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Bodo´s Power Systems®

Electronics in Motion and Conversion

October 2016

Over-Voltage Protection for RS-485 Bus Nodes

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Power Electronics


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CONTENT


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5MPF FEATURES
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It´s Show Time Again!

September started with EPE ECCE in Karlsruhe, Germany and ECCE in Milwaukee, USA.

In October there'll be numerous events focusing on power electronic and related subjects, as you'll see here on page 4. In November we'll all come together again, for almost a full week, at the bi-annual electronica in Munich. While not all power electronic specialists will be exhibiting, it is nevertheless the biggest show in Europe, and attracts people from all over the globe. Two weeks later, in Nuremberg, the SPS IPC Drives will start. November will be busy!

So what's been happening in our industry? Infineon successfully integrated IR and recently acquired Wolfspeed. So they now have strong expertise in wide band gap devices. ON Semiconductor own Fairchild now. Analog Devices bought Linear Technology. It was just a matter of time befor Linear Technology ended up at one of the bigger players. Hopefully together they will develop a good strategy for efficient power management. Renesas has started to buy Intersil. Intersil started with Harris Semiconductor analog designs for power control, multimedia and wireless. The Intersil-Zilker knowledge in digital power management set the stage for strong growth. Let's stay tuned to see how Renesas integrates Intersil into their strategy. It will be nice to see them all at electronica and to understanding their many developmental strategies.

For the engineers amongst us, it will be great to see the progress in semiconductor development. We look forward to passive components improving to better match wide

band gap semiconductors. Higher switching speeds and higher temperatures are two challenges for passives. Coils, transformers and capacitors need to improve, and we need better packaging for WBGs to allow for higher operating temperatures.

Bodo's Power Systems reaches readers across the globe. If you are using any kind of tablet or smart phone, you will now find all of our content on the new web-site www.eepower.com. If you speak the language, or just want to have a look, don't miss our Chinese version: www.bodospowerschina.com

My Green Power Tip for October:
We're looking forward to the Thanksgiving Holiday in October in Canada and November in the USA. Let us celebrate the spirit of these holidays to bring us together, and put aside the crazy, polarizing, and dangerous statements that receive so much publicity these days.

Best Regards

Events

CWIEME Chicago,
MI USA; October 4-6
http://www.coilwindingexpo.com/chicago

INTELEC 2016,
Austin TX, October 23-27
http://www.intelec.org/

SEMICON Europa 2106,
Grenoble, France, October 25-27
http://www.semiconeuropa.org/

Power Electronics 2016
Moscow, Russia, October 25-27
http://expoelectronica.primexpo.ru/en/

ESARS-ITEC 2016,
Toulouse, France November 2-4
http://www.esars-itec.org

Power Integrity Seminar 2016,
Munich Germany, November 3,
https://www.omicron-lab.com/trainings-events/1-day-workshop-with-steve-sandler-new-ways-to-vrm-modelling.html

electronica 2016,
Munich, Germany, November 8-11
http://electronica.de

sps ipc drives 2016,
Nuremberg, Germany, November 22-26
http://www.mesago.de/de/SPS/home.htm

Power Electronics Conference "The Power Awakens"

The conference offers a unique possibility to get an overview of important areas of power electronics, including applications, advanced devices, reliability, and testing.
As part of SEMICON Europa, the conference offers the possibility to establish connections among device designers/manufacturers, equipment suppliers, and system integrators, all while discovering the latest processes, materials and integration methods. The power awakens in Grenoble as one of the emerging hot-spots of power-semiconductor research in Europe.

Target Audience
Device designers, power-semiconductor material specialists, designers of power electronic systems, process specialists, foundry service providers and others interested in the growing field of power electronics.
SEMICON Europa takes place Oct. 25-27, 2016 Alpexpo, Grenoble, France

http://www.semiconeuropa.org/

APEC 2017 Sponsors Continue Student Travel Support Program

The joint sponsors of the Applied Power Electronics Conference have announced the continuation of the popular Student Attendance Travel Support Program of up to $1,000 to cover part of the travel and conference expenses for as many as 40 students to attend APEC 2017 in Tampa, FL, March 26-30, 2017. Up to $40,000 in travel support will be distributed.
Now in its 12th year, this popular program, initiated by the Power Sources Manufacturers Association (PSMA), is now jointly underwritten by PSMA and the other co-sponsors of the APEC Conference – the IEEE Power Electronics Society (PELS) and the IEEE Industry Applications society (IAS).
The recipients will be chosen by the APEC 2017 Student Travel Support Committee. Application forms are available at APEC Attendance Travel Support Application.
Applications must be made and received by the Committee by October 21, 2016.
The recipients will be notified by the Committee by November 21, 2016.

http://www.apec-conf.org/

EpiGaN Nominated for EU Innovation Radar Prize 2016

EpiGaN, the leadingEuropean supplier of commercial-grade 150mm- and 200mm- GaN-on-Silicon epi-wafers for 600V HEMT (High Electron Mobility Transistor) power semiconductors, was shortlisted along with 10 top-ranking European digital innovators participating in EU-funded projects such as Horizon 2020 to compete for the EU Innovation Radar Prize 2016 in the category "Industrial and Enabling Tech". Public voting for the winner is open through the month of August.
EpiGaN offers top performing innovative GaN-on-Si and GaN-on-SiC material solutions that are enabling the next generation of efficient power electronics, RF power and sensor devices and systems. GaN (gallium nitride) technology allows for drastic energy loss savings, more compact and lighter power conversion systems, such as power supplies, and is also in demand for its superior performance in wireless communication.
The Innovation Radar has been established recently as a data-driven initiative by the EU Commission to identify high-potential European digital innovations and the key drivers behind them participating in its FP7, CIP, and Horizon 2020 projects. The EU Innovation Radar supports and guides digital innovators in assessing and highlighting their innovative and economic potentials in the world market. EpiGaN was nominated because of its development of large diameters GaN-on-Si epiwafer that enable a new generation of power electronics with much higher conversion efficiencies.
EpiGaN is developing and manufacturing AlGaN/GaN structures on 200mm Si substrates at the 600V node to enable its customers to successfully position themselves in rapidly growing market segments, addressing the power switching market, including power supply, solar inverter, power supply for data centers, and next, for electric vehicles.

www.epigan.com

Alpha's Ravi Bhatkal Elected to iNEMI Global Board of Directors for a Second Term

Alpha Assembly Solutions, the world leader in the production of electronic soldering and bonding materials, announces the election of their Vice President of Energy Technologies, Ravi M. Bhatkal, Ph.D., to the iNEMI Global Board of Directors for a second term.
The iNEMI organization roadmaps the future technology requirements of the global electronics industry, identifies and prioritizes technology and infrastructure gaps, and helps eliminate those gaps through timely, high-impact deployment projects.
iNEMI Board of Directors provide direction on the policy, strategy and direction of the consortium. "It is an honor and privilege to serve on the iNEMI Board of Directors for a consecutive term", said Dr. Bhatkal, of Alpha Assembly Solutions, part of the MacDermid Performance Solutions group of businesses. "iNEMI is unique in its scope and approach, and adds significant value to the electronics industry through its activities. I am delighted to do my part to help advance the industry".
Dr. Bhatkal was elected by the Council of Members and will serve another three-year term. Currently, the Board is made up of 12 voting directors who are executives from a cross-section of member companies (OEMs, EMS providers, ODMs and suppliers).
As Vice President of Energy Technologies for Alpha Assembly Solutions, Dr. Bhatkal has responsibility for new business creation in the alternative energy and energy efficiency value chains, including the identification and evaluation of new opportunities.

http://alphaassembly.com/

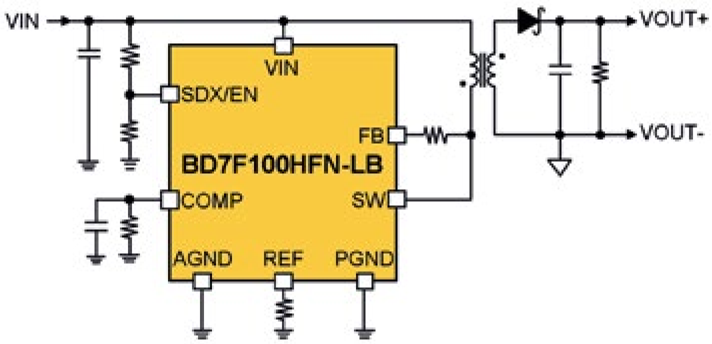
ISOLATED FLYBACK CONVERTERS

High Security | High Reliability | Compact Solution

BD7F100HFN-LB is an optocoupler-less isolated Flyback DCDC converter with primary side control. It uses ROHM's State Of Art Adapted-Type ON-Time Control Technology to control the output voltage. No additional parts to keep the loop stable are required.

Key Features

- No need for external isolation
- Adaptive ON-Time Control technology
- Built-in N-Channel MOSFET
- High-speed load response
- High efficiency at light load mode
- Enable and Soft-Start functions
- Output current compensation function



Specifications

- Supply Voltage: 3V to 40V
- SW Terminal Operating Voltage: 50V (Max)
- Built-in MOSFET Current Limit: 1.25A (Typ)
- Switching Frequency of 400kHz (Typ)
- ±1.5% of Reference Voltage Accuracy
- Quiescent Current: 0μA (Typ)
- Operating Ambient Temperature: -40°C to +85°C
- Complete protection functions UVLO, OCP, TSD
- Package HSON8IC

Applications

Isolated power supply for ndustrial application
Gate driver circuit secondary power supply
Power supply for optocoupler, digital isolations
Isolated interface drivers (RS-485/232, CAN) power supply

Visit us!
electronica 2016
Hall A5, Stand 542
Munich, 8.-11. November

Announcing Global Agreement

Richardson RFPD, Inc. announced a global agreement with Power Integrations, Inc. Under the terms of the agreement, Richardson RFPD will distribute Power Integrations' SCALE™ IGBT drivers for the high-power market, as well as its new SCALE-iDriver™ integrated circuits, on a worldwide basis.

Power Integrations' family of SCALE gate drivers reduce component count, enhance efficiency and improve reliability. They are suitable for a range of clean-power applications from 30 kW to 1 GW, including high-efficiency industrial motors, renewable energy, electric transportation, and DC transmission.

The new family of SCALE-iDriver ICs are galvanically-isolated single-channel gate driver ICs that range in output current from 2.5 A to 8 A, the industry's highest output current without an external booster.

www.power.com

Featuring Material Set Combinations for Greatest Reliability

Alpha Assembly Solutions, the world leader in the production of electronic soldering and bonding materials, will feature material set combinations to include paste, tacky flux, cored wire and wave soldering flux that are tested with one another to achieve greater reliability at the SMTA Guadalajara Expo to be held October 5th-6th in Guadalajara, Jal. MÉXICO.

"Assembly manufacturers spend a lot of time trying to find the best material set for their assembly process," said Zeus Moncivais, Regional Sales Manager for Alpha Assembly Solutions - Mexico, part of the MacDermid Performance Solutions group of businesses. "Alpha has done the testing and can provide proven material pairings that will produce the greatest reliability, thus saving our customers a great deal of trial and error while increasing their throughput and time-to-market."

Located at Booth #107, Alpha will also showcase their vast product portfolio of innovative materials and solutions for the electronics assembly industry.

www.AlphaAssembly.com

To Acquire Select Product Portfolio

Littelfuse, Inc., announced it has entered into definitive agreements to acquire the product portfolio of transient voltage suppression («TVS») diodes, switching thyristors and insulated gate bipolar transistors ("IGBT") for automotive ignition applications from ON Semiconductor Corporation for a combined purchase price of $104 million. This portfolio has annualized sales of approximately $55 million. The transactions are expected to close in late August, 2016.

Littelfuse also plans to invest approximately $30 million in its semiconductor fabrication locations to enhance its production capabilities, add significant capacity to its China fabrication facility and transfer the production of the acquired portfolio. The transfers will occur over the next few years, as the company works with customers on their timing and requirements. The expected productivity gains from this investment will drive long term profitable growth across the company's semiconductor business.

http://www.littelfuse.com/

Opening Global R&D Centre

Knowles (UK), formally Syfer Technology, have recently made the short step to new hi-tech facilities at the Hethel Engineering Centre, Norwich. The move reflects Knowles' commitment to R&D for its portfolio of ceramic based electronic components manufactured in Far East and North American facilities.

Knowles (UK) R&D facility has a traceable history back to the 1940's under Erie Electronics Ltd, before becoming Syfer Technology in the 80's. The company was later acquired by the US Dover Corporation and is now part of Knowles, a recent spin-off from Dover. The spin-off also included the DLI, Novacap and Voltronics brands manufacturing Capacitor products, EMI filters and EMC solutions that span high reliability, military specification and space level through to volume commercial products. Over the years the R&D facility has led the industry with leading technical advances producing products of the highest

quality, utilising superior materials that made Syfer a global leader. Today the remit is to support all four operating brands in developing new products.

www.knowlescapacitors.com

Renesas to Acquire Intersil

Renesas Electronics Corporation and Intersil Corporation announced they have signed a definitive agreement for Renesas to acquire Intersil for US$22.50 per share in cash, representing an aggregate equity value of approximately US$3.2 billion. The transaction has been unanimously approved by the boards of directors of both companies. Closing of the transaction is expected in the first half of 2017, following approval by Intersil shareholders and the relevant governmental authorities.

Renesas anticipates that near- and long-term revenue expansion opportunities combined with the modest anticipated cost efficiencies associated with greater scale will eventually generate synergies of US$170 million. The transaction is expected to immediately increase both gross and operating margins and be accretive to Renesas' non-GAAP earnings per share and free cash flows after closing

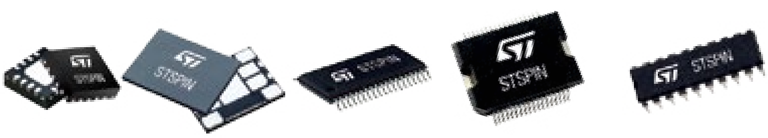
https://www.renesas.com

http://www.intersil.com

STSPIN™ motor drivers:
Leading in integration, performance and efficiency

- Scalable and flexible for all motor types with wide operating voltage, current and temperature range
- Extensive diagnostics and fully-protected to reduce the number of external components, cost and complexity
- Up to 256 micro-steps per step for unprecedented smooth and silent motion
- Multiple package options to meet a variety of design challenges from board space to thermally harsh environments
- Comprehensive development environment for fast and easy design

Motor type	System partitioning	Typical power range (W)	Part number
Stepper	Controllers	20 - 1000	L6480, L6482, L6506, L297
	Drivers with integrated FETs	< 10	STSPIN220
		5 - 90	L6472, L6470, L6474, L6228, L6208
		40 - 600	POWERSTEP01
Brushed DC		< 10	STSPIN240
		5 - 90	L6227, L6226, L6225, L6207, L6206, L6205, L6203, L6201, L298, L293, L2293Q
BLDC		< 10	STSPIN230
		5 - 60	L6229, L6235, L6234, L6230



For further information
and full design support,
visit us at **www.st.com/stspin**

electronica Attracts Founders

Contacts to possible investors, sales support or legal advice: Start-ups that want to turn a good idea into a successful company face a number of challenges. From November 8–11, 2016, the electronica Fast Forward start-up platform offers founders extensive start-up assistance. Thanks to this event, the World's Leading Trade Fair for Electronic Components, Systems and Applications has its finger on the pulse of the times. During the first few weeks, several founders and aspiring entrepreneurs have registered to present their innovative ideas at electronica.

Whether a start-up already has a product and a business plan or a creative mind is still working on an idea is not important for them to participate in electronica Fast Forward. Founders and developers can submit their projects in the "Idea", "Prototype" and "Start-up" categories and have a chance to be part of the start-up initiative. A panel of judges evaluates the submissions and selects the best participants from all of the competitors, who then have a chance to present themselves to an international audience. A number of different start-ups from more than 17 countries including China, Russia, Australia and the United States have already applied with promising solutions. Because interest is so high, electronica has extended the application deadline to October 4, 2016.

www.electronica.de/en

New Ways to VRM Modelling

Achieve reliable system performance - Speed up design cycles - Reduce cost. Exclusive 1-day Simulation and Design Workshop with Steve Sandler. Steve is the founder and chief engineer of AEi Systems LLC and CEO of Picotest. He is an internationally renowned author and expert for power integrity and power supply design.
His knowledge is based on over 35 years' experience in the design and analysis of power conversion equipment for military and space applications.

Workshop Overview

The Voltage Regulator Module (VRM) is a fundamental element in the power distribution network (PDN) with system level consequences. Very few, if any, manufacturers provide high fidelity VRM models and much of the data required to construct your own model is not published.
This workshop introduces a measurement based, non-linear VRM model that simulates very quickly while supporting both, switching and AC characteristics. The model includes PSRR and input impedance, including negative resistance, as well as VRM output impedance and supports PCB influences via EM simulation.
In this compact one day workshop you will be guided through all measurement and simulation steps required to generate a measurement based VRM model. Each theory step is followed by practical measurements performed in small groups.
Thursday, November 3rd, 2016, 08:30 - 17:00
Microchip Application Center
Osterfeldstraße 82; 85737 Ismaning (Munich); Germany
Workshop price: € 249,- net including:
Workshop documentation; Snacks, Lunch & Drinks
Steve Sandler's book on Power Integrity (€ 90,- value)

Participants

This workshop targets experienced power electronics engineers that want to widen their knowledge about VRM modelling and power integrity.

Seat Limitation

Due to the combination of theory and hands-on demonstrations, the number of seats is limited. Seats will be awarded based on "first come first served" basis according to the reception of the payment.
Feel free to download the event flyer:
https://www.omicron-lab.com/trainings-events/1-day-workshop-with-steve-sandler-new-ways-to-vrm-modelling.html

www.omicron-lab.com

Founders Girvan Patterson and John Roberts to Retire

GaN Systems Founders Girvan Patterson (left) and John Roberts to Retire.

Ten years after launching a gallium nitride (GaN) semiconductor company, and leading the company to #1 in the world of GaN power transistors, the two Ottawa-based co-founders of GaN Systems, President Girvan Patterson and CTO John Roberts have announced their retirement. Having achieved their goal of building GaN Systems into the world leading manufacturer of GaN power transistors and supplier to more than 500 customers, serial entrepreneurs Patterson and Roberts will leave their operating roles. Mr. Patterson will retain a position on GaN Systems' Board of Directors, while Mr. Roberts will remain available to the company as an emeritus contributor.
Prior to starting GaN Systems, both Roberts and Patterson had extensive experience in the semiconductor industry, each with remarkable careers launching new companies. John Roberts was founder and president of two Ottawa-based semiconductor companies, Calmos (later Tundra) Semiconductor, and SiGe Semiconductor. Girvan Patterson founded graphics workstation producer Orcatech, and co-founded Plaintree Systems, a pioneer in Ethernet switching. Together, Messieurs Patterson and Roberts have taken a dozen companies either to an IPO or acquisition.

www.gansystems.com

Industrial and Automotive Innovation at electronica 2016

At electronica 2016 in Munich, Germany, Texas Instruments will showcase innovation focused on the future of industrial automation and automotive technologies, such as the following:

Visit the industrial demo showcasing TI technology enabling Industry 4.0 and the smart factory of tomorrow leveraging TI solutions for real-time Ethernet with EtherCAT. Take a closer look at sensor communications with wired technologies like IO-Link and wireless technologies such as Bluetooth® low energy, and explore how TI's gallium nitride (GaN) technology significantly increases the efficiency in variable speed motor drives.

Discover innovation in automotive all in one car. The Evolution Car 2020 (EvoCar) showcases how TI's automotive portfolio, including DLP® products, enables advanced driver assistance systems (ADAS), infotainment, haptic feedback and adaptive LED headlights. Come and take a ride.

www.ti.com/electronica

Free Seminar Full-day DC-DC Power Seminars throughout Europe

System architects and designers of power subsystems can learn about important considerations for efficiency, heat dissipation, size, protection, and isolation in a new full-day seminar hosted by Maxim Integrated Products, Inc. and co-sponsored by Maxim's distribution partner Avnet Silica.

With wide input voltages so often used in industrial, medical, communications, and other end applications, meeting power system performance and size goals is critical. Held September through October, these free, full-day seminars cover power system design from theoretical foundation to solving attendees' most challenging design problems.

Registration is open and seating is limited: http://tinyurl.com/Maxim-Integrated-Power-Seminar.

Agenda:

- Understanding the theory of wide-input non-isolated and isolated DC-DC switching regulators
- Designing with wide-input synchronous buck regulators and power modules
- Designing DC-DC isolated topologies that eliminate opto-couplers
- Practical sessions and demonstrations showing usage of simulation tools and system protection

www.maximintegrated.com

Win a 2D/3D Touch and Gesture Development Kit from Microchip

Bodo´s Power Systems is giving readers the chance to win a 2D/3D Touch and Gesture Development Kit (DV102014) from Microchip. This kit is worth $249.00!

The 2D/3D Touch and Gesture Development Kit the industry's first development kit for integrated 2D projective-capacitive touch (PCAP) and 3D gesture recognition on displays.

The kit will provide designers with easy access to Microchip's patented 2D and 3D GestIC® sensing technology, allowing them to easily integrate 2D multi-touch and 3D hand gesture recognition into their display applications. The use of electric-field based technology now enables hand and finger gestures to be tracked both on the display surface as well as above at a distance of up to 20 cm. In addition, the development kit provides an easy-to-use, 'out-of-the-box' experience that requires no code development. Parameterisation, diagnostics and optional settings are done through Aurea 2.0, a free downloadable graphical user interface (GUI).

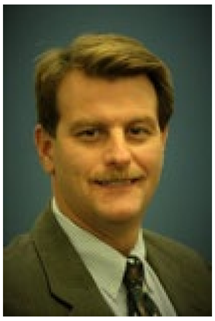
The 2D/3D Touch and Gesture Development Kit features Microchip's latest PCAP controller, the MTCH6303, with the MGC3130 3D gesture controller. It includes an eight-inch transparent touch sensor to enable rapid prototyping for widely available displays. The MTCH6303 provides multi-touch coordinates with a five-finger scan rate of 100 Hz. In addition, it has an integrated multi-finger surface gesture suite which makes it a good fit both for Operating System (OS) driven applications as well as embedded systems without an operating system.

The MGC3130 with Microchip's award-winning GestIC technology was the first electrical-field-based 3D gesture controller to offer low-power, precise, and robust hand position tracking at 200 Hz. In addition, GestIC technology uses advanced Hidden Markov Models to ensure that the recognition rate for 3D hand gestures is above 95%. Free-space hand gestures are universal, hygienic and easy to learn, making them ideal for display applications.

For the chance to win a Microchip a 2D/3D Touch and Gesture Development Kit, go to http://www.microchip-comps.com/bodo-oct16

The Rise of Global Chinese Power Supply Manufacturers

By Todd Hendrix, Director of Sales & Marketing, Enargy Power Corp, San Ramon, CA



Less than a decade after Jack Kilby and Robert Noyce invented the integrated circuit, American electronics companies began exporting manufacturing operations to low cost regions around the world, in predominantly Singapore, Taiwan, Malaysia and The Philippines in Asia. At the time mainland China was not easily accessible due to language, cultural and political issues. These first Western companies were labor intensive businesses focused on primarily on PCB assembly.

Western subsidiaries grew to employ thousands of workers and gradually developed a skilled workforce with a level of technological expertise. In the 1980's indigenous local companies began to spring up in the surrounding region to provide passive components and other electronic sub-assemblies. The founders and management were aspiring entrepreneurs---former local employees who had gained experience working for the subsidiaries of America and European companies. During the decade many related supply industries grew to generate more and more of the local content. Soon we saw the rise of the ODM business model for consumer electronics products.

In the 1990's we saw Asian companies in Japan and Korea begin to build brands that competed globally and became readily available and well known in the West. Sony and Sharp in Japan, as well as Samsung and LG in Korea were indigenous component suppliers producing simple discrete semiconductors under their own brand.

Taiwanese and Hong Kong companies began setting up factory operations in mainland China by leveraging their close cultural contacts. Soon western companies followed suit and set up manufacturing operations in southern China, which was easily accessible from Hong Kong. The first Asian power supply company, Astec, based in Hong Kong, grew by purchasing the internal power supply operations of Hauwei and Nortel, but was later acquired by Emerson Network Power, thus coming full circle back to being an American owned company. Soon Taiwan spawned Delta which became a global force in their own right in ODM power supplies; Acer and Asus also rose as global brands in PC's.

Chinese electronics and power supply companies have existed for many years. However, they have focused their efforts internally on Chinese equipment OEM's. The effort, time and cost required the address a global marketplace was seen as a daunting undertaking from their perspective. Recently companies with visionary leadership and westward-looking experience are now embracing the challenge of 'going global'.

Only in recent years have we seen the creation of global brands from Chinese companies, which is a dichotomy given that the majority of the electronics sold around the world are produced in China! Why? One reason may be that there is no pure Mandarin expression for "marketing": it has been lumped into one expression with "sales' that always pulls along the term marketing. However, global brand building requires a high degree of marketing, and there is a growing appreciation and understanding of the marketing aspect of business separate from the sales function.

During the late 2000's Chinese companies began to show interest in expanding their business internationally. Initial efforts were focused on under-developed countries and regions, such as India, South America and Africa, where the lower cost advantages of unknown brands were more attractive to poorer consumers than higher cost brand names. The US and Europe proved more difficult to penetrate due to consumers' higher level of brand consciousness as well as expectations for product service and support that Chinese companies did not have in place.

In this decade of the 2010s we are beginning to see the emergence of the first Chinese brands in the US and Europe. Some of these successes have been possible due to the acquisition of unattractive operations of western companies, such as Lenovo's purchase of IBM's PC division; Lenovo, albeit built on the IBM base, now ranks as the number one PC supplier in the world. Likewise, Haier is now building their brand globally in white goods.

In the last few years the rise of indigenous Chinese companies entering the global market for merchant power supplies is evident. My company, Enargy Power Corp. based in Shenzhen, has committed to developing business outside of China by establishing marketing and sales operations in San Ramon in northern California, with future plans to develop European operations. Our goal is to be the first Chinese power supply company to expand and establish design resources locally in North America to address custom low and medium power solutions to address the needs of medical and industrial equipment applications.

The rise of global Chinese power supply manufacturers will not happen overnight, and will require years of dedicated commitment and perseverance; but it will happen eventually as they learn how to harness the engineering, organizational, management and marketing talent that has been developed by working for, and collaborating with, foreign operations on the mainland.

www.EnargyPower.com

Over-Voltage Protection for RS-485 Bus Nodes

Robustness and reliability have made RS-485 the industrial workhorse over the past 40 years. Its large differential signal swing of 1.5V minimum and reliable operation over a wide common-mode voltage range of -7V to +12V have catapulted the RS-485's widespread deployment. Initially used as a communication network in laboratory instrumentation, RS-485 has spread to control networks in industrial and building automation, PLC networks on the factory floor, process control, commercial heating, ventilation and air-conditioning systems, seismic networks, traffic monitoring systems, and alarm indication systems in oil rigs, coal mines and the petro-chemical industry.

By Thomas Kugelstadt, Principle Applications Engineer, Intersil

Along with the growth and widespread use of RS-485 came an increasing demand for greater robustness such as:

- Higher output voltage swing to ensure higher noise margin
- Wider common-mode range to allow for larger ground potential differences between remote bus nodes
- Increased tolerance to electrostatic discharges caused by field personal
- Stand-off capability or protection against persistent over-voltages far beyond the maximum transceiver supply level specified in datasheets

It is the latter point this article focuses on: RS-485 transceiver protection against large over-voltages. First, we'll discuss the difference between over-voltage and transient protection. Then, we'll look at what it takes for an over-voltage protected (OVP) transceiver to be successful, and how meaningful an integrated transceiver is versus a discrete solution using a standard transceiver. Finally, we'll compare some performance characteristics of a 20Mbps high-speed OVP transceiver to an inferior OVP version.

Over-Voltage Protection versus Transient Protection

The 24V and 48V DC supplies in industrial and telecom systems are commonly distributed through the same conduits as the data lines of an RS-485 network. Figure 1 shows a number of causes for over-voltage occurrences.

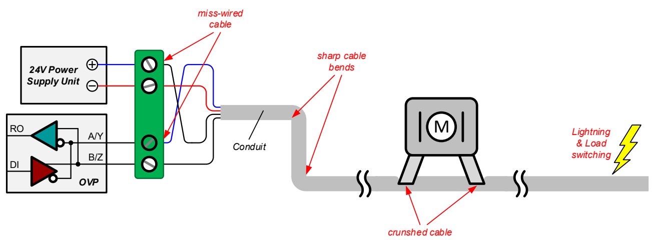


Figure 1: Multiple causes for over-voltage faults when data lines share the same conduit as DC power lines

If a DC supply shares the same connector or screw terminal block with the data lines of an adjacent bus node circuit, miss-wiring faults can occur that connect one or more supply conductors with the transceiver bus terminals.

Another failure cause is the layout of the conduit. Sharp bends often violate the minimum cable radius specified for data and supply cables. Over time, the increased mechanical pressure on the cable will cause a break in the insulation, causing shorts between power and data lines. This can also happen when machinery or equipment is placed against a conduit, thus crunching the cable. The duration of over-voltage events can last for minutes and up to weeks until their causes are eliminated.

Much shorter over-voltage events, such as over-voltage transients, can occur due to load switching activity in the power distribution system and lightning strikes, which induce high surge currents and voltages into the data lines.

Engineers new to over-voltage protection often assume that adding external transient voltage suppressors (TVS) to a non-fault protected, standard transceiver ensures protection against short- and long-term over-voltages. This is not true because the maximum power the TVS can absorb decreases with increasing transient duration, which is shown in Figure 2.

