

EMI Conducted Interference

INTERPOINT® PRODUCT 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 inductor/capacitor (LC) section which reduce the differential spectral noise components to about 30 mA p-p. This is adequate for many applications. However, where lower noise currents or compliance to a military standard (MIL-STD) or other EMI specification is required, an external power line filter will generally be required.

Interpoint power converters are fabricated on ceramic substrates, which function as printed circuit boards providing a low thermal impedance to the header of the converter on which the substrate is mounted. The converter's header, which is the bottom of its chassis, is made from thick steel and provides heat sinking for the power components internal to the converter. 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. Due to the overlapping surface areas, the power train components and traces will all have parasitic capacitances to the converter's header (chassis). The magnitude of capacitance depends on the area occupied on the substrate. The high dv/dt associated with switching power conversion causes common mode (CM) currents— $I_{CM} = C(dv/dt)$ —to be pumped in and out of the chassis corresponding to each transition of the MOSFET 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 converter, the most common Interpoint topology.

These common mode currents exit the converter's chassis and return back to the converter through the input lines, or visa-versa. We will use the converter's main switching power MOSFET as an

example of a power component driving CM currents into the chassis. The MOSFET is one of many components that will drive current into the converter's chassis. The MOSFET is mounted to the substrate. The overlapping surface areas of the MOSFET and converter's header create the plates of a capacitor. The ceramic substrate electrically isolates the MOSFET from the chassis and acts as the dielectric for the capacitor. Every time the power MOSFET turns off, the main transformer will reset and drive the drain voltage of the MOSFET high. During the reset transition the transformer's magnetizing and leakage inductance will drive current into the parasitic capacitance created by the overlapping surface areas shown as "parasitic capacitance" in Figure 2 on page 2. The current taking the path of the parasitic capacitance will come from the bottom, "dot" end, of the transformer. Any current exiting the bottom of the transformer will have to return to the top, no "dot", end of the transformer. In order to get to the top, no "dot", end of the transformer, the current entering the parasitic capacitance will exit the chassis, enter the line impedance stabilization networks (LISN) through its earth ground connection, travel through the LISN, then enter the converter through the converter's power lines. The spectrum analyzer connects to ports across R_1 and R_2 in Figure 2 on page 2. Since the input impedance to the spectrum analyzer is 50 ohms R_1 and R_2 can be considered as 50 ohms. Any current passing through the 50 ohm impedance will be measured by the spectrum analyzer depending on which port the spectrum analyzer is connected to. The port not connected to the spectrum analyzer should be terminated with a 50 ohm impedance. When the MOSFET turns off (not shown in the diagram), the capacitance will discharge and the CM current will exit the converter through the power lines, through the LISN and re-enter the converter through its chassis. If testing to MIL-STD-461C a LISN will not be used but the concept of where the current flows and how it is measured is similar. Another source of CM noise is through any primary to secondary capacitance

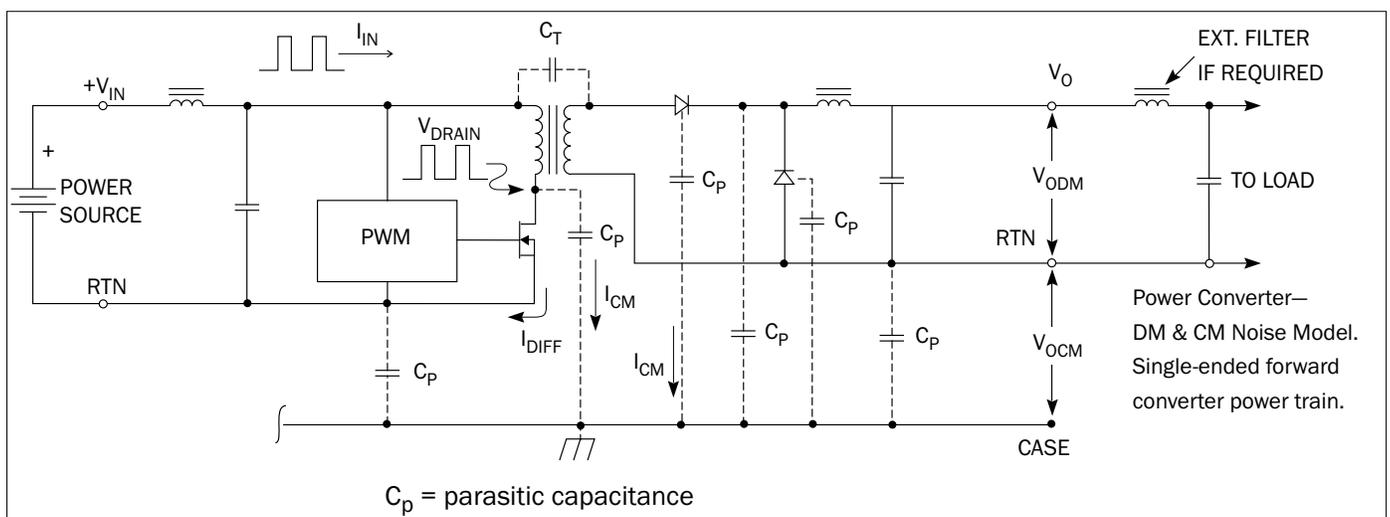


FIGURE 1: EMI PROBLEM DEFINITION

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in the main transformer. Any current traveling from the primary of the transformer to the secondary of the transformer will find its way back to the primary side of the converter. Having capacitors from output common to chassis and input common to chassis will help redirect this current back to the primary where it belongs. Any currents that take this path will not flow through the test equipment and will not be measured by the spectrum analyzer. These CM currents flowing through the chassis and transformer complicate and add another dimension to the problem of EMI suppression.

The previous discussion of CM currents relates to the transitions of the MOSFET. However while the MOSFET is in the on-state, the output load is seen looking into the transformer primary, which causes a rectangular pulse of current to flow in the primary. Ideally, this pulse of current should enter at the positive line and entirely return through the return line to the 28 volt 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 of Figure 1 on page 1. This is a differential low pass filter section which works for noise currents in one line with respect to the other.

Since there are common mode components in the chassis which belong in the power lines, we have a component of noise current which is in phase in both input lines. This requires a different type of filter, one which uses a balun and a bypass capacitor from at least the 28 volt return line to case. This is known as a common mode filter section and needs to be a part of any low noise power line filter. The balun acts as a low impedance in series with the power lines for differential mode (DM) currents due to the phasing of the balun as shown in Figure 3 on page 3. With this phasing the net ampere turns will be zero for DM currents as current going into the dot on one winding will have an equal amount of current exit the dot of the other winding. On the other hand they will be a high impedance to current in phase with both windings as current going into the dot on one winding will also have the same current entering the dot on the other winding. In this case the ampere turns will not cancel and will see the high inductance of the balun. The bypass capacitor acts as a low impedance between the return line and case, allowing CM currents to return to the converter where they belong. Any current taking this path will not flow through the test equipment and will not be measured by the spectrum analyzer.

