RICHTEK®

RTQ2105-QA

36V$_{IN}$, 3A, High Efficiency Synchronous Buck Converter with Low Quiescent Current

1 General Description

The RTQ2105-QA is a 3A, high-efficiency, current-mode synchronous buck converter optimized for automotive applications. The device operates with input voltages from 4V to 36V and is protected from load dump transients up to 42V, easing input surge protection design. The device can program the output voltage between 0.8V to V_{IN}. The low quiescent current design with the integrated low $R_{DS(ON)}$ power MOSFETs achieves high efficiency over the wide load range. The peak current mode control with simple external compensation allows the use of small inductors and results in fast transient response and good loop stability.

The wide switching frequency of 300kHz to 2200kHz allows for efficiency and size optimization when selecting the output filter components. The ultra-low minimum on-time enables constant-frequency operation even at very high step-down ratios. For switching noise-sensitive applications, it can be externally synchronized from 300kHz to 2200kHz. The optional spread spectrum frequency modulation further helps system designers with better EMC management.

The RTQ2105-QA offers precise constant current and constant voltage regulation. It is ideally suited for USB power delivery or charging super-capacitors. For applications that use long cables, the RTQ2105-QA offers cable drop compensation to maintain accurate regulation at the end of the long cable.

The RTQ2105-QA provides complete protection functions such as input undervoltage-lockout, output undervoltage protection, overcurrent protection, and over-temperature protection. Cycle-by-cycle current limit provides protection against shorted outputs, and soft-start eliminates input current surge during start-up. The RTQ2105-QA is available in a WET- WQFN-24SL 4x4 (W-Type) package.

The recommended junction temperature range is –40°C to 150°C.

2 Features

- **AEC-Q100 Grade 1 Qualified**
- **Wide Input Voltage Range: 4V to 36V**
- **Maximum Output Current: 3A**
- **Peak Current Mode Control**
- **Integrated 70mΩ Switch and 70mΩ Synchronous Rectifier**
- **Low Quiescent Current: 40μA**
- **Fast 60ns Minimum Switch On-Time**
- **Ultra-Short 65ns Minimum Switch Off-Time**
- **Adjustable and Synchronizable Switching Frequency: 300kHz to 2.2MHz**
- **Selectable PSM/FPWM at Light Load**
- **Optional Spread Spectrum Frequency Modulation for Low EMI**
- **Externally Adjustable Soft-Start**
- **Power-Good Indication**
- **Enable Control**
- **Adjustable Output Voltage with Cable Drop Compensation for V_{OUT} = 5V Applications**
- **Constant Current (CC) and Constant Voltage (CV) Regulation**
- **0.8V ± 1.5% CV Reference Accuracy**
- **Adjustable Current Limit**
- **Adjacent Pin-Short Protection**
- **Built-In UVLO, UVP, OTP**
- **Junction Temperature Range: –40°C to 150°C**

3 Applications

- Automotive Systems
- Car Camera Module and Car Cockpit Systems
- Connected Car Systems
- Point of Load Regulator in Distributed Power Systems
- Digital Set Top Boxes
- Broadband Communications
- USB Power Chargers

4 Ordering Information

RTQ2105 □□-QA

Grade
QA: AEC-Q100 Qualified and Screened by High Temperature

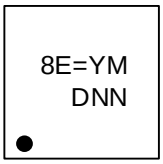
Package Type(1)
QWT WET-WQFN-24SL 4x4 (W-Type)

Lead Plating System
G: Richtek Green Policy Compliant(2)

Note 1.

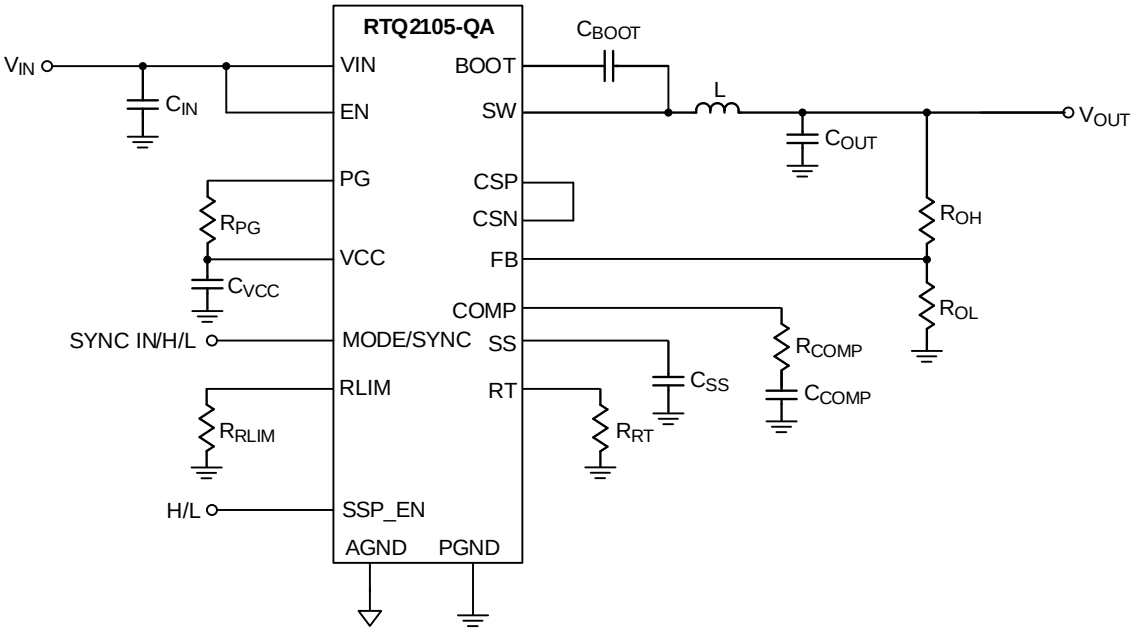
- Marked with (1) indicated: Compatible with the current requirements of IPC/JEDEC J-STD-020.
- Marked with (2) indicated: Richtek products are Richtek Green Policy compliant.

5 Marking Information



8E= : Product Code
YMDNN : Date Code

6 Simplified Application Circuit



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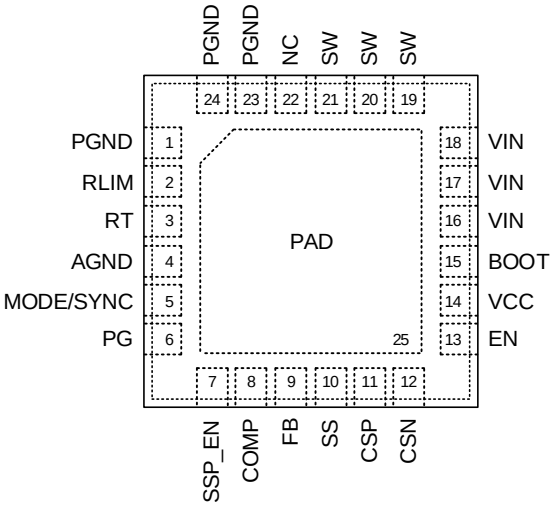
7 Pin Configuration

(TOP VIEW)



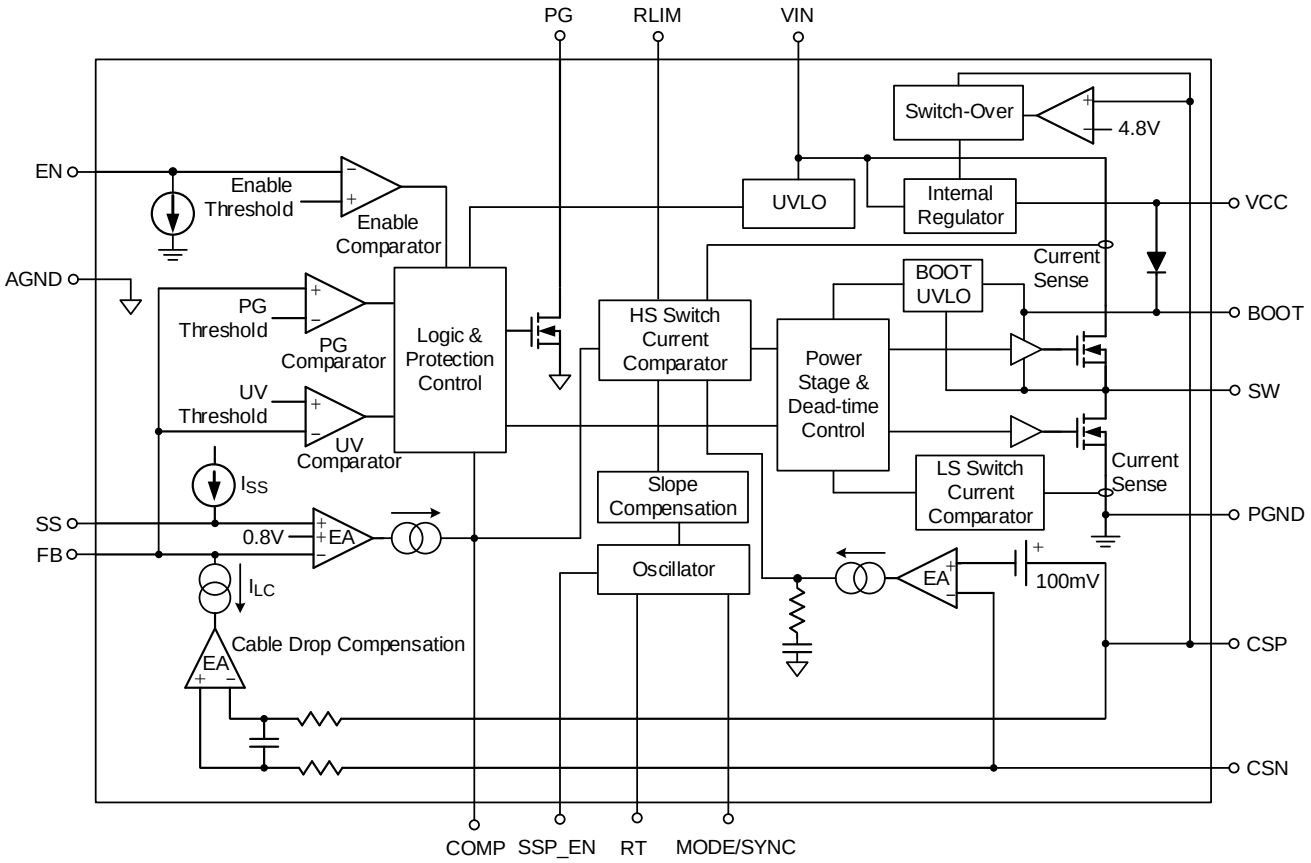
WET-WQFN-24SL 4x4

8 Functional Pin Description

Pin No.	Pin Name	Pin Function
1, 23, 24	PGND	Power ground. Connect this pin to the negative terminals of the input capacitor and output capacitor.
2	RLIM	Current limit setup pin. Connect a resistor from this pin to ground to set the current limit value. The recommended resistor value is ranging from 33kΩ (for typ. 5.5A) to 91kΩ (for typ. 2.2A).
3	RT	Oscillator frequency setup pin. Connect a resistor from this pin to ground to set the switching frequency. The recommended resistor value is ranging from 174kΩ (for typical 300kHz) to 21kΩ (for typical 2.2MHz).
4	AGND	Analog ground.
5	MODE/ SYNC	Mode selection and external synchronous signal input. Ground this pin or leave this pin floating enables the power saving mode operation at light load. Apply a DC voltage of 2V or higher, or tie to VCC for FPWM mode operation. Tie to a clock source for synchronization to an external frequency.
6	PG	Open-drain power-good indication output. Once soft-start is finished, PG will be pulled low to ground if any internal protection is triggered.
7	SSP_EN	Spread spectrum enable input. Connect this pin to VCC to enable spread spectrum. Float this pin or connect it to Ground to disable spread spectrum.
8	COMP	Compensation node. Connect external compensation elements to this pin to stabilize the control loop.
9	FB	Feedback voltage input. Connect this pin to the midpoint of the external feedback resistive divider to set the output voltage of the converter to the desired regulation level. The device regulates the FB voltage at a feedback reference voltage, typically 0.8V.
10	SS	Soft-start capacitor connection node. Connect an external capacitor between this pin and ground to set the soft-start time. Note that, for proper device operation, it is essential that the minimum soft-start time (t_{SS_min}) exceeds 500μs.
11	CSP	Current sense positive input. The CSP pin should be tied to the CSN pin and be left floating if the switch-over function is not needed.

Pin No.	Pin Name	Pin Function
12	CSN	Current sense negative input. The CSN pin should be tied to the CSP pin and be left floating if the switch-over function is not needed.
13	EN	Enable control input. A logic-high enables the converter; a logic-low forces the device into shutdown mode.
14	VCC	Linear regulator output. VCC is the output of the internal 5V linear regulator powered by VIN. Decouple with a 1μF, X7R ceramic capacitor from VCC to ground for normal operation.
15	BOOT	Bootstrap capacitor connection node to supply the high-side gate driver. Connect a 0.1μF, X7R ceramic capacitor between this pin and the SW pin.
16, 17, 18	VIN	Power input. The input voltage range is from 4V to 36V. Connect input capacitors between this pin and PGND. It is recommended to use a 4.7μF, X7R and a 0.1μF, X7R capacitors.
19, 20, 21	SW	Switch node. SW is the switching node that supplies power to the output and connects the output LC filter from SW to the output load.
22	NC	No internal connection.
25 (Exposed pad)	PAD	Exposed pad. The exposed pad is internally unconnected and must be soldered to a large PGND plane. Connect this PGND plane to other layers with thermal vias to help dissipate heat from the device.

9 Functional Block Diagram



10 Absolute Maximum Ratings

(Note 2)

- VIN Voltage, VIN ---------- –0.3V to 42V
- SW Voltage, V_{SW} ---------- –0.3V to 42V
- <50ns ---------- –5V to 46.3V
- BOOT Voltage, V_{BOOT} ---------- –0.3V to 48V
- BOOT to SW Voltage, $V_{BOOT-SW}$ ---------- –0.3V to 6V
- EN, CSP, CSN, SS Voltage ---------- –0.3V to 42V
- Other Pins ---------- –0.3V to 6V
- Lead Temperature (Soldering, 10 sec.) ---------- 260°C
- Junction Temperature ---------- 150°C
- Storage Temperature Range ---------- –65°C to 150°C

Note 2. Stresses beyond those listed under “Absolute Maximum Ratings” may cause permanent damage to the device. These are stress ratings only, and functional operation of the device at these or any other conditions beyond those indicated in the operational sections of the specifications is not implied. Exposure to absolute maximum rating conditions may affect device reliability.

11 ESD Ratings

(Note 3)

- ESD Susceptibility
  HBM ---------- 2kV

Note 3. Devices are ESD sensitive. Handling precautions are recommended.

12 Recommended Operating Conditions

(Note 4)

- Supply Input Voltage ---------- 4V to 36V
- Output Voltage Range ---------- 0.8V to VIN*

VIN*: $0.99 \times (V_{IN} - I_{OUT}(max) \times R_{DSON(max)})$

- Junction Temperature Range ---------- –40°C to 150°C
- Ambient Temperature Range ---------- –40°C to 125°C

Note 4. The device is not guaranteed to function outside its operating conditions.

13 Thermal Information

(Note 5 and Note 6)

Thermal Parameter		WET-WQFN-24SL 4x4	Unit
θ_{JA}	Junction-to-ambient thermal resistance (JEDEC standard)	38.26	°C/W
$\theta_{JC(Top)}$	Junction-to-case (top) thermal resistance	34.75	°C/W
$\theta_{JC(Bottom)}$	Junction-to-case (bottom) thermal resistance	3.32	°C/W
$\theta_{JA(EVB)}$	Junction-to-ambient thermal resistance (specific EVB)	34.7	°C/W
$\Psi_{JC(Top)}$	Junction-to-top characterization parameter	4.6	°C/W
Ψ_{JB}	Junction-to-board characterization parameter	19.2	°C/W

Note 5. For more information about thermal parameter, see the Application and Definition of Thermal Resistances report, AN061.

Note 6. $\theta_{JA(EVB)}$, $\Psi_{JC(TOP)}$, and Ψ_{JB} are measured on a high effective-thermal-conductivity four-layer test board which is in size of 70mm x 100mm, four-layer PCB with 2 oz. Cu on the outer layers and 1 oz. Cu on the inner layers. Cu. Thermal resistance/parameter values may vary depending on the PCB material, layout, and test environmental conditions.

14 Electrical Characteristics

(V_{IN} = 12V, T_J = –40°C to 125°C, unless otherwise specified.)

Parameter	Symbol	Test Conditions	Min	Typ	Max	Unit
Supply Voltage						
VIN Supply Input Voltage	V_{IN}		4	--	36	V
Undervoltage-Lockout Rising Threshold	V_{UVLO_R}	V_{IN} rising	3.6	3.8	4	V
Undervoltage-Lockout Falling Threshold	V_{UVLO_F}	V_{IN} falling	2.7	2.85	3	V
Shutdown Current	I_{SHDN}	V_{EN} = 0V	--	--	5	μA
Quiescent Current	I_Q	V_{EN} = 2V, V_{FB} = 0.82V, not switching	--	40	50	μA
Constant Voltage Regulation						
Reference Voltage for Constant Voltage regulation	V_{REF_CV}	4V < V_{IN} < 36V, PWM, T_A = T_J = 25°C	0.792	0.8	0.808	V
		4V < V_{IN} < 36V, PWM, T_J = –40°C to 125°C	0.788	0.8	0.812	
Enable Voltage						
EN Input Voltage Rising Threshold	V_{EN_R}	V_{EN} rising	1.15	1.25	1.35	V
EN Input Voltage Falling Threshold	V_{EN_F}	V_{EN} falling	0.9	1.05	1.15	
Current Limit						
High-Side Switch Current Limit 1	I_{LIM_H1}	R_{LIM} = 91kΩ	1.87	2.2	2.53	A
High-Side Switch Current Limit 2	I_{LIM_H2}	R_{LIM} = 47kΩ	3.52	4	4.48	A
High-Side Switch Current Limit 3	I_{LIM_H3}	R_{LIM} = 33kΩ	4.84	5.5	6.16	A
Negative Inductor Peak Current Limit	$I_{LIM\_PEAK\_NEG}$	From drain to source	--	2	--	A