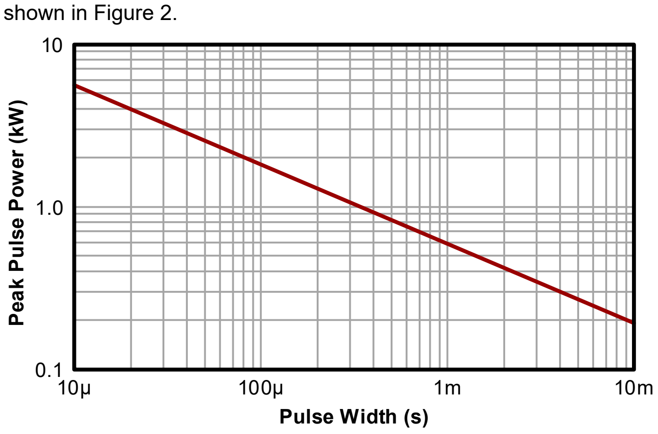


Figure 2: Peak pulse power versus pulse duration characteristic for a 600W TVS

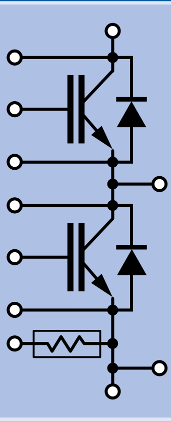
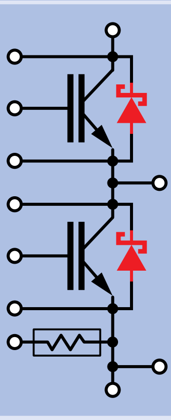
The diagram in Figure 2 shows a 600W TVS rated at 1ms pulse width. Note that the time axis ranges from 10µs to 10ms with power levels of 6000W and 200W respectively. From this characteristic, it

High Power next Core(HPnC)

with Fuji Electric's X-Series - 7G IGBT

Main features

- $T_{j(op)}$= 175 °C
- Low internal inductance <10nH
- High tracking index CTI>600
- High power cycling capability
- Low CTE base plate
- AlN Substrate
- Easy paralleling
- Modular concept, low number of variants

Technology	V_{CE}	I_C	V_{iso}
All Silicon Module Si-IGBT & Si-Diode 	1700 V	1000A	4.0kV
	3300 V	450A	6.0kV
Hybrid Module Si-IGBT & SiC-Diode Hybrid SiC 	1700 V	1000A	4.0kV
	3300 V	450A	6.0kV

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should be clear that exposing a TVS to long-term over-voltages would fry the device.

Therefore, to protect your bus nodes against the wide range of over-voltages, you need fault-protected transceivers, such as Intersil's ISL3245xE family. These transceivers provide protection against DC over-voltages of up to ±60V and transient over-voltages of up to ±80V.

Integrated versus Discrete Fault Protection

Occasionally the question arises: Why not use a non-fault protected, standard transceiver and a few discrete low-cost transistors with sufficient high voltage breakdown for over-voltage protection? The answer is simple: A discrete solution adds more cost and development time, and it consumes more space than a fault-protected transceiver.

Let's assume the function of the fault-protected, half-duplex transceiver in Figure 3 is to be accomplished with a discrete design using a standard transceiver. First, the transmit path and the receive path must be separate to allow for the implementation of a boosted output stage with high standoff voltage. This requires the use of a full duplex transceiver. The output stage could be realized with four discrete transistors or an integrated h-bridge whose control inputs require the conversion from RS-485 bus signals into TTL or CMOS logic levels. This would require a drive logic circuit between the transceiver and the discrete output stage.

In the receive path, a discrete voltage limiter, consisting of Zener diodes and series resistors, must be implemented to limit the bus voltage during an over-voltage event, otherwise it remains transparent.

Figure 4 shows the current limiting function of the ISL3245x family of fault-protected transceivers that operate over the wide common-mode range of ±20V.

Here, the first fold-back current level of 63mA ensures that the driver never folds back when driving loads within the entire 40V common-mode voltages. The very low second fold-back current setting of 13mA minimizes power dissipation if the driver is enabled when a fault occurs. This current limiting scheme ensures that the output current never exceeds the RS-485 specification, even at the common mode and fault condition voltage range extremes.

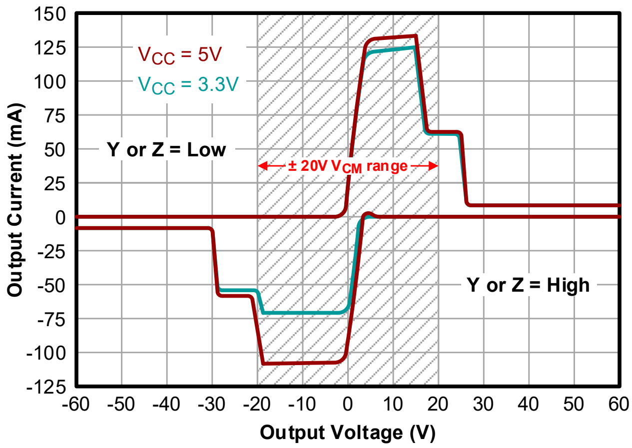

Figure 4: Driver output current limiting versus over-voltage

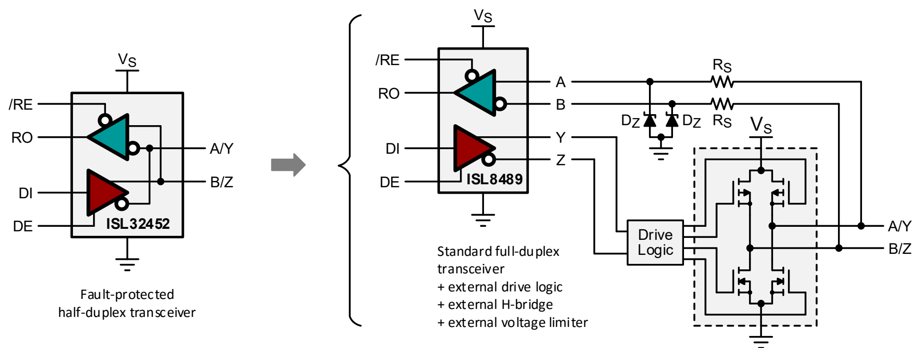

Figure 3: Integrated versus discrete over-voltage protection designs

In the event of a major short-circuit condition, the transceivers also provide a thermal shutdown function that disables the drivers whenever the die temperature becomes excessive. This eliminates any power dissipation and allows the die to cool. The drivers automatically re-enable after the die temperature drops by 15°C. If the fault condition persists, the thermal shutdown/re-enable cycle repeats until the fault is cleared. Receivers stay operational during thermal shutdown, and fault-protection is active regardless of whether the driver is enabled, disabled, or the IC is powered down.

Figure 3 shows that the discrete solution already becomes cumbersome by merely providing the basic functions for over-voltage protection, while still lacking a current limiter, which is a vital component for over-voltage protection.

Current limiting is a critical function during over-voltage events when the driver is actively driving the bus. Because the enabled driver presents a low-impedance connection to ground, bus currents flowing through the driver become huge, damaging the device if they are not limited.

Current Limiting of Fault-Protected Transceivers

Fault-protected transceivers with common-mode ranges wider than specified in the RS-485 standard require double fold-back current limiting within the driver stage.

Adding Lightning Protection to Fault-Protection

The energy of over-voltage transients caused by lightning can easily exceed the transceiver's fault protection and must be absorbed by external TVS diodes. Two conditions need to be satisfied when adding external TVS devices to a fault-protected transceiver:

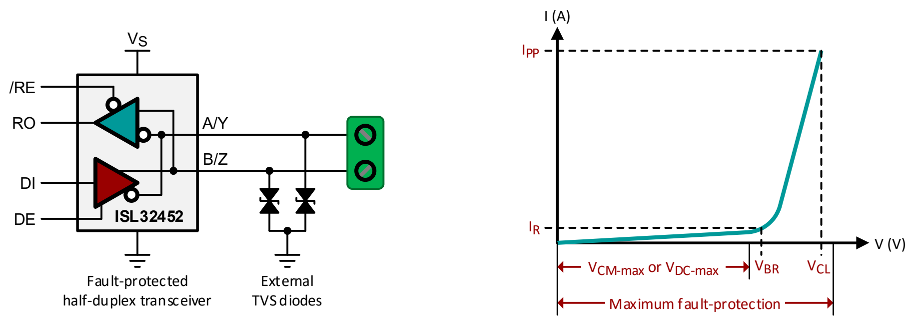

Figure 5: TVS V-I characteristic in comparison to V_{CM}-max and V_{DC-max}

The TVS breakdown voltage must be 1V higher than the highest common-mode voltage of the application or the maximum DC-supply, whichever is higher. For applications only exposed to the standard common-mode range of -7V to +12V, $V_{BR\text{-}min} \geq 13V$, for bus lines running adjacent to DC-power lines with 24V nominal supply, $V_{BR\text{-}min}$ should be ≥ 31V, as 24V systems are known for excursions of up to 30V.

The peak clamping voltage of the TVS must be below the transceiver's maximum fault-protection levels.

Figure 5 shows the respective circuit as well as the TVS switching characteristics with breakdown and clamping voltages, V_{BR} and V_{CL}, and compares them to the maximum common-mode, DC-voltage, and fault-protection levels.

Performance Comparison

Fault-protected transceivers with a wide supply voltage range enables designers to use the same device in 5V and in 3.3V low-voltage systems, which reduces logistics and can lead to an attractive price break for higher volumes.

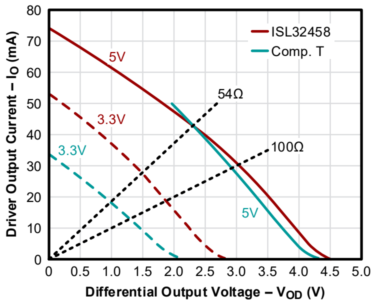

Figure 6: At 3.3V supply, the ISL32458E provides a 1.5V output compared to the 0.9V of Competitor T

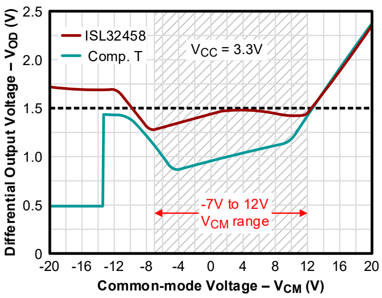

Figure 7: Competitor T's output drops further when measured across its specified common-mode range

Not all 3V-to-5V transceivers, however, provide sufficient drive capability at low supply. Figure 6 and Figure 7, for example, compare the output drive capability of the ISL32458E 20Mbps high-speed transceiver with a competing 10Mbps device, denoted as Competitor T, which also claims operation down to 3V.

The typical characteristic of driving a purely differential load (Figure 6) already discloses the inferior output drive capability of Competitor T.
At V_{CC} = 3.3V Competitor T struggles to provide 50% of the 1.5V minimum V_{OD} specified in RS-485, even without the burden of common-mode loading.
In strong contrast, ISL32458E delivers a solid 1.5V across the 54Ω differential load.

When measured across the much narrower common-mode range (Figure 7), Competitor T's V_{OD} comes nowhere near the 1.5V minimum (dotted line) for the entire range. ISL32458E deviates only slightly down to 1.3V at the extremes of the standard common-mode range and quickly regains drive strength towards the outer limits of ±20V.

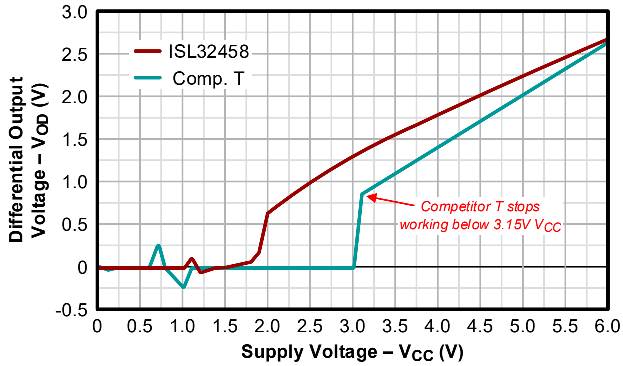

Figure 8: ISL32458E stops operating below 2V, which is 1V less supply than Competitor T's 3.15V

Another shortcoming of so-called 3V-to-5V transceivers is that they do not necessarily operate down to 3V. Figure 8, for example, shows that the Competitor T device stops operating at 3.15V, which is only 5% below the nominal 3.3V level. This of course requires a tighter tolerance of the linear regulator providing the transceiver supply voltage.

In comparison, the entire ISL3245xE family starts operating at a minimum supply of around 2V, thus not only ensuring true 3V operation, but also allowing for a relaxed tolerance specification of the voltage regulator.

Conclusion

System designers are no longer required to choose between robust fault tolerance and high performance in RS-485 and RS-422 transceivers, as the ISL32458E and ISL32459E offer both. These transceivers feature ±60V over-voltage and ±15kV ESD tolerance, while including operation over 3V to 5.5V supply voltages. They also operate up to a 20Mbps data rate, and provide a ±20V common-mode voltage range. In addition the ISL32459E provides a cable-invert function.

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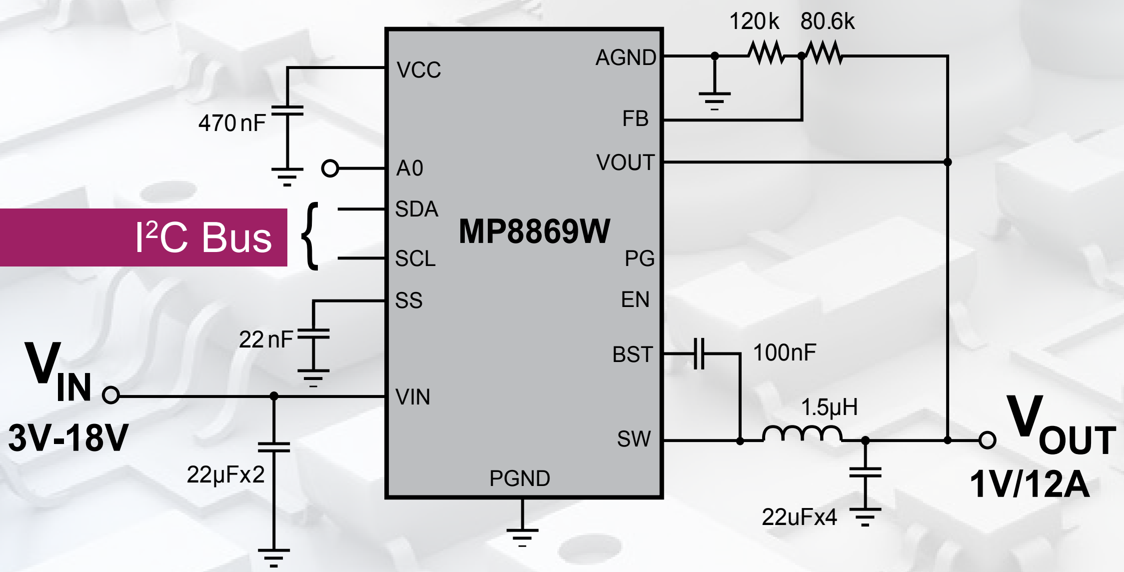
OVP-Transceivers with 3V to 5V Supply and High ESD

Device Name	OVP (V)	CMVR (V)	Data Rate (Mbps)	Unit Loads	Half/Full Duplex	High ESD	Cable Invert	Package Type
ISL32450E	±60	±20	0.25	1/4	Full	Yes	No	MSOP10, SOIC14
ISL32452E	±60	±20	0.25	1/4	Half	Yes	No	MSOP8, SOIC8
ISL32453E	±60	±20	1	1/4	Full	Yes	No	MSOP10, SOIC14
ISL32455E	±60	±20	1	1/4	Half	Yes	No	MSOP8, SOIC8
ISL32457E	±60	±20	0.25	1/4	Half	Yes	Yes	MSOP8, SOIC8
ISL32458E	±60	±20	20	1/4	Half	Yes	No	SOIC8
ISL32459E	±60	±20	20	1/4	Half	Yes	Yes	SOIC8

Digital Telemetry & Programming

3-18V, 12A Regulator with I²C



Key Highlights

- 5% Accuracy Output Voltage and Output Current Monitoring
- Programmable Current Limit
- Programmable Frequency
- Selectable PWM/PFM Mode
- 1% Internal Reference Accuracy
- I²C Programmable Output Range from 0.6V

Applications

- Enterprise SSDs
- Networking
- Servers
- Set-Top Boxes
- Smart TVs

CoolSiC™ MOSFET - a Revolution to Rely on

Silicon Carbide (SiC) switches are increasingly popular for power converters and other power-related applications. The latest technologies deliver higher efficiency, faster switching frequencies, reduced heat dissipation and space savings – benefits that, in turn, also lead to overall lower costs.

By Dr. Maximilian Slawinski and Marc Buschkuehle, Infineon Technologies AG

Launched at PCIM in May 2016 under the banner of 'A revolution to rely on', Infineon's latest CoolSIC™ MOSFETs offer designers unsurpassed power density and performance. Here we look at this technology and introduce an integrated CoolSIC-based reference design.

CoolSiC applications now and in the future

Certain application segments will always be early adopters of any new technology. Depending on a number of factors, including the system value, others will follow when the cost/performance ratio is attractive enough to migrate to a new technology.

SiC Schottky diodes are already used extensively in high-end power supplies. Looking forward, Infineon believes solar inverters and boost circuits will gain the most from MOSFET technology, quickly followed by Uninterruptible Power Supplies (UPS) and chargers. Traditional segments such as motor drives, traction and, on a longer time scale, automotive applications are expected to embrace a large scale migration to the new semiconductor technology in the future.

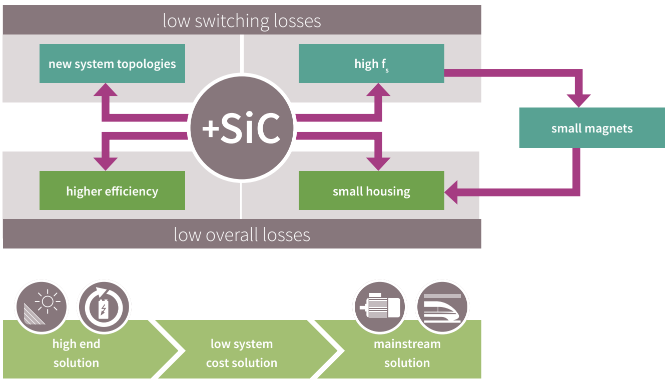


Figure 1: SiC benefits and target applications

The most significant trend in solar inverter designs is the increasing power density based on the reduction of switching losses, thereby enabling smaller heatsinks. Associated with this are the higher operation frequencies that enable smaller magnetic components.

CoolSiC MOSFET solutions

CoolSiC MOSFET devices with an on-resistance ($R_{DS(ON)}$) of 45 mΩ are an important step for SiC in power semiconductors. However, electrical performance is only part of the story. For fast switching IGBTs and SiC transistors, the design of the package is equally as important. That's why Infineon has developed a broad portfolio of packages.

By using the 4-Pin TO-247 (IMZ120R045M1) instead of the standard 3-pin TO-247 device (IMW120R045M1) it is possible to have a dedicated Kelvin source pin for the gate emitter connection. As a result, the gate control voltage is not impeded by the load current flowing in the power source pin. This improves turn on losses by up to 40%.

These discrete leaded products are ideal for fast switching applications and are expected to be implemented very quickly with the earliest applications being photovoltaic (PV) string inverters, chargers and UPS equipment.

Alongside discrete packages, the popular and flexible Easy1B power module is used to implement a half bridge configuration and booster solutions optimized for PV. These booster modules include 1200 V CoolSiC Schottky diodes and low VF silicon diodes for bypass and inverse-polarity protection. The flexible pin grid of Easy modules makes PCB layout easy and offers <10 nH stray inductance. This is a factor-of-five improvement over previous solutions and represents a valuable step in power module design. The booster module is the size of a matchbox and is designed to be used in MPP tracker systems up to 20 kW.

Lead products

Schematic	Type	R_{DSON}	V_{DS}	Package
Single switch	IMW120R045M1	45 mOhm	1200 V	TO247-3pin
Single switch	IMZ120R045M1	45 mOhm	1200 V	TO247-4pin
Half bridge with NTC	FF11mR12W1M1_B11	11 mOhm	1200 V	Easy1B PressFIT
	FF23mR12W1M1_B11	23 mOhm	1200 V	
Booster with NTC	DF11mR12W1M1_B11	11 mOhm	1200 V	
	DF23mR12W1M1_B11	23 mOhm	1200 V	

Selectively sampling on request.

Figure 2: CoolSiC MOSFET lead product overview

Efficiency, power density and cost are all key drivers behind the use of alternative semiconductor materials for power diodes and transistors. Hybrid power modules using silicon IGBTs and SiC Schottky diodes have become mainstream in power converters with high switching frequencies (>10 kHz) such as solar string inverters and UPS systems. Infineon addresses these requirements with a broad portfolio of CoolSiC Schottky Diodes in various discrete packages and several hybrid power modules using EasyPACK™ and EconoPACK™

The next big step towards higher switching frequencies and conversion efficiencies can be achieved by the use of full SiC solutions that combine SiC diodes with transistors.

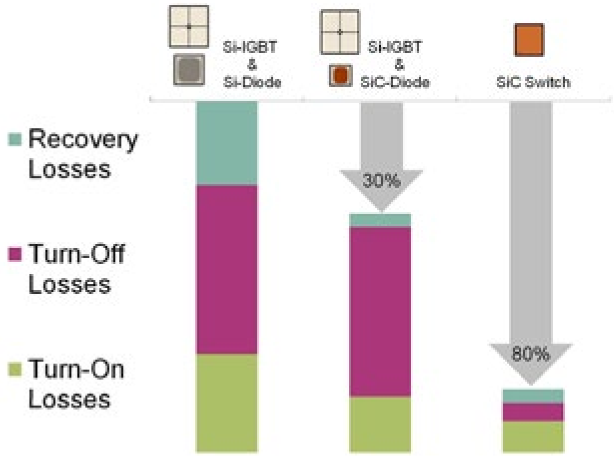


Figure 3: Evolution of loss reduction by the use of SiC power devices

The additional loss reduction of about 50 % resulting from the transition to full SiC is the result of lower turn-off and on-state losses.

Many of the currently available SiC MOSFETs require special drivers in combination with non-standard driving voltages, leading to low market adoption. The new Infineon CoolSiC MOSFET uses standard IGBT driving voltages (e.g. -5 V/15 V or -0 V/15 V). This allows the use of standard gate driver IC's such as the EiceDRIVER™ Compact and enables wide spread adoption of SiC MOSFETs into power electronic systems.

Switching behavior

The switching behavior of the new devices has been investigated by testing a 45 mΩ CoolSiC MOSFET in a TO-247 4-pin package driven with -5 V/15 V. The device exhibits very clean switching with almost no oscillations during switch on. This benign behavior allows easy and fast implementation of the CoolSiC into existing systems using standard EiceDRIVER IGBT drivers.

For the evaluation of the switching behavior at the system level, a DC/DC converter using a CoolSiC half-bridge module and an EiceDRIVER Compact with 6A output (1EDI60H12AH) was built. The module is configured in a bidirectional buck-boost converter topology and has an ohmic resistance ($R_{DS(ON)}$) of 23 mΩ at room temperature (FF23MR12W1M1_B11).

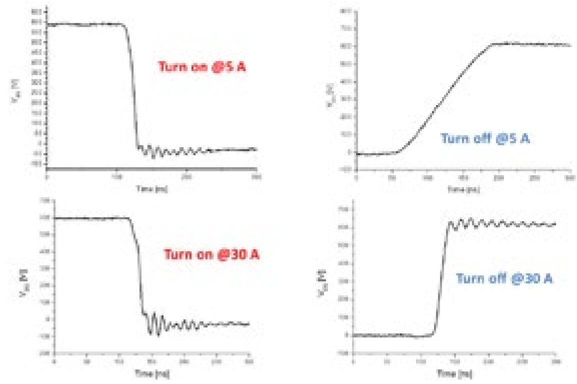


Figure 4: Turn on and turn off of the CoolSiC module within a DC/DC converter at 5 A and 30 A

Double pulse measurements were performed to measure the dv/dt at a gate resistance of 1 Ohm and different drain currents. The upper charts in Figure 4 illustrate the switching at 5 A where the turn on and turn off behavior shows only slight oscillations. The dv/dt at turn-off has a relatively low value of 5 kV/µs. The lower charts show the switching of the CoolSiC MOSFET module at 600 V and 30 A.

The dv/dt at turn-off is 34 kV/µs, which is much higher than for 5 A. The strong dependence of the switching behavior on the drain current is a result of the large output capacitance of the SiC MOSFET. Large displacement currents in the output capacitance of the device accelerate the turn-off by a dynamic increase of the turn-off voltage.

Further measurements of dv/dt and the peak voltage are shown in Table 1.

Current [A]	dv_{on}/dt [kV/µs]	$\Delta V_{DS,\,on}$ [V]	dv_{off}/dt [kV/µs]	$\Delta V_{DS,\,off}$ [V]
5	40	-71	5	+24
10	52	-79	11	+33
20	50	-89	22	+40
30	55	-97	34	+52

Table 1: dv/dt and V_{peak} of V_{DS} at different currents

Controllability

A further advantage of the CoolSiC MOSFET is the controllability of dv/dt and di/dt by the adjustment of external gate resistors. The results of testing an FF45mR12W1M1_B11 CoolSiC Easy1B half-bridge module can be seen below.

Increasing the value of the external gate resistors can reduce the dv/dt and di/dt of the MOSFET. This is a simple way to reduce electromagnetic noise, if necessary, or to fulfill application specific requirements.

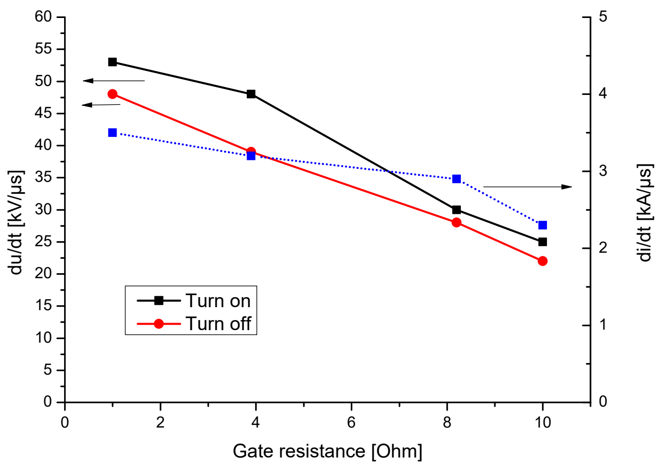


Figure 5: Controllability of the CoolSiC MOSFET via external gate resistors and the corresponding di/dt and du/dt levels

This simple dv/dt control allows the device switching performance to be tailored to match the requirements of a specific application. A common example relates to the restrictions regarding dv/dt limits for motor winding insulation (e.g. 5 kV/µs). With the right choice of gate resistors these restrictions are easy to meet with the CoolSiC MOSFET.

The key challenge for the gate driver PCB layout is to maintain a very low inductance around the gate source loop between the driver IC, the module gate and the auxiliary source pins. This can be achieved through the principle of a "strip line" design by using planes, and wide traces arranged in parallel paths. This enables a significant electromagnetic noise reduction especially in the context of hard switching with R_g=1 Ω.

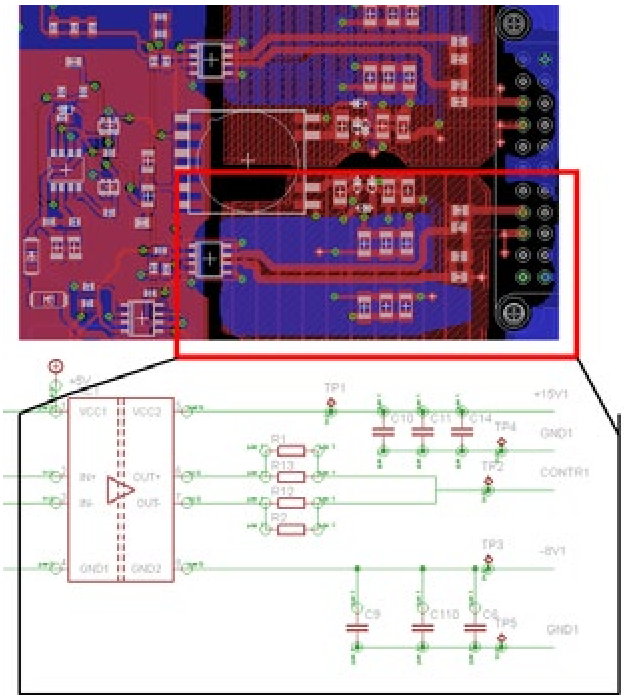


Figure 6: Example of a simple driver circuit for the CoolSiC MOSFET using Infineon's new compact driver 1EDI60H12AH

To obtain maximum performance from CoolSiC MOSFETs it is critical to pay special attention to the PCB layout, especially minimizing inductive loops. The demonstration driver circuit is part of an evaluation system, designed to measure the performance of the CoolSiC module. This system has been run at operating frequencies up to 500 kHz.

Summary and Conclusion

Full access to the additional switching performance available when using Infineon SiC MOSFETs can be obtained with standard components and topologies.

Due to the standard IGBT driver voltages and the benign switching behavior, Infineon CoolSiC MOSFETs enable very low design effort, especially with respect to driver development. Coupled with easy controllability of the dv/dt through the use of external gate resistors (simplifying EMI management and ensuring excellent switching performance), the result is a lower cost of development for high-performance power converters based on CoolSiC devices.

With the introduction of the CoolSiC MOSFET technology the beginning of a new era of power electronics in terms of power conversion efficiency and power density has begun. The new CoolSiC MOSFET enables engineers to develop smaller and more efficient PV inverters, UPS and charger systems that use higher switching frequencies and smaller heatsinks resulting in up to 80 % lower switching losses compared to silicon IGBT based solutions.

www.infineon.com/power

Digital Control in High Power AC-DC Converters with PFC

Design of High Power(1 kW and above) AC-DC converters featuring Power Factor Correction (PFC) is challenging task with issues like ripple currents in input power line, inrush current at start-up, switching high currents at inductive load, Electro Magnetic Interference (EMI), current balancing in polyphase topology, fast reaction on load change, efficient protection, and many others. Microcontroller (MCU) based Digital Control architecture along with high power discrete components from IXYS allow solve most of critical issues on the way to build high power AC-DC converters with PFC.

By Anatoliy Tsyrganovich, Leonid Neyman, and Abdus Sattar, IXYS Corporation

Digital control has significant advantage compared with traditional hardware implementation of high power converters:

- Programmable algorithm overlooking all aspects of conversion including inrush control, current balancing, over-current/over/voltage protection, and so on;
- Easy adjustment to the load variations by maintaining high efficiency in wide load range;
- Capability to accommodate different user requirements with few corrections in software and unaltered hardware;
- Replicable design that do not require sample to sample adjustment to get expected result

To graphically demonstrate all these advantages, High Power Digitally Controlled Two-phase AC-DC Converter with PFC (Converter) was designed. It utilizes two Zilog's 8-bit Z8F6481 MCUs and IXYS hardware components including novel X/X2 class trench MOSFETs.

Converter was tested with up to 1000 W and higher load. The load limitation was imposed by input power line rating and inductor's capability. This design can be extended to substantially higher conversion power, if needed.

Figure 1 illustrates a functional diagram of the Converter. Power factor correction is implemented using two-phase interleaved boost type conversion architecture. For Phase 1, it is executed with inductor L1, switching device T1, and diode D1. For Phase 2, boost conversion is accomplished with L2, T2, and D2. High Power factor is achieved by adjusting a peak current value with respect to rectified input voltage, which is used as a reference, Vin. Peak currents are captured with sense resistors Rs1 and Rs2 and amplified to the range of 2 V.

