Input EMI Filter Design for Offline Phase-Dimmable LED Power Supplies

Reproduced from
2012 Texas Instruments Power Supply Design Seminar
SEM2000, Topic 1
TI Literature Number: SLUP298

© 2012, 2013 Texas Instruments Incorporated

Power Seminar topics and online power-training modules are available at:
ti.com/psds
Input EMI Filter Design for Offline Phase-Dimmable LED Power Supplies

James Patterson and Montu Doshi

ABSTRACT

Power supplies developed for LED-based solid-state lamps must incorporate input EMI filters to comply with government regulations and agency standards. The insertion of a passive EMI filter is known to interfere with external TRIAC-based phase dimmers and cause visible light flicker when dimming. In this paper, we will investigate different EMI filter configurations suitable for LED drivers, analyze differential- and common-mode noise-attenuation techniques, and outline a simple SPICE-based EMI filter design procedure. We will also explain how the EMI filter and external dimmer interact and propose design methods to ensure reliable system operation.

I. INTRODUCTION

Input EMI filters are used in conjunction with the power stage to reduce the noise generated by high-frequency switching operation in order to satisfy regulatory limits as defined in the U.S. Federal Communications Commission (FCC) Title 47 Code of Federal Regulations (CFR) Section 15 (USA), Europe’s Comité International Spécial des Perturbations Radioélectriques (CISPR) 15 (Europe) and other equivalent regional standards.

Although numerous EMI filter topologies are described in literature for conventional AC-DC power supplies [1], their use in integrated LED lamps and downlight drivers is limited due to stringent enclosure-size and solution-cost requirements. Further, the use of a damping network for input EMI filters in AC-DC power-factor-corrected (PFC) converters is not considered critical given the slow regulation bandwidth of the voltage feedback loop. Any disturbance caused by filter resonance does not impact the output regulation of the PFC converter as it is damped by the feedback loop.

However, inserting an undamped EMI filter into a dimmable LED lamp can create visible light flicker and poor light-output quality [2]. This degradation in performance is caused by the interaction between the triode for alternating current (TRIAC)-based dimmer circuit and the EMI filter resonance. As a result, a different approach to filter design is necessary to meet the performance requirements of phase-dimmable LED drivers.

In this paper, we highlight some of the major differences between conventional EMI filter topologies and phase dimmer-compatible EMI filter topologies. The choice of input filter configuration is based on product parameters such as maximum power, solution size, cost, isolation requirements, enclosure specification and type of input power line connection. In addition, the interaction between the EMI filter and the external dimmer has to be characterized and is a key parameter within the design. In this paper, we will provide a detailed analysis of these factors, a step-by-step EMI filter design guideline using SPICE simulation software, and experimental validation.

II. OFFLINE PHASE-DIMMABLE LED DRIVERS

Offline LED drivers are used across a broad range of applications and power levels. The most common application space is the consumer replacement bulb market. The output power can range from 4 W to 20 W in an LED bulb. The AC input voltage can vary from 100 Vac in Japan to 270 Vac in commercial buildings in the U.S.

There are also products that merge the bulb with the decorative trim of the recessed can light in another type of retrofit market, with a different form factor and extended power levels. Finally, there are new LED recessed can lights and other downlights that have power levels in excess of 40 W. This extensive design space can make the design quite challenging. In addition to the large design space, the requirement to interface with...
legacy TRIAC-based phase dimmers can seem impossible.

Designers have realized that the design space must be narrowed significantly in order to produce high-performing solutions. In general, universal input is not targeted, as it is not cost-effective to design phase-dimmable solutions that work across such a large range. Two typical voltage ranges are 90 Vac to 135 Vac (which covers Japan and North America) and 185 Vac to 256 Vac (which covers Europe and Asia, for the most part). Some high-voltage designs extend to 305 Vac to cover the 277-Vac U.S. commercial market; however, phase dimming is rarely used in these applications.

Designers have realized that the design space must be narrowed significantly in order to produce high-performing solutions. In general, universal input is not targeted, as it is not cost-effective to design phase-dimmable solutions that work across such a large range. Two typical voltage ranges are 90 Vac to 135 Vac (which covers Japan and North America) and 185 Vac to 256 Vac (which covers Europe and Asia, for the most part). Some high-voltage designs extend to 305 Vac to cover the 277-Vac U.S. commercial market; however, phase dimming is rarely used in these applications.

The replacement bulb market tends to use single-stage power-supply solutions (as shown at the top of Figure 1) due to their cost and space-effectiveness. These solutions are two-wire systems where the input does not have a chassis ground. The “bulb + trim” market is a mixture of single- and two-stage power supplies depending on the required performance and size constraints. This is also a two-wire system.

The downlight market leans toward two-stage designs like those shown at the bottom of Figure 1, mostly because the power supplies are reused for multiple load configurations. The performance of a single-stage design with varying load is usually not acceptable. The downlight can be either a three-wire system with a chassis ground connection or a two-wire system, depending on the construction of the power supply.

While all of these applications and configurations have similar concerns around filter design, this paper will focus on the two-wire input, which is more unique to offline LED drivers. In addition, we will focus on designing filters for realistic offline phase-dimmable applications where cost, size and performance are critical. You can extend these same design principles to three-wire systems and power supplies in general.

![Typical Bulb Supply](image1)

**Figure 1 – Typical offline LED driver configurations.**

The replacement bulb market tends to use single-stage power-supply solutions (as shown at the top of Figure 1) due to their cost and space-effectiveness. These solutions are two-wire systems where the input does not have a chassis ground. The “bulb + trim” market is a mixture of single- and two-stage power supplies depending on the required performance and size constraints. This is also a two-wire system.

