Power Supply Design Seminar

Understanding and Optimizing Electromagnetic Compatibility in Switchmode Power Supplies

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Understanding and Optimizing Electromagnetic Compatibility in Switchmode Power Supplies

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ABSTRACT

Electromagnetic noise issues have long been a concern with switching power supplies, as their high-frequency switching of voltage and current generate higher order harmonics that have the potential to cause interference with system operation. While international specifications and standardized testing procedures have at least provided a greater understanding of the problems, optimum solutions must still come from techniques that are applied during the design process. This topic attempts to describe the more significant causes of EMI in power supplies and offer design techniques to minimize their impact.

I. INTRODUCTION

While switchmode power supplies are well known and appreciated for their significant benefits in reductions to both physical size and internal power loss, the threat of generating electromagnetic interference (EMI) from the high frequencies inherent in their fast-changing voltage and current waveforms has always been a serious concern. As advances in power supply technology have increased switching frequencies, and allowed these supplies to be used in closer proximity to the systems they power, these concerns – and the demands that designers react to them – have only heightened. Thus, designing for electromagnetic compatibility (EMC) has become a requirement every bit as important as meeting a given set of power conversion performance specifications, and it is the purpose of this discussion to address some of the more important principles and techniques involved.

First, note that while we have used the terms EMI and EMC interchangeably, it should be clear that one is the inverse of the other. The accepted definition of EMI is: *the interference of one piece of electronic equipment on the operation of another by means of electromagnetic energy transfer*.

And electromagnetic compatibility (EMC) is the absence of EMI. Two things follow from this definition. The first is that EMC is a *system-level* consideration. While it is common to test for electromagnetic noise generation from a power supply as a stand-alone box, the ultimate standards that have to be met apply to the system as a whole with the power supply as an internal component or subsystem. And the second aspect of this definition of EMI is that it involves three elements:

- A generator of electromagnetic energy: (a *source*).
- Transmission of that energy between equipments: (a *coupling means*).
- A receptor circuit whose operation is negatively impacted by the transmitted energy: (a *victim* circuit).

All three elements must be present for EMI to take place – remove any one and there can be no interference. However, while shielding or separation may solve a specific interference problem by attacking the coupling or susceptibility part of the system, the preferable approach is to remove the problem at the source – and that will be the thrust of this discussion.

II. UNDERSTANDING EMI

While it may at times seem otherwise, EMI is not a *black art*. Although often neglected in basic circuit design education, and appearing sometimes to be quite complex in practice, the basic principles for both causing and correcting EMI are relatively simple. Fundamentally, it requires a recognition of the fields caused by rapidly changing currents and voltages. While these characteristics are quantitatively described by Maxwell's equations, we need only to know that electronic noise may be induced by coupling between circuit elements through the action of either a magnetic or an electric field.

A magnetic field will cause a changing current in a conductor to induce a voltage in another according to:

$$e = M \bullet di/dt$$

where M is the mutual inductance between the source and the victim.

Similarly, an electric field will cause a changing voltage on a surface to induce a current to flow in another conductor according to:

$$i = C \bullet dv/dt$$

where C is the capacitance coupling the source to the victim.

These equations tell us that where we have rapidly changing currents – as in the conductors in series with power switching devices – we can expect to see an induced voltage across other conductors coupled by a mutual inductance. And where there is a high dv/dt – as on the drain contacts of the power switching FETs – any parasitic capacitance can couple an induced current into another path. While it may not be very helpful, it should probably be noted that one of the most obvious ways to reduce generated EMI would be to slow down the switching transitions, but the cost in increased switching losses typically makes this a less viable solution.

One important point to keep in mind on this subject is that EMI is a very low energy phenomenon! Because it can take an extremely small amount of energy, induced at the right place in the victim, to upset system performance, noise specification limits have been established at very small values. For example, at 1 MHz, it takes only 20 nW of measured EMI power to fail FCC requirements. This could be experienced with energy coupling of just a few parts per billion. Thus an easy trap to fall into is to discount some possible noise sources as too insignificant to be worth considering.

III. MEASURING EMI

When we describe a potential noise-generating system, the transmission of the noise out of the system is by either or both of two separate coupling paths: as radiated energy from the system itself by way of magnetic or electric fields (as described above), or as conducted energy flowing in either the input, output, or control lines connecting the system under evaluation to the "rest-of-the-world", where these lines can then become secondary radiators. These *conducting* and *radiating* EMI noise sources are considered as separate and distinct and are typically specified separately when evaluating the external characteristics of a definable system. One helpful characteristic, however, is that the efficiency of noise coupling is frequency dependent - the higher the frequency, the greater the potential for radiated EMI, while at lower frequencies, problems are more likely to be caused by conducted noise. There seems to be universal agreement that 30 MHz is the crossover frequency between conducted and radiated EMI. Most regulating agencies throughout the world have thus established measurement standards specifying the evaluation of conducted EMI at frequencies up to 30 MHz, with a separate set of standards applicable above 30 MHz for radiated EMI. We will discuss these two types of noise sources separately, but before doing so, some additional comparative information might be helpful.

Conducted noise is primarily driven by current but is measured as a noise voltage by using a 50-Ω current shunt. Moreover, with our discussion herein limited to power supplies, the FCC position is that only the ac input power lines are of concern, since it is here where noise currents could most readily couple to other systems through the power distribution network. While the maximum frequency of interest for conducted noise is 30 MHz, the minimum frequency limit can vary. In the United States and Canada, that limit is usually 450 kHz but many international specifications place the lower limit at 150 kHz. And some telecom specs require testing all the way down to 10 kHz.

With radiated noise specifications starting with a lower limit of 30 MHz, the upper limits can range from a few hundred MHz to 1 GHz or more. And while conducted noise can be evaluated with not much more laboratory equipment than a spectrum analyzer and a coupling device, radiated noise requires the measurement of magnetic or electric fields in free space, causing the testing to become much more complex. For this reason, radiation testing is usually performed by a separate, specialized test facility, where variables inherent in the testing environment may be very closely defined and controlled.

And speaking of variables, a non-trivial decision that needs to be made right at the beginning, is which specification to use. We have already alluded to the fact that each country has its own set of specifications for EMI, but within this distinction, there are additional determinates defined by product type and usage. For example, some specs define power supplies as "digital devices" while others combine them with "information technology" specifications. System usage also defines "Classes" where Class A designates industrial and commercial applications, and Class B includes residential usage. Class B limits are typically more stringent under the assumptions that systems made for the home are likely to be in closer proximity, and that residential users typically have fewer options available for dealing with EMI.

In the United States, the Federal Communications Commission (FCC) owns the responsibility for the control of electromagnetic interference through the Code of the Federal Register (CFR), Title 47. Within this document, Part 15 controls Information Technology Equipment (ITE), Part 18 covers Industrial, Scientific and Medical Equipment (ISM), and Part 68 regulates equipment connected to a telephone network.

International EMI specifications have been led by the International Electrotechnical Commission (IEC), which has published a European Union generic standard for emissions (EN50081), and the French-led Comité International Spécial des Perturbations Radioélectriques (CISPR). This latter body has issued their specification - CISPR Publication 22 - which is gaining world-wide acceptance and, in so doing, is applying pressure for FCC acceptance of the same standards.

