

Bias Supply Design for Isolated Gate Driver Using UCC25800-Q1 Open-Loop LLC Transformer Driver



ABSTRACT

For the electrical vehicle (EV) and hybrid electric vehicle (HEV), isolated gate drivers are widely used in the traction inverter, and on-board chargers. These isolated gate drivers require their isolated bias supplies to be low cost, high reliability, low EMI, and high CMTI (Common Mode Transient Immunity). UCC25800-Q1 open-loop LLC transformer driver is designed to meet these design challenges and provides a robust, cost effective, and low EMI solution for isolated gate driver bias. This application note introduces the LLC topology and the benefits of using LLC for isolated gate driver bias supplies. It also provides the design guidelines to simplify the design process.

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1 Introduction

UCC25800-Q1 is an open-loop LLC transformer driver. By open-loop control, fixed output to input voltage gain can be achieved through the transformer turns ratio. The open-loop control also provides a simple design and superior noise immunity. The LLC topology utilizes the transformer leakage inductance as its resonant component, allowing the converter to use a transformer with a larger leakage inductance but a much smaller primary-side to secondary-side parasitic capacitance (inter-winding capacitance).

Other topologies, such as flyback or push-pull, require minimum leakage inductance to improve the efficiency, reduce the voltage and current stress, as well as minimize the noise created by the converter. In turn, this type of transformer suffers from larger inter-winding capacitance. When they are used in the gate driver bias supply applications, the high dv/dt from the inverter power stage can be coupled through the transformer inter-winding capacitance to the low-voltage side. This creates a much more severe EMI noise issue. It also puts more challenges on the common-mode transient immunity (CMTI). The open-loop LLC transformer driver solves this issue and provides a low-noise, robust solution for the isolated gate driver bias supplies.

2 Open-Loop LLC Converter Operation Principle

Different from the typical PWM converters, LLC converters adjust the output voltage through varying its switching frequency. It is often called a PFM (Pulse Frequency Modulation) converter. As shown in [Figure 2-1](#), the LLC converter has three resonant elements, the resonant inductor (L_r), the magnetizing inductor (L_m), and the resonant capacitor (C_r). In isolated bias supply designs, the transformer leakage inductance, and the magnetizing inductor can be used as part of the resonant circuit. The only external resonant component is the resonant capacitor.

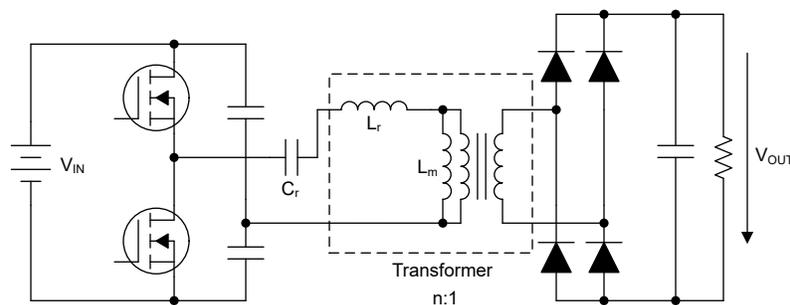


Figure 2-1. LLC Resonant Converter

The resonant frequency of the LLC converter is defined by the series resonance between the resonant inductor (L_r) and the resonant capacitor (C_r), as shown in [Equation 1](#)

$$f_0 = \frac{1}{2\pi\sqrt{L_r C_r}} \quad (1)$$

When its switching frequency is equal or below the resonant frequency, the operation waveforms of LLC converter can be found in [Figure 2-2](#).

When the switching frequency is equal to the resonant frequency, it can be observed that the transformer primary-side current is a sinusoidal shape. The secondary-side current is also a sinusoidal shape but with some phase shift to the primary side current. The phase shift is caused by the transformer magnetizing current. The output current is equal to the rectified transformer secondary-side current. In this operation mode, the resonant tank impedance is equal to zero, and the input and output voltages are connected virtually through the transformer. When the switching frequency moves away from the resonant frequency, the impedance of the resonant tank increases. As a result, the output voltage reduces. However, this only holds true when the switching frequency is higher than the resonant frequency, because the magnetizing inductor never participates the resonant and it doesn't influence the characteristic performance of the resonant circuit.

When the switching frequency is below the resonant frequency, it can be observed that the sinusoidal current becomes discontinuous. In the duration where the sinusoidal shape stops, transformer secondary-side current is equal to zero. During this period, the magnetizing inductor becomes part of the resonant circuit, but the resonant frequency is so low that the current shape appears linear. During this period, the magnetizing inductor stores

more energy and transfers it to the secondary side through the resonant capacitor in the following half switching cycle. Therefore, the LLC is able to achieve higher voltage gain.

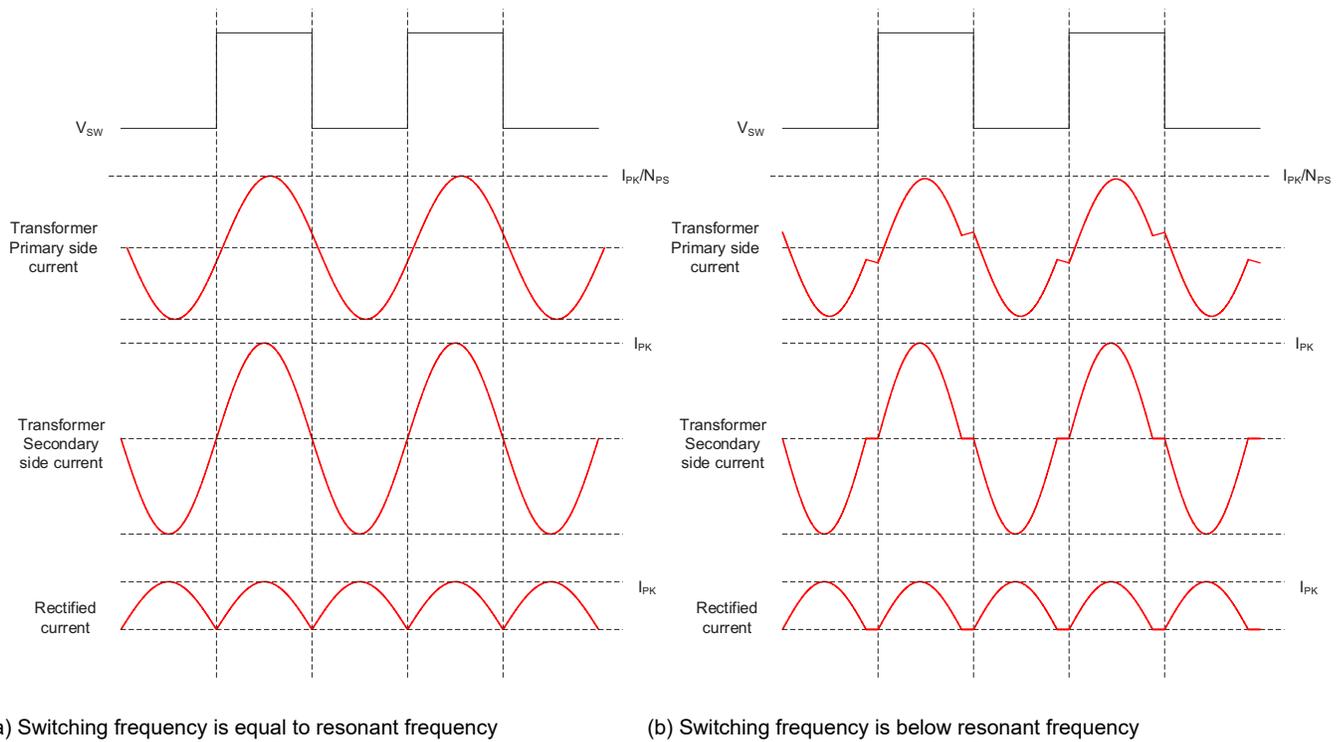


Figure 2-2. LLC Converter Operation Waveforms

The LLC converter voltage gain refers to the relationship between its output voltage and its input voltage. It is defined as Equation 2. In this equation, n is the transformer primary side to secondary side turns ratio and the $\frac{1}{2}$ comes from the half bridge configuration.

$$M = \frac{n \times V_{OUT}}{V_{IN}/2} \quad (2)$$

As shown in Figure 2-3, the voltage gain is affected by both switching frequency and load. In this set of curves, the switching frequency is normalized with the resonant frequency, which is defined in Equation 1, and Equation 3. The load can be normalized with the characteristic impedance and defined as Equation 4. Here, f_s is the switching frequency.

$$f_n = \frac{f_s}{f_0} \quad (3)$$

$$Q_e = \frac{\sqrt{L_r/C_r}}{n^2 \frac{8R_{load}}{\pi^2}} \quad (4)$$

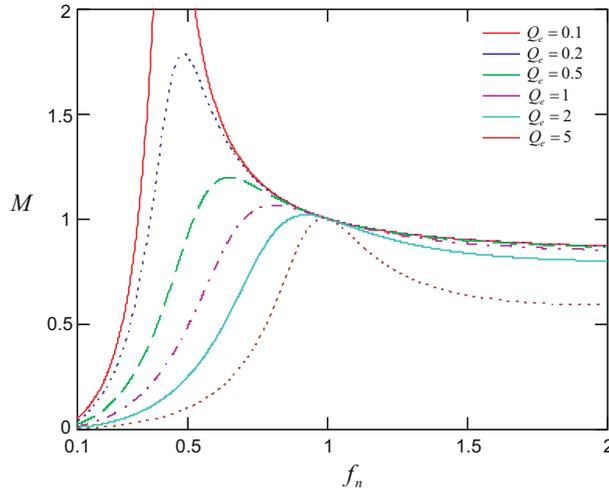


Figure 2-3. LLC Converter Voltage Gain

From this set of gain curves, at the resonant frequency, regardless of load conditions, the converter has a gain equal to 1, which means the relationship between the input and output voltage is only determined by the transformer turns ratio. This can be simply understood by the impedance of the resonant tank (L_r and C_r) is equal to zero at the resonant frequency. The input and output voltages are directly connected together virtually through the transformer.

