International Rectifier

APPLICATION NOTE

Application Note AN-1195

Primary Side Regulated LED Driver using the IRS2983

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1. Introduction

The IRS2983 SO-8 LED controller IC is an excellent solution for controlling Flyback circuits in LED applications. The IC features primary side power regulation, which in a single stage isolated Flyback or Buck-Boost converter with a fixed load enables a universal input voltage design to be realized with minimal component count, low cost and high reliability. The IC includes all necessary circuitry to control the Flyback converter on and off times, regulate the input power, provide AC current at the mains input for high PF and low THD and protect against over-current and over-voltage fault conditions. The IC also features a high voltage startup pin, eliminating the need for a VCC startup resistor and reducing overall power dissipation. The IRS2983 may also be used with secondary side regulation utilizing an opto-isolator, to supply an output with a wide voltage range. This is discussed in AN-1192. During the design of the Flyback circuit special care should be taken when generating the circuit schematic, selecting component values and ratings, constructing the Flyback inductor and laying out the PCB. This application note provides detailed design information to reduce design time and avoid circuit problems that can occur due to incorrect component value or rating selection, incorrect programming of IC parameters, and noise susceptibility. Helpful information is included for designing the control circuit, designing the IC supply circuitry and Flyback inductor, and using the IC protection features. Surge and EMI considerations are presented and PCB layout guidelines are also included to help avoid potential noise problems that could cause malfunction or poor performance.

2. Flyback Converter Overview

The Flyback converter is derived from the Buck-Boost converter with the addition of isolation using an inductor with mutually coupled coils. The simplified Flyback circuit (Figure 1) includes the Flyback inductor (LFB) with mutually-coupled coil windings (Np, Na and Ns), the MOSFET switch (MFB) and the output diode (DFB). The inductor winding polarities are such that during the off-time of the switch, the primary current transfers to the secondary winding (Ns) to maintain the same magnetic flux in the core in the same direction and the current flows through the output diode charging the output capacitor (COUT) and driving the DC output. When the magnetic flux (ϕ_m) in the core reaches zero it is detected via an auxiliary winding (Na) on the inductor and the switch is turned on again linearly increasing the input current and core magnetizing flux. During this interval the DC output is entirely supplied by the energy stored in the output capacitor. The cycle then repeats (Figure 2). Since the converter is operating in critical conduction mode (CrCM) the switching cycle occurs at the zero crossings of the inductor current.

Figure 3 shows idealized switching waveforms of the MOSFET and the Auxiliary and Secondary winding voltages. Note that the Auxiliary and Secondary voltages are in phase with each other due to the orientation of the windings and that they also exhibit a voltage swing below ground during turn-on of the primary switch. Care should be taken when selecting the blocking diode voltage rating for the Auxiliary and Secondary windings to account for this behavior.

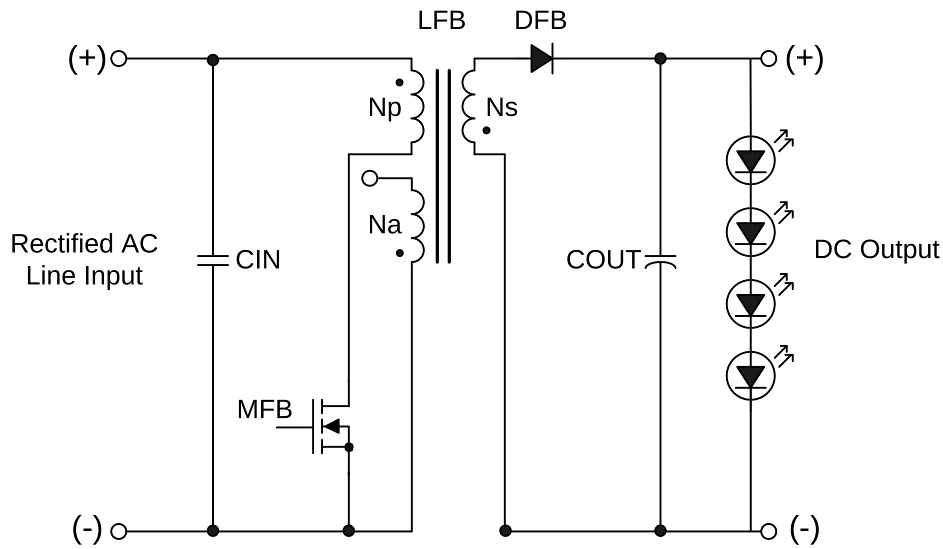


Figure 1: A Simplified Flyback Circuit.

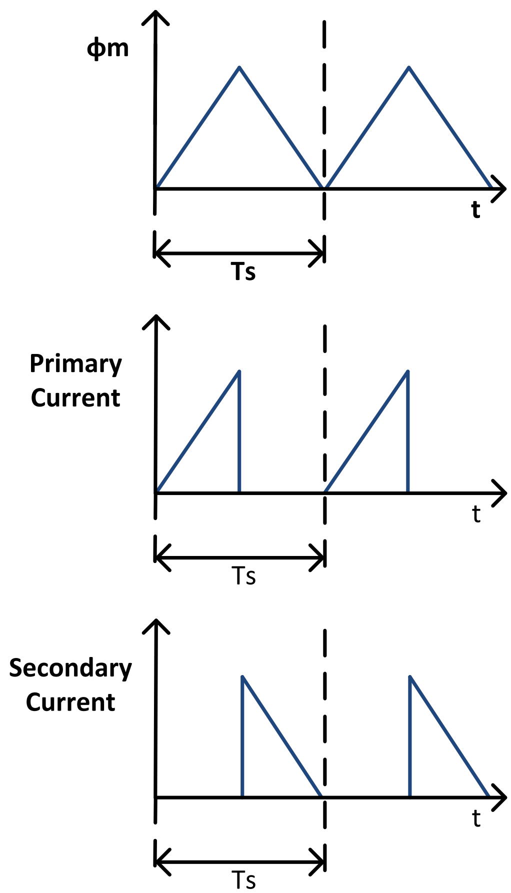


Figure 2: Idealized Flyback inductor Waveforms in CrCM

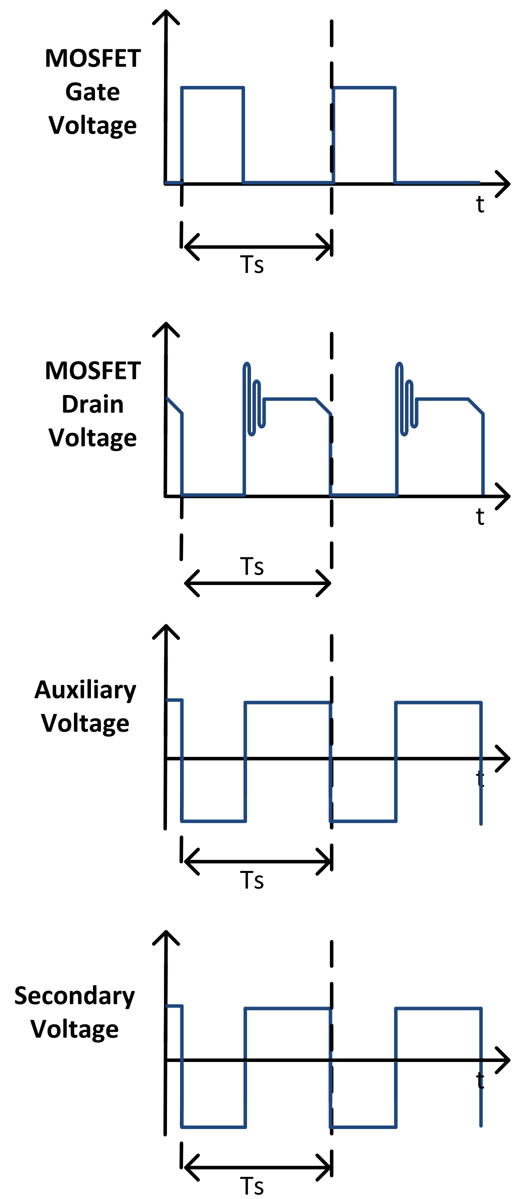


Figure 3: Idealized Switching Waveforms in CrCM

Operating the Flyback converter without a bulk input capacitor (only a high frequency capacitor is used) allows the input current to form a triangular-shape that follows the envelope of the rectified AC line voltage (Figure 4). This triangular current is smoothed by the input filter components to produce a sinusoidal input current that follows the shape of the AC line voltage, which inherently produces a high PF and low THD.

The switching frequency of the converter varies across each half-cycle of the input voltage with the maximum switching frequency occurring near the zero-crossing of the rectified AC line voltage and the minimum switching frequency occurring at the peak. The closed loop maintains a constant on-time while the off-time of the converter varies to maintain CrCM operation.

