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National Semiconductor

January 21, 2010

LM3421
LM3421Q1 LM3421Q0
LM3423
LM3423Q1 LM3423Q0

N-Channel Controllers for Constant Current LED Drivers

General Description

The LM3421/23 are versatile high voltage N-channel MosFET controllers for LED drivers . They can be easily configured in buck, boost, buck-boost and SEPIC topologies. This flexibility, along with an input voltage rating of 75V, makes the LM3421/23 ideal for illuminating LEDs in a large family of applications.

Adjustable high-side current sense voltage allows for tight regulation of the LED current with the highest efficiency possible. The LM3421/23 uses Predictive Off-time (PRO) control, which is a combination of peak current-mode control and a predictive off-timer. This method of control eases the design of loop compensation while providing inherent input voltage feed-forward compensation.

The LM3421/23 devices include a high-voltage startup regulator that operates over a wide input range of 4.5V to 75V. The internal PWM controller is designed for adjustable switching frequencies of up to 2.0 MHz, thus enabling compact solutions. Additional features include "zero current" shutdown, analog dimming, PWM dimming, over-voltage protection, under-voltage lock-out, cycle-by-cycle current limit, and thermal shutdown.

The LM3423 also includes an LED output status flag, a fault flag, a programmable fault timer, and a logic input to select the polarity of the dimming output driver.

The LM3421Q1/23Q1 are AEC-Q100 grade 1 qualified and LM3421Q0/23Q0 are AEC-Q100 grade 0 qualified.

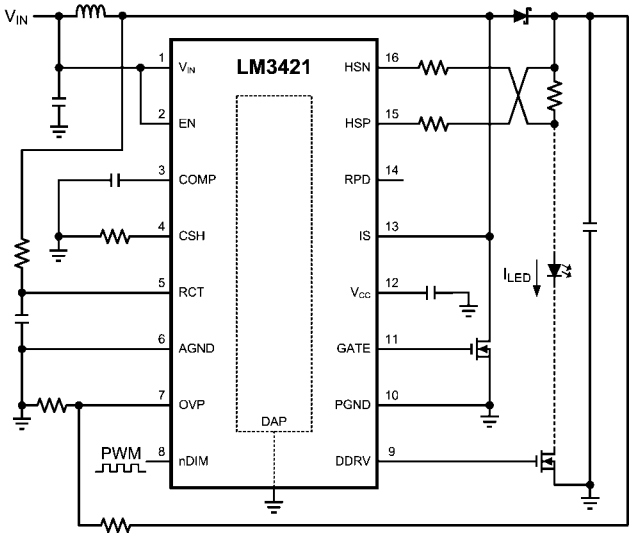
Features

- LM3421Q1/LM3423Q1 are Automotive Grade products that are AEC-Q100 grade 1 qualified (-40°C to +125°C operating junction temperature) and similarly LM3421Q0/LM3423Q0 are AEC-Q100 grade 0 qualified (-40°C to +150°C operating junction temperature)
- V_{IN} range from 4.5V to 75V
- High-side adjustable current sense
- 2Ω, 1A Peak MosFET gate driver
- Input under-voltage and output over-voltage protection
- PWM and analog dimming
- Cycle-by-cycle current limit
- Programmable switching frequency
- "Zero current" shutdown and thermal shutdown
- LED output status flag (LM3423/23Q1/23Q0 only)
- Fault status flag and timer (LM3423/23Q1/23Q0 only)

Applications

- LED Drivers - Buck, Boost, Buck-Boost, and SEPIC
- Indoor and Outdoor Area SSL
- Automotive
- General Illumination
- Constant-Current Regulators

Typical Boost Application Circuit



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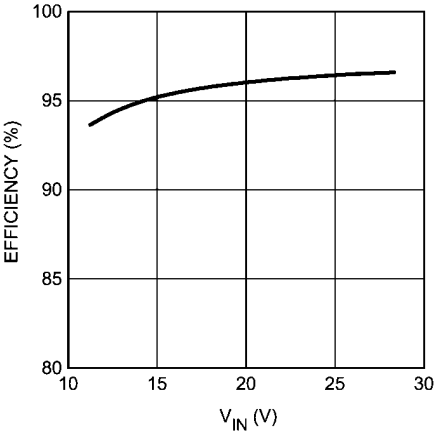

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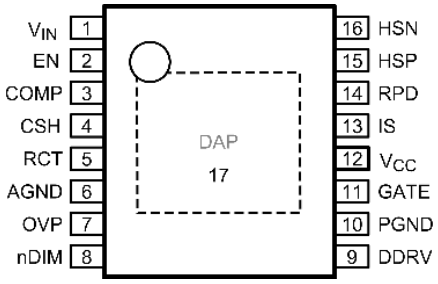
Boost Evaluation Board
9 Series LEDs at 1A

Connection Diagrams

Top View



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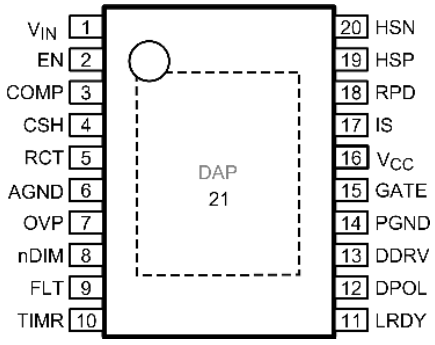
16-Lead TSSOP EP
NS Package Number MXA16A

Top View



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20-Lead TSSOP EP
NS Package Number MXA20A

Ordering Information

Order Number	Spec.	Package Type	NSC Package Drawing	Supplied As	Features
LM3421MH	NOPB	TSSOP-16 EP	MXA16A	92 Units, Rail	
LM3421MHX	NOPB	TSSOP-16 EP	MXA16A	2500 Units, Tape and Reel	
LM3423MH	NOPB	TSSOP-20 EP	MXA20A	73 Units, Rail	
LM3423MHX	NOPB	TSSOP-20 EP	MXA20A	2500 Units, Tape and Reel	
LM3421Q1MH	NOPB	TSSOP-16 EP	MXA16A	92 Units, Rail	AEC-Q100 Grade 1 qualified. Automotive Grade Production Flow*
LM3421Q1MHX	NOPB	TSSOP-16 EP	MXA16A	2500 Units, Tape and Reel	
LM3423Q1MH	NOPB	TSSOP-20 EP	MXA20A	73 Units, Rail	
LM3423Q1MHX	NOPB	TSSOP-20 EP	MXA20A	2500 Units, Tape and Reel	
LM3421Q0MH	NOPB	TSSOP-16 EP	MXA16A	92 Units, Rail	AEC-Q100 Grade 0 qualified. Automotive Grade Production Flow*
LM3421Q0MHX	NOPB	TSSOP-16 EP	MXA16A	2500 Units, Tape and Reel	
LM3423Q0MH	NOPB	TSSOP-20 EP	MXA20A	73 Units, Rail	
LM3423Q0MHX	NOPB	TSSOP-20 EP	MXA20A	2500 Units, Tape and Reel	

*Automotive Grade (Q) product incorporates enhanced manufacturing and support processes for the automotive market, including defect detection methodologies. Reliability qualification is compliant with the requirements and temperature grades defined in the AEC-Q100 standard. Automotive grade products are identified with the letter Q. For more information go to http://www.national.com/automotive.

Pin Descriptions

LM3423	LM3421	Name	Description	Function
1	1	V_{IN}	Input Voltage	Bypass with 100 nF capacitor to AGND as close to the device as possible in the circuit board layout.
2	2	EN	Enable	Connect to AGND for zero current shutdown or apply > 2.4V to enable device.
3	3	COMP	Compensation	Connect a capacitor to AGND to set the compensation.
4	4	CSH	Current Sense High	Connect a resistor to AGND to set the signal current. For analog dimming, connect a controlled current source or a potentiometer to AGND as detailed in the Analog Dimming section.
5	5	RCT	Resistor Capacitor Timing	External RC network sets the predictive “off-time” and thus the switching frequency.
6	6	AGND	Analog Ground	Connect to PGND through the DAP copper pad to provide ground return for CSH, COMP, RCT, and TIMR.
7	7	OVP	Over-Voltage Protection	Connect to a resistor divider from V_O to program output over-voltage lockout (OVLO). Turn-off threshold is 1.24V and hysteresis for turn-on is provided by 23 µA current source.
8	8	nDIM	Dimming Input / Under-Voltage Protection	Connect a PWM signal for dimming as detailed in the PWM Dimming section and/or a resistor divider from V_{IN} to program input under-voltage lockout (UVLO). Turn-on threshold is 1.24V and hysteresis for turn-off is provided by 23 µA current source.
9	-	FLT	Fault Flag	Connect to pull-up resistor from VIN and N-channel MosFET open drain output is high when a fault condition is latched by the timer.
10	-	TIMR	Fault Timer	Connect a capacitor to AGND to set the time delay before a sensed fault condition is latched.
11	-	LRDY	LED Ready Flag	Connect to pull-up resistor from VIN and N-channel MosFET open drain output pulls down when the LED current is not in regulation.
12	-	DPOL	Dim Polarity	Connect to AGND if dimming with a series P-channel MosFET or leave open when dimming with series N-channel MosFET.
13	9	DDRV	Dim Gate Drive Output	Connect to the gate of the dimming MosFET.
14	10	PGND	Power Ground	Connect to AGND through the DAP copper pad to provide ground return for GATE and DDRV.
15	11	GATE	Main Gate Drive Output	Connect to the gate of the main switching MosFET.
16	12	V_{CC}	Internal Regulator Output	Bypass with 2.2 µF–3.3 µF ceramic capacitor to PGND.
17	13	IS	Main Switch Current Sense	Connect to the drain of the main N-channel MosFET switch for R_{DS-ON} sensing or to a sense resistor installed in the source of the same device.
18	14	RPD	Resistor Pull Down	Connect the low side of all external resistor dividers (V_{IN} UVLO, OVP) to implement “zero-current” shutdown.
19	15	HSP	LED Current Sense Positive	Connect through a series resistor to the positive side of the LED current sense resistor.
20	16	HSN	LED Current Sense Negative	Connect through a series resistor to the negative side of the LED current sense resistor.
DAP (21)	DAP (17)	DAP	Thermal PAD on bottom of IC	Star ground, connecting AGND and PGND. For thermal considerations please refer to (Note 2).

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/ Distributors for availability and specifications.

V_{IN}, EN, RPD, nDIM	-0.3V to 76.0V -1 mA continuous
OVP, HSP, HSN, LRDY, FLT, DPOL	-0.3V to 76.0V -100 µA continuous
RCT	-0.3V to 76.0V -1 mA to +5 mA continuous
IS	-0.3V to 76.0V -2V for 100 ns -1mA continuous
V_{CC}	-0.3V to 8.0V
TIMR	-0.3V to 7.0V -100µA to +100µA Continuous
COMP, CSH	-0.3V to 6.0V -200 µA to +200 µA Continuous
GATE, DDRV	-0.3V to V_{CC} -2.5V for 100 ns V_{CC}+2.5V for 100 ns -1 mA to +1 mA continuous
PGND	-0.3V to 0.3V -2.5V to 2.5V for 100 ns
Maximum Junction Temperature	Internally Limited
Storage Temperature Range	–65°C to +150°C
Maximum Lead Temperature (Solder and Reflow) (Note 3)	260°C
Continuous Power Dissipation	Internally Limited
ESD Susceptibility (Note 4)	
Human Body Model	2 kV
Charge Device Model	500V CSH pin 750V all other pins

Operating Conditions (Note 1)

Operating Junction Temperature Range	
LM3421, LM3421Q1, LM3423, LM3423Q1	–40°C to +125°C
LM3421Q0, LM3423Q0	–40°C to +150°C
Input Voltage V_{IN}	4.5V to 75V

Electrical Characteristics (Note 1)

Specifications in standard type face are for T_J = 25°C and those with **boldface type** apply over the full **Operating Temperature Range** (T_J = –40°C to +150°C for LM3421Q0/LM3423Q0, T_J = –40°C to +125°C for all others). Specifications that differ between the two operating ranges will be identified in the **Temp Range** column as Q0 for T_J = –40°C to +150°C and as Q1 for T_J = –40°C to +125°C. If no temperature range is indicated then the specification holds for both Q1 and Q0. Minimum and Maximum limits are guaranteed through test, design, or statistical correlation. Typical values represent the most likely parametric norm at T_J = +25°C, and are provided for reference purposes only. Unless otherwise stated the following condition applies: V_{IN} = +14V.

Symbol	Parameter	Conditions	Temp Range	Min (Note 5)	Typ (Note 6)	Max (Note 5)	Units
STARTUP REGULATOR							
V_{CCREG}	V_{CC} Regulation	I_{CC} = 0 mA		**6.30**	6.90	**7.35**	V
I_{CCLIM}	V_{CC} Current Limit	V_{CC} = 0V		**20**	25		mA
I_Q	Quiescent Current	EN = 3.0V, Static	Q1		2	**3**	
			Q0			**3.5**	
I_{SD}	Shutdown Current	EN = 0V			0.1	1.0	µA
V_{CC} SUPPLY							
V_{CCUV}	V_{CC} UVLO Threshold	V_{CC} Increasing			4.17	**4.50**	V
		V_{CC} Decreasing		**3.70**	4.08		
V_{CCHYS}	V_{CC} UVLO Hysteresis				0.1		
EN THRESHOLDS							
EN_{ST}	EN Startup Threshold	EN Increasing	Q1		1.75	**2.40**	V
			Q0			**2.75**	
		EN Decreasing		**0.80**	1.63		
EN_{STHYS}	EN Startup Hysteresis				0.1		
R_{EN}	EN Pulldown Resistance	EN = 1V	Q1	**0.45**	0.82	**1.30**	MΩ
			Q0			**1.80**	
CSH THRESHOLDS							
	CSH High Fault	CSH Increasing			1.6		V
	CSH Low Condition on LRDY Pin (LM3423)	CSH increasing			1.0		

Symbol	Parameter	Conditions	Temp Range	Min (Note 5)	Typ (Note 6)	Max (Note 5)	Units
OV THRESHOLDS							
OVP_{CB}	OVP OVLO Threshold	OVP Increasing		**1.185**	1.240	**1.285**	V
OVP_{HYS}	OVP Hysteresis Source Current	OVP Active (high)	Q1	**20**	23	**25**	µA
			Q0			**26**	
DPOL THRESHOLDS							
$DPOL_{THRESH}$	DPOL Logic Threshold	DPOL Increasing		**2.0**	2.3	**2.6**	V
R_{DPOL}	DPOL Pullup Resistance				500	**1200**	kΩ
FAULT TIMER							
V_{FLTTH}	Fault Threshold		Q1	**1.185**	1.240	**1.285**	V
			Q0			**1.290**	
I_{FLT}	Fault Pin Source Current		Q1	**10.0**	11.5	**13.0**	µA
			Q0			**13.5**	
ERROR AMPLIFIER							
V_{REF}	CSH Reference Voltage	With Respect to AGND		**1.210**	1.235	**1.260**	V
	Error Amplifier Input Bias Current			-0.6	0	0.6	µA
	COMP Sink / Source Current		Q1	**22**	30	**35**	
			Q0			**36**	
	Transconductance				100		µA/V
	Linear Input Range	(Note 7)			±125		mV
	Transconductance Bandwidth	-6dB Unloaded Response (Note 7)		0.5	1.0		MHz
OFF TIMER							
	Minimum Off-time	RCT = 1V through 1 kΩ	Q1		35	**75**	ns
			Q0			**90**	
R_{RCT}	RCT Reset Pull-down Resistance		Q1		36	**120**	Ω
			Q0			**125**	
V_{RCT}	V_{IN}/25 Reference Voltage	V_{IN} = 14V	Q1	**540**	565	**585**	mV
			Q0			**590**	
f	Continuous Conduction Switching Frequency	2.2 nF > C_T > 470 pF			$25/(C_T R_T)$		Hz
PWM COMPARATOR							
	COMP to PWM Offset			**700**	800	**900**	mV
CURRENT LIMIT (IS)							
I_{LIM}	Current Limit Threshold			**215**	245	**275**	mV
	I_{LIM} Delay to Output		Q1		35	**75**	ns
			Q0			**90**	
	Leading Edge Blanking Time			**115**	210	**325**	
HIGH SIDE TRANSCONDUCTANCE AMPLIFIER							
	Input Bias Current				11.5		µA
	Transconductance			**20**	119		mA/V
	Input Offset Current			**-1.5**	0	**1.5**	µA
	Input Offset Voltage			**-7**	0	**7**	mV
	Transconductance Bandwidth	I_{CSH} = 100 µA (Note 7)		250	500		kHz
GATE DRIVER (GATE)							
$R_{SRC(GATE)}$	GATE Sourcing Resistance	GATE = High			2.0	**6.0**	Ω
$R_{SNK(GATE)}$	GATE Sinking Resistance	GATE = Low			1.3	**4.5**	

Symbol	Parameter	Conditions	Temp Range	Min (Note 5)	Typ (Note 6)	Max (Note 5)	Units
DIM DRIVER (DIM, DDRV)							
$nDIM_{VTH}$	nDIM / UVLO Threshold			**1.185**	1.240	**1.285**	V
$nDIM_{HYS}$	nDIM Hysteresis Current		Q1	**20**	23	**25**	µA
			Q0			**26**	
$R_{SRC(DDRV)}$	DDRV Sourcing Resistance	DDRV = High			13.5	**30.0**	Ω
$R_{SNK(DDRV)}$	DDRV Sinking Resistance	DDRV = Low			3.5	**10.0**	
PULL-DOWN N-CHANNEL MosFETS							
R_{RPD}	RPD Pull-down Resistance		Q1		145	**300**	Ω
			Q0			**350**	
R_{FLT}	FLT Pull-down Resistance		Q1		145	**300**	
			Q0			**350**	
R_{LRDY}	LRDY Pull-down Resistance		Q1		135	**300**	
			Q0			**350**	
THERMAL SHUTDOWN							
T_{SD}	Thermal Shutdown Threshold	(Note 7)	Q1		165		°C
			Q0		210		
T_{HYS}	Thermal Shutdown Hysteresis	(Note 7)			25		
THERMAL RESISTANCE							
θ_{JA}	Junction to Ambient (Note 2)	16L TSSOP EP			37.4		°C/W
		20L TSSOP EP			34.0		
θ_{JC}	Junction to Exposed Pad (EP)	16L TSSOP EP			2.3		°C/W
		20L TSSOP EP			2.3		

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur, including inoperability and degradation of device reliability and/or performance. Functional operation of the device and/or non-degradation at the Absolute Maximum Ratings or other conditions beyond those indicated in the Operating Ratings is not implied. The recommended Operating Ratings indicate conditions at which the device is functional and the device should not be operated beyond such conditions. All voltages are with respect to the potential at the AGND pin, unless otherwise specified.

Note 2: Junction-to-ambient thermal resistance is highly board-layout dependent. The numbers listed in the table are given for an reference layout wherein the 16L TSSOP package has its EP pad populated with 9 vias and the 20L TSSOP has its EP pad populated with 12 vias. In applications where high maximum power dissipation exists, namely driving a large MosFET at high switching frequency from a high input voltage, special care must be paid to thermal dissipation issues during board design. In high-power dissipation applications, the maximum ambient temperature may have to be derated. Maximum ambient temperature ($T_{A\text{-}MAX}$) is dependent on the maximum operating junction temperature ($T_{J\text{-}MAX\text{-}OP}$ = 125°C for Q1, or 150°C for Q0), the maximum power dissipation of the device in the application ($P_{D\text{-}MAX}$), and the junction-to ambient thermal resistance of the package in the application (θ_{JA}), as given by the following equation: $T_{A\text{-}MAX} = T_{J\text{-}MAX\text{-}OP} - (\theta_{JA} \times P_{D\text{-}MAX})$. In most applications there is little need for the full power dissipation capability of this advanced package. Under these circumstances, no vias would be required and the thermal resistances would be 104 °C/W for the 16L TSSOP and 86.7 °C/W for the 20L TSSOP. It is possible to conservatively interpolate between the full via count thermal resistance and the no via count thermal resistance with a straight line to get a thermal resistance for any number of vias in between these two limits.

Note 3: Refer to National's packaging website for more detailed information and mounting techniques. http://www.national.com/analog/packaging/

Note 4: The human body model is a 100 pF capacitor discharged through a 1.5 kΩ resistor into each pin. The applicable standard is JESD22-A114C.

Note 5: All limits guaranteed at room temperature (standard typeface) and at temperature extremes (bold typeface). All room temperature limits are 100% production tested. All limits at temperature extremes are guaranteed via correlation using standard Statistical Quality Control (SQC) methods. All limits are used to calculate Average Outgoing Quality Level (AOQL).

Note 6: Typical numbers are at 25°C and represent the most likely norm.

Note 7: These electrical parameters are guaranteed by design, and are not verified by test.

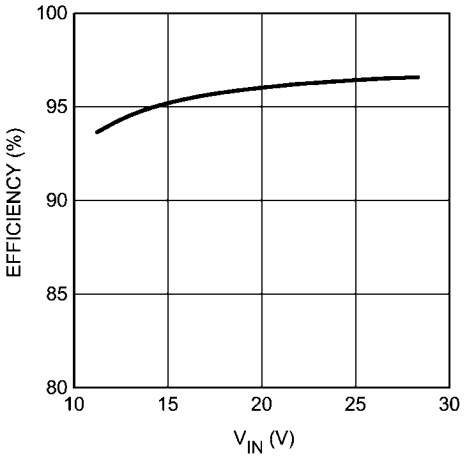
Note 8: The measurements were made using the standard buck-boost evaluation board from AN-2010.

Note 9: The measurements were made using the standard boost evaluation board from AN-2011.