The downlight market leans toward two-stage designs like those shown at the bottom of Figure 1, mostly because the power supplies are reused for multiple load configurations. The performance of a single-stage design with varying load is usually not acceptable. The downlight can be either a three-wire system with a chassis ground connection or a two-wire system, depending on the construction of the power supply.

While all of these applications and configurations have similar concerns around filter design, this paper will focus on the two-wire input, which is more unique to offline LED drivers. In addition, we will focus on designing filters for realistic offline phase-dimmable applications where cost, size and performance are critical. You can extend these same design principles to three-wire systems and power supplies in general.

![Typical Downlight Supply](image2)

**Figure 2 – CRM PFC flyback with buck secondary.**

![Primary Switch Current](image3)

**Figure 3 – CRM PFC flyback waveforms.**
III. ELECTROMAGNETIC NOISE IN OFFLINE LED DRIVERS

A constant current or power source is used to drive high-brightness LEDs because of their nonlinear voltage to current (V-I) device characteristics. An LED driver is required to properly convert the energy at the available input source to a suitable output current or power source to drive an LED array. In offline LED lighting applications, a PFC AC-DC rectifier circuit is used as a driver to convert the available input AC mains voltage into a regulated DC output suitable for driving LEDs. You can design the output of the PFC stage to directly drive the LEDs or as an input to a secondary stage that drives the LEDs.

The decision of whether to use a single- or two-stage topology will depend on a cost vs. performance trade-off, as described previously. Regardless, the input filter will see the same PFC input stage; therefore, the filter design is agnostic to the presence of the secondary stage, if it exists. In this paper, we focus on a critical conduction mode (CRM) PFC flyback AC-DC converter, shown in Figure 2, because of its popularity as a cost-effective solution for low to medium power levels [3] [4].

The operational waveforms of the CRM flyback converter are illustrated in Figure 3. When the transistor, QSW, is turned on, the transformer current rises to a peak threshold. The transistor, QSW, is then turned off while the output diode, DSW, turns on; the transformer current falls until it reaches zero. At this point, QSW is turned on and the cycle continues as in any style of hysteretic control. To achieve high power factor, a sinusoidal reference voltage is provided to the control loop to create a sinusoidal-varying average input current.

The inherent high-frequency switching operation results in a large rate of change of voltages (dV/dt) and currents (di/dt) within the circuit and generates significant electromagnetic (EM) noise. This EM noise can propagate through surrounding media and cause undesirable interference with other electronic devices. Incorporating EMI filters and other noise-mitigation techniques is necessary to achieve electromagnetic compatibility (EMC) between various electronic devices.

The interference mechanisms can be broadly classified into conducted or radiated phenomena, as shown in Figure 2, based on the noise propagation medium and the significant coupling paths [5]. As the coupling efficiency is dependent on the noise frequency, you can surmise that conducted EMI is caused by low-frequency components (< 30 MHz), while radiated EMI is generated by high-frequency components (> 30 MHz).

The EMI analysis can be simplified by treating conducted and radiated noise sources as distinct and separate over a specified range of frequency spectra. Different circuit schemes and techniques are required to attenuate the undesired frequency components and achieve system-level compliance.

The radiated EMI is a system-level phenomenon and is dependent on the entire LED luminaire, including the LED driver, heat sink, enclosure and optical elements. The radiated EMI signature is greatly influenced by parasitic elements that act as RF antennas and amplify the noise generated by the LED driver. Therefore, investigating radiated noise requires a careful analysis of parasitic components, printed circuit board (PCB) layout and lamp enclosure design. Often, the techniques used to attenuate radiated EMI are based on selective attenuation of RF frequencies using tuned filters and reducing the RF noise amplitude generated by the switching operation. The dependence of radiated EMI on system assembly and operation makes it difficult to develop a systematic approach to the design and development of filters.

In contrast to radiated EMI, conducted EMI is strongly influenced by the LED driver operation. It is therefore possible to analyze the noise generated and transmitted by the LED driver through the input and output conductors and to attenuate it using a range of passive low-pass filters. Note that the passive filters used to attenuate conducted EMI have minimal impact on radiated EMI noise. In this paper, we will focus exclusively on conducted EMI filters and their impact on LED driver performance.
The conducted EMI limits for LED-based lighting products are specified in FCC CFR 47 Section 15 and CISPR 15. Although both specifications are harmonized for the frequency range of 150 kHz to 30 MHz, the CISPR limits alone extend to a lower frequency range and therefore require a different measurement setup.

Most agencies are content to validate only the 150-kHz to 30-MHz range, so this paper will focus on these limits. In addition, the standards also specify two different types of measurements corresponding to the quasi-peak limit and the average limit, as shown in Figure 5. These limits represent scans performed with two separate detectors within the spectrum analyzer to measure the DUT conducted noise.

In practice, it is often possible to verify EMI performance by measuring only the peak signature of the conducted noise source using a simple peak amplitude detector within the spectrum analyzer. As peak measurement is always greater than quasi-peak or average value, compliance to an EMI standard can then be directly inferred by comparing only the peak magnitude to specified limits.

The LED driver is considered EMI-compliant when the peak magnitude is below the specified average measurement limit. Or, if the peak magnitude is below the quasi-peak limit, only an additional scan using an average detector is necessary. Both quasi-peak and average scans are required when the measured peak magnitude is greater than both defined limits, resulting in additional costs and time. As a compromise, designers will typically try to design for a peak scan passing the quasi-peak limits so that they can avoid the necessity of a quasi-peak scan and provide an adequate margin for component variation in their systems.

