Application Note - Interpoint

Crane Aerospace & Electronics Power Solutions

EMI-Conducted Interference

CRANE AEROSPACE & ELECTRONICS

EMI-Conducted Interference

APPLICATION NOTE

Although the concepts stated are universal, this application note was written specifically for Interpoint products.

Switching power converters are natural generators of input narrowband spectral noise currents having components at the fundamental switching frequency and higher order harmonics. Most Interpoint Power Converters have partial input line filters consisting of a single LC section which reduce the differential spectral noise components to about 1 mA RMS or less. This is adequate for many applications, but where lower noise currents or compliance to MIL-STD or other EMI Specifications is required, an external power line filter will generally be required.

Interpoint power converters are fabricated on ceramic substrates which are back-metallized and soldered to the metal package mounting base. The ceramic material is usually aluminum oxide having a dielectric constant of about 8. The substrate thickness is in the range of 0.5 mm to 1.0 mm. The power train components and traces will all have parasitic capacitances to the mounting base, the magnitude depending on the thickness and area occupied on the substrate. The high dV/dT associated with the switching power conversion causes Common Mode Currents, $I_{cm} = C(dV/dT)$, to be pumped in and out of the case corresponding to each transition of the FET power switch and rectifiers. This is shown in Figure 1, where the narrowband noise Spectral behavior and Power Train Functional Schematic are shown for a Single Ended Forward Power Converter, the most common Interpoint topology. These parasitic Common Mode currents, which belong in the return line and not the case, complicate and add another dimension to the problem of EMI Suppression.

Referring to Figure 1, the output load is seen looking into the transformer primary, and causes a rectangular pulse of current to flow in the primary when the FET is turned on. Ideally, this pulse of current should enter at the positive line and entirely return through the negative line to the 28V source. If this occurs, the noise will be entirely differential and can be attenuated with additional LC filter sections like the partial input filter shown at the circuit input. This is a Differential Low Pass Filter section which works for noise currents in one line with respect to the other. However, since we have Common Mode components in the case which belong in the return line, we have a component of noise current which is in phase in both input lines. This repre-

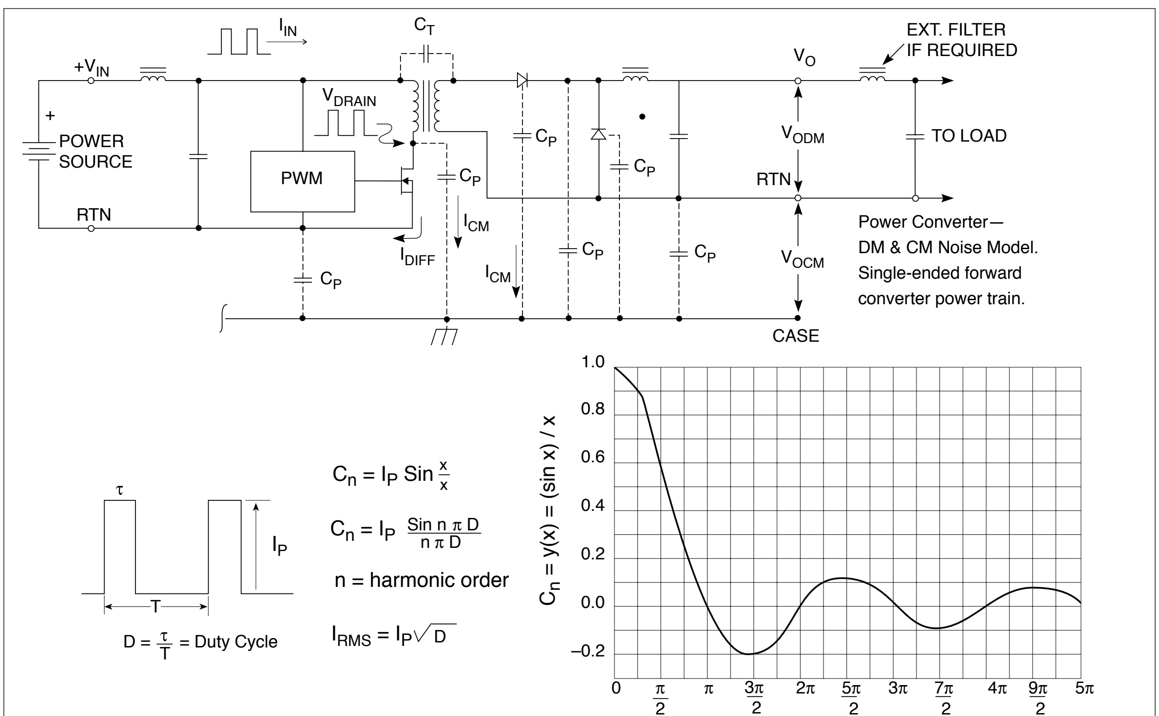

FIGURE 1: EMI PROBLEM DEFINITION

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sents noise in both lines with respect to the case, which requires a different type of filter, one which uses a balun and a bypass capacitor from at least the return line to case. This is known as a Common Mode Low Pass Filter Section and needs to be a part of any low noise power line filter. The balun acts as a high impedance in series with the power line for current components in phase in the 2 windings. When the Baluns are phased as shown in Figure 2, this will be the case, and the net DC ampere turns will be zero since the DC goes into the dot on one winding and out of the dot on the other. The bypass capacitor acts as a low impedance between the return line and case, effectively rebalancing the power lines and attenuating the Common Mode Noise Spectrum.