Typical Performance Characteristics

T_A=+25°C and V_{IN} = 14V unless otherwise specified

Boost Efficiency vs. Input Voltage
V_O = 32V (9 LEDs) (Note 9)



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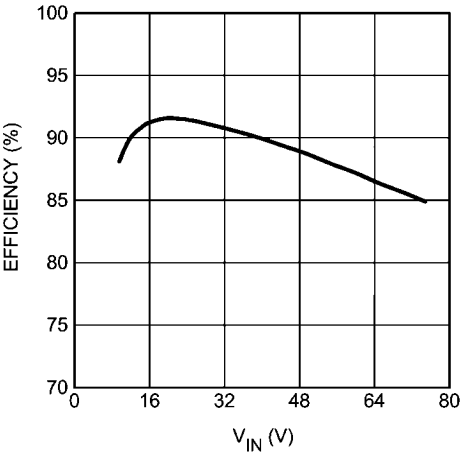
Buck-Boost Efficiency vs. Input Voltage
V_O = 21V (6 LEDs) (Note 8)



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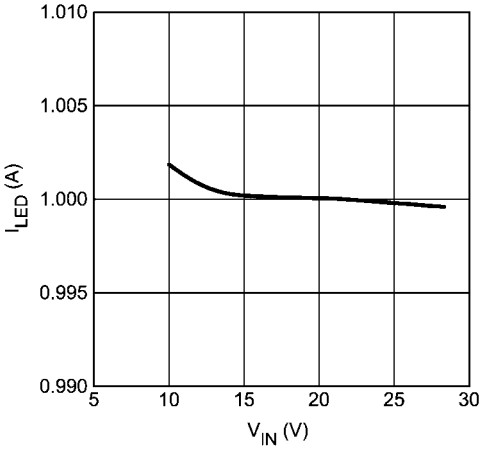
Boost LED Current vs. Input Voltage
V_O = 32V (9 LEDs) (Note 9)



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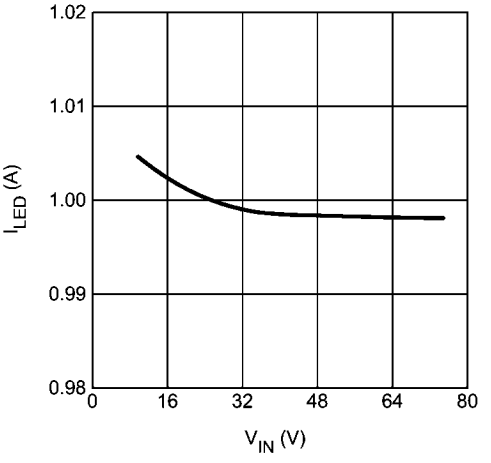
Buck-Boost LED Current vs. Input Voltage
V_O = 21V (6 LEDs) (Note 8)



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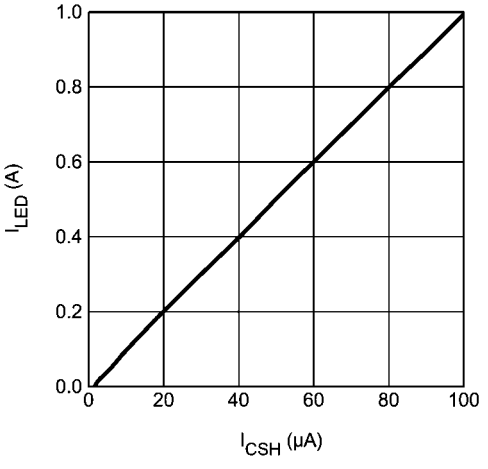
Analog Dimming
V_O = 21V (6 LEDs); V_{IN} = 24V (Note 8)



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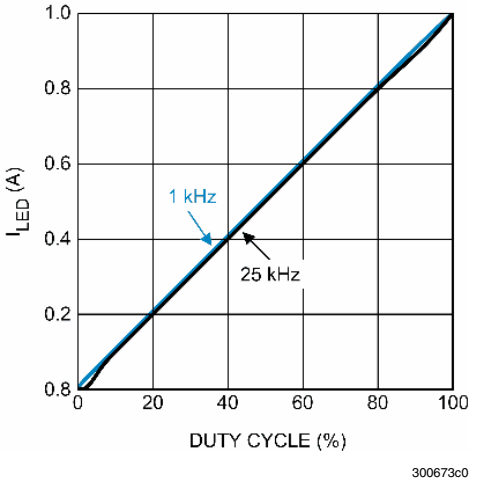
PWM Dimming
V_O = 32V (9 LEDs); V_{IN} = 24V (Note 9)



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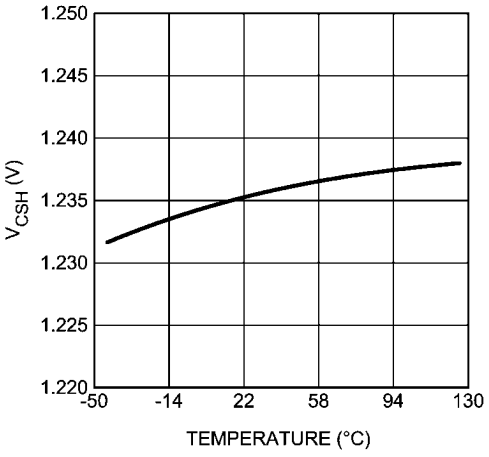
V_{CSH} vs. Junction Temperature



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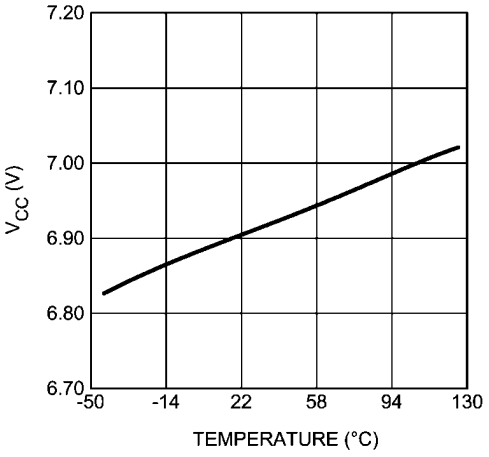
V_{CC} vs. Junction Temperature



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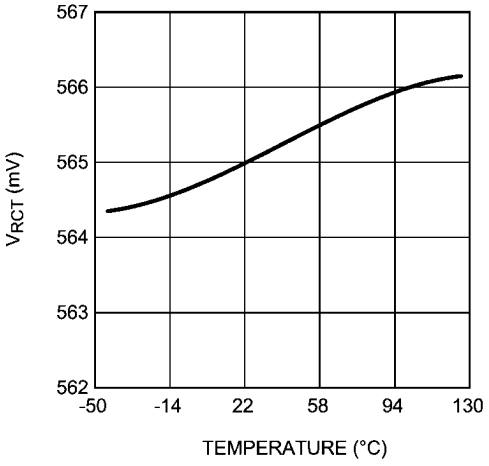
V_{RCT} vs. Junction Temperature



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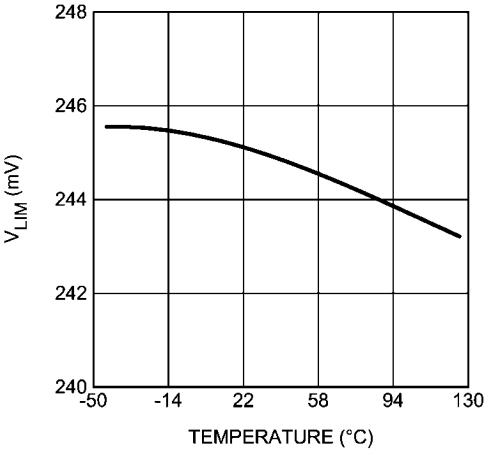
V_{LIM} vs. Junction Temperature



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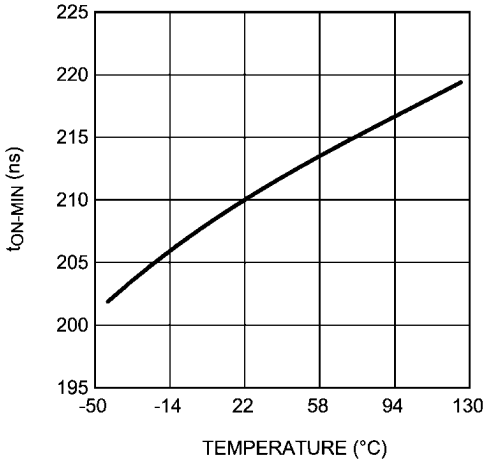
t_{ON-MIN} vs. Junction Temperature



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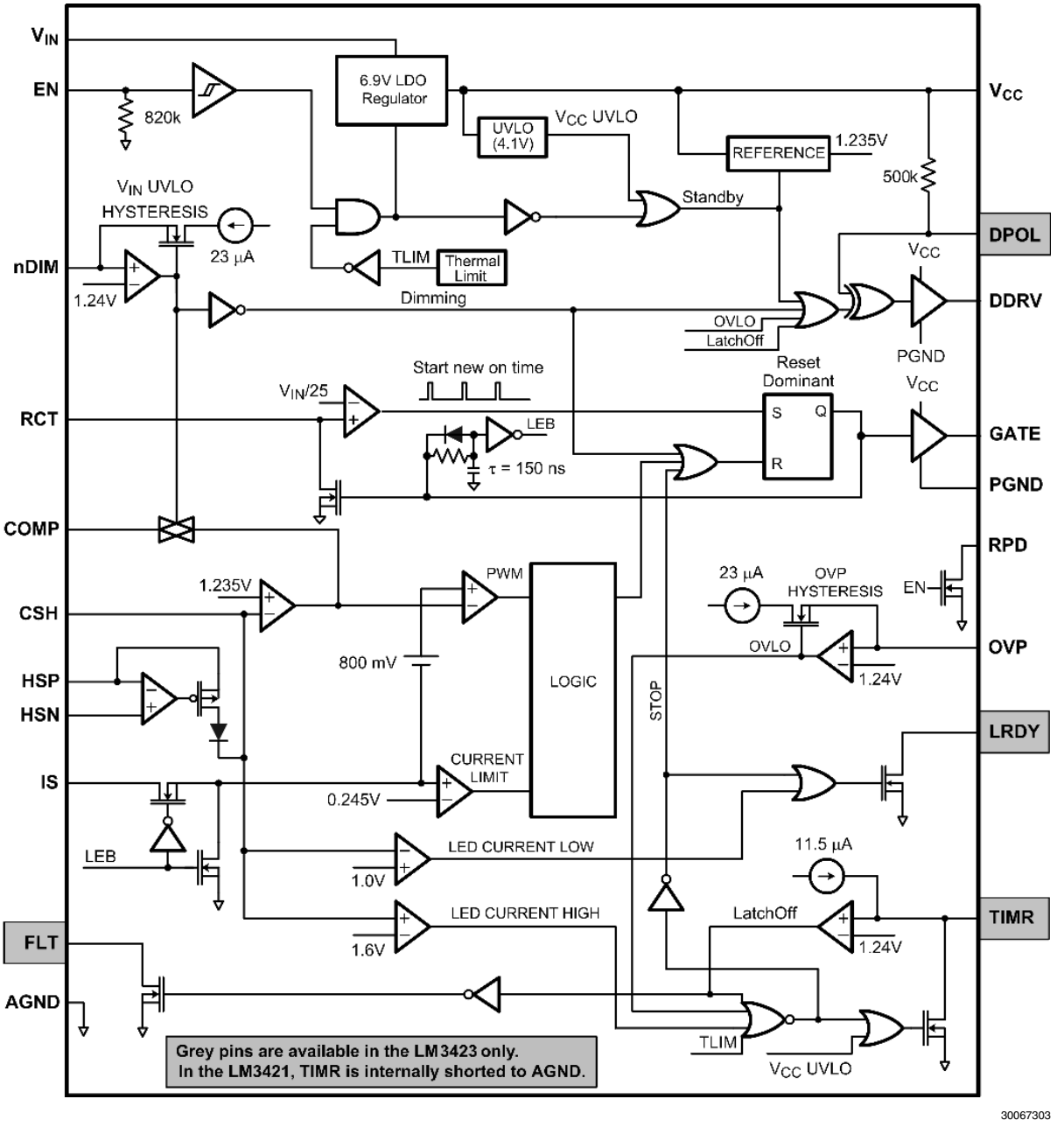
Block Diagram



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Theory of Operation

The LM3421/23 are N-channel MosFET (NFET) controllers for buck, boost and buck-boost current regulators which are ideal for driving LED loads. The controller has wide input voltage range allowing for regulation of a variety of LED loads. The high-side differential current sense, with low adjustable threshold voltage, provides an excellent method for regulating output current while maintaining high system efficiency. The LM3421/23 uses a Predictive Off-time (PRO) control architecture that allows the regulator to be operated using minimal external control loop compensation, while providing an inherent cycle-by-cycle current limit. The adjustable current sense threshold provides the capability to amplitude (analog) dim the LED current and the output enable/disable function with external dimming FET driver allows for fast PWM dimming of the LED load. When designing, the maximum attainable LED current is not internally limited because the LM3421/23 is a controller. Instead it is a function of the system operating point, component choices, and switching frequency allowing the LM3421/23 to easily provide constant currents up to 5A. This controller contains all the features necessary to implement a high efficiency versatile LED driver.

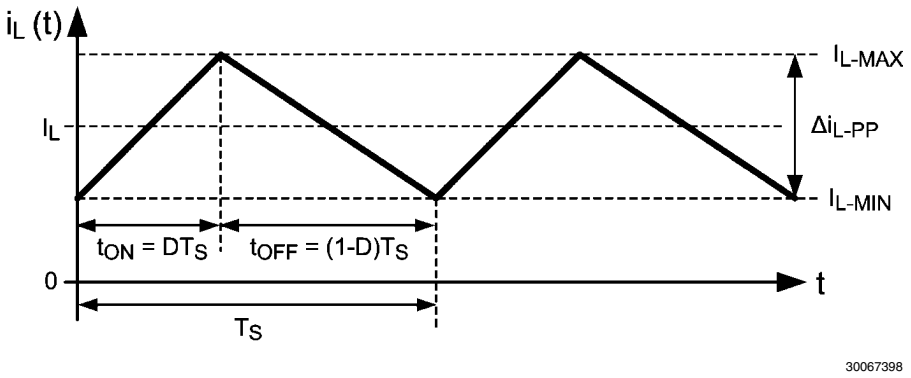


FIGURE 1. Ideal CCM Regulator Inductor Current $i_L(t)$

CURRENT REGULATORS

Current regulators can be designed to accomplish three basic functions: buck, boost, and buck-boost. All three topologies in their most basic form contain a main switching MosFET, a recirculating diode, an inductor and capacitors. The LM3421/23 is designed to drive a ground referenced NFET which is perfect for a standard boost regulator. Buck and buck-boost regulators, on the other hand, usually have a high-side switch. When driving an LED load, a ground referenced load is often not necessary, therefore a ground referenced switch can be used to drive a floating load instead. The LM3421/23 can then be used to drive all three basic topologies as shown in the Basic Topology Schematics section. Other topologies such as the SEPIC and flyback converter (both derivatives of the buck-boost) can be implemented as well.

Looking at the buck-boost design, the basic operation of a current regulator can be analyzed. During the time that the NFET (Q1) is turned on (t_{ON}), the input voltage source stores energy in the inductor (L1) while the output capacitor (C_O) provides energy to the LED load. When Q1 is turned off (t_{OFF}), the re-circulating diode (D1) becomes forward biased and L1 provides energy to both C_O and the LED load. Figure 1 shows the inductor current ($i_L(t)$) waveform for a regulator operating in CCM.

The average output LED current (I_{LED}) is proportional to the average inductor current (I_L) , therefore if I_L is tightly controlled, I_{LED} will be well regulated. As the system changes input voltage or output voltage, the ideal duty cycle (D) is varied to regulate I_L and ultimately I_{LED}. For any current regulator, D is a function of the conversion ratio:

Buck

$$D = \frac{V_O}{V_{IN}}$$

Boost

$$D = \frac{V_O - V_{IN}}{V_O}$$

Buck-boost

$$D = \frac{V_O}{V_O + V_{IN}}$$

PREDICTIVE OFF-TIME (PRO) CONTROL

PRO control is used by the LM3421/23 to control I_{LED}. It is a combination of average peak current control and a one-shot off-timer that varies with input voltage. The LM3421/23 uses peak current control to regulate the average LED current through an array of HBLEDs. This method of control uses a series resistor in the LED path to sense LED current and can use either a series resistor in the MosFET path or the MosFET $R_{DS\text{-}ON}$ for both cycle-by-cycle current limit and input voltage feed forward. D is indirectly controlled by changes in both t_{OFF} and t_{ON}, which vary depending on the operating point.

Even though the off-time control is quasi-hysteretic, the input voltage proportionality in the off-timer creates an essentially constant switching frequency over the entire operating range for boost and buck-boost topologies. The buck topology can be designed to give constant ripple over either input voltage or output voltage, however switching frequency is only constant at a specific operating point .

This type of control minimizes the control loop compensation necessary in many switching regulators, simplifying the design process. The averaging mechanism in the peak detection control loop provides extremely accurate LED current regulation over the entire operating range.

PRO control was designed to mitigate “current mode instability” (also called “sub-harmonic oscillation”) found in standard peak current mode control when operating near or above 50% duty cycles. When using standard peak current mode control with a fixed switching frequency, this condition is present, regardless of the topology. However, using a constant off-time approach, current mode instability cannot occur, enabling easier design and control.

Predictive off-time advantages:

- There is no current mode instability at any duty cycle.
- Higher duty cycles / voltage transformation ratios are possible, especially in the boost regulator.

The only disadvantage is that synchronization to an external reference frequency is generally not available.

SWITCHING FREQUENCY

An external resistor (R_T) connected between the RCT pin and the switch node (where D1, Q1, and L1 connect), in combination with a capacitor (C_T) between the RCT and AGND pins, sets the off-time (t_{OFF}) as shown in Figure 2. For boost and buck-boost topologies, the V_{IN} proportionality ensures a virtually constant switching frequency (f_{SW}).

For a buck topology, R_T and C_T are also used to set t_{OFF}, however the V_{IN} proportionality will not ensure a constant switching frequency. Instead, constant ripple operation can be achieved. Changing the connection of R_T in Figure 2 from V_{SW} to V_{IN} will provide a constant ripple over varying V_{IN}. Adding a PNP transistor as shown in Figure 3 will provide constant ripple over varying V_O.

The switching frequency is defined:

Buck (Constant Ripple vs. V_{IN})

$$f_{SW} = \frac{25 \times (V_{IN} - V_O)}{R_T \times C_T \times V_{IN}}$$

Buck (Constant Ripple vs. V_O)

$$f_{SW} = \frac{25 \times (V_{IN} \times V_O - V_O^2)}{R_T \times C_T \times V_{IN}^2}$$

Boost and Buck-boost

$$f_{SW} = \frac{25}{R_T \times C_T}$$

For all topologies, the C_T capacitor is recommended to be 1 nF and should be located very close to the LM3421/23.

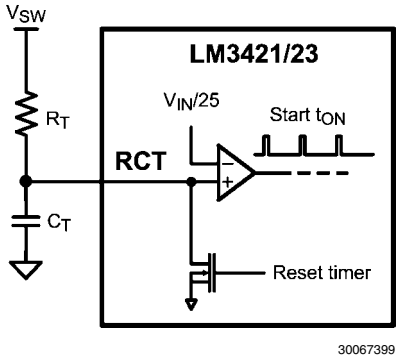


FIGURE 2. Off-timer Circuitry for Boost and Buck-boost Regulators

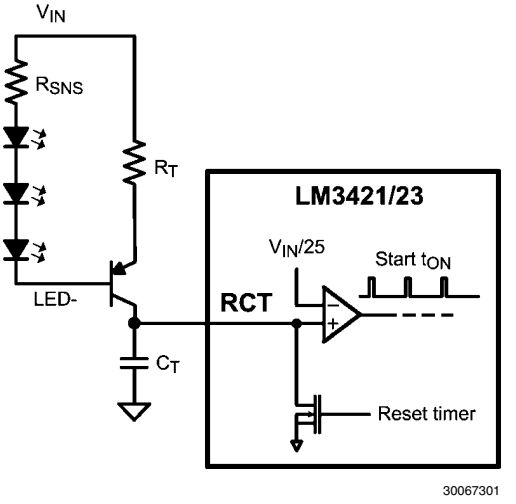


FIGURE 3. Off-timer Circuitry for Buck Regulators

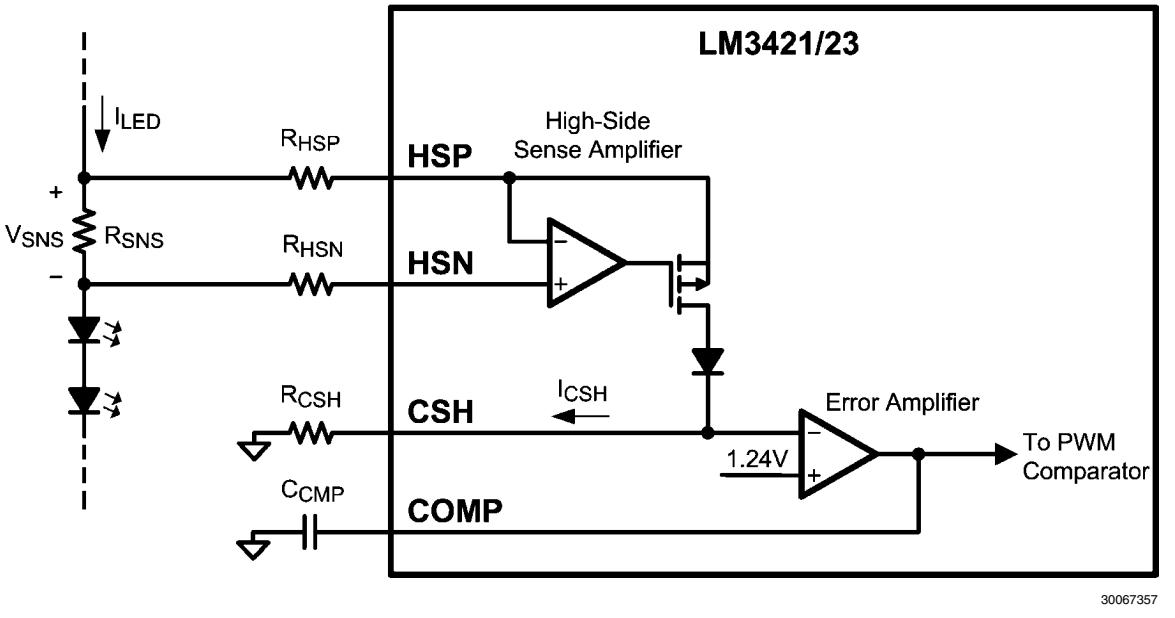


FIGURE 4. LED Current Sense Circuitry

AVERAGE LED CURRENT

The LM3421/23 uses an external current sense resistor (R_{SNS}) placed in series with the LED load to convert the LED current (I_{LED}) into a voltage (V_{SNS}) as shown in Figure 4. The HSP and HSN pins are the inputs to the high-side sense amplifier which are forced to be equal potential (V_{HSP}=V_{HSN}) through negative feedback. Because of this, the V_{SNS} voltage is forced across R_{HSP} to generate the signal current (I_{CSH}) which flows out of the CSH pin and through the R_{CSH} resistor. The error amplifier will regulate the CSH pin to 1.24V, therefore I_{CSH} can be calculated:

$$I_{CSH} = \frac{V_{SNS}}{R_{HSP}}$$

This means V_{SNS} will be regulated as follows:

$$V_{SNS} = 1.24V \times \frac{R_{HSP}}{R_{CSH}}$$

I_{LED} can then be calculated:

$$I_{LED} = \frac{V_{SNS}}{R_{SNS}} = \frac{1.24V}{R_{SNS}} \times \frac{R_{HSP}}{R_{CSH}}$$

The selection of the three resistors (R_{SNS}, R_{CSH}, and R_{HSP}) is not arbitrary. For matching and noise performance, the suggested signal current I_{CSH} is approximately 100 µA. This current does not flow in the LEDs and will not affect either the off-state LED current or the regulated LED current. I_{CSH} can be above or below this value, but the high-side amplifier offset characteristics may be affected slightly. In addition, to minimize the effect of the high-side amplifier voltage offset on LED current accuracy, the minimum V_{SNS} is suggested to be 50 mV. Finally, a resistor (R_{HSN} = R_{HSP}) should be placed in series with the HSN pin to cancel out the effects of the input bias current (~10 µA) of both inputs of the high-side sense amplifier.