At the time of this writing, noteworthy differences between the FCC and CISPR specifications include, in addition to some limit value differences, a lower frequency range for CISPR conducted noise measurements, and radiation tests made at a fixed distance of 10 meters vs 3 to 30 meters for FCC testing. By extrapolating the FCC radiation limits to 10 meters (linearly proportional to 1/distance) the comparisons can be better illustrated with the frequency plots shown in Figs. 1 and 2. [Ref. 2]
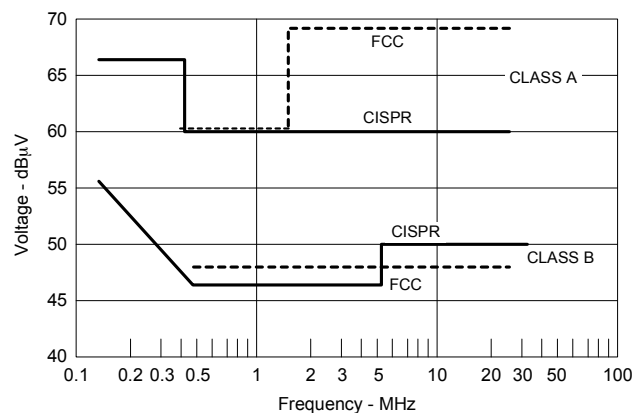

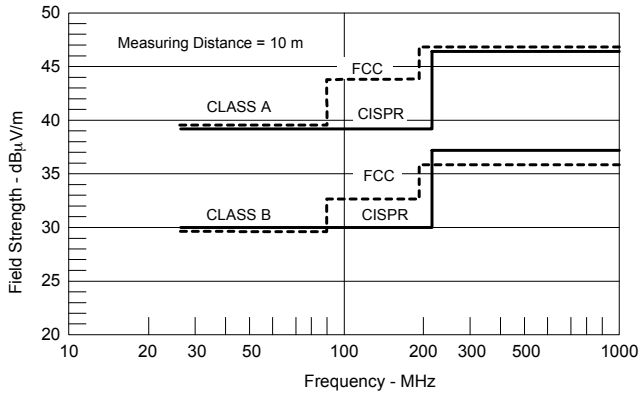Fig. 1. Conducted noise limits,
(FCC Part 15 vs. CISPR 22).

Fig. 2. Radiated noise limits,
(FCC Part 15 vs. CISPR 22).

Note that the units in these specifications are given as dBμV for direct measurements of conducted noise and dBμV/m for field strength measurements of radiated noise as sensed with an antenna. These are decibel values giving the ratio between the actual voltage measurement and one microvolt. The calculation is:

$$dB\mu V = 20\log_{10}\left[\frac{V}{1.0\mu V}\right]$$

and it applies to both volts and volts/meter.

Finally, with all the variables we have been discussing, we have not mentioned one that you might consider quite important - the operating conditions of the equipment under test. The reason is that this raises so many other questions that it defies definition. For example, do you test at the highest input voltage where dV/dt will likely be at a peak, or at the lowest voltage, maximum load where dI/dt will probably be highest? And in the power circuitry, there is one current path when the power switch is on and another when it is off, so duty cycle is likely to be a variable. As a result, in most cases the operating conditions are left to whatever the manufacturer defines as "typical" but suppliers should recognize this as a potential issue in correlating data.

IV. CONDUCTED EMI

We now will delve more specifically into the causes and design implications of each of the two major categories of EMI, focusing initially on conducted noise as measured on the power supply's input power feed lines. To observe

conducted noise on a power line, a device is needed to separate the high-frequency noise signals from the input current, and that device is called a Line Impedance Stabilization Network, or LISN, and shown schematically in Fig. 3.
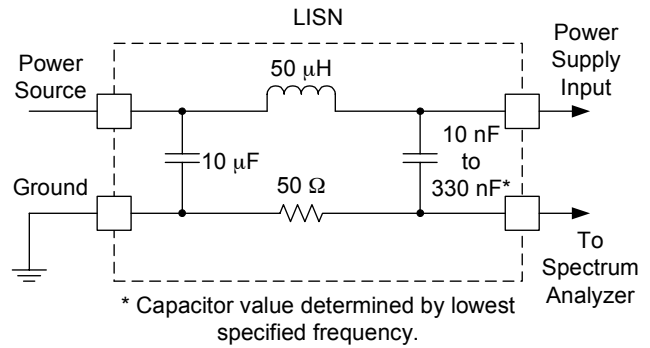
Fig. 3. A 120/240 VAC, 60 Hz LISN circuit. All measurements are made with respect to ground.

This device allows the use of a spectrum analyzer to measure the noise current through a 50-Ω source impedance while isolating the measurement from any high-frequency shunting which might exist in the power distribution network. Typically, a LISN network is added to each of the input power lines and the noise signals are measured with respect to ground.

Conducted noise at the power supply's input can be further sub-categorized into two separate modes of current flow: differential mode (DM) noise, as measured between the power feed and its return path; and common mode (CM) noise which is measured between each of the power lines and ground. The contributors to these two modes are inherent to the basic operation of a switching power supply. The action of the internal power switches causes rapid dI/dt changes in the differential current at both the input and outputs of the power supply, as illustrated in Fig. 4. Of course, input and output filters ideally would eliminate any high frequency noise external to the power supply, but neither can do the job completely. So residual ripple and switching spikes exist as a differential mode noise source with current flow bidirectionally into one terminal and out the other.
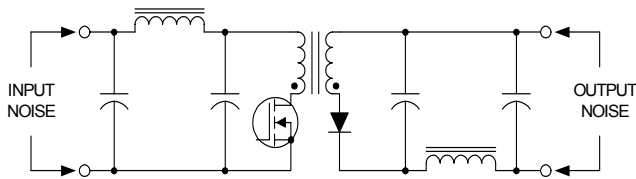
Fig. 4. Differential mode (DM) noise current is produced by normal switching action.

There are also sources of rapidly changing voltage within the power supply which can couple noise through parasitic capacitance to earth ground, some of which are shown in Fig. 5. This type of noise in the ground path, which can be seen as common mode noise on all power supply terminals, is measured with respect to ground.
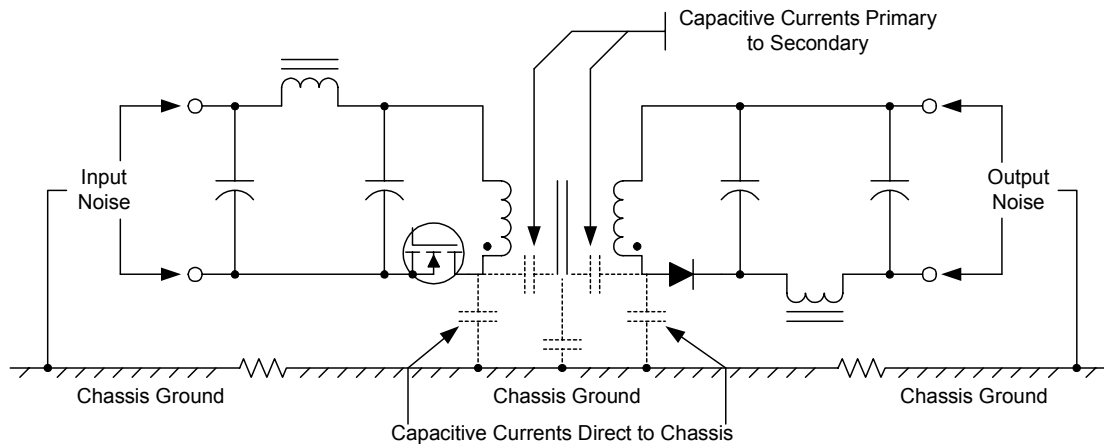
The paths for both DM and CM noise currents at the power supply's input are shown in very simplified form in Fig. 6, which also shows the application of two LISN devices, in series with both the power line input and its return.