Due to high current values, the inrush current at start-up is digitally controlled and the bulk capacitor's C pre-charge is performed in pulse mode by switching device T3, along with inductors L1 and L2. The digital inrush current control concept utilized here was published in [1].

Current sense resistor Rs3 is used to monitor bulk capacitor's C current and the load current. It is also responsible for the overload protection functionality. Load current is sensed during zero crossing of the input AC voltage, when the inductor is discharged and capacitor C is the only source of energy provided to the load. At this moment, the capacitor's current is equal to the load current. This allows connecting the load to the rectifier's ground directly and eliminating the current sensing resistor in between.

The overload protection is continuously active with the comparator set to monitor the load current. In an event of overload, the comparator shuts down conversion. The MCU can be programmed to wait for a predetermined period of time and attempts to restart the converter. Additional programmable features include the number of restarts and the period between attempts.

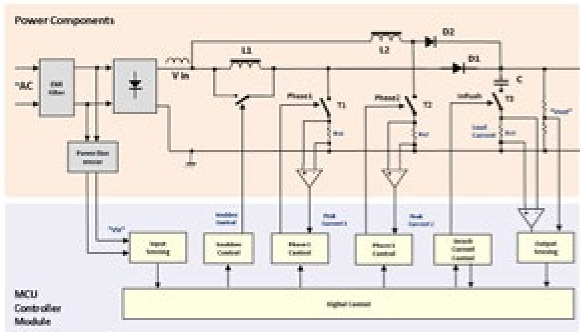


Figure 1: Functional Diagram for MCU controlled AC-DC Converter with PFC

Overvoltage protection is based on measuring output voltage during the AC line zero crossing time frame and comparing it to the programmable overvoltage reference. If overvoltage occurs, the device is shut down to prevent high output voltage in case of low load.

Principle of Operation

The following operational principles are considered for this design:

- Peak inductor current modulation was chosen over other PFC topologies aiming at high power conversion.
- Quasi-resonant mode is used as a version of critical conduction mode, allowing inductor voltage to drop to its lowest value in resonant process with strain capacitance after the inductor is discharged into a load.
- The second conversion phase is set as slave with respect to the first "master" phase.

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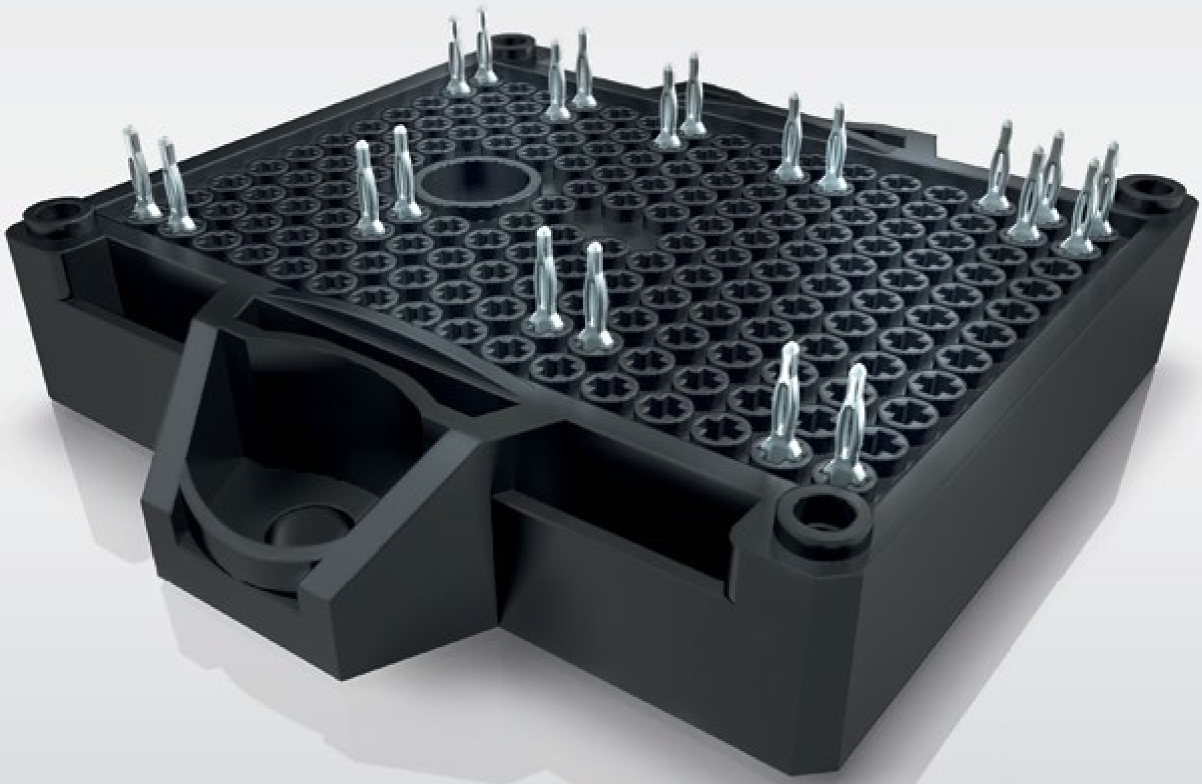

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- The balance of phase currents is performed using hardware and software.
- Low power consumption (less than half of peak power) forces the second phase into sleep mode.
- Pulse density modulation is used in addition to amplitude modulation to maintain efficiency of conversion at low loads.
- An active snubber is used to reduce stray oscillation after the inductor is discharged into the load.
- A Look-Up Table (LUT) is used to simplify calculations performed by the 8-bit MCU.
- Prediction control is performed with LUT assistance. Prediction is set to about 95% of expected output parameters.
- Digital feedback is used to keep output parameters in the predetermined range.

Peak Inductor Current Modulation – The Peak inductor current modulation architecture allows average line current to copy a form of the AC line input voltage. In this design, a rectified and scaled input sine wave is used as a reference. It applies to the analog input of the DAC, while digital DAC's input is used to set amplitude of the inductor current in respect to output power required. A signal from DAC's output provides reference voltage for peak current modulation.

Switching operations are paused near the AC input voltage zero crossing area to improve converter's efficiency by avoiding switching at low input voltages, when the voltage cannot be boosted to the required output level.

Figure 2 demonstrates inductors current/voltage waveforms at AC input of 120 V RMS and~700 W load. The peak inductor current amplitude is 12 A and peak voltage is 400 V. The bottom of the scope snapshot is a zoom-in of the top portion of the area between two vertical white lines.

Quasi-resonant Mode – The MCU allows maintain quasi-resonant mode by tracking the point of the inductor's discharge and providing the necessary delay before turning the inductor's current on again. The delay value may have some variance, because the inductor voltage has a dull bottom plateau, but inductor current changes smoothly as shown in Figure 3.

Quasi-resonant mode of operation is maintained at high power until power consumption approaches half of peak power, after which pulse density modulation is used to keep high efficiency of conversion.

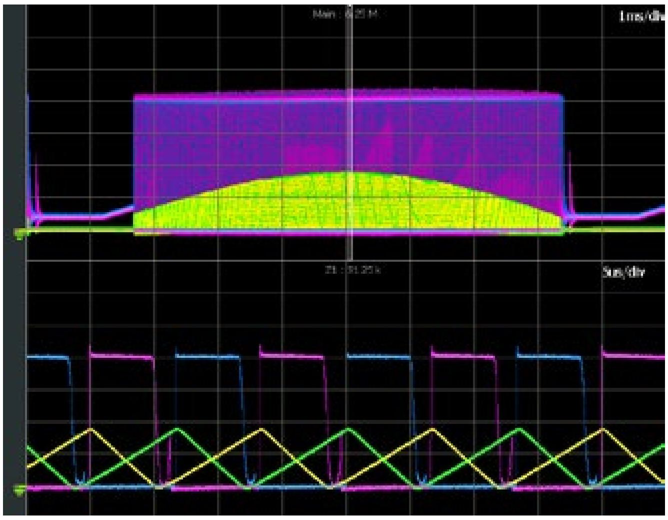

Figure 2: General Waveforms of Two-phase MCU-controlled PFC Converter
Legend:
Blue and Magenta – First and second inductor voltages. Scale -100V/div
Green and Yellow – First and second inductor currents. Scale – 5A/div
Lower half of the picture is a zoom of the top portion at the position of vertical white lines

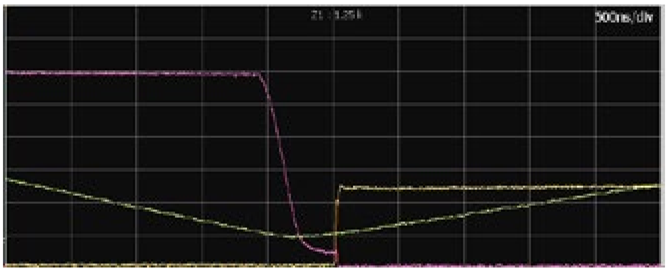

Figure 3: Quasi-resonant Mode Waveforms
Legend:
Magenta – First inductor voltage; Green – First inductor charging current; Yellow – Gate voltage for first phase MOSFET

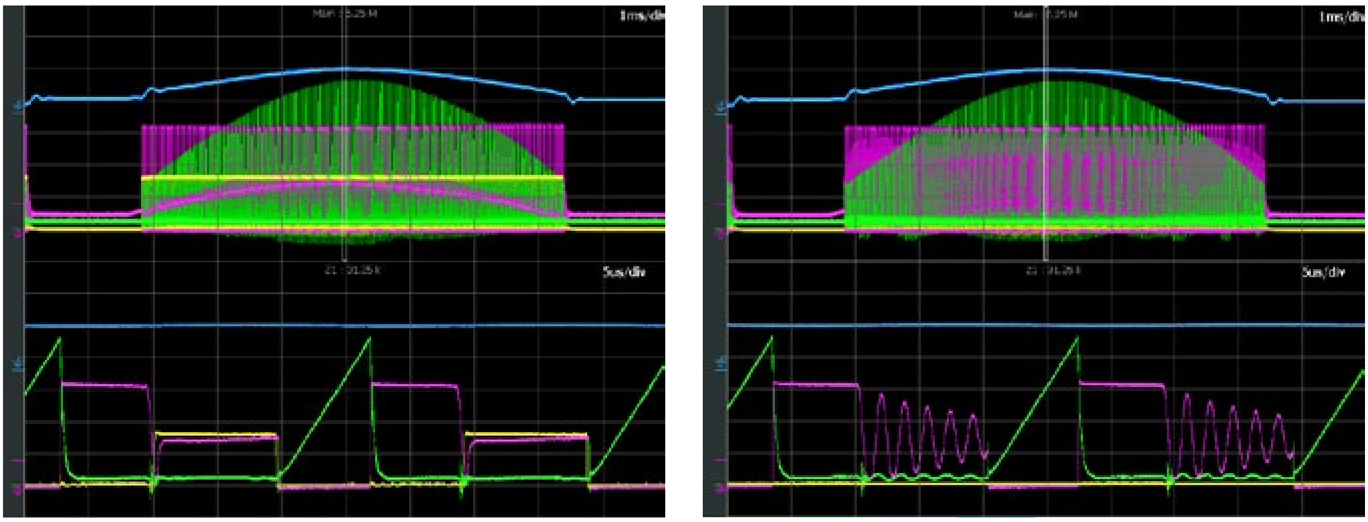

Figure 4: Inductor Voltage with Snubber Active (left) and Snubber Disabled (right)
Legend: Magenta –inductor voltage; Green – inductor charging current; Yellow – Gate voltage for MOSFET; Blue – power line current
Lower half of the picture is a zoom of the top portion at the position of vertical white lines

Second Conversion Phase is synchronized with the master phase. Synchronization is implemented by delaying start of the second phase conversion cycle with respect to the first phase timing.

Balance of Phase Currents to reduce ripples in the input power line is achieved by varying delay time of the start of second phase conversion. MCU adjusts delay time for every next conversion cycle based on previous conversion cycle timing. Peak inductor current for second phase is controlled with the same reference that is used for primary phase that allows keep them identical.

Low Power mode is determined as less than half of the nominal load consumption. Power consumption is monitored by measuring load current during AC input voltage zero crossing, where conversion is halted, but output voltage VOUT is maintained in the specified range. If load current drops below half of the maximum rated value, the second phase conversion is halted.

Pulse Density Modulation is used to further reduce generated power, when power consumption falls more than half of the rated power.

Active Snubber is used at low power consumption and low pulse density during the time frame from inductor’s discharge to the start of the next conversion cycle. Snubber timing is set by the MCU. An illustration of active snubber performance is shown in Figure 4.

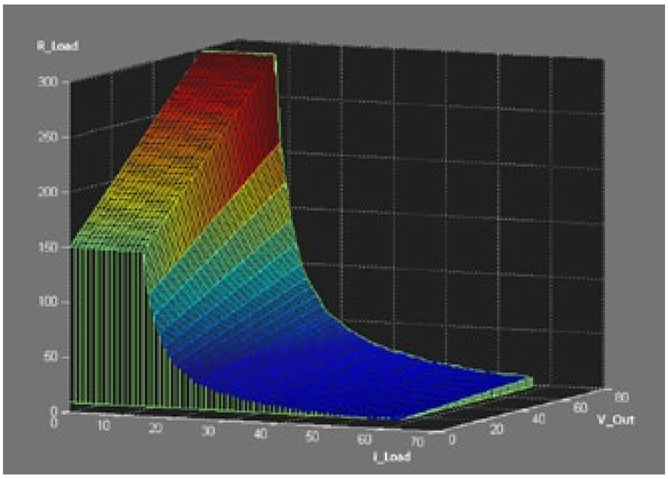

Figure 5: Two-dimensional Representation of R_Load in LUT

Look-Up-Table (LUT) simplifies MCU operations to speed up the algorithm. The data for LUT is converted to an 8-bit representation. The main LUT data contains load impedance, which is estimated by measuring output voltage VOUT and load current iLoad. Load impedance is more informative because it is estimated not only in steady state, but also in transition modes, when VOUT and iLoad are not settled yet. For simplicity of MCU operation, the load impedance is calculated as resistive parameter R_Load.

The R_Load = f (iLoad, VOUT) graph is represented in Figure 5, where iLoad is the horizontal axis. The axis values are in 8-bit representation for R_Load, 7 bit for VOUT, and 6 bit for iLoad. At low iLoad values R_Load is limited, and VOUT is controlled with feedback. If there is no load VOUT is maintained by burst of conversion pulses or may be programmed to go in shut down mode.

The LUT based algorithm is als used in prediction control.

Prediction Control sets initial values for peak inductor current DACdata supplied to DAC digital input and for delay of the next conversion pulse (indicated as Delay) to modulate pulse density.

Empirically obtained DACdata is a convoluted result of the influence of input voltage VIN and load power consumption POUT. VIN influences converted power as inverse square function 1/VIN^2. The input voltage range is set from 105 V to 140 V and scaled down to 1.72 V - 2.4 V for ADC measurement as shown in Figure 6. The power load is set from 60 W to 1070 W. The actual power values used are shown in the colored legends in the top right corner of Figure 6.

The prediction LUT is constructed as a two-dimensional LUT representing DACdata = f (VIN, R_Load). Prediction LUT is shown in Figure 7 in actual 8-bit data values along the X axis for VIN and Y axis for R_Load. Prediction control is designed to accommodate for about 95 % of control values for DAC and Delay needed to maintain output power. The Feedback Loop provides then precise correction of DACdata to keep output voltage within 1% range.

Another prediction LUT is used for delay prediction as a function of R_Load, Delay = f (R_Load). The Delay is zero at low R_Load (high power), then Delay steadily increases with reduction of power consumption (increase of R_Load) as shown in Figure 8.

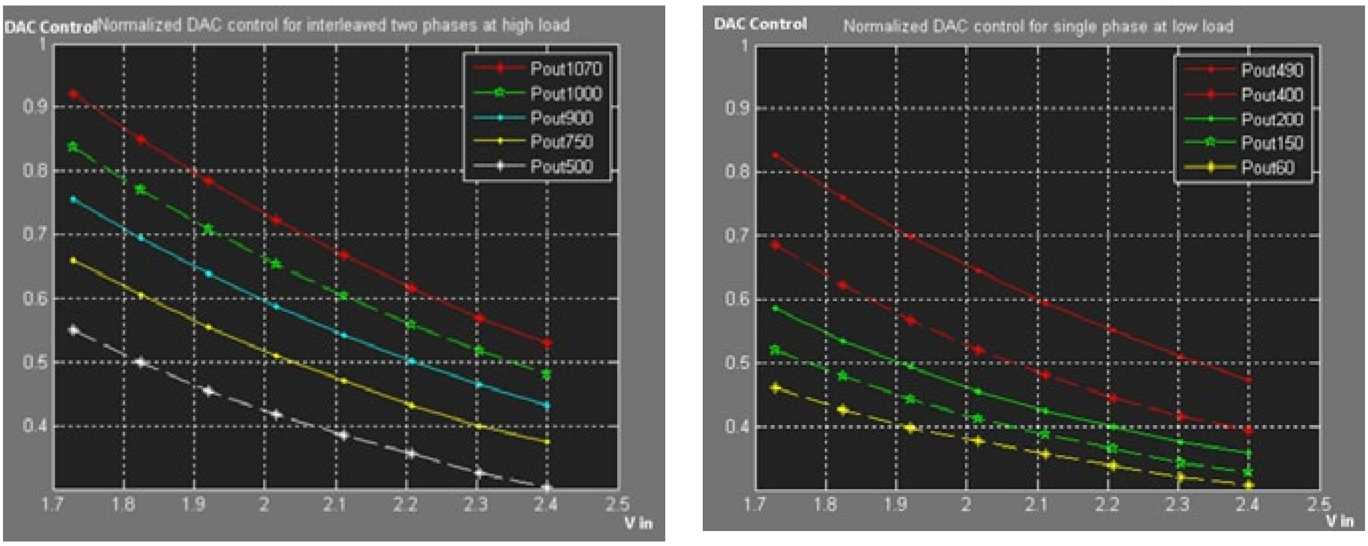

Figure 6: DAC Control Sets for Two-phase (left) and Single-phase (right) Conversion Modes

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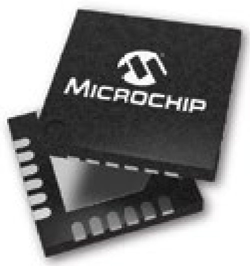

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- Power MOSFETs

Digital Feedback – Digital feedback finalizes the conversion algorithm as it modifies DACdata loaded as prediction value with feedback control value. The overall algorithm of feedback and its addition to DACprediction value is shown in Figure 9. The feedback control value is calculated once per AC half period at zero crossing, to be loaded into registers as a conversion time for following half of the sine wave period. The calculations are performed at 16-bit resolution in order to have smooth DAC control, which is of 12-bit resolution. The feedback data consist of two terms: integral term iTerm and differential term dTerm. The differential term acts as leading control to increase the stability of the control loop while reducing the settle time. KD and KI constants are used to balance the ratio between terms, while KER sets the feedback gain.

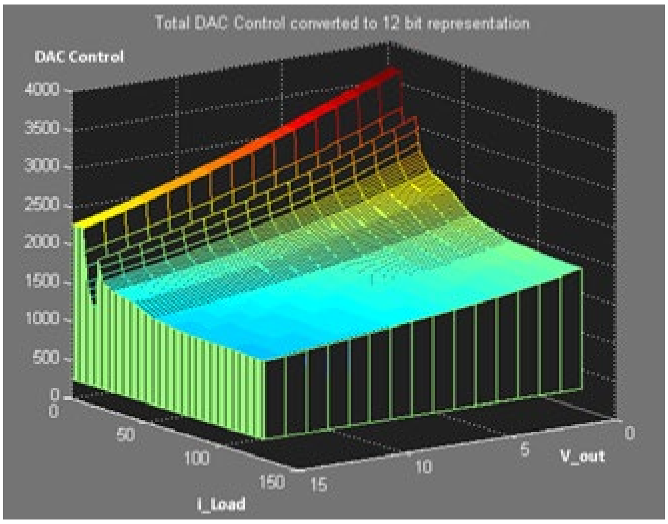

Figure 7: Prediction of DAC Control as a Function of VIN and R_Load

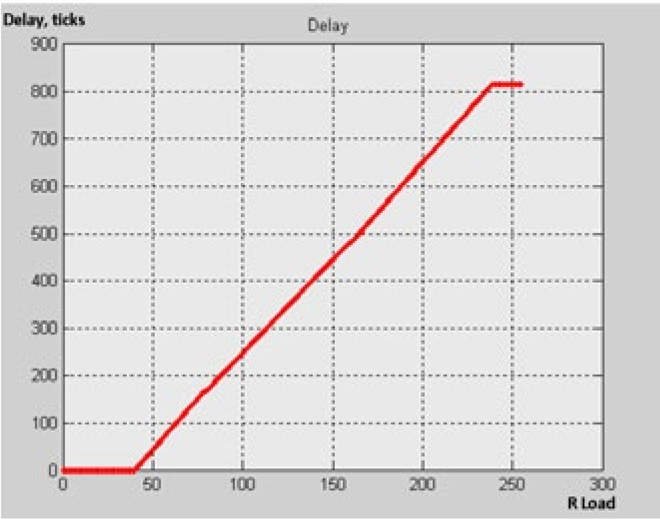

Figure 8: Prediction of Delay as a Function of R_Load

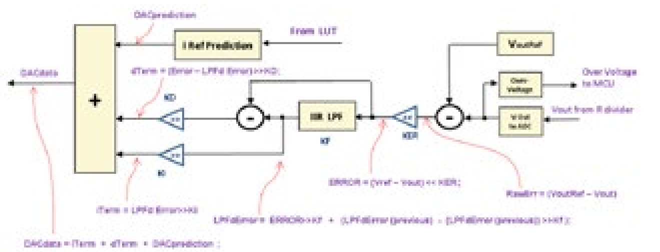

Figure 9: Feedback and Prediction Control Algorithm

The feedback loop control is set over prediction control to increase accuracy of the output voltage. Prediction data and feedback data are calculated around zero crossing and applied for next half wave of the input voltage. At the time around zero crossing, there is no active conversion and the only source of load current is a bulk capacitor. The current value is measured by MCU and is used then as index for retrieving data from LUT.

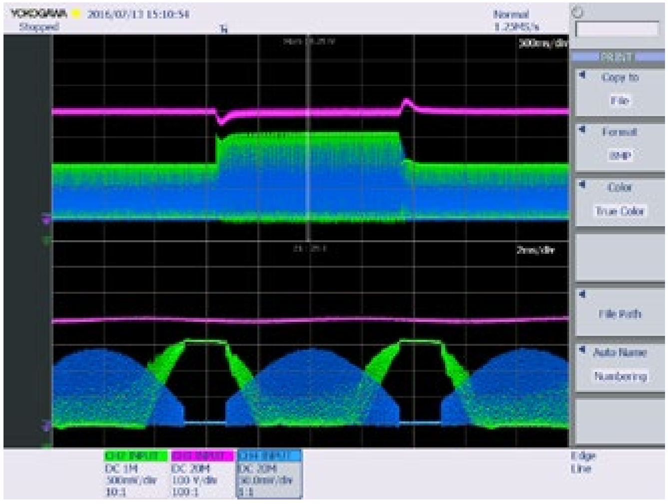

Figure 10: Scope snapshots depicting feedback reaction on a load boost from 60% to 100%
Legend: Blue – First inductor current. Scale -5A/div
Green – Load Current measured with amplifier. Scale – 0.8A/div
Magenta – output voltage. Scale – 100V/div
Lower half of the picture is a zoom of the top portion at the position of vertical white lines

Output voltage undershoots and overshoots at the time of rising and falling of load current as it is shown on Figure 10. Over/under shoots of the output voltage are at 7% range with duration of 100 ms. Output voltage stays steady after transitions.

Hardware Implementation

Discussed Converter's hardware implementation is shown in Figure 10. It consists of an MCU Module and a Main Power Board. The MCU Module is implemented as an add-on device. It contains connectors for MCU programming, because the MCU should be programmed before powering the entire system.

The Main Power Board is a two-layer surface-mount board that provides easy access to test points. The Converter may be powered from a 50 Hz or a 60 Hz 105 V to 140 V AC source. Auxiliary power supply of +3.3 V for the MCU and 12 V for the gate drivers on the Main Power Board requires.

Figure 11: MCU Module (left) and Main Power Board with MCU Module (right)

Testing Converter

The AC power line current was measured with a current probe at 20MHz bandwidth. The scope snapshot of the power line current is depicted on Figure 12. The total AC power line current value is sum of currents into two conversion phases involved, which results in AC power line current hovering over the peaks of single inductor current. Use of EMI filter centered at conversion frequency allows smooth current ripples. Overshoots at the beginning and end of conversion period are caused by EMI filter. The Converter offer following features:

- Conversion power, maximum: 1.06 kW @120V power line
- Input Voltage: 105 – 140 V AC 50/60 Hz
- Output Voltage: 400 V ± 2%, programmable
- Output Current: 2.65A @400V
- Output Voltage Ripples: <6% at full load
- Input Current Ripples: <2%
- Load Variation Range: >10x
- Device Conversion Frequency : 100–120 kHz at high power
- Programmable Overload, Overvoltage, and Brownout Protection
- Digital Inruch Current Control
- Soft Start Mode
- Power Good Status

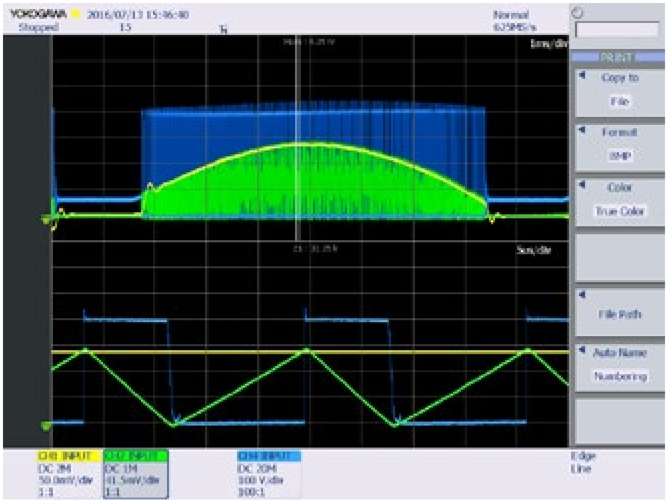

Figure 12: Scope snapshots depicting power line current along with inductor current and voltage
Legend: Yellow – Power line Current. Scale – 5A/div, Green – Inductor current Scale -5A/div, Blue – Inductor voltage. Scale – 100V/div Lower half of the picture is a zoom of the top portion at the position of vertical white lines. The current is shown only for one conversion phase.

Efficiency of the Converter was measured at 110V and 130V power line voltages. For given high power and relatively low power line voltage, efficiency is quite reasonable. It is expected to reach efficiency to about 98% going ahead with extending input voltage range to 220V -240V due to increasing output power at the same conducting losses. That can be seeing from Figure 13 where increase of efficiency by 1% is resulted in by changing input voltage from 110V to 130V.
At high power, two conversion phases are active. Efficiency slightly reduces with decreasing load until one of the active phases becomes disabled. Then the trend repeats for single conversion phase at low power. Converter's efficiency is maintained within +/-1% variations at 10X load variation.

Discussions

Digital control in power conversion provides flexibility in contemplating overall algorithm, designing dedicated features, and tailoring the design to customer needs. Use of MCU allows achieve high efficiency in wide range of loads and input voltages, while maintaining stable output voltage. Efficiency is achieved due to optimization between peak inductor (switching MOSFET) current and conversion pulse density. Power consumption is monitored by MCU, which disables second phase to reduce switching losses.
Inductor current balancing for two phase interleaved conversion is controlled by MCU maintaining identical peak current value and calculating start position for second phase current thus reaching ultimate current balancing state.
Prediction of conversion parameters speeds up settling of optimal peak current values and pulse density by programming conversion pulses delay for the next AC line half sine wave based on previous one. At 95% of predicted value to required one, the task for feedback control is greatly simplified leading to faster settling and high precision of the output voltage.
Use of fast MOSFETs and diodes based on latest X/X2 class designs from IXYS allows achieve fast switching at high current leading to increased efficiency.
Peak current modulation controlled by MCU allows modulate current from power line to be close to reference with precision of processing capabilities of MCU hence introduced harmonics are at low level.

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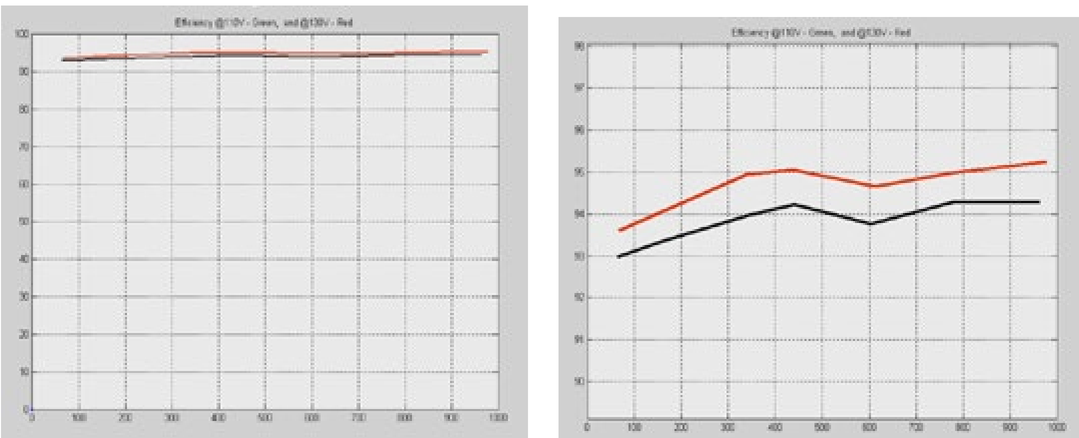


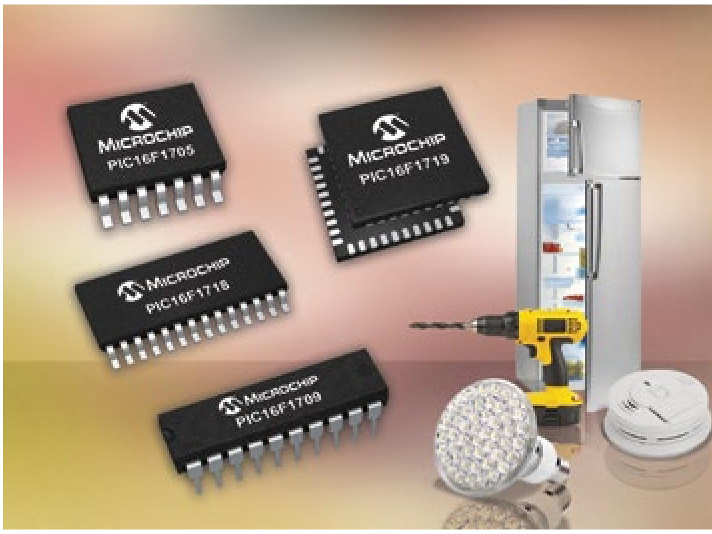
Figure 13: Efficiency of Converter in 10X load range (left) and zoom of upper area (right).
Legend: Red – Converter efficiency at 130V power line; Black – Converter efficiency at 110V power line; Horizontal axis – power, Watt

Combining Peripherals to Improve Resolution

Combing the CLC and NCO to implement a high-resolution PWM

Many applications can function with pulse width modulation (PWM) resolutions of less than eight bits, but in some cases, such as when dimming lamps, a higher resolution is required due to the sensitivity of the human eye.

