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EMC design of Switching Power Converters Part 4 - Design techniques for HF isolating transformers

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EMC design of Switching Power Converters

Part 4 – Design techniques for HF isolating transformers

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Contents

Issues 93 and 94 of The EMC Journal carried the first two of these “Stand Alone” articles [13] [42] 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* power converter applications, including: consumer, household, commercial, computer, telecommunication, radiocommunication, aerospace, automotive, marine, medical, military, industrial, power generation and distribution, in products, systems or installations.

And I will also cover hybrid & electric automobiles, electric propulsion/traction; “green power” (e.g. LED lighting); and power converters for solar (PV), wind, deep-ocean thermal, tidal, etc.

This Stand Alone article addresses the circuit design issues associated with the high-frequency (HF) isolating transformers.

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.

Before I make a start on the title subject of this article, I must return to Section 2.2 in Part 2 [42], which very briefly mentioned Ćuk converter topologies by simply referencing [23].

[23] is a Wikipedia page that only describes the non-isolated Ćuk topology, and doesn’t do justice to the very wide range of Dr Ćuk’s resonant-mode converter topologies. However, [23] partially redeems itself by referring to what I have listed below as [43] – a page in the website of www.boostbuck.com dedicated to showing how easy it is to use the Ćuk topologies to beat the pants off all other DC-DC converter topologies.

I can do no better than to copy the words on the www.boostbuck.com homepage:

> "The purpose of this web site is to show Power Electronics Engineers how to design the Boostbuck (Ćuk) Converter easily and painlessly.
>
> "The motivation is to encourage general use of this Optimum Topology to improve performance of industry designs. To that end, the optimality of this family of 4 converters is shown plainly vis. a vis. many of the other common topologies currently in use.
>
> "Along the way, a number of very sweeping generalizations are made to simplify the design process. These are each justified in turn, and save the engineer much time spent chasing after popular, but unfruitful, design approaches."

I should also mention here that one of the strong points of Ćuk converter topologies is their ability to combine the input and output inductors into one smaller component whilst simultaneously reducing noise emissions and ripple voltage, thus also reducing the size of the filter and DC storage capacitors.

Isolating Ćuk converters can combine *all three* magnetic components: the input and output inductors *plus* the isolating transformer; into a single smaller component – whilst also reducing noise emissions from the input and the voltage ripple at the output.

It seems that the reason that most people don't use Ćuk topologies is the difficulty they have in designing the integrated magnetic components, but www.boostbuck.com has a web page that claims to make this an easy task.

What made me revisit the issue of Ćuk converters, prompting me to sing their praises here, is an article by the good Dr Ćuk himself in the latest edition of Power Electronics Europe magazine [44].

This article extends his DC-DC converter topologies into AC-DC converters that do not require an input rectifier – thereby reducing the typical AC-DC converter's fourteen switching devices (transistors and rectifiers) and three magnetic components, to just four switching devices and one magnetic component, as shown in Figure 3A.

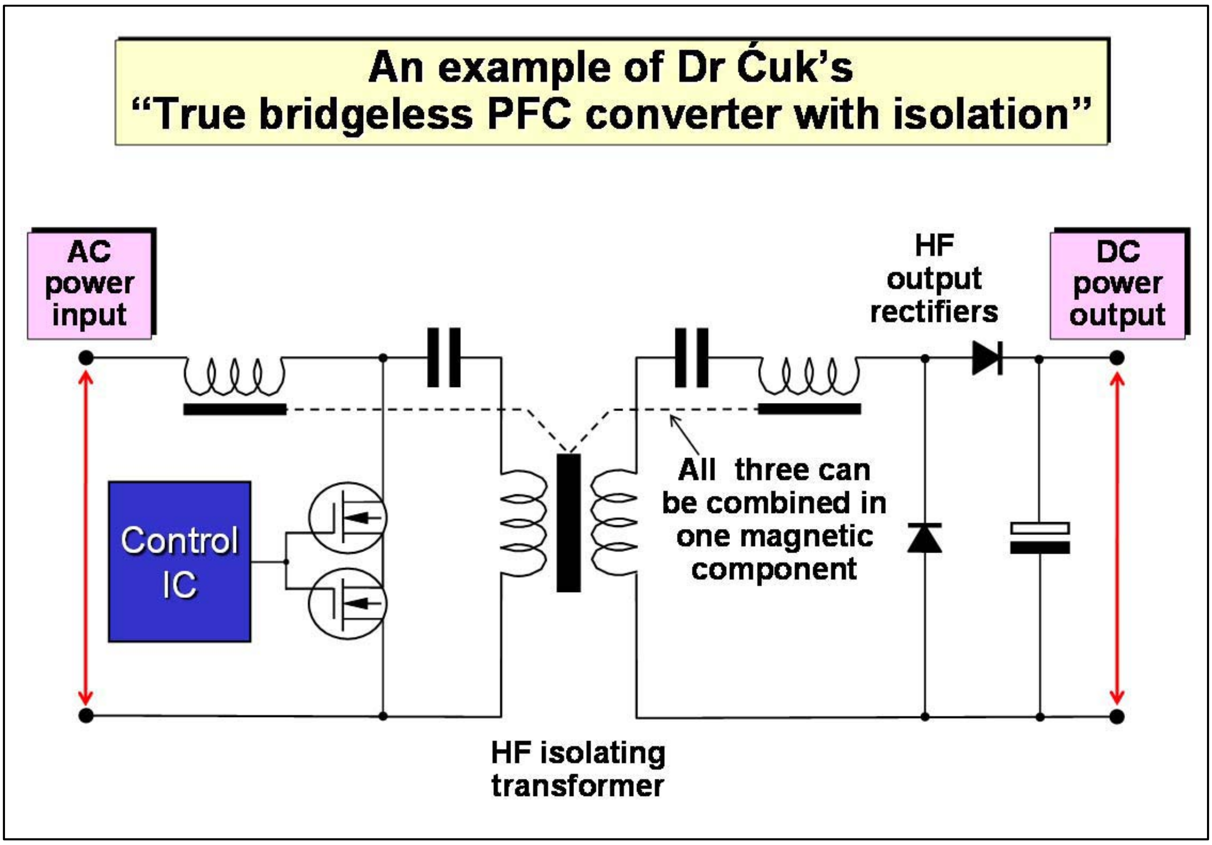

Figure 3A Example of Dr Ćuk's new converter topology

[44] also describes how using three such converters on a three-phase AC supply can eliminate the need for any energy storage/smoothing capacitance, saving even more cost and weight.

All single-phase AC-DC converters deliver pulsating DC power (a full-wave rectified sinewave) at 100Hz (or 120Hz in the US and other "60Hz" countries) because a single-phase AC mains supply provides a 50/60Hz sinewave with an unavoidably pulsating AC power. As a result, they need a lot of energy storage or "smoothing" to provide the load with a DC output that has acceptable levels of voltage ripple.

However, a three-phase mains power supply actually provides a continuous, constant AC power, but using a three-phase bridge rectifier sums three individual 100/120Hz full-wave-rectified sinewaves, each shifted 120° in phase. The result is a DC voltage, right enough, but one with a considerable level of AC ripple consisting of half sinewaves at 300/360Hz.

So, providing the load with a DC voltage that has an acceptable level of voltage ripple again needs an energy storage or "smoothing" capacitor – although it does not need to be nearly as large, for a given value of ripple voltage, as if the same load power was being delivered from a single-phase mains supply.

However, the three-phase Ćuk converter described in [44] has no bridge rectifiers and needs no energy storage/smoothing capacitors, thereby reducing cost, weight and size even further. Constant three-phase mains power in, constant DC power out.

A downside of this approach may be that because there is no "hold-up" energy provided by large energy storage/smoothing capacitors, the usual dips, dropouts and imbalances in the three-phase public mains electricity supply can harm the DC output's power quality. But where the three-phase supply has a high enough quality (e.g. when powered from a dedicated generator, such as an aircraft's jet engine's 110Vac generator) this might not be important.

I understand that Dr Ćuk's converters were used on the Space Shuttle (may it Rest In Peace) because of their high efficiency and small size and weight, so maybe aerospace applications are a good application for using this new three-phase AC-DC converter topology without energy storage/smoothing capacitance.

4 EMC design of HF isolating transformers

4.1 Introduction to the example

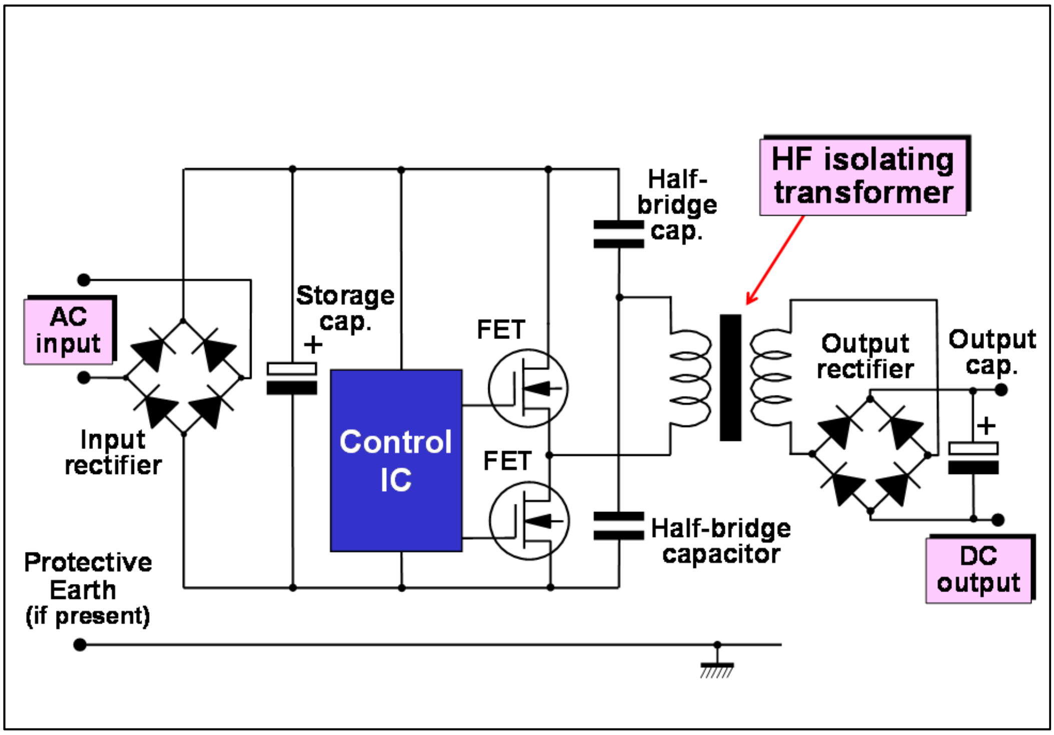

Figure 3B Example of an AC-DC SMPSU converter

This is the (very) basic circuit schematic that will be used in the discussions below, and in several later articles in this series.

