"Measuring, Suppressing and Filtering Common Mode Emissions in Switched Mode Power Supplies"

Author: Mark Nave Date: 1989

Summary: *Introduction to characterizing CM noise and suppressing it. HFPC Paper.*

MEASURING, SUPPRESSING, AND FILTERING COMMON MODE EMISSIONS IN SWITCHED-MODE POWER SUPPLIES

Mark J. Nave
EMC Services, Inc.
11833 93rd Avenue North
Seminole, FL 34642

ABSTRACT

This paper reviews methods and tools for measuring, diagnosing, suppressing, and filtering common mode emissions. Basic theory for predicting common mode emissions is reviewed, and the dependence of emissions on voltage waveform rise times is evaluated. Common mode emissions are reviewed from the EMI perspective with emphasis on suppression by design. Design techniques are reviewed which limit the generation of EMI. Filtering techniques are reviewed, and implications on real components' effect on filtering are discussed. A novel test technique to diagnose common mode core saturation and to determine sources is demonstrated.

PREDICTION OF COMMON MODE NOISE

It is very useful to have a technique to be able to predict the noise generated by the charging and discharging of parasitic capacitances to chassis. The classic example of these parasitics is the switching transistor insulator capacitance to heatsink. From this prediction the designer can determine whether or not corrective action may be necessary, and incorporate it at the design level where the options are both greater in number and cheaper. Figure 1 illustrates the aforementioned transformer voltage, switch voltage, and common mode current waveforms. The voltage waveforms of the transformer and the switch are drawn with exaggerated rise and fall times for clarity. The relationship between V_{ce} of the switch and the common mode current $Icm = \frac{C_p \bullet dV_{cg}}{dt} = \frac{2 \bullet Vs}{\tau_r}$ is apparent by inspection of Figure 1.

To get a feel for the numerical values, consider a SMPS with a 100 kHz switching rate and a 100 ns rise time. Let the insulator capacitance by 100 pF, and V_s = 100 volts, so V_{ce} = 200 Volts. Plugging in values, I_{cm} = 200 mA, as depicted in Figure 2. to derive the frequency domain relationships (see Appendix A), only the magnitude of the current pulse is considered, and the current waveform is considered to be a periodic, symmetrical, rectangular pulse as in Figure 2.[1] Under these assumptions, the envelope of the spectrum can be readily solved. To solve for the envelope pedestal in narrowband units:

$$I_{cm}(f) = 2 \bullet I_{cm}(t) \frac{\tau'}{T'} = 2\,(.2)\,\frac{100\text{ns}}{5\mu\text{s}} = 78\ \text{dB}\mu\text{A} \qquad (1)$$

