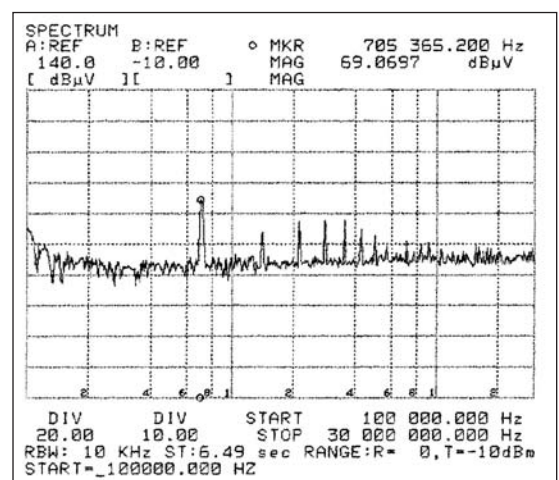


Figure 1: Conducted input noise for zero-current-switching converter with a common mode choke (left) and conducted input noise for pulse-width modulation converter with filter (right). Both 48V input, 5V output, 30A

Among other differences (because ZCS converters have sinusoidal waveforms rather than square waveforms), the lack of sharp edges and lower harmonic content results in much less excitation of the parasitic elements, resulting in less noise. With the pulse-width modulation approach, the input voltage is switched at a constant frequency (usually several hundred kilohertz) to create a pulse train. The width of the pulses is adjusted to provide the necessary power to the load at the correct voltage. At full load, the current waveform looks much like a square wave (see **Figure 2**).

Designing filters

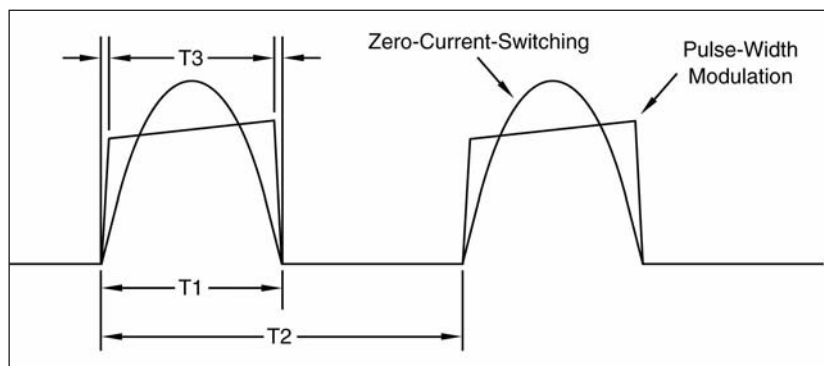
Many designers intuitively assume that it's easier to design a filter for a fixed frequency converter than for a variable frequency converter. In reality, the opposite is true. The perception is, in all likelihood, attributable to the term "fixed frequency", which is actually a misnomer. Both topologies have frequency elements that are more or less fixed and frequency elements that vary as a function of operating conditions.

Figure 2 compares the waveforms of the current flowing through the main switch. In a module using a quasi-resonant topology, the pulse width (T1) is fixed, while the repetition rate (T2) is variable. In a module using PWM, the opposite is true; the repetition rate is fixed and the pulse width is variable. Each of these topologies generates characteristic noise spectra as shown in **Figure 3**.

In the variable-frequency design, however, there are no high-frequency components associated with the leading and falling edges of the current waveform (T3), because it is essentially a half-wave rectified sine wave. The spectral content of the variable frequency waveform is lower in amplitude and contained in a narrower band.

In PWM converters, most of the energy is at the fixed frequency and odd multiples (harmonics) of it. A 100kHz PWM converter will have most of its conducted noise at 100kHz and some at 300kHz and 500kHz. They also have significant harmonics at or above 10-30MHz, due to the shape of the current waveform, i.e. high di/dt that excite parasitic elements within the converter. The input conducted filter has to be sized to handle maximum power at 100kHz. In the fixed frequency waveform, the spectral content is higher in amplitude and spread over a broader range of harmonics.

It is clear that an effective first step to minimise noise generated by the DC-DC converter is to select a topology – such as zero-current switching – that is inherently lower in common mode noise. Incidentally, some products should be avoided in noise-sensitive applications. Control devices mounted on copper plates, for example, create par-



asitic capacitance from primary referenced control devices to secondary referenced control devices through the copper base, resulting in high common mode noise.

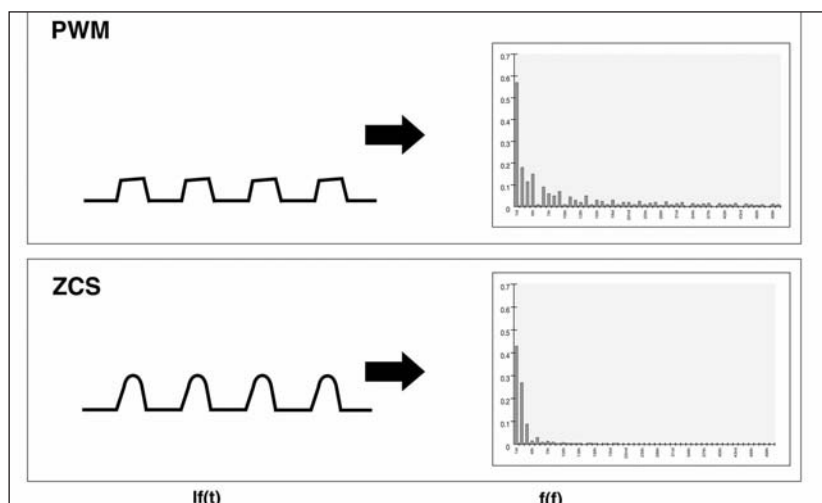
Passive EMI filter examples

Although component power modules usually incorporate some internal input and output filtering, additional external filtering is often needed to meet either system requirements or agency specifications. For example, FCC and European agencies specify the allowable levels of power supply noise that may be conducted back into the AC line. Many designers tackle these issues on their own, but most DC-DC converter manufacturers provide detailed application notes and offer the assistance of a knowledgeable, experienced and easily accessible applications engineering staff. In addition, some DC-DC converter suppliers also offer AC front-ends and EMI filters as modular accessories. These filters not only save time, they also are a means of risk prevention, as well. The EMI filter is designed to work with the supplier's converter modules, and, assuming proper layout, the combination is certified to meet the specified EMC directives.

In the US and Europe, conducted noise emissions are governed by the Class A and Class B limits of

Figure 2: Current waveforms produced by zero-current switching and pulse-width modulation

Figure 3: Current waveforms and spectral content for PWM (top) and ZCS (bottom). Note: waveforms not drawn to scale



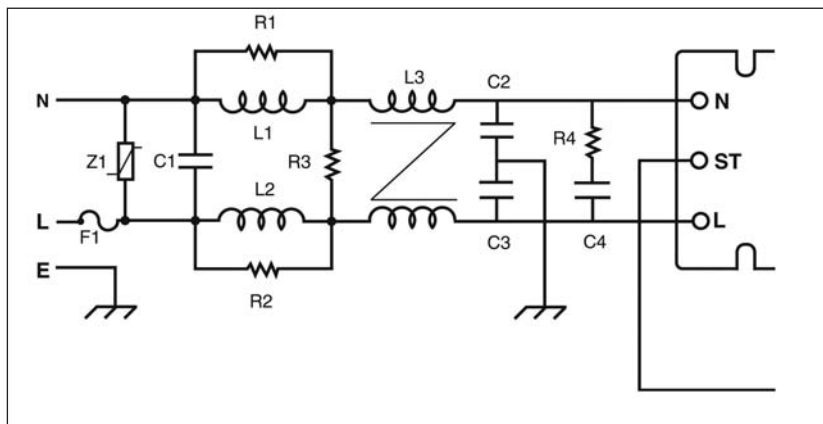


Figure 4: Input EMI filter for EN55022, Level B compliance

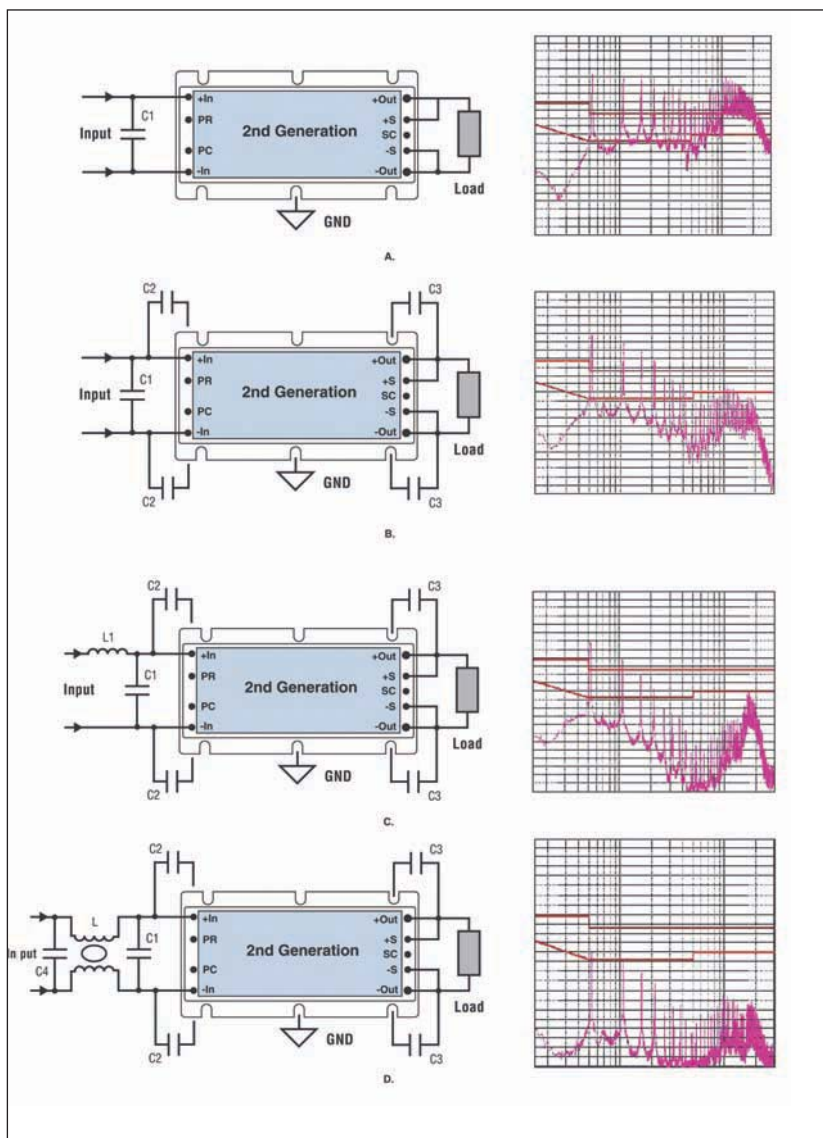


Figure 5: 48V, 150W DC-DC converter with associated noise spectrum with the use of (a) differential-mode capacitor, (b) bypass capacitors, (c) differential-mode inductor, (d) common-mode filter (with differential mode choke eliminated)

both FCC and VDE standards. In the US, the FCC requires compliance with Class A for equipment operating in factory settings and Class B – the stricter standard – for equipment destined for home use. In Europe, all countries require that equipment for both home and factory use meets the VDE Class B standard.

Most switching power supplies today operate in the frequency range between 100kHz and 1MHz. Usually, the dominant peaks in the conducted noise spectrum reflected back to the power line correspond to the fundamental switching frequency and its harmonic components. Conducted emissions standards such as EN55011 and EN55022 set quasi-peak and average limits on conducted noise reflected from the input of converters or power supply systems, back to the source over the frequency range of 150kHz to 30MHz. In order to comply, all of the conducted noise (the peaks in the spectrum) must fall below the specified limits.

EMI filters are most often constructed in a single package (with configurations similar to that shown in Figure 4). The EMI filter is a through-hole filter with a common mode choke and Y-capacitors (line-ground), plus two additional inductors and an X-capacitor (line-line). Transient protection is provided by Z1. This filter configuration provides sufficient attenuation to comply with the Level-B conducted emissions limit.

Nevertheless, capacitors, inductors and filters (both active and passive) are commonly used in power supply designs to attenuate the amount of conducted noise, both common-mode and normal-mode. First, the effects on the noise spectrum of adding individual components or filters are shown leading up to the result with a full common-mode filter. Then, one approach to meeting EMI compliance in a contemporary application is discussed.

The 48V input DC-DC converter shown on the left in Figure 5a has a differential mode capacitor (C1) on the input. This single-mode electrolytic capacitor, 120μF, 100V, is used to ensure low input impedance, stability and good transient response. It's an energy reservoir for the converter. To reap the most benefit, the capacitor must be as close to the input pins of the module as possible.

The module alone and that one capacitor provide a baseline from which to start. The spectra in Figure 5a on the right shows the harmonic content of the noise and the EMI limits, A and B levels, for this converter and differential mode capacitor combination. These measurements were made at 100% load, nominal line for a 48V, 150W DC-DC converter. With this differential mode capacitor only, the converter is clearly not meeting the limits, but the power component is not designed to meet any specific EMI limits.

The effect of adding bypass caps to the converter and differential mode capacitor combination, shown in **Figure 5b** is rather dramatic. Notice the bypass cap on each input pin to the base plate, which is ground and each output pin to the base plate. These electrolytic capacitors are 4700pF, 100 V Y caps that are commonly used in the industry. The Y caps are very effective in attenuating the type of noise that the power component generates.

The 48V design with 100% load generates a little higher noise than a 3.3V design with a 50% load would, for example, but, nevertheless, the spectrum in Figure 5b shows some significant improvement.

Even with the addition of a 27 μ H differential inductor (L1), **Figure 5c** shows that the 48V design is still not compliant at the lower frequencies, where noise is still present above the B limits.

