RICHTEK®

RT6220

23V, 6A, 500kHz, ACOT® Synchronous Buck Converter

1 General Description

The RT6220 is a synchronous buck converter with Advanced Constant On-Time (ACOT®) mode control, which provides a very fast transient response with no external compensators. The RT6220 operates from 4.5V to 23V input voltage, provides complete protection functions including Current Limit Protection, Undervoltage Protection (UVP) and Overvoltage Protection (OVP). This IC also provides a 1.5ms internal soft-start function and an open-drain power-good indicator. The recommended junction temperature range is –40°C to 125°C, and the ambient temperature range is –40°C to 85°C.

2 Ordering Information

RT6220□□□

Package Type(1)
QUF: UQFN-16L 3x3 (FC) (U-Type)

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

PWM Operation / VOUT Protection
A: DEM/Latch
AH: DEM/Hiccup
BL: FCCM/Latch
BH: FCCM/Hiccup

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.

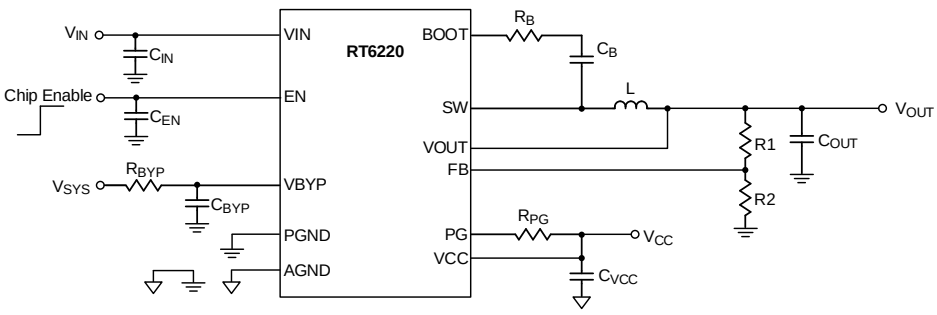
3 Features

- **4.5V to 23V Input Voltage Range**
- **Adjustable from 0.6V to 5V Output Range**
- **Up to 98% Duty for 2S Battery Application**
- **500kHz Switching Frequency**
- **ACOT® Mode Performs Fast Transient Response**
- **Integrated MOSFETs**
  - **31mΩ of High-Side MOSFET**
  - **20mΩ of Low-Side MOSFET**
- **Supports MLCC Output Capacitors**
- **Internal Soft-Start (1.5ms Typical)**
- **Built-In OVP/UVP/Current Limit Protection**
- **Power-Good Indicator**
- **Over-Temperature Protection**

4 Applications

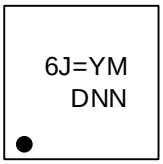
- Laptop Computers
- Tablet PCs
- Networking Systems
- Servers
- Personal Video Recorders
- Flat Panel Television and Monitors
- Distributed Power Systems

5 Simplified Application Circuit



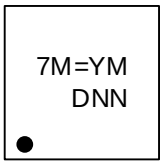
6 Marking Information

RT6220AGQUF



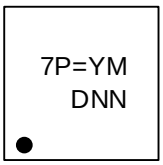
6J= : Product Code
YMDNN : Date Code

RT6220BLGQUF



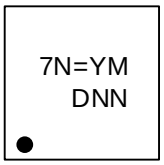
7M= : Product Code
YMDNN : Date Code

RT6220AHGQUF



7P= : Product Code
YMDNN : Date Code

RT6220BHGQUF

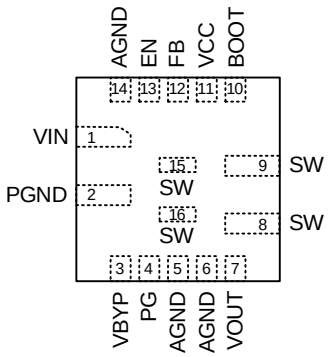


7N= : Product Code
YMDNN : Date Code

Table of Contents

7 Pin Configuration

(TOP VIEW)

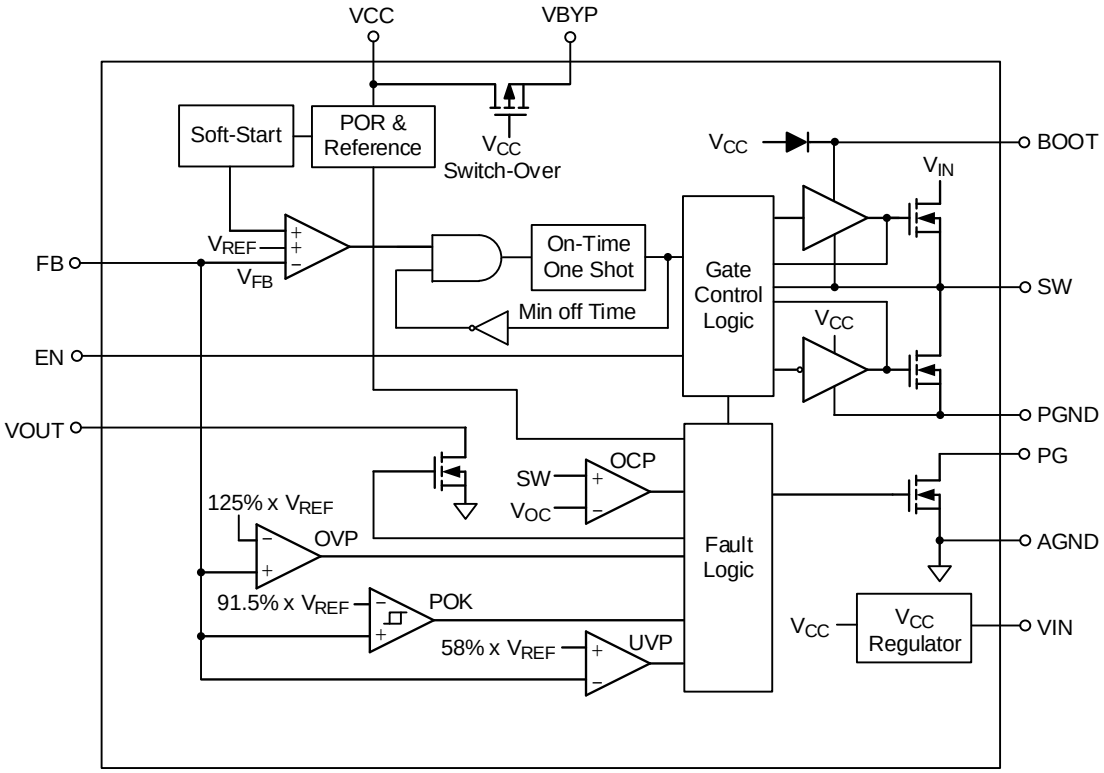


UQFN-16L 3x3 (FC)

8 Functional Pin Description

Pin No.	Pin Name	Pin Function
1	VIN	Power input connected to the high-side MOSFET drain. Place 2pcs 10μF MLCC decoupling capacitors near the input pin.
2	PGND	Power ground. Connect the power ground pin with a wide and thick trace, adding thermal vias for better heat dissipation.
3	VBYP	Switch over input supply voltage for VCC. A low-pass filter should be connected to AGND if V_{BYP} is applied, the recommended RC filter value is R_{BYP} = 5.1 and C_{BYP} = 2.2μF. If VBYP is not used, then connect this pin to AGND. Do not connect to the VCC pin.
4	PG	This pin should be connected to a pull high voltage with a 100kΩ resistor. Recommend to pull high by VCC (5V). DO NOT pull high to an external voltage, which is higher than VCC (5V).
5, 6, 14	AGND	Analog ground.
7	VOUT	Output voltage sense input. An internal discharging circuit is connected to this pin.
8, 9,15, 16	SW	Switch node.
10	BOOT	Bootstrap supply for high-side gate driver. A capacitor is needed to drive the power switch's gate above the supply voltage. It is connected between the SW and BOOT pins to form a floating supply across the power switch driver. The recommended design value is R_{BOOT} = 2.2Ω and C_{BOOT} = 0.1μF.
11	VCC	5V linear regulator output for the internal control circuit. Bypass VCC to AGND with a 2.2μF capacitor. VCC can only supply internal circuits. Do not connect to external loads.
12	FB	Feedback voltage input.
13	EN	Enable control input. Do not leave this pin floating. The slew rate of EN is recommended to be slower than 4.8V/μs. Users should add an RC circuit to avoid glitch noise.