By Cobus Van Eeden, Microchip Technology



A conventional PWM uses a timer to produce a regular switching frequency, and then uses a ripple counter to determine how many clocks the output is held high before the pulse ends. The output pulse width is adjusted to produce various duty cycle settings.

Calculating resolution

The effective resolution (measured in bits) of a PWM can be calculated by taking the base-2 logarithm of the number of pulse width settings possible.

For a device running at 16MHz, the smallest duty cycle adjustment increment would be 62.5ns (one system clock). If the PWM is configured to run at a switching frequency of 200kHz (switching period of 5µs), 100 per cent duty cycle will be achieved when the duty cycle register is set to 80 clocks (80 x 62.5ns = 5µs). This would make the effective PWM resolution only slightly more than six bits, as there are 80 steps to choose from. This is because one system clock divides into one period 80 times.

Knowing there are 80 possible duty cycle steps, a precise value for the resolution of the PWM can be calculated.

A PWM running from a 16MHz clock, which has a ten-bit duty cycle register, will start losing resolution due to this limitation at a switching frequency of 15.6kHz. For higher PWM switching frequencies, the duty cycle will reach 100 per cent before all of the steps in the ten-bit duty cycle register have been used, and for all the remaining values the output will simply remain at 100 per cent duty cycle. The frequency at which this point is reached can also be calculated.

In most PWM applications, the PWM is switched at a much higher frequency than the output can ever change. By filtering this PWM signal using a low-pass filter, the desired output is obtained. The filter removes the high-frequency switching components of the PWM by essentially calculating the average value of the PWM signal, and presents this as the output.

For example, in a switching power supply, the output voltage will be directly proportional to the duty cycle. The consequence of this is the smaller the adjustment to the PWM duty cycle, the smaller the resulting change to the output, leading to more precise control of the output.

From a control systems point of view, making small adjustments to the output effectively lowers the quantisation gain introduced by the PWM. In control systems, this lowering of the gain helps ensure the stability of the system.

Design

As shown in Figure 1, in principal, a PWM is created by the combination of two parameters – a repeating trigger, which determines how often the switching period or switching frequency are pulsed, and a single-pulse generator, which determines how wide the pulse is (the duty cycle).

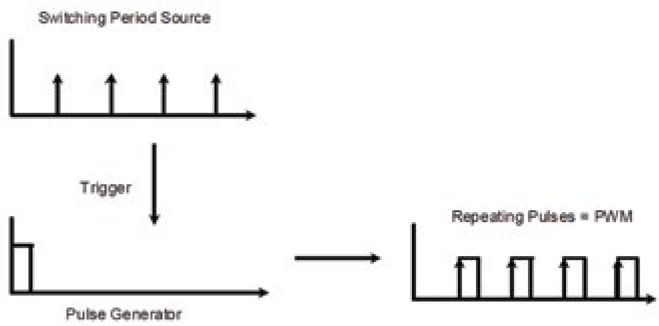


Figure 1: PWM construction

To achieve an increase in the effective PWM resolution, the numerically controlled oscillator (NCO) peripheral on a microcontroller, such as a Microchip PIC device, can be used to create a monostable circuit, which is one that gives a single pulse of fixed duration when triggered. The NCO will generate a signal that varies between two values in a defined proportion, creating an average pulse width, which

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is somewhere between two system clocks. The PWM signal pulse width will vary through jitter and dither by one clock period, with the proportion or ratio of the variation precisely determined by the NCO configuration.

In any application where the output is producing an average value, such as average power transfer to the load in SMPS or lighting applications, the variation in pulse width will be perfectly acceptable, because the average pulse width is accurately controlled.

By itself, the NCO peripheral cannot produce a PWM signal, but its behaviour can be changed by adding some logic using a configurable logic cell (CLC) to produce a PWM output. This can be achieved by using the conventional PWM as a clock source to trigger the PWM period, and use the NCO to determine the pulse width.

Any number of clock sources can be used, for example timers or even external signals, and in some applications an external trigger can be used to start the pulses, such as a zero-current detection circuit for power supplies. A simplified block diagram of this is shown on Fig. 2.

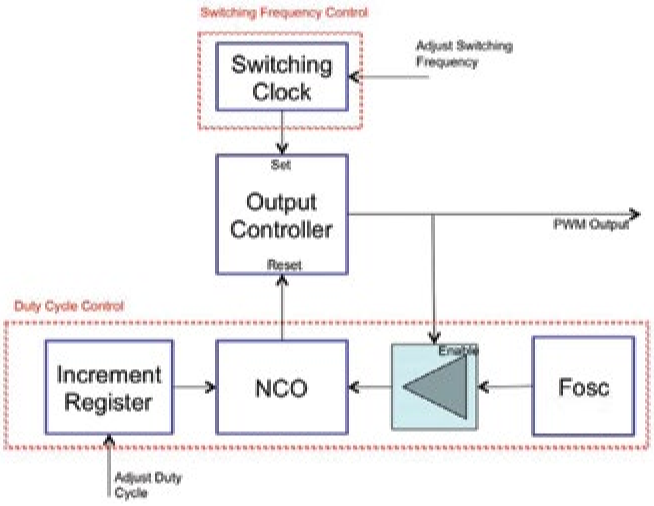

Figure 2: NCO-based principle of operation

The control logic in the CLC is used to set an output when the switching clock indicates it is time for the next pulse, and clear this output to complete the pulse once the NCO overflows.

Implementation

An implementation of this design using the NCO and CLC is shown in Fig. 3. For this design, the NCO is placed in pulse frequency mode, in which a short pulse is produced when the NCO overflows.

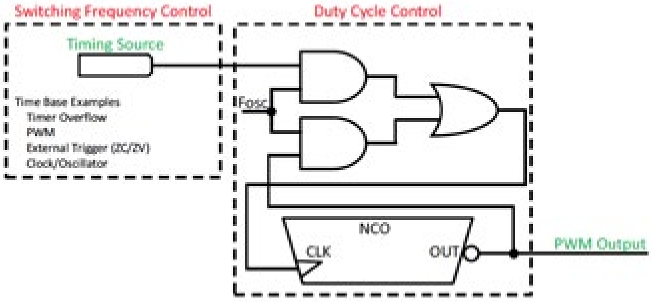

Figure 3: PWM implementation using CLC and NCO

When the system begins, the NCO output is low because it is waiting for enough clocks to count until it overflows and produces a pulse. This low output signal is inverted so the PWM output becomes high. This will supply a high-speed clock back into the NCO clock pin.

The PWM output will remain high until the accumulator overflows and the NCO output changes. This will cause it to stop producing the clock needed to run the NCO. At this point, the NCO is stuck high until it can get the clocks needed to finish its pulse. The PWM output is now low.

The timing source will then pulse high when the next period begins, feeding the high-speed clock back to the NCO.

The NCO uses these few clocks to finish the pulse, and then the output toggles back to the low position where it starts the process over from the beginning. The amount of time it takes the NCO to overflow will depend on the remainder left in the accumulator after the last overflow, as well as the increment register. Due to the accumulation of remainders, the pulse will sometimes be one system clock shorter than usual. By controlling how often this happens by setting the increment register, the exact average pulse width can be controlled.

The calculation of the pulse width will be according to the NCO overflow frequency calculation. The average overflow frequency of the NCO will determine the average output pulse width produced.

The NCO is designed to give linear control over frequency. The control over pulse width is subsequently not linear. The result is that the effective resolution of the PWM is not constant over the entire range from 0 to 100 per cent duty cycle.

For every duty cycle setting, the effective resolution at this particular point can be calculated and then plotted on a graph. This curve will look different depending on what the switching frequency is, the pulse width being adjusted independently from the switching frequency.

Conclusion

Conventional PWMs start losing effective resolution at relatively low switching frequencies. For applications where the switching frequencies have to be fairly high, and having as much PWM resolution as possible at these frequencies is necessary, the NCO can be used with the CLC to create a very high-resolution PWM output.

The smallest incremental change in pulse width achievable by a conventional PWM with a 16MHz system clock speed would be 62.5ns. If the fastest available PWM clock is a quarter the frequency of the oscillator (Fosc), then this increases to 250ns.

On the same device, a PWM with an incremental pulse width change of as little as 15ps can be constructed using this technique.

Even if the requirement is not primarily high resolution, this method may still be attractive for a number of applications, adding an additional PWM to the capability of the device, or having a constant on-off time variable frequency PWM, where the pulse is triggered externally as required when doing zero current switching in high-efficiency power converters.

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Technology for High Current, Low Insertion Force, Low Resistance and Long Cycle Life Power Connectors

Methode has developed a new class of patented power connector interface coined as "PowerBud®", which successfully overcomes key limitations of conventional power connectors. The PowerBud technology lowers both contact resistance and contact normal force without increasing connector volume, a capability that counters conventional wisdom. The resulting connectors exhibit lower insertion force, lower temperature rise, lower power loss and higher cycle life than conventional high current connectors.

By Russ Larsen, Methode Power Solutions Group

The high current (>50A) connector design presents many design challenges. At the top of the list is the need to minimize contact resistance. Lower resistance has the beneficial effect of lowering I2R power losses, lowering contact temperature, which in turn leads to higher reliability. The lower temperature rise also allows the connector pairs to be positioned closer together in a connector housing, minimizing the connector size.

Minimizing power loss is critical for supporting the Green initiatives that work to combat adverse human impact on the earth. The energy used, which is ultimately burned up as connector heat, is wasted energy.

Low insertion force is also highly important. Insertion force is simply the mechanical force necessary to mate or un-mate the connector. Reducing insertion force has the beneficial effect of reducing contact surface wear, a major contributing cause of connector failure. With lower insertion force, collective use of multiple contacts and multiple connectors is more easily achieved. Previous efforts concentrated almost entirely on developing improved contact surface coatings and materials. Other design efforts modified the pin shape to minimize the insertion force.

This paper discusses the design tradeoffs and innovative methods to optimize current rating, contact resistance and insertion force, and how those tradeoffs led to the development of an innovative new power connector technology.

Conventional Connector Challenges

Manufacturers finely polish and plate mating contact surfaces to increase contact surface area, leading to a general misconception that current flows through the entire mated surface area. However, the actual percentage of the area that actually makes contact with the mating connector is very small.

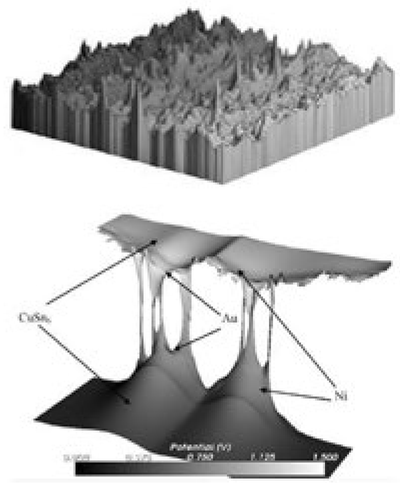

Figure 1:
Top: Surface topography showing asperities on a polished, gold-plated beryllium copper contact surface.
Bottom: Contact x/z cross-sectional view taken across AU contact interface showing potential values represented as grey scale and height.

Figure 1 shows that a polished gold-plated mating surface, viewed on a microscopic level, consists of peaks and valleys called asperities. The electrical current is concentrated and passes through the asperities, which are in actual contact. The x/z cross-sectional view taken across Au contact interface shows potential values represented as gray scale and height.

The actual points of contact between mated surfaces can have a relatively high resistance and therefore a relatively high voltage drop. Each of these points of contact has a finite resistance.

One way to minimize the overall contact resistance is to have many points of contact. By placing lots of contacts "in parallel", junction resistance is reduced.

Manufacturers have adapted to the limited amount of contact area using different methods to maintain low contact resistance, including:

- Increasing the size of the mating contacts; this yields many more microscopic points of contact. The result is a larger, more costly connector.
- Increasing the "normal force" pressing the two mating surfaces together; this slightly deforms the asperities thereby increasing contact surface area. The result is a connector with high friction force that is more difficult to mate, or an expensive connector mechanism to provide the additional force after mating.
- Using a manufacturing process to reduce the surface asperities.

The need to mechanically force the mating surfaces together has led to many design compromises. Since copper is one of the very best, reasonably-priced electrical conductors (excluding gold, silver and other exotic materials), it would be a good choice for the mating parts of the connector. However, copper has poor mechanical spring properties. If both mating surfaces were pure copper, the connector would also require an additional spring to maintain copper-to-copper contact. In the connector world, that yields an expensive product.

A more practical solution is to choose a material with both spring and conductive qualities such as copper beryllium or copper tin alloy. While less conductive than pure copper, these copper alloys are easily fabricated into a part that serves both as spring and conductor. This solution is widely used today in low cost connectors.

Methode Develops the PowerBud® Technology

Design Engineers at Methode developed the PowerBud, a new power connector interface with superior qualities compared to conventional, commercial power connectors. The PowerBud connection interface is an evolution of power connector technology that yields an interface with exceptionally low contact resistance as well as very low insertion force without a commensurate volume increase.

PowerBud connector performance is based on maximizing the number of discrete points of contact rather than attempting to increase the contact surface area or polish the mating surfaces to a finer degree.

Furthermore, the conductors are proprietary high performance copper alloy that is substantially better than more commonly used copper beryllium alloys to minimize resistance. Minimal resistance allows the PowerBud technology to handle relatively large currents with very low voltage drop.

The PowerBud connection interface embodies a design approach using massively parallel points of contact having a mechanical design that is adaptable to mass production. As show in Figure 2, there are two rows of conductors arranged one over the other. The high performance copper alloy is easily fabricated using automated processes.

A single beam conductor electrical path can be modeled as a resistor in a matrix demonstrating the cumulative parallel paths.

By arranging all individual conductors in a circular assembly results in massively parallel contact points, significantly lowering overall connector resistance. Low contact resistance means less heat generated under high current loads and less power loss.

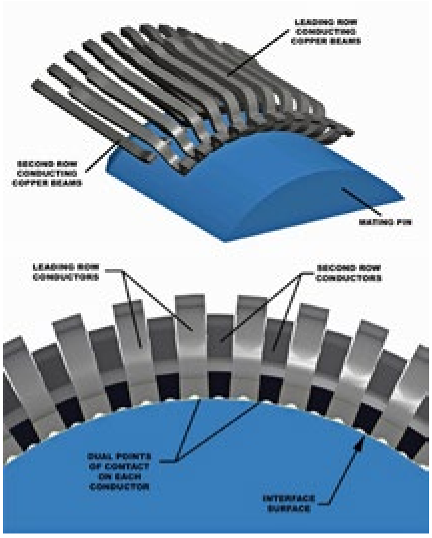

Figure 2: PowerBud mechanical construction showing two rows of contact bearns having dual contact points on each beam at the interface.

Figure 3: New type of connector connection interface

Figure 4: MQuad connector on the top and embedded PowerBud connectors on bottom

In addition, each copper alloy conductor beam includes a slight indentation in the finger-tips to create dual contact points, adding to the massively parallel contact points. This design approach, which is illustrated in Figure 3, has yielded a new type of connector connection interface with exceptionally low contact resistance as well as very low insertion force.

PowerBud Versions

The MQuad is similar to and slightly smaller than the PQ panel connector. MQuad panel connectors are also designed for blind-mate applications. As depicted in the top portion of Figure 4, each connector half floats under shoulder mounting hardware and self-aligns to the mating connector half. Contact terminations can be crimp wire, wired lugs or busbar attached. The MQuad is smaller than the PQ, and uses a 6.4mm pin rated at 100A @ 600VAC per contact with a 30ºC temperature rise.

The embedded PowerBud connectors, as seen in the bottom portion of Figure 4, have knurled outer side walls allowing direct press-fit insertion into busbars, printed circuits boards and FusionLugsTM. Installation or press-in is accomplished using any flat surface and does not require any special tooling. The current rating is dependent on the physical size of the pin and the heat-sinking capability of the mounting medium.

Conclusion

The PowerBud is a new class of power connector offering lower voltage drop resulting in lower temperature rise and lower insertion force than competitive connectors and much less than conventional connectors.

PowerBud allows more current to pass through a connector that occupies a small volume, potentially reducing package footprint. The lower voltage drop can eliminate the need for a local voltage regulator module. The lower temperature rise reduces system thermal load.

The PowerBud is suitable for systems that require connectors capable of handling hundreds of amps of current. It is particularly suitable for systems that require the connector to be mated and unmated frequently, or systems in need of multiple connectors.

www.methode.com

MiniSKiiP® with a Si_3N_4 AMB Substrate

Market trends indicate that low thermal resistance, long life and easy handling are the attributes engineers value most in power modules. Low thermal resistance is an advantage because it can serve to boost the module's power rating.

By Vince Botyánszki, Technology Engineer, Vincotech Ltd., Kossuth Lajos u. 59, H-2060 Bicske (Hungary)

Long service life benefits most industrial applications, but longevity is all the more important to solar inverters because panels are so durable and can keep delivering returns for decades to come. Lifetime and thermal resistance are connected: Lower thermal resistance reduces the ΔT during operation, thereby mitigating the module's exposure to that thermal stress. And handling ease is a cost and reliability enhancer: An easier-to-handle module is also safer and cheaper to mount.

The MiniSKiiP® design features a 0.38 mm Al_2O_3 direct copper bonding (DCB) substrate, Wacker P12 thermal interface material (TIM), and spring-loaded contacts. This recipe serves up exactly what engineers want—good thermal resistance, long life and handling ease. Thermal resistance needs to be improved first to keep pace with market demand. Around 50% of the thermal resistance of power module without a baseplate is attributable to the power module, and the other 50% to the TIM, so it is a good idea to improve both components' thermal resistance.

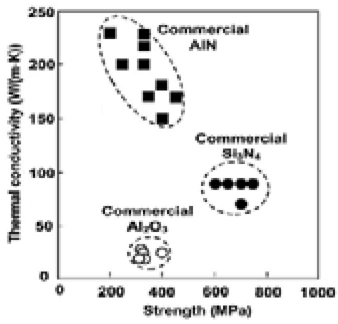

Figure 1a: Flexural strength vs. thermal conductivity [1]

The ceramic in the substrate is the key determinant of the power module's thermal resistance. AlN and Si_3N_4 substrates conduct heat better than Al2O3 substrates (Figure 1a). An AlN DCB substrate does not last as long as that of a Al_2O_3 DCB (Figure 1c). The thin Al_2O_3 layer added to the AlN surface to make it suitable for the DCB technology's oxide-based ceramics (Figure 1b). Si_3N_4 is not option with DCB technology, but AlN and Si_3N_4 ceramics work with AMB (active metal brazing) technology. The brazing layer between the copper and ceramic can absorb stress much better, so AMB substrates 'live' a lot longer than DCB substrates. The combination of AMB technology with a Si_3N_4 ceramic yields the longest life because the ceramic's high fracture toughness. Although Si_3N_4 does not conduct heat as well as the AlN ceramic, its greater flexural strength does allow AMB technology to be used to arrive at a 0.32 mm thick layer. The thinner ceramic compensates for the lower heat conductivity, so the two substrate materials end up with nearly the same thermal resistance (Figure 1d). The thermal resistance of 0.5 mm Si_3N_4 AMB copper can be equal to that of 0.3 mm AlN AMB copper, but the MiniSKiiP® housing is designed for substrates with 0.3 mm copper on the layout side. The thicker 0.5 mm copper layer compresses spring contacts, so either the contacts or housing must be modified. The difference between 0.3 and 0.5 mm copper's thermal resistance is not all that significant, so it is easier to just use 0.3 mm copper.

TIM exhibits very low heat conductivity compared to other structural materials used in power modules, so the TIM layer has to

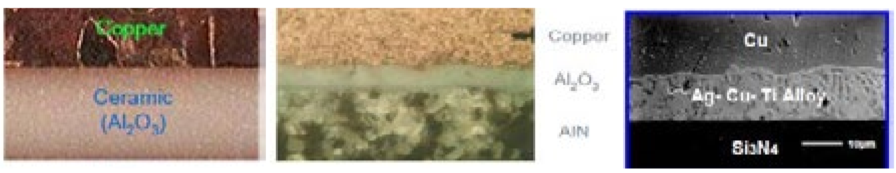

Figure 1b: Cross sections of Al_2O_3 DCB, AlN DCB and Si_3N_4 AMB [2,3]

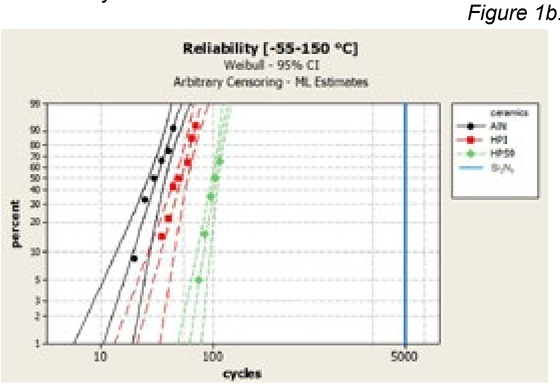

Figure 1c: Lifetime properties of substrates [4]

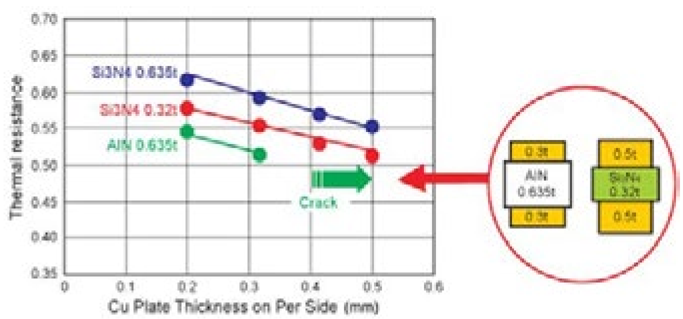

Figure 1d: Comparison of AMB substrates' thermal resistance [5]

be as thin as possible. TIMs consist of a soft matrix typically made of polymers and a highly heat-conductive solid material. These thermal interface materials have to fill the minute pockets below the power module so the solid material's grains have to be fine and the ratio of matrix and solid material has to be optimized for flow capability. TIMs' heat conductivity can be improved with highly conductive solids such as metals and a phase change matrix that makes the most of latent heat between the solid and liquid phase. Comprised of an aluminum solid and phase-change matrix material, PSX-P TIMs are four times more heat conductive than P12 TIMs made of ceramic solid material and a non-phase change silicone matrix.

The PSX-P TIM, a solid material, is attached to the backside of the power module and heat treatment is applied after it is fastened in place. The torque on the screws may decrease as the spreading material fills tiny pockets in the honeycomb structure and the TIM layer gets thinner. Hence, the heat treatment and subsequent retightening of screws entails a much greater handling effort.

Easy handling is one of the three priorities, so Rth measurements were taken to assess the effect of heat treatment and retightening. The results indicate that Rth is stable before and after spreading and retightening at the lower and upper torque limits. The solid TIM with the honeycomb structure is thrust up against the heat sink so it touches the surface, creating a thermal bridge that assures heat conductivity. The spreading starts when the power module is first used, but it improves the Rth only slightly because the pockets are so small. The TIM's thickness decreases after spreading, causing the screws to loosen, but the distance in the small gap between the module and heat sink is kept by the matrix's surface tension.

A MiniSKiiP® design with a Si_3N_4 AMB with 0.3, 0.32 and 0.3 mm layers and PSX-P phase change TIM exhibits better thermal resistance and lifetime properties, while remaining just easy to handle a Al_2O_3 DCB with 0.3, 0.38 and 0.3mm layers and P12 TIM. And that makes it an appealing proposition for customers.

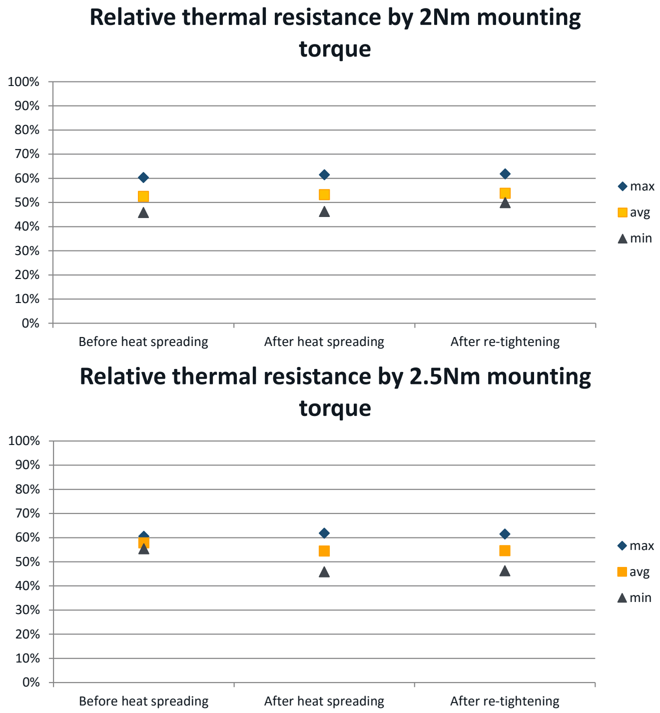

Figure 2: Rth measurement before and after heat treatment and retightening

Substrate technology and ceramic type	DCB - Al_2O_3	AMB - Si_3N_4
Top side copper thickness	0.3 mm	0.3 mm
Ceramic thickness	0.38 mm	0.32 mm
Backside copper thickness	0.3 mm	0.3 mm
TIM type	P12	PSX-P
Average R_{th} (IGBT)	100%	53%

Figure 3: Comparison Table

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4 Courtesy of Rogers Curamik'

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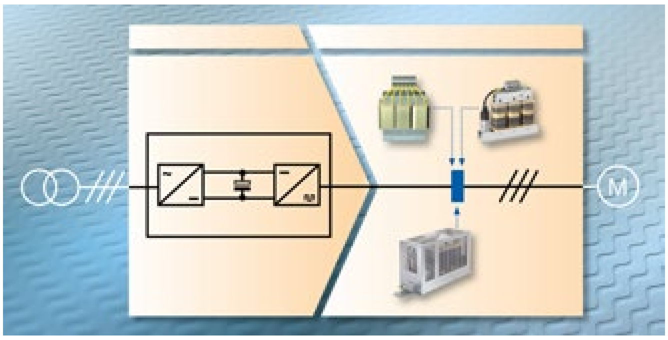
EMC Filters

Interference-free converter output



EMC problems in frequency converters occur not only at the input, but also at the output. To ensure reliable elimination of any kind of interference at the output as well, TDK offers a wide range of EPCOS filters and reactors.

By Carsten Jürgens, Director Product Marketing Power EMC Filters, TDK

Frequency converters are used for regulating the speed of asynchronous motors. In applications, in which the converter and motor are connected by a longer cable, parasitic capacitances occur between the conductors and to ground. In addition, the rise time of the square-wave pulses of the converter output voltage is in the range from 5 to 10 kV/µs, which causes high-frequency currents in the cable on every switching operation. These have a whole range of negative impacts:

- The superimposition of high-frequency currents on the cable reduces the amount of current available for operating the motor. To compensate for this, the converter must therefore be dimensioned for a higher power rating.
- Interference currents on long cables can be so great that the over-current protection switch of the converter is tripped.
- High-frequency currents, with their high switching frequency content, cause losses both in the cable and in the motor.
- As a proportion of the high-frequency currents are conducted to ground, they cause asymmetric interference. If unshielded motor cables are used, this would consequently result in inadmissibly high interference fields. For this reason, expensive shielded motor cables are generally used.

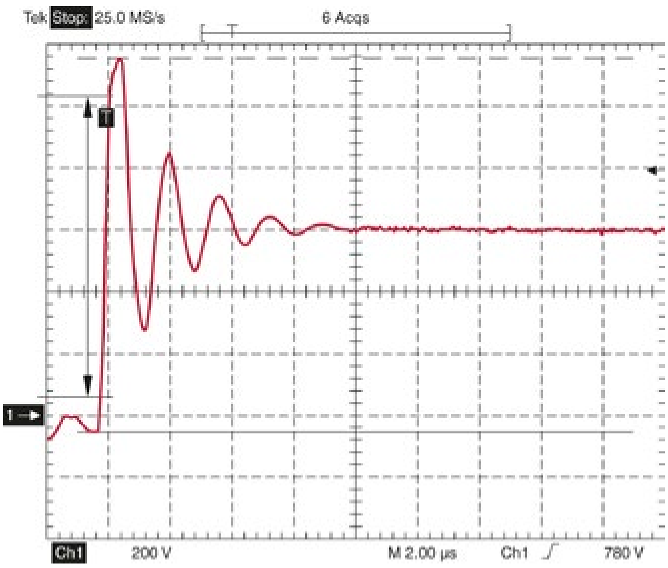


Figure 1: Voltage overshoot caused by motor cable

Long motor cables exhibit a relatively high inductance and cause large voltage overshoots in the case of steep rising edges.

- Moreover, the steep rising edge of the converter voltage excites parasitic oscillating circuits that consist of cable and motor capacitances as well as line inductances. Their transient characteristics overlay the converter voltage. On the motor side, this primarily results in momentary voltage overshoots which can far exceed the rated motor voltage (Fig. 1) and exert a load on the motor insulation due to partial discharges, which in turn can cause a motor to fail.
-

Overall, this creates the following problems at the converter output:

- Very high-frequency reactive currents in the motor cable
- EMC problems
- Overvoltage at the motor caused by steep voltage gradient and long motor cable
- Damage to the motor insulation
- Bearing damage due to leakage currents through the motor bearings
- Motor noises

Output filters permit effective interference suppression

For effective interference suppression, essential factors such as cable length, spectrum of interference frequency, motor type, or even rated power, are decisive. The amount of effort actually necessary and reasonable for suppressing interference at the converter outputs must be decided on a case-by-case basis.

dv/dt choke

EPCOS dv/dt chokes, also called motor chokes, are typical components for suppressing interference caused by long motor cables. The entire motor current flows through these serial chokes. Steep current and voltage rises at the output of the frequency converter are evened out by the inductance, while the parasitic capacitances of the motor cable are charged and discharged less powerfully. Motor chokes are used mainly for protecting motor windings against voltage spikes.

