Measurement and Filtering of Output Noise of DC-DC Converters

INTERPOINT® APPLICATION NOTE

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

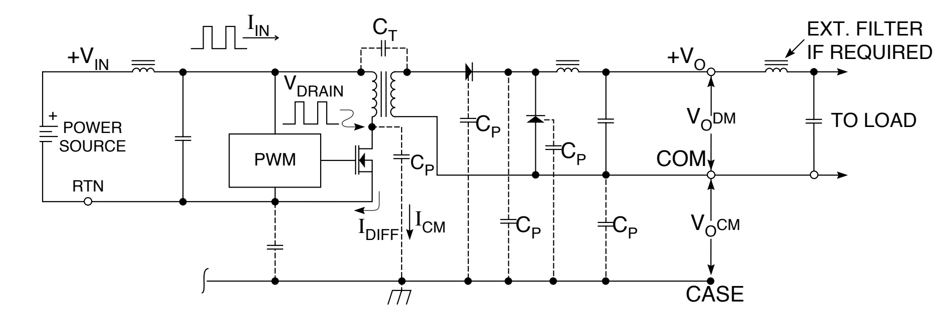
In order to determine if an external output filter is needed for a dc-dc converter, an understanding of the nature of output noise and its proper measurement will be helpful. Differential mode (DM) and common mode (CM) noise are discussed.

Defining & Measuring Output Ripple and Noise

In order to determine if an external output filter is needed for a switching power converter, an understanding of the nature of output ripple and noise and its proper measurement technique will be helpful. The two types of noise, differential mode (DM), and common mode (CM), are discussed in the following paragraphs with the aid of Figures 1 through 45.

Switching power converters are natural generators of narrowband spectral noise which will be present at the fundamental switching frequency and its higher order harmonics. For push-pull type converters it is possible to have a sub-harmonic component at one-half the switching frequency due to imperfect balance. The noise spectrum typically has both DM and CM components while the ripple component is DM. DM noise, also known as normal mode noise, occurs between the converter's output and its output return line. See Figure 1 (V_{ODM}). The DM component of the measurement is comprised of ripple voltage and noise. Figure 2 represents the output ripple voltage while Figure 3 represents ripple voltage along with the noise component. The ripple voltage occurs at the switching frequency for Interpoint converters, and is mainly determined by the changing current in the converter's output filter inductor (for a flyback converter the inductor and main transformer are one), the value of filter capacitors, their equivalent series resistance (ESR) and to a certain extent the load resistance. For a discontinuous fly-back converter the amplitude of the ripple voltage is proportional to the output current. With a forward converter the ripple voltage only varies slightly with current, assuming the converter's output inductor is always operating in the continuous conduction mode and the inductance does not significantly change with current. The noise component of the measurement occurs due to the fast transitions of the power components which cause ringing due to resonance of inductance and parasitic capacitance of the power components and their traces. Since the noise is initiated from the switching transitions of the power components, which is a very small portion of the on time or off time, the noise occurs at a much higher frequency than the ripple voltage. The noise component of the measurement will ride on top of the ripple as shown in Figure 3. The noise portion of the measurement will increase with increasing load current as more energy is stored in any inductance in the path of the current. Once the switching transition starts, or ends, the inductance will resonate with any capacitance in its path.

The CM component of the measurement is part of the noise spectrum and occurs in each of the output lines. If isolated, it can also occur between the EMI reference (the power supply metal case). One source of CM noise is due to CM currents being pumped through parasitic capacitances between the power



To measure differential output noise, first eliminate common mode noise (V_{OCM}).
1.) Connect input return and output common to case.
2.) Connect input return to case and a capacitor ≥ 100 X C_T between input return and output common (~10000 pF).
3.) Connect a cap ≥ 100 X C_T between input return and output common (~10000pF).

Note: 1.) will be most effective; 3.) least effective.
C_P's ~ <1 pF
C_T ~ 50 to 100 pF

Figure 1: Power Converter DM & CM Noise Model
Single-Ended Forward Converter Power Train

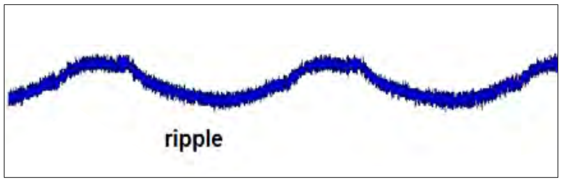


Figure 2: Output Ripple

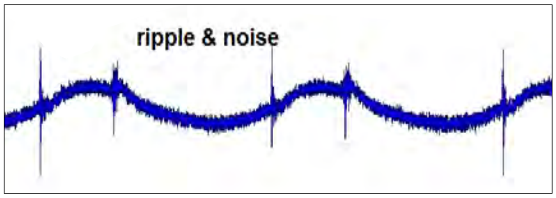


Figure 3: Output Ripple and Noise

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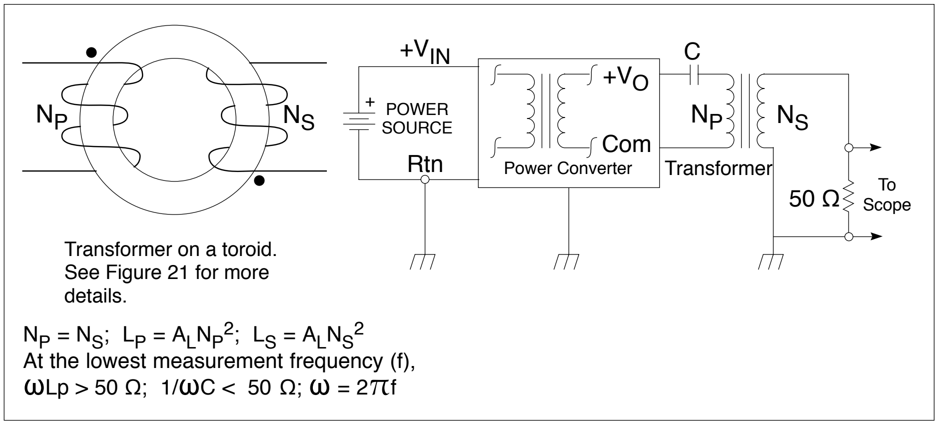

Figure 4: Using a Transformer to Suppress CM Noise

train components (and traces) and the case. These parasitic capacitances, generally less than a picofarad, are a function of the dielectric constant of the ceramic substrate, which is on the order of 8, and the overlapping surface areas of the power components (and their traces) and the chassis. The substrate is back metallized and soldered to the metal base of the converter (header) with the power train and other circuits on the substrate top side. The parasitic capacitances are then proportional to the product of dielectric constant, the overlapping surface areas and inversely proportional to the substrate thickness.

Although the power train components generate CM current components into the case due to the high dv/dt across their parasitic capacitances to the case, the CM output noise problem is mainly due to the transformer inter-winding capacitance, C_T. Refer again to the model of Figure 1, C_T is distributed between the primary and secondary windings, which forms a high ratio capacitive divider with the output-to-case parasitics. Much of the high level pulses at the primary side power MOSFET drain appear as secondary noise common to both output lines. For an input line voltage of 28 volts the peak-to-peak voltage excursion on the power switch drain will be approximately 50 volts. In the absence of any CM suppression, a large AC voltage will be commutated to all output lines of the secondary winding by the inter-winding capacitance, C_T, and is referenced to the input power return. A magnitude of 20 volts peak to peak is a typical amplitude seen at the output when referenced to the input power return. This CM noise voltage can be suppressed by forming a capacitive divider with the inter-winding capacitance, C_T, and implemented by connecting a capacitor from output common to input power return. If C_T is 50 picofarads, then connecting a 50 nanofarad ceramic capacitor from output common to input return will reduce the CM noise from 20 volts to about 20 millivolts.

The DM ripple will generally have a magnitude of less than 50 millivolts peak to peak for forward converters, and typically more than 50 millivolts peak to peak for fly-back converters. However, if the CM noise is not suppressed as suggested above, it can have a magnitude of several volts, and may appear as DM mode noise due to poor CM rejection of the noise measurement equipment. CM noise is generally not important in most system applications but can cause misleading and excessive DM noise measurements. An oscilloscope is usually used to make noise measurements and may have a DM bandwidth of 100 MHz or more, but will usually have poor common mode rejection over this bandwidth. The power converter CM output noise spectrum will have a bandwidth of several tens of MHz, and the magnitude can be significantly higher than the DM spectrum. An accurate DM noise measurement will usually require that the CM noise spectrum be suppressed prior to making the measurements. One method of suppression is to connect both input and output commons to the case and use the case as a single point ground. This is practical for bench measurements but may not be practical in the system. A good quality ceramic capacitor from input common to case and from output common to case, should suffice if the unit is mounted on a PC card and direct connections are not possible. The capacitors should be greater than 10,000 pF. The exact value can be determined experimentally. Any traces leading to, or away from the capacitors, should be as short and wide as possible as trace inductance will degrade the high frequency performance of the capacitors. Further, if the circuit is not well damped the Q created by the capacitance and trace inductance can amplify any noise at the LC resonant frequency. Ideally the capacitors should be surface mount ceramic capacitors. This should reduce the CM spectrum magnitude but may increase the DM noise spectrum, which can easily be filtered by a small value DM ceramic capacitor connected directly across the converter's output terminals.

A transformer, shown in Figure 4, can also be used to reject CM noise for measurement purposes. The transformer should have minimal inter-winding capacitance and a 1:1 turns ratio. A toroid, as shown, with segmented windings or a pot core with separated windings can be used. Excellent results for DM noise measurements on the bench can be achieved with the transformer of Figure 4. The transformer should be terminated in a low impedance of 50 ohms.

The DM output ripple and noise is measured between the two output terminals. Refer to Figure 1. The measurement must be carried out at the pins using short probe leads to reduce the pickup of radiated noise. If an oscilloscope with a differential bandwidth of 100 MHz is used, along with a 10 x probe, it is

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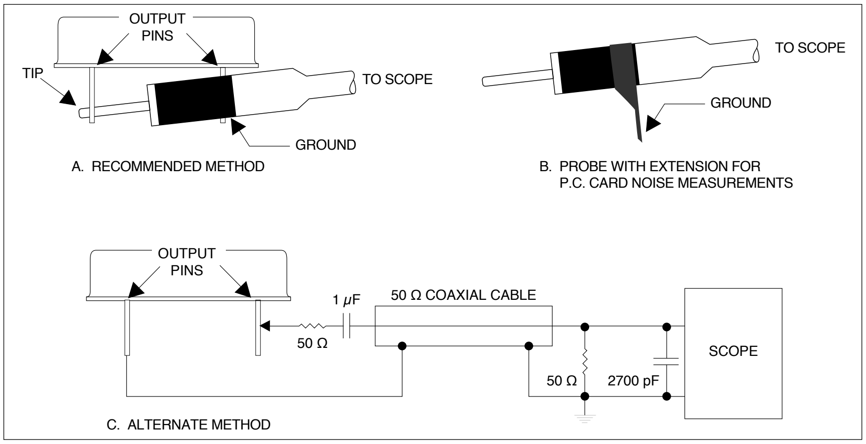

FIGURE 5: NOISE MEASUREMENT METHODS

important to be sure the probe connections to the test sample are good over this bandwidth. The probe ground lead, which is about 6 inches long, has a lead inductance of about 100 nanohenries (close to 20 nH per inch) This inductance equates to 10 ohms reactive at 16 MHz, and increases at a 6 dB/octave rate with frequency. The probe ground is not an acceptable ground connection for noise measurements, but rather is an excellent antenna for radiated noise pickup as well as a high impedance for any circulating ground loop currents to flow through (exactly what we don't want). To improve the situation, remove the ground lead and probe clip and use the probe tip and barrel ground lead as shown in Figure 5, "A. Recommended Method". To make noise measurements on a PC card where the power converter is installed, add the shortest possible lengths of wire to either the tip or the barrel ground to reduce series lead inductance. Using a rectangular cross section conductor for this purpose, as shown in Figure 5 "B. Probe with Extension", is preferred as it reduces lead inductance as compared to round wire.

