1.5A, 18V, 500kHz, ACOT™ Step-Down Converter

General Description

The RT6211A/B is a high-efficiency, monolithic synchronous step-down DC/DC converter that can deliver up to 1.5A output current from a 4.5V to 18V input supply. The RT6211A/B adopts ACOT architecture to allow the transient response to be improved and keep in constant frequency. Cycle-by-cycle current limit provides protection against shorted outputs and soft-start eliminates input current surge during start-up. Fault conditions also include output under voltage protection, output over current protection, and thermal shutdown.

Ordering Information

RT6211A/B □□□

- Package Type
  E : SOT-23-6
- Lead Plating System
  G : Green (Halogen Free and Pb Free)
- UVP Option
  H : Hiccup
- A : PSM Mode
  B : PWM Mode

Note :

Richtek products are :

- RoHS compliant and compatible with the current requirements of IPC/JEDEC J-STD-020.
- Suitable for use in SnPb or Pb-free soldering processes.

Features

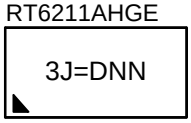
- **Integrated 230mΩ/130mΩ MOSFETs**
- **4.5V to 18V Supply Voltage Range**
- **500kHz Switching Frequency**
- **ACOT Control**
- **0.8V ± 1.5% Voltage Reference**
- **Internal Start-Up into Pre-biased Outputs**
- **Compact Package : SOT-23-6 pin**
- **High/Low Side Over-Current Protection and Hiccup**
- **V_{OUT} Range 0.8V to 6.5V**

Applications

- Set-Top Boxes
- Portable TVs
- Access Point Routers
- DSL Modems
- LCD TVs

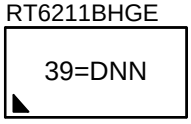
Marking Information

RT6211AHGE



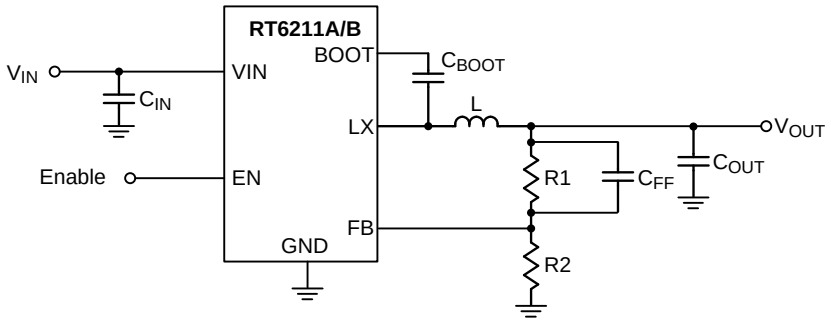
3J= : Product Code
DNN : Date Code

RT6211BHGE



39= : Product Code
DNN : Date Code

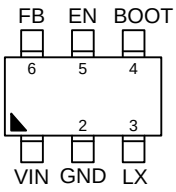
Simplified Application Circuit



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Pin Configurations

(TOP VIEW)

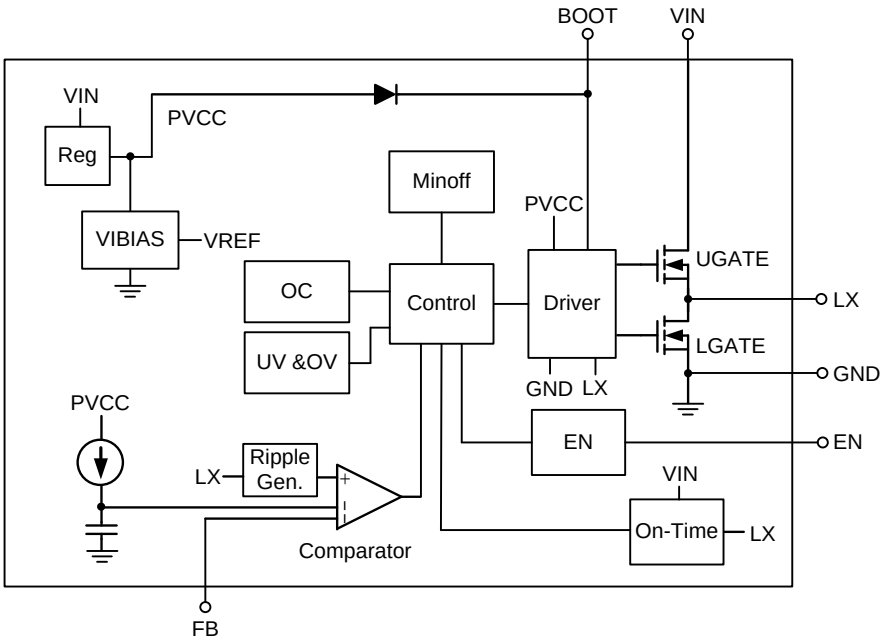


SOT-23-6

Functional Pin Description

Pin No.	Pin Name	Pin Function
1	VIN	Power Input. Supplies the power switches of the device.
2	GND	System Ground. Provides the ground return path for the control circuitry and low-side power MOSFET.
3	LX	Switch Node. LX is the switching node that supplies power to the output and connect the output LC filter from LX to the output load.
4	BOOT	Bootstrap Supply for High-Side Gate Driver. Connect a 100nF or greater capacitor from LX to BOOT to power the high-side switch.
5	EN	Enable Control Input. Floating this pin or connecting this pin to logic high can enable the device and connecting this pin to GND can disable the device.
6	FB	Feedback Voltage Input. This pin is used to set the desired output voltage via an external resistive divider. The feedback voltage is 0.8V typically.

