EMC Course Notes 2024

# Filtering and Transient Protection

# Filtering

Filtering plays an important role in EMC compliance. Virtually every electronic product on the market today employs various forms of filtering to meet both emissions and immunity requirements. There are filters on AC power inputs, filters on DC power distribution buses, and filters on most external signal inputs and outputs. Some filters are a single component, while others have several components.

There are two important aspects to filter design for EMC. First, a filter topology and component values must be selected that will provide sufficient attenuation in the correct frequency band(s). Second, and equally important, a filter must be properly laid-out and positioned in a manner that accounts for component parasitics and unintentional coupling paths. Filters that look great in the simulations can perform very poorly in a real product if the layout is not done well.

Filter performance is typically characterized by insertion loss. Insertion loss is the ratio of the power reaching a load with and without the filter in place as a function of frequency. At any given frequency,

$$\text{Insertion Loss in dB} = 10\log\left|\frac{\text{Power Delivered to Load Without the Filter in Place}}{\text{Power Delivered to Load With the Filter in Place}}\right|. \tag{9.1}$$

If the load is a resistance, the insertion loss can also be calculated as 20 times the ratio of the voltages across the load,

$$\text{Insertion Loss in dB} = 20\log\left|\frac{\text{Voltage Across the Load Without the Filter in Place}}{\text{Voltage Across the Load With the Filter in Place}}\right|. \tag{9.2}$$

The definition in (9.2) can also be applied to reactive loads (i.e., loads with a significant capacitance or inductance) in situations where the voltage, rather than the power, is the primary concern.

## Low-Pass Filtering

Most filtering for EMC is low-pass filtering. This is largely because power distribution and a majority of digital signaling takes place at low frequencies, while unwanted coupling tends to become stronger with increasing frequency.

EMC filters are usually 1$^{st}$, 2$^{nd}$, or 3$^{rd}$ order filters. The order of a filter refers to the order of the filter transfer function and is generally equal to the minimum number of reactive components required to build the filter. Higher-order filters do a better job of attenuating frequencies just outside the passband of a filter, but they require more components. The more components a filter uses, the more space it requires and the more vulnerable it is to parasitic interactions that can undermine the filter's performance.

### Series Resistors

Most digital signal terminations and "high-impedance" terminations are capacitances. In nearly all cases, the best way to provide low-pass filtering to a capacitive load is with a series resistor. For the circuit in Figure 9.1, the load voltage, $V_L$, without the series resistor is,

$$V_L = V_S \left| \frac{1}{1 + j\omega R_S C_L} \right| . \tag{9.3}$$

The same voltage with the series resistor is,

$$V_L = V_S \left| \frac{1}{1 + j\omega \left( R_S + R_{series} \right) C_L} \right| . \tag{9.4}$$

The insertion loss is therefore,

$$IL(\omega) = 20 \log \left| \frac{1 + j\omega \left( R_S + R_{series} \right) C_L}{1 + j\omega \left( R_S \right) C_L} \right| . \tag{9.5}$$

At low frequencies ($\omega \ll 1 / \left( R_S + R_{series} \right) C_L$), the insertion loss is approximately 0 dB. At high frequencies ($\omega \gg 1 / \left( R_S + R_{series} \right) C_L$), the insertion loss is approximately, $20 \log \left| \frac{\left( R_S + R_{series} \right)}{\left( R_S \right)} \right|$.

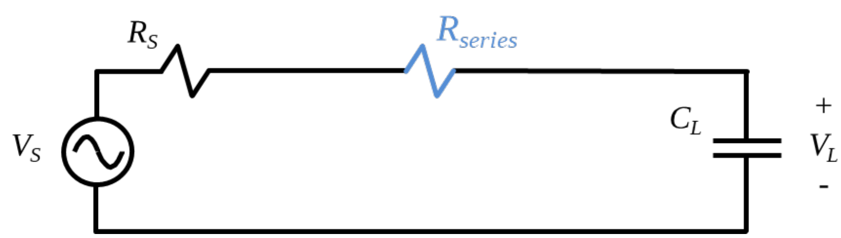



**Figure 9.1.** A series resistor filter.

As illustrated in Figure 9.2, the series resistor attenuates high frequencies by changing the cutoff frequency of the original RC network. For switching circuits, the series resistor slows the transition time, which reduces the power in the upper harmonics. Adding a series resistor allows the circuit designer to take control of the signal bandwidth.

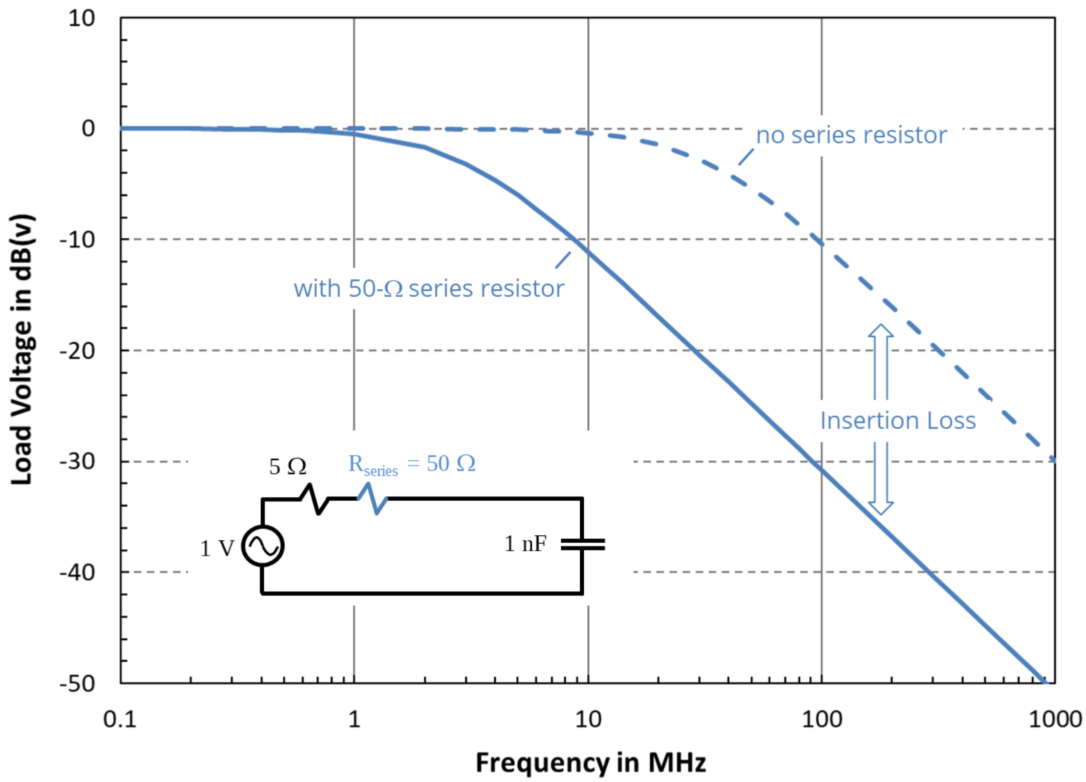



**Figure 9.2.** Effect of a series resistor on a capacitive load voltage.

Note that we could have achieved a similar insertion loss using a capacitor in parallel with the load, but series resistors have several advantages compared to parallel capacitors. A key advantage is that a series resistor reduces the amount of high-frequency current drawn from the source, while a parallel capacitor increases this current. The performance of parallel capacitors is also limited by their parasitic connection inductance.

Ferrite beads or cores are essentially frequency-dependent resistors. It is possible to provide low-pass filtering to a capacitive load using a series ferrite. But again, series resistors have several advantages compared to ferrites. The main advantage is that resistors have a well-defined behavior over a wide range of frequencies. Ferrites, on the other hand, may look inductive at some frequencies setting up the possibility of a resonance with the capacitive load. Ferrites also tend to be larger and more expensive than equivalent resistors for low-pass filter applications.

## Shunt Capacitors

For resistive loads, shunt capacitors (capacitors in parallel with the source and load) are by far the most common EMC filter. The basic configuration is shown in Figure 9.3.

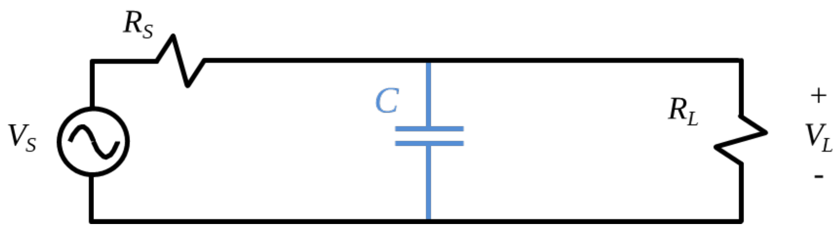



**Figure 9.3.** A shunt capacitor filter.

At low frequencies, where the capacitor impedance is high relative to the source or load impedance, the capacitor has little effect. At high frequencies, current is diverted through the shunt capacitor away from the load.

## Insertion Loss

In the circuit shown above, the capacitor is in parallel with the load resistance, $R_L$. This parallel combination is equal to the product of the impedances divided by the sum:

$$Z_{C||R_L} = R_L \,||\left(\frac{1}{j\omega C}\right) = \frac{R_L\left(\frac{1}{j\omega C}\right)}{R_L + \left(\frac{1}{j\omega C}\right)} = \frac{R_L}{1+j\omega R_L C}. \tag{9.6}$$

Using a voltage divider in the above circuit to express the load voltage, $V_L$, in terms of the open-circuit source voltage, $V_s$ gives:

$$V_L = \left(\frac{Z_{C||R_L}}{Z_{C||R_L} + R_S}\right) V_s . \tag{9.7}$$

We can rewrite (9.7) as a voltage ratio $V_L / V_s$, and expand the factor inside the parentheses using (9.6) to obtain:

$$\frac{V_L}{V_s} = \frac{\left(\frac{R_L}{1+j\omega R_L C}\right)}{\left(\frac{R_L}{1+j\omega R_L C}\right) + R_S} = \frac{R_L}{R_L + R_S + j\omega R_S R_L C} , \tag{9.8}$$

where we multiplied the numerator and denominator by $(1+j\omega R_L C)$ to simplify the expression. To change (9.4) into a form that is more intuitive for understanding insertion loss, we can multiply the numerator and denominator by $1/(R_S + R_L)$ to obtain the form:

$$\frac{V_L}{V_s} = \frac{R_L/\left(R_S + R_L\right)}{1 + j\omega\left(\frac{R_S R_L}{R_S + R_L}\right)C}. \tag{9.9}$$

Notice that the term inside the parenthesis in the denominator of this expression is the parallel combination of the source and load resistances. Defining $R_P = R_S \,||\, R_L$, we can re-write (9.9) as:

$$\frac{V_L}{V_s} = \frac{R_L/\left(R_S+R_L\right)}{1+j\omega R_P C} = \frac{R_L/\left(R_S+R_L\right)}{1+j\omega R_P C}. \tag{9.10}$$

Or the insertion loss as,

$$\text{IL}\,(\omega) = 20\log\left|\frac{V_{L(no\,filter)}}{V_{L(with\,filter)}}\right| = 20\log\left|1+j\omega R_P C\right| \tag{9.11}$$

where $R_P$ is the parallel combination of the source and load resistances, and $C$ is the shunt capacitance.

Figure 9.4 shows the insertion loss of a 1-nF shunt capacitor in a 50-Ω system plotted as a function of frequency. The 3 dB cutoff frequency is $1/2\pi R_P C$ or 6.4 MHz. Note that insertion loss is often plotted upside down, with 0 dB at the top and increasing values moving down the vertical axis. This more intuitively mimics the behavior of the load voltage for a constant amplitude source.

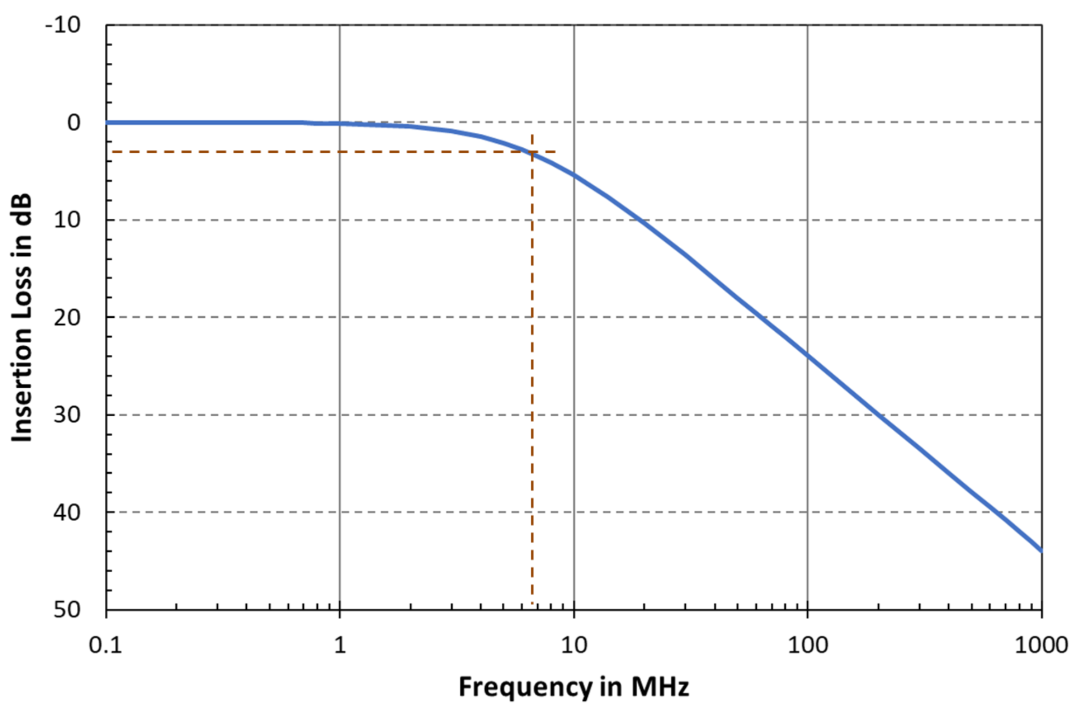



**Figure 9.4.** Insertion loss of a 1-nF shunt capacitor filter with 50-Ω source and load.

## Quick Calculation of Insertion Loss

At frequencies where the filter provides significant attenuation (e.g., >10 dB), $\omega RC >> 1$ and the expression for insertion loss in (9.11) reduces to,

$$\begin{aligned}\text{IL}(\omega) &= 20\log\left|1+j\omega R_P C\right| \\ &\approx 20\log\left(\omega R_P C\right) \\ &= 20\log\left|\frac{R_P}{Z_C}\right|.\end{aligned} \tag{9.12}$$

In other words, the insertion loss is the ratio of the parallel combination of the source and load resistances to the impedance of the shunt capacitor. So, for example, if you want 20 dB of attenuation at a particular frequency, then the capacitor should have an impedance at that frequency that is 1/10$^{th}$ of the parallel combination of the source and load impedances. If you want 40 dB of attenuation, the ratio should be 1/100. This calculation can be done quickly and doesn't involve working with complex numbers.

