Texas Instruments

Application Report
SNVA900–January 2020

Extending the Input Voltage Range of the LM5180-Q1 PSR Flyback for VDA 320 (LV 148)

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ABSTRACT

The LM5180-Q1 device is a primary-side regulated (PSR) flyback converter with the input voltage range from 4.5 V to 65 V and the output power up to 7 W. The control scheme samples feedback voltage directly from the primary-side. This eliminates the need for an optocoupler or auxiliary winding type of the feedback. Properties and easy implementation of the LM5180-Q1 make it an attractive choice for the trending 48-V automotive systems. However, the 48-V systems typically refer to the VDA 320 (LV 148) test specification that specifies the voltage range up to 70 V for a short period of time. This application report describes a circuit that extends the input voltage range of the LM5180-Q1 device, or similar, for VDA 320 (LV 148) compliance.

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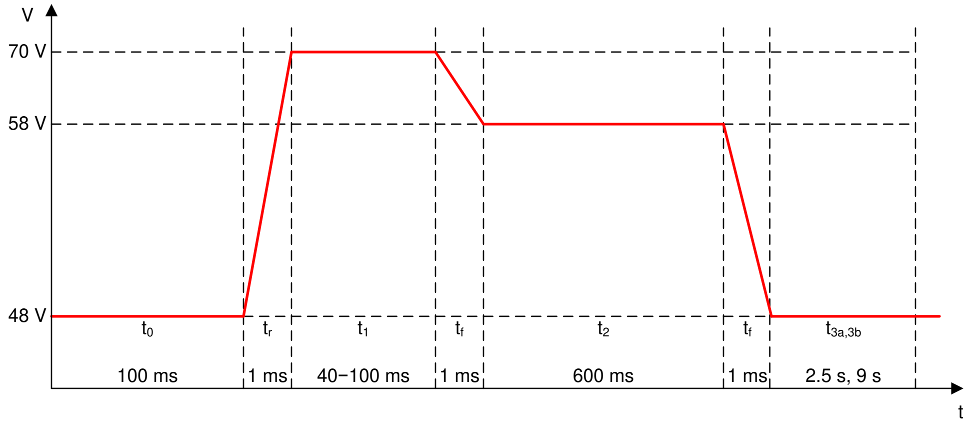
Trademarks

1 Introduction

The VDA 320, also known as LV 148, is practically the only publically available test specification for the trending 48-V automotive systems. This test specification became a base for ISO21780:2019(E) which is now in finalization. It specifies various test conditions for loads connected to the 48-V net. One of the many more or less interesting tests is the *E48-02 Transient Overvoltage* test. Figure 1 shows the test waveform.



Not to scale, 3 × t_{3a} = 2.5 s short test, 1000 × t_{3b} = 9 s endurance test

Figure 1. VDA 320 E48-02 Test Waveform

As Figure 1 shows, the voltage on the 48-V rail goes up to 70 V and stays there for 40 ms, some OEMs require even 100 ms. The device under test (DUT) has to survive this event with *Functional Status A* – the DUT must perform all the functions. Note that clamping using TVS or Zener diodes is impractical for any higher loads. Simply said, integrated circuits connected directly to the 48-V rail must withstand 70 V under all conditions. Additionally, the VDA 320 specifies a long-term overvoltage test when the 48-V rail remains at 60 V for 60 minutes.

2 Pre-regulator for the LM5180-Q1

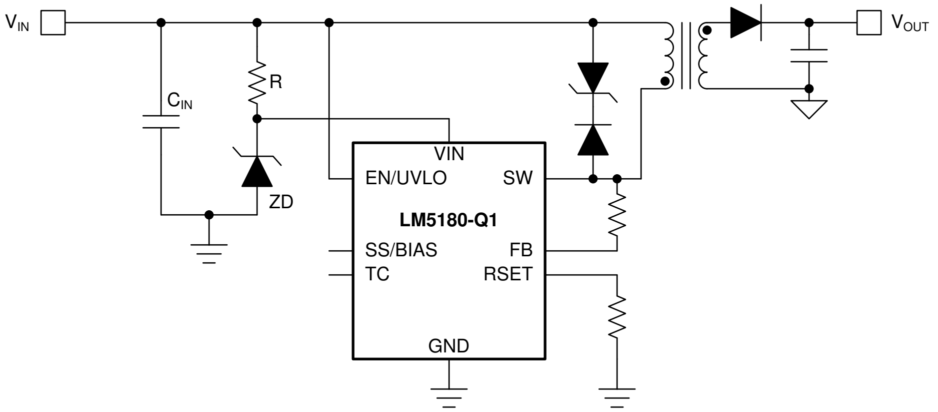


Figure 2. Zener Diode Protecting the LM5180-Q1

The easiest method for extending the input voltage range is adding a Zener diode (ZD) and a resistor R to the VIN pin as Figure 2 shows. However, the Zener regulator limits the voltage only for the VIN pin but does not protect the switch node pin SW. Although the switch node is rated for 100-V it needs to withstand the sum of input and the reflected voltage plus the leakage inductance voltage spike.

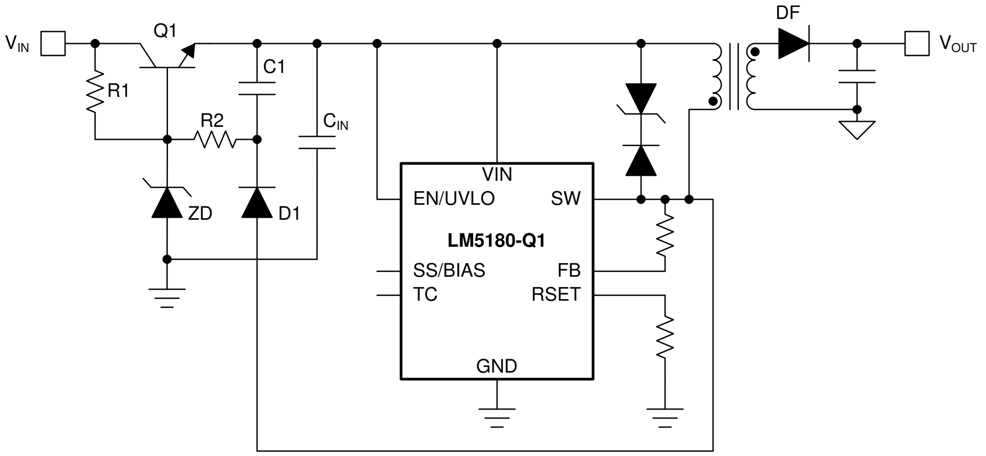


Figure 3. Pre-regulator Protecting the LM5180-Q1 Including the Switch-Node

Figure 3 shows a more advanced yet simple circuit (pre-regulator) that extends the input voltage range of the LM5180-Q1 device. This pre-regulator offers more headroom for the reflected voltage and protects the switch node too. It passes voltages below the desired limit (for example, approximately 62 V) but starts limiting voltages above the limit. This behavior is beneficial because power dissipation during normal operation is significantly reduced to the bare minimum.

The circuit operates in three modes with regards to the input voltage:

Start-Up Phase – Current from the 48-V rail flows to the base of the transistor (Q1) through the resistor R1. The Q1 starts conducting and current from the 48-V rail charges the input capacitor. This phase ends when the voltage on the input capacitor exceeds the UVLO rising threshold and the LM5180-Q1 starts switching.

Pass-Through Phase – The LM5180-Q1 is switching normally. The diode (D1) rectifies the pulsating voltage from the switch node SW and charges the capacitor C1. The voltage on C1 corresponds to the reflected voltage reduced by the voltage drop across the diode D1. The voltage on the capacitor C1 is positive with reference to the emitter of the transistor Q1. The resistor R2 further biases the base of the transistor Q1. This way the bipolar transistor enters saturation. The voltage drop across the collector-emitter V_{CE} is $V_{CE(SAT)}$ and the transistor Q1 operates as a switch.

Pre-regulator Phase – The Zener diode (ZD) starts sinking current when the input voltage exceeds the zener voltage V_Z. The voltage on the output of the pre-regulator is limited to the Zener voltage raised by the base-emitter voltage drop.

3 Design Procedure Example

This design procedure uses the LM5180-Q1 dual-output EVM as an example. Table 1 lists the selected parameters of the EVM. The design objective is increasing the input voltage range to 70 V for maximum load $I_{OUT1} = –I_{OUT2}$ = 0.2 A. The input EMI filter is removed.