**V. CHARACTERIZATION OF CONDUCTED NOISE SOURCES**

It is crucial to understand the origin and characteristics of conducted noise in order to correlate the measured EMI signature with circuit parameters and design effective EMI filters. In general, the conducted noise measured by LISN is considered to be the sum of common-mode (CM) noise and differential-mode (DM) noise, where
CM noise refers to the current flowing through the common earth ground of the test system and DM noise refers to the current flowing in and out of the input line and neutral wires [5]. The flow of CM and DM current through the test setup is graphically illustrated in Figure 4.

![Figure 6 – Model of CM and DM noise sources.](image)

Figure 6 shows an equivalent model of the characteristics of DM and CM noise generated by an isolated flyback LED driver. The DM current is generated because of the switching action of the circuit, as described in Section III. During the positive half-cycle, the switching current, $I_{SW}$, flows into the flyback circuit through the line terminal and returns through the neutral terminal. The high-frequency content of the switching current also flows through two 50-Ω sensing resistors (in series) inside the LISN. It is important to note that a reversal of current direction during the negative half-cycle has no impact on the voltage measured by LISN.

The generation of CM current is dependent on the parasitic capacitances present between the high dV/dt nodes and the earth ground. Therefore, to understand CM noise characteristics, it is essential to recognize the dominant parasitic capacitance present in a power supply and to identify the high-frequency current loop they create.

In an isolated flyback PFC circuit, two such dominant parasitic capacitances are present: one between the FET switch node and earth ground, and one between the primary and secondary winding of the transformer. As these parasitic capacitances are linked with the high dV/dt noise generated by the switching transition, they inject significant current into the earth ground plane. The current is sensed by the LISN as the loop is completed through the 50-Ω resistors connected between the ground plane and power-supply inputs.

A radio frequency (RF) addition/subtraction circuit and access to individual LISN resistor (associated with line and neutral inputs) are required to accurately measure the DM and CM components of the conducted noise [6]. However, based on the nature of the noise sources, it is common to correlate low-frequency measurements ($< 500$ kHz) with DM noise and high-frequency measurements ($> 500$ kHz) with CM noise, as shown in Figure 5. This assumption also simplifies circuit analysis and filter design, as described in the next section.
VI. TYPICAL FILTERING METHODS FOR DM AND CM NOISE

While there are many different types of DM filters, the most common passive input filter configuration for DM noise attenuation is a simple LC filter, as shown in Figure 7. One stage yields a second-order filter with 40-dB/decade rolloff while two cascaded stages yields 80-dB/decade rolloff, as shown in Figure 8. Power level and size constraints guide the choice between second- and fourth-order filters. DM filter design is fairly straightforward; we will provide a simple design procedure in the following sections of this paper.

CM filtering is another challenge entirely. Within the power supply industry, the most common input filtering technique uses a CM choke (Figure 9). Unfortunately, this is less useful for LED drivers, as we will show in the following sections of this paper.

Figure 9 shows another method of CM attenuation where the DM filter is split into two DM chokes (inductors), with one placed in each leg of the line. This method is useful in two-wire systems.

CM noise mitigation also consists of carefully controlling dV/dt rates that induce current flow through the parasitic capacitances to chassis ground (Figure 6). In a two-wire system, the ground plate of the test setup acts as the chassis ground, as shown in Figure 6.

Perhaps the biggest problem comes from the parasitic capacitance between the heat sink and the ground plate of the test setup. The heat sink can be large, and high dV/dt rates associated with the switching action of the converter can cause high currents through the capacitance to the ground plate. This is exacerbated by the capacitance across the transformer.

The best practice to control these CM noise sources is to carefully construct the transformer by minimizing the interwinding capacitance. In addition, most designers place a Y-rated capacitor from primary ground to secondary ground (as shown in Figure 10) to equalize the potential of the two isolated sides and control the flow of current. Carefully managing the edge rates of switching signals and minimizing parasitic components through PCB layout are also paramount.

Figure 7 – Second- and fourth-order LC filters.

Figure 8 – Attenuation of different LC filters.

Figure 9 – CM choke and split DM filter configurations.

Figure 10 – Y capacitor across transformer.
VII. SELECTING AN INPUT EMI FILTER FOR TWO-WIRE SYSTEMS

The method of selecting an input filter configuration for an LED driver is different than for a standard switching power supply. There are many possible configurations, and the best selection is based on a variety of specifications.

Perhaps the most important specification arises from the input configuration of the system. While most traditional power supplies have a three-wire input connection containing a chassis ground, LED drivers frequently are two-wire systems, as we mentioned before; this is especially true in the replacement bulb market.

With a standard power supply that has a three-wire input, a CM choke is usually the most effective way to attenuate CM noise. Traditionally the CM inductance, coupled with capacitors to chassis ground from each line, provides excellent CM attenuation. The DM attenuation is then provided by a large filter capacitor across the line and the leakage inductance in the CM choke. In these standard power supplies, this large capacitance is tolerable.

![Figure 11 – Second-order filter with CM choke.](attachment:image)

A CM choke and large capacitor, shown in Figure 11, can be used in two-wire systems as well. However, the CM choke provides less attenuation to the CM noise in a two-wire system. It is not possible to connect capacitors from each line to the chassis ground in a two-wire system, and is not nearly as advantageous. The CM noise flows through the parasitic capacitances between the ground plate below the measurement test setup and the driver, eventually completing the loop through the LISN.

The CM noise mitigation has to be provided by other means such as dV/dt control, careful transformer construction and nontraditional filter configurations. Since LED applications are frequently very size-constrained, the elimination of the CM choke is fortuitous.

So what are suitable input-filter configurations for LED applications? Unfortunately, using a standard second-order LC filter implementation is not usually adequate to suppress the CM noise in the system. Instead, Figure 12 shows a second-order LC filter with the DM inductors split between both legs of the AC line. In this case, the design adds one more magnetic component, but if it is the same value as the other inductor, the capacitor can be halved for the same corner frequency of the filter.