Note that both CM and DM modes of current are present in each LISN but, with the polarities shown, the power line LISN measures CM + DM, while the LISN on the neutral return measures CM − DM. Both of these quantities are vector sums and, where necessary, there are circuit networks that can be used to separate CM and DM signals,[Ref. 9] but the specifications typically do not differentiate. The total noise at each input must be measured because, with the possibility of multiple current paths within the power supply, there is no reason to assume that the values of the CM and DM contributions at the two inputs are identical.

Fig. 5. Common mode (CM) noise occurs when transients from switching voltages are coupled to earth (or chassis) ground through parasitic capacitances.
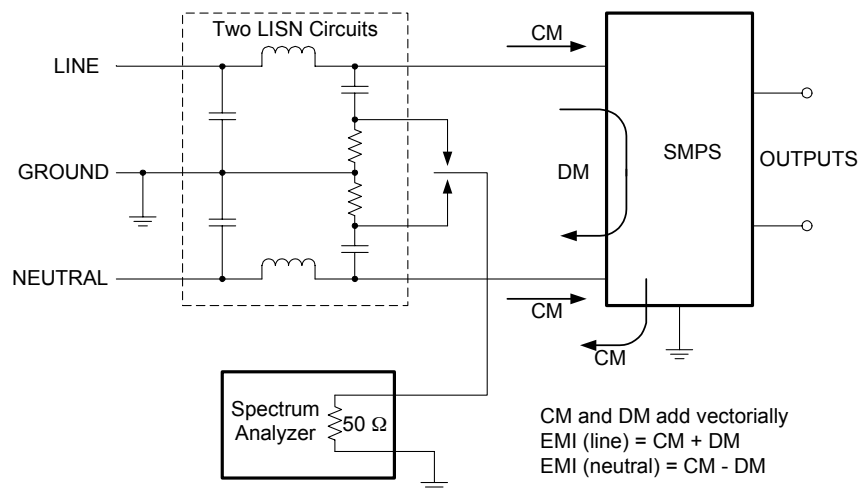
Fig. 6. DM and CM currents at the power supply's input.

V. MINIMIZING CONDUCTED DM NOISE

The first line of defense against conducted noise is the use of filters, but often this is easier said than done. The practical aspects of effective filter design are reviewed in Figs. 7 and 8, which should be referred to in the following discussion. *Some confusion could be created by the nomenclature used in Fig. 7. We will analyze the filter performance in terms of voltage attenuation between the V_{IN} terminal, which is connected to* the power source, and Vout which is the input to the power supply. In reality the filter provides *current* attenuation between current at the Vout terminal into the power supply, and the current at the filter's Vin terminal from the power source, as measured through a LISN 50-Ω source impedance. However, either a voltage or a current analysis will yield the same attenuation performance.
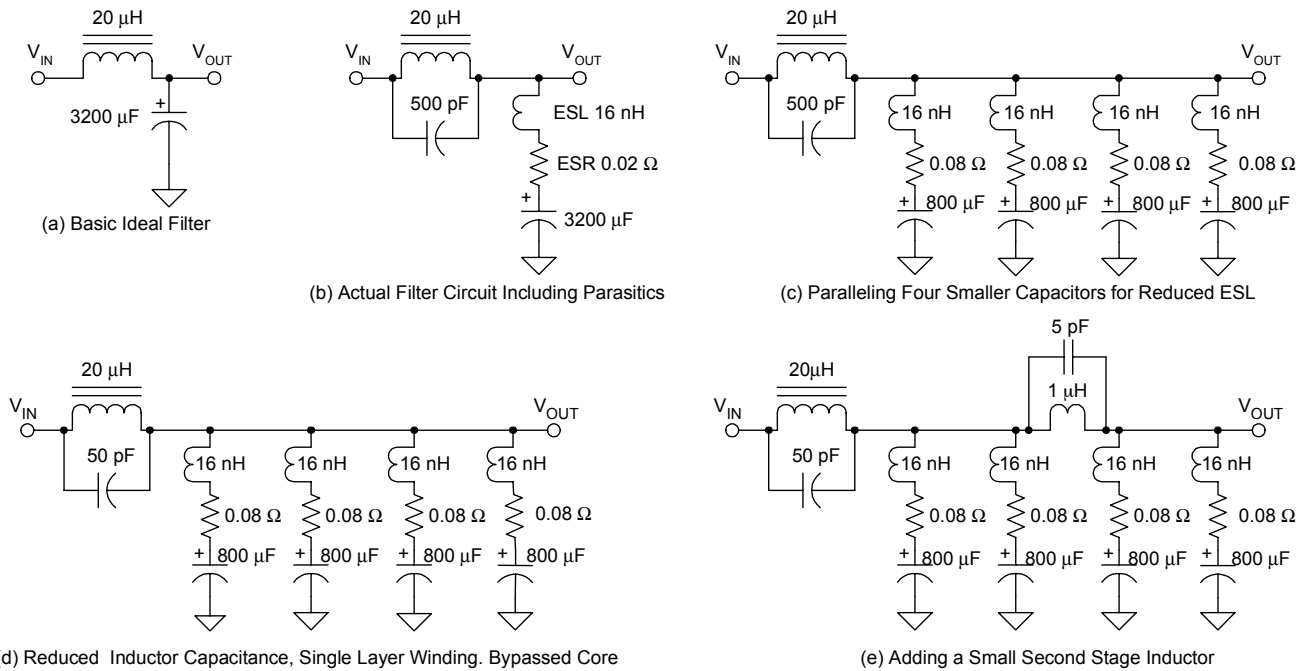
Fig. 7. DM noise minimized by a filter design that pays careful attention to component parasitics.
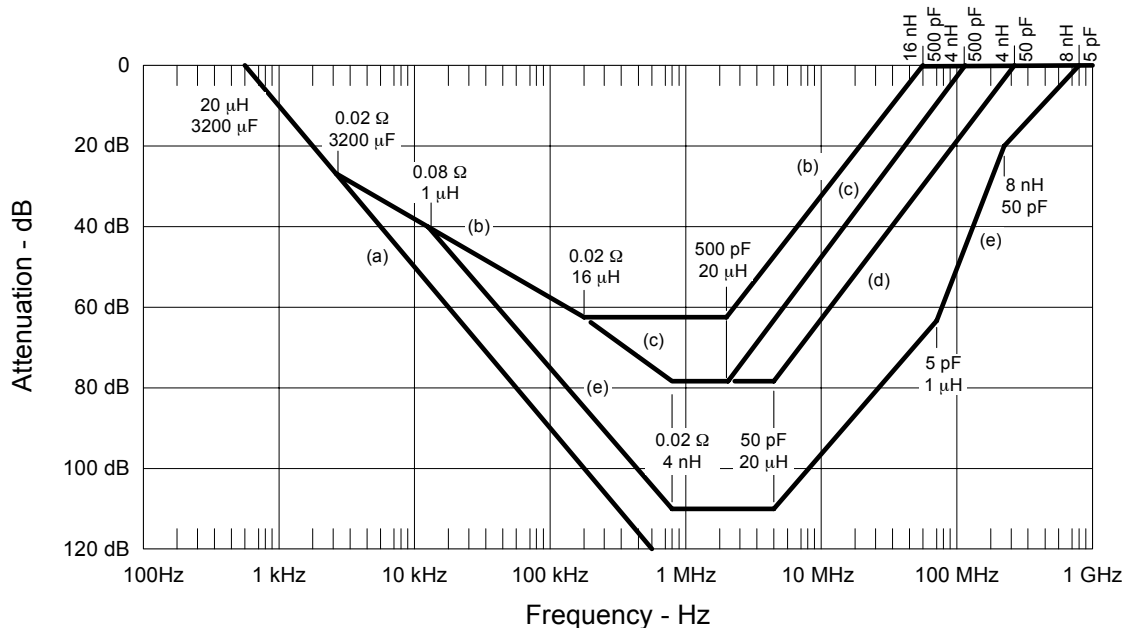
Fig. 8. DM filter frequency attenuation vs. circuit configuration.

