

## **4.2-W GU10 LED Lighting Driver Using Primary-Side Regulation**

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### **ABSTRACT**

This application note introduces a low-wattage LED lighting driver solution using TI off-line primary-side sensing controller TPS92310. By using constant on-time flyback topology with primary sensing control, the solution can achieve high efficiency and good line and load regulation. The reference design PMP4325 is a suitable form factor for GU10 replacement LED lamp (30 mm x 18 mm x 10 mm), which can support universal AC line input and three or four LEDs in serial output with a constant output current of 350 mA. The experiment demonstrates that this solution has good line and load regulation, high efficiency, and overall protection for LED lighting.

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## 1 Theory Operation

### 1.1 The TPS92310 Controller

Single-stage flyback is a highly-attractive topology for LED lighting with low power rating. Single-stage flyback is widely used for LED lighting for the following reasons:

- Low total BOM with electrical isolation
- High power factor using specific control architectures, such as constant on-time control
- Small form factor compared to other two-stage topology

Although single-stage flyback has benefits for LED lighting, some issues still require resolution. These issues are:

- High power factor
- Primary-Side Regulation (PSR) with good line and load regulation
- Open or short protections of LEDs

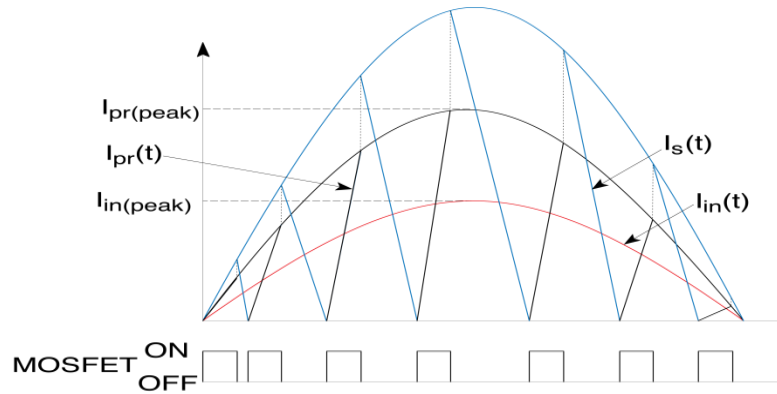
The TI TPS92310 controller is a single-stage primary-side sensing AC/DC controller to drive a constant current for high-brightness LEDs. It runs at transition mode (TM) with zero-current-detection. The ON-Time ( $T_{ON}$ ) is nearly constant within a half-cycle of the line voltage. Hence, it has inherent power factor correction (PFC) because the peak current of the primary winding is following with the input line voltage profile. The  $T_{ON}$  is adjusted so that the LED current is regulated at a preset level which is set by an external-sense resistor. The  $T_{ON}$  is also used to design for controlling a flyback, boost and buck-boost converter operating in TM with fixed on-time control to achieve high PF, and also for controlling a buck converter in TM with peak current control for general LED driver.

Primary-side sensing does not require opto-coupler and secondary-side circuitry and hence a low component count and compact PCB solution. Additionally, the controller provides features such as cycle-by-cycle current limit, output short-circuit protection, output overvoltage protection (OVP), or open LED protection, short-LED protection, and thermal shutdown; all of which provide protections for LED lighting.

### 1.2 Constant On-time Control

In a conventional boost PFC converter, TM with constant on-time control is generally used to keep input current in-phase with input voltage for high PF and low THD.

For single-stage flyback topology with TM, it is not the natural PFC because the duty cycle and frequency will always change during the switching cycle. So the power factor and THD is not optimal at this condition. Fortunately, the fixed or constant  $T_{ON}$  with TM single-stage flyback can still achieve a high PF and low THD; as shown in Figure 1, the average input current is an approximate sinusoidal wave with the same phase of input voltage.



**Figure 1. Current Waveform During  $T_{ON}$  and  $T_{OFF}$**

The TPS92310 controller is configured on constant on-time control mode in this design, the turn-on time of the switch can be fixed if a large capacitor is connected to the COMP pin to filter 100-Hz line ripple in a single stage flyback application. However, to reduce the size of the board, this reference design is not a single stage without PFC function, so a small compensation capacitor is used only to keep the stabilization of control loop, the turn-on time is almost fixed because of stable DC input voltage for flyback.

### 1.3 Constant current control with Primary Side Sensing

The primary current, secondary current and  $V_{ds}$  voltage are shown in Figure 2, according to this, the average output current  $I_o$  can be calculated as following equation (1).

$$I_o = \frac{I_{s\_pk} \times T_{OFF}}{2 \times (T_{ON} + T_{OFF} + 2 \times T_{dly})} = \frac{I_{p\_pk} \times N \times T_{OFF}}{2 \times (T_{ON} + T_{OFF} + 2 \times T_{dly})} \quad (1)$$

Where:

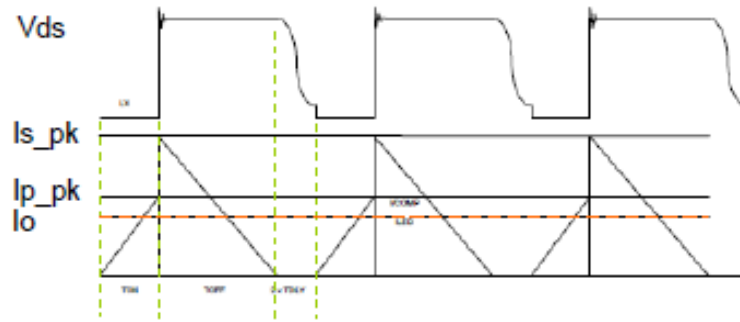
$2 \times T_{dly}$  = half time of ringing period on MOSFET drain

N = transformer turn ratio of primary winding to secondary winding

$I_{p\_pk}$  = primary current

$I_{s\_pk}$  = secondary current

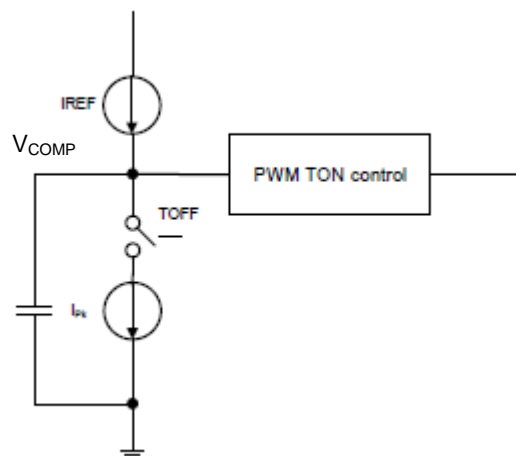
$I_o$  = average output current (LED current)



**Figure 2. Current and  $V_{ds}$  Voltage Waveform**

To regulate the output current, the converter used a PWM control circuit, as shown in Figure 3. This circuit includes charge and discharge operation modes. The charge operation mode is controlled by an internal reference current  $I_{REF} \times \text{time}$  ( $T_{ON} + T_{OFF} + 2T_{DLY}$ ). The discharge operation mode is controlled by a  $T_{OFF}$  switch and  $I_{pk}$  current source, which is proportional to the primary side peak current. COMP voltage level can represent a gate drive  $T_{ON}$ .

During normal operation, if discharge  $Q(I_{pk} \times T_{OFF})$  is greater than charge  $Q(I_{REF} \times (T_{ON} + T_{OFF} + 2T_{DLY}))$ , the COMP pin voltage decreases, as a result gate output  $T_{ON}$  decreases at the next cycle. Otherwise, if the charge  $Q(I_{REF} \times (T_{ON} + T_{OFF} + 2T_{DLY}))$  is larger than discharge  $Q(I_{pk} \times T_{OFF})$ ,  $V_{COMP}$  increases, gate driver output  $T_{ON}$  increases at the next cycle. If the charge  $Q$  equals the discharge  $Q$ , the  $V_{COMP}$  voltage will be stable. So when a capacitor of large capacitance is connected to the COMP pin for filtering 100-HZ line ripple, a fixed on-time is generated during the half sinusoidal cycle, so a PF correction is implemented. A small capacitor will be connected to the COMP pin when it only works on flyback topology without PFC to maintain loop stability.



**Figure 3. Charge and Discharge Block Diagram**

The controller implements primary current feedback and regulation to maintain constant output LED current. Figure 4 shows a block diagram of the TPS92310 controller; the red virtual line presents a main control loop.