Table 1. LM5180-Q1 EVM Characteristics

Parameter	Test Conditions		MIN	TYP	MAX	UNIT
Model type	LM5180EVM-DUAL BSR052A					
Input voltage range V_{IN}	Operating		10	48	65	V
Input current, no load $I_{IN(NL)}$	$I_{OUT1} = I_{OUT2} = 0$	V_{IN} = 24 V		1.4		mA
		V_{IN} = 48 V		1.1		
		V_{IN} = 65 V		1		
Output voltage, V_{OUT1}	I_{OUT1} = 5 mA to 200 mA		14.7	15.0	15.3	V
Output voltage, V_{OUT2}	I_{OUT1} = 5 mA to 200 mA		–7.5	–7.7	–7.9	V
Output current, I_{OUT1}, $–I_{OUT2}$	V_{IN} = 12 V		0		100	mA
	V_{IN} = 24 V		0		220	
	V_{IN} = 48 V		0		300	
	V_{IN} = 65 V		0		350	
Switching frequency	V_{IN} = 24 V, $I_{OUT1} = –I_{OUT2}$ = 100 mA			350		kHz
Transformer ratio	1:1:0.52 (N_P, N_{S1}, N_{S2})					-
Full-load efficiency, η_{FULL}	$I_{OUT1} = –I_{OUT2}$ = 200 mA	V_{IN} = 24 V		88		%
		V_{IN} = 48 V		87.5		
		V_{IN} = 65 V		87		
Input capacitance C_{IN}	V_{IN} = 0 V			4.7		µF

1. Calculate EVM maximum input current I_{IN} for minimum and maximum input voltage ($V_{IN(VIN-MIN)}$, $V_{IN(VIN-MAX))}$ as per Equation 1. Approximate efficiency based on known data.

$$I_{IN} = \frac{\left(V_{OUT1} \times I_{OUT1}\right) + \left(V_{OUT2} \times I_{OUT2}\right)}{\eta_{FULL} \times V_{IN}}$$

$$I_{IN(VIN-MIN)} = \frac{(15 \times 0.2) + (-7.7 \times -0.2)}{0.88 \times 10} = 0.52\ \text{A}$$

$$I_{IN(VIN-MIN)} = \frac{(15 \times 0.2) + (-7.7 \times -0.2)}{0.87 \times 65} = 0.08\ \text{A} \tag{1}$$

2. Calculate the maximum power dissipation on the pre-regulator ($P_{PRE-REG}$) in the pre-regulator phase. The maximum input voltage of the pre-regulator is set to $V_{IN(PRE)}$ = 70 V.

$$P_{PRE-REG} = \left(V_{IN(PRE)} - V_{IN}\right) \times I_{IN(VIN-MAX)} = (70 - 65) \times 0.08 = 0.4\ \text{W} \tag{2}$$

3. Calculate the inrush current based on the input capacitance of the flyback converter and fastest transients on the 48-V rail. As an example, Equation 3 calculates the inrush current for the fastest transient from Figure 1.

$$I_{IN(INRUSH)} = \frac{C_{IN} \times \Delta V_{IN}}{t_r} = \frac{47\ \mu\text{F} \times (70 - 48)}{1\ \text{ms}} = 0.103\ \text{A} \tag{3}$$

4. Find an NPN transistor that satisfies parameters calculated in the previous steps. Additionally:
    - The collector-emitter breakdown voltage $V_{(BR)CEO}$ must be higher than 80 V.
    - The collector-emitter saturation voltage $V_{CE(SAT)}$ must be as low as possible to minimize power dissipation in the pass-through phase.
    - The DC current gain h_{FE} is as high as possible.
    - The package can withstand power dissipation.
    - Operating conditions do not exceed safety operating area (SOA) of the transistor.

    The selected transistor is the **NSS1C201L**, 100-V, 3.0-A, low VCE(SAT) NPN transistor.
5. Find a Zener diode that limits the output voltage of the pre-regulator to a maximum of 65 V as per Equation 4. Consider manufacturing tolerances and temperature drift. The selected Zener diode is the **BZX384-B62**. This is a 62-V Zener diode with 2% tolerance. The Zener voltage (V_Z) varies from 60.8 to 63.2 V at I_Z = 2 mA. Note that Zener diodes with a higher breakdown voltage also have a higher temperature coefficient. See the *Zener Theory and Design Considerations* handbook (HBD854/D), available online from onsemi.com, for more details on applications of Zener diodes.

$$V_{IN\,(PRE-REG)} = \left(V_{Z(ZD)} + V_{BE(Q1)}\right) \leq 65\text{ V} \tag{4}$$

6. Calculate the resistor R1. Select its value so that it allows current (I_Z) between 1 to 2 mA through the Zener diode at $V_{IN(PRE-REG)}$ = 70 V. A lower value increases the current through the Q1 during the start-up phase. However, it also increases the current through the Zener diode during the pre-regulator phase. Make sure you do not exceed the power rating of the diode. The selected resistor value is **R1 = 7.5 kΩ**.

$$R1 = \frac{V_{IN(PRE-REG)} - V_{Z(ZD)}}{I_{Z(ZD)}} = \frac{70-62}{0.001} = 8\text{ k}\Omega \tag{5}$$

7. Select the diode D1. Practically any 100-V rated silicon diode will work. The selected is **BAT46WJ,115**.
8. Measure or calculate the reflected voltage on the switch node as per Equation 6. Where V_F is the forward voltage drop on the diode FD ≈ 0.6 V.

$$V_{REF} = \frac{V_{OUT1} - V_{OUT2} + 2 \times V_F}{\frac{N_{S1} + N_{S2}}{N_P}} = \frac{15.0 - (-7.7) + 2 \times 0.6}{\frac{1 + 0.52}{1}} = 15.72\text{ V} \tag{6}$$

9. The capacitor C1 will charge periodically to the reflected voltage lowered by the voltage drop on the diode D1 as per Equation 7.

$$V_{C1} = V_{REF} - V_{F(D1)} = 15.72 - 0.6 = 15.12\text{ V} \tag{7}$$

10. Interpolate the minimum DC current gain of the Q1 from the data sheet. We estimate h_{FE} = 100 for I_C = 500 mA. Calculate the resistor R2 as per Equation 8. The selected resistor value is **R2 = 2.2 kΩ**.

$$R2 \leq \frac{V_{C1} - V_{BE(Q1)}}{\frac{I_{IN(VIN-MIN)}}{h_{FE(Q1)}}} = \frac{15.12 - 0.65}{\frac{0.53}{100}} = 2.73\text{ k}\Omega \tag{8}$$

11. Calculate the capacitor C1 per Equation 9. The capacitor C1 needs to bias the base-emitter junction during the off time. We consider the absolute worst case when the LM5180-Q1 device operates at minimum switching frequency $f_{sw(min)}$ = 12 kHz with nearly zero duty cycle. We assume that the voltage on C1 does not drop more than 20% of the maximum value. Multi-layer capacitors suffer from DC voltage derating when effective capacitance reduces with voltage bias. For this reason we select **C2 = 220 nF, X7R, 50 V**.

$$C1 \leq \frac{\frac{V_{C1} - V_{BE(Q1)}}{R2} \times \frac{1}{f_{SW(MIN)}}}{0.2 \times V_{C1}} = \frac{\frac{15.72 - 0.65}{2200} \times \frac{1}{12000}}{0.2 \times 15.72} = 181\text{ nF} \tag{9}$$

4 Further Improvements, Considerations, and Testing

There are scenarios when input voltage suddenly lowers or the input even shorts to ground. The input capacitor of the flyback converter may not be able to discharge fast enough. During this event, the collector voltage is lower than the emitter voltage. This essentially applies reverse voltage to the base-emitter junction. Exceeding emitter-base breakdown voltage $V_{(BR)EBO}$ permanently damages the transistor Q1. Placing a diode D2 across the collector-emitter terminals protects the transistor Q1 from the damage. See Figure 4, for the circuit details.

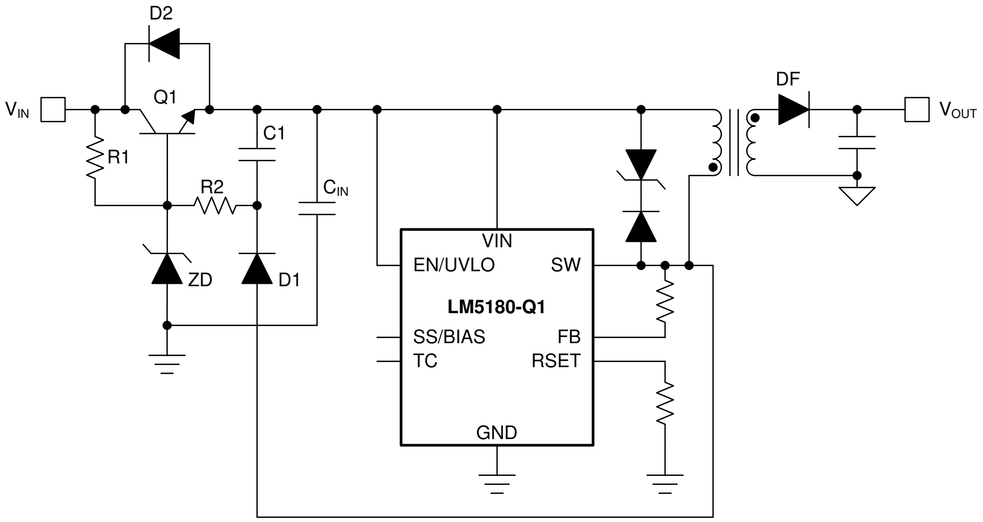


Figure 4. Protecting the Base-Emitter Junction With the Diode D2

Section 3 has already mentioned a temperature drift of the ZD. The temperature coefficient depends on the Zener voltage V_Z. In general, Zener diodes below $V_Z \approx 5$ V have a negative temperature coefficient whereas higher Zener voltages have a positive temperature coefficient. The most thermally stable Zener diodes are in the 4.5 to 5.5 V region.

