AND9094/D

NCL30002 High Power Factor Buck LED Driver

ON Semiconductor®

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APPLICATION NOTE

Introduction

LED lighting is becoming firmly established as a preferred light source due to high efficacy, long life, and no mercury. As general incandescent bulbs are phase out globally, LED light source adoption opens the door to a long life alternative without the main drawbacks of CFL. The only real drawback to incandescent bulbs is their rather pitiful efficacy which is in the range of 7−15 lumens per watt although it does have near ideal power factor (PF = 1). The efficacy of state of the art LEDs is about 125 lumens per watt. So today, LEDs are 10 times better than incandescent lamps or will produce the same lumens for one tenth the power. To further put this into perspective, if a light source could convert 100% of this electrical power to light, it would make 640 lumens per watt.

Overview

LEDs have simpler driver requirements than arc discharge lamps but still require drive electronics. LEDs behave much like voltage sources with a low series resistance so a regulated current drive is ideal and multiple LEDs can be connected in series to increase light output. A significant market is the bulb retrofit where the ancient A19 screw base is the most common lamp base. The A19 base is great for incandescent lamps but poses packaging problems for LED lamps and drive electronics. Some of the challenges for the drive electronics are compact size, low cost, high efficiency, and high power factor. New guidelines like the U.S. ENERGY STAR® program impose requirements for power factor and overall lamp/luminaire efficacy so it is crucial that the drive electronics be as efficient as possible in providing constant current drive. For ENERGY STAR bulbs over 5 W, the minimum power factor is 0.7. Commercial LED luminaires have more stringent power factor requirements greater than 0.9. The power requirements vary globally. In South Korea for example, the minimum PF requirement for integral bulbs is 0.85 and this increases to ≥ 0.9 for input power over 5 W.

For high powered applications like fluorescent light ballasts, It is common to use a dedicated boost front end converter stage to deliver power factor greater than 0.99. This is much higher performance than required in an LED bulb or down-light and adds an additional power conversion stage. Other topologies can also provide power factors greater than 0.9 if they are optimized for the application. Recall that incandescent and halogen bulbs are purposefully designed and optimized around one specific line voltage. In fact if the line voltage varies the brightness of the bulb will increase or decrease. This in itself is a huge degree of freedom for the power supply designer who may be use to designing laptop or low power wall plug supplies which are designed to work globally and operate from 90−264 Vac on both 50 Hz and 60 Hz systems. This application note describes an optimized power factor corrected buck topology that can achieve high power factor suitable for these low power LED lighting applications while achieving very high conversion efficiency with minimal parts count.

Design Target Specification

Let’s consider the following target design specifications and evaluate using a critical conduction mode (CrM) buck topology to implement a constant current driver.

- Series string of 8 LEDs @ ~3 V each
- LED Current – 750 mA
- LED Ripple Current < ±35%
- Line Voltage – 100 Vac – 132 Vac
- Line Frequency – 60 Hz
- Efficiency – ≥ 88%
- Power Factor ≥ 0.9
- Isolation – Not Required
- Small Volume

Topology Choice

The buck makes an excellent choice when the LED string voltage is much less than the line voltage. There are two common operating modes, continuous conduction mode (CCM) and discontinuous mode (DCM) based on the state of the current through the inductor. A special case of the DCM operation is critical mode conduction where the controller turns back on as soon as the inductor current goes to zero. CrM offers some specific advantages.

1. CrM is soft switching − eliminating the EMI caused by force commutation of the buck diode found in continuous conduction mode operation
2. Magnetics can be very compact
3. Variable frequency offers inherent spread spectrum EMI performance

There are some challenges with the buck topology that must be addressed.

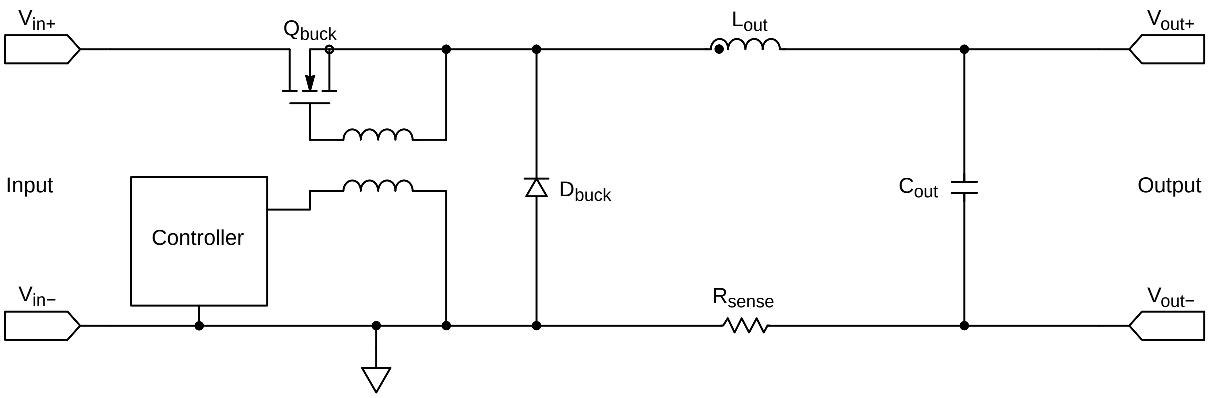

Figure 1. Conventional Simplified Buck Converter

In the traditional construction of a buck converter, the switching transistor is floating with respect to the circuit common while the output is referenced to the circuit common. This makes measuring output voltage and current convenient at the expense of a complex floating gate drive.

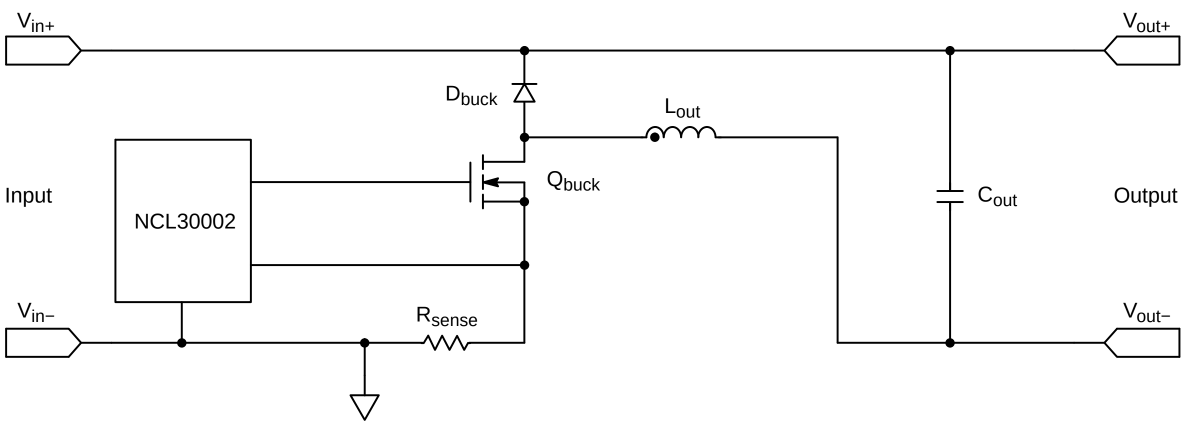

Figure 2. Inverted or Reverse Buck Converter

This is also a buck converter with a rearrangement of the circuit elements. Here the switch drive is referenced to the common of the control IC while the output is now referenced to the + input making voltage and current measurements floating. We can measure the peak current in the switch which is the same as the peak current in the inductor. One of the unique features of CrM operation is that the peak to average current in the inductor is always 2:1. Since we have a way to measure the peak current, we also have a way to control the average current in the output. This is not the case with a continuous conduction peak current sense implementation since the LED forward voltage variation and inductor variation induce additional error terms in attempting to delivery an average LED current.

To achieve high power factor, we cannot have any significant energy storage in the input circuit. So this buck converter must operate from the rectified line. This means the buck converter will operate in 3 different modes.

1. “Zero” Input Current (I_{in} = 0) − Buck converters cannot deliver power when Vin ≤ Vout. The “dead time” where no current flows around the zero crossing is dependent on the line voltage and the load voltage.
2. Constant On-Time (T_{on} = constant) − This is the same as the boost converter. Constant T_{on} forces the peak current to be proportional to the input voltage which is critical to improved PF.
3. Constant Peak Current (I_{peak} = constant) – In this region, the peak inductor current and thus the LED current is limited. In this region, the unique nature of the CrM buck means that the average output current is equal to half the peak current. Also the off time is fixed in this mode since the peak current and the output voltage are virtually constant.

