These days switching power supplies are nearly ubiquitous and used throughout every electronic device. They are valued for their small size, low cost, and efficiency. However, they have the major drawback in that their outputs can be noisy due to the high switching transients. This has kept them out of high performance analog circuits where linear regulators have ruled the roost. It has been shown that in many applications an appropriately filtered switching converter can replace a linear regulator for production of a low noise supply. Even in those demanding applications where an extremely low noise supply is required, there is probably a switching circuit somewhere upstream in the power tree. Therefore, there is a need to be able to design optimized, damped multistage filters to clean up the output from switching power converters. In addition, it is important to realize how the filter design will affect the compensation of the switching power converter.

In this article, boost circuits will be used for the example circuits, but the results will be directly applicable to any dc-to-dc converter. Shown in Figure 1 are the basic waveforms in a boost converter in constant-current mode (CCM). The issue that makes an output filter so important for a boost or any of the other topologies with discontinuous current mode is the fast rise and fall in the current time in Switch B. This tends to excite parasitic inductances in the switch, the layout, and the output capacitors. The result is that in the real world the output waveforms look much more like Figure 2 rather than Figure 1, even with a good layout and ceramic output capacitors.

![Figure 1. Basic voltage and current waveforms for a boost converter.](image1)

![Figure 2. Typical measured waveforms of a boost converter in DCM.](image2)
The switching ripple (at the switching frequency) caused by the change in 
charge of the capacitor is very small compared to the undamped ringing 
of the output switch, which we will refer to as output noise. Generally, this 
output noise is in the 10 MHz to 100+ MHz range, well beyond the self- 
resonant frequency of most ceramic output capacitors. Therefore adding 
additional capacitors will do little to attenuate the noise.

There are a couple of reasonable choices for different types of filters to 
filter this output. This article will illustrate each type of filter and give 
a step-by-step process to a design. The equations are not rigorous and 
some reasonable assumptions are made to simplify them somewhat. 
There is still some iteration required since each component will affect 
the values of the others. The ADIsimPower design tools get around this 
problem by using linearized equations for component values, like cost 
or size, to do an optimization before actual components are selected, 
and then optimize the outputs once real components are chosen from the 
database of thousands of parts. However, for a first pass at a design, 
this level of complexity is not necessary. With the provided calculations 
and possibly using a SIMPLIS simulator like the free ADIsimPE,” or some 
bench time in the lab, a satisfactory design can be found with a minimal 
amount of effort.

Before designing the filter, consider what is achievable with a single stage 
filter RC or LC filter. Typically with a second stage filter it is reasonable 
to get the ripple down to a few hundred μV p-p and the switching noise 
down below 1 mV p-p. A buck converter can be made somewhat quieter 
to get the ripple down to a few hundred μV p-p and the switching noise 
ΔVRIP. The approximate response time of the converter to an 
instantaneous change of the output load. 

The simplest type of filter is just an RC filter as shown attached to the 
output of a low current ADP161x-based boost design shown in Figure 3. 
This filter has the advantage of low cost and will not need to be damped. 
However, due to power dissipation it is only useful for very low output 
current converters. For this article, ceramic capacitors with small ESR 
are assumed.

Design Process for an RC Second Stage 
Output Filter

Step 1: Choose C1 based on assuming the value output ripple at C1 is 
approximately ignoring the rest of the filter; 5 mV p-p to 20 mV p-p 
is a good place to start. C1 can then be calculated using Equation 1.

\[ C_1 = \frac{I_{pp}}{8 F_{SW} \Delta V_{OUT}^{\text{rip}} - \Delta I_{\text{step}} R_{ESR}} \]  (1)

Step 2: R can be chosen based on power dissipation. R must be much 
larger than RESR for the capacitors and for this filter to be effective. 
This limits the range of output currents to something less than 50 mA or so.

Step 3: C2 can then be calculated from Equation 2 through Equation 6. 
A, a, b, and c are just intermediate values to simplify calculation and have 
no physical meaning. These equations assumes R << RLOAD and the ESR 
for each capacitor is small. These are both very good assumptions and 
introduce little error. C2 should be the same or larger than C1. The ripple 
in Step 1 can be adjusted to make this possible.

\[ A = \frac{\Delta V_{OUT}^{\text{rip}}}{\Delta I_{\text{step}} R_{LOAD}} \]  (2)

\[ a = A R_{LOAD}^2 C_1^2 \omega^4 + A R_{LOAD}^2 \omega^2 \]  (3)

\[ b = 2 A R_{LOAD}^2 \omega^2 C_1 - 2 A R_{LOAD} C_1 \omega^2 \]  (4)

\[ c = A R_{LOAD} \omega^2 C_1 + A - 1 \]  (5)

\[ C_2 = \frac{-b + \sqrt{b^2 - 4ac}}{2a} \]  (6)