## Effect of Parasitic Inductance

Of course, any connection of a shunt capacitor to a circuit involves a connection inductance. At high frequencies, this connection inductance can overwhelm the impedance of the capacitor and reduce the insertion loss of the filter significantly.

Figure 9.5 shows the calculated insertion loss of shunt capacitors in three circuits. The first circuit has a 50-Ω source and a 50-Ω load. The second circuit has a 2-Ω source and a 1000-Ω load. The third circuit has a 0.2-Ω source and a 1000-Ω load. In each circuit, a shunt capacitor was added with a value designed to give the filter a 3-dB cut-off frequency at 1 MHz. Note that lower impedance circuits require larger valued capacitors. In each case, the capacitor was modeled as having a 4-nH connection inductance. At frequencies above the self-resonance of the capacitor, the inductance overwhelms the capacitance.

The useful bandwidth of shunt-capacitor low pass filters is greatly diminished in circuits with low-impedance sources or loads. In the 50-Ω circuit, the attenuation of the filter is at least 20 dB between 10 MHz and 100 MHz. With a 2-Ω source impedance, there is only 20 dB of attenuation at frequencies between 6 MHz and 13 MHz. With a 0.2-Ω source impedance, 20 dB of attenuation is only obtained at or near 2.8 MHz.

At the highest frequencies where they are effective, the performance of shunt capacitors will always be determined by their connection inductance. It's important to minimize this inductance, especially in filters that are designed to work at frequencies at 10s of MHz or higher.

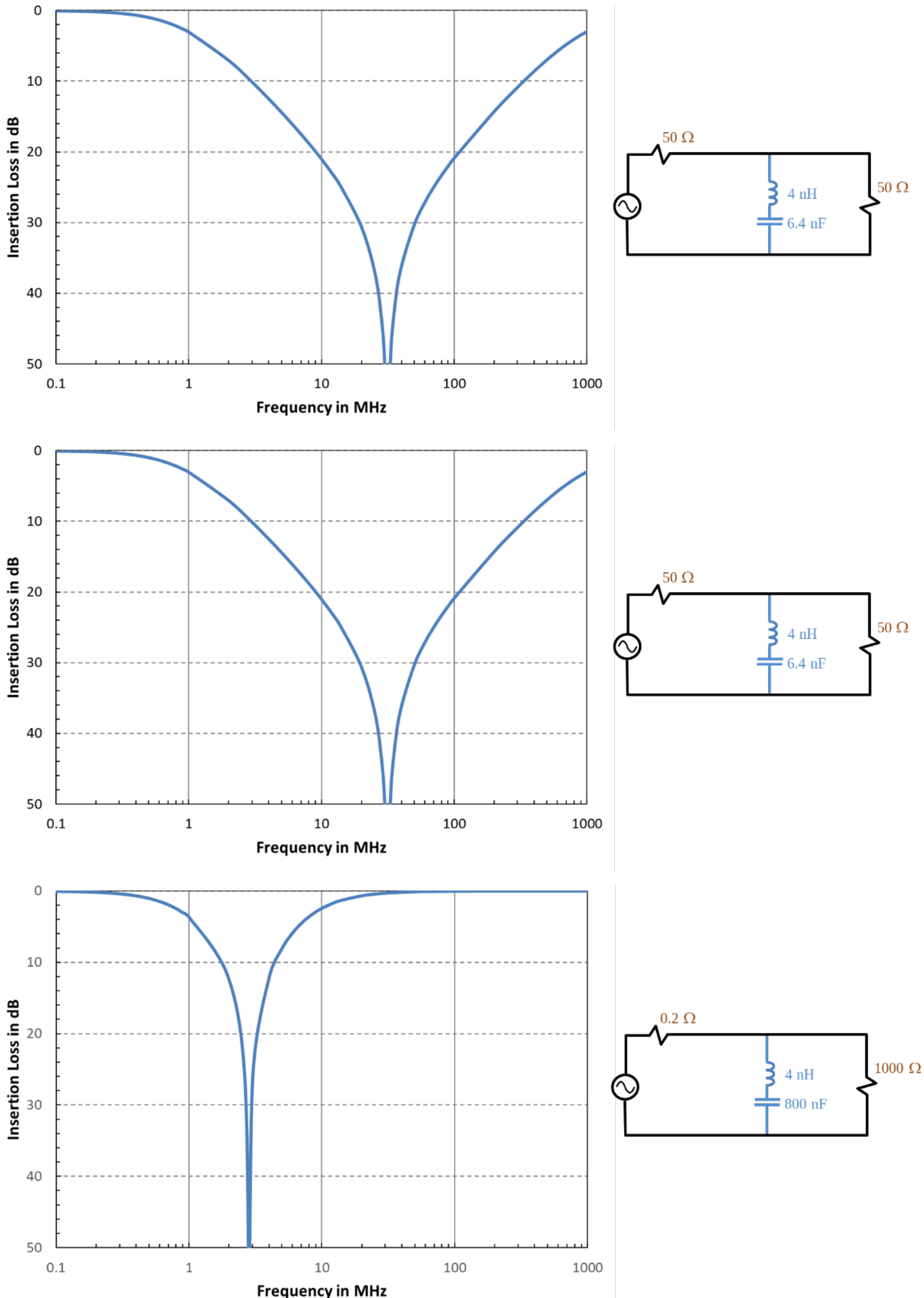



**Figure 9.5.** Insertion loss of three shunt capacitor filters with a 4-nH connection inductance.

## Connection Inductance

It's important to point out that inductance is a property of loops, and the quantity commonly referred to as a "connection inductance" is actually a mutual inductance between the current loop on the input side of the capacitor filter and the current loop on the output side of the capacitor filter. Figure 9.6 illustrates a circuit board signal source driving a load through two traces with a shunt capacitor in between. The high-frequency current flows through the capacitor causing a magnetic flux to wrap the body of the capacitor and the connecting traces. This time-varying flux couples Loop 1 to Loop 2 and induces a voltage in the second loop.

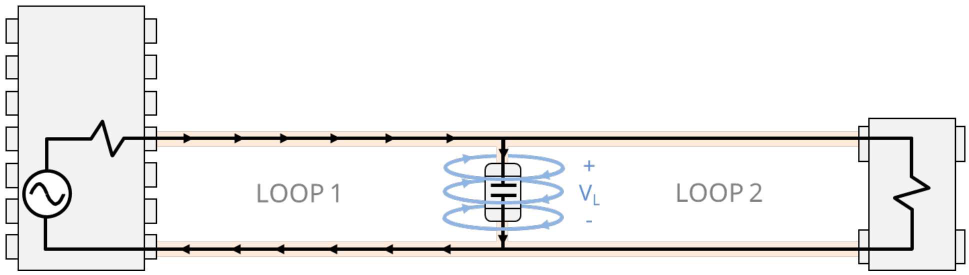



**Figure 9.6.** Current in a filter capacitor creates a magnetic flux that couples noise from the input side to the output side of the filter.

At high frequencies, the voltage appearing across the capacitor is $V_L = j\omega M I_C$, where $M$ is the mutual inductance created by the flux wrapping the capacitor. This is identical to the voltage that would be created by a self-inductance with the same value as $M$. So, from a modeling standpoint, it is often convenient to view this as a self-inductance. However, when laying out the filter, it is important to recognize that it is the mutual inductance between the input and output that must be minimized.

Figure 9.7 illustrates a two-capacitor filter. The two capacitors are connected in parallel, so in a circuit model where each capacitor was assigned a self-inductance, we would expect two identical capacitors to exhibit half the connection inductance that one capacitor had. In the circuit model, the high-frequency insertion loss of two capacitors would be 6 dB higher than the insertion loss of one capacitor.

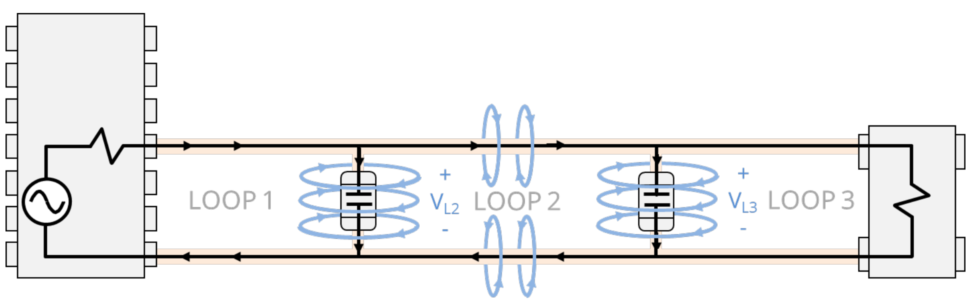



**Figure 9.7.** Current in a filter capacitor creates a magnetic flux that couples noise from the input side to the output side of the filter.

However, recognizing that the inductance is actually a mutual inductance, it becomes clear that the insertion loss associated with two capacitors in parallel depends on the spacing between them. If the capacitors are side-by-side, much of the flux that wraps the first capacitor will also wrap the second capacitor. In this case, the second capacitor will provide little added benefit. However, if there is a significant distance between the two capacitors, the insertion loss can be much higher.

In Figure 9.7, we observe that flux coupling the first capacitor induces a voltage in the second loop. Only a fraction of this voltage is dropped across the second capacitor. The rest is dropped across the signal trace inductance. If the distance between the two capacitors is much greater than the length of the capacitor and its connecting traces, only a small fraction of the voltage across the first capacitor is coupled to the third loop.

Figure 9.8 shows the results of a measurement comparing a one-capacitor low-pass filter to a two-capacitor low-pass filter in the same circuit[1]. The capacitors are connected to a 20-mm microstrip trace carrying the signal from Port 1 to Port 2. The total capacitance in the two filters is about the same, so they exhibit the same insertion loss at low frequencies. At high frequencies, the one-capacitor filter exhibits an insertion loss consistent with a connection inductance of about 1.4 nH. If this were a self-inductance, we might expect two capacitors to have an effective inductance of about 0.7 nH and the insertion loss to be 6 dB higher. But with a 6-mm spacing between the two capacitors, the high-frequency insertion loss is 14 dB higher.

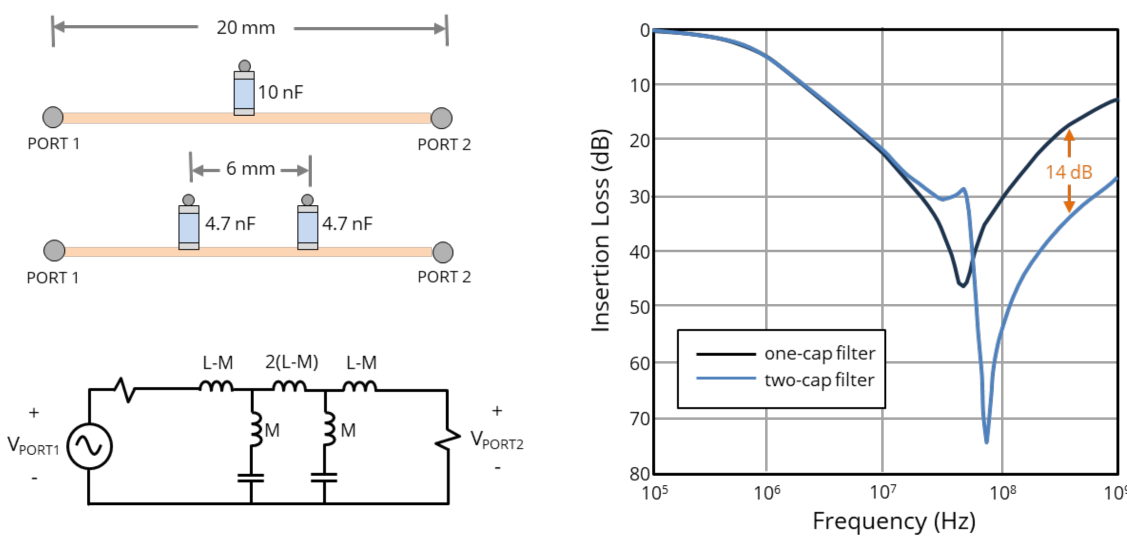



**Figure 9.8.** Insertion loss of one-capacitor and two-capacitor filters.

Two-capacitor low-pass filters can perform much better than one-capacitor filters at frequencies where both capacitors are beyond their self-resonant frequency. In general, the greater the spacing between them, the better they perform (as long as the spacing is less than a half-wavelength). Note, however, that it's important to choose capacitors that have the same nominal value. If the two capacitors have different values, one of them will be self-resonant at a lower frequency than the other one. In between the two self-resonant frequencies, the inductance of one capacitor will resonate with the capacitance of the other.

[1] T. Zeeff et al., "Analysis of simple two-capacitor low-pass filters," *IEEE Transactions on Electromagnetic Compatibility*, vol. 45, no. 4, Nov. 2003, pp. 595-601.

The plot in Figure 9.8 shows a hint of this. The two-capacitor insertion loss plot exhibits a slight peak at about 48 MHz. This is the result of the two capacitors not having exactly the same self-resonant frequency.

It's possible to connect a filter capacitor in such a way that the magnetic flux wrapping the capacitor is partially canceled by flux in the opposite direction from other parts of the connecting structure. Various schemes have been proposed in the literature, though very few of them have been used in actual products. As Figure 9.9 illustrates, these schemes can be reasonably effective. However, the additional space and the necessary precision of the mounting geometry rarely justify the gains in the insertion loss. Typically, a couple of well-mounted shunt capacitors provide sufficient insertion loss with less effort.

