SEMTECH

30V, 300mA Output Micropower Step-Down Switching Regulator

POWER MANAGEMENT

Features

- Input Voltage Range: 3V to 30V
- Low Quiescent Current: Drawing 19μA from V_{IN} when Stepping Down from 12V to 3.3V at No Load
- High Efficiency from 12V Input to 5V Output
  > 80% at 650μA
  > 85% at 10mA - 300mA
- Up to 300mA Continuous DC Output Current
- Integrated Power Switch and Schottky Diodes
- Low Output Ripple
- <1μA Shutdown Current
- Hysteretic Current-Mode Control
- Cycle-by-Cycle Current Limiting
- Alternating Between Micropower Idling and Switching States at Light Loads to Conserve Power
- Output Short-Circuit Protection
- Solution Footprint as Small as 50mm^2
- Low-Profile 3mm x 2mm MLPD 8-Lead Package

Applications

- Portable Equipment
- Notebook Computers
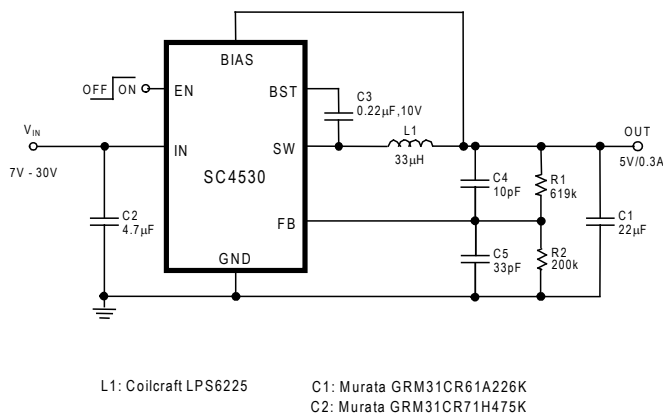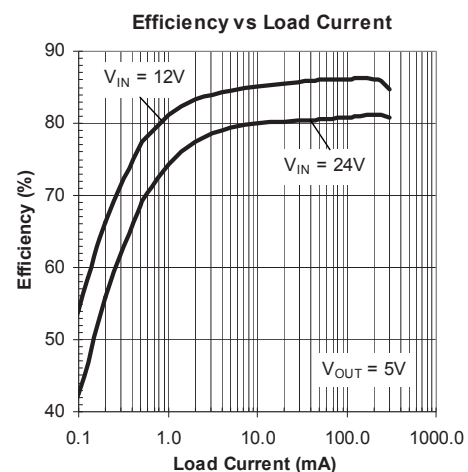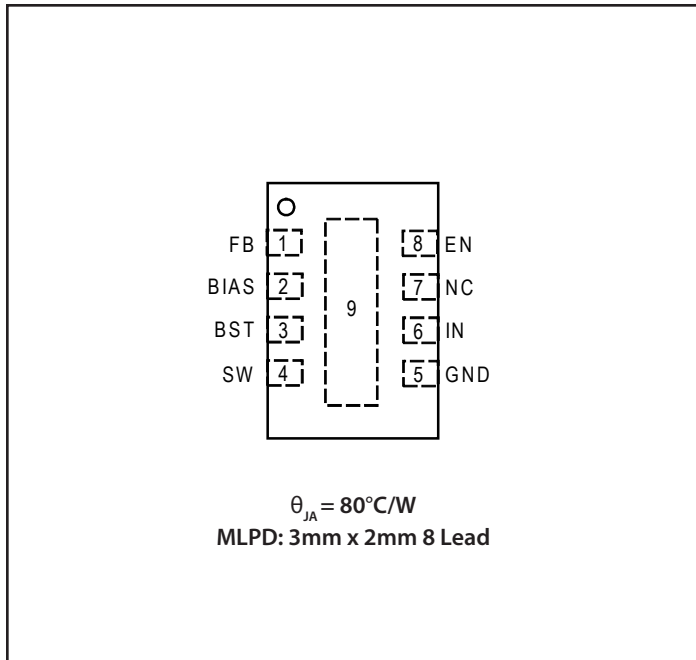- Distributed Supplies
- Backup Power Supplies

Description

The SC4530 is a micropower hysteretic current-mode step-down switching regulator capable of providing up to 300mA of output current from 3V to 30V input voltage range. It is designed to provide very high standby efficiency while simplifying design.

At light loads, the SC4530 switches only as needed to maintain regulation, while idling most of the time, to improve efficiency. Typical quiescent currents from V_{IN} and BIAS are 7μA and 26μA respectively. The control scheme produces less than 10mV of FB voltage ripple at light loads. The SC4530 automatically switches to continuous-conduction mode at heavy loads.

The SC4530 has integrated power devices and on-chip control circuitry, simplifying design and enabling a solution footprint as small as 50mm^2. Only an inductor and a few passive components are needed to complete a DC-DC regulator. The inductor current hysteretic control of SC4530 makes it inherently short-circuit robust. The wide input voltage range enables the device to operate from a variety of input sources, including single- or multi-cell batteries, system rails and wall transformers.