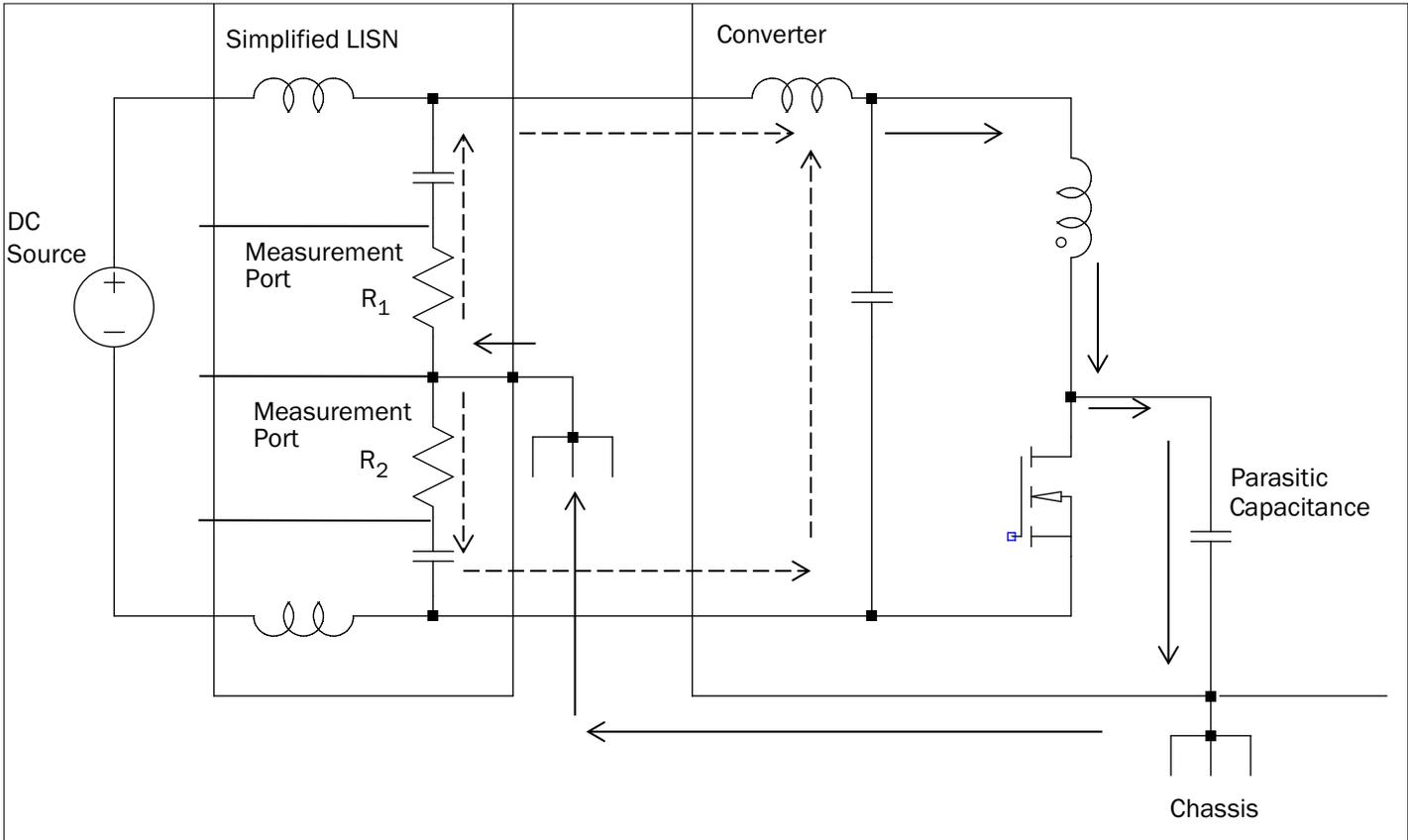


FIGURE 2: PATH OF CM CURRENT

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Interpoint EMI filter modules are designed to comply with the conducted interference limits of MIL-STD-461C, method CE03. Some models also comply with MIL-STD-461 D/-E/-F method CE102. These filters have topologies similar to those shown in Figure 3, and can be used with Interpoint power converters within the filter's 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's negative input impedance characteristic must be considered. For these devices, $P_{IN} = (V_{IN})(I_{IN})$, 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

likely with malfunction and probable damage to the PWM converter. To avoid this, the source impedance must be less than that seen looking into the PWM converter at the lowest possible input voltage, and at maximum load. The converter's internal input filter should not be considered as part of the converter's input impedance. The converter's internal input filter should actually be considered as part of the source impedance.

Referring to Figure 4, the curve $(V_{IN})(I_{IN}) = P_{IN}$, is plotted using a value of 20 watts for P_{in} , and it is a hyperbola. If the derivative of V_{IN} with respect to I_{in} 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 converter's negative input impedance, which can result in full rail to rail oscillation of the converter's input voltage.

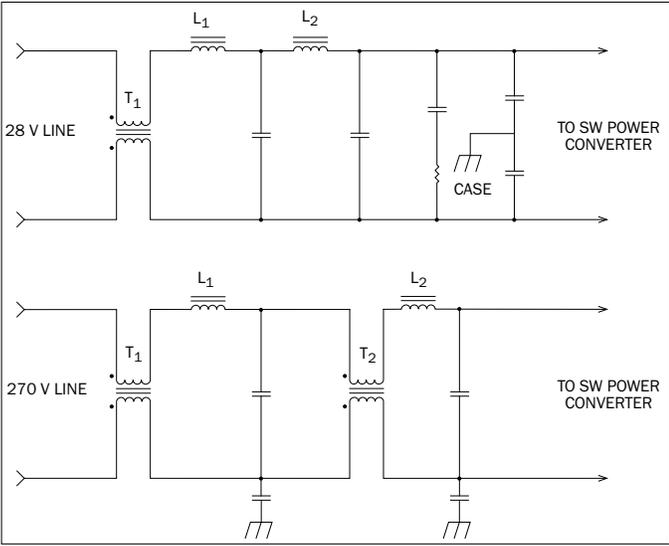


FIGURE 3: POWER LINE FILTER EXAMPLES

product catalogs. In some cases an RC damping network, discussed later, may be needed across the power lines, between the converter and filter. The Applications Engineering Department [powerapps@craneae.com] can assist in determining if a damping network is needed.

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. The source impedance can be considered as the impedance looking from the input power pins of the converter back to the dc source. This would include the dc source and anything between the dc source and converter. If the source impedance, which includes an external EMI filter, is greater than the power converter's negative input impedance, oscillation is

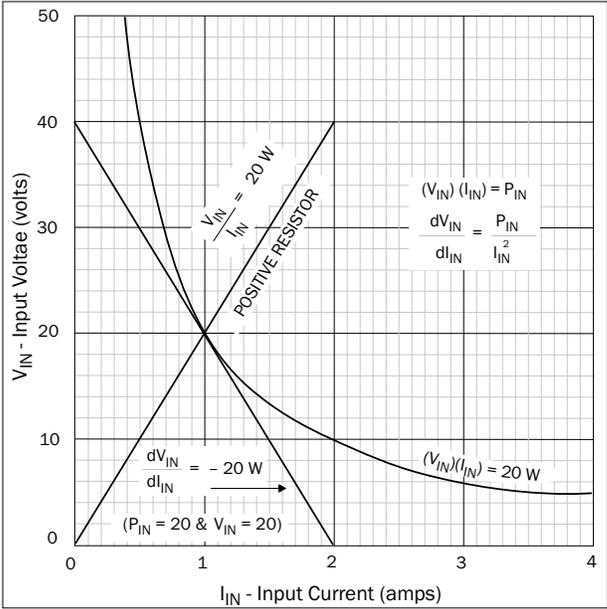


FIGURE 4: INCREMENTALLY NEGATIVE INPUT Z

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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 band-width which will probably be underdamped. Near the output filter's resonance frequency, the negative input impedance will be less than the dc predicted value, the exact value depending on the output filter's resonant impedance, the Q of the converter's output filter, the load resistance, load capacitance and the gains of the feedback loop and feed-forward compensation if implemented. Reducing the value of the dc impedance by a factor of three and assuming the lowest input voltage, the converter's maximum output power capability should be adequate to eliminate impedance problems. An example of comparing the source impedance against the converter's input impedance is addressed in Figure 18 on page 13.