The sense resistor (R_{SNS}) can be placed anywhere in the series string of LEDs as long as the voltage at the HSN and HSP pins (V_{HSP} and V_{HSN}) satisfies the following conditions.

$$V_{HSP} < 76V$$
$$V_{HSN} > 3.5V$$

Typically, for a buck-boost configuration, R_{SNS} is placed at the bottom of the string (LED-) which allows for greater flexibility of input and output voltage. However, if there is substantial input voltage ripple allowed, it can help to place R_{SNS} at the top of the string (LED+) which limits the output voltage of the string to:

$$V_O = 76V - V_{IN}$$

Note that he CSH pin can also be used as a low-side current sense input regulated to 1.24V. The high-side sense amplifier is disabled if HSP and HSN are tied to AGND (or V_{HSN} > V_{HSP}) .

ANALOG DIMMING

The CSH pin can be used to analog dim the LED current by adjusting the current sense voltage (V_{SNS}). There are several different methods to adjust V_{SNS} using the CSH pin:

1. External variable resistance : Adjust a potentiometer placed in series with R_{CSH} to vary V_{SNS}.
2. External variable current source: Source current (0 µA to I_{CSH}) into the CSH pin to adjust V_{SNS}.

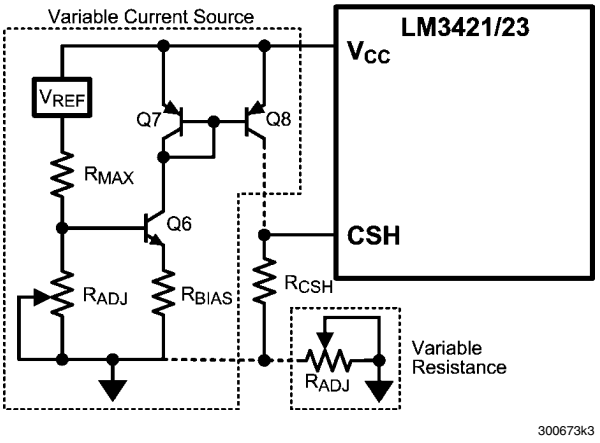


FIGURE 5. Analog Dimming Circuitry

In general, analog dimming applications require a lower switching frequency to minimize the effect of the leading edge blanking circuit. As the LED current is reduced, the output voltage and the duty cycle decreases. Eventually, the minimum on-time is reached. The lower the switching frequency, the wider the linear dimming range. Figure 5 shows how both CSH methods are physically implemented.

Method 1 uses an external potentiometer in the CSH path which is a simple addition to the existing circuitry. However, the LEDs cannot dim completely because there is always some resistance causing signal current to flow. This method is also susceptible to noise coupling at the CSH pin since the potentiometer increases the size of the signal current loop.

Method 2 provides a complete dimming range and better noise performance, though it is more complex. It consists of a PNP current mirror and a bias network consisting of an NPN, 2 resistors and a potentiometer (R_{ADJ}), where R_{ADJ} controls the amount of current sourced into the CSH pin. A higher resistance value will source more current into the CSH pin causing less regulated signal current through R_{HSP}, effectively dimming the LEDs. V_{REF} should be a precise external voltage reference, while Q7 and Q8 should be a dual pair PNP for best matching and performance. The additional current (I_{ADD}) sourced into the CSH pin can be calculated:

$$I_{ADD} = \frac{\left(\frac{R_{ADJ} \times V_{REF}}{R_{ADJ} + R_{MAX}}\right) - V_{BE\text{-}Q6}}{R_{BIAS}}$$

The corresponding I_{LED} for a specific I_{ADD} is:

$$I_{LED} = (I_{CSH} - I_{ADD}) \times \left(\frac{R_{HSP}}{R_{SNS}}\right)$$

CURRENT SENSE/CURRENT LIMIT

The LM3421/23 achieves peak current mode control using a comparator that monitors the main MosFET (Q1) transistor current, comparing it with the COMP pin voltage as shown in Figure 6. Further, it incorporates a cycle-by-cycle over-current protection function. Current limit is accomplished by a redundant internal current sense comparator. If the voltage at the current sense comparator input (IS) exceeds 245 mV (typical), the on cycle is immediately terminated. The IS input pin has an internal N-channel MosFET which pulls it down at the conclusion of every cycle. The discharge device remains on an additional 210 ns (typical) after the beginning of a new cycle to blank the leading edge spike on the current sense signal. The leading edge blanking (LEB) determines the minimum achievable on-time (t_{ON-MIN}).

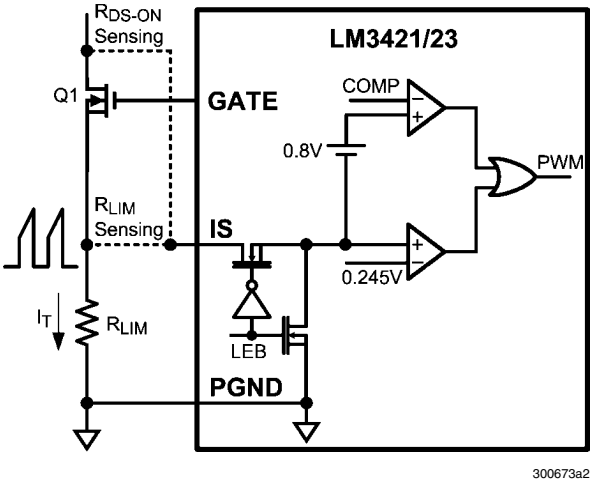


FIGURE 6. Current Sense / Current Limit Circuitry

There are two possible methods to sense the transistor current. The R_{DS-ON} of the main power MosFET can be used as the current sense resistance because the IS pin was designed to withstand the high voltages present on the drain when the MosFET is in the off state. Alternatively, a sense resistor located in the source of the MosFET may be used for current sensing, however a low inductance (ESL) type is suggested. The cycle-by-cycle current limit (I_{LIM}) can be calculated using either method as the limiting resistance (R_{LIM}):

$$I_{LIM} = \frac{245\ mV}{R_{LIM}}$$

OVER-CURRENT PROTECTION

The LM3421/23 devices have a secondary method of over-current protection. Switching action is disabled whenever the current in the LEDs is more than 30% above the regulation set point. The dimming MosFET switch driver (DDRV) is not disabled however as this would immediately remove the fault condition and cause oscillatory behavior.

ZERO CURRENT SHUTDOWN

The LM3421/23 devices implement "zero current" shutdown via the EN and RPD pins. When pulled low, the EN pin places the devices into near-zero current state, where only the leakage currents will be observed at the pins (typical 0.1 µA). The applications circuits, frequently have resistor dividers to set UVLO, OVLO, or other similar functions. The RPD pin is an open drain N-channel MosFET that is enabled only when the device is enabled. Tying the bottom of all resistor dividers to the RPD pin as shown in Figure 7 allows them to float during shutdown, thus removing their current paths and providing true application-wide zero current shutdown.

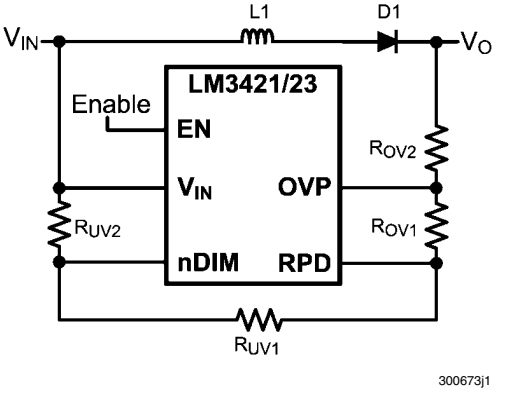


FIGURE 7. Zero Current Shutdown Circuit

CONTROL LOOP COMPENSATION

The LM3421/23 control loop is modeled like any current mode controller. Using a first order approximation, the uncompensated loop can be modeled as a single pole created by the output capacitor and, in the boost and buck-boost topologies, a right half plane zero created by the inductor, where both have a dependence on the LED string dynamic resistance. There is also a high frequency pole in the model, however it is near the switching frequency and plays no part in the compensation design process therefore it will be neglected. Since ceramic capacitance is recommended for use with LED drivers due to long lifetimes and high ripple current rating, the ESR of the output capacitor can also be neglected in the loop analysis. Finally, there is a DC gain of the uncompensated loop which is dependent on internal controller gains and the external sensing network.

A buck-boost regulator will be used as an example case. See the Design Guide section for compensation of all topologies.

The uncompensated loop gain for a buck-boost regulator is given by the following equation:

$$T_U = T_{U0} \times \frac{\left(1 - \frac{s}{\omega_{Z1}}\right)}{\left(1 + \frac{s}{\omega_{P1}}\right)}$$

Where the uncompensated DC loop gain of the system is described as:

$$T_{U0} = \frac{D' \times 500V \times R_{CSH} \times R_{SNS}}{(1+D) \times R_{HSP} \times R_{LIM}} = \frac{D' \times 620V}{(1+D) \times I_{LED} \times R_{LIM}}$$

And the output pole (ω_{P1}) is approximated:

$$\omega_{P1} = \frac{1+D}{r_D \times C_O}$$

And the right half plane zero (ω_{Z1}) is:

$$\omega_{Z1} = \frac{r_D \times D'^2}{D \times L1}$$

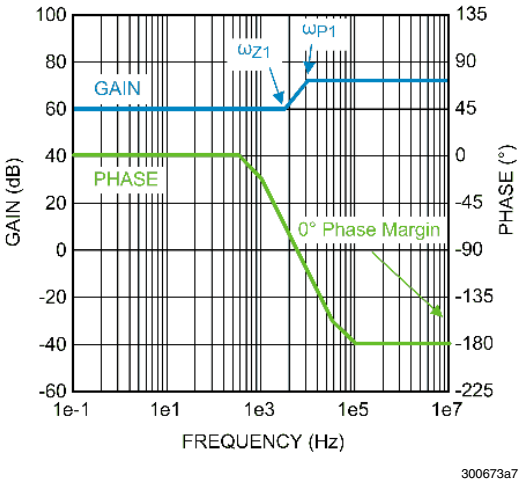


FIGURE 8. Uncompensated Loop Gain Frequency Response

Figure 8 shows the uncompensated loop gain in a worst-case scenario when the RHP zero is below the output pole. This occurs at high duty cycles when the regulator is trying to boost the output voltage significantly. The RHP zero adds 20dB/decade of gain while loosing 45°/decade of phase which places the crossover frequency (when the gain is zero dB) extremely high because the gain only starts falling again due to the high frequency pole (not modeled or shown in figure). The phase will be below -180° at the crossover frequency which means there is no phase margin (180° + phase at crossover frequency) causing system instability. Even if the output pole is below the RHP zero, the phase will still reach -180° before the crossover frequency in most cases yielding instability.

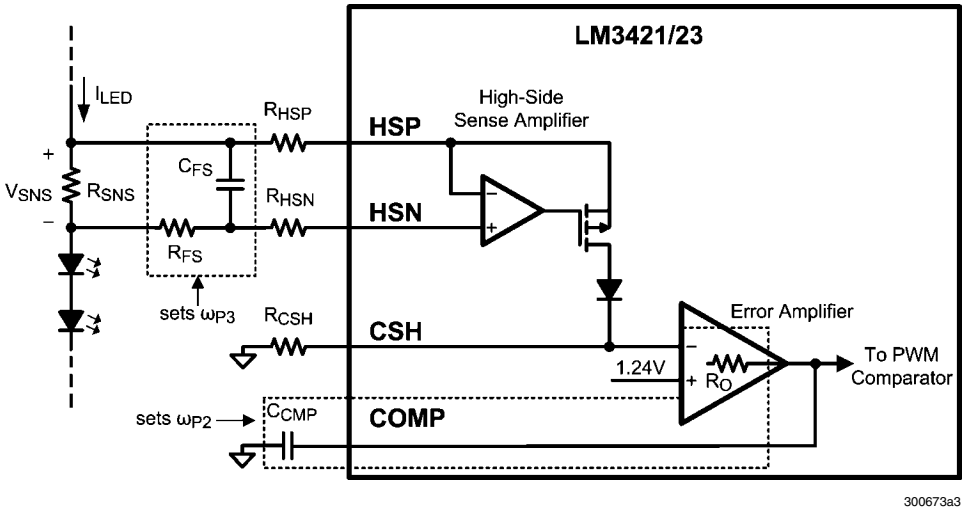


FIGURE 9. Compensation Circuitry

To mitigate this problem, a compensator should be designed to give adequate phase margin (above 45°) at the crossover frequency. A simple compensator using a single capacitor at the COMP pin (C_{CMP}) will add a dominant pole to the system, which will ensure adequate phase margin if placed low enough. At high duty cycles (as shown in Figure 8), the RHP zero places extreme limits on the achievable bandwidth with this type of compensation. However, because an LED driver is essentially free of output transients (except catastrophic failures open or short), the dominant pole approach, even with reduced bandwidth, is usually the best approach. The dominant compensation pole (ω_{P2}) is determined by C_{CMP} and the output resistance (R_O) of the error amplifier (typically 5 MΩ):

$$\omega_{P2} = \frac{1}{5e^6\Omega \times C_{CMP}}$$

It may also be necessary to add one final pole at least one decade above the crossover frequency to attenuate switching noise and, in some cases, provide better gain margin. This pole can be placed across R_{SNS} to filter the ESL of the sense resistor at the same time. Figure 9 shows how the compensation is physically implemented in the system.

The high frequency pole (ω_{P3}) can be calculated:

$$\omega_{P3} = \frac{1}{R_{FS} \times C_{FS}}$$

The total system transfer function becomes:

$$T = T_{U0} \times \frac{\left(1 - \frac{s}{\omega_{Z1}}\right)}{\left(1 + \frac{s}{\omega_{P1}}\right) \times \left(1 + \frac{s}{\omega_{P2}}\right) \times \left(1 + \frac{s}{\omega_{P3}}\right)}$$

The resulting compensated loop gain frequency response shown in Figure 10 indicates that the system has adequate phase margin (above 45°) if the dominant compensation pole is placed low enough, ensuring stability:

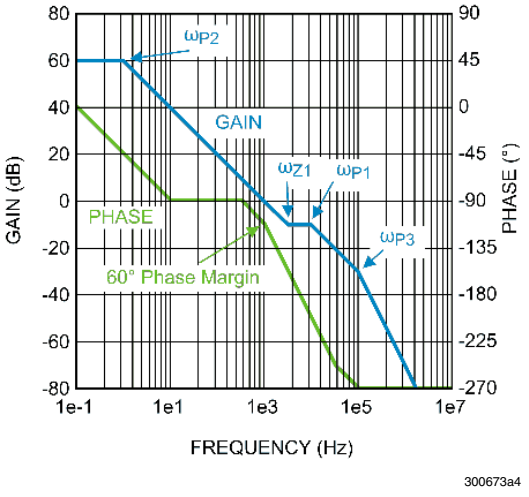


FIGURE 10. Compensated Loop Gain Frequency Response

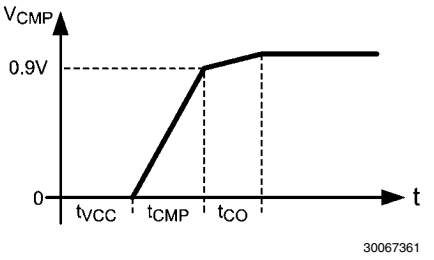


FIGURE 11. Start-Up Waveforms

START-UP REGULATOR

The LM3421/23 includes a high voltage, low dropout bias regulator. When power is applied, the regulator is enabled and sources current into an external capacitor (C_{BYP}) connected to the V_{CC} pin. The recommended bypass capacitance for the V_{CC} regulator is 2.2 µF to 3.3 µF. The output of the V_{CC} regulator is monitored by an internal UVLO circuit that protects the device from attempting to operate with insufficient supply voltage and the supply is also internally current limited. Figure 11 shows the typical start-up waveforms for the LM3421/23.

First, C_{BYP} is charged to be above V_{CC} UVLO threshold (~4.2V). The C_{VCC} charging time (t_{VCC}) can be estimated as:

$$t_{VCC} = \frac{4.2V}{25\ mA} \times C_{BYP} = 168\Omega \times C_{BYP}$$

C_{CMP} is then charged to 0.9V over the charging time (t_{CMP}) which can be estimated as:

$$t_{CMP} = \frac{0.9V}{25\ \mu A} \times C_{CMP} = 36\ k\Omega \times C_{CMP}$$

Once C_{CMP} = 0.9V, the part starts switching to charge C_O until the LED current is in regulation. The C_O charging time (t_{CO}) can be roughly estimated as:

$$t_{CO} = C_O \times \frac{V_O}{I_{LED}}$$

The system start-up time (t_{SU}) is defined as:

$$t_{SU} = t_{VCC} + t_{CMP} + t_{CO}$$

In some configurations, the start-up waveform will overshoot the steady state COMP pin voltage. In this case, the LED current and output voltage will overshoot also, which can trip the over-voltage or protection, causing a race condition. The easiest way to prevent this is to use a larger compensation capacitor (C_{CMP}), thereby slowing down the control loop.

OVER-VOLTAGE LOCKOUT (OVLO)

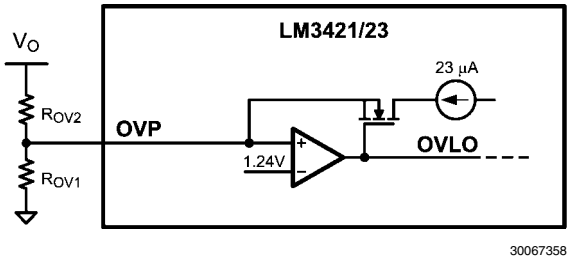


FIGURE 12. Over-Voltage Protection Circuitry

The LM3421/23 can be configured to detect an output (or input) over-voltage condition via the OVP pin. The pin features a precision 1.24V threshold with 23 µA (typical) of hysteresis current as shown in Figure 12. When the OVLO threshold is exceeded, the GATE pin is immediately pulled low and a 23 µA current source provides hysteresis to the lower threshold of the OVLO hysteretic band.

If the LEDs are referenced to a potential other than ground (floating), as in the buck-boost and buck configuration, the output voltage (V_O) should be sensed and translated to ground by using a single PNP as shown in Figure 13.

The over-voltage turn-off threshold ($V_{TURN-OFF}$) is defined:

Ground Referenced

$$V_{TURN-OFF} = 1.24V \times \left(\frac{R_{OV1} + R_{OV2}}{R_{OV1}} \right)$$

Floating

$$V_{TURN-OFF} = 1.24V \times \left(\frac{0.5 \times R_{OV1} + R_{OV2}}{R_{OV1}} \right)$$

In the ground referenced configuration, the voltage across R_{OV2} is V_O - 1.24V whereas in the floating configuration it is V_O - 620 mV where 620 mV approximates V_{BE} of the PNP.

The over-voltage hysteresis (V_{HYSO}) is defined:

$$V_{HYSO} = 23\,\mu A \times R_{OV2}$$

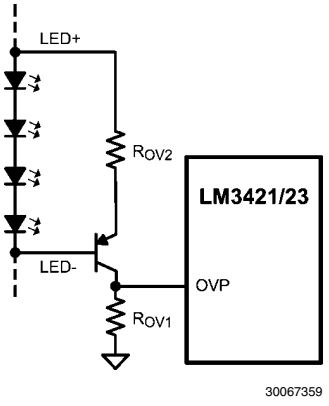


FIGURE 13. Floating Output OVP Circuitry

The OVLO feature can cause some interesting results if the OVLO trip-point is set too cose to V_O. At turn-on, the converter has a modest amount of voltage overshoot before the control loop gains control of I_{LED}. If the overshoot exceeds the OVLO threshold, the controller shuts down, opening the dimming MosFET. This isolates the LED load from the converter and the output capacitance. The voltage will then discharge very slowly through the HSP and HSN pins until V_O drops below the lower threshold, where the process repeats. This looks like the LEDs are blinking at around 2 Hz. This mode can be escaped if the input voltage is reduced.

INPUT UNDER-VOLTAGE LOCKOUT (UVLO)

The nDIM pin is a dual-function input that features an accurate 1.24V threshold with programmable hysteresis as shown in Figure 14. This pin functions as both the PWM dimming input for the LEDs and as a V_{IN} UVLO. When the pin voltage rises and exceeds the 1.24V threshold, 23 µA (typical) of current is driven out of the nDIM pin into the resistor divider providing programmable hysteresis.

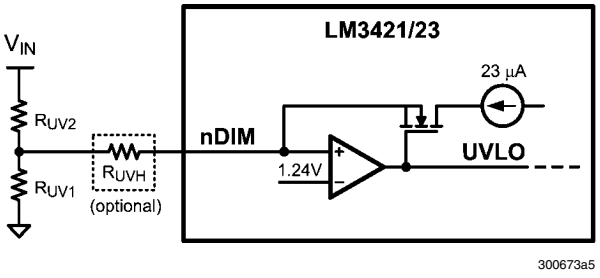


FIGURE 14. UVLO Circuit

When using the nDIM pin for UVLO and PWM dimming concurrently, the UVLO circuit can have an extra series resistor to set the hysteresis. This allows the standard resistor divider to have smaller resistor values minimizing PWM delays due to a pull-down MosFET at the nDIM pin (see PWM Dimming section). In general, at least 3V of hysteresis is preferable when PWM dimming, if operating near the UVLO threshold.

The turn-on threshold ($V_{TURN-ON}$) is defined as follows:

$$V_{TURN\ ON} = 1.24V \times \left(\frac{R_{UV1} + R_{UV2}}{R_{UV1}} \right)$$

The hysteresis (V_{HYS}) is defined as follows:

UVLO only

$$V_{HYS} = 23\,\mu A \times R_{UV2}$$

PWM dimming and UVLO

$$V_{HYS} = 23\,\mu A \times \left(R_{UV2} + \frac{R_{UVH} \times (R_{UV1} + R_{UV2})}{R_{UV1}} \right)$$

When "zero current" shutdown and UVLO are implemented together, the EN pin can be used to escape UVLO. The nDIM pin will pull-up to V_{IN} when EN is pulled low, therefore if V_{IN} is within the UVLO hysteretic window when EN is pulled high again, the controller will start-up even though $V_{TURN-ON}$ is not exceeded.

PWM DIMMING

The active low nDIM pin can be driven with a PWM signal which controls the main NFET and the dimming FET (dimFET). The brightness of the LEDs can be varied by modulating the duty cycle of this signal. LED brightness is approximately proportional to the PWM signal duty cycle, (i.e. 30% duty cycle ~ 30% LED brightness). This function can be ignored if PWM dimming is not required by using nDIM solely as a V_{IN} UVLO input as described in the Input Under-Voltage Lockout section or by tying it directly to V_{CC} or V_{IN}.