Selecting the Zener diode the way that the output never exceeds maximum input voltage 65 V is important. Additionally, the Zener diode should not limit the output below 60 V for the long-term overvoltage test. Figure 5 shows the worst case Zener voltage variation for given temperature coefficients S_Z and the initial accuracy of V_Z.

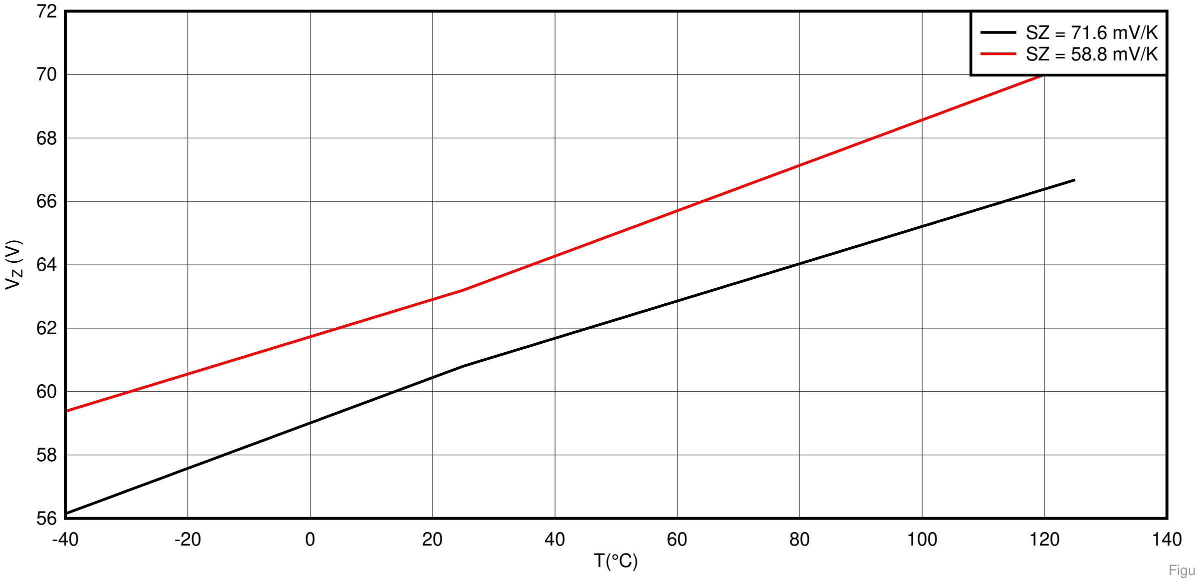


Figure 5. BZX84-B62 Zener Voltage V_Z Variation due to the Manufacturing Span and Temperature Drift

As the figure shows, it is impossible to maintain the stable output voltage of the pre-regulator in the desired 60- to 65-V region.

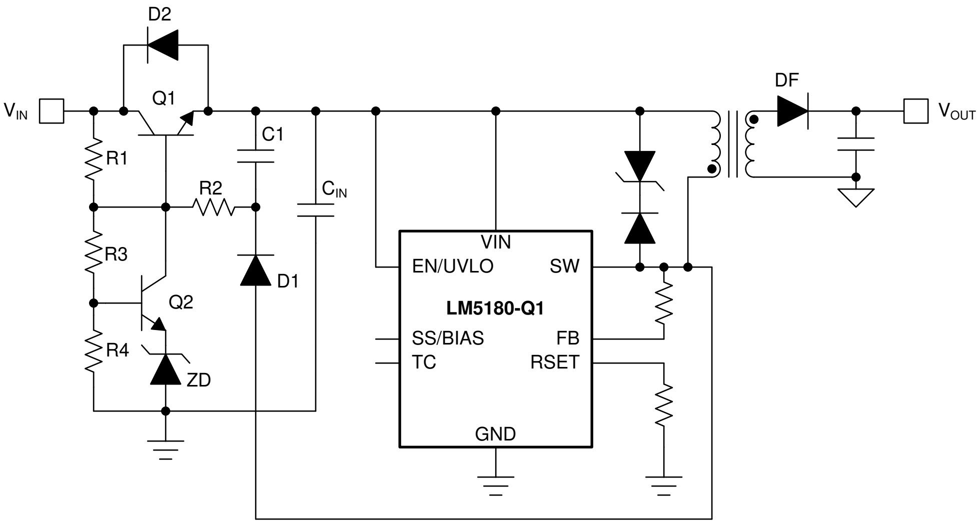


Figure 6. Improved Thermal Stability of the Pre-regulator

Figure 6 shows an improved circuit that uses an additional bipolar transistor for temperature compensation. The base-emitter junction has a temperature coefficient of approximately –2 mV/K. The idea is selecting a Zener diode with an opposite temperature coefficient. For example, BZX384-B6V2 has a typical SZ = 2.3 mV/K. This means that temperature drift nearly compensates. This keeps the voltage on the resistor R4 stable. The collector voltage is then as per Equation 10. The formula can omit the base current I_B for $I_B \ll I_{R4}$.

$$V_{C(Q2)} \cong \frac{V_{BE(Q2)} + V_{ZD}}{R4} \times (R3 + R4) \qquad (10)$$

Figure 7 shows the variation of the base voltage from the circuit shown in Figure 6. Note that the plot considers only the variation of the working voltage V_Z and temperature coefficients S_Z.

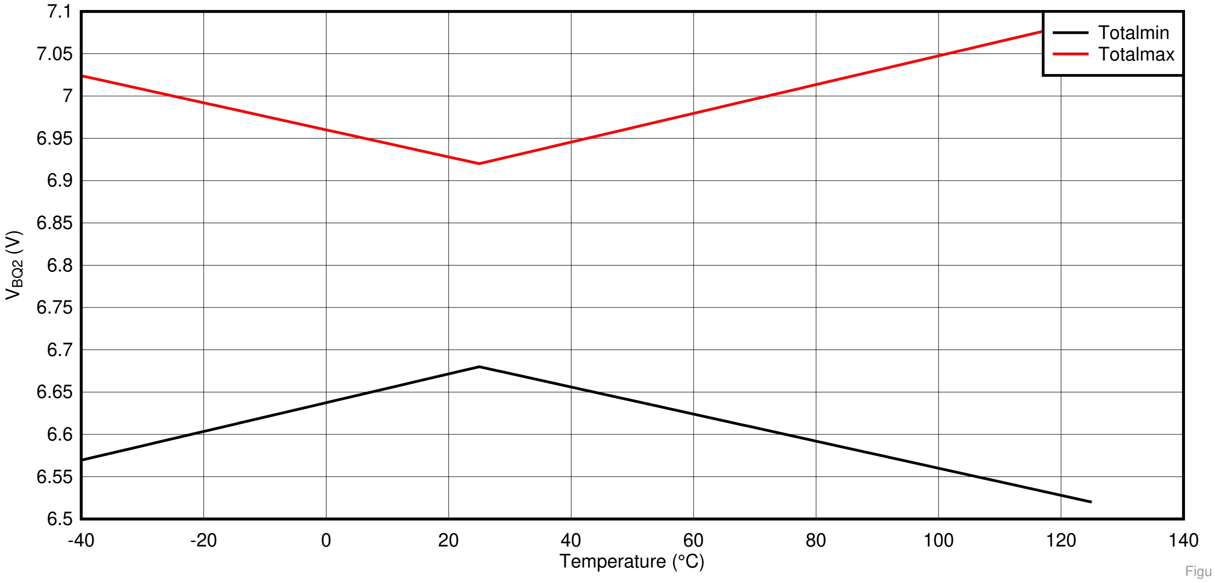


Figure 7. BZX384-B6V2 and NSS1C201L Base-Voltage Temperature Drift From Figure 6 ($I_{E2} = I_{ZD}$ = 5 mA)

Figure 8 shows the output voltage of the pre-regulator (emitter voltage of Q1) that considers temperature coefficients and tolerances of transistors Q1, Q2, and the diode ZD. The plot does not consider resistors tolerance and drift. The resistance ratio R3:R4 is 8:1.