Figure 5d shows the next stage. We're adding a common mode choke. The differential mode choke is eliminated because the common mode choke does have differential mode inductance. The common mode inductor accentuates the capabilities of the Y capacitor. That's because it provides high impedance to common mode noise being conducted out of the converter, therefore, the noise follows the path of least resistance to ground, which is through these Y caps.

The spectrum of the 48V converter is just peeking over the top of the B limit, so, a little more filtering would be needed on a 48V converter design. The noise spectrum of a 3.3V converter with a common mode filter would be below the B limit both at 50% and full load.

Active EMI filter example

The trend toward smaller devices with more functionality in smaller spaces continues unabated in the electronics industry. As spaces shrink, the potential interference between devices increases as systems contain more functions in densely packed boards and racks. As frequencies rise and voltage levels fall, the control of conducted EMI becomes an even more important design task. EMI control, a complex design task that is highly dependent on many design elements, makes use of filters, both passive and active, to manage conducted noise.

Active EMI filters, in contrast to passive solutions, reduce the volume of the common-mode choke, allowing the filter to be packaged in a 1" x 1" x 0.2" package, and provide a low profile, surface mount device. Smaller size saves valuable board real estate and the reduced height enhances airflow for better heat management.

Active EMI filters (labeled QP1 in **Figure 6**) are available that attenuate conducted-mode and differential-mode noise over the frequency range of

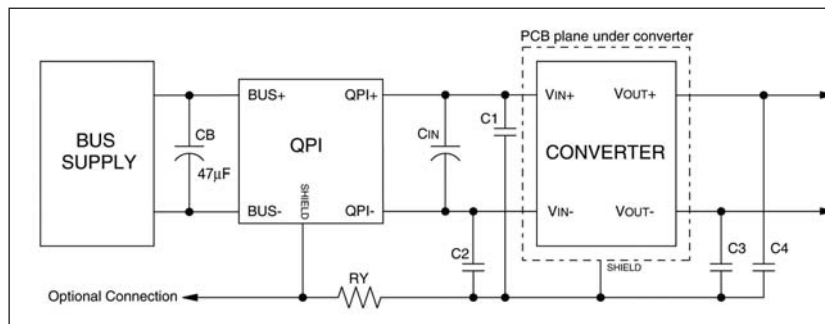


Figure 6: Typical diagram for an active EMI filter (labeled QP1) for a DC-DC converter. Values of C_{in} and C_1 , C_2 , C_3 and C_4 are those normally recommended by the converter manufacturer

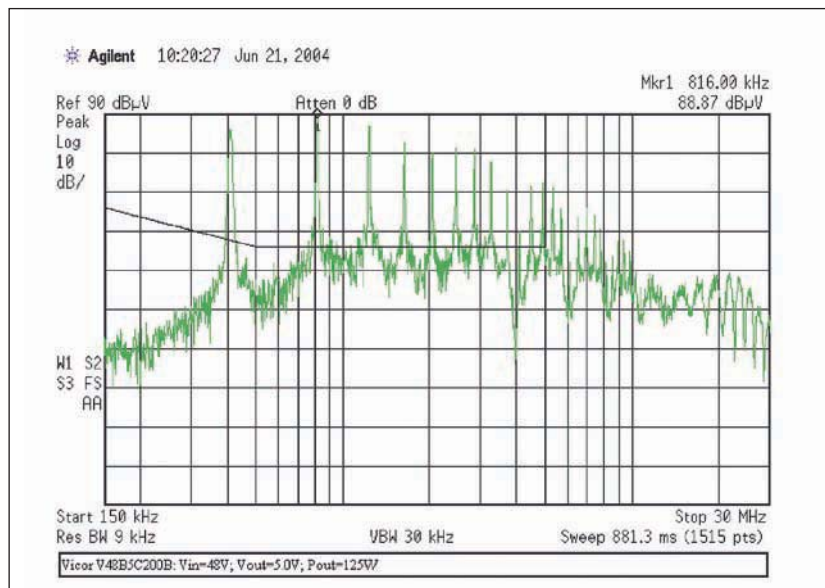
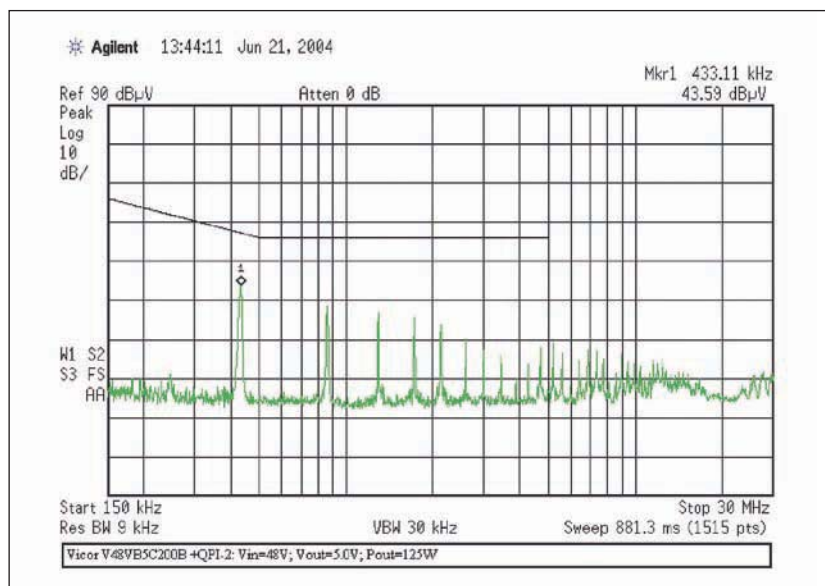


Figure 7: Conducted EMI profiles of a DC-DC converter with no filter (above) and with an active EMI filter (below)



150kHz to 30MHz, required by the conducted emissions standard EN55022 (CISPR22).

Figure 7 shows before and after plots of a DC-DC converter noise profile to demonstrate the performance of an active filter. The plots were taken using the standard measurement technique and set up as defined in CISPR22. The results show the total noise spectrum for a standard DC-DC converter under load compared to the EN55022 Class B quasi-peak detection limit. The plot shows that an active filter is effective in reducing the total conducted noise spectrum to well below the required limits.

Designers should be aware that to select and qualify an EMI filter for conducted noise, they must test the filter in their product under the set-up and conditions specified in the applicable EMI standards. Filter selection or design must be based on the pre-filtered noise magnitudes and the frequency spectrum of concern. A product's conducted noise profile includes differential and common-mode noise. It may also include radiated noise, depending on the EUT shielding and cabling screening in the measurement setup. The IEC (International

Electrotechnical Commission) special committee on radio interference specification CISPR 16-2-1 describes the methods of measurement for conducted disturbances.

Filter performance in the application is highly dependent on the input bus and load impedance and cannot be extrapolated from zero bias 50-Ohm insertion loss data alone. Final noise performance is a complex function of filter elements, equipment grounding and noise source impedances, which vary in magnitude and phase over the frequency spectrum of interest.

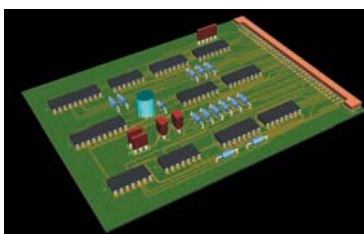
The active EMI filters provide active common-mode attenuation of conducted noise over the EN55022 range from 150kHz to 30MHz by sensing the common mode current flowing in the bus lines and creating a low impedance at the shield plane to re-circulate the noise back to the generating source. When connected as shown in Figure 6, the control loop will actively drive the shield pin and reduce the common mode current in the bus lines to values approaching the common-mode current ratio attenuation curve in Figure 7.

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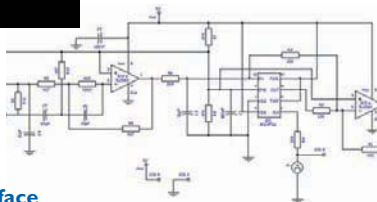
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RoHS

WHAT'S ALL THE FUSS ABOUT?

RoHS (the Restriction of the use of certain Hazardous Substances) and WEEE (Waste Electrical and Electronic Equipment) is the equivalent of Y2K for the electronics industry. The upcoming EU environmental directives are the most significant developments in electronics legislation to happen in many years and will completely revolutionise the way electrical and electronic products are designed, sold, recovered and recycled. Worryingly, many design engineers are still not fully aware that the upcoming legislation will affect them. For those who are, many questions remain unanswered. Complicated exemption rules, uncertainty about how the directives will be enforced, obsolescence and component availability has left engineers unsure of what they need to do and when. The clock is ticking. With only several months to go there's no time to lose in the transition to RoHS. If compliant components aren't already part of the design cycle it could well be too late.

Q: Will WEEE registration be required for producers of board-level products (e.g. printed circuit assemblies aimed at hobbyists, PC add-in cards, etc.)?

Steve Ridgeway, UK

A: This is a difficult question and currently under debate in the EC. PC cards are within scope but differences of opinion relate to whether or not they are finished products or components/accessories. This is important because if they are finished goods they must comply with RoHS. If not, they will be spare parts. Germany has decided they are finished goods but not all member states are of the same opinion.

Q: At the subscription levels currently being suggested, WEEE registration might not be financially viable for very small businesses that occasionally produce bespoke items with electrical/electronic content. What is the advice in this situation?

Steve Ridgeway, UK

A: Unfortunately, there is no de minimus level in the directive, so small companies as well as large have to register, otherwise the company may face a fine for non-compliance. It looks like a flat fee will be payable regardless of company turnover.

Q: Are cables covered by RoHS?

A: The current view of the Commission is that certain types of cable – whether included in EEE or sold separately – do fall within RoHS. However, the DTI takes a different position and is currently in discussion with the Commission over this.

Q: Can I import for own use?

A: No. The EC's 'Guide to the implementation of directives based on the

New Approach and the Global Approach' declares that imports for own use are still considered as 'placed on the market'.

Q: Can I manufacture non-compliant EEE in-house for own use?

A: Yes. Products built for own use are, generally, not considered as being 'placed on the market' (The EC's 'Guide to the implementation of directives based on the New Approach and the Global Approach').

Q: Will I be able to use non-RoHS compliant material for capacity expansion or an upgrade in electrical/electronic products put on the market before 1 July 2006?

A: In principle, the use of non-RoHS compliant material in EEE products put on the market before 1 July 2006 for the purposes of capacity expansion/upgrades is allowed, providing that the EEE is not put on the market as a new product.

Q: Does RoHS and WEEE apply to ink cartridges?

A: The cartridge itself does not fall under the definition of EEE. It is considered to be a consumable and therefore the RoHS directive does not apply. However, if the cartridge is inside a discarded printer, it is considered part of the WEEE because it is a consumable that is part of the printer at the time of discarding.

Q: How do I find out about the latest new exemptions?

A: It is best to seek advice from DTI policy regarding the progress of exemptions. A full list of exemptions can also be found on the Farnell InOne website.

Q: What is the official definition for a 'Producer'?

A: The Directive defines a 'producer' as a party who:

- Manufactures and sells EEE in an EU Member State under their own brand
- Sells under their own brand equipment produced by another manufacturer
- Imports or exports EEE on a professional basis into an EU Member State.

Q: Does the wheelie bin symbol replace the producer identification marking?

A: No. The symbol requirement is in addition to the producer identification marking, which must appear on the product itself.

Q: I understand that the Directive applies to medical equipment with the exemption of all implanted and infected products. What are infected products?

A: Infected products are understood to be products that have come into contact with blood or other biological contaminants prior to end-of-life.



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.

The New Standard of Reliability for Power Amplifiers

Craig Andrews from Silicon Laboratories analyses how to calculate reliability in power amplifiers and what the best methods may be

Manufacturers of power amplifiers (PAs) have tried to address reliability issues over the years with differing results. Great improvements in factory yield were achieved by integrating power control into GSM PAs in 2001. Since innovations often come in five-year cycles, the next level of innovation is now due.

The thermal demands on mobile phones are increasing due to high duty-cycle transmission and the integration of antennas within the phone packaging. Industry changes make it necessary to eliminate the thermal sensitivity of PAs.

Further improvements in PA reliability are needed. Traditional reliability estimation methods rely on limited thermal modelling in combination with mean time to failure (MTTF) statistics, where reliability is typically quoted as MTTF at a specific temperature. This method is inadequate since the junction temperature is unknown during actual operation. It is necessary to move to the next level of quality to ensure the long-term reliability of PAs.

This paper will present MTTF calculation methods, thermal model construction and a circuit that removes the uncertainty of device lifetime. This paper will also suggest future steps to improve PA reliability.