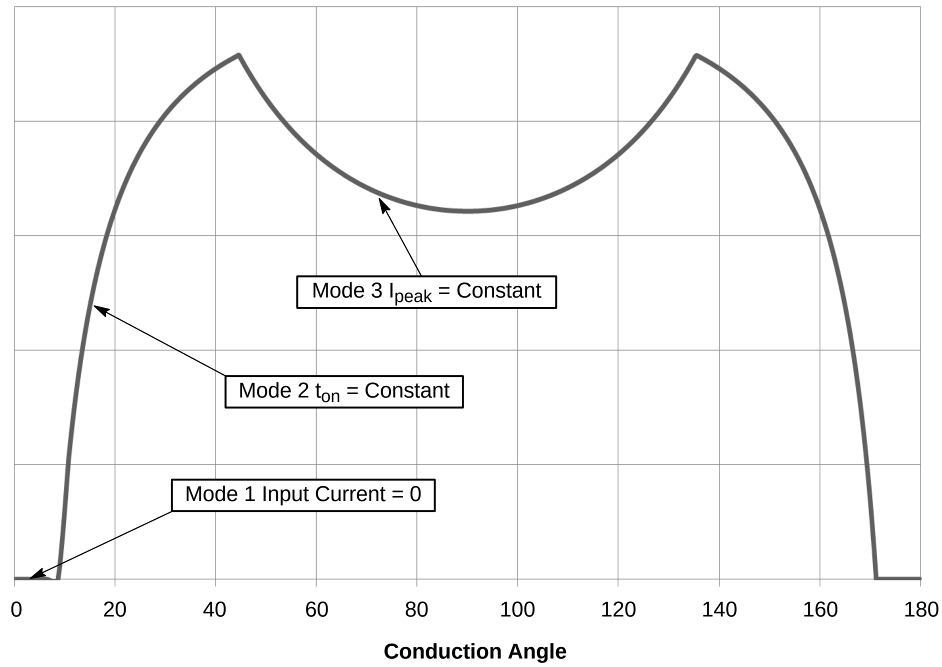

Figure 3. Theoretical Average Input Current over One Half Line Cycle (Conduction Angle)

This creates a typical waveform which does not look very sinusoidal. However this waveform has a power factor greater than 0.9 even with increased distortion. Note that mode 2 is the mode that a tradition constant on time boost PFC operates over the entire line cycle. We could have done that with this buck implementation as well resulting in a more ideal power factor; however, the peak current would be higher resulting in a larger buck inductor. Since size is an important parameter, we choose to limit the peak and accept increased distortion since it is not necessary in this application to achieve a power factor near 1.

The transition between modes 2 and 3 is the critical point in determining the power factor. If operating time in mode 3 becomes too large (this happens if Ton_max is large), the peak current at the transition grows raising the RMS input current and reducing the power factor. The current shown in Figure 4 has a power factor of 0.8 which still is acceptable in many residential lighting applications.

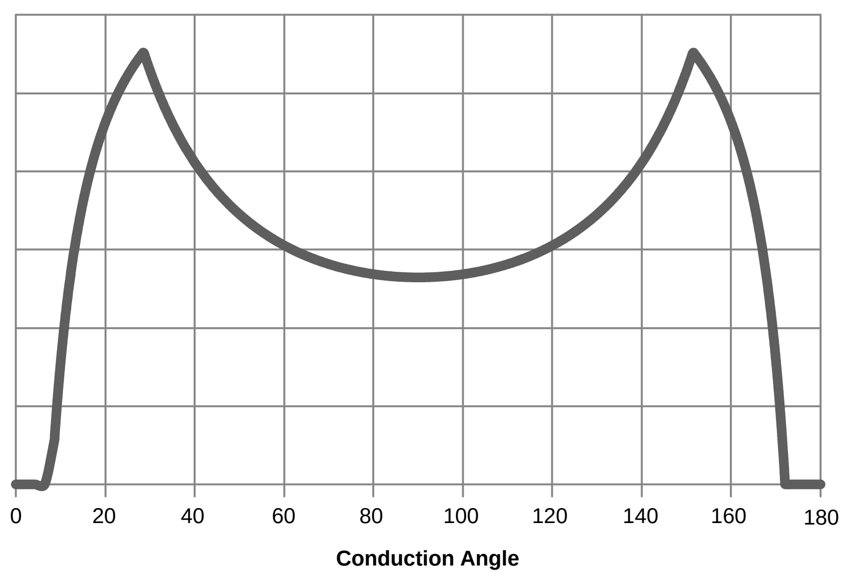

Figure 4. The "Bat" Effect due to Large T_{on} Across One Half Line Cycle

Another change between modes 2 and 3 is that the converter transitions from positive impedance in mode 2 to negative impedance in mode 3. In mode 2, the current increases as the voltage increases. So the impedance is like a normal resistor or the incandescent bulb we are trying to emulate. Mode 3 shows decreasing current with increasing voltage. The incremental value of the negative impedance goes down as the voltage decreases. If the negative impedance of the converter is smaller than the positive impedance of the input circuit (primarily the EMI filter), the converter will oscillate against the positive impedance of the input circuit. So expanding mode 3 over a higher percentage of the line cycle not only degrades power factor but also increases the likelihood of oscillation.

If the time in mode 3 (Peak Current Mode) is no more than 60% of the half cycle (or 108°) then the power factor will be more than 0.9. This is influenced by the output voltage of the LED string. Table 1 explains the effects of operating time in the 3 modes.

Table 1. EFFECT OF TIME IN VARIOUS MODES ON POWER FACTOR AND OUTPUT CURRENT

	Time in Mode 1		Time in Mode 2		Time in Mode 3	
	More	Less	More	Less	More	Less
Power Factor	Lower	Higher	Higher	Lower	Lower	Higher
Output Current	Lower	Higher	Lower	Higher	Higher	Lower

Operation in Mode 3 allows for maximum power delivery with the smallest Ipeak for a given inductance. This is an advantage over a constant Ton type PFC control where the peak current will be higher for a given power output.

Design

Now that we have established the basic operating modes, let’s work on the detailed design. Let’s start with the basic schematic using the NCL30002 and describe the operation of each section.

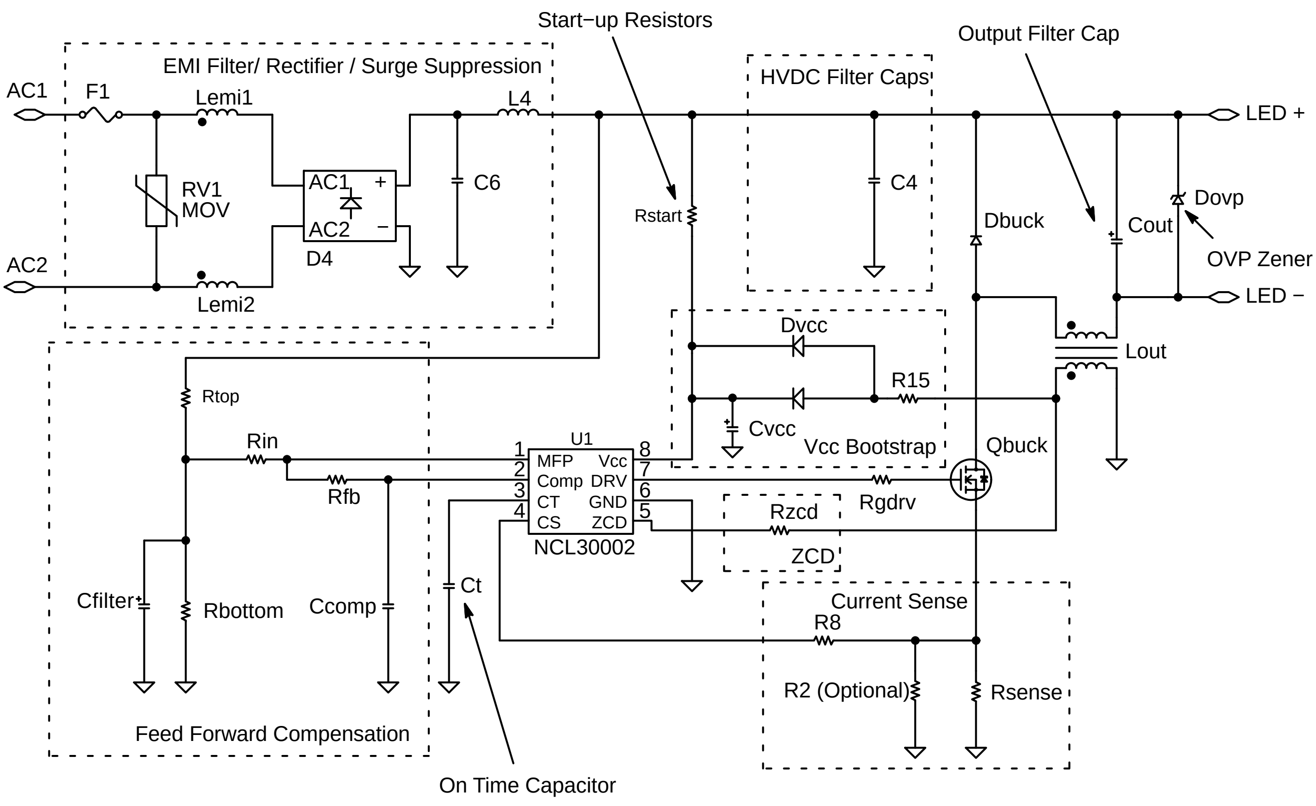

Figure 5. Simplified PFC Buck Application

Start Up Resistors

The NCL30002 draws very low current at startup. V_{CC} current is 2−3 mA typically in normal operation. Even currents this small will dissipate significant power if supplied by the high voltage DC bus. So to maximize efficiency, we use trickle start along with a bootstrap winding off the output inductor. The start-up time is dependent on the V_{CC} filter capacitor and the start-up resistors. The start resistors charge the V_{CC} capacitor to the start threshold of the NCL30002 which is about 12.5 V dc. The NCL30002 starts switching which discharges the V_{CC} capacitor. If all goes well, the bootstrap winding provides the V_{CC} power before the V_{CC} capacitor reaches the under voltage lock out. The V_{CC} capacitor needs to supply power for about 35 ms to allow time for the output voltage to reach steady state. The bootstrap voltage is proportional to the output voltage by the turns ratio of the bootstrap winding to the main inductor winding. We can calculate the size of the V_{CC} capacitor as follows:

$$\text{Start Threshold} = 12.5\ \text{V}$$

$$\text{Lock Out Threshold} = 10\ \text{V}$$

$$\text{IVCC}\ \ 2.6\ \text{mA}$$

$$C = \frac{1 \times dT}{dV} = \frac{2.6\ \text{mA} \times 35\ \text{ms}}{2.5\ \text{V}} = 36.2\ \mu\text{F}$$

We will choose a standard value of 35 μF. Excessively large V_{CC} capacitors simply delay the circuit from starting. In the typical application, rapid starting is expected. The start-up delay time is a trade-off with power dissipation. Let's calculate the resistance required to start our circuit in 1s at our low line of 100 Vac. The peak of the 100 Vac line is 141 Vdc. Since the NCL30002 start threshold is 12.5 Vdc, we can make the simplifying assumption that the V_{CC} capacitor charges from a current source = 141 Vdc/Rstart. We find Rstart for 1 second start delay.

$$dT = \frac{C \times dV}{I}$$

$$1\text{s} = \frac{35\ \mu\text{F} \times 2.5 \times \text{Rstart}}{141\ \text{V}}$$

$$\text{Rstart} = \frac{1\text{s} \times 141\ \text{V}}{(35\ \mu\text{F} \times 12.5\ \text{V})} = 322\ \text{k}\Omega$$

After the NCL30002 starts, the start-up resistor still supplies current and therefore continues to dissipation power. The power dissipation is approximately equal to:

$$\frac{\text{Vhigh\_line2}}{322\ \text{k}\Omega} = \frac{1322}{322\ \text{k}\Omega} = 54\ \text{mW}$$

The dissipation will in fact be slightly less because the Vcc subtracts from the HV DC but this represents worst case and is therefore a good safe approach to use. To reduce the start up time to 330 ms the Rstart is reduced to 1/3 the value (100 kΩ) resulting in the dissipation is increasing to 174 mW. Reducing the V_{CC} capacitor will allow faster start times without sacrificing power dissipation in Rstart. This is a trade off for higher V_{CC} voltages which permit larger ripple on V_{CC}.

HV DC Filter Capacitors

The HV DC filter capacitors are required to filter the switching current on the HV DC bus. It must be enough for adequate filtration but not so much that is hurts the power factor. For a 120 V ac nominal mains voltage, 30 nF/W is a conservative figure of merit. The capacitors must be of a low ESR type. Metalized polypropylene or metalized polyester is preferred.

Output Filter Capacitor

Since there is no energy storage on the HV DC, all filtration of the output current is done by the output filter components. In this case, a capacitor is the only output filter component. Filtering the 120 Hz ripple will determine the size for the output capacitor. The high frequency filtering is easily handled by a much smaller capacitor but to limit the maximum LED current we need to smooth the 120 Hz component. As mentioned earlier, LEDs act like a voltage source with a series resistor so this can be modeled. Figure 6 illustrates based on Cree XLamp® XP−E LEDs.

Our curve fitting algorithm shows that our string of 8 LEDs can be modeled as a 23 V source with a 1.62 Ω series resistor. It's the series resistance that helps us determine the output capacitor size. For example if we want only ±20% current ripple @ 120 Hz, then the capacitor's impedance must be 40% of the series resistance.

Let's try ±35% ripple which means 750 mA ±250 mA which keeps us under the 1 A DC maximum rating of this particular LED.

$$\text{Cout} = \frac{1}{(1.62\ \Omega \times 70\% \times 2 \times \pi \times 120\ \text{Hz})} = 1170\ \mu\text{F}$$

We will select the next higher standard capacitor value which is 1200 μF. This amount of capacitance means that an electrolytic capacitor is the practical choice. In this topology it is possible to implement a design without the use of electrolytic capacitors; the tradeoff is high ripple which influences the LEDs selection and the intended drive current. For example with ±100% current ripple the average drive current would need to be < 500 mA.

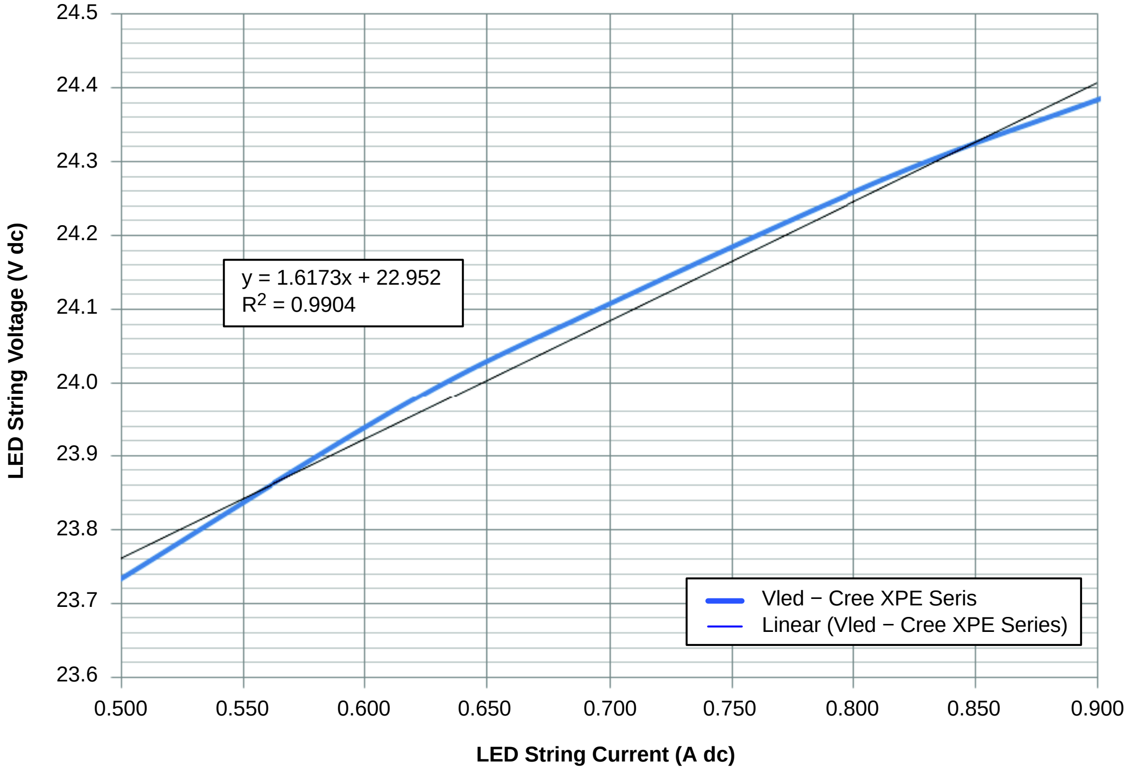

Figure 6. Typical V−I Curve for an 8 LED String

V_{CC} Bootstrap

The NCL30002 get its V_{CC} power from the bootstrap winding. The V_{CC} voltage is the image of the output voltage scaled by the turns ratio. The fact that LEDs act like a voltage source is beneficial here because the output voltage (and therefore V_{CC}) will remain stable over a wide range of output currents. The V_{CC} must be more than 10.2 V and less than 20 V under all conditions. Even for a fixed number of LEDS, the voltage will have a range due to the LED voltage thermal drift and initial forward voltage tolerance of the LEDs. In our example, Vout is 22 to 26 Vdc. So at the high end of V_{CC}, the turns ratio can be no more than

$$\left(\frac{20\ V_{CC}}{26\ V_{OUT}}\right) = 0.77$$

and at the low end no less than

$$\left(\frac{10.2\ V_{CC}}{22\ V}\right) = 0.46.$$

To maximize our design margins, we will pick the geometric mean of the 2 values:

$$\text{Turns Ratio} = \sqrt{(0.77 \times 0.46)} = 0.6$$

So our nominal $V_{CC} = 24 \times 0.6 = 14.4$ V dc

We have placed a small resistor in series with the diodes for the V_{CC} supply. This reduces peak charging of the Vcc capacitor and helps reduce EMI.