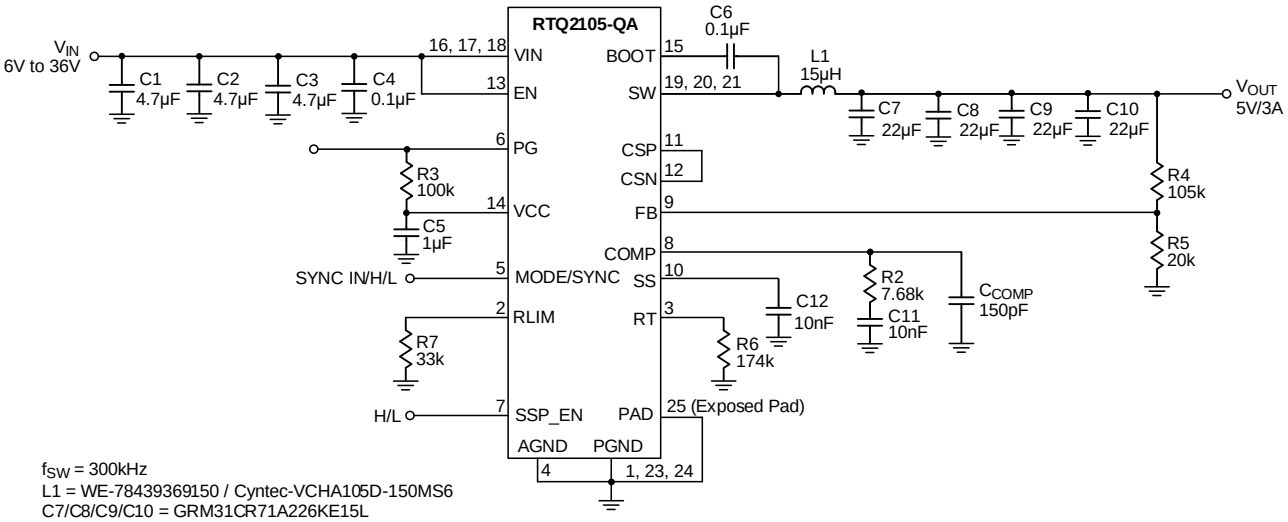
Parameter	Symbol	Test Conditions	Min	Typ	Max	Unit
Switching						
Switching Frequency 1	f_{SW1}	R_T = 174kΩ	264	300	336	kHz
Switching Frequency 2	f_{SW2}	R_T = 51kΩ	0.88	0.98	1.08	MHz
Switching Frequency 3	f_{SW3}	R_T = 21kΩ	1.98	2.2	2.42	MHz
SYNC Frequency Range		MODE/SYNC Pin = external clock	0.3	--	2.2	MHz
SYNC Switching High Threshold	V_{IH_SYNC}	MODE/SYNC Pin = external clock	--	--	2	V
SYNC Switching Low Threshold	V_{IL_SYNC}	MODE/SYNC Pin = external clock	0.4	--	--	V
SYNC Switching Clock Duty Cycle	D_{SYNC}	MODE/SYNC Pin = external clock	20	--	80	%
Minimum On-Time	t_{ON_MIN}		--	60	80	ns
Minimum Off-Time	t_{OFF_MIN}		--	65	80	ns
Internal MOSFET						
On-Resistance of High-Side MOSFET	R_{DSON_H}		--	70	130	mΩ
On-Resistance of Low-Side MOSFET	R_{DSON_L}		--	70	130	mΩ
High-Side Switch Leakage Current	I_{LK_H}	V_{EN} = 0V, V_{SW} = 0V	--	--	1	μA
Soft-Start						
Soft-Start Current	I_{SS}		4.8	6	7.2	μA
Power-Good						
Power-Good Voltage Threshold	V_{TH_PGLH1}	V_{FB} rising, % of V_{REF_CV}, PG from low to high	85	90	95	%
	V_{TH_PGHL1}	V_{FB} rising, % of V_{REF_CV}, PG from high to low	--	120	--	
	V_{TH_PGHL2}	V_{FB} falling, % of V_{REF_CV}, PG from high to low	80	85	90	
	V_{TH_PGLH2}	V_{FB} falling, % of V_{REF_CV}, PG from low to high	--	117	--	
Power-Good Leakage Current	I_{LK_PG}	PG signal good, V_{FB} = V_{REF_CV}, V_{PG} = 5.5V	--	--	0.5	μA
Power-Good Sink Current Capability	I_{SK_PG}	PG signal fault, I_{PG} sinks 2mA	--	--	0.3	V
Error Amplifier						
Error Amplifier Transconductance	gm	−10μA < I_{COMP} < 10μA	665	950	1280	μA/V
COMP to Current Sense Transconductance	gm_{CS}		4.5	5.6	6.7	A/V
Cable Drop Compensation						
Cable Drop Compensation Current	I_{LC}	V_{CSP} – V_{CSN} = 100mV, 5V < V_{CSP} and V_{CSN} < 6V	--	2	--	μA
		V_{CSP} – V_{CSN} = 50mV, 5V < V_{CSP} and V_{CSN} < 6V	--	0.95	--	
Constant Current Regulation						
Reference Voltage for Constant Current regulation	V_{REF_CC}	V_{CSP} – V_{CSN} 3.3V < V_{CSP} and V_{CSN} < 6V	--	100	--	mV

Parameter	Symbol	Test Conditions	Min	Typ	Max	Unit
Spread Spectrum						
Spread Spectrum Range	f_{SS}	Spread spectrum option only	--	+6	--	%
Over-Temperature Protection						
Over-Temperature Protection Threshold	T_{OTP}		--	175	--	°C
Over-Temperature Protection Hysteresis	T_{OTP_HYS}		--	15	--	°C
Output Undervoltage Protection						
Output Undervoltage Protection Threshold	V_{UVP}	UVP detect	0.35	0.4	0.45	V

16 Typical Application Circuit

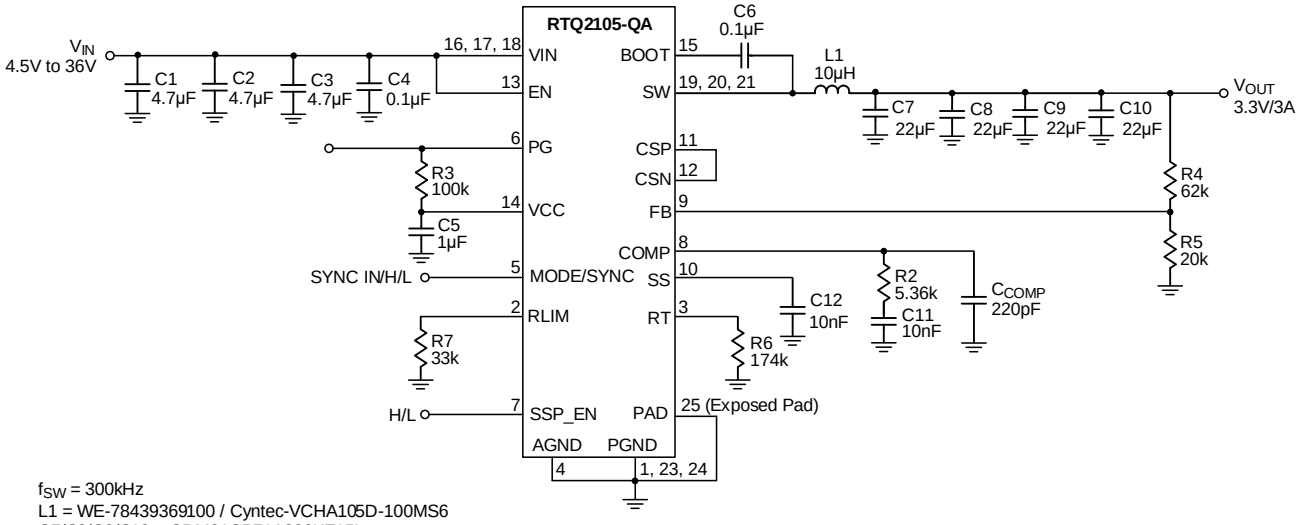
16.1 Buck Circuit without Switch-Over Function

16.1.1 300kHz, 5V, 3A Buck Converter



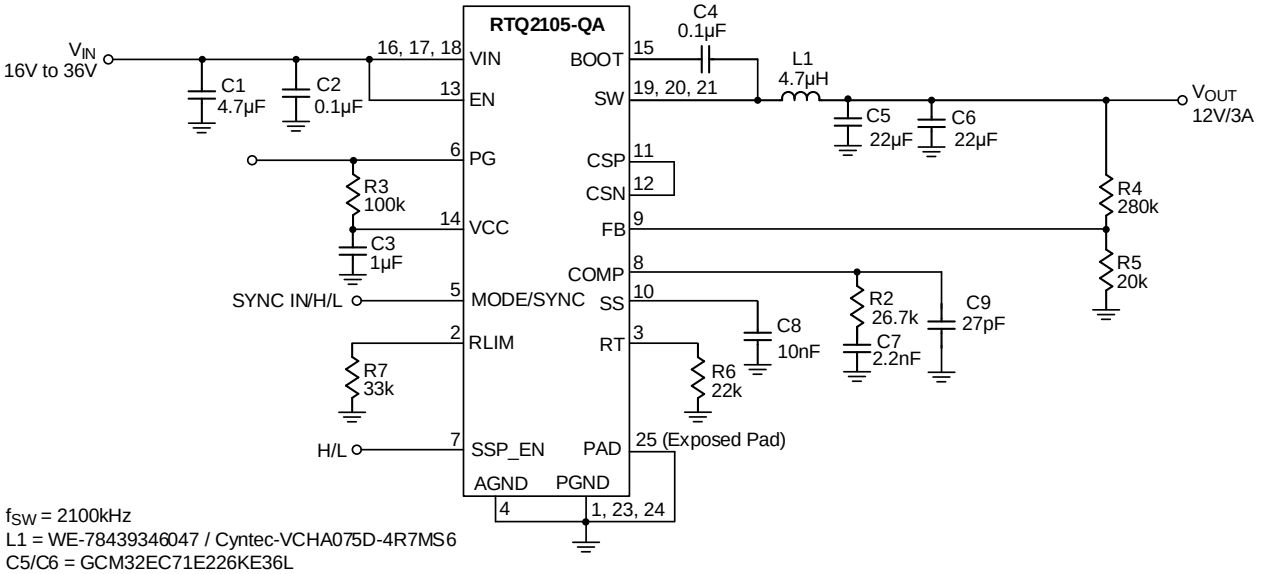
f_{SW} = 300kHz
L1 = WE-78439369150 / Cyntec-VCHA105D-150MS6
C7/C8/C9/C10 = GRM31CR71A226KE15L
C1/C2/C3 = GRM31CR71H475KA12L

16.1.2 300kHz, 3.3V, 3A Buck Converter



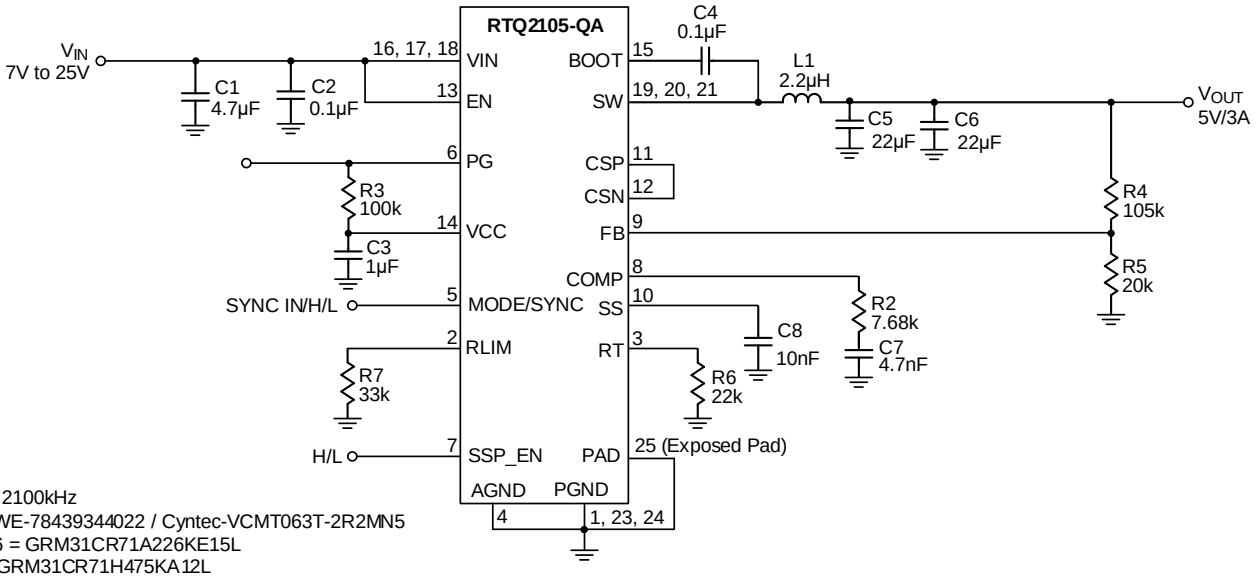
f_{SW} = 300kHz
L1 = WE-78439369100 / Cyntec-VCHA105D-100MS6
C7/C8/C9/C10 = GRM31CR71A226KE15L
C1/C2/C3 = GRM31CR71H475KA12L

16.1.3 2100kHz, 12V, 3A Buck Converter



f_{SW} = 2100kHz
L1 = WE-78439346047 / Cyntec-VCHA075D-4R7MS6
C5/C6 = GCM32EC71E226KE36L
C1 = GRM31CR71H475KA12L

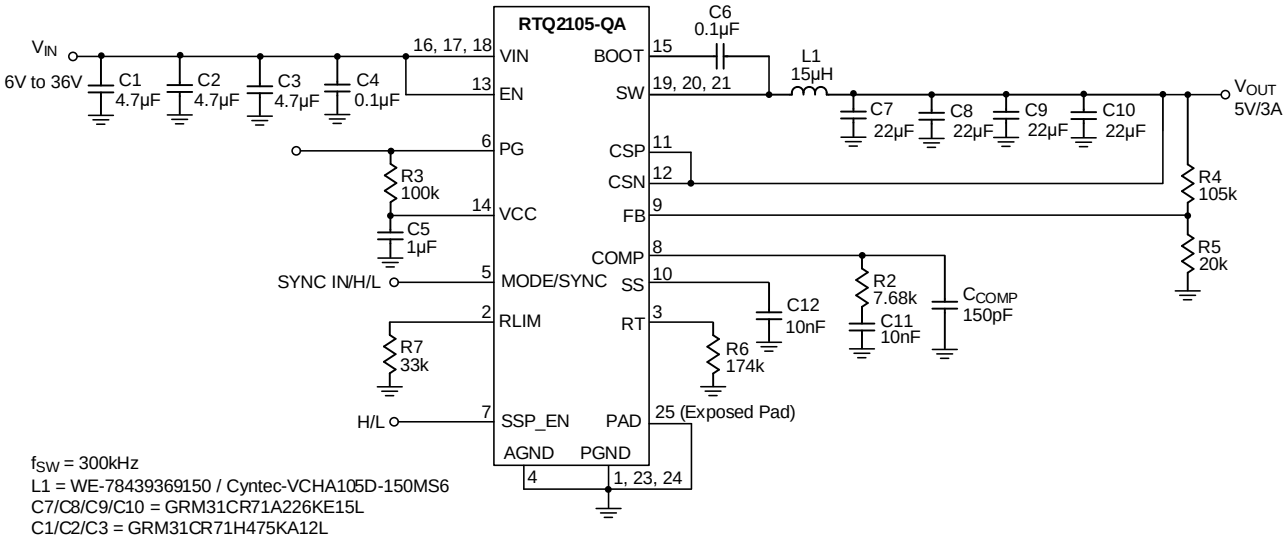
16.1.4 2100kHz, 5V, 3A Buck Converter



f_{SW} = 2100kHz
L1 = WE-78439344022 / Cyntec-VCMT063T-2R2MN5
C5/C6 = GRM31CR71A226KE15L
C1 = GRM31CR71H475KA12L

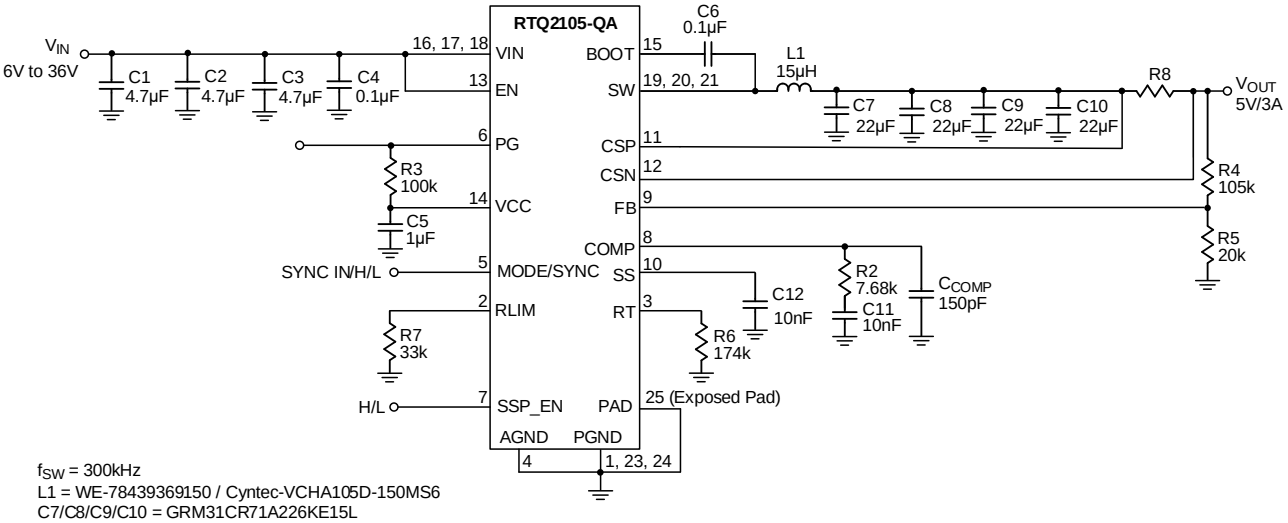
16.2 Buck Circuit with Switch-Over Function (for V_{OUT} regulated within 4.8V to 6V only)

16.2.1 300kHz, 5V, 3A Buck Converter



16.3 Buck Circuit with Cable Drop Compensation and Average Current Limit (for V_{OUT} = 5V only)

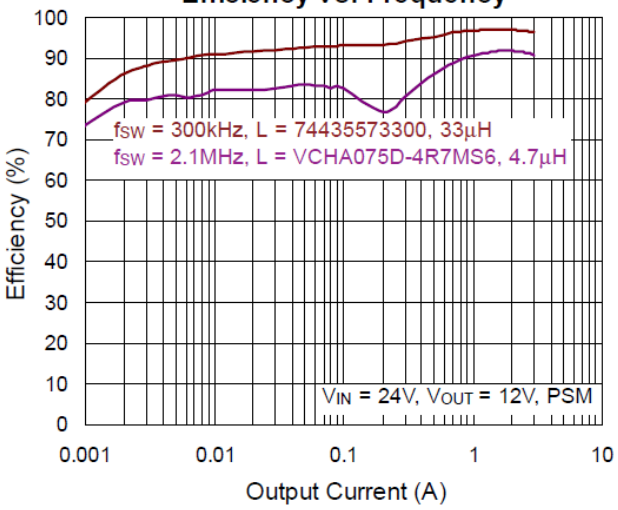
16.3.1 300kHz, 5V, 3A Buck Converter



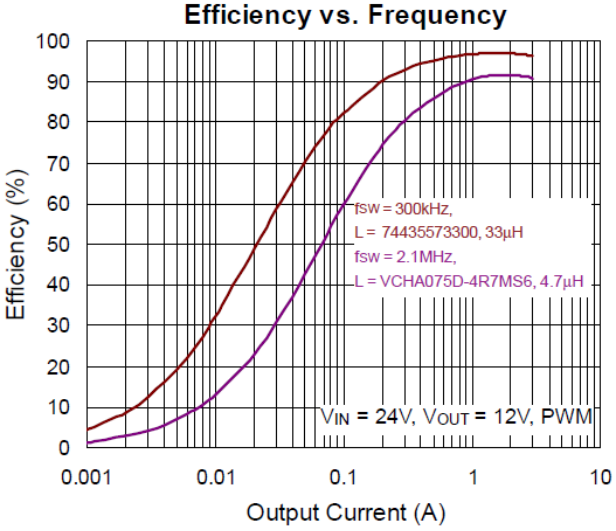
17 Typical Operating Characteristics

Unless otherwise specified, the following conditions apply: V_{IN} = 12V, T_A = 25°C. Specified temperatures are ambient.