**TEST SETUPS : MIL-STD-461C, CE03
VERSUS 461D, CE102**

The test setups for the conducted interference tests of both MIL-STD-461C and 461D are shown in Figure 5 on page 4 and Figure 6 on page 5. MIL-STD-461E and F are very similar to 461D. For our discussion we will use 461D as a comparison to 461C. For MIL-STD-461C, Figure 5, 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, and 100 µA is 40 dBµA. The power line termination of the converter being tested is through one meter long parallel lines which are terminated to the ground plane with high quality 10 µF capacitors. The total inductance of the parallel lines is about 2 µH. As such, the test sample termination is a low impedance, and compatible with all but very large PWM switching power converters.

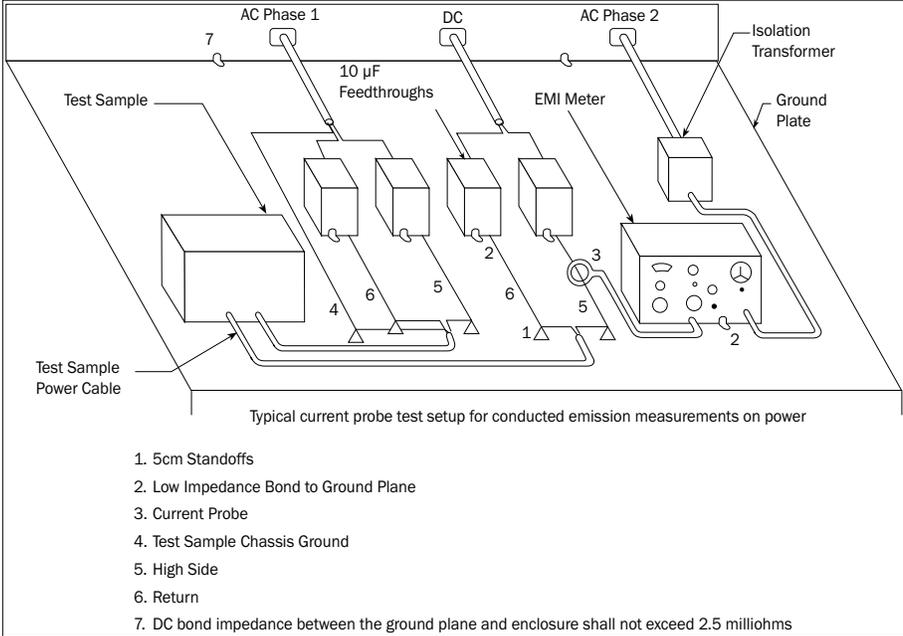


FIGURE 5: MIL-STD-461C CONDUCTED EMISSIONS TEST SETUP PER 462C

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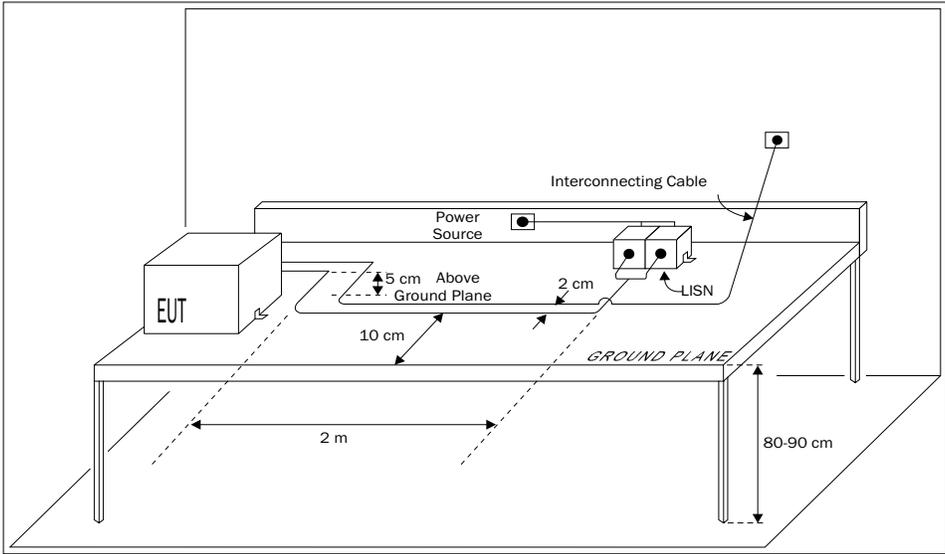


FIGURE 6: MIL-STD-461D CONDUCTED EMISSIONS TEST SETUP PER 462D

Referring to Figure 6, MIL-STD-461D uses a LISN rather than current probes to make the interference current measurements. The LISN has a 50 μH inductor in series with each power line (total of 100 μH) such that the impedance termination is not low as in the 461C case. The LISN schematic and transfer function are shown in Figure 7. 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 dB μV rather than dB μA . The scale of MIL-STD-461D is offset by 34 dB with respect to 461C, as shown on Figure 8 on page 6, where the 461C limit line is increased by 34 dB and plotted in dB μV on the 461D graph against the 461D limit line. The 34 dB correction is derived from $20 \times \text{LOG}50$, 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 filters designed to only comply with 461C, a marginal up to an out of spec condition can be expected below 1.3 MHz for conducted interference measurements made per MIL-STD-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 CM inductance of 100 to 200 μH per winding will be sufficient, and the balun can be bifilar wound.

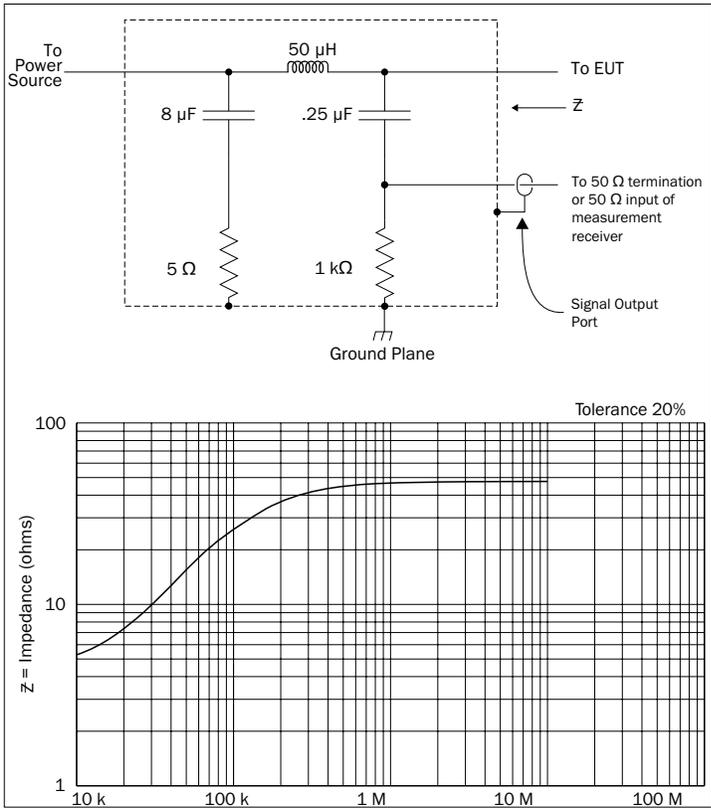


FIGURE 7: MIL-STD-461D LISN

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Where MIL-STD-461D, E or F 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. You can find these documents at <https://landandmaritimeapps.dla.mil/programs/milspec>. The stability problem due to a 50 µhenry inductor in series with each power line (100 µH total) during testing requires special attention if the maximum power capability of the switching power converter is greater than about 10 watts for a 28 volt application. If less than 10 watts, nothing may be required except for adding a shunt capacitor. Some 461 standards allow the use of a LISN with lower inductor values which is desirable for most applications. 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 volts the level at which there is no stability problem is below about 200 watts as compared to 10 watts at 28 volts.