Figure 3. ADP161x low output current boost converter design with an added RC filter on the output.
For higher current supplies it is beneficial to replace the resistor in the pi filter with an inductor as shown in Figure 4. This configuration gives very good ripple and switching noise rejection in addition to low power loss. The issue is that we have now introduced an additional tank circuit that can resonate. This can result in oscillations and an unstable power supply. Therefore, the first step to designing this filter is to choose how to damp the filter. Figure 4 shows three viable damping techniques. Adding $R_{LT}$ has the advantage of adding little extra expense or size. The damping resistor typically has little to no loss and can be small for even large power supplies. The drawback is that it significantly reduces the effectiveness of the filter by reducing the parallel impedance with the inductor. Technique 2 has the advantage of maximizing filter performance. If an all ceramic design is desired, $R_b$ can be a discrete resistor in series with a ceramic capacitor. Otherwise a physically large capacitor with a high ESR is required. This additional capacitance ($C_b$) can add significant cost and size to the design. Damping Technique 3 looks very advantageous since the dampening capacitor $C_2$ is added to the output where it might help somewhat with transient response and output ripple. However, this is the most expensive technique since the amount of capacitance required is much larger. In addition, the relatively large amount of capacitance on the output will lower the frequency of the filter resonance, which will reduce the achievable bandwidth of the converter—therefore Technique 3 is not recommended. For the ADIsimPower design tools we use Technique 1 because of the low cost and relative ease of implementing it in an automated design process.

Another issue that needs to be dealt with is compensation. It may be counterintuitive, but it is almost always better to put the filter inside the feedback loop. This is because putting it in the feedback loop helps damp the filter somewhat, eliminates dc load shift and the series resistance of the filter, and gives a better transient response with less ringing. Figure 5 shows the Bode plot for a boost converter with an LC filter output added to the output.

The feedback is taken before or after the filter inductor. The thing that is most surprising to people is how much the open-loop Bode plot changes even when the filter is not “in” the feedback loop. Since the control loop is affected with or without the filter in the feedback loop, one might as well compensate for it appropriately. In general this will mean scaling back the target crossover frequency to a maximum of a fifth to a tenth of the filter resonant frequency ($F_{RES}$).

$$F_{RES} \approx \frac{1}{2\pi \sqrt{\frac{1}{L_{ESR} C_1 C_2}}}$$  \hspace{1cm} (7)
Step 3: As described previously, adding the filter will affect the compensation of the converter by reducing the achievable crossover frequency (\(F_c\)). For a current-mode conversion, the maximum achievable \(F_c\) is the lesser of 1/10 of the switch frequency, or 1/5 the \(F_{res}\) of the filter as calculated in Equation 7. Fortunately, most analog loads do not require an exceptionally high transient response. Equation 9 calculates the approximate output capacitance (\(C_{cap}\)) required on the output of a converter to provide for a specified transient current step.

\[
C_{cap} = \frac{I_{step}}{\pi F_s (\Delta V_{out} - I_{step} R_{ESR})} \quad (9)
\]

Step 4: Set \(C_3\) as the minimum of \(C_{cap}\) and \(C_i\).

Step 5: Calculate the approximate damping filter resistance using Equation 10 and Equation 11. These equations are not absolutely accurate, but they are the closest thing to a closed form solution without the need to use extensive algebra. The ADIsimPower design tools calculate \(R_{res}\) by calculating the open-loop transfer function (OLT) of the converter with the filter and with the inductor shorted out. \(R_{res}\) values are then guessed until the peak of the converter OLT with the filter is only 10 dB above the OLT of the converter with the inductor shorted. This technique can be used in a simulator like ADIsimPower or in the lab using a spectrum analyzer.

\[
a = \frac{L_{filt} I_{load} C_t}{\Delta V_{out}^2} - L_{filt} \omega_o - R_{load} C_t L_{filt} \omega_o \quad (12)
\]

\[
b = \frac{R_{load} I_{filt} L_{filt} C_t}{\Delta V_{out}^2} - R_{filt} + R_{filt} L_{filt} \omega_o^2, \quad (13)
\]

\[
c = \frac{R_{filt} R_{load} \omega_o - R_{filt} R_{load} C_t L_{filt} \omega_o^3}{\omega_o^2} \quad (14)
\]

\[
d = -\frac{R_{load} R_{filt} \omega_o^2}{\omega_o^2} \quad (15)
\]

\[
C_2 = \frac{a^2 + b^2}{c^2 + d^2} \quad (16)
\]