Zero Crossing Detection

The bootstrap winding also provides valuable timing information for the NCL30002. When the voltage collapses to zero on the bootstrap winding, the current in the inductor has gone to zero. A resistor tied from the bootstrap winding to the ZCD input (Pin 5) provides this signal. The voltage signal is clamped on this pin and the clamp current must be less than ±10 mA. Maximum clamp current occurs during the switch on time at high line. The 10 mA limit is an absolute maximum for the NCL30002. A limit of ±5 mA is more practical target.

$$\text{Iclamp\_max} = \frac{(\text{Vin\_max} - \text{Vout\_min}) \times \text{Turns Ratio}}{\text{Rzcd}}$$

$$\text{Rzcd} = \frac{(\text{Vin\_max} - \text{Vout\_min}) \times \text{Turns Ratio}}{\text{Iclamp\_max}}$$

$$\frac{(132 \times 1.414 - 22) \times 0.6}{0.005\ \text{A}} = 20\text{k}$$

Small values of Rzcd provide minimum time delay and improved immunity from parasitic capacitance, however, the price for this is higher power dissipation. The Rzcd should be placed as close to Pin 5 of the NCL30002 as possible. Any parasitic capacitance on the connection to Pin 5, will cause delays in restarting the next switch cycle leading to variations in output current.

Feed Forward Control

As mentioned earlier, the inverted configuration of the buck converter makes output voltage and current measurements difficult since they are floating with respect to the control ground. We do not really need to know the output voltage (from a control point of view) since it is set by the LEDs themselves. In case of an open LED string fault, we will need some type of over voltage protection but we will discuss that later. Since we do measure the peak switch current at the source of the FET, we control the peak output current very accurately. The average output current is one half the peak current in Mode 3. The 2:1 relationship of peak and average current is a unique feature of CrM operation. The peak current threshold of the NCL30002 is tightly trimmed to achieve very accurate peak current regulation.

The set up for operation in Mode 2 is the critical part of feed forward control. Setting Ton_max to be a fixed value gives poor power factor and current regulation over line changes. The ideal relationship of Ton_max versus Vin is:

$$Ton\_max \times Vin = Constant$$

You will recognize this has a hyperbolic relationship between Ton_max and Vin. This is not easily implemented with simple circuitry. While it is possible to implement using a multiplier or Programmable Gain Amplifier, neither of these options is very cost effect given the application. When we plot the relationship of Ton_max versus Vin, we see that a simple straight line curve fit is a good approximation as shown in Figure 7.

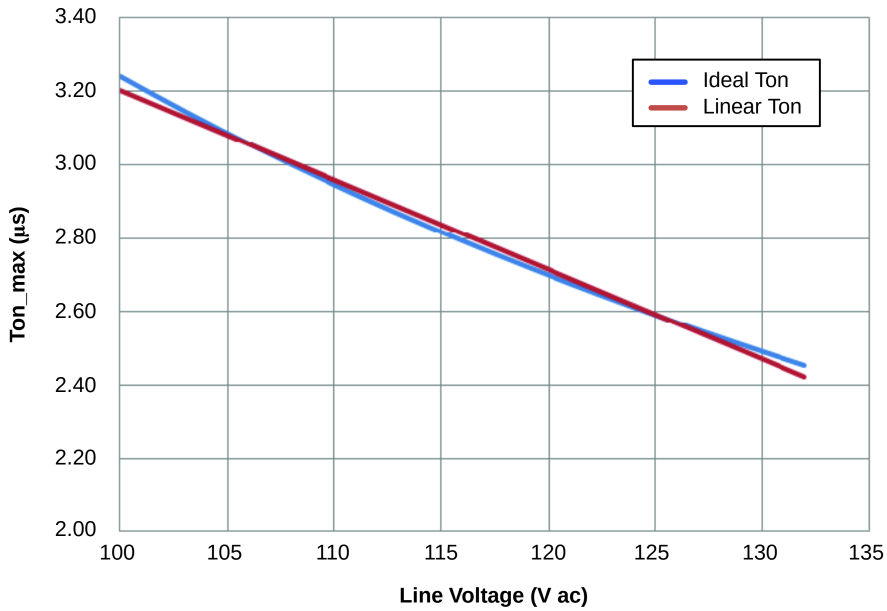

Figure 7. Ideal Ton_max vs Linear Ton_max

At this point the Microsoft Excel® based online Design Guide for the NCL30002 will become quite useful as much of the calculations are already done for you. We will discuss the principles and let the Design Guide do the heavy calculations.

From our Design Guide we get:

$$m = -0.02348\ \mu s/V \ \& \ b = 5.46\ \mu s$$

FUNCTION BOX

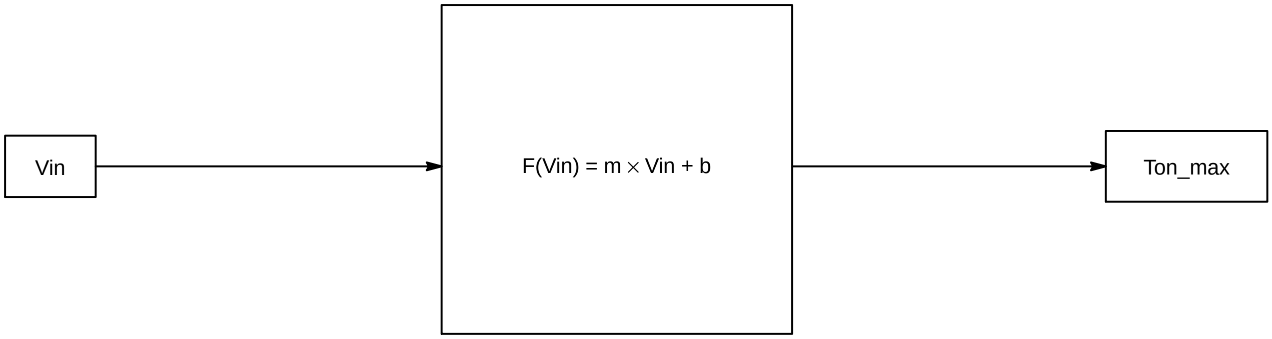

Figure 8.

Here are the hardware functions that are built in to the NCL30002 to implement our function box.

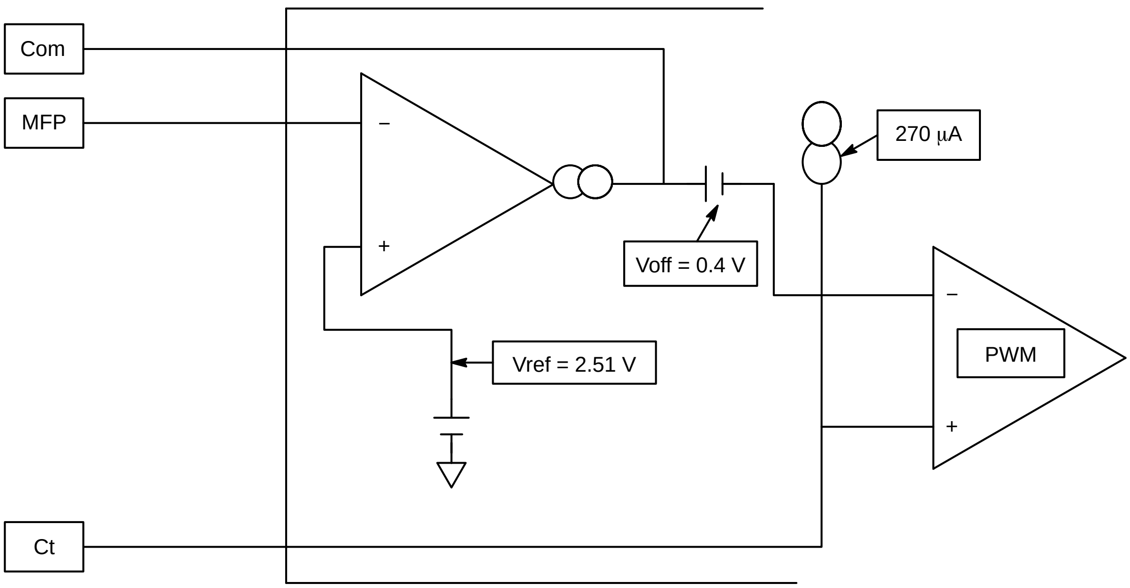

Figure 9. Simplified Block Diagram of Ton Generator

In Figure 10, you see that we are using the OTA like a traditional operational amplifier. The OTA will behave like an Op Amp if Rfb × Gm >> 1. This required Rfb to more than 100 kΩ (higher is better). The gain of the amplifier will be much less than 1. This makes the input resistor to the inverting input much higher than Rfb (approx 100x Rfb). Rin will be in order of 10 MΩ or greater. This is an impractical value because small leakage currents will cause big errors. We will do some prescaling of Vin and filtering of the 120 Hz before the amplifier. In our function box, Vin is expressed in Vrms but our circuit will use the average value of the voltage. So there is a scale factor between RMS and average that we must account for. Fortunately our Design Guide Tool has already taken these factors into consideration.