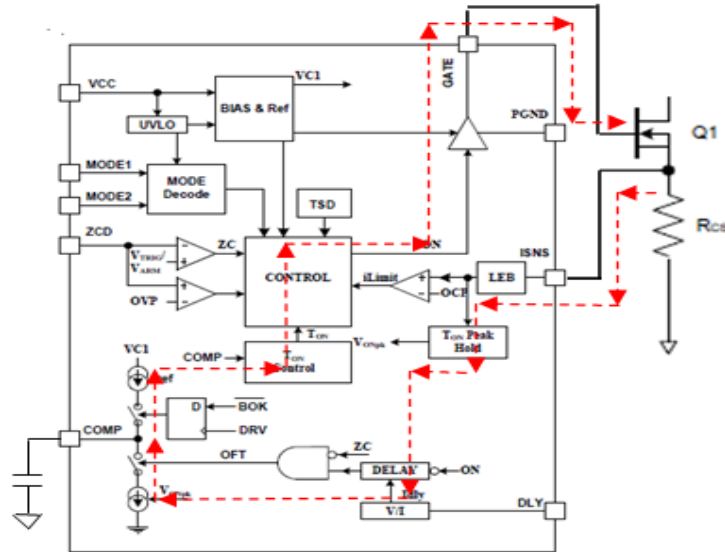


Figure 4. TPS92310 Block Diagram

#### 1.4 ZCD Detection, Delay Set and Output Overvoltage

The zero cross detect (ZCD) pin detects transformer auxiliary winding for zero current. When the ZCD voltage is lower than  $V_{ZCD}(TRIG)$  level, the internal RS flip-flop passes a ZCD signal to IDLY delay block to trigger the next switching cycle. The dual-level detect on this pin (ARM/TRIG) ensures the switching FET turns ON at zero current of the secondary side of the isolation transformer. Figure 5 shows a typical switching waveform on the DRAIN of the switching FET. The controller also offers a blank time of 300 ns for ZCD detection to avoid any possible ringing impact.

To reduce EMI and switching loss during converter operation, a DLY pin of the TPS92310 controller is provided. Connecting an external resistor can control the delay timer easily. Through the IDLY pin, the converter ensures the transformer winding has zero current instead of the main switching FET turning ON. The preset delay timer value must be considered against the resonance frequency between the primary inductance of the isolation transformer and the switching FET drain charge. The  $T_{dly}$  can be calculated by the following equation.

$$T_{dly} = \frac{1}{2} \times \pi \times \sqrt{L_p \times C_{oss}} \quad (2)$$

Where:

$L_p$  = transformer primary inductance

$C_{oss}$  = MOSFET output capacitance

$T_{dly}$  is used to control discharge time of  $V_{COMP}$ , so it must be set by external resistor connected on DLY pin, as shown in Figure 6.

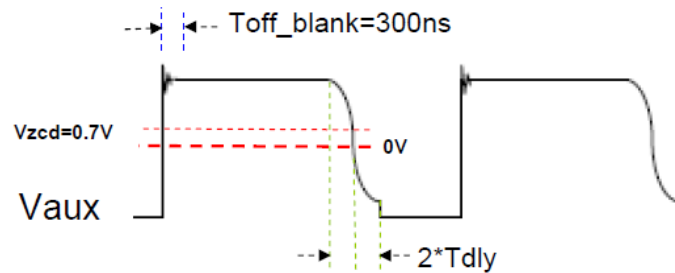


Figure 5. Typical Switching Waveform

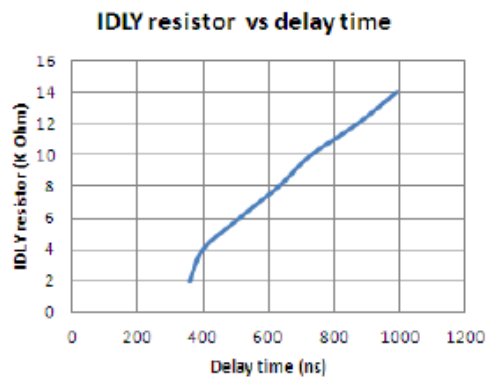


Figure 6.  $T_{dly}$  Set Curve

The ZCD pin is also used as output overvoltage protection. The positive voltage on auxiliary winding presents an output LED voltage and can be detected by external divided resistor, as shown in Figure 7. The overvoltage on the ZCD pin exceeds the OVP threshold by three cycles. The drive output should be shut down, and the controller implements a restart mode. The OVP voltage can be calculated by the following equation.

$$V_{O\_OVP} = \frac{N_s}{N_a} \times \frac{V_{OVP}}{R_L} \times (R_U + R_L) - V_D \quad (3)$$

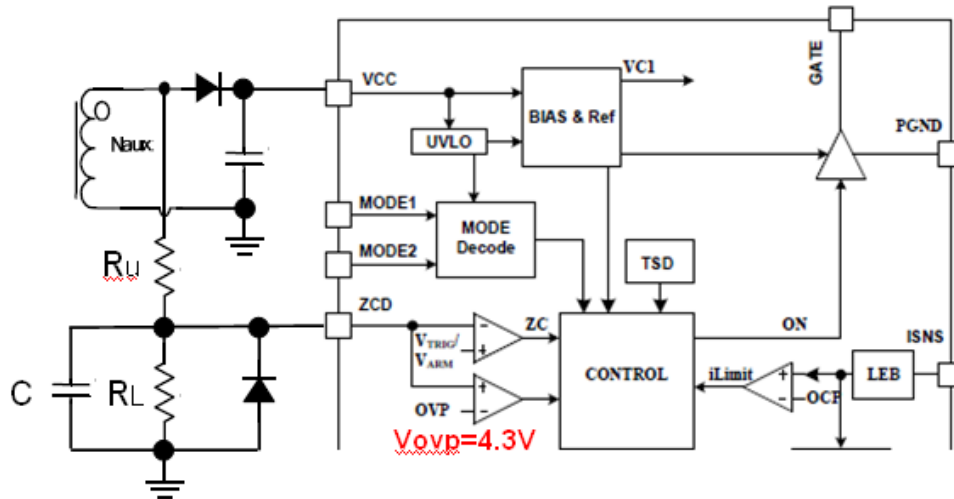
Where:

$N_s$  = turns of auxiliary winding

$N_a$  = turns of output winding

$V_D$  = forward voltage of output rectifier

The negative voltage on auxiliary winding represents a reflected voltage of input voltage, therefore power dissipation on the resistor is considered when selecting  $R_U$ . A current flow of 0.2 to 0.5 mA is suitable. A diode is connected to the ZCD pin to clamp the negative voltage below 1 V. A small capacitor, C, is always connected between the ZCD pin and GND to eliminate possible ringing impact, ensure accurate OVP, and obtain a suitable valley switching on.



**Figure 7. ZCD Pin Connect Circuit**

## 1.5 Output Short-circuit Protection

The TPS92310 controller operates at voltage mode control, and a cycle-by-cycle limit is needed for OCP and SCP. The controller provides two constant on-time modes with different OCP thresholds (0.64 V and 3.4 V) in this isolated flyback. A sensing voltage of primary current can be calculated with the following equation:

$$V_{isns} = \frac{2 \times REF}{\eta} \times \frac{V_{LED}}{V_{LED} + V_D} \times \frac{V_{in\_min} + V_{or}}{V_{in\_min}} \quad (4)$$

Where:

REF = 0.14 by controller

$V_{LED} = 12$  V

$V_D = 0.8$  V

$V_{in\_min} = 127$  Vdc

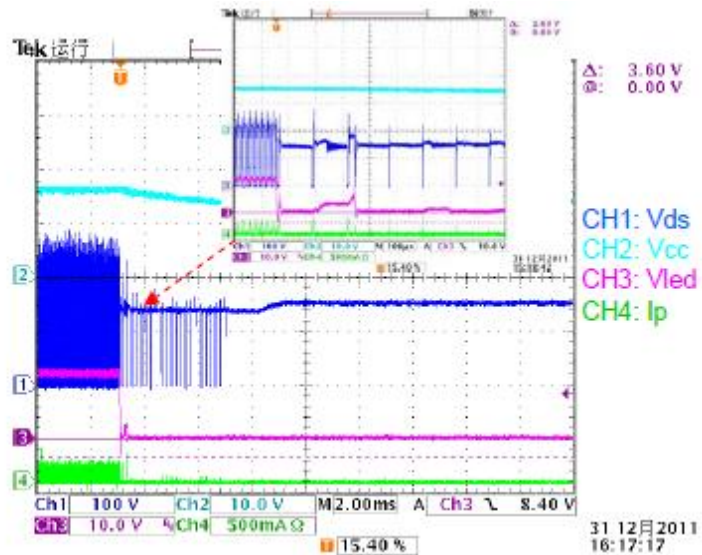
$V_{or} =$  about 85 V in this design, equals  $N_x(V_{LED} + V_D)$

$\eta$  = efficiency is evaluated about 0.8 at low line

For this conventional flyback design,  $V_{isns}$  is about 0.53 V.



Because of fixed  $V_{in\_min}$  voltage and almost fixed  $V_{or}$  designing voltage, the  $V_{isns}$  is almost constant when there is a different LED voltage. This detecting voltage is lower than the OCP threshold, so constant on-time mode with 0.64 V OCP threshold can be configured for perfect output short-circuit protection. This mode can be used in all conventional flyback designs. To avoid the ringing interference for ZCD detection during output short-circuits, a small capacitor must be connected between the ZCD pin and GND to avoid false ZCD detection; a 10-pF capacitor is suitable in this design. Figure 8 show the output short-circuit waveform.