9 Functional Block Diagram



10 Absolute Maximum Ratings

(Note 2)

- VIN to PGND ---------- –0.3V to 27V
- SW to PGND
  - DC ---------- –1V to 27.3V
  - AC(<30ns) ---------- –5V to 28V
- BOOT to PGND
  - DC ---------- –0.6V to 33.3V
  - AC(<30ns) ---------- –5V to 34V
- BOOT to SW ---------- –0.3V to 6V
- EN, FB to AGND ---------- –0.3V to 6V
- VBYP to AGND ---------- –0.3V to 5.3V
- VCC, PG, VOUT to AGND ---------- –0.3V to 6V
- PGND to AGND ---------- –0.3V to 0.3V
- Power Dissipation, P_D @ T_A = 25°C
  - UQFN-16L 3x3 (FC) ---------- 2.33W
- Package Thermal Resistance (Note 3)
  - UQFN-16L 3x3 (FC), $\theta_{JA(EVB)}$ ---------- 43°C/W
  - UQFN-16L 3x3 (FC), $\Psi_{JC(Top)}$ ---------- 0.1°C/W
  - UQFN-16L 3x3 (FC), θ_{JA} ---------- 70°C/W
  - UQFN-16L 3x3 (FC), $\theta_{JC(Top)}$ ---------- 15°C/W
- Lead Temperature (Soldering, 10 sec.) ---------- 260°C
- Junction Temperature ---------- 150°C
- Storage Temperature Range ---------- –65°C to 150°C
- ESD Susceptibility (Note 4)
  - HBM (Human Body Model) ---------- 2kV

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.

Note 3. For more information about thermal parameters, see the Application and Definition of Thermal Resistances report, AN061.

Note 4. Devices are ESD sensitive. Handling precautions are recommended.

11 Recommended Operating Conditions

(Note 5)

- Supply Input Voltage, VIN ---------- 4.5V to 23V
- Ambient Temperature Range ---------- –40°C to 85°C
- Junction Temperature Range ---------- –40°C to 125°C

Note 5. The device is not guaranteed to function outside its operating conditions.

12 Electrical Characteristics

(V_{IN} = 12V, T_A = 25°C, unless otherwise specified.)

Parameter	Symbol	Test Conditions	Min	Typ	Max	Unit
Supply Current						
Shutdown Current	I_{SHDN}	V_{EN} = 0V	--	2.5	5	μA
Quiescent Current	I_{Q_NSW}	RT6220A/AH, V_{EN} = 2V, no switching	--	100	130	μA
Quiescent Current	I_{Q_NSW}	RT6220BL/BH, V_{EN} = 2V, no switching	--	110	150	μA
BOOT to SW Leakage Current						
BOOT to SW Leakage Current	$I_{B\_S\_LK}$	V_{BYP} = 5V, V_{EN} = 0V	--	--	2.5	μA
Switch On-Resistance						
On-Resistance of High-Side MOSFET	R_{DSON_H}	$V_{BOOT} - V_{SW}$ = 5V	--	31	--	mΩ
On-Resistance of Low-Side MOSFET	R_{DSON_L}		--	20	--	mΩ
High-Side MOSFET Leakage Current	I_{HS_LK}	V_{IN} = 12V, V_{EN} = 0V	--	--	1	μA
Current Limit						
Current Limit	I_{LIM}	Valley current of low-side switch	7.6	--	11.4	A
Switching Frequency and Minimum Off Timer						
Switching Frequency	f_{SW}		450	500	550	kHz
Minimum Off-Time	t_{OFF_MIN}		--	200	--	ns
Protections						
Output Overvoltage Protection Threshold	V_{OVP}	With respect to output voltage	120	125	130	%
Overvoltage Protection Delay Time	t_{DLY_OVP}		--	5	--	μs
Output Undervoltage Protection Threshold	V_{UVP}	With respect to output voltage	53	58	63	%
Undervoltage Protection Delay Time	t_{DLY_UVP}		--	5	--	μs
Reference and Soft-Start						
Reference Voltage	V_{REF}		0.594	0.600	0.606	V
Soft-Start Time	t_{SS}	Form EN high to PG high	1	1.5	2	mV
Enable and UVLO						
EN Threshold Voltage	V_{EN}		1.25	1.35	1.45	V
EN Threshold Hysteresis	V_{EN_HYS}		50	200	250	mV
EN Input Current	I_{EN}	V_{EN} = 2V	--	1	--	μA
EN Input Current	I_{EN}	V_{EN} = 0V	--	0	--	μA
VCC Undervoltage-Lockout Rising Threshold	$V_{VCC\_UVLO\_R}$		3.8	4.2	4.45	V
VCC Undervoltage-Lockout Threshold	$V_{VCC\_UVLO\_HYS}$		75	400	650	mV

Parameter	Symbol	Test Conditions	Min	Typ	Max	Unit
VCC Regulator						
VCC Regulator	V_{VCC}		4.805	5	5.295	V
VCC Bypass Switch Turn-on Voltage	$V_{BYP\_VCC\_ON}$	V_{BYP} rising edge	4.4	4.6	4.8	V
VCC Bypass Switch Switchover Hysteresis	$V_{BYP\_VCC\_HYS}$		150	200	400	mV
VCC Bypass Switch-on Resistance	R_{BYP_VCC}		--	3	5	Ω
Power-Good Indicator						
Power-Good Voltage Threshold	V_{PG}	VOUT rising	86.5	91.5	96.5	%
Power-Good Voltage Hysteresis	V_{PG_HYS}	VOUT falling	--	10	--	%
Power-Good Low to High Delay Time	t_{DLY_PG}		--	0.5	--	ms
Power-Good Sink Current Capability	V_{SINK_PG}	Sink 4mA	--	--	0.4	V
Power-Good Leakage Current	I_{PG_LK}	V_{PG} = 5V	--	--	100	nA
Over-Temperature Protection						
Over-Temperature Protection Threshold	T_{OTP}	T_J rising	135	150	--	°C
Over-Temperature Protection Hysteresis	T_{OTP_HYS}		--	25	--	°C