The problem with inductors in series with the switching converter's input is outlined in Figure 9. The example is for a single ended forward supply with partial input filter. The series inductor which can represent the inductor internal to a LISN, or any added inductance, 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, increasing as a 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 beyond the dip. At the frequency where the sloping line, $j\omega L$, crosses the horizontal input impedance line, oscillation can 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 (source impedance) from $j\omega L$ to a lower constant value, R_D . This is shown on the graph of Figure 9. The shunt R & C form a resonant circuit with the line inductor. Above resonance, the impedance seen looking from the converter back to V_S becomes R_D , the damping resistor value. The converter's input filter will be part of the source impedance but will typically not have an effect until a much higher frequency and is not shown in this plot.

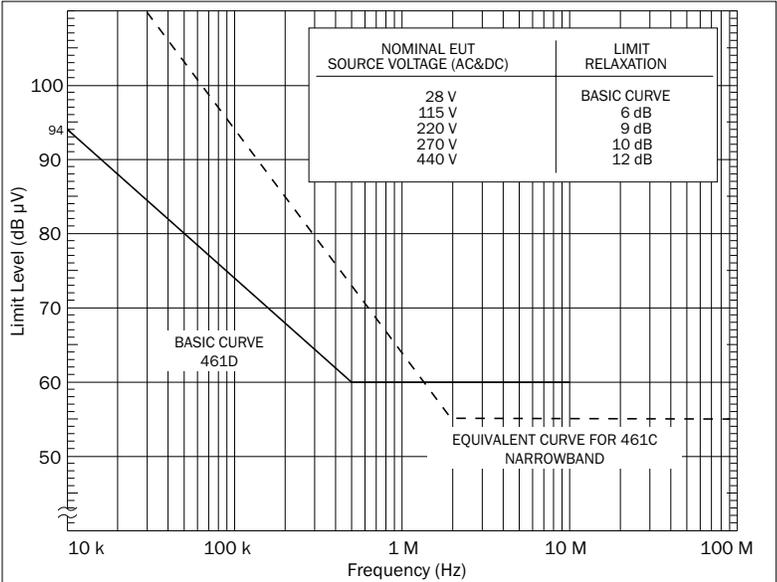


FIGURE 8: MIL-STD-461C vs. MIL-STD-461D LIMIT LINES

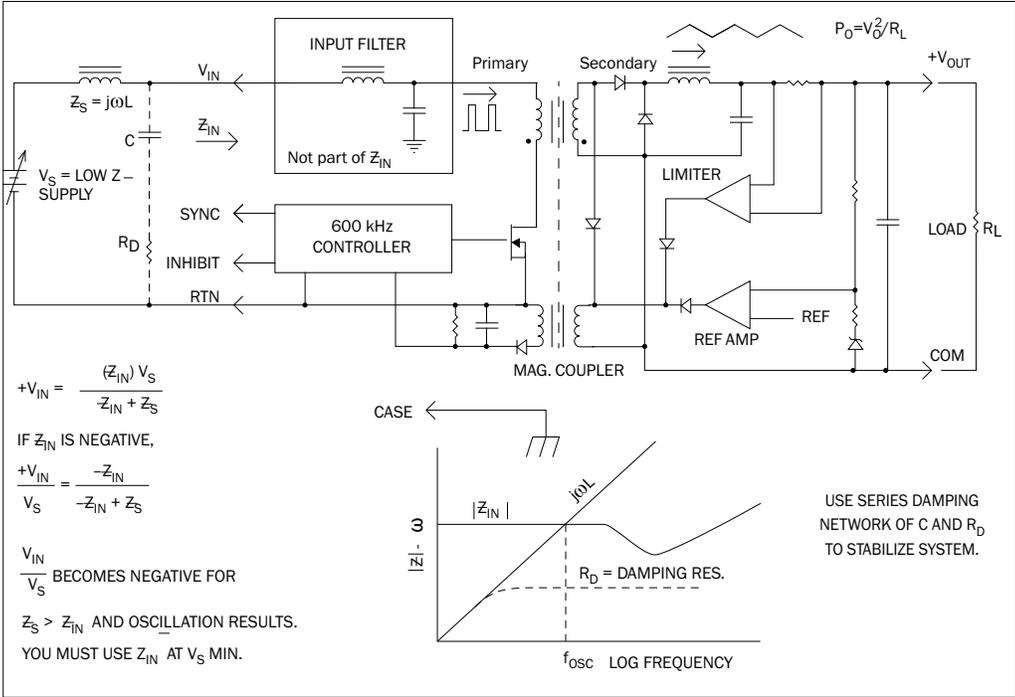


FIGURE 9: PWM SWITCHING POWER SUPPLY STABILITY EXAMPLE

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How to determine the R & C value is outlined below.

1) Determine the power converter input impedance using the lowest line voltage and the converter's maximum output power capability. The minimum input impedance is calculated from the formula,

$$Z_{IN} = \left(V_{IN}^2 / P_O \right) \text{Eff}$$

Where:

V_{IN} = minimum input voltage

P_O = maximum output power capability

Eff = decimal efficiency

Using the maximum output power capability of the converter and the converter's minimum input voltage, divide the calculated Z_{IN} value by something like three for safety margin, in this case 9.5 dB. This is the value to consider 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. It is applicable to any situation where a large inductor is required in the power line, not just MIL-STD-461D/E/F testing. It may also be necessary to add a damping network between an Interpoint EMI filter and an Interpoint converter. Modeling the EMI filter, damping network, converter's input filter, and any other external components is highly recommended. In this case the converter's input filter should be considered as part of the source impedance. The Applications Engineering Department can provide schematics of EMI filters, and filters internal to the converters, along with some guidance to ensure stability.

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 with MIL-STD-461D/E/F. 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-STD-461 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 μF or greater, should be placed in shunt with the RC damping network. If you are required to meet a MIL-STD-461C, CS01 type of test, where there will be a significant disturbance on the power rails, make sure that your capacitor and resistor can handle the power, especially if you have to dwell at the resonant frequency of the circuit.

LAYOUT AND MULTIPLE POWER CONVERTERS ON A SINGLE FILTER

It is important that the power line EMI filter be placed as close as possible to the power converter in order to have optimum performance. The path for CM interference current is from the chassis of the power converter to the chassis of the EMI filter then back into the converter's power lines. The EMI filters contain the balun and bypass capacitors necessary to redirect the CM currents back into the converter. Any currents that take this path will not show up during EMI testing. For the current path to be effective, the cases of the converter and filter need to be connected together by the lowest inductance path possible. 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 if the plane is on the opposite side. In any case, connect the cases together with the shortest and widest possible conductors. The layout of these traces play a major role in the performance of the EMI filter.

Where the power converters and filter are separated by several inches or more, even if the wires are shielded or twisted, radiation from the power line between the two will occur due to common mode currents unless corrective measures are taken. The CM currents are components in phase in both lines such that there is incomplete cancellation of the fields and hence radiated EMI. A balun and bypass capacitor at a remote power converter will rebalance the power lines, 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 10 for details.

