

TI Designs

Driving Daytime Running Lights LEDs With Thermal Foldback Reference Design



1 Overview

The TIDA-01382 implements thermal foldback for derating current through the LEDs without using a microcontroller. This TI Design uses the following devices: the TPS92691-Q1 multi-topology LED driver in boost configuration to control the LEDs, the TLC555-Q1 LinCMOS™ timer together with the OPA2377-Q1 operational amplifier to measure and generate the accurate PWM signal by applying a feedback loop and a precision shunt regulator for setting the accurate duty cycle, and the LMT87-Q1 to implement thermal foldback. The input and output stage of the design is EMI- and EMC-filtered and can be directly supplied by a car battery.

2 Resources

TIDA-01382	Design Folder
TPS92691-Q1	Product Folder
LMT87-Q1	Product Folder
TLC555-Q1	Product Folder
OPA2377-Q1	Product Folder
TL431A-Q1	Product Folder



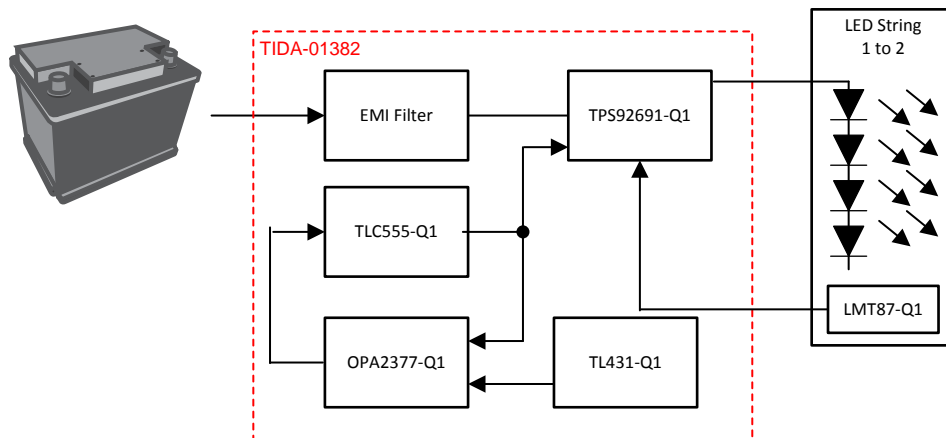
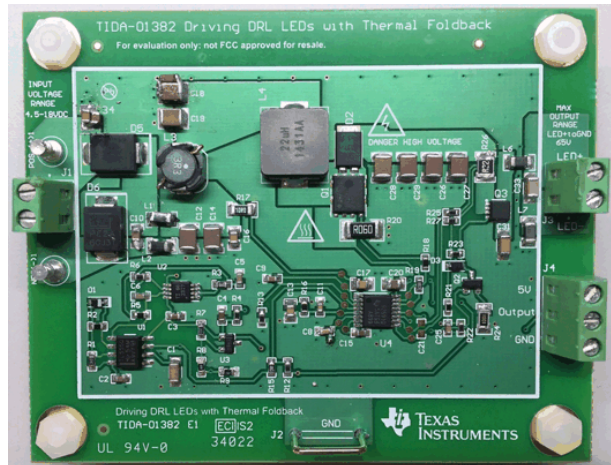
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3 Features

- Thermal Foldback
- Precision PWM Dimming
- Efficiency-Optimized Design
- Operation Through Cold Crank
- Load Dump Tolerant

4 Applications

- Automotive Front Lighting
- Automotive Daytime Running Lights
- Automotive Tail and Brake Lights



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5 System Overview

5.1 System Description

This system has been designed to be a solution to precision pulse width modulation (PWM) dimming daytime running lights and implement thermal foldback without the necessity of using a microcontroller (MCU). The design includes key peripherals like electromagnetic interference (EMI) and electromagnetic compatibility (EMC) filtering voltage conditioning (shunt regulator), thermal foldback, precision clock generation, and LED drive.

The TIDA-01382 has been designed with the following points in consideration:

- The design must be able to generate a precision PWM signal in the range of 5% to 50% duty cycle.
- The design must be able to implement thermal foldback.
- The design must satisfy power requirements for one TPS92691 device driving a string of 1 to 12 LEDs for daytime running lights
- The design operate over the full range of automotive battery conditions:
 - $V_{IN(min)}$ down to 5 V simulating a cold-cranking condition (ISO 7637-2:2004 pulse 4)
 - $V_{IN(max)}$ up to 18 V simulating the upper range of normal battery operation
- The design must survive and continue operation through:
 - Load dump (ISO 7637-2:2004 pulses 5a)
 - Double battery condition
- The output must be protected against short-to-battery and GND voltage.
- The design must optimize the individual blocks for smallest power dissipation and highest efficiency.
- The layout of the board must be set up in such a way to minimize the footprint of the solution while maintaining high performance.
- The design must provide a flexible board interface to either mate to custom board through screw terminals
- The design provide power for the TLC555-Q1, OPA2377, and LMT87-Q1.

5.2 Key System Specifications

Table 1. Electrical Characteristics

PARAMETER		COMMENTS	MIN	TYP	MAX	UNIT
SYSTEM INPUT AND OUTPUT						
V_{IN}	Operating input voltage	Battery-voltage range; outputs are functional	5	14	18	V
V_{UVLO}	Input UVLO setting	Undervoltage lockout (UVLO)	—	4.5	—	—
V_{SWMmax}	Vmax switch	Maximum switch node voltage	—	—	100	V
V_{OUT}	Output voltage	LED+ to LED– (Boost)	21	—	60	V
V_{OUT}	Output voltage	LED + to V_{IN} (Boost-to-Battery)	3	—	36	—
V_{TR}	Transient immunity	Load dump (ISO7637-2)	—	—	60	V
V_{IN_MIN}	Minimum input voltage	Cold crank (ISO 7637-2)	5	—	—	V
I_{IN}	Input current	Output at full load	—	2	—	A
I_{OUT}	Output current	Maximum current per string	—	350	365	mA
Maximum output power	—	—	—	—	25	W
PWM dimming range	240-Hz PWM frequency	—	—	20:1	—	—
LED _{Open} and short detect	LED open and short detection	—	—	Yes	—	—
LED _{Single short detect}	LED single-short detection	—	—	No	—	—
ONBOARD VOLTAGES						
V_{5V5}	Auxiliary supply, shunt regulator	TLC555, LMT87-Q1, op amp supply, and reference generation	—	5	—	V
VCC	Bias	Supply shunt regulator (TLE431-Q1)	—	5	—	V
V_{PREC_PWM}	TLC555 out	Amplitude TLC555 clock at output	—	5	—	V
CLOCKS						
f_{PREC_PWM}	Square wave frequency	Frequency at TLC555-Q1 out, U1	240	—	—	Hz
D_{OUT}	Square wave duty cycle	Duty cycle of f_{PREC_PWM}	5	—	50	%
D_{ACC}	Duty cycle accuracy	Accuracy of D_{OUT}	2%	—	—	—
f_{OSCL}	Oscillator frequency	LED driver, TPS92691-Q1	—	390	—	kHz
THERMAL						
T_A	Temperature range	Operating and ambient temperature	–40	—	125	°C
PULSE TOLERANCE						
Cold crank	Operational					
Jump start	Operational					
Jump start	Operational					
BASEBOARD						
Number of layers	Two layers, single-side populated					
Form factor	71 mm × 51 mm					

5.3 Block Diagram

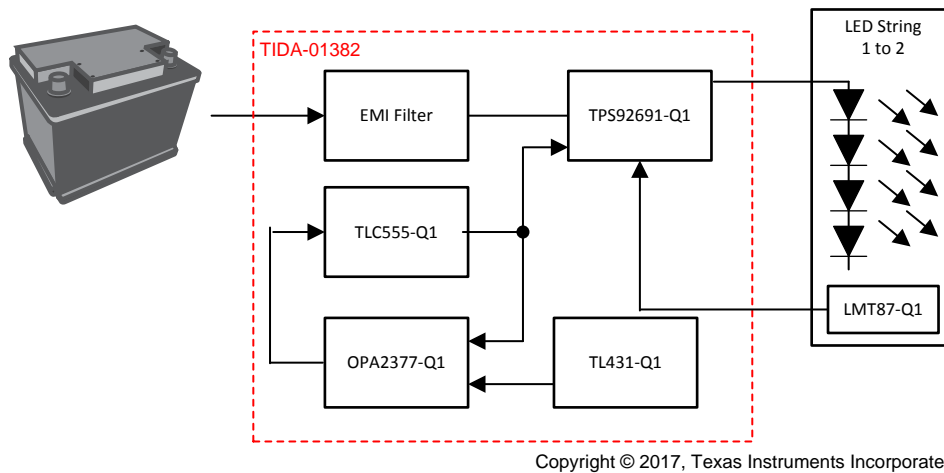


Figure 1. TIDA-01382 Block Diagram

5.4 Highlighted Products

5.4.1 LMT87-Q1

The LMT87-Q1 is a precision CMOS integrated-circuit temperature sensor with an analog output voltage that is linearly and inversely proportional to temperature. Its features make it suitable for many general temperature sensing applications. It can operate down to a 2.7-V supply with a 5.4- μ A power consumption. Package options including a through-hole TO-92 package allows the LMT87-Q1 to be mounted onboard, off-board, to a heat sink, or on multiple unique locations in the same application. A class-AB output structure gives the LMT87-Q1 a strong output source and sink current capability that can directly drive up to 1.1-nF capacitive loads. This means it is well suited to drive an analog-to-digital converter sample-and-hold input with its transient load requirements. It has accuracy specified in the operating range of -50°C to 150°C . The accuracy, three-lead package options, and other features also make the LMT87-Q1 an alternative to NTC or PTC thermistors