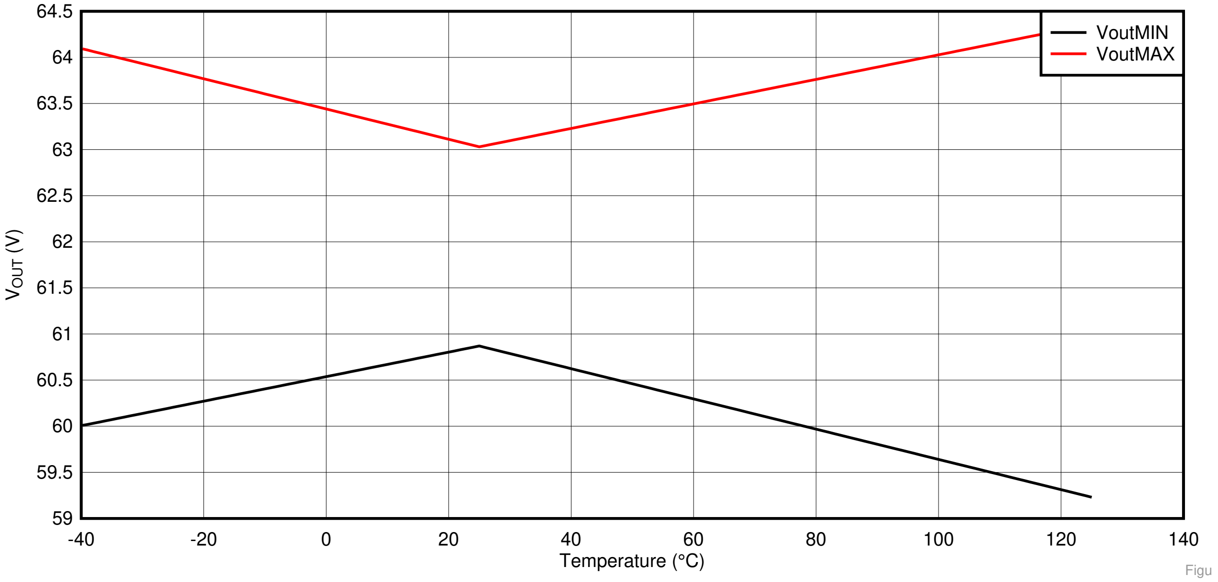


Figure 8. BZX384-B6V2 and NSS1C201L Pre-regulator Output Temperature Drift (I_{C2}, I_{ZD} = 5 mA, I_{E1} = 500 mA)

The result is already satisfactory but requires careful selection of components and a good understanding of parameters in the data sheet.

Probably the most elegant solution, that ensures output voltage accuracy, uses a 431-type of a shunt regulator as shown in Figure 9. The 431 shunt regulator requires the transistor Q2 due to the maximum cathode-anode voltage which is typically 36 V. The benefit of this circuit is that the shunt regulator maintains the stable voltage on the resistor R5. This way it eliminates the base-emitter voltage drop for both transistors Q1, Q2 and their thermal drift because it senses the feedback directly from the output of the pre-regulator. This is also the only circuit that can use N-FET transistors in place of Q1 and Q2. The variation of the threshold voltage V_{TH} is too high for previously described circuits with a Zener diode. Additionally, an N-FET transistor in the place of Q1 reduces the power dissipation in the pass-through mode due to a low on-resistance $R_{DS(on)}$. Replacing the Q2 with an N-FET eliminates the base current, therefore the error voltage on R3.

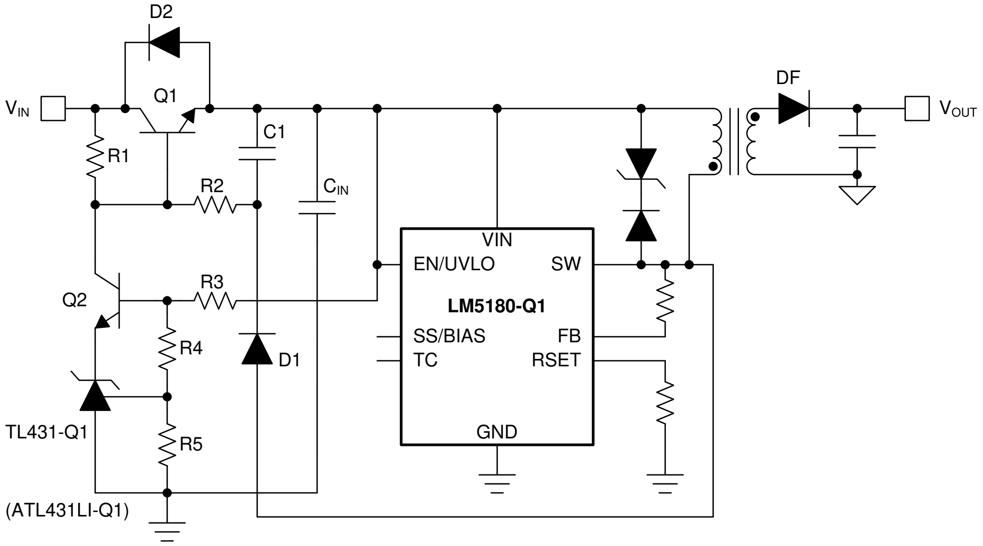


Figure 9. Output Voltage Stabilization Using the 431 Shunt Regulator

The ATL431LI-Q1 device comes in 0.5- and 1-percent tolerance grade. The reference voltage is typically 2.5 V with only 27 mV deviation over the complete temperature range. Table 2 summarizes the performance of the circuit. The table can omit the base current $I_{B(Q2)}$ for $I_{B(Q2)} \ll I_{R5}$.

Table 2. ATL431LI-Q1-Based Pre-regulator Output Characteristics

Parameter	Test conditions	MIN	TYP	MAX	Unit
Reference voltage (V_{R5})	$I_{E(Q2)}$ = 1 mA	2475	2500	2525	mV
Deviation of the reference voltage over the full temperature range	$I_{E(Q2)}$ = 1 mA		10	27	mV
Reference voltage (V_{R5MAX}, V_{R5MIN})	Over complete temperature range –40 to 125°C and the initial accuracy	2448		2552	mV
Gain (R3 + R4) / R5	For V_{CLAMP} = 62 V, 1% tolerance resistors	24.55	24.8	25.05	-
Output voltage of the pre-regulator	Overtemperature range and manufacturing tolerances	60.1	62	63.9	V

5 Test Setup

Figure 10, Figure 11, and Figure 12 show the circuit used for test measurements. The circuits connects to the LM5180EVM-DUAL evaluation board. The transistor Q2 is N-FET to mitigate the base current. A 5-W, 220-Ω resistor between V_{OUT1} and V_{OUT2} terminals act as a load. The input filter has been removed from the EVM to reduce the inrush current. Instead, the circuit implements two optional 4.7 =µF, 100-V, and X7R capacitors, each at the input and output of the pre-regulator. The circuit has been tested with the TL431A-Q1 and ATL431LI-Q1 devices. The ATL431LI-Q1 device offers a higher bandwidth, better temperature stability, and lower operating currents for the similar price point.