The first point to remember is that if it is differential mode noise we are attempting to combat, the filter must connect across the differential lines. In other words, in Fig. 7(a) - which shows an ideal LC filter in the power line where V_{IN} is the power source and V_{OUT} is the power input connection to the power supply - the capacitor negative terminal must connect to the power return line - not ground! And the second point is that there is no such thing as an *ideal* filter that would yield the *ideal* attenuation curve shown in Fig. 8(a). The *actual* filter will include the effects of parasitic capacitance across the inductor, and parasitic ESL and ESR in series with the capacitor, as shown in Fig. 7(b). Using typical values for these parasitic components, the curve in Fig. 8(b) shows that now the filtering action is much less than ideal, and "bottoms out" with little more than 60 dB of attenuation.

Attacking first the capacitor, it can usually be shown that paralleling several smaller capacitors, instead of using one large one, may reduce the parasitic values and achieve some improvement such as shown in Figs. 7(c) and 8(c).

Recognizing that greater attention to the inductor may also result in reduced parasitics, Fig. 9 indicates some of these considerations. Parasitic capacitance can be reduced if the design accommodates a single layer winding that achieves maximum spacing between the start and finish ends of the coil. However, capacity from winding to core must also be considered. Since the core is probably also a conductor, stray capacitance to the core can have the same effect of shunting the winding. By careful control of the winding process, the inductor's parasitic capacity can often be significantly reduced, yielding further improvement in filter performance as shown in Figs. 7(d) and 8(d).
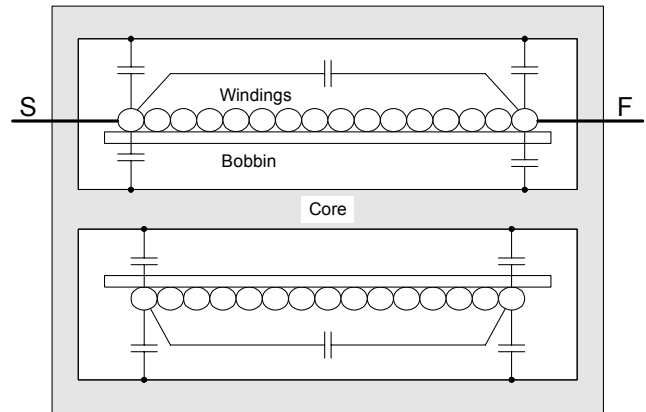
Fig. 9. A single-layer winding reduces capacity but coupling to the core must still be considered.

Often, a significant benefit, at minimal added cost − is to add a small "polishing" inductance between some of the paralleled capacitors. This essentially makes the circuit a two-stage, or second-order, filter yielding a performance gain as shown in Figs. 5(e) and 6(e).

Before leaving the design of this filter, its performance should also be examined from a resonance perspective. A simple L-C filter is resonant at its natural frequency:

$$f_r = \frac{1}{2\pi\sqrt{LC}}$$

and this has the potential to introduce three additional problems:

1. A step application of input voltage could cause the capacitor voltage to ring to a value that could approach twice the input voltage, possibly damaging a following converter.

2. If there is a potential for high-frequency noise on the input power bus, any component of that noise at the filter's resonant frequency is amplified by the "Q" of the filter.

3. The effective output impedance of the filter, if undamped, rises at the resonant frequency, raising the specter of oscillations with the input impedance of a following converter.

For any or all of these reasons, filter damping may be important. While there are many approaches to optimizing filter damping, a good starting point is to add a series R-C across the filter's capacitor as shown in Fig. 10.
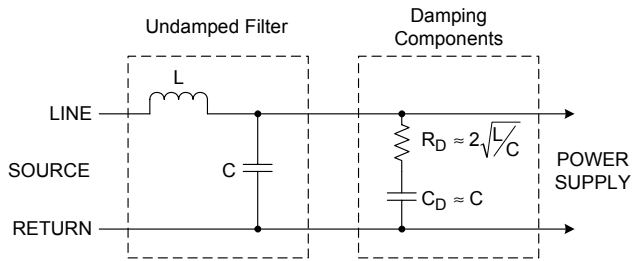
Fig. 10. A damping R-C network may be necessary to minimize problems caused by filter resonance.

VI. MINIMIZING CM GROUND NOISE

The above discussion has been with respect to optimizing a filter for DM noise, but the same considerations can apply to a filter intended to reduce CM conducted noise. As stated earlier, CM noise is largely created by parasitic capacitance to ground. A simplified example of one of the major problem sources and its potential solution begins with Fig. 11.
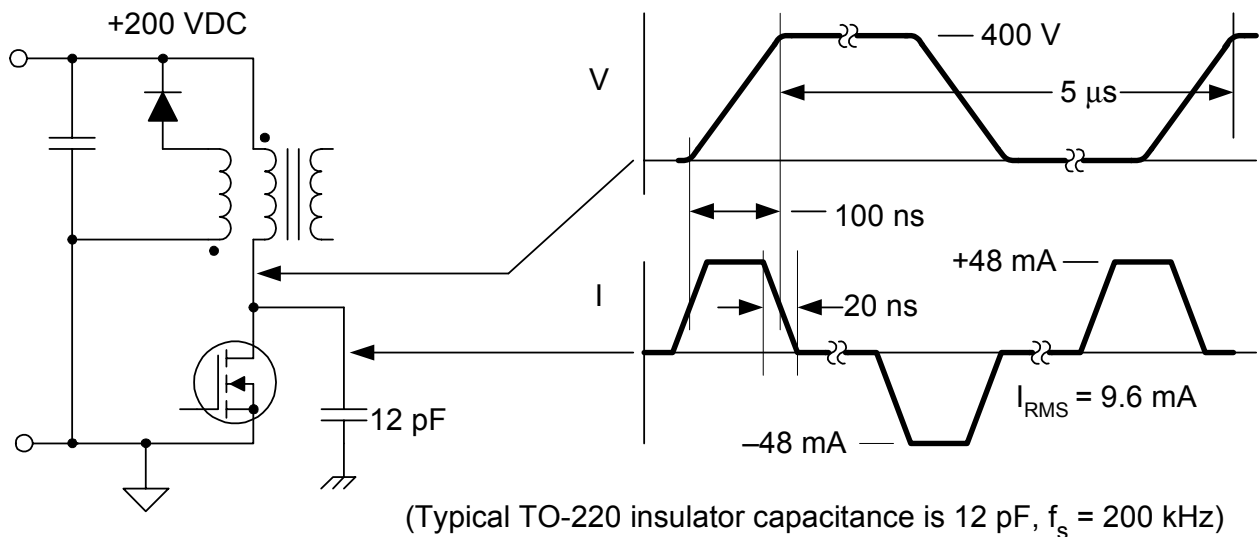
In this example, we show a single-ended power stage with a switching FET alternating between driving a transformer primary and blocking its reset voltage of 2 V_{IN}. With a switching frequency of 200 kHz, one might expect a FET drain voltage swing of 400 V with a rise and fall time of 100 nsec as shown in the figure. If, for example, the FET is in a TO-220 package mounted on a grounded heat sink with an insulating washer, the parasitic capacitance from drain to ground could be in the 12 – 32 pF range. Using the lower number, and solving for I = C(dV/dt) results in a peak current of +/- 48 mA – a not insignificant amount of injected ground noise. Fig. 12 illustrates how this injected current completes its path back to the power lines through the two 50-Ω LISN resistors in parallel, thereby creating a noise voltage at each LISN output.

