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EMC design of Switching Power Converters - Parts 2 and 3

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EMC design of Switching Power Converters

—Parts 2 and 3 —

The circuit design of high-frequency switching power converters

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Contents

33Issue 93 of the EMC Journal carried the first of these "stand alone" articles [13] on the EMC design of switch-mode and PWM power converters – my attempt to cover the entire field including DC/DC and AC/DC converters, DC/AC and AC/AC inverters, from milliwatts (mW) to tens of Megawatts (MW).

In this series I aim to address *all* applications, including: consumer, household, commercial, computer, telecommunication, radiocommunication, aerospace, automotive, marine, medical, military, industrial, power generation and distribution, in products, systems or installations.

And of course I will not be forgetting the current "hot topics" of hybrid & electric automobiles, electric propulsion/traction; "green power" (e.g. LED lighting; and power converters for solar (PV), wind, deep-ocean thermal, tidal, etc.).

[13] introduced a number of general topics and a number of generic figures showing the various "building blocks" of single-phase converters. Converters for three-phase inputs and outputs (or any other number of phases) can be developed from those building blocks.

This article addresses the circuit design issues associated with the high-frequency power "switchers" (for DC outputs) and "choppers" (for pulse-width-modulated (PWM) AC and DC outputs) themselves. Later "stand alone" articles will address input and output rectifiers; components such as transformers and storage/decoupling capacitors; PCB layout; wiring; and EMI mitigation such as shielding and filtering.

I generally won't repeat material already published in the EMC Journal [14], or in my recently-published books based on those articles [15], so that you don't get bored by repetition. But I will provide the appropriate references.

2 Circuit design of high-frequency (HF) "switchers"

I am using the term "switcher" in this article to refer to HF switch-mode power converters intended to provide DC outputs (e.g. flyback circuits), see [16], and the term "choppers" to refer to HF switch-mode power converters that produce pulse-width-modulated (PWM) outputs, that can be used to provide AC and/or DC outputs.

This section 2 deals only with HF switchers.

2.1 Compromising EM emissions and heat dissipation

Power converter designers want to use ever-lower rise/fall times to reduce thermal losses and so increase conversion efficiency, but as shown in Figure 2A this increases the energy in the higher-frequency harmonics, and so increases radio frequency (RF) emissions, especially at the RF resonances of the circuits and their mechanical structures.

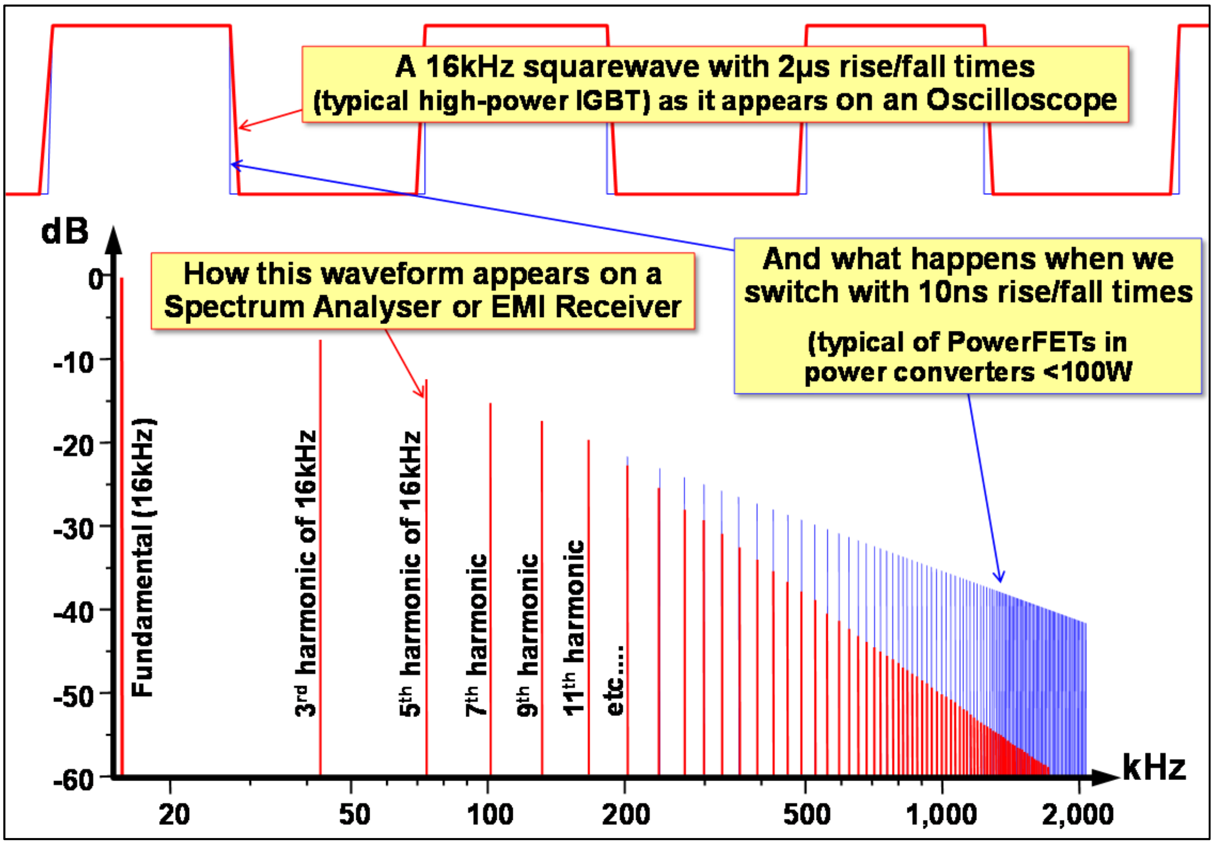

Figure 2A More RF noise is created when switching edges are faster

Also, beyond a certain point, “sharper” switching edges reduce thermal efficiency, because they increase the height of the voltage overshoots that can damage devices and increase thermal dissipation. Reducing the height of the overshoots with snubbers replaces the thermal losses in the devices with thermal losses in the snubbers.

So too-fast switching edges can be bad for thermal efficiency, and they are certainly bad for EMC, often requiring increases in the costs of EMI filtering and/or shielding.

So overall cost-effectiveness and efficiency is a question of finding the optimum rise/fall times, not simply using the lowest. There is some sort of tradeoff required, as Figure 2B tries to indicate.

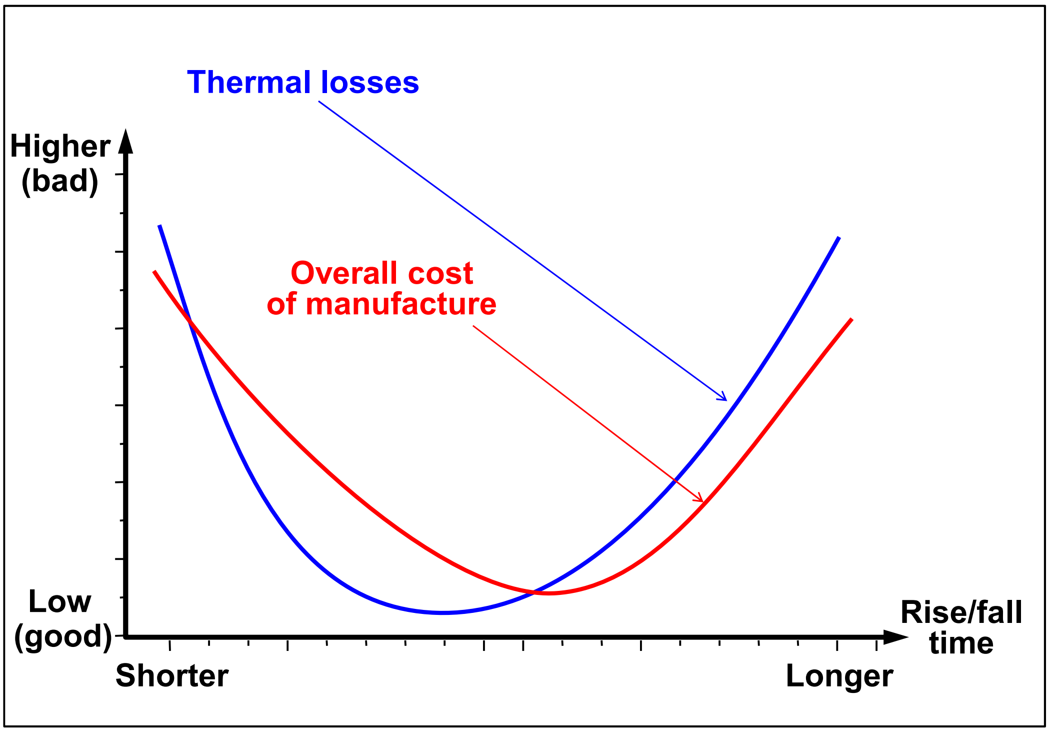

Figure 2B Cost-effectiveness and efficiency: finding the optimum rise/fall times

I have solved several EMI problems that were holding up the marketing of new products, simply by adding low-value (up to around 100Ω) metal-oxide resistors in the gates of switching FETs. The FETs did not run noticeably hotter, and the problem emissions all went away.