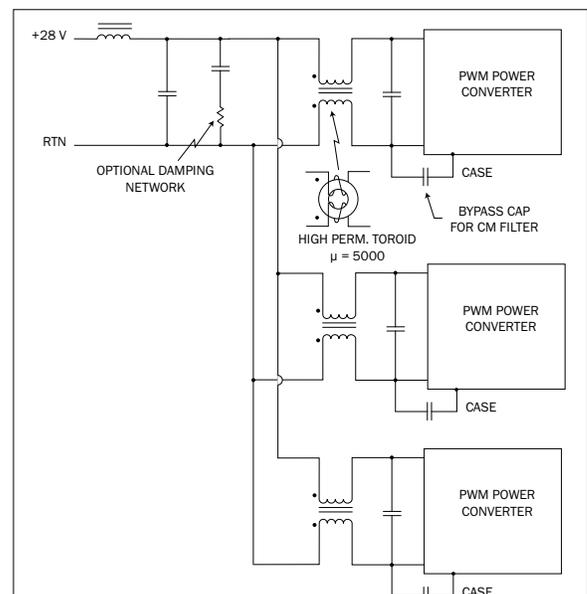


FIGURE 10: MULTIPLE CONVERTERS WITH A SINGLE FILTER

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Where multiple unsynced power converters are used with a single filter, products of modulation (beat frequencies) will occur at the various difference in 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-STD-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 10 on page 7 where a segment wound CM choke and DM capacitor are placed at the input of each converter. The leakage inductance from the CM choke will provide the differential mode (DM) attenuation needed to isolate the converters from each other at their switching frequency.

DESIGNING AN EMI POWER LINE FILTER

Designing a power line EMI filter will involve low pass filter topologies similar to those of Figure 3 on page 3. Referring to Figure 3 on page 3, the upper figure is typical of 28 volt 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 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, as shown in Figure 12 on page 9, rather than adding series damping resistance. The circuit of Figure 3 on page 3, as configured, is a second order common mode and fourth order differential filter.

The lower figure is useful for 270 volt 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. A series damping resistor may be allowable due to the lower line currents at 270 volts as compared to 28 volts. 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 on page 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 the 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 MTR (40) data sheet is 50 mA p-p. The first component value would then be about $50/(2 \times 1.414) = 18$ mA RMS. This is 18000 μ A, or 85.1 dB μ A. At the 625 kHz switching frequency, the MIL-STD-461C limit line value is about 33 dB μ A. If 12 dB of margin is used, the required attenuation at 625 kHz is $(85.1 - 33 + 12) = 64.1$ dB. Note that dB μ A is dB above 1 μ A, 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 of a second order low pass filter is outlined in Figure 11. The power converter generates differential spectral noise represented by a current source (I_N) which terminates across the filter capacitor. The object is to force the ratio of (ac line current I_L) / (noise current I_N) toward zero as a function of increasing frequency. This will readily occur since the inductor impedance, $j\omega L$, increases with frequency while that of the capacitor, $1/(j C)$, 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.

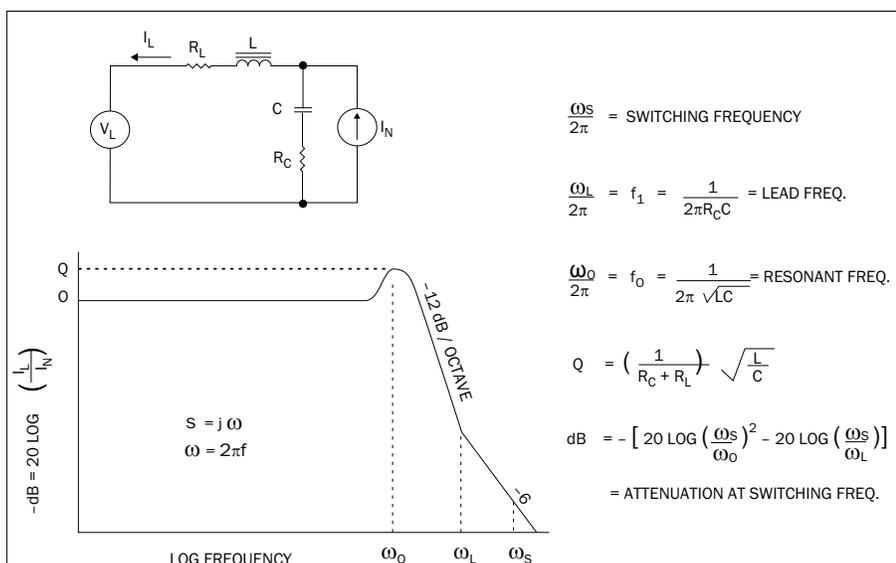


FIGURE 11: LOW PASS FILTER, PRACTICAL FORM

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The series resistances are the inductor's copper and core 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 its ESR plus trace resistance. For a solid tantalum capacitor, this zero will occur 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 circuit of Figure 11 on page 8 can be modeled in the Simulated Program for Integrated Circuit Engineering (SPICE). The results will be very similar to the transfer function shown in Figure 11 on page 8 depending on component values chosen and ESR of components. The ESR of the components and traces are important in determining the attenuation and Q of the filter. At resonance, the filter response is Q times the low pass response. Above resonance, the asymptotic response becomes a function of the ratio of $(\omega/\omega_0)^2$ and reduces to $20\text{LOG}(\omega/\omega_0)^2$, having a slope of -12 dB/octave or -40 dB/decade. Beyond the lead frequency created by the ESR of the capacitor, the response becomes $20\text{LOG}(\omega/\omega_L)$, changing the slope by +6 dB/octave, or + 20 dB/decade. Refer to the graph of Figure 11 on page 8 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 is 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 approach 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 12. 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, $\omega_D = 1/(R_D C_1)$, is set below the resonant frequency such that R_D functions as an ac connected shunt resistor. The ESR of a tantalum capacitor is typically not negligible and needs to be considered as part of R_D and must be able to handle the power if there are any prolonged disturbances at the resonant frequency. Ideally the value of R_D should be large enough to swamp out variations in the capacitor's ESR over temperature and production. Using a ceramic capacitor for the filter element places the lead frequency 2 or more octaves beyond a typical switching frequency, which results in better filter performance. Formulas for calculating the R, L, and C values are given in Figure 12.