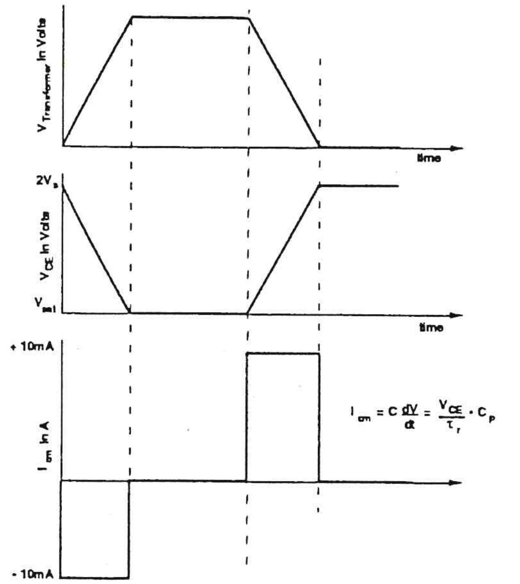

Figure 1: Switch Collector/Common Mode Current Waveforms

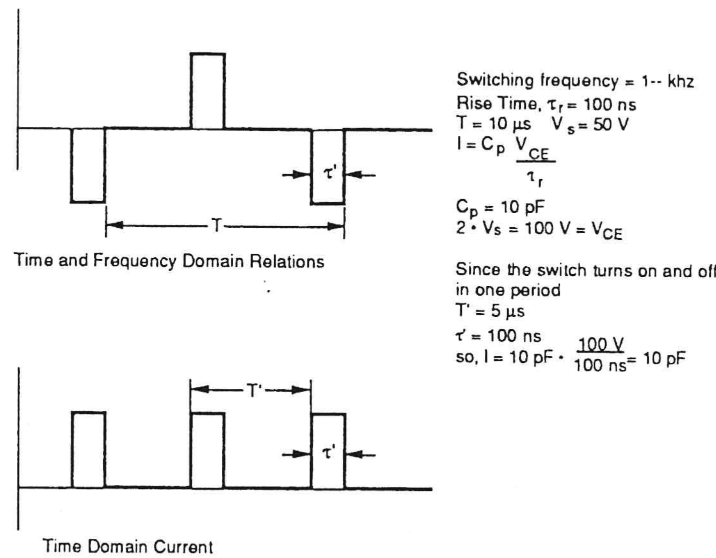

Figure 2: Common Mode Waveform

Where A is the amplitude of the current, τ' is the pulse width of the common mode current (100 ns), T' is the period of the magnitude of the common mode current, 5µs. These values are labeled in Figure 2. The corner frequency is $\frac{1}{\pi\tau'}$ = 3.18 MHz.

After the corner frequency, the frequency domain spectrum decreases at 20 dB/decade. Figure 3 shows the frequency domain prediction.

If we compare the predicted frequency domain spectrum to specification limits, say MIL-STD-461, we find the common mode emissions exceed the specification limits from roughly 0.3 MHz to 50 MHz! Figure 4 shows the comparison of the predicted common mode emissions to MIL-STD-461.

If comparison to commercial limits is desired, the pedestal current may be multiplied by 25Ω (or add 28 dBΩ). Doing this gives 106 dBµV, orders of magnitude above all commercial limits.

SUMMARY RELATIONS

The coupling mechanism has been described, and the solution developed, yet some simplification is still possible. Indeed, to be practical, simplification is necessary. By substituting the derived values for current directly into the formula for the pedestal, the following relations can be found: [2]

Pedestal = 4 • C_p • V_{ce} • F_o (current)

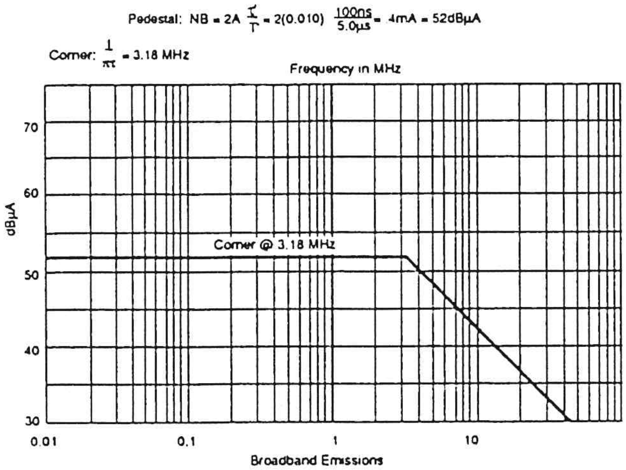

Figure 3: Current Spectrum Calculation

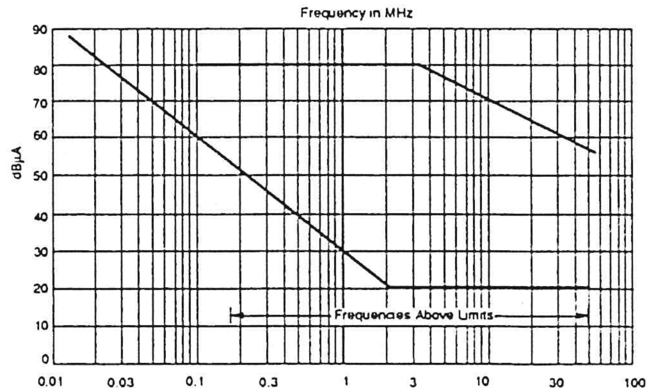

Figure 4: Limit for CE03 Narrowband Emissions

or Pedestal = $100 \bullet C_p \bullet dV \bullet F_o$ (voltage)

$$\text{Corner frequency} = \frac{1}{\pi\tau_r}, \text{ smps}$$

Where: C_p is the insulator capacitance V_{ce} is twice the input voltage, F_o is the switching frequency, and τ_r is the rise time of the switching transistor.