MTTF or activation energy

Projected reliability in PAs uses a widely-accepted statistical method. Unfortunately, manufacturers use different parameters to project reliability. Some manufacturers quote only activation energy while others quote only MTTF. However, reliability is a multi-dimensional problem that cannot be quoted based on a single-dimensional property. Therefore, it is important to understand reliability calculations in order to understand product reliability. The Arrhenius equation is the basis for calculating device reliability:

$$\ln\left(\frac{t_2}{t_1}\right) = \frac{E_a}{k} \cdot \left(\frac{1}{T_2} - \frac{1}{T_1}\right) \quad (1)$$

where:

t₁ = reference time to failure at reference T₁

t₂ = time to failure at T₂

E_a = activation energy (unique to process)

T₁ = reference temperature for failure time t₁ (temperature in Kelvin)

T₂ = temperature to calculate failure time t₂ (temperature in Kelvin)

Equation 1 calculates median time (t₂) of failure, based upon a known median time (t₁) of failure, failure activation energy (E_a) and failure temperature (T₁). Activation energy is the amount of energy required to induce a specified failure mechanism in a semiconductor technology. Equation 1 can be manipulated to produce **Equation 2**, which expresses lifetime in terms of MTTF:

$$MTTF(T_2) = MTTF(T_1) \cdot e^{\frac{E_a}{k} \cdot \left(\frac{1}{T_2} - \frac{1}{T_1}\right)} \quad (2)$$

The temperature dependence on reliability can be seen by graphing Equation 2 as a function of temperature. **Figure 1** shows MTTF plots for two different manufacturing processes. Both processes have the same MTTF at 125°C. The only difference between the two processes is the activation energy. One

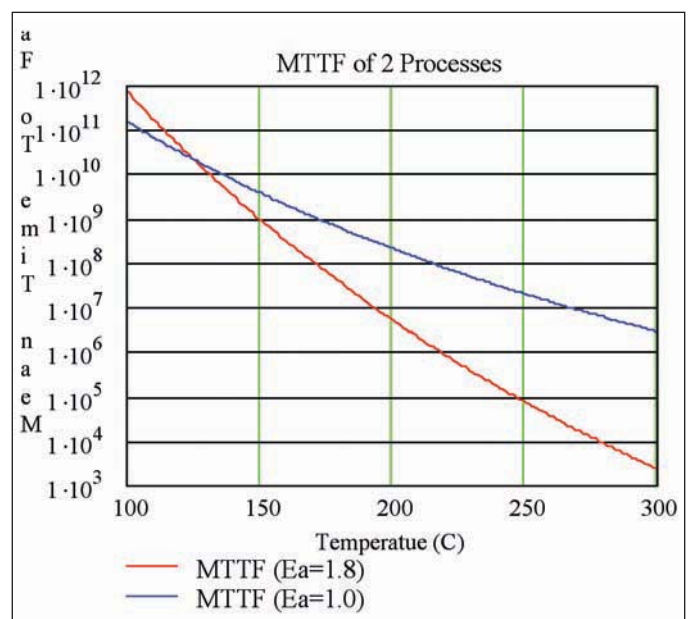


Figure 1: MTTF vs temperature

process has an activation energy of 1.0eV and the other process has an activation energy of 1.8eV.

After looking at Figure 1, it is clear that reliability is highly dependent on the device operating temperature. To put it in

perspective, when operating at 150°C, the process with the lower activation energy outlasts the other process by three billion hours of operation even though the process with higher activation energy has longer lifetimes at temperatures below 125°C. The slopes of the lines in Figure 1 are the activation energies for each process. Thus, higher activation energy results in a steeper slope with respect to temperature, showing an increased failure rate.

Some engineers characterise the quality of a process by the activation energy. Others simply quote MTTF at a specific temperature. From the previous analysis, it is clear that both views represent an overly simplistic approach to reliability. One must consider the device's operating temperature before making a judgment on the optimal activation energy. Likewise, MTTF must be quoted at the operating temperature under worst-case conditions. For reliability estimation, one must consider the device's operating temperature and the inherent failure rate of the process.

Junction temperature

The next step is to look at how junction temperatures in PAs are affected by their operating conditions and packaging.

A common technique for analysing peak junction temperature is a thermal model analogous to Ohm's law. Anything that creates heat is modelled as a current source with the units of watt. All materials are given a thermal resistance with the units of °C/W. These materials also have the ability to store heat, which is called the thermal capacity (J/°C) and are modelled as capacitors.

Figure 2 shows the model of a packaged single die PA. Looking at the static case where the capacitance is ignored, the

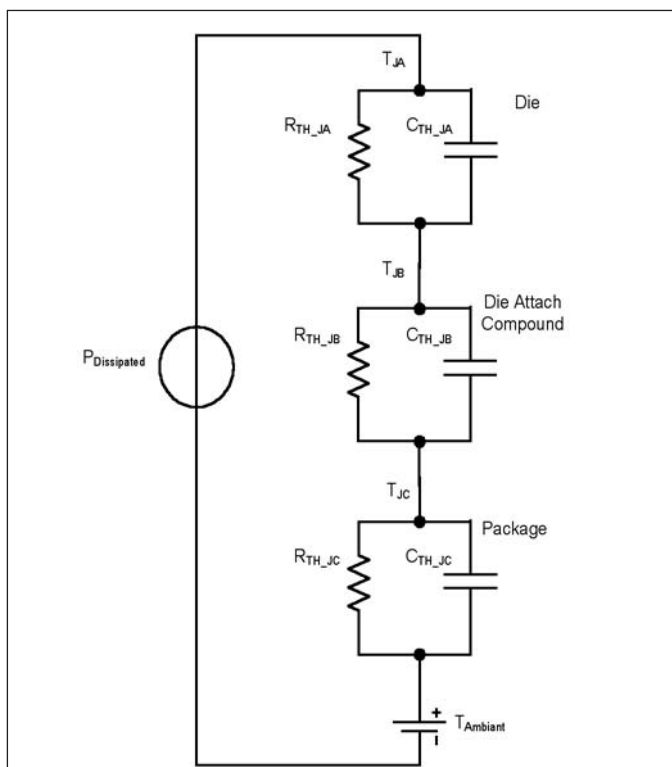


Figure 2: Thermal schematic model of a single-die PA

junction temperature is the sum of the ambient temperature, T_A , and the product of the dissipated power and the system thermal resistance as shown in **Equation 3**. In the dynamic case, the thermal capacity would have to be considered.

$$T_{JA} = P_{diss} \cdot (R_{th_ja} + R_{th_jb} + R_{th_jc}) + T_A \quad (3)$$

Equation 3 can be evaluated using values from the RF3220 WLAN PA. $R_{th} = 76^\circ\text{C/W}$; $P_{diss} = 0.997\text{W}$; $T_A = 85^\circ\text{C}$. These values will yield a junction temperature of 160.8°C. As shown, the junction temperature is dependent upon dissipated power. If the bias current, output power or efficiency change, the dissipated power will also change. This will result in a different junction temperature.

These conditions are for a PA in a package with good thermal conductivity operating at an output power of 250mW. Since the output power levels for WLAN are low, it is easier to achieve acceptable temperatures that result in longer life.

All of the previous calculations assumed a 50Ω load condition. An example of worst-case conditions is shown in **Figure 3** where a mismatch (non 50Ω load) at the antenna causes power to be reflected back to the PA. Under high mismatch conditions, very little power is actually radiated. Since physical laws dictate that energy is conserved, the power reflected back to the PA is dissipated in the form of heat.

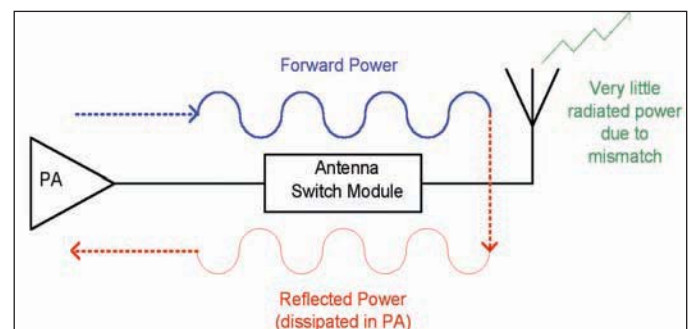


Figure 3: Power flow in mismatch conditions

Conversely, if the PA is heating due to reflected power, the phone is not radiating as much power. It is important to note that under these conditions, the phone will have trouble maintaining the call due to insufficient radiated power.

The sensitivity of this condition is dictated very much by the device technology. Thermal conductors, such as silicon, dissipate heat very efficiently due to their low thermal impedance. Thermal insulators, such as GaAs, tend to heat more under mismatch conditions due to their poor thermal conductance.

Thermal management

To fully appreciate the importance of thermal management, it is important to consider a GSM PA operating under worst-case conditions. This will help clarify how elevated temperatures impact device reliability.

The following case should be considered, as it will produce a worst-case situation:

- GSM PA transmitting 34.5dBm (for highest dissipated power)
- 50% efficiency (actual efficiency can be worse at high supply voltage)

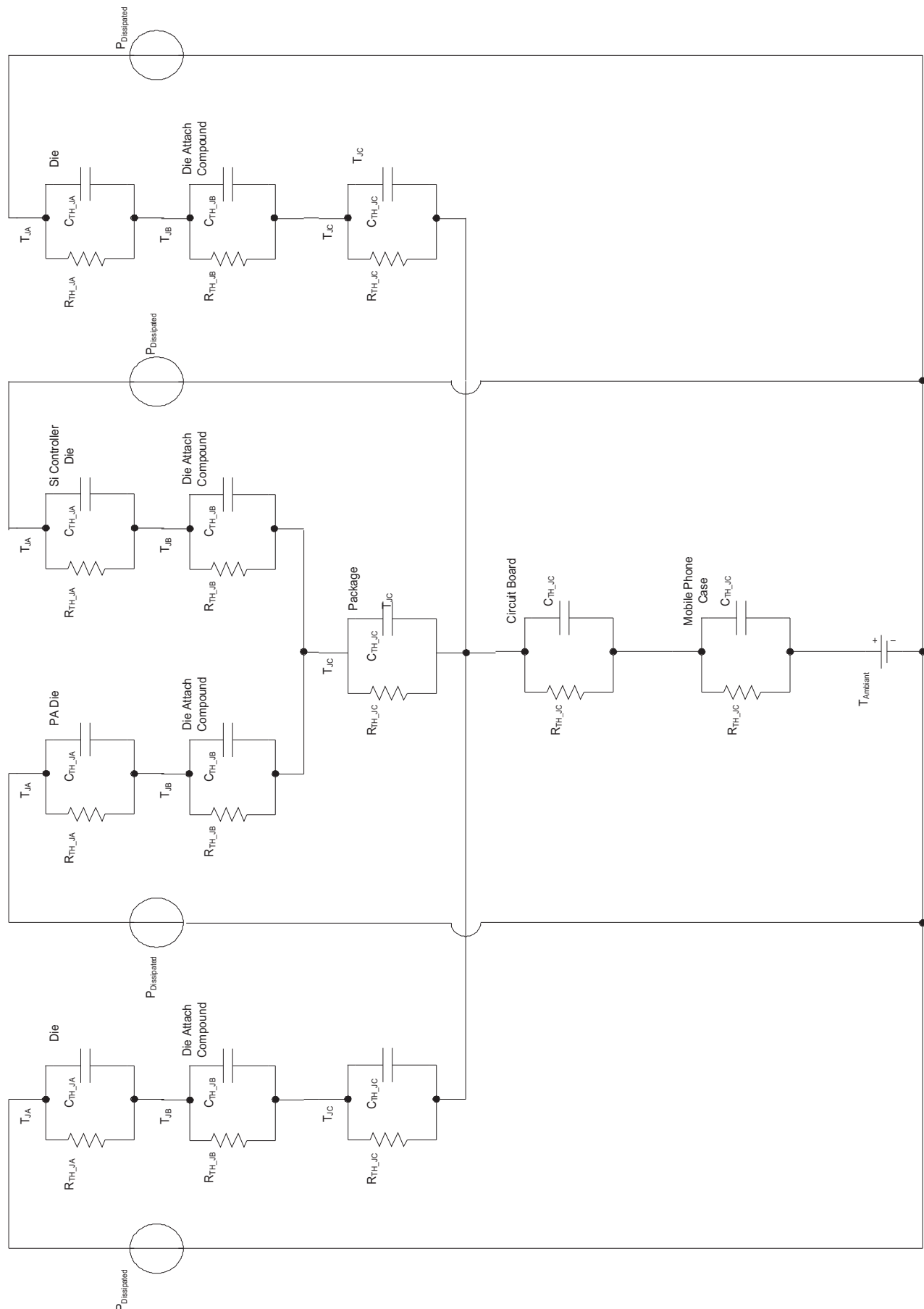


Figure 4: Thermal model for mobile phone with four heat sources

- Laminate package (for highest thermal resistance)
- 10:1 output impedance mismatch (worst-case load impedance)
- 50% duty cycle (GPRS class 12)
- Ambient temperature of 85°C

Industry leaders have known that laminate substrates are problematic since at least 1999. It is best to perform the calculation with the thermal impedance of laminate; however, since PA manufacturers do not publish laminate thermal impedances, the following calculation will be based on a lead-frame package. The data from the RF3220 will be used since it is a heterojunction bipolar transistor (HBT) in a lead-frame package.

Using Equations 2 and 3, the MTTF under ideal conditions is calculated to be 8×10^5 years. Under typical conditions, it is 1.9×10^3 years. Calculating using worst-case conditions, the MTTF is reduced to about one year. Using the thermal impedance for laminate and the dynamic thermal model, the MTTF degrades even further.

An important note is that the MTTF reports a mean time to failure. To know when the worst-case failure would occur, one would need to consider the spread. Standard quality procedures dictate a process capability (Cpk) of 1.5, which corresponds to a spread of 4.5 standard deviations. For the process being discussed, the standard deviation of the time to failure is 0.6 on a logarithmic scale. Since one would subtract 4.5 standard deviations on a logarithmic scale, one divides on a linear scale. This would mean that all of the aforementioned MTTF values reported would have to be divided by $4.5 \times 10^{0.6}$. The worst time failure is actually 17.92 times worse than the average failure rates. These results show how a PA lifetime can go from thousands of years of life to less than a month, by considering the worst-case conditions and the distribution of lifetime due to the process.

The discussion of thermal modelling presented thus far does not address the true environment where the PA will be used. The real environment has the PA mounted on a circuit board in an enclosed package with high thermal resistance and on other heat generators coexisting with the PA.

To accurately model the junction temperature for the PA in a mobile phone, one must construct a model is shown in **Figure 4**. There must be a power source for each device that will dissipate heat.