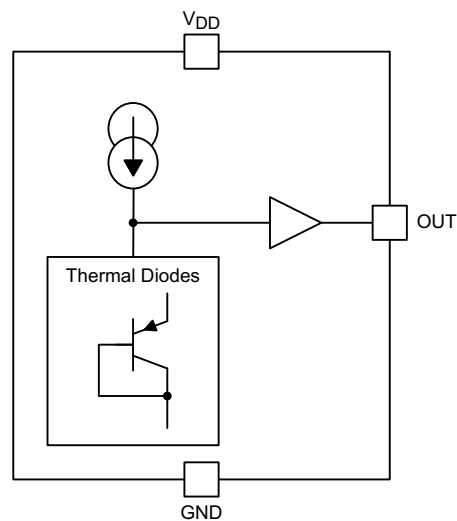
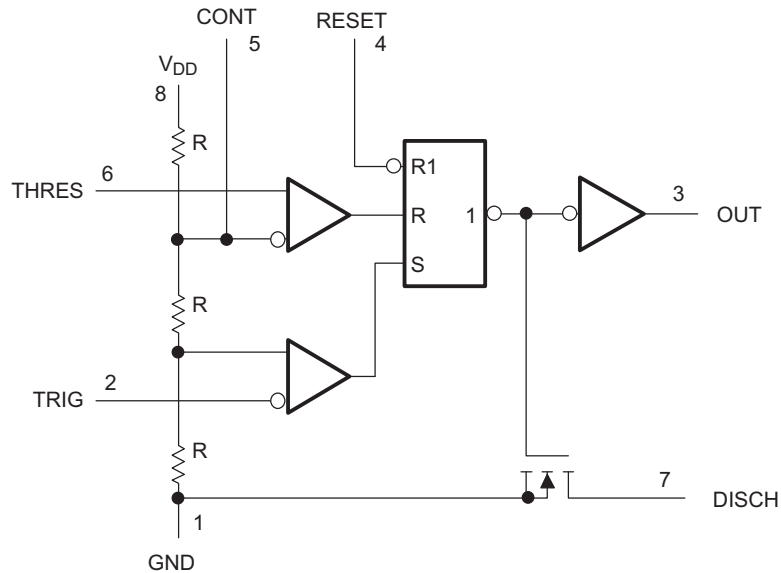


Figure 2. LMT87-Q1 Functional Block Diagram

5.4.2 TLC555-Q1

The TLC555 is a monolithic-timing circuit fabricated using the TI LinCMOS process (see [Figure 3](#)). The timer is fully compatible with CMOS, TTL, and MOS logic and operates at frequencies up to 2 MHz. Because of the high input impedance of this device, it uses smaller timing capacitors than those used by the NE555 device. As a result, more accurate time delays and oscillations are possible. Power consumption is low across the full range of power-supply voltage. The advantage of the TLC555-Q1 is that it exhibits greatly reduced supply-current spikes during output transitions. Although the CMOS output is capable of sinking over 100 mA and sourcing over 10 mA, the main reason the TLC555-Q1 is able to have low current spikes is because of its edge rates. This feature minimizes the requirement for the large decoupling capacitors required by the NE555.



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Figure 3. TLC555-Q1 Functional Block Diagram

5.4.3 OPA2377-Q1

The OPA2377-Q1 is a wide-bandwidth CMOS amplifier that provides very low noise, low input bias current, and low offset voltage while operating on a low quiescent current of 0.76 mA (typical).

The OPA2377-Q1 operational amplifier (op amp) is optimized for low voltage, single-supply applications. The exceptional combination of AC and DC performance makes the device ideal for a wide range of applications, including small signal conditioning and active filters. In addition, this part has a wide supply range with excellent power supply rejection ratio (PSRR), which makes it appealing for applications that run directly from batteries without regulation.

Figure 4 shows a block diagram of the OPA2377-Q1 op amp.

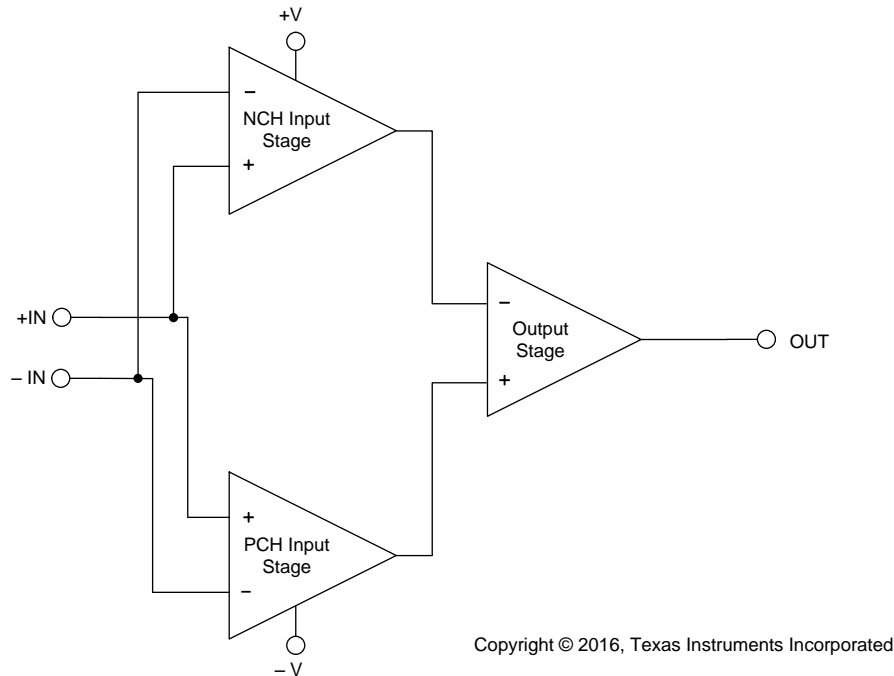


Figure 4. OPA2377-Q1 Functional Block Diagram

5.4.4 TL431-Q1

The TL431-Q1 is a three-terminal adjustable shunt regulator with specified thermal stability over applicable automotive temperature ranges (see Figure 5). The output voltage can be set to any value between V_{REF} (approximately 2.5 V) and 36 V, with two external resistors. This device has a typical output impedance of 0.2 Ω . Active output circuitry provides a sharp turnon characteristic, making this device an excellent replacement for Zener diodes in many applications, such as onboard regulation, adjustable power supplies, and switching power supplies.

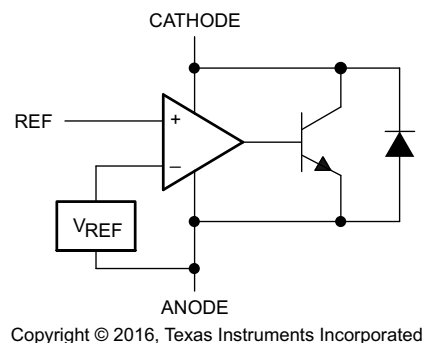


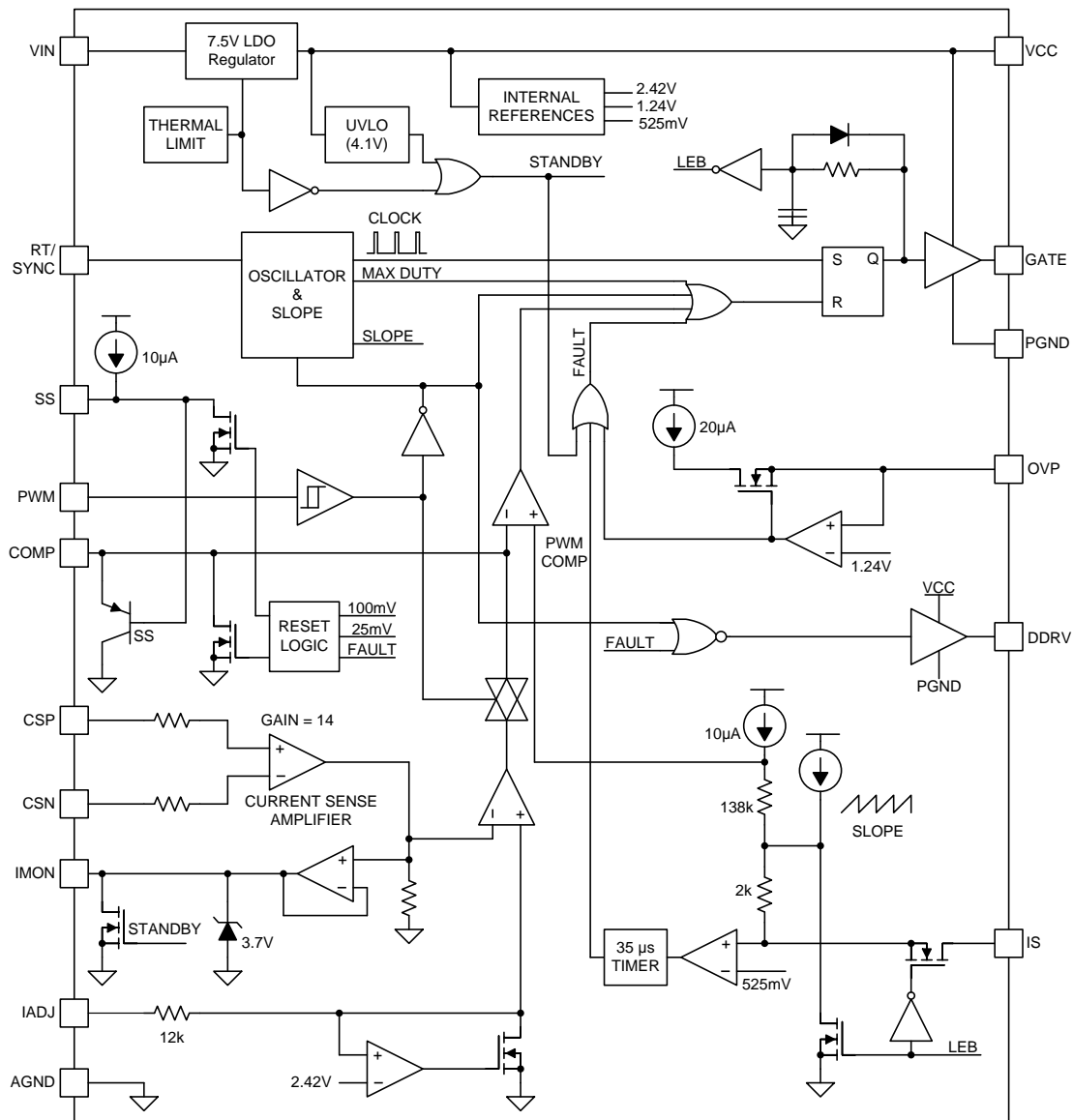
Figure 5. TL431-Q1 Functional Block Diagram

5.4.5 TPS92691-Q1

The TPS92691-Q1 is a versatile LED controller that can support a range of step-up or step-down driver topologies (see Figure 6). The device implements a fixed-frequency, peak-current-mode control technique with programmable switching frequency, slope compensation, and soft-start timing. The device incorporates a high-voltage (65-V) rail-to-rail current sense amplifier that can directly measure LED current using either a high-side or a low-side series sense resistor. The amplifier is designed to achieve low input offset voltage and attain better than $\pm 3\%$ LED current accuracy over a junction temperature range of 25°C to 140°C and output common-mode voltage range of 0 to 60 V.