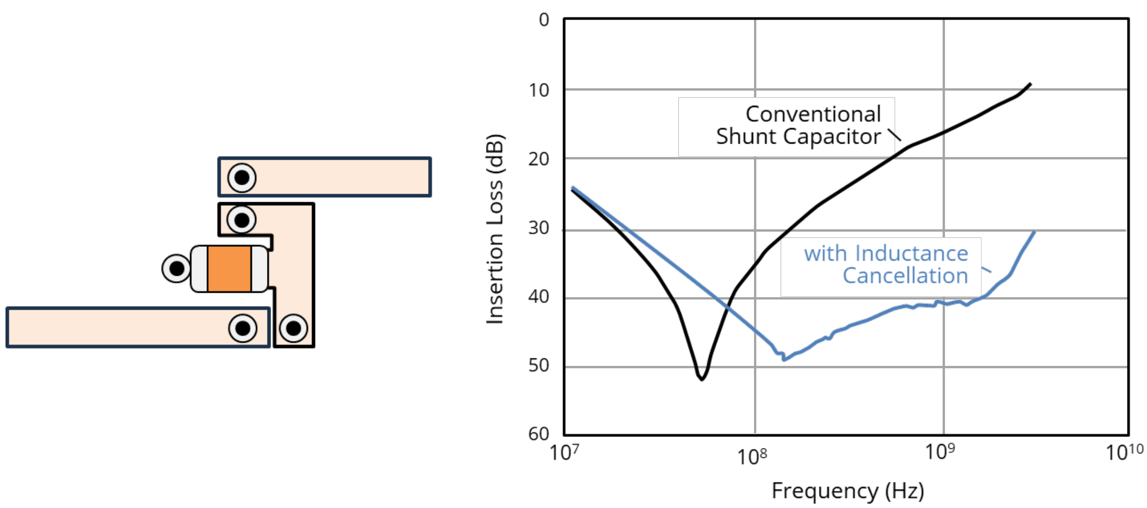



**Figure 9.9.** Shunt capacitor insertion loss with and without inductance cancellation.[2]

## Other Important Parameters

As described in Chapter 2, capacitors not only have a connection inductance, but they also have an equivalent series resistance (ESR). Although this resistance is generally very small (on the order of milliohms), the ESR sets a firm lower bound on the magnitude of the capacitor's impedance. This translates to an upper bound on the maximum obtainable insertion loss. On the other hand, the ESR provides damping that can help reduce the amplitude of parasitic oscillations.

Non-linear behavior is another consideration when choosing filter capacitors. Class 2 ceramic capacitors have capacitances that vary with the applied voltage. This can distort the applied signals creating unwanted harmonics. For this reason, Class 2 ceramic capacitors (e.g., X7R or Z5U) are typically used only in situations where the applied voltage has a strong DC component that doesn't vary significantly. Class 1 ceramics capacitors (e.g., C0G or NP0) are more stable and generally more suitable for applications where the voltage across them varies significantly.

[2] A. McDowell and T. Hubing, "A compact implementation of parasitic inductance cancellation for shunt capacitor filters on multilayer PCBs," *IEEE Trans. on Electromagnetic Compatibility*, vol. 57, no. 2, Apr. 2015, pp. 257-263.

Another important consideration is the maximum rated voltage of a capacitor. In applications where this maximum voltage may be exceeded (even briefly), capacitors are more likely to fail resulting in short or open circuits and a loss of filtering. If these failures go undetected, products may become vulnerable to electromagnetic interference without warning.

## Series Inductors

A series inductor can also be used to make a first-order low-pass filter with resistive sources and loads. The basic configuration is shown in Figure 9.10.

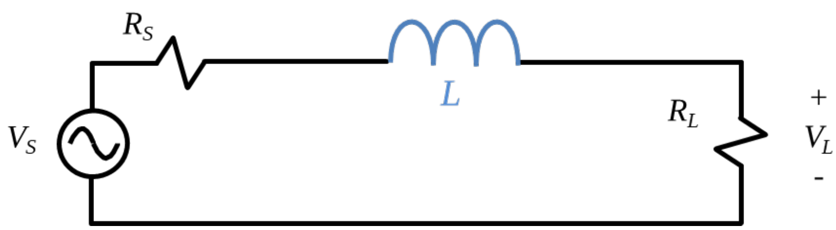



**Figure 9.10.** A series inductor filter.

Series inductors are generally more effective than shunt capacitors when the source and load resistances are low.

### Insertion Loss

Using voltage division, $V_L$ can be expressed in terms of the input voltage, $V_S$, as

$$V_L = \left( \frac{R_L}{R_S + R_L + j\omega L} \right) V_s . \tag{9.13}$$

Dividing the numerator and denominator by $R_S + R_L$ yields,

$$V_L = \frac{R_L / \left( R_S + R_L \right)}{1 + j\omega L / \left( R_S + R_L \right)} V_S . \tag{9.14}$$

The numerator in this expression times $V_S$ is the value of $V_L$ with no series inductor. So, the insertion loss is,

$$\text{IL}(\omega) = 20 \log \left| \frac{V_{L(no\,filter)}}{V_{L(with\,filter)}} \right| = 20 \log \left| 1 + \frac{j\omega L}{R_S + R_L} \right| . \tag{9.15}$$

### Quick Calculation of Insertion Loss

At the frequencies where the filter provides significant attenuation (e.g., >10 dB), $\omega L \gg R_S + R_L$ and the expression for insertion loss in (9.15) reduces to,

$$\text{IL}(\omega) = 20 \log \left| \frac{\omega L}{R_S + R_L} \right| . \tag{9.16}$$

In other words, the insertion loss is the ratio of the inductive reactance to the series combination of the source and load resistances. So, for example, if you want 20 dB of attenuation at a particular frequency, then the inductor should have an impedance at that frequency that is 10 times the series combination of the source and load impedances. If you want 40 dB of attenuation, the ratio should be 100. This calculation can be done quickly and doesn't involve working with complex numbers.

### Effect of Parasitic Capacitance

The impedance of inductors is limited by their parasitic capacitance at high frequencies. Small-valued inductors with few windings may have very little parasitic capacitance (on the order of picofarads), while large tightly wound inductors with many turns can have much higher parasitic capacitances.

### Other Important Parameters

Inductors also have an equivalent series resistance (ESR). However, in most filtering applications this resistance is not important, because it is in series with a much higher inductive reactance. The inductor resistance is more likely to be an issue in applications where the inductor is part of a high-Q resonant circuit.

Non-linear behavior is also a consideration when choosing filter inductors. All inductors have a maximum rated current. Inductors with ferrite cores can saturate if the current is exceeded. This results in a significant reduction in the nominal inductance.

Unshielded inductors may also have strong external magnetic fields. These fields can couple to nearby circuits or induce eddy currents in nearby metal planes.

## Series Ferrites

Series ferrite beads or cores are a third single-component option for low-pass filtering. Like inductors, they have little impedance at low frequencies and provide a higher impedance at high frequencies. Unlike inductors though, their impedance is mostly resistive, therefore they are less likely to resonate with other components. In fact, ferrites are commonly used to damp resonances in circuits or structures where resistors cannot be used.

One common application of ferrite beads is on DC power inputs or outputs. Power bus impedances are typically resistive with a potentially strong inductive or capacitive component that varies with frequency. Series resistors are inappropriate in this application because they reduce the DC voltage. Series inductors and shunt capacitors both run the risk of resonating with the power input or output impedance. Series ferrites allow the DC current to pass while impeding high frequencies.

Of course, it's important to choose a ferrite that will have significant impedance at the frequencies that need to be attenuated. It's also very important to ensure that the current drawn by the power input cannot exceed the saturation current of the ferrite.

Ferrite beads can be used in other applications where a high-frequency resistance is desired, and a DC resistance cannot be tolerated. Sometimes they are added to higher-order filters as a means of preventing unintended resonances with a reactive source or load impedance. As indicated earlier, ferrites should generally not be used in situations where a resistor would have worked just as well or better.

## Second-Order Low-Pass Filters

A series inductor and shunt capacitor can be combined to form a low-pass filter as shown in Figure 9.11. With two reactive components, LC filters are second-order filters. Second-order filters have two important advantages over first-order filters. One advantage is a sharper increase in insertion loss above the cut-off frequency. The insertion loss of first-order filters is proportional to frequency (20 dB/decade) above cut-off. Second-order filters can exhibit insertion losses that are proportional to the frequency squared (40 dB/decade).

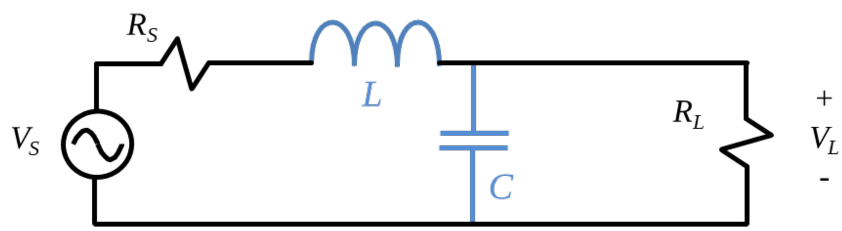



**Figure 9.11.** An LC filter.

Another advantage of LC filters is that they can be used with low-impedance sources and high-impedance loads. This situation arises often because efficient power and signal sources tend to have a low impedance, while efficient power and signal loads tend to have a high impedance. If the inductor is in series with the low-impedance source and the capacitor is in parallel with the high-impedance load, both components provide effective filtering over a wide frequency range.

Figure 9.12 shows the calculated insertion loss of an LC filter with a low-impedance source and a high impedance load. Note the 40 dB/decade fall-off at frequencies above the cut-off frequency. Compared with Figure 9.4, which shows the insertion loss of a first-order filter with the same cut-off frequency, the 40 dB/decade fall-off produces much greater attenuation at higher frequencies.

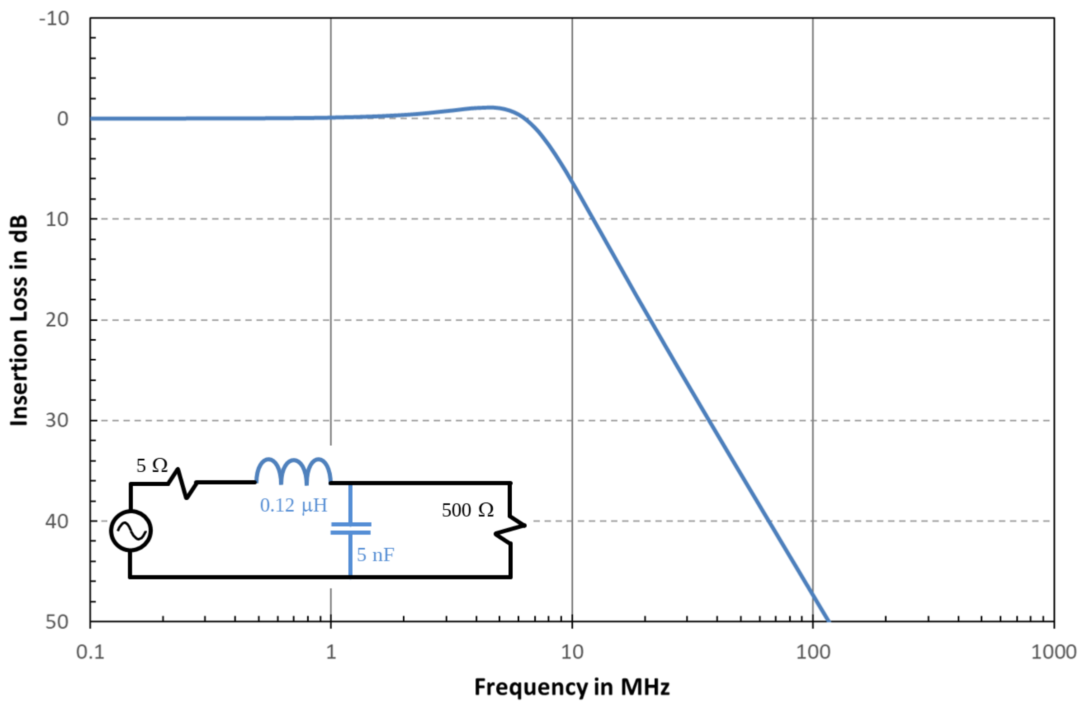



**Figure 9.12.** Insertion loss of an LC filter.

A disadvantage of LC filters is that a resonance can form between the inductor and the capacitor. This resonance can result in a negative insertion loss near the cut-off frequency. Figure 9.13 shows the insertion loss of an LC filter that has the same cut-off frequency as the filter in Figure 9.12. Note the significant negative insertion loss between 2 and 8 MHz.

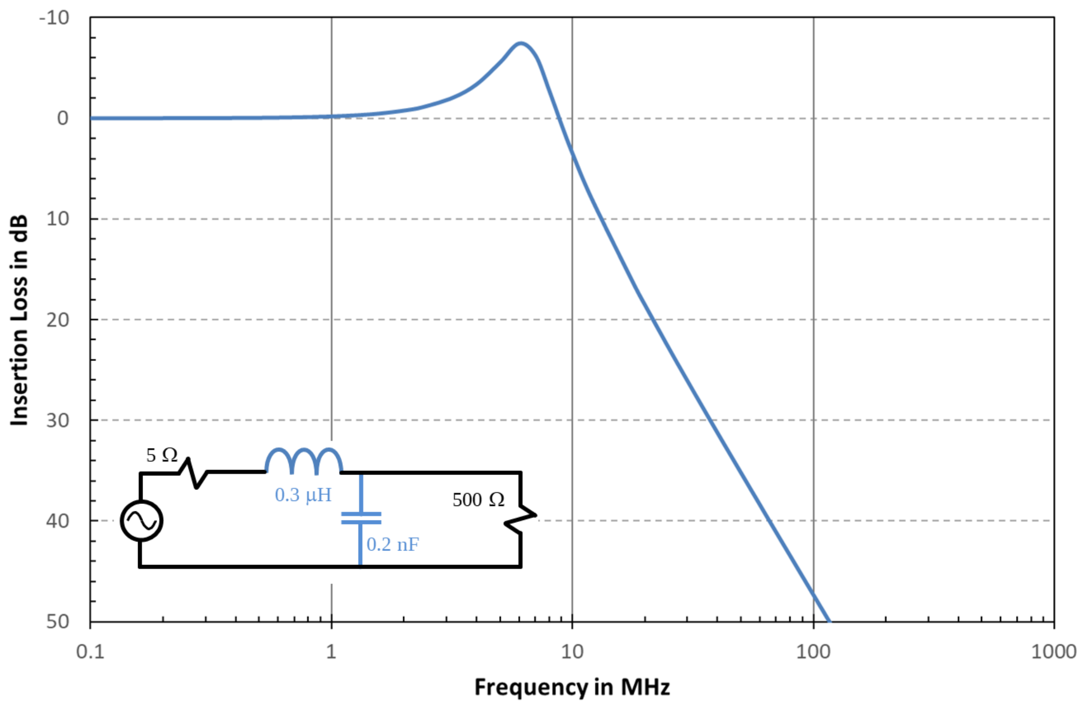



**Figure 9.13.** Insertion loss of an under-damped LC filter.

To avoid having a negative insertion loss, it is generally important to choose filter element values that will not exhibit a strong resonance. This can be accomplished by matching the filter to either the source or load resistance using the following procedure:

1. Determine the source resistance, $R_S$, the load resistance, $R_L$, and the desired cut-off frequency, $f_c$.
2. To match the filter to the source, choose $L = \frac{R_S}{2\pi f_C}$ , $C = \frac{1}{(2\pi f_C)^2 L}$.
3. Or instead, to match the filter to the load, choose $C = \frac{1}{2\pi f_C R_L}$, $L = \frac{1}{(2\pi f_C)^2 C}$.