It is an isolated half-bridge Pulse Width Modulated (PWM) converter (a "chopper") – because I had to draw *something* to use as a practical example. There are many other types of power converter, and the EMC design principles discussed below apply to them all (except where noted).

Figure 3C shows the example I will use for the 2-layer printed circuit board (PCB) layout of the converter in Figure 3B, and Figure 3D shows a sketch of this example's complete assembly.

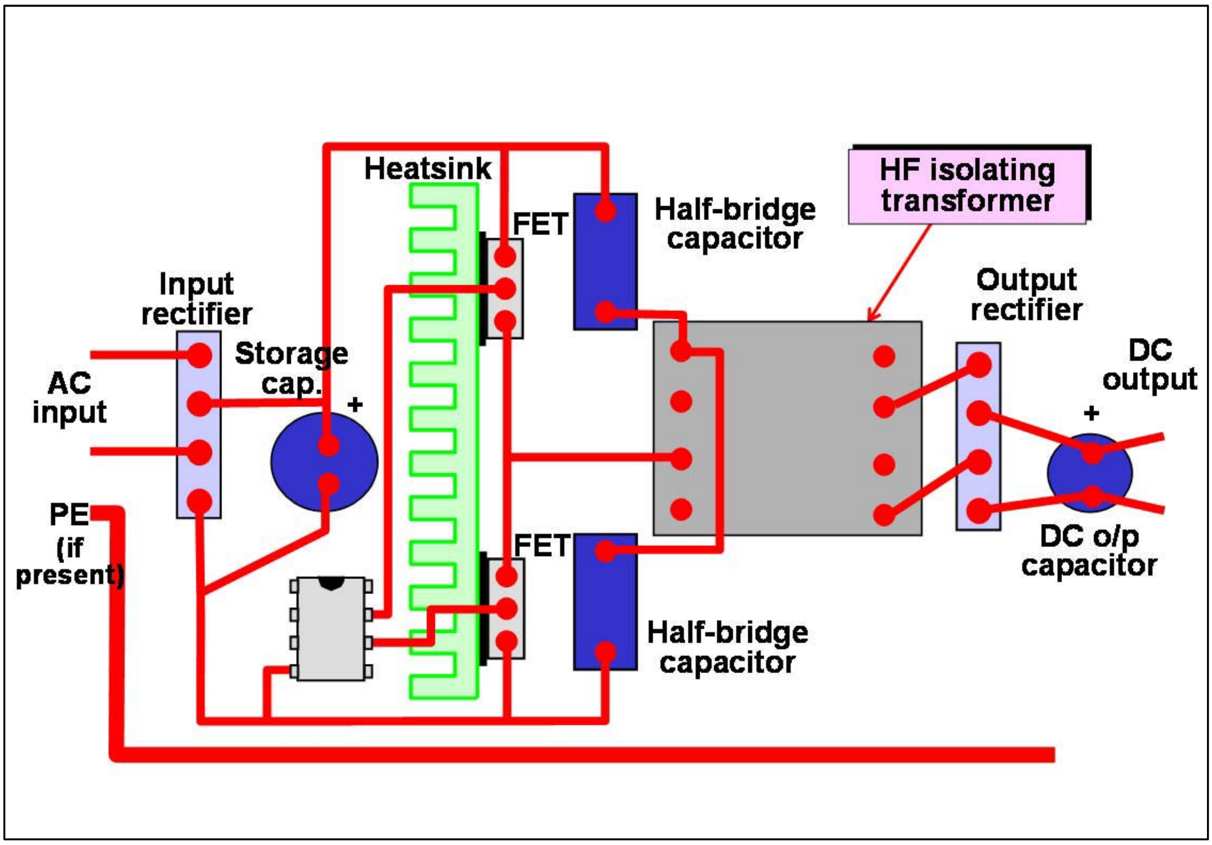

Figure 3C The basic PCB layout for the example SMPSU converter

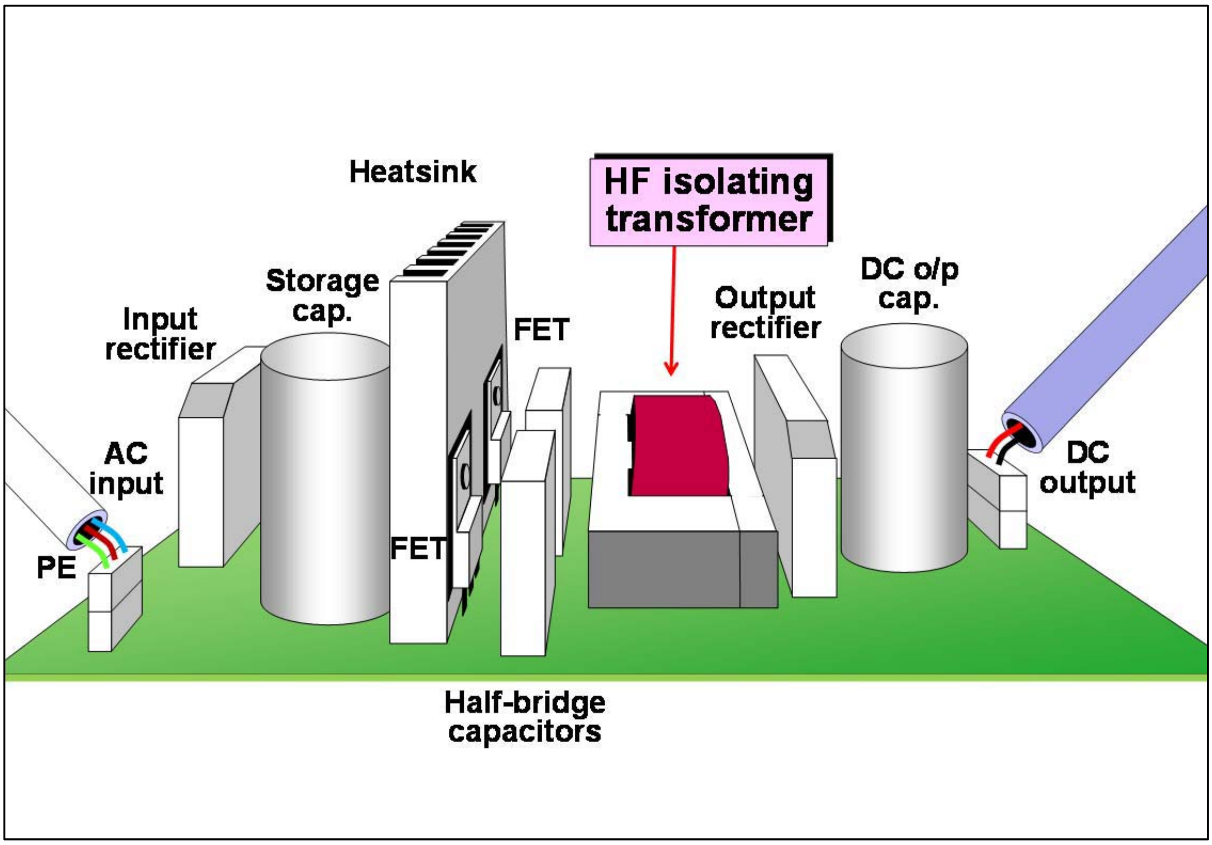

Figure 3D The basic PCB assembly for the example SMPSU converter

The following sections describe a number of good EMC design and construction practices relating to isolating transformers, with the aim of reducing the reliance on EMC mitigation methods such as filtering and shielding to a minimum, to save cost, size, weight, and time-to-market.

There are other methods than transformers for transferring electrical power across an isolation barrier, for example the “flying capacitor” method. However, isolating transformers are the most common method used, and no other methods are discussed in this article.

Reducing emissions by a few dBs *here* by applying one low-cost transformer design and construction technique, then reducing a few more dBs *there* by applying another, and so on, pretty soon adds up to a significant reduction in emissions with the lowest overall cost of manufacture, in the shortest time.

But reductions in time-to-market will not be achieved unless these good EMC design techniques are all taken into account *from the start of the project*.

I’m not saying they must all be done, only that they should all be considered individually and each one should only <u>not</u> be used if there is a <u>fixed technical constraint</u> that makes it inappropriate. In such cases, an alternative should be employed.

If EMC design is ignored until the end of a project, then *when* (not *if*) the product fails its EMC tests the timescales are usually so desperate that the bill of material (BOM) cost targets are forgotten as costly filtering and shielding “fixes” are thrown at the product until some combination of them allows it to pass, eventually. Many EMC engineers worldwide are fully employed in doing just this, on product after product. But the eventual cost and the time it takes are unpredictable, so it represents a very significant financial risk.

If only Design and EMC engineers learned to speak the financial language of their managers (it’s very easy, really, but it’s not an engineering discipline) they would soon be able to persuade them that designing using good EMC techniques from the start of a project would significantly reduce financial risks. A worked example is given in Chapter 1 of [5].

4.2 How the interwinding capacitance causes emissions

Isolating transformers have interwinding capacitance (C_{STRAY}) between each isolated winding. In this example I will consider the C_{STRAY} between a primary and a secondary, although similar principles apply to any pair of windings. Figure 3E shows where this C_{STRAY} arises.

