Power Supply Layout and EMI

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Introduction

PC-board layout determines the success or failure of every power supply project. It sets functional, electromagnetic interference (EMI), and thermal behavior. Switching power supply layout is not black magic, but is often overlooked until it is too late in the design process. Fortunately physics is on your side. Functional and EMI requirements must be met, and in a world of trade-offs in power supply unit layout, what is good for functional stability is good for EMI. Good layout from first prototyping on does not add to cost, but actually saves significant resources in EMI filters, mechanical shielding, EMI test time and PC board runs. This application note focuses primarily on nonisolated topologies, but will examine some isolated topologies as well. You will learn to make the optimum choices regarding PC-board layout for solid power supply designs.

I remember about a dozen years ago as a customer was using a switch mode power supply in a car radio for the first time many of his colleagues said that it could not be done. However, after a few things were ironed out in layout and input filtering, everything worked fine. Later a customer successfully used an LT®1940 1MHz dual step-down switching regulator, which operated in the middle of his AM band in a car radio receiver. No additional metal shielding was required for the power supply unit (PSU); it was only an issue of placement and layout. In order to get there, we need to go through some physics.

With nonisolated topologies, one of the most basic topologies is the buck regulator. EMI starts off from high di/dt loops. The supply wire as well as the load wire should not have high AC current content. So we can focus our analysis from the input capacitor, C_{IN}, which should source all relevant AC currents to the output capacitor, C_{OUT}, where any AC currents end.
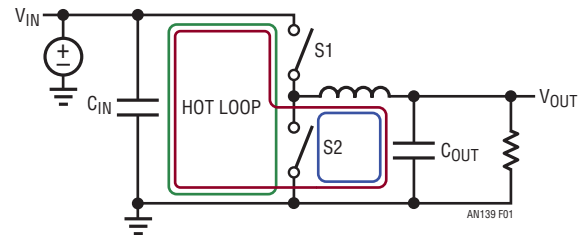

Figure 1

During the on cycle with S1 closed and S2 open, the AC current follows the red loop (Figure 1). During the off cycle, with S1 open and S2 closed, the AC current follows the blue loop. Both currents have a trapeze shape. People often have difficulty grasping that the loop producing the highest EMI is not the red nor the blue loop. Only in the green loop flows a fully switched AC current, switched from zero to I_{PEAK} and back to zero. We refer to the green loop as a hot loop, since it has the highest AC and EMI energy.

In order to reduce EMI and improve functionality, you need to reduce the radiating effect of the green loop as much as possible. If we could reduce the PC-board area of the green loop to zero and buy an ideal input capacitor with zero impedance, the problem would be solved. But we are limited to the real world. The task of engineering is to find the optimal compromise.

Let's take a look at the layout of an LT8611 buck converter (Figure 3). The LT8611 has both switches internal, so we only have to be concerned with the connection of the input capacitor.

As you can see from the schematic in Figure 2, the hot loop is not easy to spot for layout purposes.
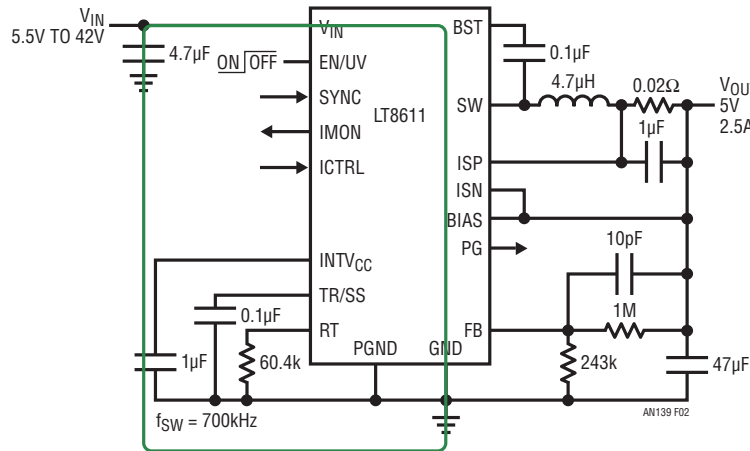
Figure 2

The green line is the hot loop in the top layer. AC current flows through the input capacitor and the switches in the part. Figure 3 shows the DC1750A LT8611 demo board. The current density in the cross cut of the hot loop will look like this (Figure 4).

How much does a copper short-circuit loop or plane under the hot loop improve the functional and EMI behavior of your circuit?
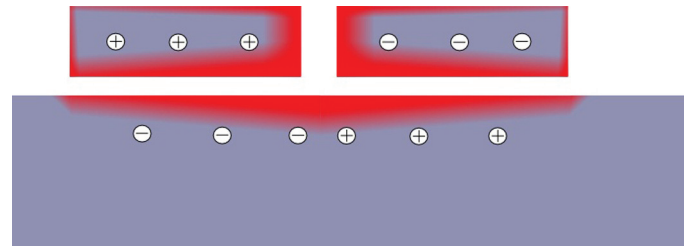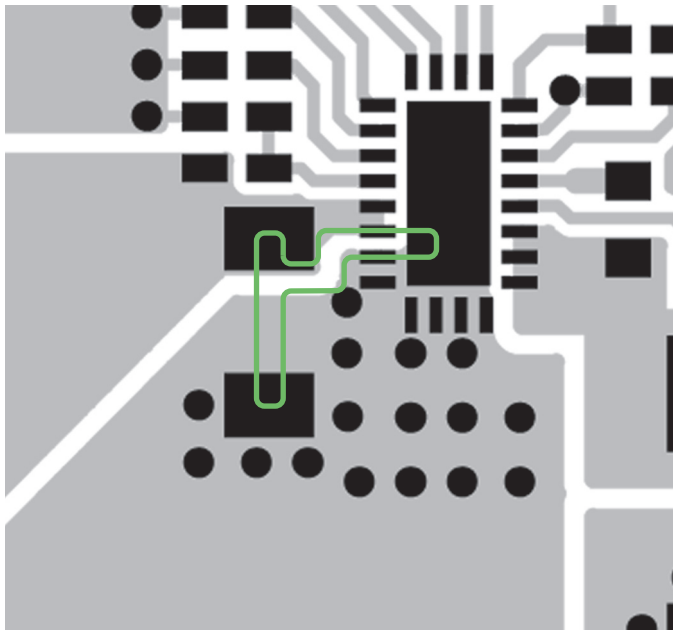

Figure 4

The result of an experiment with a 10cm × 10cm rectangular loop with 27MHz is shown in Table 1. The table gives an indication how much improvement a solid copper plane under the hot loop topside traces gives. The first line is no plane single layer.

The inductance of a single-layer loop of 187nH gets down to 13nH in the case of only 0.13mm insulation between the plane and loop traces.
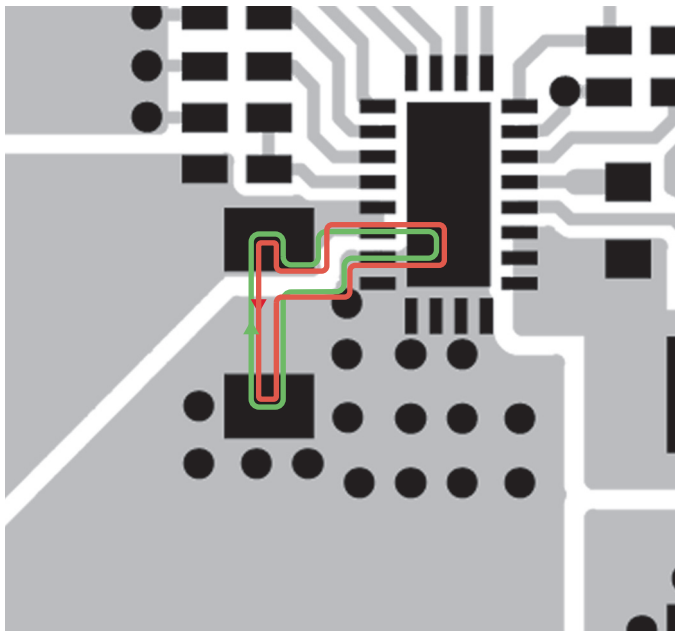

Figure 3

Table 1

d (mm)	f (MHz)	C (pF)	L (nH)		FACTOR OVER 0.12mm
	18, 4	400	187	Single-Layer Open Loop	14, 4
	21, 2	400	141	Inner Copper Short-Circuit Loop	10, 85
1, 5	38, 9	400	42	Solid Plate	3, 23
1, 5	34, 7	400	53	Rectangular Loop No Overlap	4, 08
0, 5	52, 1	400	23	Thin Rectangular	1, 77
0, 27	55	400	21		1, 61
0, 12	69	400	13	Paper	

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A solid plane on the next layer in a multilayer board (four layers or more) will have over 3× less inductance than a normal 1.5mm 2-layer board with a solid bottom plane, and over 14× less over a single-layer board. A solid plane with minimum distance to the hot loop is one of the most effective ways to reduce EMI.