Function Block Diagram



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Detailed Description

The RT6211A/B are high-performance 500kHz 1.5A step-down regulators with internal power switches and synchronous rectifiers. They feature an Advanced Constant On-Time (ACOT™) control architecture that provides stable operation with ceramic output capacitors without complicated external compensation, among other benefits. The input voltage range is from 4.5V to 18V and the output is adjustable from 0.8V to 6.3V.

The proprietary ACOT™ control scheme improves upon other constant on-time architectures, achieving nearly constant switching frequency over line, load, and output voltage ranges. The RT6211A/B are optimized for ceramic output capacitors. 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.

Constant On-Time (COT) Control

The heart of any COT architecture is the on-time one-shot. Each on-time is a pre-determined "fixed" period that is triggered by a feedback comparator. This robust arrangement has high noise immunity and is ideal for low duty cycle applications. After the on-time one-shot period, there is a minimum off-time period before any further regulation decisions can be considered. This arrangement avoids the need to make any decisions during the noisy time periods just after switching events, when the switching node (LX) rises or falls. Because there is no fixed clock, the high-side switch can turn on almost immediately after load transients and further switching pulses can ramp the inductor current higher to meet load requirements with minimal delays.

Traditional current mode or voltage mode control schemes typically must monitor the feedback voltage, current signals (also for current limit), and internal ramps and compensation signals, to determine when to turn off the high-side switch and turn on the synchronous rectifier. Weighing these small signals in a switching environment is difficult to do just after switching large currents, making those architectures problematic at low duty cycles and in less than ideal board layouts.

Because no switching decisions are made during noisy time periods, COT architectures are preferable in low duty cycle and noisy applications. However, traditional COT control schemes suffer from some disadvantages that preclude their use in many cases. Many applications require a known switching frequency range to avoid interference with other sensitive circuitry. True constant on-time control, where the on-time is actually fixed, exhibits variable switching frequency. In a step-down converter, the duty factor is proportional to the output voltage and inversely proportional to the input voltage. Therefore, if the on-time is fixed, the off-time (and therefore the frequency) must change in response to changes in input or output voltage.

Modern pseudo-fixed frequency COT architectures greatly improve COT by making the one-shot on-time proportional to V_{OUT} and inversely proportional to V_{IN}. In this way, an on-time is chosen as approximately what it would be for an ideal fixed-frequency PWM in similar input/output voltage conditions. The result is a big improvement but the switching frequency still varies considerably over line and load due to losses in the switches and inductor and other parasitic effects.

Another problem with many COT architectures is their dependence on adequate ESR in the output capacitor, making it difficult to use highly-desirable, small, low-cost, but low-ESR ceramic capacitors. Most COT architectures use AC current information from the output capacitor, generated by the inductor current passing through the ESR, to function in a way like a current mode control system. With ceramic capacitors, the inductor current information is too small to keep the control loop stable, like a current mode system with no current information.

ACOT™ Control Architecture

Making the on-time proportional to V_{OUT} and inversely proportional to V_{IN} 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 the load changes, the switch voltage drops

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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. ACOT™ uses this method, measuring the actual switching frequency (at SW) and modifying the on-time with a feedback loop to keep the average switching frequency in the desired range.

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.

ACOT™ One-shot Operation

The RT6211A/B 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 (240ns typical) so that rapidly-repeated on-times can raise the inductor current quickly when needed.

Discontinuous Operating Mode (RT6211A Only)

After soft-start, the RT6211B operates in fixed frequency mode to minimize interference and noise problems. The RT6211A uses variable-frequency discontinuous switching at light loads to improve efficiency. 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.

Current Limit

The RT6211A/B 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 it drops below the current limit level. 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 under-voltage protection level (see next section) the IC will stop switching to avoid excessive heat.

The RT6211B also includes a negative current limit to protect the IC against sinking excessive current and possibly damaging the IC. If the voltage across the synchronous rectifier indicates the negative current is too high, the synchronous rectifier turns off until after the next high-side on-time. The RT6211A does not sink current and therefore does not need a negative current limit.

Hiccup Mode

The RT6211AHGE / RT6211BHGE, use hiccup mode UVP. When the protection function is triggered, the IC will shut down for a period of time and then attempt to recover automatically. Hiccup mode allows the circuit to

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operate safely with low input current and power dissipation, and then resume normal operation as soon as the overload or short circuit is removed.

Input Under-Voltage Lockout

To protect the chip from operating at insufficient supply voltage, the UVLO is needed. When the input voltage of VIN is lower than the UVLO falling threshold voltage, the device will be lockout.

Shut-down, Start-up and Enable (EN)

The enable input (EN) has a logic-low level. When V_{EN} is below this level the IC enters shutdown mode and supply current drops to less than 6μA. When V_{EN} exceeds its logic-high level the IC is fully operational.

External Bootstrap Capacitor

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

Over-Temperature Protection

The RT6211A/B includes an over-temperature protection (OTP) circuitry to prevent overheating due to excessive power dissipation. The OTP will shut down switching operation when the junction temperature exceeds 150°C. Once the junction temperature cools down by approximately 20°C the IC will resume normal operation. For continuous operation, provide adequate cooling so that the junction temperature does not exceed 150°C.