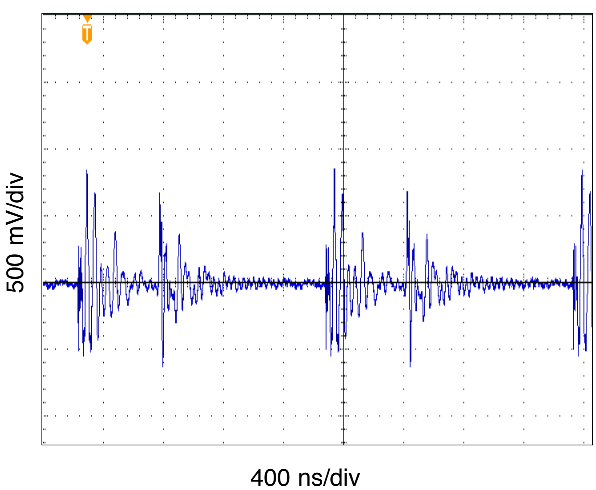

FIGURE 6: OUTPUT MEASURED WITH 6 INCH SCOPE PROBE

Alternately, a 50 ohm coaxial cable can be used in lieu of the 10 x probe. The same precautions about short leads also apply here. This setup, shown in Figure 5 "C Alternate Method", will require that the scope reading be multiplied by 2 as the two 50 ohm resistors create a voltage divider. The 50 ohm resistor shown at the scope end of the cable can be a 50 ohm terminal that interfaces between the cable and the scope, or it could be the created by the 50 ohm scope setting. A 2700 pF capacitor can be used to reduce scope bandwidth to 2 MHz to allow observing the lower frequency components only.

Figures 6 through 16 illustrate the problems with

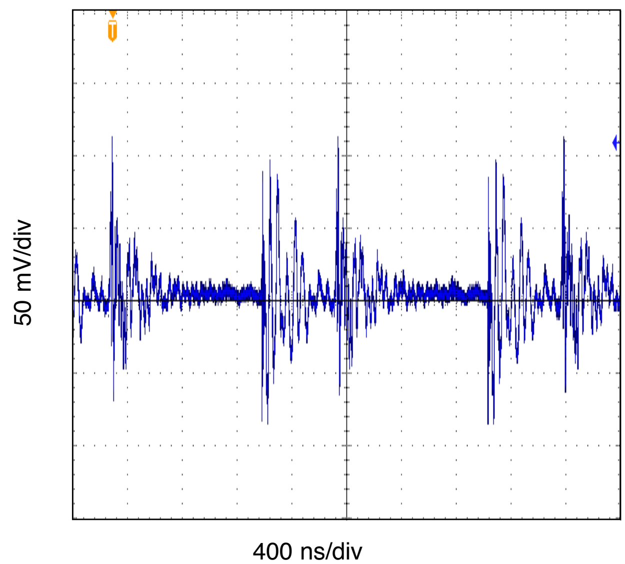

FIGURE 7: OUTPUT MEASURED USING PIG TAIL IN FIGURE 8

common measurement techniques. These scope captures measure the output noise of a single 15 volt output, 30 watt, dc to dc converter (MTR2815S), operating at full load. The scope was set to full bandwidth (100 MHz) and unless stated otherwise a 500 MHz scope probe was used to make the measurements. A small copper clad board was made with the +15 volt trace and its return trace running side by side. The board was soldered to the converter's output pins and the measurements were taken on the board. The copper clad board allows us to minimize the distance between the two measurement points, and allows soldering of ceramic capacitors close to the converter's output terminals. The measurements on the board were made as close to the converter's output pins as possible. The vertical voltage scale on the scope has been adjusted in the different scope captures to

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FIGURE 8: PIG-TAIL USED IN FIGURE 9 MEASUREMENT

allow the waveforms to fill the screen as much as possible. This should be considered when comparing results.

The measurement in Figure 6 uses the six inch ground lead that is standard with most scope probes. In this case the ripple voltage is not visible and only the noise component is present. For this measurement the ground lead was kept close to the scope probe to help reduce the noise. The output noise measures 1.47 volts peak to peak. Figure 7 shows measurement results using a pigtail style ground lead (shown in Figure 8) that

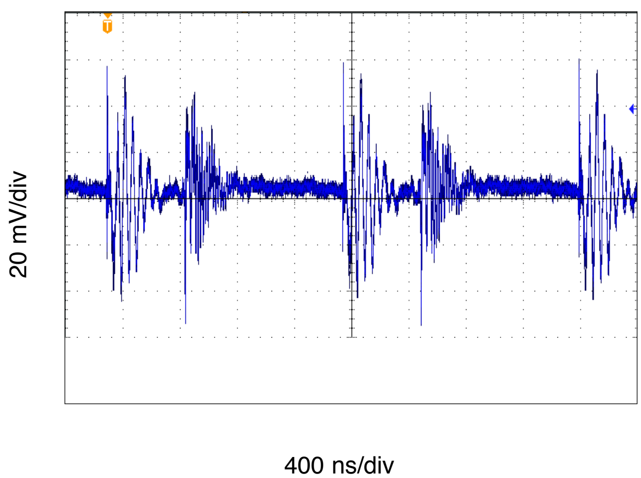

FIGURE 9: OUTPUT MEASURED USING BARREL OF SCOPE PROBE AS GROUND

is commonly used in the industry. The results show 190 millivolts peak to peak, which is much better than the 1.47 volts peak to peak results using the long ground lead, but still somewhat misleading. Figure 9 uses the barrel of the scope probe as the ground connection. A copper tab was also soldered to the ground trace of the copper clad board so that the connection from the ground trace to the barrel of the scope can be made with very low inductance and minimizes the loop. This represents the technique shown in Figure 5 “B. Probe with Extension” and with this technique the measurement has decreased from 190

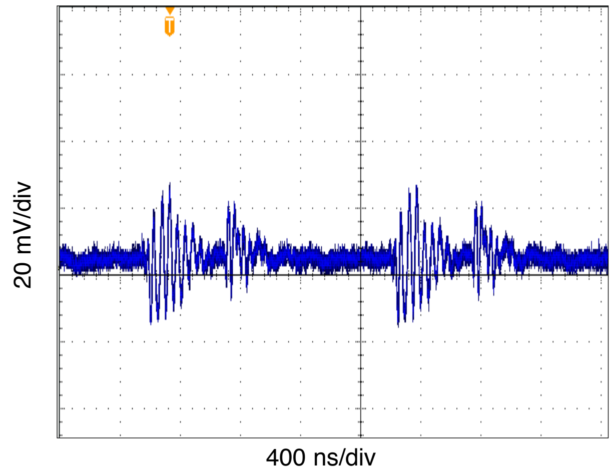

FIGURE 10: GROUND TRACE CM NOISE

millivolts peak to peak in Figure 7 to 112 millivolts peak to peak in Figure 9.

Since no external components/connections were used to suppress the CM noise, these measurements reflect both CM and DM noise. To get a feel for the CM component of the noise, a measurement was taken with the barrel of the scope probe touching the copper tab (the tab soldered to the ground trace), and the tip of the scope probe also touching the ground trace.

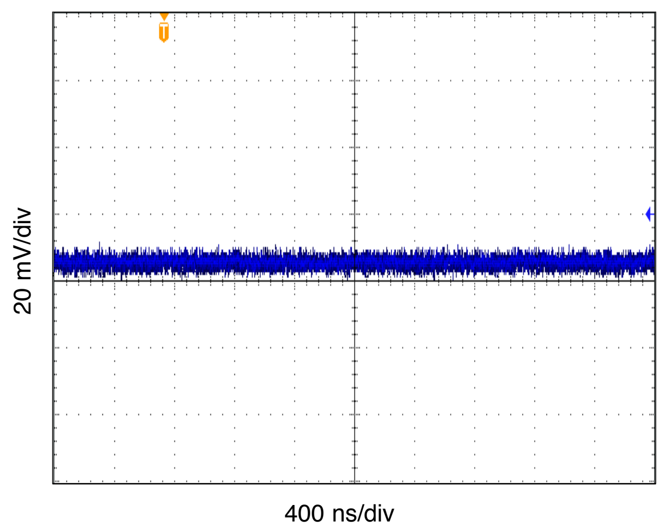

FIGURE 11: GROUND TRACE CM NOISE SUPPRESSED

In order to optimize the measurement, the distance between the measurement points was made as small as possible. Since we are only measuring the ground trace, no noise should be present. The results, in Figure 10, show the presence of switching and the noise measuring 41.6 millivolts peak to peak. This is due to the CM noise. The next step would be to suppress the CM noise then retake the data. In order to suppress the CM noise