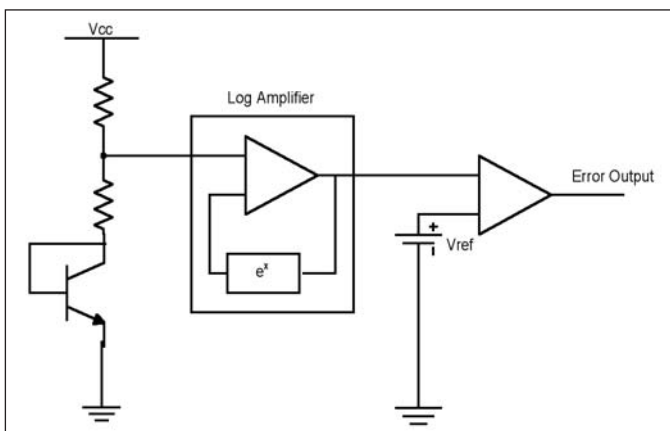


Figure 5: Over-temperature protection circuit

It is easy to see that the thermal model for a mobile phone is too complex to estimate using a simple calculation. A three-dimensional thermal simulator must be used to accurately predict the maximum junction temperature. Thus, accurately predicting the lifetime of PAs in a mobile phone becomes nearly impossible.

The accuracy of the results is further complicated when the PA is designed in a thermally resistive technology, such as gallium arsenide (GaAs). The thermal impedance of GaAs is three times larger than that of silicon (Si). Furthermore, the thermal impedance of GaAs is also temperature dependent, where the thermal impedance increases as the junction temperature increases. This could be very problematic if the thermal model is extracted under ideal conditions. Detailed analysis has shown that a number of device issues contribute to the complexity of this model.

- Spacing of emitter fingers greatly impacts thermal coupling
- Self-heating nature of GaAs HBT
- Bias circuits can cause part of the device to run hotter than some other of its parts
- Variations in bias current
- Longer emitter fingers introduce a greater thermal gradient across the device and can result in thermal runaway on only a limited portion of the emitter finger
- Asymmetrical device geometries are required to maintain uniform temperature distribution

The complexity of thermal models and the extreme sensitivity to heat show why PAs continue to have the highest defect rate of the various components in mobile phones. It is essential to improve PA reliability. There are two methods of addressing this issue. The first is to do complex modelling to accurately predict the junction temperature under worst-case conditions. However, modelling does not fix the problem, it only predicts the failure. The second method is to design the PA in such a way that these temperatures are never achieved. The latter should be the new reliability standard for PAs.

Building temperature sensors

An over-temperature protection circuit is an easy way to address this concern. The temperature sensor is placed on the die in such a way that it detects the maximum junction temperature. There are several ways to build a temperature sensor. An example of a temperature sensor is shown in **Figure 5**. The logarithmic amplifier senses the voltage, which changes based on the temperature-varying current through the transistor. The output of the logarithmic amplifier is linear with respect to temperature. Finally, an error output comes from a comparator circuit. When a pre-determined temperature is exceeded, the error output changes to indicate that the PA should be turned off. The output of this circuit can be used to drive a disabling circuit that limits the maximum temperature.

This circuit guarantees that the PA will never enter an unsafe region that would cause rapid degradation of the device's lifetime. It eliminates the need for complex modelling and provides a way of reducing the defect rate of PAs. With the co-existence of silicon in most PA modules, it should be the standard that PAs also integrate protection circuits to ensure reliability.

The placement of the temperature sensor is dependent upon the device technology. In a technology that has low thermal resistance, such as silicon, the placement is not critical since the heat spreads evenly. However, in insulating technologies, such as GaAs, it is essential that the temperature sensor be placed near the hottest portion of the die.

If the PA does not have a thermal protection circuit, the modelling must be approached very carefully. As discussed, there are many elements the model must include to be accurate. Furthermore, the estimation of MTTF for certain processes is inconsistent.

Despite the advances that have been made, it is important to continue to look forward and think about the next steps in improving the quality of PAs. As mobile phones integrate more functionality, advances in the RF section must allow the development effort to focus more on added features and not the RF section. Some possible efforts that could contribute to this are:

1. Ensure that all data sheet specifications are guaranteed across all conditions (temperature, voltage, etc.)
2. Improve the moisture sensitivity (MSL) rating of PAs.
3. Implement over-voltage protection circuits that protect the PA during mismatch.

New PA standard

Modern standards of quality dictate that devices should be designed to avoid any mode that could be destructive to the device. This standard has been applied to all other circuits except PAs.

If one chooses a PA without a thermal protection circuit, the MTTF data from corner lots that include extreme combinations of bias current, output power and efficiency should be provided by the supplier. Additionally, PA suppliers should provide MTTF data from corner lots taken under worst-case conditions including 50% duty cycle and at the worst case current in a 10:1 VSWR. Suppliers should also provide confidence intervals, including confidence of the mean and standard deviation, which offer statistical calculation of the confidence of the data. This should include standard deviations for all MTTF data. Lastly, suppliers should also offer thermal impedance and thermal capacity of the complete package referenced to the hottest point on the die.

Collecting this extensive data is difficult, thus, an easier and more effective approach should be adopted. It should be required that PAs monitor their own temperature and reduce operation when acceptable conditions are exceeded. This should be the new standard for PA reliability.



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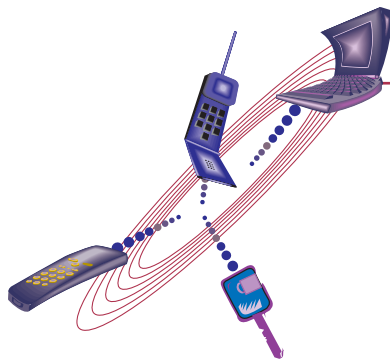
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The R&TTE Directive – An Opening Too Far?

By Mike Brookes

The European Commission (EC) launched the Radio and Telecommunications Terminal Equipment Directive (R&TTE), or Directive 1999/5/EC, in 2000. It had – and still has – the laudable objectives of making the European Telecommunications market more open while simultaneously reducing time and costs of entry for manufacturers.

Prior to R&TTE, the telecommunications regulator for each EU member state (in the UK, the Radio Communications Agency, now Ofcom) issued national performance standards, compliance with which was mandatory, obtained by submitting samples of products to an approved test house for testing against the appropriate standard. Success resulted in the issue of a Type Approval Certificate. This gave the supplier legal means to introduce his product to the targeted national market.

The procedure was criticised by industry as slow, inefficient and expensive, since every EU member state demanded compliance with its own stan-

dards and regulations. Additionally, National Type Approval inevitably invited criticism by importers of protectionism to local producers.

Since a major objective of the European Commission was to develop an open 'common market', elimination of such potentially restrictive practice was a priority.

National radio performance standards were eliminated with the formation of ETSI (European Telecommunications Standards Institute), wherein technical committees formed from representatives of national regulators and industry produced EU-wide standards, with the accent on 'harmonisation'.

In parallel, the R&TTE Directive introduced procedures to make product placement on the overall EU market: simple, quick and cheap.

R&TTE introduced the concept of voluntary standards in which compliance with its 'essential requirements' alone was needed, together with the concept of 'presumption of conformity'.

Presumption of Conformity gives a manufacturer (or an

importer) of declaring that his product complies with the harmonised ETSI standard, thus allowing him freedom to launch a product into the whole EU, without the need to undergo Type Approval tests.

There is a requirement on such "self-declaration" suppliers to maintain technical files, supporting the product against the possibility of inspection from a regulator. There are severe penalties for wrongful declaration, but, in a budget-conscious world, where costs of maintaining inspection capability (UK, Radio Investigation Service) are high, inevitably there are those that will take the risk.

Thus, the EU market is wide open but the regulators have lost their main tool – the Type Approval Certificate – for identifying and tracking down suppliers exploiting dishonestly the freedom given to the industry.

At the same time, other world regions, notably the US and Far East, maintain Type Approval procedures. It is, thus, easier for unscrupulous suppliers to dump inferior products in Europe than it is to

enter their own markets.

Unless these regions adopt an R&TTE-like procedure, and there are few signs of this to rebalance the situation, Europe is likely to become a happy hunting ground where local, conventional and honest suppliers are disadvantaged against non-compliant imports, while the regulators are impotent to support them.

This situation could lead to poor quality souring the market against radio products, particularly SRDs, which are increasingly used in consumer products, including toys.

Maybe a small step backwards is needed. No one wants to return to National Type Approval but maybe the re-introduction of a European Type Approval Certificate, valid in all member states and issued by any notified body (a test house – in this context) will provide ammunition for regulators and protection for bona fide suppliers.

Reader's views would be much appreciated on this subject, which is currently under discussion in the ECC (European Communications Committee).

Wireless Software Solutions Firmware revision 2.1 Jan 05

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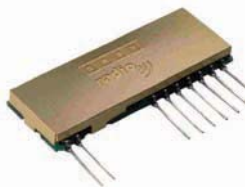
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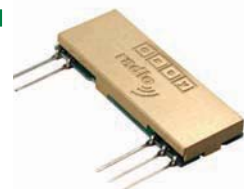
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You can do it the hard way or the easy way when it comes to creating a wireless link for your system, says **Barry Gillibrand** of LPRS



So, the fact that you have made it past the title and are reading this, means that you either thought, “Hey! That looks cool or “With easyRadio, Rodney, next year we’ll be millionaires!”

Well maybe not. At least you must have a little bit of interest in delving that bit deeper into the world of radio.

All too common, the sales meeting echoes time and again: “If only did this or that, we’d sell buckets-load more”. As usual, a tired, stressed and usually ‘already up to his neck in projects’ designer is faced with that dreaded question: “How long will it take?”

For the instances of “this or that” being a wireless link, easyRadio has been developed to give a more positive answer to that very question, and this article aims to help you to see the reasons why it really doesn’t have to be that difficult to enter the world of radio comms.

Getting started

If you have developed products with wireless links before, you will already know that integrating radio links into your application presents an obvious decision: make one or buy one.

For the gurus of this world, the first one will cause no lack of sleep, particularly if you have a team of designers around you as well. However, this option very often amounts to making a product on its own, so for the rest of us, the second option is undoubtedly the easier route.

Ready made modules also come in two varieties, dumb modules with which you have to control the power and full data modulation and a protocol system that can be reliably reversed within the receiver, without interrupting the actual function your product is supposed to do in the first place. Two-way communications naturally require both systems at each end.

Then there are ‘intelligent’ modules where data is uploaded/downloaded in some form or another and this is where easyRadio comes into its own. Using well-supported communications found within most budget microcontrollers in the form of the much-loved RS232 UART interface. With this technology, gone are the days of time and resource-wasting routines to handle the mundane task of actually getting the data to where it is supposed to be. EasyRadio does for an application what visual basic did for Windows development.

Easy does it

Obviously, every project is going to be different and no ‘how to’ style article is going to give you an exact solution to your requirements.

So what we’ll look at is a general, overall introduction to making a simple two-way radio link.

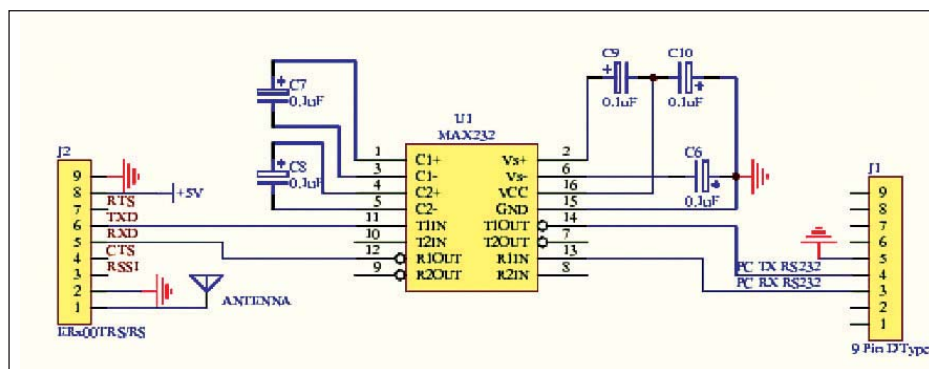
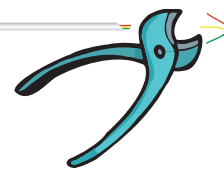


Figure 1: Schematic of a development board

We could just use a complete easy Radio evaluation kit, but that would be all too easy. Instead, setting up a microcontroller to link to a PC seems a bit more useful. On the PC end, however, we will use one of the boards from an evaluation kit, but it is simply to save time in putting together a board with an RS232 level shifter (MAX232) on it and a plug and power source, which is all it basically is. If you want to make one, you could use the schematic in **Figure 1**.

Incidentally, there are lots of development boards around these days. One of my favourites is the trusty millennium board from Elgar (Bluebird), where you have lots of sockets for your PIC processors, switches, LEDs, LCD, keypad, RS232 shifter and proto board.

An easy radio module would fit in there nicely. As I only have one to hand, we will use this development board for the microcontroller end, and the easyRadio evaluation kit PCB for the PC end.

I've decided to use a PIC16F627, although any PIC with a UART would be good for this, as would any from Atmel, Texas Instruments, Motorola and others. It just happened to be in my armoury of parts to have a go with (and it works on the Millennium board).

The modules on each end of the link are both ER400TRS-02. These are transceivers that operate on the 433-434MHz bands and have all the capabilities within to encode/decode, whatever

RS232 data you throw at them. The '02' refers to a second generation of devices, which have eradicated the first generation issues of loss of setting in the EEPROM and frequency consistency.

Easy PIC software

What the application will do is quite simply an echo of whatever is transmitted from the PC back to the PC, with the text "Echoed from the PIC" added to it.

OK, so it's not the most useful application, but it demonstrates a point and it's quite easy to implement. For the code I've also used the CCS C compiler and the source code may be downloaded for reference from www.lprs.co.uk.

First of all, we have to check that it actually works without the radio modules in circuit. Then, we can see how easy it is to convert it over to a wire-free system using the easyRadio modules.