The reduction of capacitance is always helpful in offline LED drivers. The configuration in Figure 12 is suitable for lower-power applications, or applications that have reasonable size constraints. Frequently, however, the 40-dB/decade rolloff of a single LC stage is not steep enough to be practical.

As an example, a high-power bulb might need as much as 80-dB attenuation to pass applicable regulatory limits. Using a fairly large AC capacitor of 1 µF, the filter would need a very large DM inductance of 10 mH! Conversely, using a two-stage LC filter (fourth order) with 80-dB/dec rolloff, an equivalent filter could be implemented with two 1-mH inductors and two 100-nF capacitors. While there are two more components, the solution size will be smaller and the cost cheaper. In addition, it is doubtful that a 10-mH inductor with the appropriate current rating and size even exists.

Given this, Figure 13 shows a fourth-order LC filter commonly used in LED bulb applications. The two DM inductors are placed in opposite legs, which helps CM attenuation but is not nearly as effective as splitting one of the inductors between two legs in a single stage. Given a small-enough system with well-controlled edge rates, the configuration in Figure 13 can be very effective for two-wire inputs.
Figure 14 illustrates a fourth-order filter with the split inductance on the first stage. This is very effective for two-wire LED applications. It is obviously more costly with the third inductor, yet the solution size is usually still reasonable. This filter yields excellent DM attenuation and fairly good CM attenuation as well. In addition, this system will have the lowest overall capacitance for a fourth-order filter, which will become significant later in this discussion.

Figure 12 – Second-order filter with CM configuration.

Figure 13 – Fourth-order filter.

Figure 14 – Fourth-order filter with CM configuration.

VIII. CHALLENGES WHEN IMPLEMENTING FILTERS

There are many non-idealities in the realm of input filter design. While the basic theory is an excellent starting place for the design, there are a few challenges in implementing these filters, including line and neutral imbalance, near-field coupling, and self-resonant frequency of the magnetic components.

The line and neutral imbalance is caused by an impedance mismatch at the input of the filter. Balanced configurations as shown in Figure 12 and Figure 14 do not exhibit this type of mismatch, but unbalanced configurations like the filter shown in Figure 13 will have a deviation at higher frequencies, as shown in Figure 16. This makes the effective filter size larger than necessary to compensate for the worst of the two paths. To solve this problem, a capacitor, Cin, added to the input can balance the impedance mismatch (Figure 15). This capacitor does not come for free, however, and leads us to conclude that a balanced configuration is preferable.

Figure 15 – Balancing an unbalanced filter configuration.

Figure 16 – Conducted scan with line and neutral imbalance.
The second common challenge is dealing with near-field magnetic coupling. When any magnetic is used in an electronic system, the possibility of coupling unwanted noise exists. For input filters, there is a risk of the high-frequency switching noise bypassing one or more of the DM stages. The CM choke is one of the biggest culprits, as it is unshielded and fairly large in size. Unshielded DM inductors can be problematic as well, and you must take great care to orient the magnetic fields of unshielded devices such that substantial coupling does not occur.

Unfortunately, the size constraints in LED applications frequently prevent the necessary configuration of components. Figure 18 shows a conducted scan of an LED driver using a CM choke at the input as in Figure 17. The CM choke orientation was vertical and the large increase around 250 kHz was ultimately due to near-field coupling. When we replaced the choke with a horizontal version, the anomaly was completely removed, and the 80-dB/decade rolloff of the filter was well under regulatory limits.

Finally, the self-resonant frequency (SRF) of the filter inductors can cause circulating currents that appear in the EMI signature at the SRF, as shown in Figure 20 (800 kHz in this example). This problem is fairly simple to solve by using a fairly high-valued parallel resistance (Figure 19). This is usually sized by trial and error to reduce the SRF component to a low-enough value to pass regulatory limits, while maintaining as high a resistance as possible. In some cases it is not necessary to use the resistor at all.
IX. PHASE-DIMMING IMPLICATIONS

Phase dimming has been around for more than 50 years and millions of traditional phase dimmers are installed worldwide. The large majority of these devices are very simple TRIAC-based forward-phase dimmers, as shown in Figure 21. A simple RC circuit is used to trigger a diode for alternating current (DIAC), which immediately triggers the TRIAC to begin conduction. The RC is adjustable through a potentiometer that ultimately sets the amount of delay before the TRIAC fires, as shown in Figure 22. Therefore, as the RC time constant increases, the conduction interval (θ) and the average energy delivered to the load decreases.

![Figure 21 – Standard TRIAC-based forward-phase dimmer.](image1)

![Figure 22 – Typical forward-phase-cut waveform.](image2)

This circuit worked very well for traditional incandescent light sources acting as simple resistors. But as fluorescent lighting solutions emerged, it became obvious to the industry that the TRIAC dimmer was not a good solution when switching converters were present. Two major problems were evident: the fast dV/dt due to the rising edge of the conduction interval and the minimum holding current requirement of the TRIAC.

The industry created a reverse phase dimmer, which removed the fast dV/dt associated with the rising edge of the conduction interval in forward-phase dimmers. These dimmers were microprocessor-based and used semiconductor switches like MOSFETs and insulated gate bipolar transistors (IGBTs) instead of TRIACs so that the minimum holding current requirement was also eliminated.

Unfortunately, these dimmers were – and still are – much more expensive. Since the emergence of solid-state lighting (SSL) applications, the industry has decided that it is imperative that LED solutions be backward-compatible with TRIAC-based dimmers. So compatibility with TRIAC dimmers is now one of the biggest challenges for LED driver designers.
The steep rising edge of the TRIAC dimmer causes a major problem with the system as it has been designed so far. Although the steady-state response of the filter is adequate to comply with conducted EMI regulations, the operation of the system in the time domain is another story entirely.