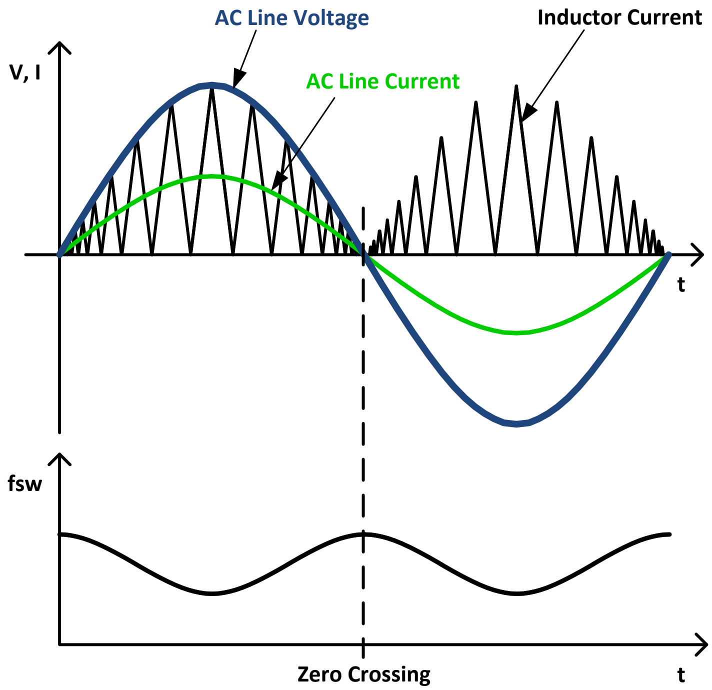


Figure 4: Flyback inductor and smoothed AC line current during one cycle of the AC line voltage

3. A Single-Stage Isolated Flyback Circuit with the IRS2983

The IRS2983 includes a primary side power control circuit for isolated Flyback converters that operate in critical conduction (CrCM) or discontinuous (DCM) modes. For this application note critical conduction mode will be used as the primary regulation method. Figure 5 below shows a typical application circuit schematic. This circuit includes the input voltage sensing components (RIN, RDC and VDC), the primary side current sensing components (RCS, RF, and CF), the zero-crossing detection circuit (RZX1 and RZX2), the VCC supply circuit (DVCC, RVCC, DZ and CVCC), the MOSFET snubber circuit (RSN, CSN and DSN), the Flyback inductor (LFB), the primary switch circuit (MFB and RG) and the output components (DFB, COUT and ROUT).

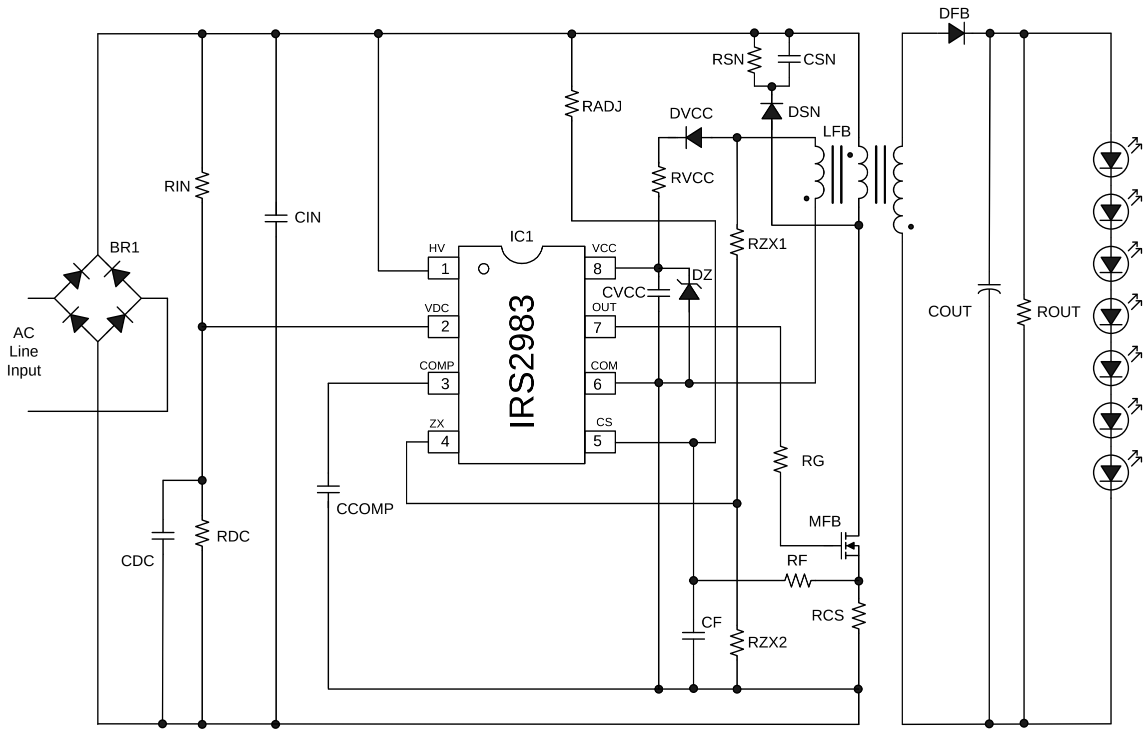


Figure 5: Typical Application Circuit Schematic

The IRS2983 includes a High Voltage regulator at the HV pin which enables rapid startup and reduces power consumption by eliminating the need for startup resistors. In order for the IRS2983 to begin switching the VCC supply voltage needs to be raised above its under voltage lockout level (UVLO). On startup the HV regulator supplies the micro-power startup current to VCC until UVLO is reached. At this point the OUT pin starts switching and driving the external MOSFET. An auxiliary winding from the Flyback inductor provides a voltage which can be used to supply VCC after the startup phase and the HV regulator shuts off. The auxiliary winding is also used to feed the IC zero-crossing information to the ZX pin. Regulation mode begins when the internal multiplier condition is satisfied and the loop is closed, assuming no current limit or fault conditions have occurred. This sequence is shown in the state diagram for the IRS2983 in Figure 6 below.

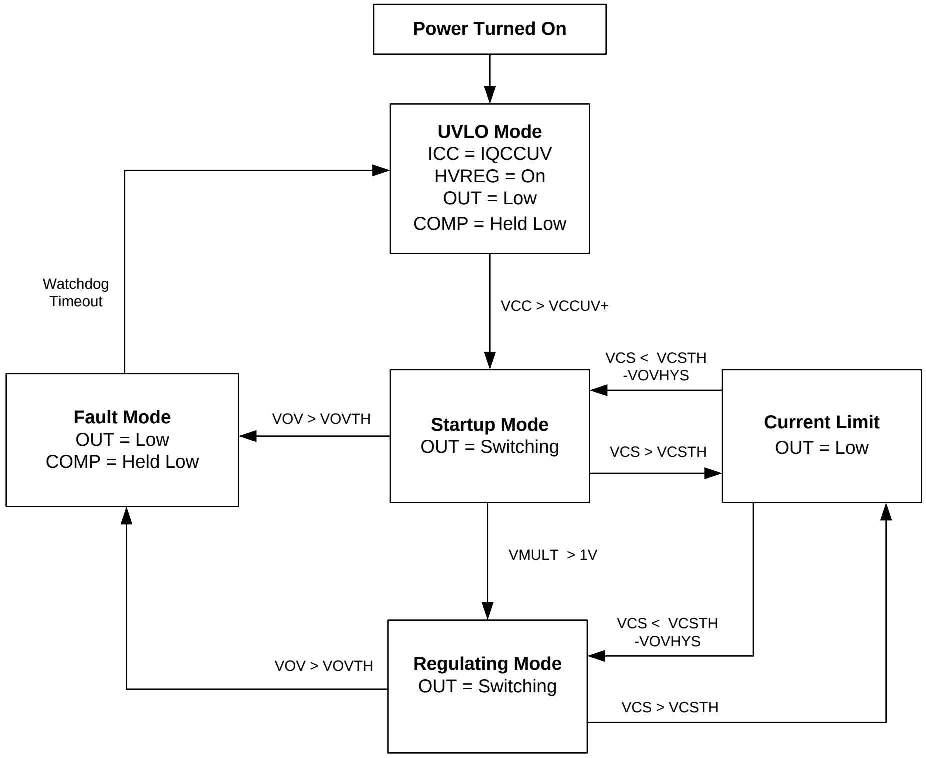


Figure 6: IRS2983 State Diagram.

In order to eliminate opto-isolators and other components necessary for isolated feedback, the IRS2893 is capable of regulating the LED output current indirectly by calculating and controlling the input power of the Flyback converter. Since the LED load voltage is essentially constant for a fixed load the power consumed is approximately proportional to the current. The IRS2983 senses the input voltage and current and uses these quantities to calculate the input power. This is then regulated against an accurate fixed reference to provide a regulation of the LED current within +/-5% over line voltage variation from 120 to 230VAC. The loop consists of the input voltage sensing pin VDC and input current sensing pin CS. The input voltage sensing is done via a simple resistor divider producing a voltage proportional to the rectified AC line voltage within the range of 0 to VDCMAX at the VDC pin. The CS pin contains an averaging circuit which takes the ramp waveform at the CS pin, proportional to the input current sensed by the current sense resistor RCS. These two signals are then multiplied together to produce a voltage proportional to the converter input power and regulated against an accurate 1V source within the IC. Regulation is performed by means of a transconductance error amplifier that uses an external capacitor connected to ground at the COMP pin to realize an integrator to provide a stable error voltage used to control the converter on-time.

LED light output (expressed in Lumens) is proportional to applied power rather than current. This means that LEDs of the same type with higher forward voltage produce more Lumen output if driven at the same current. The IRS2983 uses constant power regulation so that current is reduced for higher LED voltages resulting in a more uniform light output level over tolerances in LED load voltage.