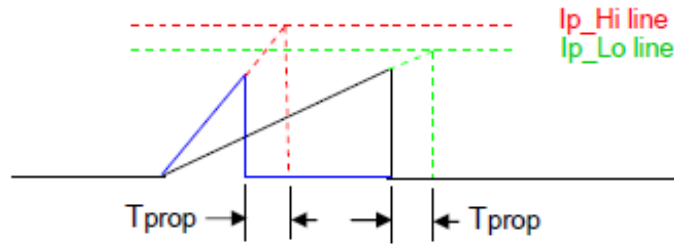


**Figure 8. Output Short-circuit Protect (SCP) Waveform**

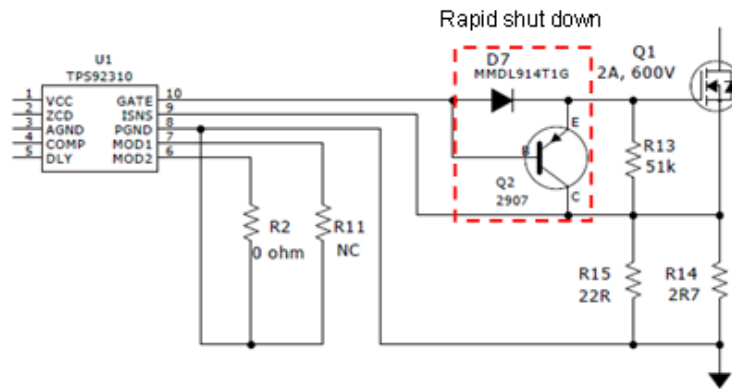
### 1.6 External Line Regulation Compensation

There is a different peak current at high line and low line because of controller inherent propagation delay, as shown in Figure 9. The high-line input voltage results in a much higher current difference than does the low line input voltage with the same propagation delay. According to Equation 1, the error on the input current sensing impacts LED current, and line regulation is not very well. There are two ways to improve line regulation when input voltage is changed from low line to high line:

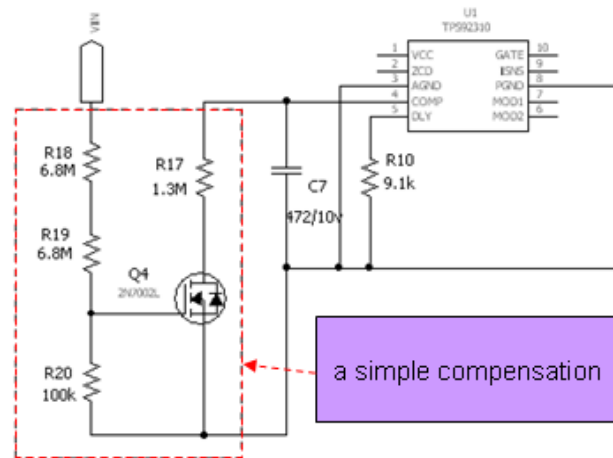
1. Add a fast shut-down circuit (shown in Figure 10). It can reduce the MOSFETs turning on delay and improve the current tolerance of 5 mA at 230 Vac in this design.
2. Add an input voltage detection circuit (shown in Figure 11) to reduce on-time at high line; it can reach the desired high current precision by adjusting R17 at 110 Vac and 230 Vac line. R19, R19, and R20 decide the inflection point of LED current. Figure 12 shows a line regulation ratio curve with external compensation.



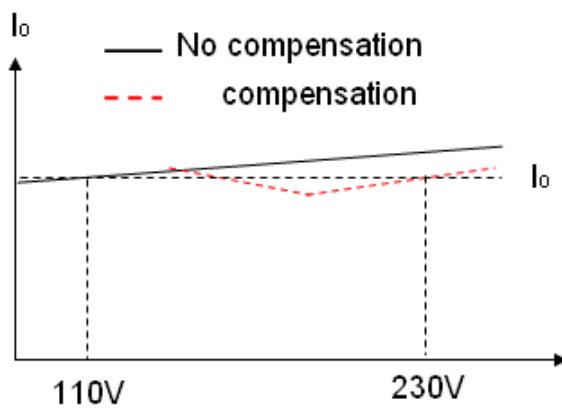
**Figure 9. Inherent Propagation Delay**



**Figure 10. Rapid Shut-down Circuit**



**Figure 11. External Line Regulation Compensation Circuit**



**Figure 12. Line Regulation Compensation Curve**

## 2 Transformer Design

According to the previous description, to use an external SCP circuit, the  $V_{isns}$  must be set below 0.6 V.

$$V_{isns} = R_{CS} \times I_P \quad (5)$$

$$R_{CS} = \frac{N \times 0.14}{I_{LED}} \quad (6)$$

$$\text{So } I_P = \frac{V_{isns} \times I_{LED}}{0.14 \times N} \quad (7)$$

$$D_{MAX} = \frac{V_{or}}{V_{in\_min} + V_{or}} = \frac{N \times (V_{LED} + V_D)}{V_{in\_min} + N \times (V_{LED} + V_D)} \quad (8)$$

$$I_P = \frac{2 \times I_{av}}{D_{MAX}} = \frac{2 \times P_O}{\eta \times V_{in\_min} \times D_{MAX}} = \frac{2 \times V_{LED} \times I_{LED}}{\eta \times V_{in\_min} \times D_{MAX}} \quad (9)$$

Combining equations 7, 8, and 9, yields

$$N = \frac{\eta \times V_{isns} \times V_{in\_min}}{0.28 \times V_{LED}} - \frac{V_{in\_min}}{V_{LED} + V_D} \quad (10)$$

Where:

$V_{isns}$  = sensing voltage of primary current (set below 0.6 V if using external SCP circuit, otherwise no limit)

$R_{CS}$  = current sense resistor

$N$  = transformer turn ratio of primary winding to output winding

$I_P$  = primary peak current

$V_{or}$  = primary reflection voltage of secondary voltage

$I_{LED}$  = LED current

$V_{LED}$  = LED voltage

$\eta$  = evaluated power supply efficiency

$V_D$  = forward voltage of output rectifier

$V_{in\_min}$  = minimum input DC voltage, usually simple to 1.3  $V_{ac\_min}$

The transformer specification can be calculated as the following equations:

$$L_P = \frac{V_{in\_min} \times D_{MAX}}{I_P \times F_{S\_MIN}} \quad (11)$$

$$N_P = \frac{L_P \times I_P}{\Delta B_{MAX} \times A_e} \quad (12)$$

$$N_{out} = \frac{N_P}{N} \quad (13)$$

$$N_{aux} = \frac{N \times (V_{aux} + V_{D\_aux})}{V_{LED} + V_D} \quad (14)$$

Where:

$L_p$  = primary inductance

$N_p$  = the turns of primary winding

$N_{out}$  = the turns of output winding

$N_{aux}$  = the turns of auxiliary winding, usually less than calculated value because of influence of spike voltage

$D_{MAX}$  = maximum duty cycle, calculated by equation 2

$F_{S\_MIN}$  = setting minimum switching frequency at low line

$\Delta B_{MAX}$  = selecting maximum working flux density

$A_e$  = effective core area

$V_{aux}$  = selecting Vcc voltage

$V_{D\_out}$  = forward voltage of auxiliary rectifier

Finally, RMS current and peak voltage on primary MOSFET can be selected and secondary rectifier for MOSFET and rectifier and build transformer can be selected according RMS current and bobbin windows.

### 3 Experimental Results

#### 3.1 Electrical Performance Specifications

Table 1. PMP4325 Electrical Performance Specifications

PARAMETER	TEST CONDITIONS	MIN	TYP	MAX	UNITS
<b>Input Characteristics</b>					
Voltage range		85		265	Vrms
<b>Output Characteristics</b>					
Output voltage, $V_{out}$	Output current = 350 mA	8	12	13	V
Output load current, $I_{out}$			350		mA
Output current ripple	$V_{out} = 12\text{ V}$ , $I_{out} = 350\text{ mA}$ , $V_{in} = 230\text{ Vac}$		105		mApp
<b>Systems Characteristics</b>					
Efficiency	Universal input, $V_{out} = 12\text{ V}$		78		%