UVP Protection

The RT6211A/B detects under-voltage conditions by monitoring the feedback voltage on FB pin. The function is enabled after approximately 1.7 times the soft-start time. When the feedback voltage is lower than 50% of the target voltage, the UVP comparator will go high to turn off both internal high-side and low-side MOSFETs.

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Absolute Maximum Ratings (Note 1)

- Supply Input Voltage ---------- –0.3V to 20V
- Switch Voltage, SW ---------- –0.8V to (V_{IN} + 0.3V)
  <10ns ---------- –5V to 25V
- Switch Node Voltage, LX ---------- –0.3V to (V_{IN} + 0.3V)
- BOOT Pin Voltage ---------- (V_{LX} – 0.3V) to (V_{IN} + 6.3V)
- Other Pins ---------- –0.3V to 6V
- Power Dissipation, P_D @ T_A = 25°C
  SOT-23-6 ---------- 0.48W
- Package Thermal Resistance (Note 2)
  SOT-23-6, θ_{JA} ---------- 208.2°C/W
- Lead Temperature (Soldering, 10 sec.) ---------- 260°C
- Junction Temperature ---------- 150°C
- Storage Temperature Range ---------- –65°C to 150°C
- ESD Susceptibility (Note 3)
  HBM (Human Body Model) ---------- 2kV

Recommended Operating Conditions (Note 4)

- Supply Input Voltage ---------- 4.5V to 18V
- Ambient Temperature Range ---------- –40°C to 85°C
- Junction Temperature Range ---------- –40°C to 125°C

Electrical Characteristics

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

Parameter	Symbol	Test Conditions	Min	Typ	Max	Unit
Supply Voltage						
VIN Supply Input Operating Voltage	V_{IN}		4.5	--	18	V
Under-Voltage Lockout Threshold	V_{UVLO}		3.6	3.9	4.2	
Under-Voltage Lockout Threshold Hysteresis	ΔV_{UVLO}		--	340	--	mV
Supply Current						
Supply Current (Shutdown)	I_{SHDN}	V_{EN} = 0V	--	--	6	µA
Supply Current (Quiescent)	I_Q	V_{EN} = 2V, V_{FB} = 0.85V	--	0.8	--	mA
Soft-Start						
Soft-Start Time			--	1000	--	µs
Enable Voltage						
EN Rising Threshold	V_{EN_Rising}		1.38	1.5	1.62	V
EN Falling Threshold	$V_{EN_Falling}$		1.16	1.28	1.4	

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Parameter	Symbol	Test Conditions	Min	Typ	Max	Unit
Feedback Threshold Voltage						
Feedback Threshold Voltage	V_{FB_TH}	4.5V ≤ V_{IN} ≤ 18V	0.788	0.8	0.812	V
Internal MOSFET						
High-Side On-Resistance	$R_{DS(ON)_H}$	V_{BOOT} − V_{LX} = 4.8V	--	230	--	mΩ
Low-Side On-Resistance	$R_{DS(ON)_L}$		--	130	--	
Current Limit						
Current Limit	I_{LIM}	Valley Current	2	2.5	--	A
Switching Frequency						
Switching Frequency	f_{OSC}		420	520	620	kHz
On-Time Timer Control						
Maximum Duty Cycle	D_{MAX}		85	88	--	%
Minimum On Time	$t_{ON(MIN)}$		--	60	--	nS
Minimum Off Time	$t_{OFF(MIN)}$		--	240	--	
Output Under Voltage Protections						
UVP Trip Threshold		UVP Detect	45	50	55	%
		Hysteresis	--	10	--	
Thermal Shutdown						
Thermal Shutdown Threshold	T_{SD}		--	150	--	°C
Thermal Shutdown Hysteresis	ΔT_{SD}		--	20	--	

Note 1. Stresses beyond those listed “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 2. θ_{JA} is measured at T_A = 25°C on a high effective thermal conductivity four-layer test board per JEDEC 51-7.

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

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

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Typical Application Circuit

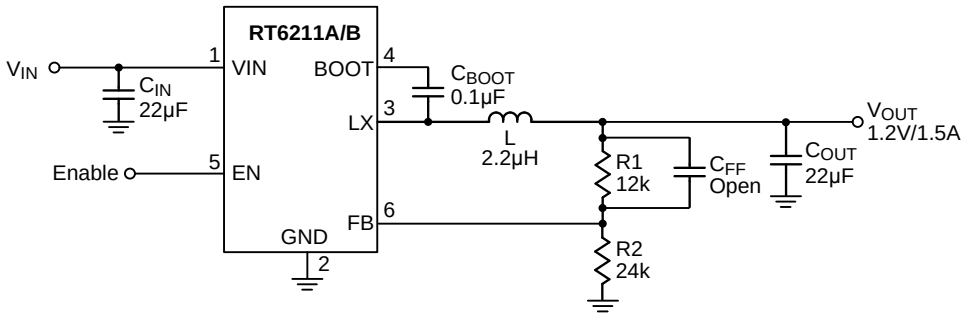


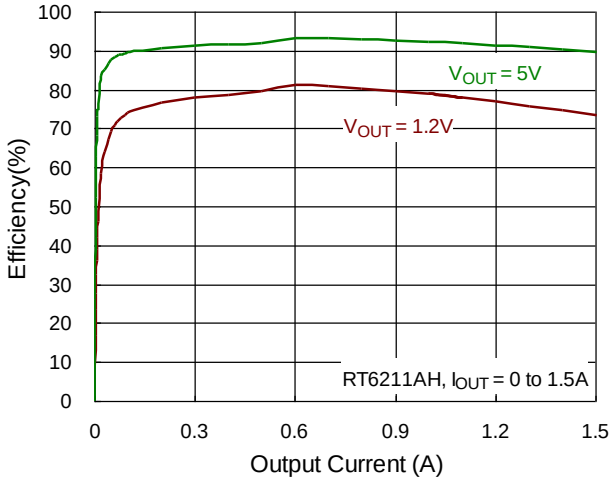
Table 1. Suggested Component Values (V_{IN} = 12V)