If the LLC converter operates with a fixed switching frequency equal to the resonant frequency, the LLC converter is able to deliver a fixed voltage gain, with different load conditions. With a fixed input voltage, a fixed output voltage can be achieved.

3 Primary-side and Secondary-side Resonant Configurations

In Figure 2-1, the resonant capacitor is on the primary side. From its gain curves, it can be observed when the switching frequency is below the resonant frequency, the converter voltage gain rises. This occurs because the energy that is stored in the magnetizing inductor is transferred to the secondary side in each half-switching cycle. The lower the switching frequency, the more energy is stored in the magnetizing inductor. Therefore the voltage gain keeps going up.

When the resonant capacitor is moved on the secondary side, the gain curves are changed. One example of using secondary-side resonant is shown in Figure 3-1.

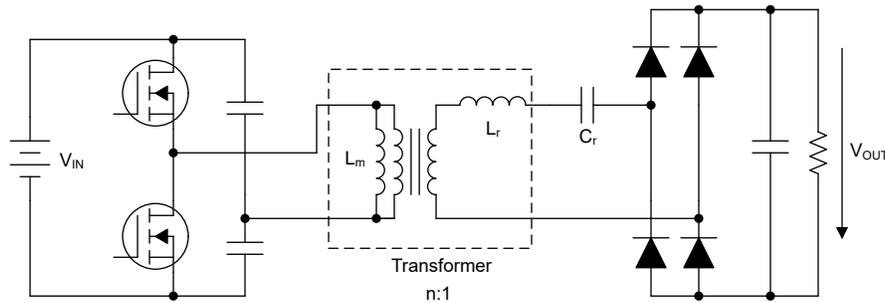
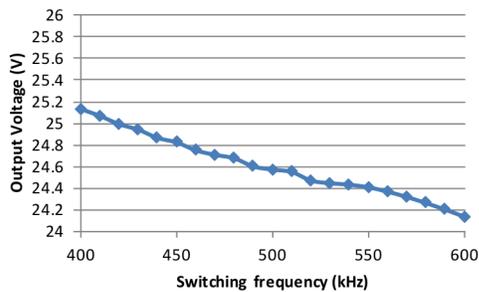
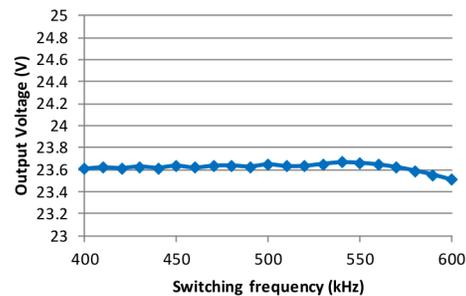


Figure 3-1. LLC Converter Using Secondary-side Resonant

Putting the resonant capacitor on the secondary side, when the switching frequency is below the resonant frequency, the energy stored in the magnetizing inductance can no longer be transferred to the secondary side. Instead, the energy is fed back to the input source. Due to this behavior, the voltage gain of the secondary-side resonant becomes flat when the switching frequency is below the resonant frequency, as shown in Figure 3-2.



(a) Primary-side Resonant



(b) Secondary-side Resonant

Figure 3-2. Output Voltage Comparison between Primary-side Resonant and Secondary-side Resonant Configurations with 550-kHz Resonant Frequency, 24-V Input Voltage, and 1:1 Transformer Turns Ratio

Based on these curves, we can see if the LLC converter is operating with a fixed switching frequency and that frequency is slightly below the resonant frequency, the voltage gain is fixed, regardless of the frequency or the load condition, which means with a fixed input voltage, we get a fixed output voltage. This property also helps to maintain constant output voltage with the tolerances on the resonant components, which is equivalent to the switching frequency variation.

4 Rectification Methods

Besides the transformer turns-ratio, the LLC converter voltage gain is also affected by the rectification structure. In this section, a few rectification methods are presented and designers can select the one according to their cost structure from the components. Because the primary-side circuit remains the same, the description only focuses on the secondary-side circuit.

4.1 One Resonant Capacitor, Voltage Doubler

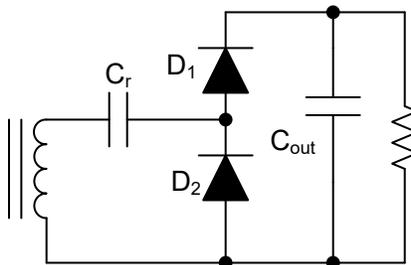


Figure 4-1. One Resonant Capacitor, Voltage Doubler

In this configuration, only one capacitor C_r is needed for the resonant capacitor. There is a DC voltage offset on the resonant capacitor and the voltage doubling is achieved through the transformer voltage in series with the DC offset of the resonant capacitor. When the transformer secondary-side voltage is positive, it adds together with the resonant capacitor offset voltage, diode D_1 is conducting, and the transformer delivers energy to the output. When the transformer voltage is negative, D_1 is off and D_2 is conducting. In this way, the output capacitor C_{out} is refreshed in one half of the switching cycle. In the other half switching cycle, the output capacitor gets no energy from the transformer and supplies the load using its stored energy. The output capacitor C_{out} should be much larger (> 10 times) than the resonant capacitor C_r . Or in other words, the output capacitor C_{out} can be considered as a high-frequency short at the switching frequency.

4.2 Two Resonant Capacitors, Voltage Doubler

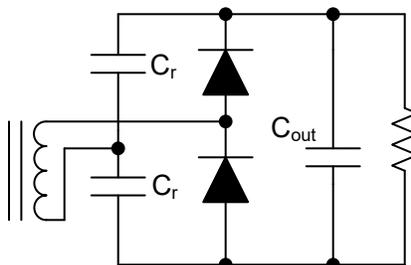


Figure 4-2. Two Resonant Capacitors, Voltage Doubler

In this configuration, the resonant capacitor is split into two capacitors. The equivalent resonant capacitance is the sum of the two capacitances. Both of the resonant capacitors carry a DC offset equal to half of the output voltage. The output capacitor gets energy for each half of the switching cycle. The output capacitor C_{out} still needs to be much larger (> 10 times) than the resonant capacitor. Or in other words, the output capacitor C_{out} can be considered as a high-frequency short at the switching frequency.

4.3 One Resonant Capacitor, Full-wave Rectifier

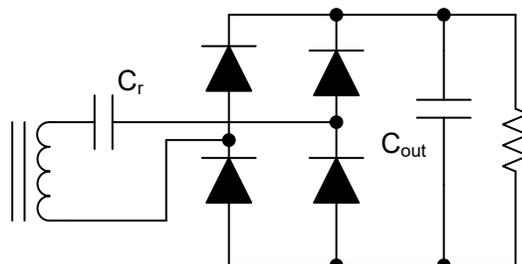


Figure 4-3. One Resonant Capacitor, Full-wave Rectifier

In this case, the full-wave rectifier is used. The output voltage gain is halved compared with the one using voltage doubler. The output capacitor C_{out} still needs to be much larger (> 10 times) than the resonant capacitor. Or in other words, the output capacitor C_{out} can be considered as a high-frequency short at the switching frequency

Other than these three rectification methods, the center-tap method can also be used. However, the center-tap method can only be used with primary side resonant and it is less preferred.

It can also be shown that, with all these three rectification methods, the transformer secondary side is in series with a capacitor. Together with the primary side DC blocking capacitor, the transformer saturation caused by DC offset can be avoided.

[Table 4-1](#) summarizes the difference among these three rectification methods. In this table, "n" represents the transformer primary-side to secondary-side turns-ratio ($N_P : N_S$).

Table 4-1. Rectification Methods Comparison

Rectification Method	Resonant Capacitance	Voltage Gain
One resonant capacitor, voltage doubler	C_r	$1/n$
Two resonant capacitors, voltage doubler	$2C_r$	$1/n$
One resonant capacitor, full-wave rectifier	C_r	$1/2n$

5 LLC Transformer Design Steps

Given very few external components, the open-loop LLC converter design is mainly designing the LLC transformer. The LLC transformer design is a simple process that involves the selection of a suitable transformer turns ratio, volt-second rating, the transformer structure, and the AC resistance. The leakage inductance and magnetizing inductance would be the byproducts of the design, and the circuit can work with the parameters achieved.