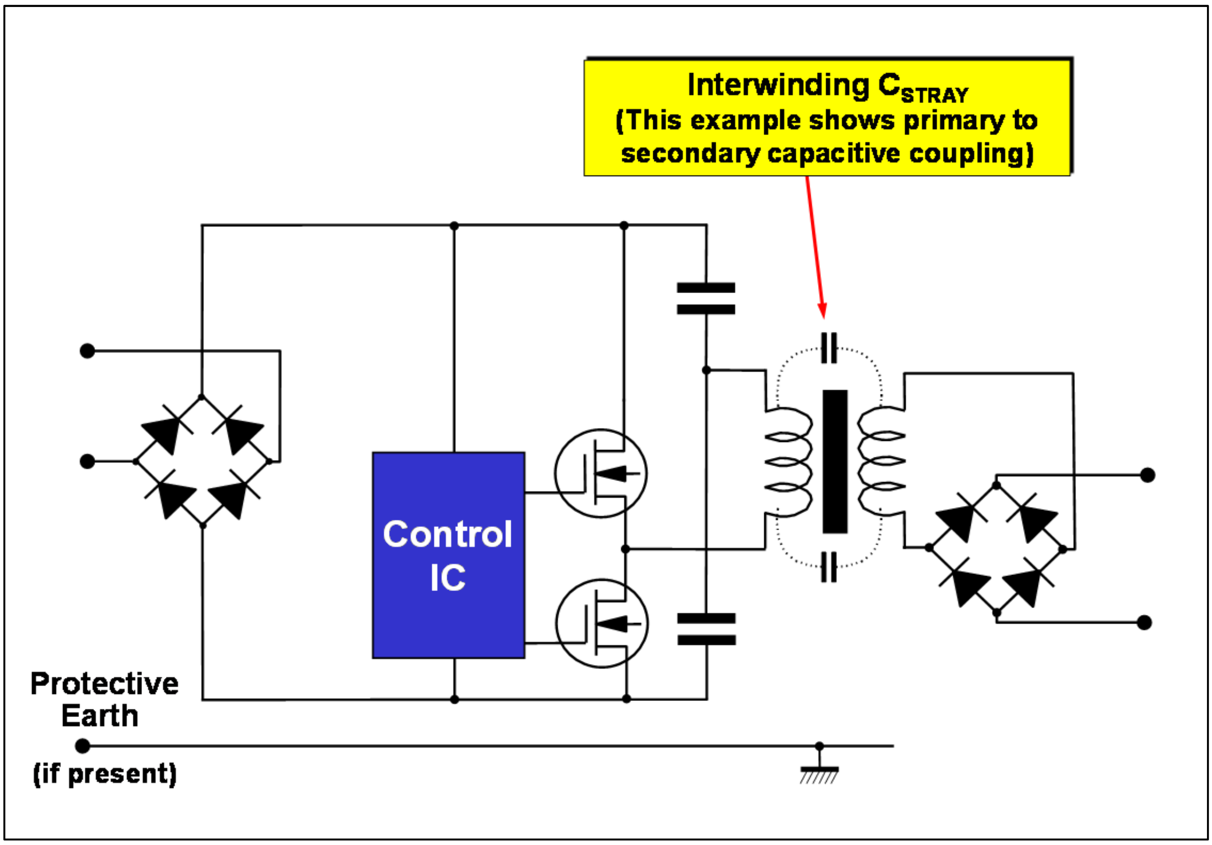

Figure 3E The stray interwinding capacitance (C_{STRAY}) in the HF isolating transformer

This C_{STRAY} injects primary switcher noises into the DC output circuit, as common-mode (CM) noises that flow in the DC output circuit and its loads. It also injects the DC output's HF rectifier's switching noises as CM noises into the primary switcher circuit, which then flow into the power supply lead and power distribution network in the building or vehicle.

As was already mentioned in an earlier part of this series (section 2.5 of [42]) ***all*** currents – including stray currents – always *always* always flow in closed loops. For more on this natural phenomenon see [4], or chapter 2 of [5] (which copies the text of [4]).

[45] and [46] describe the consequences of this law of physics/nature for any/all types of electronic design at any scale (including the fact that "earthing" or "grounding" or connecting to "chassis" *cannot* make unwanted noise vanish as if the "earth", "ground" or "chassis" was some sort of infinite sink for electrical current).

[4], [45] and [46] also show that any current loop naturally takes the path with least impedance, hence naturally creating the most compact pattern of electric (E) and magnetic (H) fields, automatically giving the best EMC (for emissions and immunity) that is possible from a given design and its physical construction.

When we understand this fundamental EMC principle, we can see that instead of trying to suppress stray CM currents with costly mitigation techniques such as filtering and shielding, we can save cost and time (and make our EMC lives much easier) by simply creating lower-impedance paths – with their even more compact E and H fields – available to all of the noise currents.

These lower-impedance paths must be very local to the noise source (always a semiconductor device), because the dominant constituent of the impedance of any current path is generally its inductance, which is directly related to the area of the current loop.

EMC textbooks and guides have for years been saying words to the effect of "keep all current loops small", because keeping the wanted, differential-mode (DM) current loops small reduces their generation of stray CM noise. We now see that keeping the stray CM current loops small is also important, by reducing the extent of their stray CM E and H fields and thus reducing emissions and improving immunity.

Effectively, this important approach manages the shapes and extents of both wanted and stray E and H fields, so that the noisy switcher circuits experience less coupling with their external electromagnetic environment.

Where a current (whether wanted or stray) has a choice of loops to flow in, it will naturally divide amongst them in the inverse ratio of their impedances – the loop with the lowest impedance will automatically carry more current than the others. So a key EMC design technique is to identify where stray capacitive coupling is occurring and – in each case – provide a local current loop with a small area and low impedance.

So let's now apply these basic principles to the interwinding CM noises in our example half-bridge PWM chopper!

Figure 3F attempts to show the paths of the two stray CM noise currents being considered here, from the noise generator (a switching device) through the interwinding C_{STRAY} and the circuit on the other side of the isolation transformer, eventually coupling back to form closed loops through stray conducted and radiated currents in the world outside of our example circuit.

What this means in practice is that primary switcher noise that flows through the transformer's C_{STRAY} and flows in the DC output, will loop back through the air and other conductors outside of the product and eventually return via the power supply lead and be measured as conducted emissions. This is why shielding and/or filtering the DC output circuits and their loads can often reduce the conducted noise that is measured on the power supply cable.

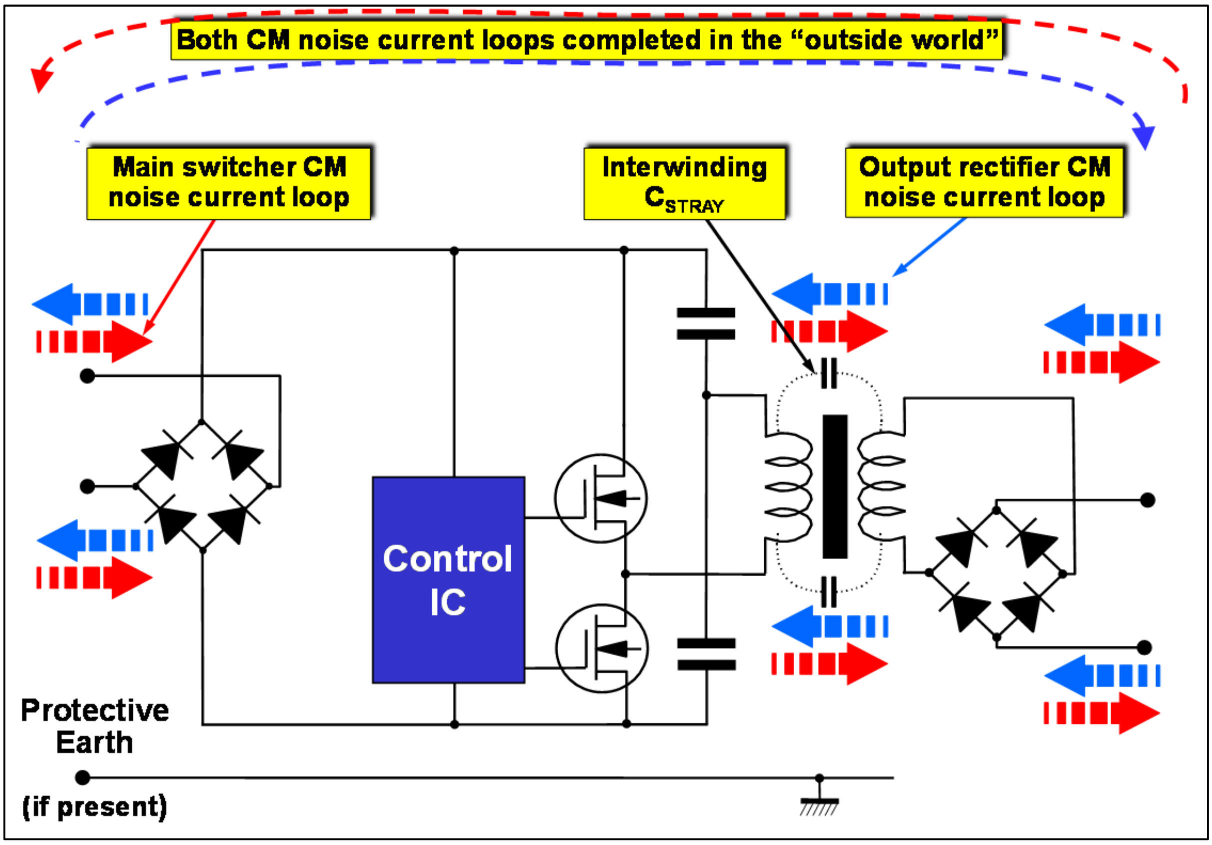

Figure 3F The CM current paths that flow through the transformer's C_{STRAY}

These stray conducted and radiated currents in the outside world are, of course, the conducted and radiated emissions measured by EMC tests, which for Regulatory Compliance we need to keep below the limit lines in the relevant test standards over the operational lifetime of the power converter.

However, we might need to ensure emissions are much lower than the standards' limits to prevent actual interference with other parts of the product that uses the power converter, or nearby equipment that is especially sensitive. My story below about powering an analogue professional audio mixing desk from a switch-mode power supply is relevant here.

That shielding and/or filtering DC circuits and their loads can reduce power supply-borne noise is not comprehensible unless one understands that all currents, even stray noise ones, have to flow in closed loops.

But how does the LISN (or AMN, AN, etc.) used to measure conducted noise in the power supply cable even detect a noise current that is coming from the power supply side? Surely it filters out all noise coming from the "wrong side"?

This confusion is all my fault, for describing currents as if the wanted signal (or unwanted noise) starts off at the source and travels all the way around the current loop before eventually returning to its source and completing the loop. This is a low-frequency circuit designer's way of looking at the issues, and I find that it helps me to visualise where the stray currents are actually flowing in a given physical construction. *But it is not what actually happens!*

All electrical currents (wanted, or noise) are real energy, so the Law of Conservation of Energy applies. What actually happens is that as the source emits its current, it also emits an antiphase return current. At each step forward in time the send and return currents progress a little further around their loop, maintaining a total energy of zero (Conservation of Energy means we can't create it, so our total energy must always be zero). Eventually, the positive and negative phase currents meet at the furthest end of the loop from the source, and cancel out.