THE EFFECT OF RISE TIME ON COMMON MODE EMISSIONS

It is often stated in literature that one means of controlling common mode emissions is to slow down the rise-time of the switching transistor. It has been shown that common mode emissions are independent of rise time. The pedestal region of the envelope of the common mode emissions in frequency domain has been shown to be

$$V_{cm}(j\omega) = 100 \bullet V_m \bullet C_p \bullet f_o$$

Note that the collector voltage rise time is not an explicit, nor an implicit variable of this expression. To understand this apparent anomaly, consider the time-domain wave form of the common mode emissions. In converting this time-domain wave form to a frequency-domain wave form, the envelope is expressed as the product of the maxima of three terms: the pedestal term, the bandwidth term, and the phase term [1], or

Pedestal•Bandwidth•Phase =

$$2A\delta \bullet 1 \bullet 2 = 4 \bullet V_{cm}(t) \bullet \frac{\tau'}{T}$$

$V_{cm}(t)$ is the amplitude of the common mode time-domain voltage

τ' is the pulse width of the common mode pulse ($=\tau_r$)

T is the period

The amplitude, A, of the voltage as measured in time-domain at one LISN during the transition is:

$$V_{cm}(t) = 25 \bullet C_p \bullet \frac{V_m}{\tau_r}$$

where,

V_m is the peak collector voltage,

C_p is the collector to chassis parasitic capacitance, and

τ' is the common mode wave form pulse width.

The pulse width, τ', is equal to the collector voltage rise time, τ_r substituting, we have:

$$4A \frac{\tau'}{T} = 4\,(25 \bullet C_p \bullet \frac{V_m}{\tau_r}) \frac{\tau_r}{T} = 100 \bullet C_p \bullet V_m \bullet f_o$$

Note that the rise time of the collector voltage wave form divides out of the expression for the pedestal. Hence, the pedestal amplitude, in frequency-domain, is independent of the collector voltage rise time.

The pedestal is proportional to the product of the Amplitude, A, and the pulse width, $\tau' = \tau_r$. It happens that as the rise time decreases, the amplitude decreases, but the pulse width increases inversely. Therefore, the product $A \bullet \tau'$ stays constant. There are, therefore, only 3 methods of source suppression:

1) Reduce C_p, the parasitic capacitance (a function of the insulator chosen),
2) Reduce V_m, the maximum collector voltage (a function of topology and duty cycle), and
3) Increase the period/reduce the switching frequency (this may result in larger magnetics components).

EMI SUPPRESSION VS SWITCHING FREQUENCY

It has been shown that the CM EMI generated is proportional to switching frequency, i.e., increasing at 20 dB/decade with switching frequency. Since the FCC and VDE conducted emissions limits are piece-wise flat, we may properly conclude that as the switching frequency increases, the required attenuation increases. To keep the generated EMI constant, the parasitic capacitance to chassis must be reduced inversely to switching frequency. Parasitic capacitance to chassis is typically a result of heatsinking. Less parasitic capacitance to chassis implies less coupling to chassis, which in turn implies less heatsinking to chassis. However, as the switching frequency goes up, the losses go up, and more heatsinking is required. Hence the parasitic capacitance to chassis will necessarily increase, and the burden for EMI compliance will fall onto the power line filter. The resonant converter, with its low switching losses, may be the topology to break this catch 22.

SOURCE SUPPRESSING CM NOISE

Figure 5 shows the primary side of a converter with the collector-heat sink capacitance. The most obvious method of reducing the capacitance to chassis is to float the heatsink. This technique effectively puts a small value capacitance in series with the larger insulator capacitance. With a physically small heatsink, this technique can be effective. An additional problem that may arise is radiation from the heatsink. The isolated heatsink acts in a manner analogous to a 1/4 wave antenna above a ground plane. A method for preventing this radiation is to tie the heatsink to a fixed potential; positive bulk or bulk return. The result of tying the heatsink to either bulk rail is to effectively put the insulator capacitance across the transformer or across the transistor. This extra capacitance will slow the rise or fall time. The effective schematic for these implementations is shown in Figure 5.

Another method of suppressing the generation of EMI at the source for the switching transistor is to use a "screened" insulator. This insulator is actually two insulators sandwiching a copper conductor forming a "screen." When the screen is connected back to bulk return, an equivalent schematic to Figure 6 results, and the path for common mode emissions is back through bulk return, effectively "shorting out" the LISNs.

Other sources of CM emissions are: primary-to-secondary capacitance of the converter transformer, commutating and rectifying diodes, base drive isolation transformers, and radiation coupling to the filters.[3]

LIMITATIONS OF FILTERING CM NOISE

Many engineers confuse the ideal L-C filter with the LC filter used in a common mode choke. The first problem is that the attenuation of an ideal filter increases at 40 dB/decade. The common mode filter attenuation does not maintain a 40 dB/decade increase. The common mode choke is usually designed to have the highest permeability core available with the fewest turns possible.[3] Typical high permeability cores have a perm. vs frequency characteristic that rolls off above 80 kHz. This means that the attenuation of the CM filter increases at 40 dB/decade until 80 kHz, after which the attenuation increases at 20 dB/decade. Because of safety requirements the "Y" capacitors are constructed with a high lead inductance. Above the self-resonance of the capacitor the filter attenuation decreases at 20 dB/decade. Because of this limitation of the permeability-bandwidth product of the core, there is a limitation to the filtering capabilities of the core. [4] Since the peak impedance of the core is reached at the corner frequency of the permeability of the core (80 kHz for Magnetics Inc. H material), this corner frequency is the optimal switching frequency of the converter to optimally utilize the filter.