Figure 23 shows a system-level block diagram of an LED driver, EMI input filter and TRIAC dimmer. The TRIAC firing acts as a very large step input to the EMI filter. Because the filter is undamped, the step function excites the natural resonances of the filter; subsequently the input current has large ringing, as shown in Figure 24. Unfortunately, this ringing can cause the input current to fall below the minimum holding current requirement of the TRIAC, which will cause the device to commutate off and misfire. This can lead to visible flickering of the light output.

The solution is fairly simple. The filter needs adequate damping to attenuate the resonances properly so that the input current looks more like Figure 25. There are several well-known styles of filter damping, including resistor-capacitor (RC) damping, inductor-resistor (LR) parallel damping and LR series damping, shown in Figures 26, 27 and 28 [4].
The use of either LR damping network is simply too large and costly for most LED applications. Therefore, RC damping is the method of choice. As a rule of thumb, the damping capacitor is typically around three times the value of the filter capacitor and the resistor is sized to the critical resistance. This is the biggest reason that you will need to minimize total capacitance.

These damping capacitors are effectively in parallel with the filter capacitors at the line frequency; therefore the total input capacitance grows quickly with an increase in filter capacitance. This has many adverse effects in an LED system, including power-factor degradation, violation of the TRIAC minimum hold requirement during the latter half of the line cycle, and greater power loss in the damper circuits.

The basic design choices for an EMI input filter in TRIAC-based phase-dimming applications are:

1. Maximize inductance and minimize capacitance.
2. Dampen each filter capacitor with an RC damping network.
3. Choose a balanced topology when possible.
4. Implement PCB circuit and component orientation carefully.

**X. INPUT FILTER DESIGN PROCEDURE**

SPICE-based simulation tools can be very helpful when designing DM input filters. The following DM filter design will use mathematical calculations and a set of SPICE simulations. The procedure is as follows:

1. Estimate the magnitude of the DM noise source.
2. Determine the necessary attenuation to pass applicable regulatory limits.
3. Solve for the appropriate filter components to provide the calculated attenuation.
4. Solve for the necessary damping components to ensure phase-dimming compatibility.
5. Validate all results using SPICE-based simulations.

The first step is to estimate the magnitude of the primary DM noise source at the worst-case frequency in the system. Since CISPR 15 Class B limits begin at 150 kHz, it is commonly the third harmonic that is the worst-case noise source. Using the CRM PFC flyback topology, the peak switch current occurring at the peak of the AC line is known to be the main DM noise source. In SPICE, designers can create this triangular waveform with a current source, $I_1$, as shown in Figure 29. $I_1$ is connected to a dummy voltage source, $V_1$, to provide a complete circuit for the simulation.

![Figure 29 – Simulation schematic for estimating DM noise.](image)

![Figure 30 – Peak primary switch current, $I_1$.](image)

Figure 30 shows the programmed triangular current waveform represented by $I_1$. This model corresponds fairly well to the peak primary switch current of a 15-W 120-Vrms offline LED driver application. The peak current is 1 A and the switching frequency is 50 kHz at a duty cycle of 50 percent.
Given the known attenuation, in this case 69 dB, choosing a filter topology is now possible. As mentioned before, SSL applications are frequently size- and cost-constrained; therefore a fourth-order DM filter is usually necessary. Using a fourth-order LC filter, the corner frequencies can be calculated in Equation 3 as:

$$f_C = \frac{150 \text{ kHz}}{10 \text{ dB}} = \frac{150 \text{ kHz}}{20 \text{ dB}} = 18.8 \text{ kHz}$$

(3)

While typical filter design methodologies lead designers to choose corner frequencies that are far apart, this is not the case with phase-dimmable applications. Because we are trying to achieve minimal damping, the best practice is to make both corner frequencies identical so that the filter resonances are very close together. The stability concerns from traditional filter designs are not relevant, since the converter is a PFC operating at very low bandwidth.

Knowing the corner frequencies and assuming the largest inductance possible given size and cost constraints, 1 mH in this example, the filter capacitors can be calculated in Equation 4 as follows:

$$C = \frac{1}{(2 \pi \times f_c)^2 \times L} = \frac{1}{(2 \pi \times 20 \text{ kHz})^2 \times 1 \text{ mH}} = 72 \text{ nF}$$

(4)

*Figure 31 – SPICE schematic of fourth-order LC filter.*

<table>
<thead>
<tr>
<th>Harmonic Number</th>
<th>Frequency [kHz]</th>
<th>Magnitude [A]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>50</td>
<td>0.3774</td>
</tr>
<tr>
<td>2</td>
<td>100</td>
<td>0.1592</td>
</tr>
<tr>
<td>3</td>
<td>150</td>
<td>0.1085</td>
</tr>
<tr>
<td>4</td>
<td>200</td>
<td>0.0796</td>
</tr>
<tr>
<td>5</td>
<td>250</td>
<td>0.0642</td>
</tr>
<tr>
<td>6</td>
<td>300</td>
<td>0.0531</td>
</tr>
<tr>
<td>7</td>
<td>350</td>
<td>0.0457</td>
</tr>
<tr>
<td>8</td>
<td>400</td>
<td>0.0398</td>
</tr>
<tr>
<td>9</td>
<td>450</td>
<td>0.0355</td>
</tr>
<tr>
<td>10</td>
<td>500</td>
<td>0.0318</td>
</tr>
</tbody>
</table>

*Table 1 – Fourier output.*

To obtain the harmonic magnitudes of the relevant frequencies, perform a Fourier analysis using the SPICE .FOUR command in combination with the .TRAN transient command. Most SPICE-based simulators use the last period of the transient simulation output to perform the Fourier analysis. The results of the Fourier analysis are located in the output file.