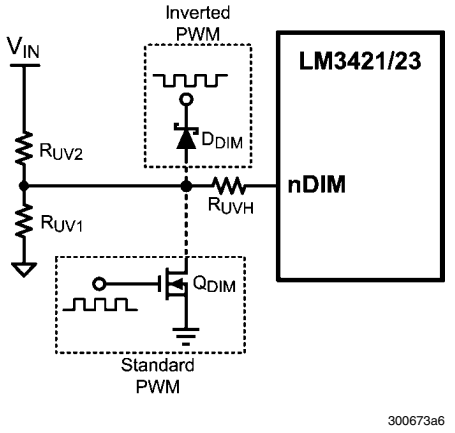


FIGURE 15. PWM Dimming Circuit

Figure 15 shows how the PWM signal is applied to nDIM:

1. Connect the dimming MosFET (Q_{DIM}) with the drain to the nDIM pin and the source to AGND. Apply an external logic-level PWM signal to the gate of Q_{DIM}.
2. Connect the anode of a Schottky diode (D_{DIM}) to the nDIM pin. Apply an inverted external logic-level PWM signal to the cathode of the same diode.

The DDRV pin is a PWM output that follows the nDIM PWM input signal. When the nDIM pin rises, the DDRV pin rises and the PWM latch reset signal is removed allowing the main MosFET Q1 to turn on at the beginning of the next clock set pulse. In boost and buck-boost topologies, the DDRV pin is used to control a N-channel MosFET placed in series with the LED load, while it would control a P-channel MosFET in parallel with the load for a buck topology.

The series dimFET will open the LED load, when nDIM is low, effectively speeding up the rise and fall times of the LED current. Without any dimFET, the rise and fall times are limited by the inductor slew rate and dimming frequencies above 1 kHz are impractical. Using the series dimFET, dimming frequencies up to 30 kHz are achievable. With a parallel dimFET (buck topology), even higher dimming frequencies are achievable.

When using the PWM functionality in a boost regulator, the PWM signal can drive a ground referenced FET. However, with buck-boost and buck topologies, level shifting circuitry is necessary to translate the PWM dim signal to the floating dimFET as shown in Figure 16 and Figure 17. If high side dimming is necessary in a boost regulator using the LM3423, level shifting can be added providing the polarity inverting DPOL pin is pulled low (see LM3423 ONLY: DPOL, FLT, TIMR, and LRDY section) as shown in Figure 18.

When using a series dimFET to PWM dim the LED current, more output capacitance is always better. A general rule of thumb is to use a minimum of 40 µF when PWM dimming. For most applications, this will provide adequate energy storage at the output when the dimFET turns off and opens the LED load. Then when the dimFET is turned back on, the capacitance helps source current into the load, improving the LED current rise time.

A minimum on-time must be maintained in order for PWM dimming to operate in the linear region of its transfer function. Because the controller is disabled during dimming, the PWM pulse must be long enough such that the energy intercepted from the input is greater than or equal to the energy being put into the LEDs. For boost and buck-boost regulators, the minimum dimming pulse length in seconds (t_{PULSE}) is:

$$t_{PULSE} = \frac{2 \times I_{LED} \times V_O \times L1}{V_{IN}^2}$$

Even maintaining a dimming pulse greater than t_{PULSE}, preserving linearity at low dimming duty cycles is difficult.

The second helpful modification is to remove the C_{FS} capacitor and R_{FS} resistor, eliminating the high frequency compensation pole. This should not affect stability, but it will speed up the response of the CSH pin, specifically at the rising edge of the LED current when PWM dimming, thus improving the achievable linearity at low dimming duty cycles.

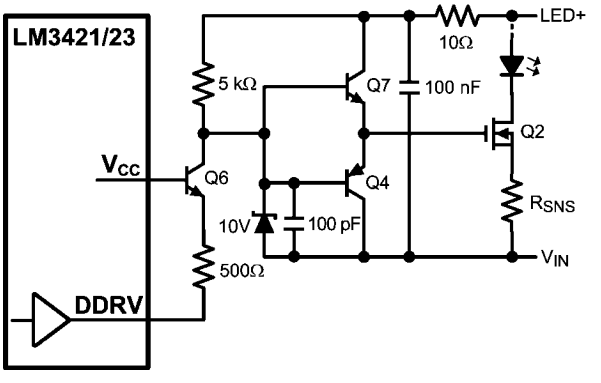


FIGURE 16. Buck-boost Level-Shifted PWM Circuit

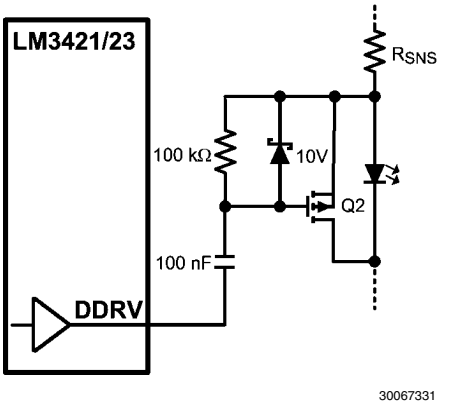


FIGURE 17. Buck Level-Shifted PWM Circuit

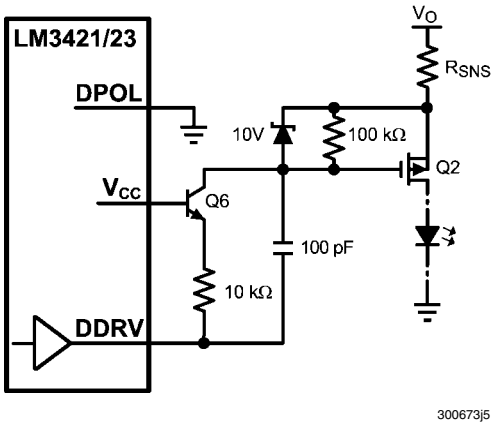


FIGURE 18. Boost Level-Shifted PWM Circuit

LM3423 ONLY: DPOL, FLT, TIMR, and LRDY

The LM3423 has four additional pins: DPOL, FLT, TIMR, and LRDY. The DPOL pin is simply used to invert the DDRV polarity . If DPOL is left open, then it is internally pulled high and the polarity is correct for driving a series N-channel dimFET. If DPOL is pulled low then the polarity is correct for using a series P-channel dimFET in high-side dimming applications. For a parallel P-channel dimFET, as used in the buck topology, leave DPOL open for proper polarity.

Among the LM3423's other additional pins are TIMR and FLT which can be used in conjunction with an input disconnect MosFET switch as shown in Figure 19 to protect the module from various fault conditions.

A fault is detected and an 11.5 µA (typical) current is sourced from the TIMR pin whenever any of the following conditions exist:

1. LED current is above regulation by more than 30%.
2. OVLO has engaged.
3. Thermal shutdown has engaged.

An external capacitor (C_{TMR}) from TIMR to AGND programs the fault filter time as follows:

$$C_{TMR} = \frac{t_{FLT} \times 11.5\,\mu A}{1.24V}$$

When the voltage on the TIMR pin reaches 1.24V, the device is latched off and the N-channel MosFET open drain FLT pin transitions to a high impedance state. The TIMR pin will be immediately pulled to ground (reset) if the fault condition is removed at any point during the filter period. Otherwise, if the timer expires, the fault will remain latched until one of three things occurs:

1. The EN pin is pulled low long enough for the V_{CC} pin to drop below 4.1V (approximately 200 ms).
2. The TIMR pin is pulled to ground.
3. A complete power cycle occurs.

When using the EN and OVP pins in conjunction with the RPD pull-down pin, a race condition exists when exiting the disabled (EN low) state. When disabled, the OVP pin is pulled up to the output voltage because the RPD pull-down is disabled, and this will appear to be a real OVLO condition. The timer pin will immediately rise and latch the controller to the fault state. To protect against this behavior, a minimum timer capacitor (C_{TMR} = 220pF) should be used. If fault latching is not required, short the TMR pin to AGND which will disable the FLT flag function.

The LM3423 also includes an LED Ready (LRDY) flag to notify the system that the LEDs are in proper regulation. The N-channel MosFET open drain LRDY pin is pulled low whenever any of the following conditions are met:

1. V_{CC} UVLO has engaged.
2. LED current is below regulation by more than 20%.
3. LED current is above regulation by more than 30%.
4. Over-voltage protection has engaged
5. Thermal shutdown has engaged.
6. A fault has latched the device off.

Note that the LRDY pin is pulled low during startup of the device and remains low until the LED current is in regulation.

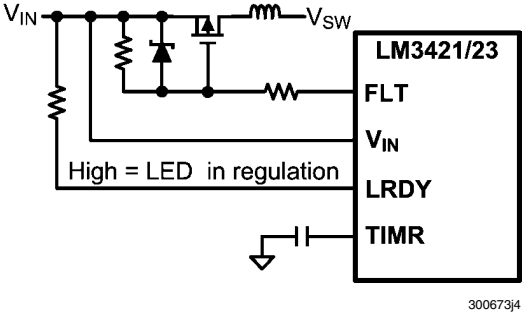


FIGURE 19. Fault Detection and LED Status Circuit

Design Considerations

This section describes the application level considerations when designing with the LM3421/23. For corresponding calculations, refer to the Design Guide section.

INDUCTOR

The inductor (L1) is the main energy storage device in a switching regulator. Depending on the topology, energy is stored in the inductor and transfered to the load in different ways (as an example, buck-boost operation is detailed in the Current Regulators section). The size of the inductor, the voltage across it, and the length of the switching subinterval (t_{ON} or t_{OFF}) determines the inductor current ripple ($\Delta i_{L\text{-}PP}$). In the design process, L1 is chosen to provide a desired $\Delta i_{L\text{-}PP}$. For a buck regulator the inductor has a direct connection to the load, which is good for a current regulator. This requires little to no output capacitance therefore $\Delta i_{L\text{-}PP}$ is basically equal to the LED ripple current $\Delta i_{LED\text{-}PP}$. However, for boost and buck-boost regulators, there is always an output capacitor which reduces $\Delta i_{LED\text{-}PP}$, therefore the inductor ripple can be larger than in the buck regulator case where output capacitance is minimal or completely absent.

In general, $\Delta i_{LED\text{-}PP}$ is recommended by manufacturers to be less than 40% of the average LED current (I_{LED}). Therefore, for the buck regulator with no output capacitance, $\Delta i_{L\text{-}PP}$ should also be less than 40% of I_{LED}. For the boost and buck-boost topologies, $\Delta i_{L\text{-}PP}$ can be much higher depending on the output capacitance value. However, $\Delta i_{L\text{-}PP}$ is suggested to be less than 100% of the average inductor current (I_L) to limit the RMS inductor current.

L1 is also suggested to have an RMS current rating at least 25% higher than the calculated minimum allowable RMS inductor current ($I_{L\text{-}RMS}$).

LED DYNAMIC RESISTANCE

When the load is a string of LEDs, the output load resistance is the LED string dynamic resistance plus R_{SNS}. LEDs are PN junction diodes, and their dynamic resistance shifts as their forward current changes. Dividing the forward voltage of a single LED (V_{LED}) by the forward current (I_{LED}) leads to an incorrect calculation of the dynamic resistance of a single LED (r_{LED}). The result can be 5 to 10 times higher than the true r_{LED} value.

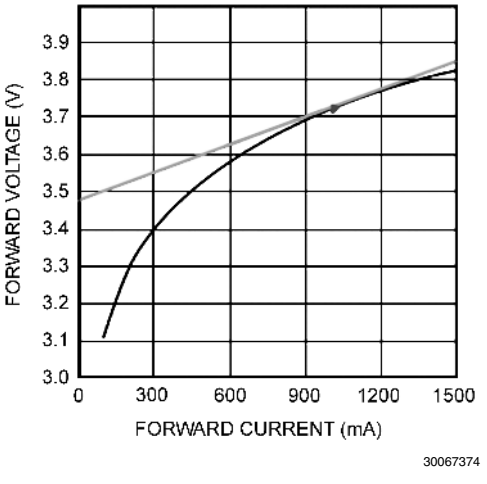


FIGURE 20. Dynamic Resistance

Obtaining r_{LED} is accomplished by refering to the manufacturer's LED I-V characteristic. It can be calculated as the slope at the nominal operating point as shown in Figure 20. For any application with more than 2 series LEDs, R_{SNS} can be neglected allowing r_D to be approximated as the number of LEDs multiplied by r_{LED}.

OUTPUT CAPACITOR

For boost and buck-boost regulators, the output capacitor (C_O) provides energy to the load when the recirculating diode (D1) is reverse biased during the first switching subinterval. An output capacitor in a buck topology will simply reduce the LED current ripple ($\Delta i_{LED\text{-}PP}$) below the inductor current ripple ($\Delta i_{L\text{-}PP}$). In all cases, C_O is sized to provide a desired $\Delta i_{LED\text{-}PP}$. As mentioned in the Inductor section, $\Delta i_{LED\text{-}PP}$ is recommended by manufacturers to be less than 40% of the average LED current ($I_{LED\text{-}PP}$).

C_O should be carefully chosen to account for derating due to temperature and operating voltage. It must also have the necessary RMS current rating. Ceramic capacitors are the best choice due to their high ripple current rating, long lifetime, and good temperature performance. An X7R dieletric rating is suggested.

INPUT CAPACITORS

The input capacitance (C_{IN}) provides energy during the discontinuous portions of the switching period. For buck and buck-boost regulators, C_{IN} provides energy during t_{ON} and during t_{OFF}, the input voltage source charges up C_{IN} with the average input current (I_{IN}). For boost regulators, C_{IN} only needs to provide the ripple current due to the direct connection to the inductor. C_{IN} is selected given the maximum input voltage ripple ($\Delta v_{IN\text{-}PP}$) which can be tolerated. $\Delta v_{IN\text{-}PP}$ is suggested to be less than 10% of the input voltage (V_{IN}).

An input capacitance at least 100% greater than the calculated C_{IN} value is recommended to account for derating due to temperature and operating voltage. When PWM dimming, even more capacitance can be helpful to minimize the large current draw from the input voltage source during the rising transistion of the LED current waveform.

The chosen input capacitors must also have the necessary RMS current rating. Ceramic capacitors are again the best choice due to their high ripple current rating, long lifetime, and good temperature performance. An X7R dieletric rating is suggested.

For most applications, it is recommended to bypass the V_{IN} pin with an 0.1 µF ceramic capacitor placed as close as possible to the pin. In situations where the bulk input capacitance may be far from the LM3421/23 device, a 10 Ω series resistor can be placed between the bulk input capacitance and the bypass capacitor, creating a 150 kHz filter to eliminate undesired high frequency noise.

MAIN MosFET / DIMMING MosFET

The LM3421/23 requires an external NFET (Q1) as the main power MosFET for the switching regulator. Q1 is recommended to have a voltage rating at least 15% higher than the maximum transistor voltage to ensure safe operation during the ringing of the switch node. In practice, all switching regulators have some ringing at the switch node due to the diode parasitic capacitance and the lead inductance. The current rating is recommended to be at least 10% higher than the average transistor current. The power rating is then verified by calculating the power loss given the RMS transistor current and the NFET on-resistance (R_{DS-ON}).

When PWM dimming, the LM3421/23 requires another MosFET (Q2) placed in series (or parallel for a buck regulator) with the LED load. This MosFET should have a voltage rating equal to the output voltage (V_O) and a current rating at least 10% higher than the nominal LED current (I_{LED}) . The power rating is simply V_O multiplied by I_{LED}, assuming 100% dimming duty cycle (continuous operation) will occur.

In general, the NFETs should be chosen to minimize total gate charge (Q_g) when f_{SW} is high and minimize R_{DS-ON} otherwise. This will minimize the dominant power losses in the system. Frequently, higher current NFETs in larger packages are chosen for better thermal performance.

RE-CIRCULATING DIODE

A re-circulating diode (D1) is required to carry the inductor current during t_{OFF}. The most efficient choice for D1 is a Schottky diode due to low forward voltage drop and near-zero reverse recovery time. Similar to Q1, D1 is recommended to have a voltage rating at least 15% higher than the maximum transistor voltage to ensure safe operation during the ringing of the switch node and a current rating at least 10% higher than the average diode current. The power rating is verified by calculating the power loss through the diode. This is accomplished by checking the typical diode forward voltage from the I-V curve on the product datasheet and multiplying by the average diode current. In general, higher current diodes have a lower forward voltage and come in better performing packages minimizing both power losses and temperature rise.

BOOST INRUSH CURRENT

When configured as a boost converter, there is a “phantom” power path comprised of the inductor, the output diode, and the output capacitor. This path will cause two things to happen when power is applied. First, there will be a very large inrush of current to charge the output capacitor. Second, the energy stored in the inductor during this inrush will end up in the output capacitor, charging it to a higher potential than the input voltage.

Depending on the state of the EN pin, the output capacitor would be discharged by:

1. EN < 1.3V: no discharge path (leakage only).
2. EN > 1.3V, the OVP divider resistor path, if present, and 10µA into each of the HSP & HSN pins.

In applications using the OVP divider and with EN > 1.3V, the output capacitor voltage can charge higher than $V_{TURN-OFF}$. In this situation, the FLT pin (LM3423 only) is open and the PWM dimming MosFET is turned off. This condition (the system appearing disabled) can persist for an undesirably long time. Possible solutions to this condition are:

- Add an inrush diode from V_{IN} to the output as shown in Figure 21.
- Add an NTC thermistor in series with the input to prevent the inrush from overcharging the output capacitor too high.
- Use a current limited source supply.
- Raise the OVP threshold.

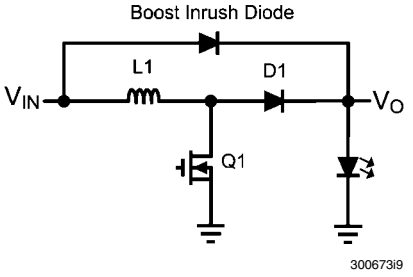


FIGURE 21. Boost Topology with Inrush Diode

CIRCUIT LAYOUT

The performance of any switching regulator depends as much upon the layout of the PCB as the component selection. Following a few simple guidelines will maximimize noise rejection and minimize the generation of EMI within the circuit.

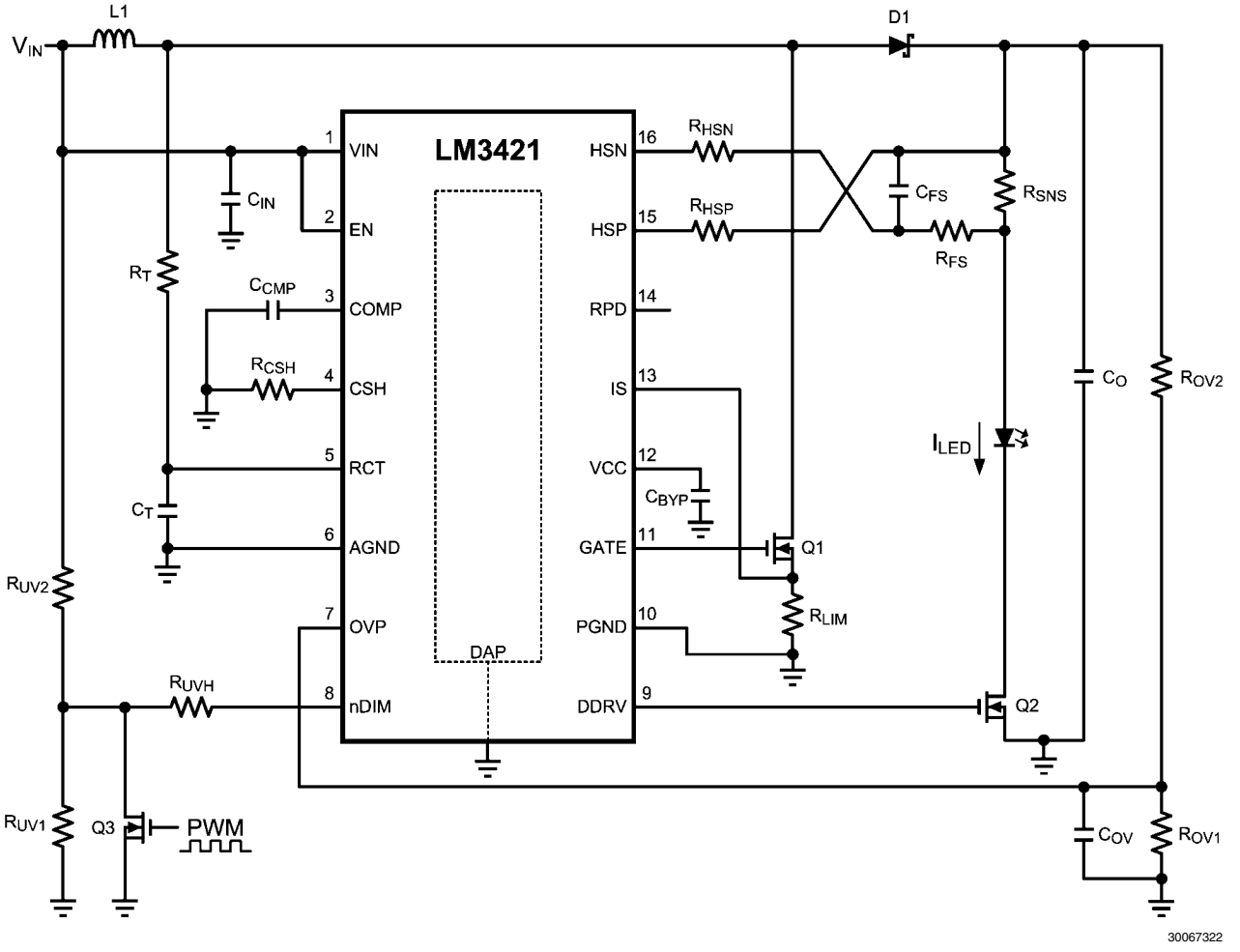
Discontinuous currents are the most likely to generate EMI, therefore care should be taken when routing these paths. The main path for discontinuous current in the LM3421/23 buck regulator contains the input capacitor (C_{IN}), the recirculating diode (D1), the N-channel MosFET (Q1), and the sense resistor (R_{LIM}). In the LM3421/23 boost regulator, the discontinuous current flows through the output capacitor (C_O), D1, Q1, and R_{LIM}. In the buck-boost regulator both loops are discontinuous and should be carefully layed out. These loops should be kept as small as possible and the connections between all the components should be short and thick to minimize parasitic inductance. In particular, the switch node (where L1, D1 and Q1 connect) should be just large enough to connect the components. To minimize excessive heating, large copper pours can be placed adjacent to the short current path of the switch node.

The RT, COMP, CSH, IS, HSP and HSN pins are all high-impedance inputs which couple external noise easily, therefore the loops containing these nodes should be minimized whenever possible.