#### 3.2 Reference Design Schematic

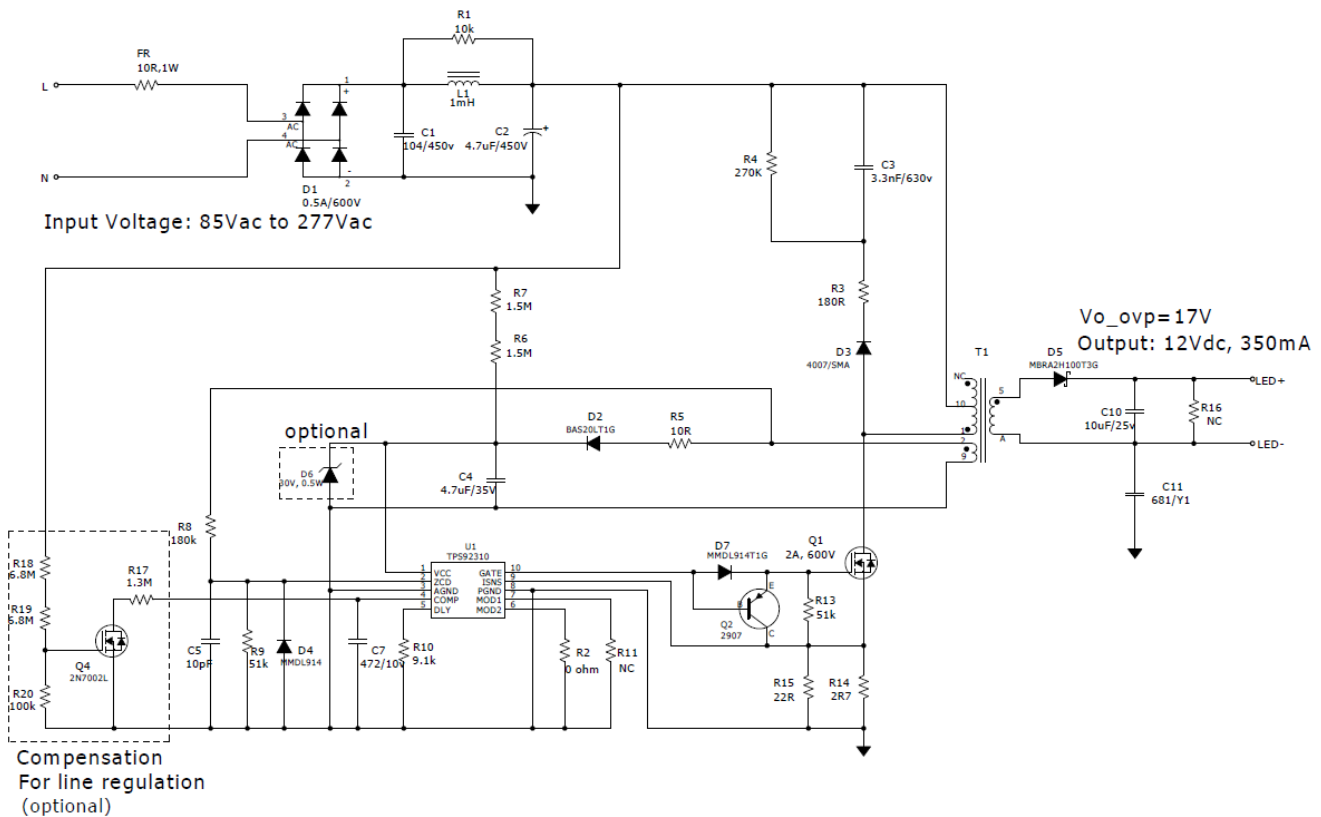


Figure 13. PMP4325 Reference Design Schematic

### 3.3 PMP4325 PCB Layout

This reference design is implemented on a double-sided PCB that is dimensionally compatible with the GU10 LED lamp and similar applications. To meet different requirements, two versions of PCB layout file are provided:

1. Released demo board without output SCP line regulation compensation circuit.
2. PCB file provided to some customers who requires a good SCP and line regulation functions.

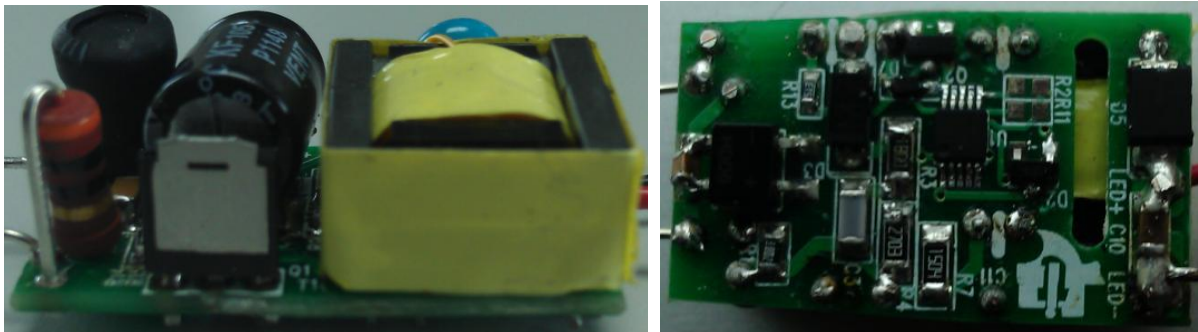


Figure 14. Component Side and Solder Side of Demo Board

#### 3.3.1 PCB Layout Without SCP and Line Regulation Compensation Circuit

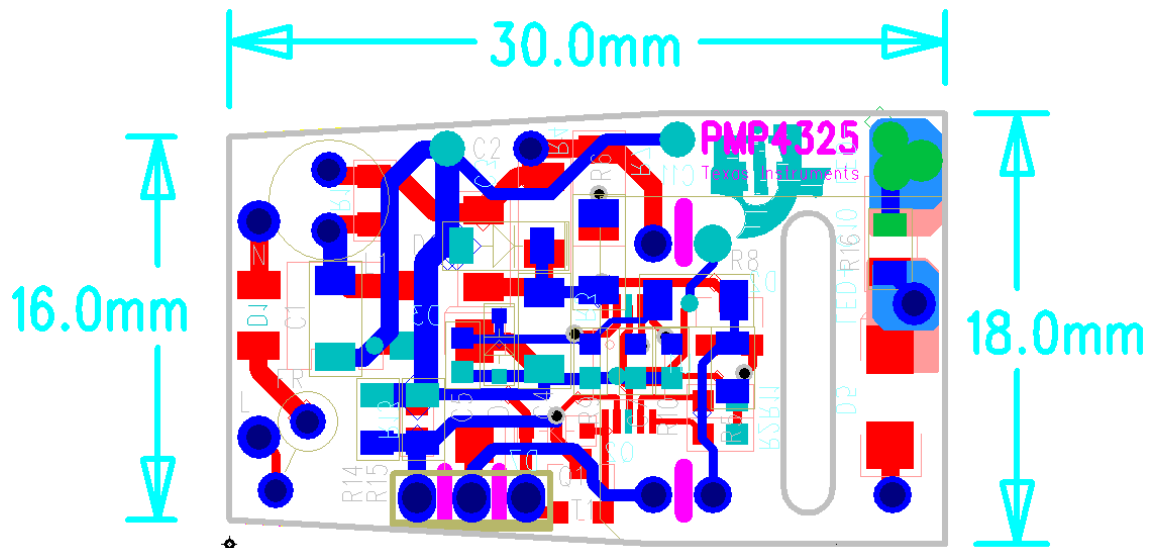


Figure 15. PCB Layout of Released Demo Board

### 3.4 Electrical Performance

Figure 16 through Figure 18 gives typical performance curves for the PMP4325 9-V and 12-V, 350-mA LED driver.