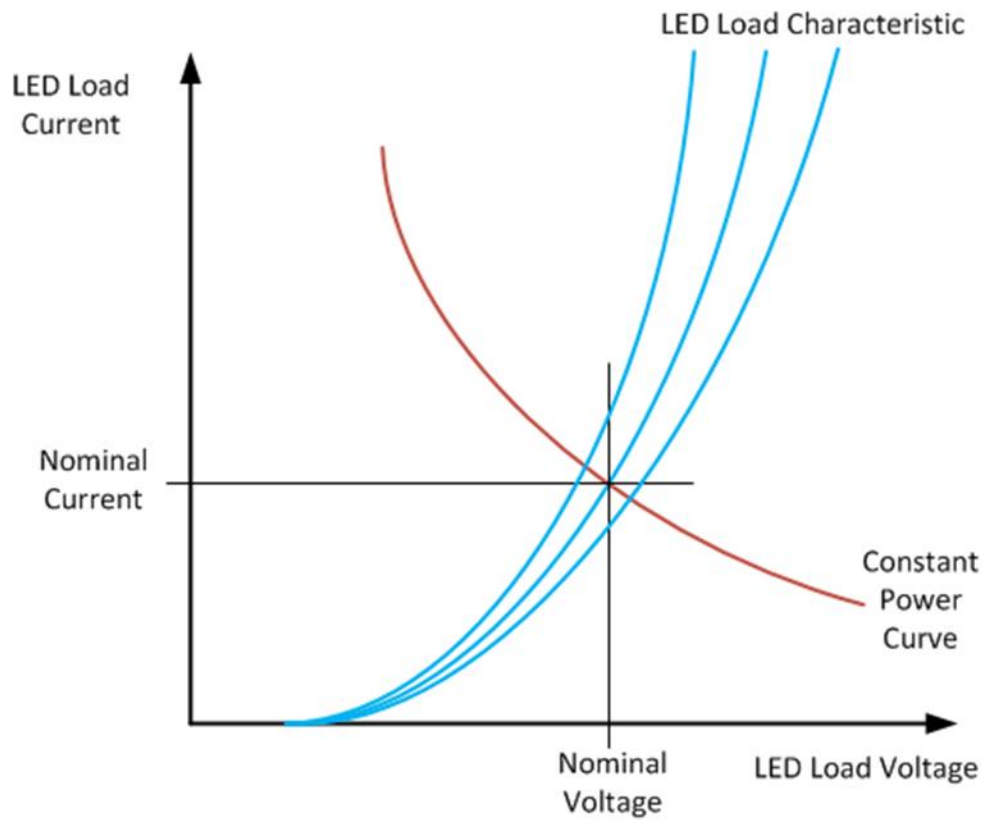


Figure 7: LED Load Characteristics

4. Design Equations

A detailed design process flow is presented below with equations for programming and selecting the IC component values and selecting the power components. The following design equations are for a rectified offline Flyback converter supplying a fixed load and without an input bulk capacitor as shown in Figure 5.

4.1 Define the Flyback Converter Operating Parameters

Below is a table of the required input parameters that should be selected prior to starting the design process. Typical values are given for some of these parameters.

PARAMETER	DESCRIPTION	UNIT
V_{ACMIN}	Minimum AC Line Voltage	VRMS
V_{ACMAX}	Maximum AC Line Voltage	VRMS
D_{MAX}	Maximum Duty Cycle. The recommended maximum value is 0.5	
f_{MIN}	Minimum operating frequency. Typically 65kHz	Hz
η	Converter Efficiency. Typically 85%	%
V_{OUT}	Nominal Output Voltage	V
I_{OUT}	Nominal Output Current	A
V_{AUXMAX}	Maximum Auxiliary Voltage. The recommended value is 18-20V	V
I_{AUXMAX}	Maximum Auxiliary Current. The recommended value is 30-50mA	A

Table 1: Required Operating Parameters

4.2 Calculate the Maximum Output Power

The maximum output power can be calculated with the following equation:

$$P_{OUTMAX} = P_{SEC} + P_{AUX} = V_{OUT} \cdot I_{OUT} + V_{AUXMAX} \cdot I_{AUXMAX} \qquad \text{[W]}$$

4.3 Calculate the Primary Inductance Value

$$L_{PRI} = \frac{V_{ACMIN}^2 \cdot \eta \cdot D_{MAX}{}^2}{\sqrt{2} \cdot P_{OUTMAX} \cdot f_{MIN}} \qquad \text{[H]}$$

4.4 Calculate the Turns Ratios for Flyback Inductor

The Flyback converter input/output ratio is identical to the Buck-Boost converter it is derived from, multiplied by the coils turns-ratio linking the primary and secondary side windings of the Flyback:

$$\frac{V_{OUT}}{V_{IN}} = \left(\frac{N_S}{N_P}\right) \cdot \left(\frac{D}{1-D}\right)$$

The secondary/primary turns ratio can be calculated:

$$\frac{N_S}{N_P} = \frac{V_{OUT} + V_F}{\sqrt{2} \cdot V_{ACMIN}} \cdot \frac{1-D_{MAX}}{D_{MAX}}$$

where V_F is the forward voltage of the output diode DFB, typically 0.7V.

The auxiliary/primary turns ratio, linking the primary side and auxiliary side windings, can also be calculated:

$$\frac{N_A}{N_P} = \frac{V_{AUXMAX} + V_F}{\sqrt{2} \cdot V_{ACMIN}} \cdot \frac{1 - D_{MAX}}{D_{MAX}}$$

where V_F is the forward voltage of the auxiliary diode DVCC, typically 0.7V.

The auxiliary/secondary winding ratio can be expressed as:

$$\frac{N_A}{N_S} = \frac{V_{AUXMAX} + V_F}{V_{OUT} + V_F}$$

4.5 Calculate the Peak Primary Inductor Current

$$I_{PK} = \frac{\sqrt{2} \cdot V_{ACMIN} \cdot D_{MAX}}{L_{PRI} \cdot f_{MIN}} \qquad \text{[A]}$$

Note: The Flyback inductor must not saturate at I_{PK} over the specified operating temperature range. Proper core size and air-gap should be considered in the inductor design.

4.6 Calculate the Reflected 'Flyback' Voltage

The VDS_{MAX} rating of the MOSFET should be greater than the peak input voltage plus the reflected output voltage:

Peak Input Voltage = $\sqrt{2} \cdot V_{ACMAX}$ [V]

Reflected 'Flyback' Voltage = $\dfrac{V_{OUT} + V_F}{\frac{N_S}{N_P}}$ [V]

Required MOSFET voltage = $VDS_{MAX} > \sqrt{2} \cdot V_{ACMAX} + \dfrac{V_{OUT} + V_F}{\frac{N_S}{N_P}}$ [V]

Please note that this only calculates the maximum DC level of the MOSFET when the switch is in an "OFF" state. However, due to the leakage inductance of the Flyback inductor, ringing occurs on each switching transition to the "OFF" state which can cause high voltage overshoots. Adequate margin should be added to the VDS_{MAX} rating of the MOSFET. For Universal Input applications where the input voltage reaches 265VAC, a 650-700V MOSFET is recommended. The voltage that the MOSFET is required to block may also be reduced by reducing the max duty cycle from the recommended 0.5, which has the effect of reducing the primary to secondary turns ratio.

4.7 Set Overvoltage Protection and Zero Crossing Detection

The ZX input is a multifunction input used for detecting the inductor zero crossing and for setting the output over-voltage level, both of which are sensed from the Auxiliary winding.

The recommended maximum current flowing into the ZX pin is 1mA.

In order to calculate the resistor dividers for ZX, the upper resistor must first be calculated. Total current is limited to below the 1mA recommended max.

Calculate R_{ZX1}. I_{ZX} may be selected to be between 0.5-1mA.

$$R_{ZX1} = \frac{V_{AUX}}{I_{ZX}} \quad \text{[Ohms]}$$

For Overvoltage Protection, the desired output overvoltage level, V_{OUTOV}, is set by choosing the value of R_{ZX2} to exceed the overvoltage threshold V_{OVTH} at the ZX pin. V_{OUTOV} should be higher than the nominal V_{OUT} so as to not interfere with normal operation.

$$V_{OUTOV} = \frac{V_{OVTH} \cdot N_S \cdot (R_{ZX1} + R_{ZX2})}{N_A \cdot (R_{ZX2})} \quad \text{[V]} \qquad V_{ZX} \geq V_{OVTH} @ V_{OUTOV}, \quad V_{ZX} \leq V_{CC}$$

For Zero Crossing Detection, the Vzx voltage should exceed the Vzx+ threshold at the minimum load voltage.

$$V_{ZX} = \left(\frac{N_A}{N_S}\right) \frac{R_{ZX2} \cdot V_{OUTMIN}}{(R_{ZX1} + R_{ZX2})} \quad \text{[V]} \qquad V_{ZX} \geq V_{ZX+} @ V_{OUTMIN}, \quad V_{ZX} \leq V_{CC}$$

4.8 Calculate the over-current resistor value

$$R_{CS} = \frac{V_{CSTH}}{I_{TRIP}} \quad \text{[Ohms]}$$

where V_{CSTH} = 1.25V

A recommended value for the trip current is $I_{TRIP} = 1.2 \cdot I_{PK}$

The R_{CS} power rating can be approximated:

$$PR_{CS} > \left(\frac{P_{OUT}}{V_{AC\,MIN} \cdot \eta} \right)^2 \cdot R_{CS} \quad \text{[W]}$$

4.9 Set the Primary Side Power Regulation

The IRS2983 averages the voltage at the CS pin, which is proportional to the converters input current and multiplies it with the VDC voltage, which is proportional to the rectified input voltage to produce a voltage proportional to the converter input power. The multiplier equation can be specified as:

$$V_{CSAVG} \cdot V_{DC} \cdot K_{MULT} = M_{OUT}$$

The multiplier output is regulated against an accurate 1V source within the IC. KMULT is the multiplier gain factor.