V_{OUT} (V)	R1 (kΩ)	R2 (kΩ)	L (µH)	C_{OUT} (µF)	C_{FF} (pF)
1.05	10	32.4	2	22	--
1.2	20.5	41.2	2	22	--
1.8	40.2	32.4	3.3	22	--
2.5	40.2	19.1	4.7	22	22 to 68
3.3	40.2	13	6.5	22	22 to 68
5	40.2	7.68	6.5	22	22 to 68

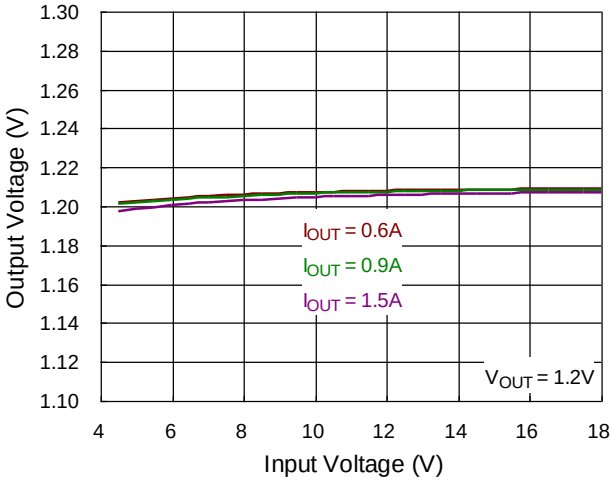
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Typical Operating Characteristics