In one case excessive radiated noise at 177MHz turned out to be caused by the 177th harmonic of a very tiny DC/DC converter exciting a resonance in the metal structure of a product. Identified in a few minutes by a close-field probe and portable spectrum analyser (see later), and fixed a few minutes later by cutting the gate track and soldering in a series resistor.

This is a good reason for not using the internal power switching device in a DC/DC converter "module". Buying a module might appear to make sense economically, and save space and assembly time, but if the module switches faster than is necessary the only way to meet the EMC requirements might be to add a filter that has almost the same size, weight and cost. Hardly cost-effective.

I have seen this exact problem occur on an automobile seat-heater, switching tens of amps of battery current with a variable pulse-width to control the temperature, with a pulse frequency of around 10Hz. Because of the horrible way that automobile manufacturers use the chassis for the battery current return from items like seat heating elements, every time the PWM power controller switched the car radio made a loud click, so it buzzed away happily when the seat heaters were on.

To save cost, the supplier of the heated seat controller had used a packaged power control module, with no way to slow down the switching speed of the power devices, whose harmonics (spaced 10Hz apart because of the pulse rate) went all the way up into the FM radio band, and even FM modulation wasn't enough to eliminate the noise.

The only options were to use a circuit that provided for control of switching speed (switching at 10Hz, even a 100 microsecond rise and fall time wouldn't require a heatsink), but that would have increased the cost by about as much as the only other option – fitting a filter. Using a twisted-pair wire to supply the heating element would also have worked, but wasn't permitted by the vehicle manufacturer.

Of course, it was the usual situation – salespeople had negotiated the sale of the heated seat controller to the automaker based on the use of the low-cost module, but with no consideration of EMC, and towards the end of the project, with delivery schedules imminent, they found that the only way to comply with the automaker's EMC specification (part of the contract) was to add so much cost that they would make a loss overall.

(Salesmen usually don't care about such mistakes, because they are paid a bonus on the value of the order, not on its profitability. I know a company where they tried to make the salespeople care about profitability by tying their bonuses to the profit made on the contracts they agreed. They all resigned.)

An important note on voltage and current measurement

Where mains-powered ("off-line") converters are being measured, the ordinary probes used with ordinary oscilloscopes are unsuitable because the 'scopes use grounded chassis for safety.

Never use ordinary probes by removing the safety earth connection to the 'scope, or by using a battery-powered 'scope.

The chassis of the 'scope will be at mains voltage and very hazardous indeed. I have had some nasty shocks from 'scopes used in this way, even where the person who removed the safety earth had thoughtfully provided a hazard warning label.

Even if using such a "floating" 'scope, the very high capacitance of the probe's screen will tend to distort the measured waveforms and can cause large project delays as non-existent problems are chased.

So – when working on mains-powered converters – for safety reasons always use high-voltage-isolated differential probes, rated for the peak surge over-voltage and highest frequency you want to measure, and checked as *really being* compliant with all of the relevant parts of IEC/EN 61010 at the peak surge voltage it requires, which it specifies depending on the application.

Of course, there are hand-held battery-powered products sold specifically for safely measuring mains-powered circuits, and some of them provide some oscilloscope functions. These do not need to use separate HV differential probes, but must only be used with the insulated probes provided with them.

Note that CE marking and EU Declarations of Conformity prove nothing at all because they only have legal relevance for Customs Officers and have no engineering value. And "buying in good faith" is a bad joke when your life depends on your purchase being safe enough. It is strongly recommended to see a favourable safety assessment from an independent safety assessment agency, e.g. TUV, BSI, Semko, etc., for the exact model purchased, and also to check with the issuing agency that it is not a forgery. Plus it is important to take all necessary precautions to avoid purchasing counterfeit instruments or probes.

2.2 Circuit topology and design

It is best to choose a circuit topology that is as EMC benign as possible, with "clean" waveforms that change as gently as possible (given thermal dissipation considerations, see 2.1) for both voltage and current – i.e. low dV/dt and low dI/dt – at all times throughout the cycle.

This usually means using topologies (circuit design schemes) with exotic names such as: Resonant Mode [17] [18]; Quasi-Resonant-Mode [19]; Single Ended Primary Inductance Converter (SEPIC) [20]; Zero Current Switching (ZCS) [21]; Zero Voltage Switching (ZVS) [22]; Ćuk [23]; Critical Conduction Mode, etc.

I am not going to attempt to describe the many and various circuit topologies, because this would go way beyond my comfort zone – but I understand that resonant topologies that combine ZVS with "continuous conduction mode" (CCM) generally have lower emissions.

However, in multiphase/interleaved designs it may be that "boundary conduction mode (BCM)" has some advantages. [24] and [25] claim this is so for power factor correction boost converters (see 2.10 below) and it may also be true for other converter applications.

There are very many application notes, magazine articles and conference papers on switcher topology, most of which focus on thermal efficiency. But some of them (e.g. [26], [26a]) also address EMI. So the best I can do is suggest doing some research to find the topology that best suits the needs of your current project whilst also reducing emissions. Application Notes from Fairchild Semiconductors and Unitrode (now part of Texas Instruments) are good places to start looking, as is [28].

Pulse-skipping and 'hiccup mode' types of circuits tend to have higher emissions, and this can mean that a power converter in lightly-loaded or standby mode can emit at much higher levels than they do when operating normally. [26a] discusses some of the relevant issues.

This came to light some years ago when an EMC test lab testing a television, found that it met its emissions specifications when displaying a TV programme, but failed them miserably when in standby mode. The problem is that the EMC test standards for most products used to state that they shall be tested "when operating as normal" – but what is normal for a TV? Usually, we consider normal operation of a TV to be when it is displaying a TV programme, but in fact most TVs spend most of their time in standby, so maybe standby is "normal operation"?

Anyway, I understand that most of the standards have now been modified (or will be) to use phrases like this one, from CISPR 22 2006: "The operational conditions of the EUT shall be determined by the manufacturer according to the typical use of the EUT with respect to the expected highest level of emission."

But the real reason we engineers bother at all with EMC is that products with poor EMC immunity don't work very reliably in real life, and products with high emissions tend to interfere with our customers' equipment and annoy them. So, to help achieve financial success we always design and test all of the modes of converter operation for EMC.

If, for some reason, we decided not to use a benign converter topology, and instead chose to use flyback-type buck or boost converters because (for example) we had an application note that looked as if it told us all we needed to know (they usually don't), we can try to reduce the emissions from these very-noisy circuits by adding small soft-ferrite chokes (e.g. 0.5µH ferrite beads) to limit the dI/dt [29], and/or add small capacitors (e.g. 100pF) in series with damping resistors to limit the dV/dt (called "snubbers" and discussed later).

A more sophisticated approach might be to use the 'gate charge' model to design our PowerFET or IGBT gate drive circuits, so that they switch quickly enough, but don't cause such high levels of emissions. This approach is described in International Rectifier's Application Note 944 [30], which takes account of the "Miller Effect" from C_{DG} (PowerFETs) or C_{GC} (IGBTs).

Some manufacturers offer power switching devices that are less affected by Miller Effect, e.g. [31] and so switch "cleaner" than other devices with the same rise/fall times, causing lower emissions.

Circuit simulators are a good way of discovering "nasties" in the waveforms, especially when the simulation models for the components and interconnections take account of their stray "internal" resistance, capacitance and inductance, even if only first-order approximations.

I well remember purchasing a costly circuit simulator for Marconi Instruments in the mid-1980s, to help us design a 150W DC-AC switch-mode converter, only for us to abandon it when it showed switching waveforms that oscillated rail-to-rail (150V peak-to-peak) seven times at every switching edge. We simply assumed that the simulator was giving a "garbage" output, which was a bit disappointing given what we'd spent on it, and training to use it. It took us six months of wrestling with the design to discover that our switcher did indeed switch seven times where it was only required to switch once.

We figured out the stray circuit elements (often called the "hidden schematic") that caused this bad behaviour, and found that the simulator had been right all along! If only we had known enough to believe its output, we could have saved 6 months of work!

2.3 Pulse width and EMC

If we try to make a pulse width that is very narrow, for example to increase the range of control of output voltage, we might find that we get pulses that are narrower than the switching devices' turn-on and turn-off times combined.

The result is that the transistors will not be turned fully on or fully off when they are commanded to change state again, and this can cause excessive emissions of EMI. It can also cause large cross-conduction currents that increase device heating by more than expected, leading to unreliability in the field and increased warranty costs.

It is an easy problem to identify, though, because the high levels of emissions it causes will correlate with the situations where the control loop is demanding very narrow pulse widths

2.4 Snubbing

Stored energy in devices and components has to be dissipated after a switching operation, and 'stray' Cs and Ls in inductors, conductors, and power switching semiconductors cause resonances. The result is voltage overshoots and ringing following the switching edge, wasted power, and levels of emissions increased by (typically) between 20 and 40dB at those frequencies.