LED current can be independently modulated using either analog or PWM dimming techniques. A linear analog dimming response with a 15:1 range is obtainable by varying the voltage from 140 mV to 2.25 V across the high impedance analog adjust (IADJ) input. PWM dimming of LED current can be achieved by modulating the PWM input pin with the desired duty cycle and frequency. Use the DDRV gate driver output to enable series FET dimming functionality to obtain over a 1000:1 contrast ratio.

The TPS92691-Q1 supports continuous LED status check through the current monitor (IMON) output. This feature allows for LED short circuit or open circuit detection and protection. Additional fault protection features include VCC UVLO, output OVP, switch cycle-by-cycle current limit, and thermal protection.



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Figure 6. TPS92691-Q1 Functional Block Diagram

6 System Design Theory

6.1 PCB and Form Factor

This TI Design is not intended to fit any particular form factor. The specific and primary objective of the design with regards to the PCB is to make a solution that is compact, while still providing a way to test the performance of the board. [Figure 7](#) shows a 3D rendering of the board.

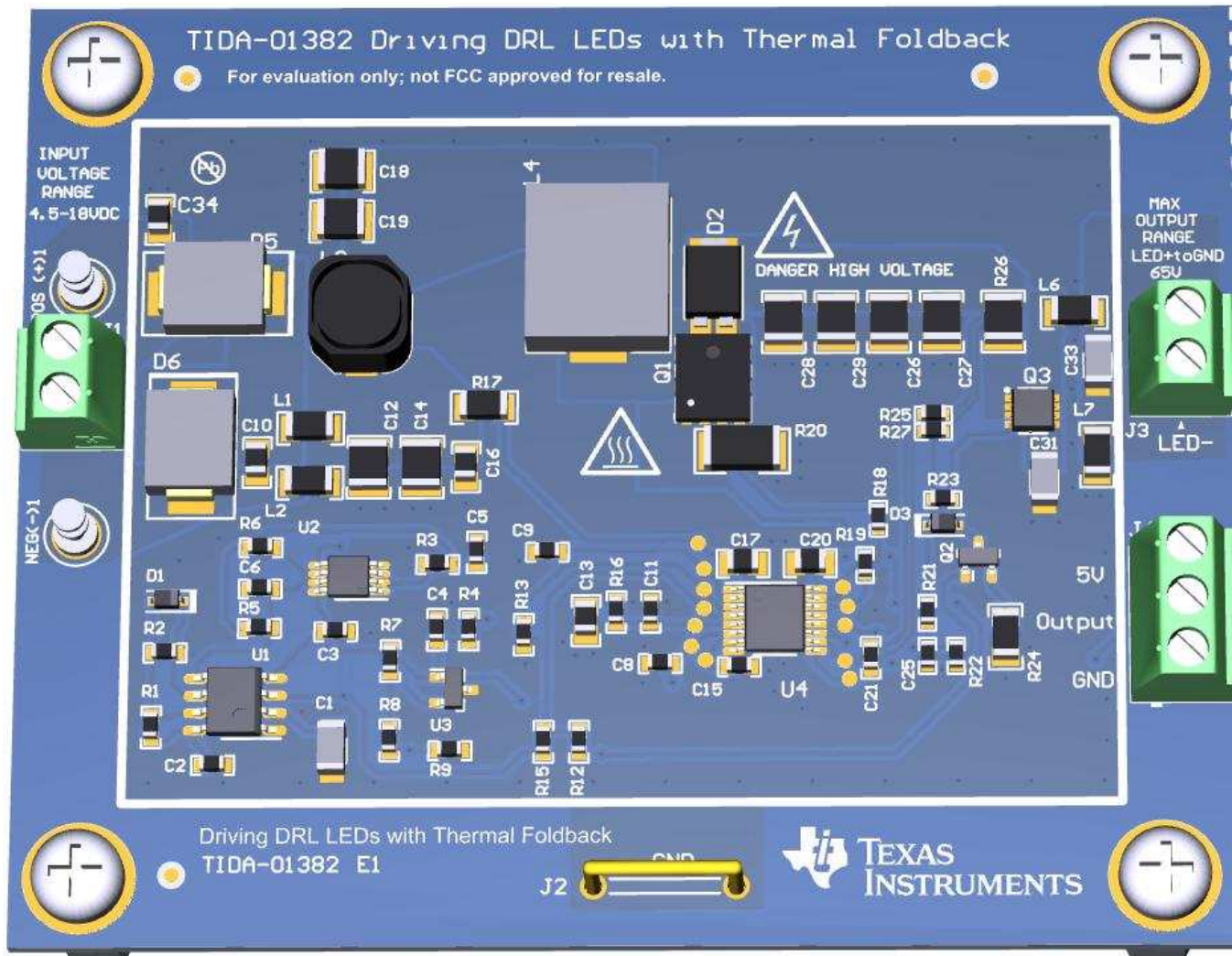


Figure 7. 3D Render of TIDA-01382 Board

In a final-production version of this TI Design, several techniques may be used to reduce the size of the solution:

- Test points, headers, sockets, standoffs, and banana plugs can be removed because they do not service a direct function for the board.
- The number, size, and value of capacitors in the system can be optimized.
- The application may not require an input-conducted emissions EMI (PI) filter.

6.2 Optimizing Board Performance Based on LED String Voltage and Current

The default board schematic has been configured to operate over a wide range of LED currents (45.4 to 730.5 mA) and string configurations (1 to 20 LEDs). The driver operation, efficiency, and transient response can be improved by reconfiguring the schematic for a given LED current and LED string forward-voltage drop. The LED current sense resistor ($R_{CS} = R26$) value can be calculated based on the maximum allowable differential voltage of 172 mV, which is achieved by pulling the IADJ pin to VCC through an external resistor. The slope compensation voltage can be adjusted by changing the switch current sense resistor, $R_{IS} = R20$, based on the maximum expected LED stack voltage. The proportional integral compensation network can be tuned to achieve high bandwidth and desired phase margin for a specified range of input and output voltages. See the TPS92691-Q1 data sheet [2] for more details.

6.3 Switching Frequency

The switching frequency of the design is set for 390 kHz set by R13 and is below the AM band. The equation to choose R13 based on a switching frequency, f_{sw} (Hz), is shown in Equation 1:

$$R13 = 1.432 \times \frac{10^{10}}{(f_{sw})^{1.047}} \quad (1)$$

6.4 Output Overvoltage Protection (OVP)

As automotive environments are harsh, good designs incorporate protection features against harsh electrical transient scenarios. This TI Design incorporates output overvoltage protection (OVP) in the event of an output open circuit. This OVP prevents damaging the output capacitors, the rectifier diode, the switching field effect transistor (FET), the CSP pin, and the CSN pin of the device. Pins V_{IN} , V_{CSP} , and V_{CSN} have a maximum voltage of 65 V, which means the OVP should be set so that the maximum output voltage does not reach 65 V. Also the resistors must be set to allow a hysteresis that will not go above that 65 V. The following equations show how to choose the values for the resistor divider going into the OVP pin.

$$R8 = \frac{V_{OV(HYS)}}{(20 \times 10^{-6})} = \frac{10 \text{ V}}{(20 \times 10^{-6})} = 500 \text{ k}\Omega \quad (2)$$

$$R9 = R8 \times \left(\frac{1.24}{(V_{O(OV)} - 1.24)} \right) = 500 \text{ k}\Omega \times \left(\frac{1.24}{(62 - 1.24)} \right) = 10 \text{ k}\Omega \quad (3)$$

6.5 Current Monitoring (IMON)

If current monitoring is desired, the IMON pin on the TPS92691-Q1 can be read by an ADC of the MCU and then the device can be shut off when a fault is detected. The IMON pin represents 14 times the voltage $V_{CSP-CSN}$. If the TPS92691-Q1 is not interfaced with an MCU for overcurrent protection, there is a discrete way to enable this feature. If current monitoring would rather be done discretely, see the TPS92691-Q1 data sheet.

6.6 Thermal Foldback

Temperature is a concern when driving the DRL LEDs at high currents in an automotive environment. Take care at high temperatures so as to not exceed the LED operating temperature requirements. For this reason, the current going through the LEDs must be decreased when the LED temperature begins to increase in order to cool down the LEDs before they are damaged. The way this is done is by using the LMT87-Q1 TI Temperature Sensor fed into the "IADJ" pin of the TPS92691-Q1. The output of the LMT87-Q1 is approximately modeled as Equation 4.

$$V = -13.6 \text{ mV}/^\circ\text{C} \times T + 2637 \text{ mV} \quad (4)$$

Where V is the output voltage and T is the temperature. Find more information about the output of the LMT87-Q1 at its [product page](#).

The "IADJ" pin of the TPS92691-Q1 acts as a current limiter for the LEDs. The output current going through the LEDs is read by shunting R26 in [Figure 22](#) to the current sense amplifier, "CSP" and "CSN". This voltage drop, $V_{(CSP-CSN)}$, is amplified by a voltage-gain factor of 14, which is internally forced to the "IADJ" pin voltage. Therefore, to force a current through the LEDs, the "IADJ" pin voltage is divided by 14 and then divided by R26 as shown in [Equation 7](#). The equation for temperature and current is shown in [Equation 8](#).

$$I_{LED} = \frac{V_{(CSP-CSN)}}{R26} \quad (5)$$

$$V_{(CSP-CSN)} = \frac{V_{IADJ}}{14} \quad (6)$$

$$I_{LED} = \frac{\left(\frac{V_{IADJ}}{14}\right)}{R26} \quad (7)$$

$$I_{LED} = \frac{\left(\frac{(-13.6 \text{ mV}/^{\circ}\text{C} \times T + 2637 \text{ mV})}{14}\right)}{R26} \quad (8)$$

[Equation 8](#) is obtained by substituting [Equation 4](#) into [Equation 7](#). In this TI Design, R26 is chosen to be 560 mΩ. This means that the relationship of current through the LEDs versus temperature looks like [Figure 8](#).