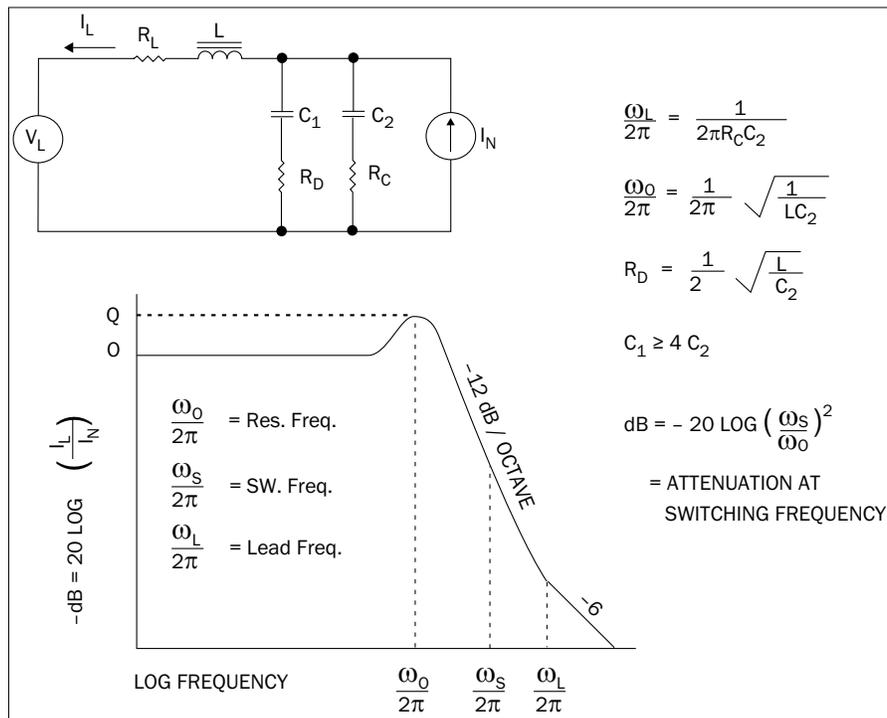


FIGURE 12: LOW PASS FILTER WITH SHUNT DAMPING

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It is difficult to calculate the amount of CM 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, $\mu = 5000$ or more for the balun, and don't worry about the permeability starting to roll off below a megahertz because the material will become very lossy at the higher frequencies, compensating for this effect. Use a CM inductance of 100 μH to 500 μH per winding, and a good ceramic capacitor to bypass the power line and its 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 lower parts solution. When using a 5000 μ core you can expect the leakage inductance (DM inductance) of a segment wound CM inductor to be approximately 1% of the CM inductance. The amount of leakage inductance achieved will depend on how close the +28 volt coil and 28 volt return coil are spaced from each other. For coils that are very close to each other the leakage inductance will be less. If there is good spacing between the coils the leakage inductance will be close to 1%.

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. See Figure 3 on page 3 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. Use inductor values less than 4 μH for the circuit of Figure 11 on page 8, and less than 40 μH for the shunt damping circuit of Figure 12 on page 9, 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. Use SPICE and optimize the models to ensure a low Q. Including the parasitic ESRs in your model is important. 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. In your SPICE model, include any other impedances such as inrush limiters, damping networks, LISNs used for testing, and any other circuitry. Note that low filter impedances require small ratios of L/C in the design of the EMI filter and the converter's internal input filter should be considered as part of the source impedance. Spacing the resonant frequency of the EMI filter away from the resonant frequency of the converter's output filter (not to be confused with converter's input filter) will help meet the impedance requirements.

The higher voltage example at the bottom of Figure 3 on page 3, 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 a lower parts count. 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 μH . The best means of verification is to make a sample and measure it with a short circuit test.

POWER LINE FILTER DESIGN EXAMPLE

The MTR (40) is used in these examples.

Assume that we have an MTR2815S (28 volts in, 15 volts out) single ended, 30 watt, forward converter operating at an 18 watt output load and a requirement to meet the conducted emissions of MIL-STD-461C, method CE03. If the application for the converter called for full load, the calculations would need to reflect the higher power. At 28 volts in and at an efficiency of 84%, the dc input current can be calculated as $I = 18 / (28 * 0.84) = 0.77$ amps. If we assume a 50% duty cycle, close enough for practical purposes, the peak current will be twice the dc, or 1.54 amps. The first spectral component at the 625 kHz clock frequency will be $(0.45)(1.54) = 0.69$ amps RMS. The dB μA level will be $20\text{LOG}(690,000) = 117$ dB μA . The MIL-STD-461C CE03 limit line is at 33 dB μA at 625 kHz. In the absence of any power line filter in the MTR2815S, the required attenuation in dB μA will be $(117 \text{ dB } \mu\text{A} - 33 \text{ dB } \mu\text{A}) = 84$ dB.

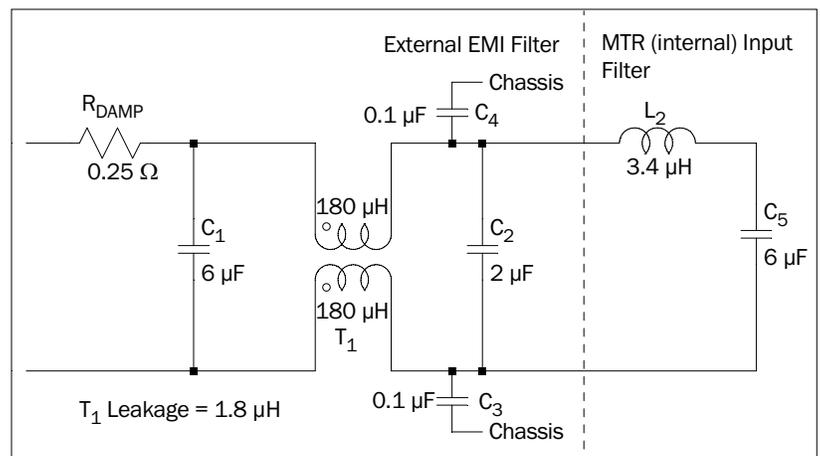


FIGURE 13: INPUT POWER LINE FILTER EXAMPLE (MTR 40)

EMI Conducted Interference

INTERPOINT® PRODUCT APPLICATION NOTE

The MTR2815S has a second order input filter comprised of a 3.4 μH inductor and 6 μF ceramic capacitor. This filter section has a resonant frequency of 35 kHz, and due to the ceramic capacitor, the lead frequency associated with its ESR will be well beyond the switching frequency of 625 kHz. The additional attenuation due to this section is $20\text{LOG}(625/35)^2 = 50$ dB. Then the additional attenuation required from the external line filter is $(84 \text{ dB } \mu\text{A} - 50 \text{ dB}) = 34$ dB.

The filter will include a segment wound balun for the CM filter, with its leakage inductance functioning as the differential inductor. Refer to Figure 13 on page 10, where the MTR2815S internal filter is shown with the proposed external line filter.

In order to keep the magnetics to a practical size, the filter will be implemented using a 0.375 inch O.D. toroid with a permeability of 3000. Ideally a permeability of 5000 should be used but the 3000 permeability was what was immediately available. The measured common mode inductance with 9 turns per winding is 180 μH , and the leakage inductance measured 1.8 μH using segmented windings. A 5000 permeability core will provide higher CM inductance for the same number of turns. Starting with a 2 μF ceramic capacitor, a practical value, we get a resonant frequency for the differential filter of 84 kHz.

$$\frac{1}{2\pi\sqrt{1.8 \mu\text{H} * 2 \mu\text{F}}} = 84 \text{ kHz}$$

The additional attenuation from T_1 and C_2 will be $20\text{LOG}(625/84)^2 = 35$ dB. We predicted above that 34 dB would be required which does not leave any margin. Since the power level is moderate, 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 6 μF capacitor was added after the 0.25 ohm resistor and, since there is approximately 2 μH of inductance in the 461C test set-up, it creates an additional two poles at 46 kHz which will add an additional $20\text{LOG}(625/46)^2 = 45$ dB of margin.

$$\frac{1}{2\pi\sqrt{2 \mu\text{H} * 6 \mu\text{F}}} = 46 \text{ kHz}$$

A 0.1 μF capacitor to bypass the return line to case, and a 0.1 μF cap to bypass the positive 28 volt input line to case places the CM resonant frequency at 38 kHz, more than a decade below the switching frequency which should be adequate. The key for good CM attenuation is the CM capacitors chosen plus creating the lowest impedance possible from the converter's chassis to the CM capacitors and from the CM capacitors to the input pins of the converter. CM noise will dominate at high frequency and beyond about 10 MHz many traces will appear as small inductors which will present a high impedance to the CM capacitors. The trace inductance can also resonate with the CM capacitors which can create peaking at their resonance. If there are problems at

high frequency take a hard look at your layout. The EMI test results of this filter versus the MIL-STD-461C CE03 limit line are shown in Figure 14 and Figure 15, where the margin at the switching frequency is about 20 dB. A 0.015 μF , ceramic cap was also connected from output common of the converter to the converter's chassis in order to achieve these results.