The components of Series B86301U* (Figure 2) are designed for a rated voltage of 520 V AC and, depending on type, for current capabilities of between 8 A and 1500 A. They are suitable for motor cables with a length of up to about 100 m at motor frequencies of between 0 Hz and 400 Hz. The smaller series types are designed for a maximum clock speed of 16 kHz or, in the case of versions with current capabilities of higher than 500 A, just 2.5 kHz. The design of the chokes complies with the IEC 60076-6 standard and all series types are manufactured with the UL-approved T-EIS-CF1 insulation system.

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Sine-wave filters

If greater demands are made on the interference suppression, sine-wave filters are recommended. These are designed as LC filters but, unlike the motor chokes, their limit frequency lies between the output frequency and the frequency converter clock speed.

As the sine-wave filter mainly suppresses the symmetrical interference between the lines, the interference acting on the phase-to-ground voltage is hardly reduced at all. Therefore, the motor leads still require shielding. Sine-wave filters reduce the motor noise and the eddy current losses and permit the use of motor leads much longer than 100 m.

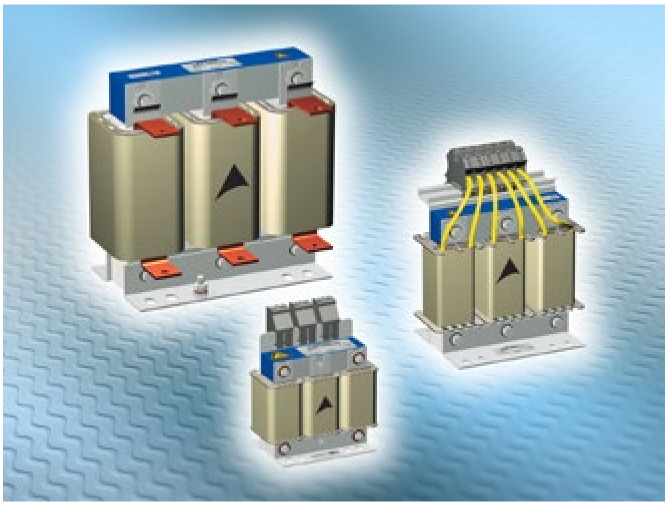

Figure 2: Motor chokes suppress voltage spikes The extensive range of EPCOS motor chokes covers a range of currents from 8 A to 1500 A. These prevent voltage spikes in the motor windings and thereby extend the service life of the motors.

Typical sine-wave filters are the components of the EPCOS B84143V*R227, R229 and R230 series. These are designed for continuous currents of between 4 A and 320 A at rated voltages from 520 V to 690 V and, depending on type, the permissible clock speed of the converters is between 1.8 kHz and 16 kHz.

SineFormer®

Although motor chokes and sine-wave filters reduce voltage peaks on the motor cables, they have barely any effect in the case of interference acting on the phase-to-ground voltage. . That's why shielded motor cables remain a necessity. Moreover, motor bearing currents are hardly reduced with the above measures alone.

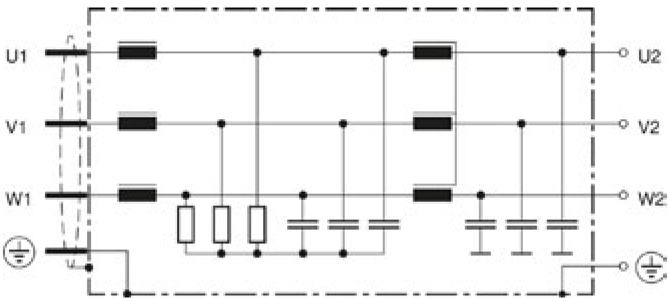

Figure 3: Circuit diagram of EPCOS SineFormer The SineFormer attenuates both symmetrical and asymmetrical interference. This eliminates the need for expensive shielded cables and protects the motor.

In order to reduce the asymmetric interference on the motor cable to such an extent that the cable requires no shielding, it is necessary to use an EMC sine-wave filter, which consists of a sine-wave filter supplemented by a current-compensated choke and capacitors to ground. The EPCOS SineFormer (Figure 3) is based on this circuit design.

Overview of benefits

Technical benefits of the EMC concept with SineFormer

- Reduction of the dv/dt to <500 V/µs
- Reduction of the noise generated by the motor
- Significant reduction of the eddy current losses
- Considerable reduction of the motor bearing currents
- Prevents coupling of interference from the motor cable with other network and signal cables
- Better EMC performance than shielded cables
- Radio interference emission within limits specified by standards
- Best possible reduction of the interference (conducted and radiated) in comparison with other output filter solutions
- No feedback to the converter's DC link is necessary

Economic advantages of the EMC concept with SineFormer

- Unshielded motor cables can be used, enabling installation overheads to be reduced, the service life of the motor to be extended significantly and the costs for motor cables to be minimized
- Motor size can be reduced
- Longer motor cables are possible (up to 1000 m unshielded have been measured)
- No maintenance costs, as the SineFormer is designed without any forced ventilation
- Compact filters (no modular system) with reduced volume and weight
- Requirements on the line filters can be reduced
- Increased system availability
- Suitable for retrofitting

The use of SineFormer in combination with unshielded cables is – depending on the cross section and cable length – usually more economical than the use of shielded cables. In installations with cable lengths of over 100 m the cost savings by using an unshielded cable instead of a shielded cable more than compensates the additional cost of SineFormer. If the costs of the SineFormer and unshielded cable are compared with those of a sine-wave filter and a shielded cable, then the point at which SineFormer represents the more cost-effective solution is already reached at less than 50 m. This does not

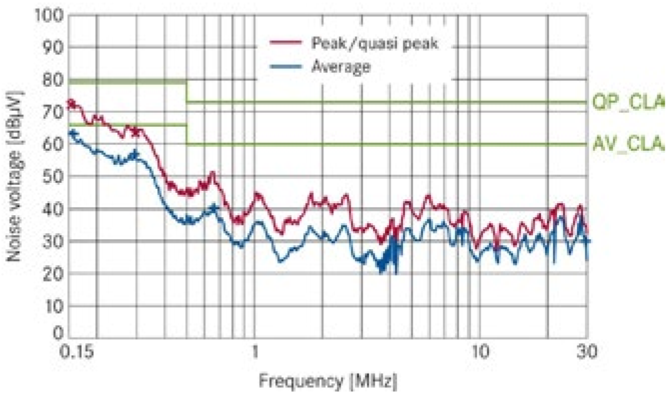

Figure 4: Noise voltage measurement on EPCOS SineFormer. Despite the unshielded cables, the permissible limits are observed.

even taken into account the fact that shielded cables incur additional mounting costs.

Figure 4 shows how well the SineFormer technology works: Even when power cables and unshielded motor cables are laid across each other, the limit values are met with a safe margin (in this case according to EN 61800-3 Category C2). The fact that virtually no coupling occurs is proof of the excellent effect of this filter technology. Overall, this enables system costs to be reduced and the plant availability to be increased.

Figure 4: Noise voltage measurement on EPCOS SineFormer. Despite the unshielded cables, the permissible limits are observed.

B84143V*R127 SineFormers are also suitable for retrofitting – i.e. when EMC problems caused by the motor cable do not become apparent until the drive is put into service. Attention should still be paid in every case to the suitable design of the line-side EMC filter, e.g. with the aid of the new compact B84243A* series of line filters. Limit Class C2 can be achieved according to EN 61800-3 in combination with SineFormer and with unshielded cables up to 300 m.

Rated voltage:	520 V AC up to 180 A, 600 V AC (320 A)
Rated current:	6 A to 320 A
Converter pulse frequency:	4 kHz to 8 kHz up to 180 A, 2.5 kHz to 3 kHz at 320 A
Approvals:	UL and cUL up to 180 A (except 6 A and 45 A)

Table: Technical parameters of the EPCOS SineFormer

www.sineformer.com

Simulating the Conducted EMI of Switching Power Converters

Simulation of electromagnetic interference (EMI) eliminates the 'black magic' of EMI design, enables increased power density with filter optimization, and reduces the risk of product delays. This article shares a methodology for model development and simulation of common mode and differential mode conducted EMI. An example simulation of a 1MHz switching power converter illustrates model development and compares the results with measurements.

By Mike Walters, Walters Power Electronics, LLC

All power converter products must meet EMI certification before release to production. EMI verification is typically one of the last product certification tests due in part to the high cost of an EMI certification lab and the need to evaluate the production version. Failing EMI compliance delays product introduction. One way to avoid an introduction delay is to overdesign the EMI filter to insure EMI compliance. However, this option increases the size of the filter, which lowers the converter's power density. An optimized filter maximizes the power density, and EMI simulation helps reduce the risk of product delay.

The EMI filter is only one piece of the high-density power converter puzzle. Recent promotion of wide-bandgap power devices, including SiC and GaN, promise increased power density with higher switching frequencies. However, when the switching frequency is within the conducted EMI frequency band, the size of the EMI filter is often larger than expected [1]. The faster switching edges of the wide-bandgap semiconductors generate a rich spectrum of high frequency noise requiring use of unfamiliar EMI components. Design engineers must comprehend each component's details on EMI performance to take advantage of higher switching frequencies.

Compounding the issue, an EMI filter may not be effective at the high switching frequencies due to the parasitic characteristics of the components. The impedance characteristics of the components change across the frequency band required for EMI certification. At the upper frequencies, inductors tend to behave as capacitors and capacitors tend to behave as inductors. It is not surprising that design engineers refer to EMI design as 'black magic'.

Optimization of the EMI filter increases the risk of delay for product introduction. Consider the traditional trial-and-error development process where the design engineer methodically evaluates the first prototype and insures the product meets the efficiency and thermal goals before evaluating EMI at a certified lab. If the product fails to meet the EMI limits, the engineer modifies the power converter and reevaluates the efficiency and thermal properties before scheduling another test secession at the EMI facility. The development cycle, including design, fabrication, and evaluation, repeats until the product meets the EMI limits. Each iteration adds development cost and delays the product's introduction.

This article advocates using a simulation model to optimize and evaluate the EMI filter during the design phase of the power converter's development. A simulation model helps the designer understand the conducted noise sources and coupling and can help quantify limits for each of the filter's parameters. Design iterations with simulation are much faster and less expensive than trial-and-error hardware iterations. Simulation helps the designer quickly evaluate their filter optimization ideas and reduces the risk of product delay.

The next section briefly describes the noise sources in switching power converters and the EMI standards for conductive emissions. This section also describes a generic simulation model that adds elements of the test setup given in the standards. The article further describes the method to extract the EMI signature. An example illustrates the construction of a simulation model for an active clamp flyback converter operating at 1MHz and compares the simulation results with published results. The final sections describe the method to estimate some of the missing model parameters and offers concluding remarks.

EMI noise sources and standards

Conducted EMI of switching power converters contain both common mode (CM) and differential mode (DM) components as illustrated in Figure 1. The high slew rate of the switching node generates CM noise in a switching power converter and couples the noise through the transformer's inter-winding capacitance, Cps to earth ground. The impedance from the power return to earth ground completes the CM noise path. The input current of a switching converter is the DM noise source. DM noise current flows in both power leads.

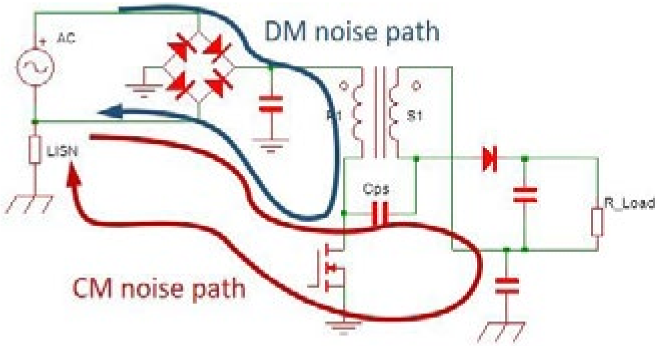

Figure 1: Conducted noise paths

International communities have established EMI limits and testing standards. In the US, the FCC specifies limits for conducted emissions over the frequency range from 450 kHz to 30 MHz. In Europe, the International Special Committee on Radio Interference (CISPR) specifies limits for conducted emissions over the frequency range from 150 kHz to 30 MHz. Engineers design the EMI filter to attenuate the converter's switching noise to below the limits of the frequency band specified by the standards.

The conductive EMI testing standards specify a test set up that includes a line impedance stabilization network (LISN) between the input source and the power converter to isolate the noise of the power converter. Figure 2 shows the LISN schematic added to a generic switching converter. The standards specify a LISN with 50μH inductors in series with both the power and return lines. The 0.1μF capacitors couples the noise to the EMI receiver reference to earth ground. The 50Ω resistors model the input impedance of the EMI receiver.

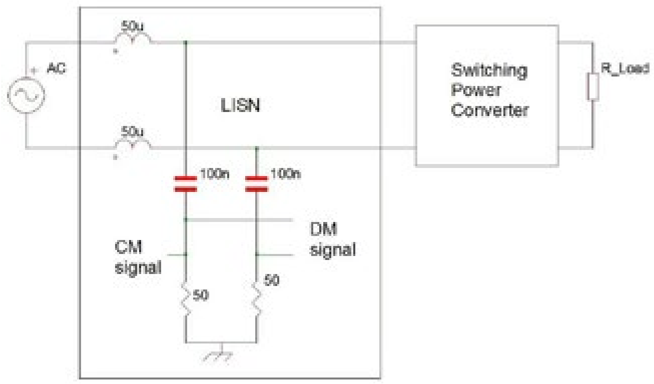

Figure 2: Conducted EMI testing with a LISN

Model and simulation approach

The EMI receiver measures the noise by sweeping an intermediate frequency (IF) filter over the specified frequency band and captures the noise amplitude with a quasi-peak detector [2]. The IF filter is a band-pass filter with a 9 kHz bandwidth at -6dB. Conceptually, a quasi-peak detector works like a peak detector followed by a lossy integrator, where the integration charge time constant is 1ms and the discharge time constant is 160ms. Quasi-peak detector readings are always less than or equal to the peak detection.

An EMI simulation runs a transient analysis of the power converter and collects time domain data. After the simulation run, Fourier methods transform the time domain data into frequency domain components. The Fourier data reports the peaks at each frequency. Rather than add complexity of a quasi-peak detector model to the simulation, you can choose to scale the peak frequency data to the RMS value [3].

Simulating EMI in a switching power converter can be challenging. Large data files result from simulating the switching converter over several AC line cycles while resolving the edges of the switching waveforms. Furthermore, a simulation schematic rich in parasitic elements and the desire to monitor numerous signals both increase the size of the data collected. Nyquist criteria requires time domain data resolved to at least 16ns to transform the data to 30MHz. However, an even smaller time increment may be need to preserve the fidelity of the switching edges. Nanosecond resolution for each waveform over tens of milliseconds of line input leads to large data files and can challenge the simulator's memory limitations and slow down the simulation time.

How do you decide which parameters of the switching converter to include in the model? Which parasitic elements do you place in your model? Do you use the manufacture's models for their semiconductors? You will need to use your engineering judgment to answer these questions.

Be careful using a supplier's model for their semiconductors. Manufacturers rarely state the limitations, validation criteria, or purpose of the model. This places the burden of understanding the model on the user. The design engineer can examine the Spice deck of the model and decipher the cryptic code or run a set of test simulations to assess the model's performance. I find both of these efforts tedious and usually end up building my own model. Below I describe the switch model development for a soft switching converter.

EMI simulation example

The following example uses the SIMPLIS simulation tool to evaluate EMI. SIMPLIS is specifically developed for switching power converters and uses piecewise linear (PWL) modeling and simulation techniques [4]. SIMPLIS is much faster than SPICE for switching converter simulations and does not suffer with convergence issues. SIMPLIS data points are not equally spaced in time due to the PWL techniques. However, discrete Fourier analysis requires equally spaced time domain data points. SIMPLIS provides the *spectrum* function to simplify frequency content extraction from the time domain data. The *spectrum* function first interpolates the data to equally spaced data in time, then performs a discrete Fourier analysis with the Hanning window function, and returns the magnitude of the frequency components. The result is both fast and accurate.

As an example, let us consider the EMI characteristics of an off-line active clamp flyback converter. CPES (Center for Power Electronics Systems) published the EMI results of their investigation into the development of a 65W active clamp flyback converter with a switching frequency of 1 MHz [5]. Their investigation showed the addition of a shield winding to the main transformer reduces both the CM and DM noise. Another CPES publication details a single-stage EMI filter design used to meet the EMI standards [6]. Figure 3 shows the schematic of the active clamp flyback converter and EMI filter.

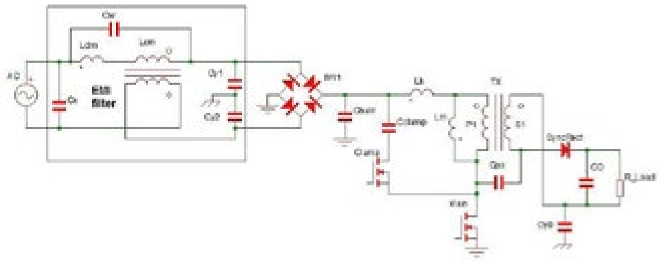

Figure 3: Off-line active clamp flyback converter with EMI filter

Figure 3 also shows several parasitic components. In addition to the primary and secondary windings, the main transformer includes the magnetizing inductance Lm, the leakage inductance Lk, and the primary to secondary capacitance, Cps. The single magnetic component of the EMI filter includes the common mode inductance, Lcm, the differential mode inductance, Ldm, and winding capacitance, Cw.

The goal of this simulation example is to reproduce the findings of the CPES researchers with a simple simulation model. The active clamp flyback converter operates with zero voltage switching. This simplifies the model of the MOSFETs and synchronous rectifier as explained below. The simulation uses CPES component values where disclosed and reasonable estimates for the remaining components. Table 1 lists the component values disclosed in the CPES publications.

Figure 4 shows the resulting simulation schematic. The earth ground (green wire) is the Earth port connection in the schematic. Ideal voltage-controlled voltage-sources with unity gain (E1 and E2) monitor the CM and DM mode noise. The bridge rectifier, BR includes non-linear junction capacitance. An ideal diode models the synchronous rectifier. The simulation operates open loop to help manage data file size by eliminating the large time constant associated with a feedback loop. The pulse generator sets the 1MHz switching frequency and the pulse width sets the output voltage. Asymmetrical delays and logic functions drive the switches and allow for dead time adjustment.

The model for the main and clamp MOSFETs in this example use a simple voltage-controlled switch, ideal diode, and a single switch node capacitor. Justification for this simple model comes from the zero-voltage switching (ZVS) operation. Prior to MOSFET turn-on, current flows in the body diode and the voltage is near zero. At turn-off of either MOSFET, the transformer primary current flows in the non-linear output capacitance of both MOSFETs. During the turn-off transition, the voltage increases across one MOSFET while decreasing across the other MOSFET. Likewise, the non-linear capacitance increases during the transition across one of the MOSFETs and decreases across the other MOSFET. You can think of the non-linear capacitors in anti-parallel that, in effect, reduces the non-linear behavior. This simulation uses a single switch node capacitor, Cswitch with a value of twice the CO(er) stated on the MOSFET datasheet. The transformer current at turn-off and Cswitch sets the slew rate of the switch node and is the CM noise source. This simple model results in faster simulation when compared to a vendor's MOSFET model with non-linear output capacitance.

The SIMPLIS *spectrum* function operates on CM and DM transient simulation noise data from 13.34ms to 30ms. This allows the transient simulation to reach steady state and captures the noise over a full AC cycle. Each simulation adjusts the pulse width of the pulse generator to realize 19.5V at the output. Additionally, the MOSFET voltage and current probes allow confirmation of ZVS operation for each switching cycle over the line cycle.

The CPES researchers first explored the impact of a transformer shield without the EMI filter. To compare their results with simulation, the schematic of figure 4 removed the EMI filter (Cx, Cy1, Cy2, Cw1, Cw2, Lcm, Ldm1, and Ldm2) and directly connected the LISN to the bridge rectifier, BR.

Component	Value	Notes
Cclamp	100nF	
Turns ratio	P1:S1 = 10:2	
R_{DS}(on)	0.29	transphorm TPH3202LS
$C_{O(er)}$	36pF	transphorm TPH3202LS
Cbulk	94uF	Rubycon BXW 2x 47uF
Cx	130nF	muRata GA3 2x 56pF
Lcm	1.4mH	
Ldm	30uH	
Cw	15pF	
Cy1 = Cy2	1nF	

Table 1: CPES active clamp flyback components

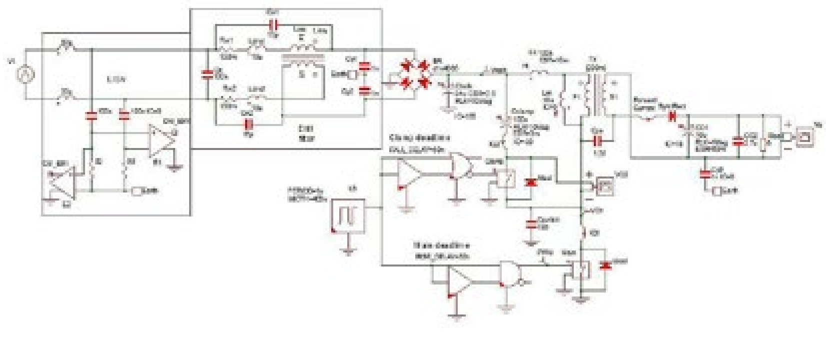

Figure 4: Example simulation schematic

Simulation Results

Figures 5 and 6 shows the measurement and simulation results of the transformer shield without the EMI filter. The measured results are from the CPES publications (see references 5 and 6). The simulation assumed a value of 400pF for Cps to model the capacitance without a shield and effectively removed Cps with a value of 1.5fF to model the effect of adding a transformer shield.

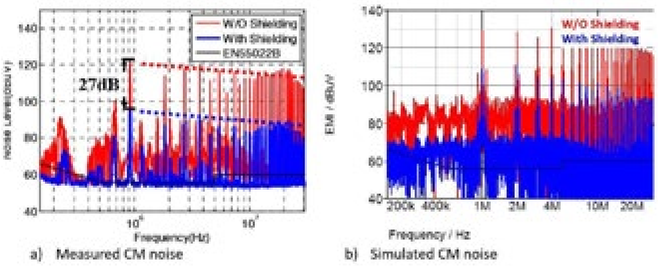

Figure 5: Common mode noise with and without shield winding and no EMI filter

Comparing measured and simulation results, figure 5 shows similar reductions of the switching frequency noise with a transformer shield. Adding a shield reduces the simulation CM noise by 21dBuV compared with 27dBuV measured by the CPES researchers.

Figure 6 shows the transformer shield reduces the simulation DM noise by 6dBuV compared with 21dBuV measurement at the switching frequency. At higher multiples of the switching frequency, the simulation results show reductions of 10dBuV and 18dBuV (2MHz and 4MHz respectively). The CPES researchers attribute the DM noise reduction to CM/DM noise transformation due to the non-linear junction capacitance of the bridge rectifier.

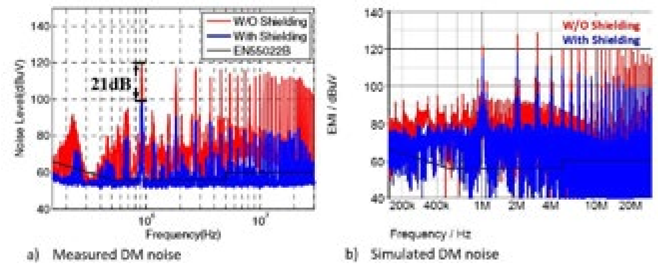

Figure 6: Differential noise with and without shield winding and no EMI filter

Given the measured noise with a shield winding, the CPES researchers chose a single stage EMI filter design. Their results compared the resulting noise with and without the EMI filter. The single-stage filter uses a single magnetic choke for both Lcm and Ldm. The simulation model for the choke (shown in figure 4) includes 1.4mH for Lcm, and

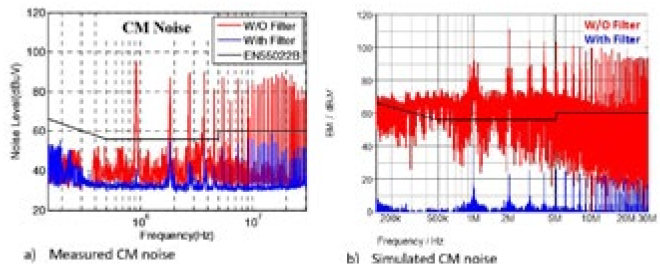

Figure 7: Common mode noise with and without an EMI filter

splits Lcm into two, one for each winding (Ldm1 and Ldm2). Splitting Ldm maintains impedance symmetry in the power and return lines. Finally, the model includes the winding capacitance, Cw1 and Cw2 across the choke's power and return terminals.

Figures 7 and 8 shows the measurement and simulation results of adding the EMI filter to a transformer with a shield winding. The simulation compares data from the full simulation with EMI filter shown in Figure 4 with the simulation results without the EMI filter from above. Note the difference of range for the vertical axis for the simulation results.

Figure 7 shows the effectiveness of the EMI filter for CM noise for both the measured and simulation. Compared to the measured EMI filter attenuation of 55dBuV at the switching frequency, the simulation results show an attenuation of 73dBuV.

Figure 8 shows the effectiveness of the EMI filter for DM noise for both the measured and simulation. Comparing the measured EMI filter attenuation of 38dBuV at the switching frequency, the simulation results show an attenuation of 87dBuV. Figures 7 and 8 show the EMI filter attenuates both the CM and DM noise below the EN55022 class B limits. However, the simulation results with EMI filtering are below the measured values.

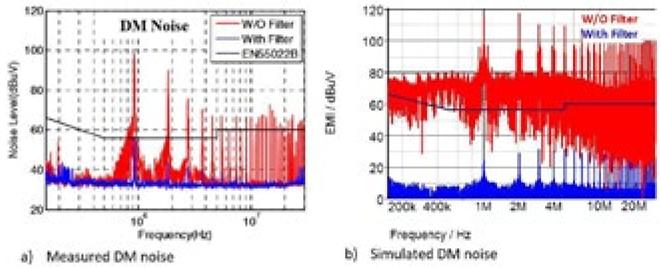

Figure 8: Differential mode noise with and without an EMI filter

Discussion

Although the simulation results support the CPES conclusions, the spectral amplitude does not exactly match the measured results. The difference is likely because the simulations used estimated parameters for a few of the critical components. CPES did not publish the values for Cps or Lm. The CPES researchers likely added a gap to the core; adjusting the magnetizing inductance to achieve ZVS. Additionally, the two CPES transformers, with and without shield winding, may have different core gaps and magnetizing inductance.

I estimated values for Lm and Cps using the ZVS criteria. CPES disclosed the core structure and material for the main transformer. The calculated maximum Lm without a core gap and 10 primary turns is approximately 74uH. The worst-case for ZVS is a transformer without a shield because the magnetizing current flows in both Cswitch and Cps. CPES value for Cps is unknown, but a similar planar transformer reported Cps values ranging between 120pF and 550pF [7]. Assuming a Cps of 400pF without a shield winding, the simulation iterated the magnetizing inductance and monitored the voltage and current waveforms of the switches for ZVS. This process found a magnetizing inductance value of 10uH necessary to achieve ZVS. Simulations both with and without shield windings assumed the same magnetizing inductance, Lm of 10uH.

Other accuracy improvements of the simulation are possible, but with diminishing return for the effort. You could chose to add the PCB impedances, distributed winding capacitances, and permeability falloff with frequency to the model. You could estimate these parameters or use finite element electromagnetic analysis. Of course, finite element modeling requires creation of a physical model and needs accurate material characteristics. Estimation of the parameters introduces uncertainty of the model's validity. At some point, the time and effort to improve the simulation accuracy exceeds the time and effort to simply build and measure the hardware.

Regardless of the absolute accuracy, the simple simulation model used in this article supports the findings of the CPES researchers. The simulation results show a reduction of EMI noise for a transformer with a shield winding and the CM/DM noise transformation due to the non-linear junction capacitance of the bridge rectifier. In addition, the simulation shows a single-stage EMI filter attenuates both the CM and DM noise below the EN55022 class B limits.

Conclusion

This article describes a method for simulating conducted EMI in switching power converters. The method highlights model development with a LISN and a procedure for extracting the EMI signature from a transient time-domain simulation. It also identifies the critical noise sources and component parameters in the noise paths. The example simulation achieves reasonable results with a simple MOSFET model and avoids the complexity of non-linear output capacitance. Historically, power converters operated with a switching frequency below 150 kHz to avoid a large EMI filter. Wide band-gap devices operating at high switching frequency promise higher power density, but only with an optimized EMI filter. Simulating EMI enables quick filter optimization compared with hardware design iterations and multiple trips to a certified EMI facility. Engineers can use simulation to identify the boundaries of real and parasitic components on EMI performance. The identified boundaries help the designer optimize the EMI filter and select or design the proper components. Simulation enhances the design engineer's understanding of conducted EMI to help eliminate the 'black magic' of EMI design.

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About The Author

Mike Walters has over 35 years of power electronic experience in the fields of LED lighting, computer power, and military aerospace. He has filed over 60 US patents with Cree, Intersil, International Rectifier, IBM, & GE. Mike is currently the proprietor of Walters Power Electronics, LLC specializing in power supply system architecture, system & product simulation, competitive analysis, integrated circuit definition as well as analog/digital control & circuit design.

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Monolithic Current and Temperature Sensing in Automotive Qualified Bare Die IGBTs

650V automotive-qualified bare die IGBTs are aimed at power module manufacturers who are developing advanced solutions for automotive traction inverters.

By Fabio Necco and Roy Davis, Fairchild

The IGBT product family PCGAXX0T65DF8 includes versions for 160, 200 and 300A. Monolithic temperature and current sensing allows direct on-chip measurement of junction temperature and collector-to-emitter current. The technique offers several advantages compared to off-chip temperature and current sensing. The article describes the sensing circuitry and outlines the advantages for traction inverters in hybrid/electrical vehicles (H/EV).

Traction inverters for H/EVs require power semiconductor devices with breakdown voltages in the range of 650V. Fairchild's new product family of AEC-Q101 automotive-grade qualified bare die IGBTs is designed for applications in today's and next-generation hybrid/plug-in hybrid, fuel-cell and battery-powered electric vehicles. We will refer to these vehicles simply as H/EV.