It is not necessary or desirable to try to match both the source and the load simultaneously. If one of the two resistances is unknown, match the other one. The theoretical insertion loss will be the same either way. However, typically for low-impedance sources, matching the filter to the source yields more convenient component values.

Note that when the filter is matched to the source, $\sqrt{L/C} = R_S$. When the filter is matched to the load, $\sqrt{L/C} = R_L$. When $\sqrt{L/C} < R_S$, the filter is overdamped. When $\sqrt{L/C} \ll R_S$, the inductor no longer contributes, and the insertion loss becomes that of an RC filter.

The filter also becomes overdamped when $\sqrt{L/C} > R_L$. When $\sqrt{L/C} \gg R_L$, the capacitor no longer contributes, and the insertion loss becomes that of an RL filter. The filter is underdamped and exhibits a negative insertion loss near cut-off when $R_S < \sqrt{L/C} < R_L$.

Of course, as indicated in Figure 9.12, even matched filters exhibit a slight negative insertion loss near the cut-off frequency. If necessary, this can be eliminated by choosing a slightly lower value of inductance when matching the source, or a slightly lower value of capacitance when matching the load. Either way, the other component's value should be increased so that the product of L and C is still, $LC = \frac{1}{(2\pi f_C)^2}$.

## Third-Order Low-Pass Filters

Pi-filters and T-filters, like those illustrated in Figure 9.14, are third-order filters. They are capable of exhibiting 60 dB/decade fall-offs above the cut-off frequency, and they can work with both high- and low-impedance sources and loads. However, they can also exhibit resonances and must be designed to ensure that these resonances will be damped.

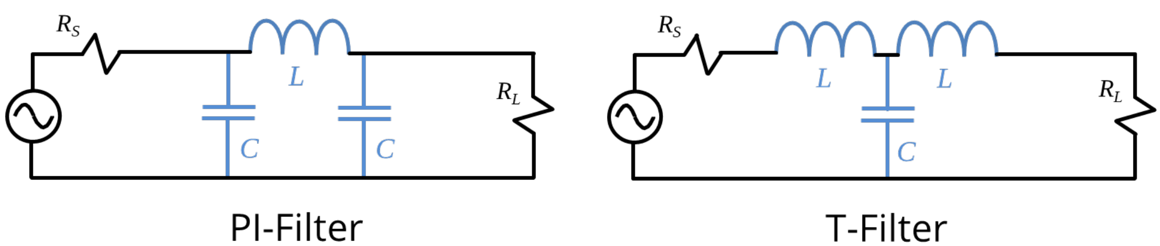



**Figure 9.14.** Pi-filter and T-filter topologies.

One way of ensuring that the resonances are sufficiently damped is to match the filter to the source and the load resistances. The general procedure for designing a matched pi-filter is described here:

1. Determine the source resistance, $R_S$, the load resistance, $R_L$, and the desired cut-off frequency, $f_c$.
2. Choose the input capacitor value to match the source resistance at cut-off, $C_{in} = \frac{1}{2\pi f_C R_S}$.
3. Choose the output capacitor value to match the load resistance at cut-off, $C_{out} = \frac{1}{2\pi f_C R_L}$.
4. Choose the inductor value to be equal to $L = \frac{1}{(2\pi f_C)^2 C_{series}}$, where $C_{series} = \frac{C_{in} C_{out}}{C_{in} + C_{out}}$.

Figure 9.15 shows an example of a pi-filter matched to a 5-Ω source and a 500-Ω load. Compared to the plots in Figure 9.4 and Figure 9.12, the 60-dB roll-off provides significantly greater attenuation at high frequencies.

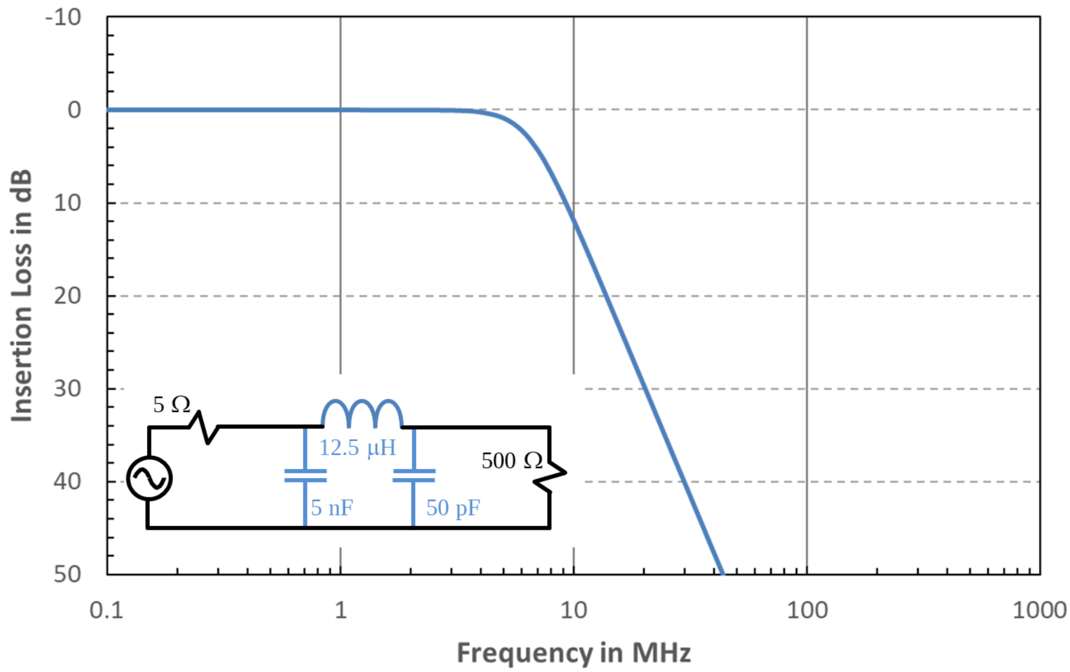



**Figure 9.15.** Insertion loss of a pi-filter matched to a 5-Ω source and 500-Ω load.

A significant disadvantage of higher-order filters is the impact that component parasitics can have on the insertion loss. It is possible to design a filter that works great in the simulations but fails miserably when it is implemented in a product. The best way to minimize the design risk is to keep the filter as simple as possible while still meeting the insertion loss requirements. Filter designers should always estimate the value of the more significant parasitics and include them in any models of the filter performance.

When either the source or load resistance is not known (or includes a significant reactance) it is not possible to design a matched filter. A filter that is designed for a 50-Ω system may exhibit strong resonances if either the source or load resistance is not 50 Ω.

Figure 9.16 compares the insertion loss of a 50-Ω matched filter in a 50-Ω system with the same filter's insertion loss in a 5-Ω/500-Ω system. There is a strong resonance near the cut-off frequency in the unmatched system.

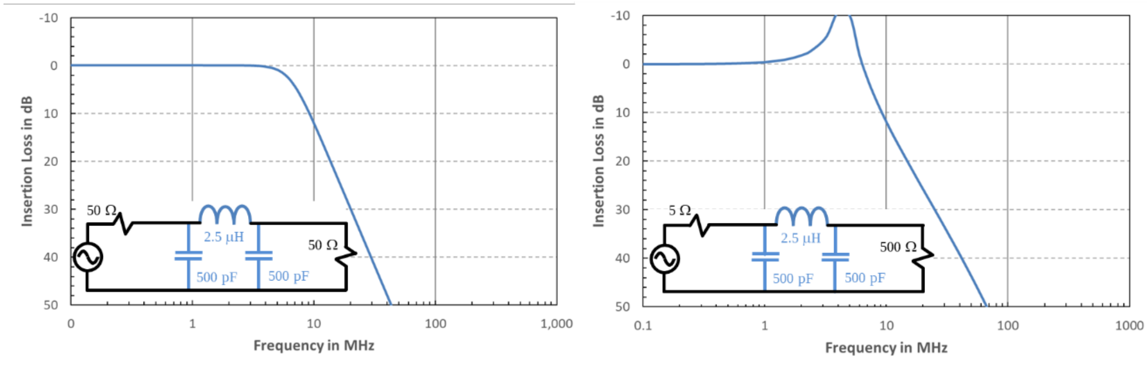


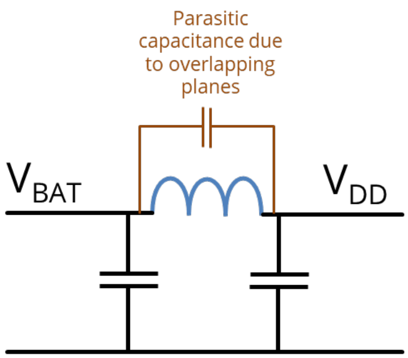

**Figure 9.16.** Insertion loss of a matched 50-Ω pi-filter in two circuits.

If the source or load impedance has a significant capacitive component, that should be included in the pi-filter calculation. If either the source or load has a significant inductive component, a pi-filter is probably not the right choice. When filtering an inductive load, a T-filter, or possibly a lower-order filter, is generally more appropriate. Ultimately, it's important to ensure that the reactance of source and load impedances won't resonate with the filter elements.

## Filter Location and Layout

Locating and laying out a filter is just as important as selecting its component values. Filter designers need to ensure that the components selected are appropriate to the application and that they will maintain their nominal values over the entire bandwidth of the filter. It's also important to ensure that the components are well-mounted to minimize their own parasitics as well as their mutual capacitance and mutual inductance with other components.

In addition, it's very important to ensure that noise does not find a path that essentially bypasses the filter. Figure 9.17 illustrates this in a circuit board layout with overlapping $V_{BAT}$ and $V_{DD}$ planes. The parasitic capacitance between the planes provides a path around the filter for high-frequency noise currents.



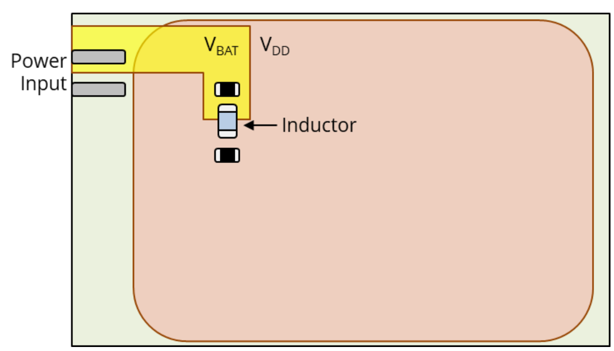




**Figure 9.17.** Overlapping planes may allow noise to bypass the filter.

While problems like this are often obvious once they are pointed out, this is a very common problem in circuit board layouts. Unless the board designer is specifically looking for these sneak noise paths, they are unlikely to be noticed. For this reason, it is a good idea to adopt the following procedure when reviewing any board design with a filter.

1. Using the board layout viewer color all the nets on the board the same color (e.g., red).
2. Find the filter in the schematic and label one side “quiet” and one side “noisy.” It doesn’t matter which label goes on which side. Ground (if there is one) is neither noisy nor quiet, it is simply the reference by which we quantify the noise voltage on the other nets.
3. On the side of the filter with the fewest nets, color all those nets a contrasting color (e.g., yellow). Color the ground (if there is one) green.
4. Now search layer by layer for any places where red and yellow nets come near each other or overlap. Those places represent areas where noise can couple from the noisy side to the quiet side without having to pass through the filter.
5. Note that nets that occupy the same board area but are isolated from each other by an intervening ground plane do not present a potentially significant coupling path.
6. Be sure to consider coupling between components mounted on the surface of the board. These are not visible in the board layout files but may be responsible for significant coupling.

In most cases, problematic coupling issues become obvious as soon as they are observed. If there is any doubt, the isolation can be improved, or a simulation can be performed to quantify the amount of potential coupling.

## High-Pass Filtering

Any of the low-pass filter topologies that work with resistive sources and loads discussed above can be converted to a high-pass filter by replacing inductors with capacitors and capacitors with inductors. If the replacement elements have the same reactance at the cut-off frequency, the resulting filter will have the same cut-off frequency. The insertion loss would be near zero above the cut-off and roll-off below cut-off. The same rules would apply regarding parasitics and self-resonances.

In practice, for EMC-related purposes, high-pass filtering using anything other than a first-order blocking capacitor is rare. Blocking capacitors are used to isolate circuits that operate with different DC bias voltages. They are also often used to establish a high-frequency connection between two “ground” conductors that require low-frequency isolation for safety purposes. Isolated grounds are common in medical products, some aerospace systems, and high-voltage industrial controls.

When conductors labeled “ground” are isolated, it is generally to control the flow of low-frequency (e.g., kHz or lower) currents. However, a small amount of high-frequency voltage between two large conductors can facilitate radiated emissions and immunity problems, so it is often necessary to establish a good high-frequency connection between isolated grounds.

Figure 9.18 illustrates two methods for establishing a high-frequency connection between isolated metal planes on a circuit board. If the planes are on the same layer, several capacitors can be used to bridge the gap between the planes. The connection inductance of each capacitor will be on the order of several nanohenries, so many capacitors distributed along the length of the gap are typically required to obtain a reasonably good high-frequency bond.