5.1 Transformer Turns Ratio Selection

When an LLC converter operates at its resonant frequency, the impedance of the resonant tank is equal to zero. The input voltage and output voltage are virtually connected together. Given this property, we can easily determine the transformer turns-ratio needed. Using the example shown in [Figure 5-1](#), the primary side uses a half bridge. Therefore, the transformer primary side only sees half of the input voltage. The secondary side uses a voltage doubler (Here, the resonant capacitor is the sum of two capacitors). Therefore, the transformer secondary side only sees half of the output voltage. In this case, considering the converter operates close to the resonant frequency, the voltage gain is equal to 1. We can get [Equation 5](#).

$$V_{IN} = n \times V_{OUT} \quad (5)$$

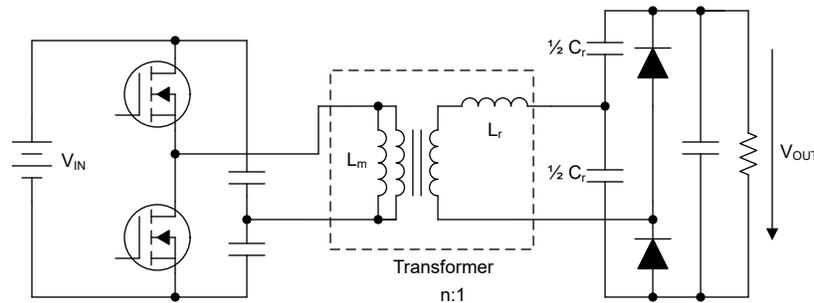


Figure 5-1. LLC Resonant Converter with Secondary-side Resonant and Voltage Doubler

Considering the voltage drop on the diodes, the transformer and the primary-side switches, we can set the transformer turns-ratio n using [Equation 6](#). The extra 1 V can be adjusted as needed to account for the actual voltage drop or extra voltage needed for the post regulator.

$$n = \frac{V_{IN}}{V_{OUT} + 1V} \quad (6)$$

5.2 Transformer Volt-second Rating Calculation

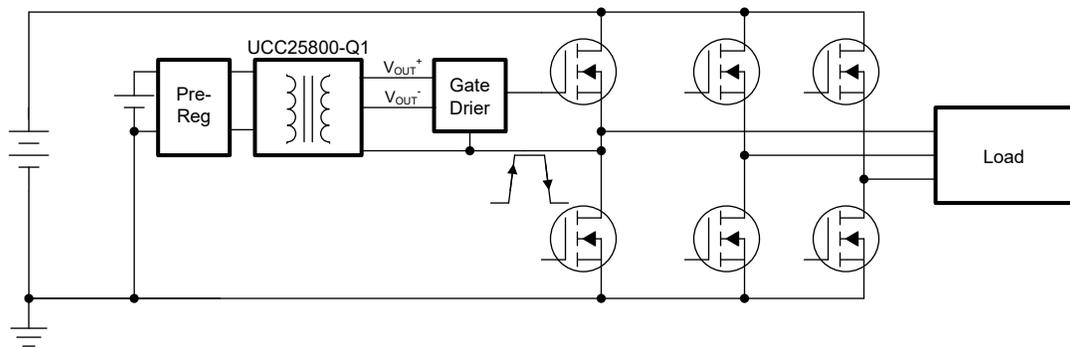
Besides the transformer turns ratio, the volt-second (VS) rating is another important parameter. Given that the transformer primary side sees the half of the input voltage and the flux goes to both the first and third quadrant, the transformer volt-second rating should be calculated based on a quarter of the switching cycle and half of the input voltage. It can be determined by [Equation 7](#). Here f_s is the switching frequency. Some design margin is recommended to consider the tolerance on the switching frequency, as well as the input voltage.

$$VS = \frac{V_{IN}}{8f_s} \quad (7)$$

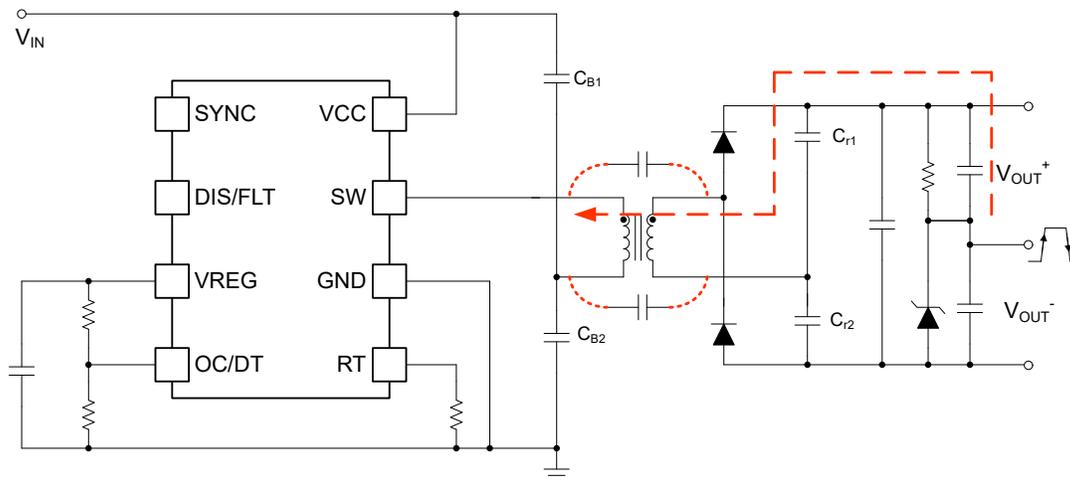
5.3 Transformer Construction

When the UCC25800-Q1 based bias supply is used in the inverter applications, especially for the high-side switches, the high dv/dt on the inverter switch node can couple through the bias supply transformer and causes extra EMI noise, as demonstrated in [Table 5-1](#).

Table 5-1. Noise Coupling Through Bias Supply

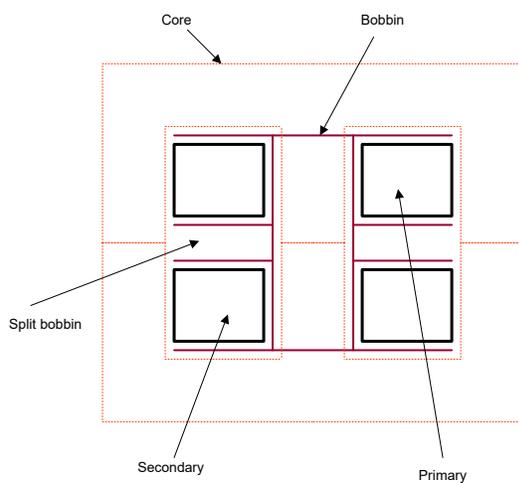


(a) UCC25800-Q1 based gate driver bias supply in inverter application

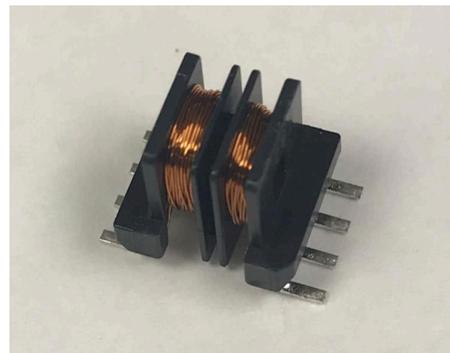


(b) Noise coupling path from inverter power stage to bias supply

To minimize this noise coupling, it is desired to minimize the transformer primary-side to secondary-side capacitance. By physically distancing the primary-side winding and secondary-side winding, a transformer with split chamber bobbin can be used to achieve a minimum parasitic capacitance and simple manufacturing. The split chamber bobbin transformer is illustrated in Figure 5-2.



(a) Transformer Structure



(b) Transformer split chamber bobbin

Figure 5-2. Transformer Construction for Minimum Parasitic Capacitance

5.4 Transformer Winding Selection

The transformer winding design starts with estimating the RMS currents of the primary-side and secondary-side windings. As discussed earlier, at the resonant frequency, the primary-side and secondary-side currents are both sinusoidal. There is a phase shift between these two currents, due to the magnetizing current. If the magnetizing current is ignored, the primary-side current and secondary-side current are scaled by the transformer turns-ratio. Since the output current is equal to the average value of the rectified transformer secondary-side current, the RMS current of the transformer windings can be easily estimated by Equation 8 and Equation 9. Considering the tolerances on the switching frequency and component values, an extra 20~30% design margin is recommended for these currents.

Secondary-side winding RMS current:

$$I_{\text{rmsS}} = \frac{\pi}{\sqrt{2}} I_0 \quad (8)$$

Primary-side winding RMS current:

$$I_{\text{rmsP}} = I_{\text{rmsS}}/n \quad (9)$$

Even though the converter gain is equal to the transformer turns-ratio when the switching frequency is equal to the resonant frequency, the output voltage will be slightly lower than the theoretical value due to the loss elements in the converter, including the diode voltage drop (V_f), the primary-side switch on-state resistance (R_{dson}), the transformer AC resistance R_{ac} , the resonant capacitor ESR (R_{ESR}), as well as the diode ESR (R_{Diode}). The overall output voltage can be estimated based on Equation 10. Given the transformer AC resistance affects the voltage gain, to improve the load regulation, it is desired to minimize the transformer AC resistance.

$$V_{\text{OUT}} \cong \frac{V_{\text{IN}}}{n} - 2V_f - \frac{\pi^2}{2} \times \left(\frac{R_{\text{dson}}}{n^2} + R_{\text{ac}} + R_{\text{ESR}} + R_{\text{Diode}} \right) \times I_0 \quad (10)$$

The transformer AC resistance should be measured from the secondary side, with primary side shorted, and at the resonant frequency, as show in Figure 5-3.

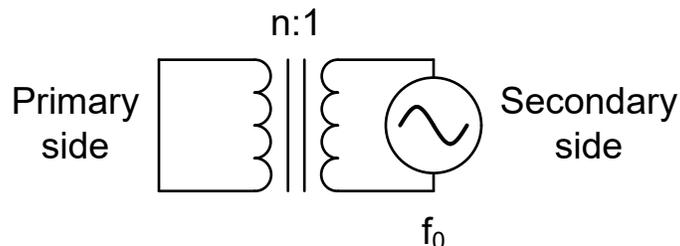


Figure 5-3. Transformer AC Resistance Measurement

Ideally, this impedance should be as low as possible. Based on the equation, we can estimate the voltage drop caused by the load current and estimate how much AC resistance can be allocated for the transformer. The R_{dson} is coming from UCC25800-Q1 and it can be estimated using 0.3 Ω . The ESR of the capacitor can be ignored if the NP0 capacitor is used. The X7R capacitor would have a larger ESR. The diode resistance can be estimated based on the diode forward voltage drop curves, normally it is about 0.3 Ω .