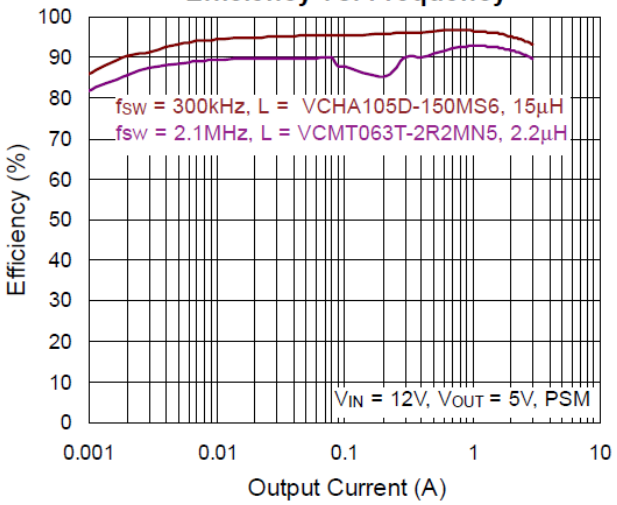
Efficiency vs. Frequency



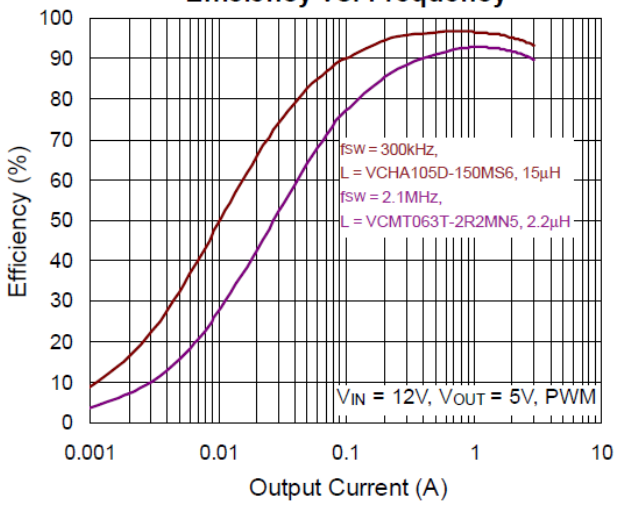
Efficiency vs. Frequency



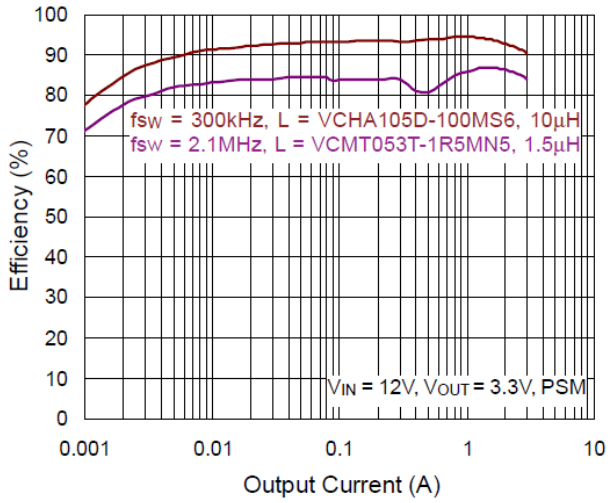
Efficiency vs. Frequency



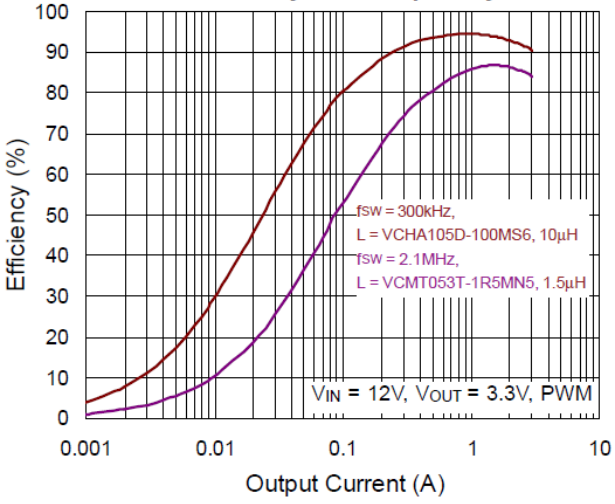
Efficiency vs. Frequency



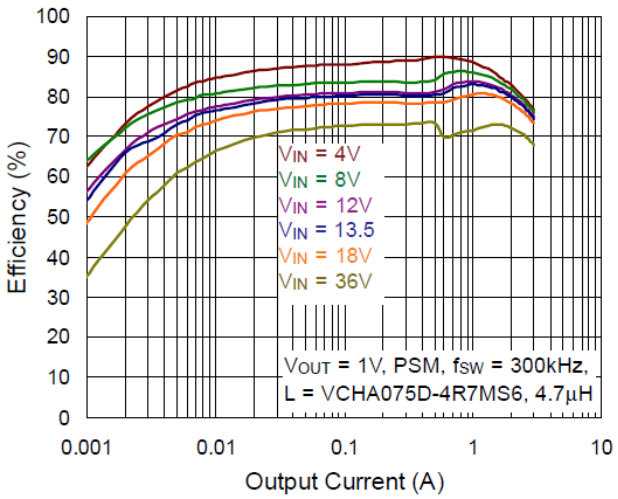
Efficiency vs. Frequency



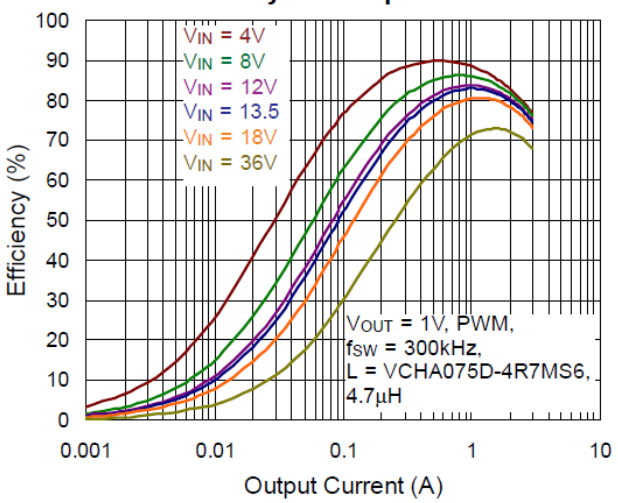
Efficiency vs. Frequency



Efficiency vs. Output Current



Efficiency vs. Output Current



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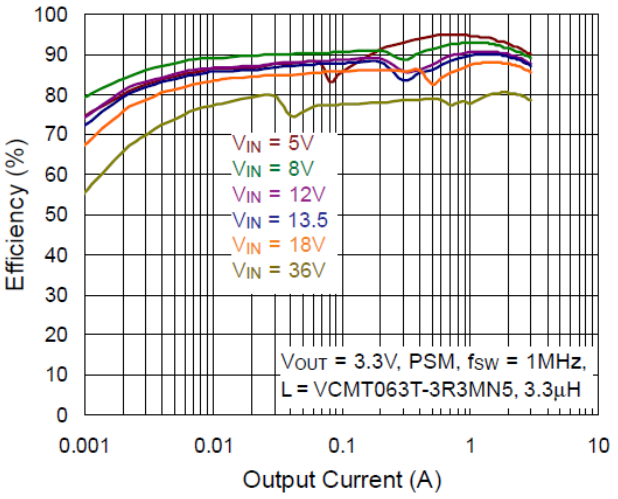
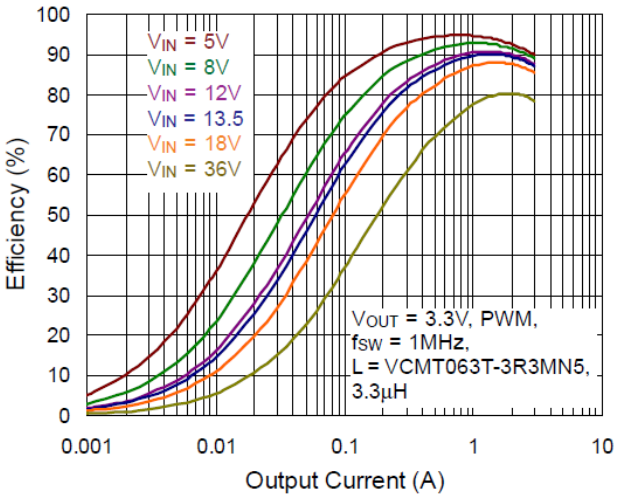
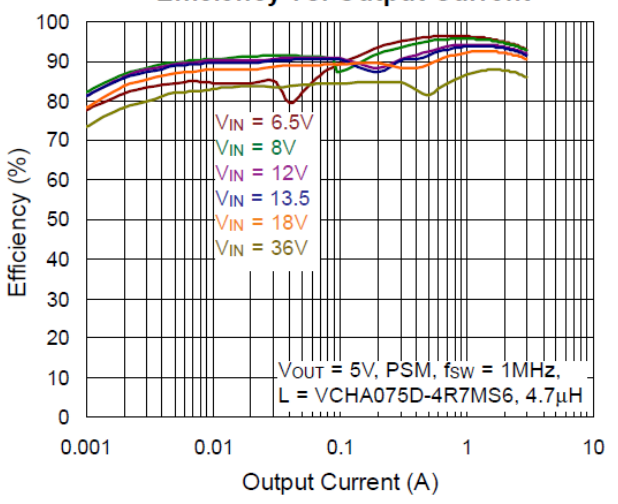


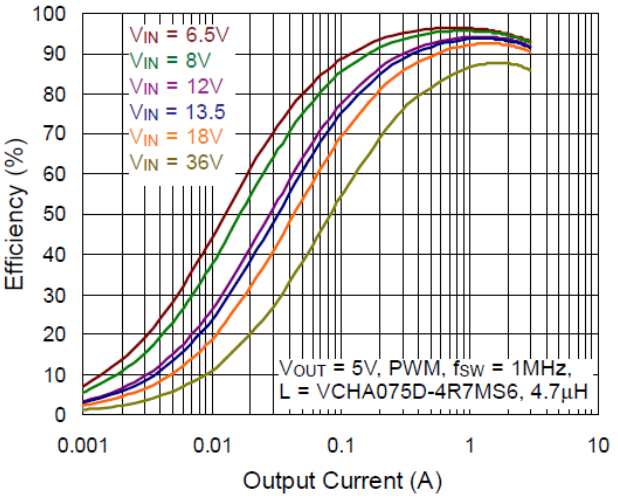
Efficiency vs. Output Current



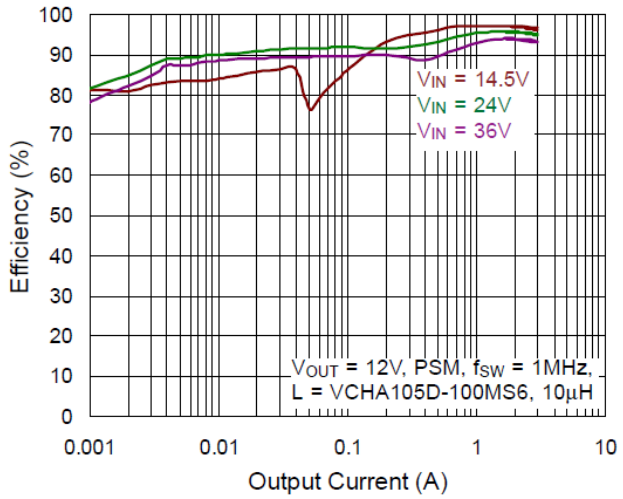
Efficiency vs. Output Current



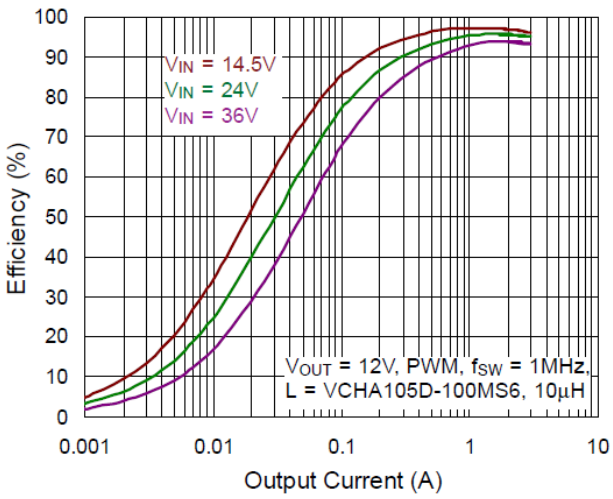
Efficiency vs. Output Current



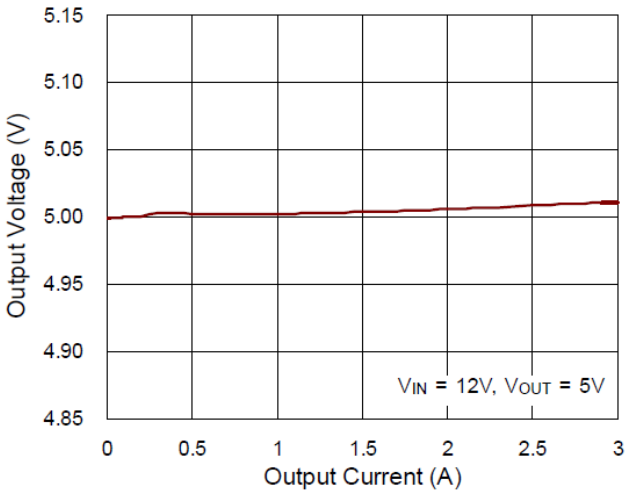
Efficiency vs. Output Current



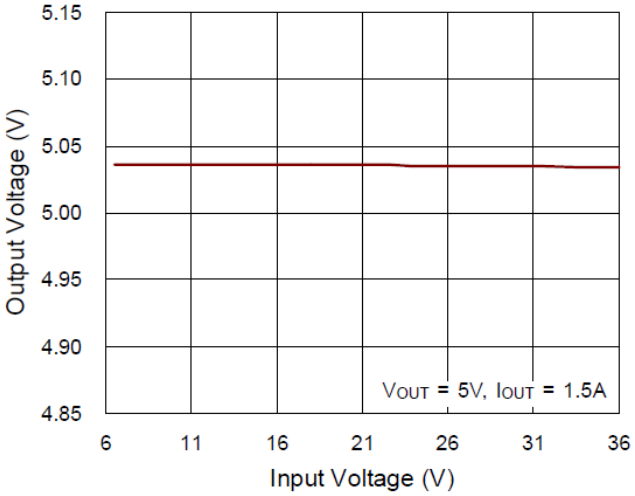
Efficiency vs. Output Current



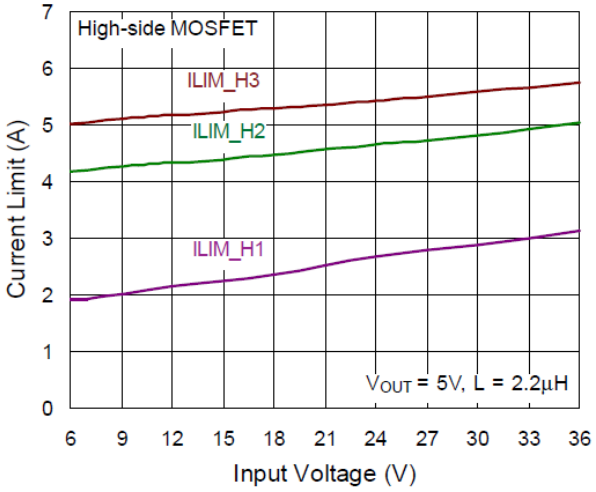
Output Voltage vs. Output Current



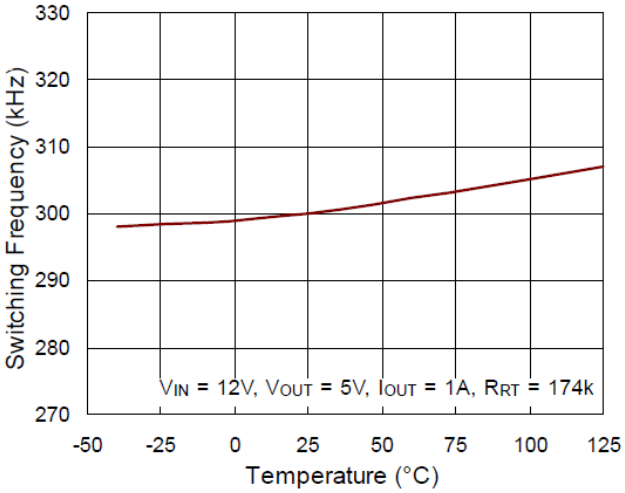
Output Voltage vs. Input Voltage



Current Limit vs. Input Voltage



Switching Frequency vs. Temperature



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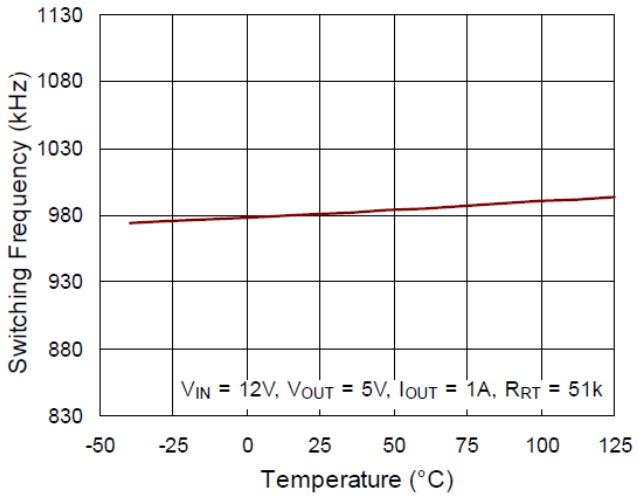
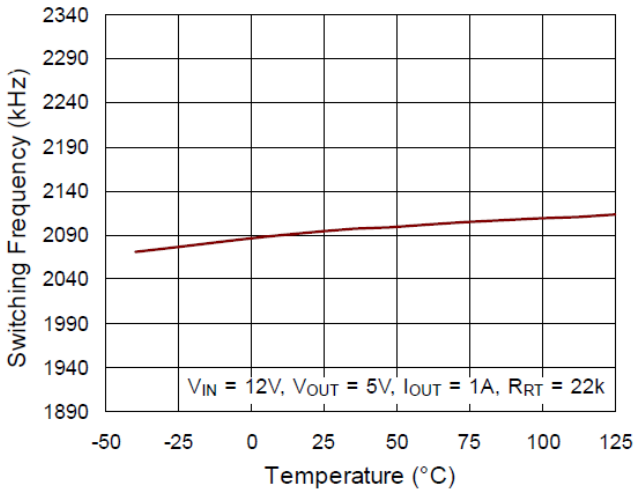
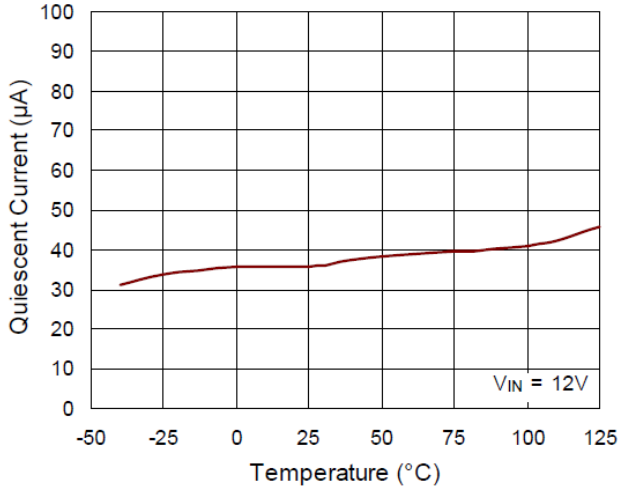