Fairchild's bare die automotive-qualified IGBTs are based on the third generation of Field Stop Trench IGBT technology and are matched with a soft fast recovery diode qualified to automotive-grade standards with additional features and options, such as changing gate pad size and location to accommodate different diameters of aluminum wire, resizing the die and customizing the breakdown voltage.

The new bare die PCGA160T65NF8, PCGA200T65NF8, and PCGA300T65DF8 IGBTs can be available with or without integrated monolithic temperature and current sensing circuitry, in three current ratings of 160, 200 and 300A. Their breakdown voltage of 650V is guaranteed over a junction temperature range from -40 to +175 °C.

Bare die IGBTs are typically used by power module manufacturers who are designing their own solutions for H/EV traction inverters to achieve a high level of power integration and reliability, or special power interconnections. The overarching objective is to push the power limits beyond standard module products.

Power Device Challenges

When dealing with power semiconductors, designers are confronted with the following challenges:

- power loss
- thermal management
- short-circuit, overcurrent/overvoltage and overtemperature protection
- current measurement

Power losses are affected by the VCEon value of an IGBT, and by its switching behavior (turn-on and turn-off time) and switching frequency. These properties are impacted by the characteristics of the IGBT technology, the gate driving circuitry, the package stray inductance, and the thermal management system.

Since power losses can only be minimized but never fully eliminated, thermal management must aim to remove the heat created by the losses from the semiconductor. The best way to remove this heat is to improve the thermal conductivity between the silicon and the outside world. In recent, more advanced power modules the users are adopting sintering techniques for the top and bottom of the power device combined with double-sided cooling to improve thermal conductivity.

The next problem that designers are facing is protecting the IGBT against overtemperature, overvoltage and overcurrent. Overvoltage can be limited by controlling the amount of the stray inductance in the current path by the appropriate design and controlling the rate of change of the current.

Overcurrent and overtemperature are undesired occurrences that can happen in the life of the power semiconductor. Being able to detect and act upon them in time can extend the life of an inverter. Current measurement is required by the motor control system to control the amount of current and torque delivered by the motor.

Temperature Sensing

Monolithic integrated temperature sensing in bare die IGBTs is accomplished by measuring the forward voltage drop (VF) of a string of polysilicon diodes, which are fabricated monolithically on the same die as the IGBT. Because the VF value of a diode shows a well known and linear dependence on Tj (Equation 1 and Figure 1). The method of monolithic current sensing proves to be the best way to measure the IGBT's junction temperature.

$$V_F = \frac{K \cdot T}{q} \ln \frac{I_F}{I_0} = S_F \cdot T_J \qquad (1)$$

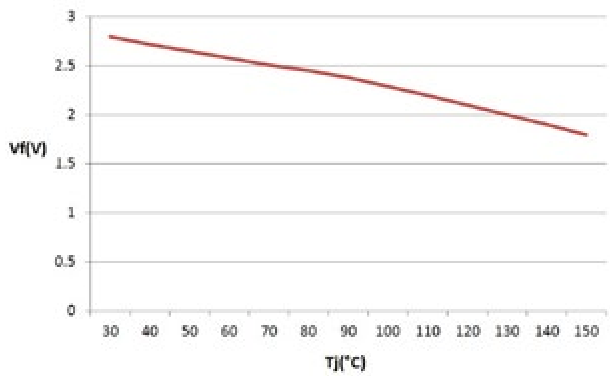

Figure 1: Dependency of voltage drop VF on the junction temperature Tj

The temperature sensor requires the diode to be forward-biased with precise constant current. The resulting voltage drop must be sensed and conditioned via an interface circuitry. Since the temperature diode is placed monolithically on silicon along with the power device, it is capacitive-coupled to the high voltage switching node of the IGBT.

The interface circuit must be designed to read the small temperature-dependent VF value, reject the switching voltages and pass the signal across an isolation barrier.

There are many ways of filtering the signal and passing it across the isolation barrier. An example is the use of an isolated amplifier and an analog-to-digital converter. This is combined with synchronous sampling to avoid the capturing of samples occurring with transient noise spikes. (Fairchild is preparing an Application Note dealing with the design of such an interface.)

Current Sensing

Monolithic integrated current sensing is accomplished by measuring the current of a small IGBT in parallel with the main IGBT and then multiplying it by a known scaling factor. IGBTs consist of thousands of cells in parallel via a metallization area on the top of the device. The small IGBT represents a portion of these cells that has been left disconnected from the rest of the cells and is made available as a current mirror.

The basic concept of the current sense function is shown in Figure 2. A separate emitter connection provides a fraction of the main collector current (IS), which is connected to an external resistor (RS). This connection generates a voltage drop, which is proportional to the sense current. The sense resistor voltage is used to determine the current, and by knowing the main collector current (IC), a “sense ratio” (Ratio) is computed, which is given by Equation 2.

$$Ratio = \frac{I_C \cdot R_S}{V_S} \qquad (2)$$

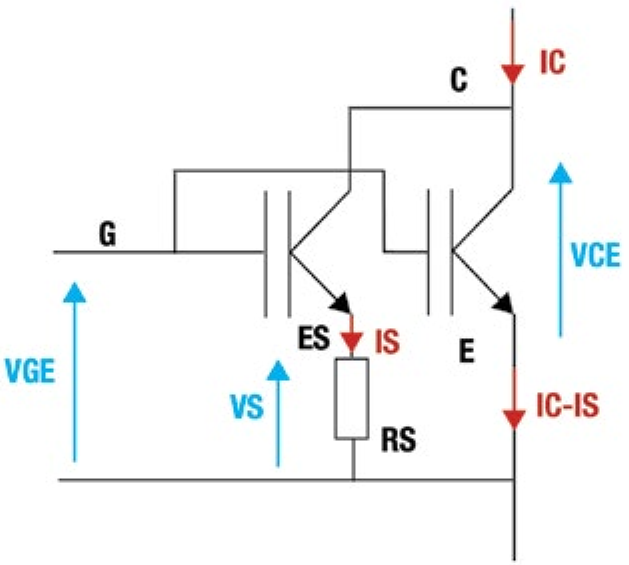

Figure 2: The basic concept of the current sense function

The sense resistor (RS) can be replaced by an op-amp circuit that is designed to directly amplify the sense current in a current-follower configuration, thereby eliminating the sense emitter voltage bias.

The current sense ratio is temperature dependent, but, when combined with junction temperature sensing, it is expected that this interface will yield the necessary information to compensate for the junction temperature and produce a more accurate measure of the main collector current.

A bit more difficult is to correct the current dependency. This causes a limitation of the accuracy at low current, where low current are considered current levels below 10 percent of the full current range. The main limitation of monolithic current sensing is indeed the accuracy at a low current level. More details will be provided in the upcoming application note.

As mentioned above, monolithic current sensing can be put to various uses in an H/EV inverter. The most simple use is overcurrent protection, which may be accomplished easily with a comparator circuit. Such use can augment or replace the traditional desaturation protection.

A much more challenging and potentially valuable use is motor control, which is feasible with additional interface circuitry and intelligent processing in combination with on-die temperature sensing. This application of the monolithic current sense function is in active development at Fairchild. Figure 3 shows some initial results achieved with the on-die current sensing.

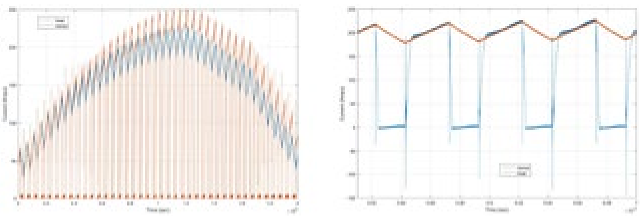

Figure 3: Some initial results achieved with on-die current sensing

Bare Die Layout for Sensing

An example of the bare die layout showing the temperature sense and current sense pads on the top side of the IGBT is given in Figure 4.

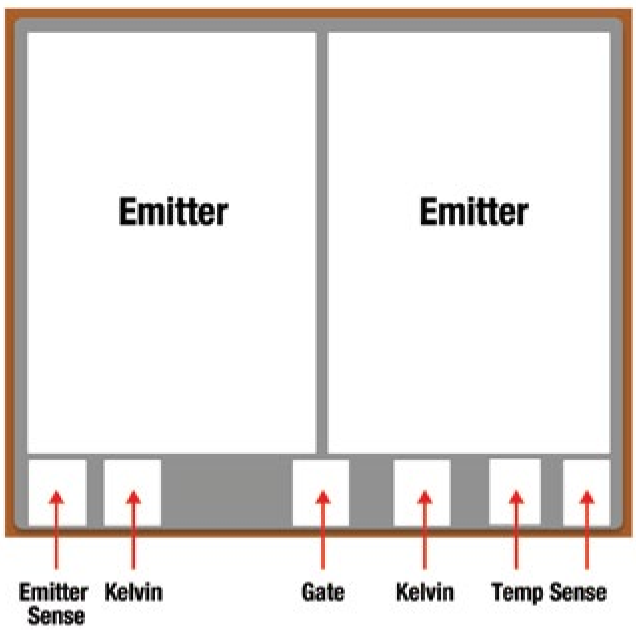

Figure 4: Layout of temperature and current sense pads on the top side of the IGBT

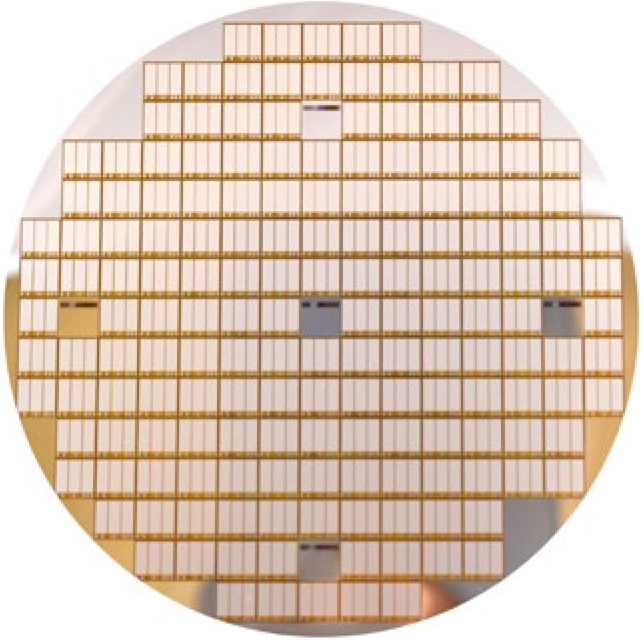

Figure 5: Wafer with bare-Die IGBTs as presented at PCIM Europe, 2016

In conclusion, monolithic integrated current sense and temperature sense circuitry provides a reliable way of measuring the junction temperature of an IGBT in power module applications, and of the collector current - without additional sensors. These techniques can provide substantial advantages for module manufacturers. It simplifies the sensing of key parameters, reduces component count, achieves a faster response to hazardous operating conditions, allows a more precise determination of junction temperature, leads to better utilization of the power silicon, and increases reliability. All of these characteristics will help inverter manufacturers or vehicle OEMs to achieve a more competitive position in the marketplace.

https://www.fairchildsemi.com/applications/automotive/hev

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Rapid Power Prototyping Speeds up the Development of Custom High Performance Power Converters

Development of high performance power electronic systems is a complex task involving the careful design of many different interdependent subsystems such as power stage, gate drivers, fault protection, digital circuits and embedded software.

By Dr. Iain Mosely, Converter Technology Ltd., Pangbourne, UK

Introduction

Most companies are resource limited and under increasingly tight product development timescales. In order to lower the barrier to development and adoption of novel power electronic systems, Converter Technology has developed a rapid power prototyping technology platform to significantly reduce the time needed in the early stage of a project to get to first functional prototype.

Modular Concept

The modular concept shown in Figure 1 allows the first prototype of a new converter to be realised by using custom high performance power electronic modules. System partitioning of the modules has been chosen carefully to separate the key functionality onto individual PCBs with a common interface. The mechanical arrangement allows a fully functional system 'stack' to be built which includes high performance digital control, advanced gate drive/protection and power devices. This approach can prove out system performance at a very early stage in the project. Availability of early stage prototype hardware is critical to de-risking overall project plans.

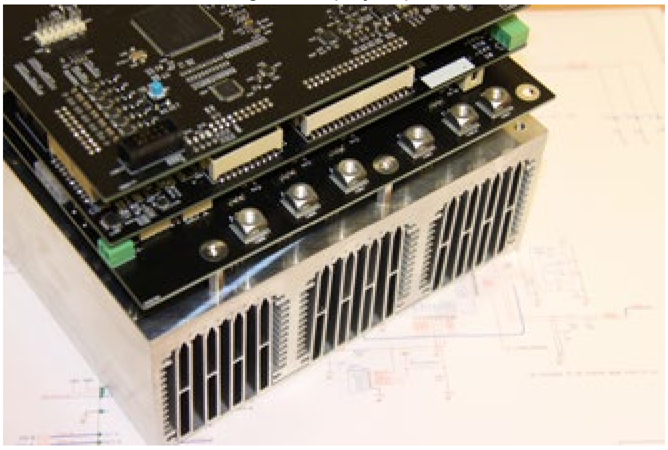

Figure 1: Modular Concept

A number of key features have been implemented into the platform from the outset: -

- Highly modular architecture to allow for rapid system configuration
- SIC MOSFET based power boards currently support up to 15kW in a single system. Systems can be paralleled to provide higher power levels. Power boards with other device types are currently under development.
- Custom advanced digitally addressable intelligent gate drive board suitable for driving many device types including MOSFETS, IGBT's and new wideband gap transistors.
- DC link capability to 800V
- Control board based around high performance 32-bit Delfino processors. The function and behaviour of the stack is defined in software and allows reconfiguration as the core part of most common isolated and non-isolated topologies including bidirectional two or four quadrant converters. Examples include three phase AFE/inverter, multiphase buck or boost, resonant converters such as phase shifted full bridge or LLC and hard-switched converters such as full bridge.
- The control board includes three 12-bit DACs which allow digital values within the control system to be output for display on a scope in real time to aid system debug and analysis.
- Robust isolated CANBUS communications to ease integration into Labview automated test environment.
- Extensive fault monitoring and protection to allow the power stage to independently look after its own environment if key operating stress levels are exceeded. This protection prevents bugs in software during application development from damaging the power stage.

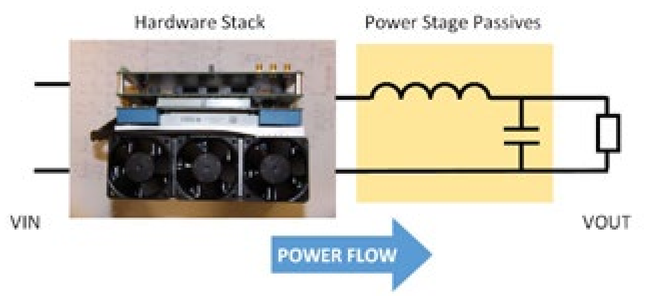

Figure 2: Synchronous buck with external inductor and capacitor

In addition to early stage prototyping, the platform acts as a useful tool to aid in depth learning about how different power topologies operate in practice. Since the behaviour is software defined, the impact of switching frequency, duty cycle and phase relationship can be easily explored in real time on real hardware. Closed loop current or voltage control can be implemented along with custom soft-start algorithms.

Example Application – HV Synchronous Buck Converter

As an example of the flexible capability of modular technology, a stack was configured to explore the complex factors impacting conversion efficiency in a high voltage synchronous buck converter. The present

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stack system power board supports three Cree SIC MOSFET half bridge legs so, for this set of tests, two of the switching legs were disabled and one leg configured to operate as a synchronous buck converter. An external power inductor and capacitor were used to complete the Synchronous buck power stage as shown in Figure 2.

In this example, the synchronous buck converter is operated under hard-switching conditions and of particular interest was the behaviour of the power conversion efficiency as a function of both load current and switching frequency. The CANBUS port was used to allow a custom Labview script to vary the power stage switching frequency as part of an automated test.

For the test, the input supply rail was set to 400V, the demand output set to 300V and efficiency was measured from input to output as a function of load current from 0 to 10A and switching frequency from 20kHz to 180kHz. The ability to automate the data capture allows many data points to be captured in a short timeframe. In the example here, 4000 individual efficiency measurements were taken over the range of load current and switching frequency in a two-hour period. Figure 3 shows the behaviour of the efficiency as a function of the two key variables whilst Figure 4 shows an equivalent contour plot highlighting regions of constant efficiency.

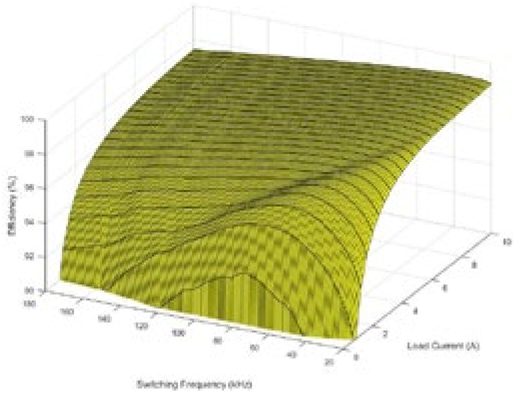

Figure 3: Efficiency as a function of two variables

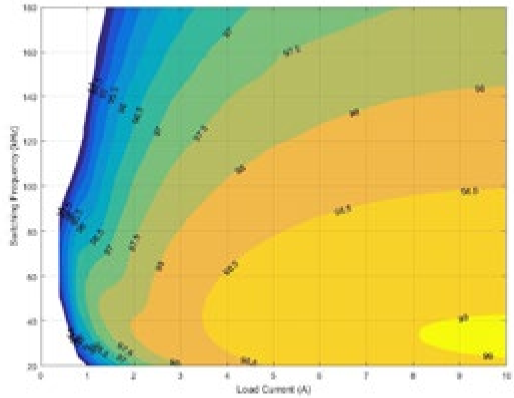

Figure 4: Regions of constant efficiency

The actual shape of the surface plot shows an interesting region at lighter load currents where the efficiency peaks and then falls away. Consideration of the operating waveforms indicates that the region corresponds to the point whereby the forced synchronous operation results in a negative inductor current for part of the operating waveform. With bipolar inductor current, each switching transition becomes lossless due to natural commutation of the node voltage during the dead-time. This provides a temporary boost to the efficiency before the reduction in load current dominates the efficiency profile and efficiency falls again. The boundary conduction mode point is a function of switching frequency and load current as shown in equation 1.

$$IOUT = \frac{Vin \cdot \delta \cdot (1 - \delta)}{2 \cdot Fs \cdot L}$$

Where Vin = Input Voltage (V), δ = duty cycle (p.u.), Fs = Switching Frequency (Hz) and L = Output Inductance (H)

For this particular example, the island of highest efficiency of 99% occurs towards the maximum load current and at the lower end of the operating frequency range. Performance over the wider range of operating conditions is excellent. Figure 5 shows the clean operating waveforms possible with SiC devices with the switching node voltage shown on Channel 1 (Yellow) and inductor current on Channel 2 (Red).

Overall efficiency of the converter is impacted by all power loss mechanisms and these mechanisms are functions of the particular operating point of the converter. For example, wide bandgap devices such as SIC or GaN have low body capacitance and natively high performance body diodes with negligible reverse recovery charge. This makes these parts suitable for high frequency operation since reverse recovery and capacitive losses are minimal. IGBT's and traditional MOSFETS would most likely show a stronger system level degradation in efficiency at higher frequencies which would change the location of the island of efficiency to a different point in the solution space.

Figure 5: Clean operating waveforms possible with SiC

IGBT switching loss is very much higher than MOSFETs and wide bandgap power devices so will tend to have an island of maximum efficiency at much lower frequency. However, at high load currents, the conduction loss of an IGBT can be much lower than that due to the Rdson of a MOSFET. For high current applications, IGBT based converters would show an efficiency plot which may exceed that of other technologies but only if the frequency was low and the demand current high enough to allow the conduction behaviour of the IGBT to dominate the loss mechanisms.

The loss mechanisms in the passives will be dominated by the inductor loss. Core loss and winding loss are strongly dependant on switching frequency and the ratio of DC to AC currents flowing. In general, a given inductor loss profile will reduce as the frequency comes down. However, at a certain point, the increased flux density variation will begin to dominate and the losses will begin to rise. The point of transition will be dependent on many aspects of the inductor design and their impact on overall efficiency will depend on how much other external loss mechanisms change with frequency.

Conclusions

Rapid prototyping of custom power electronic systems using modular technology can significantly reduce risk in new system development. Furthermore, the flexibility of software defined function within the modular system allows the system to be re-tasked easily and can form the heart of a useful learning and experimentation platform. Converter Technology has used this approach now with multiple applications including high power LLC systems. Further applications under development using the platform include an Active Front End and multi-phase converters.

The system has proved robust and significantly enhances the productivity of the hardware teams using it.

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The Flyback Dc-Dc Converter

The best bet for most SMPS"

The Flyback converter combines lowest cost with best performance and is the optimum choice for offline SMPS up to more than 250 W. The simplicity of the circuit diagram is deceptive, it is highly complex. This article is a practical guide to optimum designs and lists the pitfalls.

By Dr.-Ing. Artur Seibt, Vienna

1. The facts establishing the superiority of the flyback (FB) converter.

The name is a misnomer to start with and a reason why the FB is underestimated more often than not. The apparent simplicity of its circuit diagram veils its potential well as its complexity. There is more to it. A proper name would be Storage Converter. The name stems from the deflection/hv "flyback" transformers of previous tv sets. Already then the design of the transformer turned out to be a major task: defective transformers used to be one major cause of tv set failures, it took designers years before they succeeded in understanding and taming flyback transformers and to make them reliable components!

In principle, nothing has changed: FB design is as demanding as ever. Even renowned concerns still have to recall products due to power supply design faults like burning transformers. Over decades, an abundance of literature has accrued, however most publications treat just one design aspect, many are purely theoretical and useless for the designer, some are plainly false. Outright dangerous are publications of some semiconductor companies which try to persuade designers to use their ic's by belittling statements like : "... the design phase of a FB has almost turned to child's play ... or "... even a novice can...". In a 2011 sales promotion paper for FB offline control ic's it says: "The output voltage is defined by two external resistors and the turns ratio of the transformer!" If a company has not understood the function of a FB and of the transformer, can its expertise and its products be trusted? The same semiconductor company calls 125 C " an operating junction temperature". Such statements are fatal: FB design requires profound professional knowledge and deep insight into the properties of active and passive components, especially transformer design and materials. The transformer is the major component which decides the SMPS performance. Most important: circuit and transformer design can not be separated, the latter can not be left up to a transformer manufacturer.

The purpose of this article is a comprehensive explanation of the FB and solid practical advice for the design engineer with special regard to young engineers who will, as a rule, not receive such knowledge during their education. As will be shown, the FB combines lowest cost and best performance, especially in its most important application in offline dc/dc converters, so its study is well worth the effort in order to arrive at optimum designs and sidestep the many pitfalls along the way. Too many FB designs are far from optimum but function somehow anyway, why? The stunning truth is that even grossly wrongly designed transformers can "function" and disguise this fact.

Advantages:

1. FB's offer by far the most cost-effective solutions. A FB requires the minimum number of components. As a rule, the inductive components head the BOM, the FB transformer combines the functions of transforming, energy storage and galvanic isolation in one component, all other converters need at least two.
2. FB's are the fastest dc/dc converters, especially when operated with current-mode control (invented by the author at Tektronix 1969, US Patent). It is a widespread misconception that SMPS are by nature slower than linear regulators, if that is so the SMPS is just wrongly designed. The regulation quality of a professionally designed FB SMPS is excellent and by far sufficient but for a few most exacting applications. This translates directly into a cost, space and efficiency advantage because no postregulators are necessary!
3. Current-mode control also makes the FB independent of its input voltage without help from the regulation loop, thus creating wide-range input converters which accept all line voltages, i.e. 90 to 254 V. The dependence of the output voltage on the line voltage is essentially zero.
3. FB's can deliver multiple additional output voltages at lowest parts count and cost. An output requires only a winding, a diode and a capacitor; if no galvanic isolation between secondaries is needed, taps on a single winding suffice, reducing the cost to a diode and a capacitor. With intelligent transformer design, those voltages can be so well stabilized that postregulators can be dispensed with also on these outputs, again saving cost, space and losses.
4. The FB can offer galvanic isolation and generate any number of voltages below and above the input voltage. Also autotransformers and simple chokes may be used

Disadvantages and limitations:

1. High ripple current load on windings, active devices and the capacitors, caused by the two-phase operation: during the primary charging phase the load must be fully supplied out of the capacitors, FB's are unsuited for the generation of small voltages at high currents. Such requirements are best met by designing the FB for a higher voltage like 24 V and by adding a simple buck. The ripple can be reduced by interleaving two or more FB's with a multiphase clock. The higher ripple voltage requires an output filter. The step-up or boost converter suffers from the same disadvantage. Many if not most power supply failures are due to overstressed electrolytics which are literally "too hot to the touch" because the designer lacked the knowledge about capacitor life and and probably also how to correctly measure the ripple currents. Ripple currents in multiple-output SMPS are tricky because they are interdependent

and their waveforms and rms values do not at all correspond to what the textbooks assume.

2. High voltage stress on the active components; stresses are a matter of transformer design and the choice of active components. This applies to the simplest form with one switching transistor; the two-transistor switch clamps the transistor voltages to the input voltage.
3. Difficult and critical transformer design, requiring expert knowhow. Its core material is driven close to saturation requiring high quality ferrite. Leakage inductances and capacitances are critical, the high voltages and short rise and fall times stress the insulation materials, skin and proximity effects have to be reckoned with.
4. The high voltages and rates of rise in an offline FB SMPS as well as the resonances cause emi which requires damping and often a more expensive emi filter than with other converters.

2. What is so special about the FB transformer?

Transformer design will be treated in ch. 6, but already here the reader shall be alerted to its overwhelming importance:

1. The transformer determines the operating mode, not the circuit designer nor any fancy ic!
2. In the first place it is not a transformer, it only looks like one; in fact it is a storage choke with - mostly - separate charging and discharging windings. The energy is stored in the core's air gap. Unlike a true transformer, the turns ratios between primary and secondaries have no bearing whatsoever to the primary and secondary voltages ! Numbers of turns and turns ratios are determined by many conflicting parameters. In order to reclaim the full energy stored during charging minimal leakage inductance is crucial; any energy remaining in the leakage inductance gives rise to high overvoltage spikes and ringing across the active components requiring damping.
3. The fact that numbers of turns and turns ratios have nothing to do with primary and secondary voltages has surprising consequences. A FB transformer may be grossly wrongly designed, this can go by unnoticed, because the regulation loop will always enforce the desired output voltage. Overvoltage due to false turns ratios will destroy active components upon turn-on. Unless the smell of melting insulation points to a problem, the transformer may seem to function properly. Core saturation may be delayed until warm-up under full load and then destroy the switch. After replacement of the transistor the FB will function again, so the designer will attribute the failure to a bad transistor - until warm-up. This is the explanation why FB transformers "designed" by software or trial and error mostly function.

   There is no such thing as a "24 V FB transformer" as can be read in semiconductor firms' design papers. A FB transformer, in sharp contrast to a 50 Hz transformer, can output any voltage at any current provided that neither the core nor the windings nor other components become overstressed by just changing the loop's reference divider, i.e. one resistor.
4. And it is also a transformer. During the first phase, the charge, the applied primary voltage is transformed by the associated turns ratio to the secondary and stresses the rectifier diode by the sum of the output and the transformed voltages plus spikes. In turn,

during the second phase, the discharge, the conducting secondary diode places its output voltage across the secondary winding which is transformed by the inverse turns ratio into the primary stressing the switching transistor by the sum of the input and the transformed “reflected” voltages plus spikes.

5. Optimum transformer design is always an iterative process, there is no such thing as a set of formulas or a simulation software which will yield the final winding prescription. Again it must be warned against most semiconductor firms’ design recipes. One such firm, e.g., recommended in writing an operating ferrite flux density of 4300 Gauss (0.43 T). This is already the saturation value of standard ferrites at 25 C.This means that a transformer designed along these guidelines will immediately destroy the switch and then burn.

3. Basic circuits and methods of control.

There are 2 methods of control (VMC = voltage mode control and CMC = current mode control) and 2 operating modes (DCM = discontinuos mode and CM = continuous mode) , hence there are 4 versions of the FB converter. This is the iron rule of the FB:

► The method of control is chosen by the designer, but the operating mode is solely determined by the transformer.

From this rule it follows that circuit and transformer designs must never be separated, the transformer design can not be left up to an external supplier.

3.1 Voltage Mode Control (VMC).

Fig. 1 shows the basic FB with Voltage Mode Control (VMC). it consists of: transformer = storage choke, primary switch, secondary rectifier diode, control circuit. The control circuit turns the switch on, the input voltage is applied across the primary winding, the primary inductance is charged up by the sawtooth current given by: i = L x dv/dt. Due to the polarity the secondary diode is off, the load is supplied during this time out of the capacitor. Before saturation is reached, the control circuit turns the switch off. The current in an inductance can not change abruptly, it will continue to flow, increasing the voltage across the primary and the switch and reversing the voltage across the diode, forward biasing it. The charge stored will now be emptied into the capacitor and the load.

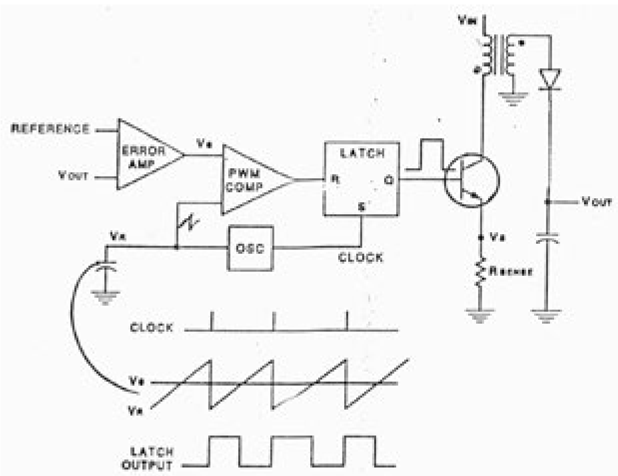


Fig. 1. Basic FB circuit with voltage mode control.

► The FB and its relative, the boost converter, in sharp contrast to other converters, are two-phase converters, i.e., energy is not continuously transferred, at no time is there a direct energy transfer by transformer action.