(Typical TO-220 insulator capacitance is 12 pF, f_s = 200 kHz)

Fig. 11. A typical CM noise source is capacitive coupling of the switching FET's drain voltage waveform to a chassis-grounded heat sink.
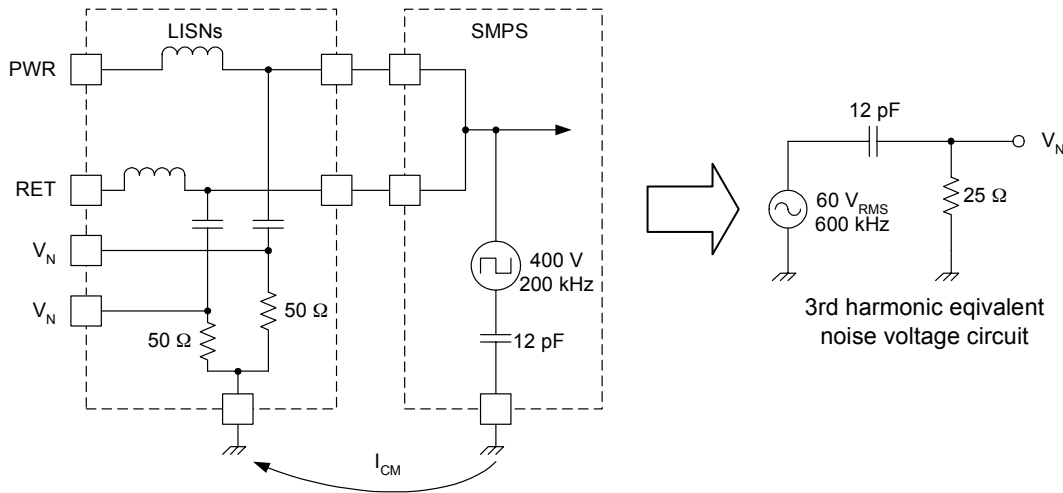
Fig. 12. Current induced in chassis ground has a return path through the LISN output impedance.

The solution for this problem is the insertion of a CM filter in series with the current path and its definition is most easily accomplished by using the equivalent voltage noise circuit also shown in Fig. 12. This equivalent circuit is generated by starting with the 400-V peak-to-peak waveform at the FET's drain terminal. Assuming a 50% duty-cycle square wave, the rms value of this signal is 200 V at the switching frequency of 200 kHz. Since 200 kHz is below the FCC minimum specified frequency of 450 kHz, we can ignore the fundamental and look to the third harmonic - 60 V_{RMS} at 600 kHz - as the most significant contributor to the noise spectrum. This signal yields a noise voltage across the paralleled LISN resistors of approximately 68 mV which, while seemingly a fairly small value, must be compared with the FCC noise limit of 1.0 mV for Class A products at 600 kHz. (The limit for Class B is even lower at 250 µV.)

To reduce 68 mV to less than 1.0 mV, we need to insert a filter that will produce an attenuation of 37 dB at 600 kHz. One way to accomplish this is with the addition of a series common mode inductor. Working backwards, we can calculate the required inductance from the reactive impedance we need at 600 kHz as shown in Fig. 13. However, completing this analysis would show that, while the required inductance would be 419 mH, we cannot have more than 0.17 pF parasitic capacitance across the inductance. Not likely!
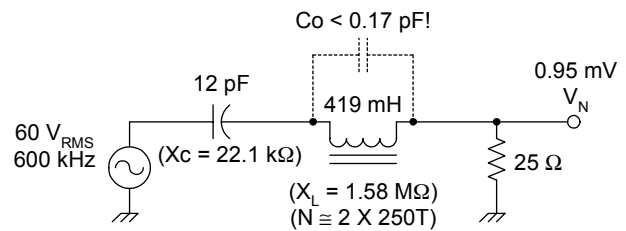
Fig. 13. Achieving 37 dB attenuation with a series inductor requires an unrealisticly low parasitic capacitance.

We will then try a shunt capacitor as shown in Fig. 14 to lower the impedance of the 25 Ω load to get under the 1.0 mV limit. The same calculations in this case yield a requirement for 760 nF, which could have a parasitic series inductance of 93 nH and still meet the attenuation requirement. While this at least looks viable, we run into another problem in that if the input to this power supply is ac line voltage, there are usually safety specifications that limit input line capacitors to less than 10 nF.
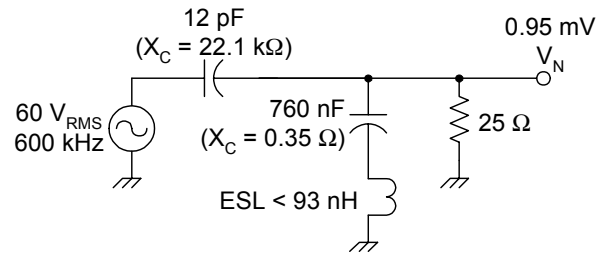
Fig. 14. Achieving 37 dB attenuation with a shunt capacitor requires an unacceptably large value.

So we finally come to the best solution for a CM input filter, which includes both inductance and shunt capacitance, but with reasonable values for each, even considering expected parasitic values. This solution is shown in Fig. 15 and is typically implemented with two windings on a single core wound in opposition such that the flux caused by the power supply's dc input current will cancel and not contribute to core saturation. Of course, in this application the shunt capacitors are connected to ground instead of differentially.
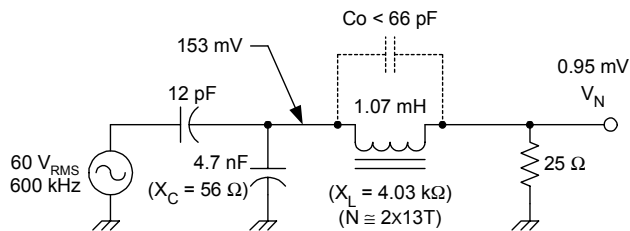
Fig. 15. Using both a shunt capacitor and a series inductor achieves a solution with practical values for both.

A CM input filter may also have to be damped to prevent problems at resonance, but since the maximum capacitance is limited, the usual approach is to divide the required capacitance into two capacitors, and then place an appropriately sized resistor in series with one of them.

Fig. 16 shows a possible input filter configuration combining both DM and CM filters. In this figure, the input load power is shown as flowing from left to right, from the ac line to the input rectifiers of the power supply. The noise signal, however, is flowing from right to left, from internal sources within the supply back toward the ac line terminals, which are the external input power connections.

Working from right to left, C_{d1} and L_{d1} represent the main DM filter. C_{c1} and L_{c1} (2 of each) form the CM filter for ground noise, with the two windings of L_{c1} built onto a single core. C_{d2} works with the leakage inductance between these windings to form a second, higher-frequency DM filter element. Ln and Cn form a notch filter at the switching frequency with Rn providing both damping and some spreading of the notch width. Finally, C_{d3} and C_{c2} (2) are often used for a last cleanup, right at the terminals where a pair of LISN devices would be connected to evaluate noise performance. These last capacitive elements work against the undefined ac line impedance so their performance, while helpful, is difficult to predict. Additional components, not shown in this figure, could be damping networks, transient protectors, fuses, and other application-dependent devices.

Fig. 16. Multiple filter components are combined in this example of conductive noise reduction circuitry.

VII. ADDITIONAL CM NOISE SOURCES

Before leaving the subject of CM conducted noise, we should mention that the drain terminals of the power switches are certainly not the only place where high dV/dt signals might introduce ground noise. Heat sinks are another potential problem area, as safety requirements typically do not allow them to be charged to a high voltage potential. If the heat sink is small enough that it can be enclosed within the power supply's case, then connecting it electrically to the circuit common rather than chassis ground may be an acceptable solution. Using an insulated intervening bracket connected electrically to circuit common may allow the accommodation of larger, grounded heat sinks by shunting capacitively coupled noise to the common rail where it can more easily be attenuated. These options are shown in Fig. 17.