The RDC resistor sets the power regulation as follows:

$$R_{DC} = \frac{R_{IN}}{\left(\frac{K \cdot P_{IN} \cdot R_{CS}}{1.11} \right) - 1} \quad \text{[Ohms]}$$

Two 680kOhm resistors in series may be used for RIN.

4.10 Select CDC capacitor

The equation for RDC assumes a DC level at the VDC input which is proportional to the full wave rectified AC signal. The DC level is achieved with the addition of the capacitor CDC.

A CDC capacitor of 1 µF is recommended to maintain an adequate DC level across the entire AC line for the VDC pin.

4.11 Select COMP capacitor

A COMP capacitor of 1 µF is recommended. Choosing too large of a COMP capacitor may cause a noticeable delay in the turn-on of the LED load. This value also enables the soft-start feature for the application circuit.

4.12 Select the Output and Auxiliary Diodes

For the output and auxiliary diodes the proper blocking voltage must be selected for the peak input line voltage.

For the Auxiliary VCC diode, the required blocking voltage can be calculated:

$$V_{DVCC} > \sqrt{2} \cdot V_{ACMAX} \cdot \frac{N_A}{N_P} + V_{AUXMAX} \quad [V]$$

For the Output diode, the required blocking voltage can be calculated:

$$V_{DFB} > \sqrt{2} \cdot V_{ACMAX} \cdot \frac{N_S}{N_P} + V_{OUT} \quad [V]$$

4.13 Output Resistor

A resistor is added across the output of the converter to discharge the capacitor during converter power-off.

The maximum output voltage of the converter under open load condition depends on the over voltage protection level, V_{OUTOV}, calculated in Step 7.

The output load resistor can be calculated:

$$R_{OUT} = \frac{V_{OUTOV}{}^{2}}{P_{MAX\_ROUT}} \quad [Ohms]$$

The max resistor wattage under open load condition may be selected to be around 0.1W.

4.14 Output Capacitor

The output capacitor affects the output current ripple. A larger capacitor reduces the output current ripple, but adds cost and size.

The equation below can be used as a starting point for selecting an output capacitor.

$$C_{OUT} = 1.7 * I_{OUT} \cdot 1000 \quad [uF]$$

Round down to the nearest standard capacitance value. The output capacitor may be decreased or increased from this starting point depending on the ripple requirement and cost factor.

4.15 Select Offset Resistor (Optional)

Although the IRS2983 already features a low output current variation over the entire line, for some applications tighter output current variation may be required over a wider input voltage range. A feed forward resistor (RADJ) may be added to the circuit as shown in the schematic in Figure 5. A typical value for this resistor is 1Mohm, but it varies depending on the input voltage range, converter properties and the output current regulation accuracy required. Output current tends to rise with increasing input voltage so this resistor will compensate over a wide input range. Also note that adding this resistor will affect the primary side regulation point.

5. Flyback Inductor Construction

A detailed design process is presented below for constructing the Flyback inductor with multiple windings.

5.1 Determine the Primary Winding Wire Size

The RMS current through the primary winding can be calculated as:

$$I_{PRMS} = I_{PK}\sqrt{\frac{D_{MAX}}{3}} \quad \text{[A]}$$

The required primary winding wire area is:

$$A_{WPRI} = \frac{I_{PRMS}}{J_{MAX}} \quad [\text{mm}^2]$$

where J_{MAX} is the current density of a wire, which is typically 4-6 A/mm^2.

With the wire area calculated, see the Appendix 2 chart to select the closest wire size in AWG.

5.2 Determine the Secondary Winding Wire Size

For the secondary winding, the peak current can be calculated as

$$I_{SPK} = \frac{2 \cdot I_{OUT}}{1 - D_{MAX}} \quad \text{[A]}$$

The RMS current through the secondary winding can be expressed as:

$$I_{SRMS} = I_{SPK} \cdot \sqrt{\frac{1-D_{MAX}}{3}} \qquad [A]$$

The required secondary winding wire area is:

$$A_{WSEC} = \frac{I_{SRMS}}{J_{MAX}} \qquad [mm^2]$$

For the secondary winding which carries the LED load current we can use two wire conductors, or strands, of one-half the area to avoid using a single large gauge of wire, thus reducing the skin effect.

$$A_{WSECSTRAND} = \frac{A_{WSEC}}{2} \qquad [mm^2]$$

With the strand wire area calculated, see Appendix 2 to select the closest wire size in AWG.

5.3 Determine the Auxiliary Winding Wire Size

For the auxiliary winding, the maximum current can be calculated as

$$I_{APK} = \frac{2 \cdot I_{AUXMAX}}{1-D_{MAX}} \qquad [A]$$

The RMS current through the auxiliary winding can be expressed as

$$I_{ARMS} = I_{APK} \cdot \sqrt{\frac{1-D_{MAX}}{3}} \qquad [A]$$

The required auxiliary winding wire area is:

$$A_{WAUX} = \frac{I_{ARMS}}{J_{MAX}} \qquad [mm^2]$$

With the wire area calculated, see Appendix 2 to select the closest wire size in AWG.

5.4 Select the Inductor Core and Calculate the Turns

We must select the core shape and size for the Flyback inductor. The table below may be used to help select the EE type core size. For other types of cores please consult the manufacturer's datasheet.

Output Power, Offline Flyback	Core Type
0-10W	E13, E16
10-20W	E19, E20
20-30W	E20, E22
30-50W	E22, E25
50-70W	E25, E30
70-100W	E30, E35, E36, E40

Table 2: EE Core Selection Guideline

After a core is selected determine the following specifications of the core from the manufacturer's datasheet:

SYMBOL	PARAMETER	UNIT
A_e	Effective area of core	mm^2
l_e	Effective length of core	mm
$\sum(l/A)$	Core factor	mm^{-1}

Table 3: Inductor Core Specifications

Select an air gap length for the core, g , in mm. A starting value of 0.5mm may be used.

Calculate the effective permeability of the core:

$$\mu_e = \frac{\mu_i}{1 + \left(\frac{g \cdot \mu_i}{l_e}\right)}$$

where the initial permeability of the core, $\mu_i = 2000$ for N27 or equivalent.

Calculate the inductance factor:

$$A_L = \frac{\mu_o \cdot \mu_e \cdot 10^6}{\sum(l/A)} \qquad \text{[nH/turn}^2\text{]}$$

where $\mu_o = 4 \cdot \pi \cdot 10^{-7}$

Calculate the number of turns for the primary and round up to the nearest number:

$$N_{PRI} = \sqrt{\frac{L_P}{A_L}}$$

Calculate the maximum flux density:

$$B_{MAX} = \frac{N_{PRI} \cdot I_{PMAX} \cdot A_L}{A_e} \quad \text{(T)}$$

Check that the maximum flux density meets the criteria $B_{MAX} \leq 0.3T$. If it exceeds this value, increase the air gap length of the core, g. It is recommended that the air gap length be chosen to get a flux density value from 0.25-0.3T. This will save space by reducing the number of turns required on the core, but not exceed the maximum flux density requirements.

Calculate the number of turns for the secondary and auxiliary windings, using the ratios obtained in section 4, rounding up to the nearest number:

$$N_{SEC} = N_{PRI} \cdot \frac{N_S}{N_P}$$

$$N_{AUX} = N_{PRI} \cdot \frac{N_A}{N_P}$$

5.5 Check if the Windings will fit the Winding Area

Now, check if the windings will fit the winding area of the core selected,

$$\frac{1}{K_U} \cdot \left[\left(N_{PRI} \cdot A_{WPRI}\right) + \left(N_{SEC} \cdot A_{WSEC}\right) + \left(N_{AUX} \cdot A_{WAUX}\right)\right] \leq A_W$$

where K_U is the fill factor, typically 0.3-0.7. This depends on the core and the winding method used. Since the wires are circular, and not wound perfectly tight, some space of the winding area will be wasted. Determining what fill factor to use for which core will require some engineering trial and error but an initial value of 0.7 is recommended.

If the area used by the windings is greater than the available winding area, either reduce the number of windings by decreasing the inductance (increasing the minimum frequency) or increasing the flux density, or select a larger core.

5.6 Flyback Inductor Construction

Flyback inductor construction is critical in the converter design process and a poorly designed inductor will reduce converter efficiency. In order to design a proper Flyback inductor with low leakage an example is given in the diagram in Figure 8, which includes the winding orientations designated by the dots. This figure shows the Primary turns split into two separate windings. The Secondary and Auxiliary windings are sandwiched between the split primaries as shown in Figure 9. This winding stack-up is critical in order to reduce the leakage inductance inherent in a Flyback inductor. Three layers of insulation tape are used between the primary and secondary sides of the windings in order to pass safety standards for electrical strength requirements. The secondary side winding should also be of triple insulated wire. This type of construction will achieve leakage inductances of ideally <2% of the primary inductance, depending on the core used and winding method.