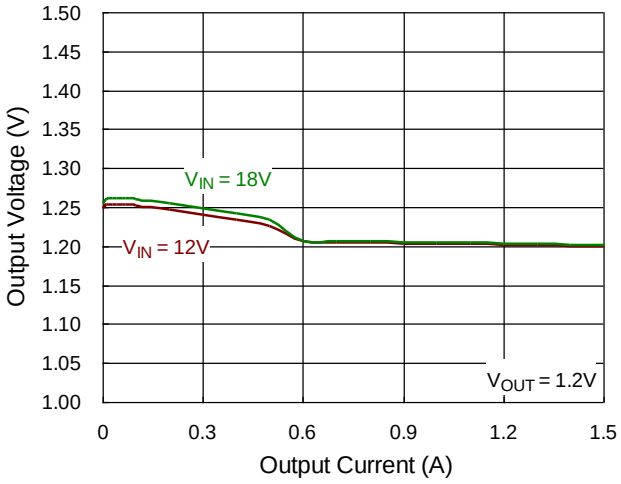
Efficiency vs. Output Current



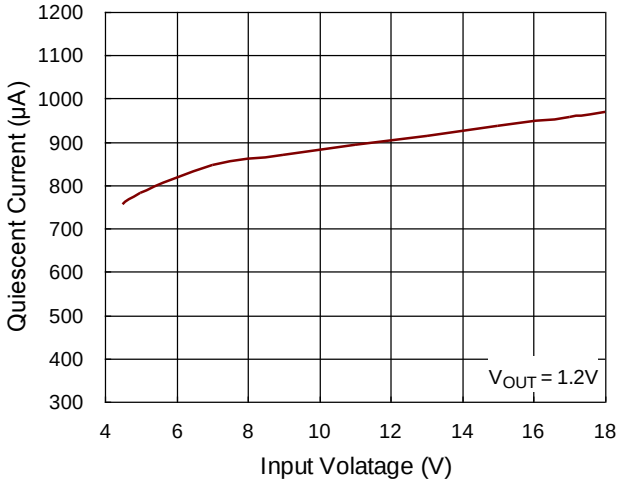
Output Voltage vs. Input Voltage



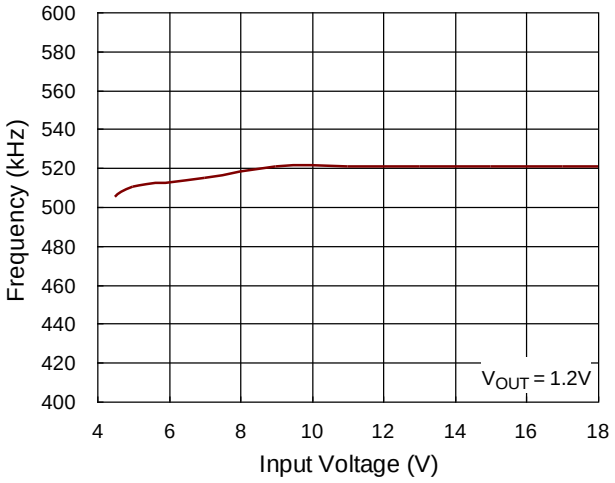
Output Voltage vs. Output Current



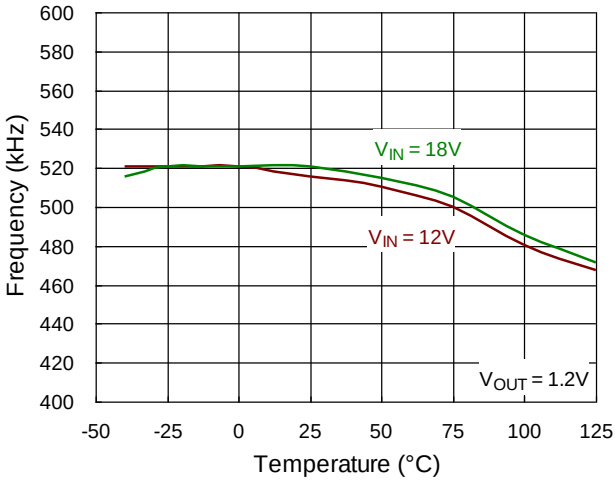
Quiescent Current vs. Input Voltage



Frequency vs. Input Voltage

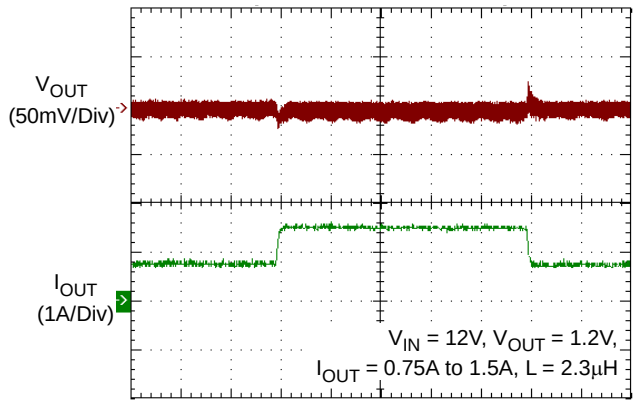


Frequency vs. Temperature

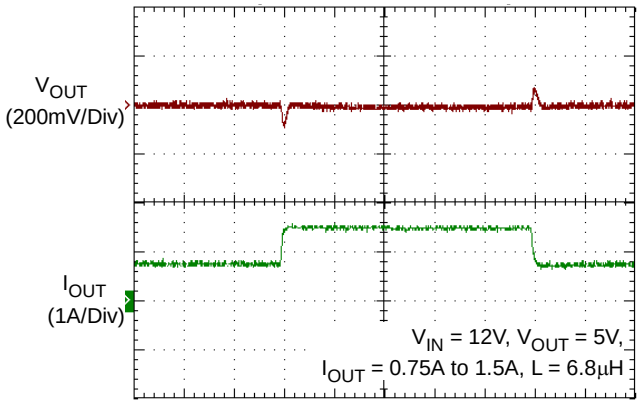


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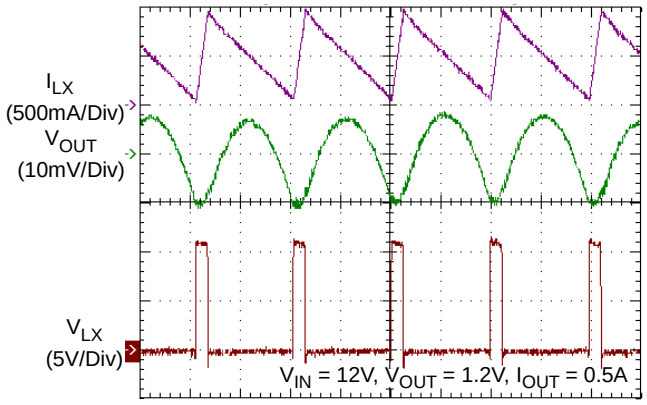
Load Transient Response