Interpoint power line filters are designed to comply with the conducted interference limits of MIL 461C, method CE03, and D0-160C Categories A or Z, for tests run with clamp-on measuring devices. These filters have topologies similar to those shown in Figure 2, and are unconditionally stable when used with Interpoint power converters within their maximum line current rating. Multiple power converters can be used with a single filter provided that the filter line current rating is not exceeded and layout constraints covered later in this note are observed. When determining maximum DC input line current, the PWM power converter negative input impedance characteristic must be considered. For these devices, it is approximately true that $P_{in} = (V_{in})(I_{in})$, and $I_{in} = (P_{in}) / (V_{in})$, when they are operated at constant load. It is important to determine the maximum DC input current at maximum load and minimum input line voltage. The input power, P_{in}, is equal to the output power, P_{out}, divided by the decimal efficiency. Once maximum input current is known, the appropriate filter can be selected from the Interpoint catalogue.

Potential EMI Problems and Conflicts

The incrementally negative input impedance characteristic of PWM switching power converters can be a problem during testing as well as in actual system applications if care is not exercised in terminating their input. If the power source impedance is greater than or equal to the power converter negative input impedance, oscillation is assured with malfunction and probable damage to the PWM converter. To avoid this, the source input termination must be less than that seen looking into the PWM

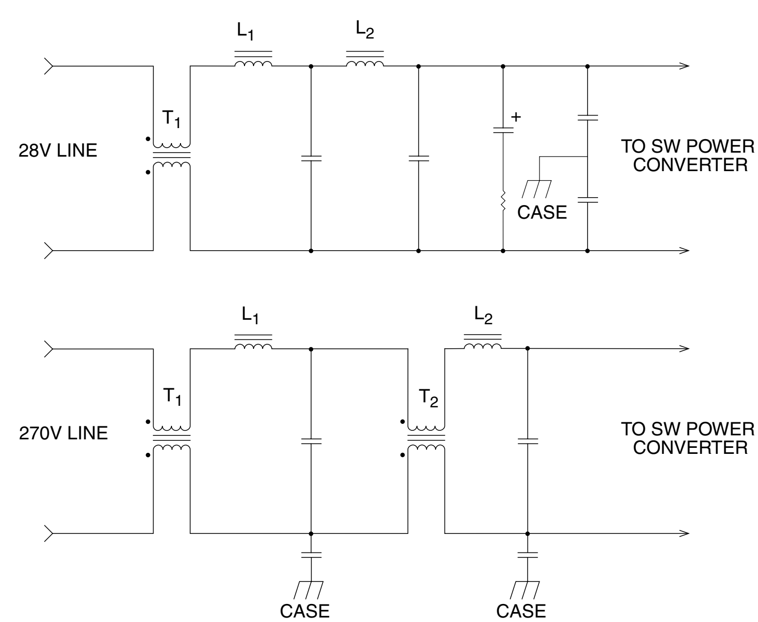

Figure 2: Power Line Filter Examples

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converter at the lowest possible input voltage, and at maximum load.

Referring to Figure 3, the curve $(V_{in})(I_{in}) = P_{in}$ is plotted using a value of 20 for Pin, and is a hyperbola. If the derivative of Vin with respect to Iin is taken – the definition of input impedance – the result is negative. Assigning a value of 20 volts for V_{in} and plotting the resultant tangent to the hyperbola is clearly a negative slope. In contrast, a straight line with a positive slope for a normal 20 ohm resistor is also shown. Referring to the hyperbola, it is easy to see that a small decrease in V_{in} can result in the current increasing at an ever increasing rate as V_{in} continues to decrease, eventually resulting in instability. The crossover point is when the source impedance becomes equal to the negative input impedance, which can result in full rail to rail oscillation.

The preceding example illustrates the basic problem as it applies at DC, but in reality it is more complex. The power converter may have a second order output filter within its control loop bandwidth which will probably be underdamped. At its resonance, the negative input impedance will be less than the DC predicted value, the exact value depending on the filter Q. Reducing the DC value by like 4X should be adequate in most cases to cover this. Assuming a control loop bandwidth of like 25 kHz will be adequate for most cases. If the customer is going to design their own filter, they should include the power converter input filter as part of the design. The component values as well as answers to other questions can be obtained from the Interpoint applications staff.

MIL-STD-461C Versus 461D Conducted Interference Test Setups

The test setups for the conducted interference tests of both MIL 461C and 461D are shown in Figures 4 and 5. For MIL 461C, the interference current measurements are made with a current probe, the units of measurement being dBμA, where dBμA is defined as dB above 1 μA. For example, 10 μA is 20 dBμA. The power line termination of the unit being tested is through 1 meter long parallel lines which are terminated to the ground plane with high quality 10 μfd capacitors. The total inductance of the parallel lines is about 2 μhenries. As such, the test sample termination is a low impedance, and compatible with all but very large PWM switching power converters.

Referring to Figure 5, MIL 461D uses Line Stabilization Networks, LISN's, rather than current probes to make the interference current measurements. The LISN has a 50 μhenry inductor in series with the power line such that the impedance termination is not low as in the 461C case. There is an LISN in each power line, and the LISN schematic and transfer function are shown in Figure 6. The LISN is basically a high pass filter for the measuring instruments 50 ohm input resistance. The interference currents, within the measurement bandwidth, return through the 50 ohm resistor such that the measurement units are dBμV rather than dBμA. The scale of MIL 461D is offset by 34 dB with respect to 461C, as shown on Figure 7, where the 461C limit line is increased by 34 dB and plotted in dBμV on the 461D graph against the 461D limit line. The 34 dB correction is derived from 20LOG50, where 50 ohms is the measurement instrument's input impedance. Below about 1.3 MHz, the 461C limit line is more lenient than that of 461D, while at higher frequencies the opposite is the case. For power converters designed to comply with 461C, a marginal to out of spec condition can be expected below 1.3 MHz for conducted interference measurements made