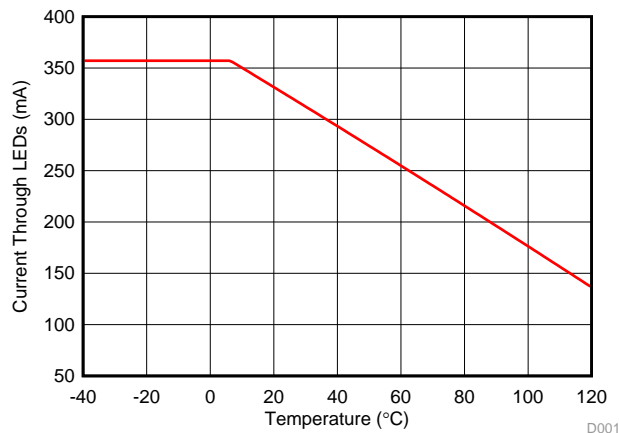


Figure 8. Current Through LEDs versus Temperature

Note that current is a first-order representation of the brightness of the LEDs, which is what is needed to be controlled for this TI Design. Find more information about the LED selection and some of the difficulties in [Section 7.2](#). Also the LMT87-Q1 is measuring the ambient temperature of the LED board. Therefore, because the junction temperature of the LEDs is the specification that could be violated, analysis must be done to determine the relationship between the ambient temperature where the LMT87-Q1 and the junction temperature of each of the LEDs so that the thermal foldback will not allow them to go beyond the maximum junction temperature. This relationship is application specific because it is dependent on a number of factors such as distance, copper thickness, heat sinks of LEDs, power dissipated through the LEDs, and so on. This calculation would also have to be done if thermal foldback was implemented with NTCs. However, with NTCs not being linear, even more complicated math would be needed to determine the ambient temperature, while the LMT87-Q1 can just be fed back into the driving device.

For this TI Design, the basic circuitry to implement thermal foldback was used as shown in Figure 9.

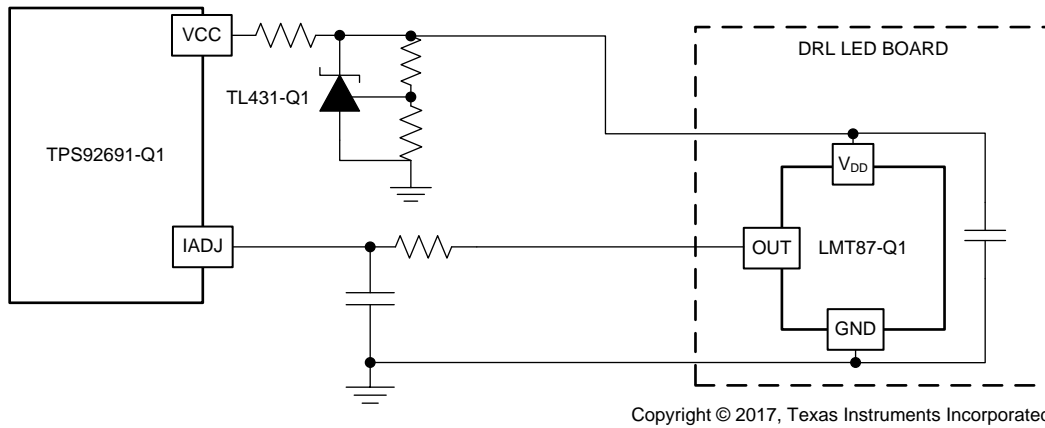


Figure 9. TIDA-01382 Thermal Foldback Circuit

6.6.1 Changing Thermal Foldback Response

Sometimes requirements call for moving the thermal response in various ways. Some of the most common changes are: deciding when to start derating current through the LEDs, changing the slope, and holding at a constant current at higher temperature. There are various tradeoffs associated with making these configurable highlighted in the following subsections.

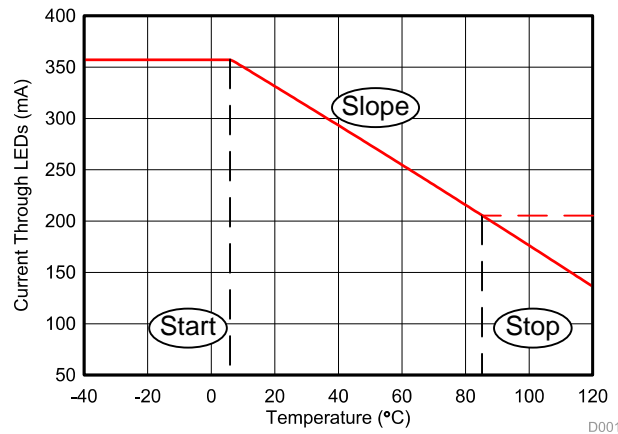
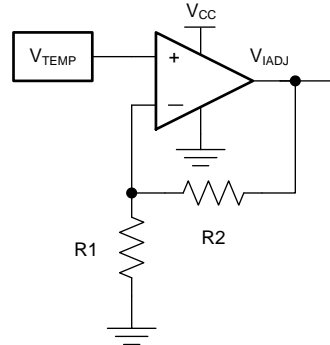


Figure 10. Thermal Foldback Possibilities

6.6.1.1 Changing Starting Point for Thermal Foldback

The TPS92691-Q1 has an internal clamp on the "IADJ" pin at 2.55 V and an absolute maximum input voltage at the pin of 8.8 V. Because there is room there to scale up the input into the "IADJ" pin, the simplest way to change the starting point for thermal foldback is introducing a non-inverting amplifier, as shown in Figure 11, with a gain found using Equation 9.



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Figure 11. Non-Inverting Amplifier

$$\frac{V_{OUT}}{V_{IN}} = \frac{1 + R2}{R1} \quad (9)$$

With an output voltage from the LMT87 of 3.16 V at -40°C , the maximum scaling factor possible is 2.5 in order to not violate the maximum voltage at the "IADJ" pin. Therefore, if a scaling factor of 2.5 was used, the temperature at which the current would start to derate would be 117°C . Also if no scaling is used, the temperature at which the current starts to derate is 7°C . Equation 10 and Equation 11 show how to choose at what temperature the derating should start.

NOTE: This is a first-order linear approximation. If being exact is required, consult the [LMT87-Q1](#) datasheet.

Temperature range from -40°C to T:

$$I_{LED} = \frac{2550 \text{ mV}}{(14 \times R26)} \quad (10)$$

Temperature range from T to 125°C :

$$I_{LED} = \frac{K \times (2637 \text{ mV} - 13 \text{ mV} \times T)}{(14 \times R26)} \quad (11)$$

Where "T" is the temperature when the current will start derating and K is the scaling factor introduced by the non-inverting amplifier. The scaling factor needed is determined by Equation 12:

$$K = \frac{2550 \text{ mV}}{(2637 \text{ mV} - 13 \text{ mV} \times T)} \quad (12)$$

Using these equations, the derating is set so that it occurs at whatever temperature chosen $\pm 3^{\circ}\text{C}$. In order to build this application, choose an op amp specifically for this non-inverting amplifier. There are several factors that will effect which op amp would be a good candidate. The input offset specification is the key factor on how accurately the current would reflect the temperature. This would mean an op amp like the [OP07](#) or [LT1013](#). Another aspect would be a low common-mode input voltage with rail-to-rail output so that the V_{TEMP} could go down to GND and the output of the op amp would accurately reflect that. Op amps such as the [TLV341](#) or the [OP314](#) would work in this case.

6.6.1.2 Changing Slope of Thermal Foldback

A limitation of this TI Design is that the thermal response will always follow the slope of the LMT87-Q1, -13.6 mV . However, thanks to TI's broad portfolio of [temperature sensors](#), changing the slope of the line is simple. Specifically, the [LM94021-Q1](#) can configure its slope from one of the four shown in [Figure 12](#).

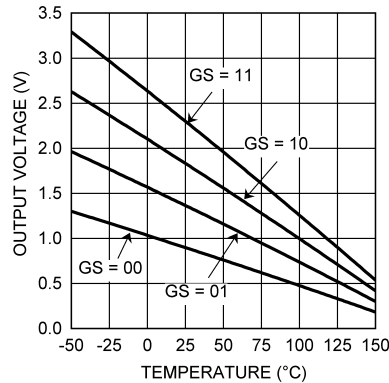
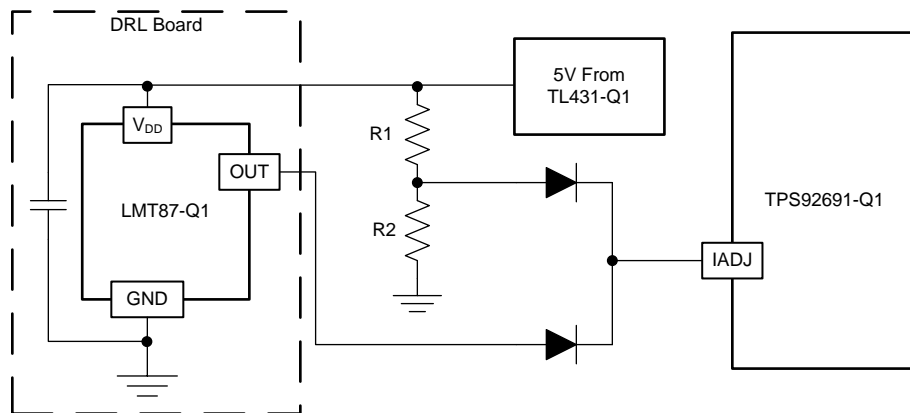


Figure 12. Different Slopes of LM94021-Q1

6.6.1.3 Constant Current at High Temperatures

The current design will derate the current according to temperature across the entire temperature range. However, in many applications, thermal foldback needs to clamp the current through the LEDs at a certain point to ensure there is a minimum current through the LEDs regardless of the temperature. This clamp will ensure that the brightness does not decrease past the required minimum brightness as in [SAE J1383](#). The easiest way to achieve this is by diode OR-ing the output voltage from the LMT87-Q1 and a fixed voltage as shown in [Figure 13](#).



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Figure 13. Current Clamp for High Temperatures

This would allow the thermal foldback to occur as intended until it reaches a specified threshold where it keeps the current through the LEDs constant. This threshold allows the graph of current versus temperature to look like [Figure 14](#).

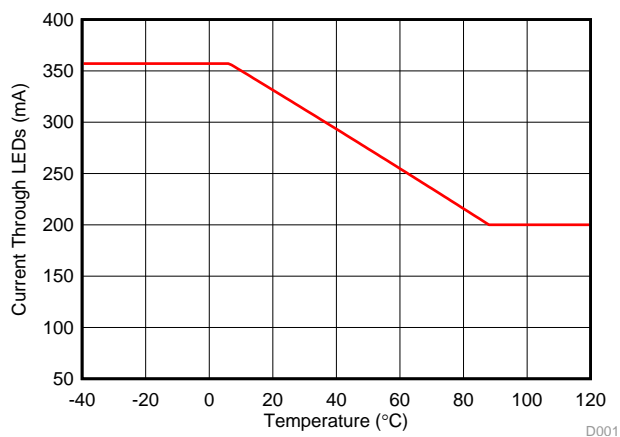


Figure 14. Current versus Temperature With Clamp

Equation 13 show how to calculate the resistor values needed to clamp at a given current.