SEPARATION OF CM FROM DM

There are three basic applications for separating common mode (CM) and differential mode (DM) noise. The first two reasons find application in filter design. Power line filter

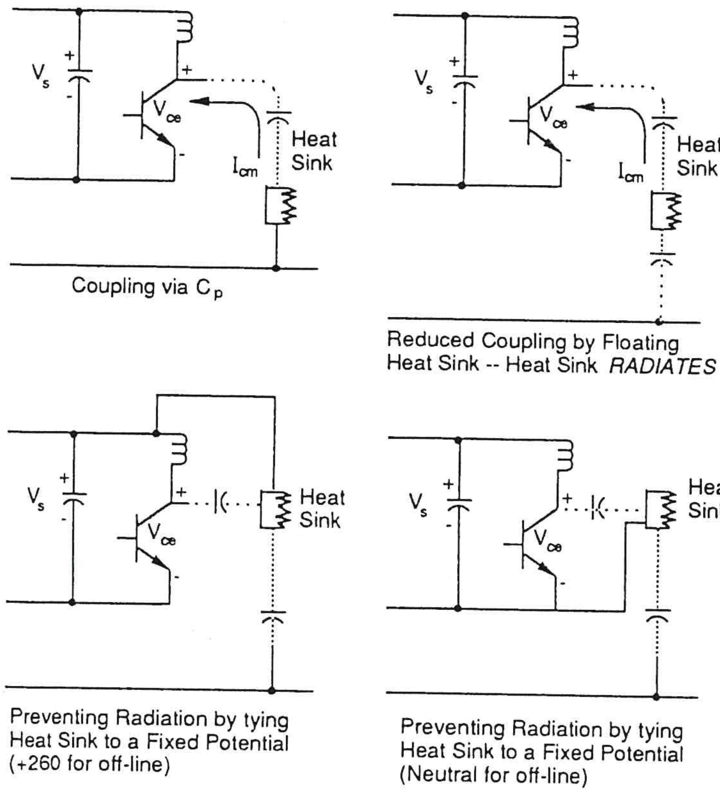

Figure 5: Reducing Coupling: Cp

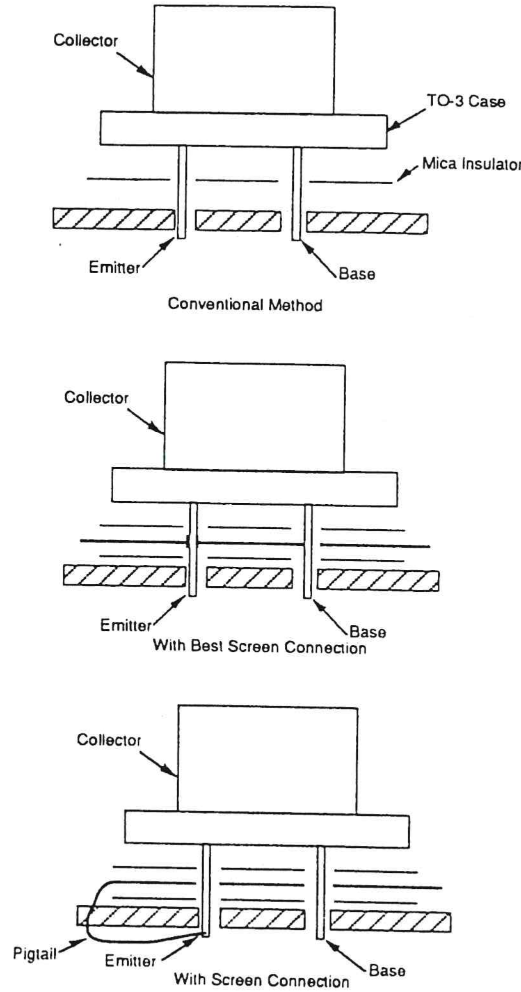

Figure 6: TO-3 Case Mounting Methods with a Screen

design is customarily done by segregation CM and DM noise and filtering them in separate stages. As a rough approximation, CM and DM stages are independent. Therefore, it would be very convenient to design a common mode filter for the common mode noise, and a differential mode filter for differential mode noise. The problem is that measuring the EMI on a LISN gives the total noise, the vector sum of the CM and DM parts. LISN MATE enables the filter designer to measure only CM emissions while designing the CM stage, and subsequently to measure only the DM emissions while designing the DM stage.[5]

The second application is also in the area of filter design, and is a method of measuring when a CM core is in saturation. There is currently no other method of performing this task.