Table 1 shows the results for the simulation setup in Figure 29. You can quickly see that the worst-case harmonic magnitude within the FCC (CISPR 15) limits is 108.5 mA at the third harmonic, 150 kHz.

The magnitude can be converted into dBuV using Equation 1:

$$V_{\text{LISN}(150 \text{ kHz})} \text{ (dBuV)} = 20 \times \log \left( \frac{0.1085 \text{ A} \times 50 \text{ } \Omega}{10^{-6}} \right) = 135 \text{ dBuV}$$

(1)

Then the noise magnitude is compared to the FCC limit at that frequency to determine the necessary attenuation (Equation 2):

$$F_A(150 \text{ kHz}) \text{ (dB)} = 135 \text{ dBuV} - 66 \text{ dBuV} = 69 \text{ dB}$$

(2)

*Figure 32 – Frequency response of fourth-order filter.*
Since the LED applications are frequently in very high-temperature environments, such as the interior of a bulb, the capacitors should be de-rated appropriately. For this design, we chose 100 nF for each LC filter stage.

With the preliminary filter design complete, SPICE can validate the performance of the design. The SPICE model shown in Figure 31 consists of a 1-V AC voltage source, V1, connected to the input of the fourth-order LC filter; the output is connected to I1 from the previous simulation. A SPICE .AC analysis provides the frequency response of the filter impedance as shown in Figure 32. As expected, the attenuation is more than adequate at 150 kHz.

As previously mentioned, CM attenuation of the DM filter can be greatly improved by adding another inductor, L1b, in the opposite leg of the first LC stage. This is also helpful because the capacitor can be halved if the total inductance of the first stage is doubled. Using this new CM-improving configuration (shown in Figure 33), and realistic capacitors and inductors that have equivalent series resistance (ESR) within the model, designers can resimulate to see the effective change in the frequency response of the filter.

Looking at Figure 34, the DM attenuation of the more realistic filter is still acceptable within the desired frequency range (150-500 kHz). However, there are several notable differences outside the frequencies of interest. At lower frequencies, the two large resonances have shifted due to the reduced capacitance and increased inductance of the first filter stage. The second deviation occurs at higher frequencies, where the non-ideal filter response occurs due to modeled ESR [7]. This deviation highlights the frequency range where CM noise is dominant and is evidence why other CM filter techniques are necessary.

While the overall DM attenuation is adequate for this design, the two filter resonances are problematic when dealing with phase dimmers. In order to eliminate ringing due to the step input of the TRIAC dimmer, both filter capacitors will be damped, as previously mentioned. Traditional damper design involves solving for the optimal damper capacitance and resistance to attenuate the Q factor of the LC filter output impedance. Because EMI filters in phase-dimmable LED drivers are subject to very large step inputs, the analysis does not follow the traditional method. And since the output impedance of the TRIAC-based phase dimmer varies greatly from dimmer to dimmer, a mathematical solution is not feasible.

Instead, a general rule of thumb is to size the damper capacitor to be three times the size of the filter capacitor. Mathematical calculations of the damping resistor also yield suboptimal results for the same basic reason. There is a large variability of the dimmer output impedance and the exact dV/ dt of the resulting step function created by the rising edge of the conduction angle. In the case of the resistor, we typically start with between 200 and 1,000 ohms and iterate within the experimental process.

After choosing the components, we can perform an AC analysis on the damped circuit (Figure 35). Figure 36 reveals that both resonances are significantly dampened as expected. Other than the damped resonance, the rest of the filter response remains unchanged as expected.
Finally, we simulated the time-domain response of the filter to see if the damping was adequate to prevent TRIAC misfires at the rising edge. To model the TRIAC behavior in phase-dimming applications, a simple step function serves as an input to the filter. Using the same SPICE filter models from Figures 33 and 35, we simply change the voltage source to a DC step function – in this example, from 0 V to the peak AC voltage of 177 V.

Looking at the results of the simulation in Figure 37, the input current waveform of the undamped filter shows large ringing; this violates the minimum hold current condition and would certainly cause the TRIAC to misfire in a real application. The damped response, on the other hand, shows adequate attenuation of the ringing to maintain the TRIAC’s minimum holding current requirement.

In practice, you may want to provide even more headroom to account for other non-idealities, and you may need more damping. This can be accomplished by increasing the damping capacitor of the first stage and resimulating until you meet the desired specification.

Figure 35 – RC damped fourth-order filter.

Figure 36 – Frequency response of damped filter.

Figure 37 – Time-domain step response.
XI. EXPERIMENTAL VALIDATION

To validate the SPICE-based design methodology, we built a 15-W 120-Vrms offline phase-dimmable LED driver, implementing the same fourth-order filter and performing both time- and frequency-domain measurements. Figure 38 shows the experimental schematic, which includes the fourth-order filter with a diode bridge between the two stages as it is frequently implemented. Figure 39 shows the main switch current waveforms of this particular LED driver stage, which are close to the simulated parameters of 1-A peak current and 50-kHz switching frequency at the peak AC voltage of 177 V.

Conversely, Figure 41 shows that using a damped filter configuration with R3, R4, C3 and C4 populated (as in Figure 38) provides ample damping to keep the TRIAC conducting immediately after the rising edge of the conduction interval. We tested one dimmer.