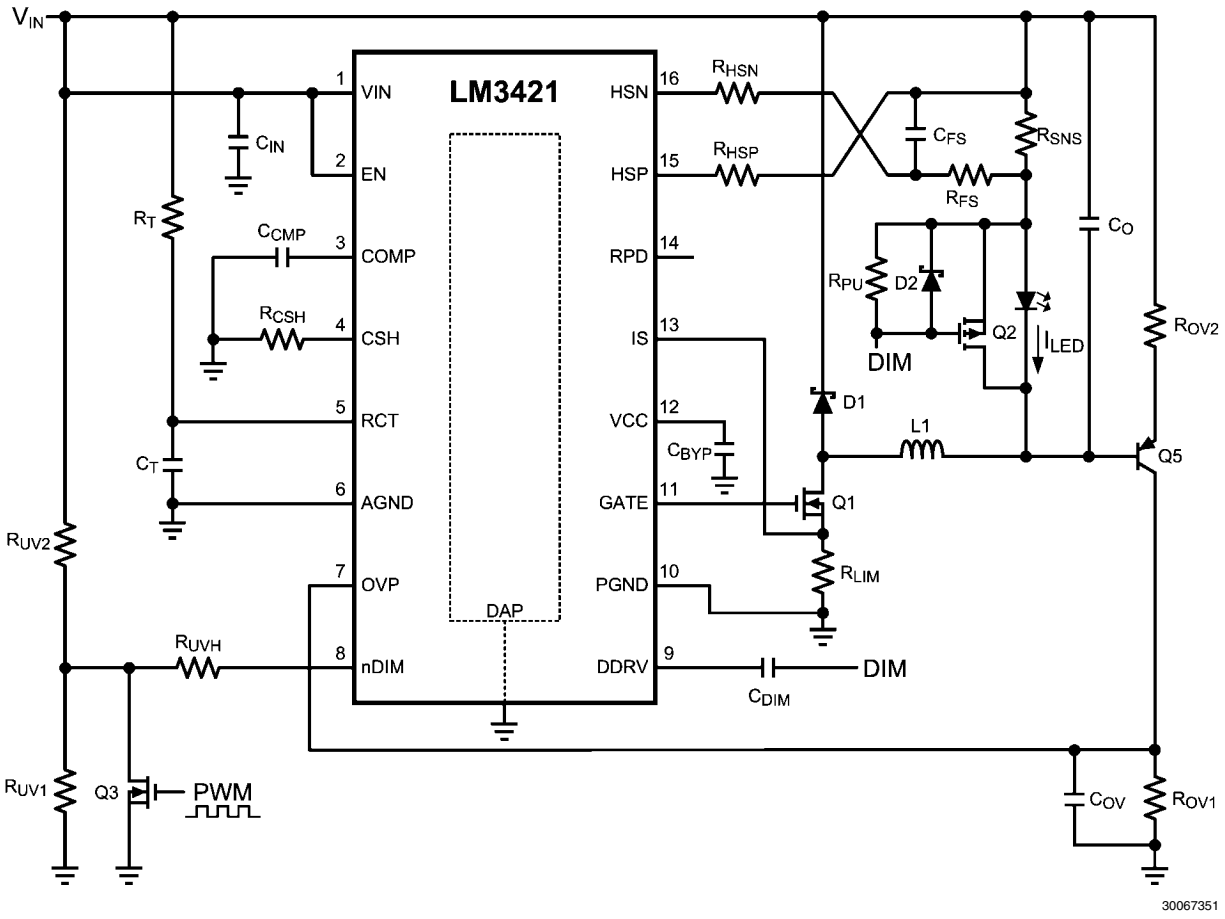
In some applications the LED or LED array can be far away (several inches or more) from the LM3421/23, or on a separate PCB connected by a wiring harness. When an output capacitor is used and the LED array is large or separated from the rest of the regulator, the output capacitor should be placed close to the LEDs to reduce the effects of parasitic inductance on the AC impedance of the capacitor.

Basic Topology Schematics

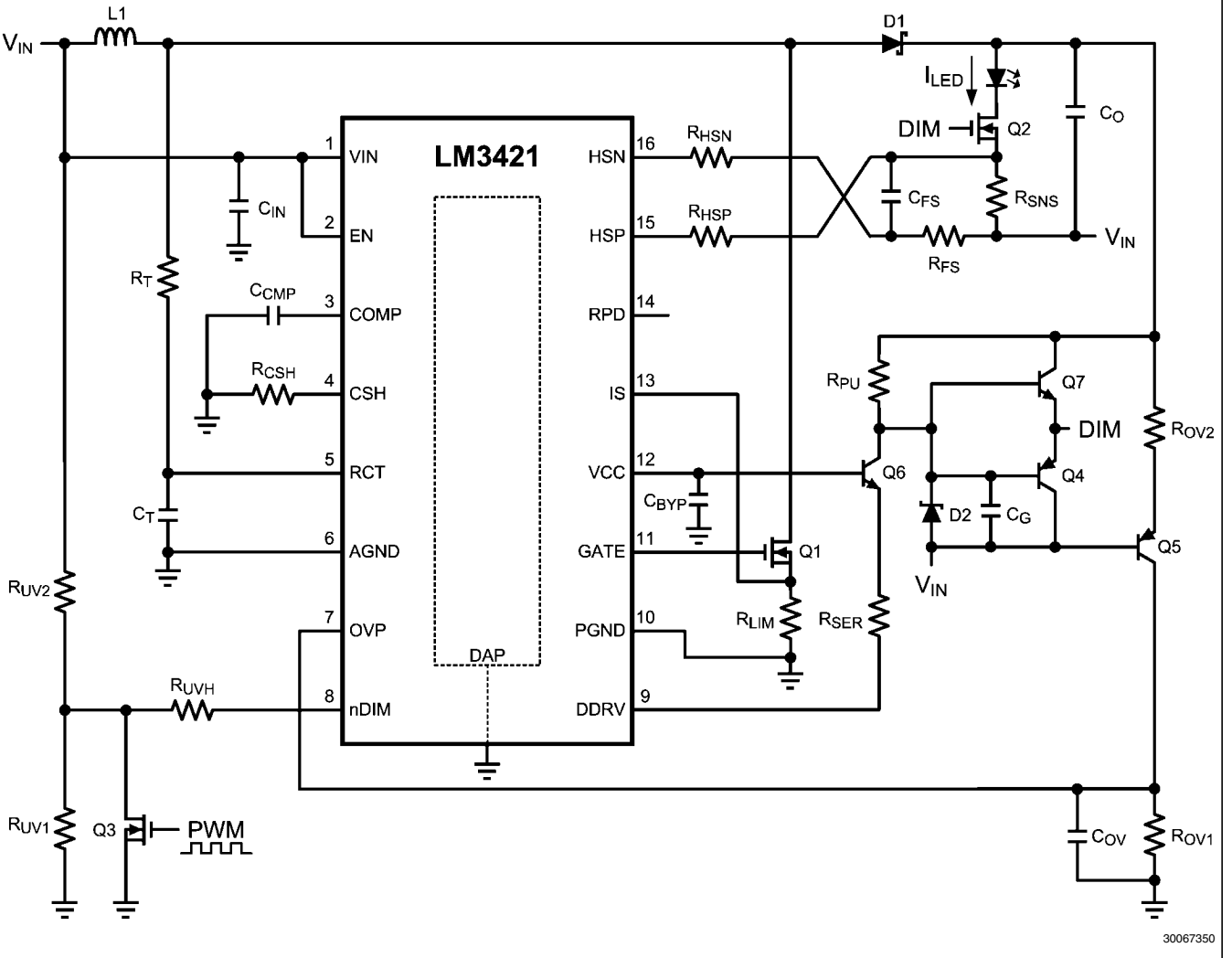
BOOST REGULATOR (V_{IN} < V_O)



BUCK REGULATOR (V_{IN} > V_O)



BUCK-BOOST REGULATOR



Design Guide

Refer to Basic Topology Schematics section.

SPECIFICATIONS

Number of series LEDs: N
Single LED forward voltage: V_{LED}
Single LED dynamic resistance: r_{LED}
Nominal input voltage: V_{IN}
Input voltage range: V_{IN-MAX}, V_{IN-MIN}
Switching frequency: f_{SW}
Current sense voltage: V_{SNS}
Average LED current: I_{LED}
Inductor current ripple: Δi_{L-PP}
LED current ripple: Δi_{LED-PP}
Peak current limit: I_{LIM}
Input voltage ripple: Δv_{IN-PP}
Output OVLO characteristics: $V_{TURN-OFF}$, V_{HYSO}
Input UVLO characteristics: $V_{TURN-ON}$, V_{HYS}

1. OPERATING POINT

Given the number of series LEDs (N), the forward voltage (V_{LED}) and dynamic resistance (r_{LED}) for a single LED, solve for the nominal output voltage (V_O) and the nominal LED string dynamic resistance (r_D):

$$V_O = N \times V_{LED}$$

$$r_D = N \times r_{LED}$$

Solve for the ideal nominal duty cycle (D):

Buck

$$D = \frac{V_O}{V_{IN}}$$

Boost

$$D = \frac{V_O - V_{IN}}{V_O}$$

Buck-boost

$$D = \frac{V_O}{V_O + V_{IN}}$$

Using the same equations, find the minimum duty cycle (D_{MIN}) using maximum input voltage (V_{IN-MAX}) and the maximum duty cycle (D_{MAX}) using the minimum input voltage (V_{IN-MIN}). Also, remember that D' = 1 - D.

2. SWITCHING FREQUENCY

Set the switching frequency (f_{SW}) by assuming a C_T value of 1 nF and solving for R_T:

Buck (Constant Ripple vs. V_{IN})

$$R_T = \frac{25 \times (V_{IN} - V_O)}{f_{SW} \times C_T \times V_{IN}}$$

Buck (Constant Ripple vs. V_O)

$$R_T = \frac{25 \times (V_{IN} \times V_O - V_O^2)}{f_{SW} \times C_T \times V_{IN}^2}$$

Boost and Buck-boost

$$R_T = \frac{25}{f_{SW} \times C_T}$$

3. AVERAGE LED CURRENT

For all topologies, set the average LED current (I_{LED}) knowing the desired current sense voltage (V_{SNS}) and solving for R_{SNS}:

$$R_{SNS} = \frac{V_{SNS}}{I_{LED}}$$

If the calculated R_{SNS} is too far from a desired standard value, then V_{SNS} will have to be adjusted to obtain a standard value.

Setup the suggested signal current of 100 µA by assuming R_{CSH} = 12.4 kΩ and solving for R_{HSP}:

$$R_{HSP} = \frac{I_{LED} \times R_{CSH} \times R_{SNS}}{1.24V}$$

If the calculated R_{HSP} is too far from a desired standard value, then R_{CSH} can be adjusted to obtain a standard value.

4. INDUCTOR RIPPLE CURRENT

Set the nominal inductor ripple current (Δi_{L-PP}) by solving for the appropriate inductor (L1):

Buck

$$L1 = \frac{(V_{IN} - V_O) \times D}{\Delta i_{L-PP} \times f_{SW}}$$

Boost and Buck-boost

$$L1 = \frac{V_{IN} \times D}{\Delta i_{L-PP} \times f_{SW}}$$

To set the worst case inductor ripple current, use V_{IN-MAX} and D_{MIN} when solving for L1.

The minimum allowable inductor RMS current rating (I_{L-RMS}) can be calculated as:

Buck

$$I_{L-RMS} = I_{LED} \times \sqrt{1 + \frac{1}{12} \times \left(\frac{\Delta I_{L-PP}}{I_{LED}}\right)^2}$$

Boost and Buck-boost

$$I_{L-RMS} = \frac{I_{LED}}{D'} \times \sqrt{1 + \frac{1}{12} \times \left(\frac{\Delta I_{L-PP} \times D'}{I_{LED}}\right)^2}$$

5. LED RIPPLE CURRENT

Set the nominal LED ripple current (Δi_{LED-PP}), by solving for the output capacitance (C_O):

Buck

$$C_O = \frac{\Delta i_{L-PP}}{8 \times f_{SW} \times r_D \times \Delta i_{LED-PP}}$$

Boost and Buck-boost

$$C_O = \frac{I_{LED} \times D}{r_D \times \Delta i_{LED-PP} \times f_{SW}}$$

To set the worst case LED ripple current, use D_{MAX} when solving for C_O. Remember, when PWM dimming it is recommended to use a minimum of 40 µF of output capacitance to improve performance.

The minimum allowable RMS output capacitor current rating (I_{CO-RMS}) can be approximated:

Buck

$$I_{CO-RMS} = \frac{\Delta i_{LED-PP}}{\sqrt{12}}$$

Boost and Buck-boost

$$I_{CO-RMS} = I_{LED} \times \sqrt{\frac{D_{MAX}}{1-D_{MAX}}}$$

6. PEAK CURRENT LIMIT

Set the peak current limit (I_{LIM}) by solving for the transistor path sense resistor (R_{LIM}):

$$R_{LIM} = \frac{245\ mV}{I_{LIM}}$$

7. LOOP COMPENSATION

Using a simple first order peak current mode control model, neglecting any output capacitor ESR dynamics, the necessary loop compensation can be determined.

First, the uncompensated loop gain (T_U) of the regulator can be approximated:

Buck

$$T_U = T_{U0} \times \frac{1}{\left(1 + \frac{s}{\omega_{P1}}\right)}$$

Boost and Buck-boost

$$T_U = T_{U0} \times \frac{\left(1 - \frac{s}{\omega_{Z1}}\right)}{\left(1 + \frac{s}{\omega_{P1}}\right)}$$

Where the pole (ω_{P1}) is approximated:

Buck

$$\omega_{P1} = \frac{1}{r_D \times C_O}$$

Boost

$$\omega_{P1} = \frac{2}{r_D \times C_O}$$

Buck-boost

$$\omega_{P1} = \frac{1+D}{r_D \times C_O}$$

And the RHP zero (ω_{Z1}) is approximated:

Boost

$$\omega_{Z1} = \frac{r_D \times D'^2}{L1}$$

Buck-boost

$$\omega_{Z1} = \frac{r_D \times D'^2}{D \times L1}$$

And the uncompensated DC loop gain (T_{U0}) is approximated:

Buck

$$T_{U0} = \frac{500V \times R_{CSH} \times R_{SNS}}{R_{HSP} \times R_{LIM}} = \frac{620V}{I_{LED} \times R_{LIM}}$$

Boost

$$T_{U0} = \frac{D' \times 500V \times R_{CSH} \times R_{SNS}}{2 \times R_{HSP} \times R_{LIM}} = \frac{D' \times 310V}{I_{LED} \times R_{LIM}}$$

Buck-boost

$$T_{U0} = \frac{D' \times 500V \times R_{CSH} \times R_{SNS}}{(1+D) \times R_{HSP} \times R_{LIM}} = \frac{D' \times 620V}{(1+D) \times I_{LED} \times R_{LIM}}$$

For all topologies, the primary method of compensation is to place a low frequency dominant pole (ω_{P2}) which will ensure that there is ample phase margin at the crossover frequency. This is accomplished by placing a capacitor (C_{CMP}) from the COMP pin to AGND, which is calculated according to the lower value of the pole and the RHP zero of the system (shown as a minimizing function):

$$\omega_{P2} = \frac{\min(\omega_{P1}, \omega_{Z1})}{5 \times T_{U0}}$$

$$C_{CMP} = \frac{1}{\omega_{P2} \times 5e^6}$$

If analog dimming is used, C_{CMP} should be approximately 4x larger to maintain stability as the LEDs are dimmed to zero.

A high frequency compensation pole (ω_{P3}) can be used to attenuate switching noise and provide better gain margin. Assuming R_{FS} = 10Ω, C_{FS} is calculated according to the higher value of the pole and the RHP zero of the system (shown as a maximizing function):

$$\omega_{P3} = \max(\omega_{P1}, \omega_{Z1}) \times 10$$

$$C_{FS} = \frac{1}{10 \times \omega_{P3}}$$

The total system loop gain (T) can then be written as:

Buck

$$T = T_{U0} \times \frac{1}{\left(1+\frac{s}{\omega_{P1}}\right) \times \left(1+\frac{s}{\omega_{P2}}\right) \times \left(1+\frac{s}{\omega_{P3}}\right)}$$

Boost and Buck-boost

$$T = T_{U0} \times \frac{\left(1-\frac{s}{\omega_{Z1}}\right)}{\left(1+\frac{s}{\omega_{P1}}\right) \times \left(1+\frac{s}{\omega_{P2}}\right) \times \left(1+\frac{s}{\omega_{P3}}\right)}$$

8. INPUT CAPACITANCE

Set the nominal input voltage ripple (Δv_{IN-PP}) by solving for the required capacitance (C_{IN}):

Buck

$$C_{IN} = \frac{I_{LED} \times (1 - D) \times D}{\Delta V_{IN-PP} \times f_{SW}}$$

Boost

$$C_{IN} = \frac{\Delta i_{L-PP}}{8 \times \Delta V_{IN-PP} \times f_{SW}}$$

Buck-boost

$$C_{IN} = \frac{I_{LED} \times D}{\Delta V_{IN-PP} \times f_{SW}}$$

Use D_{MAX} to set the worst case input voltage ripple, when solving for C_{IN} in a buck-boost regulator and D_{MID} = 0.5 when solving for C_{IN} in a buck regulator.

The minimum allowable RMS input current rating ($I_{CIN-RMS}$) can be approximated:

Buck

$$I_{CIN-RMS} = I_{LED} \times \sqrt{D_{MID} \times (1-D_{MID})}$$

Boost

$$I_{CIN-RMS} = \frac{\Delta i_{L-PP}}{\sqrt{12}}$$

Buck-boost

$$I_{CIN-RMS} = I_{LED} \times \sqrt{\frac{D_{MAX}}{1-D_{MAX}}}$$

9. NFET

The NFET voltage rating should be at least 15% higher than the maximum NFET drain-to-source voltage (V_{T-MAX}):

Buck

$$V_{T-MAX} = V_{IN-MAX}$$

Boost

$$V_{T-MAX} = V_O$$

Buck-boost

$$V_{T-MAX} = V_{IN-MAX} + V_O$$

The current rating should be at least 10% higher than the maximum average NFET current (I_{T-MAX}):

Buck

$$I_{T-MAX} = D_{MAX} \times I_{LED}$$

Boost and Buck-boost

$$I_{T-MAX} = \frac{D_{MAX}}{1 - D_{MAX}} \times I_{LED}$$

Approximate the nominal RMS transistor current (I_{T-RMS}) :

Buck

$$I_{T-RMS} = I_{LED} \times \sqrt{D}$$

Boost and Buck-boost

$$I_{T-RMS} = \frac{I_{LED}}{D'} \times \sqrt{D}$$

Given an NFET with on-resistance (R_{DS-ON}), solve for the nominal power dissipation (P_T):

$$P_T = I_{T-RMS}^2 \times R_{DSON}$$

10. DIODE

The Schottky diode voltage rating should be at least 15% higher than the maximum blocking voltage (V_{RD-MAX}):

Buck

$$V_{RD-MAX} = V_{IN-MAX}$$

Boost

$$V_{RD-MAX} = V_O$$

Buck-boost

$$V_{RD-MAX} = V_{IN-MAX} + V_O$$

The current rating should be at least 10% higher than the maximum average diode current (I_{D-MAX}):

Buck

$$I_{D-MAX} = (1 - D_{MIN}) \times I_{LED}$$

Boost and Buck-boost

$$I_{D-MAX} = I_{LED}$$

Replace D_{MAX} with D in the I_{D-MAX} equation to solve for the average diode current (I_D). Given a diode with forward voltage (V_{FD}), solve for the nominal power dissipation (P_D):

$$P_D = I_D \times V_{FD}$$

11. OUTPUT OVLO

For boost and buck-boost regulators, output OVLO is programmed with the turn-off threshold voltage ($V_{TURN-OFF}$) and the desired hysteresis (V_{HYSO}). To set V_{HYSO}, solve for R_{OV2}:

$$R_{OV2} = \frac{V_{HYSO}}{23\,\mu A}$$

To set $V_{TURN-OFF}$, solve for R_{OV1}:

Boost

$$R_{OV1} = \frac{1.24V \times R_{OV2}}{V_{TURN-OFF} - 1.24V}$$

Buck-boost

$$R_{OV1} = \frac{1.24V \times R_{OV2}}{V_{TURN-OFF} - 620\,mV}$$

A small filter capacitor (C_{OVP} = 47 pF) should be added from the OVP pin to ground to reduce coupled switching noise.

12. INPUT UVLO

For all topologies, input UVLO is programmed with the turn-on threshold voltage ($V_{TURN-ON}$) and the desired hysteresis (V_{HYS}).

Method #1: If no PWM dimming is required, a two resistor network can be used. To set V_{HYS}, solve for R_{UV2}:

$$R_{UV2} = \frac{V_{HYS}}{23\,\mu A}$$

To set $V_{TURN-ON}$, solve for R_{UV1}:

$$R_{UV1} = \frac{1.24V \times R_{UV2}}{V_{TURN-ON} - 1.24V}$$

Method #2: If PWM dimming is required, a three resistor network is suggested. To set $V_{TURN-ON}$, assume R_{UV2} = 10 kΩ and solve for R_{UV1} as in Method #1. To set V_{HYS}, solve for R_{UVH}:

$$R_{UVH} = \frac{R_{UV1} \times (V_{HYS} - 23\,\mu A \times R_{UV2})}{23\,\mu A \times (R_{UV1} + R_{UV2})}$$

13. PWM DIMMING METHOD

PWM dimming can be performed several ways:

Method #1: Connect the dimming MosFET (Q_3) with the drain to the nDIM pin and the source to AGND. Apply an external PWM signal to the gate of Q_{DIM}. A pull down resistor may be necessary to properly turn off Q_3.

Method #2: Connect the anode of a Schottky diode to the nDIM pin. Apply an external inverted PWM signal to the cathode of the same diode.

The DDRV pin should be connected to the gate of the dimFET with or without level-shifting circuitry as described in the PWM Dimming section. The dimFET should be rated to handle the average LED current and the nominal output voltage.