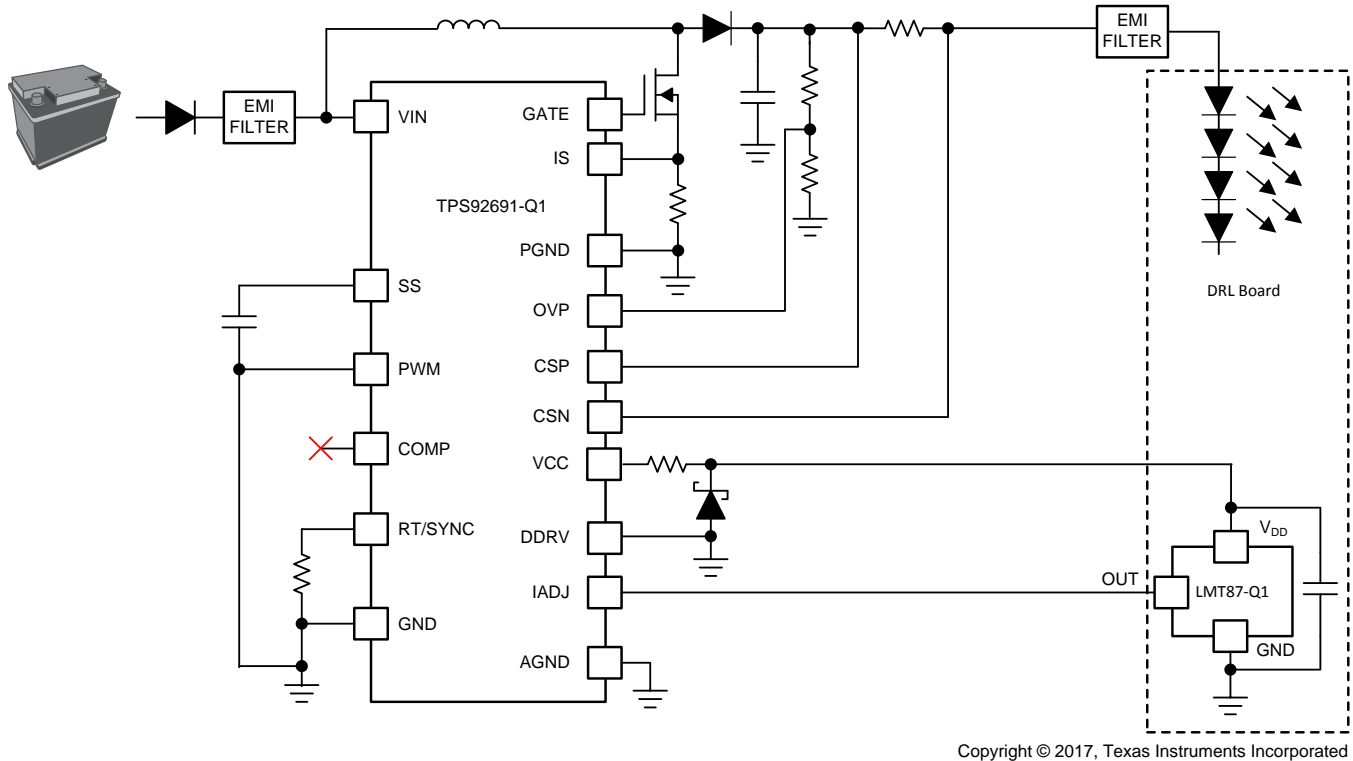
$$V_{IADJ} = I_{LED} \times R26 \times 14 \tag{13}$$

Equation 14 shows at what temperature it is going to clamp.

$$T = \frac{(2637 \text{ mV} - V_{IADJ})}{0.013 \text{ V}} \tag{14}$$

6.6.2 Thermal Foldback Without PWM Dimming

If precision PWM dimming is not required in the application, thermal foldback can still be implemented. The first way to achieve this is adding a bias resistor and a Zener diode to power the LMT87-Q1 from the output of the internal LDO of the TPS92691-Q1. These devices allow the capability of thermal foldback without PWM dimming and without an MCU. Figure 15 is a diagram of this schematic.



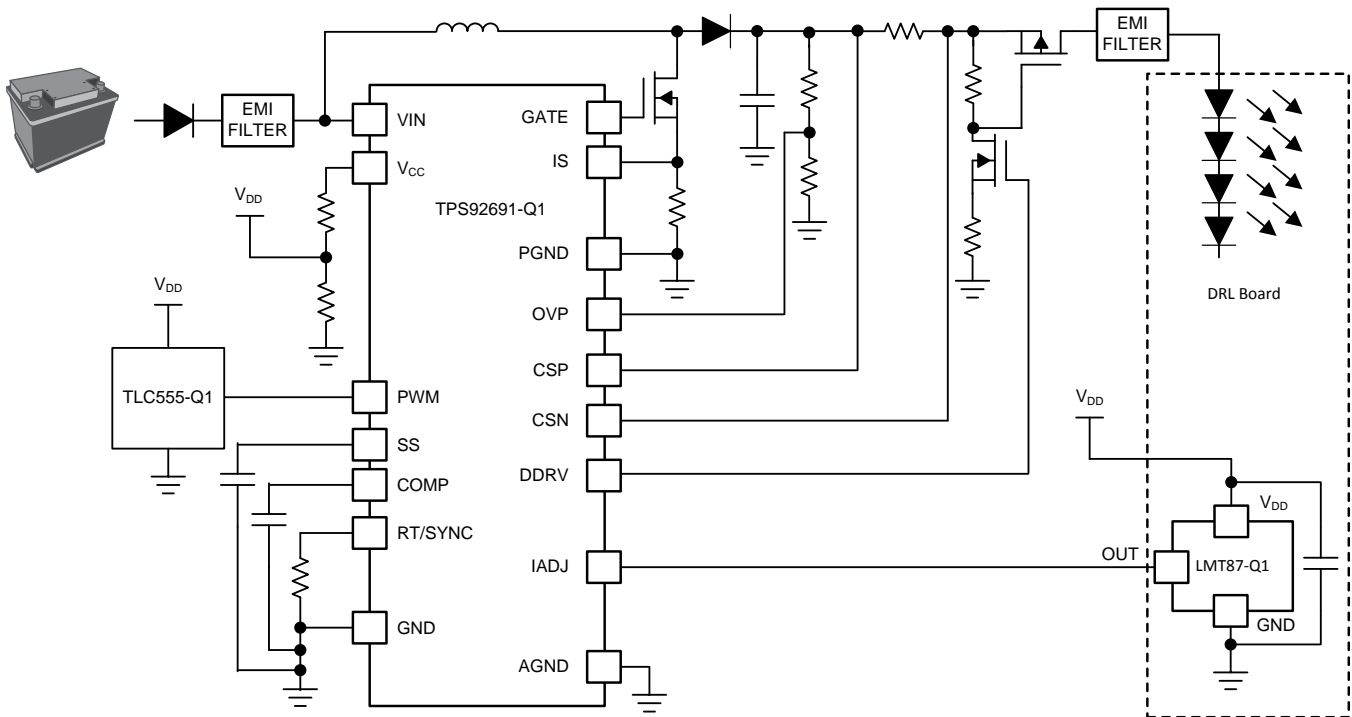
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Figure 15. Thermal Foldback Without PWM Dimming

In this application, the COMP pin is disconnected internally so it does not matter, while the DDRV pin is connected to ground. The VCC pin needs to be connected through a bias resistor to a Zener diode because the output of the internal LDO, V_{CC} , is a constant 7.5 V while the maximum input voltage for the LMT87-Q1 is 5.5 V. Because the current consumption of the LMT87-Q1 is a maximum of 9 μ A across temperature, the minimum value of the bias resistor can be calculated. All calculations must be made for a worst case scenario so that the LMT87-Q1 will be powered whenever the output of the internal LDO is regulating. The maximum V_{CC} is 8 V and the minimum voltage needed to power the LMT87-Q1, but still maintain the minimum operating temperature, is 3.5 V. Therefore, the Zener diode is chosen to be somewhere between 3.5 and 5.5 V. Next, take the minimum value of the Zener diode output with respect to its swing across temperature, and subtract it from the max V_{CC} . That value is then divided by the maximum current required to operate the LMT87-Q1 (9 μ A) plus the minimum current required to keep the Zener regulating. Equation 15 gives a minimum value of the bias resistor because it ensures the Zener diode to be regulating when the LDO is regulating as shown in Equation 15.

$$R_{BIAS} \leq \frac{(V_{CC(max)} - V_{ZENER})}{(I_{S(LMT87-Q1)} + I_{ZENER})} \quad (15)$$

If PWM dimming is required but precision is not, thermal foldback can still be achieved with this TI Design. The schematic reduces to [Figure 16](#).



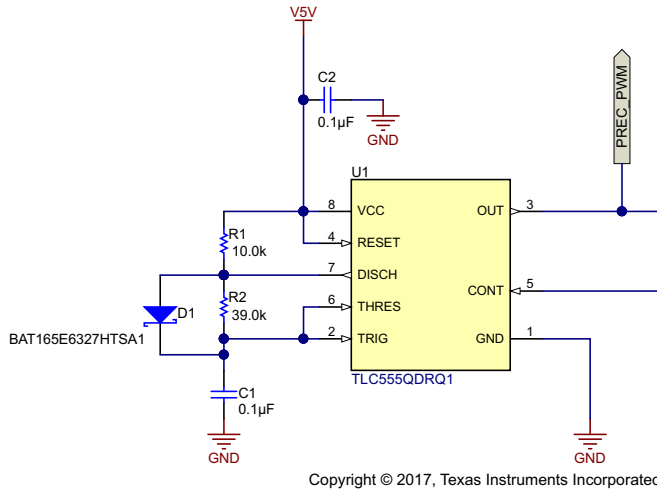
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Figure 16. Thermal Foldback With PWM Dimming

The TLC555-Q1 is generating an approximate PWM signal that goes through a low-pass filter into the PWM pin. This signal causes the DDRV pin to PWM dim the output as in the original design. However, in this application, the LMT87-Q1 is being powered by a resistor divided output of the internal LDO. This allows a reduction of parts while still implementing accurate thermal foldback.