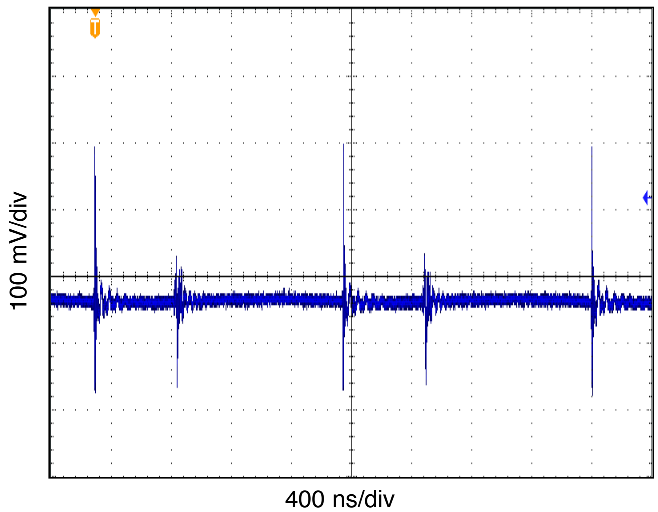

FIGURE 12: OUTPUT CM NOISE

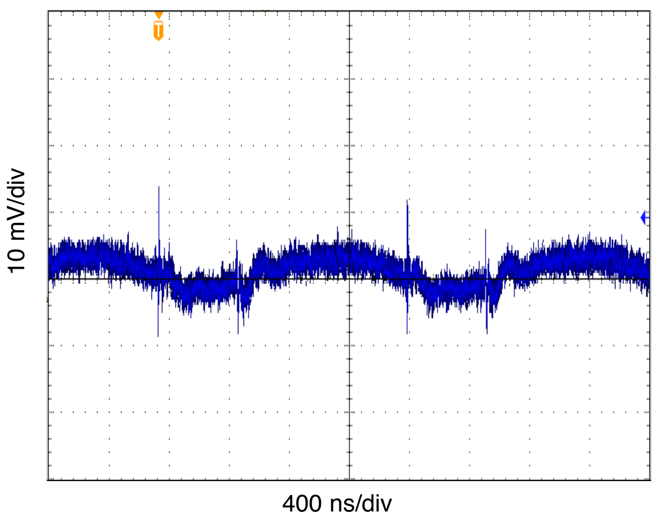

FIGURE 13: OUTPUT CM NOISE WITH DM CAP

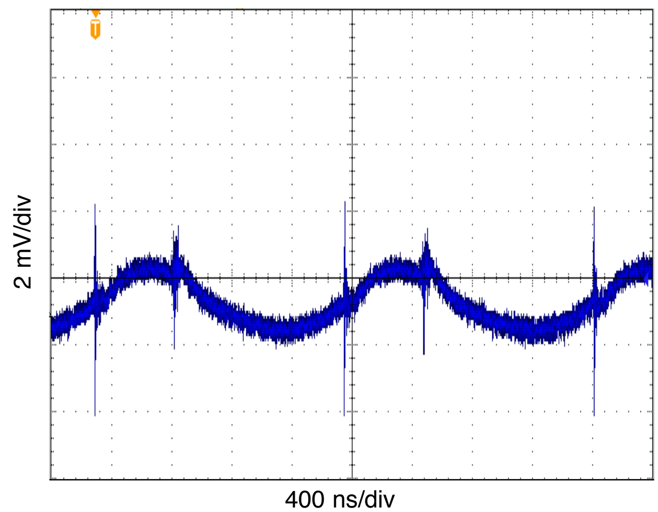

FIGURE 14: OUTPUT WITH 2 MHZ PROBE

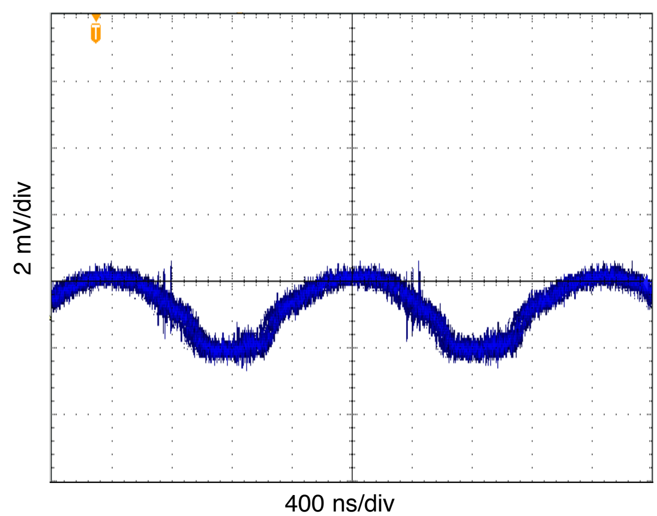

FIGURE 15: OUTPUT WITH 2 MHZ PROBE AND DM CAP

a 0.1 uF capacitor was connected from input common to the converter's chassis and a 0.1 uF capacitor was connected from output common to the converter's chassis. The connections to chassis were made with short wide traces in order to minimize the inductance in the path. This creates a low impedance from the secondary side of the transformer's inter-winding capacitance to the primary. The measurement of Figure 10 was retaken and the results are shown in Figure 11. The switching noise is no longer present and the scope results do not change much even if the converter is not powered. These results, as well as all further captures in Figure 12-16, use the barrel of the scope probe for the ground connection unless stated otherwise.

With the addition of 0.1 uF capacitors connected from input common to chassis and from output common to chassis, the DM measurement across the converter's output was retaken. The results are shown in Figure 12. Now the CM currents are suppressed during the measurement but the DM noise has increased to 364 millivolts peak to peak. The noise occurs at a very high frequency which is easily filtered with a good quality DM ceramic capacitor connected directly at the converter's output terminals. Figure 13 shows the results of the same measurement with a 0.1 uF, ceramic capacitor, connected from the converter's +15 volt rail to the 15 volt return. With a 0.1 uF cap the noise is reduced from 364 millivolts peak to peak in Figure 12 to 20.4 millivolts peak to peak in Figure 13. The 0.1 uF capacitor significantly reduced the high frequency noise but does very little for the lower frequency ripple voltage which is now prominent. In reality any noise sensitive system will have many DM, ceramic by pass capacitors, which should reduce the noise

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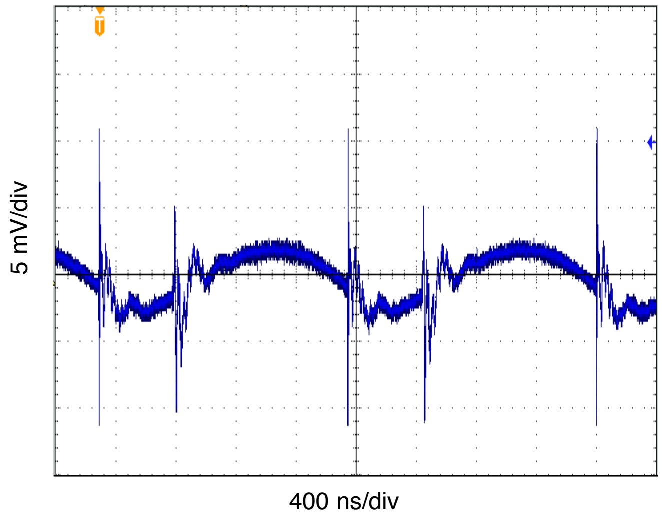


FIGURE 16: OUTPUT WITH 15 MHZ PROBE

to a low level. Having a surface mount technology (SMT) ceramic capacitor directly at the output pins of the converter will keep the high frequency current local, which will help prevent conducted noise from becoming radiated noise.

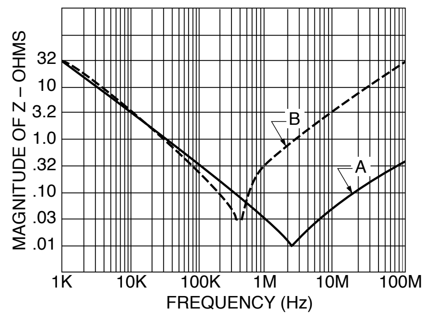
In order to verify the datasheet parameters the bandwidth of the measurement would need to be limited to the datasheet specifications. For many Interpoint converters this will be 2 MHz. Figure 14 shows the results with using a 2 MHz scope probe similar to the one shown in Figure 5 "C: Alternate Method". For this measurement no DM capacitors are used at the converter's output, but since this is a DM parameter, both input and output common are connected to chassis via 0.1 uF capacitors in order to suppress the CM noise. The results show 6.3 millivolts peak to peak but since the scope probe divides the measurement by a factor of two the true reading would be 12.6 millivolts peak to peak. Decreasing the length of the connections of the 2 MHz probe reduced the noise further. As a further step the 0.1 uF DM capacitor was added back across the converter's output rails very close to the converter's output pins. The results are shown in Figure 15. Now the noise has decreased from 12.6 millivolts to 6.6 millivolts (3.28 millivolts x 2).

Many times when measuring very low output noise, using a 10 x probe will not show enough detail and it will be hard to determine the actual ripple and noise. In this case a 1 x probe can be used. Figure 16 shows the results with a 15 MHz, 1 x probe, being used to measure output noise. In this measurement output common and input common are bypassed to chassis with 0.1 uF ceramic capacitors. Also a 0.1 uF DM capacitor is across the output rails of the converter. Since these are the same conditions as Figure 13 the waveforms in Figure 16 and Figure 13 can be compared to each other. In Figure 13 the shape of the ripple, and the noise spikes are still fairly well defined. If the ripple and noise were lower the 1 x probe would show more detail. Figure 16 is set to 5 millivolts / division but with a 1 x probe this could have been reduced to 1 millivolts / division which would have expanded the vertical scale by a factor of five. This will become much more relevant later in our design example.

FILTERING OF DIFFERENTIAL MODE OUTPUT NOISE

Filtering, if required, can be as simple as adding an external DM capacitor as illustrated above. This may work where the higher frequency components are the problem. Only high quality, low ESR, SMT ceramic capacitors are recommended. They should be connected to output common, close to the pin, to form a single point ground. The other end of the capacitor should be as close to the positive output pin as possible. Use the widest, shortest, copper traces possible and take the load connections from the capacitor termination. This is important because the capacitor becomes useless if appreciable ESR and/or ESL are added in series with it. Figure 17 shows an example of the impedance vs. frequency of a good 4.7 μF ceramic capacitor, and how it can begin deteriorating at frequencies below 1 MHz due to additional trace ESR and ESL. The increase in these parameters is what can be expected with about 2 inches of small wire, or thin traces in series with the capacitor. If using a leaded capacitor the leads will also contribute to the inductance. This is where an SMT capacitor has an advantage. It is important that the capacitor be of low impedance over the bandwidth of interest because it has to work against the power supply output impedance which will also be low.

Second-order low-pass LC filters will provide better and more predictable attenuation as a function of frequency. They are also more forgiving to imperfect layouts because the inductor is in



$$|Z_C| = \sqrt{\left(\omega L - \frac{1}{\omega C}\right)^2 + R^2}$$
$\omega = 2\pi f$

A: Capacitor only:
C = 4.7 μF
ESR = .01 Ω
ESL = .001 μH

B: With wire ESR and ESL added:
C = 4.7 μF
ESR = .03 Ω ESL = .05 μH

FIGURE 17: IMPEDANCE VS. FREQUENCY WITH A 4.7 μF CERAMIC CAPACITOR

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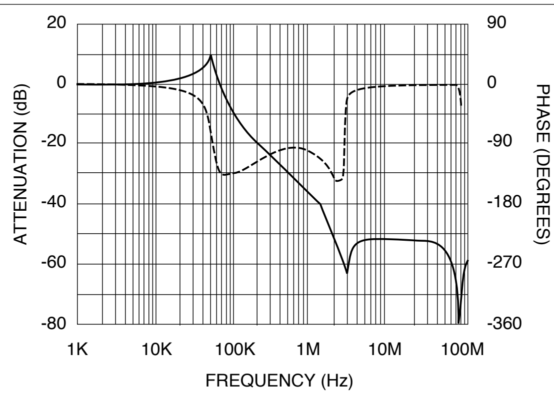
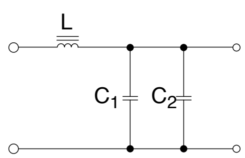
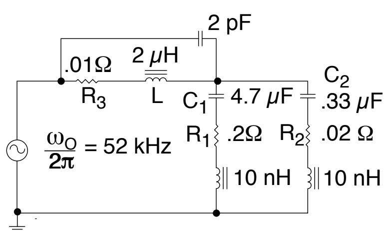


2ND ORDER FILTER WITH PARASITIC ELEMENTS



COMPUTER MODEL WITH PARASITICS



$$f_o = \frac{\omega_o}{2\pi} = \frac{1}{2\pi}\sqrt{\frac{1}{L\,(C_1 + C_2)}}$$

$$Q_o = \left(\frac{1}{R_1 + R_3}\right)\sqrt{\frac{L}{C_1 + C_2}}$$

L = 2 μH; 4½ T #25 AWG on 7X4 mm pot core with air gap for AL = 100 nH/Turn^2
C_1 = 4.7 μF tantalum
C_2 = 0.33 μF ceramic

In making a good filter, the quality of the electrical connections is everything. For the capacitors, the leads must be short. Excessive lead length will render the filter useless. Long leads in series with the inductor are permissible within reason.