On the next page is a complete C code example of how easy it is to echo the data received from the UART on a PIC.

After compiling the software, we can program the part with either an ICD2 or PICSTART Plus from Microchip. The PIC can then be connected in the board as shown in **Figure 3**.

To test it we can connect this board directly to the PC and load up the easyRadio evaluation software. It is good to get familiar now with the software as you can use it later.

Opening the Communications tab, you

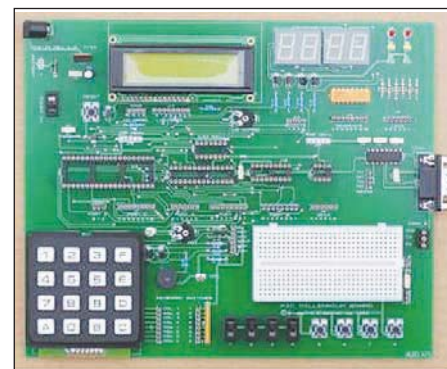


Figure 2: A development board with all the relevant sockets

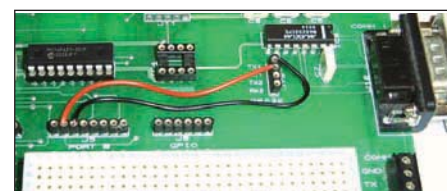


Figure 3: Connecting the PIC correctly

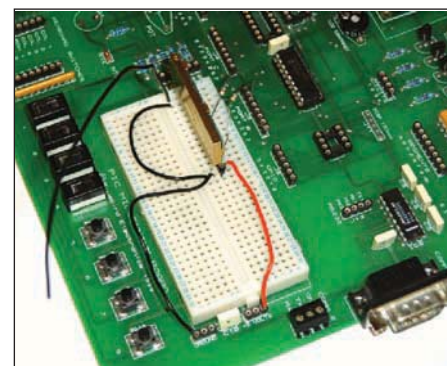


Figure 4: Demo board with the power connected

can enter the text you wish to send from the PC. When the software is running on the PIC, anything you send to it will be echoed with the added text (see **Figure 5**).

All that is needed is to swap over the PIC connections from the MAX232 to the ER400TRS instead.

As the internal power rail of the ER400TRS is on 3.3V, it is advisable to buffer the 5V PIC with 1K resistors to the module. It prevents problems with the Schmitt trigger level on the PIC UART when supplied by 5V.

With the power connected to the module as well, our demo board would be connected as in **Figure 4**.

To complete the link, the second module would be connected directly to the PC, using the easyRadio evaluation kit.

You can do exactly the same again with the software and the result should be identical as before (see **Figure 6**).

So the two-way link is up and running but there are things to be aware of when communicating with the modules.

- There is a 180 byte limit on any single transmission;
- You will have to wait a period of time while the data is being transmitted before uploading more into the buffer. Simply monitoring the BSY pin can do this.

But what makes easyRadio great is the configurability of each aspect of the radio.

- From the UART data rate to the over-air-data rate
- From the channel frequency to the power output
- From the digital separation of channels to digital encryption.

All of which can be controlled through a simple command interface.

All easyRadio commands start with the text "ER_CMD#", after which a function letter is specified, followed by the parameters required to operate it.

When an easyRadio module recognises a valid command stream, it echoes back the text. Then, as a safety mechanism against unwanted command operation, the host will upload the text "ACK" to make the module do what it has been told.

This is dead easy on a PC as the evaluation software demonstrates, but how easy is it to do on a microcontroller. The code function (again using the CCS C compiler) described in **Code 2** is all that is really necessary to do it.

So, say you wanted to change the channel number from the default (7) to channel 3. The command you would have to send is "ER_CMD#C3" (which is 9 bytes in length).

To use the above code simply make a call to the function:

```
Send_ER_Command("ER_CMD#C3",9);
```

```
#include <16f627.h>

#device *=16 ADC=10

#FUSES          INTRC,NOWDT,PROTECT
#USE DELAY(CLOCK=4000000, RESTART_WDT)
#ZERO_RAM

#use rs232(baud=19200, xmit=PIN_B2,rcv=PIN_B1,RESTART_WDT)
#include <stdio.h>
#include <stdlib.h>
#include <string.h>

int Send_ER_Command(char * Command,int Length);

char Buffer[80];

main()
{
    char    i,a,TimeOut;
    set_tris_b(0x0F); //set all PORTB as inputs
    TimeOut = 255;

    delay_ms(1000); // wait for 1 second (Only really need about 20mS)
    //strcpy(Buffer , "ER_CMD#C5");
    //Send_ER_Command(Buffer,0x9);

    while(1)
    {
        if(KBHIT())
        {
            a = 0;
            while(TimeOut--)//will stay in this loop until no more data
            {
                if(kbhit())
                {
                    Buffer[a++] = getc();
                    TimeOut = 255;
                }
            }
            puts("Echoed from the PIC:");
            for(i=0;i<a;i++)
            {
                putc(Buffer[i]);
            }
        }
    }
    //end if(KBHIT())
}
```

Code 1: Echoing the data received from the UART on a PIC

It's as easy as that. This code does all that's necessary to verify the reply from the module and to send the "ACK" string as required.

Other varieties

Other varieties of easyRadio include a dedicated transmitter module for when a one-way link is all that is required and receiver only versions to complement them. A long awaited frequency hopping solution is being developed with the same

user interface and development is moving swiftly on.

RF communications needn't be 'black art' for the digital engineer any more. This type of technology is coming of age now and should be taken seriously in any decision to introduce RF into your next project. Is it worth it? Well, try it for yourself and decide. What you'll get is the years of experience in RF data comms, conveniently wrapped up in a little gold (well, brass actually).

```

int Send_ER_Command(char * Command, int Length)
{
    int i;
    long Timeout = 0xFFFF; // Long initial wait for the first byte from the module
    char RX_BYTE;
    int Success = true;
    for(i = 0; i < Length; i++)    // Send Command byte by byte
    {
        putc(Command[i]);
    }
    //      Receive the command back from the module
    // Compare the recieved buffr to the sent command
    i = 0;    // Reset

    while(Timeout--)
    {
        if(kbhit()) // wait for a byte to appear in the buffer
        {
            RX_BYTE = getc();
            if(Buffer[i] != RX_BYTE)
            {
                Success = false;
            }
            i++;
            Timeout = 255;    // wait for end of data
        }
    }
    //end While
    if(!Success)    // if replied text does not match then return false
        return false;

    if(i == Length)
    {
        putc('A'); // Command complete
        putc('C'); // Command complete
        putc('K'); // Command complete
        return true;
    }

    return false;
}

```

Code 2: Issuing commands to the microcontroller

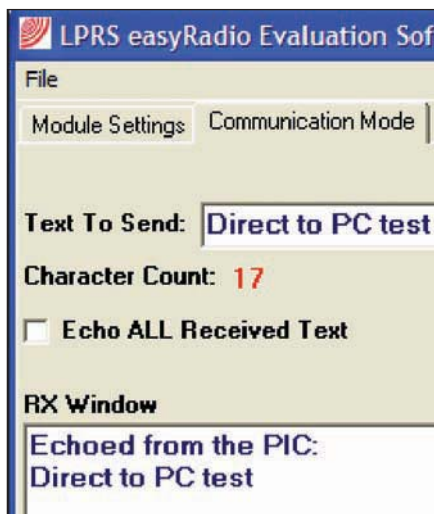


Figure 5: Message window to test the PC link

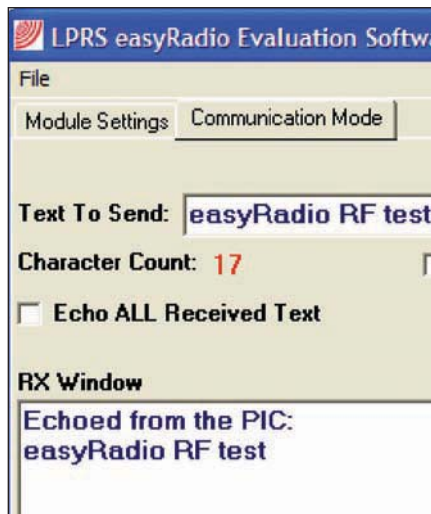


Figure 6: Test message window

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PICmicro Microcontroller CCP and ECCP

The Capture, Compare and PWM (CCP) modules that are found on many of Microchip's microcontrollers are used primarily for the measurement and control of time-based pulse signals. The Enhanced CCP (ECCP), available on some of Microchip's devices, differs from the regular CCP module in that it provides enhanced PWM functionality – namely, full-bridge and half-bridge support, programmable dead-band delay and enhanced PWM auto-shutdown. The ECCP and CCP modules are capable of performing a wide variety of tasks. The tips below describe some of the basic guidelines to follow when using these modules, as well as suggestions for practical applications.

TIP 1: Generating an analogue output

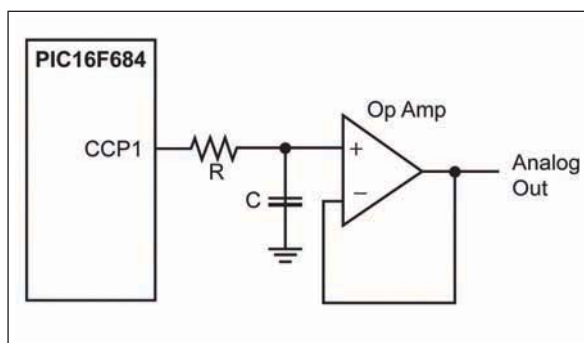


Figure 1: Low-pass filter

Pulse-width modulated signals can be used to create digital-to-analogue (D/A) converters with only a few external components. Conversion of PWM waveforms to analogue signals involves the use of an analogue low-pass filter. In order to eliminate unwanted harmonics caused by a PWM signal to the greatest degree possible, the frequency of the PWM signal (FPWM) should be significantly higher than the bandwidth (FBW) of the desired analogue signal. **Equation 1** shows this relation.

$$FPWM = K \cdot FBW \quad (1)$$

where harmonics decrease as K increases.

R and C are chosen based on the following equation:

$$RC = 1/(2\pi FBW) \quad (2)$$

Pick a value of C arbitrarily and then calculate R. The attenuation of the PWM frequency for a given RC filter is:

$$Att(dB) = -10 \cdot \log[1 + (2\pi FPWM RC)^2] \quad (3)$$

If the attenuation calculated in **Equation 3** is not sufficient, then K must be increased in Equation 1. See application

note AN538, 'Using PWM to Generate Analogue Output in PIC17C42', for more details on using PWM to generate an analogue output.

TIP 2: Boost power supply

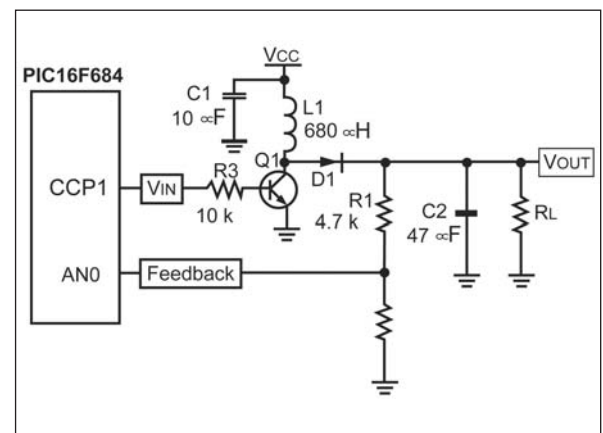


Figure 2: Boost power supply circuit

Hardware

Pulse-Width Modulation (PWM) plays a key role in boost power supply design. **Figure 2** shows a typical boost circuit. The circuit works by Q1 grounding the inductor (L1) during the high phase of the PWM signal generated by CCP1. This causes an increasing current to flow through L1 while VCC is applied. During the low phase of the PWM signal, the energy stored in L1 flows through D1 to the storage capacitor (C2) and the load. VOUT is related to VIN by **Equation 4**.

Note: Technical Brief TB053 'Generating High Voltage Using the PIC16C781/782' provides details on boost power supply design.

The first parameter to determine is the duty cycle based upon the input and output voltages. See Equation 4.

$$\frac{VOUT}{VIN} = \frac{1}{1 - D} \quad (4)$$

Next, the value of the inductor is chosen based on the maximum current required by the load, the switching frequency and the duty cycle. A function for inductance, in terms of load current, is given by **Equation 5**, where T is the PWM period, D is the duty cycle and IOUT is the maximum load current.

$$L = \frac{VIN (1 - D) DT}{2 IOUT} \quad (5)$$

The value for L is chosen arbitrarily to satisfy this equation given IOUT, a maximum duty cycle of 75% and a PWM frequency in the 10kHz to 100kHz range.

Using the value chosen for L, the ripple current is calculated using **Equation 6**.

$$I_{RIPPLE} = \frac{V_{IN} DT}{L} \quad (6)$$

I_{RIPPLE} cannot exceed the saturation current for the inductor. If the value for L does produce a ripple current greater than I_{SAT}, a bigger inductor is needed.

Note: All equations above assume a discontinuous current mode.

Firmware

The PWM duty cycle is varied by the microcontroller in order to maintain a stable output voltage over fluctuating load conditions. A firmware-implemented PID control loop is used to regulate the duty cycle. Feedback from the boost power supply circuit provides the input to the PID control.

Note: Application note AN258, 'Low Cost USB Microcontroller Programmer' provides details on firmware-based PID control.