6.7 Clock Generation (PWM)

Connecting TRIG to THRES, as Figure 17 shows, causes the timer to run as a multivibrator. The capacitor C1 charges through R1 and D1 to the threshold voltage level (approximately $0.67 V_{DD}$) and then discharges through R2 only to the value of the trigger voltage level (approximately $0.33 V_{DD}$). As Figure 18 shows, the output is high during the charging cycle ($t_{c(H)}$) and low during the discharge cycle ($t_{c(L)}$).



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Figure 17. TLC555-Q1 Clock Generation

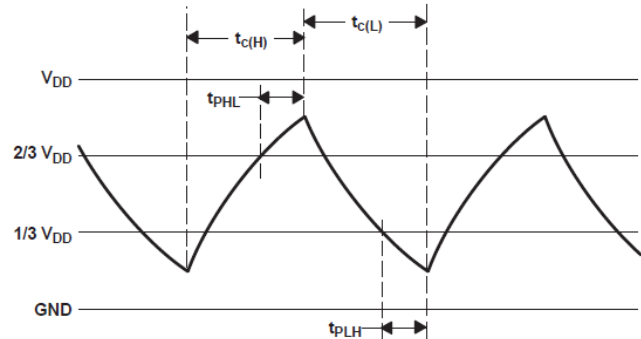


Figure 18. Trigger and Threshold Voltage Waveform

The values of R1, D1, R2, and C1 control the duty cycle, as the following equations show:

$$t_{c(H)} \approx C_T R_1 \ln 2$$

$$t_{c(L)} \approx C_T R_2 \ln 2$$

$$\text{Period} : t_{c(L)} + t_{c(H)} \approx C_T R_2 \ln 2$$

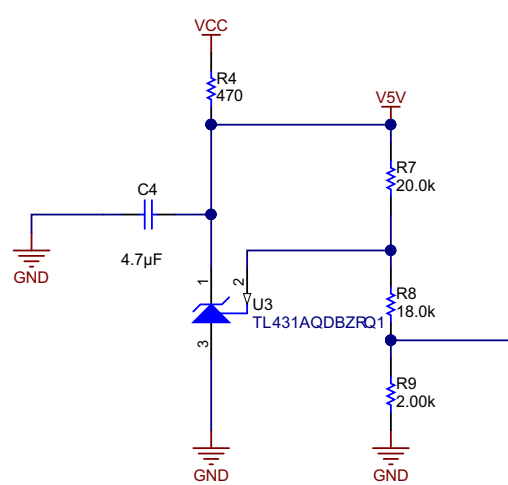
$$\text{Output waveform duty cycle} : \frac{t_{c(H)}}{t_{c(L)} + t_{c(H)}} \tag{16}$$

- $t_{c(H)} = 100 \text{ nF} \times 10 \text{ k}\Omega \times \ln 2 = 693 \text{ }\mu\text{s}$
- $t_{c(L)} = 100 \text{ nF} \times 39 \text{ k}\Omega \times \ln 2 = 2.7 \text{ ms}$
- $\text{Period} = t_{c(H)} + t_{c(L)} = 3.39 \text{ ms}$
- $\text{Output waveform duty cycle} = 20.4\%$

The preceding formulas do not allow for any propagation delay times from the TRIG and THRES inputs to DISCH. These delay times add directly to the period and create differences between calculated and actual values that increase with frequency. In addition, the internal ON-state resistance (r_{ON}) during discharge adds to R2 to provide another source of timing error in the calculation when R2 is very low or r_{ON} is very high. These errors can be canceled out by applying a feedback loop.

6.8 Onboard Supply and Setting Duty Cycle

The supply for the TLC555-Q1 device, the OPA2377-Q1 op amp, as well as the reference voltage for setting the duty cycle is derived from the precision shunt regulator (0.5%) TL431-Q1 (see Figure 19). The device is supplied by the VCC (7.5 V) output of the TPS92691 device and regulates this voltage down to a precision 5 V (V5V) set by the resistors R7, R8, and R9. Additionally, the reference voltage for the op amp, which is used to set the duty cycle, is derived from this voltage that resistor divider R8 and R9 generate.

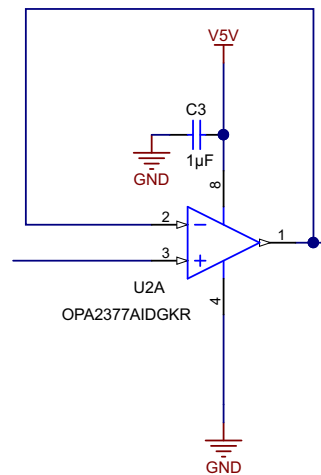


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Figure 19. TL431-Q1 Precision Voltage Generation

6.9 Buffering, Averaging, and Filtering

The OPA2377-Q1 dual op amp, low-noise, rail-to-rail input and output, and low offset, makes this device ideal for these type of applications. The output stage of the TLC555-Q1 as [Section 5.4.2](#) describes is hooked up through a 1-K resistor to the PWM input of the TPS92691 device, which dims the LEDs and also connects to a buffer stage and a second-order filter formed by the dual OPA2377 op amp (see [Figure 20](#)). The buffer stage is placed to avoid changing impedance from input to output so as not to interfere with the shape or timing of the square wave generated by the TLC555. The waveform after the buffer stage is then averaged by the next op amp stage, which forms a second-order filter that is compared against a reference voltage set by the resistor divider from the precision reference. The output of the second-order filter is hooked up to the CONT pin of the TLC555-Q1 closing the loop. The CONT input of the TLC555 allows the upper and lower trigger threshold of the timing duty cycle to be changed by varying the voltage level at this pin, which is performed by the output of the second-order filter.



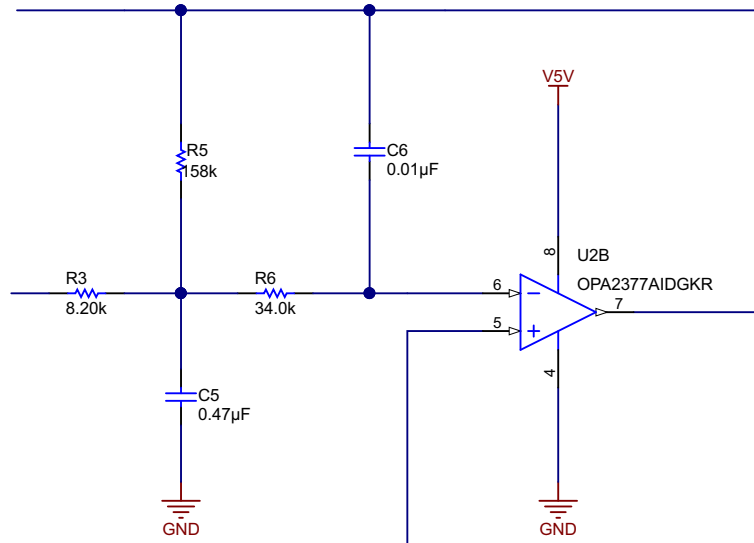
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Figure 20. Buffer

Use the following parameters for this TI Design example:

- Gain = 19.2 V/V (inverting gain)
 - Low-pass cutoff frequency = 31.67 Hz
- Second-order Chebyshev filter response with a 3-dB gain peaking in the passband

Figure 21 shows the infinite-gain multiple-feedback circuit for a low-pass network function.



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Figure 21. Second-Order Filter and Averaging

Use Equation 17 to calculate the voltage transfer function:

$$\frac{\text{Output}}{\text{Input}}(s) = \frac{-1}{R_1 R_3 C_2 C_5} \frac{1}{s^2 + \left(\frac{s}{C_2}\right) \times \left(\frac{1}{R_1} + \frac{1}{R_3} + \frac{1}{R_4}\right) + \frac{1}{R_3 R_4 C_2 C_5}} \quad (17)$$

This circuit produces a signal inversion. For this circuit, the gain at DC and the low-pass cutoff frequency are calculated by Equation 18:

$$\text{Gain} = \frac{R_5}{R_3}$$

$$f_c = \frac{1}{2\pi} \sqrt{\frac{1}{R_6 \times R_5 \times C_5 \times C_6}} \quad (18)$$

Where:

- Gain = 19.2
- $f_c = 31.67$ Hz

Software tools are readily available to simplify filter design. WEBENCH® Filter Designer is a simple, powerful, and easy-to-use active filter design program. This program enables designers to create optimized filter designs using a selection of TI op amps and passive components from TI's vendor partners.

6.10 Boost Converter

The TPS92691-Q1 controller is suitable for implementing step-up or step-down LED driver topologies (see Figure 22). This TI Design uses the boost configuration. Use the detailed design procedure of the device datasheet[2] to select component values for the TPS92691-Q1 device. This section addresses the design process for the boost converter.

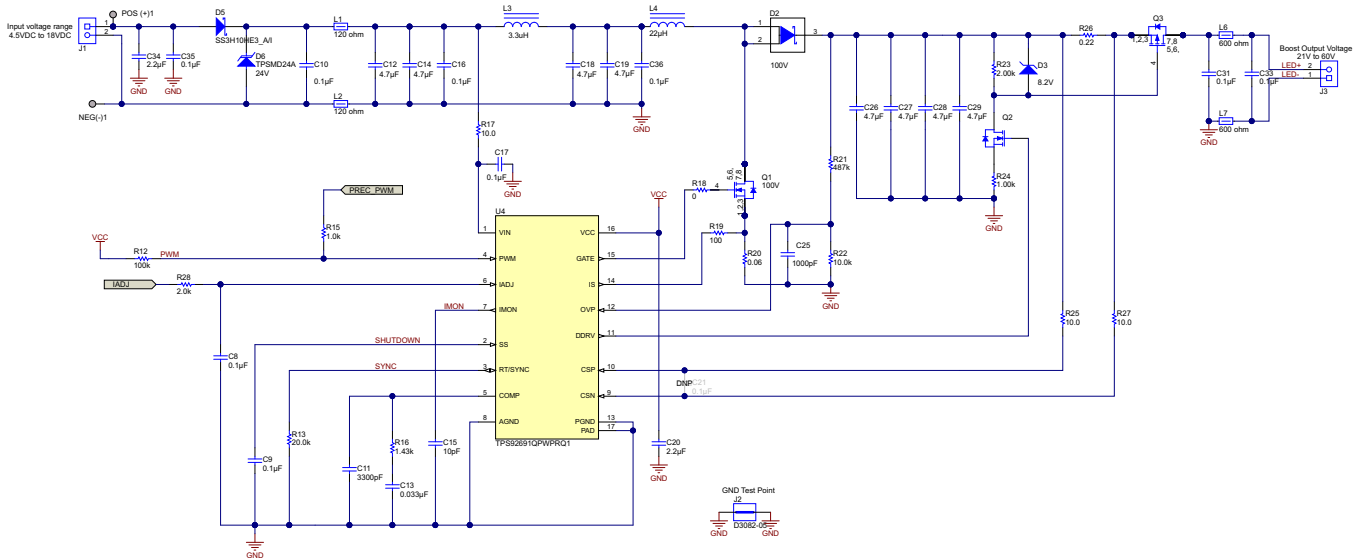


Figure 22. TPS92691-Q1 LED Driver in Boost Configuration

DC-DC converters can couple large amounts of energy (especially at the fundamental switching frequency) back through the battery inputs and into the remainder of the vehicle. This energy is produced because of the switching action of the input-current waveform that is translated into voltage noise by the equivalent series resistance (ESR) of the input capacitors that carry most of this current. A low-pass filter, placed between the input of the module and the DC-DC converters, has been added to attenuate this noise. The low-pass filter also filters incoming noise that enters the system. The low-pass filter can be designed empirically or theoretically (by calculation and simulation). The empirical approach is to design the system without the EMI filter, measure the conducted emissions with a spectrum analyzer, and compare it to the standard that must be passed. Next, calculate the attenuation required to pass at certain frequencies and place the corner frequency of the filter low enough to achieve the desired attenuation.