Where Does the Current Flow in the Plane?

The green top layer hot loop magnetic AC field produces eddy currents in the plane (Figure 5). Those eddy currents produce a mirror AC magnetic field, which is opposite the hot loop field (red trace). Both magnetic fields will cancel out. This works better the closer the mirror current is to the hot loop. Current is a round trip in the top layer. The most likely current path in the shield is the same round trip direct under the top layer. Both currents are almost the same. Since the plane current needs to be as high as the top trace current, it will produce as much voltage across the plane as is necessary to sustain the current. To the outside it will show up as GND bounce.

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Figure 5

From EMI perspective small hot loops are best. A power supply IC with integrated sync switches, optimized pinout and careful internal switch control will outperform on EMI a non-sync power supply IC with external Schottky diode. And both will outperform a controller solution with external MOSFETs.

The boost circuit can be viewed in continuous mode as a buck circuit operating backwards.

The hot loop is identified as the difference between the blue loop if S1 is closed and the red loop (Figure 6) with S1 open and S2 closed.
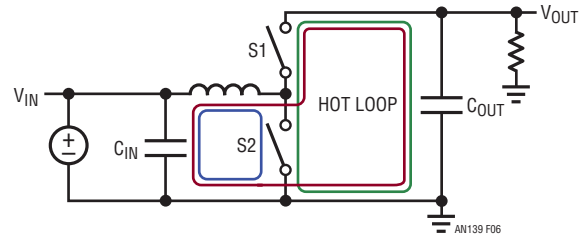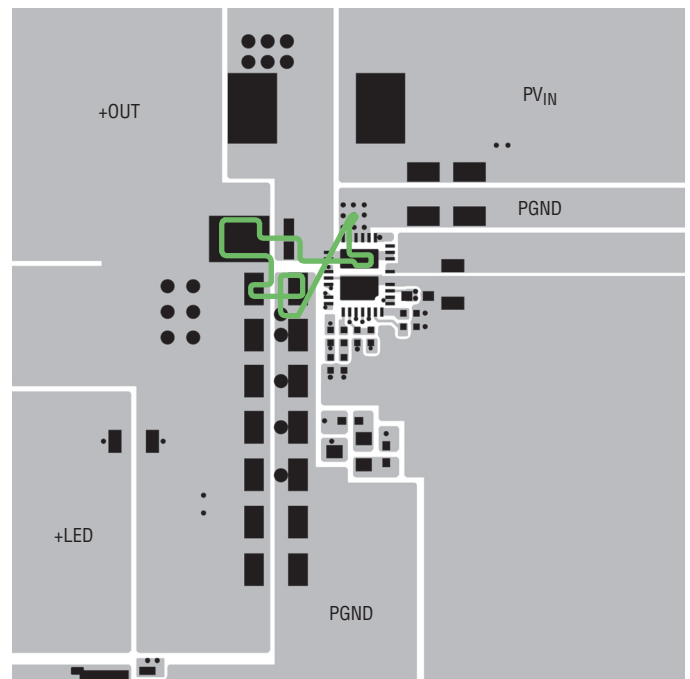

Figure 6

The hot loop of the LT3956 LED driver boost controller is shown in green (Figure 7). The second layer is a solid GND plane. The main EMI emitter is the magnetic antenna the hot loop creates. The area of the hot loop and its inductance are tightly related. If you are comfortable thinking in inductance, try to decrease it as much as you can. If you are more comfortable in antenna design, reduce the effective area of the magnetic antenna. For near field purposes, inductance and magnetic antenna effectiveness are essentially the same. See Appendices A and B for further background.

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Figure 7

The single inductor 4-switch buck-boost (Figure 8) consists of a buck circuit followed by a boost circuit. The layout will often be complicated by a common GND current shunt which belongs to both hot loops. The LTC®3780 DC1046A demo board (Figure 9) shows an elegant solution splitting the sense resistor in two parallel ones.

A bit different drawing of a SEPIC circuit (Figure 10) shows its hot loop. Instead of an active MOSFET for the top switch, a diode is often used. The LT3757 DC1341A (Figure 11) shows a good SEPIC layout. The green hot loop area is minimized and has a solid GND plane on the next layer.
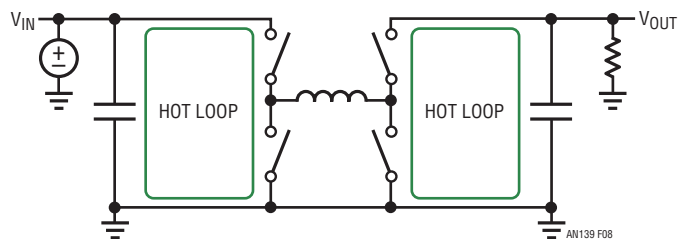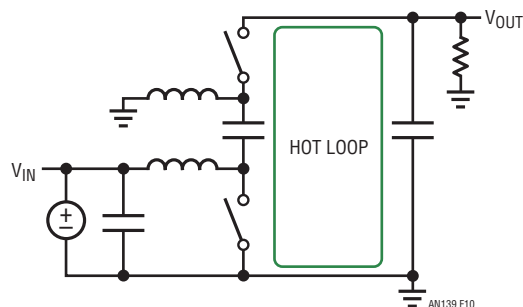

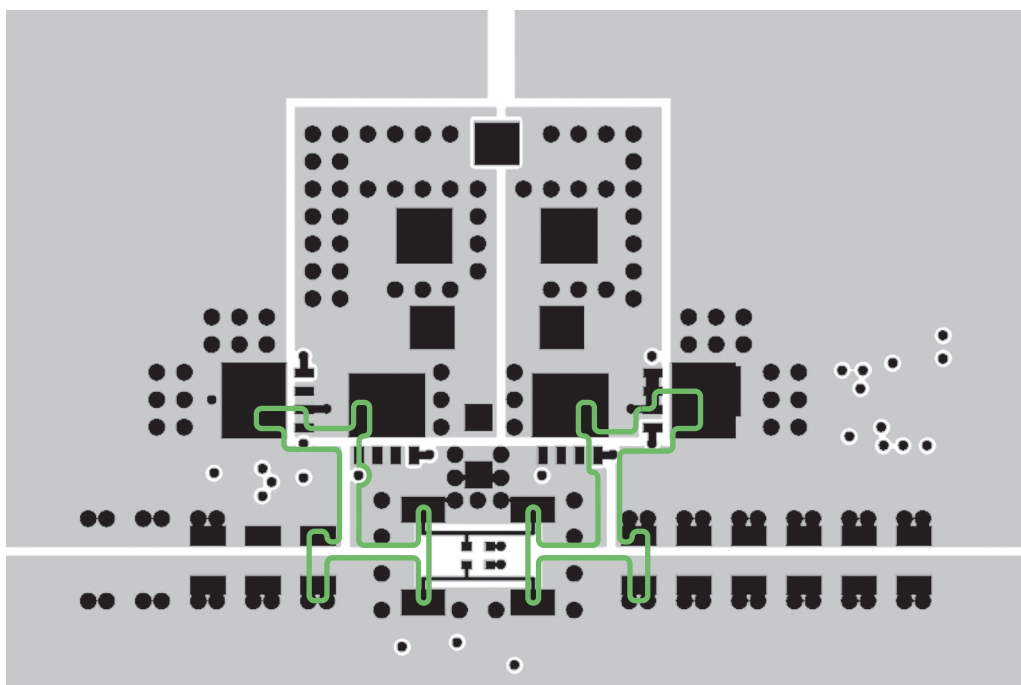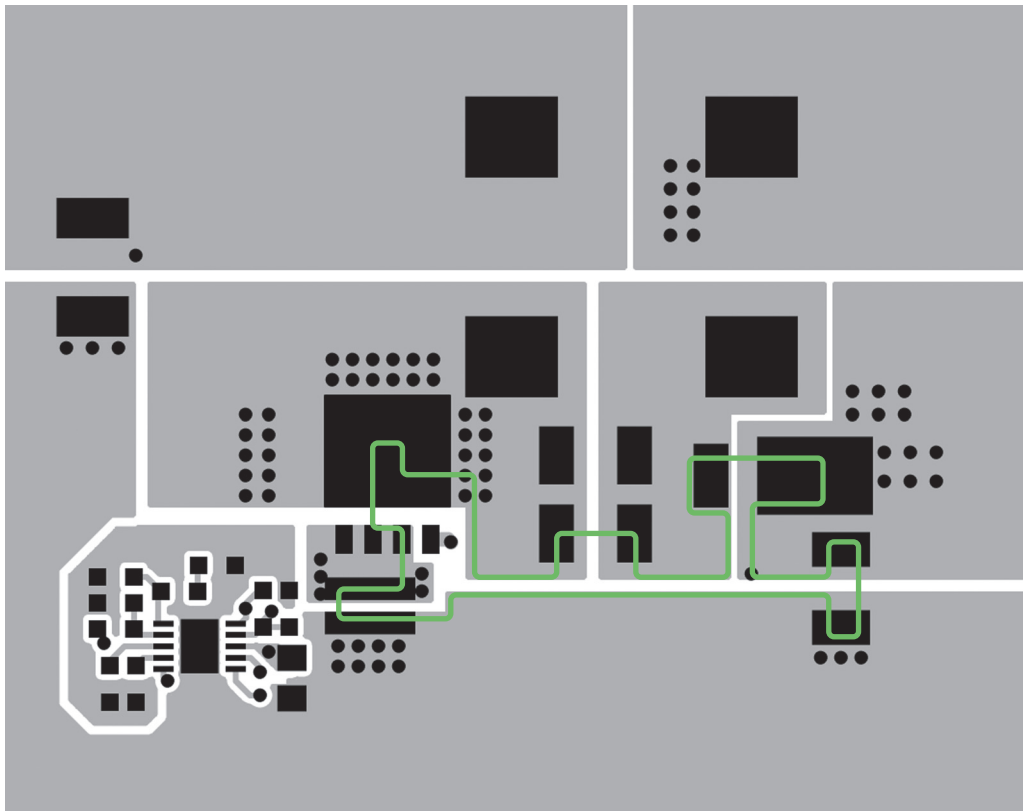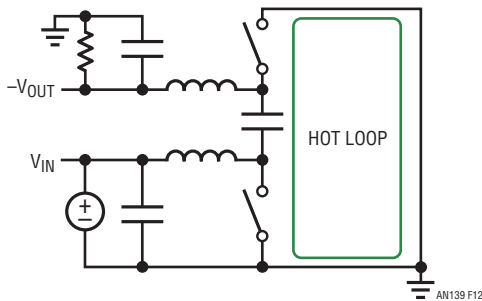Figure 8. 4-Switch Buck-Boost