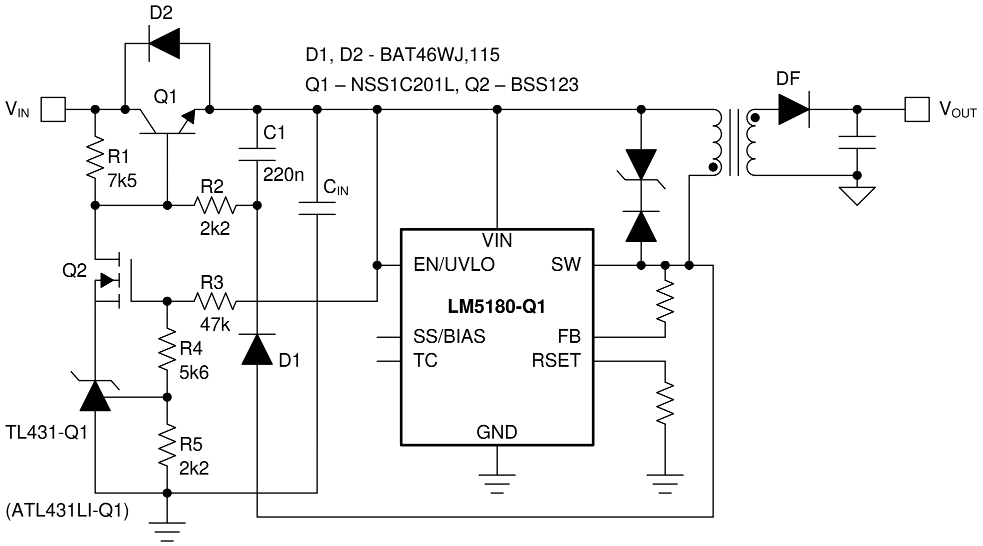


Figure 10. Circuit Tested With the TL431-Q1 (ATL431LI-Q1) Shunt Regulator

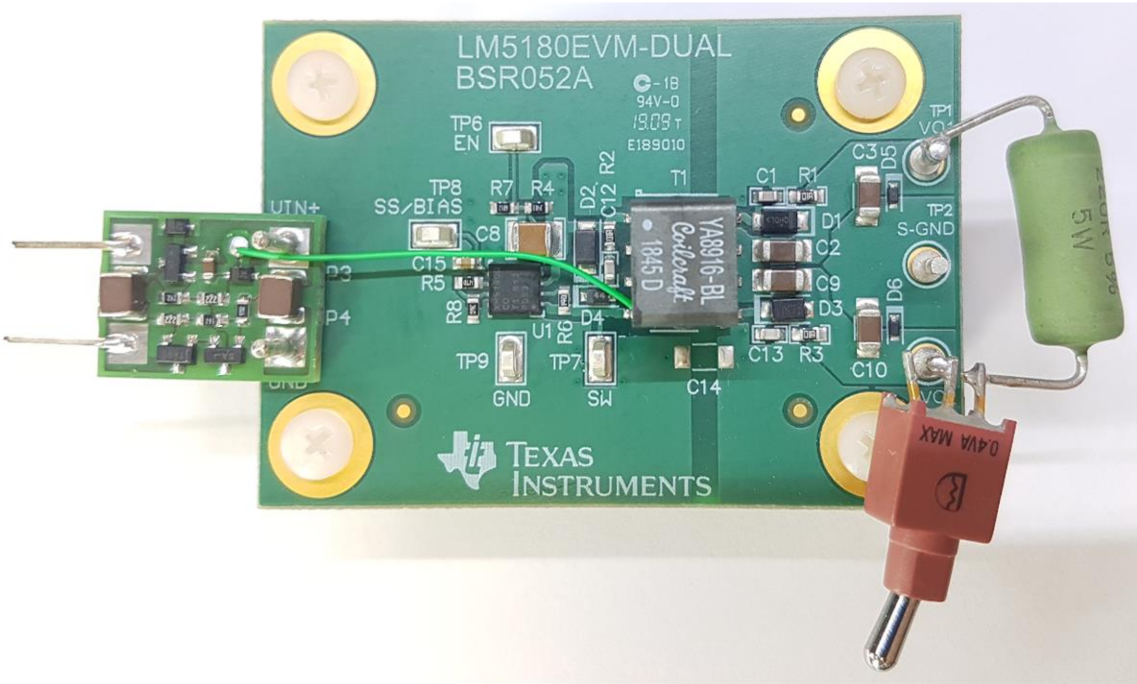


Figure 11. LM5180EVM-DUAL and the Pre-regulator Board (Top Side)

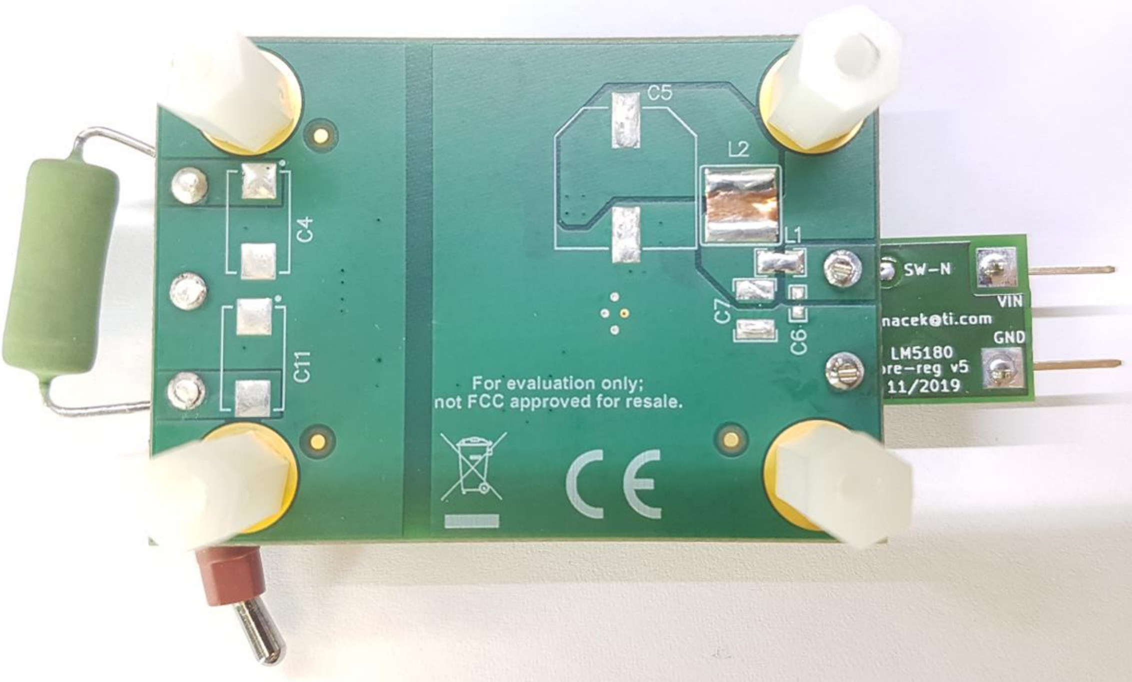


Figure 12. LM5180EVM-DUAL and the Pre-regulator Board (Bottom Side)

6 Measurements

Figure 13 shows waveforms for the basic circuit with a Zener diode shown in Figure 3. The dark blue trace, C1 is the input voltage. The pink trace C2 is the output voltage of the pre-regulator. Note that the output voltage slightly rises in the pre-regulator mode. This is due to the self-heating effect when current flows through the Zener diode. The light blue trace C3 is the emitter-collector voltage on the transistor Q1.

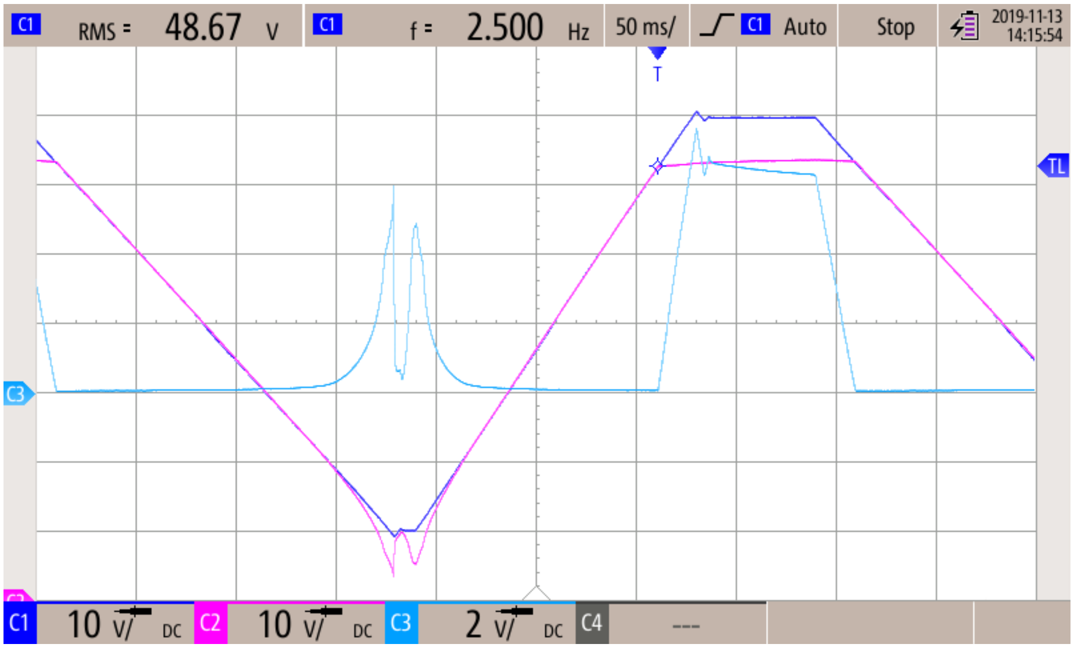


Figure 13. Waveforms for the Circuit With a Zener Diode

Figure 14 and Figure 15 show same waveforms for the circuit with the TL431-Q1 shunt voltage regulator. The output voltage is limited very close to the calculated value and remains stable under all operating conditions. The blue trace C1 is the input voltage. The pink trace C2 is the output voltage of the pre-regulator.