A FB can also be realized with a single choke or an autotransformer if no galvanic isolation is required. Note that the hardware necessary is the same as for a buck or boost: 3 components, the parts are just differently arranged. The transformer is nothing else but a choke with separate charging and discharging windings which immediately points to the importance of tight coupling resp. low leakage inductance. Also it is obvious that the two-phase operation implies that during phase 1, the charging phase, the load is fully supplied out of the capacitor, causing high ripple stresses and a residual ripple voltage; for sensitive loads a LC filter will be necessary. Due to the high frequency mostly two ferrite rods with a one layer winding will be sufficient, providing series and common mode suppression.

The classic control method VMC requires the 4 components shown if no galvanic isolation is provided. Simpler versions of the control circuit do without the flip-flop, but there is the danger that the switch is again turned on within the same cycle. A RS flip-flop controls the switch. An oscillator sets the flip-flop at the beginning of a cycle. An error amplifier compares the output voltage or a fraction thereof with a reference, a comparator compares the amplifier output with the timing sawtooth of the oscillator, if the momentary value of the sawtooth exceeds the amplifier output, it will reset the flip-flop and terminate the cycle for good. The control generates a duty-cycle-modulated switching signal. It is obvious that, within one cycle, there must also be time for charge removal. The transformer current is indirectly limited because the ic’s limit the duty cycle to 50 to 67 %. Full protection requires the additional use of the signal across RSense .

The regulation loop is fairly slow because it reacts to the output voltage which is obtained by averaging the cycles. The loop has to compensate for all influences, in particular also variations of the input voltage.

3.2 Current Mode Control (CMC).

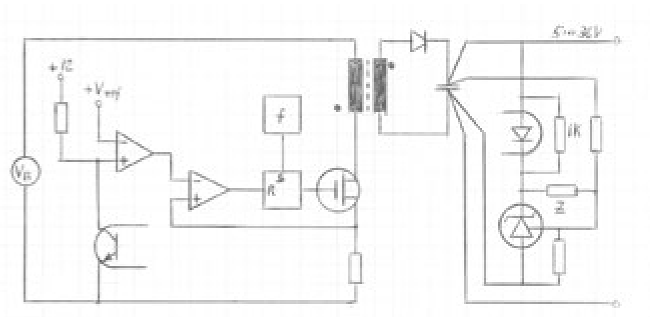


Fig. 2. FB converter with current mode control and galvanic isolation.

Fig. 2 shows a complete FB with current mode control (CMC), with the additional components needed for galvanic isolation. CMC is characterized by a second regulation loop underlying the first slow loop for keeping the output constant. Here, the flip-flop steering the switch is also set at the beginning of an oscillator cycle, but the cycle is terminated by a comparator which compares the current-proportional signal across the sensing resistor RSense with the output of the error amplifier. This loop is as fast as the clock frequency allows, in principle, this loop could follow any output variations from cycle to cycle; in practice the loop bandwidth amounts to about 1/6 to 1/5 of the clock frequency which is much faster at > 100 KHz frequencies than the loop of an ordinary linear regulator. The overlying output voltage loop feeds into the first loop through the error amplifier; its output level determines the peak current in the transformer resp. switch.

Precisely speaking the transformer current rises only up the value which is necessary to supply power to the actual load. The output power is given by:

$P_{DCM} = 0.5 \times L_p \times i_p^2 \times f$ in DCM resp. $P_{CM} = 0.5 \times Lp \times (i_{pmax}^2 - i_{pmin}^2) \times f$ in CM.

This has quite a few consequences. in the first place, variations of the input voltage have no influence because they cause only changes in the charging time resp. slope of the charging sawtooth but not on the amount of charge; this comes about automatically without that the voltage loop has to interfere which has only to respond to load changes. Thus a FB with CMC provides a wide-range input at no cost. By design it must be assured that at the lowest input voltage there is still enough time to charge up to the level required by the maximum load. At a given clock frequency this poses limits on the maximum Lp and Ls ; this will be treated in detail later.

CMC features built-in protection against transformer saturation and switch overload which in turn means that, with CMC, the smallest und least expensive transformer can be used! With VMC, the transformer must be designed for a much higher peak current at maximum ferrite temperature so it must be bigger and more expensive. This reserve is necessary because the onset of saturation is very steep and fast, so fast that many protection circuits in control ic's can not prevent switch destruction! Also thermal runaway is possible. Si mosfet's are subject to a di/dt limit, this is mentioned in the data sheets but often overlooked. They all contain a parasitic npn in parallel to the mosfet proper; its base is short-circuited to the emitter, but the current through the drain-to-base capacitance increases with dv/dt and eventually creates a voltage across the short that turns the npn on. This causes immediate destruction by secondary breakdown. The message from the mosfet manufacturers that mosfet's did not suffer from secondary breakdown has been always false. This must not be mixed up with the avalanche rating.

To the right of Fig. 2 there is the standard secondary control circuit. The inexpensive TL 431 has become a virtual standard because its performance is so excellent that more expensive solutions make no sense. It contains a reference of 2.5 V and an error amplifier: if the input voltage exceeds 2,5 V, a power transistor good for 100 mA will be turned on across the terminals cathode and anode. The LED of the optocoupler is connected in series with the TL 431 cathode, a divider steps the output voltage down to 2,5 V If the output tries to rise, the TL 431 will increase the LED current. The transistor of the optocoupler will be turned on harder and decreases the voltage at the second input of the comparator so the peak current will be reduced.

This loop will be quite fast if the compensation Z across the TL 431 is correct. A 220 p capacitor will short the TL 431 for the clock frequency, a series RC is always sufficient, a few nF and a few to some ten K. One often sees a 0.1 uF capacitor across the TL 431 - with a shudder! The TL 431 requires a minimum of 1 mA, as the LED voltage will be around 1 V, a 1 K resistor is sufficient.

Note the correct method of connections at the capacitor. The voltage pick-off is always taken from the first capacitor and never behind a LC filter if any, the delay would create stability problems and lead to still slower regulation. The ripple on the first capacitor must be kept very low, otherwise it would overdrive the amplifier which would cause nonlinear and poor behaviour of the loop, a dangerous pitfall seldomly recognized and a major cause of the belief that SMPS have poor and slow regulation. These amplifiers are lf designs in the nature of a LM 224 which can neither react to nor digest 100 KHz ripple. Any loop compensation with a capacitor bypassing the output voltage divider aggravates this problem. Long experience has proven that if the circuit design and its layout on the e.c. board is professionally done, such additional feedforward compensation is never required; to the contrary, if it seems to be necessary this is a clear sign of design or/and layout problems. Even with VMC, if the input voltage is fairly constant, e.g. by a preceding PFC, static output deviations are in the vicinity of 10-4 . Load steps e.g. from 10 to 100 % cause output drops of maybe 10-3 for a few ten µs. With CMC the performance is still better. Often is it argued that such high performance is not necessary in a power supply, this misses the point entirely: this performance comes about at no cost, automatically, and is the proof of correct design.

4. Operating modes.

4.1 DCM = discontinuous mode.
It is repeated that the designer - or his selected control ic - determines VMC or CMC, but it is the transformer which dictates absolutely whether the FB runs in DCM or CM! If the transformer design is farmed out to a transformer manufacturer, he fully interferes with the designer's job. This way never an optimum transformer can result, and the responsibility for the whole FB converter design will be split. It is the designer's job to design the transformer and to hand the manufacturer a winding prescription which the latter has to follow, period. This does not preclude that the designer will accept advice about the material properties of ferrites, wires, insulation and approved insulation systems.

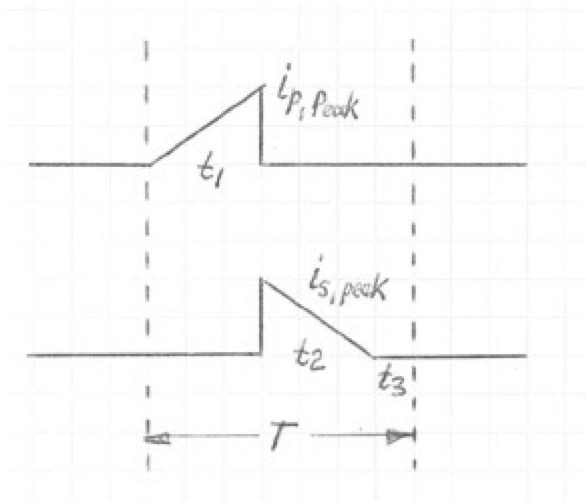


Fig. 3. Waveforms and definitions for DCM.

► DCM only if t1 + t2 > T, i.e. t3 > 0.

Contrary to many texts DCM is not guaranteed by just a small Lp .

Discharge must be completed before the start of the next cycle, t3 > 0. A changeover into CM will entail a number fof consequences

As will be discussed later, DCM operation has some advantages, so which are the conditions? In DCM the FB converter operates in pure two-phase mode, i.e. the transformer is charged and fully discharged in every cycle. After discharge, there is no current in any winding until the start of the next cycle. During charge only the primary circuit is active, during discharge only the secondary circuit. DCM is the "true" FB operating mode. These are the criteria for DCM:

These are the general formulae, losses neglected:

$i_{peak,p} = V_B / L_p \times t1$ $i_{peak,s} = V_s / L_s \times t2$
$\tau = t1/T$ = duty cycle   N: turns ratio R_L : load

$I_s = I_{output} = 0.5 \times i_{peak,s} \times t2/T = VB \times N \times [\tau (1 - \tau)/2fL_p]$

$P_{out} = 0.5 \times \tau^2 \times V_B^2 / fL_p = 0.5 \times L_p \times i_{p,\, peak}^2 \times f$

$t1 = 1/V_B \times \sqrt{(2PL_p / f)}$ $t2 = \sqrt{(2L_s / fR_L)}$
$V_{out} = V_B \times \tau \times \sqrt{(R_L / 2fL_p)}$

$I_{DC,in} = \tau \times i_{p,peak} / 2$ $I_{AC,in} = i_{p,in} \times \sqrt{(\tau/3 - \tau^2/4)}$
$I_{rms,in} = 0.577 \times i_{p,peak} \times \sqrt{\tau}$

Note: V_{out} is no function of N (nor of n_p or n_s)!

Special case: t1 = t2 = T/2.

In principle, the percentages of a period for charge and discharge can be freely chosen, but this special case is a good starting point both for DCM and CM. At this point it should be noted that only such control ic's should be used which limit the charging period to 50 to 67 %; this ensures some protection in case of overload and start, limiting the peak current and by allowing time for discharge. For a distribution of 50 % - 50 % these formulae apply:

$L_{p,max} <= V_B^2 / 8Pf = 2P/(i_{p,peak}^2 \times f)$ $i_{p,peak} = \sqrt{(2P/fLp)}$
$I_{p,rms} = 0.4\ i_{p,peak} = 1.64\ I_{average}$

$I_{p,av} = I_{p,DC} = P/V_{B,min}$

$L_{s,max} <= V_s^2 / 8Pf$ $i_{s.peak} = \sqrt{(2P/fL_s)}$
$I_{s,rms} = 0.4\ i_{s,peak} = 1.64\ I_{s,av}$

All texts known to the author state that DCM is guaranteed if $L_{p,max}$ is obyed. This is not true, because $L_{p,max,50\,\%}$ only guarantees that the charge will be terminated at T/2, nothing more!

How long the discharge will take resp, if it will be terminated within the remaining t2 is determined solely by the secondary circuit, independent of the primary circuit! Which follows from the strict two-phase operation in DCM where each phase is independent of any other. It should be borne in mind that the FB transformer is a storage choke with (mostly) separate charging and discharging windings. In principle, the energy stored can be discharged via the same winding, with associated switches; also the "secondary" winding can be used for charging; this would last as long as the discharge.

Because N has nothing to do with the primary and secondary voltages, the secondary number of turns and the discharge speed can, in principle, be chosen freely. The discharge speed is given by dis /dt = Vs /Ls which yields the discharge time t2 = is,peak /(Vs /Ls), valid for the normal situation with a very large output capacitor which averages the high pulse currents from the transformer and provides a fixed discharge voltage. is,peak is determined by the primary current at turn-off resp. Ls or N. As soon as the diode conducts, the output voltage is switched onto the secondary winding, discharge takes place against this voltage in a linear fashion.

It becomes apparent that the design task is quite complex and as a rule iterative, and that the simple "recipes" in most texts will not yield an optimum design, if at all a functioning one. No fancy ic can enforce DCM if the transformer does not allow it. Many interdependent requirements have to be fulfilled simultaneously. Assuming that he designer knows τmax because the ic used will terminate the charge at 67 %, a practical starting point for the partition of T would be 50/50 or 40/60, using the general formulae for t1 and t2 and the given values for f and P, Lp,max and Ls,max can be calculated. Whether this can be realized will show up during transformer design which may require shifting t1 and t2. In any case the following rule must be obeyed under all combinations of input voltage and load:

► DCM requires both limits Lp,max,t1 and Ls,max,t2 for t1 + t2 < T!

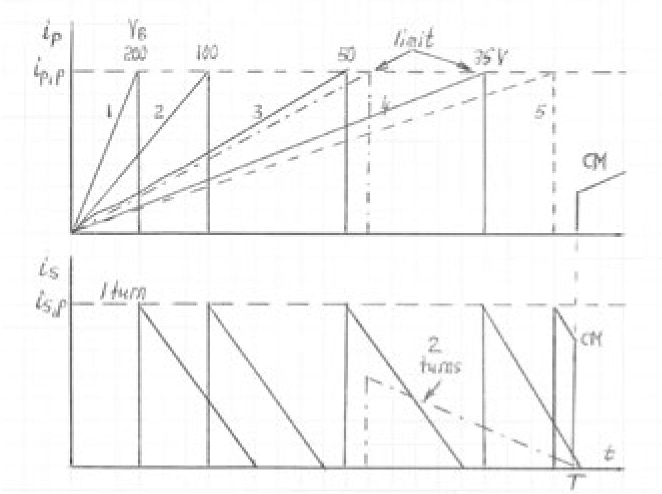


Fig. 4. Example of a FB with CMC.

Fig. 4 shows a practical example of a FB with CMC. If the load is constant, the transformer will always be charged to the same primary peak current ip,peak . The input = charging voltage has no influence on the energy stored nor the output voltage. If the input voltage is reduced only the charging time is increased; the voltage regulation loop does not need to compensate for the input voltage variations. CMC offers a wide-range input at no cost. The picture shows several cases from 1 to 5. The lower part shows the secondary current for 1 turn. The converter stays well in DCM down to an input voltage of 35 V, below it jumps into CM as shown by curve 5; here a high secondary discharge current remains at the end of the period, resulting in a corresponding primary current at the start of the next period. The transition into CM was caused here by the primary side, but it can - quite independently - also be caused by the secondary. This is shown by the dashed waveforms which apply to 2 secondary turns, increasing Ls by a factor of 4, halving $i_{s,peak}$ and doubling t2, so the transition happens just below 50 V.

This shall be further clarified by this example:
P = 10 W, A_L = 250, V_B = 360 V, f = 100 KHz, V_{out} = Vs = 5 V: From these numbers follow:

$L_{p,max}$ = 16.2 mH, $i_{p,peak}$ = 0.11 A, di/dt = 22.2 A/ms. i_s runs up to 0.11 A in t1 = 5 μs = T/2. The primary number of turns n_p = 254 follows from n = √(AL /L).

Assuming 1 secondary turn n_s = 1, L_s = 0.25 μH, N = 254:1, $i_{s,peak}$ = 28.2 A, di/dt = 5 V/0.25 uH = 20 A/μs, t2 = 1.41 μs. t1 + t2 = 6.41 μs < 10 μs, hence DCM. Iout = $I_{s,av}$ = 0.5 x $i_{s,peak}$ x 1.41/10 = 2 A x 5 V = 10 W.

Now, 1 turn is unacceptable in practice, the coupling would be miserable, so 2 turns are selected: $i_{s,peak}$ = 14.1 A because N = 127. L_s is increased 4-fold to 1 μH, di/dt = 5 V/1 μH = 5 A/μs, t2 = 2.82 μs. t1 + t2 = 7.82 μs < 10 us, still DCM.

2 turns still yield no acceptable coupling, so 4 turns are tried: $i_{s,peak}$ = 7 A, t2 = 5.6 μs, t1 + t2 = 10.6 μs > 10 us. Not feasible, would entail CM! How the problem of unacceptable coupling with only a few secondary turns is solved in practice will be described in ch. 6.

If the output voltage shall be 7 V instead of 5 V, the discharge speed increases to 7 V/4 μH = 1.75 A/μs, so t2 = 4 μs; t1 + t2 = 9 μs < 10 μs, again DCM. This example illustrates that the numbers of turn have nothing to do with the voltages, but the change to 7 V causes the primary overvoltage $V_{reflected}$ to rise from 5 V x 63.5 = 317 V to 7 V x 63.5 = 444 V which comes on top of the 360 V and stresses the switch now with 1044 V. This set of numbers is hence unacceptable in practice alone for this reason; in order to comply with the specs of standard mosfet's, e.g. 650 V for Coolmos, N must be drastically reduced so the reflected voltage stays below 650 - 360 = 290 V. This in turn requires increasing n_s which, as shown, would cause CM. However, the example also shows that the output voltage of a FB can be changed just by changing the reference divider in the regulation loop, but, in this example this would destroy the primary switch. Also the blocking voltage of the diode will be increased from 5 V + 360 V/63.5 = 10.7 V to 12.7 V.

► In order to guarantee DCM, it is necessary to check the whole range of input voltage and load whether the sum of t1 + t2 remains < T; most critical are $V_{B,min}$ and P_{max} .

1. Most noteworthy is the fact that the FB in DCM is a constant-power resp. constant-current source, note that the load does not appear in the formula for P_{out} . The charge put in must be consumed before the next cycle; if there is too low a load or none, the output voltage will rise towards infinity, in practice until some component will be destroyed. In order to realize a constant-voltage source, a feedback loop has to be added, which in turn means that a protective means like a transil zener diode has to be provided in order to save sensitive or expensive loads in case of loop failure! In some applications where constant power or current is asked for, a FB in DCM can be directly used; examples are LEDs, high-pressure gas discharge lamps.
2. There is no memory in the form of stored energy from the preceding cycle(s). Therefore the FB in DCM is extremely fast, especially with CMC. No other offline converter can compete.
3. The formula for P says that the 3 factors L_p , $i_{p,peak}^2$, f may be interchanged , but great care is advised: E.g. if $i_{p,peak}$ is reduced to ½, P will drop to ¼ which will require to increase Lp by 4 which means to double np . Θ = i x n remains unchanged. If e.g. ip,peak has to be reduced with regard to the mosfet used, an increase in np will not be possible because the available winding space will not suffice, also 4 x L_p will probably cause CM. An increase in f allows a reduction of Lp resp. np , B will be reduced, so a smaller transformer might be possible, but losses will increase.
4. The regulation loop may act upon τ (VMC), f or ip,peak . CMC influences ip,peak , so the influence is quadratic. There are varieties with constant on or off times which in general lead to variable frequency operation. The design of the loop is simple, but care must be exercised not to enter CM, because the loop may become unstable.
5. DCM suffers from the disadvantage that Bpeak and ip,peak are higher than in CM with ensuing core losses, also the best ferrites

are needed unless P is small, and that its rms currents are higher. The FB's greatest disadvantage remains in any case that it causes the highest rms currents stressing the electrolytics, the windings and the switches. The electrolytics are stressed additionally because, during the charging phase, the load must be fully supplied by them. The fact that the electrolytics are partly discharged and recharged in every cycle is especially deleterious to their life, a serious fact, too often overlooked. The load on the electrolytics requires special attention during the design, also the choice of manufacturers. There is so much black knowhow in electrolytics that it is highly risky to use unknown sources.

6. The fact that the secondary current drops to zero allows the use of diodes with standard recovery which have lower forward voltages and are less expensive. Again, involuntary transition into CM must be prevented.
7. Duty cycles > 50 % may cause instabilities by subharmonics; many texts point this out and make a great deal of this; in practice this can mostly be neglected, because it remains nearly invisible in the output voltage if the loop is properly designed, and this counts. Control ic's which switch modes back and forth create real problems.
8. High input voltages and low loads cause very short charging pulses, eventually cycles will be skipped. Properly designed SMPS will remain stable, i.e. it is by no means necessary to switch to a burst mode which causes undesired high amplitude low frequency ripple.

4.2 CM = continuous mode.

While DCM is fairly simple, CM is rather complex, and there are a multitude of more or less false descriptions around, especially from semiconductor firms. In Germany, there is a saying: "Keep to the business you know!", an advice these firms should heed, 80 % of their circuit design proposals are worthless and misleading. Frequently, CM is recommended as the "better" operating mode and, the deeper into CM, the better, a wrong advice.

.► CM mode is entered if $i_{s,min} > 0$, resp.
if $t1 + t2 > T$

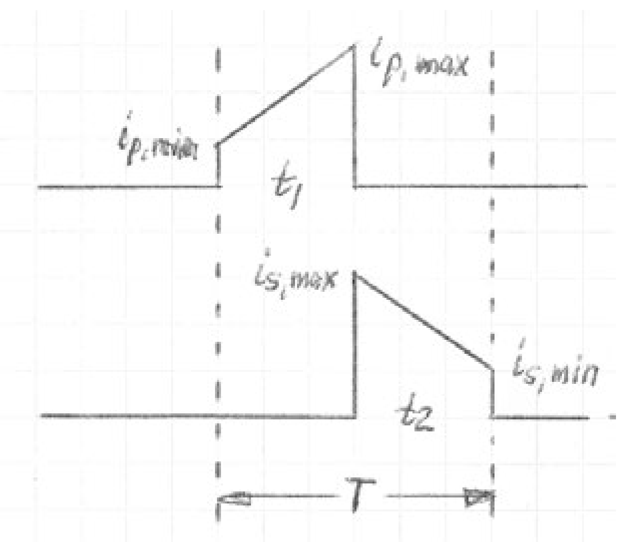


Fig. 5. Waveforms in CM

The fundamental difference to DCM is the fact that the transformer is not discharged at the end of a cycle, but there is still an energy and an associated is,min left and carried over into the following cycle which does not start at zero, but at ip,min defined by N. The waveforms change from sawtooths to trapezoids. The nature of the FB is changed with a series of consequences. As mentioned before, CM can be entered unintentionally, e.g. if a FB in DCM is overloaded, or if Lp or/and Ls were incorrectly chosen. On the other hand all FB's in CM will enter DCM at some weak load or other.

The general formulae for CM are:

$i_{p,max} - i_{p,min} = \Delta i_p = V_b / L_p \times t1$ $\qquad$ $i_{s,max} - i_{s,min} = V_s / L_s \times t2$

$I_{p,av} = I_{p,DC} = \tau \times (i_{p,max} - \Delta i_p /2)$ $\qquad$ $I_{p,rms} = (i_{p,max} - \Delta i_p /2) \times \sqrt{\tau}$

$I_{p,AC} = (i_{p,max} - \Delta i_p /2) \times \sqrt{(\tau \times (1 - \tau))}$

While in DCM each period is independent, the residual energy at the end of a period in CM means that the converter has a "memory" of the preceding period(s), precisely speaking it remembers the previous load. Basically, a FB has no knowledge of the load during its charging phase, but here a starting current $i_{p,min}$ reflects the load from the preceding cycle which is, of course, unchanged from cycle to cycle. In other words: In CM the load is present by $i_{p,min}$ in each charging phase as shown by this form of the formula for the power:

$$P_{out,CM} = 0.5 \times \tau^2 \times V_B{}^2 / fL_p + \tau \times V_B \times i_{p,min}$$

Although the FB in CM still operates in two phases charge - discharge, is reacts in each phase to the actual load. This is the property of a constant-voltage converter.

► The transition from DCM to CM means that the FB changes its nature from a constant-current to a constant-voltage converter!

5. Comparisons of both modes.

5.1 Output voltage.

The output voltage is given by:

$V_{out,CM} = V_B / N \times \tau/(1 - \tau)$ $\quad$ $\tau = t1/T$ $\quad$ N = turns ratio

$I_{out} = 0.5\ N/fL_p \times \tau\ (1 - \tau)$

This is a most remarkable result: Neither the load RL nor the output current Iout appear in this formula. VB and N are constants, so the output voltage depends only on τ/(1 - τ), it is stable as long as τ is not changed! The output voltage of a buck depends linearly on τ. The function x/(1 - x) is nonlinear:

x =	x/(1 - x) =
0.1	0.1
0.2	0.25
0.3	0.43
0.4	0.67
0.5	1
0.6	1,5
0.7	2.3
0.8	4
0.9	9
1	inf.

This shows that higher values of τ than 0.5 are impractical, in fact, most FB's in CM operate around 0.2 .. 0.3. With a constant τ the output voltage is rock-solid, of course, due to losses in all components, the regulation loop reacts to load variations by very slight changes of τ. Like the buck, variations of the input voltage can only be compensated for by changes of $\tau/(1 - \tau)$, but the loop can only affect τ. With the buck τ = Vout /Vin ; if the above formula is rearranged for τ and assumed that Vout /Vin << 1 it is reduced to: τ = N x Vout /Vin i.e. the same formula.

By far most power supplies are constant-voltage types, so this is one reason to choose CM. Note that the duty-cycle limits of the ic's no longer protect the SMPS, so CM operation requires protection e.g. by output current limiting just like any other constant-voltage supply.

In DCM load variations are compensated for by duty cycle variations as this the only means to control the energy charged resp. the power. In CM the duty cycle is constant, independent of the output power. At the transition point the duty cycle must sharply jump! The design of the regulation loop is more difficult, because in CM the loop gain increases enormously; in other words: in CM, a very slight change of the duty cycle has greater influence than in DCM.

Another fundamental difference to DCM is the load on the active components. As is the case with bucks and boosts in CM, the FB in CM switches onto a transformed secondary current at the beginning of a cycle; the secondary current still flows because, in an inductance, the current can not jump. When the primary switch turns on, it places the input voltage across the primary, forcing the diode to turn off, the transformed current will now continue to flow in the primary winding and the switch. This causes increased losses in the switch as well as in the diode, the latter must now be a Si Schottky (only at low voltages) or an ultrafast or hyperfast (only at high voltages) type, because the transistor switches onto a conducting diode which constitutes a short-circuit during reverse recovery! For output voltages > 24 V standard 600 V SiC Schottkies are superior to the best ultrafast Si diodes, inspite of their higher forward voltage; it is the total losses which count, and SiC excels also by the fact that, in contrast to Si, losses do not increase with temperature.

5.2 Output Power.

Neglecting again the losses, the power is equal to the energy which is added during the charging phase x f. In order to arrive at the formula for power in CM, the energy present at the end of the charging phase must be subtracted from the energy at the beginning which shall be called the pendulum energy because it swings back and forth between secondary and primary, causing only losses.

$$P_{out,\,CM} = 0.5 \times L_{p,}\,CM \times (i_{p,max}^2 - i_{p,min}^2) \times f.$$

For ip,max = ip,min = square wave, P is zero, for ip,min = 0 the formula becomes the one for DCM.

5.3 Which "deepness" of CM makes sense?

The formula points out that too much CM is poison, there must exist a practical limit. Are there more advantages than constant-voltage behaviour?

A look at the currents in both modes is helpful. From

$$I_{out,CM} = 0.5\ (i_{s,max} - i_{s,min}) \times \tau^2 \text{ and } V_{out,CM} = VB\ /N \times \tau/(1 - \tau)$$

another form of the equation for P can be deducted:

$$P_{out,CM} = 0.5 \times \tau \times VB \times (i_{p,max} + i_{p,min})$$

Assuming that VB and τ are the same in DCM and CM equating of both expressions yields:

$$P_{out,\,DCM} = 0.5 \times \tau \times VB \times i_{p,DCM} = P_{out,\,CM} = 0.5 \times \tau \times VB \times (i_{p,max,CM} + i_{p,min,CM})$$

With $k = i_{p,min} / i_{p,max}$ (k = 0 ... <1) $i_{p,DCM} = i_{p,max,CM}$ (1 + k).

This shows that for k > 0 (CM) the peak current is lower than in DCM which is a further advantage of CM, reducing the load on the semiconductors. In the unreal case of k → 1 it would be reduced to half, however, this would mean ip,min = ip,max resp. the primary current would change from a trapezoid to a square wave, Lp → infinity, nothing would be charged, Pout = 0. Only pendulum energy would shuttle back and forth. The appearance of pendulum energy resp. pendulum currents for all values of k > 0 creates additional losses in all components which operate at the clock frequency.

The relationship of the inductances in both modes is:

$$L_{p,CM} = L_{p,DCM} \times (1 + k)/(1 - k).$$

Reasonable values of k are 0 ... 0.5, at most 0.7. For k = 0.5 a factor of 3 results, for 0.7 the factor is 5.66. The associated peak currents ip,max,CM are 0.67 resp. 0.59 x ip,DCM . The value 0.59 is already close to the theoretical minimum 0.5.

►In practice one does not exceed the factor 2 resp. k = 0.34; this corresponds to an increase of the number of turns by the factor 1.4 and a peak current factor 0.75.

CM operation is at a cost! The inductances must be increased, the more, the lower the load limit shall be before the transition to DCM. The higher inductance makes the converter slower. The higher numbers of turns require more winding space and, in general, necessitate a bigger and more expensive transformer even though the sizes of wires may be reduced due to the lower rms currents. But the higher number of turns as well as the greater lengths of wires in a bigger transformer cause a higher resistance of the winding, so the losses in the winding will mostly be increased. Switching onto an already flowing current and perhaps also slowly recovering diodes as well as the pendulum currents create additional losses. In each design case it must be checked whether it makes sense to operate in CM. On the whole, DCM is the best choice for small powers, SMPS for higher powers will profit from CM.

5.4 Magnetic circuit considerations.

The core is driven by Θ = ipeak x n. H = ipeak x n/lcore ; B = H/Acore is the decisive parameter, but l and A are constants. Equating the equations for DCM and CM yields:

$$nCM \times i_{p,max,CM} = nDCM \times i_{p,\,DCM} \times 1/\sqrt{(1 - k2)}.$$

For k = 0.5 a factor of 1.15 appears on the right, so in CM the core is driven 15 % harder. k = 0.5 corresponds to LCM = 3 LDCM resp. 1.73 times the number of turns. The peak current was 0.67 ip,DCM , the product of 0.67 x 1.73 = 1.15. A practical maximum B = 0.3 T at 100 C in a DCM design would rise to 0.35 T which is no more acceptable, so a bigger core will also be required for this reason - in addition to 1.73 times the number of turns. With DCM this bigger core would allow a higher drive and higher power.

Will continue in November 2016

dr.seibt@aon.at

Oscilloscopes Optimize Vertical Resolution for Exceptional Signal Fidelity

Teledyne LeCroy introduced the HDO9000 High Definition Oscilloscopes, which unveiled HD1024 high definition technology that automatically optimizes vertical resolution under each measurement condition to deliver 10 bits of vertical resolution. The combination of the next-generation MAUI with OneTouch user interface with a big, bright, 15.4” touch screen takes oscilloscope efficiency, intuitiveness, and ease of use to a completely new level. The addition of the HD1024 technology to an exceptionally deep analysis toolbox enables the HDO9000 to easily uncover difficult to find signal abnormalities.