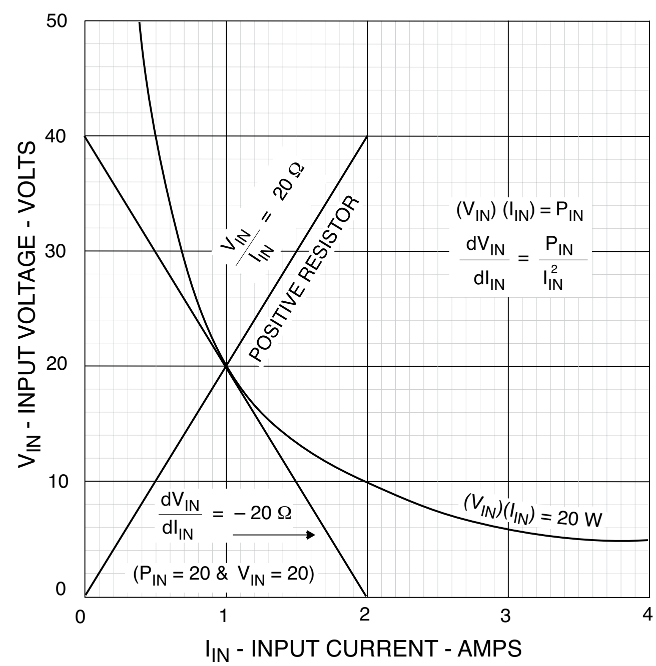

Figure 3: Incrementally Negative Input Z

per MIL-461D. For example, the difference is about 12 dB at 500 kHz. Additional rejection can best be obtained by adding a small balun in series with the power converter or line filter input. A common made inductance of 100 to 200 μhenries per winding will be sufficient, and the balun can be bifilar wound.

Where MIL-461D is a system requirement, it is important to recognize any problems early in the design since correcting them will involve adding components as suggested above. The stability problem due to a 50 μhenry inductor in series with each power

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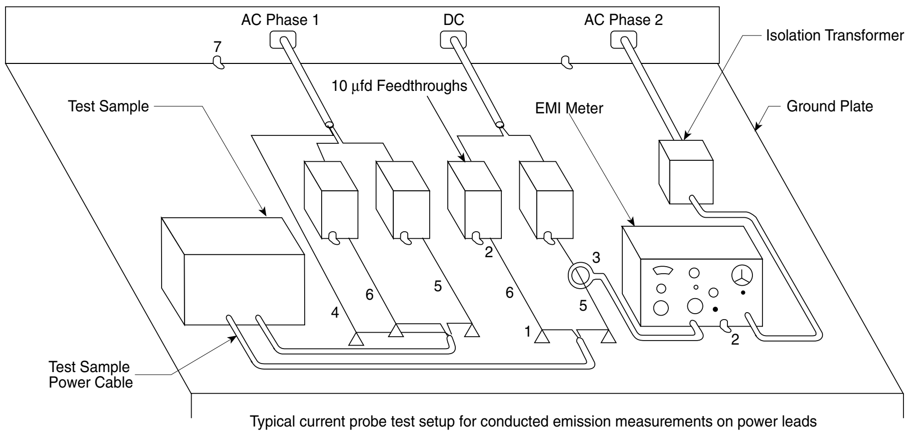

FIGURE 4: MIL-STD-461C CONDUCTED EMISSIONS TEST SETUP PER 462C

line during testing requires special attention if the maximum input power to the PWM switching power converter is greater than about 10 watts for a 28 Volt application. If less than 10 watts, nothing should be required except for adding a shunt capacitor as suggested in the previous paragraph. For a higher voltage power bus, the problem is less severe due to lower input current at a given power level. For example, at 270 VDC the level at which there is no stability problem is below about 200 watts as compared to 10 watts at 28 VDC.

The problem with inductors in series with the PWM switching supply input is outlined in figure 8. The example is for a single ended forward supply with partial input filter. The series inductor is shown in the positive input line. On the graph of Magnitude Impedance vs. Log Frequency, the inductive reactance $J\omega L$ is shown as a straight line, an increasing function of frequency. The power converter input impedance is shown as a horizontal line with a dip where the output filter natural frequency is. The control loop 0 dB gain crossing will be in the area of the dip or just beyond. At the frequency where the sloping line, $J\omega L$, crosses the horizontal input impedance line, oscillation will occur if nothing is done to prevent it. Prevention amounts to adding a damping network across the power converter input, shown in dotted lines as a series R & C. The object is to change the impedance looking back into the inductor from $J\omega L$ to a lower constant value, R_d. This is shown on the graph of Figure 8. The shunt R & C form a series resonant circuit with the line inductor. Above resonance, the impedance seen becomes R_d, the damping resistor value. How to determine the R & C values is outlined in the following.

1. Determine the power converter input impedance using the lowest line voltage involved in testing or application, and the maximum output load. The minimum input impedance is calculated from the formula,

$Z_{in} = (V_{in}^2 / P_o)$ Eff, where,

V_{in} = input voltage

P_o = output power

Eff = decimal efficiency

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Divide the calculated Z_{IN} value by something like 2 for safety margin, in this case 6 dB. This is the value to use for the damping resistor, R_d.

2. The resonant circuit $Q = (1/R_d)\sqrt{L/C}$. For a maximally flat response, a Q of 0.707 should be used to determine the minimum value of C, the damping capacitor. Substituting for Q, we get,

$$C_{min} = 2L/R_d^2$$

This is the minimum capacitor value to use for the damping network, and is applicable to any situation where a large inductor is required in the power line, not just MIL-461D testing.