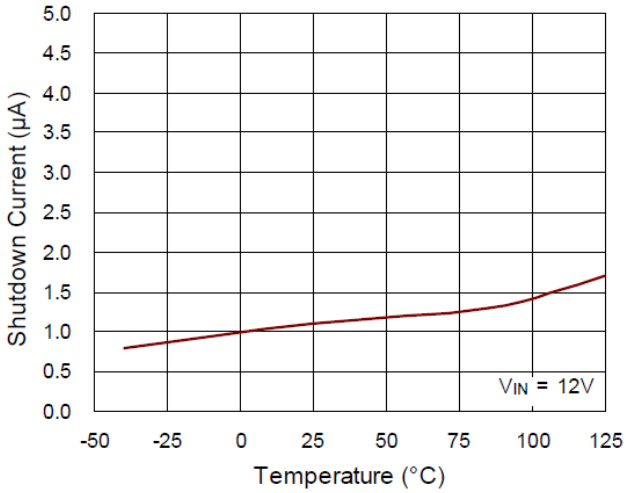
Switching Frequency vs. Temperature



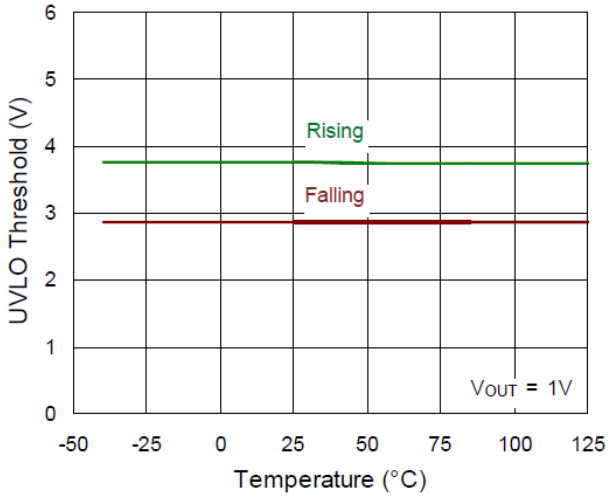
Quiescent Current vs. Temperature



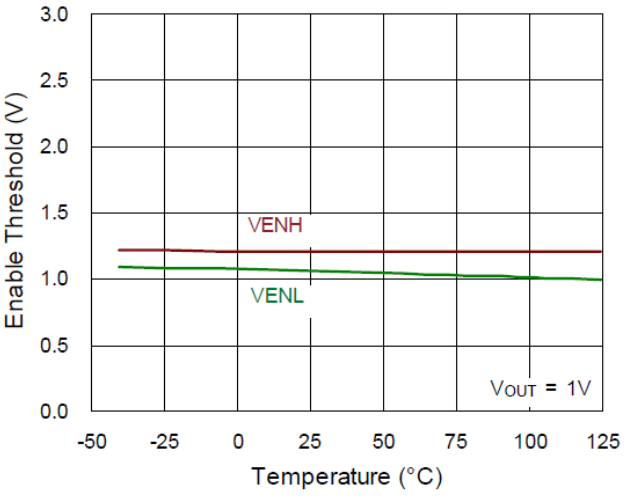
Shutdown Current vs. Temperature



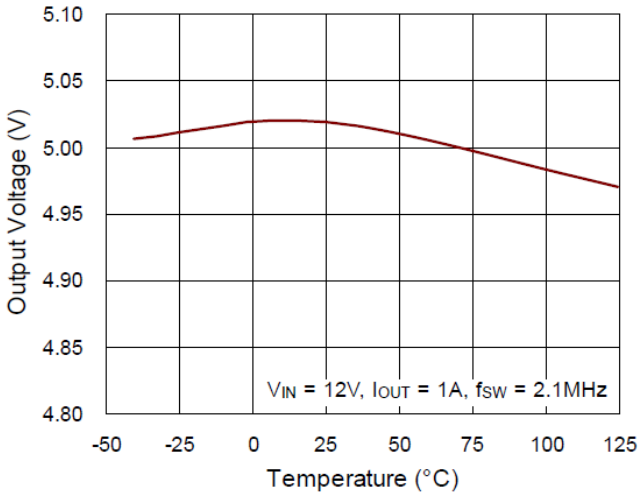
UVLO Threshold vs. Temperature



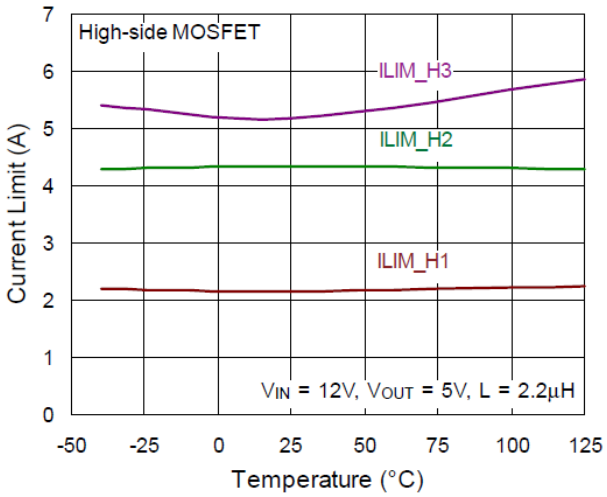
Enable Threshold vs. Temperature



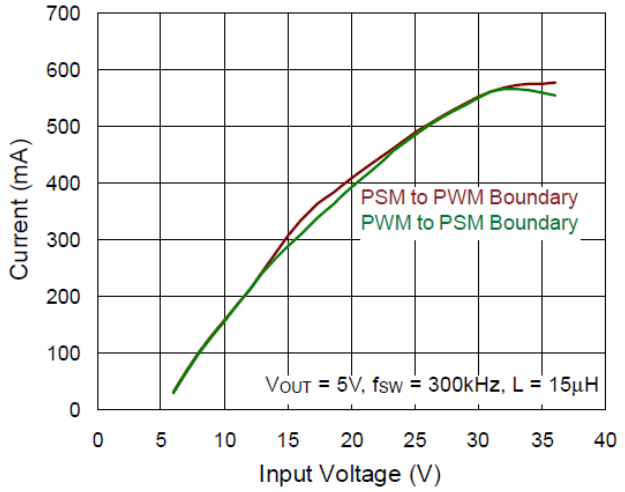
Output Voltage vs. Temperature



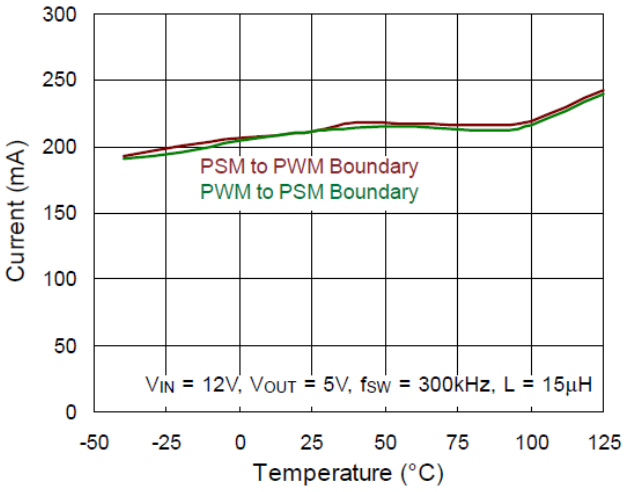
Current Limit vs. Temperature



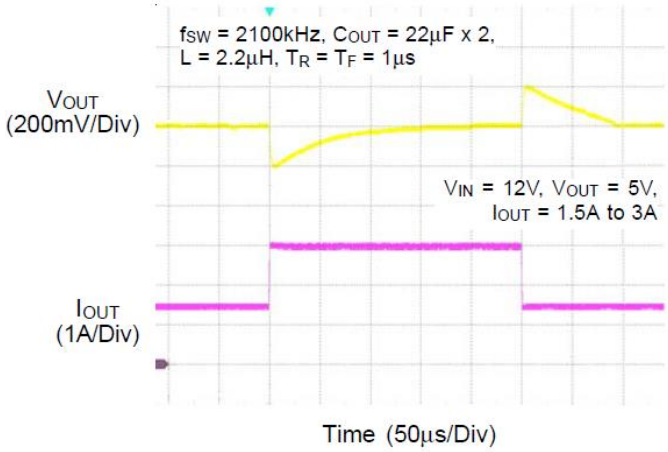
PWM/PSM Boundary vs. Input Voltage



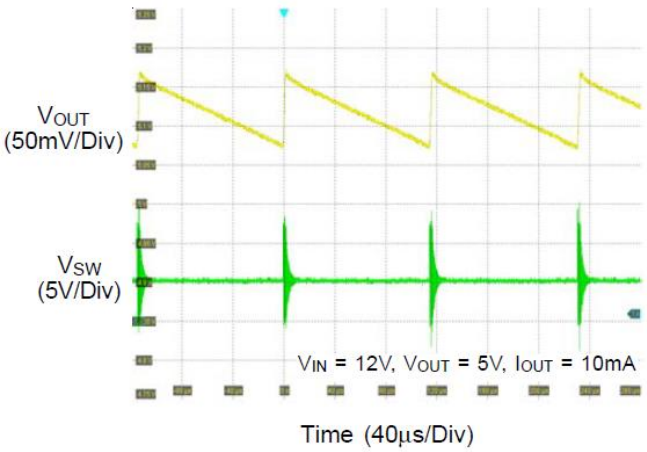
PWM/PSM Boundary vs. Temperature



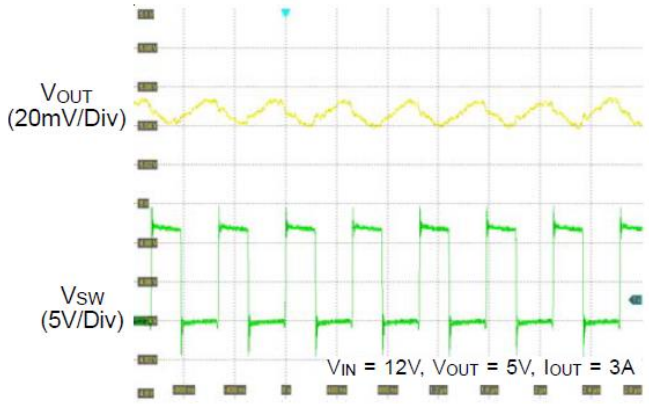
Load Transient Response



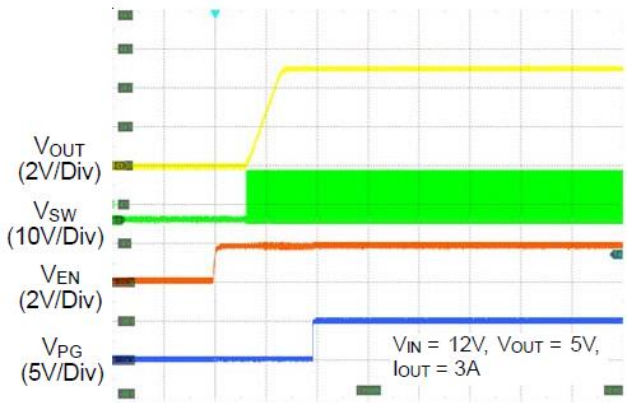
Output Ripple Voltage



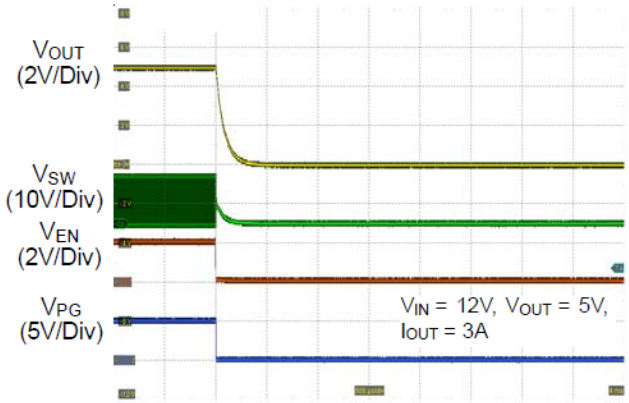
Output Ripple Voltage



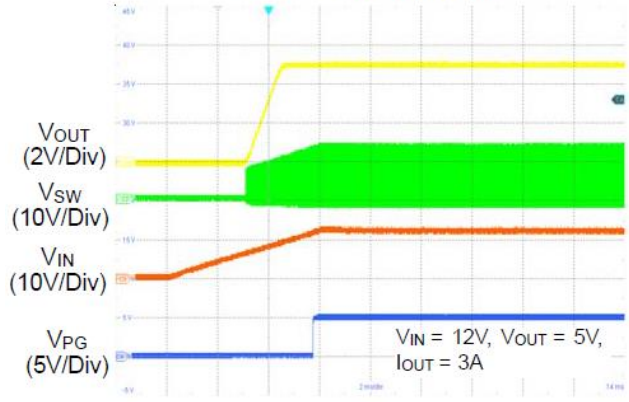
Power On from EN



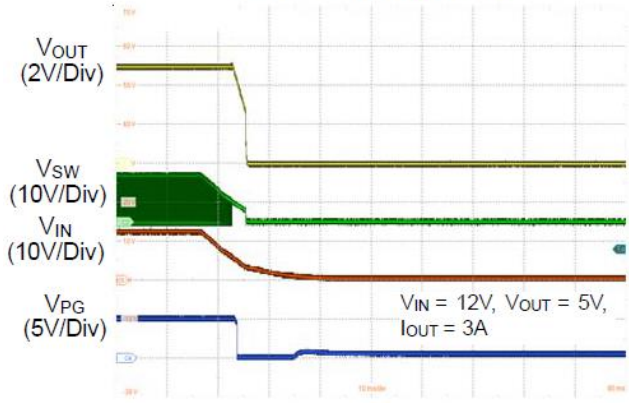
Power Off from EN



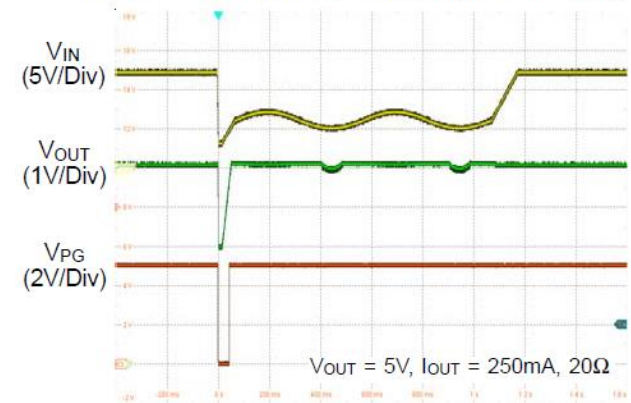
Power On from VIN



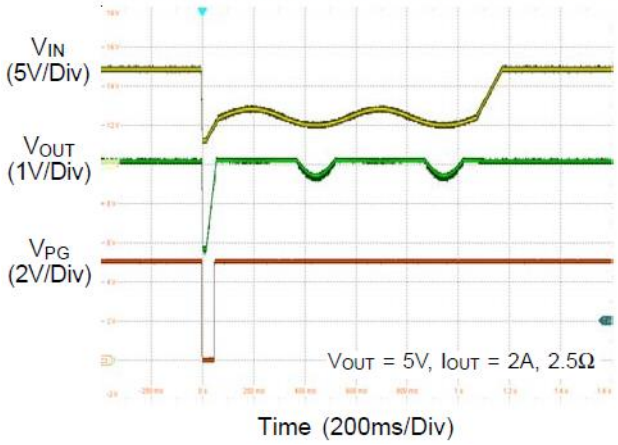
Power Off from VIN



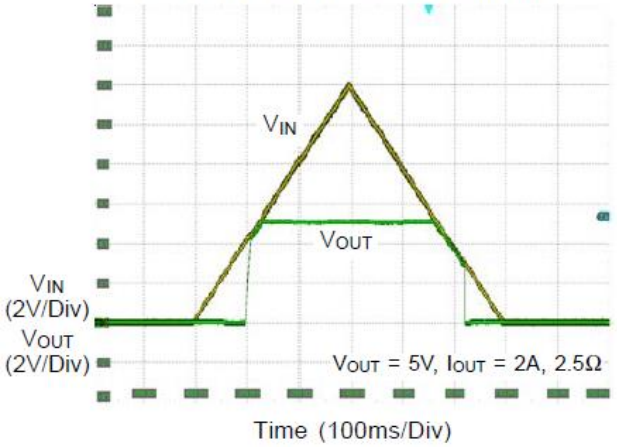
Level III of Starting Profile (ISO16750-2)



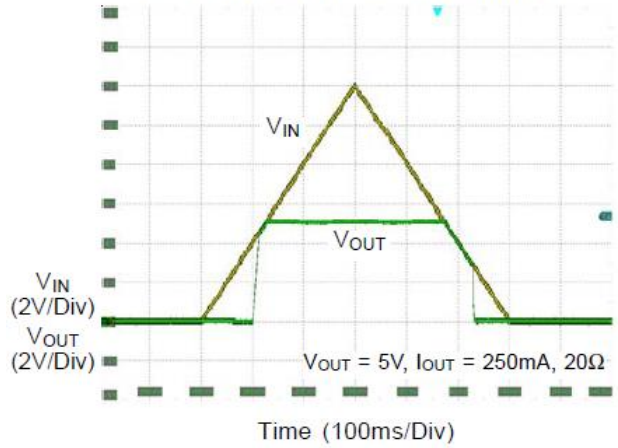
Level III of Starting Profile (ISO16750-2)



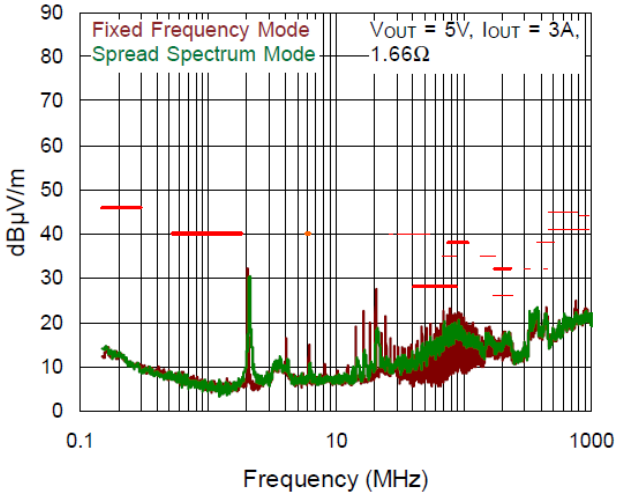
Start-Up Dropout Performance



Start-Up Dropout Performance



Radiated EMI Performance



18 Operations

18.1 Main Control Loop (CV Regulation)

The RTQ2105-QA is a high-efficiency step-down converter that utilizes the peak current mode control. An internal oscillator initiates the turn-on of the high-side MOSFET switch. At the beginning of each clock cycle, the internal high- side MOSFET switch turns on, allowing current to ramp up in the inductor. The inductor current is internally monitored during each switching cycle. The output voltage is sensed on the FB pin via the resistor divider, R1 and R2, and compared with the internal reference voltage for constant voltage control (V_{REF_CV}) to generate a CV compensation signal (V_{COMP}) at the COMP pin. A control signal derived from the inductor current is compared to the voltage at the COMP pin, derived from the feedback voltage. When the inductor current reaches its threshold, the high-side MOSFET switch is turned off and the inductor current ramps down. While the high-side MOSFET switch is off, the inductor current is supplied through the low-side MOSFET switch. This cycle repeats at the next clock cycle. In this way, the duty-cycle and output voltage are controlled by regulating the inductor current.

18.2 Constant Current (CC) Regulation

The RTQ2105-QA offers an average current control loop also. The control loop behavior is basically the same as the peak current mode in constant voltage regulation. The difference is the COMP will be clamped by the output of the internal current error amplifier when the FB voltage is below the regulation target. The output current control is obtained by sensing the voltage drop across an external sense resistor (R_{SENSE}) between CSP and CSN, as shown in Figure 1. The internal reference voltage for the current error amplifier is V_{REF_CC} (typically 100mV). If the output current increases and the current sense voltage (V_{CS}, i.e. $V_{CSP} - V_{CSN}$) equals V_{REF_CC}, the current error amplifier output will clamp the COMP lower to achieve average current control, and vice versa. Once the output current decreases and the current sense voltage is less than 100mV, the CV loop dominates the COMP again, and the output voltage goes back to the regulation voltage determined by the resistor divider from the output to the FB pin and ground accordingly.

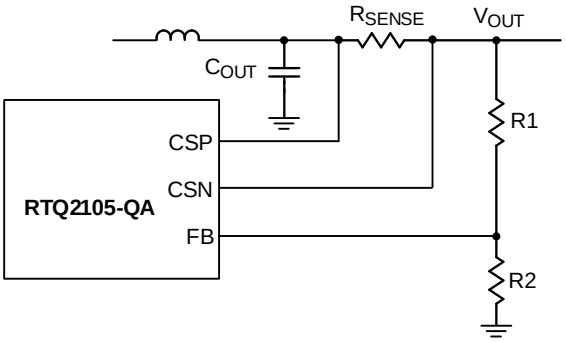


Figure 1. Average Current Setting

18.3 MODE Selection and Synchronization

The RTQ2105-QA provides a MODE/SYNC pin for Forced-PWM Mode (FPWM) and Power Saving Mode (PSM) operation selection at light load. If $V_{MODE/SYNC}$ rises above a logic-high threshold voltage (V_{IH_SYNC}) at the MODE/SYNC input, the device is locked in FPWM. If $V_{MODE/\ SYNC}$ is held below the logic-low threshold voltage (V_{IL_SYNC}) of the MODE/SYNC input, the device operates in PSM at light load to improve efficiency. The RTQ2105-QA can also be synchronized with an external clock ranging from 300kHz to 2.2MHz via the MODE/SYNC pin.

18.4 Forced-PWM Mode

Forced-PWM operation provides constant frequency operation at all loads and is useful in applications sensitive to switching frequency. This mode trades off reduced light load efficiency for low output voltage ripple, tight output voltage regulation, and constant switching frequency. In this mode, a negative current limit of I_{SK_L} is imposed to prevent damage to the low-side MOSFET switch of the regulator. The converter synchronizes to any valid clock signal on the SYNC input when in FPWM.

When constant frequency operation is more important than light load efficiency, pull the MODE/SYNC input high or provide a valid synchronization input. Once activated, this feature ensures that the switching frequency stays away from the AM frequency band, while operating between the minimum and maximum duty cycle limits.

18.5 Power Saving Mode

With the MODE/SYNC pin is floating or pulled low, that is, with a logic low on the MODE/SYNC input, the RTQ2105-QA operates in power saving mode (PSM) at light load to improve light load efficiency. In PSM, the IC starts to switch when V_{FB} is lower than the PSM threshold (V_{REF_CV} x 1.005, typically) and stops switching when V_{FB} is high enough. The IC detects the peak inductor current (I_{L_PEAK}) and keeps high-side MOSFET switch on until the I_L reaches its minimum peak current level (1A at V_{IN} = 12V, typically) to ensure that IC can provide sufficiency output current with each switching pulse. Zero-current detection is also activated to prevent I_L from becoming negative and to ensure no external discharging current from the output capacitor. During the non-switching period, most of the internal circuit is shut down, and the supply current drops to the quiescent current (typically, 40μA) to reduce the quiescent power consumption. With lower output loading, the non-switching period is longer, so the effective switching frequency becomes lower to reduce the switching loss and switch driving loss.

18.6 Maximum Duty Cycle Operation

The RTQ2105-QA is designed to operate in dropout at the high duty cycle approaching 100%. If the operational duty cycle is large and the required off time becomes smaller than the minimum off time, the RTQ2105-QA starts to enable the skip off time function and keeps the high-side MOSFET switch on continuously. The RTQ2105-QA implements the skip off time function to achieve a high duty cycle approaching 100%. Therefore, the maximum output voltage is near the minimum input supply voltage of the application. The input voltage at which the device enters dropout changes depending on the input voltage, output voltage, switching frequency, load current, and the efficiency of the design.