5.5 Transformer Inductance

As described earlier, the transformer parasitic inductance, including the leakage inductance and the magnetizing inductance can be used as part of the resonant circuit. The resonant capacitor can be designed based on these inductances to match the desired switching frequency.

5.5.1 Leakage inductance

When the split chamber bobbin is used, the primary-side winding and secondary-side winding are physically separated. The coupling between two windings is poor and large leakage inductance is created. When using the

secondary-side resonant, the method described in [Figure 5-3](#) can be used to measure the leakage inductance (L_k) from the transformer secondary side. The leakage inductance is measured on the secondary side when the primary side is shorted. Following the same principle, when the primary-side resonant is used, measure the leakage inductance from the transformer primary side with the secondary side shorted. Once the leakage inductance is measured, the resonant capacitor C_r can be selected according to [Equation 11](#). In this equation, L_k is the leakage inductance measured and f_0 is the resonant frequency. The resonant frequency can be chosen as 10% above the switching frequency to allow component tolerances. If the one resonant capacitor configuration is used, the resonant capacitor should be the calculated total resonant capacitor C_r .

$$C_{r1} = C_{r2} = \frac{1}{2}C_r = \frac{1}{2} \times \frac{1}{4\pi^2 L_k f_0^2} \quad (11)$$

5.5.2 Magnetizing inductance

Traditionally in LLC converters, the magnetizing inductance serves two purposes: to help achieve ZVS (Zero Voltage Switching) and to help boost the voltage gain up when the switching frequency is below the resonant frequency, to compensate the input voltage variation. In the open-loop LLC bias supply designs, there is no need to boost the voltage gain up. The only purpose of the magnetizing inductance is to help achieve ZVS. Based on the ZVS criteria, the design target of the magnetizing inductance can be calculated based on [Equation 12](#). In this equation, L_m is the magnetizing inductor value, t_d is the dead-time, f_{SW} is the switching frequency, and C_{SW} is the SW-pin parasitic capacitance (it has a typical value of 170 pF).

$$L_m = \frac{t_d}{8C_{SW}f_{SW}} \quad (12)$$

Furthermore, because it is a low voltage input, even without achieving the ZVS, the switching loss is not significant. We could live with the magnetizing inductance without using the air-gap.

If the magnetizing inductance is much larger than the leakage inductance (> 20 times), the transformer can use the core without an air-gap.

If the leakage inductance is a much larger percentage of the magnetizing inductance, to minimize the error caused by the magnetizing inductance tolerance, it is desired to insert an air-gap in the transformer core. In this case, the air-gap should be minimal without causing additional manufacturing cost.

5.6 Transformer Selections

To simplify the design process, different transformer vendors have designed and released standard LLC transformers for the isolated bias supply applications. In [Table 5-2](#), there is a summarized set of transformers that can be used together with UCC25800-Q1 for the typical input and output voltage for different applications. This list is not exhaustive and there are many other standard transformers that work for this type of application.

Table 5-2. List of Transformers for Typical Input and Output Voltages

Input voltage V_{IN}	Output voltage V_{OUT}	Transformer turns ratio needed	Manufacturer: Part number (Turns ratio)	Isolation rating
12 V	5 V	2:1	Wurth: 750319831 (2.2:1)	4000VAC
			Coilcraft: ZC1576-AL (2:1)	1500VAC
	12 V	1.08	Wurth: 750319331 (1:1)	4000VAC
			Wurth: 750319688 (1:1) ¹	3125VAC
			Wurth: 750319834 (1.08:1)	4000VAC
			Wurth: 750319848 (1:1) ^{1 2}	3125VAC
	15 V	1:1.33	Wurth: 750319835 (1:1.29)	4000VAC
			Coilcraft: ZC1580-AL (1:1.33)	1500VAC
	18 V	1:1.58	Wurth: 750319836 (1:1.57)	4000VAC
			Sumida: 13399-T052 (1:1.5)	1800VAC
	24 V	1:2.08	Wurth: 750319831 (1:2.2)	4000VAC
			Coilcraft: ZC1576-AL (1:2)	1500VAC

Table 5-2. List of Transformers for Typical Input and Output Voltages (continued)

Input voltage V_{IN}	Output voltage V_{OUT}	Transformer turns ratio needed	Manufacturer: Part number (Turns ratio)	Isolation rating
15 V	5 V	2.5:1	Würth: 750319831 (2.2:1)	4000VAC
			Coilcraft: ZC1582-AL (2.5:1)	1500VAC
	12 V	1.15:1	Würth: 750319833 (1.2:1)	4000VAC
			15 V	1:1.07
	Würth: 750319688 (1:1) ¹	3125VAC		
	Würth: 750319834 (1.08:1)	4000VAC		
	Würth: 750319848 (1:1) ^{1 2}	3125VAC		
	18 V	1:1.27	Würth: 750319835 (1:1.29)	4000VAC
			Coilcraft: ZC1580-AL (1:1.33)	1500VAC
24 V	1:1.67	Würth: 750319177 (1:1.67)	2500VAC	
		Coilcraft: ZC1581-AL (1:1.67)	1500VAC	
24 V	12 V	1.85:1	Würth: 750319832 (1.83:1)	4000VAC
			Renco: RLTI-1430 (1.85:1) ³	6000VAC
	15 V	1.5:1	Würth: 750319836 (1.57:1)	4000VAC
			Sumida: 13399-T052 (1:1.5)	1800VAC
	24 V	1:1.04	Würth: 750319833 (1.2:1)	4000VAC
			Würth: 750319331 (1:1)	4000VAC
			Würth: 750319688 (1:1) ¹	3125VAC
Würth: 750319834 (1.08:1)			4000VAC	
Würth: 750319848 (1:1) ^{1 2}	3125VAC			
30 V	12 V	2.31:1	Würth: 750319831 (2.2:1)	4000VAC
	15V	1.88:1	Würth: 750319832 (1.83:1)	4000VAC
			Renco: RLTI-1430 (1.85:1) ³	6000VAC
	18 V	1.58:1	Würth: 750319836 (1.58:1)	4000VAC
	24 V	1.2:1	Würth: 750319833 (1.2:1)	4000VAC

1. Low profile design
2. Dual outputs
3. 16 mm between primary-side pins to secondary-side pins

6 Negative Voltage Generation

In gate driver applications, often the bias supply needs to provide both the positive voltage and the negative voltage to the gate driver circuit. The positive voltage is to ensure full enhancement of the switching devices. The negative voltage is to provide better noise immunity to prevent shoot through, or increases the turn-off speed for lower switching loss. The LLC circuit can easily create a single output voltage. Some simple circuits can be added to split the single output voltage into positive and negative voltages.

6.1 Using a Zener Diode

The output voltage can be split using a simple Zener diode if the voltage regulation requirement is not stringent. The method is shown in [Figure 6-1](#).

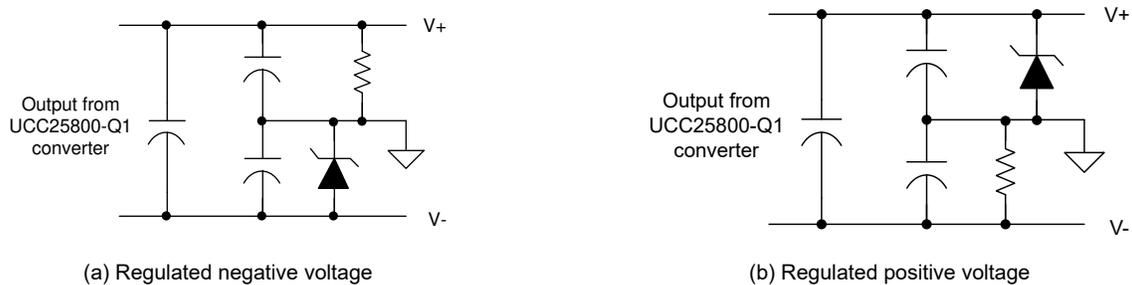


Figure 6-1. Splitting A Single Output Voltage Using A Zener Diode

In the example of regulated negative voltage, the negative output voltage is determined by the Zener diode. The remaining voltage is left to be the positive voltage. In this case, the resistor is needed to provide the bias current for the Zener diode. The positive output and negative output capacitors can be designed to be proportional to its corresponding voltage. In this case, the two output voltages would rise proportionally. Otherwise, the Zener diode side could reach its final value much earlier than the positive side. The Zener diode can also be put on the positive side so that the positive voltage is determined by the Zener diode and the negative voltage is the rest of the voltage.

6.2 Using a Shunt Regulator

The Zener diode provides a simple and cost effective solution. However, due to the tolerances on the Zener voltage, it might not meet the voltage regulation requirement. A more accurate shunt regulator can be used to improve the voltage regulation, as shown in [Figure 6-2](#).