14. ANALOG DIMMING METHOD

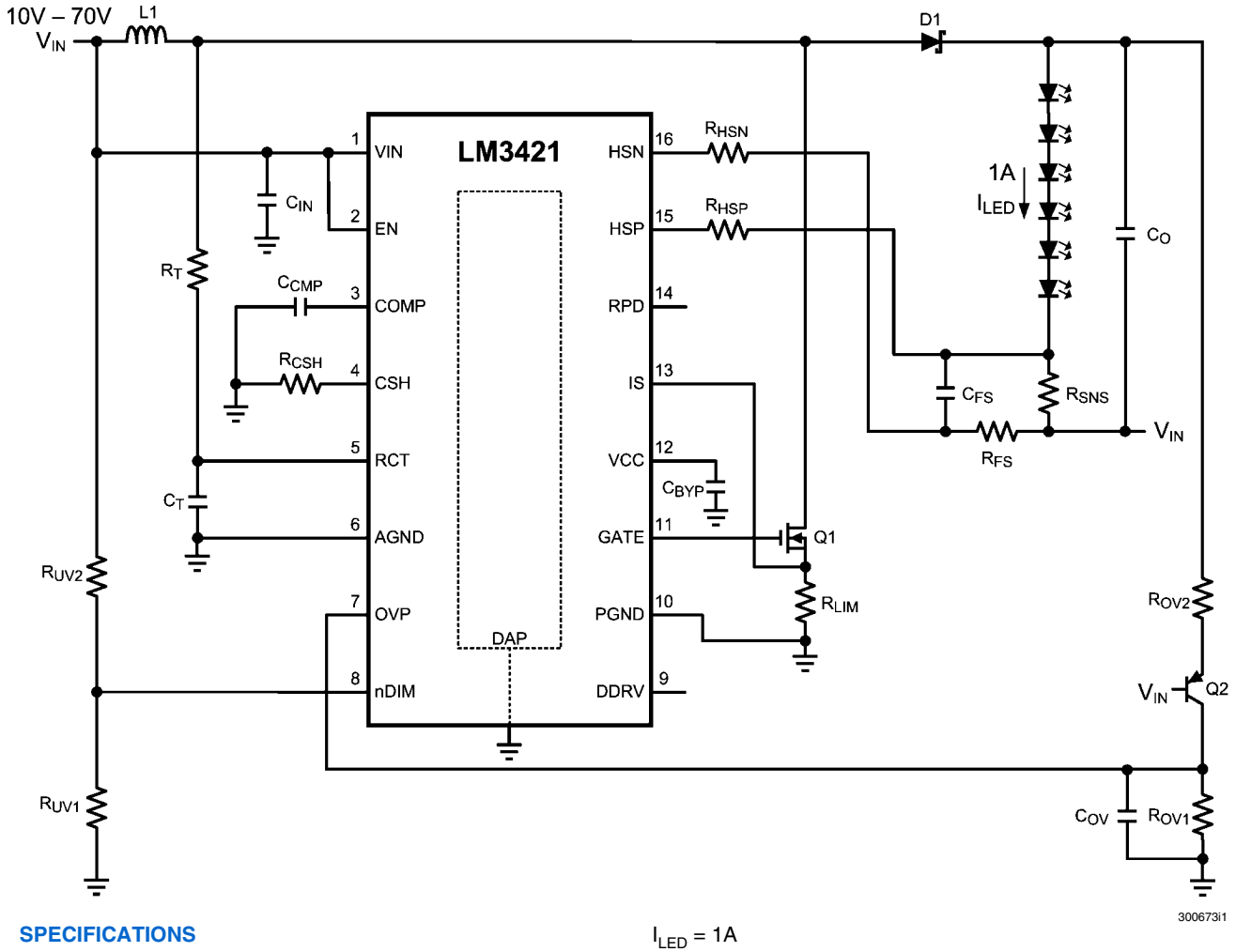
Analog dimming can be performed several ways:

Method #1: Place a potentiometer in series with the R_{CSH} resistor to dim the LED current from the nominal I_{LED} to near zero.

Method #2: Connect a controlled current source as detailed in the Analog Dimming section to the CSH pin. Increasing the current sourced into the CSH node will decrease the LEDs from the nominal I_{LED} to zero current in the same manner as the thermal foldback circuit.

Design Example

DESIGN #1 - LM3421 BUCK-BOOST Application



SPECIFICATIONS

N = 6
V_{LED} = 3.5V
r_{LED} = 325 mΩ
V_{IN} = 24V
V_{IN-MIN} = 10V
V_{IN-MAX} = 70V
f_{SW} = 500 kHz
V_{SNS} = 100 mV
I_{LED} = 1A
Δi_{L-PP} = 700 mA
Δi_{LED-PP} = 12 mA
Δv_{IN-PP} = 100 mV
I_{LIM} = 6A
$V_{TURN-ON}$ = 10V
V_{HYS} = 3V
$V_{TURN-OFF}$ = 40V
V_{HYSO} = 10V

1. OPERATING POINT

Solve for V_O and r_D:

$$V_O = N \times V_{LED} = 6 \times 3.5V = 21V$$

$$r_D = N \times r_{LED} = 6 \times 325\ m\Omega = 1.95\Omega$$

Solve for D, D', D_{MAX}, and D_{MIN}:

$$D = \frac{V_O}{V_O + V_{IN}} = \frac{21V}{21V + 24V} = 0.467$$

$$D' = 1 - D = 1 - 0.467 = 0.533$$

$$D_{MIN} = \frac{V_O}{V_O + V_{IN\text{-}MAX}} = \frac{21V}{21V + 70V} = 0.231$$

$$D_{MAX} = \frac{V_O}{V_O + V_{IN\text{-}MIN}} = \frac{21V}{21V + 10V} = 0.677$$

2. SWITCHING FREQUENCY

Assume C_T = 1 nF and solve for R_T:

$$R_T = \frac{25}{f_{SW} \times C_T} = \frac{25}{500\ kHz \times 1\ nF} = 50\ k\Omega$$

The closest standard resistor is 49.9 kΩ therefore f_{SW} is:

$$f_{SW} = \frac{25}{R_T \times C_T} = \frac{25}{49.9\ k\Omega \times 1\ nF} = 501\ kHz$$

The chosen component from step 2 is:

C_T = 1 nF
R_T = 49.9 kΩ

3. AVERAGE LED CURRENT

Solve for R_{SNS}:

$$R_{SNS} = \frac{V_{SNS}}{I_{LED}} = \frac{100\ mV}{1A} = 0.1\Omega$$

Assume R_{CSH} = 12.4 kΩ and solve for R_{HSP}:

$$R_{HSP} = \frac{I_{LED} \times R_{CSH} \times R_{SNS}}{1.24V} = \frac{1A \times 12.4\ k\Omega \times 0.1\Omega}{1.24V} = 1.0\ k\Omega$$

The closest standard resistor for R_{SNS} is actually 0.1Ω and for R_{HSP} is actually 1 kΩ therefore I_{LED} is:

$$I_{LED} = \frac{1.24V \times R_{HSP}}{R_{SNS} \times R_{CSH}} = \frac{1.24V \times 1.0\ k\Omega}{0.1\Omega \times 12.4\ k\Omega} = 1.0A$$

The chosen components from step 3 are:

R_{SNS} = 0.1Ω
R_{CSH} = 12.4 kΩ
$R_{HSP} = R_{HSN}$ = 1 kΩ

4. INDUCTOR RIPPLE CURRENT

Solve for L1:

$$L1 = \frac{V_{IN} \times D}{\Delta i_{L\text{-}PP} \times f_{SW}} = \frac{24V \times 0.467}{700\ mA \times 501\ kHz} = 32\ \mu H$$

The closest standard inductor is 33 µH therefore $\Delta i_{L\text{-}PP}$ is:

$$\Delta i_{L\text{-}PP} = \frac{V_{IN} \times D}{L1 \times f_{SW}} = \frac{24V \times 0.467}{33\ \mu H \times 501\ kHz} = 678\ mA$$

Determine minimum allowable RMS current rating:

$$I_{L\text{-}RMS} = \frac{I_{LED}}{D'} \times \sqrt{1 + \frac{1}{12} \times \left(\frac{\Delta i_{L\text{-}PP} \times D'}{I_{LED}}\right)^2}$$

$$I_{L\text{-}RMS} = \frac{1A}{0.533} \times \sqrt{1 + \frac{1}{12} \times \left(\frac{678\ mA \times 0.533}{1A}\right)^2} = 1.89A$$

The chosen component from step 4 is:

L1 = 33 µH

5. OUTPUT CAPACITANCE

Solve for C_O:

$$C_O = \frac{I_{LED} \times D}{r_D \times \Delta i_{LED\text{-}PP} \times f_{SW}}$$

$$C_O = \frac{1A \times 0.467}{1.95\Omega \times 12\ mA \times 501\ kHz} = 39.8\ \mu F$$

The closest capacitance totals 40 µF therefore $\Delta i_{LED\text{-}PP}$ is:

$$\Delta i_{LED\text{-}PP} = \frac{I_{LED} \times D}{r_D \times C_O \times f_{SW}}$$

$$\Delta i_{LED\text{-}PP} = \frac{1A \times 0.467}{1.95\Omega \times 40\ \mu F \times 501\ kHz} = 12\ mA$$

Determine minimum allowable RMS current rating:

$$I_{CO\text{-}RMS} = I_{LED} \times \sqrt{\frac{D_{MAX}}{1 - D_{MAX}}} = 1A \times \sqrt{\frac{0.677}{1 - 0.677}} = 1.45A$$

The chosen components from step 5 are:

C_O = 4 x 10 µF

6. PEAK CURRENT LIMIT

Solve for R_{LIM}:

$$R_{LIM} = \frac{245\,mV}{I_{LIM}} = \frac{245\,mV}{6A} = 0.041\Omega$$

The closest standard resistor is 0.04 Ω therefore I_{LIM} is:

$$I_{LIM} = \frac{245\,mV}{R_{LIM}} = \frac{245\,mV}{0.04\Omega} = 6.13A$$

The chosen component from step 6 is:

$$\boxed{R_{LIM} = 0.04\Omega}$$

7. LOOP COMPENSATION

ω_{P1} is approximated:

$$\omega_{P1} = \frac{1+D}{r_D \times C_O} = \frac{1.467}{1.95\Omega \times 40\,\mu F} = 19k\frac{rad}{sec}$$

ω_{Z1} is approximated:

$$\omega_{Z1} = \frac{r_D \times D'^2}{D \times L1} = \frac{1.95\Omega \times 0.533^2}{0.467 \times 33\,\mu H} = 36k\frac{rad}{sec}$$

T_{U0} is approximated:

$$T_{U0} = \frac{D' \times 620V}{(1+D) \times I_{LED} \times R_{LIM}} = \frac{0.533 \times 620V}{1.467 \times 1A \times 0.04\Omega} = 5630$$

To ensure stability, calculate ω_{P2}:

$$\omega_{P2} = \frac{\min(\omega_{P1}, \omega_{Z1})}{5 \times T_{U0}} = \frac{\omega_{P1}}{5 \times 5630} = \frac{19k\frac{rad}{sec}}{5 \times 5630} = 0.675\frac{rad}{sec}$$

Solve for C_{CMP}:

$$C_{CMP} = \frac{1}{\omega_{P2} \times 5e^6\Omega} = \frac{1}{0.675\frac{rad}{sec} \times 5e^6\Omega} = 0.30\,\mu F$$

To attenuate switching noise, calculate ω_{P3}:

$$\omega_{P3} = (\max\ \omega_{P1}, \omega_{Z1}) \times 10 = \omega_{Z1} \times 10$$

$$\omega_{P3} = 36k\frac{rad}{sec} \times 10 = 360k\frac{rad}{sec}$$

Assume R_{FS} = 10Ω and solve for C_{FS}:

$$C_{FS} = \frac{1}{10\Omega \times \omega_{P3}} = \frac{1}{10\Omega \times 360k\frac{rad}{sec}} = 0.28\,\mu F$$

The chosen components from step 7 are:

$$\boxed{\begin{aligned} C_{CMP} &= 0.33\,\mu F \\ R_{FS} &= 10\Omega \\ C_{FS} &= 0.27\,\mu F \end{aligned}}$$

8. INPUT CAPACITANCE

Solve for the minimum C_{IN}:

$$C_{IN} = \frac{I_{LED} \times D}{\Delta v_{IN\text{-}PP} \times f_{SW}} = \frac{1A \times 0.467}{100\,mV \times 504\,kHz} = 9.27\,\mu F$$

To minimize power supply interaction a 200% larger capacitance of approximately 20 µF is used, therefore the actual $\Delta v_{IN\text{-}PP}$ is much lower. Since high voltage ceramic capacitor selection is limited, four 4.7 µF X7R capacitors are chosen.

Determine minimum allowable RMS current rating:

$$I_{IN\text{-}RMS} = I_{LED} \times \sqrt{\frac{D_{MAX}}{1 - D_{MAX}}} = 1A \times \sqrt{\frac{0.677}{1 - 0.677}} = 1.45A$$

The chosen components from step 8 are:

$$\boxed{C_{IN} = 4 \times 4.7\,\mu F}$$

9. NFET

Determine minimum Q1 voltage rating and current rating:

$$V_{T\text{-}MAX} = V_{IN\text{-}MAX} + V_O = 70V + 21V = 91V$$

$$I_{T\text{-}MAX} = \frac{0.677}{1 - 0.677} \times 1A = 2.1A$$

A 100V NFET is chosen with a current rating of 32A due to the low $R_{DS\text{-}ON}$ = 50 mΩ. Determine $I_{T\text{-}RMS}$ and P_T:

$$I_{T\text{-}RMS} = \frac{I_{LED}}{D'} \times \sqrt{D} = \frac{1A}{0.533} \times \sqrt{0.467} = 1.28A$$

$$P_T = I_{T\text{-}RMS}^2 \times R_{DSON} = 1.28A^2 \times 50\,m\Omega = 82\,mW$$

The chosen component from step 9 is:

$$\boxed{Q1 \rightarrow 32A,\ 100V,\ DPAK}$$

10. DIODE

Determine minimum D1 voltage rating and current rating:

$$V_{RD\text{-}MAX} = V_{IN\text{-}MAX} + V_O = 70V + 21V = 91V$$

$$I_{D\text{-}MAX} = I_{LED} = 1A$$

A 100V diode is chosen with a current rating of 12A and V_D = 600 mV. Determine P_D:

$$P_D = I_D \times V_{FD} = 1A \times 600\,mV = 600\,mW$$

The chosen component from step 10 is:

D1 → 12A, 100V, DPAK

11. INPUT UVLO

Solve for R_{UV2}:

$$R_{UV2} = \frac{V_{HYS}}{23\,\mu A} = \frac{3V}{23\,\mu A} = 130\,k\Omega$$

The closest standard resistor is 130 kΩ therefore V_{HYS} is:

$$V_{HYS} = R_{UV2} \times 23\,\mu A = 130\,k\Omega \times 23\,\mu A = 2.99V$$

Solve for R_{UV1}:

$$R_{UV1} = \frac{1.24V \times R_{UV2}}{V_{TURN\text{-}ON} - 1.24V} = \frac{1.24V \times 130\,k\Omega}{10V - 1.24V} = 18.4\,k\Omega$$

The closest standard resistor is 18.2 kΩ making $V_{TURN\text{-}ON}$:

$$V_{TURN\text{-}ON} = \frac{1.24V \times (R_{UV1} + R_{UV2})}{R_{UV1}}$$

$$V_{TURN\text{-}ON} = \frac{1.24V \times (18.2\,k\Omega + 130\,k\Omega)}{18.2\,k\Omega} = 10.1V$$

The chosen components from step 11 are:

$R_{UV1} = 18.2\,k\Omega$
$R_{UV2} = 130\,k\Omega$

12. OUTPUT OVLO

Solve for R_{OV2}:

$$R_{OV2} = \frac{V_{HYSO}}{23\,\mu A} = \frac{10V}{23\,\mu A} = 435\,k\Omega$$

The closest standard resistor is 432 kΩ therefore V_{HYSO} is:

$$V_{HYSO} = R_{OV2} \times 23\,\mu A = 432\,k\Omega \times 23\,\mu A = 9.94V$$

Solve for R_{OV1}:

$$R_{OV1} = \frac{1.24V \times R_{OV2}}{V_{TURN\text{-}OFF} - 0.62V} = \frac{1.24V \times 432\,k\Omega}{40V - 0.62V} = 13.6\,k\Omega$$

The closest standard resistor is 13.7 kΩ making $V_{TURN\text{-}OFF}$:

$$V_{TURN\text{-}OFF} = \frac{1.24V \times (0.5 \times R_{OV1} + R_{OV2})}{R_{OV1}}$$

$$V_{TURN\text{-}OFF} = \frac{1.24V \times (0.5 \times 13.7\,k\Omega + 432\,k\Omega)}{13.7\,k\Omega} = 39.7V$$

The chosen components from step 12 are:

$R_{OV1} = 13.7\,k\Omega$
$R_{OV2} = 432\,k\Omega$

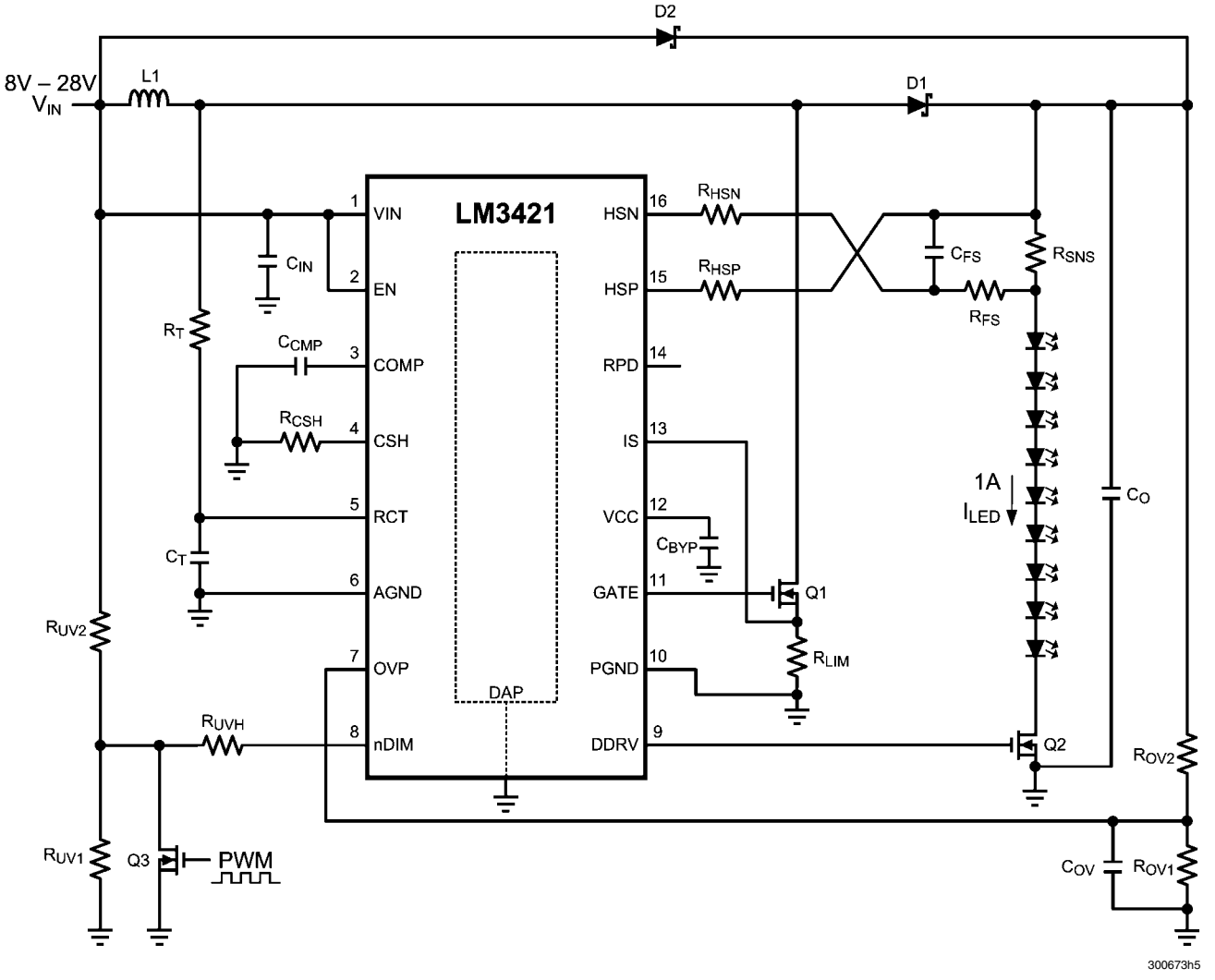
DESIGN #1 Bill of Materials

Qty	Part ID	Part Value	Manufacturer	Part Number
1	LM3421	Buck-boost controller	NSC	LM3421MH
1	C_{BYP}	2.2 µF X7R 10% 16V	MURATA	GRM21BR71C225KA12L
1	C_{CMP}	0.33 µF X7R 10% 25V	MURATA	GRM21BR71E334KA01L
1	C_{FS}	0.27 µF X7R 10% 25V	MURATA	GRM21BR71E274KA01L
4	C_{IN}	4.7 µF X7R 10% 100V	TDK	C5750X7R2A475K
4	C_O	10 µF X7R 10% 50V	TDK	C4532X7R1H106K
1	C_{OV}	47 pF COG/NPO 5% 50V	AVX	08055A470JAT2A
1	C_T	1000 pF COG/NPO 5% 50V	MURATA	GRM2165C1H102JA01D
1	D1	Schottky 100V 12A	VISHAY	12CWQ10FNPBF
1	L1	33 µH 20% 6.3A	COILCRAFT	MSS1278-333MLB
1	Q1	NMOS 100V 32A	FAIRCHILD	FDD3682
1	Q2	PNP 150V 600 mA	FAIRCHILD	MMBT5401
1	R_{CSH}	12.4 kΩ 1%	VISHAY	CRCW080512K4FKEA
1	R_{FS}	10Ω 1%	VISHAY	CRCW080510R0FKEA
2	R_{HSP}, R_{HSN}	1.0 kΩ 1%	VISHAY	CRCW08051K00FKEA
1	R_{LIM}	0.04Ω 1% 1W	VISHAY	WSL2512R0400FEA
1	R_{OV1}	13.7 kΩ 1%	VISHAY	CRCW080513K7FKEA
1	R_{OV2}	432 kΩ 1%	VISHAY	CRCW0805432KFKEA
1	R_{SNS}	0.1Ω 1% 1W	VISHAY	WSL2512R1000FEA
1	R_T	49.9 kΩ 1%	VISHAY	CRCW080549K9FKEA
1	R_{UV1}	18.2 kΩ 1%	VISHAY	CRCW080518K2FKEA
1	R_{UV2}	130 kΩ 1%	VISHAY	CRCW0805130KFKEA

Applications Information

The following designs are provided as reference circuits. For a specific design, the steps in the Design Procedure section should be performed. In all designs, an RC filter (0.1 µF, 10Ω) is recommended at VIN placed as close as possible to the LM3421/23 device. This filter is not shown in the following designs.