L will carry 3 amps at B = 2.1 kgauss. L can also be wound on a small low permeability toroid. Such as 6T of #26AWG on Magnetic Inc. MPP Core having an O.D. of 0.18 inches. MAG. Inc. PN 55147-AY

FIGURE 18: OUTPUT FILTER EXAMPLE

series with the output. The inductor can also make the layout of the filter easier since the trace and wire length in series with the inductor will not compromise filter attenuation. Keep the capacitor trace lengths as short as possible and make the load connections at the points where the capacitor is terminated. An example of a simple second-order LC filter is shown in Figure 18. The resonant frequency of 52 kHz is determined by the 4.7 μF tantalum capacitor and the 2 μH inductor. The 0.33 μF ceramic capacitor is used to compensate for the zero due to the tantalum capacitor's ESR and hence improve high frequency rejection. This type of filter has been used with Interpoint power converters in the past with excellent results. The actual attenuation and phase characteristics will be close to that of the computer model if it is designed and laid out as described in Figure 18 notes.

The simple example of Figure 18 has a disadvantage. The Q at resonance can be difficult to control because it is an inverse function of the ESR sum for L and C. Due to additional power loss and the effects on load regulation, adding resistance in series with the inductor to reduce the Q is not practical unless the output current is small.

An alternate filter design, shown in Figure 19, does not have this disadvantage and also has improved high frequency response. The example shown uses approximately the same capacitor values shown in Figure 18, but a somewhat larger inductor value. The resonant frequency is a function of the 21 μH inductor and the 0.47 μF ceramic capacitor. The 4.7 μF tantalum capacitor and series resistor function as an AC load or parallel damping network.

An external LC filter can affect load regulation. Using the smallest L value possible and keeping the inductor ESR small will help. The LC filter also acts as a series resonant circuit across the power converter output terminals and can affect the stability of the converter if the filter is under damped and the filter's resonant frequency is inside or close to the control loop bandwidth. Setting the filter's resonant frequency above 30 kHz should suffice for any Interpoint power converter. If the filter's resonant frequency is within or close to the converter's bandwidth, keeping the Q below 2 should be adequate for most Interpoint converters.

In reality the system load will most likely have many ceramic capacitors, and/or larger bulk capacitors as part of the load requirement. This capacitance will inherently become part of the filter. If the system capacitance dominates it can leave the 4.7 uF tantalum capacitor with a series damping resistor ineffective. Any load capacitors need to be included as part of the filter model when considering the attenuation and the Q of the filter. If significant bulk capacitance is part of the system load and it has significant ESR, or implements a series damping resistor, this can dampen a filter very well and in many cases can improve the stability of the converter. Sometimes the added capacitance can lower gain which can degrade the load transient response of a converter. On the other hand an un-damped filter that has a high

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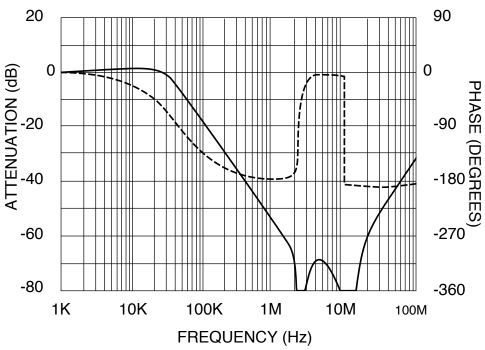

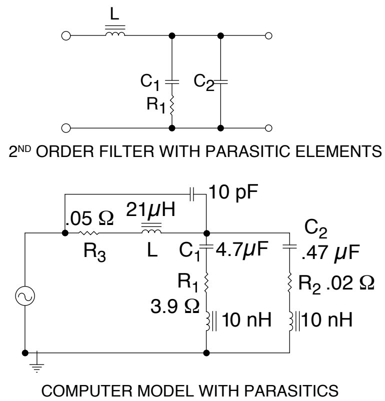

L can also be wound on a toroid such as 14T of #27AWG on MAG. Inc. Kool Mu PN 77270-A7 having an O.D. of 0.29".

L = 21 μH; 14½T #27 AWG on 9x5 mm pot core with air gap for A_L = 100 nH/Turn²
C_1 = 4.7 μF tantalum cap.
R_1 = 3.9 Ω in series with C_1 ESR.
C_2 = 0.47 μF ceramic cap.

$$f_o = \frac{\omega_o}{2\pi} = \frac{1}{2\pi}\sqrt{\frac{1}{LC_2}\left(\frac{R_1 + R_3}{R_1 + R_2}\right)}$$

$$C_1 > 4C_2$$
$$R_1C_1 >> \sqrt{LC_2}$$

$$Q_o = \frac{R_1}{\omega_o L} = R_1\sqrt{\frac{C_2}{L}}$$

L will carry 2 amps at B = 2.6 kgauss

FIGURE 19: WELL DAMPED OUTPUT FILTER EXAMPLE

Q can amplify any disturbances that occur at the filter's resonant frequency.

When using a converter with remote sense leads, and implementing an external LC filter, it is not recommended to connect the sense leads after an external filter inductor. This can add an extra two poles to the converter's control loop and cause instability.

FILTERING COMMON MODE NOISE

Common mode noise can be of importance in analog circuits where sensitive devices such as high resolution A/D converters, high gain instrumentation amplifiers, and/or video displays are involved. Common mode output noise represents noise components which are in phase in both output lines with respect to a common reference. This can lead to radiated and conducted noise problems if not corrected. The solution is to use case bypass capacitors from both input common and output common to case, and also a balun to place a high CM impedance in the input lines. The input bypass capacitor directs the case CM currents back into the return line, reducing radiated noise. A bypass capacitor can be replaced by a direct connection where permitted in the equipment spec, and the balun may not be needed. A combined DM and CM filter, using the low pass section of Figure 18 and the CM filter just discussed, is shown in the example of Figure 20. Design of the balun is covered in the paragraphs which follow.

LOW NOISE OUTPUT FILTERS FOR DC-DC CONVERTERS

Very low noise output filters can be implemented using baluns in both the input and output lines of dc-dc power converters. These are particularly effective where long lines or traces run from the power converter to the load. These filters use baluns of the types shown in Figure 21 and Figure 23. Single output power converters use a two winding balun, with a three winding balun used for dual output converters. The balun leakage

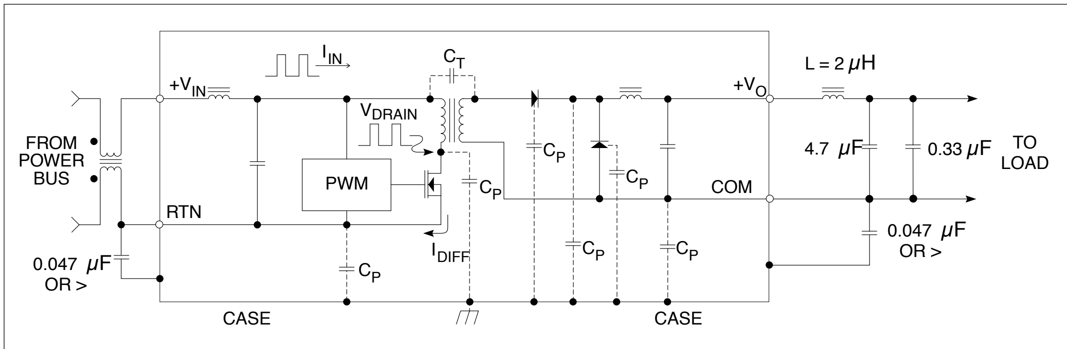

FIGURE 20: POWER CONVERTER – DM AND CM NOISE MODEL WITH EXTERNAL FILTERS

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$N = N_1 = N_2$

$L_{CM} = L_1 = L_2 = A_L N^2$

$$L_{CM} = \frac{4\pi\mu A N^2 \times 10^{-7} \text{ Henries}}{\ell_M}$$

μ = Incremental permeability
A = Cross section area in meters^2
ℓ_M = Mean circumference of toroid in meters
N = Number of turns
L_ℓ = Leakage inductance

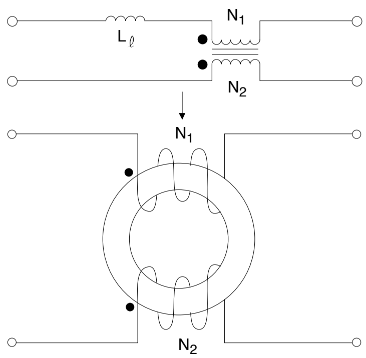
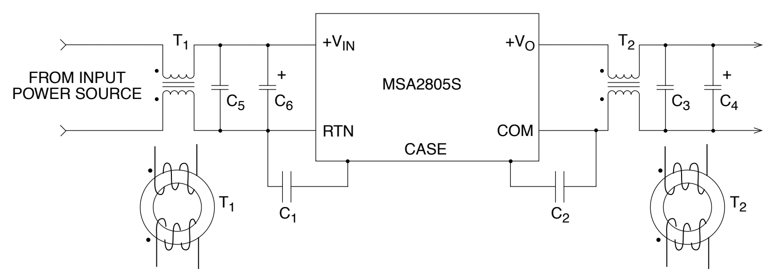


FIGURE 21: BALUN WITH TWO SEGMENTED WINDINGS

inductance, created by the separated coils, is used for the DM filter. The CM filter is implemented first with a ceramic bypass capacitor which forms a capacitive divider with the primary to secondary transformer inter-winding capacitance. Secondly the balun forms a high frequency CM filter which works with the stray capacitance of the load side of the balun. This can be particularly effective where the load is on the other end of a pair of long wires.

A low noise output filter for an isolated single output MSA2805S power converter is shown in Figure 22. Here the filter is implemented using baluns in both the input and output lines. The input balun and ceramic bypass capacitor are used to attenuate CM input currents. This can also be accomplished to a limited degree by shorting the input return to the case where permitted in the equipment specification. Shunt capacitors (C_3, C_4 and C_5, C_6) can be used to work against the balun leakage inductance to create a differential filter section. Both baluns are wound on high permeability ferrite toroids, with a 5000 perm core being the most practical, and are segment wound to create leakage inductance, each winding covering less than 180 degrees of core circumference. When segment winding is implemented, the separation of the windings prevents a short that may otherwise occur if the toroid were bifilar or trifilar wound. With a bifilar wound core the two windings will be touching in many places which could create a short that could bring down the rail.

When wound with equal turns on each winding, the net ampere turns in the core will always be zero when used as shown in Figure 21 and Figure 23. A solid tantalum capacitor, paralleled with a ceramic capacitor for improved high frequency performance, is used for the DM filter. This is the DM filter example of Figure 18, where the leakage inductance serves as the DM inductor. The leakage inductance is in the air, not the core, and therefore cannot be saturated. The circuit of

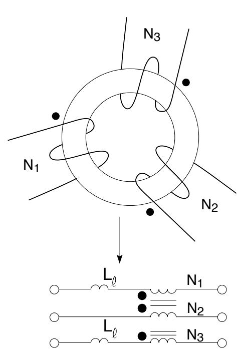


T_1 = 17T #31AWG/ winding on Ferroxcube TC 5.8/3.1/1.5-3E27 or Magnetics, Inc. J-40601-TC
C_4 = C6 = 6.8 microfarad tantalum.
C_3 = C5 = 0.27 microfarad or > ceramic cap.
C_5 & C6 are for optional diff. input filter.
T_2 = 15T / winding #28 AWG on Ferroxcube TC9.5/4.8/3.2-3E27 toroid – μ 5000 C_1 = 0.033 microfarad or > ceramic cap.
C_2 = 0.033 microfarad or > ceramic cap.

FIGURE 22: LOW NOISE OUTPUT FILTER FOR SINGLE OUTPUTS

Figure 22 was built as described on copper clad vector board using readily available components, and layout techniques to minimize trace ESL and ESR in series with the capacitors. The result was a DM and/or CM output noise level of a few millivolts peak to peak. The measurement bandwidth was 100 MHz.