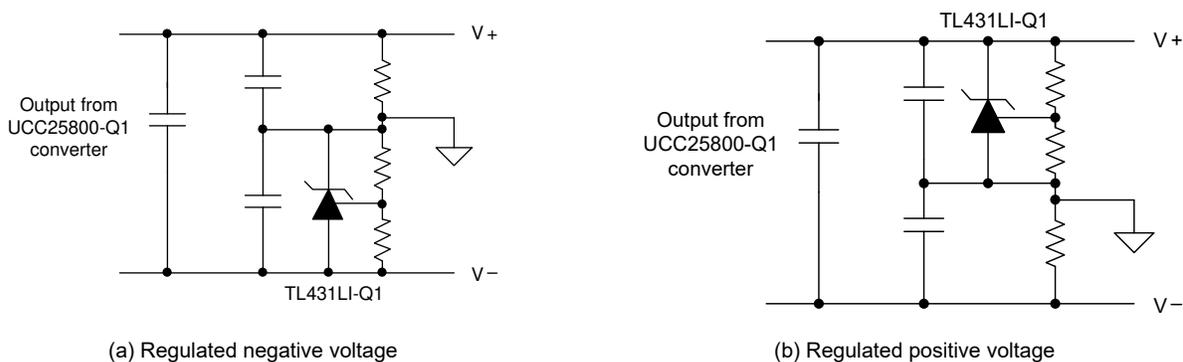


Figure 6-2. Splitting A Single Output Voltage Using A Shunt Regulator

In the example of regulated negative voltage, the negative output is regulated by the shunt regulator. The remaining of the voltage is left on the positive output.

6.3 Using a Shunt Regulator and Linear Regulator

In some cases, both the positive output and negative output need to be well regulated. In this case, an extra linear regulator can be used to improve the positive voltage rail voltage accuracy. The circuit diagram is shown in Figure 6-3. The linear regulator is added on top of the shunt regulator circuit. In this case, both the positive output and negative output are well regulated. When designing the LLC converter output voltage, the extra voltage drop on the linear regulator must be considered.

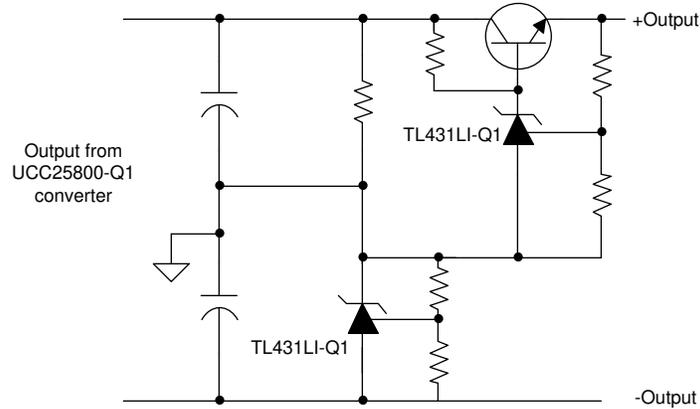


Figure 6-3. Shunt Regulator And Linear Regulator To Split The Output Voltage

7 Multiple-output Design

In electrical vehicle designs, different gate driver architectures are used based on the cost, the reliability, and the safety requirement. Normally, we see three different architectures, as shown in Figure 7-1. In the pictures, GDx block represents the gate driver circuit. GD1, GD3, and GD5 are the three high-side gate drivers that is biased from its individual ground. GD2, GD4, and GD6 are the three low side gate drivers that share the same common ground. The transformer block represents the isolated bias supply.

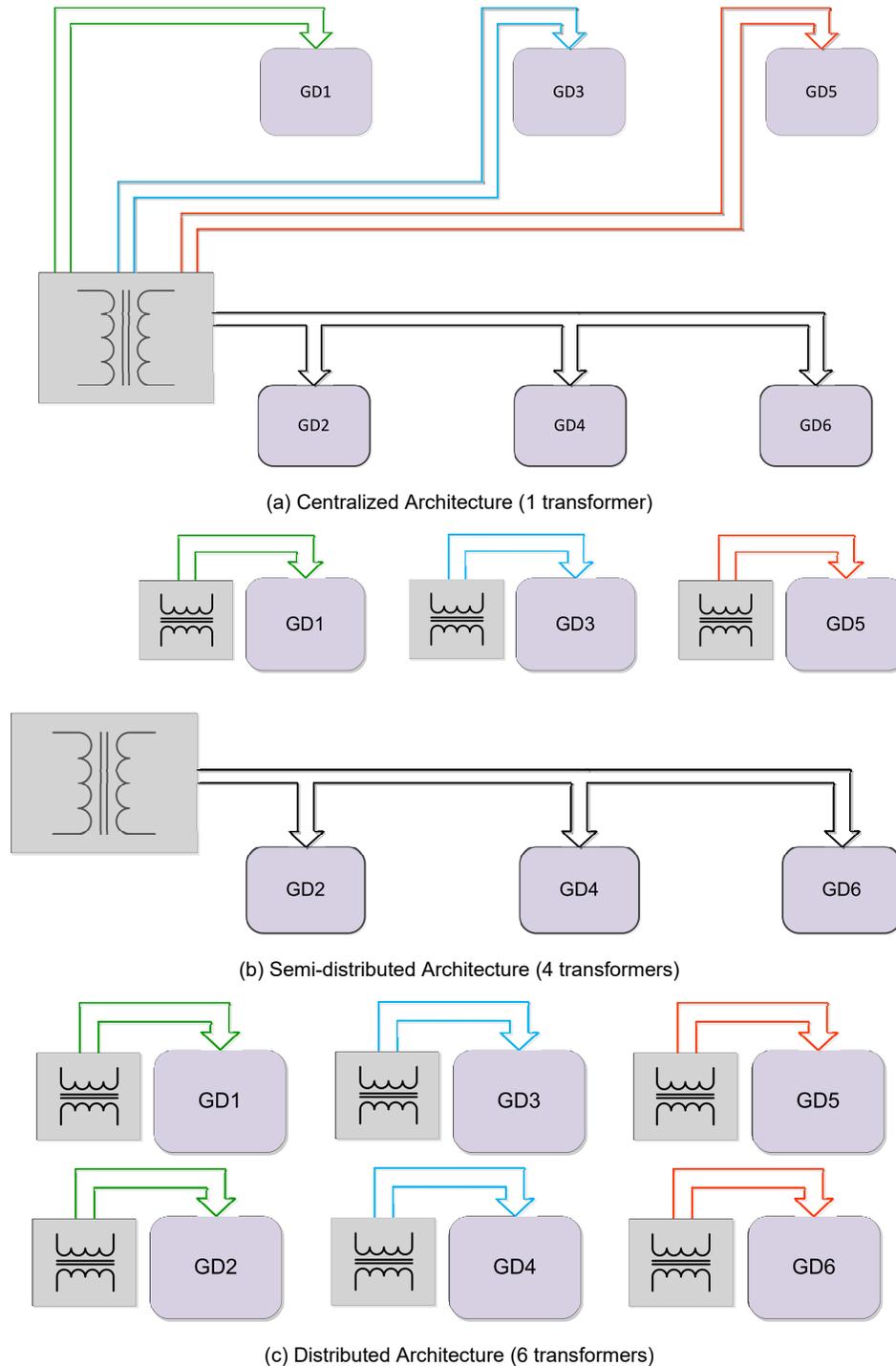


Figure 7-1. Gate Driver Bias Supply Architectures

The centralized architecture uses a single bias supply to bias all the gate driver circuits. The semi-distributed architecture biases all the three low side drivers using one bias supply, and uses three bias supplies to bias each of the high side switches. The distributed architecture biases each gate driver with an individual bias supply.

For UCC25800-Q1, multiple outputs can be easily achieved through different circuit configurations.

7.1 One UCC25800-Q1 Drives Each Output

Given the low component count, each output can be created by an individual UCC25800-Q1 circuit. This approach works well for the distributed architecture.

7.2 Transformer With Multiple Secondary-side Windings

The LLC converter is able to support multiple secondary-side windings and their output voltages are proportional to the transformer turns-ratio. Given the complicated transformer leakage inductance distribution, the primary-side resonant is preferred in this configuration, as shown in Figure 7-2. Since this uses primary-side resonant, the primary-side capacitors are the resonant capacitor and the secondary-side series capacitors are the DC-blocking capacitor for the voltage doubling purpose. The voltage-regulation tolerance is not as good as secondary-side resonant. Besides, all the secondary-side windings are on the same transformer, the noise coupling among these outputs needs to be considered.

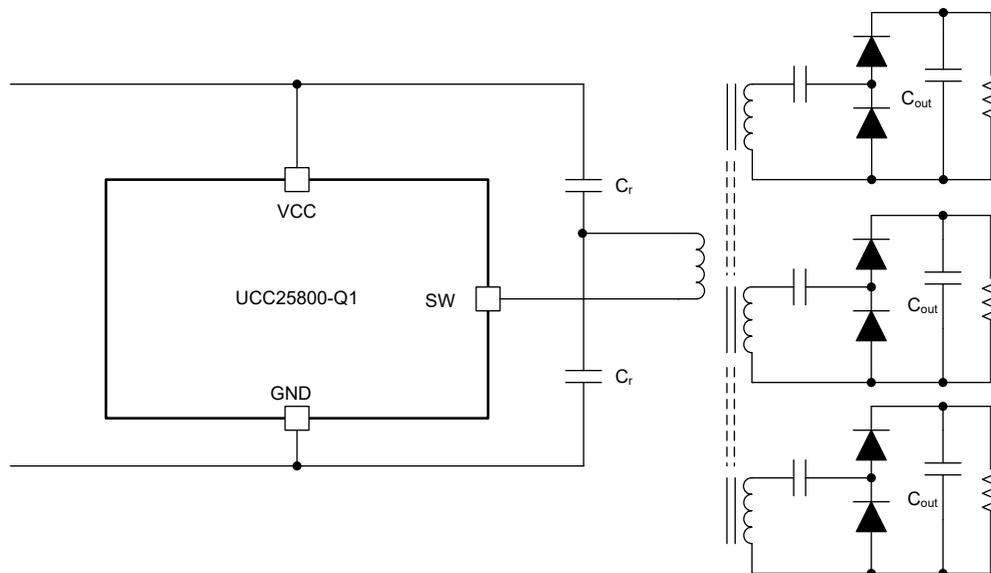


Figure 7-2. UCC25800-Q1 Drives One Transformer With Multiple Secondary-side Windings

7.3 Multiple Transformers

Multiple output voltages can also be accomplished by using one UCC25800-Q1 to drive multiple transformers. In this case, each transformer can be configured as secondary-side resonant and the tolerance on the resonant frequency can be mitigated. The secondary-side resonant capacitor can be chosen to achieve the same resonant frequency for each transformer. The example circuit configuration is shown in Figure 7-3.