TIP 3: Varying LED intensity

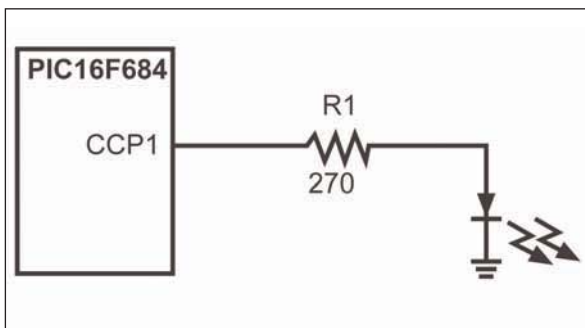


Figure 3: LED drive

The intensity of an LED can be varied by pulse width modulating the voltage across the LED. A microcontroller typically drives an LED with the circuit shown in **Figure 3**. The purpose of R1 is to limit the LED current so that the LED runs in its specified current and voltage range, typically around 1.4V at 20mA. Modulating the LED drive pin on the microcontroller will vary the average current seen by the LED and thus its intensity. LEDs and other light sources should be modulated at no less than 100Hz in

order to prevent noticeable flicker.

The CCP module, configured in PWM mode, is ideal for varying the intensity of an LED. By varying the duty cycle of the PWM signal driving the LED, adjustments to the intensity of the LED can be made. This is accomplished by varying the CCPRxL register between 0 and 0xFF.

TIP 4: Generating X-10 carrier frequency

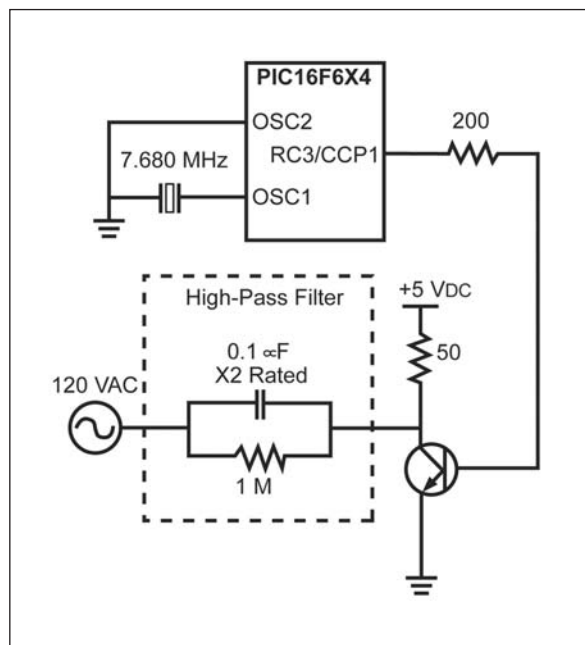


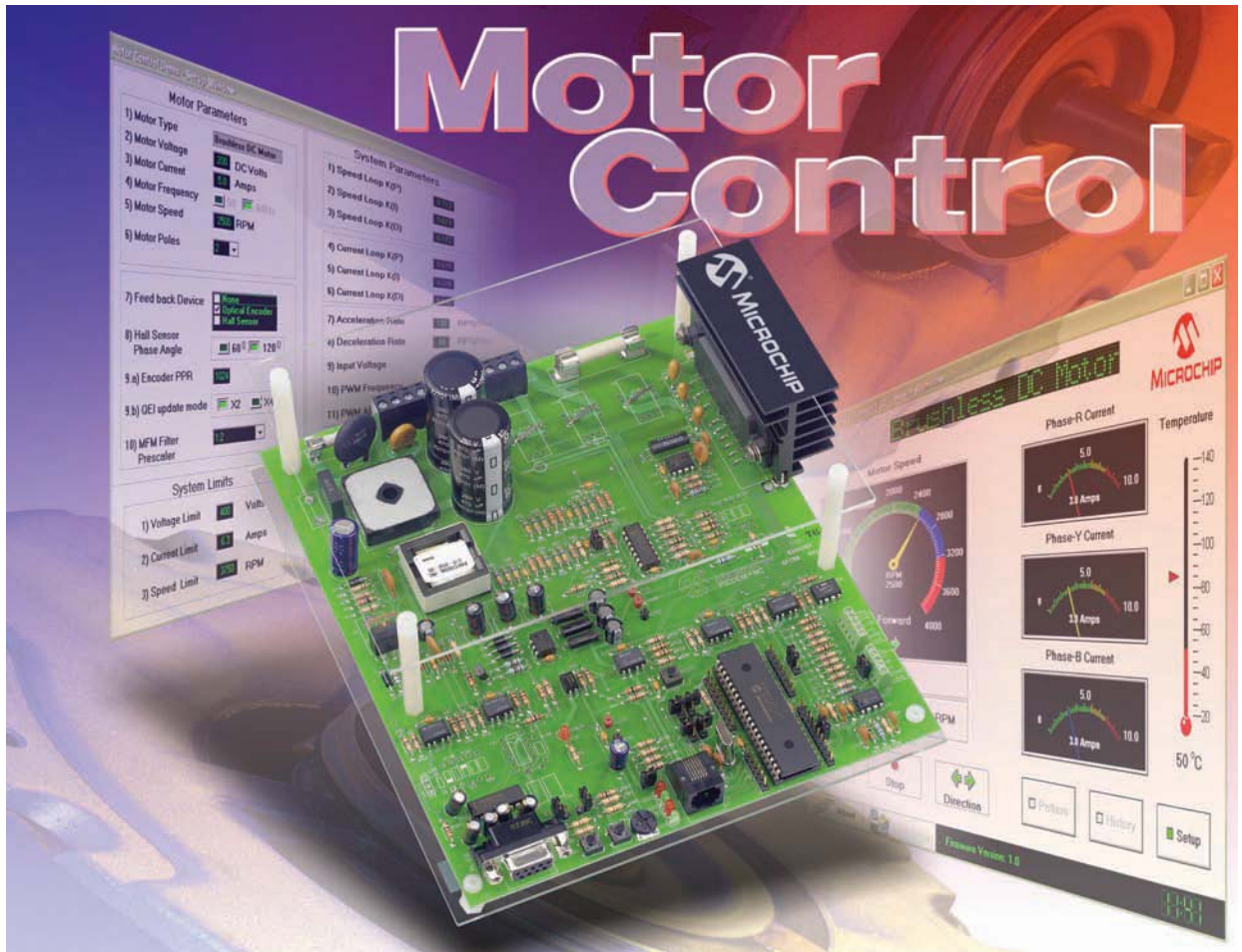
Figure 4: 120kHz carrier frequency mounted onto sinusoidal 60Hz power waveform

X-10 uses a piggybacked 120kHz square wave (at 50% duty cycle) to transmit information over 60Hz power lines. The CCP module, running in PWM mode, can accurately create the 120kHz square wave, referred to as the carrier frequency. **Figure 4** shows how the 120kHz carrier frequency is piggybacked onto the sinusoidal 60Hz power waveform.

X-10 specifies the carrier frequency at 120kHz (±2kHz). The system oscillator in Figure 4 is chosen to be 7.680MHz, so that the CCP module can generate precisely 120kHz. X-10 requires that the carrier frequency be turned on and off at different points on the 60Hz power waveform. This is accomplished by configuring the TRIS register for the CCP1 pin as either an input (carrier frequency off) or an output (carrier frequency on). Refer to application note AN236, 'X-10 Home Automation Using the PIC16F877A' for more details on X-10 and for source code for setting up the CCP module appropriately.

See overleaf for the MicroChip competition

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For those wanting something a little different from their mobile phone, check out the design-led Serene from Samsung and Bang & Olufsen. The display and microphone are placed in the lower shell, the circular keyboard and loudspeaker in the upper shell and a thumb-operated wheel in the middle, taking care of all primary operations.

It comes with a triangular charger/docking station and, if the phone is placed in the docking station or used in a table mode, it can double up as a mini-laptop as the display turns 180 degrees for convenient use. As a special feature for Bang & Olufsen customers, a DECT docking station is also available, which enables Serene to synchronise its phonebook with the domestic cordless system of telephones from the company.

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With popularity in VoIP (voice over IP) growing all the time, USRobotics has introduced a USB Internet phone – the USR809600. Integrating seamlessly with Skype's Internet telephony application, it plugs into a USB 1.1 or 2.0 port and it's ready to go. Similar to a mobile phone, it features an LCD screen offering call status and caller ID.

Designed primarily for the Skype application, the USR809600 can be used with any other instant message platform with voice capabilities. Other features include echo cancellation and noise reduction, offering call quality on a par with traditional phone calling. Available now from Insight, Inmac, Misco, PC World Business, Watford Electronics, Dabs.com, Broad Band Buyer, Ebayer, Nickknows.com and WStore.

Around £29.99



ZyXEL's AG-225H Wi-Fi Finder offers a dual-band 802.11a/b/g USB 2.0 tri-mode client adaptor and soft access point for quick and simple on-the-road access detection and connection.

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high definition LCD screen, which gives users the SSID (Service Set Identifier), channel, security settings and signal strength of all the visible wireless networks in the local area. It provides a secure wireless connection when working as a 54Mbps high-speed USB wireless adaptor and also features a built-in soft access point that can allow a laptop to instantly share wireless access with others. Available now. **Around £51 plus VAT**

www.zyxel.co.uk



The diagram illustrates a vacuum tube radio receiver circuit. It begins with a power supply section on the left, featuring a 230V mains input connected to a transformer with a 12V secondary. The secondary is connected to a 1N4002 diode rectifier. The rectified output is filtered by a 10µF tantalum capacitor (C6) and a 500R W. resistor (R6). The power supply is connected to a vacuum tube (V1) and a 1.5kV/1.625kV meter. The main circuit features a 12V input connected to a 1N4002 diode (D3) and a 10µF tantalum capacitor (C6). The output stage is a push-pull amplifier using two BC546B transistors (T1, T2), with a 10nF capacitor (C1) and a 10k resistor (R1). The circuit is powered by a 12V supply and includes a 1N4002 diode and a 10µF tantalum capacitor. The output is connected to a speaker or headphones.

This high alternating voltage feeds the existing rectifier and voltage doubler block to gen-

erate the high direct voltages for the treble and bass panels (1.5kV (ESL)/1.625kV (LE1) and 6.0kV (ESL)/6.5kV (LE1)). The high direct voltages can be measured with a 10M Ω input resistance voltmeter in series with a 990M Ω resistor network (see figure left). Meter reading will be 1/100 of the desired value. Extreme care should be taken when working with such high voltages!

The oscillator is powered by a voltage regulation circuit around VR1. Trimming of P1 sets the output voltage of the oscillator at w2: V(w2). After removing the mains transformers inside the cabinets of both types of loudspeaker, there is enough space to fix the circuit boards of the oscillator and its internal voltage regulation.

Two standard mains adaptors (one for each loudspeaker, e.g. Egston 12V/6W) feed the loudspeaker's new internal power supply with the required direct voltage (min 10V) and current (min 80mA). The existing mains input sockets have to be replaced by low voltage types (B1). D3 prevents the circuit from wrong input polarisation.

I've been testing the LE1 with a 2 x 30W valve amp (Braun CSV60, power valves: PL504) and a 2x30W transistor power amp (Douglas Self blameless type – Self on Audio, p. 223 – with output stages formed by a pair of multi emitter transistors – 2SC2565Y / 2SA1095Y – driven by BD139/BD140). Both amps did an excellent job, no hum could be heard at all, the overall sound was and still is excellent – but better than before.

Burkhard Vogel
Austria

PIC-based metronome

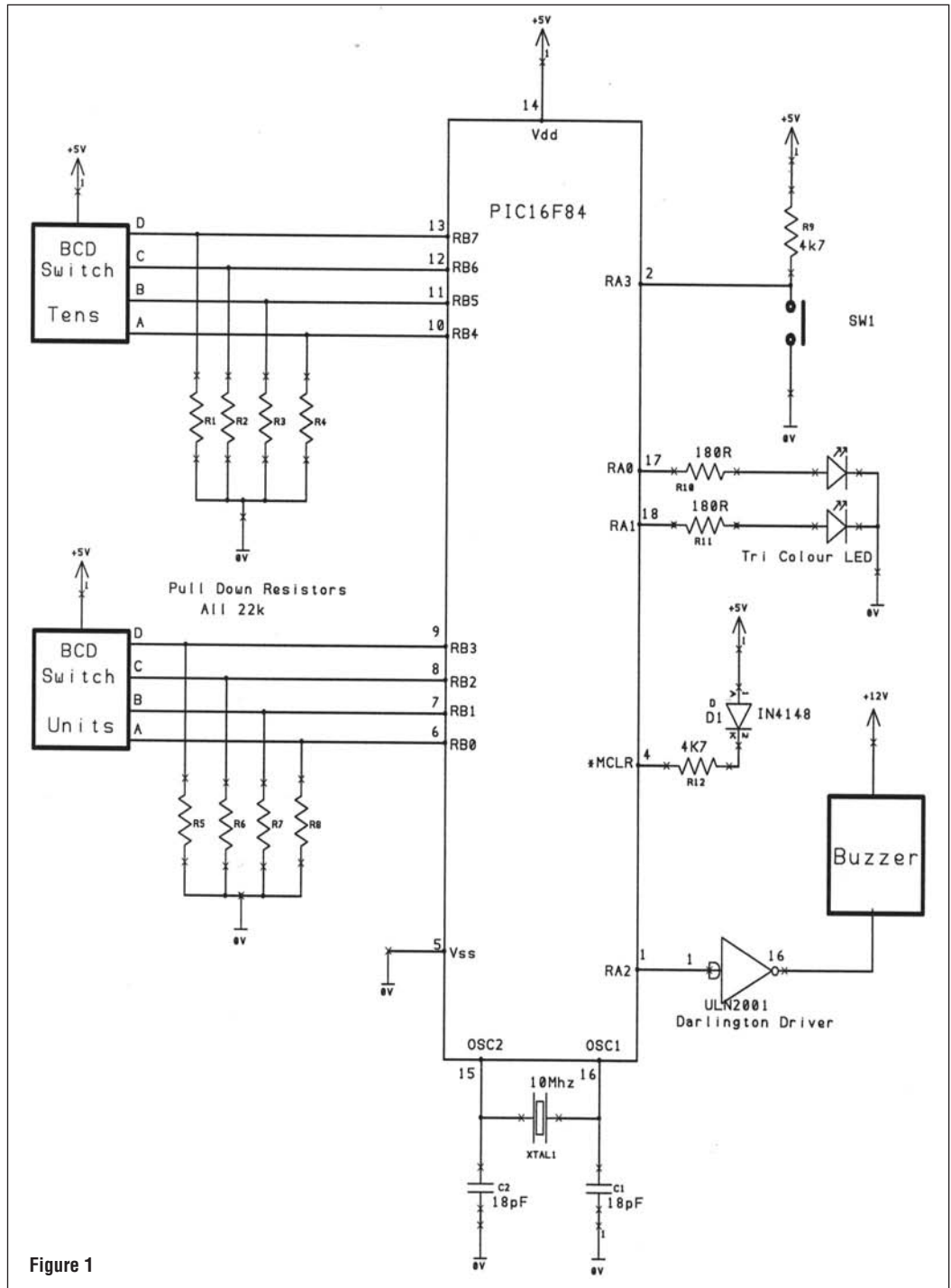


Figure 1

Part of a research study being conducted into athletic performance required athletes to perform shuttle runs between the front and back walls of a squash court. An adjustable timing system

was required, which could be altered by the researcher to provide a beep denoting the start time for each run. Since each of the athletes being tested was allocated a specific time interval between each

shuttle run, the timer had to be variable between 2.5 seconds and 4.5 seconds. Being unable to find a suitable off-the-shelf product to achieve this, a PIC-based metronome system that was

variable in 0.1-second steps between 0.3 seconds and 9.9 seconds was developed.