A low noise output filter circuit for a dual MSA power converter is shown in Figure 24. This is similar to Figure 22 except for the three winding balun and the two additional DM output capacitors. Here, the balun is also segment wound to create leakage inductance, each winding occupying less than 120 degrees of core circumference. When phased as shown, and with equal turns on each of the three windings, the net ampere turns in the core will always be zero regardless of load



$N = N_1 = N_2$

$L_{CM} = L_1 = L_2 = A_L N^2$

$$L_{CM} = \frac{4\pi\mu A N^2 \times 10^{-7} \text{ Henries}}{\ell_M}$$

μ = Incremental permeability
A = Cross section area in meters^2
ℓ_M = Mean circumference of toroid in meters
N = Number of turns
L_ℓ = Leakage inductance

The balun can be wound on a potcore by winding L_1 first, L_2 over L_1, and L_3 last

FIGURE 23: BALUN WITH THREE SEGMENTED WINDINGS

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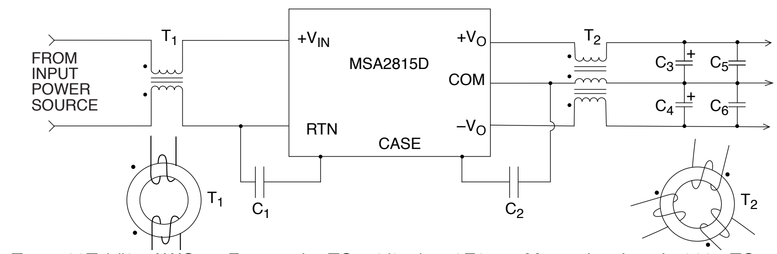

T_1= 20T / #31 AWG on Ferroxcube TC 5.8/3.1/1.5-3E27 or Magnetics, Inc. J-40601-TC
C_3 = C4 = 6.8 microfarad tantalum
C_5 = C6 = 0.27 microfarad ceramic
T_2 = 10T #28AWG on Ferroxcube TC9.5/4.8/3.2-3E27 or Magnetics, Inc. J-41003-TC
C_1 = 0.033 microfarad or > ceramic cap.
C_2 = 0.033 microfarad or > ceramic cap.

FIGURE 24: LOW NOISE OUTPUT FILTER FOR DUAL OUTPUTS

leakage inductance slightly while more space between the two coils would increase the leakage inductance slightly.

For the circuit example of Figure 24, the measured leakage inductance of the output filter balun was 1.44 μH at 0.5 MHz. This is the value of the component shown in series with two of the three windings. The calculated CM inductance is 214 μH. The filters of Figure 22 and Figure 24 as configured, have the DM inductance limited to the output balun leakage inductance. This works for most applications, but where additional DM inductance is required, an inductor can be added in series with the leakage inductance shown on the balun equivalent circuits in Figure 21 or Figure 23. The DM inductance will now be the sum of the leakage and added series inductance, which allows some added flexibility in the design of the DM output filter, such as the filter type of Figure 19 where a shunt damping network is used to control the Q. This can also allow the balun to be made on a pot or RM core by bifilar or trifilar winding since the DM filter is not now solely dependent on leakage inductance. In this case, use a high permeability ferrite material and a core pair with no air gap for the balun. The added DM inductor can be made on a small Molyperm toroid or air gapped pot core.

distribution. This circuit was also built on copper clad vector board using readily available components with about the same results as for Figure 22.

The baluns should be designed using Ferrite materials having a relatively high permeability of approximately 5000. The high permeability allows a high CM inductance with a minimal number of turns, resulting in reduced copper loss, distributed capacitance and core size. Since the balun net ampere turns are always zero, when used as in Figure 22 and Figure 24, the only thing limiting small size is copper loss/temperature rise, the availability of small cores and how physically small the balun can be made.

A simplified equivalent circuit for a two winding balun is shown on Figure 21. The leakage inductance is shown as an additional DM inductor in series with the balun. The common mode inductance is that of one winding and can be calculated knowing the core

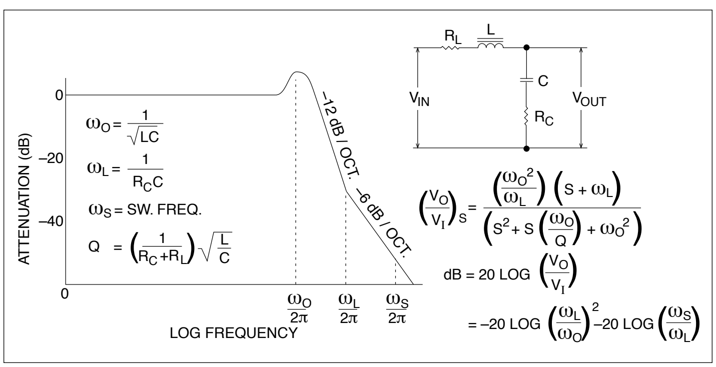

FIGURE 25: SECOND ORDER L.P. FILTER

permeability and dimensions using the equation in Figure 21. The leakage inductance is much smaller than the former, on the order of ~1% of CM inductance, and should be measured on a sample balun by shorting one winding and measuring the inductance looking into the other winding. For the circuit example of Figure 24, the measured leakage inductance of the output filter balun was 3.5 μH at 0.5 MHz. The calculated CM inductance is 480 μH. Less space between the two coils would decrease the

The input CM filter can be part of an EMI power line filter, but must be close to the power converter to be effective. Where a separate balun is used, the CM natural frequency should be placed two to three octaves or more below the switching frequency. Since this is a low pass filter, the exact frequency is not important. In Figure 22 and Figure 24, where the switching frequency is nominally 500 kHz, the CM natural frequency has been placed at around 62 kHz. A CM inductance in the 100 uH to 500 μH range should be used with the largest practical value of ceramic capacitor available.

The output CM filter case bypass capacitor should be the largest practical value which can be used. It is only necessary to bypass the output common to case since the output lines are shorted together by the DM filter capacitors. The output balun CM inductance can be the largest practical value provided that inter-winding capacitance is kept to less than approximately one picofarad. Stray capacitance on the load side of the balun, which determines the CM natural frequency with the CM inductance, can be expected to be no more than a few picofarads.

Designing the DM output filter requires knowledge of the amplitude of the power converter output ripple / noise so it can be ratioed with the required level at the filter output to determine the required filter attenuation. To determine the ripple level, use the maximum value of the device specification, which is usually at a restricted bandwidth and does not include high frequency switching spikes. It defines the envelope of the first few spectral components which need to be addressed, since the low pass filter will automatically take care of the high frequency switching components. Once the needed attenuation has been determined,

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the graph and equations of Figure 25 can be used to complete the DM output filter design. Use the maximum ESR to determine the location of the zero associated with the tantalum capacitor. The natural frequency of the filter can now be determined by calculation from the transfer function plot and the equations. Figure 18 and Figure 19 may also be helpful, but modeling in SPICE is highly recommended. Use the maximum value of ESR for your model to determine the attenuation and use the maximum and minimum value of ESR (if it is known) to determine the Q of the filter. Unless the DM filter is damped, its natural frequency should not be set inside the control loop band width of the power supply to avoid instability. This is because the output filter appears to the power supply as a series resonant circuit and will act as a crowbar on the output at its natural frequency. For most Interpoint power converters, using 30 kHz as the lower limit should be adequate although for many Interpoint models the bandwidth can be significantly lower. If a better approximation of a converter's bandwidth is needed our Applications Department can assist in providing information for a particular converter. (See contact information on page 18.) In cases where the above guidelines don't produce enough attenuation, an additional second order DM filter section can be cascaded with the first. In doing this, place the natural frequency of the second section one or more octaves above that of the first. Consider using a small Molyperm toroid for the inductor and make sure the inductance is the expected value due to the influence of the dc current on the magnetizing force and thus the core permeability. Output filters using output inductors and capacitors to provide very low CM and DM noise levels have been discussed. These filters are applicable to any switching power supply where low output noise is a requirement. The user should be aware that the output filter may cause additional dynamic regulation or other errors, and should be careful to test for these effects before committing to a final design.

EXAMPLES

DESIGN EXAMPLE 1

Assume the maximum allowable ripple and noise for a system specification is 10 millivolts peak to peak, and the converter to be used is capable of providing 15 volts at 2 amps. The specifications for the output ripple voltage on the converter datasheet are 25 millivolts peak to peak typical, and 40 millivolts peak to peak maximum at 25°C. Over temperature (-55°C to 125°C) the output ripple is 40 millivolts peak to peak typical, and 90 millivolts peak to peak maximum. Although the application will never be close to either temperature extreme, the worst case of 90 millivolts peak to peak will be used. This means the external filter will need to provide 19 dB of attenuation (20 log (90/10)) at the switching frequency of 600 kHz. For simplicity -20 dB will be used.

The system implements 58 uF of tantalum capacitance and 4 uF of ceramic capacitance as required by the load. Since the system capacitance is significant it will inherently become part of the filter. The maximum ESR of the tantalum capacitance is 0.3 ohms and the minimum ESR is assumed to be 0.1 ohms. With an ESR of 0.3 ohms, the lead frequency created by the ESR will be 9.1 kHz ($1/(2\pi \times 0.3 \times 58 \times 10^{-6})$) and the 58 uF capacitor will not be very effective at the 600 kHz switching frequency. For the design it is assumed only the 4 uF of ceramic capacitance will contribute to the filtering and an inductor will be needed to meet the attenuation specification. It is also assumed that the trace resistance and inductance in series with the ceramic capacitors are negligible. With an LC filter using high quality ceramic capacitors, the roll off at the filter's resonant frequency will be -40 dB/decade. The resonant frequency needed to provide 20 dB of attenuation at the switching frequency is determined by equation 1, where f_{SW} is the converter's switching frequency, and f_0 is the natural resonant frequency of the external filter.

Equation 1 $$\text{Attenuation} = 40 \text{ Log } \frac{f_{SW}}{f_0}$$

Substituting 600 kHz for f_{SW}, 20 dB for the attenuation, and dividing both sides of equation 1 by 40, we can simplify the equation and solve for f_0

$20 = 40 \text{ Log } (600 \text{ kHz}/f_0)$,
$\text{Log } (600 \text{ kHz}/f_0) = ½$,
$10^{½} = (600 \text{ kHz}/f_0)$,

Solve for $$f_0 = \frac{600 \times 10^3}{10^{½}},$$

Solving for f_0 we get a filter resonant frequency of 190 kHz. Now that the filter's resonant frequency and capacitance are known we can find the inductance using equation 2 and solving for L.