Figure 10. SEPIC

Figure 9

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Figure 11

Figure 12. Inverting

The inverting topology (Figure 12) is very similar to SEPIC, only the load has moved through the top switch and top inductor. Layout is very similar, and demo boards can typically be modified from SEPIC to inverting provided the IC can also regulate on negative feedback voltage like LT3581, LT3757 etc..

Flyback (Figure 13) uses separate windings on a transformer and there is only magnetic coupling between the primary and secondary windings. The current in the primary winding goes to zero at a relative high di/dt; only the energy
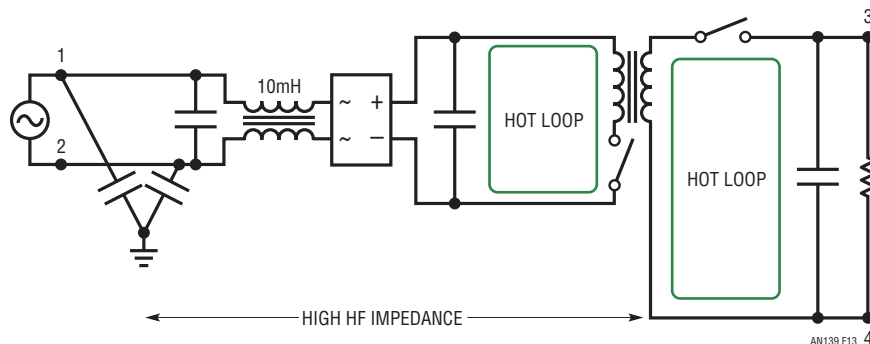
Figure 13. Isolated Offline Flyback

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stored in the leakage inductance and capacitance between windings and on the switch node slows that down. The primary and other transformer windings can be seen as fully switched current. We get two main hot loops as in the buck-boost case (Figure 8). To reduce EMI, in addition to close V_{IN} decoupling for differential mode EMI, common mode chokes are used for the likely dominant common mode EMI in this topology.

Other AC Loops

The hot loop with the main switching energy is the major source of RF energy. However for the operation of the IC and circuit, other AC carrying loops are required. All circuits need a supply for the main switch driver. In the case of the buck, it is often decoupled with the same V_{IN} capacitor as the hot loop. Other ICs use a separate voltage for the drive circuit, often referred to as $INTV_{CC}$ (Figure 14).

Make the $INTV_{CC}$ capacitor PGND and GND loop as small as possible and shield it with a solid plane in the next layer. EMI energy is on the order of 20dB lower than in the main hot loop. Any excessive inductance in the $INTV_{CC}$ loop will deteriorate IC performance. The $INTV_{CC}$ capacitor decouples beside switch drive internal sensitive and wide band circuitry like current sense comparators, reference and error amplifiers which are very often internal supplied from $INTV_{CC}$ too.
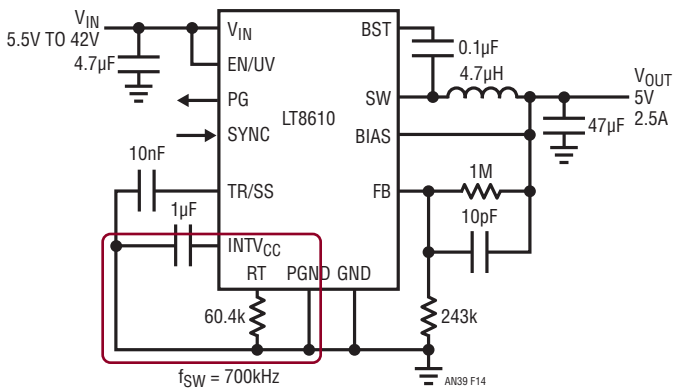

Figure 14

Figure 15 is an FFT of the current in the $INTV_{CC}$ decoupling capacitor (C2 in Figure 17).

Figure 16 is an FFT of the current in the input capacitor (C6 in Figure 17). The RF energy is over 20dB higher than in the $INTV_{CC}$ loop.
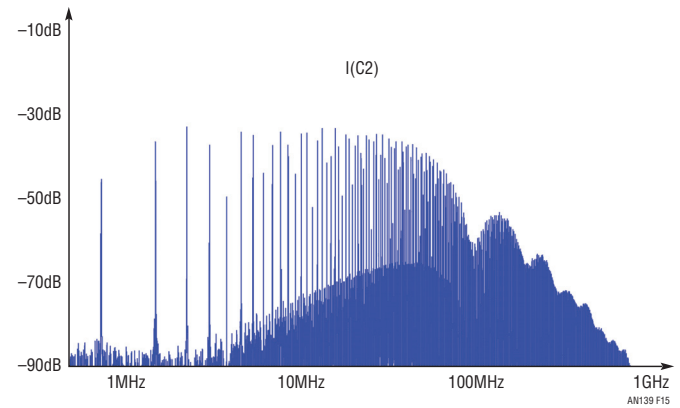
Figure 15
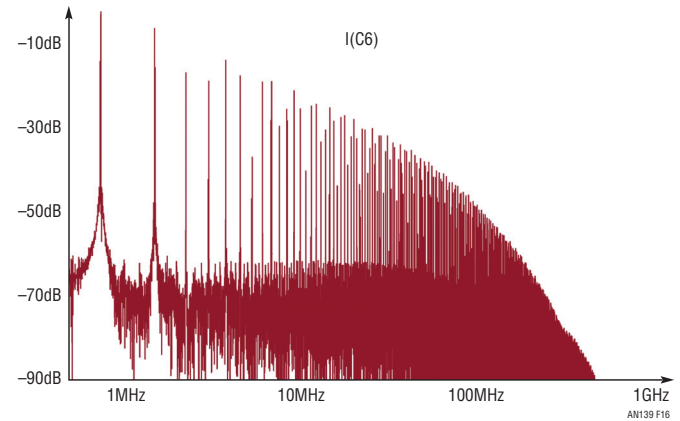
Figure 16

The LT8610 LTspice® circuit (Figure 17) L2 with a high loss of $R_{PAR} = 1\Omega$ is used to decouple the zero impedance LTspice voltage source, V1, from the input capacitor, C6. Note: LTspice switch mode regulator models are developed to simulate the functionality of the IC. Use great caution to extrapolate RF behavior because the models do not account for the internal or the external lump devices or board layout. However, it is nonetheless a great tool to get an approximation of very difficult to measure effects.