Figure 2C shows an example of a common resonance problem. The chain dotted box shows the idealised first-order model of a winding (primary or secondary) of an isolating transformer – which consists of the ideal transformer impedance (Z, the reflected impedance of the other side multiplied by the square of the turns ratio) in series with the winding's leakage inductance and intrawinding capacitance (the capacitance between the turns of wire in that winding).

Clearly this approximation to the non-ideality of the winding is very crude indeed, but it does make it clear that the leakage inductance and intrawinding capacitance are in parallel – and (as we learned on our Mother's knee) a parallel L and C have a very high impedance at their resonant frequency.

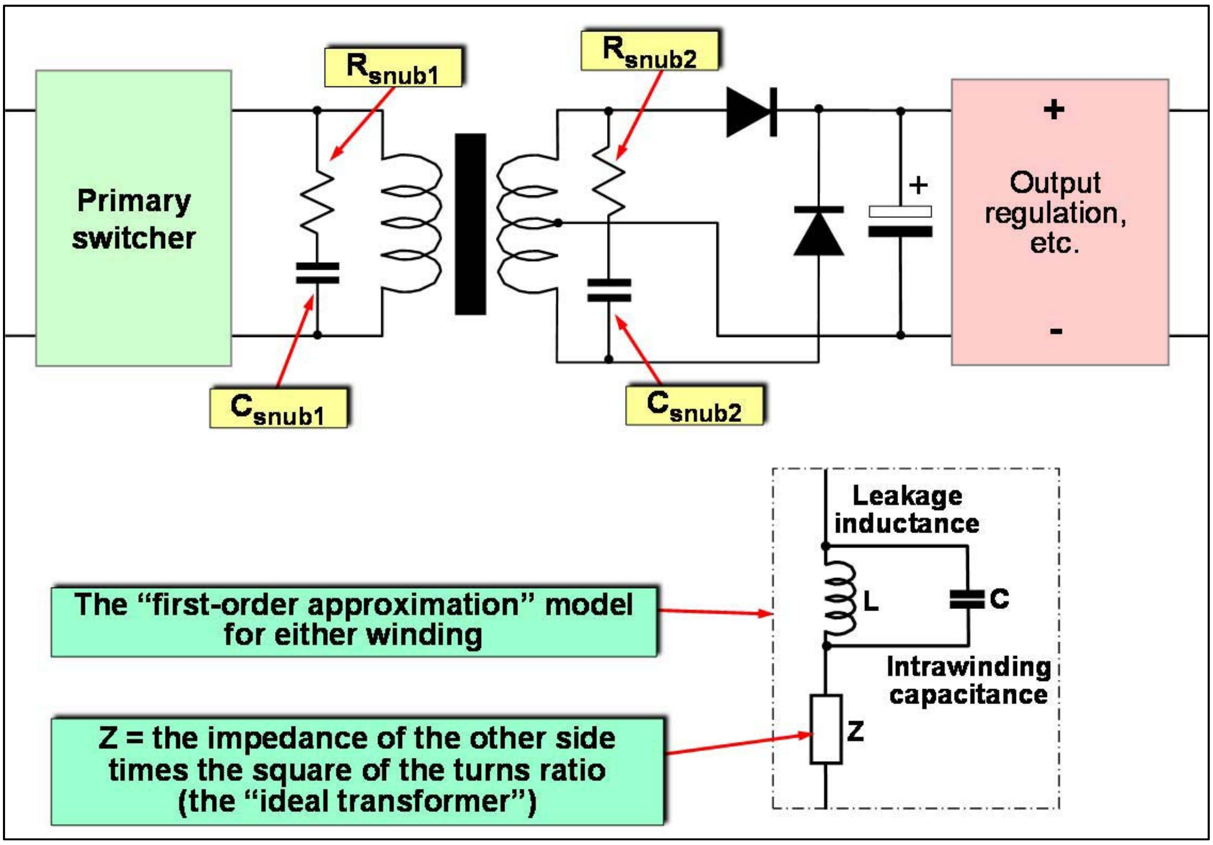

Figure 2C Snubbing primary and/or secondary circuits

As Figure 2A shows, a switched current has a wide frequency spectrum, and the components of that spectrum that lie near to this parallel LC resonance will experience a much higher impedance than the other components – and so will generate a much higher voltage than they do. Bingo! – voltage overshoot and ringing on every switching edge.

When we see unwanted ringing at, say, 16.5MHz, following the edges of our switching waveforms, it tells us we will have excessive emissions at 16.5MHz. It's as simple as that.

In real designs the situation is generally much more complex, with two or more ringing frequencies present, indicating two or more resonances in our design.

Snubbers aim to provide an alternative, lower impedance path for the currents that lie near the problem resonance frequencies, so that they don't generate such high voltages. This is often called "damping" or "damping down", and it is a loss mechanism and so has to handle some heat dissipation.

Figure 2C shows the favourite places for fitting RC snubbers, used to dampen down the parallel resonances in its isolating converter example. Snubbers are used on the primary side to dampen down the primary winding resonances that are "excited" by the fast edges and stored energy in the primary switching devices. They may also be used on the secondary side to dampen down the secondary winding resonances that are "excited" by the fast edges and stored energy in the rectifiers.

Silicon carbide (SiC) Schottky rectifiers are expensive, but – unlike the usual PN junction rectifiers –totally lack any stored energy and so can often save cost overall because they generally don't need snubbers and can also make EMI filtering easier and less costly. Some power switching devices also have lower output charge, and so cause fewer problems. (I plan to cover rectifiers in more detail in a separate issue.)

Given the switching devices, a good approach that maintains thermal efficiency is to carefully choose or design components, circuits and PCB/wiring layouts to increase the resonant frequencies in the circuit. This is so that the resonances occur where the harmonics from the switching operation (see Figure 8 in [13]) have less energy. Then "dampen" those resonances with snubbers.

Resistor-Capacitor (RC) snubbers are easy to design but lossy. Active devices such as rectifiers, Silicon Avalanche Diodes (SADs) or Metal-Oxide Varistors (MOVs) are sometimes used in snubber circuits, or instead of them, to recover some of the energy in overshoots and undershoots. Although they permit higher thermal efficiency than RC snubbers they will probably not provide such good suppression of emissions.

Any devices used for snubbing or energy-recovery should be very low-inductance types (never wirewound!) and be fitted directly across the switched terminals of the power switching devices, where the switching edges are the fastest and hence where the highest-frequency harmonics have the most energy.

Snubbers have to handle power, and in powerful converters it can be a problem to find resistors that can handle the snubbing power without adding inductance that merely "retunes" the ringing frequency rather than damping it.

2.5 Emissions from heatsinks

Switching device collectors and drains have stray capacitance to their heatsink through their thermal insulators (often 50 - 100pF), so their dV/dt injects transient currents into their heatsink, giving them a dV/dt too, radiating E-fields, and causing circulating common-mode (CM) interference currents.

I have seen a TO-220 heatsink tab on its own cause conducted emissions to be 50dB over the conducted limit at 150kHz. Connecting an additional thermal radiator to it just made things worse.

Connecting a heatsink to the safety earth/ground or 'chassis' injects the CM transient currents into it, making it noisy and often causing EMC test failure, because the safety earth/ground or chassis is not a sink for electrical noise!

Electronic designers often assume that "0Vs", "earths", "grounds" or "chassis" are sinks for electrical noise, and in my work fixing interference problems in real life I often meet people who assume that because they ran a PCB track to a chassis screw fixing point, or a wire with green/yellow insulation (or braid strap) to a safety earth/ground terminal block, or to a nearby water pipe, all the RF noise – even to several hundred MHz – should all somehow have "drained away".

But our universe doesn't work like that. The plain fact is that the laws of physics (actually, Maxwell's Equations, underpinned by the theory of Quantum Electrodynamics) means that <u>all</u> currents – including "stray" and CM noise currents – flow in closed loops. There are no exceptions, ever, to this rule (in *this* universe, anyway).

The loop that a current flows in, which at RF (say, above a few tens of kHz) can just as easily include paths through dielectrics such as heatsink thermal washers or the air (via stray capacitance and stray inductance) as paths along conductors.

All such current loops are electromagnetic antennas, and since we don't want our SMP or PWM circuit to emit excessive RF noise we call them "accidental antennas". Up to the point where their physical dimensions cause them to resonate, the larger the size of the current loop – the higher the levels of radiated and/or conducted emissions it causes.

The good news is that those same laws of physics mean that all currents of whatever type "prefer" to flow in lower-energy paths – which are the paths with the least impedance (generally the smaller-sized loops).

This simple observation of how our universe actually works shows us that what we need to do to reduce the emissions from the stray currents capacitively coupled into heatsinks, is to provide a small local loop (or loops) for the stray currents, which maintain a low impedance up to our highest frequency of concern.