The same approach for diverting what might become CM noise away from ground is applicable to transformers by the use of electrostatic shields between the windings, as shown in Fig. 18. In many power supply designs, the secondary-side output circuitry is ground referenced, and it therefore follows that any high voltage ac potential on the primary side, which is coupled through the transformer by parasitic capacitance, can become CM noise. Properly applied, electrostatic shields can prevent this by diverting the coupled noise into the primary common bus. (As an additional hint, the shield could do its job connected to either the high voltage rail or the return. As a rule of thumb, if the power FET dV/dt is greater at turnon than at turnoff, connect the shield to the return, but if turnoff is faster, then connect it to the positive rail.) Alternative shield configurations are shown in Figs. 19 and 20.
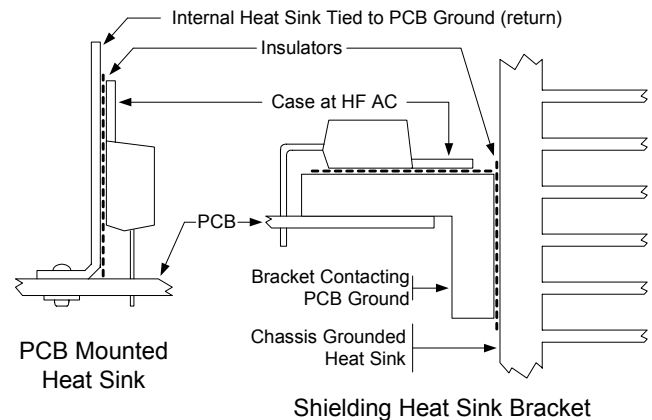
Fig. 17. Two heat sink arrangements that divert capacitively coupled noise to circuit common rather than chassis ground.

Use of Primary Shield
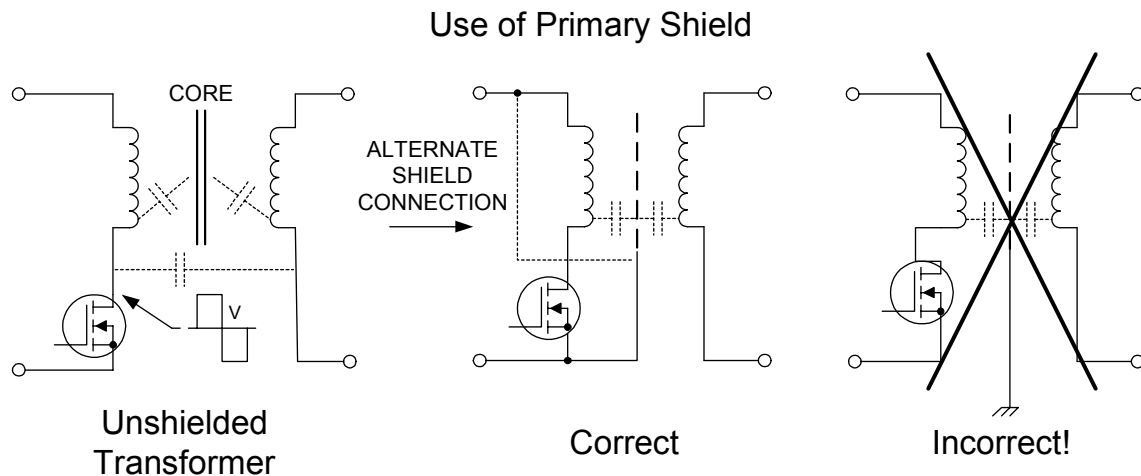
Fig. 18. Using an electrostatic shield in a transformer can minimize CM noise.

Shielding primary and
secondary windings.

Adding a third shield for safety.

Fig. 19. Alternative shield configurations.

Basic arrangement of
PRI and SEC shields.

Shielding a floating or
safety grounded core
from a HV primary.
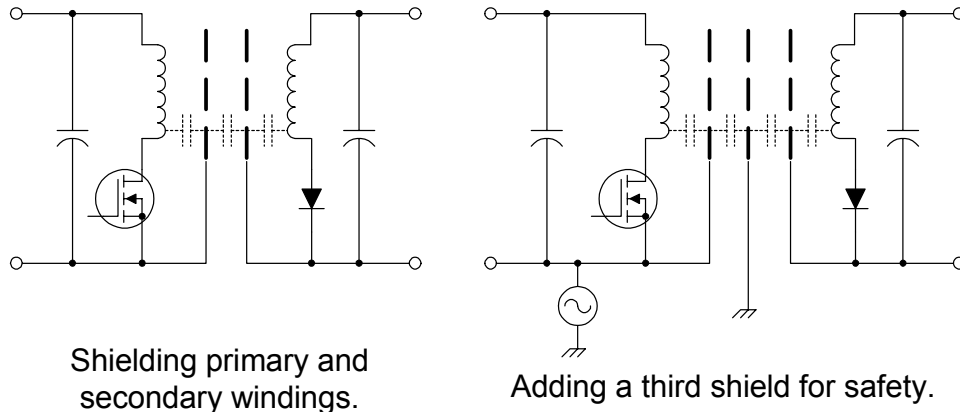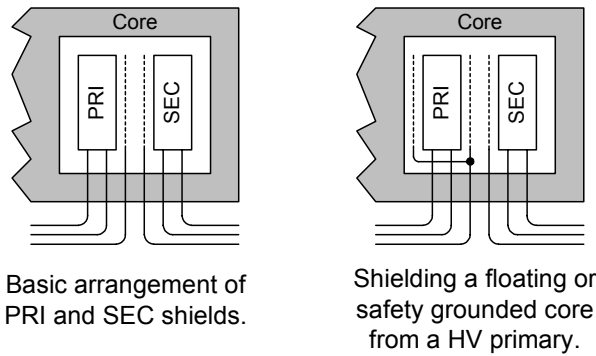
*Fig. 20. Positioning shields for maximum
effectiveness.*

VIII. RADIATED EMI

As we leave the subject of conductive EMI and move on to radiated interference, we should first emphisize that while we treat each type of noise separately, they are not unrelated. In an electronic system, particularly a switchmode power supply, EMI energy can be transformed back and forth between conducted and radiated forms - perhaps even several times - between its generation and its measurement. If the noise energy is conducted in a wire or PCB trace, an electromagnetic field is created which gives us radiated EMI. If there is then mutual inductance or capacitive coupling to another conductor, then the radiated energy is transformed back to conducted noise, but now in a different location in the system. The point is that any conductor can become an antenna, and an antenna can both send and receive radiated signals.

Testing for radiated electromagnetic compatibility is a much more complex process than for conducted noise. In the first place, because - by specification - we are looking for signals above 30 MHz, all the test instrumentation becomes more crucial. Second, the test environment must be well-controlled, which typically means using either an RF screen room to shield the test setup from any extraneous RF signals from other generators or reflections, or open-air testing in a well-defined environment. In either case, a knowledgeable operator is vital in order to obtain reliable data. The type of antenna and its distance and orientation with respect to the device under test, as well as a method to sweep all radiating angles, are all important parts of the test conditions. A schematic representation of a radiation EMI test setup is shown in Fig. 21.
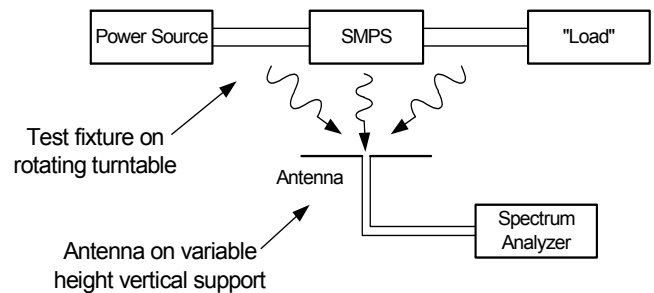
Fig. 21. Measuring radiated noise with an antenna.