Note that achieving an actual 100% output will only be possible under no-load conditions. In practical scenarios, the ideal maximum output voltage will be equal to the input voltage minus the product of the output current and the maximum high-side MOSFET turn-on resistance. Additionally, when considering a low boot voltage condition, the low-side MOSFET may be turned on for a certain duration. In this case, the actual VOUT can be expressed as V_{OUT} = 0.99 x (V_{IN} – $I_{OUT(max)}$ x $R_{DSON(max)}$). Therefore, it is advisable to allocate a sufficient design margin to ensure that the target output is maintained under all possible loading current scenarios during the system's operation.

18.7 BOOT UVLO

The BOOT UVLO circuit is implemented to ensure a sufficient voltage of the bootstrap capacitor for turning on the high-side MOSFET switch under any conditions. The BOOT UVLO usually actives at extremely high conversion ratios or when the higher V_{OUT} application operates at a very light load. For extremely high conversion ratio conditions after soft-start is finished, or when the higher V_{OUT} application operates at a very light load and PSM, the low-side MOSFET switch may not have sufficient turn-on time to charge the bootstrap capacitor. The device monitors the voltage of the bootstrap capacitor and force to turn on the low-side MOSFET switch when the voltage

of the bootstrap capacitor falls below $V_{BOOT\_UVLO\_L}$ (typically, 2.3V). Meanwhile, the minimum off time is extended to 150ns (typically), hence prolonging the bootstrap capacitor charging time. The BOOT UVLO is sustained until the $V_{BOOT-SW}$ is higher than $V_{BOOT\_UVLO\_H}$ (typically, 2.4V).

18.8 Internal Regulator and Switch-Over

The device integrates a 5V linear regulator (V_{CC}) that is supplied by VIN and provides power to the internal circuitry. The internal regulator operates in low dropout mode when V_{IN} is below 5V. The V_{CC} can be used as the PG pull-up supply but it is "NOT" allowed to power other device or circuitry. The VCC pin must be bypassed to ground with a 1μF, X7R capacitor, and it needs to be placed as close as possible to the VCC pin. Be careful to account for the voltage coefficient of ceramic capacitors when choosing the value and case size. Many ceramic capacitors lose 50% or more of their rated value when used near their rated voltage. The RTQ2105-QA implements a switch-over function to improve efficiency at all loads. The switch-over function can be enabled when the CSP and CSN pins are tied to a voltage higher than 4.8V (typically) and V_{CC} will be supplied from V_{CSP}, otherwise V_{CC} will be supplied from VIN by the internal regulator. Typically, the CSP and CSN pins can be tied to the output of the RTQ2105-QA if the output voltage is regulated at 4.8V to 6V with ±1% tolerance or they can be tied to an external power supply with a voltage range of 4.8V to 6V. If the V_{CSP} drops below

4.6V (typically), the internal VCC regulator will automatically turn on to provide power to the internal circuit blocks. The CSP pin should be tied to the CSN pin and be left floating if the switch-over function is not needed.

18.9 Enable Control

The RTQ2105-QA provides an EN pin as an external chip enable control to enable or disable the device. If V_{EN} is held below a logic-low threshold voltage (V_{EN_F}), switching is inhibited even if the VIN voltage is above the VIN undervoltage-lockout threshold (V_{UVLO_R}). If V_{EN} is held below 0.4V, the converter will enter shutdown mode, that is, the converter is disabled. During shutdown mode, the supply current can be reduced to I_{SHDN} (5μA or below). If the V_{EN} rises above the logic-high threshold voltage (V_{EN_R}) while the VIN voltage is higher than V_{UVLO_R}, the device will be turned on, that is, switching being enabled and the soft-start sequence being initiated. The current source of EN typically sinks 1.2μA.

18.10 Soft-Start

The soft-start function is used to prevent large inrush currents while the converter is being powered up. The RTQ2105-QA provides an SS pin so that the soft-start time can be programmed by selecting the value of the external soft-start capacitor C_{SS} connected from the SS pin to ground. During the start-up sequence, the soft-start capacitor is charged by an internal current source I_{SS} (typically, 6μA) to generate a soft-start ramp voltage as a reference voltage to the PWM comparator. If the output is pre-biased to a certain voltage during start-up, the device will not turn on the high-side MOSFET switch until the voltage difference between the SS pin and the FB pin is larger than 400mV ($V_{SS} - V_{FB} > 400mV$, typically). And only when this ramp voltage is higher than the feedback voltage V_{FB}, the switching will be resumed. The output voltage can then ramp up smoothly to its targeted regulation voltage, and the converter can have a monotonic smooth start-up. For soft-start control, the SS pin should never be left unconnected. After the V_{SS} rises above 2V (typically), the PG pin will be in high impedance and V_{PG} will be held high. The typical start-up waveform shown in Figure 2 indicates the sequence and the timing between the output voltage and related voltages.

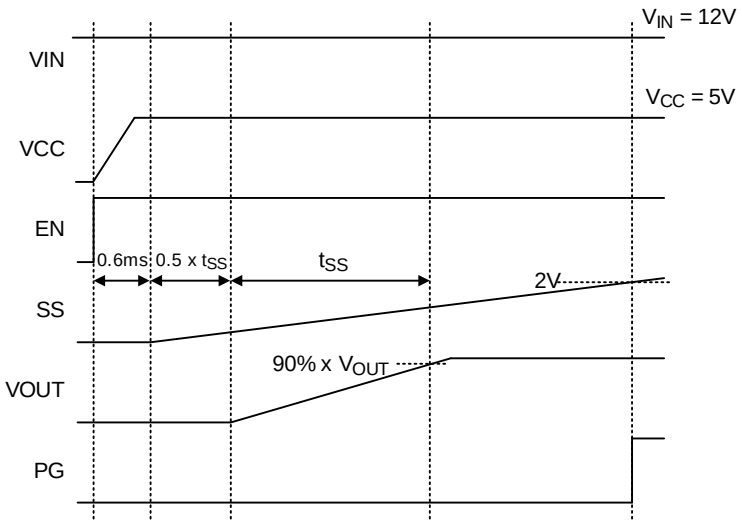


Figure 2. Start-Up Sequence

18.11 Power-Good Indicator

The RTQ2105-QA features an open-drain power-good output (PG) to monitor the output voltage status. The output delay of the comparator prevents false flag operation for short excursions in the output voltage, such as during line and load transients. Pull up PG with a resistor to V_{CC} or an external voltage below 5.5V. The power-good function is activated after the soft-start is finished and is controlled by a comparator connected to the feedback signal V_{FB}. If V_{FB} rises above a power-good high threshold (V_{TH_PGLH1}) (typically 90% of the reference voltage), the PG pin will be in high impedance and V_{PG} will be held high after a certain delay elapsed. When V_{FB} exceeds V_{TH_PGHL1} (typically 120% of the reference voltage), the PG pin will be pulled low. Moreover, the IC turns off the high-side MOSFET switch and turns on the low-side MOSFET switch until the inductor current reaches I_{SK_L} if the MODE pin is set high. If the V_{FB} is still higher than V_{TH_PGHL1}, the high-side MOSFET switch remains prohibited and the low-side MOSFET switch will turn on again in the next cycle. If the MODE pin is set low, the IC turns off the low-side MOSFET switch once the inductor current reaches zero current unless $V_{BOOT-SW}$ is too low. For V_{FB} higher than V_{TH_PGHL1}, V_{PG} can be pulled high again if V_{FB} drops back by a power-good high threshold (V_{TH_PGLH2}) (typically 117% of the reference voltage). When V_{FB} falls short of the power-good low threshold (V_{TH_PGHL2}) (typically 85% of the reference voltage), the PG pin will be pulled low. Once started up, if any internal protection is triggered, PG will be pulled low to ground. The internal open-drain pull-down device (10Ω, typically) will pull the PG pin low. The power-good indicator profile is shown in Figure 3.

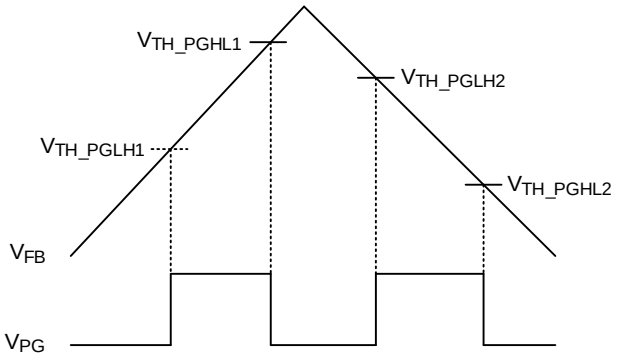


Figure 3. The Logic of PG

18.12 Spread Spectrum Operation

Due to the periodicity of the switching signals, the energy concentrates at one particular frequency and its harmonics. These levels or energy are radiated, and therefore this is where a potential EMI issue arises. The RTQ2105-QA has an optional spread spectrum function, and the SSP_EN pin can be programmed to turn on/off the spread spectrum, further simplifying compliance with the CISPR and automotive EMI requirements. The spread spectrum can be activate when the soft-start is finished and zero-current is not detected. If V_{SSP_EN} rises above a logic-high threshold voltage 2V at the SSP_EN input, the device enables spread spectrum operation. The spread spectrum is implemented using a pseudo-random sequence and uses a +6% spread of the switching frequency. For example, when the RTQ2105-QA is programmed to 2.1MHz, the frequency will vary from

2.1MHz to 2.226MHz. Therefore, the RTQ2105-QA still guarantees that the 2.1MHz switching frequency setting does not drop into the AM band limit of 1.8MHz. However, the spread spectrum cannot be active when the device is synchronized with an external clock via the MODE/SYNC pin.

18.13 Input Undervoltage-Lockout

In addition to the EN pin, the RTQ2105-QA also provides enable control through the VIN pin. If V_{EN} rises above V_{EN_R} first, switching will still be inhibited until the VIN voltage rises above V_{UVLO_R}. It ensures that the internal regulator is ready, preventing operation with not-fully-enhanced internal MOSFET switches. After the device is powered up, if the VIN voltage goes below the UVLO falling threshold voltage V_{UVLO_F}, this switching will be inhibited; if the VIN voltage rises above the UVLO rising threshold (V_{UVLO_R}), the device will resume switching. Note that V_{IN} = 3V is only design for cold crank requirements; the normal input voltage should be larger than the V_{UVLO_R} threshold to turn on.

18.14 High-Side Switch Peak Current-Limit Protection

The RTQ2105-QA includes a cycle-by-cycle high-side switch peak current-limit protection against the conditions that the inductor current increasing abnormally, even exceeding the inductor saturation current rating. The high-side MOSFET switch peak current limit of the RTQ2105-QA is adjustable by placing a resistor on the RLIM pin. The recommended resistor value ranges from 33kΩ (for typical 5.5A) to 91kΩ (for typical 2.2A). It is recommended to use 1% tolerance or better and a temperature coefficient of 100 ppm or less resistors. The inductor current through the high-side MOSFET switch will be measured after a certain delay when the high-side MOSFET switch is turned on. If an overcurrent condition occurs, the converter will immediately turn off the high-side MOSFET switch and turn on the low-side MOSFET switch to prevent the inductor current from exceeding the high-side MOSFET switch peak current limit (I_{LIM_H}).

18.15 Low-Side Switch Current-Limit Protection

The RTQ2105-QA not only implements the high-side switch peak current limit but also provides the sourcing current limit and sinking current limit for low-side MOSFET switch. With these current protections, the IC can easily control the inductor current at both side switches and avoid current runaway for short-circuit conditions.

For the low-side MOSFET switch sourcing current limit, there is a specific comparator in the internal circuitry to compare the low-side MOSFET switch sourcing current to the low-side MOSFET switch sourcing current limit at the end of every clock cycle. When the low-side MOSFET switch sourcing current is higher than the low-side MOSFET switch sourcing current limit, which is the high-side MOSFET switch current limit (I_{LIM_H}) multiplied by 0.95, the new switching cycle is not initiated until the inductor current drops below the low-side MOSFET switch sourcing current limit.

For the low-side MOSFET switch sinking current-limit protection, it is implemented by detecting the voltage across the low-side MOSFET switch. If the low-side switch sinking current exceeds the negative inductor peak current

limit ($I_{LIM\_PEAK\_NEG}$) (typically 2A), the converter will immediately turn off the low-side MOSFET switch and turn on the high-side MOSFET switch.

18.16 Output Undervoltage Protection

The RTQ2105-QA includes output undervoltage protection (UVP) against overload or short-circuited conditions by constantly monitoring the feedback voltage V_{FB}. If V_{FB} drops below the undervoltage protection trip threshold (typically 50% of the internal reference voltage), the UV comparator will go high to turn off the high-side MOSFET and then turn off the low-side MOSFET when the inductor current drops to zero. If the output undervoltage condition continues for a period of time, the RTQ2105-QA enters output undervoltage protection with hiccup mode and discharges the C_{SS} by an internal discharging current source I_{SS_DIS} (typically, 80nA). During hiccup mode, the device remains shut down. After the V_{SS} is discharged to less than 150mV (typical), the RTQ2105-QA attempts to restart. The internal charging current source I_{SS} gradually increases the voltage on Css. The high-side MOSFET switch will start switching when the voltage difference between the SS pin and the FB pin is larger than 400mV ($V_{SS} - V_{FB} > 400mV$, typically). If the output undervoltage condition is not removed, the high-side MOSFET switch stop switching when the voltage difference between the SS pin and the FB pin is 700mV ($V_{SS} - V_{FB} = 700mV$, typically) and then the I_{SS_DIS} discharging current source begins to discharge C_{SS}.

Upon completion of the soft-start sequence, if the output undervoltage condition is removed, the converter will resume normal operation; otherwise, this cycle for auto-recovery will be repeated until the output undervoltage condition is cleared.

Hiccup mode allows the circuit to operate safely with low input current and power dissipation, and then resume normal operation as soon as the overload or short-circuit condition is removed. A short-circuit protection and recovery profile is shown in Figure 4.

Since the C_{SS} will be discharged to 150mV when the RTQ2105-QA enters output undervoltage protection, the first discharging time (t_{SS_DIS1}) can be calculated as follows:

$$t_{SS\_DIS1} = C_{SS} \times \frac{V_{SS} - 0.15}{I_{SS\_DIS}}$$

The equation below assumes that the V_{FB} will be 0 in a short-circuited condition and can be used to calculate the C_{SS} discharging time (t_{SS_DIS2}) and charging time (t_{SS_CH}) during hiccup mode.

$$t_{SS\_DIS2} = C_{SS} \times \frac{0.55}{I_{SS\_DIS}}$$

$$t_{SS\_CH} = C_{SS} \times \frac{0.55}{I_{SS\_CH}}$$

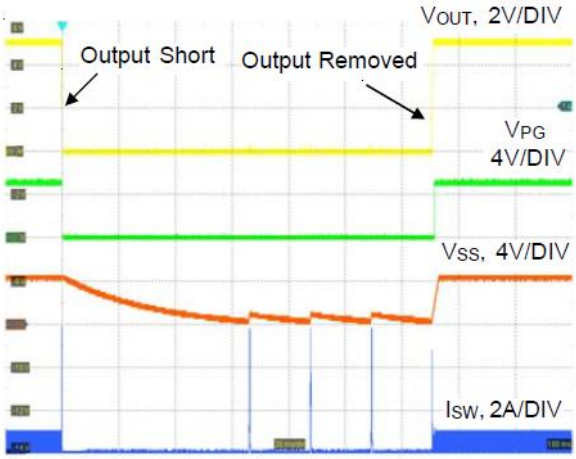


Figure 4. Short Circuit Protection and Recovery

18.17 Over-Temperature Protection

The RTQ2105-QA includes an over-temperature protection (OTP) circuitry to prevent overheating due to excessive power dissipation. The OTP will shut down switching operations when the junction temperature exceeds the over-temperature protection threshold T_{OTP}. Once the junction temperature cools down by the over-temperature protection hysteresis (T_{OTP_HYS}), the IC will resume normal operation with a complete soft-start.

18.18 Pin-Short Protection

The RTQ2105-QA provides pin-short protection for neighbor pins. The internal protection fuse will burn out to prevent IC smoke, fire, and sparks when the BOOT pin is shorted to the VIN pin.

19 Application Information

(Note 8)

The RTQ2105-QA features a general application circuit shown in Typical Application Circuit. External component selection is largely driven by the load requirements and begins with the selection of the operating mode by setting the $V_{MODE/SYNC}$ and the operating frequency using an external resistor R_T. Next, the inductor L, the input capacitor C_{IN}, and the output capacitor C_{OUT} are chosen. Following this, feedback resistors and compensation circuit are selected to set the desired output voltage and crossover frequency. Next, the internal regulator capacitor C_{VCC} and the bootstrap capacitor C_{BOOT} can be selected. Finally, the remaining external components can be selected for functions such as the EN, external soft-start, PG, inductor peak current limit, synchronization, spread spectrum, average current limit, and adjustable output voltage with cable drop compensation.

19.1 FPWM/PSM Selection

The RTQ2105-QA provides a MODE/SYNC pin for Forced-PWM Mode (FPWM) and Power Saving Mode (PSM) operation selection at light load. To optimize efficiency at light loads, the RTQ2105-QA can be set in PSM. When the VMODE/ SYNC is held below a logic-low threshold voltage (V_{IL_SYNC}) of the MODE/SYNC input, that is, with the MODE/SYNC pin floating or pulled low, the device operates in PSM at light load to improve light load efficiency. If it is necessary to keep switching harmonics out of the signal band, the RTQ2105-QA can operate in FPWM. The device is locked in PWM mode when $V_{MODE/SYNC}$ rises above a logic-high threshold voltage (V_{IH_SYNC}) of the MODE/SYNC input. The FPWM trades off reduced light load efficiency for low output voltage ripple, tight output voltage regulation, fast transient response, and constant switching frequency.

19.2 Switching Frequency Setting

The RTQ2105-QA offers adjustable switching frequency settings, and the switching frequency can be set by using an external resistor R_T. The switching frequency range is from 300kHz to 2.2MHz. Selection of the operating frequency is a trade-off between efficiency and component size. High-frequency operation allows the use of smaller inductor and capacitor values. Operation at lower frequencies improves efficiency by reducing internal gate charge and transition losses but requires larger inductance values and/or capacitance to maintain low output ripple voltage. An additional constraint on operating frequency is the minimum on-time and minimum off-time. The minimum on-time, t_{ON_MIN}, is the smallest duration of time in which the high-side switch can be in its “on” state. This time is 60ns (typical). In continuous mode operation, the minimum on-time limit imposes a maximum operating frequency, f_{SW_MAX}, of:

$$f_{SW\_MAX}(MHz) = \frac{V_{OUT}(V)}{t_{ON\_MIN}(\mu s) \times V_{IN\_MAX}(V)}$$

where V_{IN_MAX} is the maximum operating input voltage. The minimum off-time, t_{OFF_MIN}, is the smallest amount of time that the RTQ2105-QA is capable of turning on the low-side MOSFET switch, tripping the current comparator, and turning the MOSFET switch back off. The minimum off-time is 65ns (typical). If the switching frequency should be constant, the required off time needs to be larger than minimum off time. Below is the calculation of the minimum off time with loss terms consideration:

$$t_{OFF\_MIN}(ns) \le \frac{1 - \left[\frac{V_{OUT}(V) + I_{OUT\_MAX}(A) \times \left(R_{DS(ON)\_L} + DCR\right)}{V_{IN\_MIN}(V) - I_{OUT\_MAX}(A) \times \left(R_{DS(ON)\_H} - R_{DS(ON)\_L}\right)} \right]}{f_{SW}(MHz)}$$

where $R_{DS(ON)\_H}(m\Omega)$ is the on-resistance of the high-side MOSFET switch; $R_{DS(ON)\_L}$ is the on-resistance of the

low-side MOSFET switch; DCR(mΩ) is the DC resistance of the inductor.

An external resistor R_T, connected between the RT pin and ground, sets the switching frequency f_{SW}. The failure modes and effects analysis (FMEA) considerations are applied to the RT pin setting to avoid abnormal switching frequency operation under failure conditions. It includes failure scenarios of short-circuit to ground and the pin is left open. The switching frequency will be 2.35MHz (typical) when the RT pin is shorted to ground and 250kHz (typical) when the pin is left open. The equation below shows the relationship between the setting frequency and the RT value.

$$R_T(k\Omega) = 74296 \times f_{SW}^{-1.06}$$

where f_{SW} (kHz) is the desire setting frequency. It is recommended to use resistors with 1% tolerance or better and a temperature coefficient of 100 ppm or less. Figure 5 shows the relationship between switching frequency and the RT resistor.