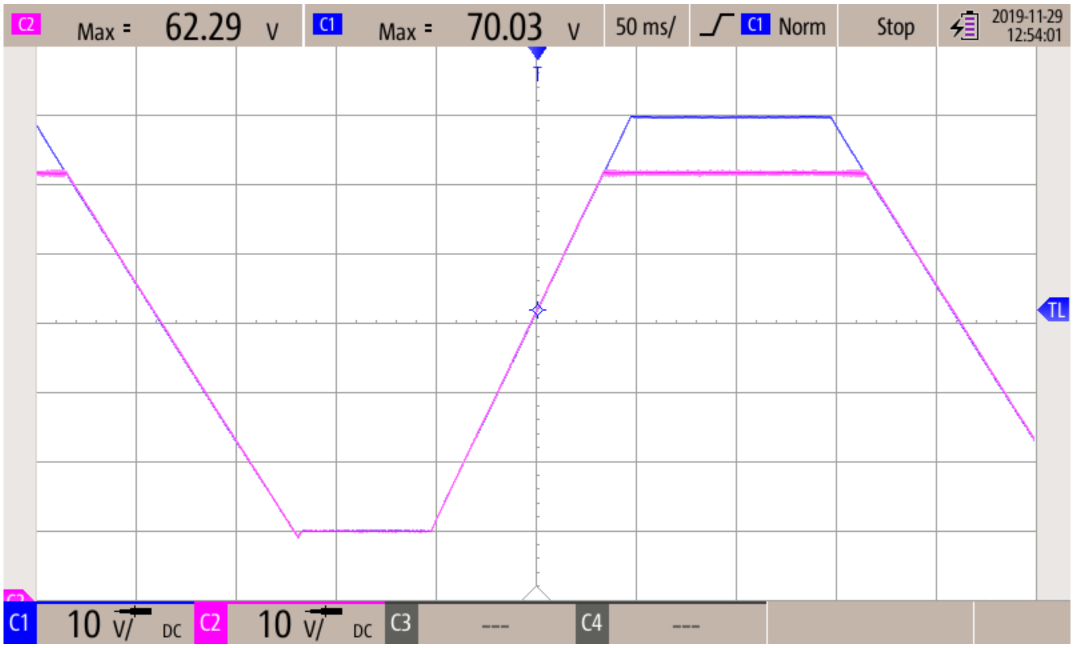


Figure 14. Waveforms for the Circuit with the TL431-Q1 Voltage Regulator (No Load)

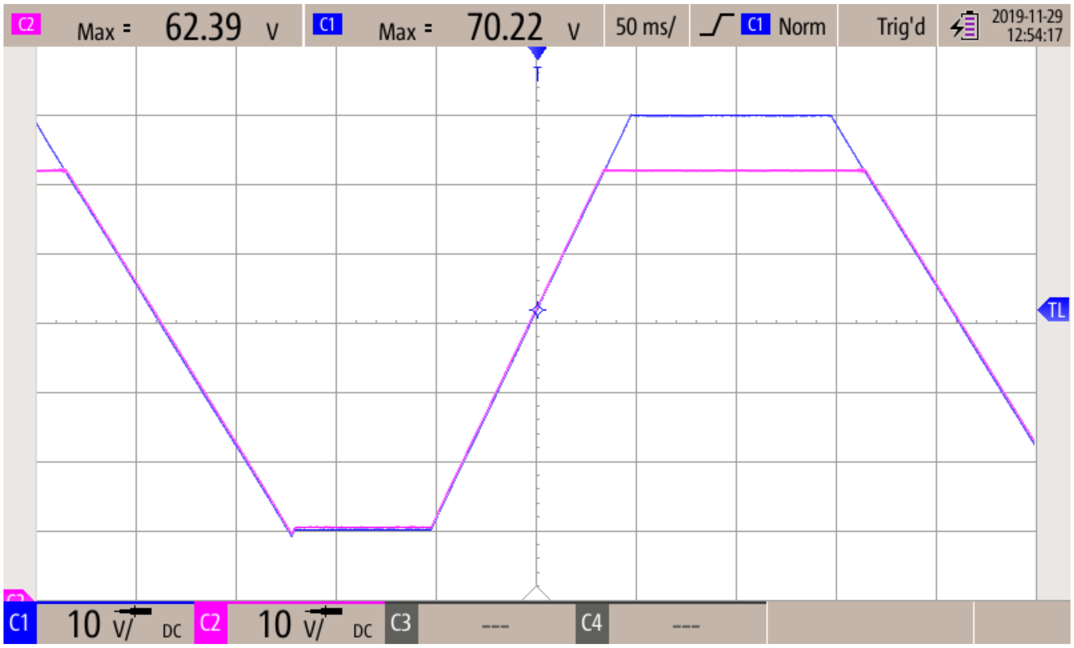


Figure 15. Waveforms for the Circuit With the TL431-Q1 Voltage Regulator (EVM Loaded With a 220-Ω Resistor)

Figure 16 through Figure 23 show the transient response of the pre-regulator circuit with the TL431-Q1 shunt voltage regulator. The dark blue trace C1 is the input voltage. The pink trace C2 is the output voltage of the pre-regulator. The light blue trace C3 is the output voltage ($–V_{OUT1}$ + V_{OUT2}) of the flyback converter (EVM).

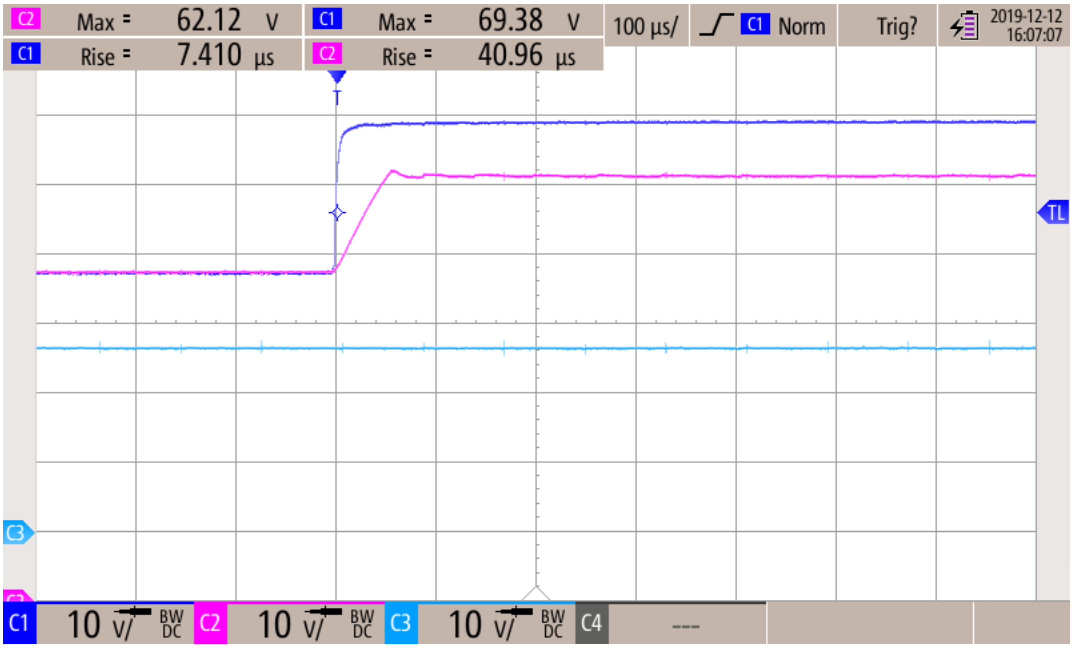


Figure 16. Transient Response From 48 V to 70 V (No Load)