13 Typical Application Circuit

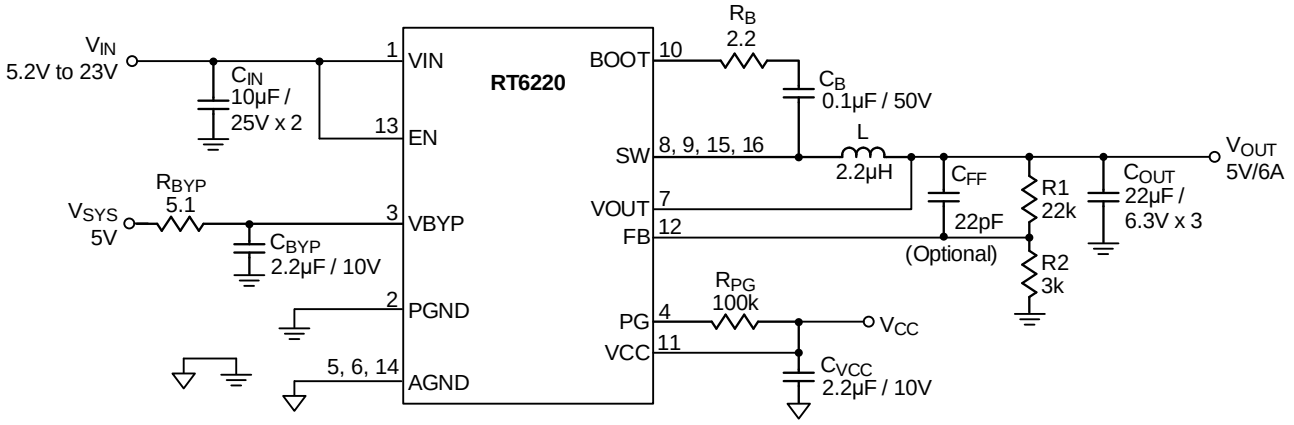


Figure 1. Typical Application Circuit for V_{OUT} = 5V

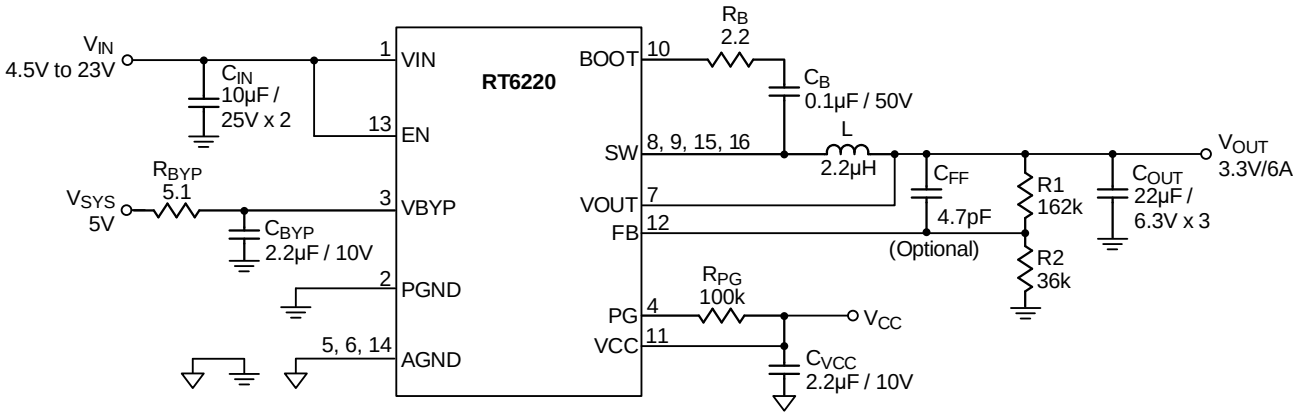


Figure 2. Typical Application Circuit for V_{OUT} = 3.3V

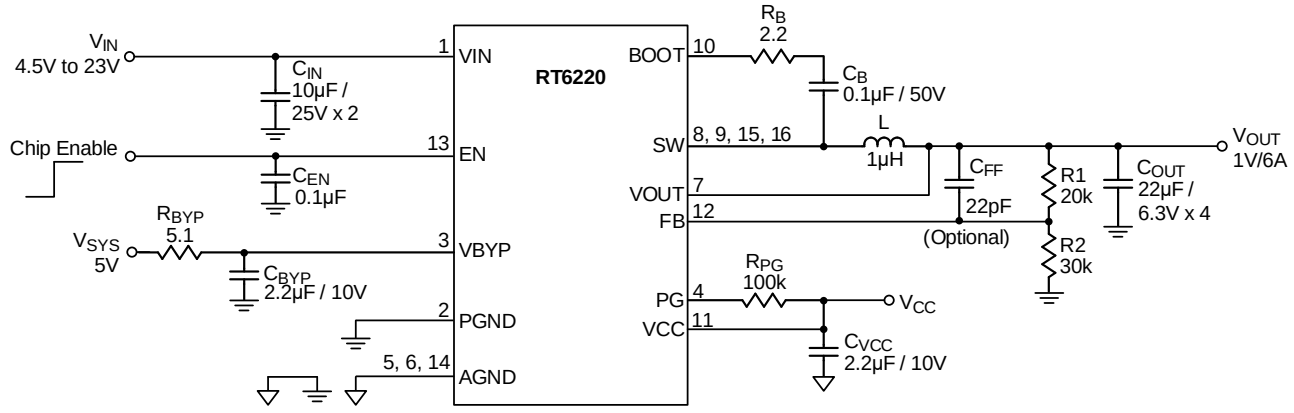
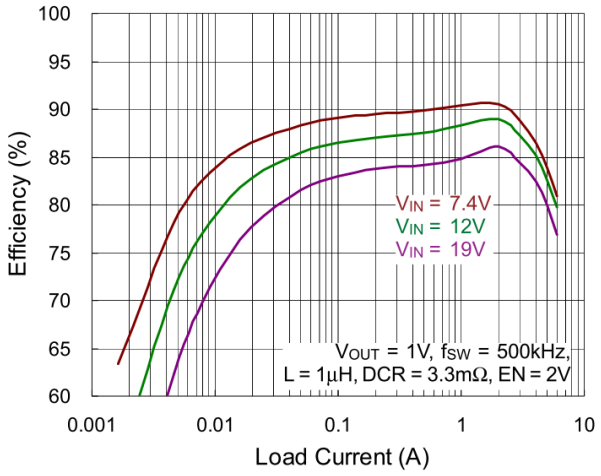


Figure 3. Typical Application Circuit for V_{OUT} = 1V

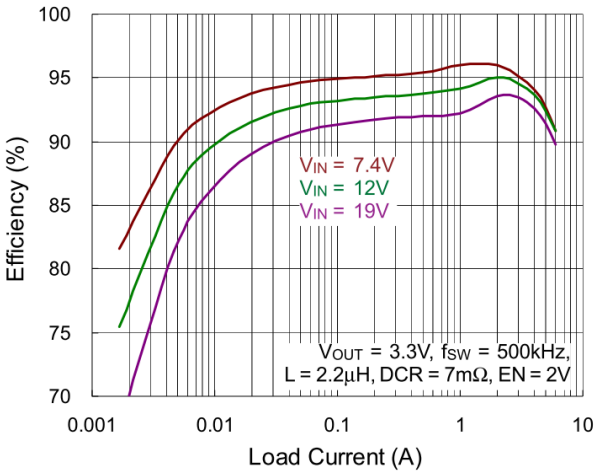
14 Typical Operating Characteristics

Performance waveforms are tested on the evaluation board of the Typical Application Circuit using the RT6220A., VIN = 12V, VOUT = 1V, L = 1µH, T_J = 25°C, unless otherwise noted.