The above defines the procedure for calculating the damping network component values. These parts may have to be included within the system enclosure to comply

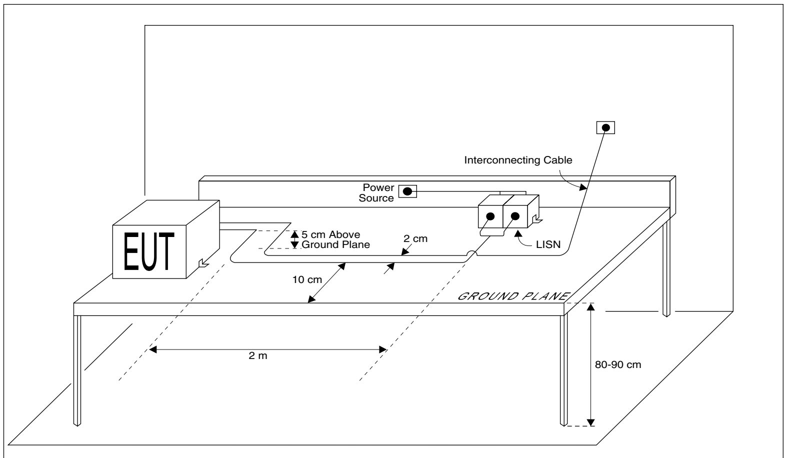

FIGURE 5: MIL-STD-461D CONDUCTED EMISSIONS TEST SETUP PER 462D

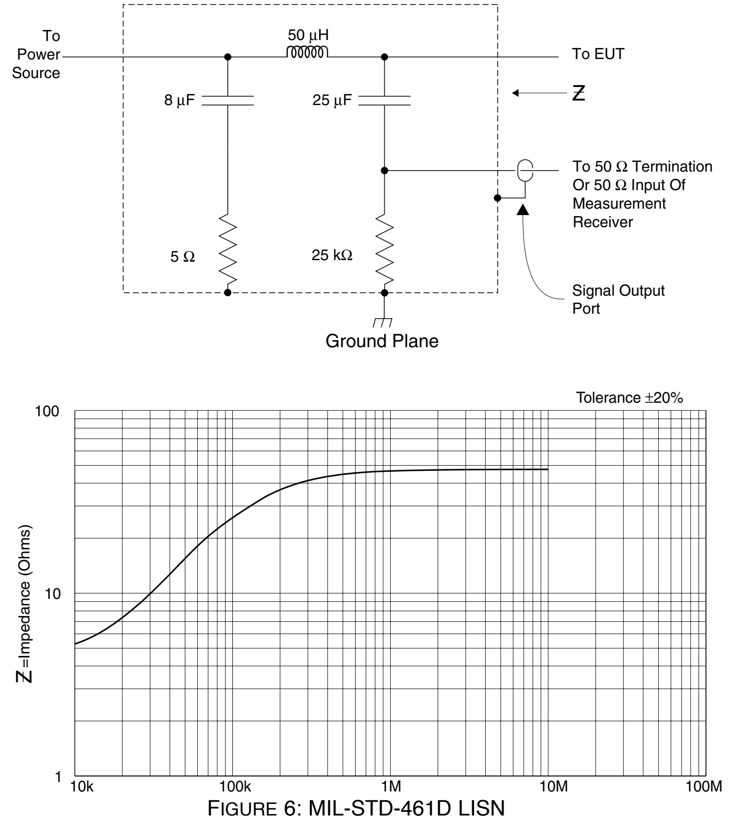

FIGURE 6: MIL-STD-461D LISN

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with MIL-461D since they act as a current shunt and will improve the low frequency CE102 conducted interference measurements. The damping resistor should be a non inductive composition or film type. A porous anode tantalum such as the Sprague type 109D or its MIL equivalent will provide a small package capacitor compatible with power line applications. Where the damping resistor is greater than 5 ohms, a small ceramic capacitor, 0.2 mfd or greater, should be placed in shunt with the RC damping network.

Layout and Multiple Power Converters on a Single Filter

It is important that the power line EMI filter be as close as possible to the power converter in order to have optimum performance. The common mode interference current generator is in the power converter (or converters). The correction circuits–balun and bypass capacitors–are in the filter. The cases need to be connected together by a low inductance means because of the filter bypass capacitors which must be terminated to the power converter case in order to steer common mode currents back into the 28 V return line. An ideal way to connect the cases together is with a small ground plane which is no more than a copper area under the parts or on the other side of a PC Card on which they may be mounted. The cases can be connected directly to the ground plane by mounting on it or by soldering to the case pins, where the plane is on the opposite side. In any case, connect the cases together with the shortest and widest possible conductors.

Where the power converters and filter are separated by several inches or more, radiation from the power line between the two will occur due to common mode currents even if the wires are shielded or twisted, unless corrective measures are taken. The common mode currents are components in phase in both lines such that there is incomplete cancellation of the fields and hence radiation. A balun and bypass capacitor at the remote power converter will rebalance the power line, reducing or eliminating the radiation. Also, if the balun is segment wound, the leakage inductance can be used with a shunt capacitor to obtain additional differential mode rejection. Refer to Figure 9 for details.

Where multiple unsynced power converters are used on a single filter, products of modulation will occur at the various difference