Equation 2 $$f_0 = \frac{1}{2\pi\sqrt{LC}},$$

Solve for L $$L = \left(\frac{1}{2\pi\, f_0}\right)^2 \frac{1}{C}$$

Substituting 190 kHz for f_0, and 4 uF for C, into equation 2,

$$L = \left(\frac{1}{2\pi\,(190 \times 10^3)}\right)^2 \left(\frac{1}{4 \times 10^{-6}}\right)$$

Measurement and Filtering of Output Noise of DC-DC Converters

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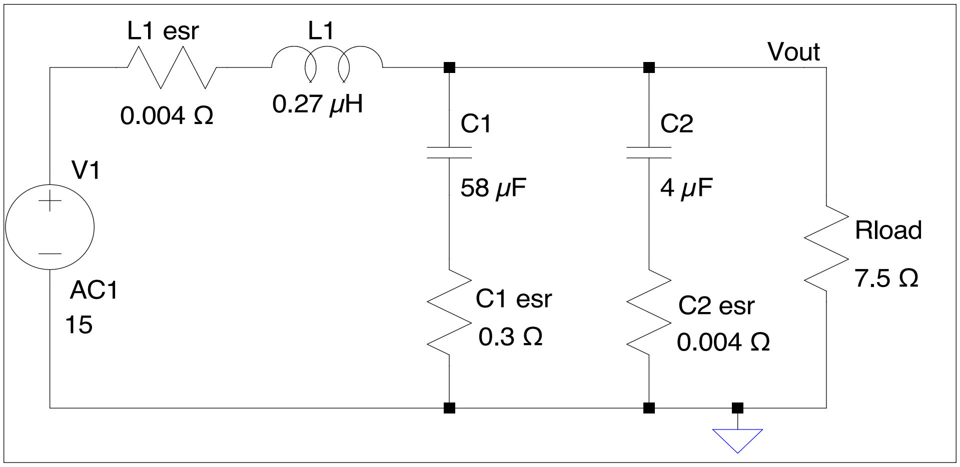

FIGURE 26: DESIGN EXAMPLE 1, OUTPUT FILTER, 0.3 OHMS ESR

the inductance needed is 0.175 uH. With an inductance of 0.175 uH, the physical size of the inductor will be very small. Molyperm cores are a good choice and are available from many manufactures. In this case an MPP core, part number 55177-A2, from Magnetics Inc, was chosen. The dimensions for this core are, OD = 0.208 inches (5.28 mm), ID = 0.073 inches (1.85 mm) and height = 0.125 inches (3.18 mm). A smaller core could have been used but the 55177-A2 was a core we had on hand. The inductance, L, is a function of the number of turns on the core, N, and the A_L value of the core ($L = N^2 \times A_L$). The A_L, value from the catalog is 67 mH/1000 turns (or 67 nH) and the desired inductance is 0.175 uH. Solving for N we get 1.6 turns which will be rounded up to two turns. Using two turns equates to an inductance of 0.27 uH which will provide margin. To make sure the core does not saturate we determine the magnetizing force (H) in oersteds. The equation used to determine the magnetizing force is $H = (0.4\,\pi\,N\,I)/L_e$ where H is in oersteds, N is the number of turns on the core, I is the current flowing through the coil, and L_e is the magnetic path length of the core in centimeters.

In this case N = 2 turns, I = 2 amps, and from the catalogue L_e is 1.06 cm. Solving for H we get 4.7 oersteds. The permeability of the 55177-A2 core is 200 u and the "Permeability versus DC Bias" curves shows that at 4.7 oersteds there is no decrease in permeability. A decrease in permeability is acceptable as long as we have accounted for the decreased inductance due to the dc current.

The filter's resonant frequency with a 0.27 uH inductor and a 4 uF capacitor is 153 kHz. Plugging 153 kHz into equation 1 gives an attenuation of 23.7 dB. The next step would be to model the results in SPICE. A schematic of the model is shown in Figure 26 and the simulation results are shown in Figure 27. The simulation results show the attenuation is -23.5 dB at 600 kHz which agrees very well with our calculations. The predicted resonant frequency is 153 kHz but due to the influence of the 58 uF of capacitance the resonant frequency is closer to 120 kHz. The Q is around 1.25 and is beyond the converter's bandwidth so it is not a concern.

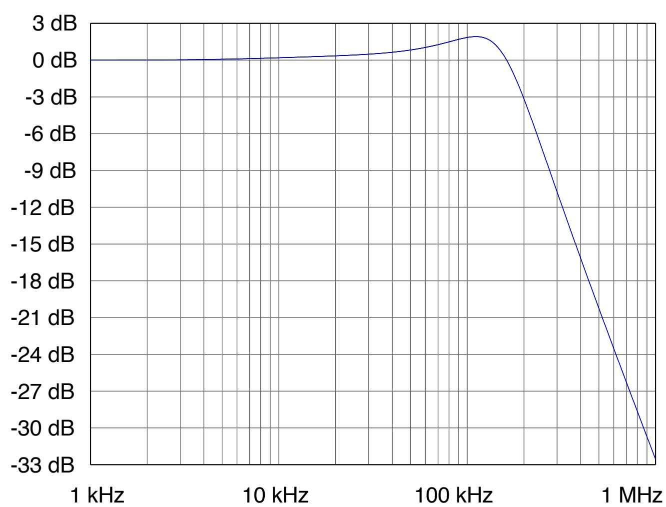

FIGURE 27: SIMULATION RESULTS OF FIGURE 26

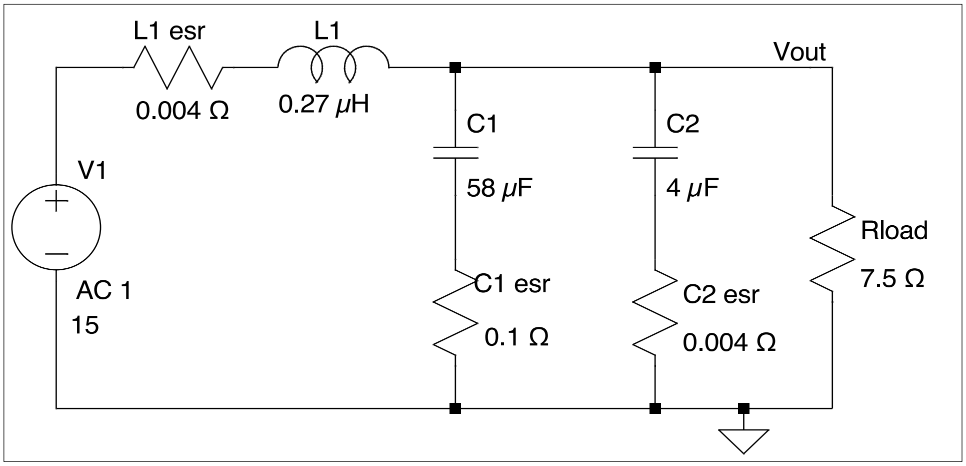

FIGURE 28: DESIGN EXAMPLE 1, OUTPUT FILTER, 0.1 OHMS ESR

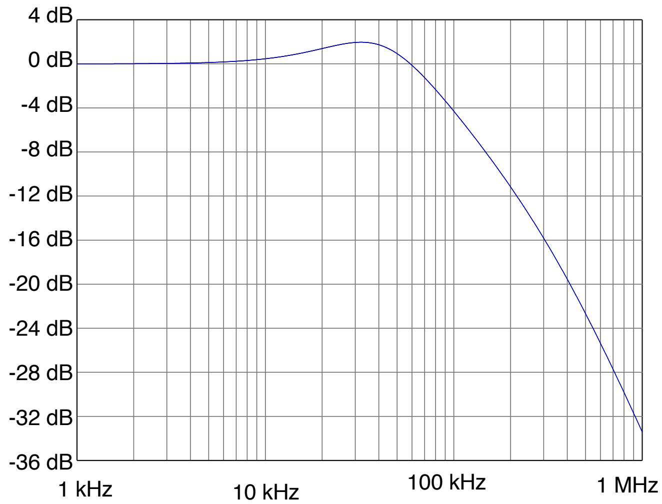

FIGURE 29: SIMULATION RESULTS OF FIGURE 28

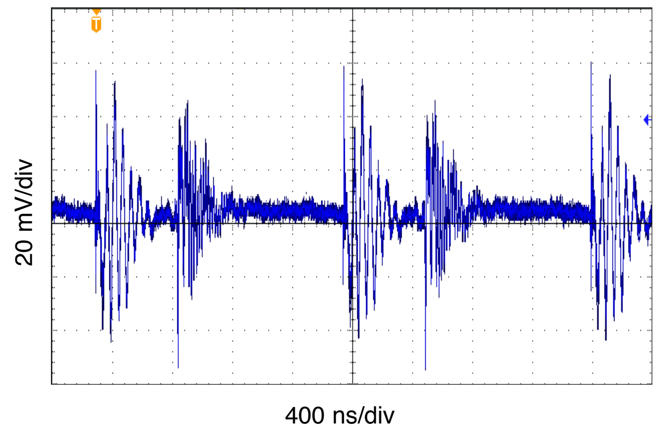

FIGURE 30: OUTPUT FULL LOAD, NO FILTER

The next step would be to model the filter using the lower capacitor ESR, which in this case is 0.1 ohms. The schematic of the model is shown in Figure 28 and the results are shown in Figure 29. With the lower ESR the attenuation is -25.3 dB at 600 kHz and the Q of the filter is still around 1.25 with a resonant frequency of 33 kHz.

A scope capture of the ripple and noise for the converter operating at full load, without any filtering, is shown in Figure 30. The results show 112 millivolts peak to peak using the barrel of the scope probe for the ground connection. The next step was to suppress the CM noise. This was accomplished by adding a 0.1 uF ceramic capacitor from input common to chassis and a 0.1 uF ceramic capacitor from output common to chassis. The results are shown in Figure 31, where the noise has increased from 112 millivolts peak to peak to 364 millivolts peak to peak. Now the CM noise has decreased but the DM noise has increased significantly. The high frequency DM noise is not a problem as the addition of our DM filter will take care of this.

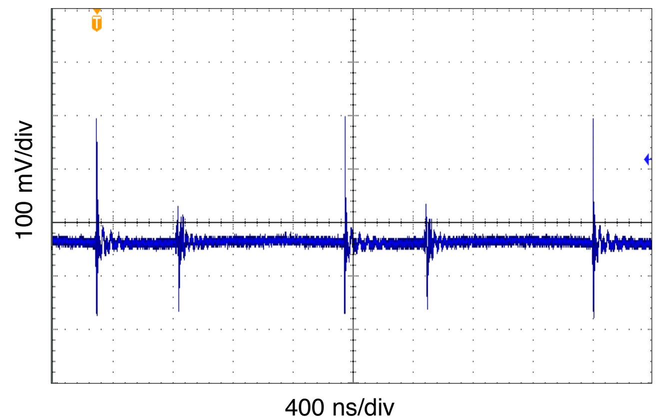

FIGURE 31: OUTPUT FULL LOAD, NO FILTER, CM SUPPRESSED

Measurement and Filtering of Output Noise of DC-DC Converters

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To check the measurement against the datasheet we use the 2 MHz scope probe shown in Figure 5 "C: Alternate Method". The results are shown in Figure 32 but need to be to be multiplied by 2 to compensate for the x ½ probe. With the multiplication factor the results are 12.6 millivolts peak to peak and are well below the datasheet specifications.

Now that the CM noise has been suppressed, we add the DM filter shown in Figure 26. The results, with the filter added, are shown in Figure 33. The results are 7.2 millivolts peak to peak, however, even without the converter running the results still shows 7.2 millivolts peak to peak. This would indicate most of what is showing up on the scope is ambient noise. To read a voltage with such small amplitudes a 1 x probe should be used. Figure 34 shows the same measurement using a 15 MHz, 1x probe. The results show 1.12 millivolts peak to peak but even without the converter running the scope showed 0.8 millivolts peak to peak. We will assume that this is an accurate measurement. This example illustrates that with only the additions of a very small inductor, very good results can be achieved. This inductor can be made with minimal time and cost and with two turns should be very easy to construct and should allow for a smaller toroid to be used.