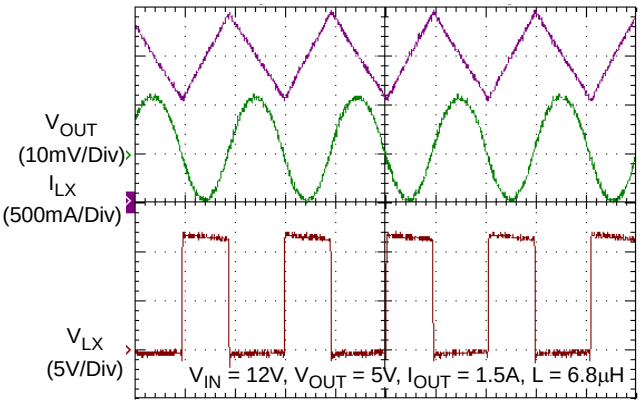
Load Transient Response



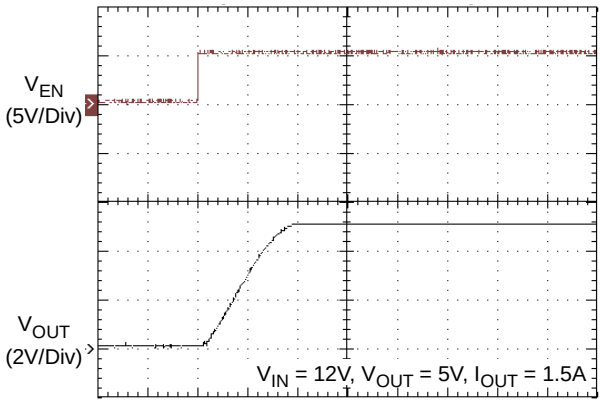
Output Ripple Voltage



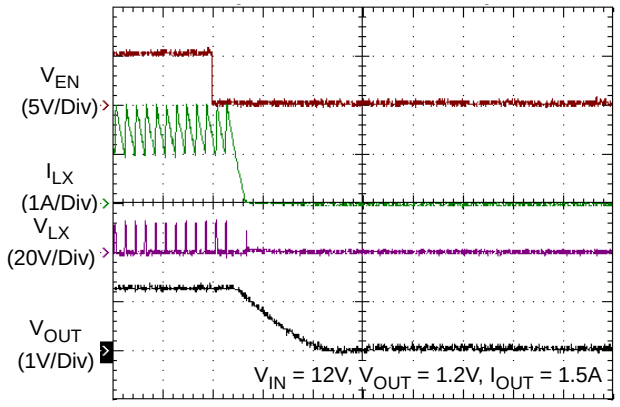
Output Ripple Voltage



Power On from EN



Power Off from EN



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Application Information

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 they can improve the circuit's transient response, but 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 transient response will be slower since more time is required to change 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% to 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 a ΔI_L as some 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.

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

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 under-voltage shutdown feature make this unnecessary for most applications.

$I_{L(PEAK)}$ should not exceed the minimum value of IC's upper current limit level or the IC may not be able to meet the desired output current. If needed, reduce the inductor ripple current (ΔI_L) to increase the average inductor current (and the output current) while ensuring that $I_{L(PEAK)}$ does not exceed the upper current limit level.

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.

Considering the Typical Operating Circuit for 1.2V output at 1.5A and an input voltage of 12V, using an inductor ripple of 0.45A (30%), the calculated inductance value is :

$$L = \frac{1.2 \times (12 - 1.2)}{12 \times 500kHz \times 0.45A} = 4.8\mu H$$

The ripple current was selected at 0.45A and, as long as we use the calculated 4.8μH inductance, that should be the actual ripple current amount. The ripple current and required peak current as below :

$$\Delta I_L = \frac{1.2 \times (12 - 1.2)}{12 \times 500kHz \times 4.8\mu H} = 0.45A$$

$$\text{and } I_{L(PEAK)} = 1.5A + \frac{0.45A}{2} = 1.725A$$

For the 4.8μH value, the inductor's saturation and thermal rating should exceed 1.725A. Since the actual value used was 4.8μH and the ripple current exactly 0.45A, the required peak current is 1.725A.

Input Capacitor Selection

The input filter capacitors are needed to smooth out the switched current drawn from the input power source and to reduce voltage ripple on the input. The actual capacitance value is less important than the RMS current rating (and voltage rating, of course). The RMS input ripple current (I_{RMS}) is a function of the input voltage, output voltage, and load current :

$$I_{RMS} = I_{OUT(MAX)} \times \frac{V_{OUT}}{V_{IN}} \sqrt{\frac{V_{IN}}{V_{OUT}} - 1}$$

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Ceramic capacitors are most often used because of their low cost, small size, high RMS current ratings, and robust surge current capabilities. However, take care when these capacitors are used at the input of circuits supplied by a wall adapter or other supply connected through long, thin wires. Current surges through the inductive wires can induce ringing at the RT6211A/B input which could potentially cause large, damaging voltage spikes at VIN. If this phenomenon is observed, some bulk input capacitance may be required. Ceramic capacitors (to meet the RMS current requirement) can be placed in parallel with other types such as tantalum, electrolytic, or polymer (to reduce ringing and overshoot).

Choose capacitors rated at higher temperatures than required. Several ceramic capacitors may be paralleled to meet the RMS current, size, and height requirements of the application. The typical operating circuit uses two 10μF and one 0.1μF low ESR ceramic capacitors on the input.

Output Capacitor Selection

The RT6211A/B are optimized for ceramic output capacitors and best performance will be obtained 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

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.

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

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

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

For the Typical Operating Circuit for 1.2V output and an inductor ripple of 0.45A, with 2 x 22μF output capacitance each with about 5mΩ ESR including PCB trace resistance, the output voltage ripple components are :

$$V_{RIPPLE(ESR)} = 0.45A \times 5m\Omega = 2.25mV$$

$$V_{RIPPLE(C)} = \frac{0.45A}{8 \times 44\mu F \times 500kHz} = 2.557mV$$

$$V_{RIPPLE} = 2.25mV + 2.557mV = 4.807mV$$

Output Capacitors Stability Criteria

The e RT6211A/B's ACOT™ control architecture uses an internal virtual inductor current ramp and other compensation that ensures stability with any reasonable output capacitor. The internal ramp allows the IC to operate with very low ESR capacitors and the IC is stable with very small capacitances. Therefore, output capacitor selection is nearly always a matter of meeting output voltage ripple and transient response requirements, as discussed in the previous sections. For the sake of the unusual application where ripple voltage is unimportant and there are few transients (perhaps battery charging or LED lighting) the stability criteria are discussed below.