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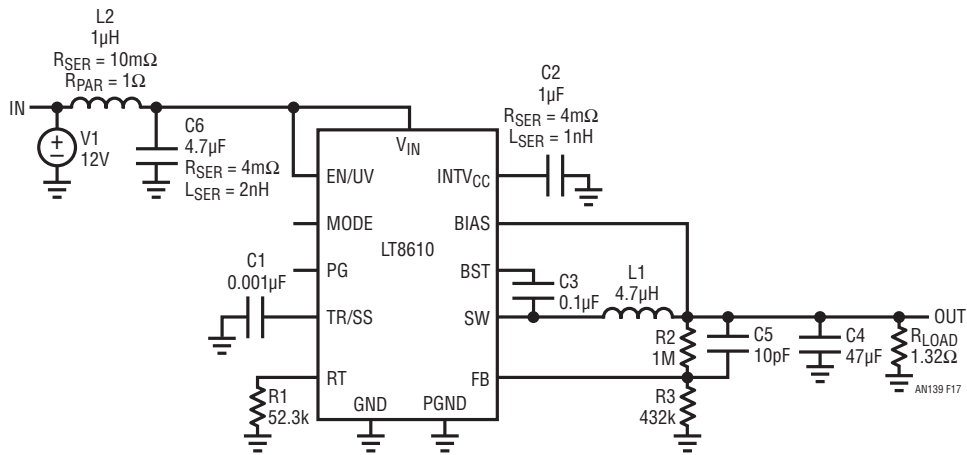
Figure 17

External Switch Drive Signals

After the main hot loop and the $INTV_{CC}$ decoupling loop, the next EMI trouble sources are often external switch drive signals. Even modern MOSFETs have single-to-dual digit nanofarad effective input capacitance. Their drivers often have drive currents in the single amp range, with rise and fall times approaching single digit nanoseconds.

Figure 18 is the FFT of the gate current into Q2 of Figure 19.

The green loop (Figure 19) shows the current path for the bottom gate current. It is supplied from the C1. Make sure that the loop is small. The Q2-source GND to C1 GND connection is most easily accomplished with a solid GND area in the next layer under the component layer.

Figure 18

Figure 19

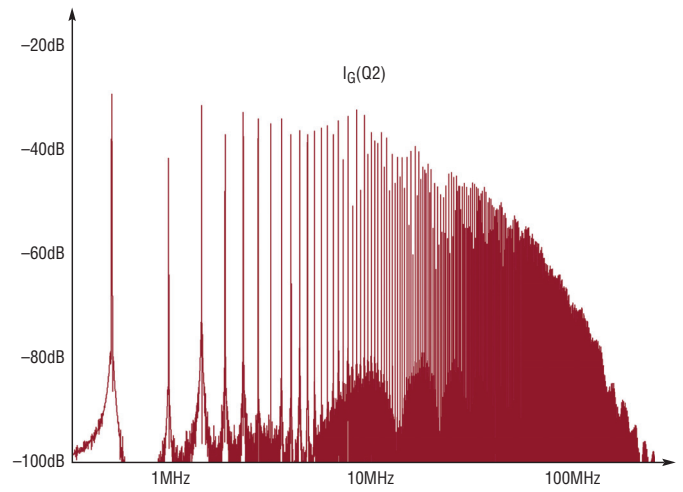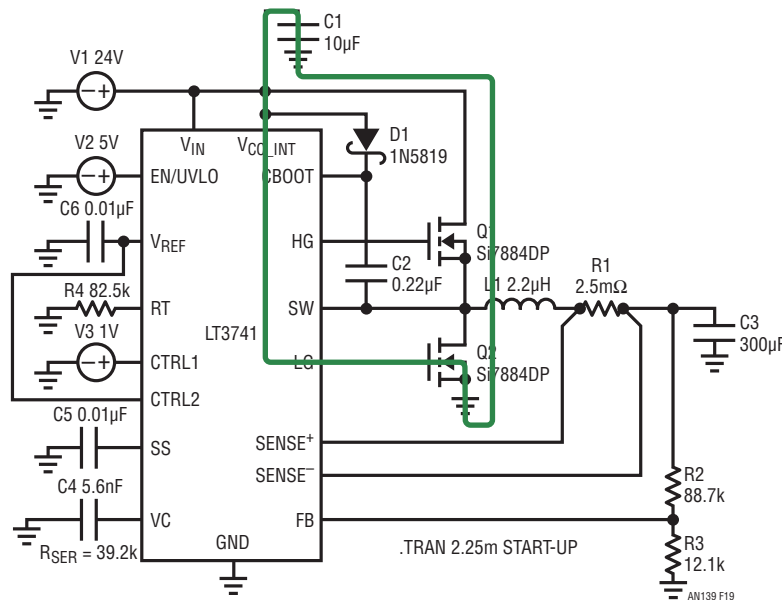The red and green loop (Figure 20) shows the top gate drive current loops. They are supplied from the boost capacitor, C2, and the return is the SW connection to the controller IC. Keep the red loop small and place the traces parallel with only a small gap. If C2 is placed close to the controller IC, the green loop will also become small.

The green loop (Figure 21) shows the recharge of the boost capacitor. If you have already made the above loops small and have placed D1 relatively close, this loop is also small.

Passive antennas, including magnetic antennas or loops, operate reciprocally, and their transmit and receive characteristics are the same. The sense amplifier in Figure 22 looks with high bandwidth at SENSE⁺ and SENSE⁻ on the small shunt voltage over R1 to terminate the top switch on cycle.

Even sub-millivolt noise will result in duty cycle jitter. Coupling to any of the former transmitting loops should be minimal. First, this loop area needs to be small as well,

Figure 20

Figure 21

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Figure 22

so make the gap between the SENSE⁺ and SENSE⁻ line minimal. Then place the loop traces on the other side of a shielding plane from the high current loops mentioned above. If enough layers are available, the SENSE⁺ and SENSE⁻ line can be on top of each other if there are not significant magnetic AC fields parallel to the PC board layers. If SENSE⁺ and SENSE⁻ filtering is used, place the filter close to the controller IC, since R1 is always low impedance and the sense inputs are higher impedance.

The main hot loop in Figure 23 consists of the external MOSFETs Q1 and Q2 and the closest low impedance decoupling capacitor C7. This is the loop with the highest RF energy in a controller solution. For a detailed discussion of controller layout, see LTC Application Note 136 [8].
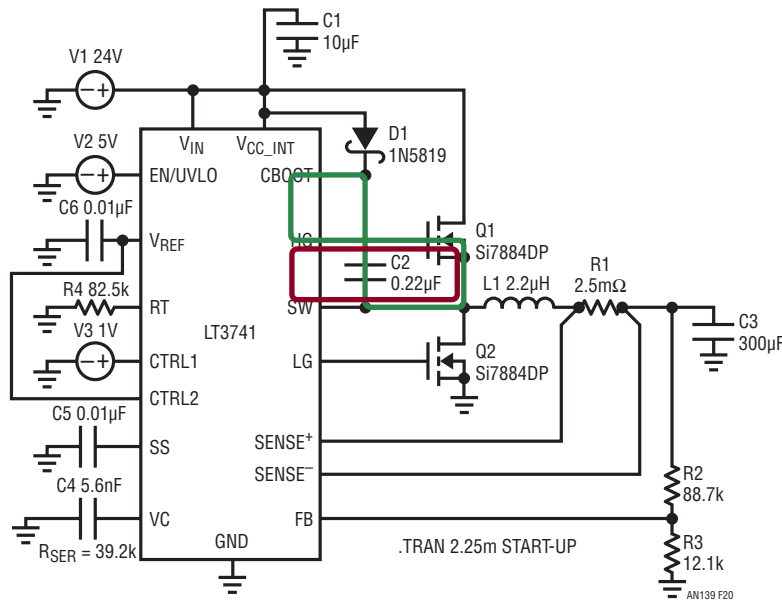

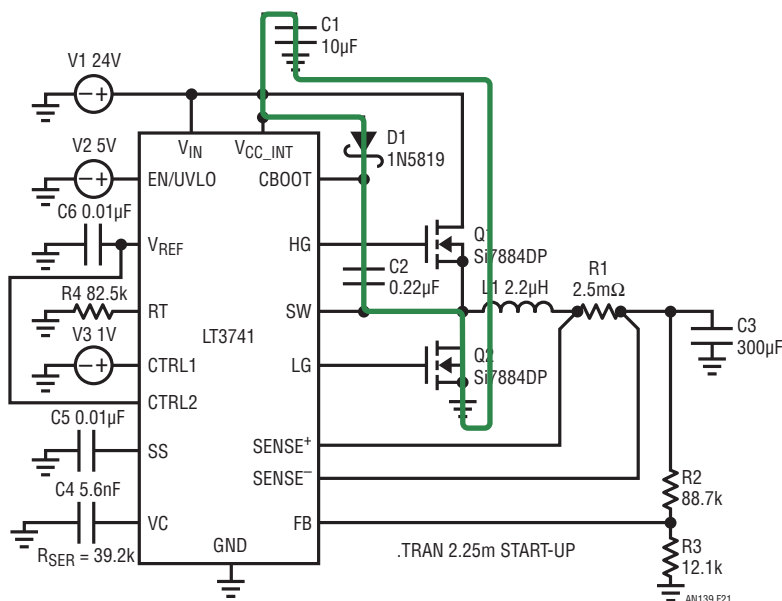**Figure 23**

How Shielding Works

DC magnetic fields go through air, FR4 epoxy dielectric and copper almost undisturbed.