As can be seen from **Figure 1**, user input to the PIC 16F84 is achieved using a pair of Binary Coded Decimal (BCD) switches. A Darlington driver IC was used to drive the buzzer and a tricolour LED used to indicate the 'Stopped' and 'Running' states. The master clear (*MCLR) signal is pulled 'high' through diode D1 and resistor R12, to ensure proper power on reset. (1) Testing was initiated by depressing push-button switch SW1.

Software for the PIC is shown in **Code 1**. At power up, the buzzer is turned off

and the tricolour LED's red segment is turned on to indicate the 'Stopped' state. When the user presses and releases the pushbutton switch (SW1), BCD switch data is read into Port B (2), and the tri-colour LED's green segment is illuminated to indicate the 'Running' state. The entered BCD value is converted to an integer (bcd_value), representing the number of 100 millisecond delays required to realise the specified time delay between successive beeps.

At the end of each beep cycle, the PIC's output pin A2 is taken 'high', allowing the Darlington driver to sink current through the buzzer.

Since the duration of a beep is 200 milliseconds, then 'bcd_value' is reduced by 2. In the event of a user keying a value less than 3 (the minimum time delay possible), then 'bcd_value' is made equal to 3. The metronome then continues to produce beeps at the predetermined time interval set up on the BCD switches. The test protocol required time intervals to be alterable 'on the fly' and push-button switch SW1 was also used for this purpose. Setting a new value on the BCD switches and holding down SW1 until the end of a beep cycle allows a new value to be read in. The red segment of the LED will illu-

minate when the new value is read in, and when the tester releases the switch, the LED will return to the green 'Running' state.

The BCD switches used in the prototype are available from Rapid Electronics, part number 78-0805.

Frank Kelly
UK

References

<http://ww1.microchip.com/downloads/en/DeviceDoc/30430c.pdf>

Custom Computer Services, Inc, PCB, PCM, and PCW PIC Compiler Reference Manual. (1999)

```

/* Metron.cpp */
/* Program for metronome. */
/* Program reads in 2 BCD digits from a BCD switch and packs */
/* them into a single variable (bcd_value) between 0 and 99. */
/* The bcd_value variable is then used to calculate the */
/* number of times that a 100ms delay is executed. */
/* Frank Kelly 23rd September 2005 */
#include <16F84.H>
#define PORTB = 6

#define HS,NOWDT,NOPROTECT

#define use delay(clock=1000000)
#define ALL_IN 0xff /* Control value for Port B */
#define MIN_TIME 3 /* Minimum time value possible */
#define BEEP_TIME 2 /* Time duration of the buzzer */
#define LO_MASK 15 /* Mask value to read lower nibble */
#define HI_MASK 240 /* Mask value to read upper nibble */

int upper; /* The upper BCD value */
int lower; /* The lower BCD value */
int bcd_value; /* The packed BCD value */
int count; /* Number of delays required */

/* Function to turn on the Red LED */
void RED_LED()
{
    output_high(PIN_A0); /* Turn on red segment.. */
    output_low(PIN_A1); /* ...and turn off green segment */
}

/* Function to turn on the Green LED */
void GREEN_LED()
{
    output_high(PIN_A1); /* Turn on green segment.. */
    output_low(PIN_A0); /* ...and turn off red segment */
}

/* Read in the BCD switches as an 8 bit binary value and convert */
/* it to its corresponding integer value */
void get_BCD()
{
    lower = PORTB; /* Read port B into lower... */
    upper = PORTB; /* ....and also into upper */

    lower = (lower & LO_MASK); /* lower contains lower nibble */
    upper = (upper & HI_MASK); /* upper contains upper nibble */
    swap(upper); /* move upper into bottom 4 bits */
    upper = (upper*10); /* multiply * 10 to represent 'tens' */
    bcd_value = (lower + upper); /* ..then add lower */
    if(bcd_value < MIN_TIME) /* If user inputs value less than 3 */
        bcd_value = MIN_TIME; /* ..make bcd value equal to 3 */
}

void switch_press()
{
    while(input(PIN_A3)) /* Wait for switch press */
        delay_ms(5); /* Delay for debouncing */
    while(!input(PIN_A3))
        delay_ms(5);
}

main()
{
    output_low(PIN_A2); /* Ensure that buzzer is off */
    set_tris_b(ALL_IN); /* Make all port B inputs */
    while(1)
    {
        RED_LED(); /* Turn on the red LED */
        switch_press(); /* Wait for the user to operate switch */
        get_BCD(); /* ..then get the BCD value */
        GREEN_LED(); /* Turn on the green LED */
        while(input(PIN_A3)) /* While switch not pressed */
        {
            count = bcd_value; /* Turn buzzer on */
            output_high(PIN_A2); /* Turn buzzer on */
            delay_ms(200); /* Turn buzzer off */
            output_low(PIN_A2); /* Turn buzzer off */
            count = (count - BEEP_TIME); /* Subtract buzzer time */

            while(count)
            {
                delay_ms(100); /* Count down number of delays */
                count--;
            }
        }
    }
}

```

Code 1

The block diagram of the unit shown in **Figure 1** depicts the overall arrangement of this system. Two laser torches are installed at the required points on one side of the path and their corresponding LDRs are placed on the other side of the path, such that the laser beams from the torches fall directly on the LDRs. This forms a



to an astable multivibrator such that it produces pulses spaced at 1 ms only, while the output of bistable is high. These pulses of astable multivibrator given to a counter come seven-segment display

The working of the circuit is as follows: Initially, when the laser



beams fall on the LDRs without interruption, this causes their resistances to move to low values, which compels the transistors of the receiving circuits to go into saturation, making the output of these circuits zero. However, when a vehicle interrupts the laser beam, the bias point of the circuit gets disturbed, forcing the transistor into cut-off mode, thus making the output of the circuit go high. The variable resistors R1 and R2 are used to calibrate the device for various intensi-

ties of surrounding light.

As per the above discussion, whenever a vehicle interrupts the first beam, a pulse is produced by first receiver circuit, which gets applied to the J-K flip-flop (74LS109) in the toggle mode as the first trigger causes it to go high. After crossing the first beam, the object interrupts the second beam and the trigger from the second receiver circuit causes the flip-flop to toggle and go low. Thus a vehicle passing through the two points in time T activates and deactivates the flip-flop. The

flip-flop produces a pulse of duration equal to the time taken (T) by the vehicle to cross the two given points, separated by a desired distance.

Since the output of the flip-flop is connected to the reset pin of IC555 in astable mode, in this interval of time T, when the output of flip-flop is high, the 555 produces pulses spaced at 1ms. The counter (CD4033) counts these pulses and the counter is incremented every 1ms. The time T taken by the vehicle to cross the required distance is displayed

on the seven-segment display in milliseconds.

Application

The circuit can find many applications. For example, it can be used to determine the speed of a vehicle or to determine the time required by an athlete to complete a race. With only a few modifications, the same circuit can be used as a stopwatch too.

Dhirender Singh
Hyderabad
India

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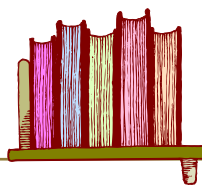
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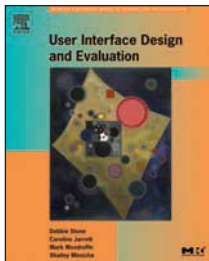
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User Interface Design and Evaluation

Debbie Stone, Caroline Jarrett, Mark Woodroffe and Shailey Minocha

Newnes (Elsevier)



The blurb states that “Whether you are a professional new to the user-centered design field, or an experienced designer who needs to learn the fundamentals of user interface design and evaluation, this book can lead the way”. The preface states, “After studying this book, you will be able to develop an UI in a flexible, iterative manner, working in close collaboration with the users.”

When this book dropped with a promising thud on to my doormat, the blurbs promised much. Sadly, I was very disappointed. Why?

First, there was the promise of a “supporting website”. Good, I thought, let’s go online and I can refer to it throughout. Sadly, the URL is not in the book.

I did a search for the publisher’s URL, but had to go through five more links before I found, not a “supporting website” but a “companion website”. I found that the documents available are intended only for lecturers and students taking the Open University course (in the UK); hardly a promising start.

A plus is the ‘flick through’ page numbering. A minus is that those colours are not repeated in the Contents for colour-coded access when flicking through. Another minus is the placing of the page numbers, which results in each left page having one wide margin. Thus, the small right margin rolls the text into the spine, which does not make it very readable.

I am sure that if the over-the-top margins were reduced it would be much lighter and easier to handle. Severe pruning would also help there, as well as clarify the ideas.

Generally, I found the book plodding and without any relieving sparks of humour. Maybe, the general problem is with having multiple authors. The plodding nature of the book is made worse by their habit of giving introductory sec-

tions in every Part and Chapter, with “what we have learnt” summaries at the end of even the shortest section. There are also many distracting little boxes alongside the text to tell you “You will learn more about X in Chapter Y”.

Overall, I consider the content to be glib and facile. Glib, as it revels in jargon words, facile as it uses them unnecessarily and even badly. One basic example is the use of the word “affordance”. They use it unqualified, but Norman (1988) states that it should always be qualified with either “perceived” or “real”.

Another example is the Exercise 2.2. It uses a photo that is so badly out of focus that the question to be answered is meaningless.

The “Discussion” is equally meaningless as no explanation is given as to why the users acted in the way the authors state.

I also feel that the Preface and blurbs should be rewritten to make it clear that the book is not really intended for reference, but just to accompany the Open University course.

Once I understood in which direction the authors were heading, I browsed for highlights. I immediately found two glaring omissions regarding input/output UIs. A very brief mention is made that there is a subject called ergonomics, but nothing to make the reader appreciate how important it is. There is no discussion of the ergonomics of UIs anywhere in the book and there’s no mention of kinesthetics either. Thus, very important design criteria are effectively ignored. (See the US Department of Labour check list for example.)

Likewise, although analysis of UI evaluations is of the utmost importance, even in the so-called case studies, the use and importance of statistical discipline in assessing evaluations is minimised.

In the section on input devices, most emphasis is placed on input to computers. The expected sketches of mobile phones and games consoles are there, but the extremely poor design of most, where the user is encouraged, or even forced, to use the thumb(s), is not even mentioned. Ignoring ergonomics again.

Throughout, I got a definite feeling of academics writing for academics to teach students to be academics. Three of the authors are academics and the fourth runs courses and tutors for the Open University.

Finally – the case studies. They should have given readers a clear insight into the necessary procedures and the end results, but are so cursory as to make very little contribution. In the only two real cases presented, the authors were not part of the teams working on the separate studies and implementations. The text is simply interviews with some of the participants giving an inadequate summary of the work done.

In all four, they aren’t any case studies (“a careful study/a detailed analysis”).

Tokairo – the authors do not even give the reason as to why the redesign was considered to be necessary. Photos show the final design – the ‘kiosk’ – in place, but the authors do not comment on the unfortunate placing that shows how very awkward it must be for a right handed user to input data.

Fast – the early background at LATCC was firmly based on work study and engineering. This is not considered.

So, to an Electronics World reader I recommend to buy a good book on work study, with side helpings of job evaluation and website design. If you buy this book you will find a few useful nuggets of information, but as the engineering input is so sparse, you won’t find many facts or data.

Anthony C Batchelor

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EWAdmin@nexusmedia.com

Move if you want to earn more

Your editorial (November issue, entitled 'Learning through light exercise') throws light on the state of engineering.

You write: "Engineering is still not as well paid as some other professions and certainly it does not command as high a status in society". How little things have changed in 25 years. In 1980 I had the chance to move to continental Europe and set out to better my career chances. It was the best move I could have made. I can easily imagine my position now if I had remained: a £30,000 job in a small department, a terraced house, a large mortgage, a second-hand car and worries for the future.

Many times I have been asked why I enjoy my position in comparison to UK-based engineers and I, basically, illustrate it with a summary of the attitude in the UK to engineering and engineers: "Doctor? Ah, do come in. Let me take your coat. Judge, welcome, please make yourself comfortable. Engineer? Your entrance is down the side here, take off your boots and wash your hands first!"