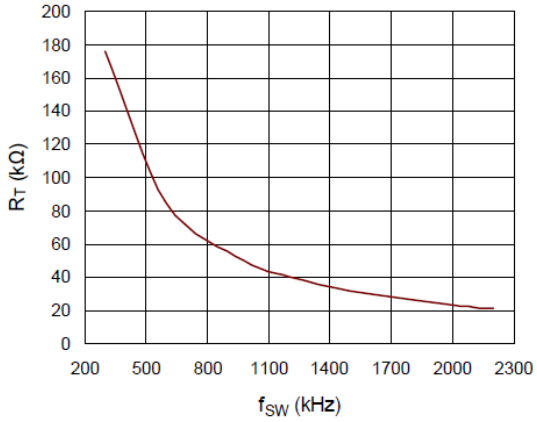


Figure 5. Switching Frequency vs. R_T

19.3 Inductor Selection

The inductor selection makes trade-offs among size, cost, efficiency, and transient response requirements. Generally, three key inductor parameters are specified for operation with the device: inductance value (L), inductor saturation current (I_{SAT}), and DC resistance (DCR). A good compromise between size and loss is a 30% peak-to-peak ripple current to the IC rated current. The switching frequency, input voltage, output voltage, and selected inductor ripple current determines the inductor value as follows:

$$L = \frac{V_{OUT} \times (V_{IN} - V_{OUT})}{V_{IN} \times f_{SW} \times \Delta I_L}$$

Larger inductance values result in lower output ripple voltage and higher efficiency, but a slightly degraded transient response. This results in additional phase lag in the loop and reduces the crossover frequency. As the ratio of the slope-compensation ramp to the sensed-current ramp increases, the current-mode system tilts towards voltage-mode control. Lower inductance values allow for a smaller case size, but the increased ripple lowers the effective current-limit threshold, increases the AC losses in the inductor and may trigger low-side switch sinking current limit in FPWM. It also causes insufficient slope compensation and ultimately loop instability as the duty cycle approaches or exceeds 50%. When the duty cycle exceeds 50%, the following condition needs to be satisfied:

$$2.1 \times f_{SW}(MHz) > \frac{V_{OUT}(V)}{L(\mu H)}$$

A good compromise among size, efficiency, and transient response can be achieved by setting an inductor current

ripple (ΔI_L) with about 10% to 50% of the maximum rated output current (3A). To enhance the efficiency, choose a low-loss inductor with the lowest possible DC resistance that fits in the allotted dimensions. The inductor value determines not only the ripple current but also the load-current value at which DCM/CCM switchover occurs. The selected inductor should have a saturation current rating greater than the peak current limit of the device. The core must be large enough not to saturate at the peak inductor current (I_{L_PEAK}):

$$\Delta I_L = \frac{V_{OUT} \times (V_{IN} - V_{OUT})}{V_{IN} \times f_{SW} \times L}$$

$$I_{L\_PEAK} = I_{OUT\_MAX} + \frac{1}{2} \Delta I_L$$

The current flowing through the inductor is the inductor ripple current plus the output current. During power-up, faults, or transient load conditions, the inductor current can exceed the above peak inductor current level calculated. In transient conditions, the inductor current can increase up to the switch current limit of the device. For this reason, the most conservative approach is to specify an inductor with a saturation current rating equal to or greater than the switch current limit, rather than the peak inductor current. It is recommended to use shielded inductors for good EMI performance.

19.4 Input Capacitor Selection

Input capacitance, C_{IN}, is needed to filter the pulsating current at the drain of the high-side power MOSFET. C_{IN} should be sized to do this without causing a large variation in input voltage. The peak-to-peak voltage ripple on the input capacitor can be estimated using the following equation:

$$\Delta V_{CIN} = D \times I_{OUT} \times \frac{1-D}{C_{IN} \times f_{SW}} + ESR \times I_{OUT}$$

where

$$D = \frac{V_{OUT}}{V_{IN} \times \eta}$$

Figure 6 shows the C_{IN} ripple current flowing through the input capacitors and the resulting voltage ripple across the capacitors. For ceramic capacitors, the equivalent series resistance (ESR) is very low, so the ripple caused by ESR can be ignored. The minimum value of effective input capacitance can be estimated using the following equation:

$$C_{IN\_MIN} = I_{OUT\_MAX} \times \frac{D(1-D)}{\Delta V_{CIN\_MAX} \times f_{SW}}$$

where $\Delta V_{CIN\_MAX} \leq 200mV$

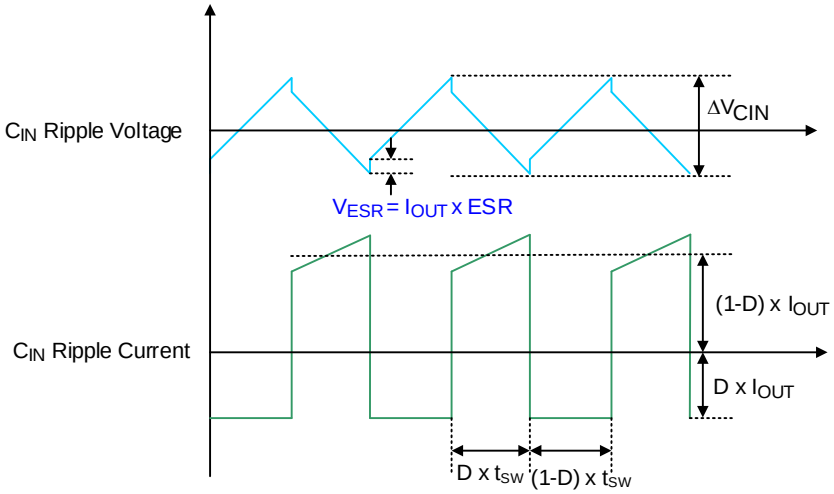


Figure 6. C_{IN} Ripple Voltage and Ripple Current

In addition, the input capacitor needs to have a very low ESR and must be rated to handle the worst-case RMS input current. The RMS ripple current (I_{RMS}) of the regulator can be determined by the input voltage (V_{IN}), output voltage (V_{OUT}), and rated output current (I_{OUT}) using the following equation:

$$I_{RMS} \cong I_{OUT\_MAX} \times \frac{V_{OUT}}{V_{IN}} \times \sqrt{\frac{V_{IN}}{V_{OUT}} - 1}$$

From the above, the maximum RMS input ripple current occurs at maximum output load, which will be used as the requirement to consider the current capabilities of the input capacitors. The maximum ripple voltage usually occurs at a 50% duty cycle, that is, VIN = 2 x VOUT. It is common to use the worse-case $I_{RMS} \cong 0.5 \times I_{OUT\_MAX}$ at $V_{IN} = 2 \times V_{OUT}$ for design. Note that ripple current ratings from capacitor manufacturers are often based on only 2000 hours of life, which makes it advisable to further de-rate the capacitor, or choose a capacitor rated at a higher temperature than required. Several capacitors may also be paralleled to meet size, height, and thermal requirements in the design. For low input voltage applications, sufficient bulk input capacitance is needed to minimize transient effects during output load changes.

Ceramic capacitors are ideal for switching regulator applications due to their small size, robustness, and very low ESR. However, care must be taken when these capacitors are used at the input. A ceramic input capacitor combined with trace or cable inductance forms a high-quality (under damped) tank circuit. If the RTQ2105-QA circuit is plugged into a live supply, the input voltage can ring to twice its nominal value, possibly exceeding the device's rating. This situation is easily avoided by placing the low ESR ceramic input capacitor in parallel with a bulk capacitor with higher ESR to damp the voltage ringing.

The input capacitor should be placed as close as possible to the VIN pin, with a low inductance connection to the PGND of the IC. It is recommended to connect a 4.7μF, X7R capacitor between the VIN pin to the PGND pin for 2.1MHz switching frequency. The larger input capacitance is required when a lower switching frequency is used. For filtering high-frequency noise, an additional small capacitor 0.1μF should be placed close to the part and the capacitor should be 0402 or 0603 in size. X7R capacitors are recommended for best performance across temperature and input voltage variations.

19.5 Output Capacitor Selection

The selection of COUT is determined by considering the voltage ripple and the transient loads. The peak-to-peak output ripple, ΔV_{OUT}, is determined by:

$$\Delta V_{OUT} = \Delta I_L \left(ESR + \frac{1}{8 \times f_{SW} \times C_{OUT}} \right)$$

where the ΔI_L is the peak-to-peak inductor ripple current. The output ripple is highest at maximum input voltage since ΔI_L increases with input voltage. Multiple capacitors placed in parallel may be needed to meet the ESR and RMS current handling requirements.

Regarding to the transient loads, the V_{SAG} and V_{SOAR} requirements should be taken into consideration when choosing the effective output capacitance value. The amount of output sag/soar is a function of the crossover frequency factor at PWM, which can be calculated using the following equation:

$$V_{SAG} = V_{SOAR} = \frac{\Delta I_{OUT}}{2 \times \pi \times C_{OUT} \times f_C}$$

Ceramic capacitors have very low equivalent series resistance (ESR) and provide the best ripple performance. The recommended dielectric type of the capacitor is X7R for the best performance across temperature and input voltage variations. The variation of the capacitance value with temperature, DC bias voltage, and switching frequency needs to be taken into consideration. For example, the capacitance value of a capacitor decreases as the DC bias across the capacitor increases. Be careful to consider the voltage coefficient of ceramic capacitors when choosing the value and case size. Most ceramic capacitors lose 50% or more of their rated value when used near their rated voltage. Transient performance can be improved with a higher value output capacitor. Increasing the output capacitance will also decrease the output voltage ripple.

19.6 Output Voltage Programming

The output voltage can be programmed by a resistive divider from the output to ground with the midpoint connected to the FB pin. The resistive divider allows the FB pin to sense a fraction of the output voltage, as shown in Figure 7. The output voltage is set according to the following equation:

$$V_{OUT} = V_{REF\_CV} \times \left(1 + \frac{R1}{R2} \right)$$

where the reference voltage of the constant voltage control V_{REF_CV} is 0.8V (typical).

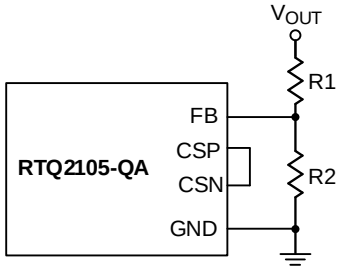


Figure 7. Output Voltage Setting

The placement of the resistive divider should be within 5mm of the FB pin. The resistance of R2 is not larger than 170kΩ for noise immunity considerations. The resistance of R1 can then be obtained as follows:

$$R1 = \frac{R2 \times (V_{OUT} - V_{REF\_CV})}{V_{REF\_CV}}$$

For better output voltage accuracy, the divider resistors (R1 and R2) with ±1% tolerance or better should be used. For general CV regulation without switch-over, cable drop compensation and CC regulation, the CSP pin should be tied to the CSN pin and be left floating to avoid the output voltage inaccuracy. Note that the resistance of R1 relates to the cable drop compensation setting. The resistance of R1 should be designed to match the needs of the voltage drop application. See Adjustable Output Voltage with Cable Drop Compensation (for VOUT = 5V only).

19.7 Compensation Network Design

The purpose of loop compensation is to ensure stable operation while maximizing the dynamic performance. An undercompensated system may result in unstable operations. Typical symptoms of an unstable power supply include audible noise from the magnetic components or ceramic capacitors, jittering in the switching waveforms, oscillation of output voltage, overheating of power FETs and so on. In most cases, the peak current mode control architecture used in the RTQ2105-QA only requires two external components to achieve a stable design, as shown in Figure 8. The compensation can be selected to accommodate any capacitor type or value. The external compensation also allows the user to set the crossover frequency and optimize the transient performance of the device. Around the crossover frequency, the peak current mode control (PCMC) equivalent circuit of a buck converter can be simplified, as shown in Figure 9. The method presented here is easy to calculate and ignores the effects of the slope compensation that is internal to the device. Since the slope compensation is ignored, the actual crossover frequency will usually be lower than the crossover frequency used in the calculations. It is always necessary to make a measurement before releasing the design for final production. Though the models of power supplies are theoretically correct, they cannot take full account of circuit parasitics and component nonlinearity, such as the ESR variations of output capacitors, the nonlinearity of inductors and capacitors, and so on. Also, circuit PCB noise and limited measurement accuracy may also cause measurement errors. A bode plot is ideally measured with a network analyzer, while Richtek application note AN038 provides an alternative way to check the stability quickly and easily. Generally, follow the following steps to calculate the compensation components:

1. Set up the crossover frequency, f_C. For stability purposes, our target is to have a loop gain slope that is −20dB/decade from a very low frequency to beyond the crossover frequency. In general, one-twentieth to one-tenth of the switching frequency (5% to 10% of f_{SW}) is recommended to be the crossover frequency. Do “NOT” design the crossover frequency over 80kHz when the switching frequency is larger than 800kHz. For dynamic purposes, the higher the bandwidth, the faster the load transient response. The downside to high bandwidth is that it increases the regulators susceptibility to board noise, which ultimately leads to excessive falling edge jitter of the switch node voltage.
2. R_{COMP} can be determined by:

$$R_{COMP} = \frac{2\pi \times f_C \times V_{OUT} \times C_{OUT}}{gm \times V_{REF\_CV} \times gm\_CS} = \frac{2\pi \times f_C \times C_{OUT}}{gm \times gm\_CS} \times \frac{R1 + R2}{R2}$$

   where gm is the error amplifier gain of trans-conductance (950μA/V) and gm_cs is COMP to current sense (5.6A/V)
3. A compensation zero can be placed at or before the dominant pole of the buck converter, which is provided by the output capacitor and maximum output loading (R_L). Calculate C_{COMP}:

$$C_{COMP} = \frac{R_L \times C_{OUT}}{R_{COMP}}$$

4. The compensation pole is set to the frequency at the ESR zero or 1/2 of the operating frequency. The output capacitor and its ESR provide a zero, and optional C_{COMP2} can be used to cancel this zero.

$$C_{COMP2} = \frac{R_{ESR} \times C_{OUT}}{R_{COMP}}$$

If 1/2 of the operating frequency is lower than the ESR zero, the compensation pole is set at 1/2 of the operating frequency.

$$C_{COMP2} = \frac{1}{2\pi \times f_P \times R_{COMP}}$$

Note 7. Generally, C_{COMP2} is an optional component to be used to enhance noise immunity.

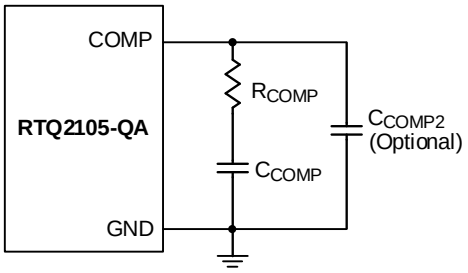


Figure 8. External Compensation Components

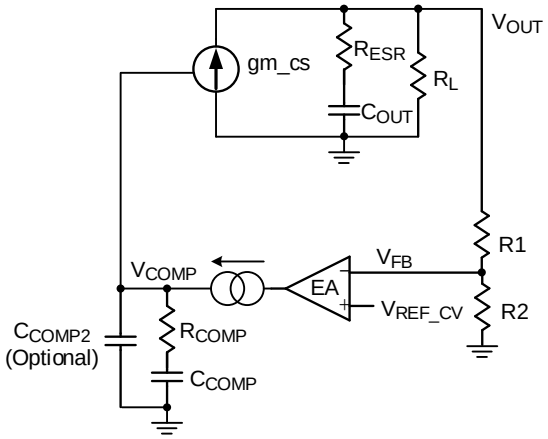


Figure 9. Simplified Equivalent Circuit of Buck with PCMC

19.8 Internal Regulator

The device integrates a 5V linear regulator (VCC) that is supplied by VIN and provides power to the internal circuitry. The internal regulator operates in low dropout mode when V_{IN} is below 5V. The VCC can be used as the PG pull-up supply but it is “NOT” allowed to power other devices or circuitry. The VCC pin must be bypassed to ground with a 1μF X7R capacitor and it needs to be placed as close as possible to the VCC pin. Be careful to account for the voltage coefficient of ceramic capacitors when choosing the value and case size. Many ceramic capacitors lose 50% or more of their rated value when used near their rated voltage.

19.9 Bootstrap Driver Supply

The bootstrap capacitor (C_{BOOT}) between the BOOT pin and the SW pin is used to create a voltage rail above the applied input voltage, V_{IN}. Specifically, the bootstrap capacitor is charged through an internal diode to a voltage

equal to approximately V_{CC} each time the low-side switch is turned on. The charge on this capacitor is then used to supply the required current during the remainder of the switching cycle. For most applications a 0.1μF, 0603 ceramic capacitor with X7R is recommended, and the capacitor should have a 6.3 V or higher voltage rating.

19.10 External Bootstrap Diode (Optional)

It is recommended to add an external bootstrap diode between an external 5V voltage supply and the BOOT pin to enhance the high-side MOSFET switch and improve efficiency when the input voltage is below 5.5V. The recommended application circuit is shown in Figure 10. The bootstrap diode can be a low-cost one, such as 1N4148 or BAT54. The external 5V can be a fixed 5V voltage supply from the system, or a 5V output voltage generated by the RTQ2105-QA. Note that the $V_{BOOT-SW}$ must be lower than 5.5V. Figure 11 shows an efficiency comparison between with and without a Bootstrap Diode.

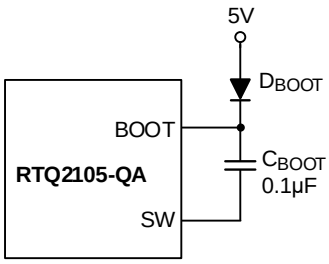


Figure 10. External Bootstrap Diode

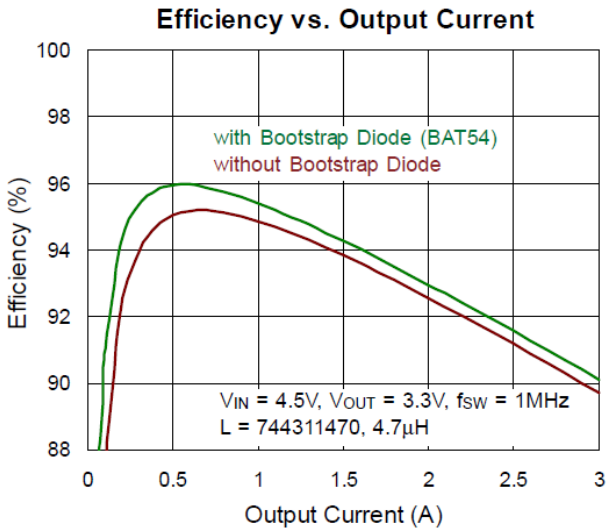


Figure 11. Efficiency Comparison between with and without a Bootstrap Diode

19.11 External Bootstrap Resistor (Optional)

The gate driver of an internal power MOSFET, utilized as a high-side MOSFET switch, is optimized for turning on the switch not only fast enough for reducing switching power loss, but also slow enough for minimizing EMI. The EMI issue is worse when the switch is turned on rapidly due to high di/dt noise induced. When the high-side MOSFET switch is being turned off, the SW node will be discharged relatively slowly by the inductor current due to the presence of the dead time when both the high-side and low-side MOSFET switches are turned off. In some cases, it is desirable to reduce EMI further, even at the expense of some additional power dissipation. The turn-

on rate of the high-side MOSFET switch can be slowed by placing a small bootstrap resistor R_{BOOT} between the BOOT pin and the external bootstrap capacitor, as shown in Figure 12. The recommended range for the R_{BOOT} is several Ωs to 10 Ωs and it can be 0402 or 0603 in size.

This will slow down the turn-on rate of the high-side MOSFET switch and the rise of V_{SW}. In order to improve EMI performance and enhance the internal MOSFET switch, the recommended application circuit is shown in Figure 13, which includes an external bootstrap diode for charging the bootstrap capacitor and a bootstrap resistor R_{BOOT} placed between the BOOT pin and the capacitor/diode connection.