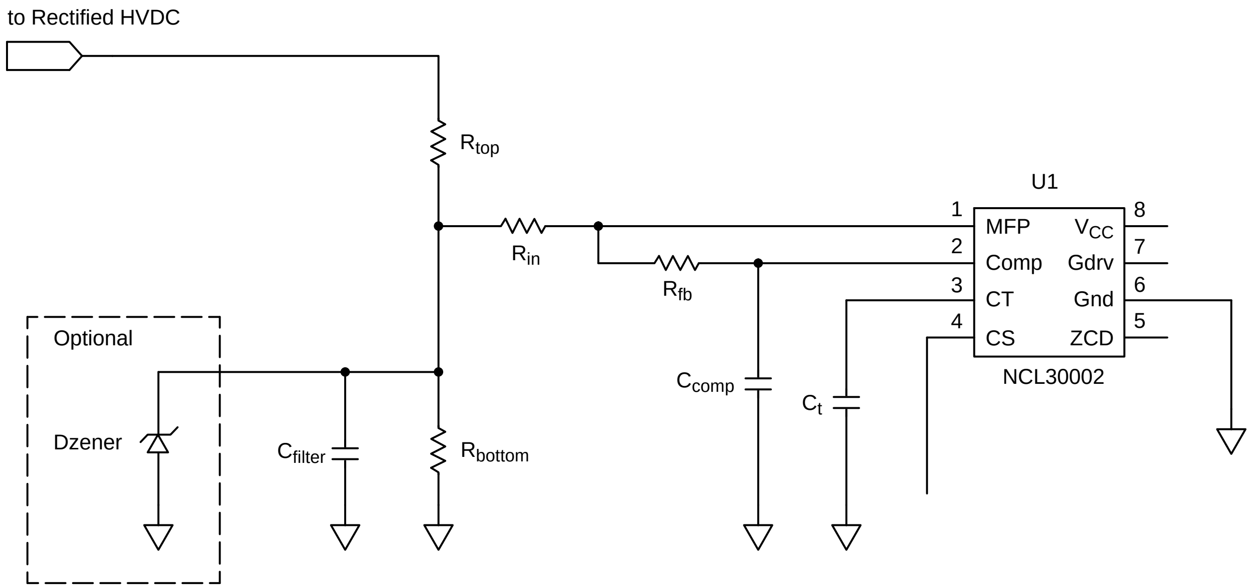

Figure 10. Partial Schematic Showing Feed Forward Control

Here are the values generated by the Design Guide for our feed forward set up.

Feed Forward Set up			Messages
Ct	474	pF	
Rsense	0.226	Ohms	
Rfb	693	K Ohms	
Rin	938	K Ohms	
Rtop	2.0	Meg Ohm	
Rbottom	64.0	K Ohms	
Cfilter	0.249	μF Min	
Dzener	4.6	Volts	Zener Required
K (Voltage Divider Constant)	0.031	K is OK	

Figure 11. View of the Feedforward Set Up from the Design Guide

Output Adjustment

Since the control is open loop, component variation will contribute to errors in the current output. Adjustments to any of the resistors in the feed forward setup will change both gain and offset. However, a bias current adjustment into the MFP pin will only affect offset without affecting gain.

Power Converter

The CrM nature of the buck converter makes the operating frequency dependent on the line, load, inductor value, and Ton_max. The Design Guide is a powerful tool that can help analyze the design trade-offs among the parameters.

Here are a few things to consider when setting up your power stage:

1. Power Factor greater than 0.9 is not really possible with a Mode 3 conduction angle greater than 108°
2. A minimum L × I will yield the smallest possible inductor
3. Delay times in the control and power components will have larger effects as the switching frequency increases much above 100 kHz and will complicate the EMI filtering design.

The peak current limit threshold is well controlled on the NCL30002. However, the desired sense resistor value may not be available. The schematic includes a parallel sense resistor that can help to make fine adjustments to the peak current and allow the use of standard component values.

Power Component Selection

Buck Diode: CrM operation allows the current in the buck diode to go to zero before the MOSFET switches on for the next cycle. This is a huge advantage over CCM since the current is not force commutated in the diode. Forced commutation requires a very fast diode to avoid high turn on losses in the diode and MOSFET as well as greatly reduced radiated EMI. The reset recovery (T_{rr}) of the diode is critical for CCM but not nearly as important for CrM. Since the T_{rr} of the diode increases with temperature, the efficiency of the CCM converter go down with increased temperature as a function of increased T_{rr}.

The average current in the buck diode is always less than the output current. The average current through the buck diode asymptotically approaches Iout as the duty cycle approaches zero. The duty cycle is low through much of the line cycle since the output voltage is much lower than the input voltage. Therefore we can use Iout as our average current rating for selection purposes. A 1 A diode does not provide much derating especially at high temperatures. A 2 A diode is the next step up in current and provides excellent derating and better thermal performance than the 1 A diode while still being small and low cost.

The maximum diode voltage stress is equal to the maximum input voltage at the maximum mains plus any surge voltage. For 132 Vac input the line peak is 191 Vdc. A 200 V diode has far too little design margin. The next step up is 400 V which has good derating for a 132 Vac application. Lower voltage diodes tend to perform better for T_{rr} and for V_f than higher voltage diodes. The MURS230T3G is an excellent choice here. The 300 V rating give a 66% voltage stress and the 2 A rating give a current stress less than 50%.

MOSFET: The power switch has the same voltage stresses as the buck diode. For 132 Vac input, a 200 Vds rating is insufficient. A 300 V device would be a good choice but there are few choices around 300 V. Even at 400 V, the MOSFET selection is limited. Fortunately, there are many choices at 500 V with low $Rds_{(on)}$ with wide selection due to demand at the 500 V voltage range.

The RMS current is in the MOSFET is 675 mA in this example. To keep efficiency high, the MOSFET needs to have an $R_{DS(on)}$ of an Ohm or less at Maximum temperature. This means that the $Rds_{(on)}$ needs to be 0.5 Ω at 25°C. The STD11NM50 meets this requirement.

Inductor: With the help of the Design Guide, we calculate a target value of 125 μH and a minimum I_{sat} of 2.1 A.

An EE16 Core will work fine and be very cost effective.

Inductor Design							
	EFD20	EFD15	EE13	EE16	RM6	Other	
Will it Fit?	Yes	No	No	Yes	Yes	Yes	
B @ Imax (Tesla)	0.30	0.30	0.30	0.32	0.30	0.30	
Area Prod Re'qd cm^4	0.026	0.026	0.026	0.031	0.026	0.026	
Area Prod Available cm^4	0.081	0.022	0.014	0.043	0.047	0.030	
Al (nH/T^{2})	157	37	25	74	157	37	
Current Density (A/cm^2)	600	600	600	600	600	600	
Fill Factor	0.5	0.5	0.5	0.4	0.5	0.5	
Turns	28	58	71	41	28	58	
Wire AWG	21	26	28	24	23	25	
DCR mOhms @ 25C	38	208	354	111	55	154	
Core Area						0.15	cm^2
Window Area						0.2	cm^2

Figure 12. Inductor Selection Guide from Design Guide

The next step is to review the corner conditions which are already entered in the Design Guide for both high line and low line conditions as well as minimum and maximum LED forward voltage.

F G H	I J	K L	M N	O P	Q R	S T	U
	Low Line		**Typical Line**		**High Line**		
	Vf Max	**Vfmin**	**Vf Max**	**Vfmin**	**Vf Max**	**Vfmin**	
Power Factor	0.977	0.961	0.955	0.955	0.967	0.950	
RMS Input Current	190	168	168	146	152	134	mA
Avg Output Current	713	735	741	759	748	764	mA
Max SW Freq	312	311	368	366	412	412	kHz
Avg SW Freq	95	84	99	91	107	94	kHz
Output Power	18.6	16.2	19.3	16.7	19.4	16.8	Watts
Ipeak Switch	2.1	2.1	2.1	2.1	2.1	2.1	A
Peak output Rect. Current	2.1	2.1	2.1	2.1	2.1	2.1	A

Statistics	**Vled Min**	**Vled Max**	
Mean Output Current	753	734	mA
Upper Limit	1.5%	1.9%	
Lower Limit	-2.3%	-2.8%	
Fsw Avg	90	100	kHz

Figure 13. View of the Calculated Results from the Design Guide

We see that the power factor should be well above 0.9 for all of our various line and output voltage conditions. The output current is close to our target and is nicely regulated over the line range. The average switching frequency is also close to the target. As expected, the switching frequency does vary depending on the line and load conditions.