AC magnetic fields are impacted only by the induced currents in a conductor—usually copper or tin, etc. So, absent from ferromagnetic material, we can focus on currents as the only source to alter or attenuate AC magnetic fields in a typical PC board environment.
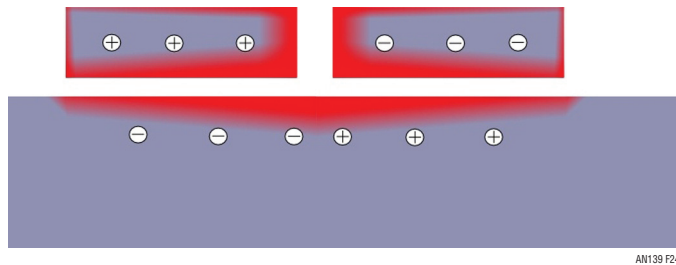

Figure 24

We know from experiments that complete conductive enclosures have a very high attenuation—easily over 100dB—over a very broad frequency range above the AM band. Cookie boxes are a popular example widely used in the R&D community to shield sensitive circuitry, to enable high frequency (HF) measurements. For HF circuits you can buy them in rectangular form which fits easier for PC boards and HF connectors like BNC, N, SMA, etc. For practical purposes HF magnetic fields do not escape closed conductive boxes as long as the wall is thicker than skin depth (see Appendix C).

If we blow up the conductive enclosure like a balloon large enough that it is a flat wall between us and the AC current, the eddy currents in the shield material need to create an AC magnetic field, which is the opposite of the original AC current behind the wall (Figure 24). The eddy currents in the shield will only be enough to cancel any magnetic AC field to the outside of the box or, in our case, in the direction shown at the lower portion of Figure 25.
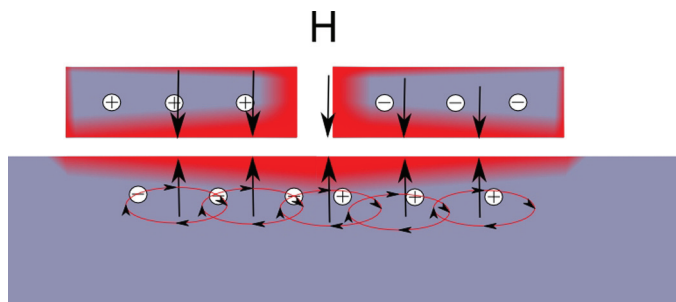

Figure 25

How Effective is a Shield?

If the board copper plane were non-conductive, it would be transparent to any magnetic field like a sheet of paper. The current in the copper is the only source of influence on the magnetic field. The energy available for eddy currents is induced in the shield. Such currents will cancel any magnetic field inside the copper shield and outside in the direction away from the inducing current, within the limits of skin and proximity effect (Figure 26).
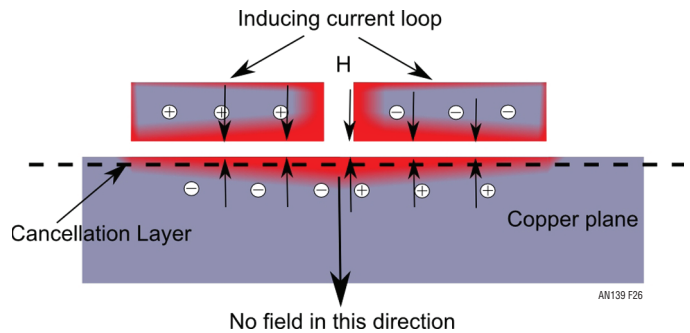

Figure 26

The cancellation layer is the layer that cancels the magnetic field from the inducing currents (i.e., from the hot loop) and induced currents in the shield. So it is sort of midway between the AC currents in opposite directions. The position of the cancellation layer can be estimated by rule of thumb and may help you visualize how the remaining AC fields look.

The closer the cancellation layer is to the induced current layer, the better the cancellation. Inductance is the integral of the magnetic field—the same magnetic field that forms the near field RF antenna. All of these effects are reduced when the cancellation layer is placed closer to the inducing current loop.

What Works Even Better?

If the current returns through the plane, the situation is somewhat different (Figure 27). Now the current in the plane is not only passive induced, but is the active return flow.
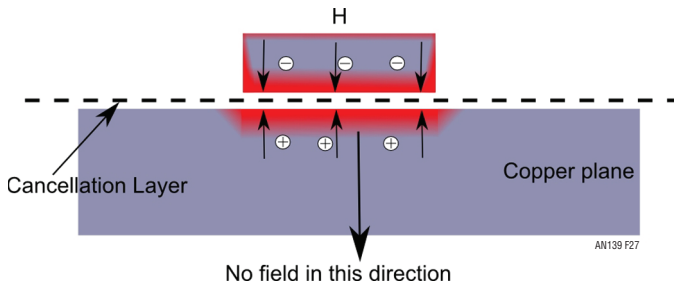

Figure 27

It is forced to be same as the current in the top trace. The benefit is that the cancellation is now closer to the inducing trace, in this case halfway between the top layer current and the returning current in the plane. Because the cancellation layer is now half the distance to the inducing current, it is safe to say that the magnetic field is now lower by at least a factor of two compared to the above situation with passive shielding only.

If you can, let the return current flow in the closest layer. Make its dielectric (isolation) as thin as practical. It is better to have the return current flow in the closest trace to the inducing current with minimum dielectric distance. That's what solid GND planes do by default.

Standard multilayer boards often have much thinner outer layer dielectrics to help reduce EMI.

Line Conducted EMI Measurements

Figure 28 is an LT8611 with some input filtering with a ferrite bead and 4.7µF ceramics line conducted from 30MHz to 400MHz. The display line of −67dBm corresponds to 40dBuV.

Figure 29 shows an LT8610 power supply 13V in, 5V out at 1A out, line conducted up to 900MHz. In Figure 29 an additional wideband amplifier (LNA) in the HF signal chain with 35dB gain is used. So the −100dBm spectrum analyzer display line corresponds to −135dBm, which is the thermal noise floor of a 50Ω system with 10kHz bandwidth at room temperature. The main HF energy needed to be filtered out of the LT8610/LT8611 is below 400MHz. +10dBuV corresponds to $3.16\mu V_{RMS}$.
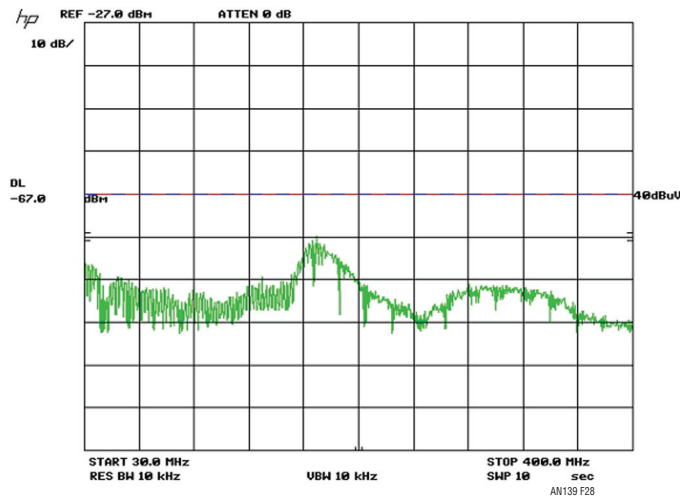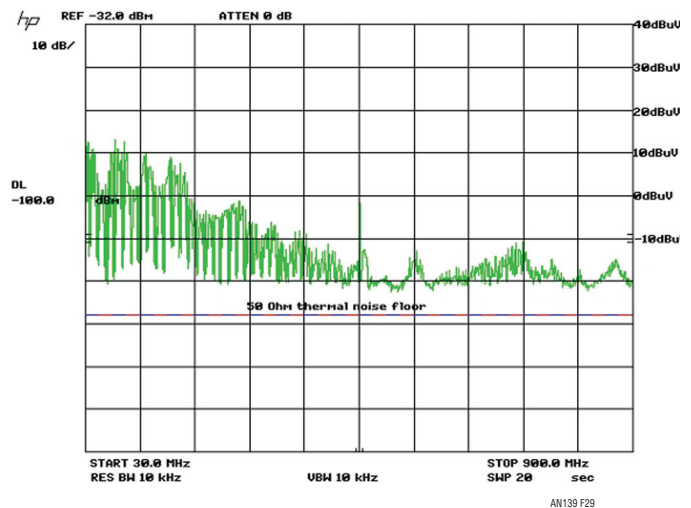

Figure 28

Figure 29

EMI Optimization

First check the layer stack. Use a PC board of four layers or more if possible. The second layer from top is typically only about 200µm distance, and a shield will cancel the hot loop much better than any shield over 1mm away in a dual layer board.