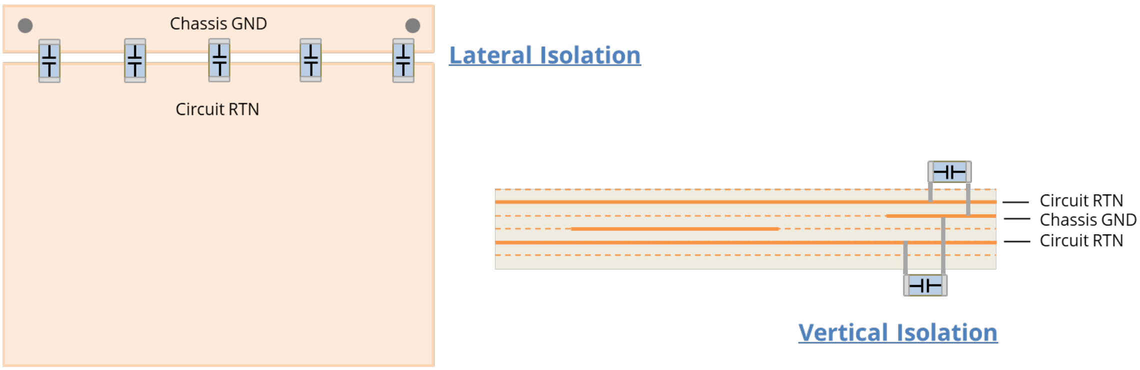



**Figure 9.18.** Isolated planes on the same layer vs. isolated overlapping planes.

A much better high-frequency connection can be obtained by locating the planes on adjacent planes with a significant overlap. This allows the bonding capacitors to have a much lower connection inductance. In addition, the parasitic capacitance between the planes can provide significant bonding at GHz frequencies where the connection inductance of the capacitors limits their effectiveness.

## Band-Pass Filtering

Band-pass filters are generally applied in situations where an RF signal has a relatively narrow bandwidth. In that case, anything picked up by the receiver outside that bandwidth is noise. High-order band-pass filters can be very selective and may be implemented in hardware or software. Hardware filters can be either passive or active.

### Simple Passive Filters

In some digital signal applications, it is necessary to block the DC component while also filtering the highest frequencies. Band-pass filters with a wide bandwidth are often formed by combining a high-pass and low-pass filter.

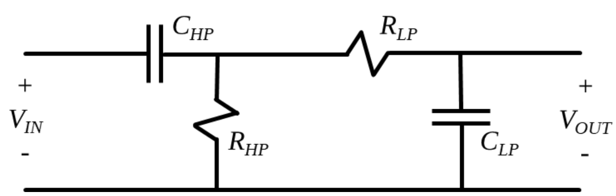



**Figure 9.19.** High-pass RC filter followed by a low-pass RC filter.

For example, the circuit in Figure 9.19 shows a high-pass RC filter followed by a low-pass RC filter. For a high-impedance termination, the low-frequency cut-off frequency is approximately,

$$f_{c-low} \approx \frac{1}{2\pi R_{HP} C_{HP}} . \tag{9.17}$$

The high-frequency cut-off frequency is approximately,

$$f_{c-high} \approx \frac{1}{2\pi R_{LP} C_{LP}} . \tag{9.18}$$

Note that when driving capacitive loads, the load capacitance can replace $C_{LP}$.

For narrower bandwidths, an RLC filter may be appropriate, as illustrated in Figure 9.20. This filter has a center frequency given by,

$$f_{center} \approx \frac{1}{2\pi\sqrt{LC}} , \tag{9.19}$$

and, for high-impedance loads, a 3-dB bandwidth given by,

$$\Delta f_{\text{3-dB}} \approx \frac{R}{L} . \tag{9.20}$$

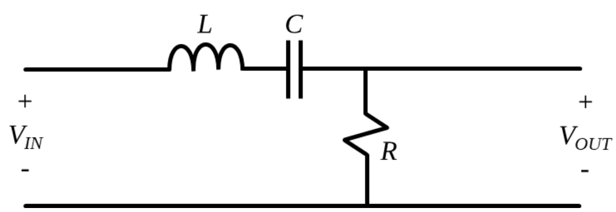



**Figure 9.20.** A series RLC bandpass filter.

## Simple Active Filters

Filters employing active devices, such as amplifiers, can be advantageous in situations where the signal-to-noise ratio is a major concern, or when very narrow bandwidths are required. There are a wide range of options for designing active filters, and numerous books, papers, and online articles describe active filter design.

An example of a simple active bandpass filter is shown in Figure 9.21. Like many active filters, frequency-dependent feedback to a high-gain amplifier determines the frequency response. In this case, the lower cut-off frequency is determined by $R_1$ and $C_1$, while the upper cut-off frequency is determined by $R_2$ and $C_2$.

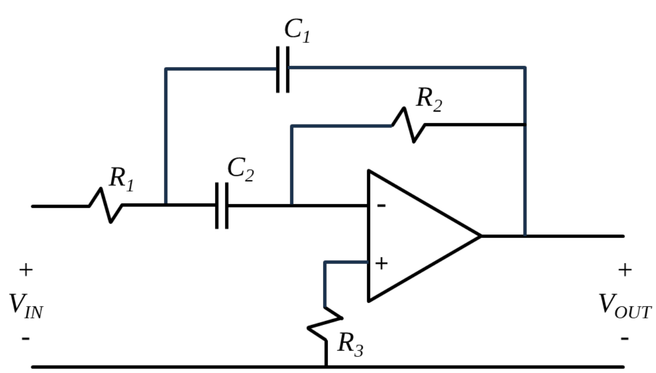



**Figure 9.21.** An active bandpass filter.

Active filters that employ high-gain amplifiers are potentially susceptible to radiated immunity problems. Small voltages directly coupled to the input of the amplifier can drive the amplifier outside its linear range. Care must be taken when laying out an active filter on a printed circuit board. The loop areas associated with the amplifier input and feedback must be kept small. The entire circuit should be located over a solid return plane, and the input and feedback components should lie flat over the plane.

## Common-Mode Filtering

So far, all of the filters discussed have been filtering the differential-mode signal on two conductors. However, in Chapter 7 we saw that two conductors are also capable of carrying a common-mode current. In many cases, it is the common-mode current that is responsible for unwanted EM coupling. In those cases, it is the common-mode current that needs to be filtered.

As discussed in Chapter 7, the term "common-mode" is commonly used to describe two very different phenomena in transmission lines with three or more conductors. The common-mode current can refer to the current propagating in one direction on the two signal conductors and returning on the other conductors. Or, it can refer to the current flowing in one direction on all the transmission conductors without a nearby return conductor.

### Filtering Balanced Three-Conductor Lines

Consider the conducted emissions noise source illustrated in Figure 9.22(a). The switching of a flyback converter generates a time-varying voltage between the hot and neutral conductors of the device's power input. This voltage appears across the 50-Ω resistances of the two LISNs used for the conducted emissions measurement. The hot and neutral conductors have the same impedance to ground in the device under test, the power cord, and the LISNs. Therefore, the system is balanced, and the switching source can be modeled using the differential-mode equivalent source shown in Figure 9.22(b).

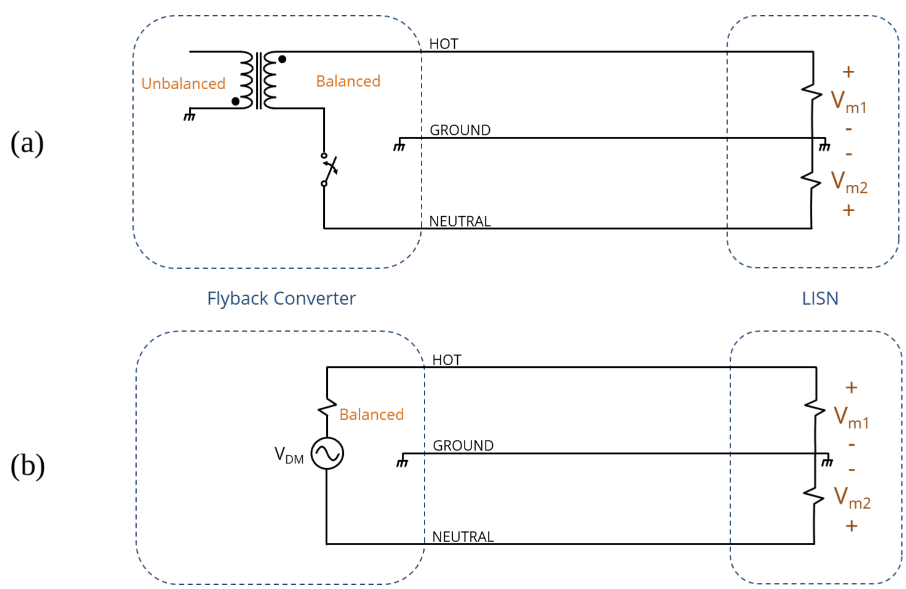



**Figure 9.22.** Model of conducted emissions generated by a flyback converter.

Note that the LISN voltages are defined relative to ground, so differential-mode noise currents flowing into the hot side of LISN 1 and out of the neutral side of LISN 2 generate measurement voltages that have the same magnitude, but opposite polarity. In this case, the differential-mode voltage measured by the LISNs is,

$$V_{\text{differential-mode}} = V_{m1} - V_{m2} = V_{m1} + V_{m1} = 2V_{m1}. \tag{9.21}$$

The common-mode voltage measured by the LISNs is,

$$V_{\text{common-mode}} = \frac{V_{m1} + V_{m2}}{2} = \frac{V_{m1} - V_{m1}}{2} = 0. \tag{9.22}$$

If $V_{m1}$ is above the conducted emissions limit, filtering may be required to reduce it. Suppose we attempt to reduce the amplitude of $V_{m1}$ using the filter capacitor shown in Figure 9.23. If this capacitor has an impedance well below 50 Ω at the measurement frequency, it diverts current away from LISN 1 and successfully reduces $V_{m1}$. However, this current is still routed through $V_{m2}$. In fact, $V_{m2}$ is likely to be higher than it was without the filter causing the product to be further over the limit.

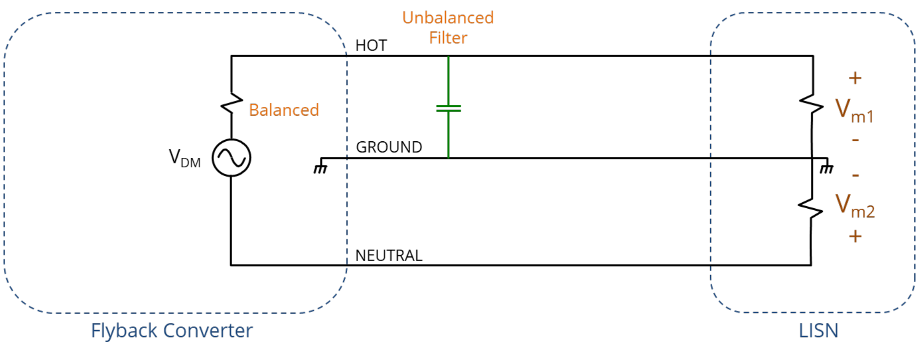



**Figure 9.23.** An unbalanced filter in a balanced system.

If the source impedance and the filter impedance are both much less than 50 Ω at the measurement frequency, the differential-mode voltage measured by the LISNs is,

$$V_{\text{differential-mode}} = V_{m1} - V_{m2} \approx 0 + V_{m2} = -V_{m2}. \tag{9.33}$$

The common-mode voltage measured by the LISNs is,

$$V_{\text{common-mode}} = \frac{V_{m1} + V_{m2}}{2} = \frac{0 - V_{m2}}{2} = \frac{V_{m2}}{2}. \tag{9.34}$$

In general, applying an unbalanced filter to a balanced system forces a mode conversion that can have undesirable consequences. For this reason, filters in balanced systems should be balanced.

Consider the balanced filter design in Figure 9.24. Both power conductors are filtered using the same impedance to ground. Both $V_{m1}$ and $V_{m2}$ are reduced by the same amount. There is no conversion from differential mode to common mode.

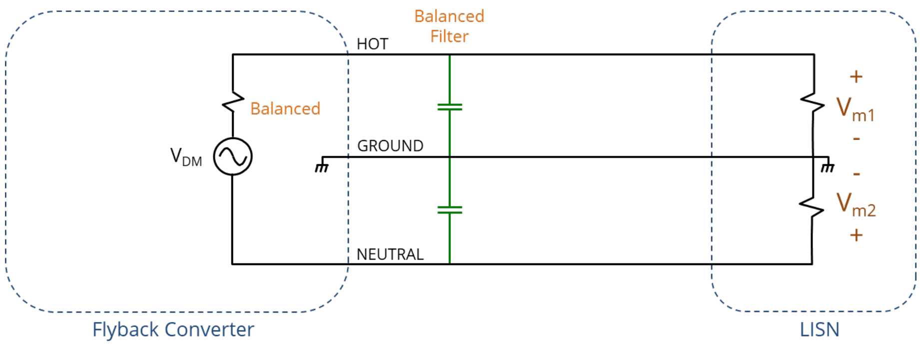



**Figure 9.24.** A balanced Y-capacitor filter in a balanced system.

Capacitors that connect balanced pairs of power conductors to ground are called Y-capacitors. However, for *differential-mode* noise in a balanced system, it is not necessary to connect any of the filter elements to ground. A single capacitor with half the nominal value of the Y-capacitors between the hot and neutral wires would provide the same insertion loss. Capacitors that connect directly between the two power conductors in a balanced system are called X-capacitors. Figure 9.25 shows an X-capacitor filter.

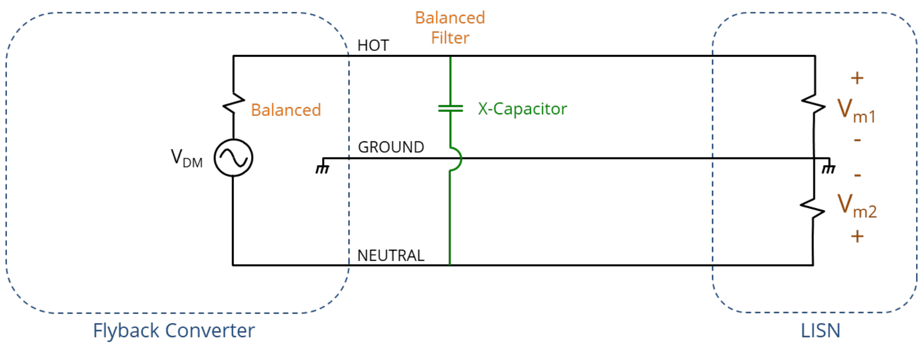



**Figure 9.25.** A balanced X-capacitor filter in a balanced system.

The source impedance of a power converter can be a very small value. In this case, shunt capacitors may not provide enough insertion loss to meet a conducted emissions requirement. To achieve higher insertion loss, second or third-order filters can be applied. Figure 9.26 shows a third-order pi-filter applied in a balanced system. Note that both sides of the filter maintain the same impedance to ground.