Protection

Over Voltage and Short Circuit Protection: The application for this LED driver is a dedicated driver for a specific bulb or luminaire configuration. So an over voltage condition would only exist if the load fails open. In this case, the requirement of protection is only a safety concern not a functional concern since the load has already failed. A simpler zener diode will prevent the voltage from rising too high in the output and possibly venting the output capacitor. The zener will eventually fail short causing a short circuit on the output. An output short is not a particularly stressful condition for the driver. The controller gets V_{CC} reflected from the load voltage through the output inductor. If the output is shorted, the bootstrap doesn't work and the controller will hiccup as the V_{CC} capacitor charge and discharges. The duty cycle is very low in this condition.

Over Temperature Protection: While not shown on the basic schematic, over temperature protection can be implemented any number of ways. One implementation is shown below and is intended to fold-back the current in the event the bulb is installed in an environment where the operating temperature will impact lifetime. Foldback is preferred over a thermal shutdown because light is still generated at a reduced level and the power dissipation is reduced proportionally as well. With the addition of a simple transistor, a few resistors and a positive temperature coefficient resistor, this can be easily implemented.

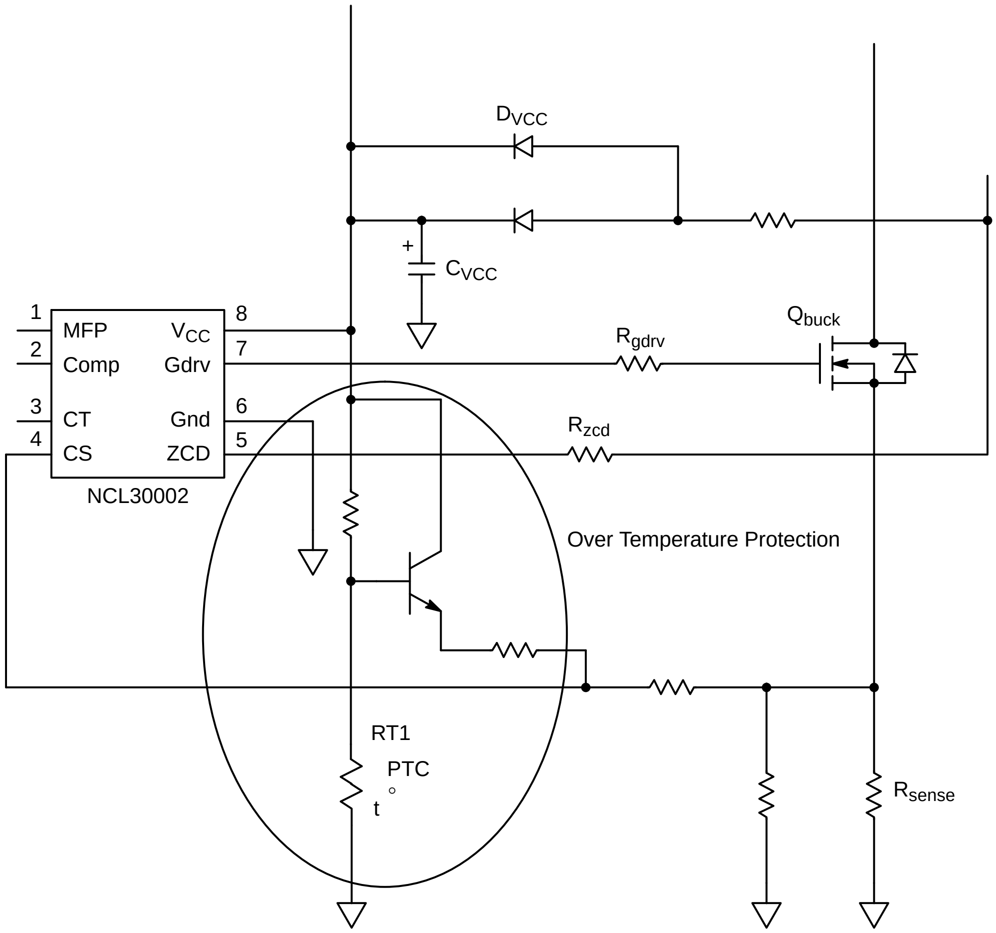

Figure 14. Over Temperature Foldback Protection

Choose the value of the base resistor and the PTC to reduce the current threshold above a specified temperature for the PTC.

EMI Filter

The EMI filter needs special consideration here due to the negative impedance of the buck converter. The negative impedance is worst at low line where the input current is highest.

During mode 3, input power is constant.

$$\mathrm{Vin} \times \mathrm{Iin} = \mathrm{Pin}$$

$$\mathrm{Iin} = \frac{\mathrm{Pin}}{\mathrm{Vin}}$$

$$\mathrm{dIin} = -\frac{\mathrm{Pin}}{\mathrm{v2}}\mathrm{dVin}$$

$$\frac{\mathrm{dVin}}{\mathrm{dIin}} = -\frac{\mathrm{Vin2}}{\mathrm{Pin}}$$

From our equation, the negative impedance is lowest at Vin = minimum and Pin = maximum. If the conduction angle is 90 degrees, then:

$$\mathrm{Vin} = 100 \times 1.414 \times \mathrm{Sin}(45^{\circ})$$

$$\mathrm{Vin} = 141.4 \times 0.707 = 100\ \mathrm{V}$$

If Pin = 22 W (considering the efficiency of the converter)

$$\mathrm{Rin} = -\frac{100\ \mathrm{V} \times 100\ \mathrm{V}}{22\ \mathrm{W}} = -454\ \Omega$$

The positive impedance of the EMI filter plus the power line impedance must be less than 454 Ω under all conditions or the power converter will oscillate.

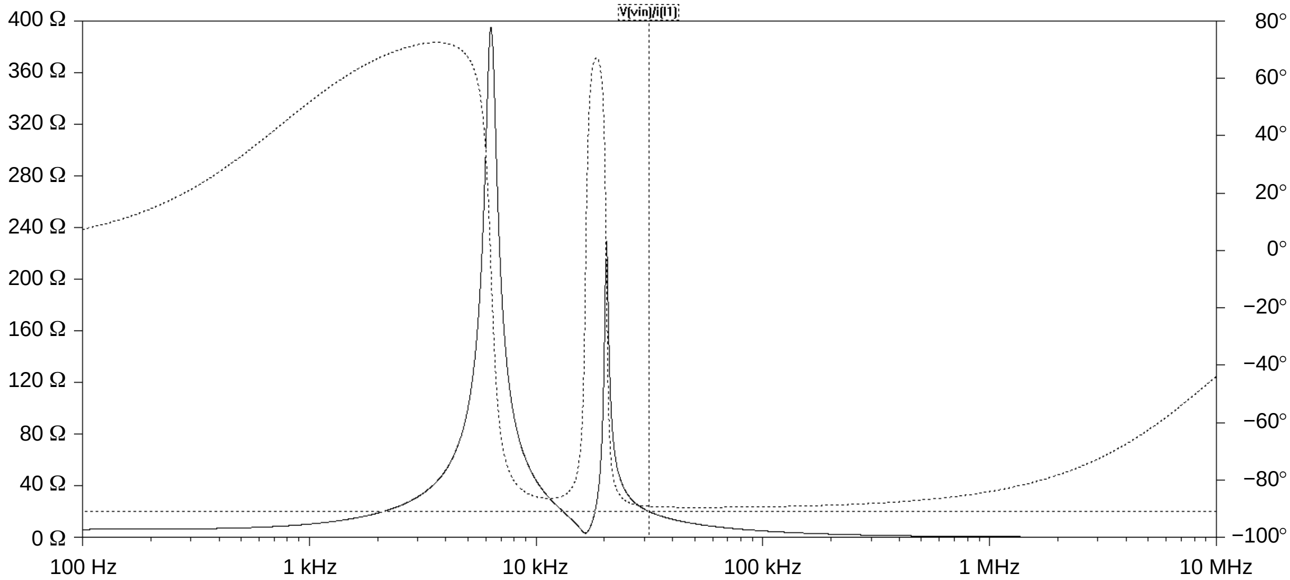

Figure 15. EMI Filter Simulation Impedance Plot

The impedance of the filter is always less than the negative impedance of the converter so the system is stable. In general, the larger X-caps and smaller series inductors tend toward lower EMI filter impedances. For compactness, the capacitors are on the DC side of the bridge rectifier so no special safety ratings are required. The filter chosen is a four pole filter which attenuates the noise to CISPR Class B levels.