### 3.4.1 Efficiency Curve at 3- and 4-LED Applications

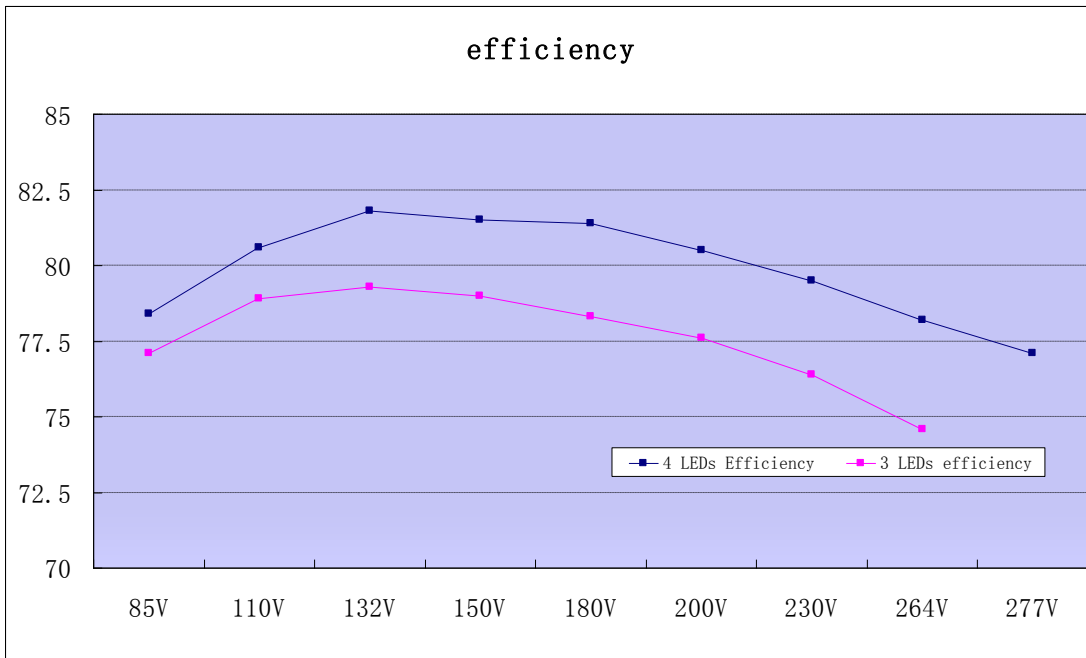


Figure 16. Efficiency Curve With 3- and 4-LED Loads

### 3.4.2 Line Regulation Curve

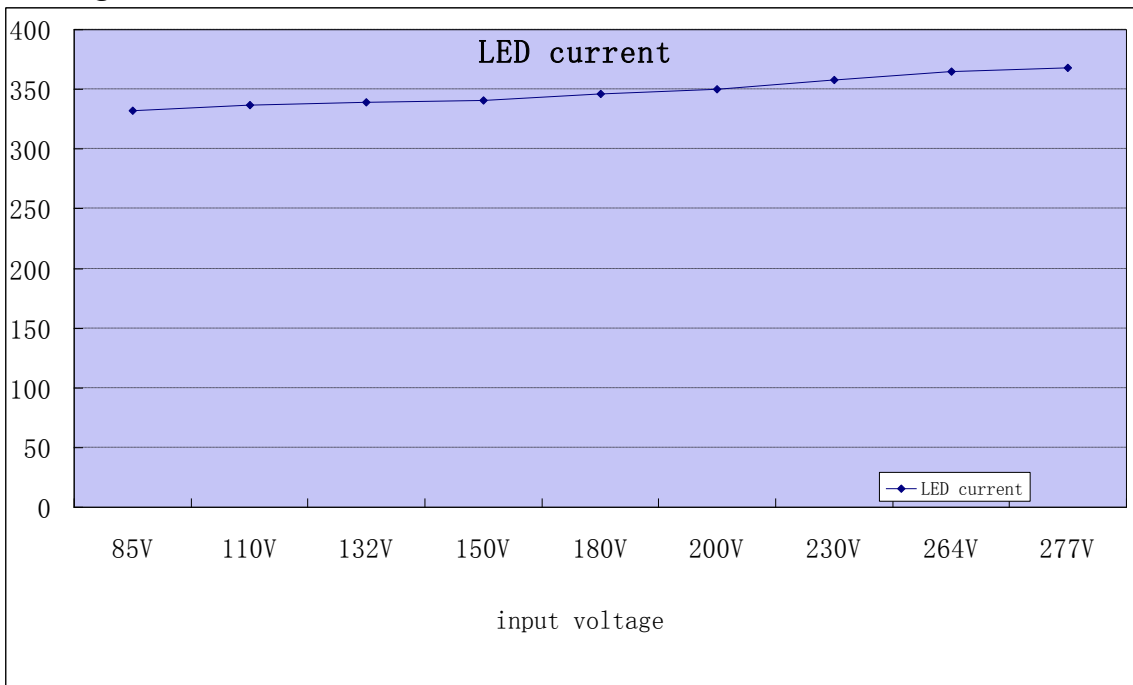


Figure 17. Line Regulation of LED Current



### 3.4.3 Line Regulation Curve With Compensation Circuit

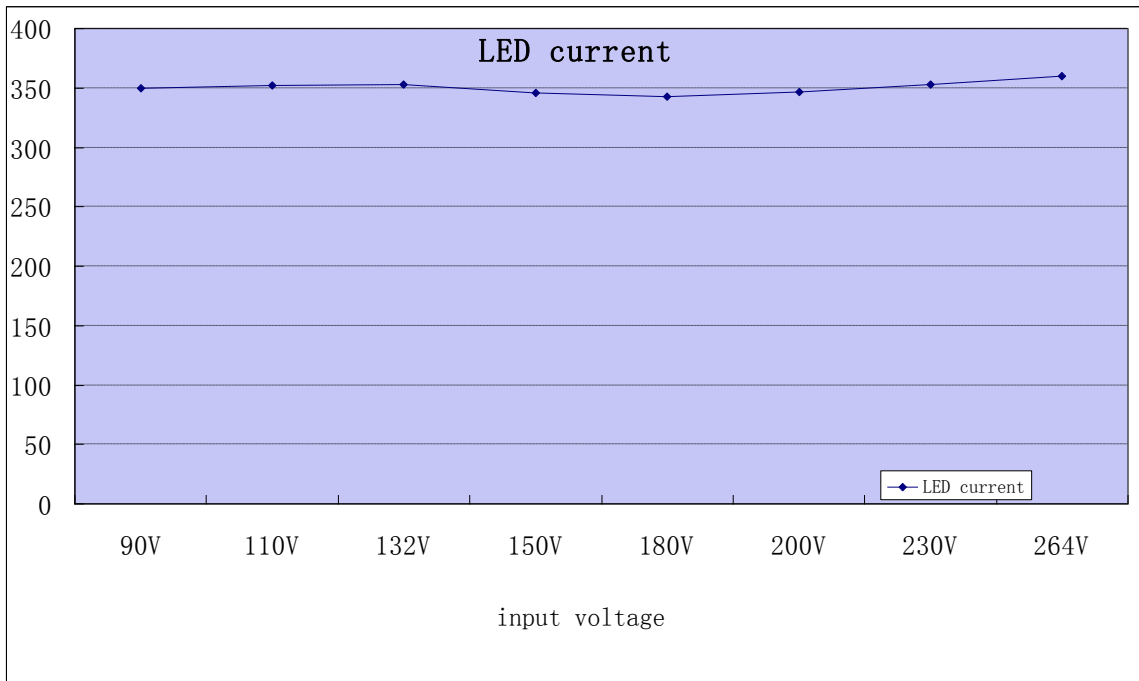
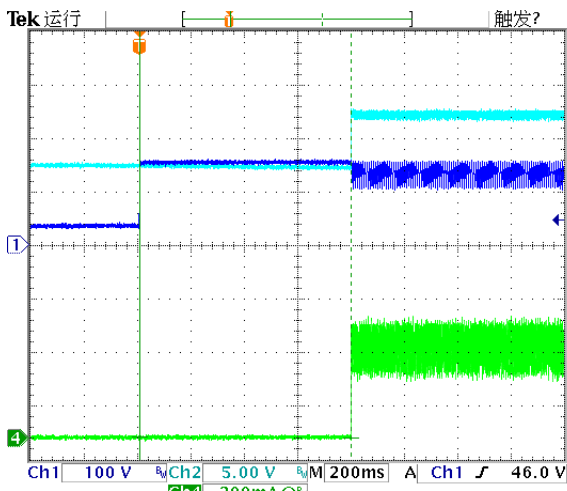


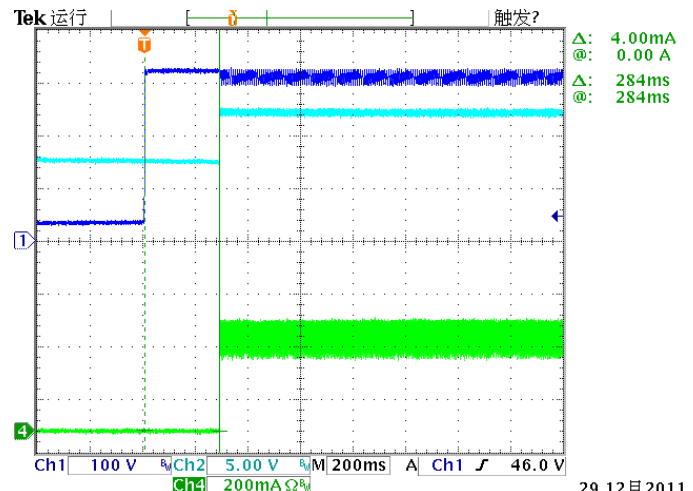
Figure 18. Line Regulation of LED Current With Compensation

### 3.4.4 Start-up Output Waveform



Ch1: Input bulk CAP voltage 100V/div  
Ch2: LED voltage, 5V/div  
Ch4: LED current 200mA/div

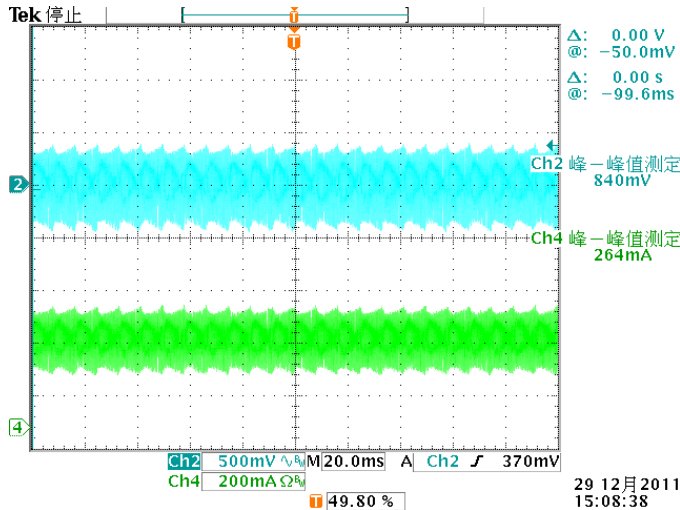
Figure 19. Start-up Test With 110 VAC



Ch1: Input bulk CAP voltage 100V/div  
Ch2: LED voltage, 5V/div  
Ch4: LED current 200mA/div