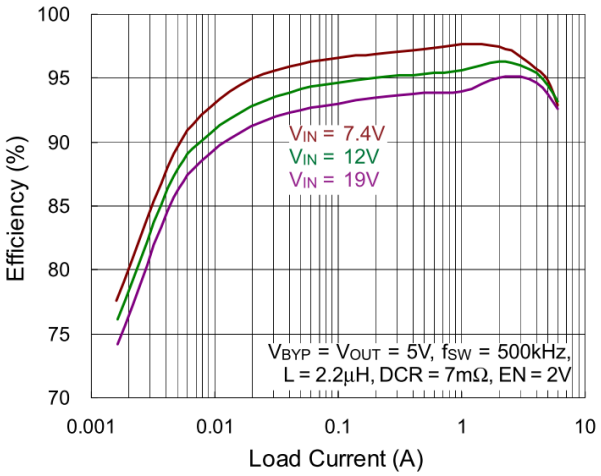
Efficiency vs. Load Current



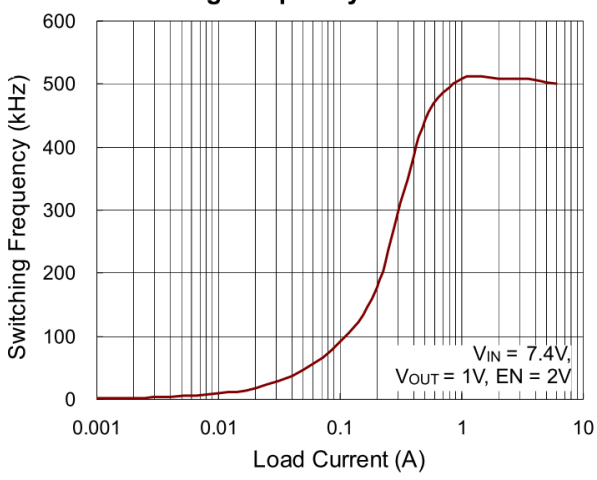
Efficiency vs. Load Current



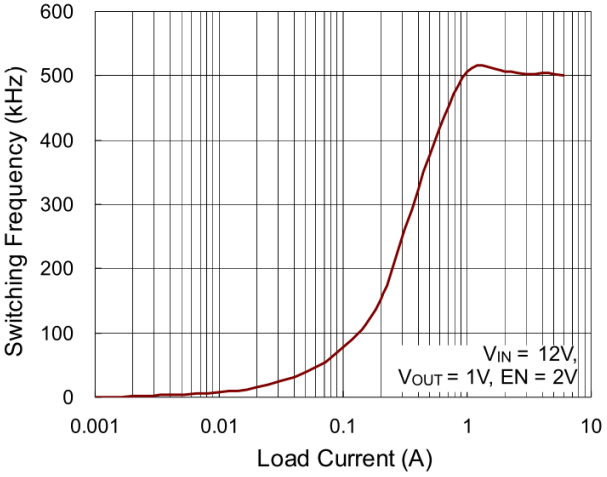
Efficiency vs. Load Current



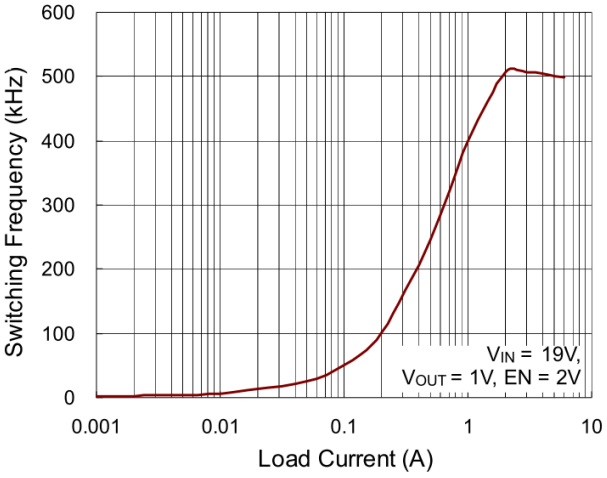
Switching Frequency vs. Load Current



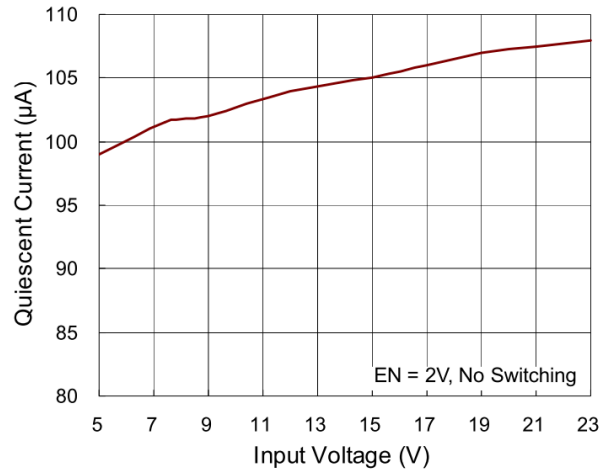
Switching Frequency vs. Load Current



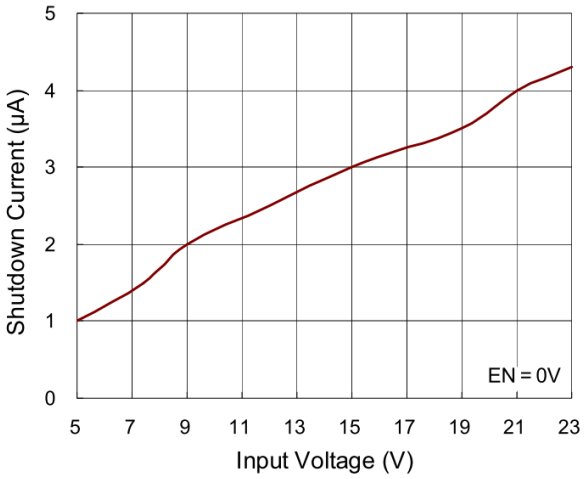
Switching Frequency vs. Load Current



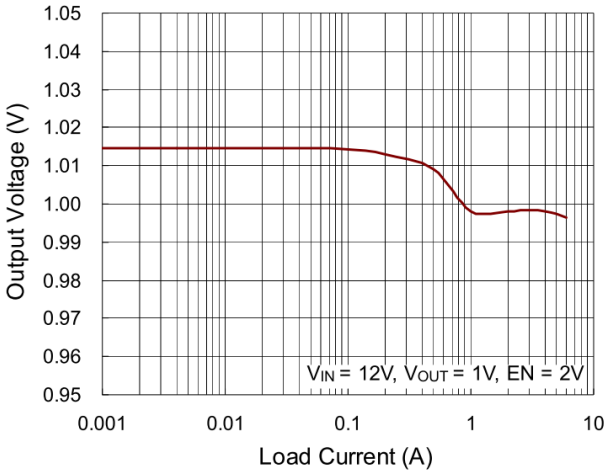
Quiescent Current vs. Input Voltage



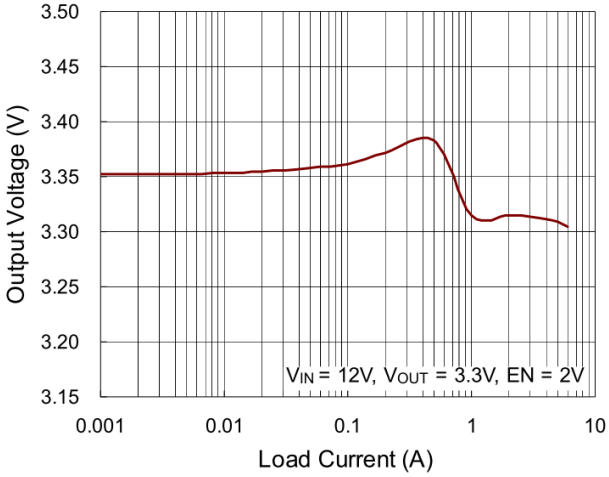
Shutdown Current vs. Input Voltage



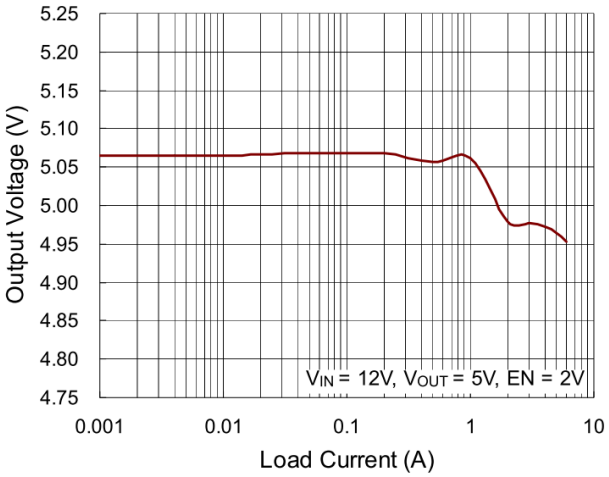
Output Voltage vs. Load Current



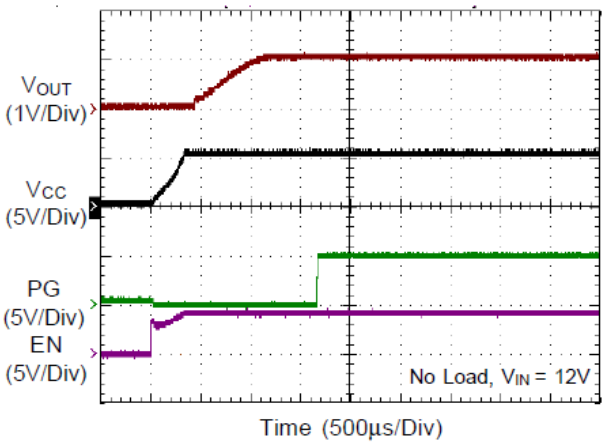
Output Voltage vs. Load Current



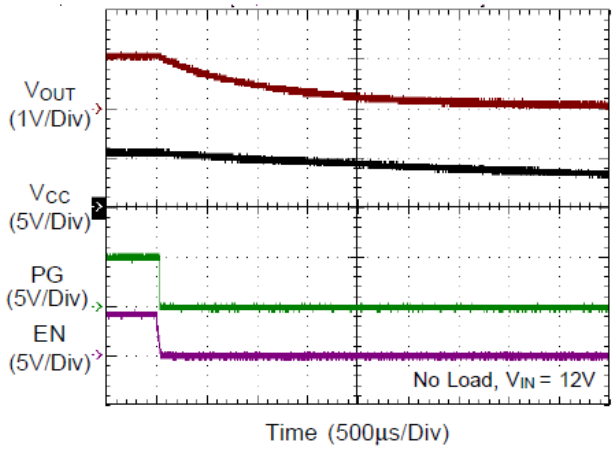
Output Voltage vs. Load Current



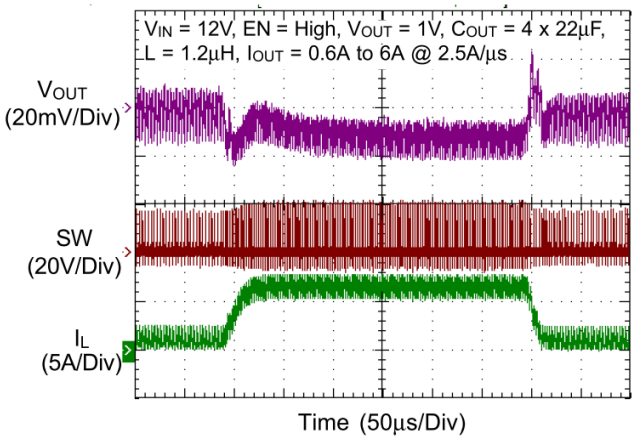
Power On Through EN



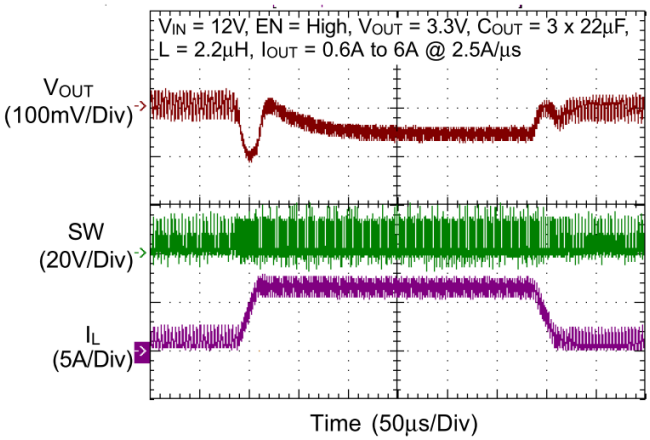
Power Off Through EN



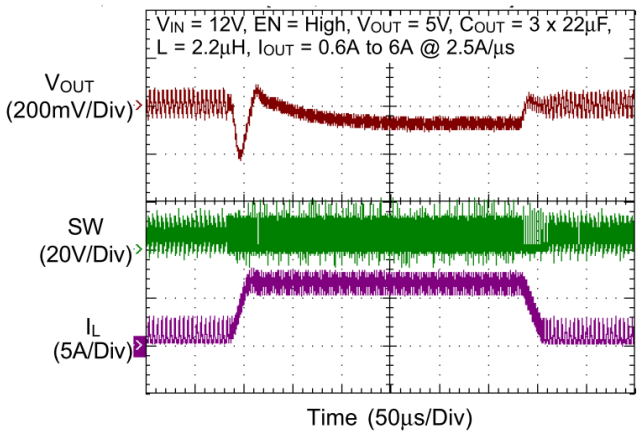
Load Transient Response



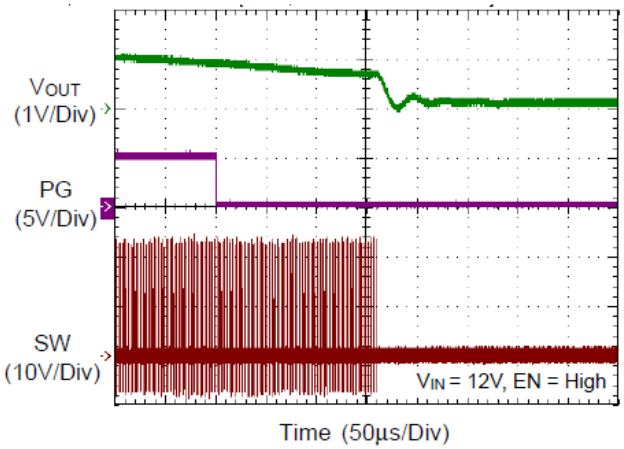
Load Transient Response



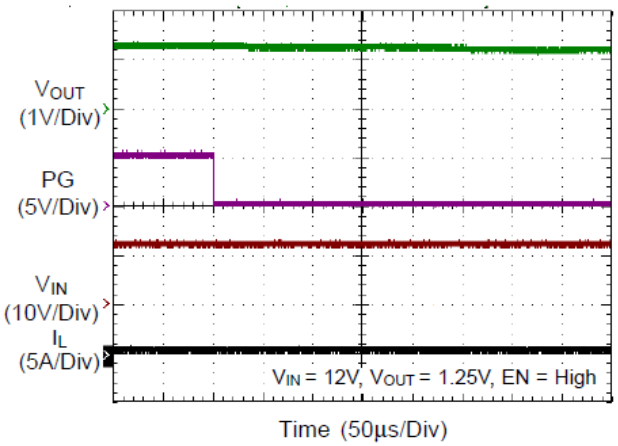
Load Transient Response



UVP



OVP



15 Operation

The RT6220 is a synchronous step-down converter with advanced constant on-time control mode. Using the ACOT® control mode can reduce the output capacitance and provide fast transient response. It can minimize the component size without additional external compensation network.