Step 7: Step 3 through Step 5 should be repeated until a well-damped filter design is calculated that meets the required ripple and transient specifications. It should be noted that these equations ignore the dc series resistance of the filter inductor \(R_{seq}\). This resistance can be quite significant for lower current supplies. It improves filter performance by helping to dampen the filter, which increases the required \(R_{filt}\) and increases the impedance of the filter. Both effects can significantly improve the performance of the filter. It can therefore be very helpful for low noise requirements to trade off a small amount of power loss in \(L_{filt}\) for improved noise performance. Core loss in \(L_{filt}\) also helps to attenuate some of the high frequency noise. Therefore, high-current-powdered iron cores can be a good choice. They also tend to be smaller and cheaper for the same current capability. ADIsimPower of course factors in both the resistance of the filter inductor in addition to the ESR of the two capacitors for maximum accuracy.

Step 8: When choosing actual components to match the calculated values, remember to derate the capacitances of any ceramic capacitors to account for dc bias!

As stated previously, Figure 4 gives two viable techniques for damping the filter. If instead of choosing a parallel resistor, a capacitor \(C_c\) can be chosen to damp the filter. This will add some cost, but it provides the best filter performance of any technique.

**Design Process for an LC filter Using an RC Damping Network (Technique 2 in Figure 4)**

Step 1: As in the previous topology, choose \(C_3\), as if there were not going to be an output filter. 10 mV p-p to 100 mV p-p is a good place to start, depending on the final target output ripple. \(C_3\) can then be calculated using Equation 8. \(C_3\) can be smaller in this topology than the previous topologies because the filter is more effective.

Step 2: As in the previous topology an inductor between 0.5 \(\mu F\) to 2.2 \(\mu F\) is chosen. 1 \(\mu F\) is a good value for converters between 500 kHz and 1200 kHz.

Step 3: As before, \(C_3\) can be chosen from Equation 16, but with \(R_{res}\) set to something large like 1 M\(\Omega\) since it will not be populated. The reason this is the same value despite \(C_3\) having an additional capacitor is that in order to provide good damping, \(R_3\) will be made large enough that \(C_3\) will not significantly reduce the ripple. Set \(C_3\) as the minimum of the calculated \(C_3\) value, \(C_{cap}\) and \(C_i\). It can be useful at this point to return to Step 1 and adjust the ripple assumed on \(C_3\) to get a calculated \(C_3\) that is closer to \(C_{cap}\) and \(C_i\).

Step 4: \(C_3\) should be set to the same value as \(C_3\). In theory you can achieve more damping of the filter using a larger capacitance, but it needlessly adds to the cost and size, and it can reduce converter bandwidth.

Step 5: \(R_3\) can be calculated from Equation 17. \(F_{res}\) is calculated using Equation 7, ignoring the presence of \(C_3\). This is a good approximation since \(R_d\) is typically large enough that \(C_3\) will have little effect on the location of the filter resonance.

\[
R_3 = \frac{1}{\pi C_3 F_{res}} \quad (17)
\]

Step 6: Now that both \(C_3\) and \(R_3\) have been calculated either a ceramic capacitor with a series resistance can be used, or a tantalum or similar capacitor with large ESR should be chosen that matches the calculated specifications.

Step 7: When choosing actual components to match the calculated values, remember to derate the capacitances of any ceramic capacitors to account for dc bias!

Another filter technique is to replace the L in the previous filter with a ferrite bead. However, this arrangement has many drawbacks that limit its effectiveness at filtering switching noise and does almost nothing for switching ripple. First is saturation. The ferrite bead will saturate at a very low level of bias current, meaning that the ferrite will give much lower impedance than shown in the zero bias curves shown in all data sheets. It may still need damping since it is still an inductor and therefore can resonate with the output capacitance. However, now the inductance is variable and poorly characterized in the very minimal data provided in most data sheets. For this reason ferrite beads are not recommended for use as a second stage filter, but can be used downstream from one to further reduce very high frequency noise.
Conclusion
This article has laid out several output filter techniques for switching power supplies. For each topology, a step-by-step design process has been devised to reduce the amount of guess and check required for filter design. The equations have been simplified somewhat so that they are useful to an engineer looking to do a quick design by understanding what is achievable from a second stage output filter.

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