"Proper" EMC engineers with good mathematical skills, and electromagnetic (EM) field solvers, analyse all electronic circuits as a huge number of very, very tiny dipoles – each one emitting EM waves in one direction whilst at the same time emitting identical *antiphase* EM waves in the opposite direction.

Tim William's excellent textbook [47] gives the basic equations for vanishingly small electric dipoles (also known as "current filaments" or "Hertzian dipoles"), and for magnetic dipoles (i.e. current loops) in its Appendix D, section D.3.8 on page 461. If you want to get into more depth, read chapter 7.1 of [48].

I tend to focus more on the practical implications for product design, which led me to write [45] and [46]. If your interest is more in visualising EM wave propagation (which, after all, is what all electricity really is, whether used as power, signals, data or noise), these two references should be more to your taste.

All circuit analysis is taught using "circuit theory" that is a gross simplification of real EM propagation (i.e. electricity), but this is never explained. Consequently, concepts based on circuit theory (including all circuit simulators) can lead us badly astray when we try to understand how best to design for EMC, and for that we have to blame our University teachers for not providing us with a complete understanding of electricity.

It's not surprising, really, that generations of circuit designers have resorted to getting a circuit functional and then – to comply with EMC requirements – stuffing it all into a costly and heavy shielded box that is fitted with costly and heavy shielded and/or filtered cable connections, then fiddling about with the shielding and filtering for as long as it took to get a design that passed all the EMC tests. We can do so much better than that, saving huge amounts of cost, time, weight, and mental anguish, when we understand what is really going on with EMC.

So, let's get back to the question of why it is that primary switcher C_{STRAY} noise currents that are emitted from our DC output and its loads and then (using an approach based on flawed circuit theory) "circle back" into the building or vehicle's power supply to complete their current loops, along the way being measured as conducted noise emissions on the power supply.

What is really happening is that our power converter is acting like a radiating dipole antenna, and as primary switcher C_{STRAY} noise current flows *out* of the converter's DC output, an exact antiphase replica of it simultaneously flows *out* of the power supply lead, into the test LISN (AMN, AN,etc.) and so gets measured as part of the converter's conducted emissions.

Whenever our grossly simplified analysis shows CM noise currents entering a LISN by its (filtered) power supply terminals to complete their loop in accordance with Maxwell's equation, what is really happening is that noise currents are entering the LISN from its product terminals.

This same "dipole effect" also means that sometimes conducted emissions measured on the power supply lead can be reduced by reducing radiated emissions (e.g. by shielding).

Figure 3G attempts to visualise the "near fields" around the converter's PCB assembly and its cables, as the stray CM currents complete their loops by flowing as conducted and radiated currents into and out of various components and cables.

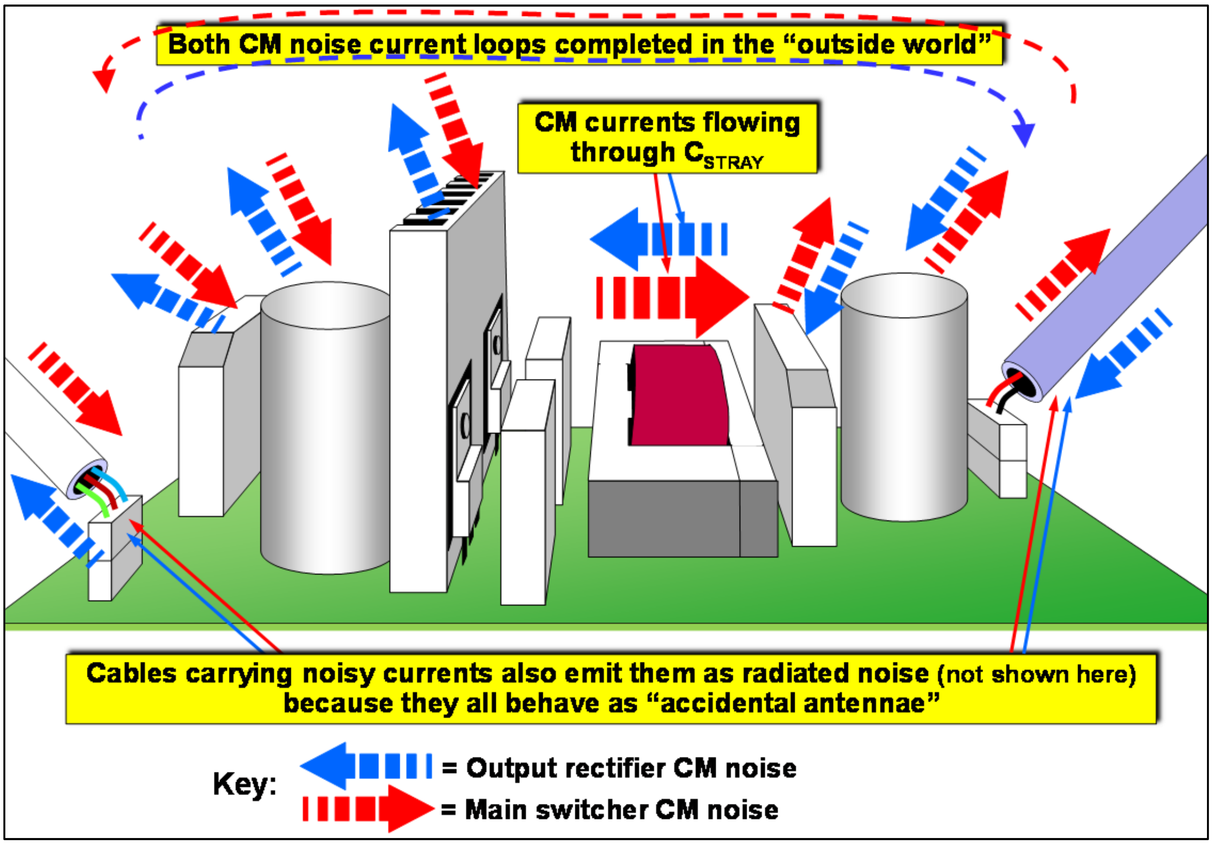

Figure 3G Visualising the CM current noise loops flowing around the converter's assembly

Because these CM noises are injected across an isolation barrier, their current loops are completed by external circuits (e.g. power source, load) and also by stray electric fields between the conductors and components on either side of the transformer, all behaving as accidental antennas, creating high levels of conducted and radiated emissions in the outside world.

The main culprits for creating emissions are the input and output cables, because conducted noises are measured directly on the power supply input cable, and because cable lengths make them very efficient "accidental antennas" for emitting radiated fields. (The unavoidable "accidental antenna" behaviour associated with any type of conductors is described in detail in [4] and Chapter 2 of [5].)

The radiated emissions from components acting as accidental antennas couples currents through E fields in the air (i.e. stray capacitances) into the power supply cable and so can increase conducted emissions. This is especially a problem for switching device heatsinks, and appropriate design solutions were discussed in Section 2.5 of [42].

CM noise via the HF transformer interwinding C_{STRAY} is difficult to filter at frequencies below 10MHz, because the stray interwinding capacitance presents a high source impedance at such frequencies.

As Chapter 5 of [5] shows, with a high noise source impedance we can't achieve much attenuation by using series CM chokes, as they work best when the noise source impedance is *much lower* than their series CM impedance.

Because the noise is CM, and is galvanically isolated by the transformer from its origin, capacitors across inputs or outputs have no effect (they only work on DM noise). Capacitors from inputs or outputs to the chassis or protective earth/ground *might* help, but adds a new accidental antenna and so might cause more problems than they solve (see 4.9 below). Remember – "earth", "ground", "chassis" or whatever *cannot* act as some sort of infinite sink for noise currents.

These CM noise currents flowing through the high impedance of C_{STRAY} create CM noise voltages between the primary and secondary circuits, which can upset voltage control feedback circuits and often make it necessary to use optoisolators (or other galvanic isolation signal techniques) with high dV/dt ratings, in the feedback signal's path.

It is best to reduce noise generation at source by keeping dV/dt and dI/dt low at every instant throughout the switching cycle, ideally by using a 'benign' type of resonant-mode power switching topology (see section 2.2. in [42], and the additional comments on Ćuk converters above) and also by using more "benign" HF output rectification techniques, which will be covered in a later "Stand Alone" article.

If further suppression of conducted or radiated emissions is required, it is best to apply the techniques described below as they might reduce (possibly even eliminate) the need to apply the more costly usual mitigation techniques: filtering and shielding.

4.3 Reducing emissions with an interwinding shield (or two, or more)

To reduce the CM noise currents flowing through an HF isolating transformer's interwinding C_{STRAY} from primary to secondary, or vice-versa, we can add one or more interwinding shields.

An electrostatic shield (usually a wide copper foil) that is wound next to a primary winding and connected to the associated primary circuit's RF Reference (one of its voltage rails), reduces that primary's CM noise current injection into the other windings, by providing the stray noise currents with a more local current loop with a much smaller area, that they will "prefer" to take. Figure 38 of [49] provides some detailed guidance on constructing a primary shield.

And a shield that is wound next to a secondary winding and connected to the associated secondary circuit's RF Reference (usually its 0V rail) reduces that secondary's HF rectifier noise CM current injection into the primary (and any other windings), also providing the stray noise currents with a more local current loop with a much smaller area.

A practical tip: split all such shields so they don't act as shorted-turns! We only want them to shield the electrostatic fields (i.e. C_{STRAY}) – not the magnetic fields that create the transformer action.

Figure 3H shows an example of using an interwinding shield to reduce the CM noise created by the primary switching circuit and flows through C_{STRAY}.