In the hot loop shield, the same current as in the top trace hot loop runs as eddy current. Keep the layer 2 shield solid. Place vias away from the hot loop for connections to GND planes you want to keep quiet. The hot loop shield cancellation currents create HF voltage across the loop, and you do not want to couple it with vias in areas you need quiet. This current decays with distance, but often remains a problem.

The challenge for filtering the input and, if required, the output is mainly in finding an area that is quiet enough. An effective way is to make a complete GND ring around the the power supply unit (PSU) connected with vias.

Filter capacitor GND return should be at the location where the V_{IN} current crosses the ring. There should be still filtering inductance in the direction of the hot loop. That can be a ferrite bead or an inductor.

Beware of magnetic coupling between filter inductors and the main inductor. The transformer action can jeopardize your attenuation. Place filter inductors at a distance from the main inductor.

Use short ceramic capacitors, i.e., 0402, or reverse geometry capacitors because the block capacitors need low equivalent series inductance (ESL) which mainly dictates their impedance. Shorter and thicker ones have lower ESL. See Appendix C for discussion.

Characteristic impedance of the strip lines you create with your V_{IN} traces is in the low single digit Ohm area. The ESL of your block capacitors should be as low as possible. Use reverse geometry capacitors or a stack of 0402 closest to the filter point and larger cases close by. Any trace length significantly increases the few hundred pH inductance your small block capacitors have. Ensure that the routing path of the V_{IN} and the return trace go through the filter capacitor pads. This avoids additional trace inductance.

APPENDIX A

Electromagnetic Field Impedance

Far field impedance is about $377\Omega = 120\pi$ or $29,9792458 \bullet 4 \bullet \pi\Omega$ for the vacuum velocity of light. Any electromagnetic wave far enough from its source (rule of thumb >wavelength/$2 \bullet \pi$) has a 377Ω relation between its magnetic and electric field. Closer to the source, it can be a perfect matched antenna, which transforms its input power source to the right 377Ω electromagnetic field. Or there is significant mismatch and the antenna starts mainly as a magnetic field source or an electric field source.

The magnetic field source has a lower impedance of 377Ω. The electric field source has a higher than 377Ω impedance. The graph Figure 30 shows that, regardless if it starts as an electric or a magnetic field source, the electromagnetic field balances itself to its far field impedance at a distance of:

$$\frac{\lambda}{2 \bullet \pi}$$

Nonisolated switch mode power supply units have primarily magnetic field sources since the impedances of the EMI-relevant loops with high di/dt are much lower than 377Ω unless you have very low current high voltage power supplies. So minimizing the AC magnetic fields on any nonisolated power supply unit will be the key to success.

Any isolated power supply unit will have AC loops with lower than 377Ω, where the same magnetic field minimization as on nonisolated PSUs will be required. However, due to the very nature of isolation, we need higher impedances between the isolation barrier. On the isolation barrier, which is mostly done with a transformer, we try to get $M\Omega$ of isolation. On the isolation barrier, the electric AC field dominates and requires a different strategy. Here we try to get as low capacitive field coupling as possible. So we try to get as much distance as possible and to minimize the size of any conductive material—i.e., copper.
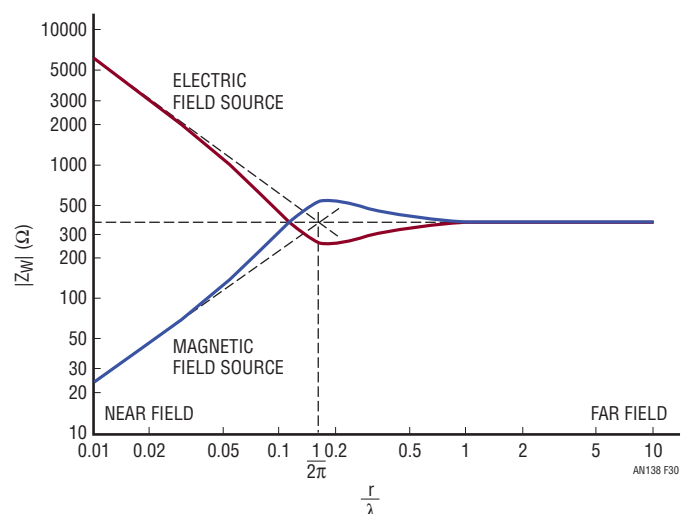

Figure 30 [1]

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APPENDIX B

Dipol Antenna Effect of the Hot Loop

When analyzing what the hot loop does, magnetic dipole antennas give a good clue.

The AC current flows around an area and creates the magnetic field part of a normal dipole antenna (Figure 31).

Magnetic antennas with loop diameters $<<\lambda$ have very low radiation resistance. Range $\mu\Omega$ to $m\Omega$.

$$R_R = 320\pi^4 \left[\frac{NF}{\lambda^2} \right]^{-2}$$

R_R = Radiation resistance in Ohm

F = Area of magnetic loop

N = Number of turns (= 1 in most layouts)

λ = Wavelength

With $\lambda = \dfrac{c}{f}$ and N = 1 for all practical layout loops

$$R_R = \frac{320\pi^4 f^4 F^2}{c^4}$$

c = Speed of light ≈ 300000km/s

f = Frequency

[2]

The radiation resistance is low ($m\Omega$) for typical dimensions of a PC-board power supply unit. Increasing the radiation resistance improves the matching and increases emitted radiation proportional to the radiation resistance. The parameter we can influence the most with layout is the area of the magnetic loop.

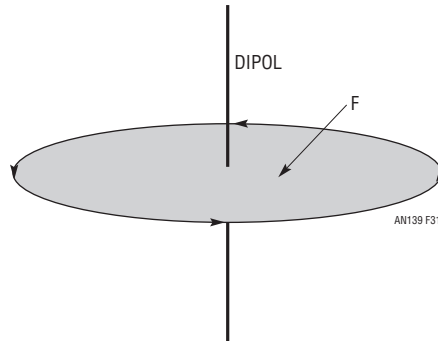The emitted radiation is proportional to the square of this area.

DIPOL

F

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Figure 31

APPENDIX C

Skin Effect

To understand the effect of shielding, we need to dig a bit into the electromagnetic properties of the materials used. Electric current, whether or not it is tied to a conductor, will always flow on the path of least impedance. For higher frequencies this will be the path of least inductance. This means the current will also flow on the pass of its lowest losses. Electric conduction material minimizes any internal magnetic AC fields by creating eddy currents that oppose AC fields internal of the conducting material. Viewed from the outside, this looks like skin effect, because the current density is forced to the outside of the conductor.

For the first model example, assume that the current, I, flows uniformly through a cylinder. This is a typical case for DC current.

If the current is DC, then it will look like that in Figure 32. Since I is constant, the resulting H is constant and I_W is zero. In the case that I has an AC content, which means there is di/dt, then the resulting magnetic field H changes. The changing magnetic field H creates induction voltage and since we are inside a conductor, the induction voltage creates an induced current I_W, often referred to as eddy current. Eddy currents will create losses. If we assume that nature minimizes these losses, the only way to minimize this is to change the original uniform current distribution and let all current flow only at the surface of the cylinder.
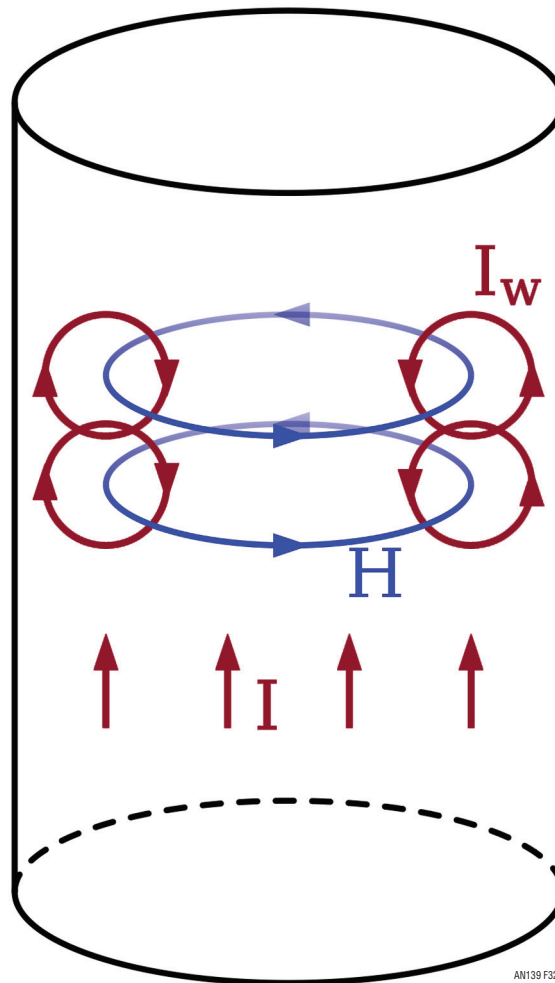

Figure 32 [3]

Now the magnetic field H is only at the surface of the cylinder. This way the eddy currents I_W are moved to the outside and the return path is cut off, omitting most of the eddy currents and its losses.