One potential escape from this problem is a clause in the FCC requirements which states that the frequency range of EMI testing is based on the highest fundamental internally generated clock frequency, and if that frequency is less than 1.705 MHz, then the maximum test frequency is 30 MHz. While this might preclude the need for radiation testing on most stand-alone power supplies, it is unlikely to be of much benefit when the power supply is combined with its load and the entire system must be evaluated for EMI.

Note that in performing system-level radiation EMI tests, the input and output connections to the power supply need to be included, and it is here where high frequency conducted EMI energy might generate an RF field and contribute to the total radiated noise. For this reason, ac line cords often include a ferrite snubber and, if the actual system load is not included, the power supply should be loaded with passive resistors rather than an electronic load unless the effects of dynamic loading are an objective of the testing program.

IX. COMBATING RADIATED EMI

The contributors to radiated EMI can also be subdivided into two categories depending upon how the energy is generated. This energy can be from an <u>electric</u> field, which is generated by dV/dt on conductive surfaces, or from a <u>magnetic</u> field which is generated by dI/dt in conductors. The nature of these fields change as the distance from the source to the point of measurement increases. At distances close to the source, the fields are determined by their cause and both electric and magnetic fields must be considered separately. This region is called the *near-field*. Beyond this region (in the *far-field*) the two fields meld together into a single electromagnetic radiation. The boundary between *near-field* and *far-field* distinctions is defined by the wavelength of the noise energy, as $\lambda/2\pi$. When you consider that the wavelength of a 1 MHz signal is 300 meters, it is usually safe to assume that most power supply concerns can be limited to near-field phenomena.

Within the near-field region, an additional assumption which is usually valid is that if the source has high voltage and low current, the resultant field will primarily be electric, while with a high current, low voltage source, a magnetic field will predominate.

An electric field is produced when switched voltages are present on surfaces such as heat sinks or magnetic cores, causing them to act as antennas. Typical locations within a power supply where this might occur are shown in Fig. 22. Electric fields can usually be shielded relatively easily by conductive enclosures, where the conductive material terminates the field by converting it to current. Of course, there must be a path for this current but, with the enclosure normally grounded, this current merely contributes to overall CM conducted noise energy where it can be addressed with filters as previously described.
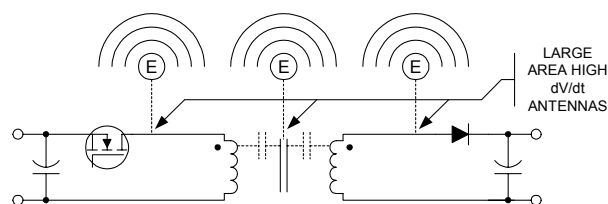
Fig. 22. Electric fields are generated by surfaces with high dV/dt acting as antennas.

Magnetic field EMI energy can emanate from a power supply as either stray fields from transformers or inductors, or as magnetic fields which are created when there is a rapidly changing current flowing in a conductive loop. It is here where internal wiring layout becomes very critical as the magnetic field from a current loop is a function of the inductance that is determined by the area enclosed by the loop. Fig. 23 shows just one example where the high-current secondary leads from a power transformer connect to the output rectifiers. While this is a flagrant illustration of "good" and "bad" practices of handling conductors with high dI/dt, these basic principles should be observed throughout the power supply.
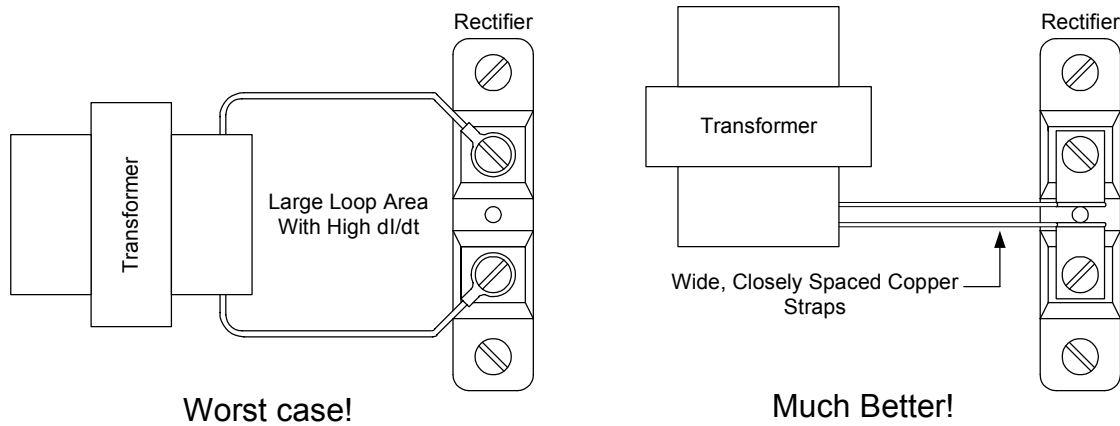
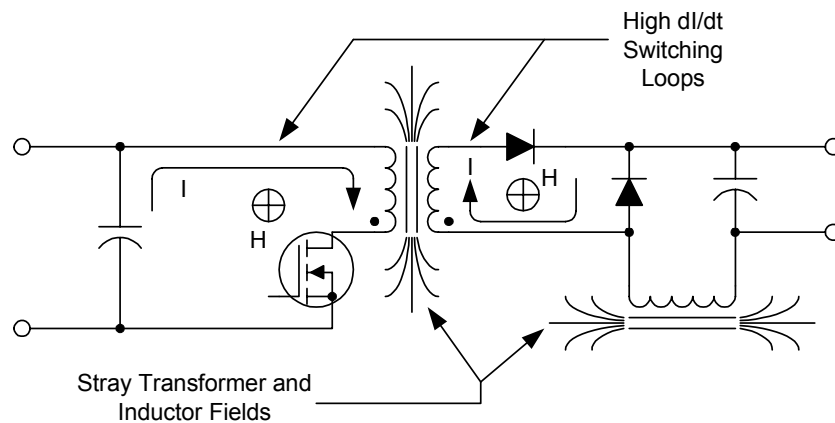*Fig. 23. Conductors with large loops and high dI/dt make excellent antennas!*

Fig. 24. Magnetic fields must be minimized by design as shielding can be difficult.

Additional power supply contributors to magnetic fields are shown in Fig. 24. Magnetic fields are not as easily shielded - it takes a magnetic material to block a magnetic field and while magnetic material in sheet form exists, it is (1), very expensive when used for this purpose, and (2), its magnetic properties fade very rapidly at the higher frequencies of EMI. However, shields of non-magnetic, conductive material can be used for magnetic fields, but the process is that the magnetic noise induces eddy currents to flow in the enclosure, which in turn generate a canceling magnetic field. The catch here is that there can be no interruption of these eddy currents - any gaps, joints, or holes in the shielding will allow a surprising amount of the magnetic field to escape, or *leak*, from the enclosure. Thus it is far better to combat the problem at its source by minimizing current loops and containing magnetic fields.

In designing transformers for switchmode usage, it is well known that leakage inductance between primary and secondary windings can be detrimental to electrical performance, but it can also contribute significantly to radiated magnetic fields. Fig. 25 shows a two-winding transformer wound on what might be an EE or EI core structure. The leakage inductance generates a transverse magnetic field between the windings and, while some of this field may be captured by the core, the rest acts as a magnetic dipole radiating out into surrounding space with an intensity which decays as the cube of the distance.