The third application is in trouble shooting. Common mode and differential mode noise are generated by different mechanisms. The first clue in determining the source of the noise is the propagation mode, CM or DM.

Hence, there are at least 3 major reasons to separate CM and DM noise. Without this capability, the engineer's task is much more difficult and time consuming.

MEASURING CM CORE SATURATION

Often, a difficult problem to detect is saturation of the CM core (total or partial). A simple test can show how much the CM filter attenuation is effected by a decrease in inductance due to bias currents. The test requires an oscilloscope and a LISN MATE (DMRN). First, the line voltage is monitored with the oscilloscope. This is the input for channel "A" of the oscilloscope. Set the oscilloscope time base to 2ms/div. The oscilloscope is triggered on channel "A." Line currents will flow during the peak portion of the AC voltage wave form. This is the region where a decrease in filter effectiveness is anticipated. The inputs of the LISN MATE are connected to the LISNs. The output of the LISN MATE is terminated in 50 Ohm, and connected to channel "B" of the oscilloscope. When the CM choke is operated in the linear region the emissions monitored on channel "B" should increase by no more than 6-10 dB during the input current surge. If the CM choke is driven into saturation, the emissions during the input surge will become increasingly large. If the CM choke is driven into hard saturation, the emissions will be at virtually the same level as without a filter, easily 40 dB higher. This test method underscores the importance of having a known, calibrated, high degree of DM rejection. If the DM noise is not attenuated to a level lower than the lowest CM signal, then the engineer is uncertain which mode is being measured. One method of defining the "uncertainty region" is to plot the composite (CM and DM) emissions, and plot another line at a level equal to the composite minus the differential mode rejection of the LISN MATE (>50 dB). Signals in the uncertainty region may be either noise type. Clearly then, the higher the differential mode rejection, the better! [5]

APPENDIX A: ENVELOPE APPROXIMATION OF A FOURIER TRANSFORM

This appendix summarizes the basic results of "Prediction of Pulse Spectral Level," by Harold L. Rehkopf, prepared for the Fourth National IRE Symposium on RFI, 28-29 June 1962.

The envelope of the Fourier transform of a periodic trapezoidal pulse is derived by finding a pedestal, the first corner frequency, f_1, and the second corner frequency, f_2. No energy exists beneath the fundamental, f_0, and 99% of the energy is contained between the fundamental and the second corner frequency.

The pedestal is determined by the peak amplitude of the time domain signal (volts or amps), and the 50% height duty cycle.

$$\text{Pedestal} = 2A\frac{\tau}{T} \qquad (A)$$

Where A is the peak amplitude, t is the 50% height pulse width, and T is the pulse period.

The first corner is also related to the 50% height pulse width:

$$f_1 = \frac{1}{\pi\tau} \tag{B}$$

The second corner frequency is related only to the rise time of the pulse, τ_r.

$$f_2 = \frac{1}{\pi\tau_r} \tag{C}$$

Figure A-1 illustrates the salient features of the time domain waveform and the frequency domain envelope.

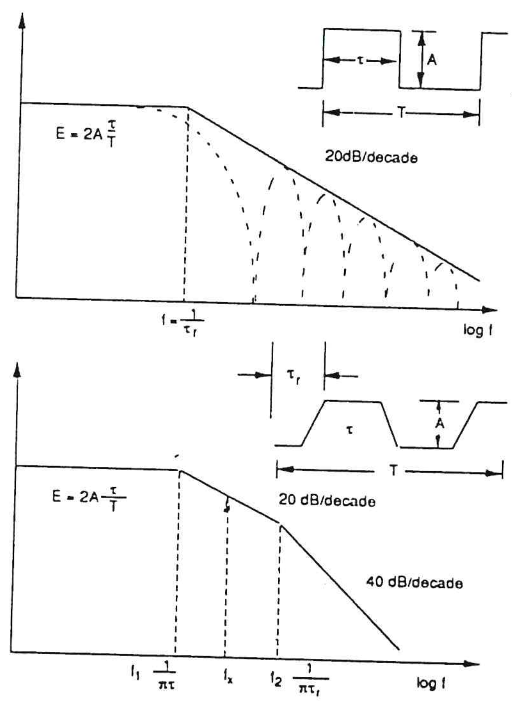

Figure A-1

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