These loops might well include conductors that we have labelled as 0V, ground, chassis, frame, protective conductor, safety earth, star point, V_{CC}, V_{DD}, 12VDC, 230VAC, whatever, but to a propagating electromagnetic wave (which is what a current loop really is, at any frequency) the label we put on a conductor does not matter, the only thing that matters is the part it plays in the impedance of its current loop (which, as I said earlier, can include dielectrics such as heat-sink thermal washers and the air).

The principle of designing small, hence local, paths for stray and other unwanted noise currents, to minimise their accidental antenna behaviour, is crucial to the cost-effective design of electronic products for both signal integrity (SI) and EMC. And not just for heatsinks!

This is because it is almost always easier, quicker, and much less costly to reduce accidental RF emissions at source, than it is to reduce them by using EMI mitigation techniques (e.g. filtering and shielding). Especially since the mitigation is usually thrown on at the end of a project when a product is failing its EMC tests and overshooting its market-launch deadlines.

For more information on this principle see [32], and for its application in systems and installations see [33]. For even more information read [4] or Chapter 2 of [5], and for the relevant in-depth analysis of Maxwell's Equations, read Chapter 2 of [34].

I discussed EMC design techniques for heatsinking in detail in [35], which is also covered by Chapter 13 of [5], so I won't repeat them here.

It is important to note that heatsinking can be combined with shielding to save assembly complexity and cost, and suitable techniques are described in [35] (and in Chapter 13 of [5]).

[35] and Chapter 13 of [5] both describe the technique of using a copper heatsink plane as a heatsink, which also makes the best low-impedance current return path for stray heatsink currents. They also mention using thicker copper plating for the plane, up to "8 ounce" or even "12 ounce" copper, plus the fact that some PCB manufacturers can laminate metal sheets or plates in the PCB's stack-up.

If the laminated sheets or plates are copper or brass, some PCB manufacturers can bond to them with the copper plating of the barrels of the via holes, to provide a Reference Plane for the circuits that has both a very low electrical resistance and a very low thermal resistance.

Low electrical resistance is of course very useful in power converters that must handle large currents, and it is also very useful for reducing interference from the systematic circulating currents, sometimes called “ground loop” currents, caused by connecting cable shields at both ends (as is required to make them provide good shielding to RF), see [36].

Other articles in this “Stand Alone” series will discuss advanced electrical solutions, such as PCBs with embedded wiring or embedded busbars, but in the context of this section we are more interested in the low thermal resistance of such a thick metal plane.

Used as a heatsink, a thick metal PCB plane helps dissipate the heat from the devices, keeping them cooler and/or allowing more power to be dissipated by board-mounted components, thereby reducing the overall cost of manufacture.

Extending the thick metal plate/plane beyond the PCB allows it to conduct the heat to a larger heatsink or a “cold wall”, making it possible to dissipate much higher levels of power on the board, and also providing a convenient method of physical assembly, saving even more cost. Figure 2D attempts to show this and should be compared with Figure 3 in [35] or Figure 13C in [5]

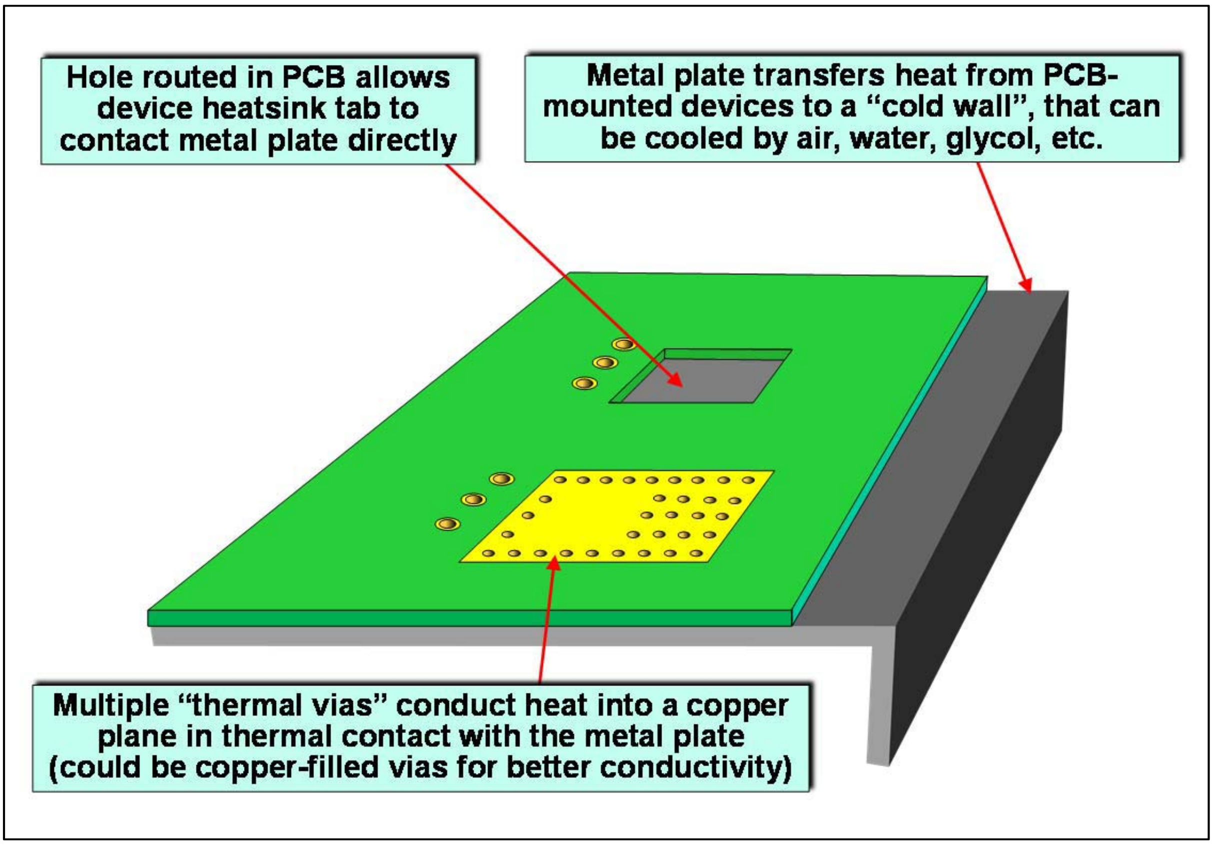

Figure 2D Example of combining a PCB with a metal heatsink plate

I mention the issue of “overall manufacturing cost” – which includes warranty costs – because this is actually important for the financial well-being and continued existence of a manufacturer, whilst the BOM cost (Bill of Materials cost) almost never matters. But every year I speak to electronic designers who are browbeaten by their managers to achieve the lowest BOM cost regardless of the effect on the overall cost of manufacture – which can have very serious repercussions up to and including loss of the company when too high a proportion of products are returned under warranty.

Regular readers of my articles will know that the current self-destructive obsession with low BOM cost is a particular bugbear of mine (see [11] and [12]).

Interestingly, in a recent series of training courses in China, covering the material in [5] and [37], to senior people in Chinese companies that are setting out to create their own brands, I made a point of saying that the lowest BOM cost had nothing to do with financial success, and what mattered was the overall cost of manufacture, and everyone there agreed as if it was – duh! – obvious.

Getting back to heatsinks and EMC, it is possible to use “plugged” vias, which are completely filled with copper, for a very much improved thermal conductivity to the PCB plane or heatsink on the other side of the board.

Copper-filled vias also stop solder paste from being sucked down the via hole and possibly causing a dry joint for the component, so can be helpful when designing with ball grid arrays (BGAs) by allowing “via-in-pad” layouts that save a lot of routing space, make it possible to use fewer layers in the stack-up, and also reduce the impedance in the BGA’s power busses, and so are good for EMC.

In May 2008, there were 18 PCB manufacturers offering copper-filled via services in the UK, and at least one of them will manufacture small quantities for a bare-board price of about 10% higher than boards with non-filled-vias.

An important issue with copper-filled vias is that their surface must be flat, so as not to trap bubbles of air when the solder paste is printed on them. During reflow soldering, such bubbles will expand and even burst, often creating a weakened or dry joint.

Prototype THP boards that will be manually soldered can have their vias filled instead with high-temperature solder to save the cost of plugging them.

2.6 Preventing self-oscillation

Switching devices can self-oscillate in themselves, causing high levels of noise emissions, and overheating. This is a problem for PowerFETs, especially paralleled PowerFETs.

I well remember a presentation in Stevenage UK given in the mid-1980s by Ed Oxner of Siliconix – one of the original inventors of PowerFETs. While experimenting with some early die in TO-3 metal cans (remember those?), he had two of them in parallel (to increase the current handling), side-by-side with their TO-3 cans (the FET drains) linked with a short copper bar. One leg of the DC supply was connected to the centre-point of the bar.

The PowerFETs were running stinking hot, but Ed could find no cause for this in the waveforms he measured around the circuit with his 1GHz oscilloscope. But he always measured the drain voltage by placing his ‘scope probe on the centre of the busbar where the DC supply was connected to the two FET drains.