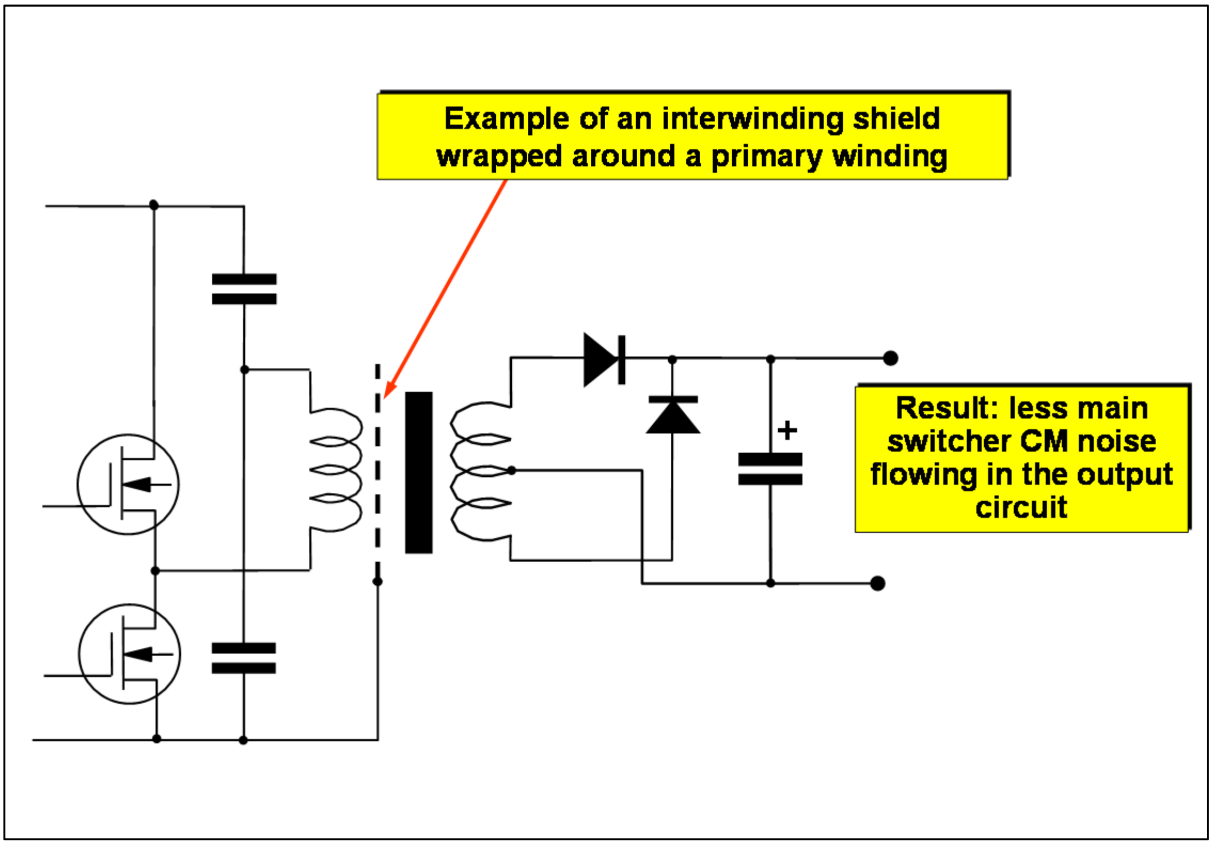

Figure 3H Example of shielding a primary winding

Interwinding shields tend to increase the primary-secondary leakage inductance, and so could increase the problems of overshoot and ringing of the switching waveforms, require more snubbing (see section 2.4 in [42]) It is possible that this extra snubbing might reduce the conversion efficiency a little.

Also, if a transformer has been designed without shields it might have no room to add them, and a larger core size might be required. Discovering this at a late stage in a project can cause all sorts of knock-on redesign of PCBs and mechanical housings to fit the larger transformer, so it is always a good idea to initially design prototype isolating transformers with two or more interwinding shields.

If the shields turn out not to be required after all, they can be left unconnected, or removed from production units. If removing the shields allows a smaller core size, this can be a cost-saving modification the next time the product is changed.

Another reason for possibly having to use a larger core size when adding interwinding shields, is that the shields can have a negative impact on the thermal properties of a transformer – the windings might be running too hot for their insulation ratings.

I am told that it is possible to use additional single-layer windings as interwinding shields, and that they shield better than foil. But I haven't experienced this type of interwinding shield yet, and wonder if the series inductance of such a "shield coil" would limit its effectiveness at high frequencies.

"Planar" transformers can easily add any number of shields at a late stage in a project without causing significant knock-on redesign issues, because each new shield simply requires an additional PCB layer.

I have found in my training courses that not everyone is familiar with planar transformers, so have included Figures 3J and 3K to show that their magnetic cores pass through holes cut in the PCB, and the PCB traces provide the windings.

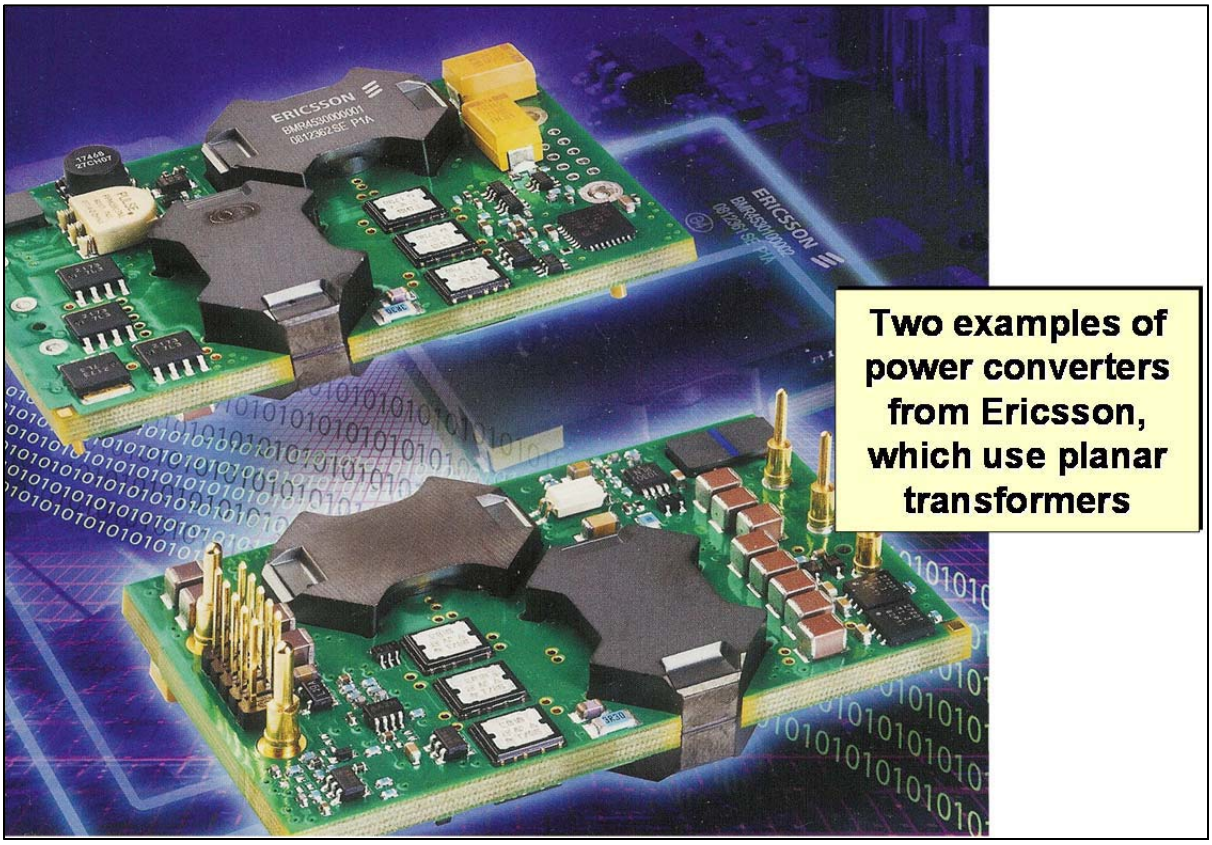

Figure 3J Example of the use of planar isolating transformers in power converters

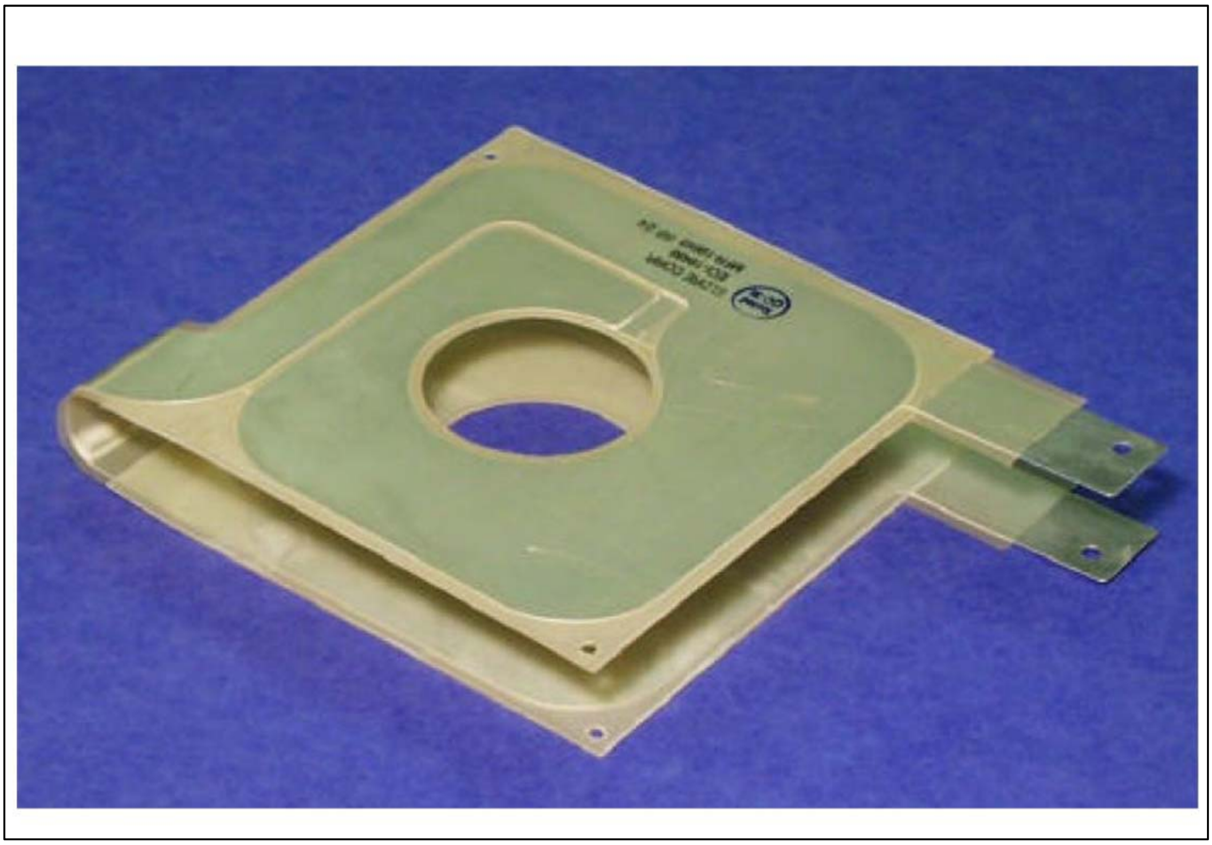

Figure 3K Example of a secondary winding for a planar isolating transformer for an electric vehicle's mains charger

For better suppression, add two or more interwinding shields, connecting the shields to the correct circuit nodes to return the various capacitively-injected stray currents to their sources via the smallest-area and therefore lowest-impedance paths.