A change in the winding procedure to interleave the primary as shown in Fig. 26 now produces two leakage fields with opposite polarities. This provides for a significant amount of cancellation and the resultant quadrapole field falls off with the fourth power of the distance, and thereby greatly attenuates the radiant energy.
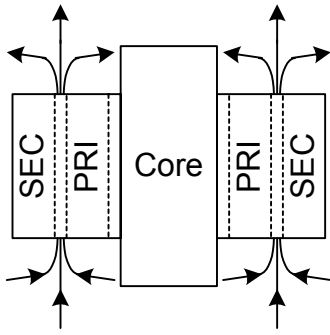
Fig. 25. The transformer leakage inductance field for a single primary and secondary is a dipole field.

Fig. 26. Sandwiched windings create opposite leakage field dipoles that tend to cancel.

Another technique for reducing stray magnetic fields from a transformer is the use of a conductive "flux strap" (also sometimes called a "belly band", "flux band", or "hum strap"). This copper band, illustrated in Fig. 27, provides a path for the eddy currents that result from the leakage inductance magnetic dipole. The current flowing in the flux strap then creates an opposing magnetic dipole which tends to cancel the original field at close proximity to the transformer.

Inductors are also potential generators of stray magnetic fields. Fig. 28 illustrates a poor inductor design with significant stray field caused by the gaps in the core which are outside the coil winding. Changing the core design so that all the gap is in the center leg, and now fully contained within the winding, reduces this source of radiated EMI. Fig. 29 shows some additional inductor designs that can create problems with stray fields.

Fig. 27. A continuous flux strap around the transformer further reduces magnetic fields.
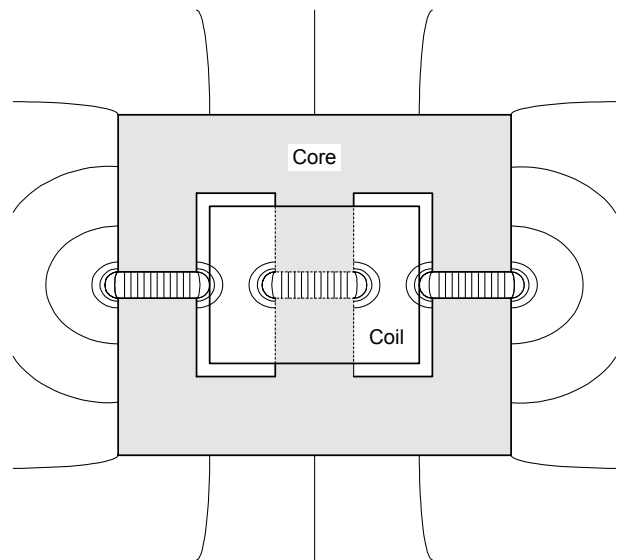
Fig. 28. Stray magnetic fields are caused when air gaps are not enclosed by the inductor winding.

Magnetic fringe field around partial toroidal winding on a distributed gap core.

Drum and rod core fields, good antenna design; poor for EMI.
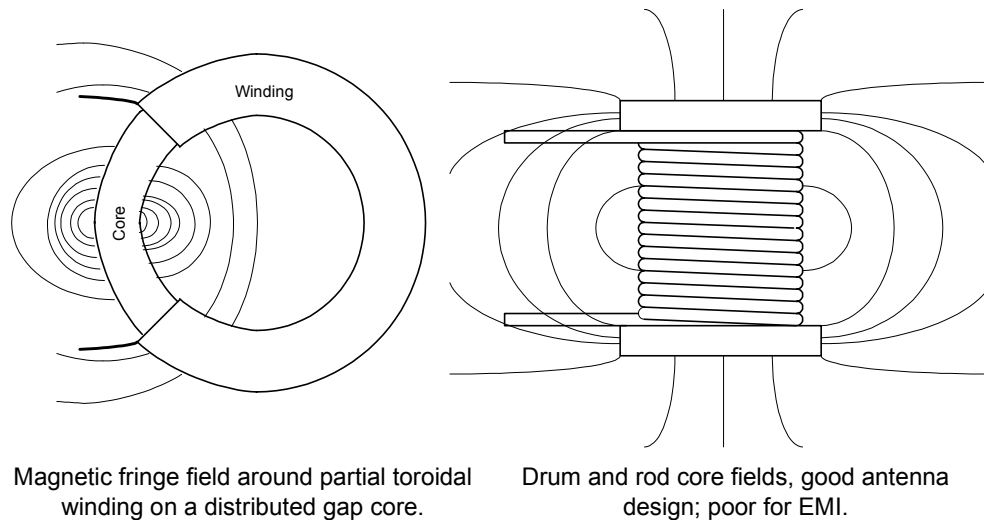
Fig. 29. Other inductor designs with strong leakage fields.

X. FINAL THOUGHTS

This Topic has attempted to cover the high points of EMI prevention by the use of design techniques applied within the power supply. We have described the differences between conducted and radiated noise, emphasizing that while the solution to conducted noise lies with effective filter design, radiated noise prevention is largely influenced by construction techniques. One issue not discussed is the other side of the coin - susceptibility to noise from external sources; but it turns out that in most cases, the best defense is a good offense. In other words, the action taken to reduce the generation of noise often also contributes to reduced susceptibility.

Finally, there is another "noise reducing" technique which some have applied to switchmode power supplies. Since EMI specifications are written to evaluate noise at specific frequencies, by using some form of random modulation of the supply's switching frequency, the noise generated by the power supply is "smeared" or spread out instead of falling at specific harmonics of a constant fundamental.[Ref. 8] Since this can greatly reduce the <u>average</u> energy content at any <u>specific</u> frequency, compliance with FCC specifications is certainly eased, but this technique is controversial over whether it is just exploiting a *loophole* in the standards or truly benefiting the system's EMC.

XI. ACKNOWLEDGMENT

XII. REFERENCES

[1] Mark Montrose, *Printed Circuit Board Design for EMC Compliance*, Second Edition, IEEE Press, Piscataway, NJ, 2000.

[2] Henry Ott, *Noise Reduction Techniques in Electronic Systems*, Second Edition, John Wiley & Sons, New York, 1988.

[3] Bruce Carsten, Design Techniques for the Inherent Reduction of Power Converter EMI, Powercon 11 Proceedings, Dallas, TX, 1984.

[4] Bruce Carsten, *Design Tricks, Techniques, and Tribulations at High Conversion Frequencies*, Second HFPC Conference Proceedings, Washington DC, 1987.

[5] Bruce Carsten, *H-Field Probes Spot Switchmode Supply EMI*, PCIM Magazine, September 2001, pp. 24-36.

[6] David Williams, *A Tutorial on EMI Characterization of Switching Regulators*, IEEE APEC'96 Conference Proceedings, San Jose, CA, pp. 333-339.

[7] Peter Bardos, *Predicting the EMC Performance of High Frequency Inverters*, IEEE APEC'01 Conference Proceedings, Anaheim, CA, pp. 213-219.

[8] D.A. Stone, B. Chambers, & D. Howe, *Easing EMC Problems in Switched Mode Power Converters by Random Modulation of the PWM Carrier Frequency*, IEEE APEC'96 Conference Proceedings, San Jose, CA pp. 327-332.

[9] T. Guo, D. Chen, & F. Lee, *Separation of the Common-Mode and the Differential-Mode Conducted EMI Noise*, HFPC 1994 Conference Proceedings, San Jose, CA, pp. 128-134.

[10] Richard Farrington, *EMI Characteristics*, SynQor Application Note 00-08-02, SynQor Power Supplies, Hudson, MA, January, 2001.

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