The equations giving the minimum required capacitance for stable operation include a term that depends on the output capacitor's ESR. The higher the ESR, the lower the capacitance can be and still ensure stability. The equations can be greatly simplified if the ESR term is removed by setting ESR to zero. The resulting equation gives the worst-case minimum required capacitance and it is usually sufficiently small that there is usually no need for the more exact equation :

$$C_{OUT} > \frac{3 \times 5.23 \times 10^{-11}}{V_{IN} \times L}$$

The worst-case high capacitance requirement is for low VIN and small inductance, so a 12V to 1.2V converter is used for an example. Using the inductance equation in a previous section to determine the required inductance :

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$$L = \frac{1.2\times(12-1.2)}{12\times500kHz\times0.4A} = 5.4\mu H$$

Therefore, the required minimum capacitance for the 12V to 1.2V converter is :

$$C_{OUT} > \frac{3\times5.23\times10^{-11}}{12V\times5.4\mu H}$$

$$C_{OUT} > 2.42\mu F$$

Feed-forward Capacitor (C_{ff})

The RT6211A/B are optimized for ceramic output capacitors and for low duty cycle applications. However for high-output voltages, with high feedback attenuation, the circuit's response becomes over-damped and transient response can be slowed. In high-output voltage circuits (V_{OUT} > 3.3V) transient response is improved by adding a small "feed-forward" capacitor (C_{ff}) across the upper FB divider resistor (Figure 1), to increase the circuit's Q and reduce damping to speed up the transient response without affecting the steady-state stability of the circuit. Choose a suitable capacitor value that following below step.

- Get the BW the quickest method to do transient response form no load to full load. Confirm the damping frequency. The damping frequency is BW.

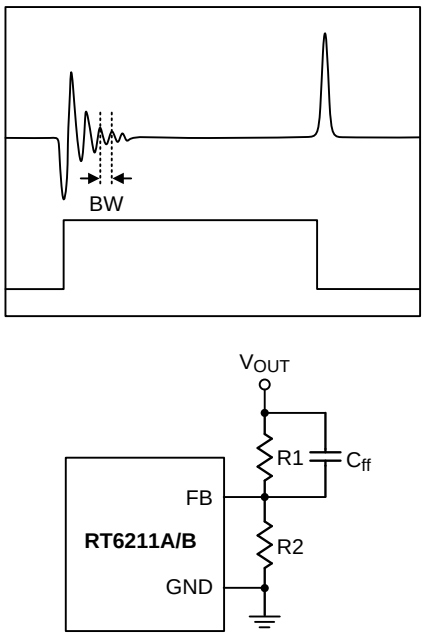


Figure 1. C_{ff} Capacitor Setting

- C_{ff} can be calculated base on below equation :

$$C_{ff} = \frac{1}{2\times3.1412\times R1\times BW\times0.8}$$

Enable Operation (EN)

For automatic start-up the high-voltage EN pin can be connected to VIN, through a 100kΩ resistor. Its large hysteresis band makes EN useful for simple delay and timing circuits. EN can be externally pulled to VIN by adding a resistor-capacitor delay (R_{EN} and C_{EN} in Figure 2). Calculate the delay time using EN's internal threshold where switching operation begins.

An external MOSFET can be added to implement digital control of EN when no system voltage above 2V is available (Figure 3). In this case, a 100kΩ pull-up resistor, R_{EN}, is connected between VIN and the EN pin. MOSFET Q1 will be under logic control to pull down the EN pin. To prevent enabling circuit when VIN is smaller than the VOUT target value or some other desired voltage level, a resistive voltage divider can be placed between the input voltage and ground and connected to EN to create an additional input under voltage lockout threshold (Figure 4).

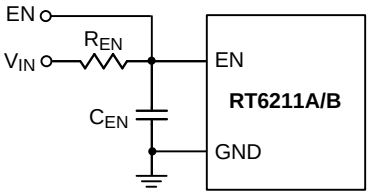


Figure 2. External Timing Control

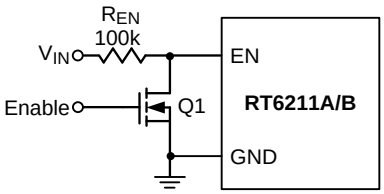


Figure 3. Digital Enable Control Circuit

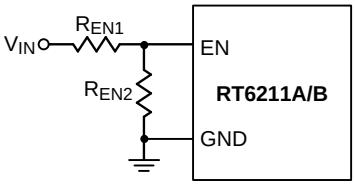


Figure 4. Resistor Divider for Lockout Threshold Setting

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Output Voltage Setting

Set the desired output voltage using a resistive divider from the output to ground with the midpoint connected to FB. The output voltage is set according to the following equation :

V_{OUT} = 0.8V x (1 + R1 / R2)

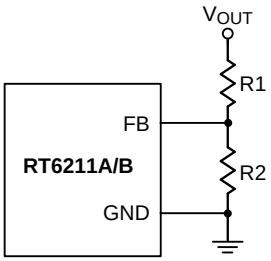


Figure 5. Output Voltage Setting

Place the FB resistors within 5mm of the FB pin. Choose R2 between 10kΩ and 100kΩ to minimize power consumption without excessive noise pick-up and calculate R1 as follows :

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

For output voltage accuracy, use divider resistors with 1% or better tolerance.

External BOOT Bootstrap Diode

When the input voltage is lower than 5.5V it is recommended to add an external bootstrap diode between VIN (or VINR) and the BOOT pin to improve enhancement of the internal MOSFET switch and improve efficiency. The bootstrap diode can be a low cost one such as 1N4148 or BAT54.