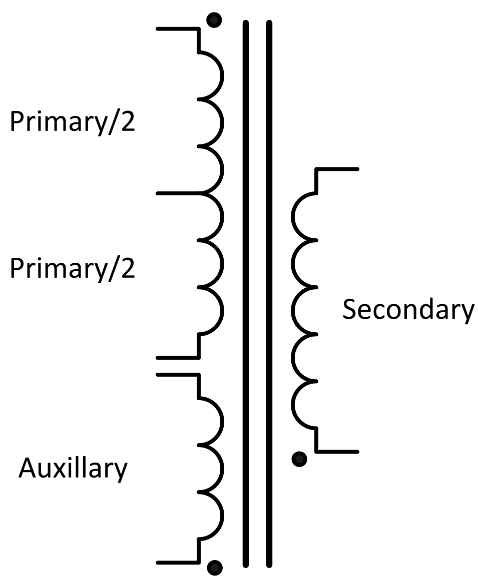


Figure 8: Flyback inductor Diagram

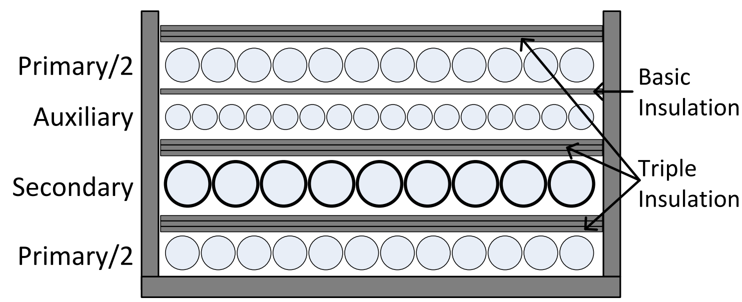


Figure 9: Flyback inductor Winding Stack-up

6. Auxiliary Supply Circuit

Two examples are given below for biasing the supply of the IC from the auxiliary winding. Figure 10 uses only a Zener diode and current limiting resistor to clamp the VCC voltage to a regulated value. The downside of this circuit is that because the auxiliary voltage varies with output voltage variations the clamping zener diode power losses are associated with the voltage dropped across the device. For a more robust solution a BJT (QVCC) is used to drop the voltage, shown in Figure 11. The benefit of this is that the BJT only draws as much current as the IC requires so the power losses are held in check even with a wide range of auxiliary voltages.

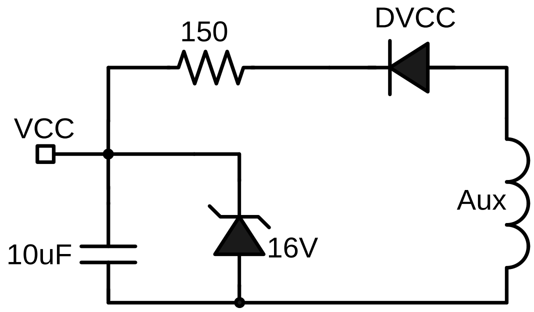


Figure 10: VCC Supply

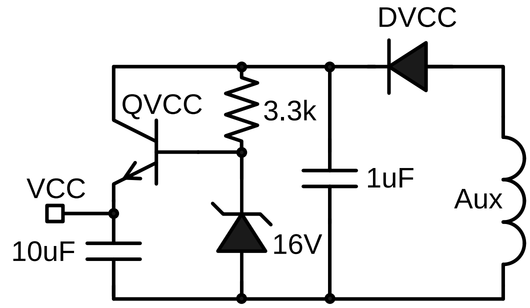


Figure 11: VCC Supply with BJT

7. Output Short Circuit Protection

Short circuit protection is achieved by allowing the auxiliary supply circuitry to collapse during this condition. In applications where the converter input voltage is >230VAC an additional 22-47pF capacitor should be added across RZX2 to filter out voltage spikes into the ZX pin caused by the leakage inductance during short circuits. The IC will go through startup again in a “hiccup” mode until the short circuit is removed, preventing damage to the circuit.

8. Snubber Circuit

A MOSFET snubber circuit is necessary for Flyback designs in order to clamp the ringing voltage caused by the leakage inductance which is inherent in the Flyback inductor. The snubber will protect the MOSFET Drain voltage from exceeding its breakdown (BVDSS) rating. Two examples of snubber networks are given in Figures 12 and 13. The most common snubber, the RCD snubber shown in Figure 12, is a cost-effective way to clamp the Drain ringing voltage. This circuit consists of a high voltage diode DSN, a high voltage capacitor CSN in the nF range, and a resistor RSN in the hundreds of kOhms range. Damping may be added to this circuit by adding another resistor (100Ohm-1kOhm) in series to the capacitor CSN to dampen the ringing further at the expense of an increased voltage transient on switch turn-off. The trade-off of this solution is calculating the optimal snubber components, which will vary in each design. Figure 13 shows another solution using a high voltage diode and TVS. This solution is simpler to implement as only the clamping voltage of the TVS needs to be selected.

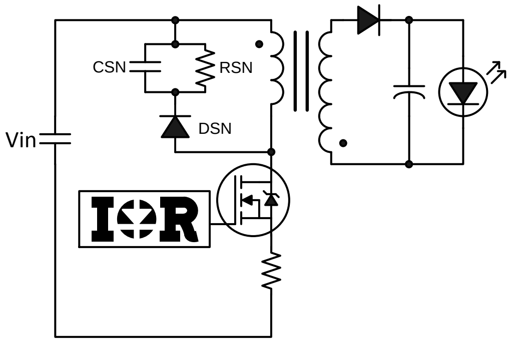


Figure 12: RCD Snubber

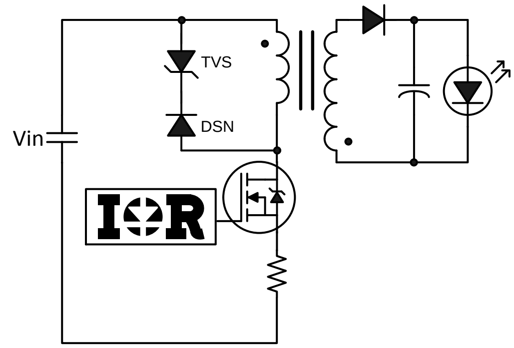


Figure 13: TVS Snubber

The results of the MOSFET Snubber circuits are shown in Figures 14-16. In this particular example, a high leakage inductance was present in the Flyback inductor due to poor construction. In Figure 14, without a snubber circuit the drain voltage of the MOSFET on turn-off reached a peak voltage of 775V! This is above the 700V rating of the MOSFET used in this example, a destructive level to the primary switch. In Figure 15, the RCD snubber is used which effectively clamped the peak voltage to 524V. Notice the large amount of ringing present due to the leaky inductor. The ringing may be damped by adding a resistor in series with CSN (be careful as this increases the peak voltage at turn-off), but a better solution is to construct the Flyback inductor as described in the previous section in order to achieve a low leakage design and avoid such excessive ringing caused by the leakage in the first place. Figure 16 shows the effectiveness of using a TVS, which clamped the peak voltage to 488V and also damped the ringing. The downside to this solution is the additional cost of the TVS.

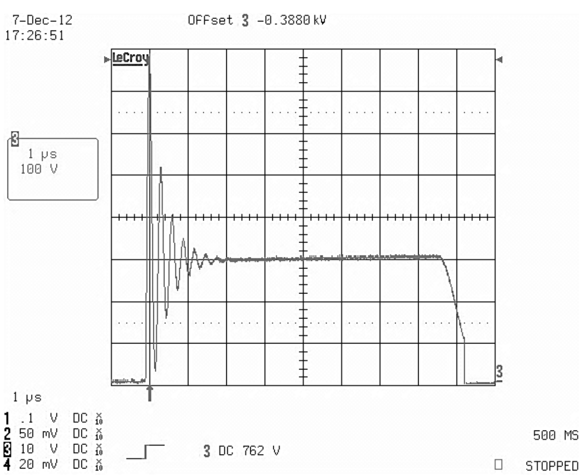


Figure 14: Drain Voltage No Snubber

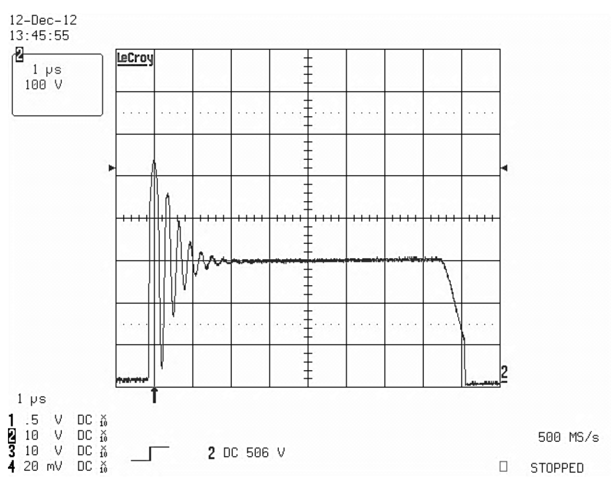


Figure 15: Drain Voltage RCD Snubber

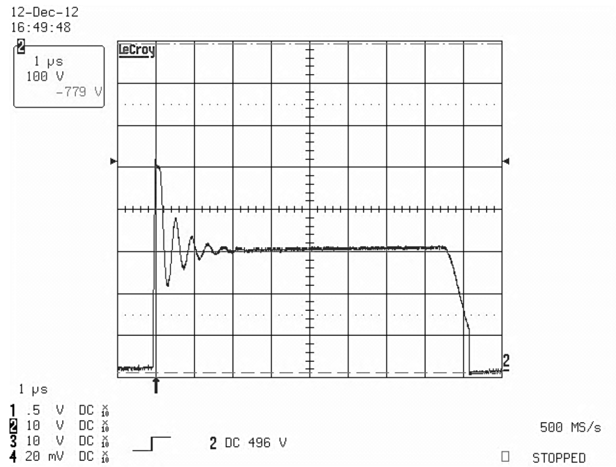


Figure 16: Drain Voltage TVS Snubber

9. EMI Considerations

In order to pass conducted EMI compliance using a Flyback converter an input EMI filter and capacitor connected from the input to output must be added. The input to output capacitor, CYC, in Figure 17 must be a Y-type capacitor with a voltage ranging equal to or greater than the maximum AC voltage rating, in order to be able to pass the isolation requirements for isolated Flyback designs. The value of this capacitor is limited by the max input voltage and leakage requirements to protective earth ground, typically 0.25mA:

$$C_{y\_MAX} = \frac{0.25mA}{V_{ACMAX} \cdot 2\pi \cdot f_{LINE}}$$

The input EMI filter for a two-terminal input consists of two X-capacitors (CX1 and CX2), a common mode choke (LCM) and differential mode inductors (LDM1, LDM2). The stray inductance of the common mode choke is generally used as the differential mode inductance shown in Figure 17. The value of these components depends on the input voltage range and power level of the converter and the required attenuation.