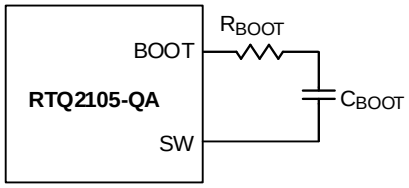


Figure 12. External Bootstrap Resistor at the BOOT Pin

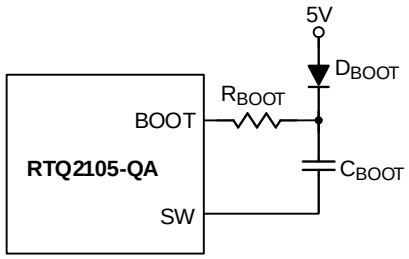


Figure 13. External Bootstrap Diode and Resistor at the BOOT Pin

19.12 EN Pin for Start-Up and Shutdown Operation

For automatic start-up, the EN pin, with high-voltage rating, can be connected to the input supply VIN directly. The large built-in hysteresis band makes the EN pin useful for simple delay and timing circuits. To add an additional delay, the EN pin can be externally connected to VIN by adding a resistor R_{EN} and a capacitor C_{EN}, as shown in Figure 14. The time delay can be calculated with the EN's internal threshold, at which switching operation begins (typically 1.25V). An external MOSFET can be added for the EN pin to be logic-controlled, as shown in Figure 15. In this case, a pull-up resistor, R_{EN}, is connected between VIN and the EN pin. The MOSFET Q1 will be under logic control to pull down the EN pin. To prevent the device from being enabled when VIN is smaller than the V_{OUT} target level or another desired voltage level, a resistive divider (R_{EN1} and R_{EN2}) can be used to externally set the input undervoltage-lockout threshold, as shown in Figure 16.

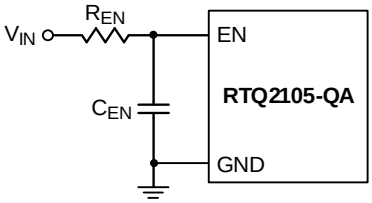


Figure 14. Enable Timing Control

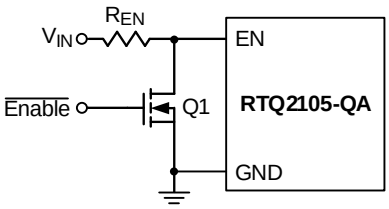


Figure 15. Logic Control for the EN Pin

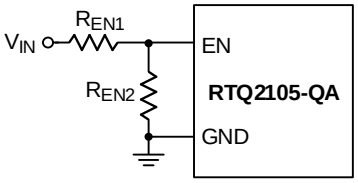


Figure 16. Resistive Divider for Undervoltage-Lockout Threshold Setting

19.13 Soft-Start Function

The RTQ2105-QA provides adjustable soft-start function. The soft-start function is used to prevent large inrush current while the converter is being powered up. For the RTQ2105-QA, the soft-start timing can be programmed by the external capacitor C_{SS} between the SS pin and ground. An internal current source I_{SS} (6μA) charges an external capacitor to build a soft-start ramp voltage. The V_{FB} will track the internal ramp voltage during the soft-start interval. The typical soft-start time, which is the time for V_{OUT} to rise from zero to 90% of the set value, is calculated as follows:

$$t_{SS} = C_{SS} \times \frac{0.8}{I_{SS}}$$

Be aware that the system design should allocate a margin for the minimum capacitance of C_{SS} to avoid the possibility of a failure to boot up within the soft-start period, which can occur due to a design with a slow compensation bandwidth. Additionally, for proper device operation, it is essential that the minimum soft-start time (t_{SS_min}) exceeds 500μs. If a heavy load is added to the output with large capacitance, the output voltage will never enter regulation because of UVP. Thus, the device remains in hiccup operation. The C_{SS} should be large enough to ensure the soft-start period ends after C_{OUT} is fully charged.

$$C_{SS} \geq C_{OUT} \times \frac{I_{SS} \times V_{OUT}}{0.8 \times I_{COUT\_CHG}}$$

where I_{COUT_CHG} is the C_{OUT} charge current, which is related to the switching frequency, inductance, high-side MOSFET switch peak current limit, and load current.

19.14 Power-Good Indicator

The PG pin is an open-drain power-good indicator output and should be connected to an external voltage source through a pull-up resistor. The external voltage source can be an external voltage supply below 5.5V, V_{CC} or the output of the RTQ2105-QA if the output voltage is regulated under 5.5V. It is recommended to connect a 100kΩ between the external voltage source to the PG pin.

19.15 Inductor Peak Current Limit Setting

The current limit of the high-side MOSFET switch is adjustable by an external resistor connected to the RLIM pin. The recommended resistor value ranges from 33kΩ (for typical 5.5A) to 91kΩ (for typical 2.2A). It is recommended

to use resistors with 1% tolerance or better and a temperature coefficient of 100 ppm or less . When the inductor current reaches the current-limit threshold, the VCOMP will be clamped to limit the inductor current. The nductor current ripple current should also be considered when setting the current limit. It is recommended to set the current limit minimum to 1.2 times the peak inductor current. The minimum current limit value can be calculated as follows:

Current limit minimum = ($I_{OUT(MAX)}$ + 1/2 inductor current ripple) x 1.2.

The current limit value can be set by connecting an external resistor R_{LIM} to the RLIM pin.

The approximate formula for the current limit value is given as follows:

$$R_{LIM}(k\Omega) = \frac{178.8}{I_{SET} - 0.2531} - 1$$

where I_{SET} is the desire current limit value (A).

The failure modes and effects analysis (FMEA) consideration should also be applied to the RLIM pin setting to avoid abnormal current limit operation under failure conditions. It includes failure scenarios such as short-circuit to ground and the pin is left open. The inductor peak current limit will be 6.2A (typical) when the RLIM pin is shorted to ground and 1.4A (typical) when the pin is left open. Note that the inductor peak current limit variation increases as the tolerance of R_{LIM} increases. If the R_{LIM} value is small, the inductor peak current limit will probably operate as if the RLIM pin is shorted to ground, and vice versa. The R_{LIM} variation range is limited from 30kΩ to 100kΩ to eliminate the undesired inductor peak current limits. When choosing an R_{LIM} outside the recommended range, make sure there are no issues by evaluating it with a real machine.

19.16 Synchronization

The RTQ2105-QA can be synchronized with an external clock ranging from 300kHz to 2.2MHz, which is applied to the MODE/SYNC pin. The external clock duty cycle must be from 20% to 80% and the amplitude should have valleys that are below V_{IL_SYNC} and peaks above V_{IH_SYNC} (up to 6V). The RTQ2105-QA will not enter PSM operation at light load while synchronized to an external clock. Instead, it will operate in FPWM to maintain regulation.

19.17 Average Current Limit

The RTQ2105-QA implements Constant Current Control to achieve an average current limit. The constant current of constant current mode control is set by an external sense resistance (R_{SENSE}). The average current is set according to the following equation:

$$\text{Average Current Limit} = \frac{V_{REF\_CC}}{R_{SENSE}}$$

where the reference voltage of constant current regulation V_{REF_CC}, is 100mV (typical) and the V_{REF_CC} variation is around ±10%. The average current limit function is recommended to operate with CSP and CSN voltages ranging from 3.3 V to 6V.

19.18 Adjustable Output Voltage with Cable Drop Compensation (for VOUT = 5V only)

The RTQ2105-QA provides a cable drop compensation function during CV regulation. If the trace from the RTQ2105-QA output terminator to the load is too long, there will be a voltage drop on the long trace, which varies with the load current. The RTQ2105-QA is capable of compensating for the output voltage drop to keep a constant voltage at the load, whatever the load current is. The compensation voltage (V_{O_OFFSET}) is based on the cable drop compensation current (I_{LC}) and the upper side resistor R1, which can be calculated using the following formula:

$V_{O\_OFFSET} = I_{LC} \times R1$

The cable drop compensation current variation is ±10%, and it is a function of the current sense voltage (V_{CS}):

I_{LC} (μA) = 21 x (V_{CS} – 0.00476)

where the current sense voltage is the voltage difference between the CSP pin and the CSN pin, which is the voltage across a current sense resistor (R_{SENSE}). Figure 17 shows the relationship between cable drop compensation current (I_{LC}) and VCS.

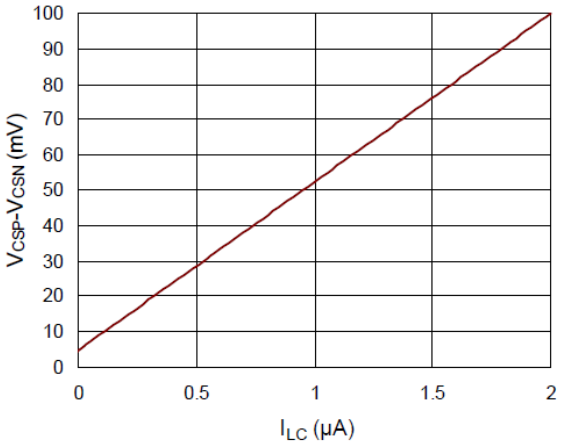


Figure 17. ILC vs. VCSP − VCSN

According to the formula above, the desired compensation voltage, which is set at the rated output current, can be calculated as follows:

V_{O_OFFSET} = 21 x (R_{SENSE} x I_{OUT} – 0.00476) x 10^{-6} x R1

where I_{OUT} is the rated output current.

Choose the R_{SENSE} with the rated load current and reserve some de-rating margin for better thermal and life considerations. The R_{SENSE} selection is suggested between 5mΩ and 30mΩ to minimize power consumption and must also be selected with an appropriate power rating. To avoid undesired CC control loop interruption, the current sense voltage should be selected to be the lower value of 100mV. If the system implements constant current control to achieve average current limit, the R_{SENSE} is set based on the average current limit equation.

Considering CV regulation with cable drop compensation situation, the desire cable drop compensation is 0.5V at a rated 3A loading, and R_{SENSE} is selected as 20mΩ, the R1 can be calculated as follows:

$$R1 = \frac{V_{O\_SFFSET}}{21 \times (R_{SENSE} \times I_{OUT} - 0.00476) \times 10^{-6}} = 431k\Omega$$

The resistance of R2 can then be obtained as follows:

$$R2 = \frac{R1 \times V_{REF\_CV}}{V_{OUT} - V_{REF\_CV}} = 80.09k\Omega$$

In this case, 431kΩ is available for the resistance of R1 and 80.6kΩ is available for the resistance of R2. The R1 and R2 values can be calculated based on above equation. If the R1 and R2 values are too high, the regulator will be more susceptible to noise, and voltage errors from the FB input current will be noticeable. Make sure the current flowing through the FB resistive divider is larger than 5x10-6. In addition, a feed-forward capacitor C_{FF} may be required to improve output voltage ripple in PSM. The power dissipation on sensing resistor will be:

$$P_{RSENSE} = R_{SENSE} \times I_{OUT}^2 = 180mW$$

Choose a current sense resistor power rated with a 50% derating rule of thumb for better heat and life considerations. A 1/2W size is sufficient for this case. Hence, the 20mΩ, 0.5W size RSENSE is determined and with the aid of the cable drop compensation feature, the RTQ2105-QA can compensate the 0.5V voltage drop to maintain excellent output voltage accuracy at the rated 3A load current. Note that the RSENSE should be connected as close to the CSP and CSN pins with short and direct traces, creating a Kelvin connection. This ensures that noise and current sense voltage errors do not corrupt the differential current sense signals between the CS and VOUT pins. The cable drop compensation function is recommended to operate with CSP and CSN voltages ranging from 3.3 V to 6V.

19.19 Thermal Consideration

In many applications, the RTQ2105-QA does not generate much heat due to its high efficiency and low thermal resistance of its WET-WQFN- 24SL 4x4 package. However, in applications in which the RTQ2105-QA is running at a high ambient temperature and high input voltage or high switching frequency, the generated heat may exceed the maximum junction temperature of the part.

The junction temperature should never exceed the absolute maximum junction temperature $T_{J(MAX)}$, listed under Absolute Maximum Ratings, to avoid permanent damage to the device. If the junction temperature reaches approximately 175°C, the RTQ2105-QA stop switching the power MOSFETs until the temperature drops about 15°C cooler.

The maximum power dissipation can be calculated by the following formula:

$P_{D(MAX)} = (T_{J(MAX)} - T_A)/\theta_{JA}$

where $T_{J(MAX)}$ is the maximum allowed junction temperature of the die. For recommended operating condition specifications, the maximum junction temperature is 150°C. TA is the ambient operating temperature, θ_{JA} is the system-level junction to ambient thermal resistance. It can be estimated from thermal modeling or measurements in the system.

The device thermal resistance depends strongly on the surrounding PCB layout and can be improved by providing a heat sink of surrounding copper ground. The addition of backside copper with thermal vias, stiffeners, and other enhancements can also help reduce thermal resistance. Experiments in the Richtek thermal lab show that simply set $\theta_{JA(EFFECTIVE)}$ as 110% to 120% of the $\theta_{JA(EVB)}$ is reasonable to obtain the allowed $P_{D(MAX)}$.

As an example, consider the case when the RTQ2105-QA is used in applications where V_{IN} = 12V, I_{OUT} = 3A, f_{SW} = 2100kHz, V_{OUT} = 5V. The efficiency at 5V, 3A is 89.7% by using Cyntec-VCMT063T-2R2MN5 (2.2μH, 15mΩ DCR) as the inductor and measured at room temperature. The core loss can be obtained from its website of 37.1mW in this case. In this case, the power dissipation of the RTQ2105-QA is

$$P_{D,\,RT} = \frac{1-\eta}{\eta} \times P_{OUT} - \left(I_O^2 \times DCR + P_{CORE}\right) = 1.55W$$

Considering the $\theta_{JA(EFFECTIVE)}$ is 38.2°C/W by using the RTQ2105-QA evaluation board with 4 layers PCB, 2 OZ. Cu on the outer layers and 1 OZ. Cu on the inner layers, the junction temperature of the regulator operating in a 25°C ambient temperature is approximately:

T_J = 1.55W x 38.2°C/W + 25°C = 84.2°C

Figure 18 shows the RTQ2105-QA $R_{DS(ON)}$ versus different junction temperature. If the application calls for a higher ambient temperature, we might recalculate the device power dissipation and the junction temperature based on a higher $R_{DS(ON)}$ since it increases with temperature. Using 60°C ambient temperature as an example, the

change of the equivalent $R_{DS(ON)}$ can be obtained from Figure 18 and yields a new power dissipation of 1.658W. Therefore, the estimated new junction temperature is

T_J' = 1.658W x 38.2°C/W + 60°C = 123.3°C

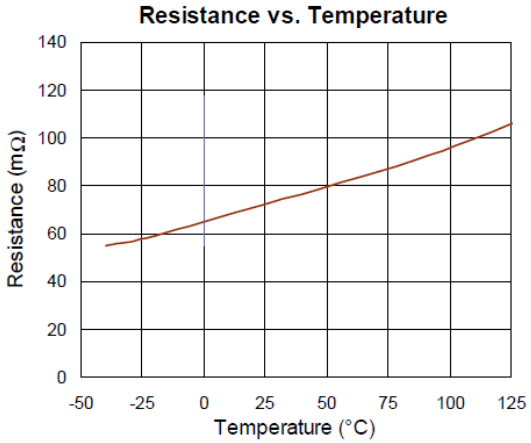


Figure 18. RDS(ON) vs. Temperature

If the application calls for a higher ambient temperature and may exceed the recommended maximum junction temperature of 150°C, care should be taken to reduce the temperature rise of the part by using a heat sink or air flow. Note that the over-temperature protection is intended to protect the device during momentary overload conditions. The protection is activated outside of the absolute maximum range of operation as a secondary failsafe and therefore should not be relied upon operationally.

Continuous operation above the specified absolute maximum operating junction temperature may impair device reliability or permanently damage the device.

19.20 Layout Considerations

When laying out the printed circuit board, the following checklist should be used to ensure proper operation of the RTQ2105-QA:

- Four-layer or six-layer PCB with maximum ground plane is strongly recommended for good thermal performance.
- Keep the traces of the main current paths wide and short.
- Place high-frequency decoupling capacitor C_{IN2} as close as possible to the IC to reduce the loop impedance and minimize switch node ringing.
- Place the C_{VCC} as close to the VCC pin as possible.
- Place the bootstrap capacitor, C_{BOOT}, as close to the IC as possible. Route the trace with a width of 20mil or wider.
- Place multiple vias under the device near VIN and PGND, and near input capacitors to reduce parasitic inductance and improve thermal performance. To keep thermal resistance low, extend the ground plane as much as possible, and add thermal vias under and near the RTQ2105-QA to additional ground planes within the circuit board and on the bottom side.
- The high frequency switching nodes, SW and BOOT, should be as small as possible. Keep analog components away from the SW and BOOT nodes.

- Reducing the area size of the SW exposed copper to reduce the electrically coupling from this voltage.
- Connect the feedback sense network behind the via of the output capacitor.
- Place the feedback components R_{FB1}/R_{FB2}/C_{FF} near the IC.
- Place the compensation components R_{CP1}/C_{CP1}/C_{CP2} near the IC.
- Connect all analog grounds to a common node and then connect the common node to the power ground with a single point.
- Minimize current sense voltage errors by using Kelvin connection for PCB routing of the CSP pin, CSN pin, and current sense resistor (R_{SENSE}).

Figure 19 to Figure 22 are the layout example using a 70mm x 100mm, four-layer PCB with 2 OZ. Cu on the outer layers and 1 OZ. Cu on the inner layers.

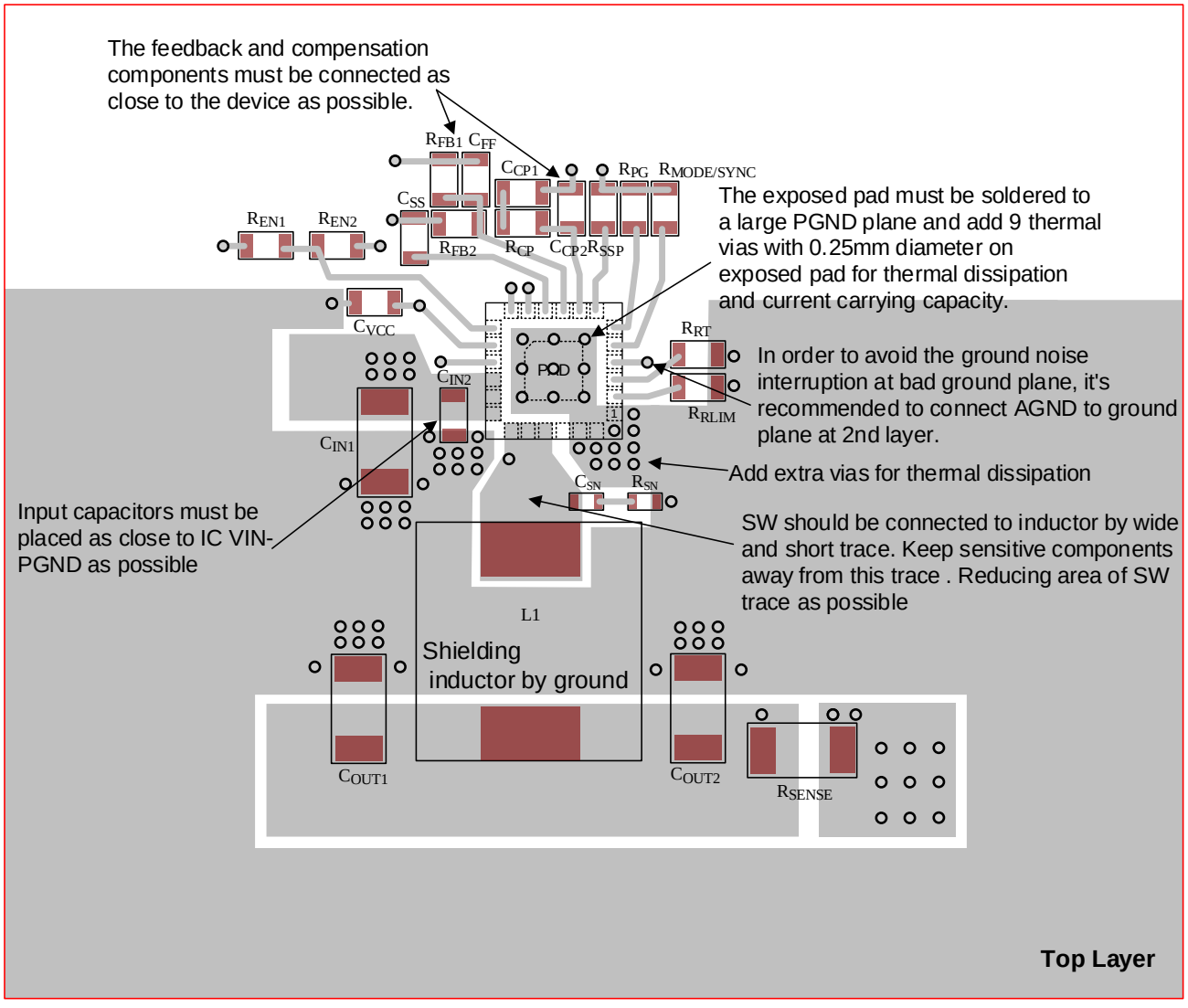


Figure 19. Layout Guide (Top Layer)