After struggling with the problem for days, he accidentally touched his probe to one of the TO-3 cans, and immediately saw hundreds of volts peak-to-peak at nearly 1GHz. Using two probes, one on each TO-3 can, he saw that they both had the same waveform, but in antiphase! The short copper busbar had enough inductance in it to cause the paralleled PowerFETs to oscillate at nearly 1GHz, but the centre of the bar was like the centre of a see-saw – no voltage on that at all!

This was a big shock, of course, high-current semiconductors that could switch so quickly that they could oscillate at GHz had never been seen before, and remember these were very early prototypes, not production versions that would be expected to have “better” (i.e. faster) switching characteristics.

Anyway, the point of this little piece of history is that PowerFETs love to oscillate and cook themselves, and since most SMP and PWM development laboratories are only equipped with oscilloscopes that measure up to 300MHz or so, designers might not be able to see that self-oscillation is the cause of the excessive heat dissipation and poor thermal efficiency.

Converter designers working with power up to, say, 1kW, therefore need to have oscilloscopes and probing systems that can measure to well over 1GHz. Converter designers working with low-voltage PowerFETs at up to, say, 100W, probably need at least 6GHz oscilloscopes and probing systems with appropriate probing techniques to be sure not to be blind-sided by self-oscillation.

Oscilloscope probing at GHz does not allow the use of long probe “earth leads” with crocodile clips, and GHz+ ‘scopes and probes are expensive (especially if high-voltage isolated and differential!) – so an easier technique might be to use a small unshielded close-field probe with a spectrum analyser.

An unshielded magnetic loop probe will detect GHz currents and voltages when all it is, is a 10mm-diameter shorted-turn of wire at the end of a 50 ohm cable.

Figure 2E is a photograph of the unshielded loop probes I have used since 1990 to solve very many real-life EMC problems, although these ones are only insulated enough to be suitable for low-voltage circuits.

I have blown up the input mixers on two (low-cost, thankfully) spectrum analysers by accidentally touching an uninsulated probe to a live circuit – it only takes a few volts! So all close-field probes must be insulated with sleeving or encapsulation, and for safety with mains-powered circuits the insulation should consist of at least two separate layers, each capable of withstanding the peak surge voltage. IEC/EN 61010 is the relevant safety standard (with different parts for instruments, probes, etc.).

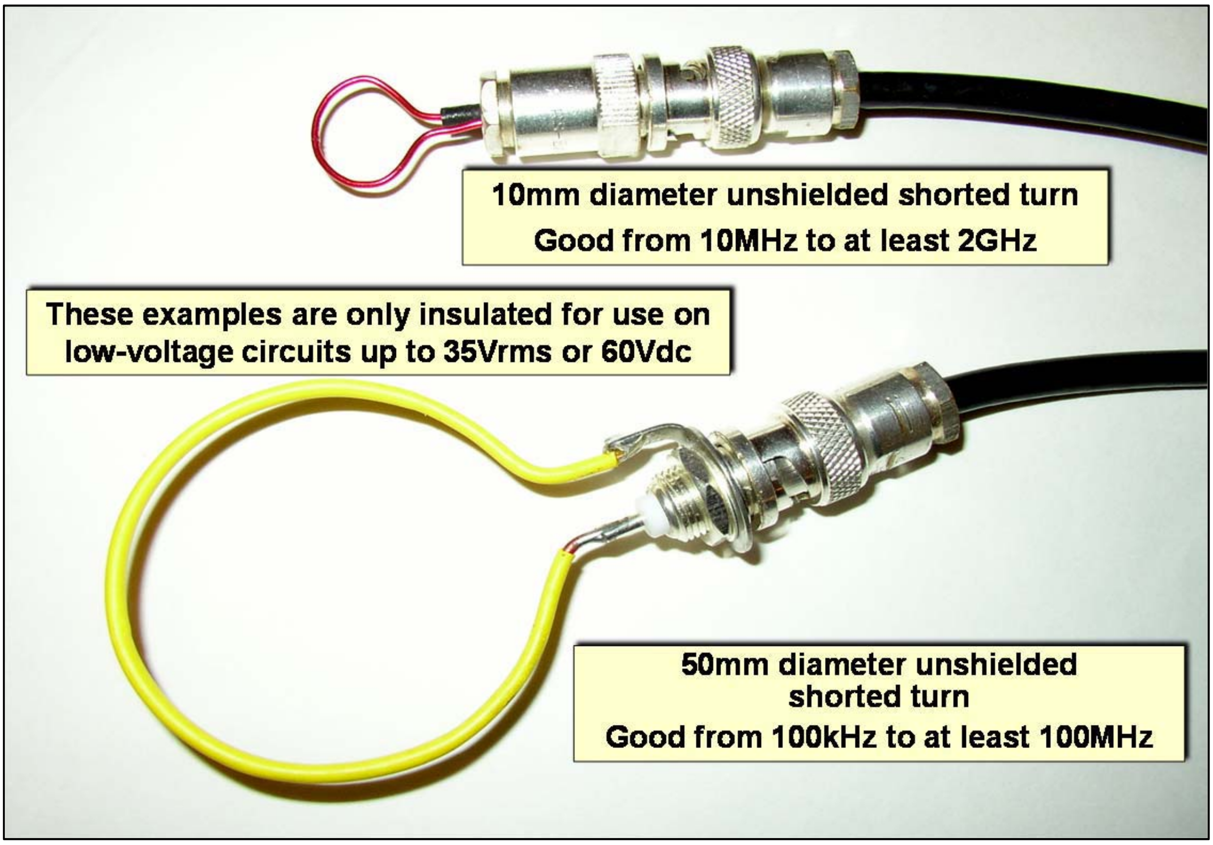

Figure 2E Two examples of low-voltage-rated home-made close-field probes

Figure 2F shows some of the low-cost spectrum analysers that can be used to great effect with such probes. For portability I use the Thurlby-Thandar shown bottom right, which cost me £862 a few years ago and measures to 1.3GHz. They also have a similar model that does up to 2.7GHz. These little spectrum analysers are very handy indeed, help solve all sorts of EMC problems quickly when used with close-field probes, and are priced so that everyone can have one (and why not)?

Figure 2F Four examples of portable spectrum analysers

Of course, if you work for a large or rich company, you absolutely must have the Rohde & Schwarz or Agilent portable spectrum analysers, pictured bottom left and top right.

The Agilent E7400 series (top left) doesn't fit in a pocket, but is light and rugged and easily portable, and includes Quasi-Peak and Average detectors that meet the CISPR specifications for all but the most difficult noise sources. Because they are a 15 year old design there are many 2nd-user examples available at very good prices.

They don't have Ethernet, Wi-Fi, Bluetooth or USB connections – and nobody now uses the GPIB or floppy disc interfaces they do provide – but in any case I've found it is generally quicker simply to photograph the display of a spectrum analyser than use their PC interface software to capture screen-shots. (Where would we be without digital cameras?)

(Incidentally, I will be providing a few hours of demonstrations and tutorials on the wonderful things one can do with close-field probes and a low-cost spectrum analyser at EMC-UK 2011, see the advertisements in this issue of the EMC Journal.)

By waving such a probe near to a power converter and seeing the spectrum analyser display high levels of noise emissions at several hundred MHz or even a GHz or two, it only takes a few *seconds* to identify a self-oscillation problem, quicker than connecting a 6GHz 'scope probe. And this technique just as quickly shows when a "fix" has worked.

I recommend checking all hoped-for fixes with the devices cooled to -20°C or less, and then again with them heated to +150°C or more, for example by using dry nitrogen temperature forcing equipment (or freezer spray and a hot-air gun). This will help ensure that the fix works with the inevitable variations of device characteristics found in serial manufacture (although it is no guarantee).

Such self-oscillations in semiconductors are generally cured by driving their gates/bases through low-value resistors (typically between 3.3 and 100 ohms), or soft-ferrite RF beads having a *resistance* (not inductive reactance) of the same order at the frequency of oscillation.

It is most important that the resistor or bead is located *immediately adjacent* to the devices' gates or bases. In the case of leaded devices, they should usually be fitted in the legs of the devices, which usually means threading ferrite beads onto the legs prior to assembly on the board.

Sometimes it seems that placing a ferrite bead in the source/emitter lead instead (or as well) prevents self-oscillation. As before, the bead must be very close to the body of the device.

Paralleled Schottky rectifiers can also self-oscillate, because at some currents they exhibit negative resistance. In this situation, the low-value metal-oxide resistors or ferrite beads must be located immediately adjacent to the body of each paralleled Schottky, slipping a ferrite bead onto the legs of leaded devices, if necessary.

2.7 Stability and decoupling

AC-DC and DC/DC power converters are generally used with decoupling capacitors on their outputs, which – by design – have very low impedance over a wide frequency range, possibly causing problems for the stability of the converters' voltage feedback loops.

Unstable converters cause high levels of emissions, so it is important to make sure they will remain stable with the full range of resistive/capacitive/inductive loading that could foreseeably occur in real life.