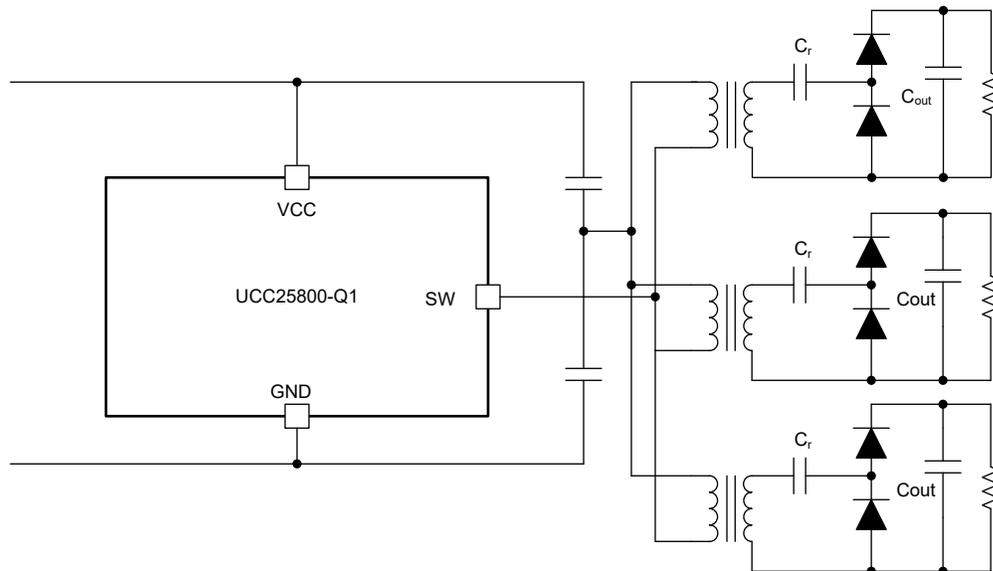


Figure 7-3. UCC25800-Q1 Drives Multiple Transformers

This implementation method is verified by modifying PMP22835 6-W 24V input, 24V output reference design. Using one primary-side circuit to drive three sets of transformers and secondary-side circuits, as shown in [Figure 7-4](#).

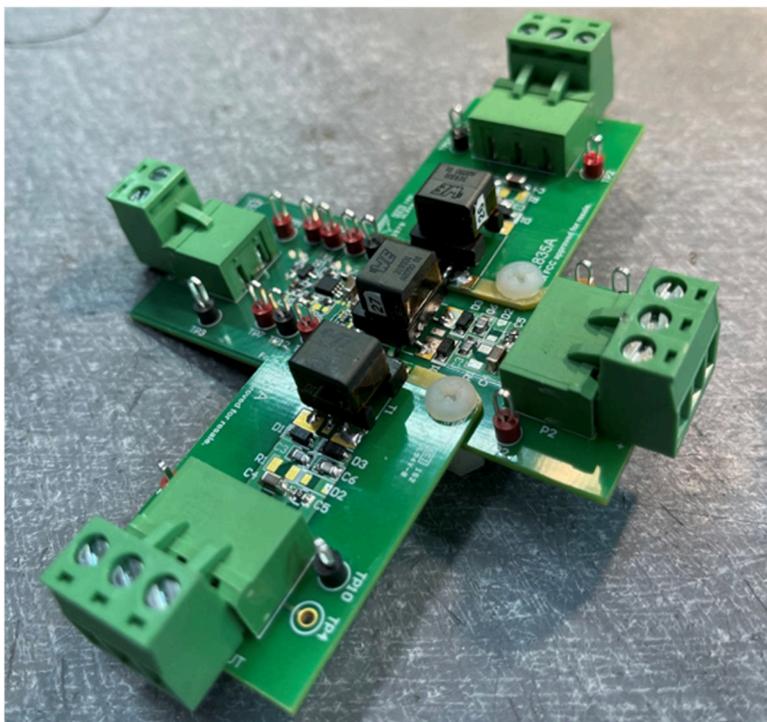


Figure 7-4. Hardware of One UCC25800-Q1 Driving Three Sets of Transformers

Three output voltages at different load conditions are summarized in [Figure 7-5](#). It can be seen with different transformers, the output voltage remains the same when they are loaded with the same load current. It can be noticed that when the load is extremely light (less than a few mA), the output voltage rises significantly. This is due to the extra energy delivered to the secondary side through charging the secondary-side diode junction capacitors. The energy delivered is limited and it can be ignored when the output voltage is loaded. However, when the output load is very light, the delivered energy might be more than the load can consume. This causes the output voltage to rise. Since the extra energy is limited, normally the gate driver IC quiescent current is large

enough to consume the energy. If the quiescent power is not sufficient to consume the energy, a Zener diode or dummy-load resistor can be used to reduce the output voltage rise. Only a few mA load can bring the output voltage to the regular levels.

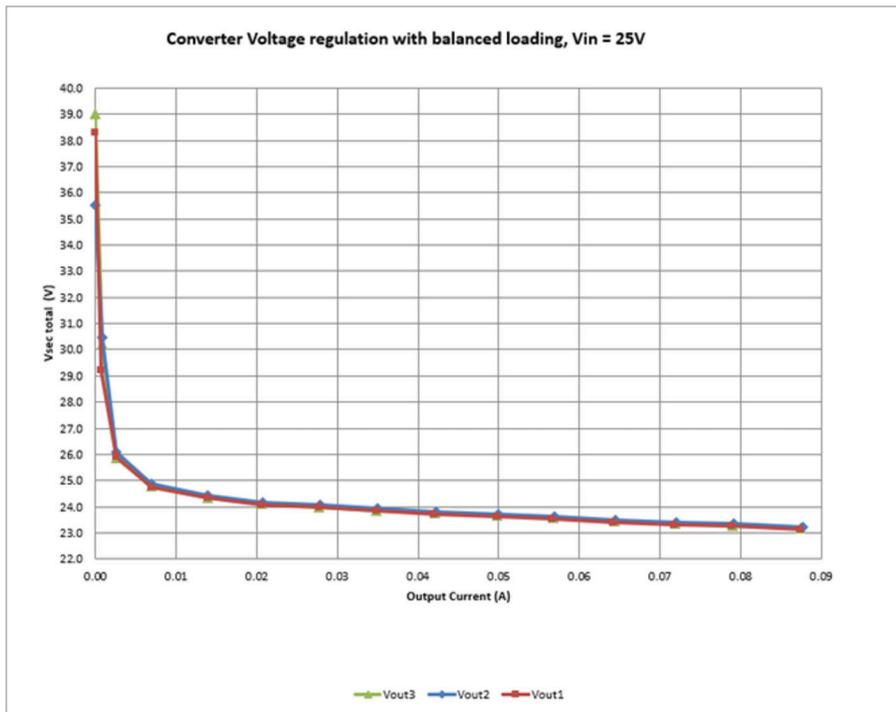


Figure 7-5. Output Voltage Regulation For One UCC25800-Q1 Driving Three Sets Of Transformers

8 EMI Performance

Traditionally, the isolated bias supply is created through PWM converters like flyback or push-pull converters. These converters have a triangular or trapezoidal shape input current, which has more harmonic content and creates more EMI noise. Besides, both of these converters typically operate as hard switching and generate higher dv/dt on the switch node, which creates more common mode noise. Furthermore, due to the topology limitations, they need to work with the transformer with fairly low leakage inductance, which means the transformer primary-side to secondary-side parasitic capacitance will be large (10 ~ 30 pF range). The high dv/dt , together with the large parasitic capacitance make the flyback and push-pull perform poorly from an EMI standpoint.

Unlike the PWM converters, the LLC converter has a sinusoidal shape current. It is also capable of achieving ZVS with lower switch node dv/dt . Furthermore, the LLC converter is able to use a transformer with split chamber bobbin and significantly reduce the transformer primary-side to secondary-side parasitic capacitance (below 2 pF). The sinusoidal shape current, lower dv/dt , and the low parasitic capacitance make the LLC a much better solution from the EMI noise point of view.

[Table 8-1](#) summarizes the parasitic capacitance comparison between different solutions.

Table 8-1. Transformer comparisons among different topologies

	LLC	flyback	push-pull
Construction	Split chamber bobbin	One chamber bobbin	Toroid
Primary-side to secondary-side parasitic capacitance	< 2 pF	~ 20 pF	~ 10pF

8.1 EMI Performance With Standalone Converter

To compare the performance of the three different topologies (LLC, flyback, and push-pull), three boards are tested as stand alone converters, instead of using them to bias the gate drivers. The LLC converter is based on the UCC25800-Q1 with 24-V input and 24-V output, 6-W maximum power. The flyback converter uses the [LM5180 EVM](#), also setup with 24-V input and 24-V output, 200mA maximum load. The push-pull converter is based on [SN6505B](#), with 5-V input, with 22-V output voltage. All three converters are connected with a 150-Ω resistor load. The setup follows the CISPR 25 Class 5 (Conducted Emission) automotive EMI test standard.