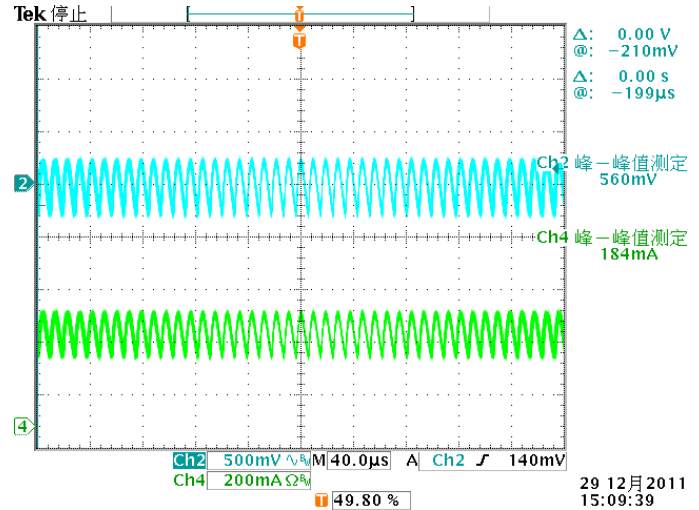
Figure 20. Start-up Test With 230 VAC

### 3.4.5 Output Ripple Voltage and Current



Ch2: LED ripple voltage 500mV/div  
Ch4: LED current 200mA/div

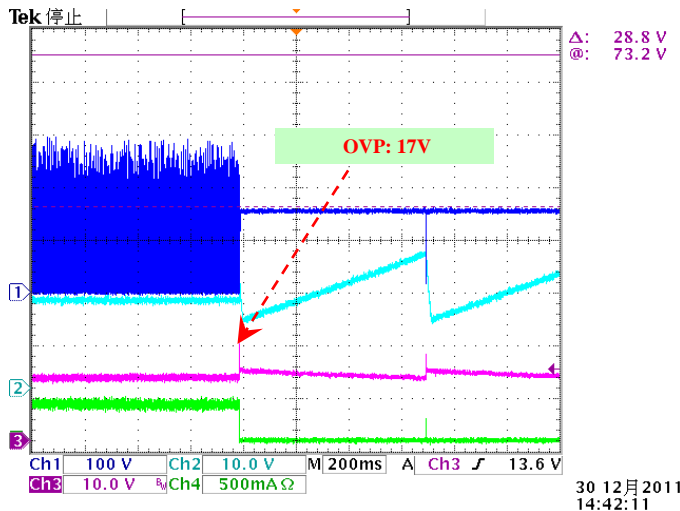
Figure 21. Output Ripple Test With 110 VAC



Ch2: LED ripple voltage 500mV/div  
Ch4: LED current 200mA/div

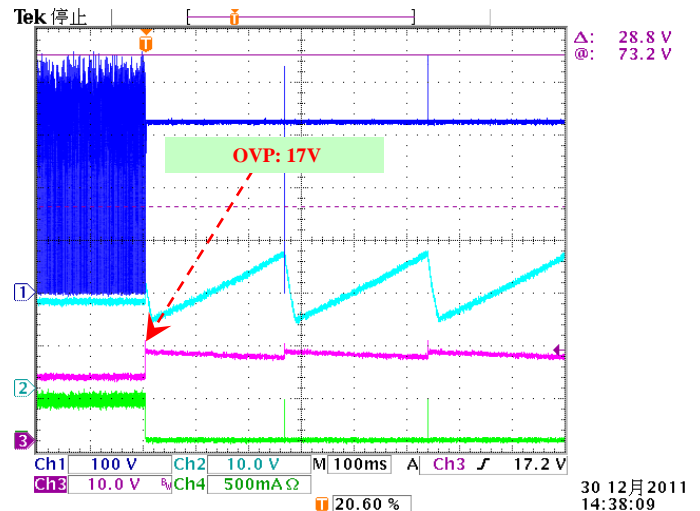
Figure 22. Output Ripple Test With 230 VAC

### 3.4.6 Output Overvoltage and Open LED Protection



Ch1: MOSFET Vds, 100V/div  
Ch2: Vcc, 10V/div  
Ch3: LED voltage, 10V/div  
Ch4: LED current, 500mA/div

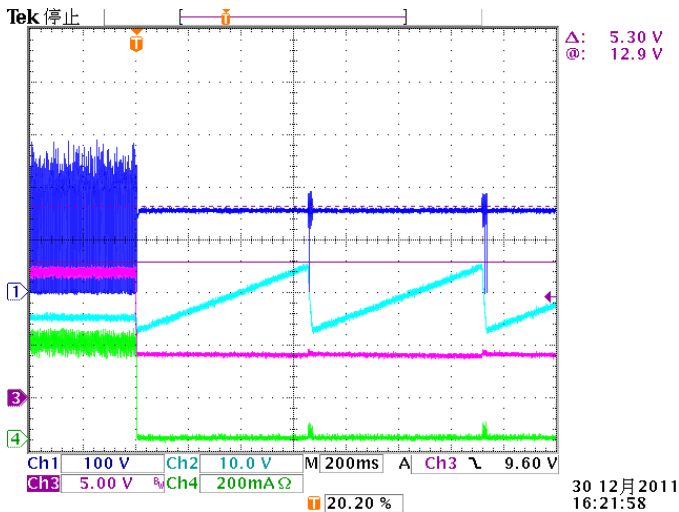
Figure 23. OVP Test With 110 VAC



Ch1: MOSFET Vds, 100V/div  
Ch2: Vcc, 10V/div  
Ch3: LED voltage, 10V/div  
Ch4: LED current, 500mA/div

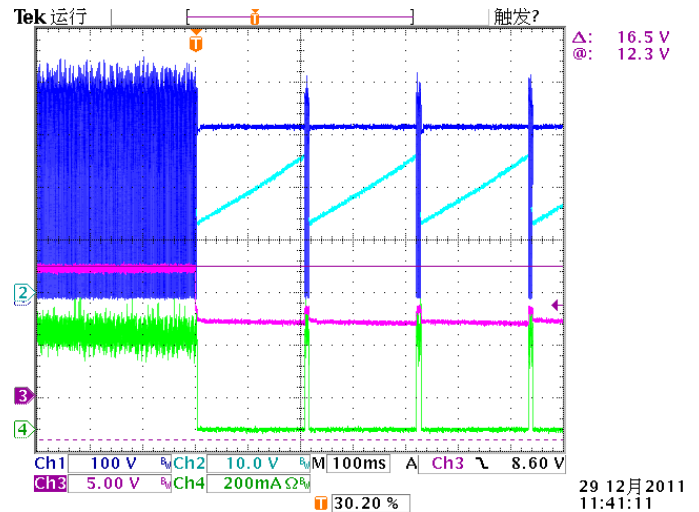
Figure 24. OVP Test With 230 VAC

### 3.4.7 Two-LED Protection



Ch1: MOSFET Vds, 100V/div  
Ch2: Vcc, 10V/div  
Ch3: LED voltage, 5V/div  
Ch4: LED current, 200mA/div

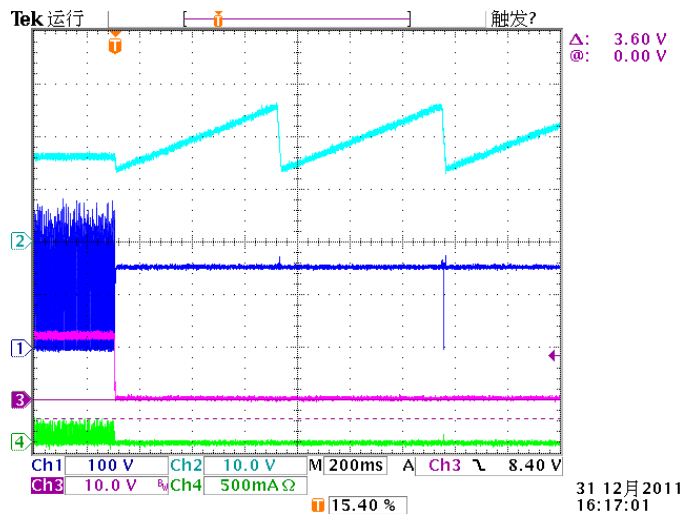
**Figure 25. Shorting Two LEDs Test With 110 VAC**



Ch1: MOSFET Vds, 100V/div  
Ch2: Vcc, 10V/div  
Ch3: LED voltage, 5V/div  
Ch4: LED current, 200mA/div

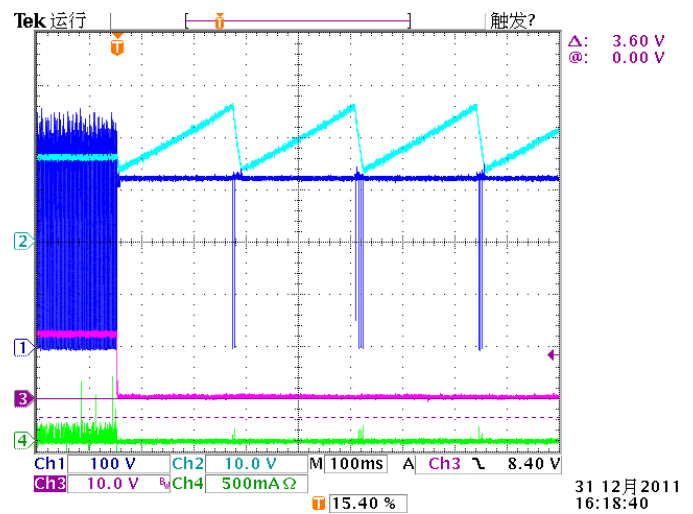
**Figure 26. Shorting Two LEDs Test With 230 VAC**