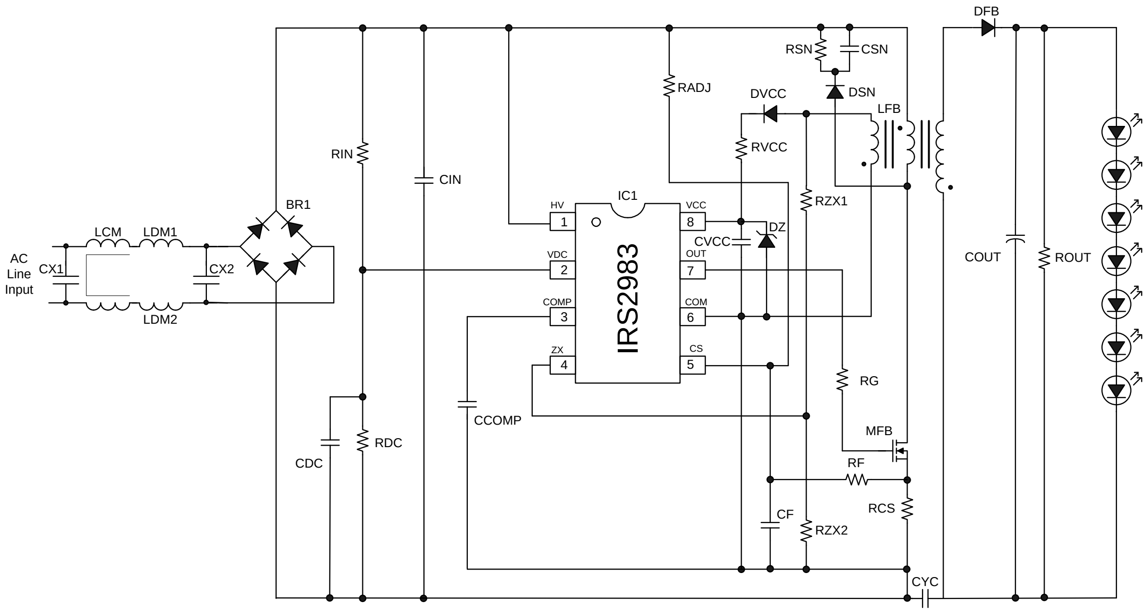


Figure 17: Typical Application Circuit Schematic with EMI Components

10. Surge Protection

Surge protection for a Flyback converter may be implemented by adding a MOV on the input as shown in Figure 18. The MOV should be rated higher than the maximum input line voltage to avoid destructive conduction during normal operation. A fuse is also added in front of the line to protect the Flyback converter and the MOV in the case of a short circuit.

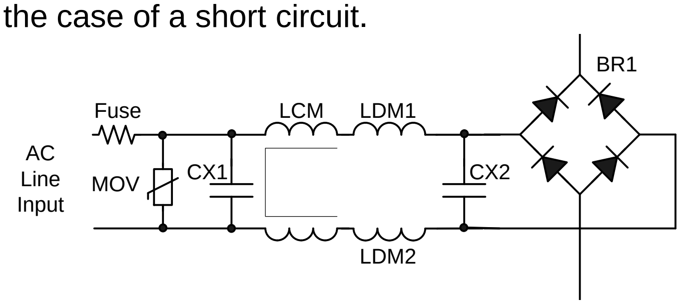


Figure 18: Surge Protection on the Input

For Flyback converters with no input bulk capacitor additional surge protection might be required, especially if surge ratings of 1kV and above are needed. In the typical application circuit schematic shown in this example, only a high frequency input capacitor is used. Additional surge protection is therefore required after the bridge as the MOV only clamps a portion of the surge voltage. In Figure 19, a voltage divider circuit is shown utilizing an additional inductor and TVS. The TVS may be selected with a specific clamping voltage offering greater protection of the rectified BUS during a surge event. The inductor (L1) is typically 1mH. This method allowed offline Flyback converters to pass 2.5kV surge testing. Another solution is to use a bulk capacitor, in the µF range, with a high voltage blocking diode as shown in Figure 20. The bulk capacitor will absorb the energy from the surge to limit high voltages on the BUS. A downside to this solution is the physical size of the capacitor needed and the high voltage that will always be present on the capacitor. A combination of these methods may also be used to realize even higher surge protection ratings.

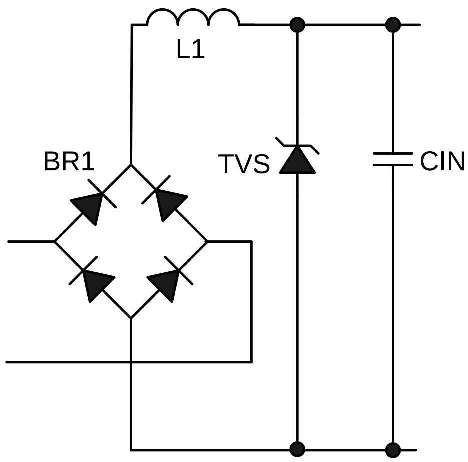


Figure 19: TVS Surge Protection

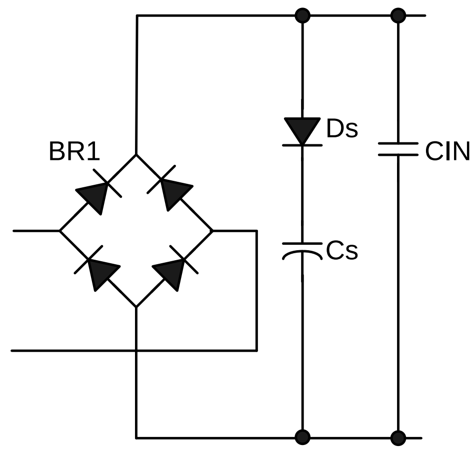


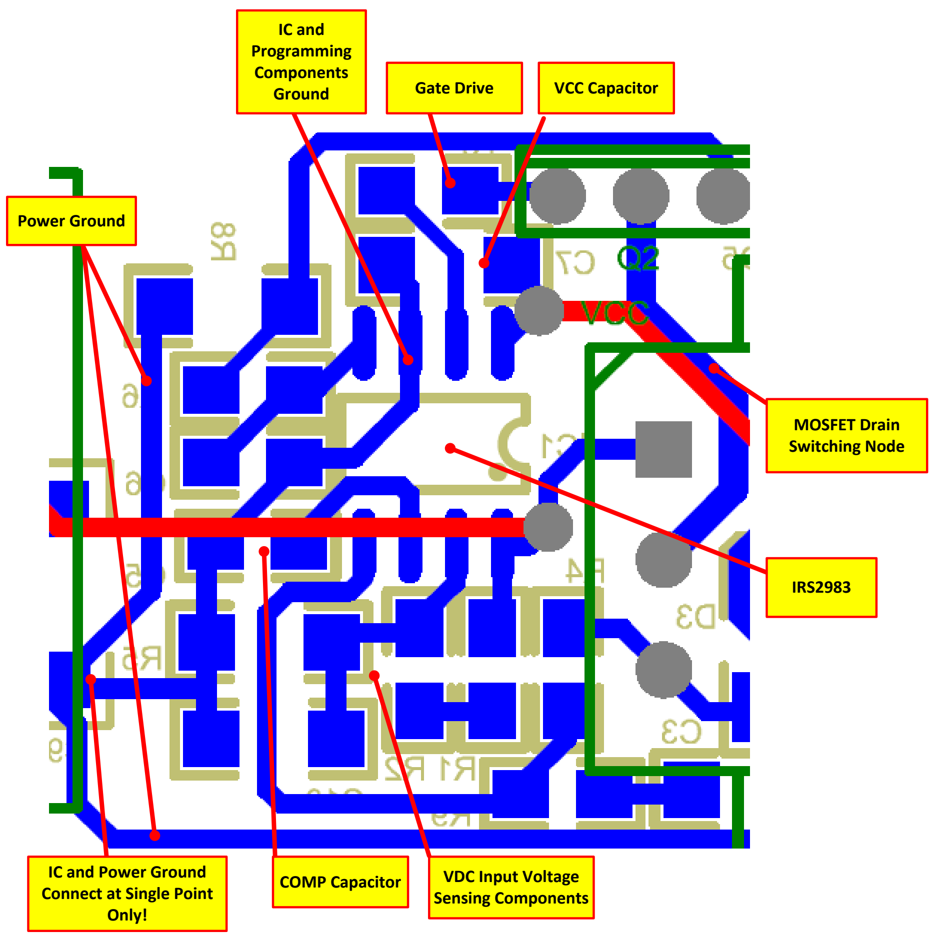
Figure 20: High Voltage Diode and Bulk Capacitor

11. PCB Layout Considerations

For correct circuit functionality and to avoid high-frequency noise problems, proper care should be taken when designing the PCB layout. Typical design problems due to poor layout can include high-frequency voltage and/or current spikes, EMC issues, latch up, abnormal circuit behavior, component failures, low manufacturing yields, and poor reliability. The following layout tips should be followed as early in the design phase as possible in order to reduce circuit problems, shorten design cycles, and to increase reliability and manufacturability:

1) Keep high-frequency, high-current traces as short as possible (drain switching node, output diode node). This will help reduce noise due to parasitic inductance of PCB traces.
2) Keep high-frequency switching nodes away from quiet or critical circuit nodes. This will help reduce noise coupling from switching nodes to other circuit nodes.
3) Place high-frequency filter capacitors directly at their IC pins (VCC pin). This will help insure the best possible filtering against high-frequency noise.
4) Do not connect power ground through IC ground or small-signal filter or programming component ground. Keep separate traces for power and IC or small-signal grounds and connect small-signal ground to power ground at a single point only. This will prevent high-frequency noise from occurring on critical small-signal nodes or IC pins which can cause circuit malfunction or failures.