Stability can be hard to measure at the output, due to the filtering effect created by the decoupling capacitors, so it is best to monitor a point within the voltage control feedback loop – usually the error voltage – to see if it oscillates (or rings too much) when the resistive part of the test load is mechanically switched from low current to high current, and when switched back again. Ideally, the peak overshoot on the feedback should be less than 60% more than the steady-state value and there should be less than 2 cycles of ringing, with any likely combination of real-life loading.

To help ensure stability in real-life, despite differences between batches of devices, repeat the switched-load stability tests with the control IC (ideally the whole circuit) circuit at -20°C or less, and then again with it heated to +150°C or more, for example by using dry nitrogen temperature forcing equipment.

An alternative to the above switched-load stability test is to load the converter with single frequency and sweep it over a range while monitoring the output voltage on an oscilloscope. This makes it easier to determine the poles and zeros of the converter's transfer function, to use in calculations or simulations. A description of a suitable method is given in [38], and can easily be adapted to other kinds of switchers.

2.8 'Spread-spectrum' techniques

Switching circuits emit all their RF leakage energy at basic switching frequency and its harmonics, and with a constant (fixed) switching frequency all of their emitted energy falls into very narrow frequency bands, which consequently measure as having high levels of emitted noise.

However, varying the basic frequency quasi-randomly, e.g. by using "spread-spectrum" clocking for switchers and "dithered" or "chaotic" PWM, spreads (smears) the emitted frequencies around, and – if the frequency spread exceeds the resolution bandwidth (RBW) of the EMC measuring receiver (e.g. CISPR specify 9kHz RBW when measuring in the range 150kHz to 30MHz) – the emissions will be measured as being at lower levels. [39] [40] and [41] may be interesting.

Figure 2G shows an example of the reduction in the peak levels of emissions achieved by using a particular kind of spread-spectrum technique. The technique has only reduced the peak levels by about 10dB, but when the Quasi-Peak or Average detectors are used, as required by most EMC emissions test standards, we could expect the measured emissions to reduce further. If the spread-spectrum technique concerned is really very effective, measured emissions could be reduced by up to another 15dB (Quasi-Peak), or 30dB (Average).

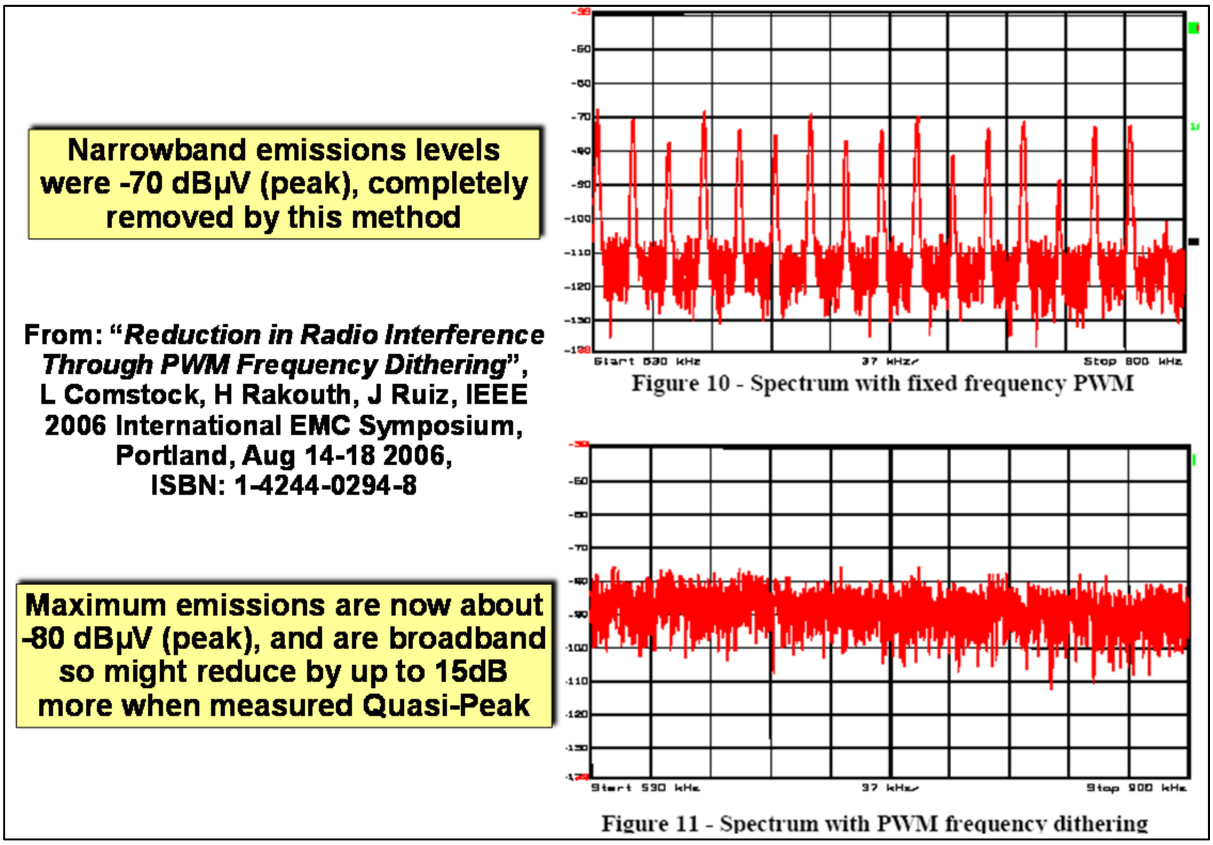

Figure 2G An example of reducing conducted emissions by spread-spectrum clocking

For example, to make a 3dB difference in the 150kHz – 30MHz band (9kHz RBW), the switching frequency would need to spread over at least 12kHz, and for a 6dB reduction at least 18kHz. For a 180kHz basic switching frequency, an 18kHz spread spectrum would be described as ±10%, but a 1.8MHz switcher would only need ±1%.

It is worth noting that this reduction is only achieved on EMC emissions tests that use integrating detectors (e.g. the CISPR Quasi-Peak, Average, and RMS detectors) and a test that uses a Peak detector (as specified by some ETSI standards) would measure just the same regardless of the frequency spreading.

Some large converters have always used spread-spectrum clocking to avoid acoustic noise and vibration that could otherwise cause mechanical damage.

Of course, spread-spectrum techniques don't reduce the instantaneous emissions at specific frequencies, and only measure lower emissions because of the integration characteristics of emissions detection instruments. So, with fixed-frequency radio channels, spread-spectrum clocking is less likely to cause high-level interference, but more likely to cause low-level interference, but it might be more likely to cause interference with certain types of digitally-modulated radio channels.

Computers and other IT equipment can now be interfered with by noise lasting less than a nanosecond, so spread-spectrum clocking might be best regarded as a trick to pass tests, rather than an effective means of reducing a product's interference in real life.

2.9 Beating and intermodulation

The frequencies (basic + harmonics) generated by an HF switcher will "beat" with the frequencies generated by other switchers, choppers, even clocked digital circuits or the sampling clocks of A/D and D/A converters, in the same product.

This beating is caused by intermodulation in semiconductor junctions and any other non-linear devices, and it creates new frequencies at the sums and differences of those frequencies and their harmonics.

Beating is especially troublesome when the intermodulation frequencies fall within the passband of an analogue circuit, where they cause whines, whistles, and pulsating noise signals, with frequencies that continually fluctuate as the original frequencies drift slightly due to temperature, load current, etc. On video signals, beating noises generally appear as bright diagonal lines or cross-hatches on the display.

Figure 2H shows a simple example with just two frequencies, 40 and 50kHz.