15.1 Internal VCC Regulator

The regulator provides 5V power to supply the internal control circuit. Connecting a 2.2μF ceramic capacitor for decoupling and stability is required.

15.2 Soft-Start Function

In order to prevent the converter output voltage from overshooting during the startup period, the soft-start function is necessary. The soft-start time is internal setting and the duration is around 1.5ms.

15.3 Current Limit Protection

The inductor valley current is monitored cycle-by-cycle via the internal switches, preventing an on-time until the current drops below the current limit.

15.4 Power-Good Indicator

After soft-start is finished, the power good output goes high. The PG pin is an open-drain output.

15.5 VCC Switch-Over

The internal regulator output will switch over to VBYP if VBYP level is higher than 4.6V.

15.6 Power Off

There is an internal discharging circuit to discharge the residual charge of output capacitor when converter is power off.

16 Application Information

(Note 6)

The RT6220 is high-performance 500kHz 6A buck regulators with internal power switches and synchronous rectifiers. It features an Advanced Constant On-Time (ACOT®) control architecture that provides stable operation for ceramic output capacitors without complicated external compensation, among other benefits. The input voltage range is from 4.5V to 23V, and the output voltage is adjustable from 0.6V to 5V.

The proprietary ACOT® control scheme improves conventional constant on-time architectures, achieving nearly constant switching frequency over line, load, and output voltage ranges. Since there is no internal clock, response to transients is nearly instantaneous and inductor current can ramp quickly to maintain output regulation without large bulk output capacitance.

16.1 ACOT® Control Architecture

In order to achieve good stability with low-ESR ceramic capacitors, ACOT® uses a virtual inductor current ramp generated inside the IC. This internal ramp signal replaces the ESR ramp normally provided by the output capacitor's ESR. The ramp signal and other internal compensations are optimized for low-ESR ceramic output capacitors.

Making the on-time proportional to VOUT and inversely proportional to VIN is not sufficient to achieve good constant-frequency behavior for several reasons. First, voltage drops across the MOSFET switches and inductor cause the effective input voltage to be less than the measured input voltage and the effective output voltage to be greater than the measured output voltage as sensing input and output voltage. When the load changes, the switch voltage drops change causing a switching frequency variation with load current. Also, at light loads if the inductor current goes negative, the switch dead-time between the synchronous rectifier turn-off and the high-side switch turn-on allows the switching node to rise to the input voltage. This increases the effective on-time and causes the switching frequency to drop noticeably.

One way to reduce these effects is to measure the actual switching frequency and compare it to the desired range. This has the added benefit eliminating the need to sense the actual output voltage, potentially saving one pin connection. The ACOT® uses this method, measuring the actual switching frequency and modifying the on-time with a feedback loop to keep the average switching frequency in the desired range.

16.2 ACOT® One-Shot Operation

The RT6220 control algorithm is simple to understand. The feedback voltage, with the virtual inductor current ramp added, is compared to the reference voltage. When the combined signal is less than the reference, the on-time one-shot is triggered, as long as the minimum off-time one-shot is clear and the measured inductor current (through the synchronous rectifier) is below the current limit. The on-time one-shot turns on the high-side switch and the inductor current ramps up linearly. After the on-time, the high-side switch is turned off and the synchronous rectifier is turned on and the inductor current ramps down linearly. At the same time, the minimum off-time one-shot is triggered to prevent another immediate on-time during the noisy switching time and allow the feedback voltage and current sense signals to settle. The minimum off-time is kept short (typically 200ns) so that rapidly repeated on-times can raise the inductor current quickly when needed.

16.3 Diode Emulation Mode (DEM)

In diode emulation mode, the RT6220 automatically reduces the switching frequency at light load conditions to maintain high efficiency. This reduction of frequency is achieved smoothly. As the output current decreases from heavy load conditions, the inductor current is also reduced, and eventually comes to the point that its current valley touches zero, which is the boundary between continuous conduction and discontinuous conduction modes. To emulate the behavior of diodes, the low-side MOSFET allows only a partial negative current to flow when the inductor freewheeling current becomes negative. As the load current is further decreased, it takes longer and longer time to discharge the output capacitor to the level that requires the next “ON” cycle. In reverse, when the output current increases from light load to heavy load, the switching frequency increases to the preset value as the inductor current reaches the continuous conduction. The transition load point to the light load operation is shown in Figure 4 and can be calculated as follows:

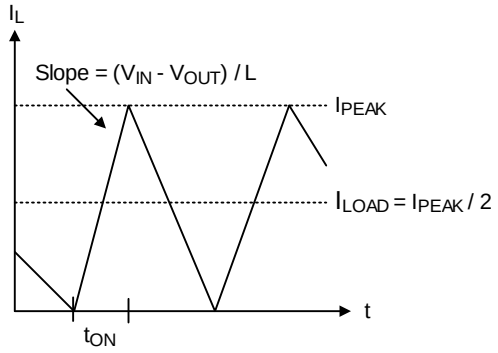


Figure 4. Boundary Condition if CCM/DEM

$$I_{LOAD} = \frac{(V_{IN} - V_{OUT})}{2L} \times t_{ON}$$

where t_{ON} is the on-time.

The switching waveforms may appear noisy and asynchronous when a light load causes diode emulation operation. This is normal and results in high efficiency. Trade-offs in DEM noise vs. light load efficiency is made by varying the inductor value. Generally, low inductor values produce a broader efficiency vs. load curve, while higher values result in higher full load efficiency (assuming that the coil resistance remains fixed) and less output voltage ripple. Penalties for using higher inductor values include larger physical size and degraded load transient response (especially at low input voltage levels).

During discontinuous switching, the on-time is immediately increased to add “hysteresis” to discourage the IC from switching back to continuous switching unless the load increases substantially. The IC returns to continuous switching as soon as an on-time is generated before the inductor current reaches zero. The on-time is reduced back to the length needed for 500kHz switching and encouraging the circuit to remain in continuous conduction, preventing repetitive mode transitions between continuous switching and discontinuous switching.

16.4 Linear Regulators (VCC)

The RT6220 includes a 5V linear regulator (VCC). The VCC regulator steps down input voltage to supply both internal circuitry and gate drivers. Do not connect the VCC pin to external loads. When PG is pulled high and BYP pin voltage is above 4.6V, an internal 3Ω P-MOSFET switch connects VCC to the BYP pin while the VCC linear regulator is simultaneously turned off.

16.5 Current Limit Protection

The RT6220 current limit is a cycle-by-cycle “valley” type, measuring the inductor current through the synchronous rectifier during the off-time while the inductor current ramps down. The current is determined by measuring the voltage between Source and Drain of the synchronous rectifier, adding temperature compensation for greater accuracy. If the current exceeds the current limit, the on-time one-shot is inhibited until the inductor current ramps down below the current limit. If the output current exceeds the available inductor current (controlled by the current limit mechanism), the output voltage will drop. If it drops below the output undervoltage protection level (see Output Overvoltage Protection and Undervoltage Protection), the IC will stop switching to avoid excessive heat.

16.6 Output Overvoltage Protection and Undervoltage Protection

The RT6220 features an output Overvoltage Protection (OVP). For the RT6220A and the RT6220BL, if the output voltage rises above the regulation level, the IC stops switching and is latched off. On the other hand, for the RT6220AH and the RT6220BH, the IC will stop switching and restart automatically after a short period, which is Hiccup mode. The RT6220 also features an output Undervoltage Protection (UVP). If the output voltage drops below the UVP trip threshold for longer than 5μs (typical), the UVP is triggered, and the IC will shut down. Likewise, for the RT6220A and the RT6220BL, the IC stops switching and is latched off. On the other hand, for the RT6220AH and the RT6220BH, the IC will stop switching and enter the Hiccup mode. To restart operation from latch off, toggle EN or power the IC off and then turn on again.

16.7 Input Undervoltage-Lockout

In addition to the enable function, the RT6220 features an Undervoltage-Lockout (UVLO) function that monitors the input voltage. To prevent operation without fully enhanced internal MOSFET switches, this function inhibits switching when the input voltage drops below the UVLO-falling threshold. The IC resumes switching when the input voltage exceeds the UVLO-rising threshold.