The HDO9000 High Definition Oscilloscopes offer 10-bit resolution, bandwidths of 1 GHz to 4 GHz and sample rates of 40 GS/s, enabling efficient and accurate debug in high definition. Two units can be synchronized to operate as 8-Channel Oscilloscope without limitations, another unique feature in these high performance oscilloscopes.

A critical element of the HDO9000 is the HD1024 technology, which provides 10 bits of vertical resolution with 4 GHz bandwidth. Waveforms are displayed clearly and crisply, unmasking fine signal details that cannot be seen on conventional 8-bit oscilloscopes. HD1024 technology enables the HDO9000 High Definition Oscilloscopes to automatically and dynamically determine the best ADC configuration under each specific measurement condition.

www.teledynelecroy.com

High Temperature Power Module for Harsh Environments

Raytheon UK’s Integrated Power Solutions (IPS) business unit in Glenrothes, Scotland, has developed a high temperature, small formfactor bridge leg power module. Aimed at high speed switching applications, the module has potential uses in the aerospace sector as it requires minimal external cooling and presents considerable weight saving opportunities within the More Electric Aircraft power system. Also, by supporting applications in harsh environments and in meeting high operating temperature demands, the module can also be used in the geothermal power and oil and gas sectors.

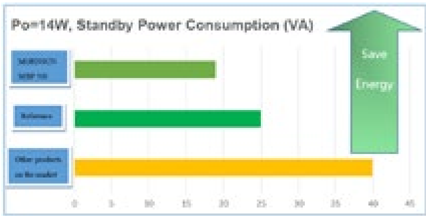
A prototype module has currently amassed more than 1,000 hours of stable operation at 300oC; a temperature at which traditional Silicon-based semiconductors cannot operate. Tests on the module, which includes two 1,200V Silicon Carbide (SiC) Bipolar Junction Transistors (BJTs), have been performed switching 500V at room temperature and switching 200V at 300°C. The BJTs are controlled by integrated base driver circuitry, fabricated using Raytheon’s propriety High Temperature Silicon Carbide (HiTSiC®) process.

“The co-location of BJT base driver circuitry and power transistors into a single high temperature module is a major industry breakthrough,” comments David Gordon, Technical Lead with Raytheon’s IPS. “For example, in many instances it is necessary to switch power stage transistors at tens of kHz and that requires getting the base driver circuitry as close as possible to the power transistors. However, in a high temperature environment that presents a problem.

www.raytheon.com

165~264VAC Input Voltage AC/DC Battery Charging MBP Series

MORNSUN recently released charging power supplies MBP300 and MBP500 series specialized for distribution automation system, which offer ultra-low standby power consumption, active PFC and high efficiency up to 86%. These series have standby power consumption as low as 25VA for 14W output power and 40VA for 20W, and meet requirements of DL/T721-2013 standard.



For FTU, DTU and RTU systems, most of time the whole monitoring system operates in no-load state. Therefore, low standby power consumption and high reliability of charging power supply are significant and critical to wholly reduce energy consumption and improve the reliability of the system.

MORNSUN MBP series provide following features:

Longer time of instantaneous power: When system fault occurs, MBP series still offer 540W output for 30S. It will help isolate the fault area and charge the safety area. Thus, the wires are protected from damage.

Longer hold-up time: In practical applications, when systems cannot be connected with the backup battery, input fault soft turn-off may cause data of the core units' missing or losing during its processing and storage. MBP series is designed with longer hold-up time (1.42S+) to avoid it.

Excellent EMS performance: Concentrated on EMS performance, design reliability and system reliability, MBP series have been tested to meet the requirements of 5KV impact compression test, surge test (level four) and pulse test, etc.

Multiple protections and various applications: MBP series provide output over-voltage, over-current and short-circuit protections, etc. With battery activation (manually or automatically through external signal), over discharge protection, external communications and control functions, they are suitable for distribution automation system (DTU/FTU), electric power, power substation, RMU, intelligent power box-type substation and intelligent switch controller applications, etc.

www.mornsun-power.com

High Isolation DC-DC Converter for High Power

Power electronics specialist, Amantys Power Electronics Limited, has announced the launch of a high isolation DC-DC converter to power high voltage gate drives for 4500V and 6500V IGBT modules. The DC-DC converter is designed to comply with international standards including EN 50155 for railway applications and IEC 61800-5-1 for variable speed motor drives.

The DC-DC converter can accept a 15V or 24V input voltage and has an output power of up to 15W making it suitable for driving several gate drives in parallel. It features low quiescent current of 50mA and a maximum current output of 1100mA.

The DC-DC converter provides a minimum of 12kVrms working voltage isolation, making it suitable for two level and three level converter topologies.

The robust design can operate from -40ºC to +85ºC and has protection features for: input reverse polarity; input overvoltage; output short circuit; and output overload. The protection features are complemented by status LEDs that indicate DC-DC operation and fault conditions.

The isolated DC-DC converter will be used in applications such as railway traction, HVDC infrastructure and medium voltage motor drives. The DC-DC converter is available for sample orders in Q4 2016.

www.amantys.com

Small-Package 60 and 100V Automotive N-Channel MOSFETs

Toshiba Electronics Europe has announced two compact, N-Channel MOSFETs for load switching in automotive applications. The 60V SSM3K341R and the 100V SSM3K361R deliver class leading low ON-resistance and are qualified according to the AEC-Q101 standard.

As a result, they are ideal for use in power management, including DC-DC converter or load switch applications which are of increasing importance for the ever developing electronic systems in vehicles.
Increased industry demand for power-saving LEDs has seen growth in requirements for the N-channel MOSFETs that are used as switches for LED drivers. Toshiba's SSM3K341R and SSM3K361R small-package MOSFETS address this demand class-leading typical RDS(ON) ratings (@VGS = 4.5V) of 36mΩ and 65mΩ respectively. In addition, both devices support a maximum channel temperature of 175°C, which enables wide spread use in a range of automotive applications.
Both new devices reduce heat dissipation resulting from turn-on losses by approximately 65% against Toshiba previous products[2]. In addition, the MOSFETs are housed in compact SOT-23F flat lead type packages. These reduce PCB footprint by approximately 64% compared to a conventional SOT-89 package, while maintaining a maximum level of 2.4W power dissipation.

http://toshiba.semicon-storage.com/de/top.html

Knowles Capacitors will be at European Microwave Week 2016 in force with a selection of new products across its brands – DLI, Novacap and Syfer

Stand number 178 will feature a new range of high temperature Multi-layer Ceramic Capacitors in 0402, 0603 and 0805 sizes with ultra-Low ESR and High Q characteristics available from the Syfer and Novacap brands. In addition DLI branded 175°C MLC devices in UL and AH

High Q Porcelain dielectrics will be shown. DLI will also be showcasing their EAR99 Filters, Power Dividers and Couplers together with the introduction of 5G specific surface mount device options. DLI Ins its brands - Novacap,single layer ceramic Milli-Caps and Opti-Caps are another highlight.
Milli-Cap behave like an Ideal Capacitor, constructed as a single piece unit with insensitive orientation. They match typical 50Ω line widths with very low series inductance and ultra-high series resonance. They are low loss, High Q parts available in 0201, 0402 and 0602 footprints. Capacitance range covers 30pF - 220nF over the ceramic dielectrics X5R: -55°C to +85°C (TCC ± 15%) and X7R: -55°C to +125°C (TCC ± 15%).
Opti-Cap Ultra Broadband DC blocking capacitors exhibit low loss frequency stability over temperature with very low series inductance. They provide resonance free DC blocking to >40GHz. Available in 0201, 0402 and 0602 footprints for SMT by solder or epoxy bonding. Typical applications include Test Equipment, Photonics, SONET, TOSA/ROSA, High Speed Data; Broadband Microwave/Millimeter Wave and Transimpedance Amplifiers.

www.knowlescapacitors.com

Unified for Pioneering Wireless Solutions

AMBER wireless GmbH, manufacturer of wireless connectivity solutions, is part of Würth Elektronik eiSos GmbH & Co. KG now. With the merger the owner-run wireless specialist founded in 1998 lays the foundation for its further growth and the globalization of its activities. Through the acquisition, Würth Elektronik eiSos considerably expands its range in growth fields, such as Internet of Things, Industry 4.0 and Smart Metering.
AMBER wireless is one of the leading manufacturers of low power ISM/SRD solutions in Europe and offers highly efficient wireless products in the 169 MHz, 433 MHz, 868 MHz, 915 MHz and 2.4 GHz frequency bands. These wireless solutions are considered to be important components for growth-intensive application areas, such as sensor networks, Internet of Things, telemetry, logistics, asset tracking, Smart Metering, medical technology, security systems, as well as Smart Home, industry and building automation.

www.we-online.com

Efficient and Robust Current Sensing Optocoupler for Servo Drives and Motors

Broadcom Limited announced a new optically isolated sigma-delta modulator device, the Avago ACPL-C799, designed for a wide range of industrial applications including servo drives and motors. The device features a low differential input range of +/-50 mV allowing designers to reduce power dissipation across a current-sensing shunt resistor in servo drive and motor applications.

Compared to previous generation products with an input range of +/-200 mV, the ACPL-C799 enables the use of a smaller shunt resistor with one-quarter the value, thereby eliminating 75 percent (75%) of shunt resistor's power losses. Despite the reduced input voltage range, the device delivers superb SNR, ENOB and offset drift performance, enabling a high precision motor control in space constrained, high temperature environments.

http://www.avagotech.com/products/optocouplers/industrial-plastic/isolation-amplifiers-modulators/sigma-delta-modulators/acpl-c799

www.broadcom.com

Robust AC Capacitors

TDK Corporation presents two series of robust EPCOS alternating current (AC) capacitors. Both the cans and the tops of the components are constructed of aluminum, making them particularly robust.

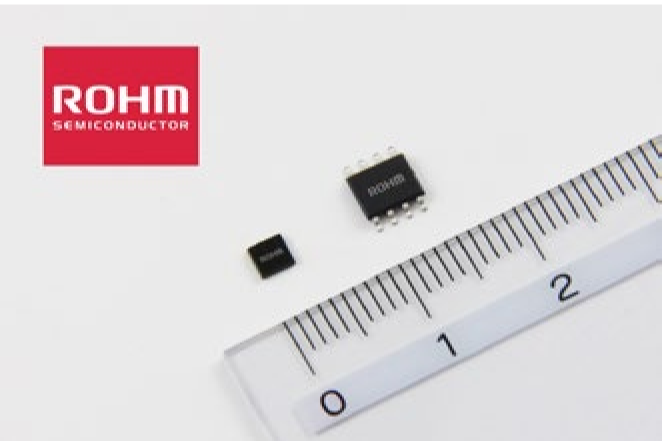
All capacitors in the B33331* and B33335* series are CE-compliant, have both VDE approval in accordance with EN/IEC 60252-1 Class A and UL approval. Thanks to their robust design and integrated overpressure disconnector, the capacitors meet the requirements of safety class S2 as specified by EN/IEC 60252-1. The rated voltage for all the new types is 450 V AC (50/60 Hz) and the permissible operating temperature is between -40°C and +85°C. This results in life expectancy of 30,000 hours.

The B33331* series covers a capacity range from 2 µF to 50 µF. Depending on the capacity, the dimensions range from 30 mm x 55 mm to 50 mm x 100 mm (d x h). The B33335* series features two capacitors installed in one enclosure. With these double capacitors, different combinations of capacitance values are offered. The current standard range covers values between 12 µF + 1.5 µF and 60 µF + 8 µF. The dimensions range from 50 mm x 57 mm to 63.5 mm x 102 mm (d x h). The double capacitors are particularly suitable for use in air-conditioning systems, in which the larger capacitance is used for the compressor motor and the smaller capacitance for the fan motor. Further capacitance and voltage values are available on request.

www.epcos.com/ac_capacitors

Isolated Power Supply Control ICs for Industrial Inverters

ROHM has recently announced the availability of isolated flyback-type DC/DC converter control ICs for inverters in high power industrial equipment such as FA/solar inverters and power storage systems. Isolation is almost always built into electronic equipment to provide protection against shock and damage. High voltage, high power industrial applications (including power storage systems and solar inverters which are seeing increased adoption) are no exception, requiring even more stringent reliability demands due to the harsher temperature conditions and the need for continuous, long-term operation.



Conventional solutions utilize a optocoupler and other components for flyback-type isolated power supply control. However, this presents challenges associated with current consumption and reliability (including circuit scale and product life). In contrast, the BD7F series reduces the number of external parts by nearly half by eliminating the need for a optocoupler, providing greater energy savings, reliability, and miniaturization. In addition, a newly developed adaptive ON time control method is used to improve load response time (a drawback of conventional isolated power supply ICs), while voltage fluctuation is limited to under 200mV, contributing to even higher reliability.

Samples, along with an evaluation board, BD7F100HFN-EVK-001, integrating the BD7F100HFN-LB, will be available for purchase. This board enable to be designed for 24V input voltage and 5V output voltage with output current range of 13.8mA to 800mA for maximum output power of 4W.

www.rohm.com

High Efficiency, High Current Buck-Boost Regulators for Portables

Intersil Corporation announced the ISL91127 and ISL91128 high efficiency buck-boost regulators. The latest in Intersil's industry-leading buck-boost family, the new devices feature 4.5A switches, best-in-class efficiency up to 96%, and a compact footprint ideal for providing system power or powering the peripherals in battery-operated devices. They offer the industry's lowest quiescent current of 30µA for superior light load efficiency.

In hand-held device applications where the input voltage may be higher or lower than the output voltage, buck-boost regulators improve efficiency and provide longer battery life compared to a boost regulator plus bypass solution. With increasing demands for smaller and smaller footprints, Intersil's ISL9112x compact QFN and WLCSP packages enable power designs that maximize efficiency while providing flexibility and ease of design.

The ISL91127 and ISL91128 operate in buck, boost or buck-boost mode, depending on the relation between input and output voltages, and provide smooth transitions between modes to prevent noise and glitches. This capability, combined with patented Intersil technology for delivering superior light load efficiency with ultra-low quiescent current, maximizes efficiency under all conditions. This is essential to improve battery life by reducing power drain and heat dissipation in portable and mobile applications.

www.intersil.com

Isolated Power Supply Control ICs for Industrial Applications

ROHM has recently announced the availability of isolated flyback-type DC/DC converter control ICs for Auxiliary Supply in High Power Industrial equipment such as FA/solar inverters and power storage systems. Following the BD7F100, the company now expands the series with the new BD7F200, covering all applications with a power requirement from 1W to 10W, and including two different packaging options. In addition, ROHM introduces a BD7F100 evaluation kit that facilitates the design of flyback isolated applications.

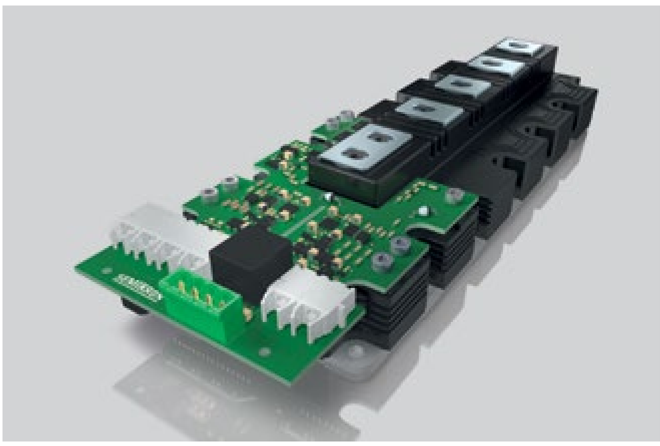
Isolation is almost always built into electronic equipment to provide protection against shock and damage. With this respect, a broad range of applications such as power metering, industrial programmable logic controllers (PLCVs), Industrial Fieldbus, Industrial automation, insulated-gate bipolar transistor (IGBT) and SiC MOSFET gate drivers require isolated DC/DC converters. These devices provide galvanic isolation, improve safety and enhance noise immunity to the systems. With them it is possible to generate multiple output voltage rails and dual polarity (positive and negative) rails. High voltage, high power industrial applications (including power storage systems and solar inverters which are seeing increased adoption) are no exception, requiring even more stringent reliability demands due to the harsher temperature conditions and the need for continuous, long-term operation.

www.rohm.com/eu

IGBT Driver with Optical Interface Saves External Sensors

SKYPER Prime O is an IGBT driver with optical interface and galvanically insulated temperature and DC link signals. SKYPER Prime O is designed and qualified with the IGBT module SEMITRANS10 and offers an Plug & Play bundle with improved SOA characteristic. This saves costly qualification and redesign loops and at the same time offers significant performance advantages in various applications. SKYPER Prime O offers galvanically isolated, highly accurate DC-link and temperature signals as a PWM signal to the controller. So very compact high power inverters can be built without costly sense circuits, cabling effort or power supplies. In addition, second source requirements are covered, both on driver and on module side. Electrical and optical interfaces are available.

http://www.semikron.com



800 V CoolMOS™ P7 Series Sets a Benchmark in Efficiency

Infineon Technologies AG introduces the 800 V CoolMOS™ P7 series. Based on the superjunction technology, this 800 V MOSFET combines best-in-class performance with exceptional ease-of-use. This new product family is a perfect fit for low power SMPS applications, fully addressing market needs in performance, ease of design, and price/performance ratio. It mainly focuses on flyback topologies which are typically found in applications like adapter, LED lighting, audio, industrial and auxiliary power.

The 800 V CoolMOS P7 series offers up to 0.6 percent efficiency gain. This translates into 2 to 8 °C lower MOSFET temperature compared to the CoolMOS C3 or to competitor parts tested in typical flyback applications. This new benchmark results from a combination of optimized device parameters: including the reduction of more than 50 percent in E oss and Q g, as well as a reduced C iss and C oss. The further improved performance enables higher power density designs through lower switching losses and better DPAK R DS(on) products. Overall, this helps customers to save BOM costs and reduce assembly efforts.

www.Infineon.com/p7

Isolated DC-DC Converters Offer High Efficiency in a Rugged, Encapsulated Package

CUI Inc. added to its line-up of isolated dcdc converters with the introduction of three encapsulated models ranging from 10 W to 30 W. The PDQ10D, PDQ15D and PDQ30D output 10 W, 15 W and 30 W of power respectively in an industry standard 1” x 1” package, making them ideal replacements for larger 1” x 2” modules. The PDQD models also feature a 4:1 input voltage range of 9~36 or 18~75 Vdc with high typical efficiencies up to 90%.

Housed in a compact, low-profile DIP package with a five-sided, shielded metal case, the rugged, encapsulated modules measure just 25.4 x 25.4 x 10.16 mm (1 x 1 x 0.40 in) and come available with single regulated output voltages of 3.3, 5, 12 and 15 Vdc or dual regulated output voltages of ±5, ±12 and ±15 Vdc.

Thanks to its wide -40 to +105°C operating temperature range and encapsulated design, the PDQD series is suitable for convection-cooled equipment and harsh environments with target applications that include telecom, industrial, remote sensor systems and portable electronics.

All PDQ-D models offer an input to output isolation of 1,500 Vdc and feature remote on/off control while single output models also offer output voltage trimming that allows for a ±10% adjustment of the nominal output. Protections for over voltage, input under voltage lockout and continuous short circuit are also included.

Visit CUI’s booth (Power Hall A2, Booth 613) at electronica 2016

www.cui.com

Linear Current Regulator and Controller Family for Automotive LED

Allegro MicroSystems Europe has announced two adjustable linear current regulators for driving automotive LED arrays. Allegro's AEC-Q100 qualified A6274 and A6284 devices include many features designed to further enhance the existing family of linear LED drivers. These low-EMI ICs integrate an optional pre-regulator gate drive to dynamically and linearly control an external P-channel MOSFET, which extends the output power capability of a linear LED driver by moving most of the power losses to an external power FET. This eliminates the need for a DC/DC switching solution and its associated inductor, diode and higher EMI. The ICs sink up to 60 mA (A6274) or 120 mA (A6284) from each of six LED pins, and may also be paralleled to drive higher current LED strings for a total of up to 720 mA. In addition the multiple ICs can be configured in parallel for large lighting systems.

LED light output can be adjusted by PWM (pulse width modulation) dimming, and the ICs include an internal PWM dimming circuit that is programmed by external resistors for accurately controlled PWM duty cycle and frequency. This eliminates the requirement for a PWM signal usually provided by a local microcontroller or 555 timer. Alternatively, an external PWM signal can be applied in particular if already available by an existing control module. The devices can be configured to change between full and dim modes for bi-level operation, as often needed for stop/tail rear combination lamps (RCL) or daytime running lights (DRL).

www.allegromicro.com

Mini Power Supplies for Smart Home / Office Applications with EN60335

Intelligently networked smart homes and smart offices often require control systems with many low power nodes, actuators and sensors that are "always on". RECOM's robust and reliable 2W and 3W AC/DC power supplies are specially designed to power smart building infrastructure 24/7 by offering compact sizes for easy installation, very low standby power consumption (35mW) and full household (IEC/EN60335), CE (LVD+EMC+RoHS2) and industrial safety certifications (IEC/EN/UL60950).

The RAC03-SER/277 is a regulated power supply designed to fit inside standard wall boxes. Its round shape and low profile of only 11mm means that it is possible to implement simple and quick smart home/smart office solutions without extensive renovation works by using the space behind switch plates. For PCB-mounting or even more space-constrained wired applications, the RAC02-SE/277 and RAC03-SE/277 modules with a footprint of only 34x22mm are also available. All three power modules deliver isolated, regulated, short-circuit and overload-protected local DC power to run smart building automation applications such as remote sensors and actuators, climate controls, touchscreen interfaces, door access controls, wireless alarm systems, garage door-openers, conferencing and entertainment systems.

www.recom-power.com

PSMA Offers Power Supply Safety & Compliance Database

The Power Sources Manufacturers Association (PSMA) announces the availability of a new on-line Safety & Compliance Database. The continually updated resource lists the many state, national and worldwide organizations currently active in establishing and maintaining safety, electromagnetic compatibility, material toxicity and environmental standards for power supplies used in commercial applications. Recognizing the database as an invaluable tool for power electronics industry professionals, the PSMA is offering it free of charge to both PSMA members and non-members alike.

Intended users of the Safety & Compliance Database are those who design power systems for products that will be offered in the global marketplace, and who therefore need to comply with current and evolving safety and standards for their target markets. The database can be searched by specific applications; giving the most recent status of standards, identifying key documents, meetings and milestones associated with each standard, and providing links to the appropriate websites of controlling organizations.

www.psma.com

Hermetically Sealed Aluminum Electrolytic Capacitor Line Expands to Higher Voltages

Cornell Dubilier Electronics, Inc. has announced significantly higher voltage values for its MLSH Slimpack series of hermetically sealed aluminum electrolytic capacitors. These unique components feature a true glass-to-metal seal that prevents dry-out of the capacitor electrolyte. As a result, they provide extraordinarily long life (5,000 hours) to more than meet the most demanding applications for military and aerospace. With the line expansion just announced, they are now available in nine values, from 120 µF to 3,200 µF, with ratings up to 250 Vdc.

The hermetic Slimpack is an offshoot of the well-established non-hermetic Flatpack series. The new MLSH Slimpack can be used as a replacement for parallel and series banks of wet tantalum capacitors for both new and existing designs. This change is especially beneficial where bulk storage is paramount. MLSH Slimpack devices measure 1.0” x 1.5” x 0.5”, weigh less and have more capacitance than a parallel bank of three or more wet tantalum capacitors. High rated capacitance is maintained at temperatures as low as -55 °C, a key requirement for power supplies used in military and aerospace applications. Top-end temperature rating is 125 °C.

All devices in the MLSH Slimpack series feature rugged stainless steel cases with a vibration rating of 80g. In most applications, there are significant cost and space savings vs. a comparably rated bank of wet tantalum capacitors. The new high-voltage MLSH Slimpacks eliminate the need for series banking.

Since its founding in 1909, CDE has been dedicated to advancing capacitor technology for new applications. The company combines innovative products with engineering expertise to provide reliable component solutions for inverters, wind and solar power, electric vehicles, power supplies, motor drives, HVAC, motors, welding, aerospace, telecom, medical equipment and UPS systems.

www.cde.com/MLSHSlimpack

All-SiC High Performance Half-Bridge Power Module & Gate Driver Combination

Wolfspeed, A Cree Company, and a leading global supplier of silicon carbide (SiC) power products, has introduced the first fully-qualified commercial power module from its Fayetteville, Arkansas location. The high-performance 62mm module represents a new generation of all-SiC power modules that enable unprecedented efficiency and power density for high current power electronics, such as: converters/inverters, motor drives, industrial electronics, and high performance electric vehicle systems. The new module allows systems designers to realize lighter weight systems that are up to 67% smaller by achieving efficiencies of over 98% and improvements in power density of up to 10 times compared to systems built with silicon-based technologies.

"Wolfspeed's launch of this fully-qualified, next-generation SiC power module represents the culmination of an engineering development program that began with the company's early SiC MOSFET module designs," explained John Palmour, Wolfspeed's chief technology officer. "Our 2015 acquisition of Arkansas Power Electronics International (APEI) enabled our team to deliver this advanced power module design on an aggressive timeline by combining leading-edge SiC device technology with the industry's most advanced wide bandgap packaging innovation. The resulting module is the first of many products that promise an accelerated disruption of the power electronics market."

Compared to conventional silicon IGBT power modules, or even previous generations of SiC MOSFET modules, the new module will offer significantly higher power density for applications in which volume and weight are critical limitations. Operating at a higher switching frequency without any compromise in system efficiency means a reduction in the number and size of magnetic and passive components required for the balance of the system. The superior thermal characteristics of SiC devices, along with the packaging design and materials, enable the module to operate at 175°C, which is a key advantage for many industrial, aerospace, and automotive applications.

http://www.wolfspeed.com/power/products

www.cree.com

Efficient 250W Baseplate Cooled Quarter-Brick DC-DC Converter

Murata is announcing a development effort for Industrial Class DC-DC converter modules. The first in this series is a 250 Watt isolated DC-DC quarter-brick converter, the ICQ series from Murata Power Solutions. These highly efficient converters, typically 92%, are believed to be the only power modules on the market to offer 250 Watts of output power from a quarter-brick format with a 4:1 (9 to 36 VDC) Vin range. The ICQ series is designed for rough service/industrial applications and incorporates features such as a specially designed enclosed package that uses baseplate cooling to improve thermal performance. The electrical and mechanical design allows the ICQ series to deliver the full 255 Watt output power with a baseplate temperature in the range of -40 to +90°C without derating. In addition, the ICQ is designed to meet the environmental stress limits for shock & vibration specified in MIL-STD-810G throughout its service life. The ICQ0120V1PC achieves its high performance by incorporating proprietary circuit architectures, synchronous rectification, advanced packaging, and thermal design approaches. Accommodating a wide 4:1 input voltage range for operation from 9 to 36 VDC, the initial ICQ delivers a regulated 12 VDC output rated at 20 A and provides a basic insulation system for galvanic I/O isolation rated at 2,500 VDC. The ICQ has been designed to meet EN60950, UL/CSA safety requirements for imbedded power modules.

The industry-standard quarter-brick package measures 39.1 mm x 60.7 mm x 12.7 mm (1.54 x 2.39 x 0.50 inches) and is ideal as a drop-in replacement for lower power less efficient converters.

www.murata.com

DC/DC Converters for Railways

RECOM introduces the RP40-FR DC/DC converter series certified for use in rolling stock as well as industrial applications requiring compliance with recognised safety standards.

EN50155 certification ensures that the RP40-FR series is ideal for harsh environments which can be encountered in railway applications. The new series is also certified according to UL/cUL 60950-1 and is therefore also ideal for applications with higher input voltages in industry, telecommunications, and distributed power supply architectures. Along with a wide 4:1 input voltage range, RECOM DC/DC converters deliver 40W of regulated output power from single or dual outputs and come with an output trim function. The DC/DC converters include on/off remote control with positive or negative logic. The series is available with rated DC inputs of 24V or 48V with 1.6kVDC insulation or 110V with 3kVDC insulation. Due to very high efficiency up to 92%, a high power density is possible – 40W is available in 2x1 inches (50.8x25.4x10.2 mm without a heat sink or 56.8x25.4x16.8 mm with a heat sink). The converters in this series are equipped with short-circuit, overvoltage, overload, and overtemperature protection and have an extended operating temperature range of -40° to 105°C. RECOM provides a three-year guarantee on the RP40-FR series.

www.recom-power.com

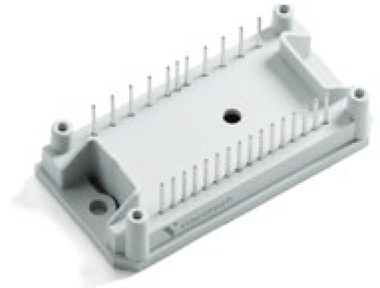
Advertising Index

IPM with CIP Topology

Vincotech, a supplier of module-based solutions for power electronics, announced the launch of a deeply integrated Intelligent Power Module for 600 V applications. The flowIPM 1B CIP 600 V features the powerful combination of a high-speed F5 IGBT with the switching performance of a MOSFET and a silicon carbide boost diode in the PFC circuit optimized for frequencies up to 150 kHz. The high-speed F5 IGBT, paired with the silicon carbide boost diode in the PFC circuit, not only delivers impressive performance, it also drives down the cost of external passive components. The current rating of this new CIP (converter + inverter + power factor correction) topology housed in an integrated power module is also impressive: 10 A @ 80° C heat sink temperature.

The deeply integrated flowIPM 1B CIP 600 V module enables manufacturers to slash their overall system's size, cost, and time to market. It also features an inverter gate drive with a bootstrap circuit for high-side power supply, as well as emitter shunts (30 mΩ) for vastly improved motion control.

The flowIPM 1B CIP 600 V modules come in 17 mm flow 1B housings. Versions in the 12 mm housing, with Press-fit pins and with



phase-change material are available on request. Samples may be sourced on demand from our usual channels.

http://www.vincotech.com

PrimePACK™ with IGBT5 and .XT

The dawning of a new era – Infineon's latest IGBT chip and interconnection technologies

Benefits

- Increasing power density by 25 % or
- 10 times longer lifetime

Main Features

- New Chip Technology: IGBT5
  - Reduction of total losses by up to 20%
  - Increased operation temperature T_{vjop} = 175° C
- New Interconnection Technology: .XT
  - Increased power and thermal cycling capabilities

www.infineon.com/PrimePACK