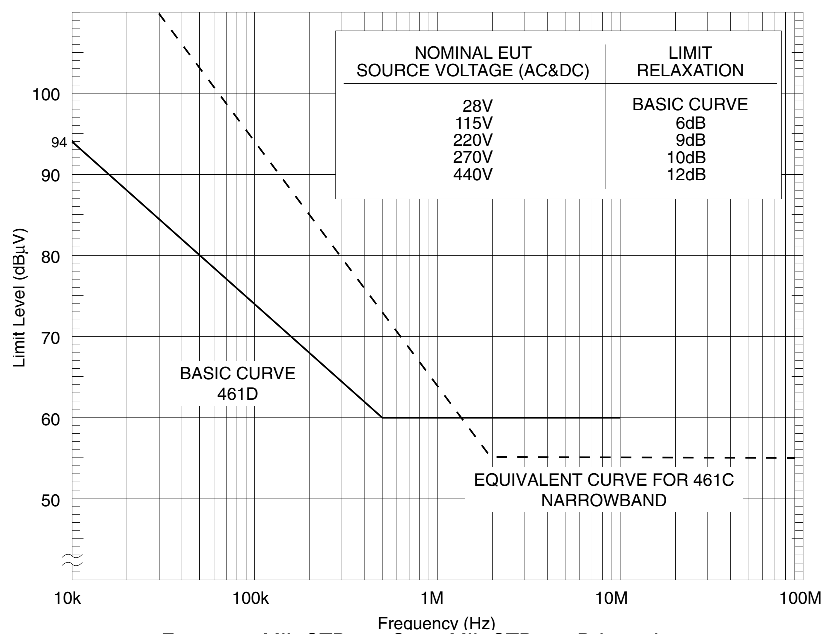

Figure 7: MIL-STD-461C vs. MIL-STD-461D Limit Lines

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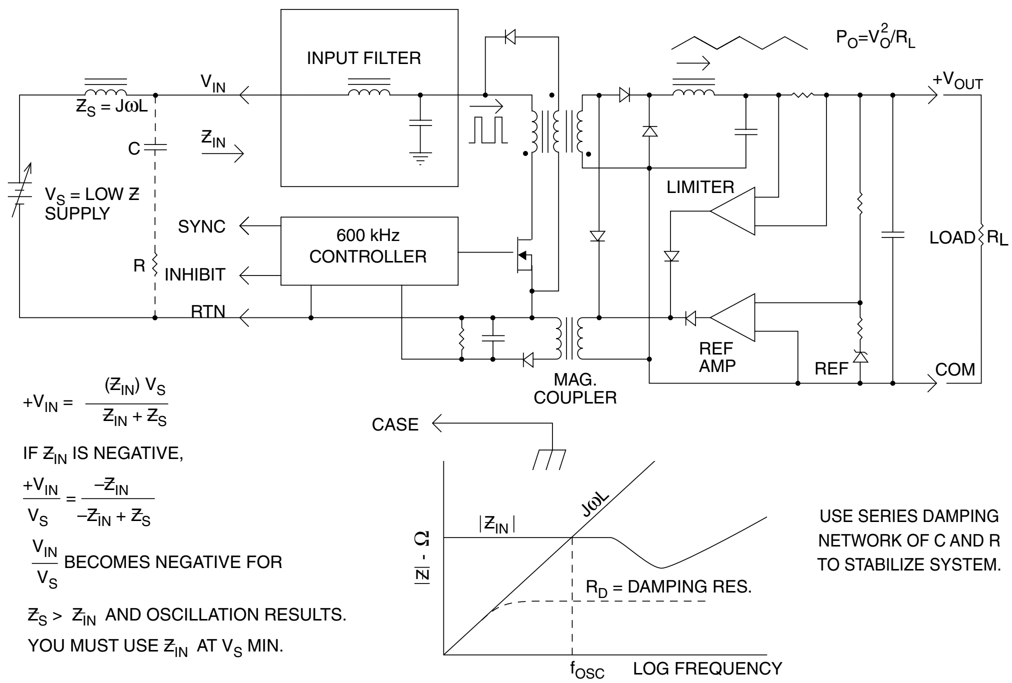

FIGURE 8: PWM SWITCHING POWER SUPPLY STABILITY EXAMPLE

frequencies. These will appear as part of the input ripple current with components at generally less than 50 kHz and with amplitudes generally below the MIL-461C CE01 and/or CE03 limit line. Where the amplitudes are objectionable, a partial EMI filter at each power converter will reduce the differential noise spectrum amplitude and also that of the modulation products. A simplified filter can then be used which is common to all. Refer to Figure 9 for details.

DESIGNING THE EMI POWER LINE FILTER

Designing a power line EMI filter will involve low pass filter topologies similar to those of Figure 2. Referring to Figure 2, the upper figure is typical of 28 V applications where a single balun and bypass capacitor and one or more differential sections are used. If the balun is segment wound on a toroid, or the windings separated by some similar other means, the leakage inductance can be used in place of L_1. This configuration may also use a series RC network for shunt damping rather than adding series resistance for this purpose. This circuit, as configured, is a second order common mode and fourth order differential filter.

The lower figure is useful for 270 V power systems where the higher dV/dT's result in larger common mode currents and hence a second balun and bypass capacitor. L_1 and L_2 can be eliminated if the balun leakage inductance can be used in their place, resulting in a total of seven parts. A series damping resistor may be allowable due to the lower line currents at 270 V as compared to 28 V. This circuit, as configured, is a fourth order common mode and a fourth order differential filter. This configuration may also be used at lower line voltages where a shunt damping network may be used rather than the series resistor to reduce power loss in some cases.

In designing your own EMI filter, the first step is to determine the amount of differential attenuation required at the switching frequency, the first noise component in the narrowband spectral series. This can be done by calculation from the input line current Fourier Series, or by estimating the value from information on the device data sheet. For a Forward Converter, the example of Figure 1 can be used. If a duty cycle of 0.5 is assumed, the Peak Pulse line current is twice the DC value, and the first spectral component RMS value is 0.45 times the Peak value. Use the maximum load and nominal input voltage to make the calculation, and subtract the EMI Specification limit value from the obtained result. Add some margin, like 12 dB, and this is where you start the filter design. If using data sheet information, use half the maximum value for the first spectral component value, and calculate the RMS value. This will probably result in a conservative answer since much margin may be built into the data sheet figures. For example, the maximum Line Ripple Current value on the MHD data sheet is 50 mA p-p. The first component value would then be about 50/(2)(1.414) = 18 mA RMS. This is 18000 μA, or 85.1 dBμA. At the 625 kHz switching frequency, the MIL-STD-461C limit line value is about 33 dBμA. If 12 dB of margin is used, the required attenuation at 625 kHz is (85.1–32 + 12) = 65.1 dB. Note that dBμA is dB above 1 μA, which becomes 20Log (current in microamps). It is not necessary to determine the value of higher order spectral components since the series declines as a function of frequency and the Low Pass Filter will attenuate them more readily than the fundamental if it is designed and fabricated properly.