DESIGN #2 - LM3421 BOOST Application



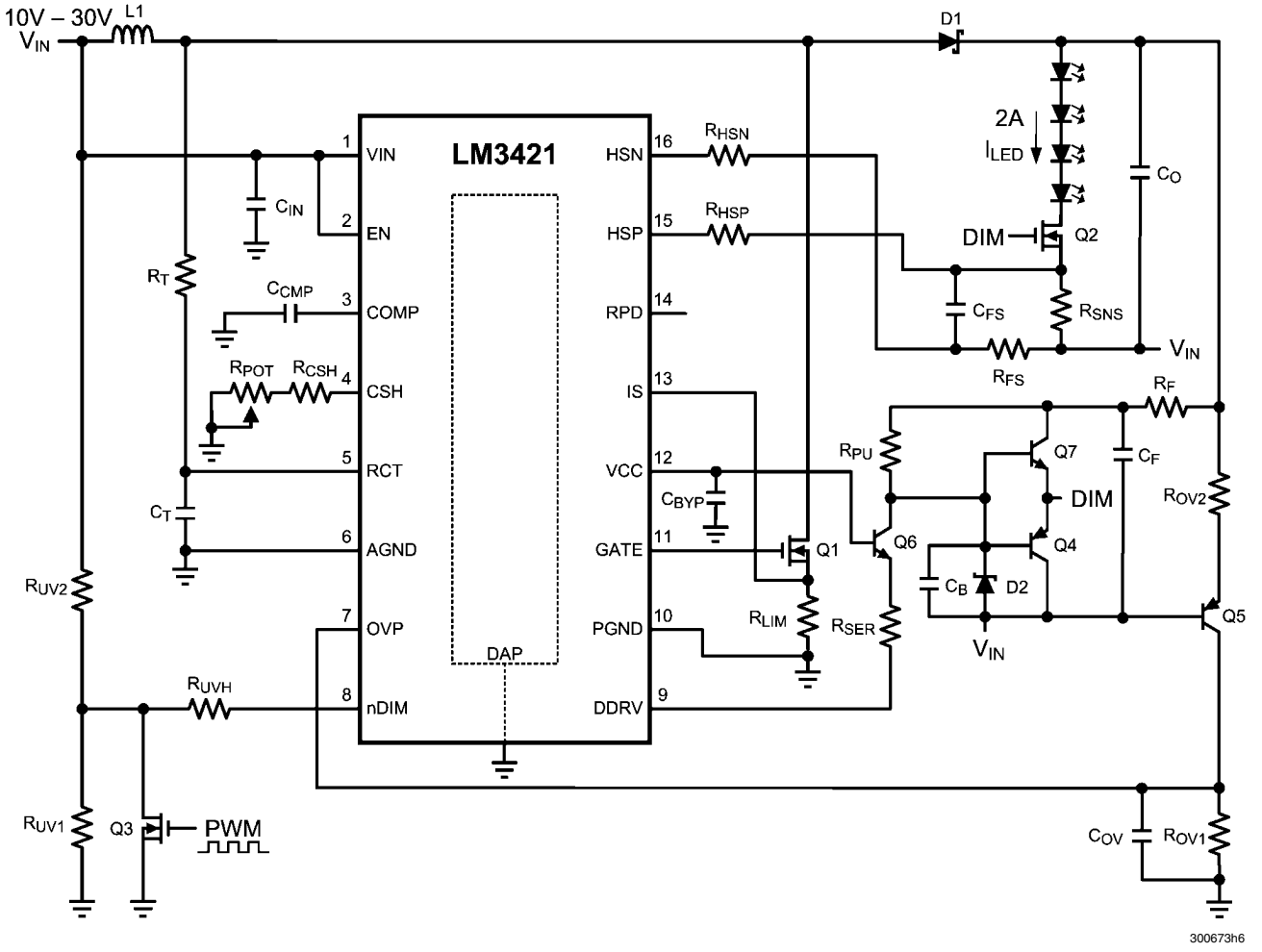
Features

- Input: 8V to 28V
- Output: 9 LEDs at 1A
- PWM Dimming up to 30kHz
- 700 kHz Switching Frequency

DESIGN #2 Bill of Materials

Qty	Part ID	Part Value	Manufacturer	Part Number
1	LM3421	Boost controller	NSC	LM3421MH
1	C_{BYP}	2.2 µF X7R 10% 16V	MURATA	GRM21BR71C225KA12L
1	C_{CMP}	0.1 µF X7R 10% 25V	MURATA	GRM21BR71E104KA01L
0	C_{FS}	DNP		
4	C_{IN}	4.7 µF X7R 10% 100V	TDK	C5750X7R2A475K
4	C_O	10 µF X7R 10% 50V	TDK	C4532X7R1H106K
1	C_{OV}	47 pF COG/NPO 5% 50V	AVX	08055A470JAT2A
1	C_T	1000 pF COG/NPO 5% 50V	MURATA	GRM2165C1H102JA01D
2	D1, D2	Schottky 60V 5A	COMCHIP	CDBC560-G
1	L1	33 µH 20% 6.3A	COILCRAFT	MSS1278-333MLB
2	Q1, Q2	NMOS 60V 8A	VISHAY	SI4436DY
1	Q3	NMOS 60V 115mA	ON-SEMI	2N7002ET1G
2	R_{CSH}, R_{OV1}	12.4 kΩ 1%	VISHAY	CRCW080512K4FKEA
1	R_{FS}	0Ω 1%	VISHAY	CRCW08050000Z0EA
2	R_{HSP}, R_{HSN}	1.0 kΩ 1%	VISHAY	CRCW08051K00FKEA
1	R_{LIM}	0.06Ω 1% 1W	VISHAY	WSL2512R0600FEA
1	R_{OV2}	499 kΩ 1%	VISHAY	CRCW0805499KFKEA
1	R_{SNS}	0.1Ω 1% 1W	VISHAY	WSL2512R1000FEA
1	R_{UV2}	10.0 kΩ 1%	VISHAY	CRCW080510K0FKEA
1	R_T	35.7 kΩ 1%	VISHAY	CRCW080535K7FKEA
1	R_{UV1}	1.82 kΩ 1%	VISHAY	CRCW08051K82FKEA
1	R_{UVH}	17.8 kΩ 1%	VISHAY	CRCW080517K8FKEA

DESIGN #3 - LM3421 BUCK-BOOST Application



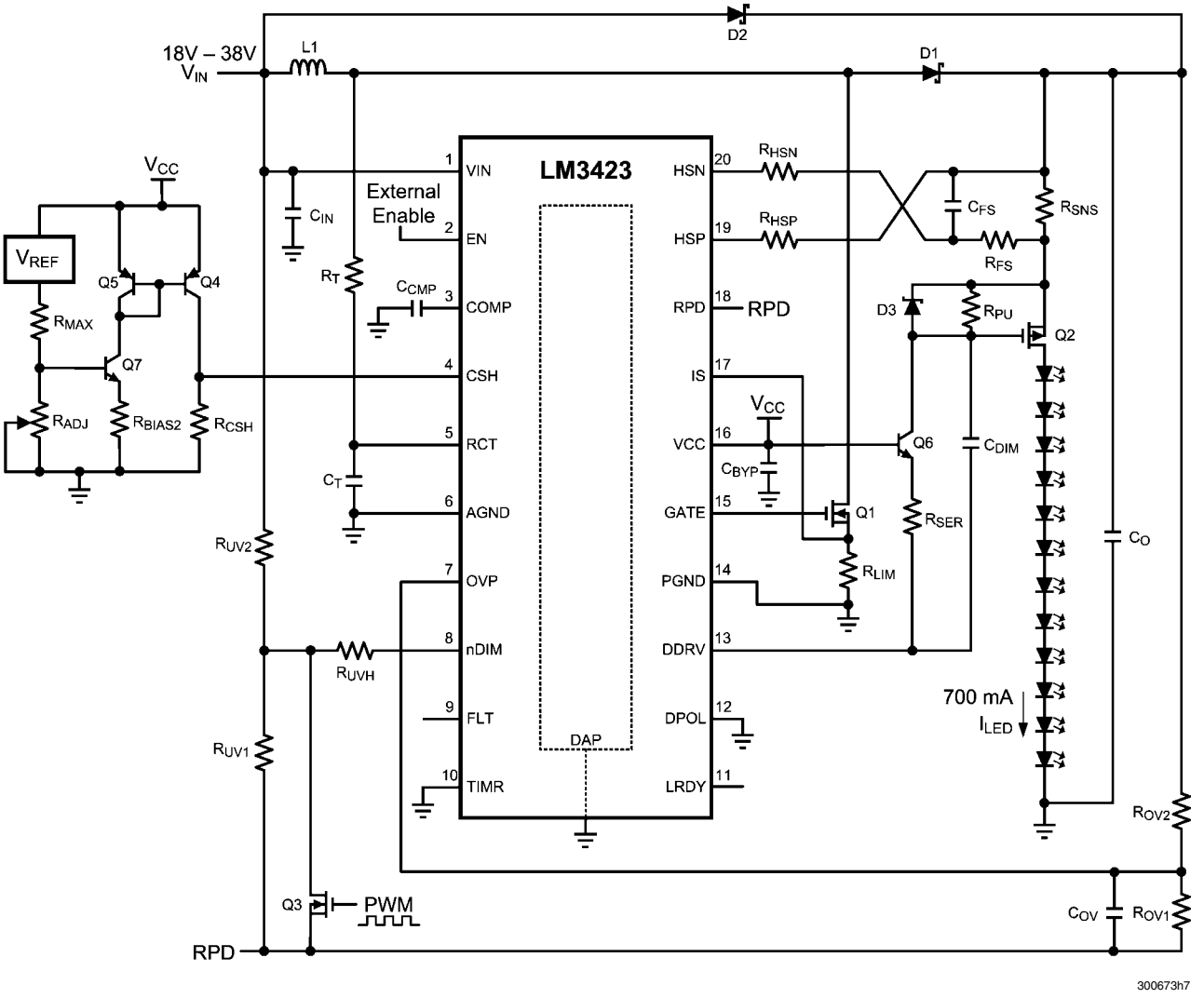
Features

- Input: 10V to 30V
- Output: 4 LEDs at 2A
- PWM Dimming up to 10kHz
- Analog Dimming
- 600 kHz Switching Frequency

DESIGN #3 Bill of Materials

Qty	Part ID	Part Value	Manufacturer	Part Number
1	LM3421	Buck-boost controller	NSC	LM3421MH
1	C_B	100 pF COG/NPO 5% 50V	MURATA	GRM2165C1H101JA01D
1	C_{BYP}	2.2 µF X7R 10% 16V	MURATA	GRM21BR71C225KA12L
3	C_{CMP}, C_{REF}, C_{SS}	1 µF X7R 10% 25V	MURATA	GRM21BR71E105KA01L
1	C_F	0.1 µF X7R 10% 25V	MURATA	GRM21BR71E104KA01L
0	C_{FS}	DNP		
4	C_{IN}	6.8 µF X7R 10% 50V	TDK	C5750X7R1H685K
4	C_O	10 µF X7R 10% 50V	TDK	C4532X7R1H106K
1	C_{OV}	47 pF COG/NPO 5% 50V	AVX	08055A470JAT2A
1	C_T	1000 pF COG/NPO 5% 50V	MURATA	GRM2165C1H102JA01D
1	D1	Schottky 100V 12A	VISHAY	12CWQ10FNPBF
1	D2	Zener 10V 500mA	ON-SEMI	BZX84C10LT1G
1	L1	22 µH 20% 7.2A	COILCRAFT	MSS1278-223MLB
2	Q1, Q2	NMOS 60V 8A	VISHAY	SI4436DY
1	Q3	NMOS 60V 260mA	ON-SEMI	2N7002ET1G
1	Q4	PNP 40V 200 mA	FAIRCHILD	MMBT5087
1	Q5	PNP 150V 600 mA	FAIRCHILD	MMBT5401
1	Q6	NPN 300V 600 mA	FAIRCHILD	MMBTA42
1	Q7	NPN 40V 200 mA	FAIRCHILD	MMBT6428
1	R_{CSH}	12.4 kΩ 1%	VISHAY	CRCW080512K4FKEA
1	R_F	10Ω 1%	VISHAY	CRCW080510R0FKEA
1	R_{FS}	0Ω 1%	VISHAY	CRCW08050000Z0EA
1	R_{UV2}	10.0 kΩ 1%	VISHAY	CRCW080510K0FKEA
2	R_{HSP}, R_{HSN}	1.0 kΩ 1%	VISHAY	CRCW08051K00FKEA
1	R_{LIM}	0.04Ω 1% 1W	VISHAY	WSL2512R0400FEA
1	R_{OV1}	18.2 kΩ 1%	VISHAY	CRCW080518K2FKEA
1	R_{OV2}	499 kΩ 1%	VISHAY	CRCW0805499KFKEA
1	R_{POT}	1 MΩ potentiometer	BOURNS	3352P-1-105
1	R_{PU}	4.99 kΩ 1%	VISHAY	CRCW08054K99FKEA
1	R_{SER}	499Ω 1%	VISHAY	CRCW0805499RFKEA
1	R_{SNS}	0.05Ω 1% 1W	VISHAY	WSL2512R0500FEA
1	R_T	41.2 kΩ 1%	VISHAY	CRCW080541K2FKEA
1	R_{UV1}	1.43 kΩ 1%	VISHAY	CRCW08051K43FKEA
1	R_{UVH}	17.4 kΩ 1%	VISHAY	CRCW080517K4FKEA

DESIGN #4 - LM3423 BOOST Application



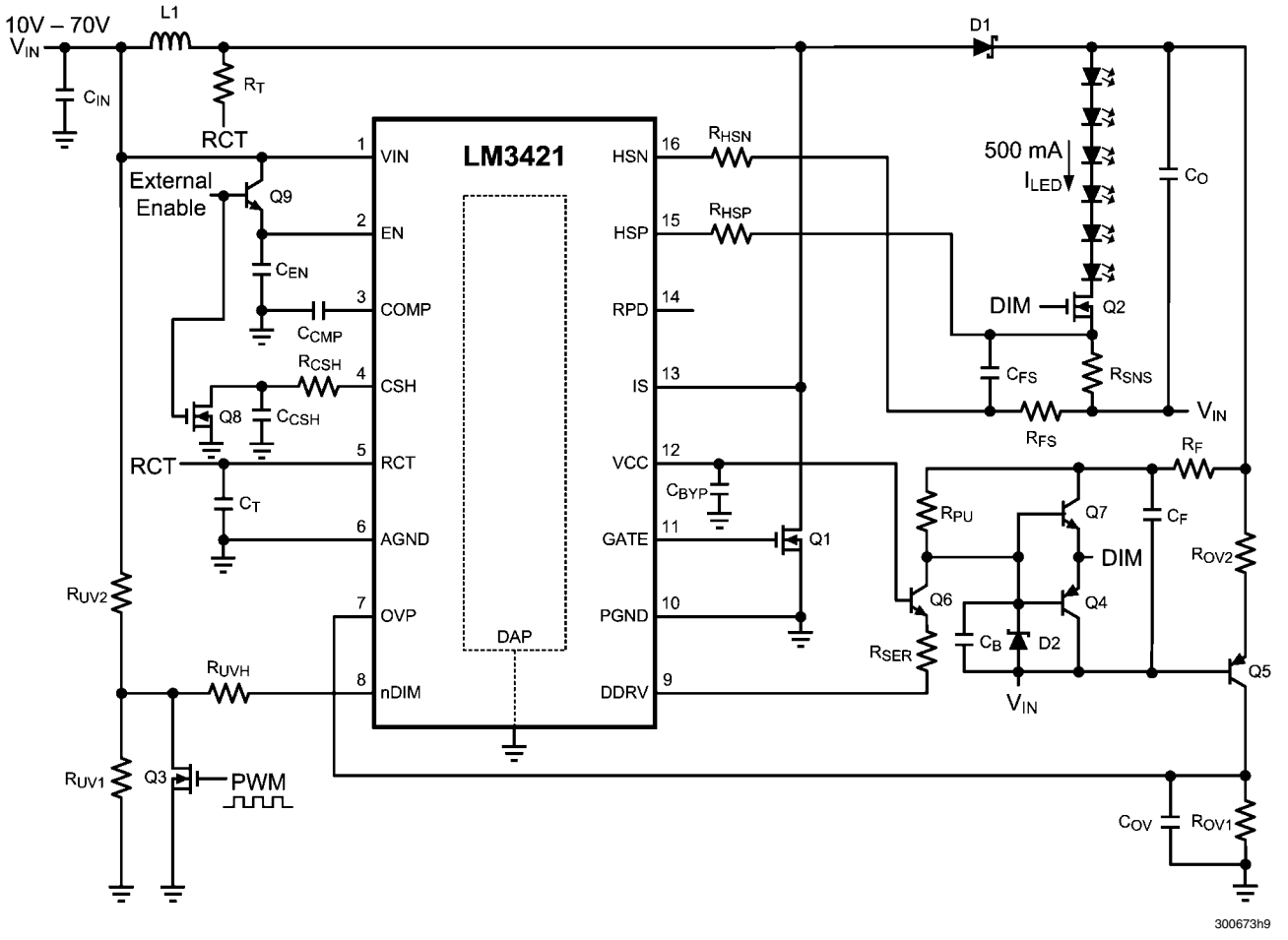
Features

- Input: 18V to 38V
- Output: 12 LEDs at 700mA
- High Side PWM Dimming up to 30 kHz
- Analog Dimming
- Zero Current Shutdown
- 700 kHz Switching Frequency

DESIGN #4 Bill of Materials

Qty	Part ID	Part Value	Manufacturer	Part Number
1	LM3423	Boost controller	NSC	LM3423MH
1	C_{BYP}	2.2 µF X7R 10% 16V	MURATA	GRM21BR71C225KA12L
1	C_{CMP}	1 µF X7R 10% 25V	MURATA	GRM21BR71E105KA01L
1	C_{FS}	0.1 µF X7R 10% 25V	MURATA	GRM21BR71E104KA01L
4	C_{IN}	4.7 µF X7R 10% 100V	TDK	C5750X7R2A475K
4	C_O	10 µF X7R 10% 50V	TDK	C4532X7R1H106K
1	C_{OV}	47 pF COG/NPO 5% 50V	AVX	08055A470JAT2A
1	C_T	1000 pF COG/NPO 5% 50V	MURATA	GRM2165C1H102JA01D
2	D1, D2	Schottky 60V 5A	COMCHIP	CDBC560-G
1	D3	Zener 10V 500mA	ON-SEMI	BZX84C10LT1G
1	L1	47 µH 20% 5.3A	COILCRAFT	MSS1278-473MLB
1	Q1	NMOS 60V 8A	VISHAY	SI4436DY
1	Q2	PMOS 70V 5.7A	ZETEX	ZXMP7A17K
1	Q3	NMOS 60V 260mA	ON-SEMI	2N7002ET1G
1	Q4, Q5 (dual pack)	Dual PNP 40V 200mA	FAIRCHILD	FFB3906
1	Q6	NPN 300V 600mA	FAIRCHILD	MMBTA42
1	Q7	NPN 40V 200 mA	FAIRCHILD	MMBT3904
1	R_{ADJ}	100 kΩ potentiometer	BOURNS	3352P-1-104
1	R_{BIAS2}	17.4 kΩ 1%	VISHAY	CRCW080517K4FKEA
2	R_{CSH}, R_{OV1}	12.4 kΩ 1%	VISHAY	CRCW080512K4FKEA
1	R_{FS}	10Ω 1%	VISHAY	CRCW080510R0FKEA
3	R_{HSP}, R_{HSN}, R_{MAX}	1.0 kΩ 1%	VISHAY	CRCW08051K00FKEA
1	R_{LIM}	0.06Ω 1% 1W	VISHAY	WSL2512R0600FEA
1	R_{OV2}	499 kΩ 1%	VISHAY	CRCW0805499KFKEA
1	R_{SNS}	0.15Ω 1% 1W	VISHAY	WSL2512R1500FEA
1	R_T	35.7 kΩ 1%	VISHAY	CRCW080535K7FKEA
1	R_{UV1}	1.43 kΩ 1%	VISHAY	CRCW08051K43FKEA
1	R_{UV2}	10.0 kΩ 1%	VISHAY	CRCW080510K0FKEA
1	R_{UVH}	16.9 kΩ 1%	VISHAY	CRCW080516K9FKEA

DESIGN #5 - LM3421 BUCK-BOOST Application



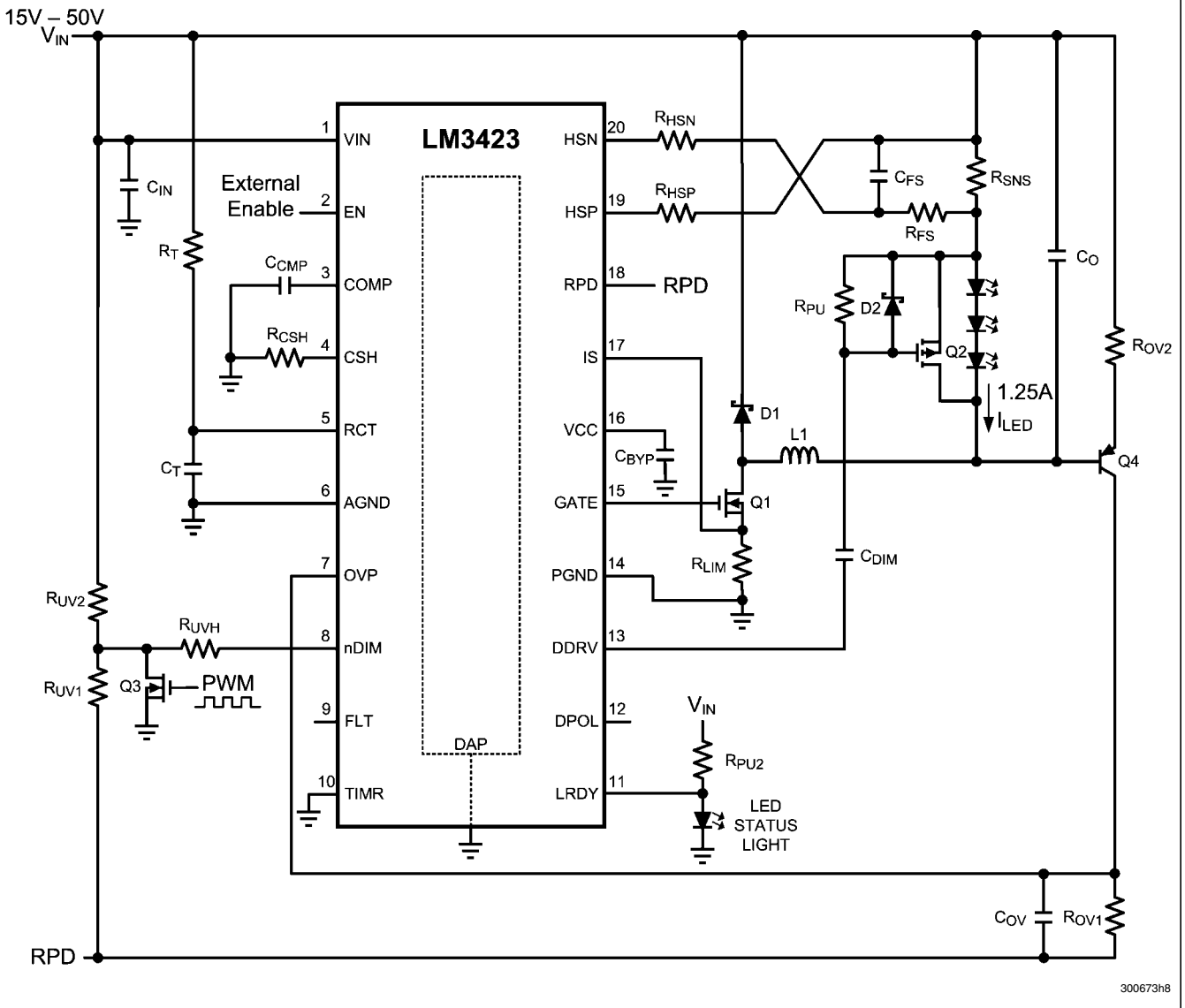
Features

- Input: 10V to 70V
- Output: 6 LEDs at 500mA
- PWM Dimming up to 10 kHz
- Slow Fade Out
- MosFET R_{DS-ON} Sensing
- 700 kHz Switching Frequency