The theoretical approach is more complicated. Ensure the assumption is that the boost converter is the problem and that the noise generated by the downstream circuitry is to be filtered by the boost inductor or capacitors.

NOTE: The main sources of noise are fundamental at the switching frequency of the boost (400 kHz) and the harmonics. If the amplitude of the noise at that frequency can be estimated and attenuated appropriately, the harmonics can also be attenuated.

The input voltage is the voltage generated by the ripple current through the ESR of the input capacitors. Because ceramic capacitors are used, this ESR is very low (approximately 3 mΩ). The peak amplitude of the input voltage ripple is approximately 2.7 mV (see Equation 19). The concern is the frequency content at 400 kHz, not the time domain.

$$3 \text{ m}\Omega \times 0.9 \text{ A} = 2.7 \text{ mV} \tag{19}$$

Use the Fourier transform of this asymmetric-triangle waveform to find the coefficients and amplitudes of each component frequency. The coefficient of the fundamental for this type of waveform is 0.8. Multiply the coefficient times the time domain amplitude to find the energy at 400 kHz (see Equation 20).

$$0.8 \times 2.7 \text{ mV} = 2.16 \text{ mV} \tag{20}$$

Using [Equation 21](#), convert the product of [Equation 20](#) to dB μ V to make it easier to analyze based on the CISPR 25 standards.

$$20 \times \log\left(\frac{2.16 \text{ mV}}{1 \mu\text{V}}\right) \approx 67 \text{ dB}\mu\text{V} \quad (21)$$

Compare the 67 dB μ V to the CISPR 25 specification and calculate how much to attenuate. The CISPR 25 specification does not define a limit at 400 kHz, but the limit at 530 kHz for Class 5 conducted emissions is 54 dB μ V (peak). An attenuation of at least 13 dB is required. Make the goal 40-dB attenuation at the switching frequency. Calculate where to place the corner frequency of the filter when attenuation at 400 kHz is known. The second-order low-pass filter has a rolloff of -40 dB per decade. Place the corner frequency at 40 kHz to attain 40 dB of attenuation at 400 kHz. The corner frequency is related to the values of the filter inductor and capacitor, calculated by using [Equation 22](#):

$$2 \times \pi \times f = \frac{1}{\sqrt{L \times C}} \quad (22)$$

Choose an L of 3.3 μ H. There is approximately 47 nF, calculating out for C. To keep the ESR low, put two capacitors in parallel and choose 4.7 μ F for C12, C14 and C18, C19. Choosing a larger value lowers the corner frequency of the filter, which provides more attenuation at 400 kHz. Also, ceramic capacitors suffer from DC bias effects and operate at a capacitance that is less than their rating. To filter the high-frequency noise content, a 100-nF capacitor is added.

7 Getting Started Hardware

7.1 Hardware

Connect the desired number of LEDs per string at the output screw terminals to get started with the TIDA-01382 board. [Figure 23](#) shows an image of the TIDA-01382 board.

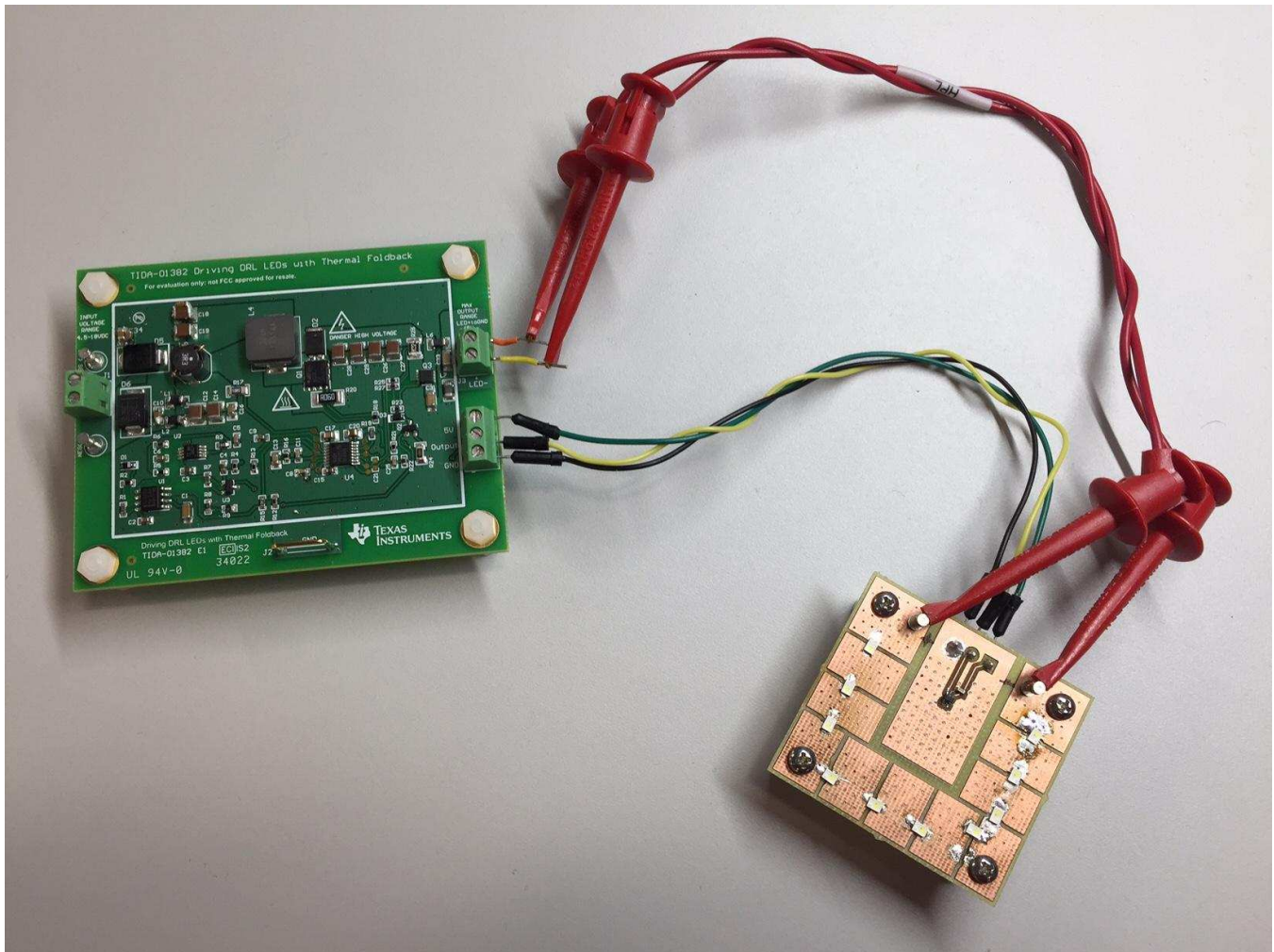


Figure 23. TIDA-01382 With LED Daughter Board

7.2 LED Selection

Selecting the DRL LEDs is a key aspect of this TI Design. OSRAM Opto Semiconductor's [SYNOS P2720](#) LEDs that can come in 300-, 700-, and 1000-mA ranges that allow them to be anywhere from 50 to 315 lumens. These LEDs allow for scalability because they all share the same package, just a different die size. Choosing which of these LEDs depends greatly on the optics that the LED will be placed in; each optic can reduce the light efficiency of the LED, which should be considered in choosing LEDs. Another aspect of LEDs is their thermal roll-off. According to the datasheet of the LEDs used in this TI Design, [KW DMLN31.SG\[6\]](#), the relative luminous flux at high temperatures can decrease to 90% which means that even though they are dissipating more energy at higher temperatures, they are actually dimmer than expected. This must be taken into consideration with the optics so that the LEDs will be able to pass any government standards for automotive lighting such as the [SAE J1383](#). All of these conditions make the equation for luminosity non-linear and heavily dependant on final application factors. Therefore for this TI Design, a first-order estimate of brightness, or the current through the LEDs, is what is used to reflect the brightness of the LEDs. For a final design, the thermal roll-off and optic reflector will need to be considered and a graph of luminosity versus temperature can be performed.

7.3 J3, LED+, LED– (Boost)

The screw-down connector (J3, marked LED+ and LED–) is for connecting the LED load to the board in the boost configuration. The leads to the LED load must be twisted and kept as short as possible to minimize voltage drop, inductance, and EMI transmission. The boost design is for approximately 6 to 20 white LEDs.

7.4 J1, POS(+), NEG(–)

The screw-down connector [J1, marked POS(+) and NEG(–)] is for connecting the board to the DC input voltage supply. One other POS(+) and NEG(–) test turret is provided on the board that can also be used.

7.5 J4, Temperature Sensor Connection

The screw-down connector (J4 marked 5V, Output, and GND) is for connecting the external temperature sensor, LMT87-Q1, to implement thermal foldback. Output is tied through a low pass filter to the IADJ pin of the TPS92691-Q1 to change the current based on the analog voltage coming from the LMT87-Q1. The LMT87-Q1 can be powered from the 5 V and connected to GND.

7.6 Duty Cycle Adjust

Per default, the duty cycle of this system has been set by resistors R8 and R9 to the minimum of 5%. If the duty cycle needs to be changed, adjust resistors R8 and R9 accordingly (see [Figure 19](#)).