External BOOT Capacitor Series Resistance

The internal power MOSFET switch gate driver is optimized to turn the switch on fast enough for low power loss and good efficiency, but also slow enough to reduce EMI. Switch turn-on is when most EMI occurs since V_{LX} rises rapidly. During switch turn-off, LX 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 (<47Ω) resistance between BOOT and the external bootstrap capacitor. This will slow the high-side switch turn-on and V_{LX}'s rise. To remove the resistor from the capacitor charging path (avoiding poor enhancement due to undercharging the BOOT capacitor), use the external diode shown in Figure 6 to charge the BOOT capacitor and place the resistance between BOOT and the capacitor/diode connection.

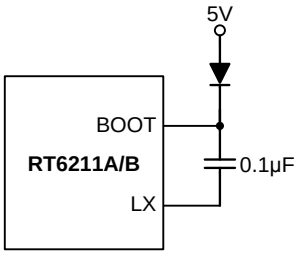


Figure 6. External Bootstrap Diode

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 :

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

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

For recommended operating condition specifications, the maximum junction temperature is 125°C. The junction to ambient thermal resistance, θ_{JA}, is layout dependent. For SOT-23-6 package, the thermal resistance, θ_{JA}, is 208.2°C/W on a standard four-layer thermal test board. The maximum power dissipation at T_A = 25°C can be calculated by the following formula :

$P_{D(MAX)}$ = (125°C – 25°C) / (208.2°C/W) = 0.48W for SOT-23-6 package

The maximum power dissipation depends on the operating ambient temperature for fixed $T_{J(MAX)}$ and thermal resistance, θ_{JA}. The derating curve in Figure 7 allows the designer to see the effect of rising ambient temperature on the maximum power dissipation.

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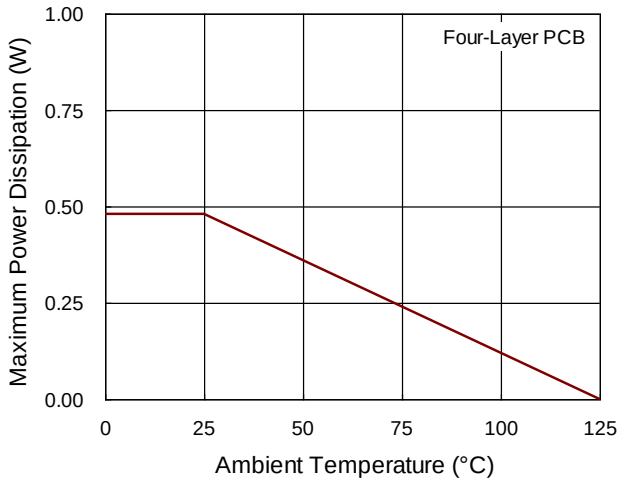


Figure 7. Derating Curve of Maximum Power Dissipation

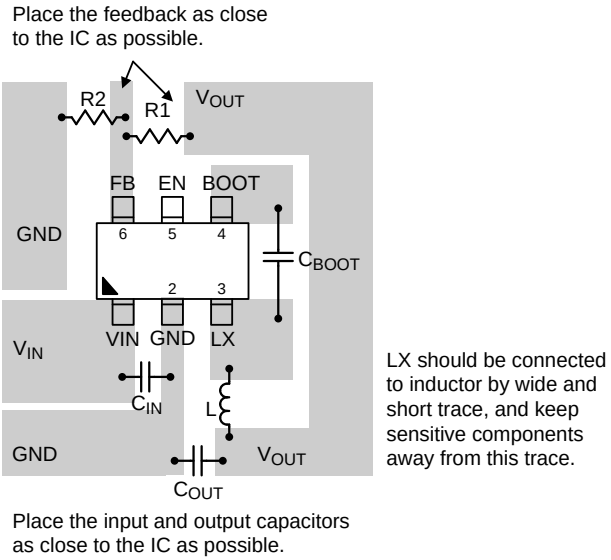
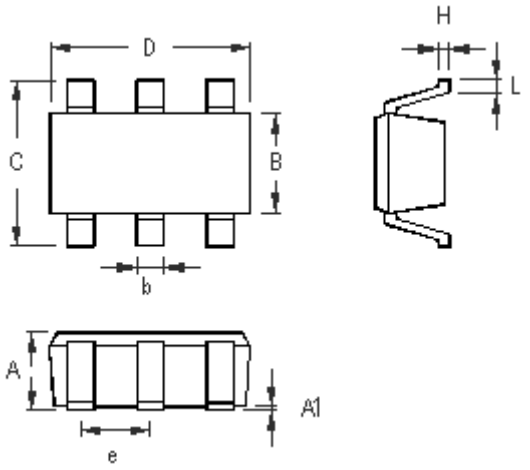


Figure 8. PCB Layout Guide

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Outline Dimension



Symbol	Dimensions In Millimeters		Dimensions In Inches	
	Min	Max	Min	Max
A	0.889	1.295	0.031	0.051
A1	0.000	0.152	0.000	0.006
B	1.397	1.803	0.055	0.071
b	0.250	0.560	0.010	0.022
C	2.591	2.997	0.102	0.118
D	2.692	3.099	0.106	0.122
e	0.838	1.041	0.033	0.041
H	0.080	0.254	0.003	0.010
L	0.300	0.610	0.012	0.024

SOT-23-6 Surface Mount Package

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