5) Reduce the distance of the power switches to their gate drive pins as much as possible (OUT). This will help reduce the parasitic inductance in the traces. This will reduce possible voltage spikes due to gate drive switching and help prevent latch up due to voltage over- or under-shoot.
6) Use a limiting resistor in between the auxiliary supply and VCC. This will help prevent damage due to high-voltage or high-current spikes from the charge pump supply that can cause electrical overstress of the IC.
7) Place critical sensing nodes (current-sensing resistor, input voltage sensing, ZX detection) as close to the IC as possible. This will help eliminate false triggering or circuit malfunction due to noise being coupled onto to sensitive control signals.
8) For isolation, provide proper spacing between the input and output traces and components of the Flyback converter as per the relevant requirements.
9) See Figure 21 for PCB layout guidelines around the IRS2983.



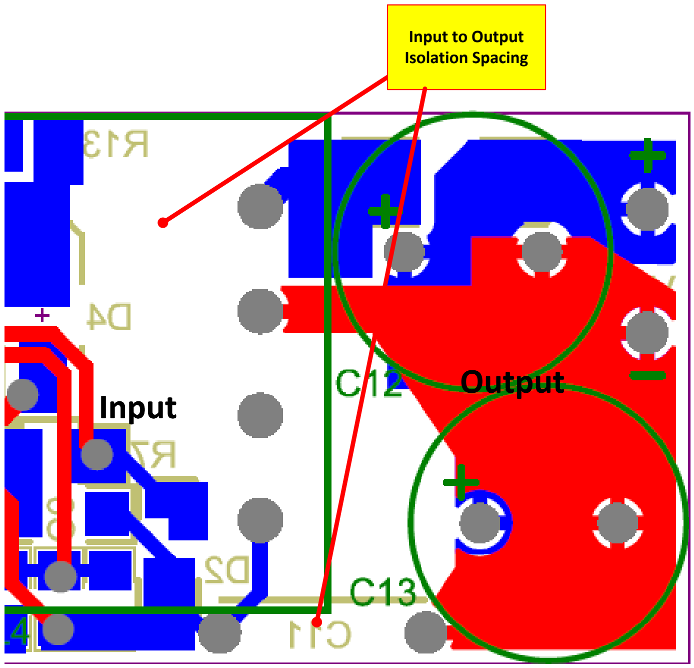


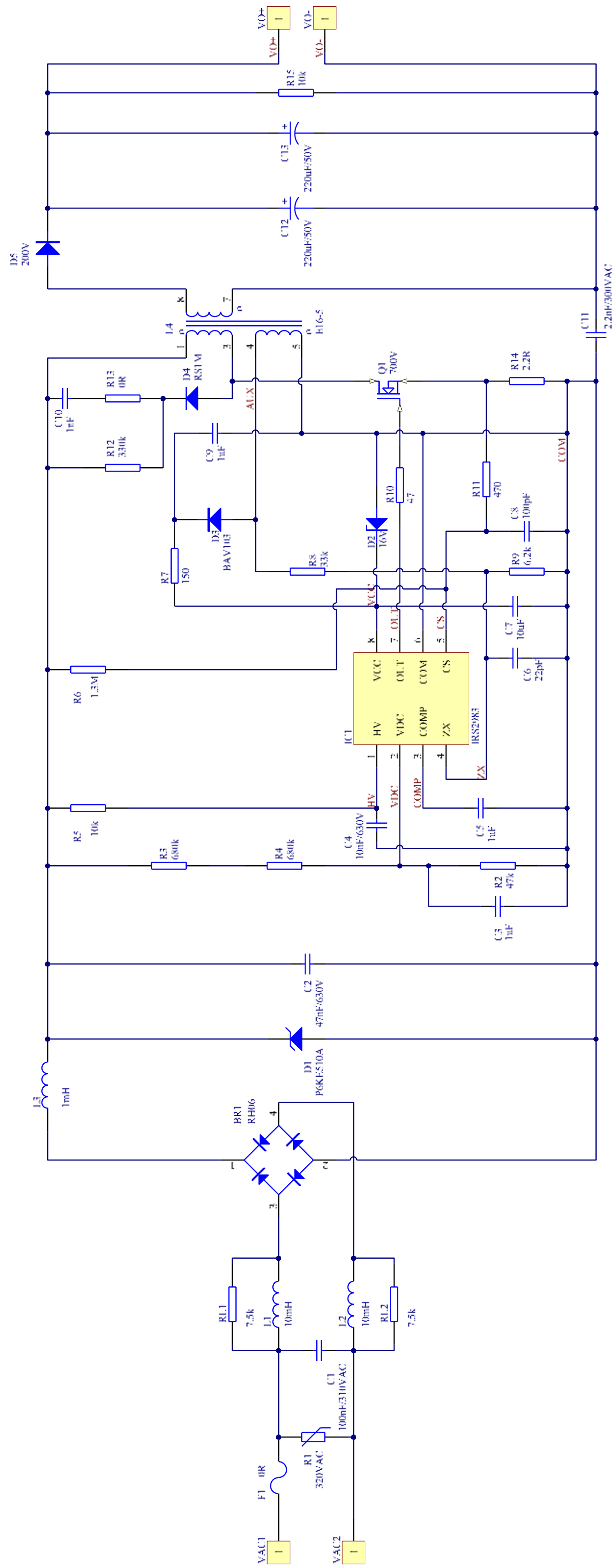
Figure 21: PCB Layout Guidelines

12. Conclusion

The information presented in this application note will help improve the design of the primary side regulated isolated offline Flyback converter and help reduce potential circuit problems. Ease of using and programming the IC, correct design and construction of the Flyback inductor, design of the VCC supply, design of the Snubber circuit, EMI and Surge considerations, and proper PCB layout guidelines help minimize design time, maximize performance, and maximize manufacturability and robustness of the final design. Finally, an excel spreadsheet design tool ("IRS2983 Design Calculator.xls") is also available that contains all of the necessary calculations described in this application note.

13. Appendix 1: Reference Design

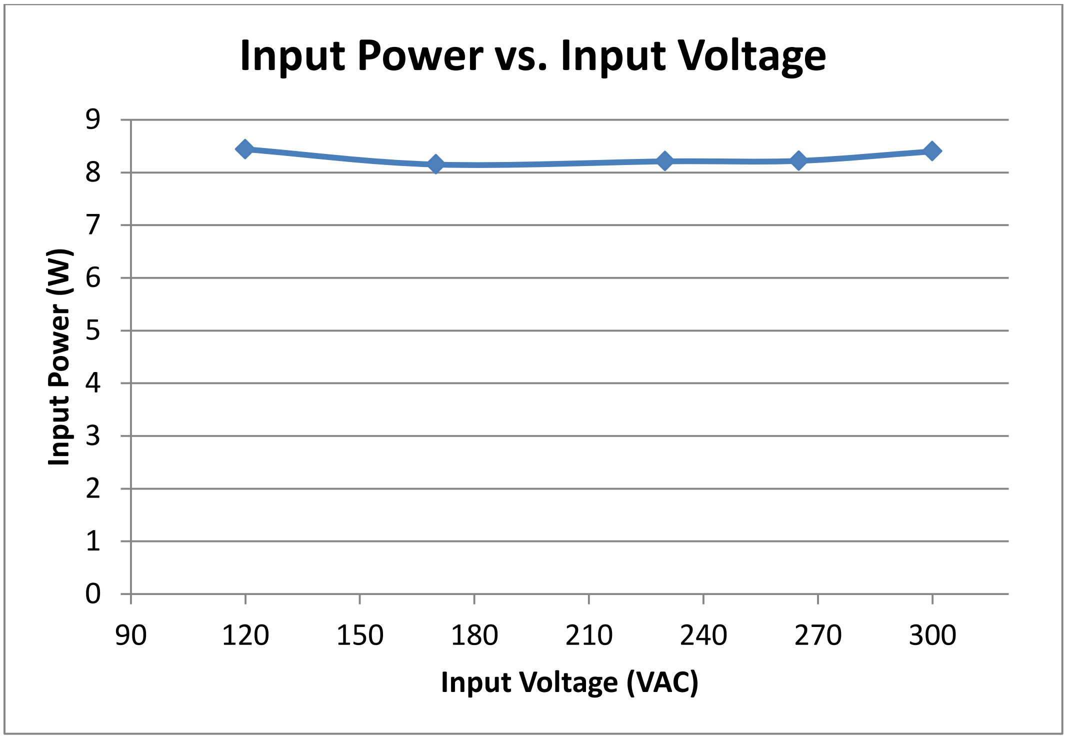
a. Schematic (PIN < 9W, VIN = 120 to 300VAC, VOUT = 24V, IOUT = 280mA)