8 Testing and Results

Several tests were conducted to verify the accuracy of the duty cycle, the thermal foldback reaction, and EMI compliance.

8.1 Duty Cycle Accuracy

1. The board ran continuously with a fixed duty cycle set and the duty cycle recorded continuously with the oscilloscope and standard deviation calculated (see [Figure 24](#)).

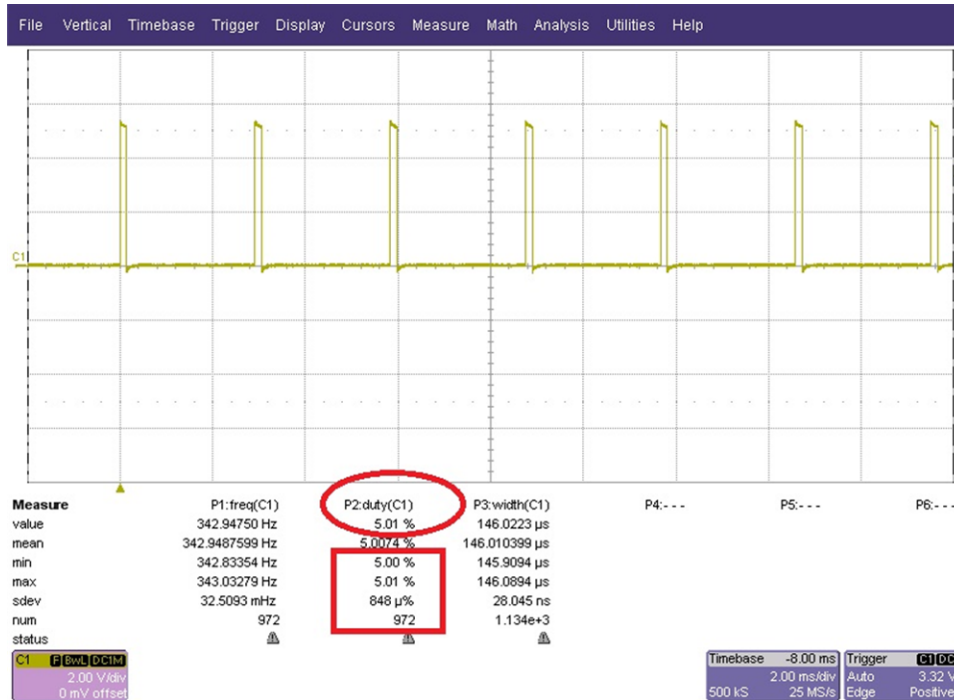


Figure 24. Duty Cycle Accuracy

The result was a very accurate duty cycle with a standard deviation of 848μ% (0.000848%).

2. Different clock devices (TLC555) from a different production batch were tested on the same board. The result was a very accurate duty cycle with a standard deviation of less than 0.5%.
3. Temperature test: -40°C to 110°C. The board ran continuously with a fixed duty cycle set and the duty cycle continuously recorded with an oscilloscope and standard deviation calculated. The result was a very accurate duty cycle with a standard deviation of less than 0.5%.

8.2 Thermal Foldback Testing

The benefit of this TI Design is how linearly and accurately the current through the LEDs follows the temperature as shown in [Figure 25](#).

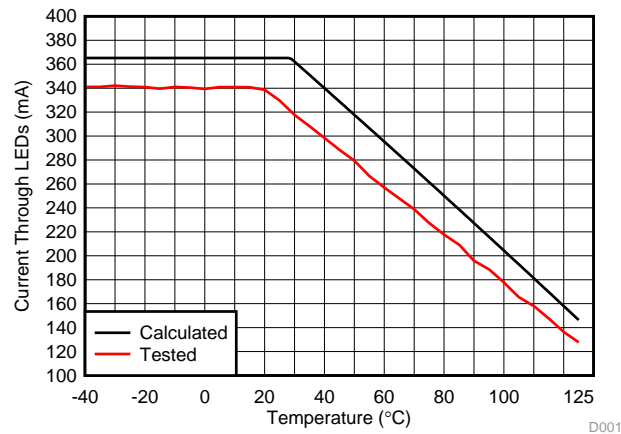


Figure 25. Thermal Foldback Tested

Note that the calculated data is based on the clamp of the IADJ pin set to 2.25 V. This can vary as per the datasheet; however, the slopes of the line when they are decreasing are different by 0.225 and the linearity of the slope is 0.9998. As discussed in [Section 6.6](#), the slope, starting point, and bottom clamp of this graph can all be manipulated in order to create the desired foldback response.

8.3 EMI Testing

In automotive lighting, EMI is a problem. Therefore, this TI Design was made so that it can be CISPR 25 compliant. In testing it was found that the design is CISPR 25 Radiated Emissions Class 4 compliant as shown in Figure 26, Figure 27, and Figure 28, and is CISPR 25 Conducted Emissions Class 5 compliant as shown in Figure 29.

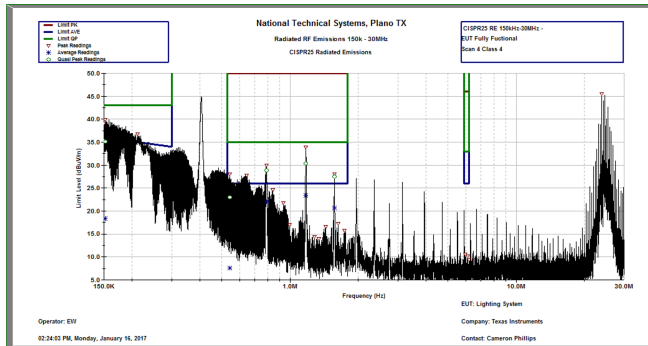


Figure 26. CISPR 25 RE 150k to 30 MHz Class 4

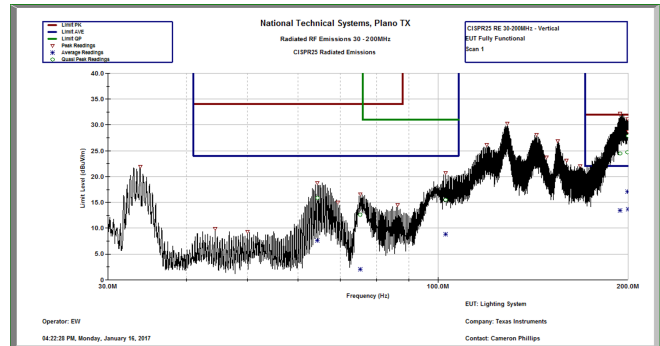


Figure 27. CISPR 25 RE 30 to 200 MHz Class 4

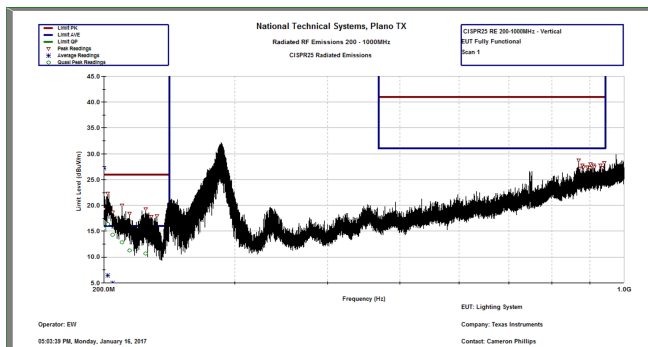


Figure 28. CISPR 25 RE 200 to 1000 MHz Class 4

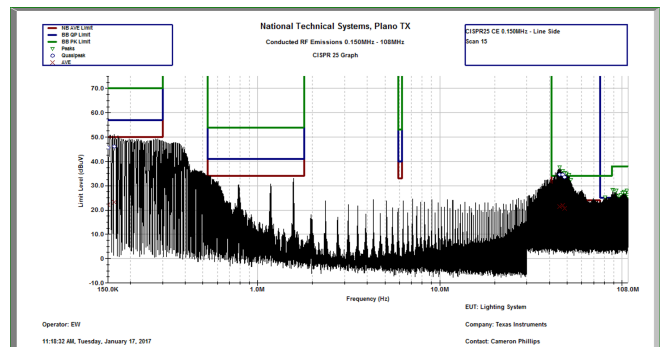


Figure 29. CISPR25 CE 0.150 to 108 MHz Class 5

8.4 Accuracy Calculation

See the [TIDA-01183](#) design to see how the calculations of accuracy were done to prove the output is within 2%.

9 Design Files

9.1 Schematics

To download the schematics, see the design files at [TIDA-01382](#).

9.2 Bill of Materials

To download the bill of materials (BOM), see the design files at [TIDA-01382](#).

9.3 PCB Layout Recommendations

9.3.1 Layout Prints

To download the layer plots, see the design files at [TIDA-01382](#).

9.4 Altium Project

To download the Altium project files, see the design files at [TIDA-01382](#).

9.5 Gerber Files

To download the Gerber files, see the design files at [TIDA-01382](#)

9.6 Assembly Drawings

To download the assembly drawings, see the design files at [TIDA-01382](#).

10 Related Documentation

1. Texas Instruments, [TLC555-Q1 LINCMOS™ TIMER](#), TLC555-Q1 Datasheet (SLFS078)
2. Texas Instruments, [TPS92691/-Q1 Multi-Topology LED Driver With Rail-to-Rail Current Sense Amplifier](#), TPS92691-Q1 Datasheet (SLVSD68)
3. Texas Instruments, [OPAx377-Q1 Low-Noise, Low Quiescent Current, Precision Automotive Grade Operational Amplifier](#), OPA2377-Q1 Datasheet (SBOS797)
4. Texas Instruments, [TL431-Q1 ADJUSTABLE PRECISION SHUNT REGULATOR](#), TL431A-Q1 Datasheet (SGLS302)
5. Texas Instruments, [TPS92691 Boost and Boost-to-Battery LED Driver Evaluation Board](#), TPS92691-Q1 User's Guide (SLVUA07)
6. OSRAM Opto Semiconductor, [SYNIOS P2720 KW DMLN31.SG Verison 1.2](#), SYNIOS P2720 Datasheet (http://www.osram-os.com/Graphics/XPic0/00192945_0.pdf/KW%20DMLN31.SG.pdf)

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11 About the Author

CAMERON PHILLIPS is an applications engineer at Texas Instruments where he is responsible for developing reference design solutions for the Automotive Body and Lighting segment. Cameron earned his bachelor of science in electrical engineering from Texas A&M University.

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