I was the first person (as far as I am aware) to use a commercial switch-mode AC-DC power supply to drive an all-analogue professional audio mixing console, way back in 1981. The analogue audio circuits were the quietest we knew how to make at the time, but were so sensitive to noise on their split-rail DC power supplies that the huge custom-made and very costly mixing desks we were making for music, TV and film studios, all required very large, heavy racks of linear power supplies (e.g. ±18V at 100A), achieving 50% efficiency if we were very lucky.

A switch-mode power converter with the same output rating was much smaller, lighter, more efficient and less costly, but nobody had ever managed to find one that didn't completely destroy the very high signal-to-noise ratios of our mixers, regardless of whatever filters were fitted to their power supply inputs and/or DC outputs.

But in 1981 I borrowed a "low-noise" switching power supply (from Advance Ltd, if memory serves) that, with the addition of a little bit of filtering on the DC side, was just perfect! So we used that instead. When I asked how they had made it so quiet, I was told that its safety isolating transformer used five (5!) interwinding shields – each one connected to a different part of its circuit.

In those days we didn't have to meet EMC emissions standards, but if we had we would have found that the "audio-quiet" switch-mode power converter had much lower emissions than all of the others.

So we shouldn't be shy about adding interwinding shields in our prototypes. They cost less and take up less space than filtering and shielding, and we can always take them out later on if we find they are not needed to pass the emissions tests.

Transformer cores also have C_{STRAY} to their windings, and since the types of ferrites used in switch-mode magnetic components are conductive, transformer cores will suffer from stray capacitive noise currents, and since they are "floating" they have a high impedance relative to the RF Reference and so experience a dV/dt and emit E fields that add to the product's emissions.

This is a similar problem to heatsink emissions (see 2.5 of [42]), with similar solutions: either connect the transformer core to the appropriate power rail (taking care of any and all safety issues), or wrap a shield around the core and connect that to the appropriate power rail instead. (As before, split the foil shield so that it doesn't create a shorted-turn!)

4.4 Reducing emissions with primary-secondary capacitors

Connecting a capacitor between reference voltages for the primary and secondary circuits, provides a low-impedance local loop current return path for the C_{STRAY} noises that have coupled across the galvanic isolation barrier from primary to secondary, and vice-versa.

Important Safety Note: Capacitors connected between primary and secondary are almost always safety-critical! See section 4.6 below.

The stray CM currents "prefer" to flow through this capacitor with its low-impedance current loop, rather than flow in the much larger loops available from the various components and cables and their E fields that I tried to visualise in Figures 3F and 3G.

Of course, nothing is ever perfect, and so because we can never truly create 0Ω current paths for the stray currents there is always *some* stray CM noise current flowing in the larger loops shown in Figures 3F and 3G, that get measured as conducted and/or radiated emissions.

For example, when using a primary interwinding shield, there will still be *some* C_{STRAY} noise that couples from primary to secondary, so providing a primary-secondary capacitor gives it a second lower-impedance loop and reduces the stray current flowing in the (high-impedance) external loops even more.

Because the C_{STRAY} noises flowing through the transformer generally have a high source impedance (due to the small values of interwinding capacitance), the low impedance of the primary-secondary capacitor makes it very effective as a filter (Chapter 5 of [5]).

Also, the low impedance of the primary-secondary capacitor has the beneficial effect of reducing the source impedances of the remaining C_{STRAY} noises that are still stubbornly flowing in the larger loops, making it possible to obtain much better suppression on the power supply input and/or DC output conductors by using series-connected CM choke filters.

Very large isolating transformers (e.g. MW ratings) have large values of C_{STRAY}, but MW-rated converters switch at lower frequencies with lower dV/dt, so their C_{STRAY} noise source impedance is still high. Very high-frequency converters (such as Analog Devices' "isoPower" devices, see 4.5) use very small isolating transformers that have very small values of C_{STRAY}, so despite their very high frequency operation the source impedance of the C_{STRAY} noise is still high.

Figure 3L shows my example half-bridge chopper fitted with two interwinding shields (one to reduce emissions of primary switcher C_{STRAY} noise, the other to reduce emissions of HF rectifier C_{STRAY} noise) and a primary-secondary capacitor.

It has some curved arrows that are my attempt to visualise how these techniques ensure that almost all C_{STRAY} noises flow locally, so that they don't contribute significantly to the measured emissions. The broad red and blue arrows try to show the paths followed by the majority of the C_{STRAY} noise currents through the interwinding shields, while the dashed red and blue arrows try to show that most of the noise currents that managed to escape capture by the shields are rounded up and herded back to their sources via the primary-secondary capacitor.

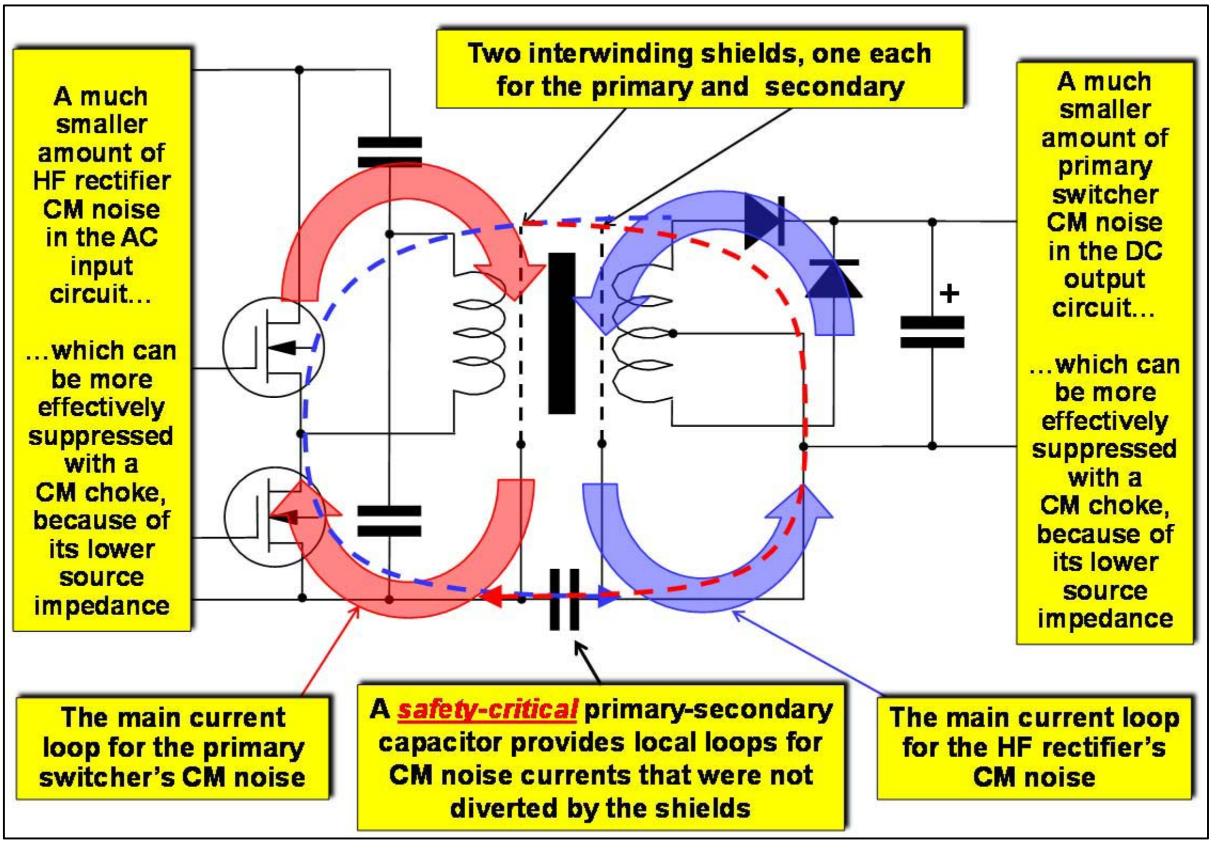

Figure 3L Example of adding a capacitor from primary to secondary circuits (also shows two interwinding shields)

If not using a benign, resonant-mode switching topology, and using an isolating transformer, it makes good sense to allow for at least two interwinding shields plus a primary-to-secondary capacitor as shown in Figure 3L.

When testing emissions, we might find that we only need one of the shields and the capacitor, or just the capacitor, or just one or more shields – but because we made provision for them we are not going to delay the project by wrestling for days or weeks with costly and expensive filtering and shielding.

It is sometimes possible to use the good EMC design techniques described in this "Stand Alone" series to completely eliminate the need for a power supply filter. But it is more likely that some power supply filtering will be needed, in which case using these good EMC practices will reduce its complexity, size, weight and cost – and also make its design much easier.

These good EMC design practices can also make shielding totally unnecessary as long as rise and fall times of the switching waveforms are no less than about 10ns.

4.5 Using "buried" PCB capacitance for frequencies above 100MHz

Generally, mains suppression capacitors (e.g. Class X or Y types) perform well up to 10MHz or more, and higher frequencies may be possible by using types with polypropylene or ceramic dielectrics. But their self-inductance plus the series inductance of the wire leads or PCB traces used to connect them into the circuit (see 4.7 below) begins to dominate their impedance above a few MHz, rendering them pretty much useless at frequencies above about 100MHz.

However, we can easily create "buried capacitors" inside a PCB's layer structure, by overlapping areas of copper plane on adjacent layers. These have such low series inductances that they are effective at providing low-impedance current loops for frequencies well-above 100MHz.

To keep the noise loop area as small as possible (hence as low-impedance as possible), the overlapping plane areas should be placed underneath – and symmetrical with – the semiconductor whose CM noise emissions are to be provided with low-impedance and local current loops.