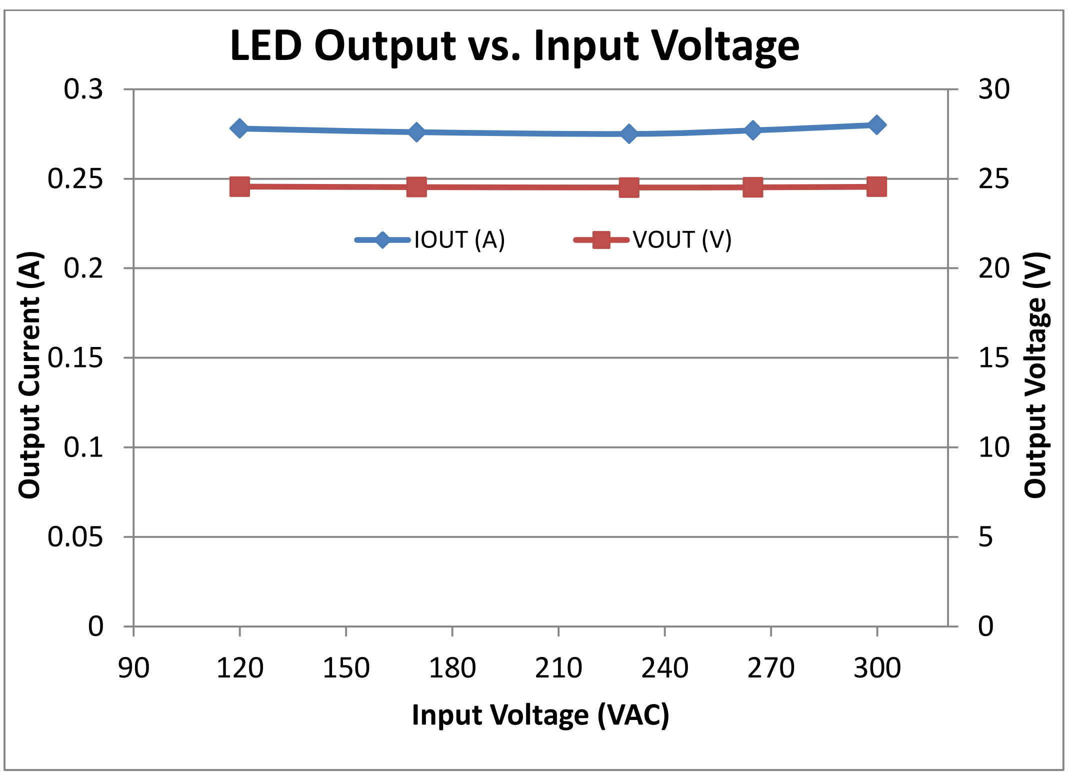


b. BOM (PIN < 9W, VIN = 120 to 300VAC, VOUT = 24V, IOUT = 280mA)

Index	Description	Part Number	Manufacturer	Quantity	Designator
1	RECT BRIDGE GP 600V 0.5A MINIDIP	RH06-T	Diodes Inc	1	BR1
2	CAP FILM 0.1UF 630VDC RADIAL	BFC233820104	Vishay BC Components	1	C1
3	CAP CER 0.047UF 630V X7R 1210	C3225X7R2J473K200AM	TDK Corporation	1	C2
4	CAP CER 1UF 25V 10% X5R 0805	C2012X5R1E105K085AC	TDK Corporation	2	C3, C5
5	CAP CER 10000PF 630V X7R 1206	C3216X7R2J103M115AA	TDK Corporation	1	C4
6	CAP CER 22PF 50V 5% NP0 0805	C0805C220J5GACTU	Kemet	1	C6
7	CAP CER 10UF 25V 20% X6S 1206	C3216X6S1E106M160AB	TDK Corporation	1	C7
8	CAP CER 100PF 50V 5% NP0 0805	C0805C101J5GACTU	Kemet	1	C8
9	CAP CER 1UF 50V 10% X7R 1206	CL31B105KBHNNNE	Samsung Electro-Mechanics America, Inc	1	C9
10	CAP CER 1000PF 630V 20% X7R 1206	C3216X7R2J102M115AA	TDK Corp	1	C10
11	CAP CER 2200PF 300VAC 20% RADIAL	VY2222M35Y5US63V7	Vishay	1	C11
12	CAP ALUM 220UF 50V 20% RADIAL	ECA-1HHG221	Panasonic Electronic Components	2	C12, C13
13	TVS 600W 510V UNIDIRECT DO-15	P6KE510A-TP	Micro Commercial Co	1	D1
14	DIODE ZENER 16V 500MW MINIMELF	ZMM5246B-7	Diodes Inc	1	D2
15	DIODE SW 200V 250MA SOD80C	BAV103-GS18	Vishay	1	D3
16	DIODE FAST RECOVERY 1KV 1A SMA	RS1M-13-F	Diodes Inc	1	D4
17	DIODE SUPER FAST REC 2A 200V SMB	ES2D-13-F	Diodes Inc	1	D5
18	RES 4.7 OHM 1/2W 5% AXIAL	NFR25H0004708JR500	Vishay BC Components	1	F1
19	Flyback Control IC	IRS2983	International Rectifier	1	IC1
20	CHOKE RF 10000UH 10% RADIAL	RL895-103K-RC	Bourns Inc.	2	L1, L2
21	INDUCTOR 1000UH .16A 10% AXIAL	AIAP-01-102-K-T	Abracon Corporation	1	L3
22	Flyback inductor EE16/8/5 Horizontal	NA	NA	1	L4
23	700V MOSFET N-Channel IPAK	NA	NA	1	Q1
24	VARISTOR 510V 1.75KA DISC 7MM	S07K320E2	Epcos Inc.	1	R1
25	RES 47K OHM 1/8W 1% 0805 SMD	ERJ-6ENF4702V	Panasonic Electronic Components	1	R2
26	RES 680K OHM 1/8W 1% 0805 SMD	ERJ-6ENF6803V	Panasonic Electronic Components	2	R3, R4
27	RES 10K OHM 1/8W 5% 0805 SMD	ERJ-6GEYJ103V	Panasonic Electronic Components	1	R5
28	RES 1.3M OHM 1/8W 5% 0805 SMD	ERJ-6GEYJ135V	Panasonic Electronic Components	1	R6
29	RES 150 OHM 1/4W 5% 1206 SMD	ERJ-8GEYJ151V	Panasonic Electronic Components	1	R7
30	RES 33K OHM 1/8W 5% 0805 SMD	ERJ-6GEYJ333V	Panasonic Electronic Components	1	R8
31	RES 6.2K OHM 1/8W 5% 0805 SMD	ERJ-6GEYJ622V	Panasonic Electronic Components	1	R9
32	RES 47 OHM 1/8W 5% 0805 SMD	ERJ-6GEYJ470V	Panasonic Electronic Components	1	R10
33	RES 470 OHM 1/8W 5% 0805 SMD	ERJ-6GEYJ471V	Panasonic Electronic Components	1	R11
34	RES 330K OHM 1/2W 5% 1210 SMD	ERJ-P14J334U	Panasonic Electronic Components	1	R12
35	RES 10 OHM 1/8W 5% 0805 SMD	ERJ-6GEYJ100V	Panasonic Electronic Components	1	R13
36	RES 2.2 OHM 1/4W 1% 1206 SMD	ERJ-8RQF2R2V	Panasonic Electronic Components	1	R14
37	RES 10K OHM 1/4W 5% 1206 SMD	ERJ-8GEYJ103V	Panasonic Electronic Components	1	R15
38	RES 7.5K OHM 1/8W 5% 0805 SMD	ERJ-6GEYJ752V	Panasonic Electronic Components	2	RL1, RL2

c. Results (PIN < 9W, VIN = 120 to 300VAC, VOUT = 24V, IOUT = 280mA)





14. Appendix 2: American Wire Gauge Table

American wire gauge (AWG)	Equivalent wire diameter (mm)	Equivalent wire area (mm^2)
50	0.0251	0.00049
49	0.0282	0.00062
48	0.0315	0.00078
47	0.0356	0.00100
46	0.0399	0.00125
45	0.0447	0.00157
44	0.0503	0.00199
43	0.0564	0.00250
42	0.0632	0.00314
41	0.0711	0.00397
40	0.0799	0.00501
39	0.0897	0.00632
38	0.101	0.00797
37	0.113	0.01
36	0.127	0.0127
35	0.143	0.016
34	0.16	0.0201
33	0.18	0.0254
32	0.202	0.032
31	0.227	0.0404
30	0.255	0.0509
29	0.286	0.0642
28	0.321	0.081
27	0.361	0.102
26	0.405	0.129
25	0.455	0.162
24	0.511	0.205
23	0.573	0.258
22	0.644	0.326
21	0.723	0.41
20	0.812	0.518
19	0.912	0.653
18	1.024	0.823
17	1.15	1.04
16	1.291	1.31
15	1.45	1.65

14	1.628	2.08
13	1.828	2.62
12	2.053	3.31
11	2.305	4.17
10	2.588	5.26
9	2.906	6.63
8	3.264	8.37
7	3.665	10.5
6	4.115	13.3
5	4.621	16.8
4	5.189	21.2
3	5.827	26.7
2	6.544	33.6
1	7.348	42.4
0	8.251	53.5

15. References

[1] IRS2983 LED Flyback Control IC Datasheet

16. Revision History

Date	Revision	Changes	Author
01/30/2014	1	Initial version	Ektoras Bakalakos