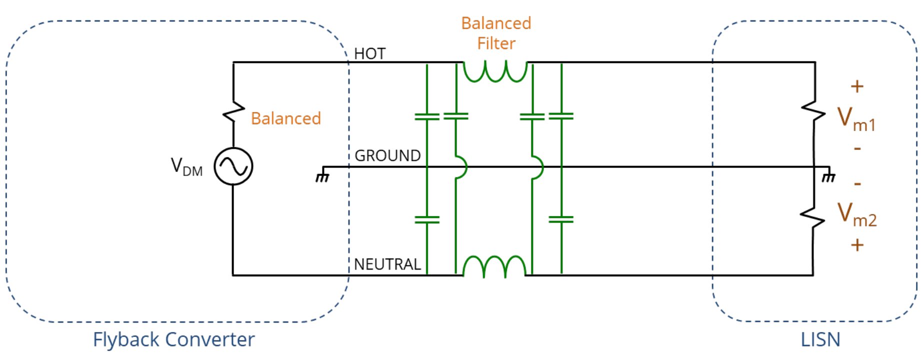



**Figure 9.26.** A balanced third-order filter.

If the source of the noise drives both power conductors relative to the LISN ground, it will have a common-mode component. Suppose the source of the conducted emissions is switching noise that drives a heatsink relative to circuit board ground on the secondary side of the transformer as illustrated in Figure 9.27.

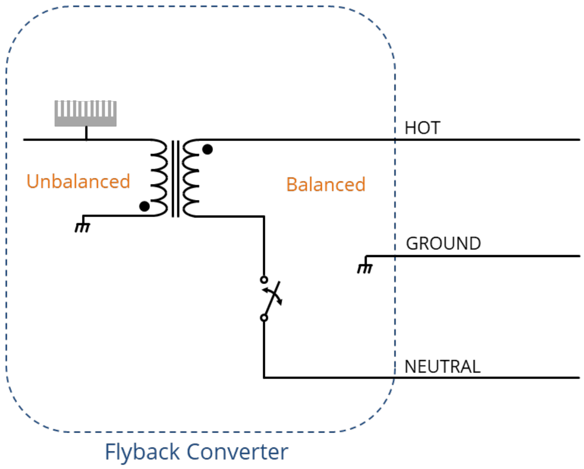



**Figure 9.27.** An unbalanced noise source in a flyback converter.

The voltage between the heatsink and circuit board generates electric field lines as illustrated in Figure 9.28. If the board and the heatsink do not have the same impedance to the chassis ground, then the number of field lines emanating from the heatsink is not equal to the number of field lines terminating on the circuit board. This is an unbalanced source relative to chassis ground, and it produces a voltage that drives the power lines relative to the chassis ground.

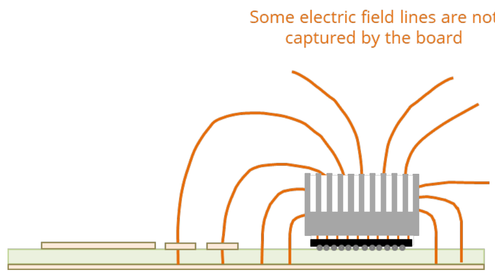



**Figure 9.28.** Electric field lines between a heatsink and circuit board.

This noise source drives the balanced power cord and LISNs with a common-mode voltage that can be modeled as illustrated in Figure 9.29. Note that the common-mode noise currents flow into the positive sides of both LISNs to the LISN ground. Common-mode currents generate voltages that have the same magnitude and the same polarity in both LISNs. Since the hot and neutral inputs still have the same impedance to ground, the conducted emissions source is still balanced.

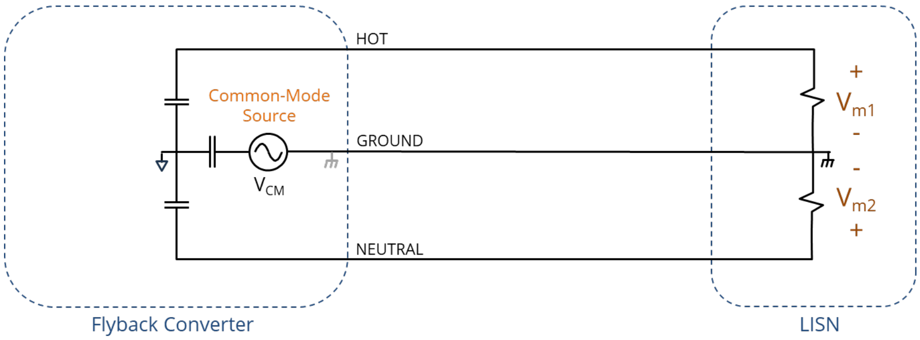



**Figure 9.29.** Model of a common-mode conducted emissions source.

Note that the balanced filters in Figures 9.24 and 9.26 attenuate common-mode noise as well as differential-mode noise. In many cases, the two filter inductors in Figure 9.26 are replaced with a common-mode choke. The common-mode choke provides a significant impedance (inductance) to common-mode currents. Differential-mode currents, on the other hand, see very little inductance.[3]

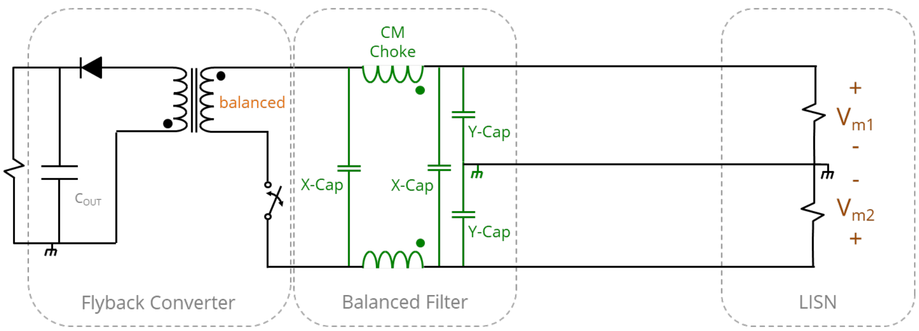



**Figure 9.30.** A common third-order filter for balanced power inputs.

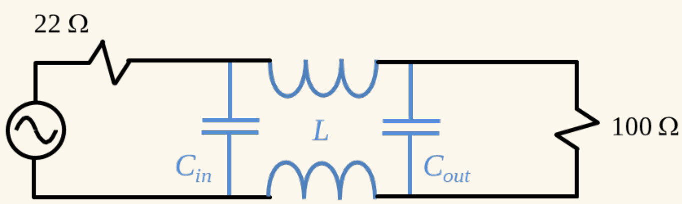
The power currents produce magnetic fields in the ferrite core of a common-mode choke that cancel each other. Because of this, common-mode chokes can handle much higher power currents than similarly sized uncoupled inductors without saturating. This means that, for a given current rating, a common-mode choke is typically much smaller than two uncoupled inductors. The filter in Figure 9.30 relies on the X-capacitors to provide most of the differential-mode attenuation. The common-mode choke and the Y-capacitors provide the common-mode attenuation.

[3] Since the coupling between the two windings of a common-mode choke is never perfect, differential currents also experience a small inductance. This inductance can be utilized to attenuate differential-mode noise to some extent.

**Example 9-1: AC Power Line Filter**

*Design a balanced differential-mode pi-filter for a 110-v, 5-A, 60-Hz AC power input with a cut-off frequency of 15 kHz. How much attenuation would we expect this filter to provide at 150 kHz? How about 15 MHz?*

We know the resistance across the output of the filter, because measurements are made with an LISN. The differential-mode load impedance is the series combination of the two 50-Ω LISNs or 100 Ω. The problem does not state the source impedance, so we will approximate the source resistance as the nominal voltage divided by the nominal maximum current. This is 110 V divided by 5 A, or 22 Ω. The resulting pi-filter model is shown below.




The value of the input capacitor that matches the source resistance at cut-off is,

$$C_{in} = \frac{1}{2\pi f_C R_S} = \frac{1}{2\pi\left(15\times10^3 \text{ Hz}\right)\left(22\ \Omega\right)} = 0.48\ \mu\text{F}.$$

The value of the output capacitor that matches the load resistance at cut-off is,

$$C_{out} = \frac{1}{2\pi f_C R_L} = \frac{1}{2\pi\left(15\times10^3 \text{ Hz}\right)\left(100\ \Omega\right)} = 0.11\ \mu\text{F}.$$

The series combination of these capacitances is,

$$C_{series} = \frac{C_{in}C_{out}}{C_{in}+C_{out}} = \frac{(0.48)(0.11)}{(0.48+0.11)} = 0.09\ \mu\text{F}$$

The total series inductance is therefore,

$$L = \frac{1}{\left(2\pi f_C\right)^2 C_{series}} = \frac{1}{\left(2\pi\times15\times10^3 \text{ rad/s}\right)^2\left(9\times10^{-8}\text{ F}\right)} = 1.25 \text{ mH}.$$

Since the filter must be balanced, this inductance can be achieved with two inductors that each have a nominal value of 625 μH, or a common-mode choke with a leakage inductance of 1.25 mH.

The insertion loss increases at a rate of 60 dB/decade above the cutoff frequency. So, at 150 kHz, we nominally expect to see 60 dB of attenuation. However, parasitics in the components and the layout tend to become more important as the nominal insertion loss becomes greater. Without modeling the parasitics, we cannot reliably predict the insertion loss at frequencies much higher than 150 kHz.

## Filtering Unbalanced Three-Conductor Lines

AC power inputs and isolated DC power inputs are typically balanced. However, many devices have unbalanced power inputs. Consider the buck converter input illustrated in Figure 9.31. Note that the impedance-to-ground of the two power phases is not the same. In this example, VBATT- is shorted to the frame ground making the power input perfectly unbalanced. As indicated by the model in Figure 9.31(b), all of the switching noise appears across $V_{m1}$ while $V_{m2}$ is zero.

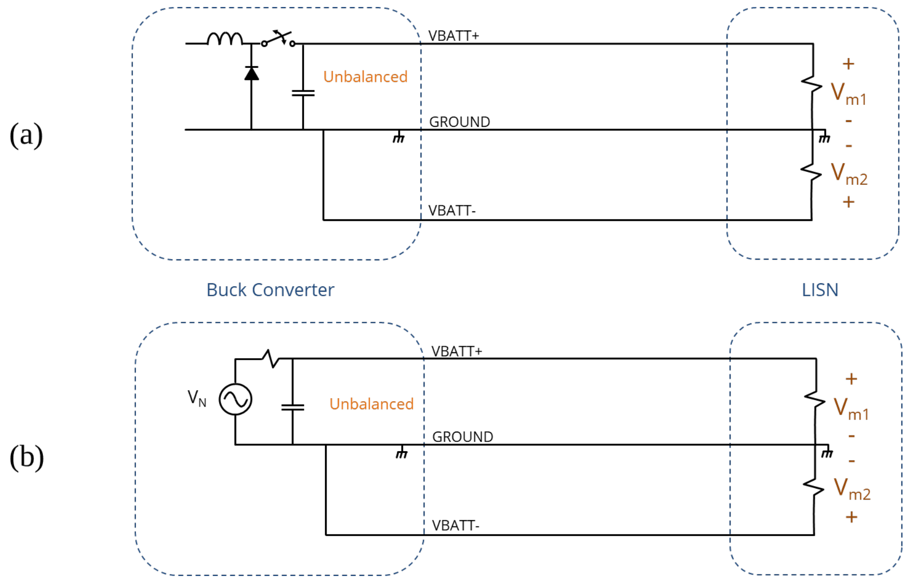



**Figure 9.31.** Model of conducted emissions generated by a buck converter.

Applying the same modal decomposition that we applied in the previous example, the differential-mode voltage measured by the LISNs is,

$$V_{\text{differential-mode}} = V_{m1} - V_{m2} = V_{m1} + 0 = V_{m1}. \tag{9.21}$$

The common-mode voltage measured by the LISNs is,

$$V_{\text{common-mode}} = \frac{V_{m1} + V_{m2}}{2} = \frac{V_{m1} - 0}{2} = \frac{V_{m1}}{2}. \tag{9.22}$$

It's tempting to conclude that both common-mode and differential-mode filtering is required. However, in this case, the modal decomposition of the LISN voltages is misleading. In this example, there is only one source, and it is driving VBATT+ relative to the chassis ground. The VBATT+ side requires filtering, the VBATT- side does not.

A common-mode choke would force noise currents to return on VBATT- instead of the GROUND wire. This would create a negative $V_{m2}$ voltage and nominally reduce the common-mode component of the noise, but it would not necessarily reduce $V_{m1}$. In most cases, a common-mode choke is not appropriate for filtering unbalanced power inputs.

Figure 9.32 shows a filter more appropriate for unbalanced power inputs like the one in Figure 9.31. The power conductor being driven by the switching noise is filtered to the chassis ground. The other power conductor (shorted to the ground) does not require any filtering as long as the short is made near the power connector. If some noise is picked up between the shorting point and the connector, then filtering (e.g. a capacitor to chassis ground) may be necessary.

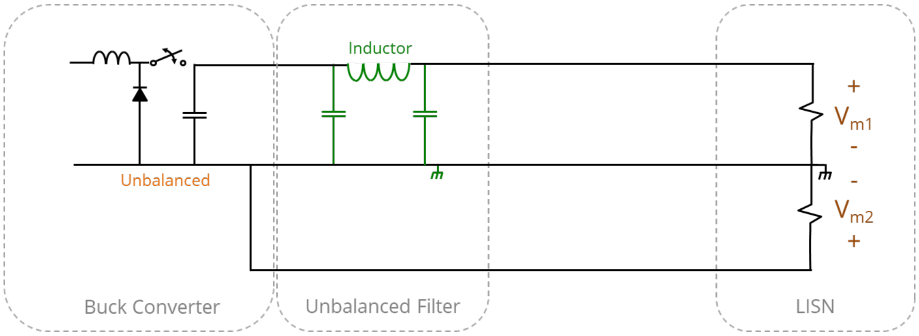



**Figure 9.32.** Filtering for an unbalanced power input.

In some devices with unbalanced power inputs, VBATT- is not shorted to the ground. For example, in devices that require DC isolation between power ground and the chassis, VBATT- might connect to the circuit board's power return plane without connecting directly to the chassis ground. In these cases, there are two approaches that generally work well for filtering the power input.