To date, we have not often had to deal with CM noise emissions above 100MHz, because of the switching speeds (and resulting harmonics, see Figure 2A of [42]) of typical power converters. However, some low-power isolating DC/DC converters operate at very high frequencies indeed, with high levels of interwinding C_{STRAY} emissions at those frequencies, and I have recently met University researchers attempting to design 200W AC-DC converters that switch at frequencies of 30MHz or more.

The Analog Devices "isoPower" ADuM5xxx product family is an example of a tiny isolating DC/DC converter that can transfer nearly a Watt of power across a 4kV isolation barrier by running its switchers at between 180 and 300MHz.

It apparently uses a resonant-mode topology that produces sinewave voltages and currents with minimal levels of harmonics. No discrete capacitors have sufficiently low series inductance (see 4.7) to be able to successfully provide a low-enough-impedance local current loop for the C_{STRAY} emissions at their switching frequencies, and [50] recommends creating an buried capacitor within the PCB.

The common PCB material FR4 has a nominal relative dielectric constant (ε_r) of about 4.3, between 1MHz and 1GHz, which means that between overlapping areas of copper planes we create a $C_{overlap}$ of about 35/d nF per square metre, where d = the overlapping planes' dielectric spacing in millimetres.

For a noise current loop that provides any significant benefits we generally need a total loop impedance of less than 1Ω (much less would be much better). For 1Ω at 180MHz we need just under 1nF, and we can create 1nF with a 10 sq cm plane overlap with a 0.3mm (7.5 thousandths of an inch) FR4 layer between the two copper plane areas.

Chapter 5.3.15 of [37] discusses techniques for achieving much greater buried capacitance

Figure 3M sketches the general concept of the idea, using the example of an isolating DC/DC converter that bridges between a PCB's 0V plane and a galvanically isolated area of 0V plane. It is a common experience to suffer excessive conducted and radiated emissions from such PCB constructions, which is why Figure 3M shows a row of discrete capacitors spread all around the circumference of the 0V plane split.

Spacing these capacitors apart by less than one-tenth of a wavelength (λ) at the highest frequency to be controlled (for emissions or immunity) helps prevent resonances that could create huge problems for EMC. (For more on the use of the "<λ/10 spacing" approach to prevent unwanted resonances in PCBs and other electronic product design aspects, see [5] and [37]).

Figure 3M shows a row of via holes being used all along one edge of the buried plane area, to connect it to the galvanically-isolated 0V plane. Via holes suffer inductance at the rate of about 1nH per mm of length, so the 0.3mm length of via hole we are using has an inductance of about 0.3nH, which has an impedance of about 0.34Ω at 180MHz. This series impedance would add significantly to the <1Ω loop impedance we are trying to create for the C_{STRAY} CM noise currents, so Figure 3M shows the use of four vias in parallel, spread uniformly along one edge of the buried plane area, to reduce this stray series inductance to less than 0.1Ω at 180MHz.

For frequencies above 1GHz, the inductance of four 0.3mm long via holes connected in parallel as shown in Figure 3M would be 0.85Ω. Clearly, if we ever want to provide local loops for C_{STRAY} currents at such high frequencies – as we no doubt will have to one day soon – we will have to be cleverer.

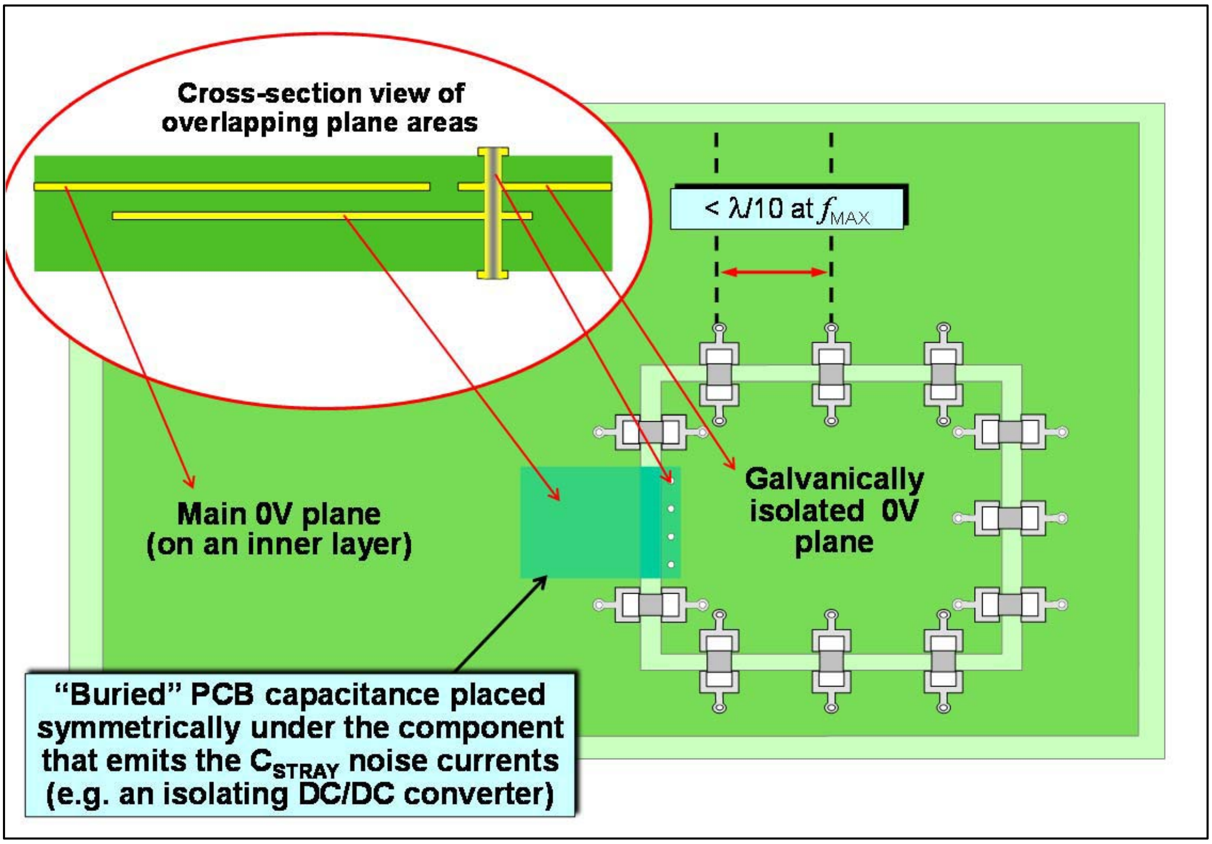

Figure 3M Creating buried PCB capacitance to control C_{STRAY} noise at 100MHz or more

4.6 Safety issues for capacitors that cross galvanic isolation barriers

Capacitors connected between the mains power input (primary) circuit and any —

- Protective conductor; safety ground; protective earth; chassis; enclosure, etc.
- Safety-isolated output circuit(s)

— must all be chosen with safety foremost. They will be "Y rated" according to the relevant standard (IEC 60384-14) for the nominal mains voltage, and their values will be chosen on the basis of the maximum permissible ground leakage current specified by the relevant safety standard listed under the appropriate EU Safety Directive.

(Some commonly used safety standards are: IEC/EN 60950; IEC/EN 61010, and IEC/EN 60335. Medical devices use IEC/EN 60601-1, which for patient-connected equipment can limit leakage currents to 10µA – making it virtually impossible to use any primary-secondary capacitors, or connect Y-rated mains filter capacitors to the protective earth or safety ground.)

I always recommend that the only capacitors that are used in any safety-critical applications (such as the primary-secondary capacitor discussed in section 4.4 above) are types that have been approved by a 3rd-party Safety Approval Body such as SEMKO, DEMKO, NEMKO, UL, VDE, BSI, TUV, etc.

But it is not sufficient to take the supplier's word for this, because the manufacturer who incorporates a component into their product assumes full responsibility under the law for any resulting safety defects. "Buying in good faith" is not a legal defence. So all safety-critical capacitors' safety approval certificates should be obtained from their vendors, and carefully checked to make sure the capacitors are rated correctly for what is required for the safety compliance of the final product.

It is commonplace to see a VDE (or other Safety Approval Body's) logo printed on a capacitor along with its ratings – let's assume for the sake of argument "230VAC, Class Y1".

But it should *never* be assumed that the presence of their logo means that Safety Approval Body has actually *approved* that capacitor as "230VAC, Class Y1".

The capacitor manufacturer might rate his part at 230VAC Class Y1 but the Safety Approval Body whose logo is proudly printed on the component might only have approved it as 120VAC Class Y2, or worse.

I have even heard of Class Y safety capacitors marked with the VDE logo, for which the logo had only been awarded for the quality of the tin-plating on its leads!

Because some capacitor manufacturers have been known to forge their 3rd-party Safety Approval documents, I also always recommend that the Safety Approval certificates obtained from the capacitor suppliers are checked – with the 3rd-party Safety Approval Body named on the certificate – as being valid.

Whenever I have done this, I have always found the Safety Approval Body's personnel to be most helpful and kind – which I suppose is not surprising because it is helping them police their market against fraud and protecting their good reputation.

Similar precautions are recommended for any/all safety-critical parts, not least the safety isolating transformer itself.

I also recommend carefully reading through all Safety Approval certificates for any "conditions of use". I have seen a US-made UL-Approved mains power supply module that included in its Approval certificate a brief statement that under fault conditions it could emit a jet of flame 12 inches long, and so must be enclosed in a fireproof enclosure when used in a final product.

Nobody had actually read the approval document. It had simply been obtained from the supplier and filed away on the assumption that because the unit was UL approved, it must therefore be perfectly safe. As a result, the power supply module was being used in the manufacturer's product with no flame protection at all, creating a serious fire hazard he was not aware of when he declared the product to be compliant with the Low Voltage Directive and affixed the CE marking to it.

Where a buried PCB capacitor (see 4.5) is used across a safety-related galvanic isolation barrier, the relevant safety standard will provide the "creepage and clearance" rules to be followed to ensure the capacitor is safe enough. For the dielectric between the plates, most (maybe all) safety standards simply require that it passes the withstand voltage tests that it specifies for "reinforced insulation". These generally apply 3kV rms at 50 or 60Hz, or 4.25kV DC, for one minute, and check that no significant current flows.