This attitude is, of course, to be expected, since in the UK, factory workers and other labourers who tighten wheel-nuts on car production lines are also called "engineers" and are best known for taking "industrial (in)action".

Until this situation changes and engineers are recognised as the professional group they are, fewer students will opt for engineering courses and of those who succeed many will leave the UK to earn the respect they deserve – just as I did a quarter of a century ago.

F.C. Trevor Gale
Netherlands



Mind the LHB that gives electric shocks

A month ago, I installed a (non-Sky) satellite system. Everything was going fine until I was fixing the dish to the wall of my garden. I was fixing the 'F' coupling to the LHB when I started getting electric shocks; not enough to cause damage but enough to make me sore. Using a Fluke 77, I measured an 82V ac. This made me check the masthead amplifier on the new wideband DTT aerial. It measured 79V ac, but as I was on the roof of my house and not on the ground I did not get an electric shock. Both units have the CE stamp.

In the 1950s, the regulations required that radios fed from a mains dropper resistor had to have no metal showing, plastic knobs had to cover the steel shafts, and aerial and earth sockets were fed via a small value capacitor.

Fifty years later we seem to be going backwards. I realise that SMPS have superseded transformer analogue supplies due to new design engineering, small transformers over heat and cause unit failures.

I would like to hear any thoughts on this subject.

D. Lucas
UK

Problem solving challenge

I want to be able to watch the weather forecasts on board my small trail-sail yacht using the 12V supply. I was told (Maplin) this combination would work through the aerial feeds, but it doesn't. When I phoned Digifusion they said it needed a Scart socket connection, but they didn't sound that convincing. The call engineer did suggest it might work through a VCR with the last connection via the aerial lead only, but space and power are very limited. However, Maplin said it should work using a Scart lead with two AV plugs that would plug into the AV lead supplied with the TV.

I tried using a Scart lead from the digital box that had the red, yellow and white video and audio plugs leading from it. I plugged this into the TV using the AV lead supplied with matching sockets, but did not have any success.

Not wishing to thrash around in the dark or to damage the equipment, can you throw any light on the matter?

I am trying to receive digital Freeview on a Saka 5" TFT LCD colour, portable, television (12V), in conjunction with a Digifusion box which has the following:



Saka TV

Socket for external aerial

AV in

AV out

No Scart

C. SYS. Auto, PAL, P3.58, N3.58, N4.43, SECAM, SECAM-L, PALM, PALN

S.SYS. M, BG, I, DK

The Freeview box is a Digifusion FRT 101 (12V)

RF in

RF out

AV Scart to VCR

AV Scart to TV

If this setup is not possible, are there either past or present boxes that will work without a Scart socket on the TV? Or could I just use the relevant circuit from an old VCR?

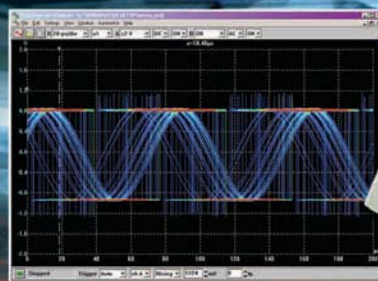
Ian Anderson
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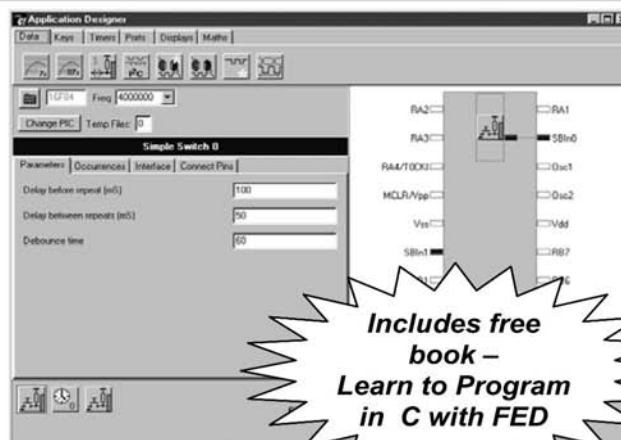
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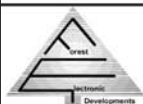
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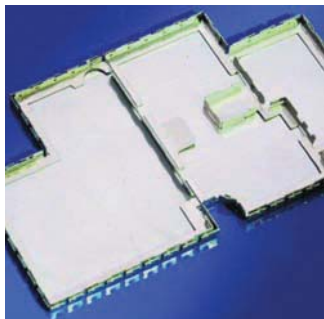
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RoHS compliant shielding is more economical

Nickel Silver is fast becoming the preferred alloy for the majority of board level shielding tasks according to Precision Micro, the Birmingham-based precision metal components manufacturer. The material is said to be replacing traditional plated steel and yellow metals across the board and certainly for those applications above 15MHz, where shielding effectiveness becomes more of a skin (conductivity) effect than a bulk effect.

Nickel Silver is less conductive than copper and brass but far stronger. It is an alloy of copper, zinc and nickel, the 'silver' terminology being descriptive of its appearance rather than its constituents. Nickel Silver is far more corrosion resistant than either steel copper or brass and exhibits excellent solderability characteristics for ease of assembly.

Although this RoHS compliant material is marginally more expensive, it doesn't need plating to improve its solderability or corrosion resistance, so invariably finished shielding cans manufactured in Nickel Silver are less expensive. There is also a time saving, generated by removing a complete process stage, which can be important when 'time to market' is critical.

www.precisionmicro.com

Development tool for a variety of 8051 MCUs

Keil recently announced the availability of its new μ Vision3 development environment and C51 version 8 optimising compiler for classic and extended 8051 microcontroller (MCU) development. The firms says that together these products provide developers with a feature-rich, integrated development environment, optimised for a wide range of 8051 MCU variants released in the past nine months, including the Analog Devices ADuC83x and ADuC84x and the Infineon XC866.

The μ Vision3 development environment features source code outlining, function navigator, template editing and incremental search. It also incorporates a configuration wizard, which speeds the generation of start-up code and configuration files. The built-in microcontroller simulator can model supported MCU devices, including the instruction set, on-chip peripherals and external signals used to stimulate it. Application behaviour can be captured using the μ Vision3 logic analyser, showing the changing state of MCU pins and peripherals alongside program variables.

Customers with valid software maintenance agreements can download the Keil C51 version 8 compiler and μ Vision3 update from the Keil web site.

www.keil.com

CompactPCI power supply

C&D Technologies has expanded its family of CompactPCI power supplies with a 3U x 8HP 325W module that combines fully featured IPMI (Intelligent Platform Management Interface) functionality with the ability to operate at significantly lower airflows.

The new CPCI325D-1 delivers full power across its operating temperature range of 0°C to +50°C with as little as 200l/m of airflow. Built-in IPMI functionality includes the basic FRU reporting capabilities



offered by other CompactPCI modules plus enhanced functionality for reporting output current, output voltage and temperature. An output inhibit function is also provided as standard.

The CPCI325D-1 is designed for compliance with the PICMG 2.11 D1.0 CompactPCI Power Interface Specification. Features such as multiphase conversion, synchronous rectification and output ORing FETs have been

used to achieve high efficiency and the lowest airflow requirements of any CompactPCI power supply on the market.

Operating from a wide 36-72VDC input voltage range, this 325W supply supports hot-swap, N+1 redundant operation for globally deployed, high-availability (HA) applications.

As a result, the module is ideally suited to the broad range of CompactPCI applications in telecommunications, networking, industrial computing, test and measurement and industrial equipment.

www.cd4power.com

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Quadrant, the Hertfordshire-based component distributor and cable assembly manufacturer, has made a major addition to its cable assembly capability by offering custom over-moulding of cable assemblies using 'D' type sub-miniature and other connectors.

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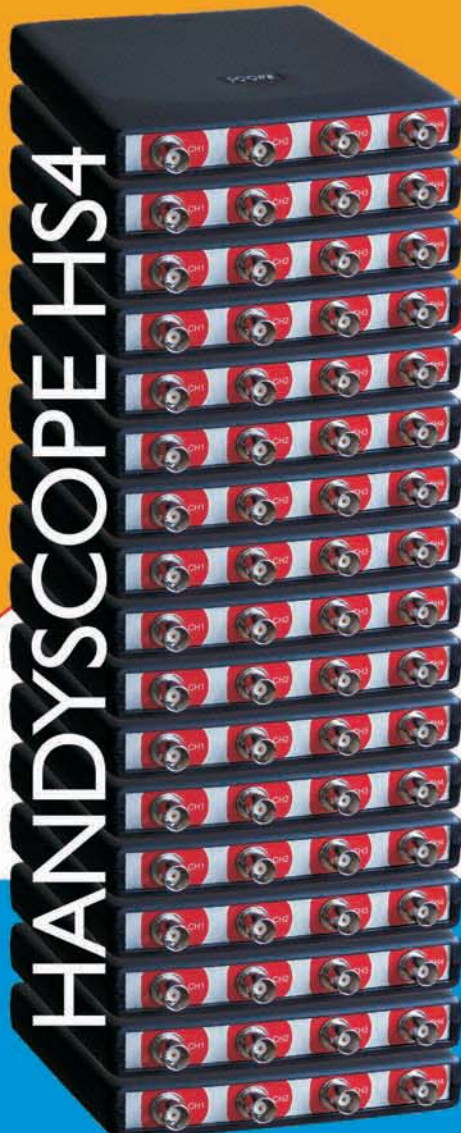
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Agilent/HP 8447F 1.3GHz Pre/Power Dual Amplifier	1250	50	Keithley 238 High Current Source Measure Unit	3350	139	Agilent/HP 8115A 50MHz Dual Channel Pulse Generator	2350	95
Agilent/HP 8449B 26.5GHz 26dB +7dBm Pre-amplifier	4500	180	Keithley 2400 200V Digital Sourceceter	2850	86	SIGNAL & SPECTRUM ANALYSERS		
Anritsu MH648A 100kHz-1.2GHz Preamplifier	650	33	Keithley 2410 1100V High Voltage Sourceceter	4150	125	Advantest R3261A 2.6GHz Spectrum Analyser	3450	104
ENI 2100L 10kHz-12MHz 100W Power Amplifier	6500	260	Kikusui DME1400 4.5 Digit Multimeter	175	18	Advantest R3265 8GHz Spectrum Analyser	5975	247
ENI 240L 20kHz-10MHz 40W RF Amplifier	3950	158	Thandar 1604 4.75 Digit Digital Multimeter	395	22	Advantest R9211A 10MHz-100kHz Dual Channel FFT Analyser	2950	89
ENI 601L 800kHz-1GHz 1.2W RF Amplifier	850	36	NETWORK ANALYSERS			Advantest U3641 3GHz RF Spectrum Analyser	3500	107
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Agilent/HP 5316A 100MHz Frequency Counter	295	18	Agilent/HP 3575A Gain/Phase Meter	1150	59	Agilent/HP 3562A 100kHz Dual Ch Dynamic Signal Analyser	2450	74
Agilent/HP 5334A/030 1.3GHz Frequency Counter	495	20	Agilent/HP 3577A 5Hz-200MHz Vector Network Analyser	3950	119	Agilent/HP 3588A/001 150MHz Spectrum Analyser	4650	140
Agilent/HP 5342A/011/002 24GHz Frequency Counter	1150	42	Agilent/HP 4195A 500MHz Vector Network/Spectrum Analyser	6950	209	Agilent/HP 53310A/001/030 2.5GHz Mod Domain Analyser	3850	160
Agilent/HP 5347A/006 20GHz Counter/Power Meter	4050	162	Agilent/HP 85032B/K05 Type N (f) Calibration Kit	750	28	Agilent/HP 8563E/006/007/008 30Hz-26.5GHz Spec Ana	14950	450
Agilent/HP 5371A 500MHz Frequency/Time Interval Analyser	1650	50	Agilent/HP 8510B 50GHz Microwave Network Analyser	4250	170	Agilent/HP 8566B 22GHz Spectrum Analyser	6950	210
Agilent/HP 5372A 500MHz Frequency/Time Interval Analyser	2575	78				Agilent/HP 8568B 100Hz-1.5GHz Spectrum Analyser	2950	89
EIP 578 26GHz Microwave Source Locking Counter	1850	75				Agilent/HP 8591A/010/021 1.8GHz Spectrum Ana With TG	3750	113
Marconi 2440 20GHz Microwave Counter	1550	47				Agilent/HP 8593E/004/041/130 22GHz Spectrum Analyser	10950	329
Philips PM6670/01 120MHz Frequency Counter Timer	495	30				Agilent/HP 8594E/041 2.9GHz Spectrum Analyser	4500	135
Racal 1991/04A 160MHz Frequency Counter	395	32	Agilent/HP 8714C 300kHz-3GHz Vector Network Analyser	6950	209	Agilent/HP 8595E 6.5GHz Spectrum Analyser	7950	317
Racal 1992/001 1.3GHz Frequency Counter	575	30	Agilent/HP 8720B/001 20GHz Vector Network Analyser	19950	802	Agilent/HP 8596E/021/101/105 12.8GHz Spectrum Analyser	8550	260
Racal 1992/04C 1.3GHz Frequency Counter	575	30	Agilent/HP 8720C 20GHz Vector Network Analyser	23950	961	Agilent/HP 8901A/001/002 1.3GHz Modulation Analyser	950	48
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Agilent/HP 3314A 20MHz Function Generator	925	30	Fluke 80i-1000S AC Current Probe	75	8	Agilent/HP 83711A/1E1 1-20GHz Synthesised CW Sig Gen	7650	230
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Agilent/HP 1670G 500MHz Timing 150MHz State 136Ch	4650	140				W&G 4106 GSM/DCS1800/PCN1900 Mobile Phone Tester	2250	70
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