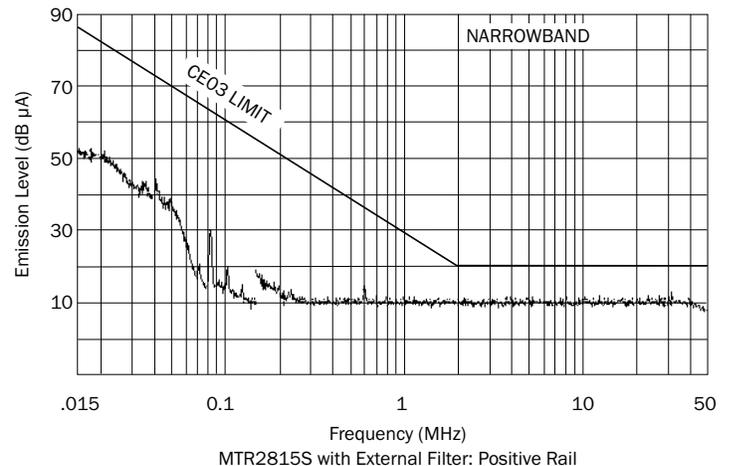


FIGURE 14: SPECTRAL NOISE TEST DATA – HIGH LINE

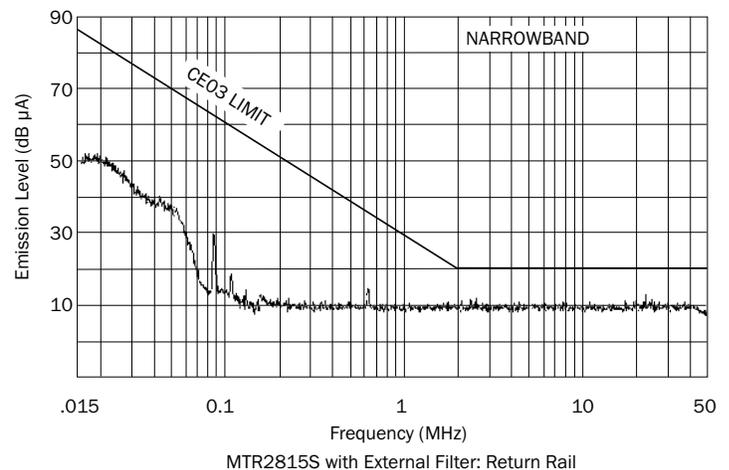


FIGURE 15: SPECTRAL NOISE TEST DATA – RETURN LINE

EMI Conducted Interference

INTERPOINT® PRODUCT APPLICATION NOTE

The SPICE model is shown in Figure 16. The results are shown in Figure 17.

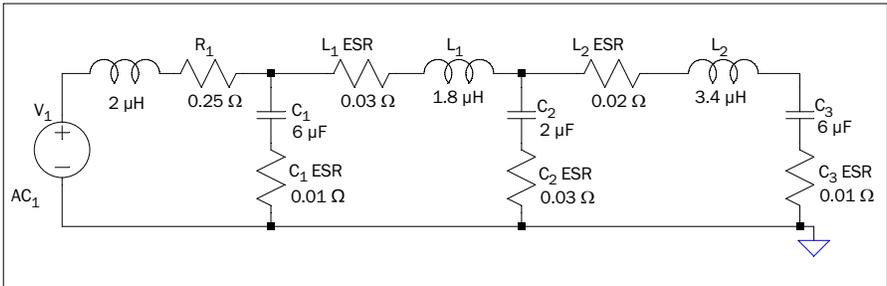


FIGURE 16: SPICE MODEL FOR PROPOSED EMI FILTER AND MTR (40) INTERNAL INPUT FILTER

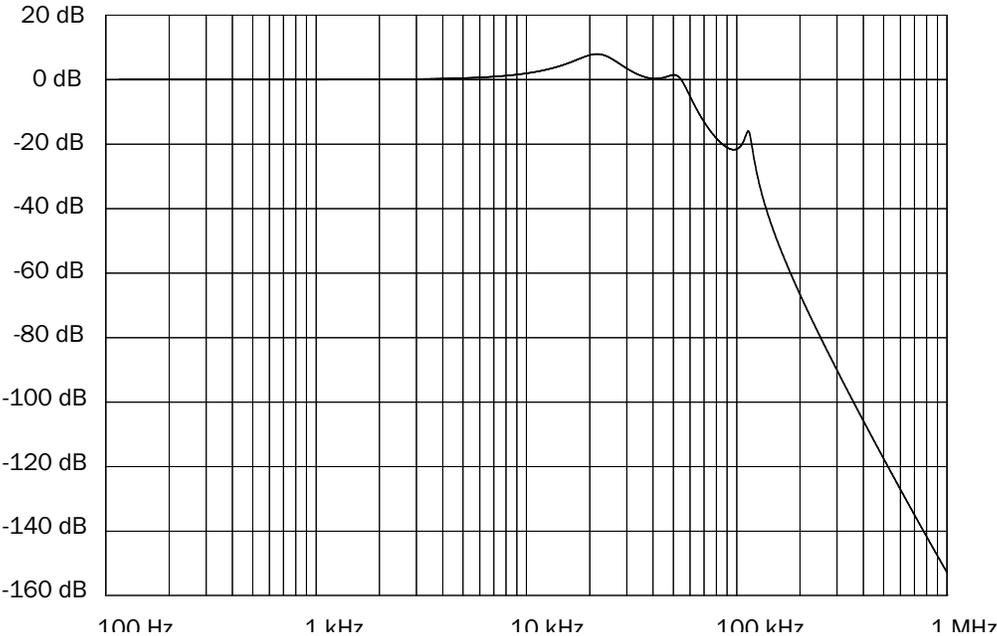


FIGURE 17: TRANSFER FUNCTION OF FIGURE 16

EMI Conducted Interference

INTERPOINT® PRODUCT APPLICATION NOTE

The impedance of the proposed EMI filter combined with the MTR input filter and the inductance of the MIL-STD-461C test set-up is compared to the MTR2815S's (MTR) open loop input impedance. The SPICE model is shown in Figure 18 and the results are shown in Figure 19.

In Figure 18 the current source is set to 1 amp so that the voltage on the Y axis of Figure 19 translates directly to impedance in ohms. The results show the MTR's open loop input impedance is lower than the filter's impedance (source impedance)—close to 5 kHz. However, using the open loop impedance of the converter is extremely conservative and in reality the gains of the feed forward compensation and feedback loop will increase the converter's

minimum input impedance relative to the output filter's resonant frequency so this should not be a problem. If we use the guidelines stated earlier on page four, the source impedance should be more than three times lower than the dc input impedance. In this example the converter is capable of 30 watts, the low line voltage for this converter is 16 volts and the efficiency is 84%.

$$(16 V^2 / (3 * 30 W)) * 0.84 = 2.4$$

Using these numbers requires the source impedance to be 2.4 ohms or less. This value should be very conservative.