16.8 Over-Temperature Protection

The RT6220 features an Over-Temperature Protection (OTP) circuitry to prevent overheating due to excessive power dissipation. The OTP shuts down switching operation when the junction temperature exceeds 150°C. Once the junction temperature cools down by approximately 25°C the IC resumes normal operation with a complete soft-start. For continuous operation, provide adequate cooling so that the junction temperature does not exceed 150°C. Note that the VCC regulator remains on as the OTP is triggered.

16.9 Enable and Disable

The enable input (EN) has a logic-low level of 1.15V. When V_{EN} is below this level, the IC enters shutdown mode and supply current drops to less than 5μA (typical). Besides, the switch-over switch is turned off and VCC LDO is also powered off. When V_{EN} exceeds its logic-high level (1.35V typical), the IC is fully operational.

16.10 Soft-Start Function

The RT6220 provides an internal soft-start function to prevent large inrush current and output voltage overshoot when the converter starts up. The soft-start (SS) automatically begins once the IC is enabled. During soft- start, it clamps the ramp of the internal reference voltage, which is compared with the FB signal. And it will correct the output voltage more accurately after soft-start. The typical soft-start duration is 1.5ms.

16.11 Power Off

When V_{EN} is pulled to GND or lower than the logic-low level of 1.15V, there is an internal discharging resistor to discharge the residual charge inside the output capacitors. Besides, the value of discharging resistor is about twenty ohms.

16.12 Power-Good Output (PG)

The power-good output is an open-drain output that requires a pull-up resistor. When the output voltage is 20% (typical) below its set voltage, PG will be pulled low. It is held low until the output voltage returns to 90% of its set voltage once more. During soft-start, PG is actively held low and only allowed to be pulled high after soft-start is over and the output reaches 90% of its set voltage and the PG low to high delay (500μs typical) has passed. There is a 2μs PG high to low delay built into the PG circuitry to prevent false triggering.

In addition, the PG open drain driver is supplied by the VCC power source or the VBYP pin voltage source in switch-over mode. When converter is powered off by EN low signal, the pull-low strength of PG open-drain driver decreases after the VCC voltage is lower than the VCC_POR threshold (3.8V typical). As a result, the PG pin is floated and pulled up by external voltage source. In consideration of PG status after EN power off, it is recommended that connecting the PG pin with a 100kΩ resistor to VCC (5V). DO NOT pull high to an external voltage, which is higher than VCC (5V).

16.13 External Bootstrap Capacitor (C_{BOOT})

Connect a 0.1μF low ESR ceramic capacitor between the BOOT and SW pins. This bootstrap capacitor provides the gate driver supply voltage for the high-side N-MOSFET switch.

The internal power MOSFET switch gate driver is optimized to turn the switch on quickly enough for low power loss and good efficiency, and slow enough to reduce EMI. Most EMI occurs during switch turn-on, as V_{SW} rises rapidly. During switch turn-off, SW is discharged relatively slowly by the inductor current during the dead-time between high-side and low-side switch on-times. In some cases, it is desirable to reduce EMI further, at the expense of some additional power dissipation. The switch turn-on can be slowed by placing a small (<10Ω) resistance between BOOT and the external bootstrap capacitor. This will slow the high-side switch turn-on and V_{SW}'s rise.

16.14 Setting Output Voltage

The output voltage of the RT6220 is adjustable and with valley control. There is an easy way to determine the output voltage only by two resistors, R1 and R2. As the feedback circuit shown in Figure 5, the relationship between V_{OUT} and V_{REF} can be derived as V_{OUT} = (1+R1/R2) x V_{REF} readily. Generally, the stability is a serious issue for converters. In order to achieve better performance on stability and transient, a feed-forward capacitor, C_{FF}, is added to increase the noise margin and transient response of loop control. However, there is a tradeoff of adding a feed-forward capacitor. An additional dc offset will be generated on output voltage due to the amplified feedback ripple by feed-forward compensator. This is not always the case that every C_{FF} makes the same value of dc offset, it is based on different pole and zero placement generated by R1, R2, and C_{FF}. For simplicity, a symbol named $V_{dc,offset}$ is supposed to be the value of dc offset. This value may slightly influence the performance (e.g., regulation or peak value of V_{OUT}) of the converter. The suggested C_{FF} is to select a pair of pole and zero to provide the maximum phase lead at the switching frequency.

$$V_{OUT,valley} = \left(1 + \frac{R1}{R2}\right) \times V_{REF} + V_{dc,\ offset}$$

$V_{OUT,valley}$ is the valley of the output voltage, and $V_{dc,offset}$ is used for describing the additional dc offset on V_{OUT}.

and the value is related to the output voltage ripple and C_{FF}.

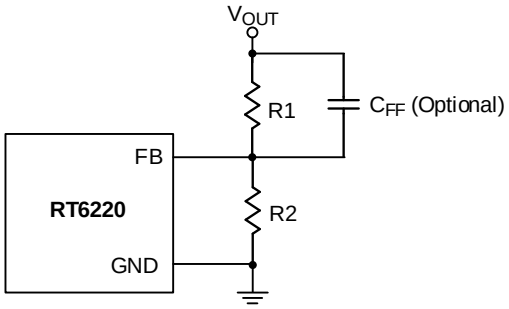


Figure 5. The Equivalent Circuit of Feedback Loop

16.15 Inductor Selection

Selecting an inductor involves specifying its inductance and also its required peak current. The exact inductor value is generally flexible and is ultimately chosen to obtain the best mix of cost, physical size, and circuit efficiency. Lower inductor values benefit from reduced size and cost and can improve the circuit's transient response. However, they increase the inductor ripple current and output voltage ripple and reduce the efficiency due to the resulting higher peak currents. Conversely, higher inductor values increase efficiency, but the inductor will either be physically larger or have higher resistance since more turns of wire are required and the transient response will be slower since more time is required to change the current (up or down) in the inductor. A good compromise between size, efficiency, and transient response is to use a ripple current (ΔI_L) about 20-50% of the desired full output load current. Calculate the approximate inductor value by selecting the input and output voltages, the switching frequency (f_{SW}), the maximum output current ($I_{OUT(MAX)}$), and estimating ΔI_L as a percentage of that current.

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

Once an inductor value is chosen, the ripple current (ΔI_L) is calculated to determine the required peak inductor current.

$$\Delta I_L = \frac{V_{OUT} \times (V_{IN} - V_{OUT})}{V_{IN} \times f_{SW} \times L} \quad \text{and} \quad I_{L(PEAK)} = I_{OUT(MAX)} + \frac{\Delta I_L}{2}$$

To guarantee the required output current, the inductor needs a saturation current rating and a thermal rating that exceeds $I_{L(PEAK)}$. These are minimum requirements. To maintain control of inductor current in overload and short-circuit conditions, some applications may desire current ratings up to the current limit value. However, the IC's output undervoltage shutdown feature makes this unnecessary for most applications.

For best efficiency, choose an inductor with a low DC resistance that meets the cost and size requirements. For low inductor core losses, some type of ferrite core is usually best. And a shielded core type, although possibly larger or more expensive, will probably give fewer EMI and other noise problems.

16.16 Input Capacitor Selection

High-quality ceramic input decoupling capacitors, such as X5R or X7R, with values greater than 20μF are recommended for the input capacitor. The X5R and X7R ceramic capacitors are usually selected for power regulator capacitors because the dielectric material has less capacitance variation and more temperature stability. Voltage rating and current rating are the key parameters when selecting an input capacitor. Generally, selecting an input capacitor with a voltage rating 1.5 times greater than the maximum input voltage is a conservatively safe

design. The input capacitor is used to supply the input RMS current, which can be calculated using the following equation:

$$I_{RMS} = \sqrt{\frac{V_{OUT}}{V_{IN}} \times \left[\left(1 - \frac{V_{OUT}}{V_{IN}} \right) \times I_{OUT}^2 + \frac{\Delta I_L^2}{12} \right]}$$

The next step is to select a proper capacitor for RMS current rating. One good design uses more than one capacitor with low Equivalent Series Resistance (ESR) in parallel to form a capacitor bank. The input capacitance value determines the input ripple voltage of the regulator. The input voltage ripple can be approximately calculated using the following equation:

$$\Delta V_{IN} = \frac{I_{OUT} \times V_{IN}}{C_{IN} \times f_{SW} \times V_{OUT}} \times \left(1 - \frac{V_{OUT}}{V_{IN}} \right)$$

The typical operating circuit is recommended to use two 10μF low ESR ceramic capacitors on the input.

16.17 Output Capacitor Selection

The RT6220 is optimized for ceramic output capacitors and the best performance will be obtained by using them. The total output capacitance value is usually determined by the desired output voltage ripple level and transient response requirements for sag (undershoot on positive load steps) and soar (overshoot on negative load steps).