DESIGN EXAMPLE 2

The previous example used a forward converter which is typically less noisy than a fly-back converter. Our next example will consider a discontinuous mode fly back converter. As in the last example, assume the maximum ripple and noise for the system specification is 10 millivolts peak to peak and the converter to be used is a flyback converter that provides 5 volts at 1 amp, and operates at 500 kHz. The systems requirements are 5 volts at 0.8 amps and the maximum temperature range the converter will see in this application is -20°C to 80°C. The output ripple specifications on the converter's datasheet are 125 millivolts peak to peak typical, and 350 millivolts peak to peak maximum at 25°C, and 525 millivolts peak to peak maximum over temperature (-55°C to 125°C). These noise levels are significantly higher than the datasheet specification used in our previous example

In this case the system has a little over 1 uF of ceramic capacitance but the capacitance is made up of many smaller capacitors sprinkled across the board and strategically located in noise sensitive areas. Since many of these small capacitors are buried in traces, which have inductance and resistance we will assume they are not part of the filter. In this case we will add 2 uF of additional ceramic capacitance after the inductor and assume the 2 uF is the only capacitance. Since the maximum output ripple specification on the datasheet is 525 millivolts peak to peak and the system requirement is 10 millivolts the attenuation needed is 34 dB (20 log (525/10)). With a switching frequency of 500 kHz, and 34 dB of attenuation, we can use equation 1 and solve for f_o which in this case is 71 kHz ($500 \times 10^3 / 10^{34/40}$). Now

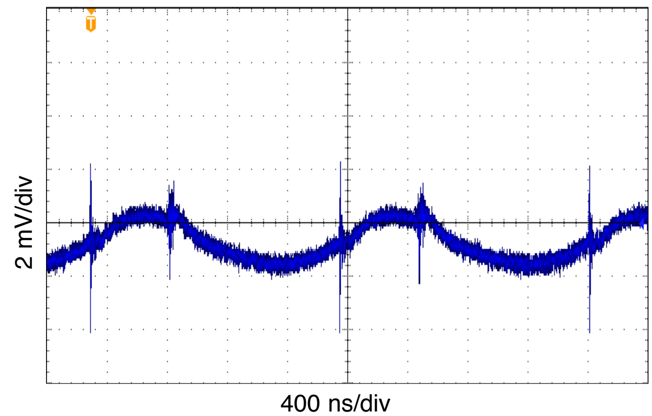

FIGURE 32: OUTPUT USING 2 MHZ X ½ PROBE

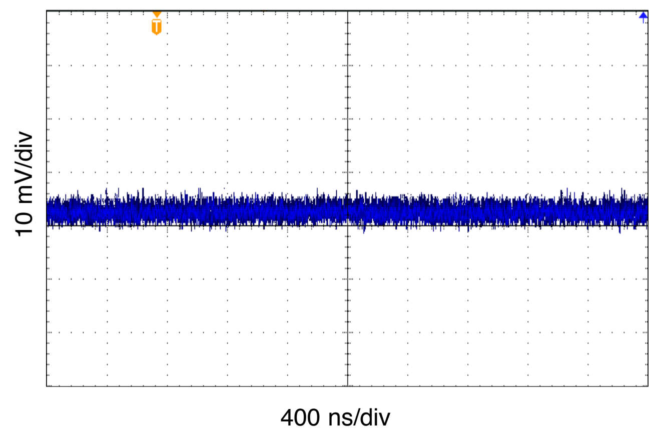

FIGURE 33: OUTPUT WITH FIGURE 26 FILTER

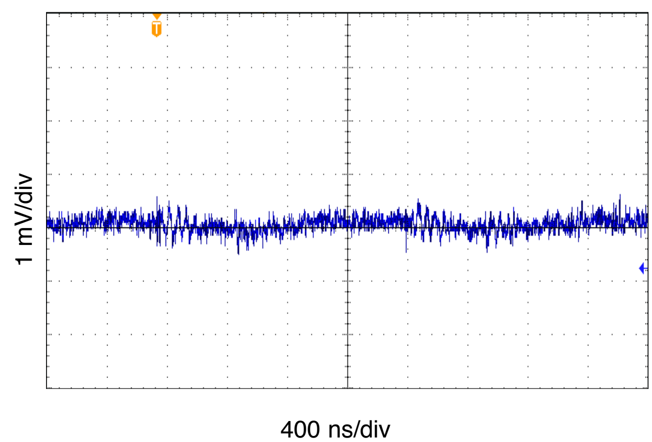

FIGURE 34: OUTPUT WITH FIGURE 26 FILTER

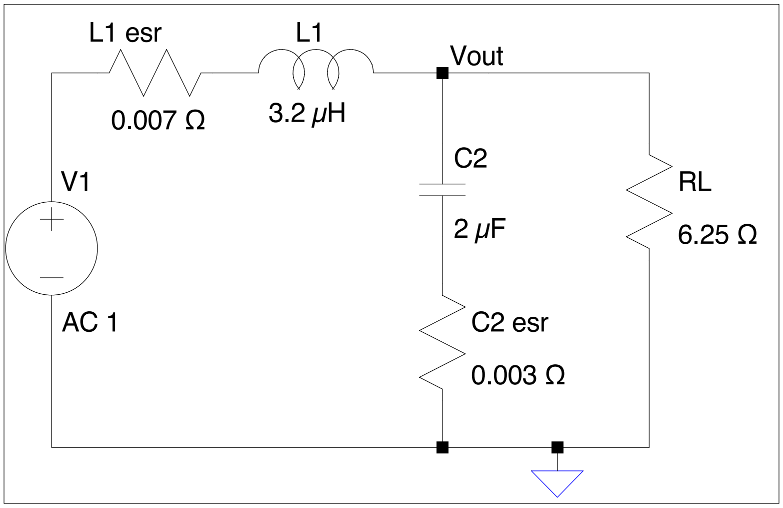

FIGURE 35: DESIGN EXAMPLE 2 OUTPUT FILTER

that we know the resonant frequency of the filter and the filter capacitance, we can use equation 2 and solve for L which is 2.5 uH. Note that if we had a specific inductor that we wanted to use we could use equation 2 and solve for C instead of L.

The same 55177-A2 core that was used in the last example will be used here. If six turns are used the inductance is 2.4 uH (6^2 x 67 nH = 2.4 uH) which is a little shy of our goal. The core will easily accommodate another turn so seven turns of 25 AWG will be used which will increase the inductance to 3.3 uH. With seven turns, and 0.8 amps of current, the inductor has a magnetizing force of 6.6 oersteds (0.4 π x 7 x 0.8) / 1.06) which puts the inductance at about 97% of calculated value, or 3.2 uH. Using Equation 2, the resonant frequency is 63 kHz and using Equation 1, the attenuation is 36 dB at 500 kHz.

The datasheet specifications for the switching frequency are 400 kHz to 600 kHz so if we want to be conservative 400 kHz should be used for the calculations. With a 3.2 uH inductor and 2 uF capacitor, the attenuation at 400 kHz is -32 dB. Since the converter's temperature is well within either extreme, and full power is not required, the filter design should be okay. Latter scope captures will show the difference between in output ripple voltage for MSA2805S operating with load currents of 1 amp and 0.8 amps. If the goal is to meet the 10 millivolts peak to peak specification under all possible conditions, then adding another turn on the core would provide 4 uH of inductance and we would meet our target of -34 dB at 400 kHz.

The next step would be modeling the filter to determine the Q of the filter and verify the attenuation. The schematic of the model is shown in Figure 35 with the results shown in Figure 36. The Q is 4.8 at 62 kHz and the attenuation is -36 dB at 500 kHz, which agrees fairly well with our calculations. The Q is very high but at 62 kHz the stability of the converter should not be an issue. However, any disturbances that occur at the filter's resonant frequency will be amplified by the filter's Q. If this is a concern, which we will assume it is, a damping network can be added. The

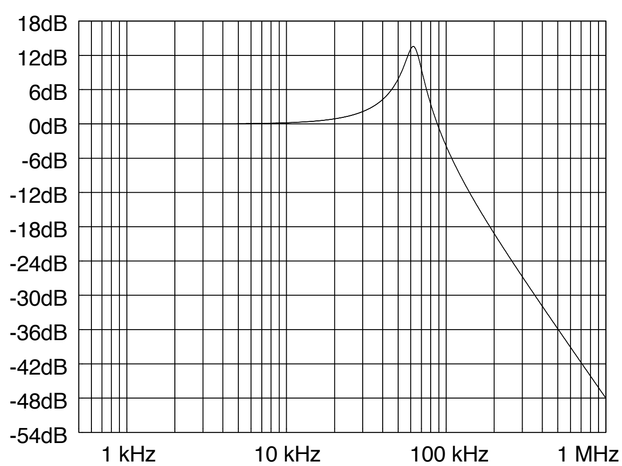

FIGURE 36: SIMULATION RESULTS OF FIGURE 35

Measurement and Filtering of Output Noise of DC-DC Converters

APPLICATION NOTE

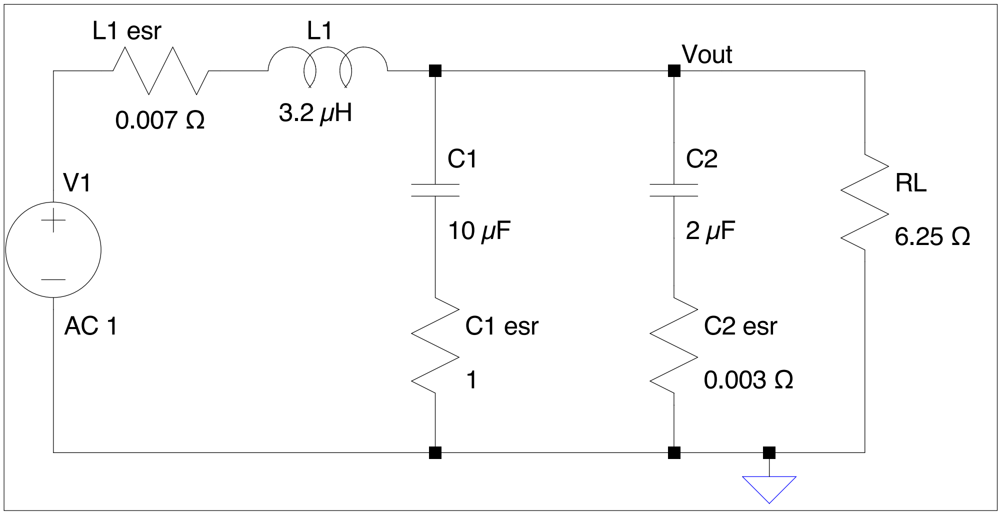

FIGURE 37: DESIGN EXAMPLE 2 OUTPUT FILTER WITH DAMPING

damping network will be a 10 uF capacitor with a series 1 ohm resistor. Verify that any capacitance used as part of a damping network, does not have significant ESR. If it does this needs to be taken into account and the damping resistor may need to be adjusted accordingly. The schematic of the model with its damping network is shown in Figure 37 and the simulation results are shown in Figure 38. The simulation shows the Q is 1.3 at around 25 kHz and the attenuation at 500 kHz is -36 dB. If the inductor's capacitance, and capacitors ESL are known, they can be added to the model for more accuracy. At 500 kHz to 600 kHz the effects should be negligible for most components.

Figure 39 shows a scope capture of the output ripple and noise for the converter without filtering of any kind. Although our example assumes 0.8 amps of output current, the data was taken at the full load capability of 1 amp. The results show 1.38 volts peak to peak. By bypassing input common to chassis and output common to chassis the results did not improve significantly. This will be revisited after implementing the filter as the noise levels will be much lower and the contribution of the CM noise will be more pronounced. To get an idea of the effects of adding ceramic capacitance to the output, a 0.47 uF ceramic capacitor was soldered on the PCB.