The schematic and definition of a Second Order Low Pass filter is outlined in Figure 10. The power converter generates differential spectral noise represented by a current source, In, which terminates across the filter capacitor. The object is to force the ratio of (AC line current) / (noise current) toward zero as a function of increasing frequency. This will readily occur since the inductor impedance, JωL, increases with frequency while that of the capacitor, 1/JωC, decreases with frequency, ideally forcing the noise current through the capacitor rather than the line inductor as the frequency becomes large compared to the filter resonant frequency.

In Figure 10, the short circuited current transfer function is derived and written as a function of S, where S = Jω or d/dt, and $\omega = 2\pi f$, the angular radian frequency. The line voltage source is assumed a short circuit for this exercise. The series resistances are the inductor's copper and other losses, and the capacitor's ESR plus trace resistance. The transfer function has a zero in the numerator and two poles in the denominator. The zero is a function of the capacitor value and it's ESR plus trace resistance. For a solid tantalum capacitor, this zero will in the area of 50 kHz. For a BX or X7R type ceramic capacitor, the zero will be upwards of 1 MHz. It is desirable to have the zero at the highest possible frequency since it modifies and spoils the filters asymptotic behavior and response. Ceramic capacitors have this advantage as well as being more tolerant to surges as compared to solid tantalums.

The transfer function in terms of the R, L, and C parameters can be compared to the transfer function standard form in terms of the Q, Lead and Resonant Frequencies. See Figure 10. By equating like terms, the expressions for Q, Lead and Resonant Frequencies can be written, as well as the formula for attenuation in dB at the switching frequency in terms of these frequencies. At resonance, the filter response is Q times the low pass response. Above resonance, the denominator asymptotic response becomes a function of the ratio of $(\omega/\omega O)^2$ and reduces to $20Log(\omega/\omega O)^2$, having a slope of -12 dB/octave or -40 dB/decade. Beyond the lead frequency the numerator response becomes 20Log(ω/ωL), having a slope of +6 dB/octave, or + 20 dB/Decade. Refer to the graph of Figure 10 where the asymptotic response is shown as having a slope of -12 dB/octave, and reducing to – 6 dB/octave after the Lead comes in.

It is desirable to damp the filter and maintain a Q of 1 or less, 0.5 being critically damped. The formula for Q suggests a small ratio of L/C since the R values cannot usually be used to control Q because of power loss or lowering of the numerator zero frequency. Often times it's not possible to maintain a low Q, and you settle for what you can get, such as a value of 3 or 4. An alternate means is to use shunt damping which involves additional parts and is covered in the following.

A second order filter section using shunt damping is shown in Figure 11. Here a ceramic capacitor, C_2, is used as the filter capacitor, and a tantalum capacitor with series resistor, C_1 and R_d, are used as the shunt damper. The angular frequency at which the damping network comes in, $\omega d = 1/R_dC_1$, is set below the resonant frequency such that R_d functions as an AC connected shunt resistor. Using a ceramic capacitor for the filter element places the Lead frequency 2 or more octaves beyond the switching frequency, which results in better filter performance. Formulas for calculating the R, L, and C values are given in Figure 11.

It is possible to calculate the amount of Common Mode attenuation required, but usually setting the CM resonant frequency at least a decade below the switching frequency will be adequate. The first CM spectral component occurs at the switching frequency. Use a high permeability core, μ = 5000 or more for the balun, and don't worry about the permeability starting to go off below a megahertz because the material will become very lossy at the higher frequencies, compensating for this effect. Use a Common Mode Inductance of 200 to 500 μhenries, and a good ceramic capacitor to bypass the return line to the power converter case. Calculate the capacitor value to place resonance at least a decade below the switching frequency, and/or use the largest practical value of capacitance. If the balun is segment wound on a toroid, windings separated, the leakage inductance can be used for the differential filter, giving a minimum parts solution.

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When designing the EMI filter, it often is impractical to use a single second order section because of the size of the parts involved. In this case use 2 second order sections and treat them as if they don't interact. They do, but the effect will be slight for the initial design estimates. See Figure 2 for circuit configurations. For 28 volt systems, use the top configuration and start with the section interfacing the power converter. Set the resonant frequency as low as practical dependent on component sizes, etc. Use inductor values less than 5 μhenries for the circuit of Figure 10, and less than 50 μhenries for the shunt damping circuit of Figure 11, where a ceramic filter capacitor is used. If a second section is required, use the balun leakage inductance for the second section, or add inductor L_1. Set the resonant frequency an octave or so above that of the low frequency section such that its resonant peak is buried in the asymptote of the latter. Optimize using Spice or Microcap models. Make sure the impedance looking into the filter output with its input shorted is low enough to be compatible with the power converters it must work with. Note that low filter output impedances require small ratios of L/C.

The higher voltage example at the bottom of Figure 2, can be designed using two identical sections with a single series damping resistor. One differential pole pair will be above, and the other below the calculated value. If the balun leakage inductance can be used for L_1 and L_2, this filter can be made with the seven parts shown. All capacitors are ceramic types. The leakage inductance of a 15 turn/winding segmented balun on a 0.5 inch O.D. toroid with a permeability of 5000 will be around 3 μhenries. The best means of verification is to make a sample and measure it with a short circuit test.

POWER LINE FILTER DESIGN EXAMPLE

Assume that we have an MHD2805S (28 VDC in, 5 VDC out) single ended forward power converter operating at a 2.5 Amp. output load and a requirement to meet the conducted emissions of MIL-461C, method CE03. At 28 VDC and at an efficiency of 82%, the DC input current can be calculated as I = (12.5) / (28)(0.82) = 0.54 Amps. If we assume a 50% duty cycle, close enough for practical purposes, the peak current will be twice the DC, or 1.08 Amps. The first spectral component at the 625 kHz clock frequency will be (0.45)(1.08) = 0.49 Amps RMS. The dBuA level will be 20Log(490,000) = 114 dBuA. The 461C-CE03 Limit Line is at 33 dBuA at 625 kHZ. In the absence of any power line filter in the MHD2805S, the required attenuation in dBuA will be (114 dBuA – 33 dBuA) = 81 dB.