### 3.4.8 Output Short-circuit Protection



Ch1: MOSFET Vds, 100V/div  
Ch2: Vcc, 10V/div  
Ch3: LED voltage, 5V/div  
Ch4: LED current, 200mA/div

**Figure 27. Output Short-circuit Test With 110 VAC**



Ch1: MOSFET Vds, 100V/div  
Ch2: Vcc, 10V/div  
Ch3: LED voltage, 5V/div  
Ch4: LED current, 200mA/div

**Figure 28. Output Short-circuit Test With 230 VAC**

### 3.5 Conducted EMI

#### 3.5.1 EMI for 4-LED GU10 Load With Y CAP

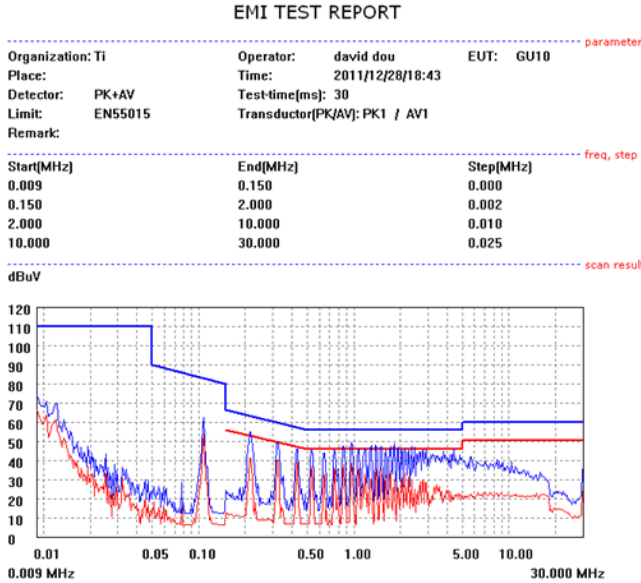


Figure 29. Conducted EMI With 230 VAC, Line

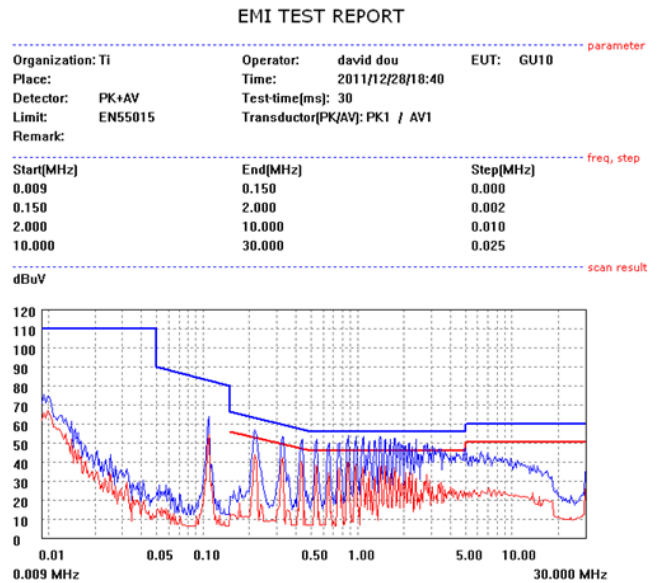


Figure 30. Conducted EMI With 230 VAC, Neutral

#### 3.5.2 EMI for 3-LED GU10 Load With Y CAP

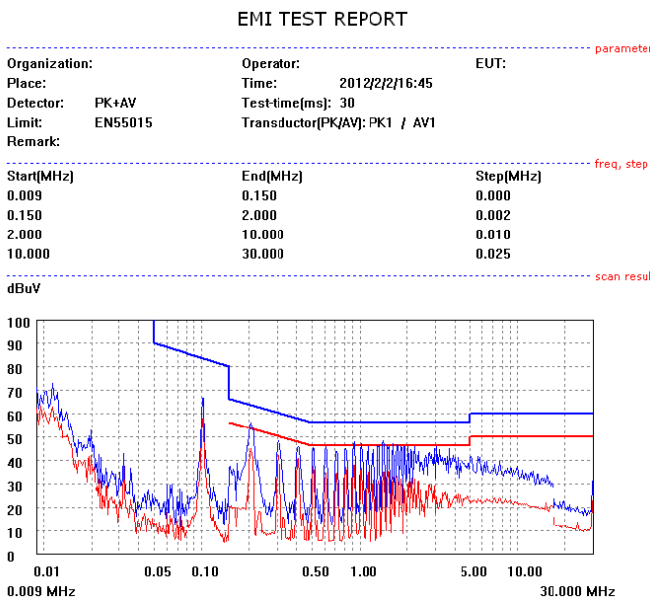


Figure 31. Conducted EMI With 230 VAC, Line

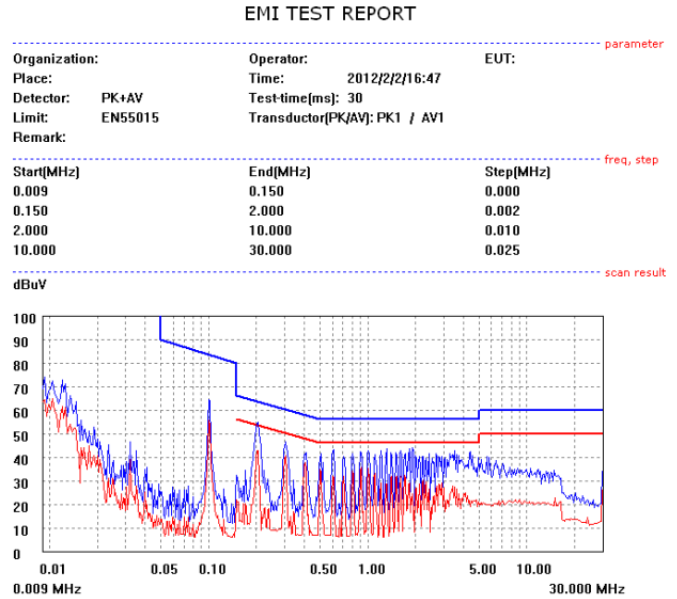
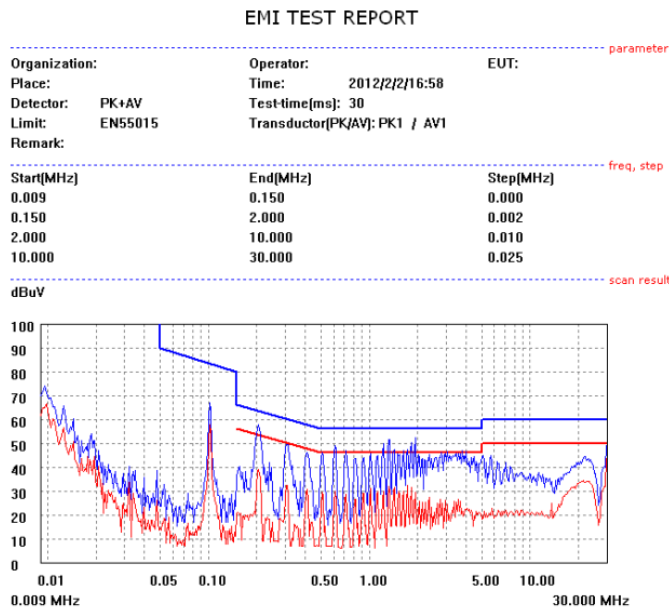
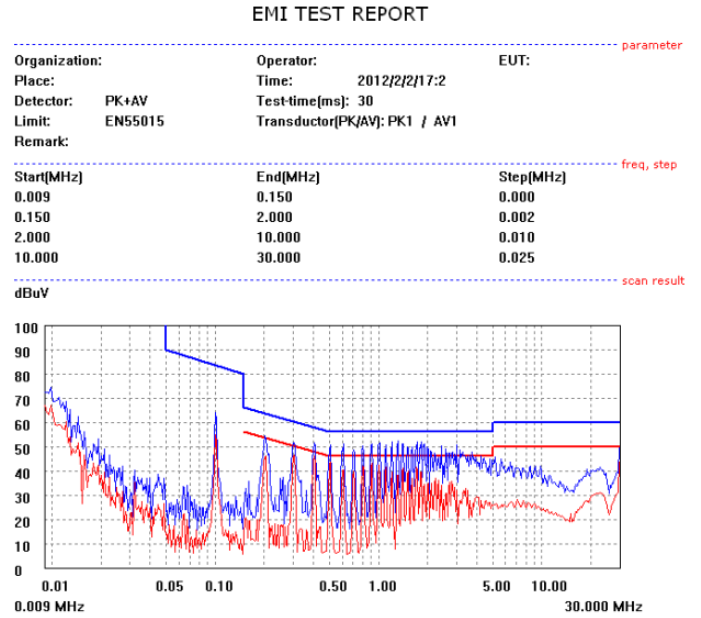