As you experiment with other dimmers, the effect of the damping network will change. The actual damping is heavily dependent on the output impedance of the dimmer and the dV/dt of the rising edge of the conduction angle. If certain dimmers cause more ringing than what was simulated, you can either increase the damper capacitor of the first stage (C3) or attempt to increase the damper resistance to change the time constant of the ring. This is not an exact science by any means and is really accomplished via trial and error.
Finally, we performed a set of conducted EMI scans to validate that the frequency response of the system is within the CISPR 15 Class B limits as designed. Figure 42 shows the result of a peak conducted scan; Figure 43 shows the result of an average conducted scan. The results are excellent. The peak scan actually passes the average limits, which eliminates the need for more costly scans as mentioned before.

It is worth emphasizing that this DM filter design has little effect on frequencies above 1 MHz (due to component parasitics) and is mostly designed to attenuate from 150 kHz and 500 kHz. The higher frequency attenuation of the scans in Figures 42 and 43 was accomplished by controlling the edge rates, PCB design and parasitic capacitances in the system.

If a more aggressive design were necessary due to cost and/or size constraints, you could remove the capacitor de-rating from the design procedure; however, if you are not using high-quality components, the absence of de-rating could cause a significant change in attenuation over the operating range.

**XII. Conclusion**

While input EMI filter design for phase-dimmable LED drivers can feel like black magic to novice designers, there are reliable methods to quickly design and validate these systems. SPICE-based simulation tools provide an excellent medium for this process, allowing designers to estimate the harmonic content of switching converters and simulate both time- and frequency-domain responses.

Though this design yields a complete EMI input filter, there is no substitute for experimental validation and iteration between simulation and experimentation. Offline phase-dimmable LED driver design requires a well-thought-out EMI filter design in order to balance the performance of the filter vs. the performance of the dimmer system.
XIII. REFERENCES


**TI Worldwide Technical Support**

**Internet**
TI Semiconductor Product Information Center
Home Page
support.ti.com

TI E2E™ Community Home Page
e2e.ti.com

**Product Information Centers**

<table>
<thead>
<tr>
<th>Region</th>
<th>Phone</th>
<th>Fax</th>
<th>Internet/Email</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Americas</strong></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Brazil</td>
<td>0800-891-2616</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Mexico</td>
<td>0800-670-7544</td>
<td>+1(972) 927-6377</td>
<td>support.ti.com/sc/pic/americas.htm</td>
</tr>
<tr>
<td><strong>Europe, Middle East, and Africa</strong></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>European Free Call</td>
<td>00800-ASK-TEXAS</td>
<td></td>
<td></td>
</tr>
<tr>
<td>(00800 275 83927)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>International</td>
<td>+49 (0) 8161 80 2121</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Russian Support</td>
<td>+7 (4) 95 98 10 701</td>
<td></td>
<td></td>
</tr>
<tr>
<td><strong>Asia</strong></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>International</td>
<td>+91-80-41381665</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Domestic</td>
<td>Toll-Free Number</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Australia</td>
<td>1-800-999-084</td>
<td></td>
<td></td>
</tr>
<tr>
<td>China</td>
<td>800-820-8682</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Hong Kong</td>
<td>800-96-5941</td>
<td></td>
<td></td>
</tr>
<tr>
<td>India</td>
<td>1-800-425-7888</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Indonesia</td>
<td>001-803-8861-1006</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Korea</td>
<td>080-551-2804</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Malaysia</td>
<td>1-800-80-3973</td>
<td></td>
<td></td>
</tr>
<tr>
<td>New Zealand</td>
<td>0800-446-934</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Philippines</td>
<td>1-800-765-7404</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Singapore</td>
<td>800-886-1028</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Taiwan</td>
<td>0800-006800</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Thailand</td>
<td>001-800-886-0010</td>
<td></td>
<td></td>
</tr>
<tr>
<td><strong>Japan</strong></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Domestic</td>
<td>0120-92-3326</td>
<td></td>
<td></td>
</tr>
<tr>
<td>International</td>
<td>+81-3-3344-5317</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Domestic</td>
<td>0120-81-0036</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Internet/Email</td>
<td>support.ti.com/sc/pic/japan.htm</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Domestic</td>
<td><a href="http://www.tij.co.jp/pic">www.tij.co.jp/pic</a></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

**Important Notice:** The products and services of Texas Instruments Incorporated and its subsidiaries described herein are sold subject to TI's standard terms and conditions of sale. Customers are advised to obtain the most current and complete information about TI products and services before placing orders. TI assumes no liability for applications assistance, customer’s applications or product designs, software performance, or infringement of patents. The publication of information regarding any other company’s products or services does not constitute TI’s approval, warranty or endorsement thereof.

© 2013 Texas Instruments Incorporated
Printed in U.S.A. by (Printer, City, State)
IMPORTANT NOTICE

Texas Instruments Incorporated and its subsidiaries (TI) reserve the right to make corrections, enhancements, improvements and other changes to its semiconductor products and services per JESD46, latest issue, and to discontinue any product or service per JESD48, latest issue. Buyers should obtain the latest relevant information before placing orders and should verify that such information is current and complete. All semiconductor products (also referred to herein as “components”) are sold subject to TI's terms and conditions of sale supplied at the time of order acknowledgment.

TI warrants performance of its components to the specifications applicable at the time of sale, in accordance with the warranty in TI's terms and conditions of sale of semiconductor products. Testing and other quality control techniques are used to the extent TI deems necessary to support this warranty. Except where mandated by applicable law, testing of all parameters of each component is not necessarily performed.

TI assumes no liability for applications assistance or the design of Buyers' products. Buyers are responsible for their products and applications using TI components. To minimize the risks associated with Buyers' products and applications, Buyers should provide adequate design and operating safeguards.