The MHD2805S does have a second order input filter having an L value of 3.4 μhenries and ceramic capacitor of 6.0 μfarads. This filter section has a resonant frequency of 35 kHz, and due to the ceramic capacitor, the Lead associated with its ESR will be well beyond the switching frequency of 625 kHz. The

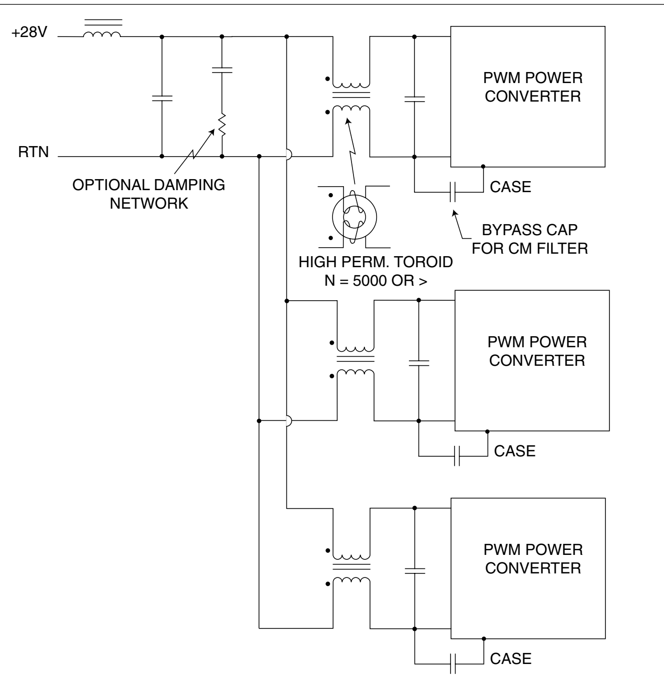

FIGURE 9: MULTIPLE CONVERTERS WITH A SINGLE FILTER

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additional attenuation due to this section is $20\mathrm{Log}(625/35)^2 = 50$ dB. Then, the additional attenuation required from the external line filter is (81 dBuA – 50 dB) = 31dB. This can probably be done with a single second order LC section, particularly if a ceramic capacitor is used for the C. We will use a loosely coupled balun for the CM filter, with its leakage inductance functioning as the differential inductor. Refer to Figure 12, where the MHD2805S internal filter and spectral noise source are shown with the proposed external line filter.

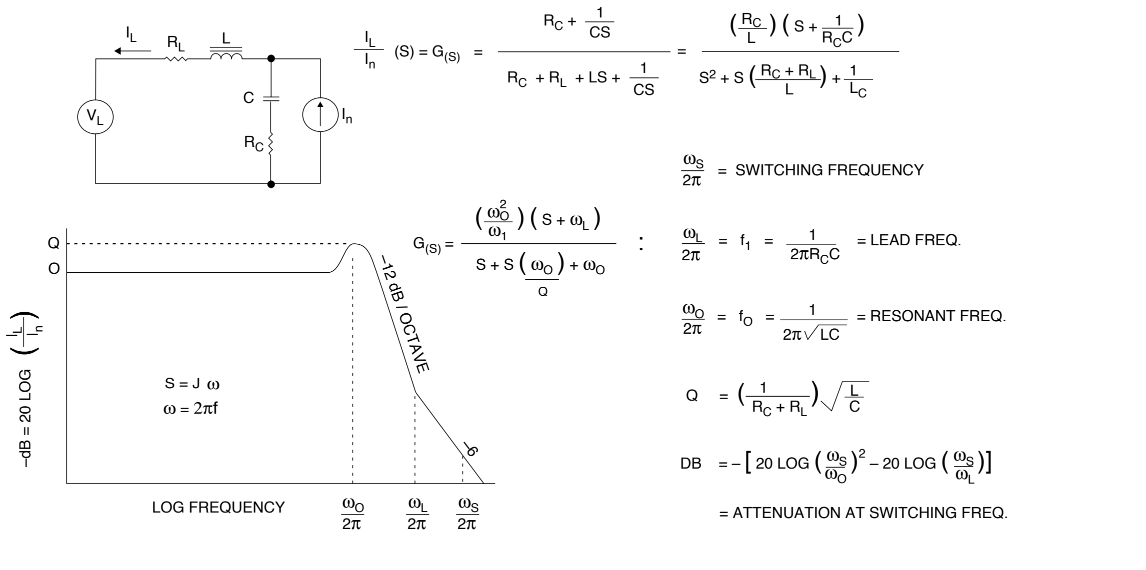

FIGURE 10: LOW PASS FILTER, PRACTICAL FORM

The filter will be implemented using a 0.375 inch O.D. toroid with a permeability of 5000. The measured common mode inductance with 13 turns per winding is 380 μhenries at 100 khz, and the leakage inductance 2.5 μhenries using segmented windings. Starting with a 2 μfd ceramic capacitor, a practical value, we get a resonant frequency for the differential filter of 71 kHz. See the formulas of figure 10. The additional attenuation from the external line filter will be $20\mathrm{Log}(625/71)^2 = 38$ dB. We predicted above that 31 dB would be required which leaves 7 dB of margin. Since the power level is low, a 0.25 ohm series resistor is used for damping. The dominant Q is 2.5 with this resistor, and about 10 without it. A Q of 2.5 is probably allowable, but 10 is too high. A 0.047 μfd capacitor, to bypass the return line to case, places the CM resonant frequency at 38 kHz, more than a decade below the switching frequency which should be adequate.