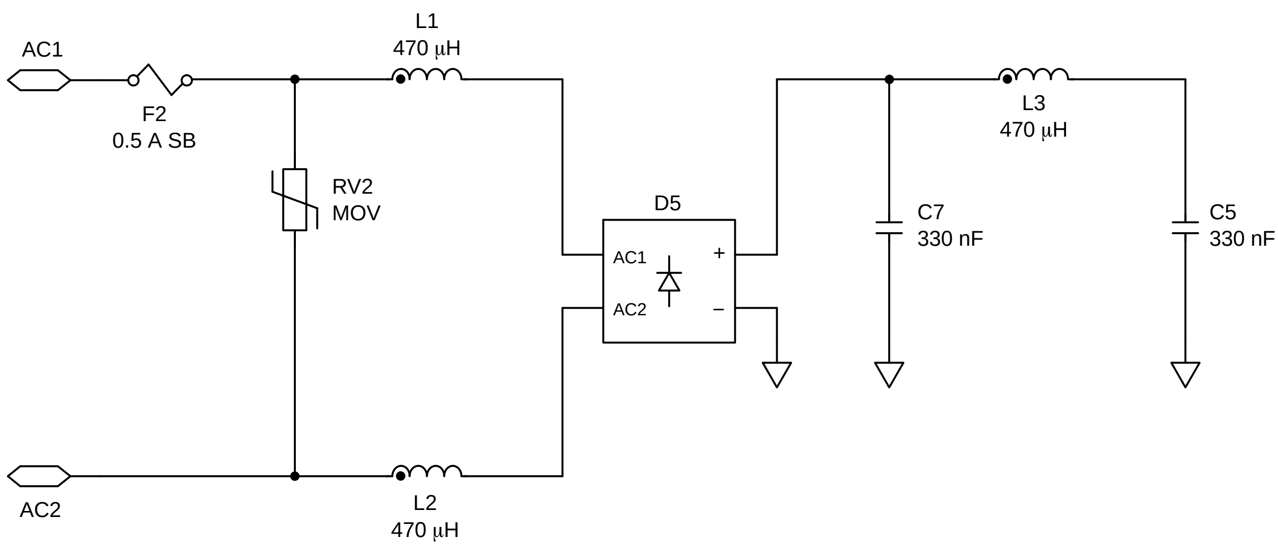

Figure 16. EMI Filter Schematic

Test Results

Let's see compare the results against the target design specifications.

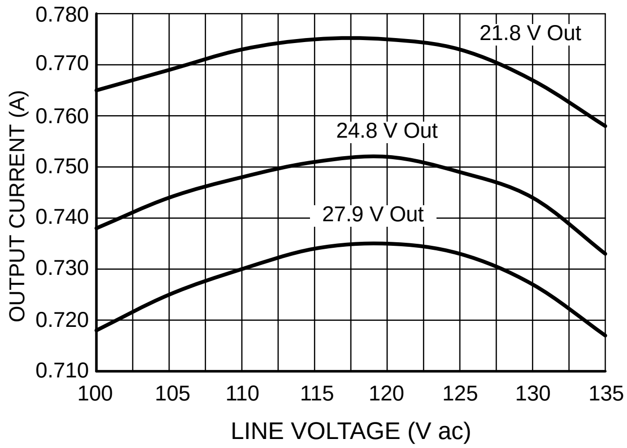

Figure 17. Output Current over Line Voltage

We can see that the current is well controlled against line changes. Conveniently with this design, the output current has a negative tracking coefficient. This limits the power in the event that the Vf is higher than expected. In this case the green and blue LED forward voltages are outside the design spec of 22−26 V dc, but the current is still tightly regulated around 750 mA.

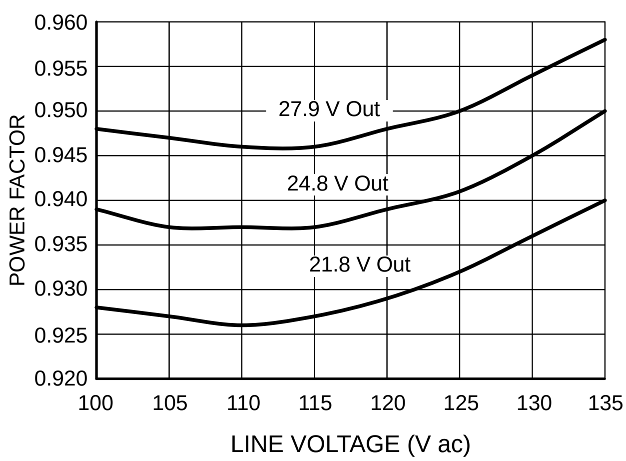

Figure 18. Power Factor over Line Voltage

The Design Guide tool does not take into account contribution of the EMI filter to the power factor but is reasonable accurate and the actual results are still well above 0.9.

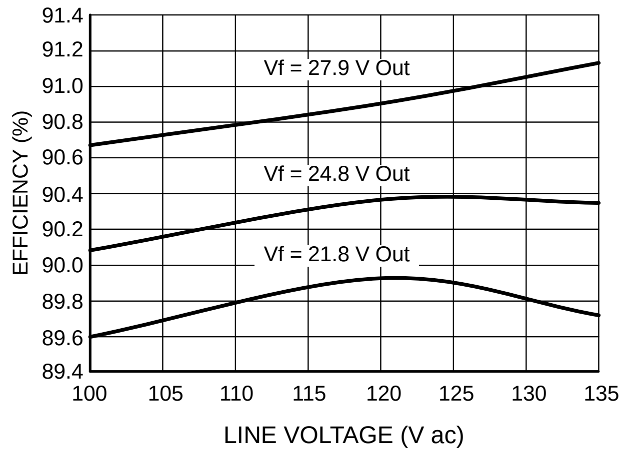

Figure 19. Efficiency over Line Voltage and Load

The efficiency improves at higher output voltages and power levels as expected but even at the lowest output power we are exceeding our target of 88% for all conditions.