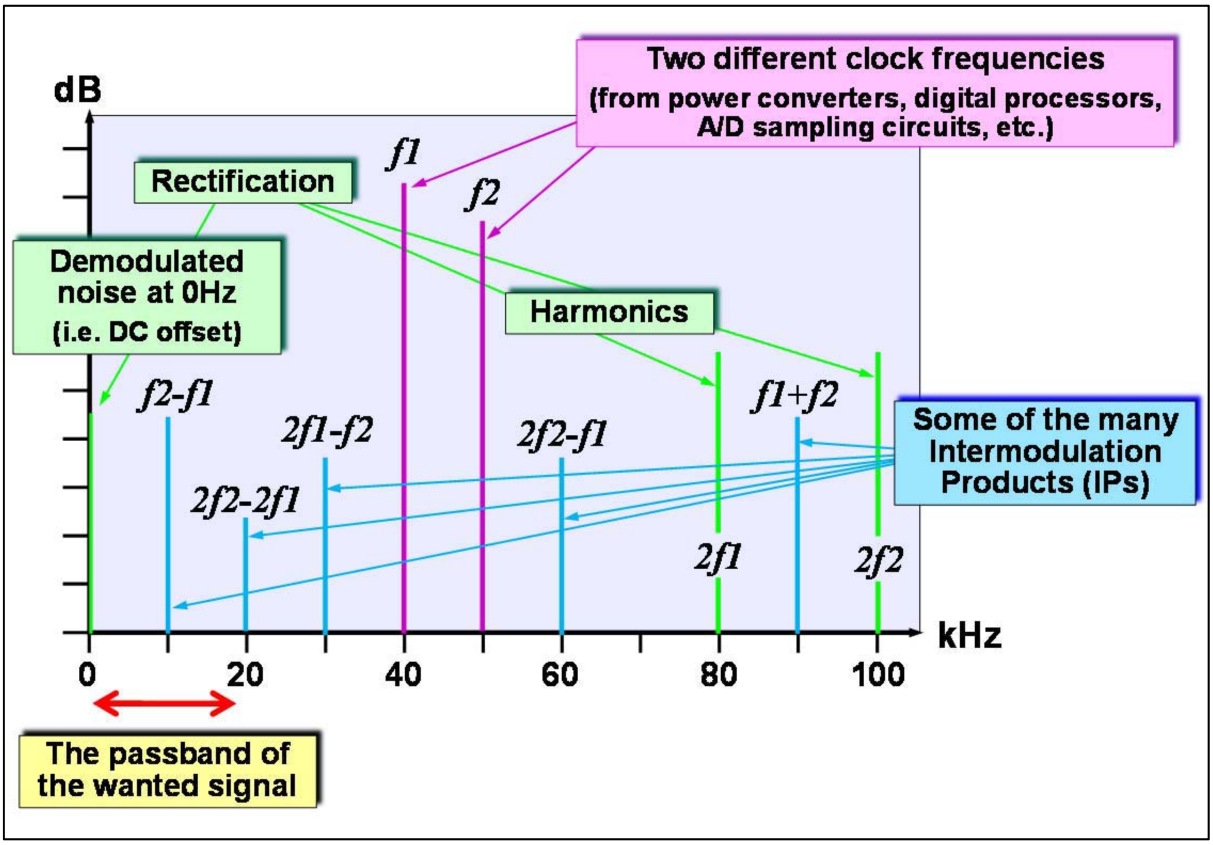

Figure 2H Example of beating (intermodulation) causing noise in the signal passband

Non-linearities in the signal-processing devices and circuits rectify each of the two clock frequencies, resulting in demodulated noise at 0Hz, generally called a DC offset or zero-shift, which can be bad news on its own. Rectification also creates harmonics, at twice, three times, four times, etc., the individual clock frequencies. Figure 2H only goes up to 100kHz so only shows the 2nd-order harmonics.

The two original clock frequencies "beat" to produce what are called the 1st-order Intermodulation Products (IPs). There are just two of these, at their sum and difference frequencies: 10kHz and 90kHz respectively. Their difference frequency of 10kHz happens to fall within the signal passband, and makes a noise in the signal at that frequency. If it is an audio signal it would be called a whistle, and if it was a lower frequency it might instead be called a whine, even (if low enough) a hum.

If either of the two clock frequencies vary, this in-band IP noise will vary by the same amount, changing the pitch of the whistle.

Figure 2H also shows three of the four 2nd-order IPs, when the 2nd-order harmonics add and subtract from each other and from the original frequencies. The missing 2nd-order IP is at 180kHz and so not visible on this scale. There are also six 3rd-order IPs, eight 4th-order IPs, ten 5th-order, and so on, not shown on this figure but generally diminishing in amplitude (unless they happen to coincide with a resonance).

With the two original frequencies chosen in this example, only $f2\text{-}f1$ lands in the wanted signal's passband, but $2f2\text{-}2f1$ is only just outside it, at 20kHz, and a small reduction in either of the clock frequencies (or both) could make it move into the signal band.

The best solution to prevent in-band IP noise is to phase-lock (synchronise) all digital clocks, sampling gates, switchers and choppers to one master clock, so that so that the beat/intermodulation frequencies remain fixed.

The individual operating frequencies can then be chosen so that all of their IPs fall *outside* the signal passband (either above or below) maximising the signal-to-noise ratio (S/N, or SNR, or THD+N) performance of the circuit.

Some HF switchers and chopper control ICs provide master clock inputs, which force the circuit to switch at the same rate. But where devices must use different frequencies, they can all be phase-locked to the same master clock by simple frequency dividers, phase-locked loop frequency dividers/multipliers, and even by fractional-N synthesisers.

There are several free intermodulation calculators available from the Internet, usually intended for use by radiocommunication companies but easily used for the type of calculation required to keep the beat frequencies out of a signal's passband.

However, phase-locking two or more "noisy" circuits can increase emission levels if it results their fundamentals or some of their harmonics coinciding at the same frequency. A solution to this can be to synchronise the various circuits and add a time-delay (a fixed phase angle) between each one, which causes the peak switching currents and peak emissions to be reduced.

The best results are achieved by running two similar converter circuits in antiphase, three similar converters in three-phase, or (to be general) N similar circuits in N-phase. This leads us into the next section.

2.10 Multiphase/interleaved converters

This has become the standard method of providing the very high current low-voltage DC supplies for microprocessor cores, which can require 0.9VDC at 100A or more.

Instead of a single high-current DC/DC converter, two or more smaller identical converters are operated in parallel and at the same frequency – with appropriate phase angles between their clocks.

The ripple voltages at the DC inputs and outputs of the multiphase/interleaved converters are lower because of current cancellation, and the ripple frequencies are higher (by N, the number of converters) – which makes them easier to filter with smaller components.

Such "multiphase" switchers need less decoupling capacitance for the same ripple voltage, at both their DC inputs or outputs, than would a single supply (shorter "load hold-up" times between charging pulses). Or they can use the same value of decoupling capacitance at the output for lower output DC ripple, or at the input for lower conducted emissions on their DC inputs (and, eventually, on their mains leads).

For example, the LTC6909 multiphase converter clock generators provides 8 clock outputs with pin-programmable phase angles from 45° to 120°, a resistor-programmable clock frequency from 12.5kHz to 6.67MHz, and can spread-spectrum by up to ±10%.

[24] recommends the use of "Interleaved BCM" in power factor correction (PFC) circuits (which I intend to cover in a separate article) to improve efficiency and reduce EM emissions.

3 Pulse-Width-Modulation (PWM) "choppers"

PWM choppers "simulate" or "synthesise" AC or DC output waveforms by varying the mark-space ratio of the HF power switching circuits. A fixed mark-space gives a DC output, and a repetitively-varying mark-space gives DC + AC. AC output (e.g. a sinewave with no DC offset) is achieved by ensuring the average value of the repetitive output voltage is zero.

Typically, the chopping frequency is generally at least ten times higher than the highest AC output frequency required, so as to produce an output waveform with an acceptably low distortion for the application.

Figure 2J shows a simple PWM chopper circuit, using an "H-Bridge" of power switching devices to provide four-quadrant control of the output AC, AC+DC, or DC.

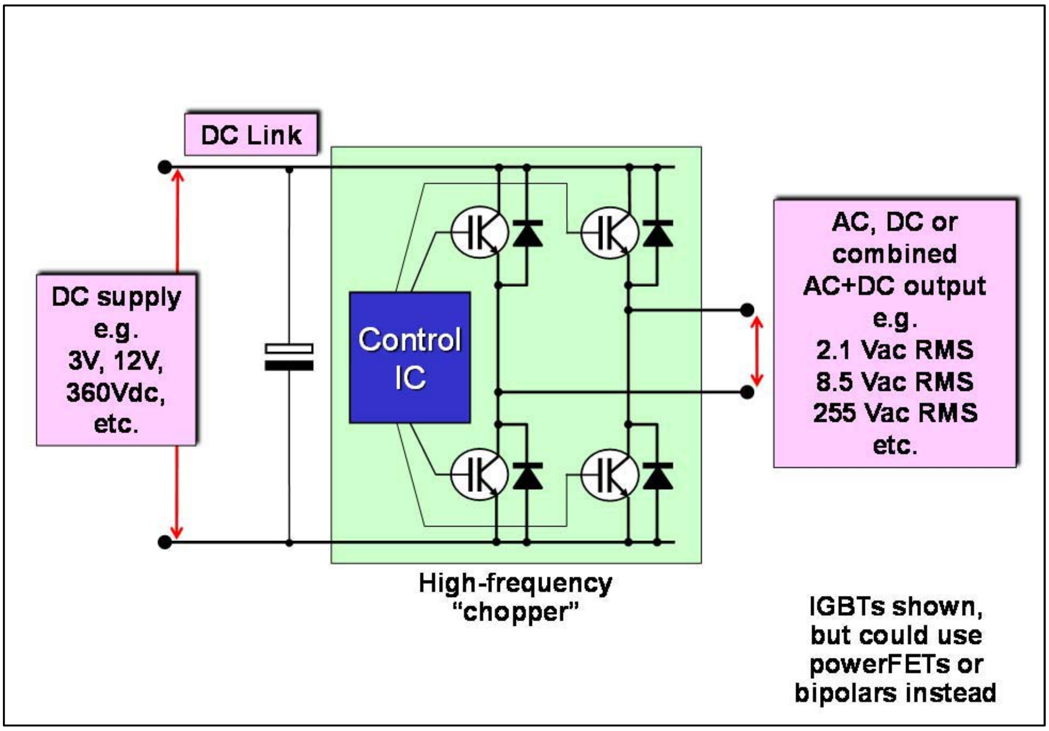

Figure 2J Example of single-phase (1ϕ) PWM chopper

Figure 2K is a simple example to illustrate how PWM works. The output voltage is "hard-switched" to achieve low heat dissipation in the power devices and the mark-space ratio of their switching waveform is varied repetitively to – in this example – produce a current that, when integrated by the inductance of the load (e.g. an electrical motor) it approximates to a sinewave.