[Figure 8-1](#) through [Figure 8-3](#) demonstrate the measured EMI performance from three different converters. The EMI filters are removed to show the noise generated by each converter itself. Due to the different bypass capacitor values, the differential-mode noise is not quite the same as the converter generated, while the common-mode and high-frequency noise should not be affected and are properly represented. From the noise measurement, it can be observed the LLC converter provides a much lower EMI noise level, especially at high-frequency range (around 20-dB reduction). As described earlier, the soft switching and lower parasitic capacitance help to achieve lower EMI noise.

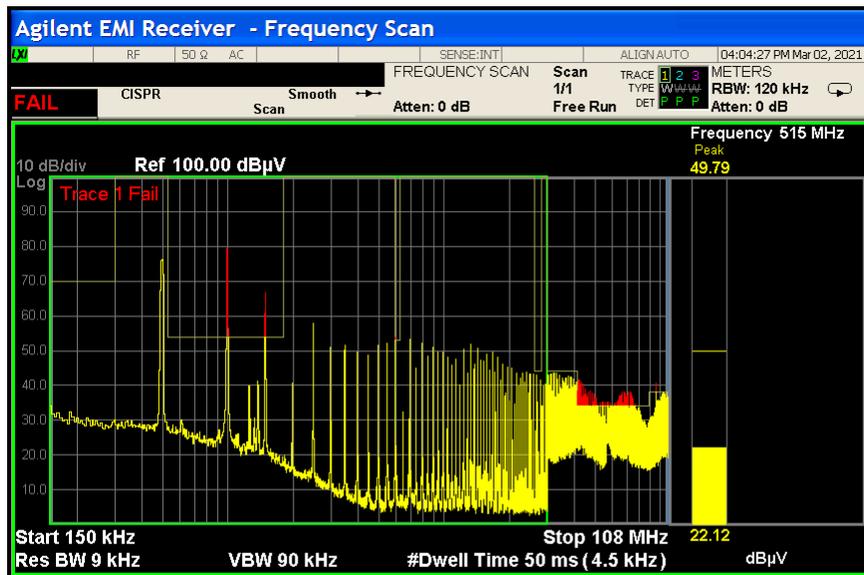


Figure 8-1. EMI Noise Measured From The LLC Converter

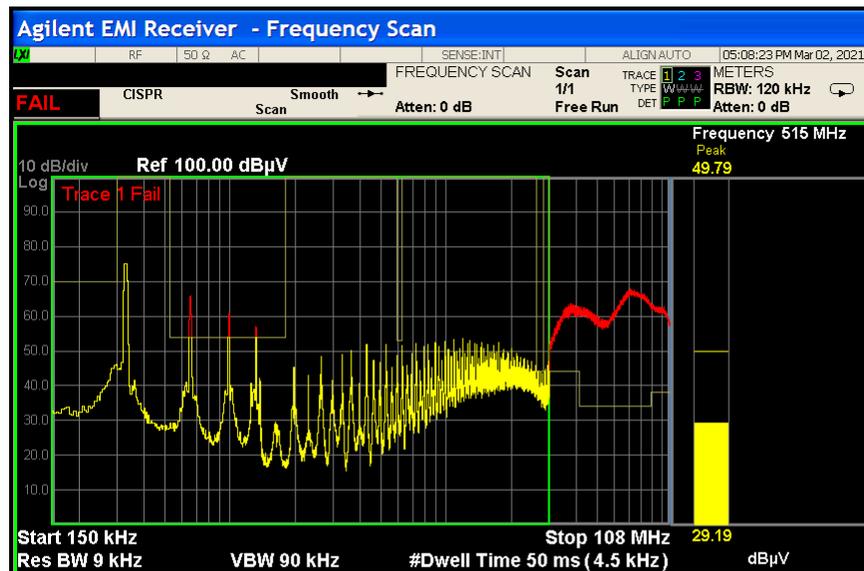


Figure 8-2. EMI Noise Measured From The flyback Converter

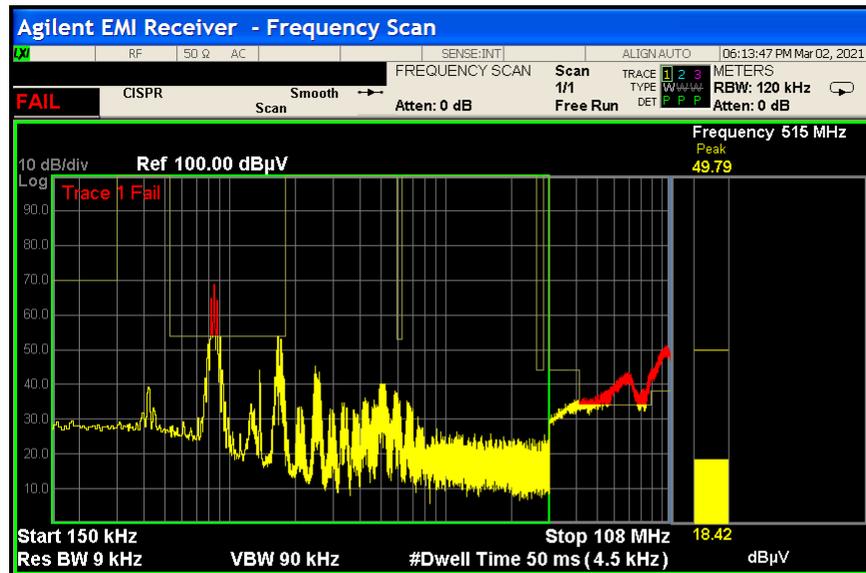


Figure 8-3. EMI Noise Measured From The push-pull Converter

8.2 EMI performance with an inverter power stage

The LLC converter and flyback converter EMI performances are also measured when they are used to bias an inverter gate drivers. A half-bridge inverter design based on Recom Power inverter [EVM R-REF01-HB](#) was biased with an LLC converter and the flyback converter, respectively. The EMI noise is measured from the 24-V input side, and the inverter is powered through a high-voltage power supply with 400-V DC output. The inverter is switched at 20 kHz with 50% duty cycle with no load. The inverter creates high dv/dt through its switch node and coupled through the transformer parasitic capacitance. This coupled noise can be measured through the LISN connected at the 24-V input. The measurement setup is shown in [Figure 8-4](#). Similar to the stand alone board setup, no EMI filter is used to measure the raw noise coming from the setup.

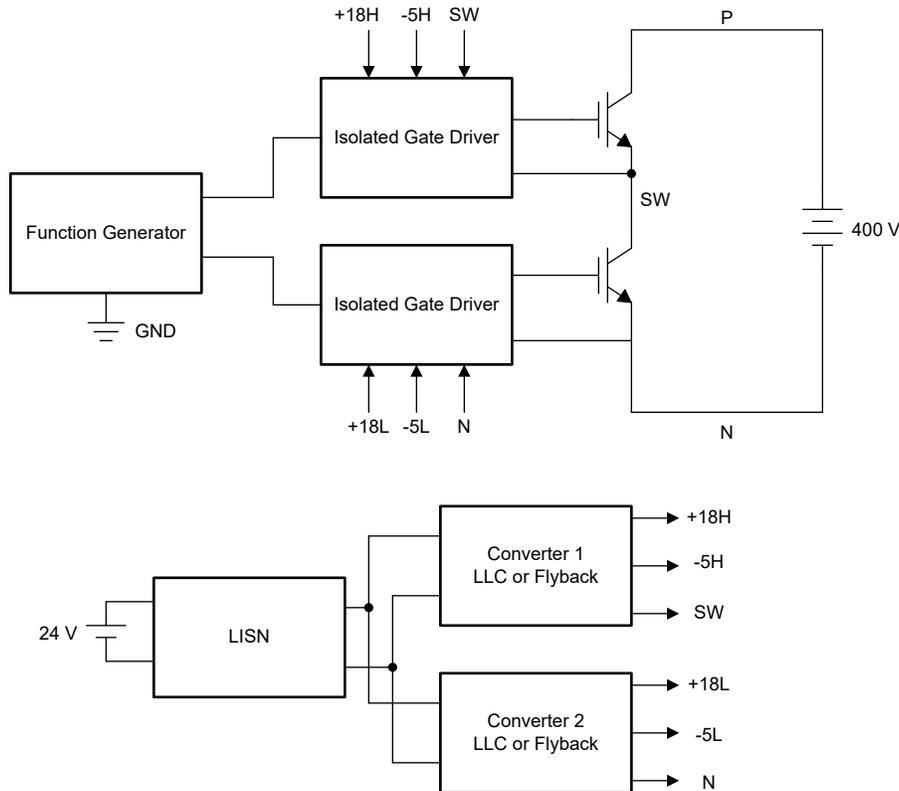


Figure 8-4. Inverter Setup For Measuring EMI Performance

The measured EMI performances is summarized in [Figure 8-5](#) and [Figure 8-6](#). The noise floor is raised due to the noise generated by the inverter power stage. However, it can be observed that the LLC converter still provides much lower EMI noise performance comparing with flyback converter, especially at the higher frequency range.