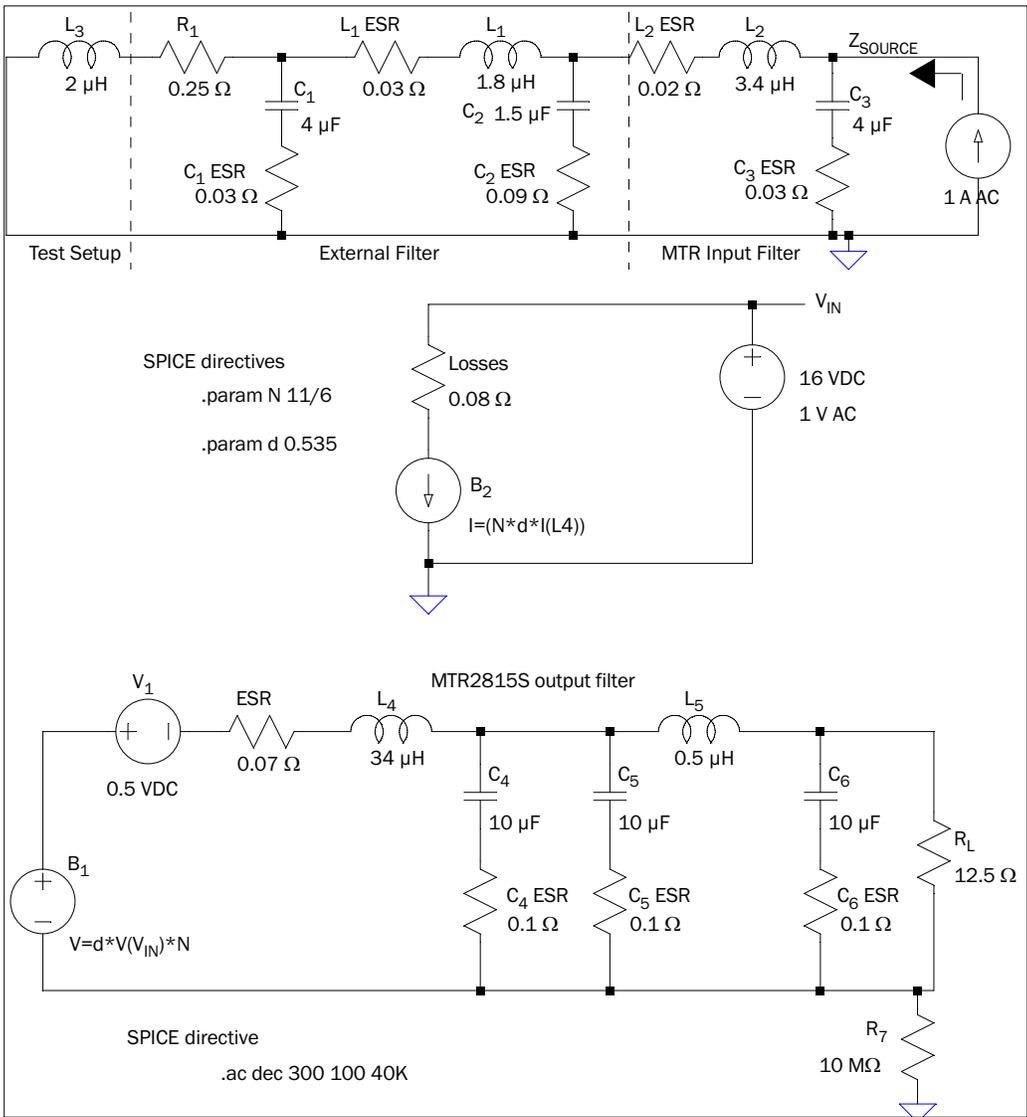


FIGURE 18: SPICE MODEL OF CONVERTERS INPUT IMPEDANCE VS. SOURCE IMPEDANCE

EMI Conducted Interference

INTERPOINT® PRODUCT APPLICATION NOTE

Referring to Figure 19, at close to 5 kHz, the converter’s input impedance reaches its minimum value of less than half an ohm. We can replace this minimum value with 2.4 ohms and, at 5 kHz, the source impedance is well below 2.4 ohms and there is significant margin.

In Figure 18 the capacitors values have decreased from the values shown in Figure 13 on page 10 and figure 16 on page 12 as the value of the ceramic capacitance decreases with applied voltage. Also the ESR has increased as this model is looking at frequencies around 25 kHz where the ESR will be much higher than at the switching frequency of 625 kHz. If any additional load capacitance is connected at the converter’s output, it should be included in the model along with its associated ESR.

Input and output filter values and the transformer turns ratio can be obtained from Applications Engineering to create an open loop SPICE model.

The filter in Figure 13 on page 10 is an example of what can be done with only a few parts. For it to work as demonstrated, the filter should be next to the power converter and built over a small ground plane. Surface mount type capacitors are preferable. Limit any trace ESR or ESL in series with the capacitors if the filter is to function properly.

Applications Engineering
email powerapps@craneae.com
or call 425-882-3100.

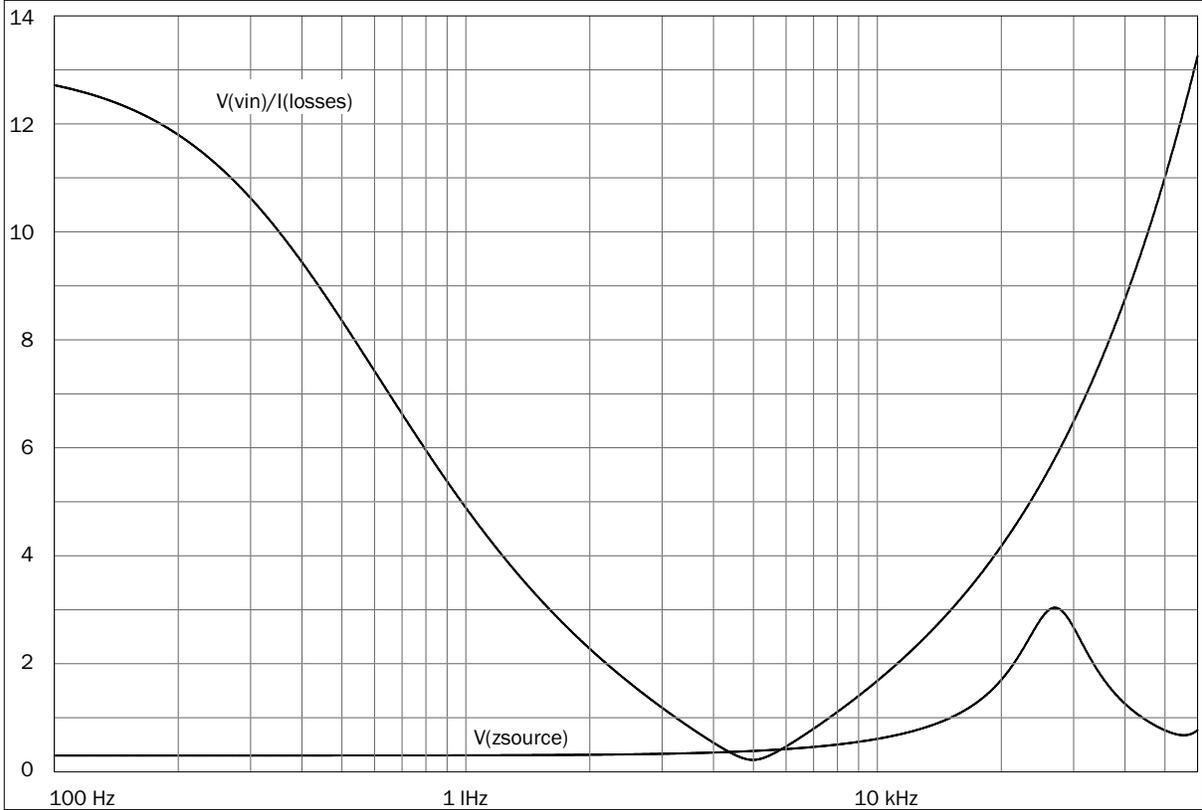


FIGURE 19: RESULTS OF SPICE MODEL IN FIGURE 18