DESIGN #5 Bill of Materials

Qty	Part ID	Part Value	Manufacturer	Part Number
1	LM3421	Buck-boost controller	NSC	LM3421MH
1	C_B	100 pF COG/NPO 5% 50V	MURATA	GRM2165C1H101JA01D
1	C_{BYP}	2.2 µF X7R 10% 16V	MURATA	GRM21BR71C225KA12L
1	C_{CMP}	1 µF X7R 10% 25V	MURATA	GRM21BR71E105KA01L
1	C_F	0.1 µF X7R 10% 25V	MURATA	GRM21BR71E104KA01L
0	C_{FS}	DNP		
4	C_{IN}	4.7 µF X7R 10% 100V	TDK	C5750X7R2A475K
4	C_O	10 µF X7R 10% 50V	TDK	C4532X7R1H106K
1	C_{OV}	47 pF COG/NPO 5% 50V	AVX	08055A470JAT2A
1	C_T	1000 pF COG/NPO 5% 50V	MURATA	GRM2165C1H102JA01D
1	D1	Schottky 100V 12A	VISHAY	12CWQ10FNPBF
1	D2	Zener 10V 500mA	ON-SEMI	BZX84C10LT1G
1	L1	68 µH 20% 4.3A	COILCRAFT	MSS1278-683MLB
2	Q1, Q2	NMOS 100V 32A	FAIRCHILD	FDD3682
1	Q3	NMOS 60V 260mA	ON-SEMI	2N7002ET1G
2	Q4, Q8	PNP 40V 200mA	FAIRCHILD	MMBT5087
1	Q5	PNP 150V 600 mA	FAIRCHILD	MMBT5401
1	Q6	NPN 300V 600mA	FAIRCHILD	MMBTA42
2	Q7, Q9	NPN 40V 200mA	FAIRCHILD	MMBT6428
1	R_{CSH}	12.4 kΩ 1%	VISHAY	CRCW080512K4FKEA
1	R_{FS}	0Ω 1%	VISHAY	CRCW08050000Z0EA
1	R_{UV2}	10.0 kΩ 1%	VISHAY	CRCW080510K0FKEA
2	R_{HSP}, R_{HSN}	1.0 kΩ 1%	VISHAY	CRCW08051K00FKEA
1	R_{OV1}	15.8 kΩ 1%	VISHAY	CRCW080515K8FKEA
1	R_{OV2}	499 kΩ 1%	VISHAY	CRCW0805499KFKEA
1	R_{PU}	4.99 kΩ 1%	VISHAY	CRCW08054K99FKEA
1	R_{SER}	499Ω 1%	VISHAY	CRCW0805499RFKEA
1	R_{SNS}	0.2Ω 1% 1W	VISHAY	WSL2512R2000FEA
1	R_T	35.7 kΩ 1%	VISHAY	CRCW080535K7FKEA
1	R_{UV1}	1.43 kΩ 1%	VISHAY	CRCW08051K43FKEA
1	R_{UVH}	17.4 kΩ 1%	VISHAY	CRCW080517K4FKEA

DESIGN #6 - LM3423 BUCK Application



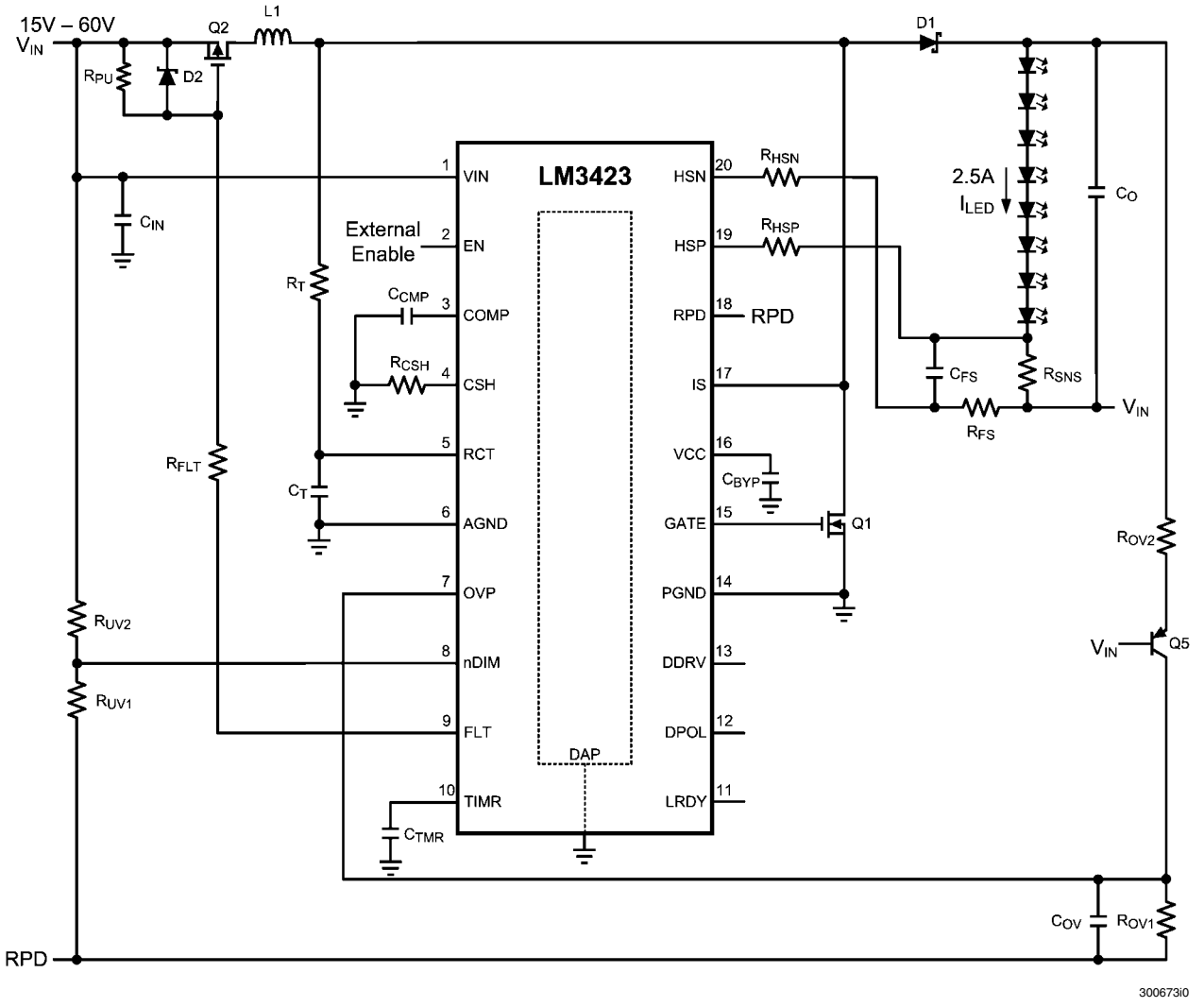
Features

- Input: 15V to 50V
- Output: 3 LEDs at 1.25A
- PWM Dimming up to 50 kHz
- LED Status Indicator
- Zero Current Shutdown
- 700 kHz Switching Frequency

DESIGN #6 Bill of Materials

Qty	Part ID	Part Value	Manufacturer	Part Number
1	LM3423	Buck controller	NSC	LM3423MH
1	C_{BYP}	2.2 µF X7R 10% 16V	MURATA	GRM21BR71C225KA12L
2	C_{CMP}, C_{DIM}	0.1 µF X7R 10% 25V	MURATA	GRM21BR71E104KA01L
0	C_{FS}	DNP		
4	C_{IN}	4.7 µF X7R 10% 100V	TDK	C5750X7R2A475K
0	C_O	DNP		
1	C_{OV}	47 pF COG/NPO 5% 50V	AVX	08055A470JAT2A
1	C_T	1000 pF COG/NPO 5% 50V	MURATA	GRM2165C1H102JA01D
1	D1	Schottky 100V 12A	VISHAY	12CWQ10FNPBF
1	D2	Zener 10V 500mA	ON-SEMI	BZX84C10LT1G
1	L1	22 µH 20% 7.3A	COILCRAFT	MSS1278-223MLB
1	Q1	NMOS 60V 8A	VISHAY	SI4436DY
1	Q2	PMOS 30V 6.2A	VISHAY	SI3483DV
1	Q3	NMOS 60V 115mA	ON-SEMI	2N7002ET1G
1	Q4	PNP 150V 600 mA	FAIRCHILD	MMBT5401
1	R_{CSH}	12.4 kΩ 1%	VISHAY	CRCW080512K4FKEA
1	R_{FS}	0Ω 1%	VISHAY	CRCW08050000OZEA
2	R_{HSP}, R_{HSN}	1.0 kΩ 1%	VISHAY	CRCW08051K00FKEA
1	R_{LIM}	0.04Ω 1% 1W	VISHAY	WSL2512R0400FEA
1	R_{OV1}	21.5 kΩ 1%	VISHAY	CRCW080521K5FKEA
1	R_{OV2}	499 kΩ 1%	VISHAY	CRCW0805499KFKEA
3	R_{PU}, R_{PU2}, R_{UV2}	100 kΩ 1%	VISHAY	CRCW0805100KFKEA
1	R_T	35.7 kΩ 1%	VISHAY	CRCW080535K7FKEA
1	R_{SNS}	0.08Ω 1% 1W	VISHAY	WSL2512R0800FEA
1	R_{UV1}	11.5 kΩ 1%	VISHAY	CRCW080511K5FKEA

DESIGN #7 - LM3423 BUCK-BOOST Application



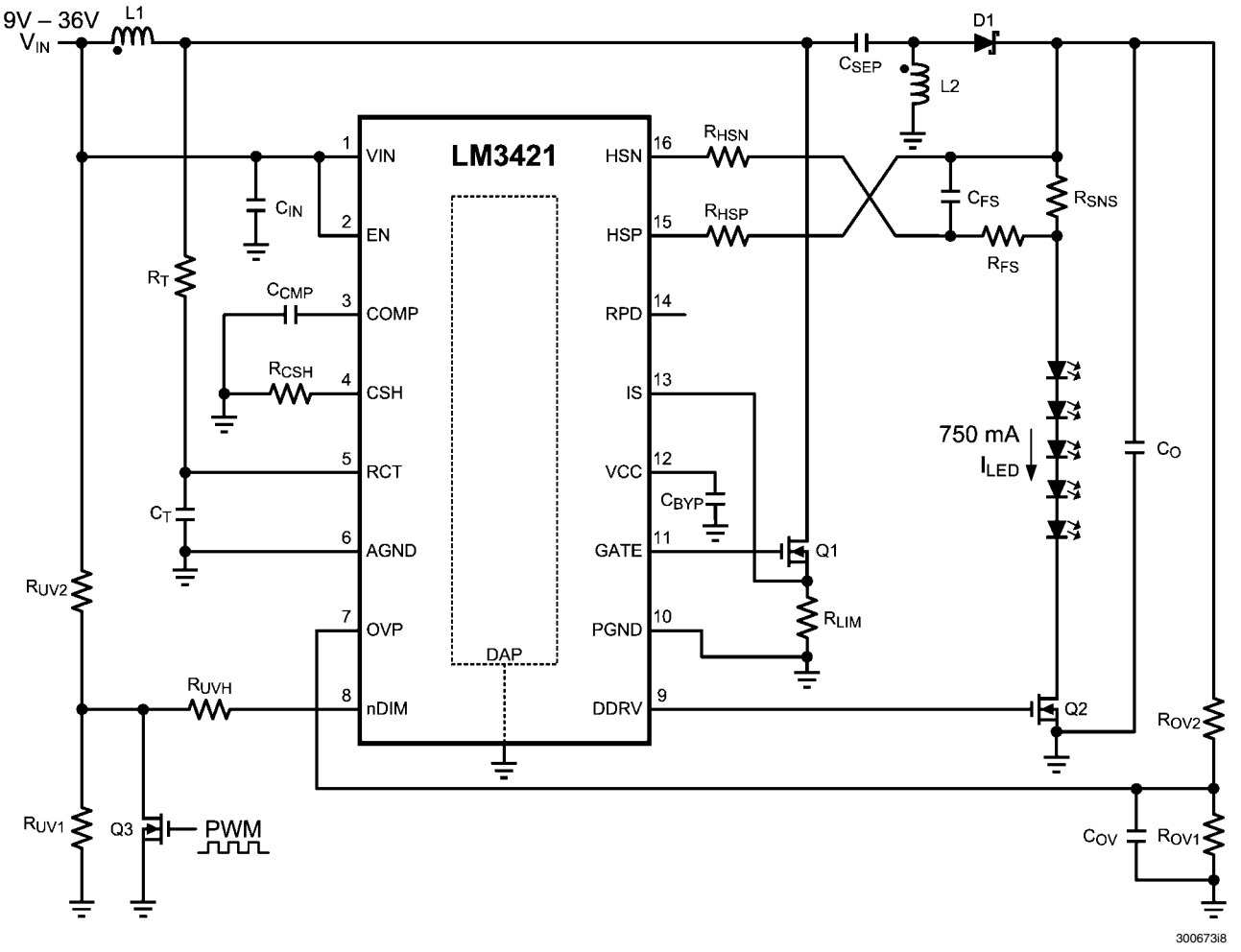
Features

- Input: 15V to 60V
- Output: 8 LEDs at 2.5A
- Fault Input Disconnect
- Zero Current Shutdown
- 500 kHz Switching Frequency

DESIGN #7 Bill of Materials

Qty	Part ID	Part Value	Manufacturer	Part Number
1	LM3423	Buck-boost controller	NSC	LM3423MH
1	C_{BYP}	2.2 µF X7R 10% 16V	MURATA	GRM21BR71C225KA12L
1	C_{CMP}	0.33 µF X7R 10% 25V	MURATA	GRM21BR71E334KA01L
1	C_{FS}	0.1 µF X7R 10% 25V	MURATA	GRM21BR71E104KA01L
4	C_{IN}	4.7 µF X7R 10% 100V	TDK	C5750X7R2A475K
4	C_O	10 µF X7R 10% 50V	TDK	C4532X7R1H106K
1	C_{OV}	47 pF COG/NPO 5% 50V	AVX	08055A470JAT2A
1	C_T	1000 pF COG/NPO 5% 50V	MURATA	GRM2165C1H102JA01D
1	C_{TMR}	220 pF COG/NPO 5% 50V	MURATA	GRM2165C1H221JA01D
1	D1	Schottky 100V 12A	VISHAY	12CWQ10FNPBF
1	D2	Zener 10V 500mA	ON-SEMI	BZX84C10LT1G
1	L1	22 µH 20% 7.2A	COILCRAFT	MSS1278-223MLB
1	Q1	NMOS 100V 32A	FAIRCHILD	FDD3682
1	Q2	PMOS 70V 5.7A	ZETEX	ZXMP7A17K
1	Q5	PNP 150V 600 mA	FAIRCHILD	MMBT5401
2	R_{CSH}, R_{OV1}	12.4 kΩ 1%	VISHAY	CRCW080512K4FKEA
1	R_{FS}	10Ω 1%	VISHAY	CRCW080510R0FKEA
2	R_{FLT}, R_{PU2}	100 kΩ 1%	VISHAY	CRCW0805100KFKEA
2	R_{HSP}, R_{HSN}	1.0 kΩ 1%	VISHAY	CRCW08051K00FKEA
2	R_{LIM}, R_{SNS}	0.04Ω 1% 1W	VISHAY	WSL2512R0400FEA
1	R_{OV1}	15.8 kΩ 1%	VISHAY	CRCW080515K8FKEA
1	R_{OV2}	499 kΩ 1%	VISHAY	CRCW0805499KFKEA
1	R_T	49.9 kΩ 1%	VISHAY	CRCW080549K9FKEA
1	R_{UV1}	13.7 kΩ 1%	VISHAY	CRCW080513K7FKEA
1	R_{UV2}	150 kΩ 1%	VISHAY	CRCW0805150KFKEA

DESIGN #8 - LM3421 SEPIC Application



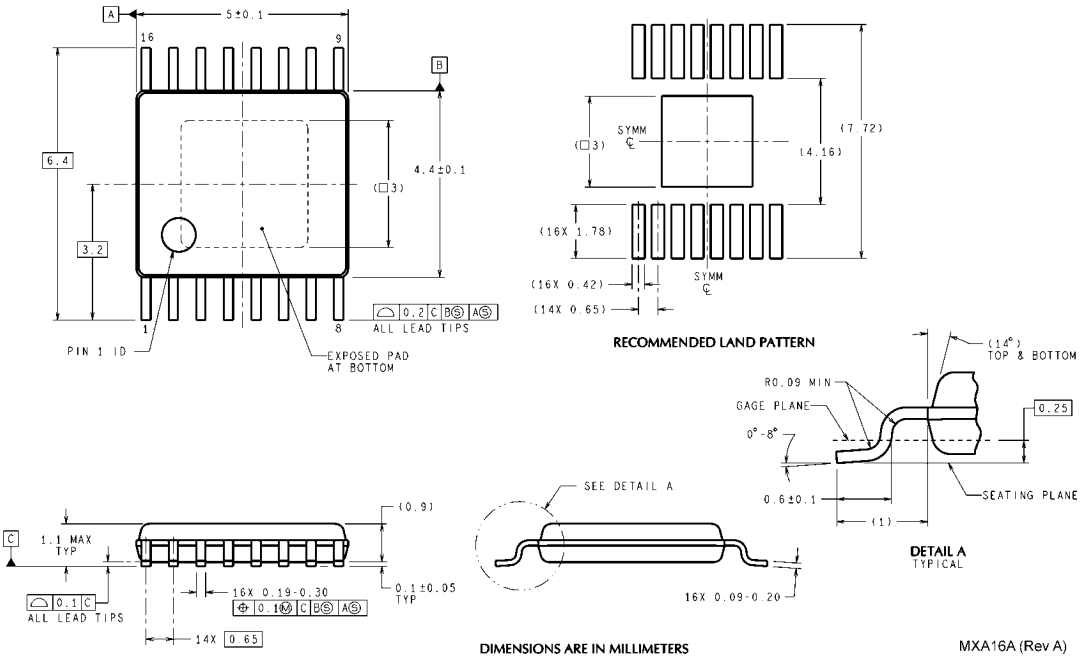
Features

- Input: 9V to 36V
- Output: 5 LEDs at 750mA
- PWM Dimming up to 30 kHz
- 500 kHz Switching Frequency

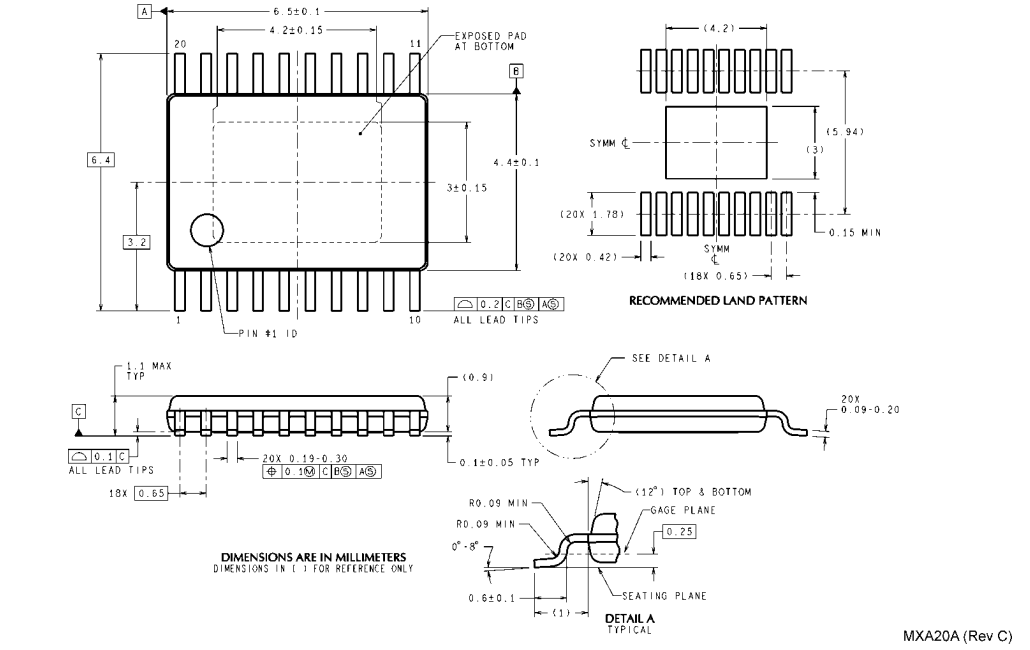
DESIGN #8 Bill of Materials

Qty	Part ID	Part Value	Manufacturer	Part Number
1	LM3421	SEPIC controller	NSC	LM3421MH
1	C_{BYP}	2.2 µF X7R 10% 16V	MURATA	GRM21BR71C225KA12L
1	C_{CMP}	0.47 µF X7R 10% 25V	MURATA	GRM21BR71E474KA01L
0	C_{FS}	DNP		
4	C_{IN}	4.7 µF X7R 10% 100V	TDK	C5750X7R2A475K
4	C_{O}	10 µF X7R 10% 50V	TDK	C4532X7R1H106K
1	C_{SEP}	1.0 µF X7R 10% 100V	TDK	C4532X7R2A105K
1	C_{OV}	47 pF COG/NPO 5% 50V	AVX	08055A470JAT2A
1	C_{T}	1000 pF COG/NPO 5% 50V	MURATA	GRM2165C1H102JA01D
1	D1	Schottky 60V 5A	COMCHIP	CDBC560-G
2	L1, L2	68 µH 20% 4.3A	COILCRAFT	DO3340P-683
2	Q1, Q2	NMOS 60V 8A	VISHAY	SI4436DY
1	Q3	NMOS 60V 115 mA	ON-SEMI	2N7002ET1G
1	R_{CSH}	12.4 kΩ 1%	VISHAY	CRCW080512K4FKEA
1	R_{FS}	0Ω 1%	VISHAY	CRCW08050000OZEA
2	R_{HSP}, R_{HSN}	750Ω 1%	VISHAY	CRCW0805750RFKEA
1	R_{LIM}	0.04Ω 1% 1W	VISHAY	WSL2512R0400FEA
1	R_{OV1}	15.8 kΩ 1%	VISHAY	CRCW080515K8FKEA
1	R_{OV2}	499 kΩ 1%	VISHAY	CRCW0805499KFKEA
2	R_{REF1}, R_{REF2}	49.9 kΩ 1%	VISHAY	CRCW080549K9FKEA
1	R_{SNS}	0.1Ω 1% 1W	VISHAY	WSL2512R1000FEA
1	R_{T}	49.9 kΩ 1%	VISHAY	CRCW080549K9FKEA
1	R_{UV1}	1.62 kΩ 1%	VISHAY	CRCW08051K62FKEA
1	R_{UV2}	10.0 kΩ 1%	VISHAY	CRCW080510K0FKEA
1	R_{UVH}	16.9 kΩ 1%	VISHAY	CRCW080516K9FKEA

Physical Dimensions inches (millimeters) unless otherwise noted



TSSOP-16 Pin EP Package (MXA)
For Ordering, Refer to Ordering Information Table
NS Package Number MXA16A



TSSOP-20 Pin EP Package (MXA)
For Ordering, Refer to Ordering Information Table
NS Package Number MXA20A

Notes

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