1. Make the power input balanced and use a balanced filter.
2. Make the power input very unbalanced and filter the side with the higher impedance to ground.

The second approach is illustrated in Figure 9.33. In this figure, VBATT+ connects to the switch and has no low-impedance path to the chassis. VBATT- connects to the current-return plane on the circuit board. This plane has a large surface area and a relatively low impedance to the chassis at high frequencies. To emphasize the imbalance, there is another plane on the board under the connector (usually labeled *chassis ground*). Several capacitors make a good high-frequency connection between the two planes. At the measurement frequencies, this power input is very unbalanced and can be filtered in the same manner as the non-isolated power input in Figure 9.22.

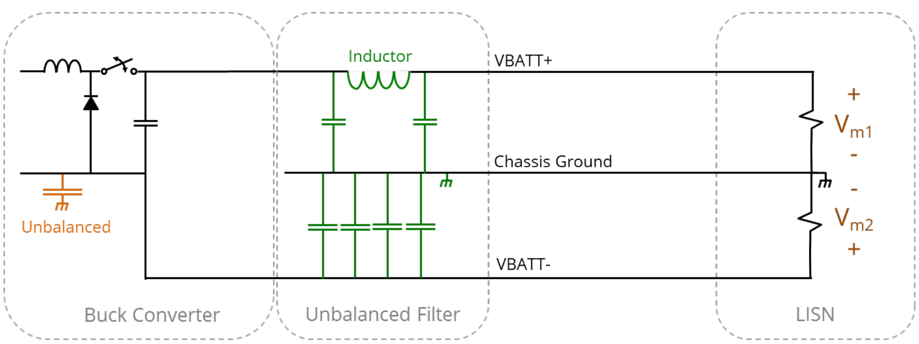


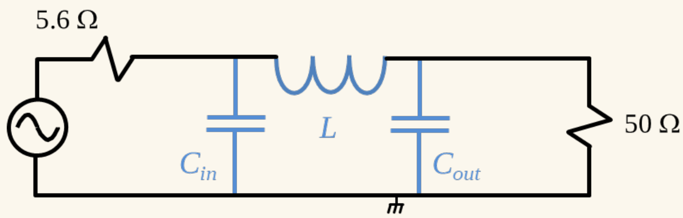

**Figure 9.33.** Filtering for an unbalanced power input with isolated chassis ground.

In most situations where power input is both isolated and unbalanced, it is best emphasize the imbalance by establishing a good high-frequency connection between the "circuit ground" and the chassis ground. As discussed in Chapter 3, a small voltage difference (e.g., 1 mV) between two relatively large conductors can result in unacceptable conducted and radiated emissions.

**Example 9-2: DC Power Line Filter**

*Design a pi-filter for a 28 V, 5 A DC power input located in a device with a metal enclosure. The power converter is unbalanced and intentionally isolated from the chassis ground. The filter should provide 60 dB of attenuation at the converter's 800-kHz switching frequency.*

Since we have an unbalanced power input in a device with a metal enclosure, we will attempt to provide a good high-frequency connection between the power return and the chassis using a chassis ground plane on the circuit board and multiple low-inductance capacitor connections. Then, we can filter the +28 V power using an unbalanced pi-filter as illustrated below.




Note that the load impedance is 50 Ω because we are filtering relative to the chassis ground. The problem does not state the source impedance, so we will approximate the source resistance as the nominal voltage divided by the maximum current. This is 28 V divided by 5 A, or 5.6 Ω. We want the cutoff frequency to be a factor of 10 lower than the switching frequency, or 80 kHz. The value of the input capacitor that matches the source resistance at cut-off is,

$$C_{in} = \frac{1}{2\pi f_C R_S} = \frac{1}{2\pi\left(80\times10^3\text{ Hz}\right)\left(5.6\ \Omega\right)} = 0.36\ \mu\text{F}.$$

The value of the output capacitor that matches the load resistance at cut-off is,

$$C_{out} = \frac{1}{2\pi f_C R_L} = \frac{1}{2\pi\left(80\times10^3\text{ Hz}\right)\left(50\ \Omega\right)} = 0.04\ \mu\text{F}.$$

The series combination of these capacitances is,

$$C_{series} = \frac{C_{in}C_{out}}{C_{in}+C_{out}} = \frac{(0.36)(0.04)}{(0.36+0.04)} = 0.036\ \mu\text{F}.$$

The total series inductance is therefore,

$$L = \frac{1}{\left(2\pi f_C\right)^2 C_{series}} = \frac{1}{\left(2\pi\times80\times10^3\text{ rad/s}\right)^2\left(3.6\times10^{-8}\text{ F}\right)} = 110\ \mu\text{H}$$

Note that most DC power inputs have their own shunt capacitance, so this should be subtracted from the input capacitance of our pi-filter.

## Filtering Balanced Two-Conductor Lines

Two conductor transmission lines can also carry a common-mode current. If the power cord doesn't have a ground wire, the common-mode current returns as a displacement current and/or a conduction current on nearby metal surfaces such as the metal table-top used in many standard EMC measurements.

Whether the power cord has two conductors or three conductors, the sources of common-mode and differential-mode noise are basically the same. From a filtering perspective, the most significant difference is that at least part of the common-mode current path includes a self- or mutual capacitance. This capacitance limits the common-mode current at low frequencies, so the filtering is mostly needed at high frequencies. Also, filters designed for resistive sources and loads may no longer be appropriate when the return path is capacitive.

Figure 9.34 illustrates appropriate filtering for products with balanced power inputs and a two-conductor power supply. In Figure 9.34(a), the product has a metal enclosure. Since the metal enclosure captures most of the field lines from an unbalanced source, common-mode filtering can return these noise currents to the enclosure with a chassis ground connection.

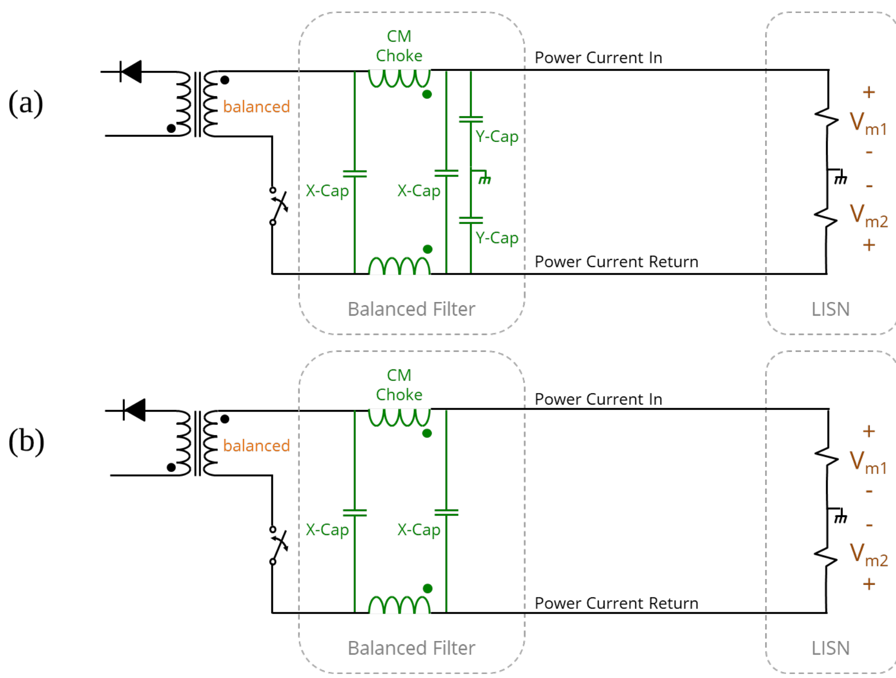



**Figure 9.34.** Filtering for a balanced power input with two-conductor power supply.

Figure 9.34(b) shows a filter for a product with no metal chassis or enclosure. In this case, there is no chassis ground and Y-capacitors cannot be used to reduce common-mode noise. This filter relies solely on the choke to provide common-mode filtering.

In products without a metal chassis, the common-mode current return path is capacitive at frequencies below the first cable resonance. Inductive chokes shift these resonant frequencies lower without necessarily providing any overall attenuation. In most situations where a common-mode choke is used without Y-capacitors to chassis ground, the ferrite material in the choke needs to provide high-frequency loss. Lossy chokes are widely available and provide attenuation similar to ferrite cores. Like ferrite cores, it is important to choose a choke that is effective at the frequencies where common-mode attenuation is required.

## Filtering Unbalanced Two-Conductor Lines

Most low-voltage DC-to-DC converters have unbalanced power inputs. If these converters are fed by balanced two-conductor power lines, half the differential switching noise voltage they create is converted to a common-mode voltage that drives the power line relative to the circuit board. In these situations, the first priority is to filter the differential-mode switching noise with an unbalanced filter. If the differential-mode noise is low enough by the time it reaches the power line, the mode-conversion due to the change in balance at the connector won't be important.

Small devices in plastic enclosures (with all wired connections on one side of the board) tend to be weak sources of common-node current because there is nothing of significant size that can be driven relative to the attached wires. In these boards, differential-mode filtering is usually all that is required. This can be accomplished with a simple pi-filter referenced to the circuit board ground as illustrated in Figure 9.35.

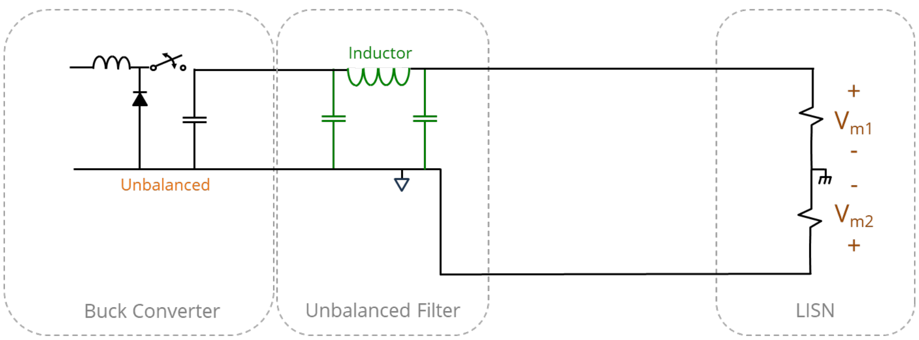



**Figure 9.35.** Filtering for an unbalanced power input with no chassis.

If the device with an unbalanced power input has a metal chassis, the chassis needs to be connected to the circuit board's PCB EMC ground near the connector as described in Chapter 8. In this case, the filter shown in Figure 9.35 is still appropriate.

Larger devices in plastic enclosures are capable of generating significant common-mode voltages directly. An example of this would be a circuit board with a very large heatsink as illustrated in Figure 9.36. In this situation, the lines of electric flux that are not captured by the board represent a self-capacitance. This is effectively a capacitance to the LISN's ground reference.

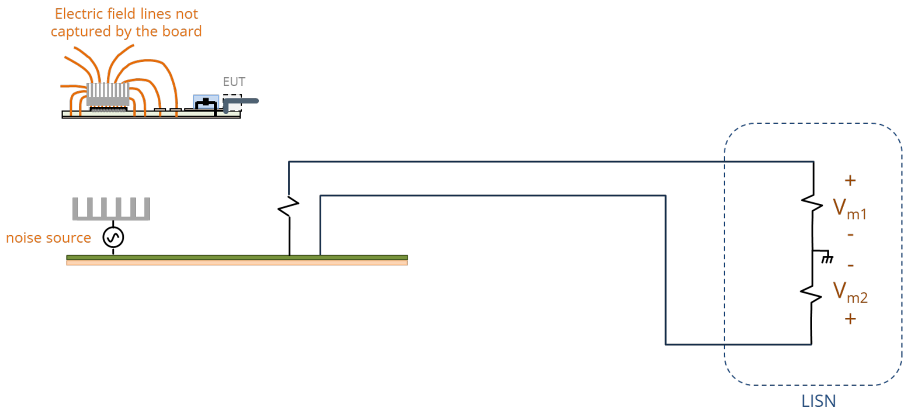



**Figure 9.36.** A common-mode source in a device with no chassis.

In this situation, the source drives the entire board relative to the LISN ground. Filtering the powerline wires back to the board doesn't help, because it is the board's potential relative to ground that drives the common-mode current. Electrical balance is not a factor, because the common-mode current is not due to mode conversion. It is the direct result of a common-mode voltage source.

Without a chassis to provide an alternative path for the noise current, there is really only one good filtering option. As illustrated in Figure 9.37, we can use Y-capacitors to give both power wires the same low-impedance to the board ground. This alone doesn't reduce the common-mode current, but it makes the power input look balanced at the noise frequencies. With a balanced power input, we can use a lossy common-mode choke to reduce the amount of common-current generated by the common-mode voltage source.

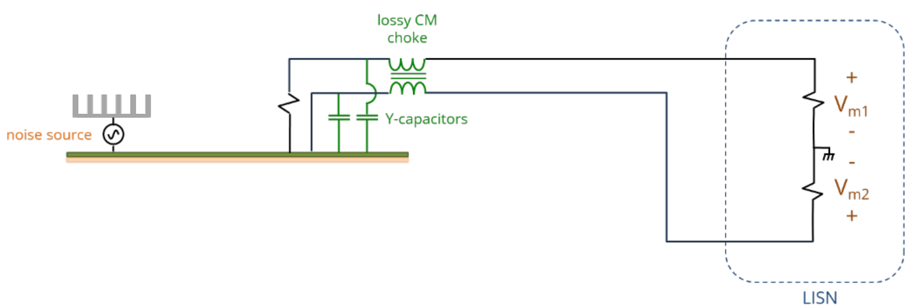


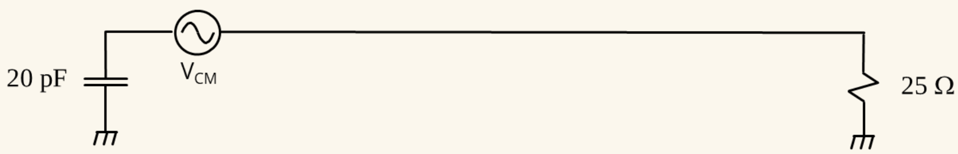

**Figure 9.37.** Filtering a common-mode source in a device with no chassis.

Note that it is very important that no Y-capacitors be placed on the LISN side of the choke. Y-capacitors in this position would allow the common-mode current to flow around the choke. Also, it is important that the choke provides a high-frequency resistance (not just an inductance) to the common-mode current. Part of the common-mode current path is already capacitive. Putting a low-loss inductance in this path could create an unwelcome resonance or simply shift an existing resonance to a new frequency.