Output ripple at the switching frequency is caused by the inductor current ripple and its effect on the output capacitor's ESR and stored charge. These two ripple components are called ESR ripple and capacitive ripple. Since ceramic capacitors have extremely low ESR and relatively little capacitance, both components are similar in amplitude and both should be considered if ripple is critical.

$$V_{RIPPLE(ESR)} = V_{RIPPLE(ESR)} + V_{RIPPLE(C)}$$

$$V_{RIPPLE(ESR)} = \Delta I_L \times R_{ESR}$$

$$V_{RIPPLE(C)} = \frac{\Delta I_L}{8 \times C_{OUT} \times f_{SW}}$$

In addition to voltage ripple at the switching frequency, the output capacitor and its ESR also affect the voltage sag (undershoot) and soar (overshoot) when the load steps up and down abruptly. The ACOT® transient response is very quick and output transients are usually small. However, the combination of small ceramic output capacitors (with little capacitance), low output voltages (with little stored charge in the output capacitors), and low duty cycle applications (which require high inductance to get reasonable ripple currents with high input voltages) increases the size of voltage variations in response to very quick load changes. Typically, load changes occur slowly with respect to the IC's 500kHz switching frequency. However, some modern digital loads can exhibit nearly instantaneous load changes and the following section shows how to calculate the worst-case voltage swings in response to very fast load steps.

The amplitude of the ESR step up or down is a function of the load step and the ESR of the output capacitor:

$$V_{ESR\_STEP} = \Delta I_{OUT} \times R_{ESR}$$

The amplitude of the capacitive sag is a function of the load step, the output capacitor value, the inductor value, the input-to-output voltage differential, and the maximum duty cycle. The maximum duty cycle during a fast transient is a function of the on-time and the minimum off-time since the ACOT® control scheme will ramp the current using on-times spaced apart with minimum off-times, which is as fast as allowed. Calculate the approximate on-time (neglecting parasitics) and maximum duty cycle for a given input and output voltage as:

$$t_{ON} = \frac{V_{OUT}}{V_{IN} \times f_{SW}} \quad \text{and} \quad D_{MAX} = \frac{t_{ON}}{t_{ON} + t_{OFF\_MIN}}$$

The actual on-time will be slightly longer as the IC compensates for voltage drops in the circuit, but we can neglect both of these since the on-time increases compensations for the voltage losses. Calculate the output voltage sag as:

$$V_{SAG} = \frac{L \times (\Delta I_{OUT})^2}{2 \times C_{OUT} \times (V_{IN(MIN)} \times D_{MAX} - V_{OUT})}$$

The amplitude of the capacitive soar is a function of the load step, the output capacitor value, the inductor value and the output voltage:

$$V_{SOAR} = \frac{L \times (\Delta I_{OUT})^2}{2 \times C_{OUT} \times V_{OUT}}$$

Most applications never experience instantaneous full load steps and the RT6220's high switching frequency and fast transient response can easily control voltage regulation at all times. Therefore, sag and soar are seldom an issue except in very low-voltage CPU core or DDR memory supply applications, particularly for devices with high clock frequencies and quick changes into and out of sleep modes. In such applications, simply increasing the amount of ceramic output capacitor (sag and soar are directly proportional to capacitance) or adding extra bulk capacitance can easily eliminate any excessive voltage transients.

In any application with large quick transients, it should calculate soar and sag to make sure that overvoltage protection and undervoltage protection will not be triggered.

16.18 Thermal Considerations

For continuous operation, do not exceed absolute maximum junction temperature. The maximum power dissipation depends on the thermal resistance of the IC package, PCB layout, rate of surrounding airflow, and difference between junction and ambient temperature. The maximum power dissipation can be calculated by the following formula:

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

where $T_{J(MAX)}$ is the maximum junction temperature, T_A is the ambient temperature, and $\theta_{JA(EVB)}$ is the junction to ambient thermal resistance.

For continuous operation, the maximum operating junction temperature indicated under Recommended Operating Conditions is 125°C. The junction-to-ambient thermal resistance, $\theta_{JA(EVB)}$, is highly package dependent. For a UQFN-16L 3x3 (FC) package, the thermal resistance, $\theta_{JA(EVB)}$, is 43°C/W on a standard JEDEC 51-7 high effective-thermal-conductivity four-layer test board. The maximum power dissipation at T_A = 25°C can be calculated as below:

$P_{D(MAX)}$ = (125°C – 25°C) / (43°C/W) = 2.33W for a UQFN-16L 3x3 (FC) package.

The maximum power dissipation depends on the operating ambient temperature for fixed $T_{J(MAX)}$ and thermal resistance, $\theta_{JA(EVB)}$. The derating curve in Figure 6 allows the designer to see the effect of rising ambient temperature on the maximum power dissipation.

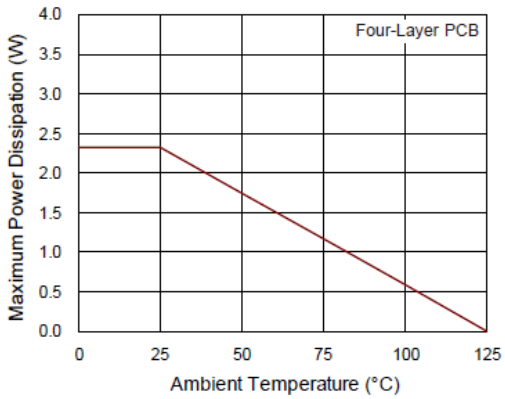


Figure 6. Derating Curve of Maximum Power Dissipation

16.19 Layout Considerations

Layout is very important in high-frequency switching converter design. Improper PCB layout can radiate excessive noise and contribute to converter instability. The following points must be considered before starting a layout using the RT6220.

- Make traces of the high current paths as short and wide as possible.
- Put the input capacitor as close as possible to the device pins (VIN and PGND).
- The SW node encounters high frequency voltage swings so it should be kept in a small area. Keep sensitive components away from the SW node to prevent noise coupling.
- The PGND pin should be connected to a strong ground plane for heat sinking and noise protection.
- Avoid using vias in the power path connections that have switched currents (from CIN to PGND and CIN to VIN) and the switching node (SW).
- The ground of VCC is recommended to connect to GND layer through a via, and the decoupling capacitor (CVCC) should be placed near the VCC pin. No via connection is recommended.

An example of PCB layout guide is shown in Figure 7 for reference.

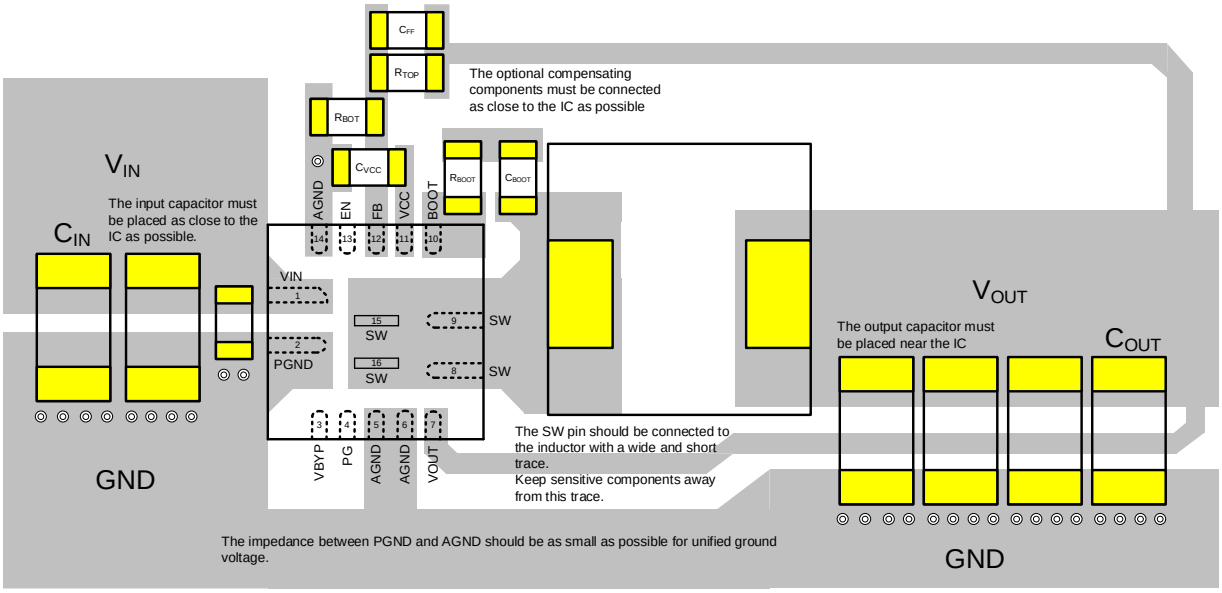


Figure 7. PCB Layout Guide

16.20 Trace Width Design

For thermal management, efficiency, and the PCB handling current capability, the trace width design is very important. According to IPC-2221 formally IDC-D-275 PWB, the following formulas can be used to calculate the trace width for printed circuit boards.