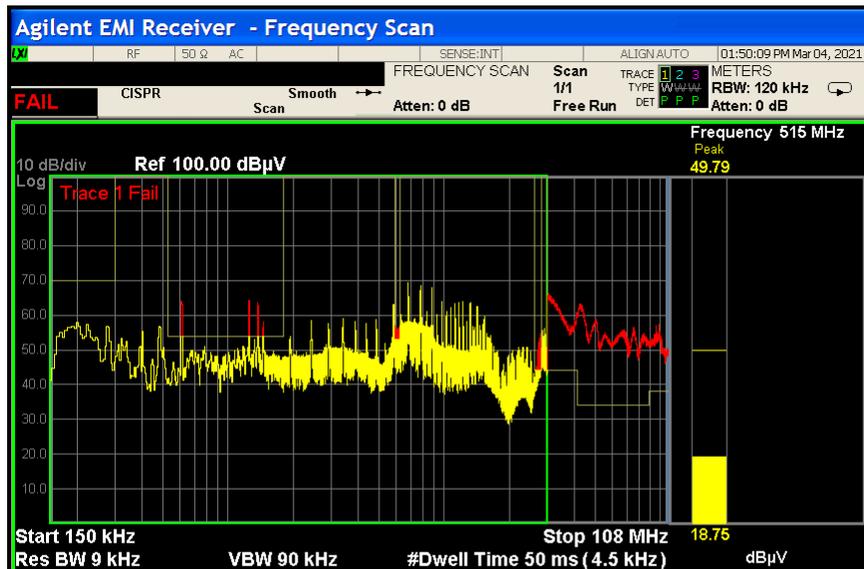


Figure 8-5. EMI Performance Measured Using LLC Converter Biasing Inverter

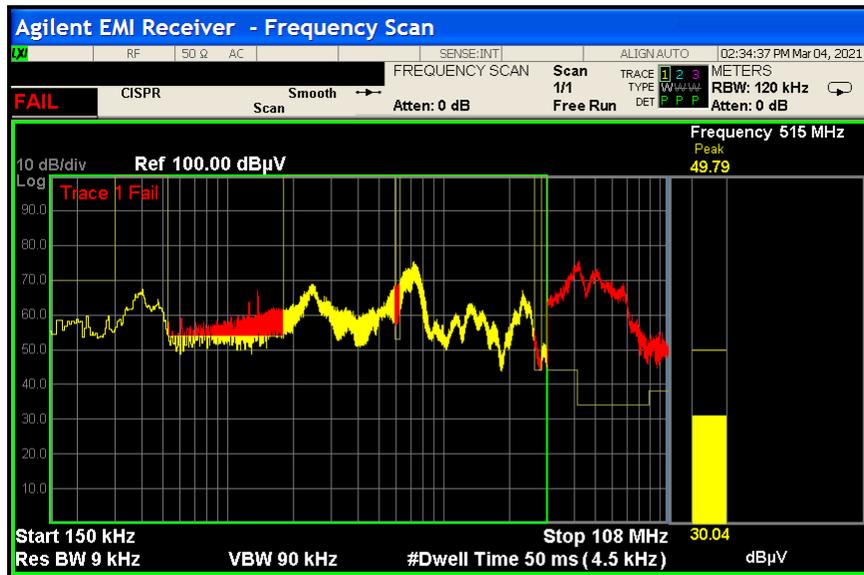


Figure 8-6. EMI Performance Measured Using flyback Converter Biasing Inverter

9 Common-Mode Transient Immunity (CMTI)

With the development of wide-bandgap power-semiconductor technology, inverters start to use SiC or GaN MOSFETs as the switching devices. These wide bandgap devices have faster switching speed and create higher dv/dt at the switch node, sometimes well above 50V/ns. This high dv/dt can couple through the transformer parasitic capacitance, as shown in [Noise Coupling Through Bias Supply](#). Common-mode transient immunity (CMTI) test becomes a standard requirement for this type of application. A UCC25800-Q1 based open-loop LLC converter uses a transformer with low parasitic capacitance. This makes the solution demonstrate excellent CMTI performance. The CMTI test setup is shown in [Figure 9-1](#).

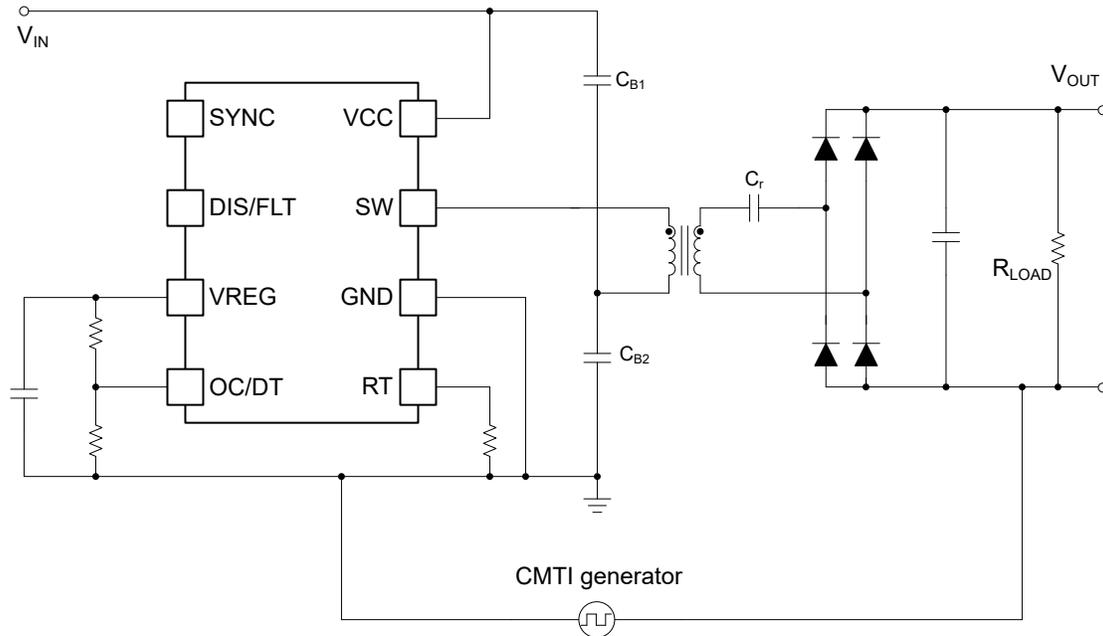
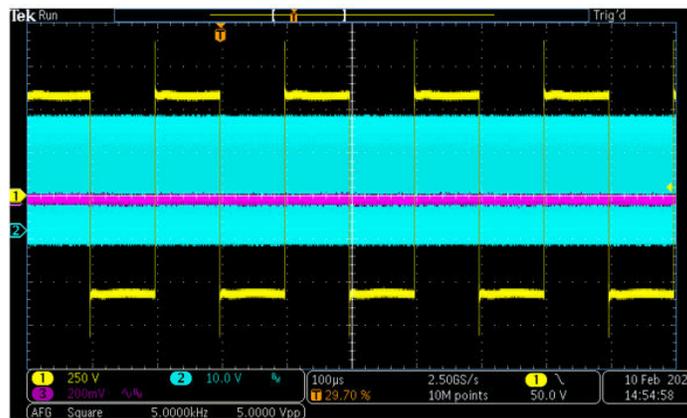


Figure 9-1. CMTI Test Setup

A CMTI generator was used to create CMTI strikes. It operates with 5-kHz switching frequency and 50% duty cycle. Its rise and fall rate are 165V/ns and 155V/ns respectively. With such a high voltage slew rate, the UCC25800-Q1 still operates normally. Its SW pin voltage remains a fixed switching frequency square wave.



(a) CMTI strikes don't interfere with converter switching (CH1: CMTI strike; CH2: UCC25800-Q1 SW-pin voltage; CH3: Not used)

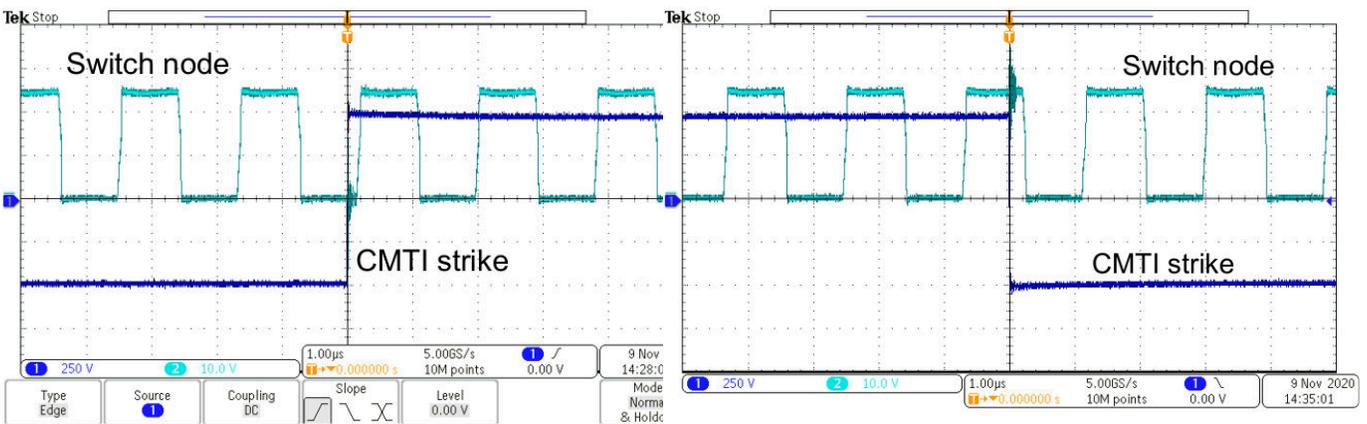


Figure 9-2. CMTI Performance of UCC25800-Q1 Based Converter

The open-loop LLC converter uses a transformer with such a low parasitic capacitance. The current injection from the high dv/dt switch node becomes greatly reduced when compared to legacy solutions. Therefore, this solution can easily pass the CMTI test.

10 Summary

The UCC25800-Q1 provides a simple solution for an open-loop LLC converter. Due to its low EMI, robust operation, and better CMTI performance, as well as the low solution cost, the open-loop LLC converter is a desirable solution for isolated bias supplies.

11 Revision History

NOTE: Page numbers for previous revisions may differ from page numbers in the current version.

Changes from Revision * (March 2021) to Revision A (December 2021)	Page
• Updated magnetizing inductance design guideline.....	12
• Update transformer selection table.....	12

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