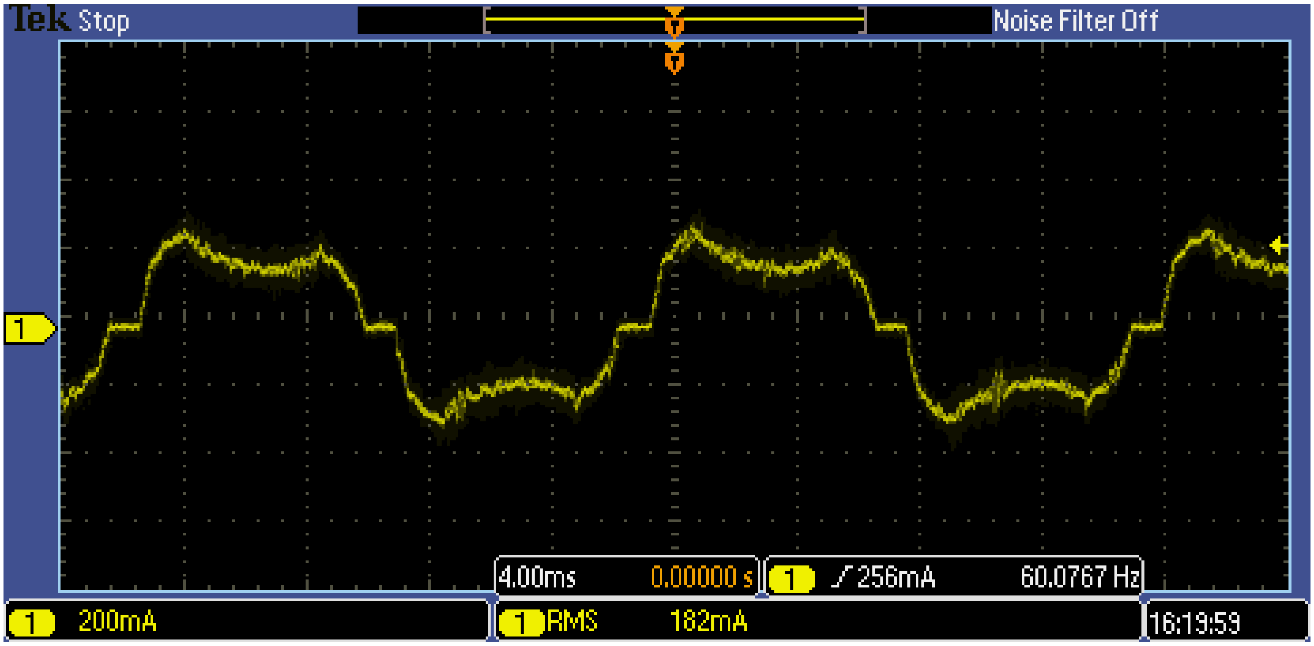

Figure 20. Typical Input Current

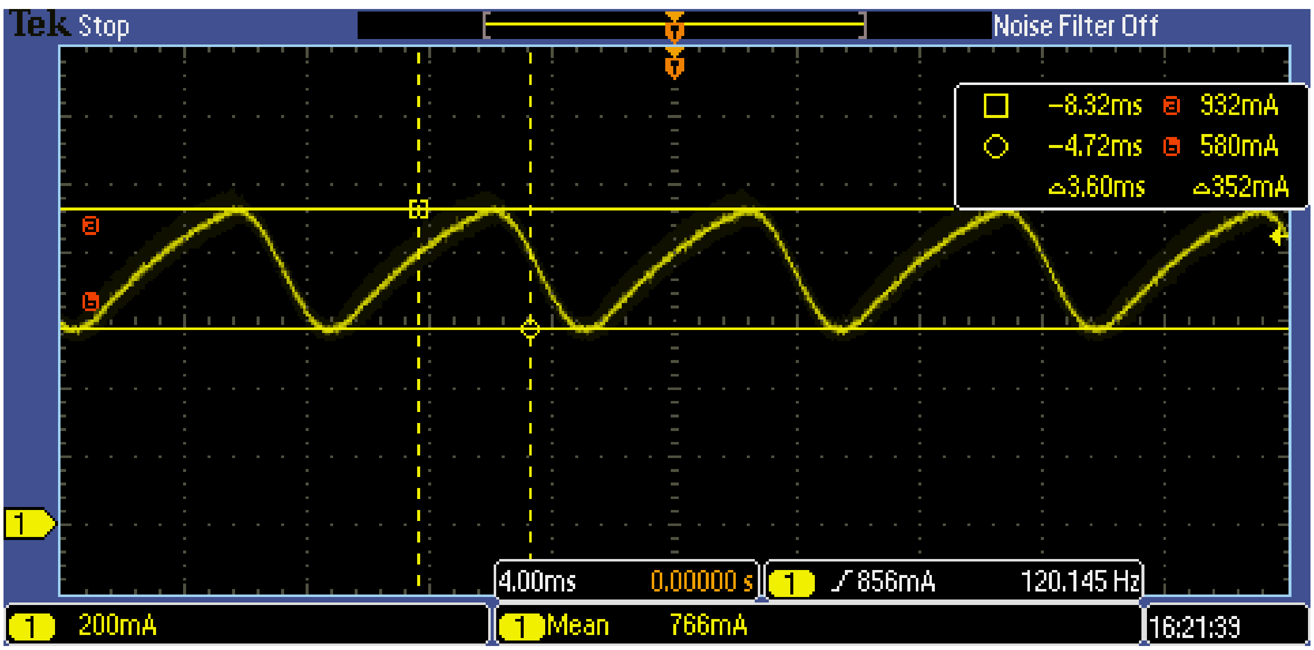

Figure 21. Typical Output Current

Conclusion

This application illustrates the design procedure for a high power factor corrected buck LED driver along with actual characterization across line and load for an optimized 8 LED driver application. This simple circuit architecture is capable of delivering load regulation better then ±5% over more than the variation expected for a single line range while delivering high power factor and high efficiency. The output ripple is well below the 1 A DC rating of the LEDs. Key protection features are incorporated in the circuit to address both open and shorted LED output faults. Finally an optional thermal fold-back circuit can be easily incorporated into the design to protect the LED bulb in the event it is inadvertently installed in an environment where its upper temperature range is exceeded.

APPENDIX: SCHEMATIC AND BILL OF MATERIALS

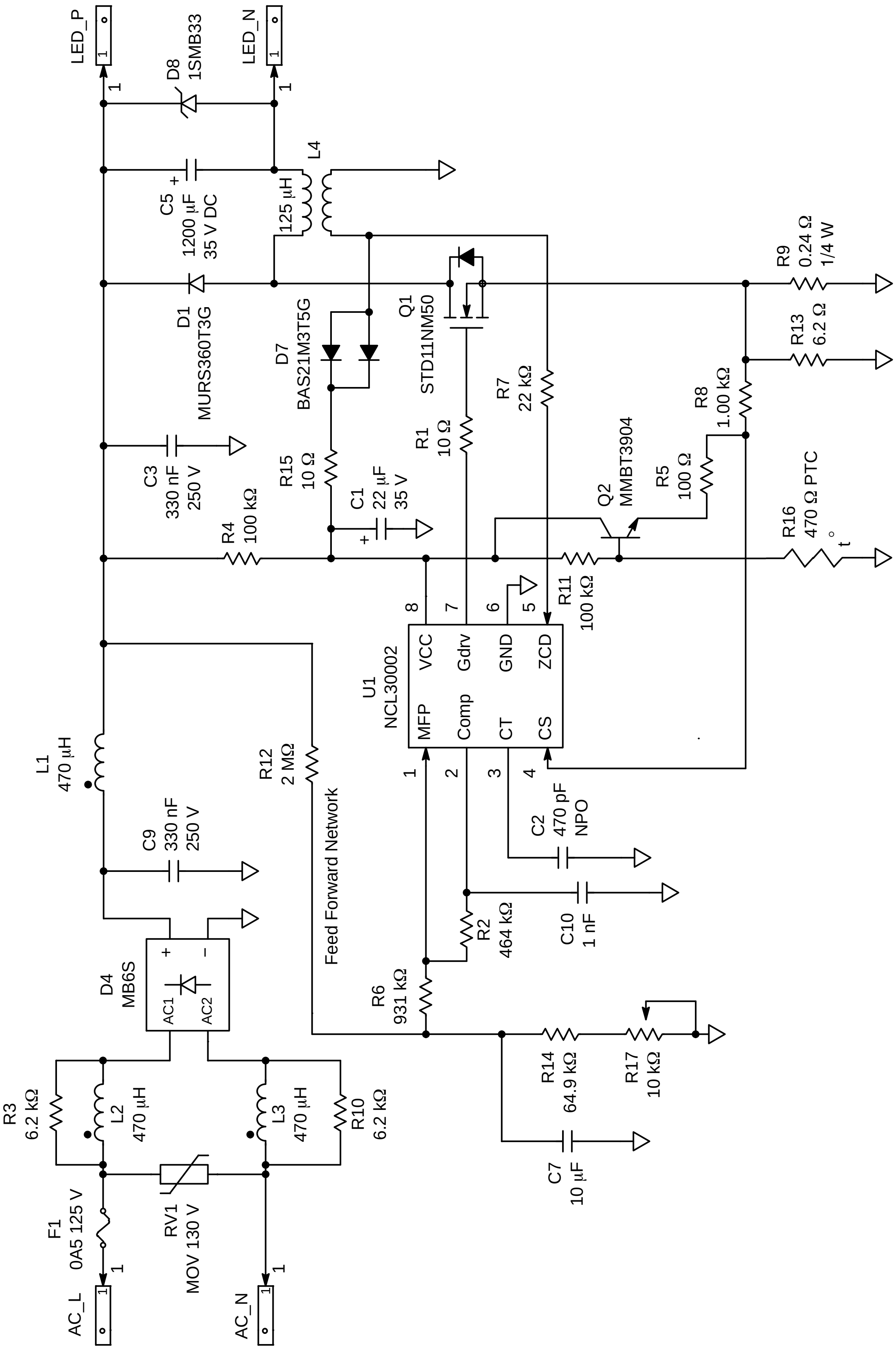

Figure 22. Schematic

Table 2. BILL OF MATERIALS

Reference	Value	Size/Footprint	Manufacturer	Part Number
C1	22 μF 35 V	1210	Various	−
C2	470 pF 50 V NP0 2%	603	Various	−
C3,C9	330 nF 250 V	Radial LS = 5 mm	Epcos	B32529C3334M***
C5	1200 μF 35 V	Radial LS = 7.5 mm	Nichicon	UHE1V122MHD6
C7	10 μF X7R 10 V 10%	603	Various	−
C10	1 nF 25 V NP0 20%	603	Various	−
D1	MURS360	SMB	ON Semiconductor	MURS360BT3G
D4	MB6S	4 Pin SMD	MCC	MB6S
D7	BAS21DW5	SC88A (SC70−5)	ON Semiconductor	BAS21DW5T1G
D8	1SMB33	SMB	ON Semiconductor	1SMB33CAT3G
F1	0.5 A Slow Blow	Axial	Littelfuse	0449.500MRT1L
L1, L2, L3	470 μH	Radial LS = 2.5 mm	Wurth	7447462471
L4	125 μH	EE16 Core	Wurth	750313017
Q1	STD11NM50	DPAK	ST	STD11NM50
Q2	MMBT3904	SOT−23	ON Semiconductor	MMBT3904
R1, R15	10 Ω 1%	603	Various	−
R2	464 kΩ 1%	603	Various	−
R3, R10	6.2 kΩ 5%	805	Various	−
R4	100 kΩ 1%	1206	Various	−
R5	100 Ω 1%	603	Various	−
R6	931 kΩ 1%	603	Various	−
R7	22 kΩ 5%	1206	Various	−
R8	1 kΩ 1%	603	Various	−
R9	0.24 Ω	1206	Various	−
R11	100 kΩ 1%	603	Various	−
R12	2 MΩ 1%	805	Various	−
R13	6.2 Ω 5%	603	Various	−
R14	64.9 kΩ 1%	603	Various	−
R16	470 Ω	603	Epcos	B59601A0115A062
R17	10 kΩ	Gull Wing Top Mnt	TTI	22AR10KLFTR
RV1	130 V	Disc 7mm	Littelfuse	V130LA2P
U1	NCL30002	SOIC−8	ON Semiconductor	NCL30002DR2G

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