We can think of forces (Figure 33) that move the current density to the outside of a conductor

This is called skin effect. The wall thickness of a pipe, which would give the same resistance at DC as a full cylinder wire at AC, is called skin depth (Figure 34). Since the forces moving the current density to the outside are a result of the induction law, the skin depth goes down with rising frequency. Skin depth goes down with rising conductivity and goes down with increasing magnetic permeability.
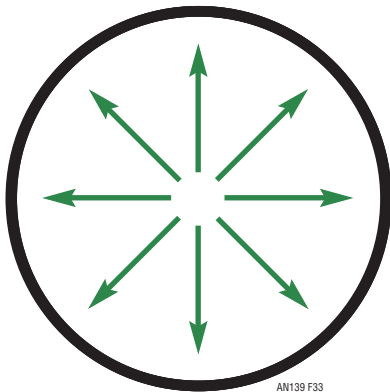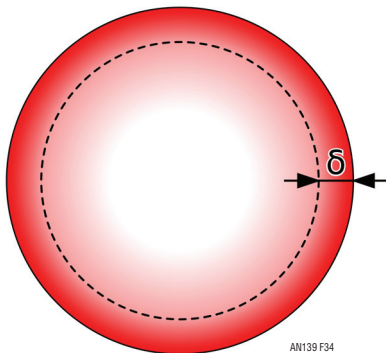
We can use another example.

Assume we have a pipe wall conducting all the current on its surface and it is the large diameter shown in Figure 35. Total inductance of a given length of this wire (1 meter or 1 foot whatever is most suitable) is the complete volume integral of its magnetic field up to infinity or whatever physical size you assume for the universe.

Now we shrink the pipe diameter to the smaller inner one shown in Figure 35. The magnetic field is still the same as the old larger pipe to infinity. However, we now have an additional magnetic field volume between the new small pipe and the prior larger pipe diameter. So the total integral of the magnetic field is now larger. This shows that the
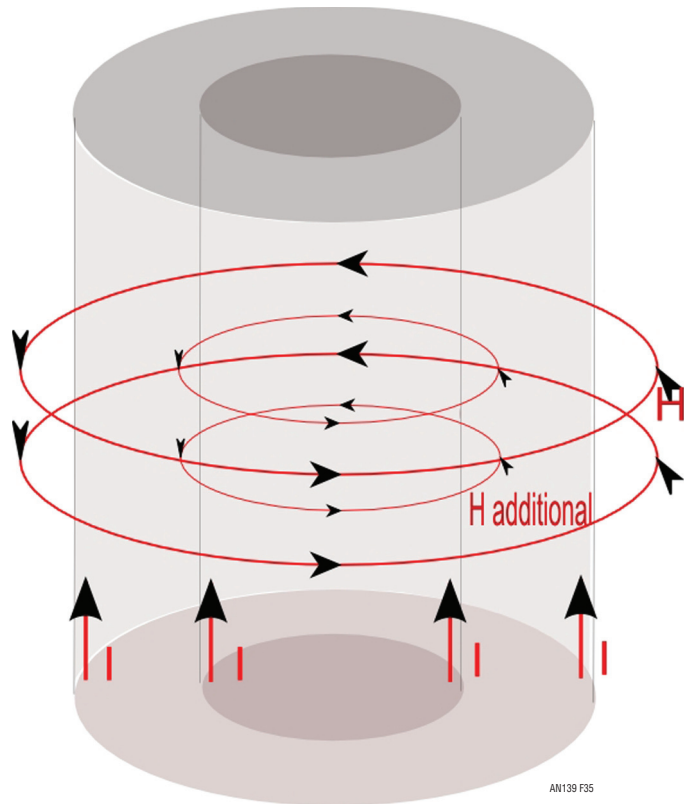

Figure 33 [3]

Figure 34 [3]

Figure 35

inductance is larger for a thinner conductor of a given length. Or we can assume that the inductance increases as more of the current moves from the outer wall to the center of the wire. If we apply a voltage over both ends of the wire, the resulting current distribution is determined by the impedance. Since the inductance in the center is higher, most current density moves to the surface. The most extreme case of skin effect we have in conductors with zero resistance is shown in superconductors. There quantum effects prevent all current from being bound to an outer layer of zero thickness. The thickness where most current is concentrated on superconductors is called London depth.

A graph of skin depth over frequency for some materials is shown in Figure 36.

We see that copper on typical PC-board material is affected from skin effect starting from the 5MHz to 50MHz range. And we can see that even highly doped silicon at the thickness typically used on ICs is only affected in the terahertz region. Copper (Cu) and gold (Au) are close together. Material with high magnetic permeability (Fe-Ni) has low skin depth even at audio frequencies. For this reason it is used to shield audio transformers.

AC current through a good conductor will push current density to the outside. The current will flow where the impedance, dominated by inductance, is lowest. With regard to the impact of skin effect on layout and components, we can derive simple guidelines. Better is short and thick or wide. Reverse geometry capacitors have lower ESL because they are shorter and thicker.
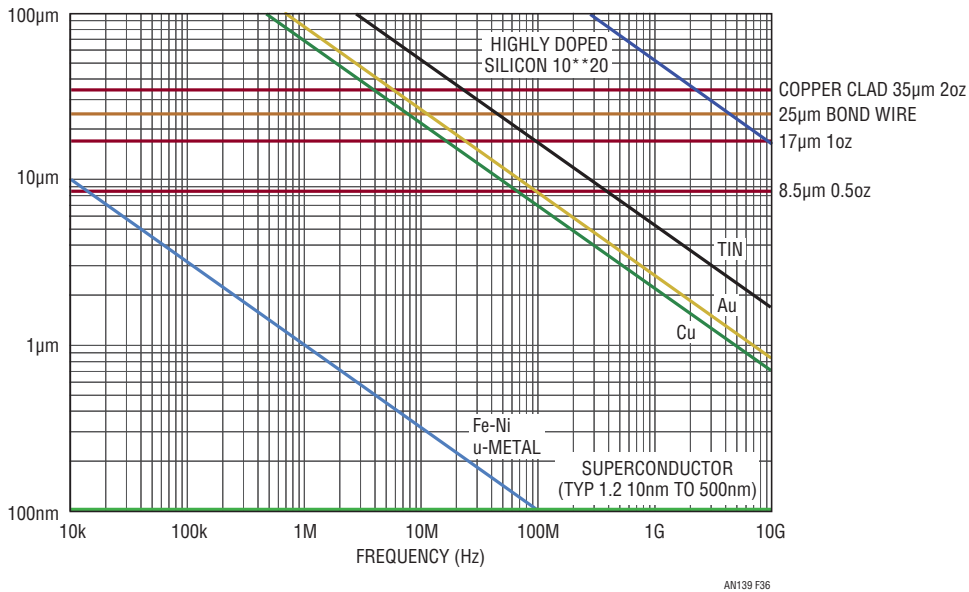

Figure 36

APPENDIX D

Proximity Effect

If we have opposite direction current the same forces will attract the currents to each other (Figure 37). This is called Proximity effect.

If we have a PC board trace on a single layer board, the high frequency current density will look like that in Figure 38.

If we add a solid plane in a second layer eddy currents will be generated in the plane, creating in total a mirror image of the current in the top trace, as shown in Figure 39.