Inner trace:

$I(Amp) = 0.015 \times \Delta T(°C)^{0.5453} \times Area(mils^2)^{0.7349}$

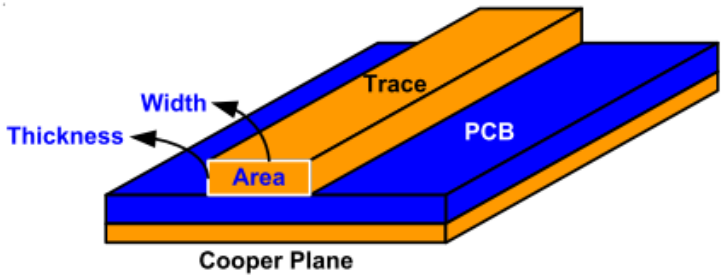
Outer trace:

$I(Amp) = 0.0647 \times \Delta T(°C)^{0.4281} \times Area(mils^2)^{0.6732}$

$$Width(mil) = \frac{Area(mils^2)}{Thickness(oz) \times 1.37\left(\frac{mil}{oz}\right)}$$

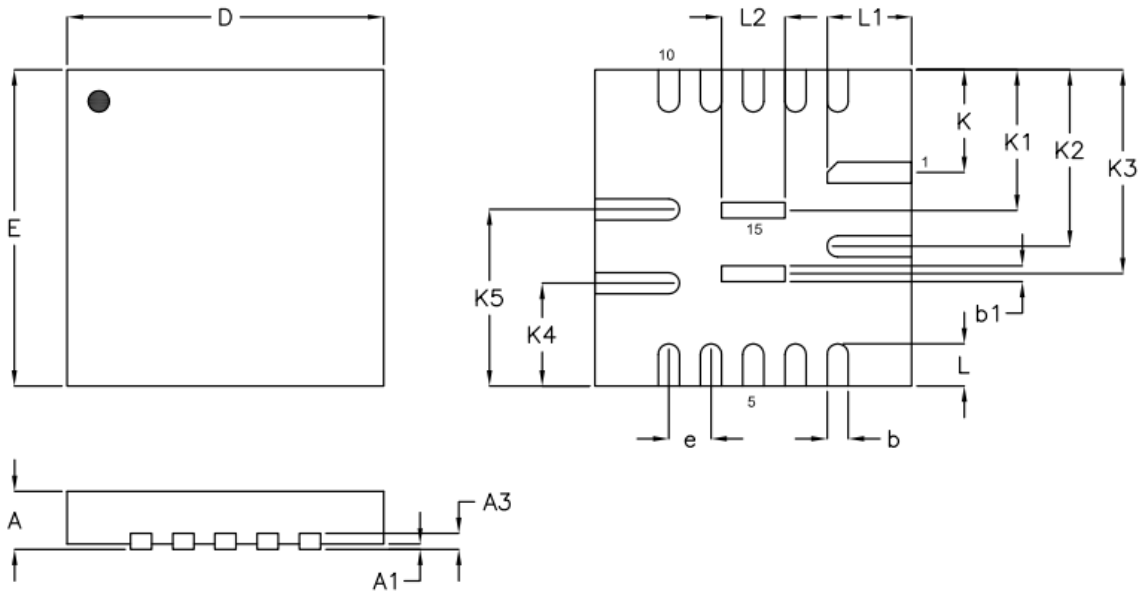
where

I(Amp) = Current, $\Delta T(°C)$ = Temperature rise, $Area(mils^2)$ = Cross-sectional area = Width x Thickness, Width(mil) = Trace width, and Thickness(oz) = Layer Cu thickness.



Note 6. 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.

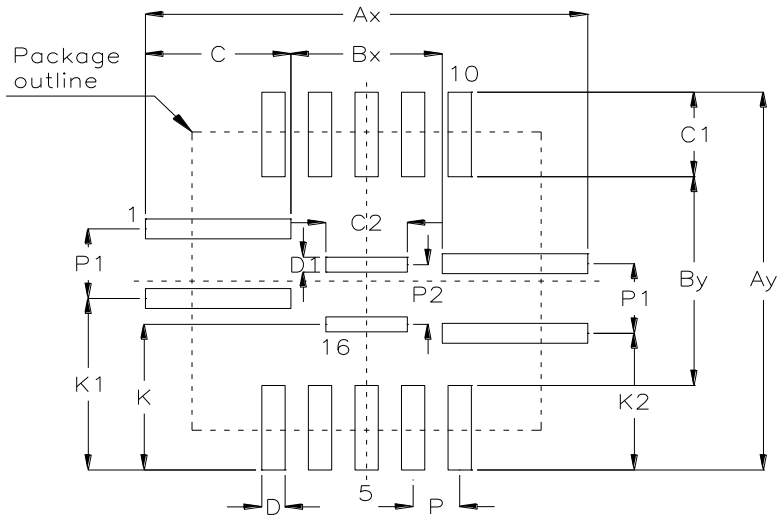
17 Outline Dimension



Symbol	Dimensions In Millimeters		Dimensions In Inches	
	Min	Max	Min	Max
A	0.500	0.600	0.020	0.024
A1	0.000	0.050	0.000	0.002
A3	0.100	0.200	0.004	0.008
D	2.900	3.100	0.114	0.122
E	2.900	3.100	0.114	0.122
b	0.150	0.250	0.006	0.010
b1	0.100	0.200	0.004	0.008
L	0.350	0.450	0.014	0.018
L1	0.750	0.850	0.030	0.033
L2	0.550	0.650	0.022	0.026
e	0.400		0.016	
K	0.975		0.038	
K1	1.335		0.053	
K2	1.675		0.066	
K3	1.935		0.076	
K4	0.975		0.038	
K5	1.675		0.066	

U-Type 16L QFN 3x3 (FC) Package

18 Footprint Information

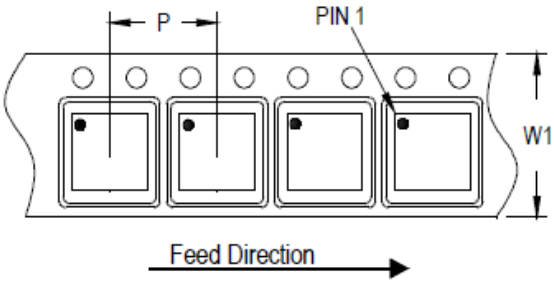


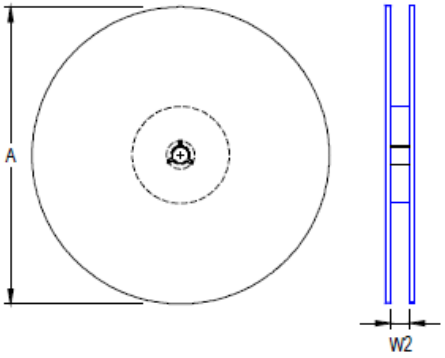
Package	Number of Pin	Footprint Dimension (mm)								Tolerance
		P	P1	P2	Ax	Bx	Ay	By	C*4	
UQFN3x3-16(FC)	16	0.400	0.700	0.600	3.800	1.300	3.800	2.100	1.250	±0.050

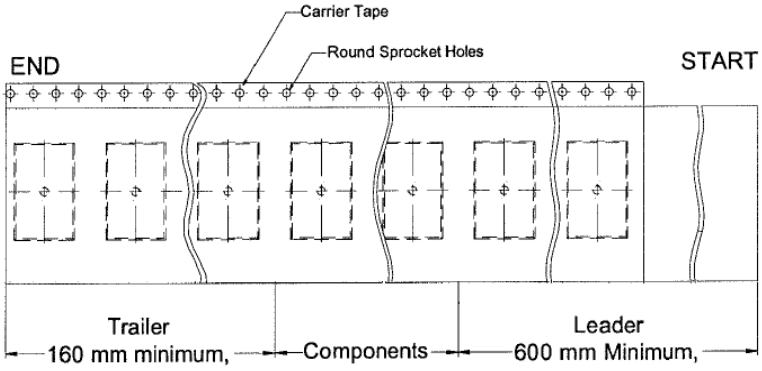
Package	Number of Pin	Footprint Dimension (mm)							Tolerance
		C1*10	C2*2	D*14	D1*2	K	K1	K2	
UQFN3x3-16(FC)	16	0.850	0.700	0.200	0.150	1.465	1.725	1.375	±0.050

19 Packing Information

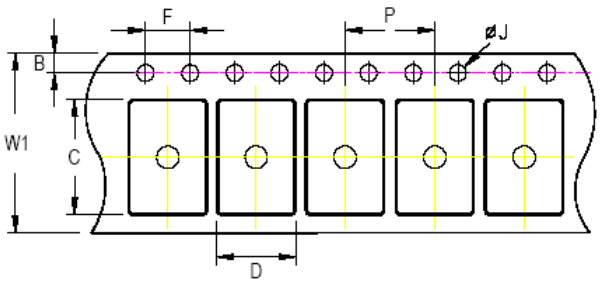
19.1 Tape and Reel Data







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 3x3	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 max.

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

19.2 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 3x3	7”	1,500	Box A	3	4,500	Carton A	12	54,000
			Box E	1	1,500	For Combined or Partial Reel.		

19.3 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}**

Richtek Technology Corporation
14F, No. 8, Tai Yuen 1st Street, Chupei City
Hsinchu, Taiwan, R.O.C.
Tel: (8863)5526789

20 Datasheet Revision History

Version	Date	Description	Item
12	2025/2/7	Modify	*General Description on page 1* *Features on page 1* *Ordering Information on page 1* *Typical Operating Characteristics on page 10* *Application Information on page 6* *Footprint Information on page 24* - Added Footprint Information *Packing Information on page 25 to 27* - Added packing information