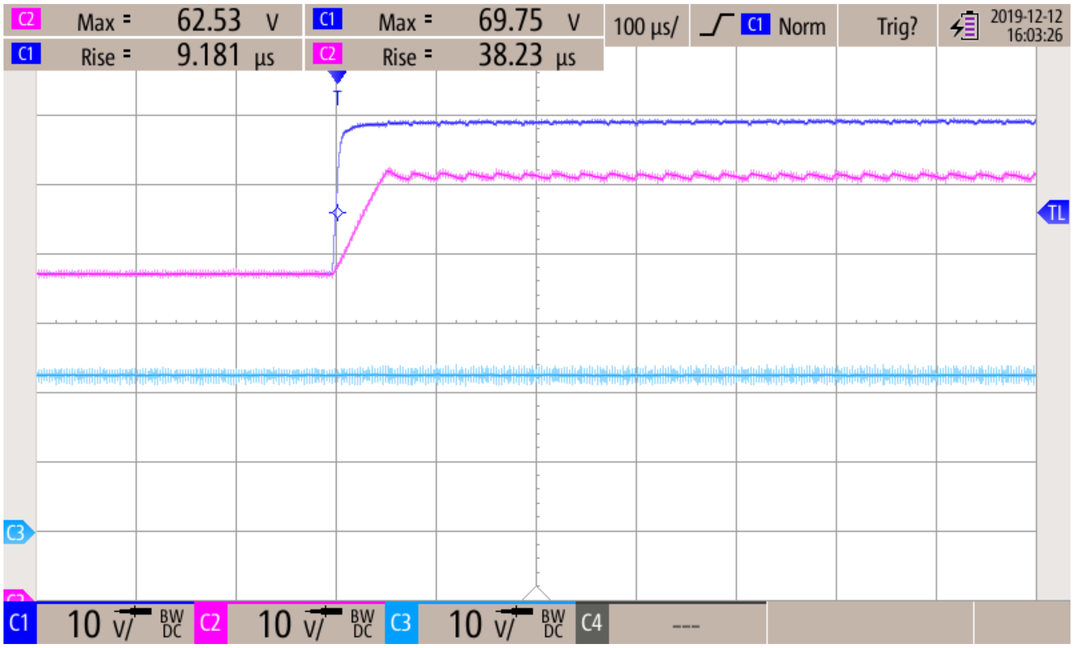


Figure 17. Transient Response From 48 V to 70 V (EVM Loaded With a 220-Ω Resistor)

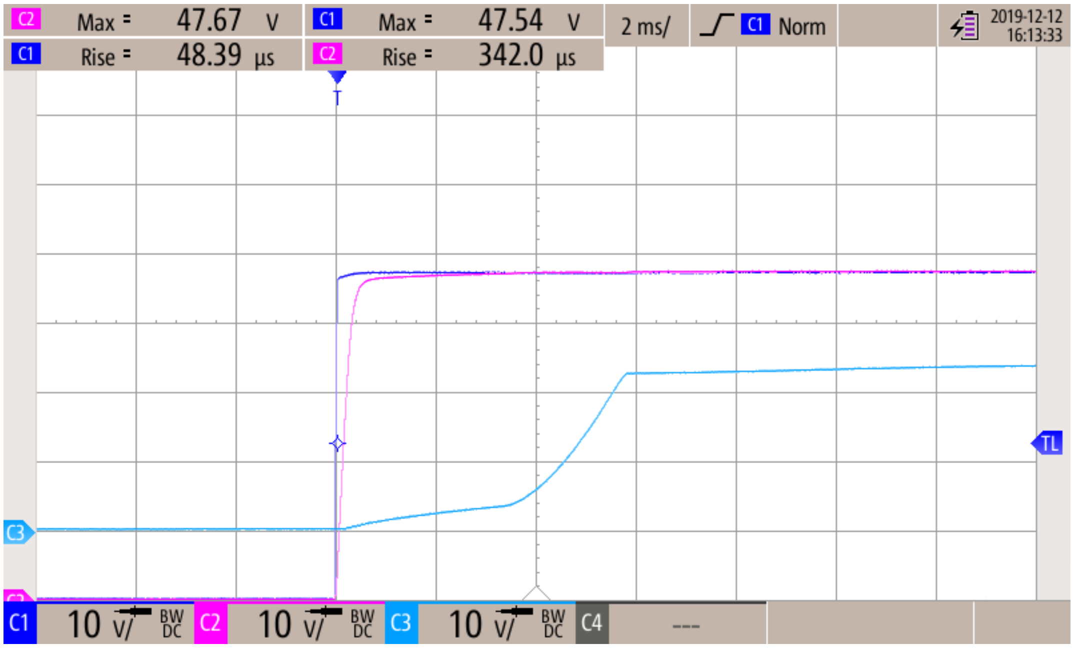


Figure 18. Cold Start From 0 V to 48 V (No Load)

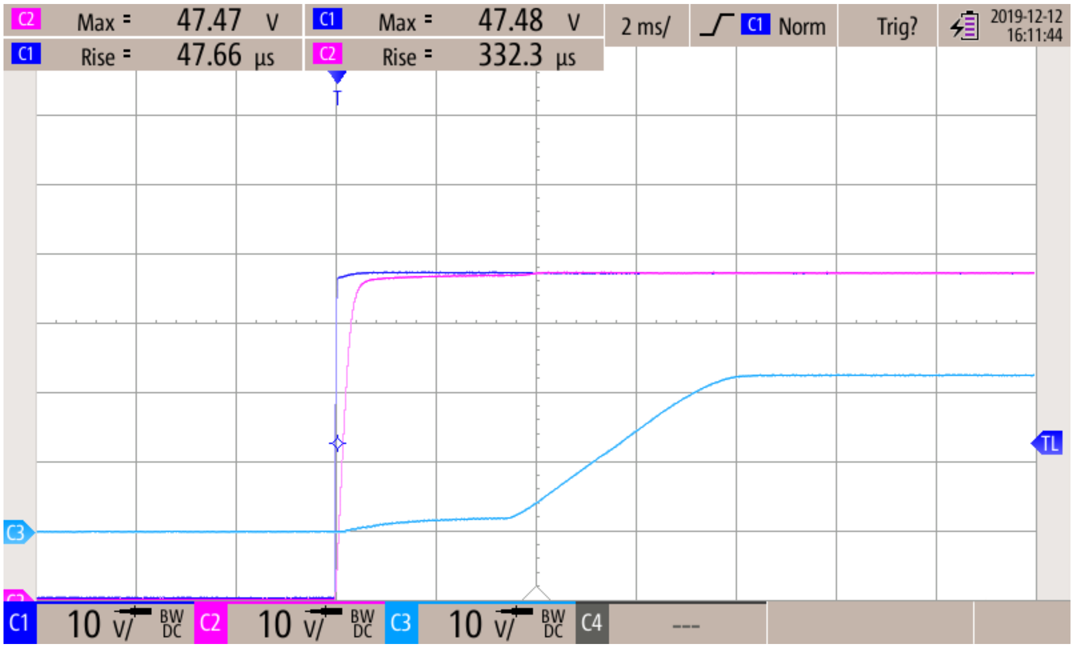


Figure 19. Cold Start From 0 V to 48 V (EVM Loaded With a 220-Ω Resistor)

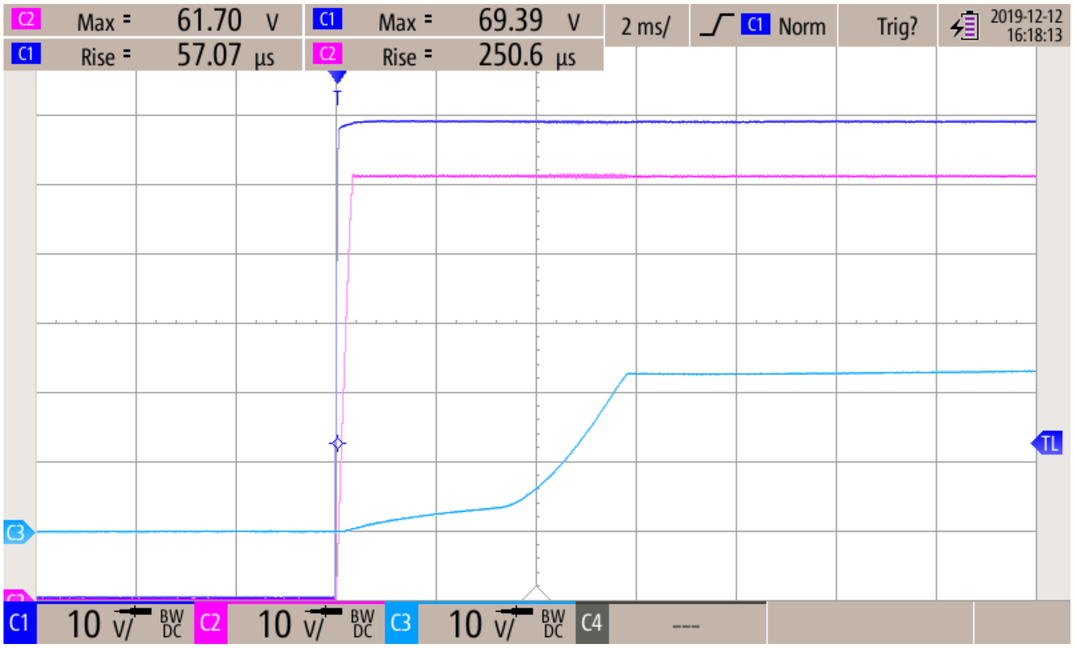


Figure 20. Cold Start From 0 V to 70 V (No Load)