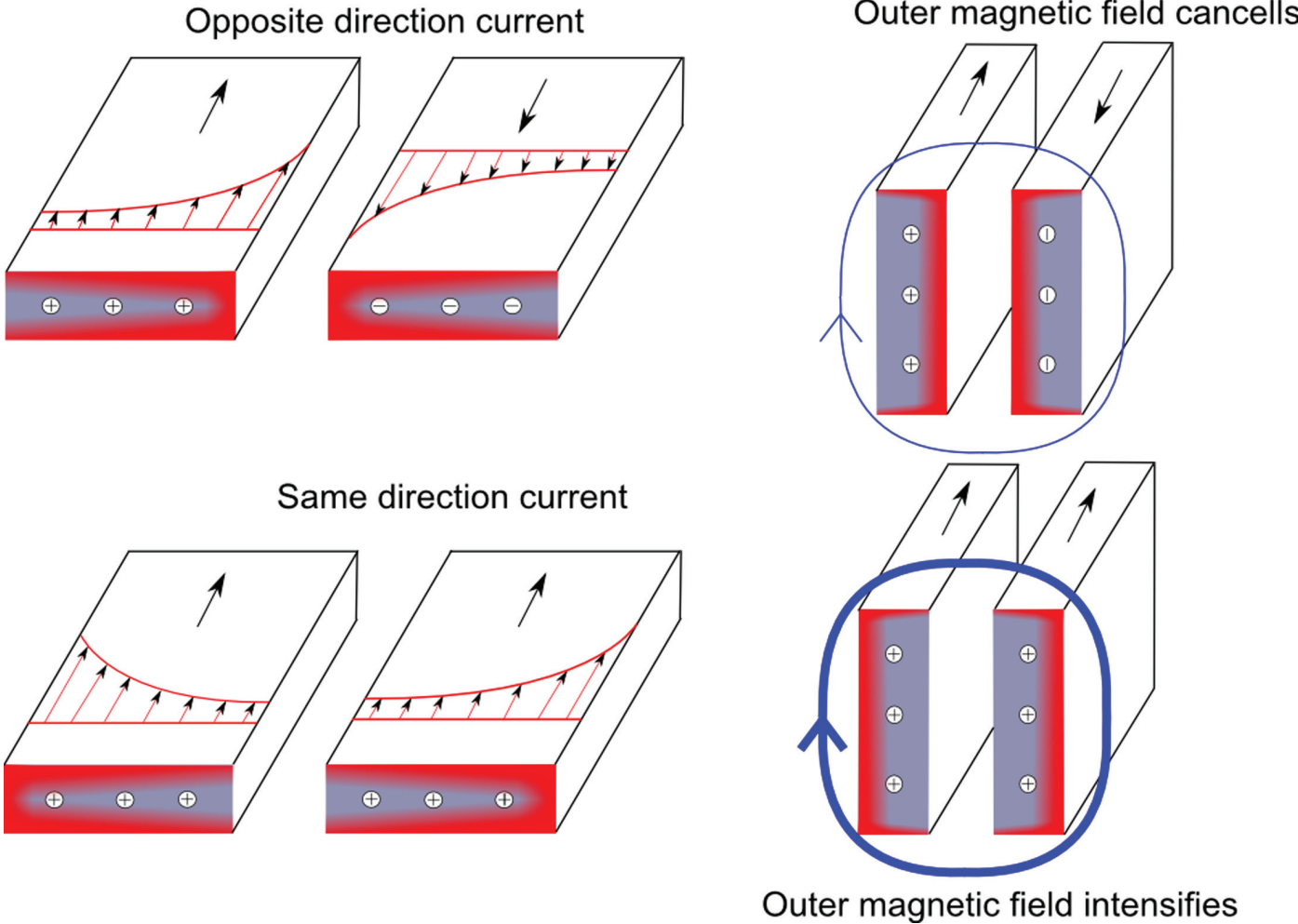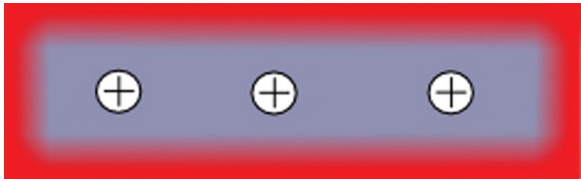On multilayer boards, a plane will shield the AC currents (Figure 40).

Opposite direction current

Same direction current

Outer magnetic field cancells

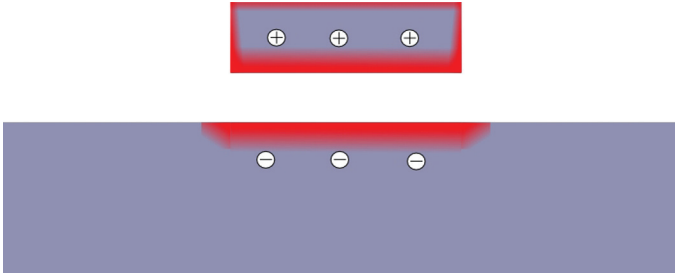Outer magnetic field intensifies

AN139 F37

Figure 37 [4]

Figure 38

Figure 39

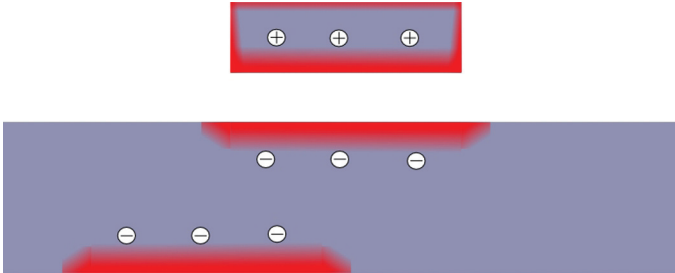**Figure 40**

APPENDIX E

EMI Measurements

Most power supplies are relatively small compared to the wavelength of the relevant EMI frequencies they produce and are measured against. If prudently designed, in current power supply technology you will find most energy below 500MHz. EMI standards must be general and apply to equipment of variable size. Those can be in the order of the wavelength of interest. So they call for line conducted measurements up to 30MHz and radiated measurements above 30MHz. However reliable radiated measurements require large anechoic chambers. Their hourly prices are high and availability is scarce. Free field is too noisy, requires a large and difficult setup and is weather dependent.

For power supply optimization work, a reasonable approach is to make line-conducted precompliance measurements up to the end of the spectrum required for radiated measurements. Since the power supply dimensions are still small against the wavelength of interest, we can assume that most energy will find its way through the V_{IN} and V_{OUT} wires, where we can measure them line conducted.

The setup is quite simple. We need a LISN (line impedance stabilizer network) or AN (artificial network), an input supply, a load and a measurement receiver.

The purpose of the LISN (AN) is to isolate the voltage source V1 from the power supply (or DUT device under test) V_{IN} (V_{DUT}^+, V_{DUT}).

This is an example of a non-symmetric LISN often used in automotive (Figure 41). Such a simple circuit can be made with L1 as air coil, or an inductor with losses can be used. Some standards specify different core types in series and a special winding scheme. However, the main purpose is to create a wideband high impedance against 50Ω for L1.
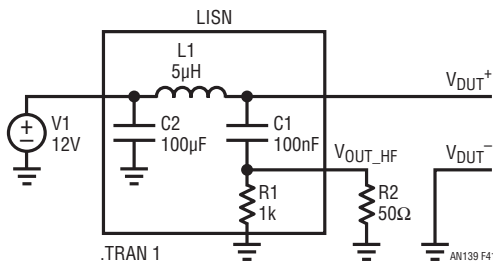
Other than the wire length inside for C1, R1 and the V_{OUT_HF} and L1 impedance, nothing limits the usable upper frequency range. So buy one or build your own. Resonances of L1 can be damped with a resistor over a part of L1 windings.

Any dedicated EMI receiver can be used, but a spectrum analyzer will usually do for precompliance work. Make sure that you use the AC-coupled input, since it provides a second barrier against blowing up your expensive mixer inside the analyzer.

From the EMI lab experts, you can expect a lengthy discussion about the required detection method from the used EMI standard, including peak, quasi peak, average with relative accurate time constants required for them. You can shortcut this discussion when your power supply operates fixed frequency in the load area of interest. At fixed frequency, only harmonics with a distance of the switching frequency can be created, a frequency comb. If the switching frequency is above the required resolution bandwidth (mostly 9kHz up to 30MHz and 120kHz above 30MHz), peak, quasi peak and average methods will yield the same results, so you can use whatever your receiver provides. Some standards allow for the 9kHz to use a 10kHz bandwidth and for 120kHz 100kHz. The error for a fixed frequency switching PSU, which operates well above 100kHz, is not relevant for our precompliance task.

If your system includes a processor it can produce currents with large fluctuations with frequency contents well within the above resolution bandwidth.

Then you need to refrain to the filter method your standard requires.

If you see components that are a fraction of the switching frequency or cannot be divided by integers to the switching frequency, check the switch node with an oscilloscope. In the time domain, you will likely see pulse skipping or subharmonic oscillation. Check the source for this behavior before proceeding further. Do not forget to unhook your scope probe because you get different results with the additional introduced probe antenna if you do EMI measurements with a probe attached.

Figure 41

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