TI does not warrant or represent that any license, either express or implied, is granted under any patent right, copyright, mask work right, or other intellectual property right relating to any combination, machine, or process in which TI components or services are used. Information published by TI regarding third-party products or services does not constitute a license to use such products or services or a warranty or endorsement thereof. Use of such information may require a license from a third party under the patents or other intellectual property of the third party, or a license from TI under the patents or other intellectual property of TI.

Reproduction of significant portions of TI information in TI data books or data sheets is permissible only if reproduction is without alteration and is accompanied by all associated warranties, conditions, limitations, and notices. TI is not responsible or liable for such altered documentation. Information of third parties may be subject to additional restrictions.

Resale of TI components or services with statements different from or beyond the parameters stated by TI for that component or service voids all express and any implied warranties for the associated TI component or service and is an unfair and deceptive business practice. TI is not responsible or liable for any such statements.

Buyer acknowledges and agrees that it is solely responsible for compliance with all legal, regulatory and safety-related requirements concerning its products, and any use of TI components in its applications, notwithstanding any applications-related information or support that may be provided by TI. Buyer represents and agrees that it has all the necessary expertise to create and implement safeguards which anticipate dangerous consequences of failures, monitor failures and their consequences, lessen the likelihood of failures that might cause harm and take appropriate remedial actions. Buyer will fully indemnify TI and its representatives against any damages arising out of the use of any TI components in safety-critical applications.

In some cases, TI components may be promoted specifically to facilitate safety-related applications. With such components, TI’s goal is to help enable customers to design and create their own end-product solutions that meet applicable functional safety standards and requirements. Nonetheless, such components are subject to these terms.

No TI components are authorized for use in FDA Class III (or similar life-critical medical equipment) unless authorized officers of the parties have executed a special agreement specifically governing such use.

Only those TI components which TI has specifically designated as military grade or “enhanced plastic” are designed and intended for use in military/aerospace applications or environments. Buyer acknowledges and agrees that any military or aerospace use of TI components which have not been so designated is solely at the Buyer's risk, and that Buyer is solely responsible for compliance with all legal and regulatory requirements in connection with such use.

TI has specifically designated certain components as meeting ISO/TS16949 requirements, mainly for automotive use. In any case of use of non-designated products, TI will not be responsible for any failure to meet ISO/TS16949.

<table>
<thead>
<tr>
<th>Products</th>
<th>Applications</th>
</tr>
</thead>
<tbody>
<tr>
<td>Audio</td>
<td><a href="http://www.ti.com/audio">www.ti.com/audio</a></td>
</tr>
<tr>
<td>Amplifiers</td>
<td>amplifier.ti.com</td>
</tr>
<tr>
<td>Data Converters</td>
<td>dataconverter.ti.com</td>
</tr>
<tr>
<td>DLP® Products</td>
<td><a href="http://www.dlp.com">www.dlp.com</a></td>
</tr>
<tr>
<td>DSP</td>
<td>dsp.ti.com</td>
</tr>
<tr>
<td>Clocks and Timers</td>
<td><a href="http://www.ti.com/clocks">www.ti.com/clocks</a></td>
</tr>
<tr>
<td>Interface</td>
<td>interface.ti.com</td>
</tr>
<tr>
<td>Logic</td>
<td>logic.ti.com</td>
</tr>
<tr>
<td>Power Mgmt</td>
<td>power.ti.com</td>
</tr>
<tr>
<td>Microcontrollers</td>
<td>microcontroller.ti.com</td>
</tr>
<tr>
<td>RFID</td>
<td><a href="http://www.ti-rfid.com">www.ti-rfid.com</a></td>
</tr>
<tr>
<td>OMAP Applications Processors</td>
<td><a href="http://www.ti.com/omap">www.ti.com/omap</a></td>
</tr>
<tr>
<td>Wireless Connectivity</td>
<td><a href="http://www.ti.com/wirelessconnectivity">www.ti.com/wirelessconnectivity</a></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Products</th>
<th>Applications</th>
</tr>
</thead>
<tbody>
<tr>
<td>Audio</td>
<td>Automotive and Transportation <a href="http://www.ti.com/automotive">www.ti.com/automotive</a></td>
</tr>
<tr>
<td>Amplifiers</td>
<td>Communications and Telecom <a href="http://www.ti.com/communications">www.ti.com/communications</a></td>
</tr>
<tr>
<td>Data Converters</td>
<td>Computers and Peripherals <a href="http://www.ti.com/computers">www.ti.com/computers</a></td>
</tr>
<tr>
<td>DLP® Products</td>
<td>Consumer Electronics <a href="http://www.ti.com/consumer-apps">www.ti.com/consumer-apps</a></td>
</tr>
<tr>
<td>DSP</td>
<td>Energy and Lighting <a href="http://www.ti.com/energy">www.ti.com/energy</a></td>
</tr>
<tr>
<td>Clocks and Timers</td>
<td>Industrial <a href="http://www.ti.com/industrial">www.ti.com/industrial</a></td>
</tr>
<tr>
<td>Interface</td>
<td>Medical <a href="http://www.ti.com/medical">www.ti.com/medical</a></td>
</tr>
<tr>
<td>Logic</td>
<td>Security <a href="http://www.ti.com/security">www.ti.com/security</a></td>
</tr>
<tr>
<td>Power Mgmt</td>
<td>Space, Avionics and Defense <a href="http://www.ti.com/space-avionics-defense">www.ti.com/space-avionics-defense</a></td>
</tr>
<tr>
<td>Microcontrollers</td>
<td>Video and Imaging <a href="http://www.ti.com/video">www.ti.com/video</a></td>
</tr>
<tr>
<td>RFID</td>
<td>e2e.ti.com</td>
</tr>
</tbody>
</table>

Mailing Address: Texas Instruments, Post Office Box 655303, Dallas, Texas 75265
Copyright © 2013, Texas Instruments Incorporated