The measurement, shown in Figure 40, has decreased from 1.38 volts peak to peak to 134 millivolts peak to peak. Note that the noise has significantly decreased leaving mainly the ripple voltage. The 0.47 uF capacitor was placed on the board directly by the converter's output pins and the measurement was taken very close to the capacitor. With the capacitor further away from the output pins, and not measuring directly at the capacitor, the noise portion of the measurement increases. This shows how a

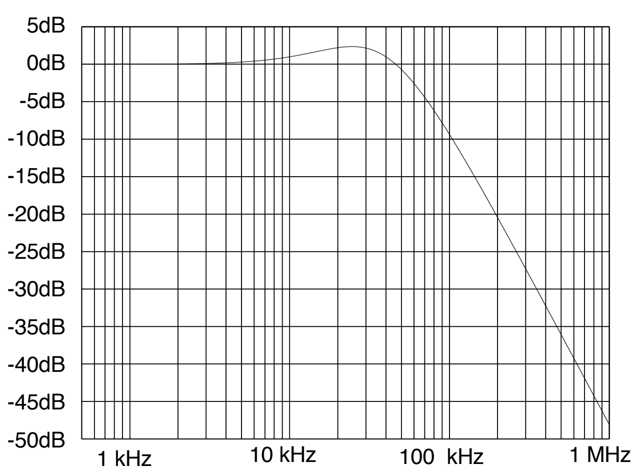

FIGURE 38: SIMULATION RESULTS OF FIGURE 37

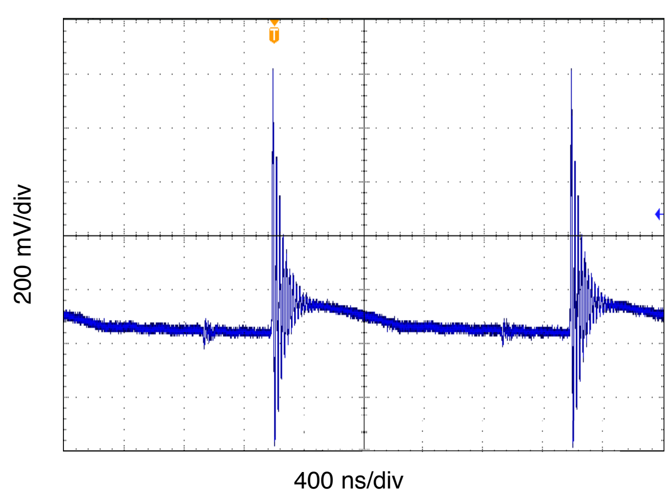

FIGURE 39: OUTPUT NOISE NO FILTER, 10 X PROBE

Measurement and Filtering of Output Noise of DC-DC Converters

APPLICATION NOTE

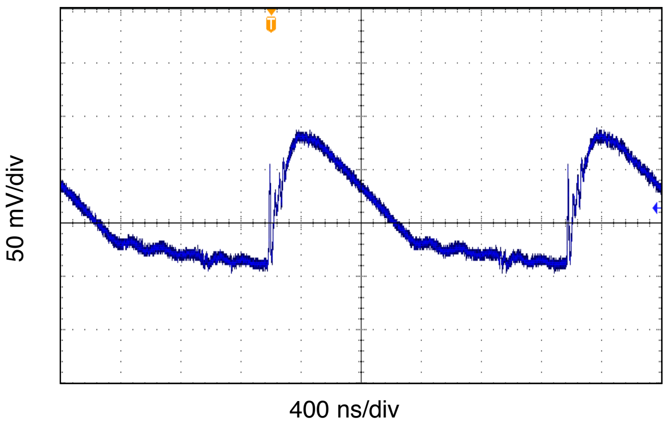

FIGURE 40: OUTPUT NOISE WITH 0.47 uF CAP, 10 X PROBE

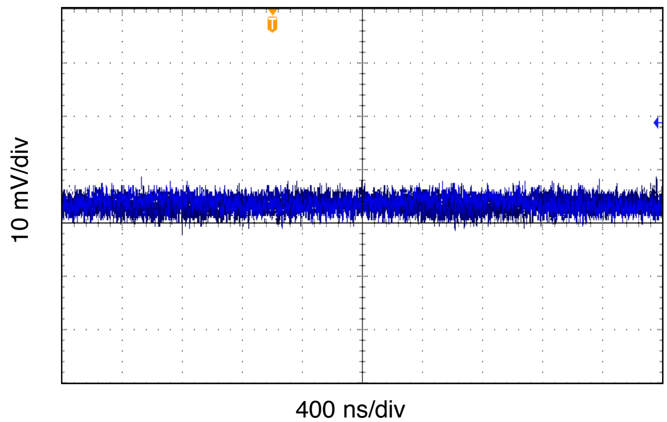

FIGURE 42: OUTPUT WITH FIGURE 37 FILTER, CM SUPPRESSED. 10 X PROBE

little capacitance with proper layout techniques can significantly improve the high frequency noise portion of the waveform.

The 0.47 uF capacitor was removed and the filter in Figure 37 was added to the converter's output. The results are shown in Figure 41. Here the lower frequency ripple is no longer present but the high frequency noise remains. The results are 40 millivolts peak to peak. To determine if this is CM noise a 0.1 uF ceramic capacitor was connected from input common to chassis and a 0.1 uF ceramic capacitor was connected from output common to chassis. The results are shown in Figure 42. This indicates that the high frequency noise in Figure 41 is CM noise. The results of Figure 42 show 8 millivolts peak to peak but even without the converter running the measurement did not change much which would indicate much of this is ambient noise. To determine the real noise level a 15 MHz, 1x scope probe was used and the data was retaken. The results, shown in Figure 43, are 3.1 millivolts peak to peak and the ripple voltage, which is missing with the 10x probe, is now very pronounced. This shows the importance of using a 1x probe when the ripple and noise are at such low levels. There is still some high frequency noise that can be cleaned up by adding another 0.1 uF capacitor from the positive 5 volt rail to chassis. The results with the additional 0.1 uF ceramic capacitor from the positive output to chassis are shown in Figure 44. Now the noise is completely gone and we are only left with the 500 kHz ripple voltage that measures 2.48 millivolts peak to peak. Now we can compare the actual attenuation with the calculated attenuation above. The 134 millivolts peak to peak measurement in Figure 40 is the 500 KHz ripple voltage without any noise.

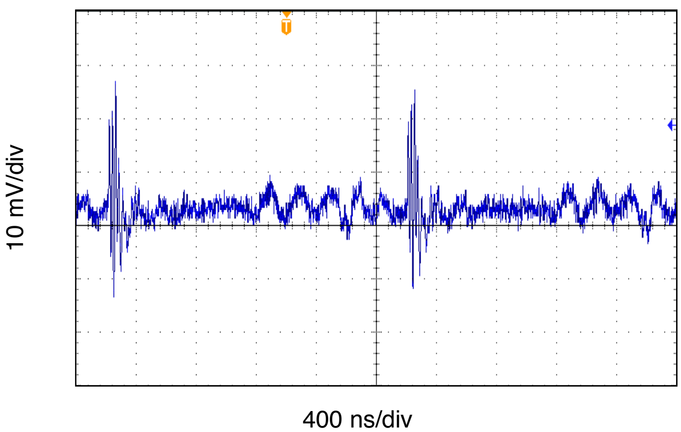

FIGURE 41: OUTPUT WITH FIGURE 37 FILTER

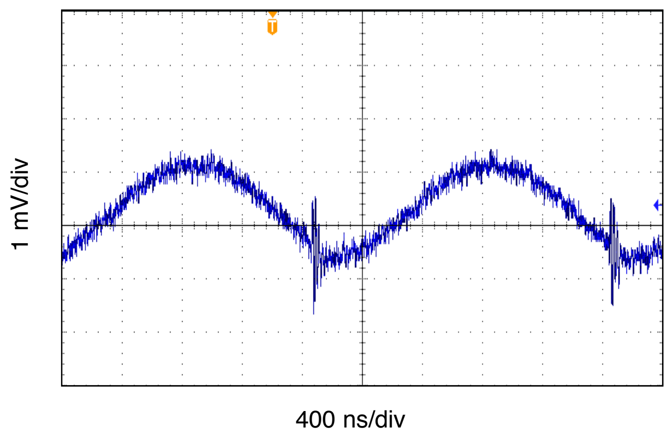

FIGURE 43: OUTPUT WITH FIGURE 37 FILTER, CM SUPPRESSED, 1 X PROBE

Measurement and Filtering of Output Noise of DC-DC Converters

APPLICATION NOTE

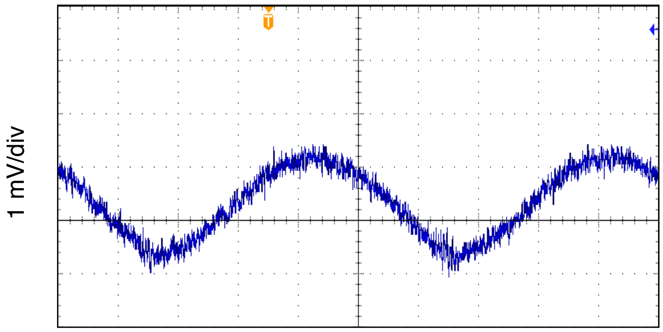

FIGURE 44: OUTPUT WITH FIGURE 37 FILTER, CM SUPPRESSED, 0.1 uF Cap, 1X Probe

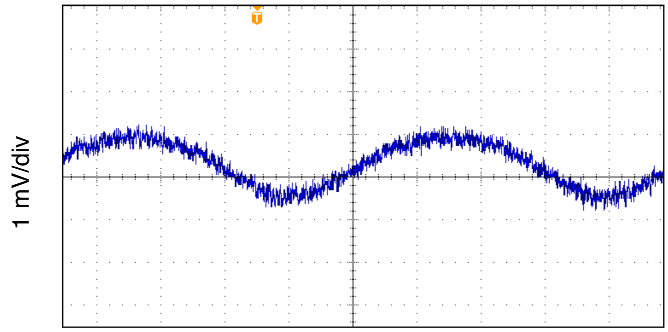

FIGURE 45: OUTPUT WITH FIGURE 37 FILTER, CM SUPPRESSED, 0.1 uF Cap, 1X Probe, Output Decreased from 1 A to 0.8 A.

Comparing this to the most recent measurement of 2.48 millivolts peak to peak would indicate a reduction of 54,

$\frac{2.48}{134}$ or -35 dB (20 log $\frac{1}{54}$) which is very close to our calculation of -36 dB. The 0.47 uF cap, used in Figure 40 to suppress the noise, most likely provided some attenuation of the ripple which would have made the measured value closer to the calculated value as the 0.47 μF was not present.

Since all of the measurements were taken at 1 amp of output current, and the application calls out for 0.8 amps, the data with the output filter and three 0.1 uf chassis bypass capacitors, was retaken at 0.8 amps. The results are shown in Figure 45 where the ripple has decreased from 2.48 millivolts peak to peak to 2 millivolts peak to peak. This is well below the 10 millivolts peak to peak specification.

Most ceramic capacitors will decrease in value with a bias voltage applied across them. This should be considered when designing an external filter.

This shows how a few components can be used to provide excellent results. Each application will be slightly different and needs to be considered on its own merits.

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