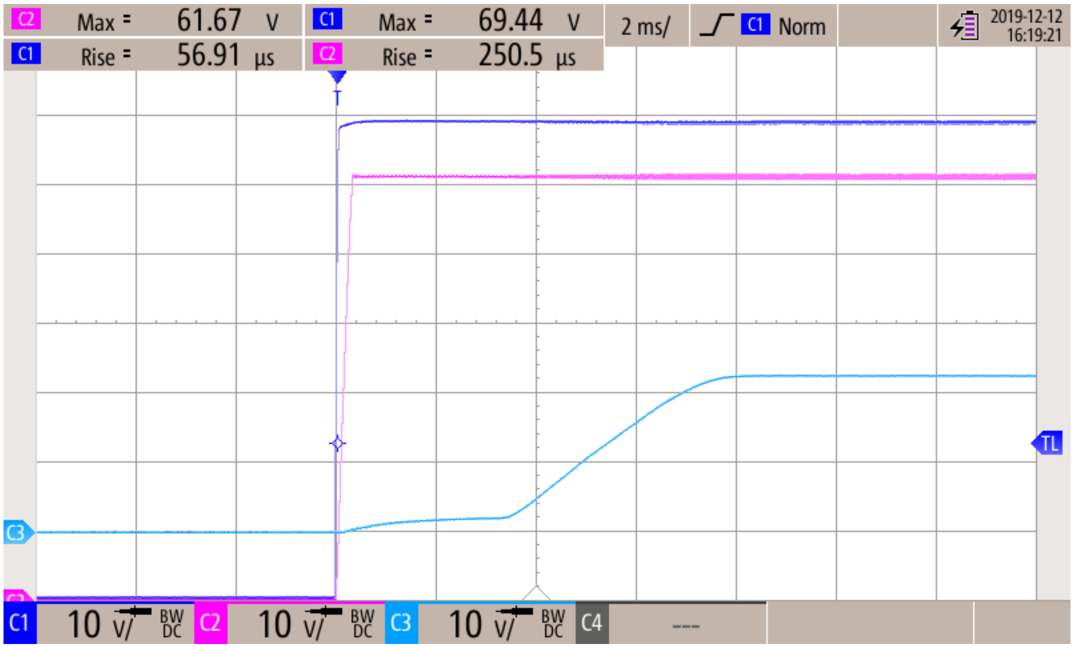


Figure 21. Cold Start From 0 V to 70 V (EVM Loaded With a 220-Ω Resistor)

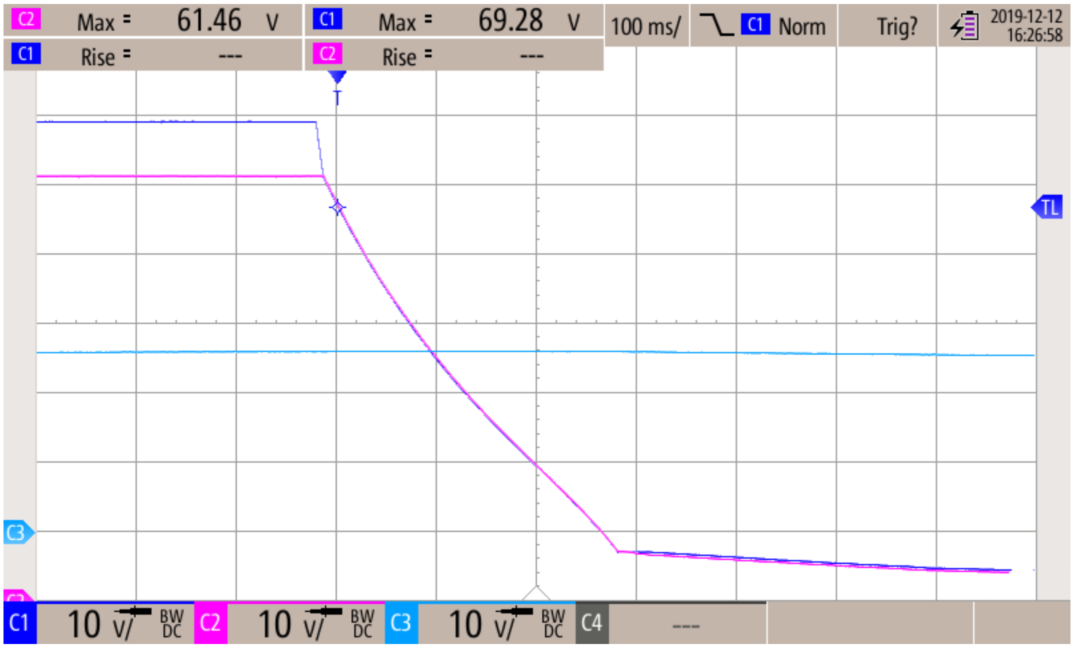


Figure 22. Turn-off Behavior From 70 V to 0 V (No Load)

Figure 23. Turn-off Behavior From 70 V to 0 V (EVM Loaded With a 220-Ω Resistor)

The circuit performance with the ATL431LI-Q1 is practically identical. The transient response is slightly better with a little overshoot. Compare Figure 17 and Figure 24.

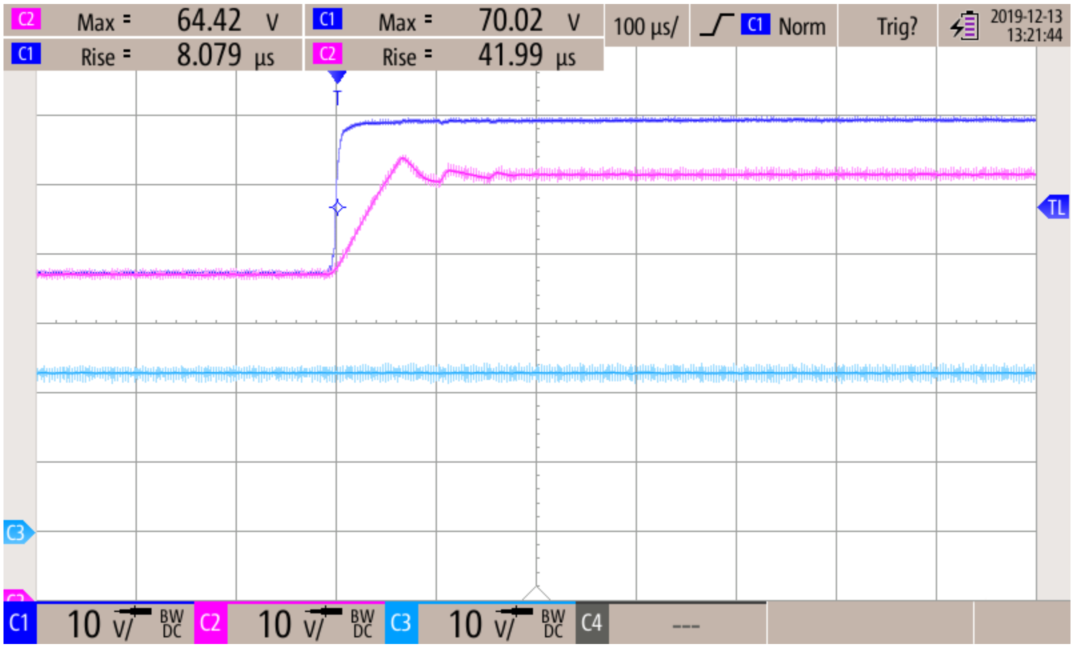


Figure 24. Transient Response From 48 V to 70 V for the Circuit With ATL431LI-Q1 (EVM Loaded With a 220-Ω Resistor)

7 Commercial Aspects

Table 3 lists parts used in the test circuit. Prices are informative, obtained from the internet for ten thousand units.

Table 3. Cost Structure Analysis (Estimated)

Reference Designator	Type	Quantity	Unit Price (USD)	Total (USD)
R1, R2, R3, R4, R5	0603, Thick Film, 1% tolerance, AEC-Q	5	0.003	0.015
C1	0603, X7R, 220 nF, 50 V	1	0.020	0.020
D1, D2	BAT46WJ,115	2	0.040	0.080
IC1	ATL431LI-Q1	1	0.170	0.170
Q2	BSS123	1	0.040	0.040
Q1	NSS1C201L	1	0.081	0.081
Total				0.406

8 Conclusion

The tested circuit extends the input voltage range of the LM5180-Q1 device and allows for powering the device directly from the 48-V rail. Extensive testing has confirmed proper behavior during various operating modes and transient events. The transistor Q1 requires careful selection for the given inrush current and safety operating area (SOA). Additionally, the circuit can be modified for use in buck converters. Additional complexity may still allow for cheaper system implementation and avoiding more expensive flyback or buck controllers.

9 References

- (December 2017). *Zener Theory and Design Considerations*. HBD854/D (Revision 1). Retrieved from https://www.onsemi.com/pub/Collateral/HBD854-D.PDF

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