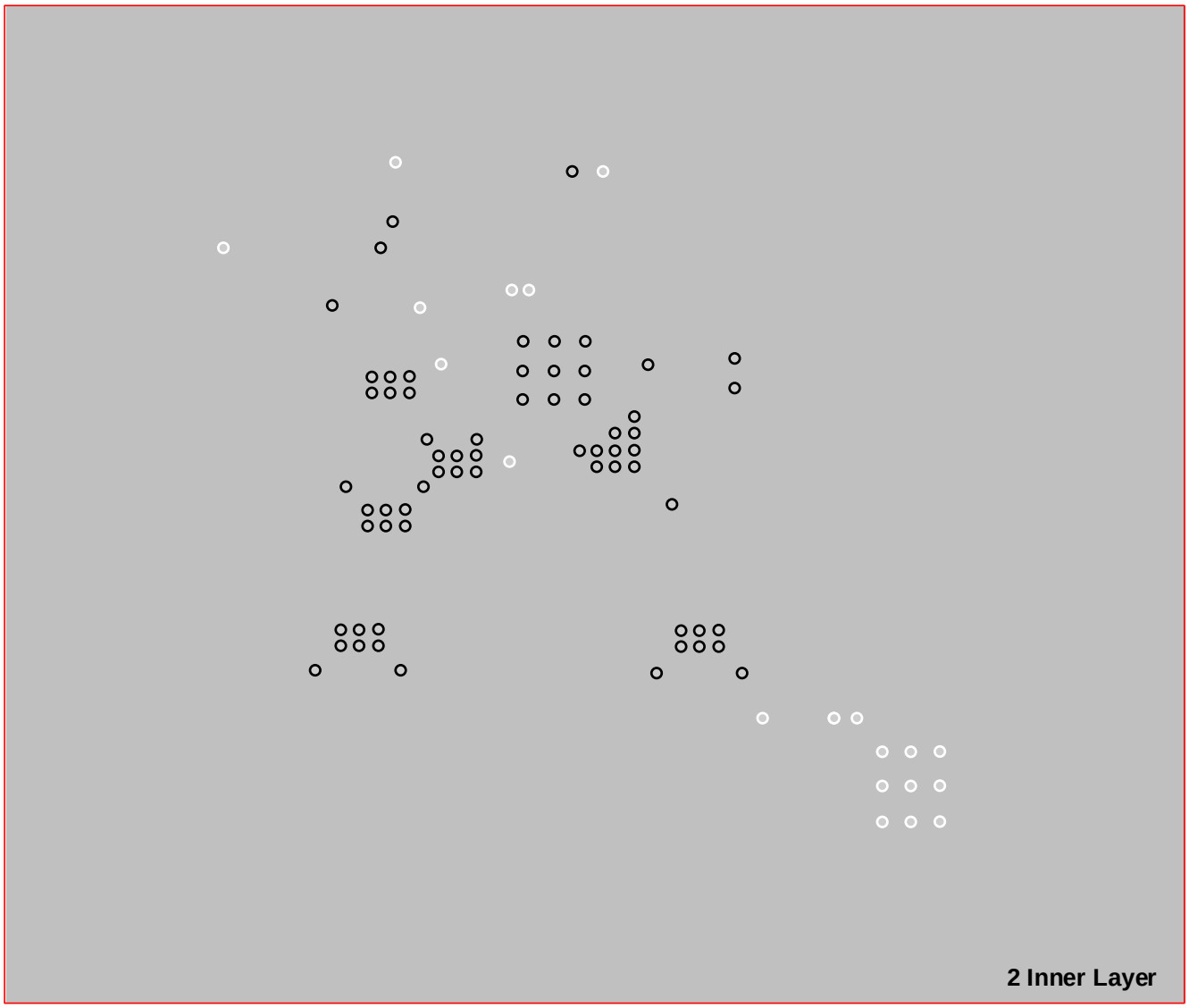


Figure 20. Layout Guide (2 Inner Layer)

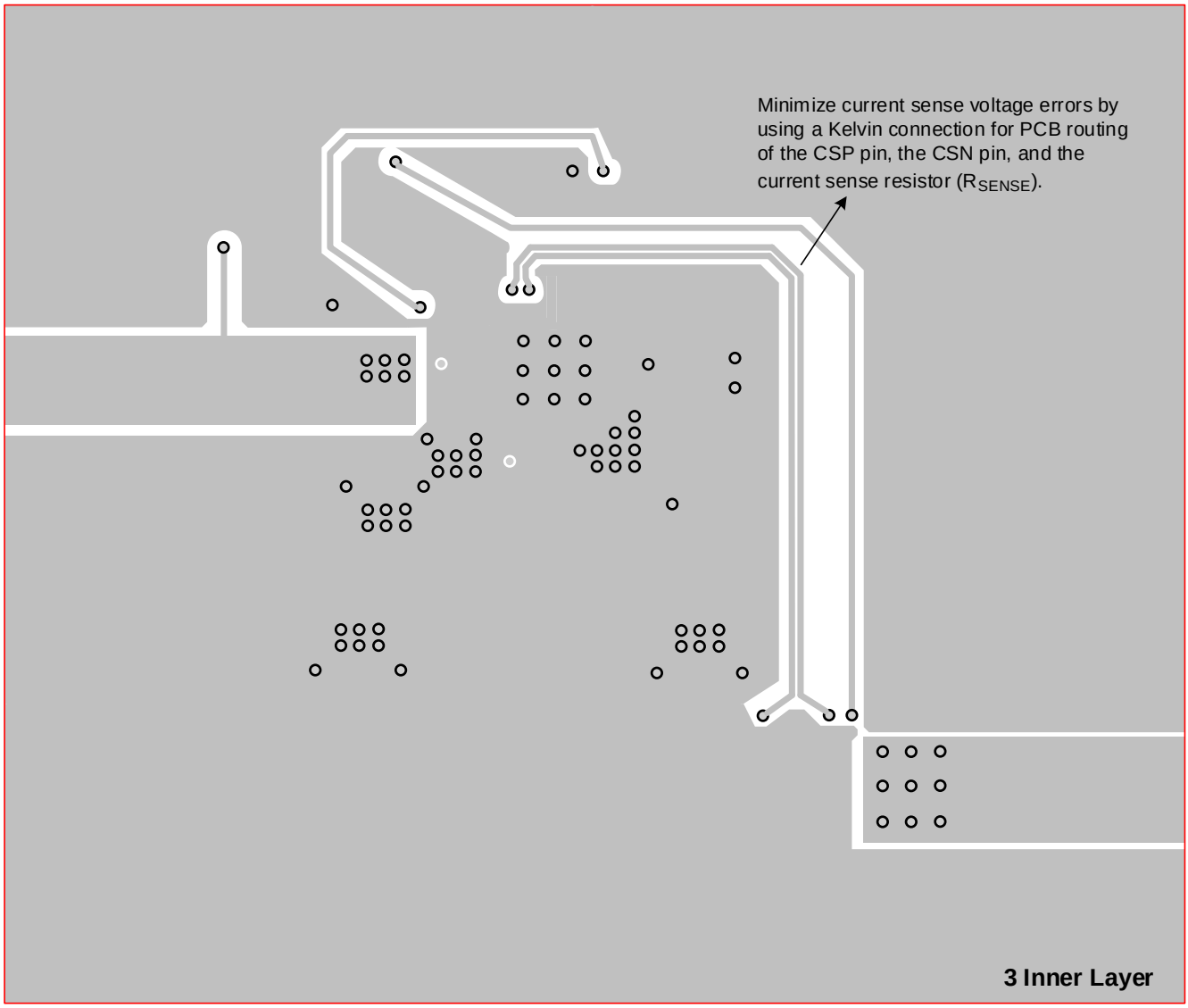


Figure 21. Layout Guide (3 Inner Layer)

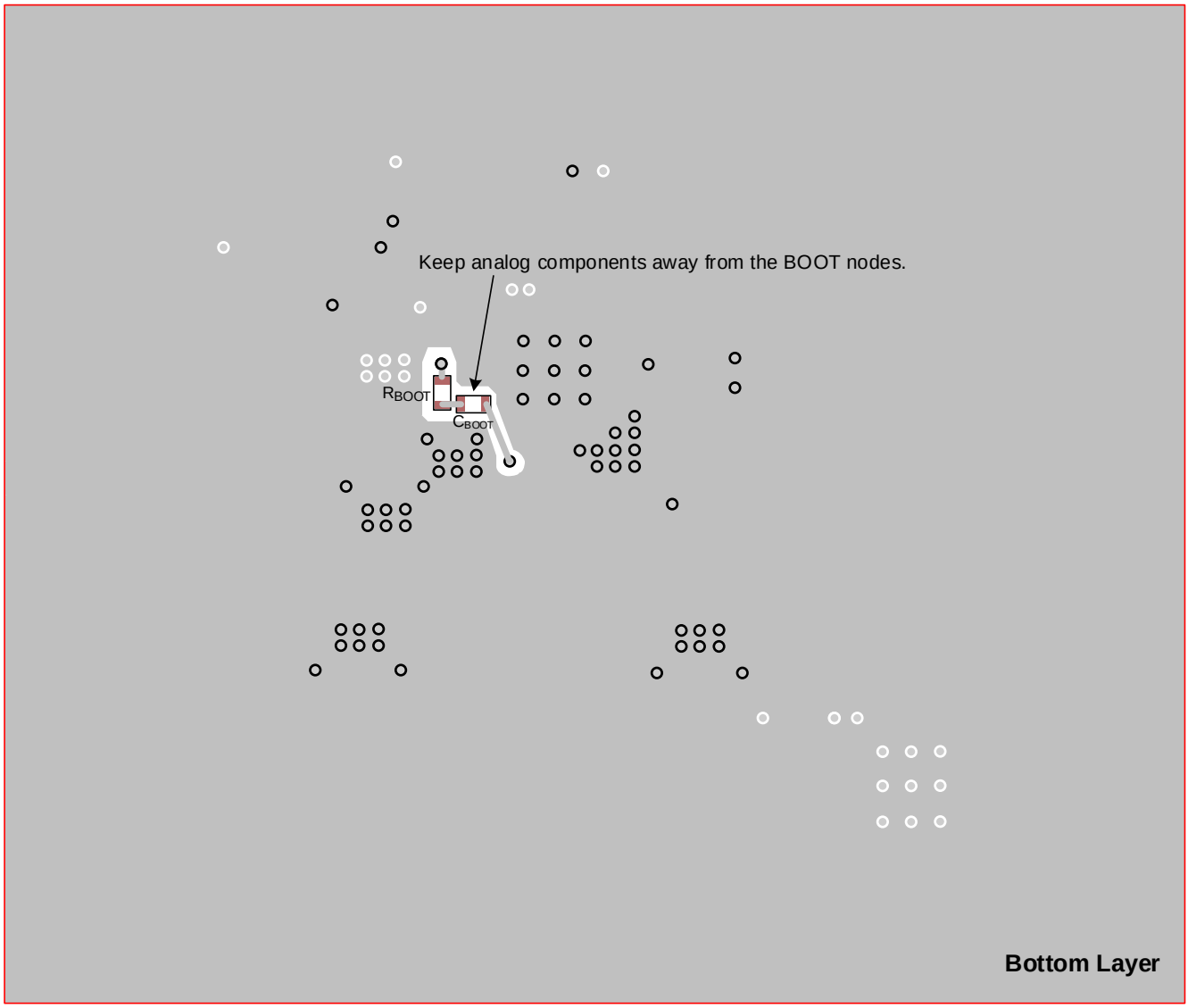
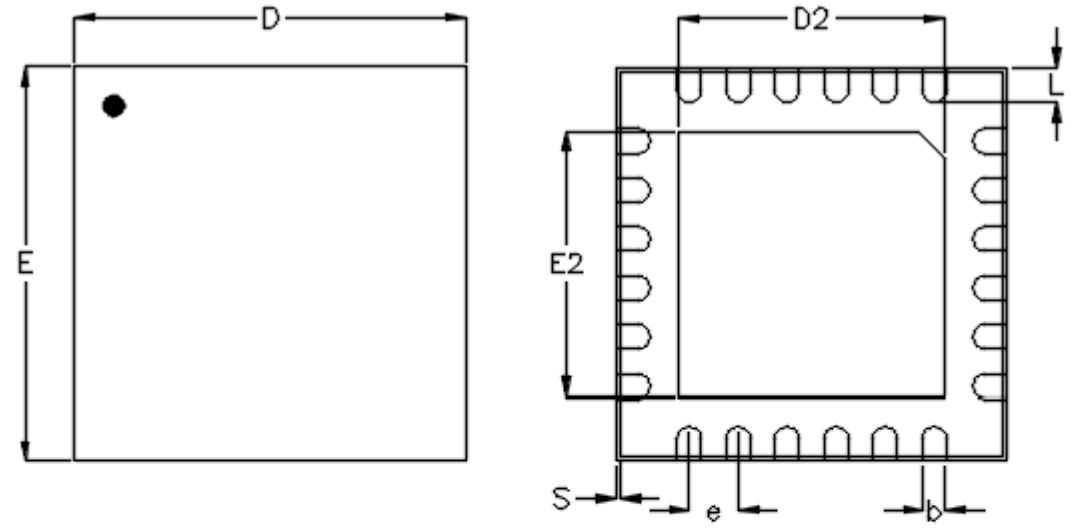


Figure 22. Layout Guide (Bottom Layer)

Note 8. The information provided in this section is for reference only. The customer is solely responsible for the designing, validating, and testing your product incorporating Richtek's product and ensure such product meets applicable standards and any safety, security, or other requirements.

20 Outline Dimension



DETAIL A

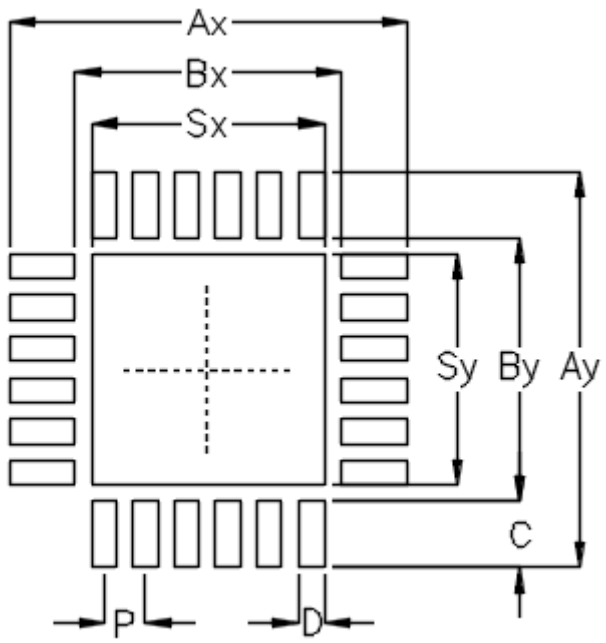
Pin #1 ID and Tie Bar Mark Options

Note : The configuration of the Pin #1 identifier is optional, but must be located within the zone indicated.

Symbol	Dimensions In Millimeters		Dimensions In Inches	
	Min	Max	Min	Max
A	0.700	0.800	0.028	0.031
A1	0.000	0.050	0.000	0.002
A3	0.175	0.250	0.007	0.010
b	0.200	0.300	0.008	0.012
D	3.900	4.100	0.154	0.161
D2	2.650	2.750	0.104	0.108
E	3.900	4.100	0.154	0.161
E2	2.650	2.750	0.104	0.108
e	0.500		0.020	
L	0.300	0.400	0.012	0.016
R	0.050	0.150	0.002	0.006
S	0.001	0.090	0.000	0.004

WET W-Type 24SL QFN 4x4 Package

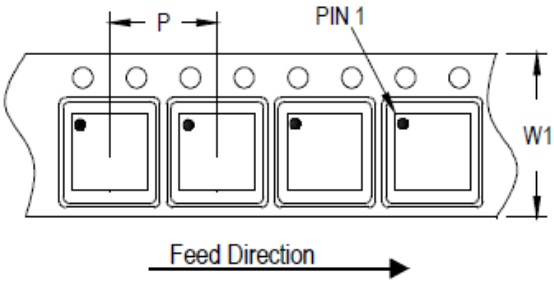
21 Footprint Information

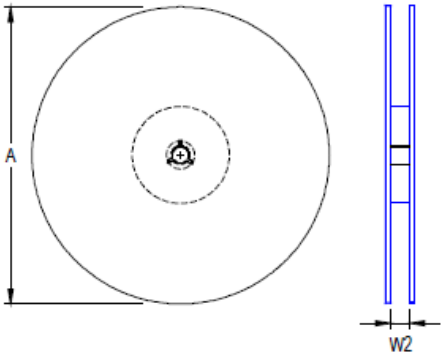


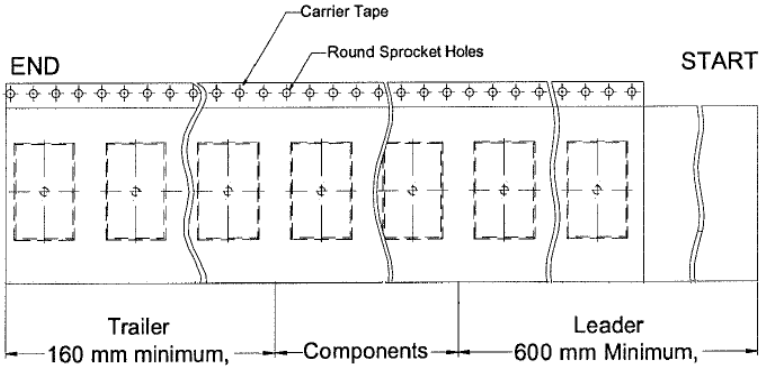
Package	Number of Pin	Footprint Dimension (mm)									Tolerance
		P	Ax	Ay	Bx	By	C	D	Sx	Sy	
WET-WQFN4x4-24S	24	0.50	4.80	4.80	3.20	3.20	0.80	0.30	2.80	2.80	±0.05

22 Packing Information

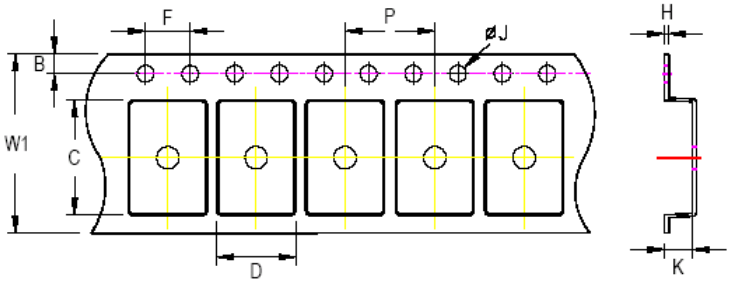
22.1 Tape and Reel Data







Package Type	Tape Size (W1) (mm)	Pocket Pitch (P) (mm)	Reel Size (A)		Units per Reel	Trailer (mm)	Leader (mm)	Reel Width (W2) Min/Max (mm)
			(mm)	(in)				
(V, W) QFN/DFN 4x4	12	8	180	7	1,500	160	600	12.4/14.4



C, D, and K are determined by component size. The clearance between the components and the cavity is as follows:

- For 12mm carrier tape: 0.5mm maximum

Tape Size	W1	P		B		F		ØJ		K		H
	Max	Min	Max	Min	Max	Min	Max	Min	Max	Min	Max	Max
12mm	12.3mm	7.9mm	8.1mm	1.65mm	1.85mm	3.9mm	4.1mm	1.5mm	1.6mm	1.0mm	1.3mm	0.6mm

22.2 Tape and Reel Packing

Step	Photo/Description	Step	Photo/Description
1	 Reel 7”	4	 3 reels per inner box **Box A**
2	 HIC & Desiccant (1 Unit) inside	5	 12 inner boxes per outer box
3	 Caution label is on backside of Al bag	6	 Outer box **Carton A**

Container / Package	Reel		Box			Carton		
	Size	Units	Item	Reels	Units	Item	Boxes	Unit
(V, W) QFN/DFN 4x4	7”	1,500	Box A	3	4,500	Carton A	12	54,000
			Box E	1	1,500	For Combined or Partial Reel.		

22.3 Packing Material Anti-ESD Property

Surface Resistance	Aluminum Bag	Reel	Cover tape	Carrier tape	Tube	Protection Band
Ω/cm^2	**10^4 to 10^{11}**	**10^4 to 10^{11}**	**10^4 to 10^{11}**	**10^4 to 10^{11}**	**10^4 to 10^{11}**	**10^4 to 10^{11}**

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23 Datasheet Revision History

Version	Date	Description	Item
03	2025/4/11	Modify	*Changed the Step-Down Converter to Buckt Converter* *General Description on page 1* *Features on page 1* *Simplified Application Circuit on page 2* - Added *Simplified Application Circuit* *Recommended Operating Conditions on page 7* *Electrical Characteristics on page 8* *Application Information on page 27 to 44* *Packing Information on page 47* - Added Tape Size “K”