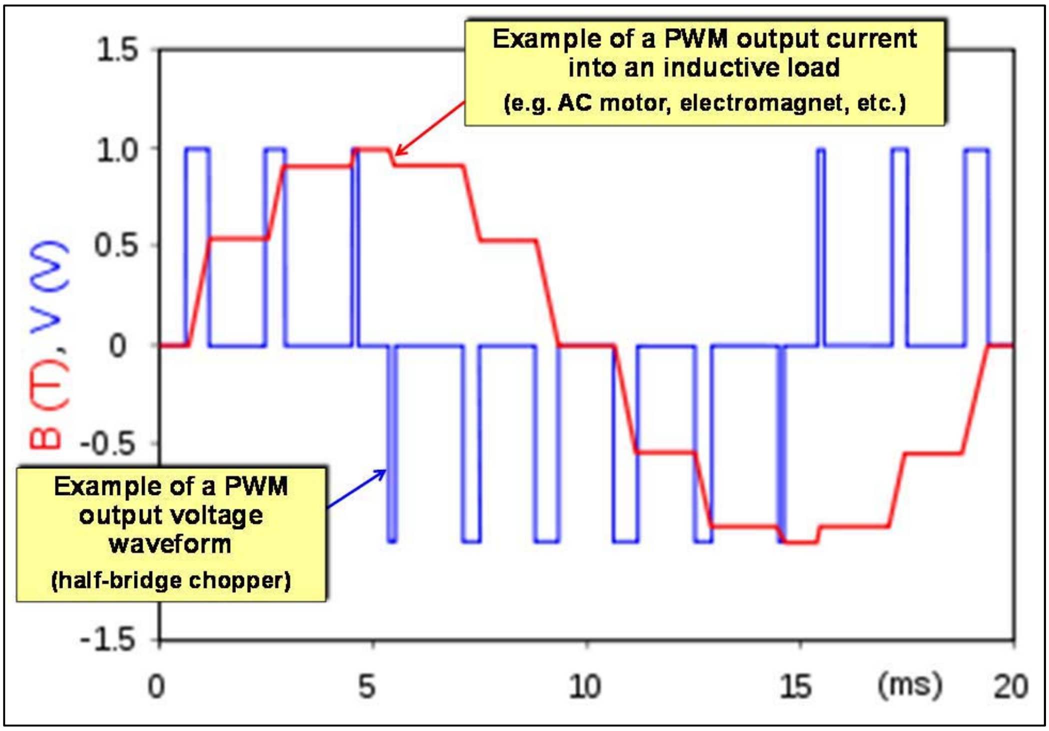

Figure 2K A simple example of PWM

Where the output requires very low distortion, e.g. for audio signals, the chopping frequency may need to be more than 100 times the highest frequency. At the other extreme, some DC-AC inverters that provide 110/230V for powering mains-powered equipment from car batteries (for example) only switch at three times their AC output frequency, because the rectifier-capacitor inputs of their intended loads (chargers for laptops and cellphones) will generally work and not be damaged by their very distorted mains waveforms.

The output PWM voltage is usually connected directly to the load, relying on the load's electrical inductance and/or mechanical and/or thermal inertia to integrate the "chopped" PWM waveform to create the desired resultant DC or low-frequency waveform in the desired parameter, for example magnetic flux, mechanical motion, heat, etc.

Unfortunately, connecting a PWM output to the load is like connecting a powerful radio transmitter, transmitting at its fundamental and many harmonics, to an electrical load instead of to a radio antenna. A 10kW PWM motor drive is really just a 10kW radio transmitter with a multifrequency output, and because the cables used, and the loads, are not impedance-matched to the PWM output there will be ringing, reflections, etc., which could cause huge amounts of conducted and radiated emissions.

It will be appreciated that it doesn't take much "leakage" from a 10kW motor drive's output circuit to cause a failure to comply with the CISPR limits for radio frequency (RF) emissions!

Figure 2L shows a real-life measurement of the output voltage and current of a 700kW variable-speed AC motor drive (a "VSD"). The measurements were made with a "power quality" meter that had a bandwidth limited to 5kHz, and since the switching frequency of the converter was around 5kHz the meter could not accurately display the switching waveform.

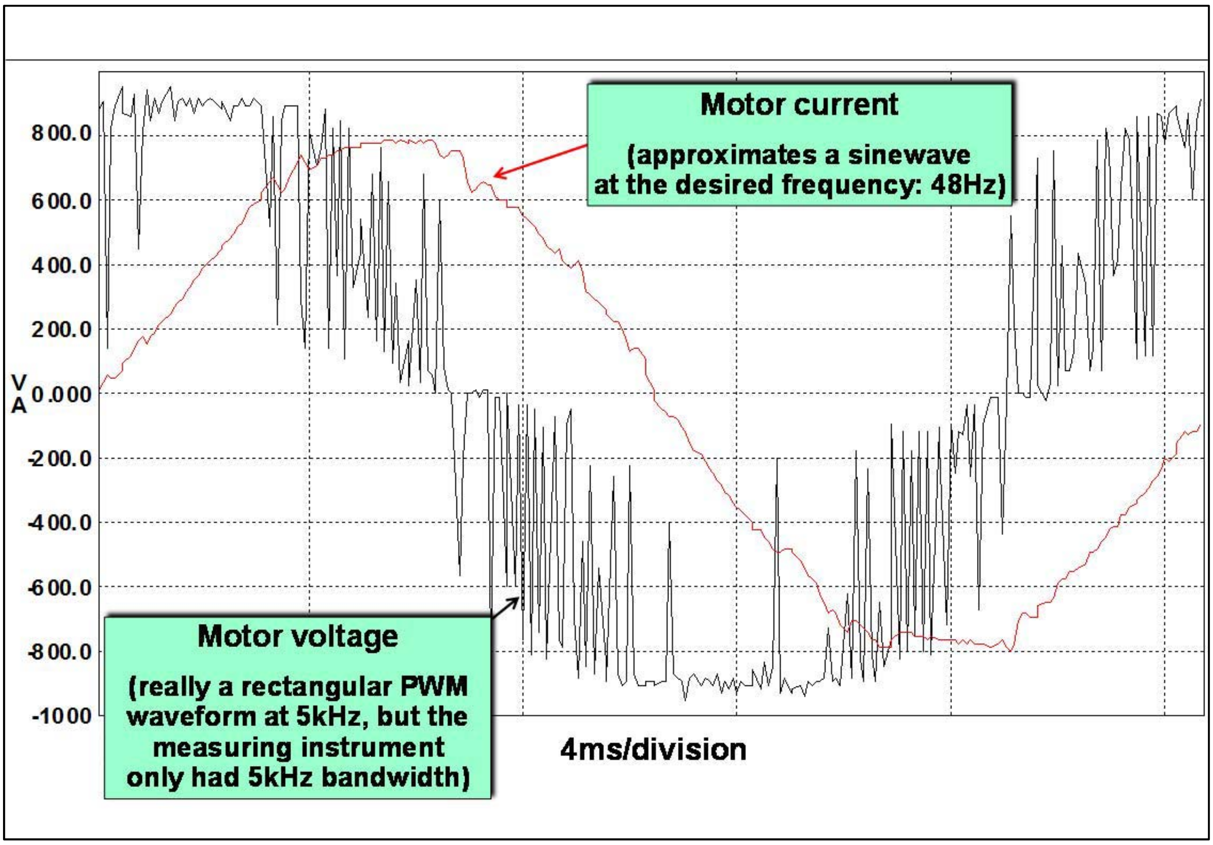

Figure 2L Example of 700kW variable-speed AC motor drive

Many of the EMC design issues discussed above in the context of HF switchers, also apply to HF choppers. For example, too-low rise and fall times can cause excessive RF emissions, overshoot and ringing can require snubbing, power switching devices can self-oscillate, feedback loops can be unstable, and multiple HF sources in one product can cause intermodulation ("beating") problems within the bandwidths of the wanted signals.

But some of the EMC design techniques aren't appropriate. For example there are circuit topologies for DC-output switchers that are designed to reduce their emissions, which can't be applied where an accurate PWM output is required.

If we try to simulate a sinewave that has peaks equal to (or very close to) the maximum DC supply ("DC Link") voltage, we can get switching pulses that are less than the minimum time required to turn-on and then turn-off the power devices. The result is that transistors will not turn fully on or off, which can cause excessive EMI, if not excessive thermal dissipation, even damage. It is easy to detect this problem, and to avoid it by not specifying such extreme mark/space ratios.

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