Figure 32. Conducted EMI With 230 VAC, Neutral

### 3.5.3 EMI for 3-LED GU10 Load Without Y CAP



**Figure 33. Conducted EMI With 230 VAC, Line**



**Figure 34. Conducted EMI With 230 VAC, Neutral**

### 3.6 Bill of Materials

**Table 2. PMP4325 Bill of Materials**

QTY	RefDes	Description	MFR	Part Number
1	D2	Switching Diode, 0.2A,/200V, SOT23	ON Semi	BAS20LT1G
1	C4	Capacitor, chip,4.7 uf, 35 V,1206	TDK	C3216X7R1V475K
1	C5	Capacitor, chip,10 pF/10 V, 10%, 0603	Std	Std
1	C7	Capacitor, chip, 4700 pF/10 V, 10%, 0603	Std	Std
1	C10	Capacitor, chip, 10uF/25 V , 10%, 1206	TDK	C3216X7R1E106M
1	C3	Capacitor, chip,3300 pF/ 630 V, 10%, 1206	TDK	C3216X7R2J332K
1	D1	Bridge Rectifier, 600 V, 0.5 A, SMD	FSC/others	MB6S
1	D5	Schottky Diode, 2 A, 100 V, SMA	ON/Diodes	MBRA2H100G
1	D3	Standard Rectifier, 1 A, 1000 V, SMA	ON/Diodes	4007/SMA
1	L1	Inductor, 1mH,0.2A, 6mm 8.5 mm	LEDpower	Custom
1	C11	Y1 CAP,680pF, Optional	Std	Std
1	T1	EPC13, 10pin Horizontal bobbin	LEDpower	Custom
1	FR	Fusible Resistor, 10R, 1W	Vitrohm	CRF251-4
1	R2	Resistor, chip, 0 ohm, 0603	Std	Std
1	R2	Resistor, chip, 1/8W, 5%, 0603	Std	Std
1	R8	Resistor, chip, 1/4W, 1%, 1206	Std	Std
1	R9	Resistor, chip, 1/8W, 1%, 0603	Std	Std
1	R10	Resistor, chip, 1/8W, 1%, 0603	Std	Std
2	R5	Resistor, chip, 1/5W, 5%, 0805	Std	Std
1	R13	Resistor, chip, 1/8W, 5%, 0603	Std	Std
1	R14	Resistor, chip, 1/5W, 1%, 0805	Std	Std
1	R15	Resistor, chip, 1/5W, 1%, 0805	Std	Std
1	R1	Resistor, chip, 1/5W, 5%, 0805	Std	Std
1	R3	Resistor, chip, 1/4W, 5%, 1206	Std	Std
2	R6 R7	Resistor, chip, 1/4W, 5%, 1206	Std	Std
1	R16	Resistor, chip, 1/5W, 5%, 0805	Std	Std
1	R4	Resistor, chip, 1/4W, 5%, 1206	Std	Std
1	C2	ECAP, 4.7 uF/400 V, 8 mm x 11.5 mm, 105° C, 5000h	Capxon	KW series
1	C1	Capacitor, chip, 0.1 uF/450 V, 1206	TDK	C3216X7T2W104M
1	D4,D7	Switching diode, 0.2 A, 100 V, SOD323	ON SemI	MMDL914T1G
1	D6	Zener, 30 V,0.5 W,SOD-123, Optional	Std	Std
1	U1	Primary current sense controller, MSOP8	TI	TPS92310
1	Q1	MOSFET, N-ch, 600 V, 2 A, IPAK	ON/FSC/Infineon	2N60
1	Q2	PNP transistor, -0.6 A, -60 V, sot23	Std	Std
1	R17	Resistor, chip,1/8w,1%,1.3M,1206	Std	Std
2	R18,R19	Resistor, chip,1/4W,1%,.6.8M,1206	Std	Std
1	R20	Resistor, chip,1/8w,1%,100K,0603	Std	Std
1	Q4	N MOSFET, 0.115A, 60 V, SOT23	Std	2N7002L

### 3.7 Transformer specification

This section contains transformer specifications for the core and bobbin, the electrical diagram, the electrical specification, and the building construction diagram.

CORE: EPC13

Core material: PC40 or equivalent

Bobbin: 10Pin horizontal bobbin, size is as following:

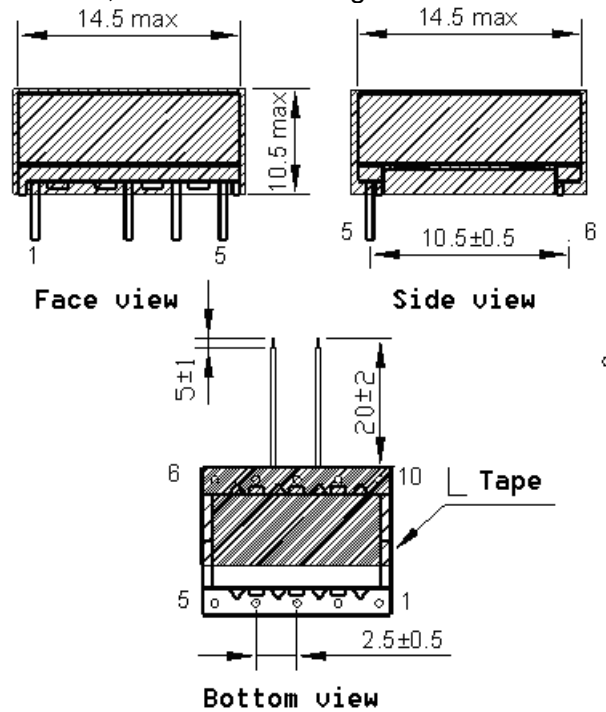


Figure 35. 10-Pin Horizontal Bobbin

### EPC13-10PIN

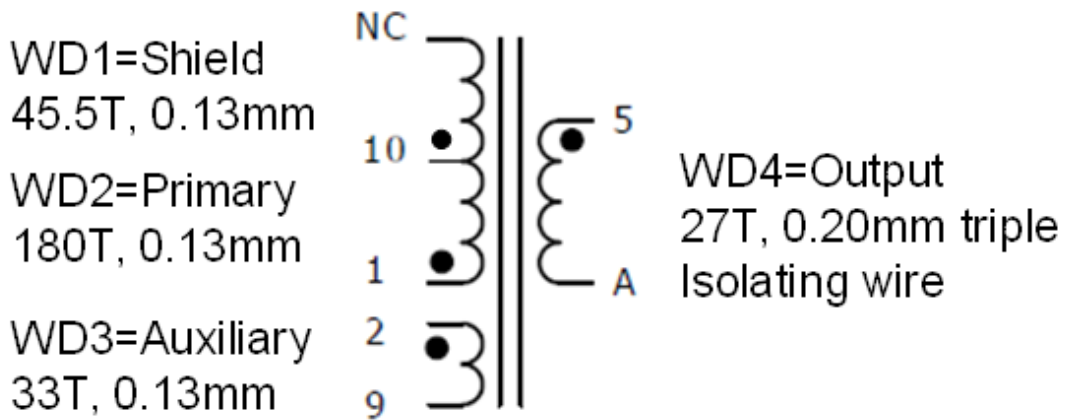


Figure 36. Transformer Electrical Diagram

Table 3. Transformer Electrical Specification

Electrical Strength	1 second, 60 Hz, from pins 1, 2, 9, 10 to 5, A	3000 V
Primary Inductance	Pins 1-10, all other windings open, measured at 10 kHz, 1V	2.6 MHz+/-10%

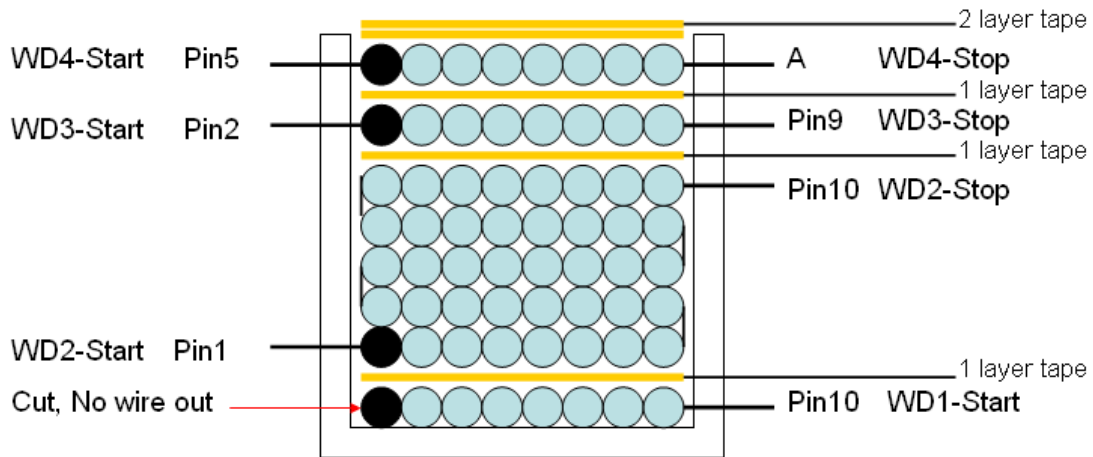


Figure 37. Transformer Building Construction Diagram

### References

1. [TPS92310](#) Off-Line Primary Side Sensing Controller with PFC datasheet, Texas Instruments



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