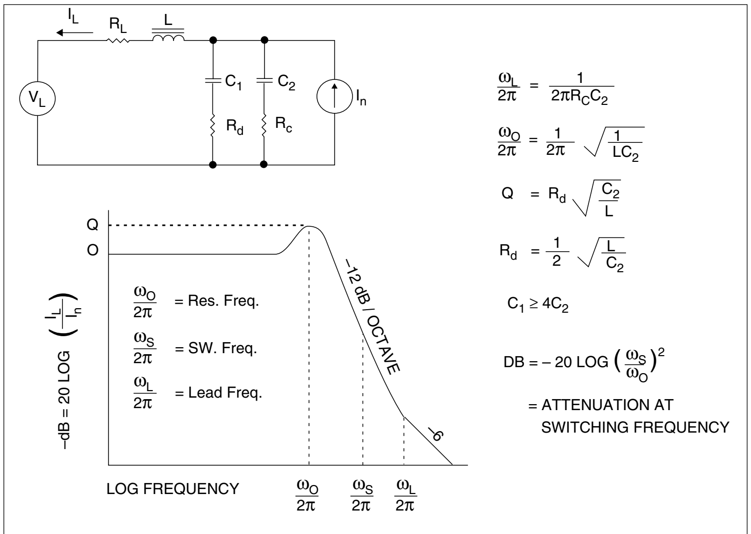

FIGURE 11: LOW PASS FILTER WITH SHUNT DAMPING

The EMI test results versus the MIL 461C-CE03 limit line are shown in Figures 13 and 14, where the margin at the switching frequency is about 10 dB. The magnitude frequency response and short circuited output impedance from MICROCAP computer models is shown in figures 15 and 16 also. This filter is an example of what can be done with only a few parts, and in most cases requires no shielding, or an additional metal box. For it to work as demonstrated, it should be next to the power converter and built over a small ground plane. Surface mount type capacitors are preferable, and you must be sure not to add any additional ESR or ESL in series with the capacitors if the filter is to function properly.

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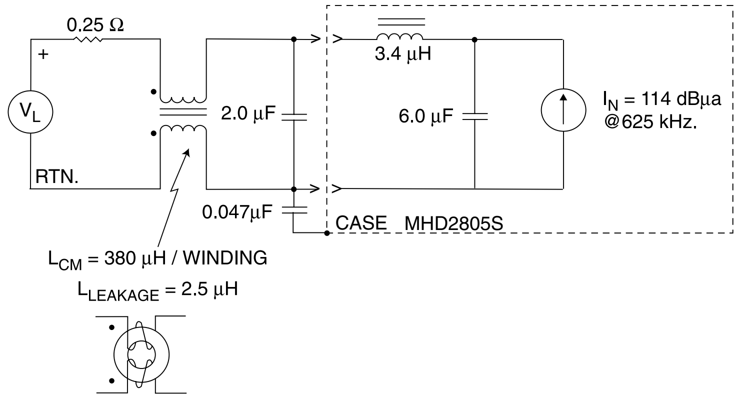

FIGURE 12: INPUT POWER LINE FILTER EXAMPLE

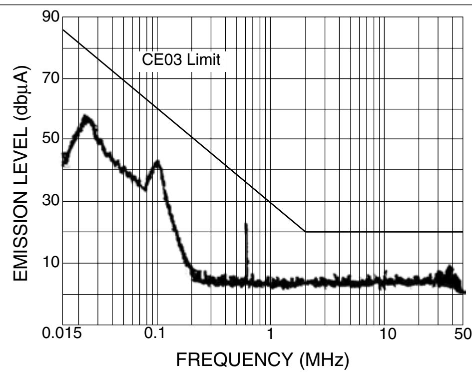

FIGURE 13: SPECTRAL NOISE TEST DATA – HIGH LINE

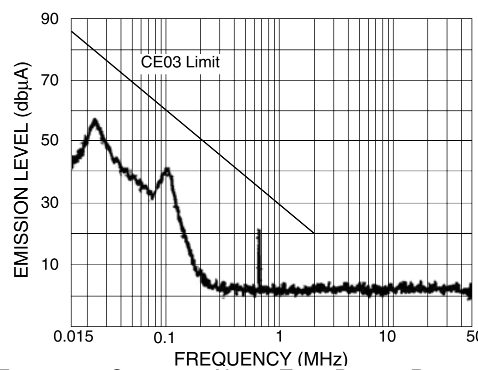

FIGURE 14: SPECTRAL NOISE TEST DATA – RETURN LINE

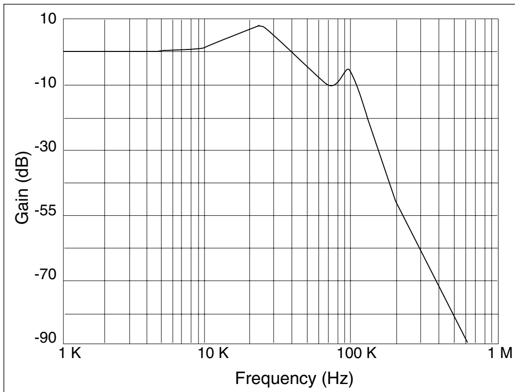

FIGURE 15:
MHD WITH EXTERNAL FILTER
MAGNITUDE FREQUENCY RESPONSE

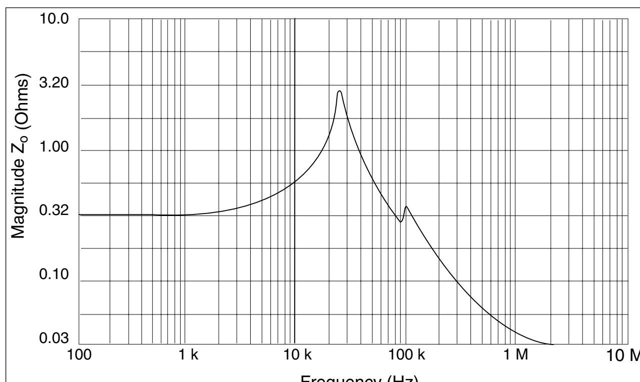

FIGURE 16:
MHD POWER LINE FILTER
MAGNITUDE OF SHORT CIRCUIT Z IN OHMS