When FR4 is dry it will withstand about 40kV per mm [51], so a 0.3mm thickness of it should easily cope with the reinforced insulation test voltages. However, FR4 tends to absorb water, depending on the humidity and temperature of its environment, which can lead to it becoming conductive, and in the presence of high voltages (say, over 50V) cause "tracking" to occur – the dendritic growth of conductive metal salts – *inside* the PCB.

Safety standards are well-used to dealing with tracking on the *surfaces* of PCBs, in mains-powered areas, because with inadequate PCB layout it can quite quickly lead to electric shock and fire hazards. However, the safety standards all ignore the possibility of *internal* tracking, so I recommend that PCBs with buried safety-critical capacitance should be subjected to simulated lifetime testing with maximum humidity and temperatures that result in the PCB absorbing the most moisture, whilst the reinforced insulation voltage withstand test is continually applied to the buried capacitor to check that it doesn't fail.

Air has a much lower breakdown voltage than FR4, and with a spacing of 0.3mm (say) it will spark-over at about 340V. So it is important to ensure that the PCB dielectrics used in boards with buried safety-critical capacitors are "void free" – something for the Purchasing and Quality Control Departments to deal with.

But there is also a design-related concern – overheating a PCB can make it delaminate and/or carbonise (char), which can open up voids inside the PCB and/or convert the material from a slightly damp insulator to quite a good conductor. So it is important to ensure that any safety-critical buried capacitors are not exposed to heat sources that could cause any damage to their dielectrics over the lifetime of the product, even in the hottest anticipated environment.

4.7 Connecting to interwinding shields, cores, and primary-secondary capacitors

Connections to interwinding shields and primary-secondary capacitors all suffer from series inductance, which limits their effectiveness at providing low-impedance current loops, especially at higher frequencies.

So it is important to keep all shield and capacitor connections as short as practical, whether they use PCB traces, leads or wires, or busbars.

It might help to reduce the inductance of a shield connection by using two connections in parallel, to opposite sides of an interwinding shield (although to avoid creating a shorted turn they should both be on the same side of the transformer core).

Every electrical/electronic structure has resonances, due to their stray capacitances interacting with their stray inductances (this is a low-frequency view, see Chapter 3 of [4] or Chapter 2.5 of [5] for more detail). "Series resonant modes" provide current loop impedances that are as low as the series resistance (often just a few milliΩ), regardless of loop area. "Parallel resonant modes" provide current loop impedances that are very high indeed, possibly several tens of kΩ.

So it can be possible to "tune" a circuit's series resonances to make noise currents more likely to follow desirable loops, and/or to tune parallel resonances to make noise currents less likely to follow undesirable loops.

For example, if the CM C_{STRAY} noise currents caused by interwinding capacitance cause just one emissions frequency to be above the limit line, adding a primary-secondary capacitor that is series-resonated with the total inductance of its current loop (including the self-inductance of the capacitor) at that exact emissions frequency will have the maximum effect at that frequency – but will probably be less effective at other frequencies.

I don't generally recommend "resonant tuning" EMC techniques, because a small design change later on can make them completely ineffective. If the change is not being done by the original "tuner" (and he/she has remembered that they "tuned" the design!), people can be unpleasantly surprised by how a small change could cause the emissions to increase by such a large amount.

Many manufacturers do not bother to redo their EMC tests when making a "small" change (e.g. a component substitution), so can be caught out badly when an entire batch of products is returned under warranty because they cause interference or don't function reliably enough. As I have often pointed out in articles and books, the real reason we do EMC engineering and comply with EMC standards is to control financial risk. Complying with the EMC Directive is a very secondary issue, by comparison, for all that it is legally required in the EU.

4.8 Everything resonates

When the stray capacitances and stray inductances associated with a local noise current loop that we have designed cause it to resonate in "parallel" ("shunt") mode, the local loop can have an impedance of several tens of kΩ, possibly even more – making it ineffective at reducing emissions at the parallel resonant frequencies.

We normally consider a wire or cable to have an impedance that increases the longer it is or the more loop area it encloses, but when it resonates – either due to stray capacitance or because of transmission-line effects – it can have a very low impedance, maybe less than an Ohm.

So a building's or vehicle's power distribution and protective earthing (safety grounding) networks can have very low-impedances at certain resonant frequencies, effectively "sucking" our converter's noise currents away from the local loops we have created for the CM noise and increasing measured emissions at those frequencies.

This is a big problem for the design of power supply filters, and sometimes it is necessary to fit an "earth-line choke" to help prevent noises being "sucked" into the building or vehicle's earth or ground structure.

Consequently, it is very important to be in control of all "accidental" resonances, if we are to cost-effectively achieve good EMC characteristics in a timely and low-risk manner, as our managers would like us to do.

Understanding resonances in sufficient detail to control them effectively during design is beyond the scope of this article, but is covered in [4], [5] and [37].

4.9 Connecting interwinding shields, cores, and primary-secondary capacitors to earth/ground/chassis/frame/etc.

The astute reader will have noticed that I have not suggested connecting anything to "earth", "ground", "chassis", "frame", or any of the other words people use to describe a protective earth or safety ground connection, or some mythical infinite sink for electrical noises.

Protective earths or safety grounds in buildings or vehicles can only provide additional "accidental antenna" effects for noise currents that were created by switching devices. These noise currents must (by the laws of physics) flow in closed loops, but the conductive structure of a building or vehicle's earth or ground is not a suitable low-impedance loop for it to flow in, because it is outside the converter.

I recently heard a very wise and experienced EMC engineer describe a site's protective earthing (safety grounding) wiring structure as an "interference distribution network" – a very true and telling observation that I wish I had made.

As for any of the other words people might use to mean an infinite sink for electrical noise – no such thing can ever exist.

Connecting a part that is carrying noisy CM currents (such as a switching device's heatsink, interwinding shield, ferrite transformer core, etc.) to the protective earth (safety ground) connection simply allows its noise currents to flow in all sorts of additional paths, most/all of which will increase emissions.

However, where costly high-performance filtering and shielding is used, it can be designed to be very effective at returning the internally-generated noise currents back to their sources – completing their current loops entirely within the product's shielded enclosure, and therefore passing emissions tests.

Very many switch-mode power converters (or products using them) have been made in the past by using such "brute force" EMC mitigation techniques. Typically, they add around 25% to the BOM cost, and to the product's volume and weight, and they can add 50% to 100% to the design timescales.

When using such a brute-force EMC-mitigated filtered and shielded construction, the internal earth/ground structure can be conveniently used to provide all the noise currents with closed loops inside the product enclosure.

Unfortunately, this has helped to perpetuate the myth that "earthing" or "grounding" ***in itself*** provides magical benefits for EMC. Which it doesn't.

Military and government projects use taxpayer's money so still tend to use brute-force EMC mitigation techniques, where the resulting long timescales and large size and weight are not problems, However, the rest of us have to be much more cost-and-time effective – which is why in this article I have not suggested closing any noise current loops that use any conductive structures that are connected to the safety ground or protective earth (or to the chassis, frame, or whatever).

Without high-performance, large, costly, weighty filtering (on all unshielded cables) and shielding of the enclosure and unfiltered cables, using earth or ground conductors inside the products as part of noise current loops, encourages the noise currents to escape and flow widely outside of the product – increasing its emissions.

I am not saying never to use an "earth" or "ground", "chassis", "frame" or whatever as part of a low-impedance local noise current loop, as this is not always practicable.

What I am saying, is that every time we use an "earth", "ground", "chassis", "frame", etc., as part of a stray noise current's loop, we tend to increase the specification required for the product's filtering and shielding – and hence increase its cost, size and weight and the difficulty of designing them to control emissions.

4.10 Construction of the isolating transformer

[52] and [49] provide EMC design advice on constructing isolating HF transformers for flyback converters. Some of their recommendations are specific to flyback topologies, whilst others are relevant to isolating HF transformers in general.

They tell us that the primary leakage inductance should be high to decrease the dI/dt of the primary switching current, thereby decreasing the harmonic content of the switching noise. The downside – increased flyback voltages at switch-off – must be dissipated as heat in snubbers or overvoltage protection circuits.

A lower intrawinding (i.e. turn-to-turn) capacitance in the primary winding results in smaller current spike at the turn-on of the switching transistor and a smaller stored energy in the resonant circuit. This can be achieved by increasing the distance between its turns.

To efficiently demagnetize the air gap and transformer core on each cycle of operation, the secondary leakage inductance must be low and the mutual inductance with the core must be high.

The interwinding capacitance C_{STRAY} causes stray high-frequency CM currents (see 4.2), which can be reduced by reducing the value of C_{STRAY}, for example by spacing the primary and secondary windings further apart. However, this decreases their mutual inductance and increases their leakage inductance so might increase H field emissions and/or may affect the resonance/ringing frequencies.

However, the CM C_{STRAY} currents can be reduced without reducing the value of C_{STRAY}, by arranging any multilayer windings so that the layers that are "closer" to connections to quieter circuit nodes (i.e. have lower dV/dt) are the closest winding layers to the other windings and/or to the core.

For example, in Figure 3L, the capacitor side of the half-bridge primary switcher has a very much lower dV/dt than the switched side, so – if the primary winding has two or more layers – the layer that is connected to the capacitors should be the one that is closest to the secondary winding in the transformer's "layer stack" (if the aim is to reduce the stray CM current flowing in the secondary).

Alternatively, if the aim is to reduce the stray CM current flowing in the core, that primary layer should be the one closest to the core.

To avoid suffering from increased emissions at the transformer's resonant frequency, it is possible to design so that its fundamental resonance frequency lies in-between any two of the switcher's emissions frequencies (the switching frequency and its harmonics).

This transformer resonance design trick can be a neat and clever way of reducing certain emissions by 20dB or more, but like all "tuning" techniques it is vulnerable to causing unexpectedly large increases in certain emissions frequencies (by 20dB or more) in the future, often resulting from what was thought to be a small design change that was not significant for EMC.

It is especially important to use a magnetic core that has its only air gap in a central limb (so, no "C-core" types). With this construction, the magnetic leakage flux from the air gap is shielded to some extent by the outer limbs of the core, with circular types (e.g. pot cores) providing better shielding than rectangular cores (e.g. H-core types).

A shorted-turn (i.e. complete loop) of wide copper tape around the *outside* of the complete transformer also helps reduce the stray H field emissions from its air gap(s). It might help to electrically connect this "stray flux band" to both halves of the transformer's ferrite core. The orientation of this stray flux band with respect to the air gap may be significant, too.

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