**Example 9-3: DC Power Line Filter**

*A heatsink on a switching transistor has a self-capacitance of 20 pF and is driven by a signal with a harmonic at 10 MHz that has an amplitude of 40 mV. Using the model in Figure 9.36, calculate the voltage measured by the LISN. Then, applying the filter in Figure 9.37, estimate the attenuation that would be obtained with a choke that provides 100 Ω of common-mode resistance at 10 MHz.*

The equivalent circuit for the configuration described above looks like this.




The 25-Ω load represents the common-mode impedance of the LISNs (two 50-Ω terminations in parallel). The common-mode current is,

$$\left|I_{CM}\right| = \left|\frac{V_{CM}}{R - j/\omega C}\right| = \left|\frac{40\times10^{-3}\ \mathrm{V}}{25\ \Omega - j/\left(2\pi\times10^{7}\ \mathrm{rad/s}\right)\left(20\times10^{-12}\ \mathrm{F}\right)}\right| = 50.2\ \mu\mathrm{A}.$$

The voltage measured by each LISN is therefore $\left|V_{meas}\right| = \left(25\ \Omega\right)\left(50\ \mu\mathrm{A}\right) = 1.25\ \mathrm{mV}$ or 62 dB(μV). Adding a common-mode choke with 100 Ω of series resistance would make the new common-mode current,

$$\left|I_{CM\,\text{with choke}}\right| = \left|\frac{V_{CM}}{R - j/\omega C}\right| = \left|\frac{40\times10^{-3}\ \mathrm{V}}{125\ \Omega - j/\left(2\pi\times10^{7}\ \mathrm{rad/s}\right)\left(20\times10^{-12}\ \mathrm{F}\right)}\right| = 49.7\ \mu\mathrm{A}.$$

In other words, the current (and the measured voltage) are reduced by,

$$20\log\frac{50.2}{49.7} = 0.087\ \mathrm{dB}.$$

If the choke had an inductance of 25 μH or less, the current would increase because the positive reactance of the choke would cancel some of the negative reactance of the capacitance. An inductance greater than 25 μH nominally reduces the common-mode current, but at 10 MHz, inductors with a nominal value this large are likely to exhibit parasitics that limit their effectiveness.

In this case, it's generally better to address the problem at the source rather than attempting to solve the problem with a filter. Options would include connecting the heatsink to the board, using a larger board, or getting the heatsink closer to the board.

## Transient Protection

In the same way that it is sometimes necessary to filter out noise at unwanted frequencies, it can be necessary to filter noise in the time domain at unwanted amplitudes. Momentary spikes (*transients*) in a received signal voltage can corrupt data, trigger a system shutdown, or even damage components. The goal of transient protection devices and circuits is to mitigate the effects of these unwelcome transients.

Voltage transients come from a wide variety of sources including electrostatic discharges, arcing in high-voltage equipment, crosstalk from switching circuits, lightning, and intentional electromagnetic interference (IEMI) sources. The amplitude, shape and energy in a transient voltage also varies greatly depending on the source. Figure 9.38 shows four voltage transient waveforms specified in four EMC test standards.

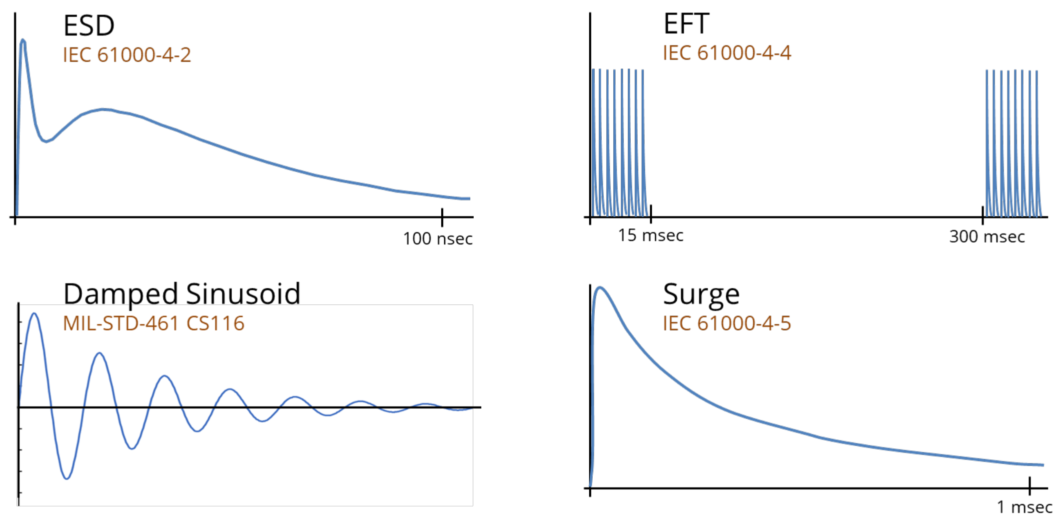



**Figure 9.38.** Voltage transients in four EMC test standards.

The IEC 61000-4-2 waveform is designed to emulate voltage transients that might typically be induced in a circuit due to a direct electrostatic discharge. These transients are characterized by very fast transitions (thousands of volts in about 1 nanosecond). On the other hand, the total energy in the transient (represented by the area under the curve) is relatively small (on the order of millijoules).

The IEC 61000-4-4 waveform emulates an electrical fast transient (EFT). It is meant to emulate the transients that might be coupled to a circuit sharing a wire harness with another circuit that switches an inductive load. These transients tend to be a little slower than ESD transients (hundreds of volts per nanosecond), but they arrive in bursts and are capable of delivering more total energy than a coupled ESD transient.

The damped sinusoidal waveform of MIL-STD-461 CS116 emulates ringing transient currents that can be induced in cables. It is a transient test, but the frequency of the ringing is varied.

The surge transient in IEC 61000-4-5 is designed to emulate relative high-energy sources such as coupling to long wires from a nearby lightning strike. The transition times are about a thousand times slower than an ESD transient, but the energy in the transient is a thousand times higher.

## Transient Protection Devices

Ideally, voltage transient protection components would have no impact on the operation of a circuit until the moment they were needed. At that point, they would instantly spring into action to limit the unwanted voltage and absorb or reflect the unwanted transient energy. However, transients come in many shapes and energy levels, and circuits have widely varying operational requirements. So, it's not surprising that a variety of transient protection devices are available with different strengths and weaknesses.

Key parameters to consider when choosing a transient protection component include:

- *Threshold Voltage* – The voltage at which the component "turns on." This should be higher than any normal signal voltage, but below any level that would damage or unacceptably disrupt the circuit.
- *Response Time* – The delay between the time the voltage exceeds the threshold and the time the device acts to limit it.
- *Terminal Capacitance* – The capacitance between the terminals of the device can affect the circuit. High-speed or high-frequency circuits require this capacitance to be small.
- *Energy Absorption* – The maximum energy in a transient that the component can handle without being damaged. Depending on the type of device, this might be specified as an instantaneous *peak power* or *peak current*.

Other parameters that might be relevant include the dimensions, operating temperature range, and leakage current.

Transient protection components can generally be grouped into four categories: Diodes, metal oxide varistors, thyristors and gas discharge tubes. General properties of these devices are summarized in Figure 9.39.

Diodes optimized for transient protection are commonly referred to as transient voltage suppression (TVS) diodes. They are basically Zener diodes with a reverse breakdown voltage that acts as the threshold voltage for the transient protection. In operation, they are connected across a power or signal line with a DC voltage. The power or signal voltage reverse-biases the diode. If a transient voltage exceeds the threshold voltage, the diode conducts enough current to clamp the voltage at the threshold. If a transient of the opposite polarity tries to pull the signal voltage below zero, the diode becomes forward biased, and the voltage is clamped at (or near) zero volts.

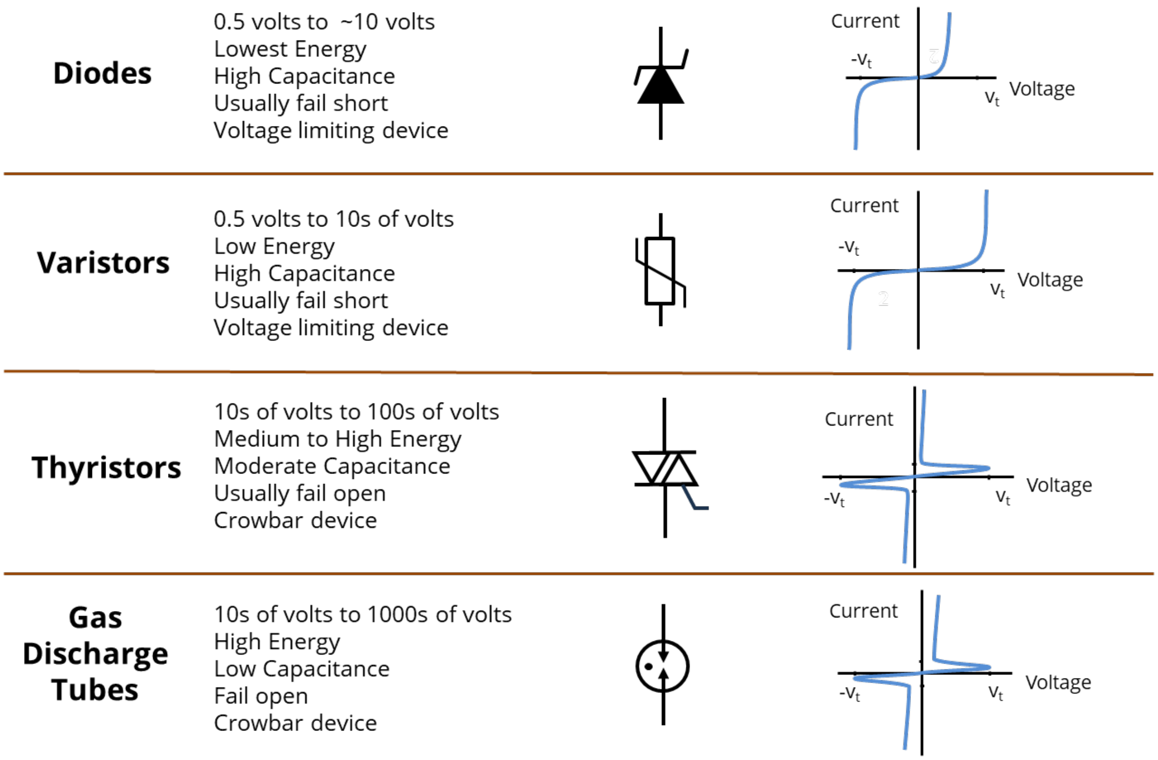



**Figure 9.39.** Transient protection devices.

Zener diodes are inherently unipolar devices that hold the voltage between zero and a threshold voltage with a given polarity. In applications where the circuit to be protected can take on both positive and negative values, two diodes can be connected in series (cathode-to-cathode). Together, they will block current of both polarities unless the threshold voltage is exceeded. Typically, both diodes are sealed in the same package and the component is referred to as a bipolar TVS diode.

TVS diodes generally have the fastest response times of any transient protection component. There is generally a trade-off between response time, capacitance, and energy absorption. Diodes with higher junction areas can typically handle more energy but have more capacitance and can be slightly slower.

Metal oxide varistors (MOVs) are made from a mixture of metal oxides compressed between two electrodes. Like diodes, they are voltage limiting devices that begin conducting current when the voltage across the electrodes exceeds a threshold. Unlike diodes, they are inherently bipolar. There are many applications where either an MOV or a bipolar TVS diode could be applied. However, generally, MOVs are better-suited for applications with high threshold voltages, such as line voltage transient protection.

Both MOVs and diodes are voltage-limiting devices that clamp the voltage at the threshold while the current rises. Thyristors and gas discharge tubes are *crowbar devices* that turn on like a switch (or dropping a crowbar across the terminals) when their threshold voltage is exceeded. Instead of clamping the voltage, the voltage drops to near zero when the current starts to flow. The power dissipated as heat in the device is the product of the voltage and the current. Since at least one of these quantities is always near zero in crowbar devices, they are able to handle transients with much higher energy.

Thyristors are semiconductor devices with an anode, a cathode, and a gate. There is normally a high resistance between the anode and the cathode, but that resistance drops to near zero when a sufficient voltage is applied to the gate. With a voltage divider the threshold voltage can be set to virtually any value higher than the gate trigger voltage.

Gas discharge tubes (GDTs) have the simplest construction of any transient protection device. They consist of two metal electrodes with a well-defined gap between them. The electrodes are enclosed in a tube of gas. When the voltage between the electrodes exceeds a threshold, the gas breaks down forming an arc. The arc conducts current causing the voltage across the terminals to drop. As long as the current continues to flow, the arc is maintained, and the terminal voltage remains low. Compared to other forms of transient protection, gas discharge tubes have slow response times. They can take microseconds to turn on. On the other hand, they have very low capacitance and can handle very high-energy transients.

In systems that are routinely exposed to high-energy transients, it is common to employ both primary and secondary transient protection. The primary protection absorbs or reflects the bulk of the transient energy. The secondary protection deals with the transient energy that gets past the primary protection while it is turning on. For example, the transient protection on a telecommunications cable at a building entrance might be configured as shown in Figure 9.40. In this circuit, the metal oxide varistor on the output side clamps the transient voltage to a manageable level until the gas discharge tube on the input fires and dissipates the transient energy.

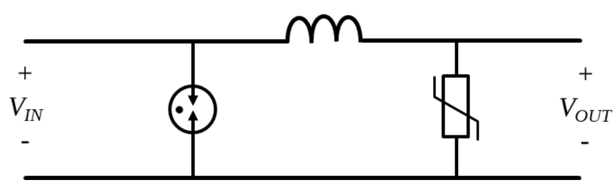



**Figure 9.40.** Primary and secondary transient protection.

Many commercial (and some residential) buildings have primary transient protection at the power service entrance. Nevertheless, voltage spikes as high as a few hundred volts are not uncommon in a building’s power distribution network. For this reason, many electronic products that are powered from wall outlets have secondary transient protection on their power inputs.