Typical Application Circuit


L1: Coilcraft LPS6225
C1: Murata GRM31CR61A226K
C2: Murata GRM31CR71H475K

Figure 1. 5V Output Step-Down Converter

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Pin Configuration

```
        ○
  FB  [1]  ┌ ┐  [8]  EN
 BIAS [2]  │ │  [7]  NC
  BST [3]  │9│  [6]  IN
   SW [4]  └ ┘  [5]  GND
```

θ_{JA} = **80°C/W**
MLPD: 3mm x 2mm 8 Lead

Ordering Information

Device	Package
SC4530WLTRT[1] [2]	MLPD-W-8 3x2
SC4530EVB	Evaluation Board

Notes:
(1)   Available in tape and reel only. A reel contains 3,000 devices.
(2)   Available in lead-free package only. Device is WEEE and RoHS compliant.

Marking Information

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  ○

  4530
  XXXX
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XXXX - Lot Number
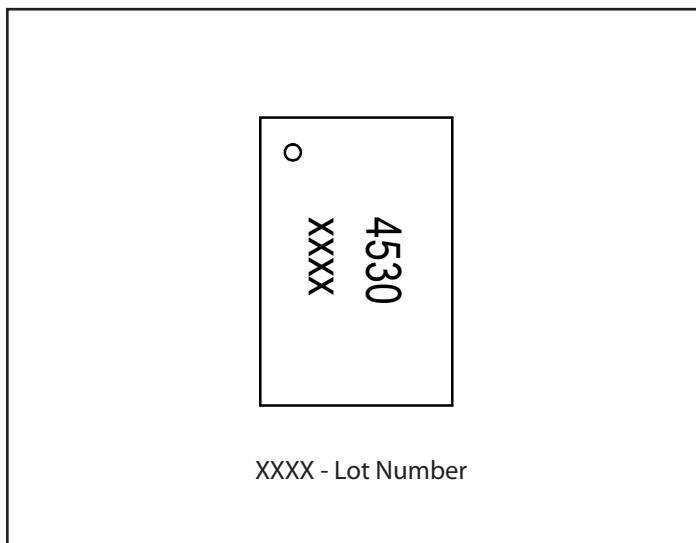
Absolute Maximum Ratings [1]

IN…………………………………………………… -0.3V to 32V

SW[2]……………………………………………… -0.6V to V$_{IN}$

BST …………………………………………………   42V

BST Above SW………………………………………   30V

FB ……………………………………………… -0.3V to 1.9V

BIAS …………………………………………… -0.3V to V$_{IN}$

EN …………………………………………… -0.3V to V$_{IN}$

ESD Protection Level [3] …………………………………… 2kV

Recommended Operating Conditions

Junction Temperature Range………………… -40°C to +125°C

V$_{IN}$ ………………………………………………… 3V to 30V

Output DC Current……………………………   up to 300mA

Thermal Information

Thermal Resistance, Junction to Ambient [4] ………… 80°C/W

Maximum Junction Temperature………………………+150 °C

Storage Temperature Range …………………-65°C to +150°C

Peak IR Reflow Temperature (10s to 30s)……………… +260°C

Exceeding the above specifications may result in permanent damage to the device or the device may malfunction. Operation outside of the parameters specified in the Electrical Characteristics section is not recommended.

NOTES:
(1) Unless noted otherwise, all voltage values in this section are with respect to ground.
(2) The transient negative voltage specification for the SW pin is -1V for 100ns.
(3) Tested according to JEDEC standard JS - 001- 2012.
(4) Calculated from package in still air, mounted to 3" x 4.5", 4-layer FR4 PCB with thermal vias under the exposed pad, per JESD51 standards.

Electrical Characteristics

Unless otherwise noted, T$_A$ = 25°C for typical values, -40°C < T$_A$ = T$_J$ < 125°C. V$_{IN}$ = V$_{EN}$ = 10V, V$_{BST}$ = 15V, V$_{BIAS}$ = 3V.

Parameter	Name	Conditions	Min	Typ	Max	Units
V$_{IN}$ Operating Range			3		30	V
V$_{IN}$ Quiescent Supply Current		V$_{EN}$ = 0.2V		0.1	0.5	µA
		Not Switching		7	11	µA
		Not Switching , V$_{BIAS}$ = 0		34	50	µA
BIAS Quiescent Supply Current		V$_{EN}$ = 0.2V		0.1	0.6	µA
		Not Switching		26	40	µA
		Not Switching , V$_{BIAS}$ = 0		0.1	1	µA
EN Pin Input High Voltage			2			V
EN Pin Input Low Voltage					0.2	V
EN Pin Current		V$_{EN}$ = 2.5V		1	2.5	µA
Feedback Voltage		V$_{FB}$ Falling	1.212	1.232	1.245	V
Feedback Voltage Line Regulation		V$_{IN}$ = 3V to 30V		0.01	0.02	%/V
FB Pin Bias Current		V$_{FB}$ = 1.25V		20	60	nA
Minimum Switch Off-time	T$_{OFF(MIN)}$			530		ns
Maximum Switch Duty Cycle	D$_{MAX}$		90	96		%

Electrical Characteristics (continued)

Unless otherwise noted, $T_A = 25°C$ for typical values, $-40°C < T_A = T_J < 125°C$. $V_{IN} = V_{EN} = 10V$, $V_{BST} = 15V$, $V_{BIAS} = 3V$.

Parameter	Name	Conditions	Min	Typ	Max	Units
Switch Current Limit	I_{LIM}	$V_{FB} = 0$	0.39	0.50	0.66	A
Inductor Current Hysteresis [1]		$V_{FB} = 0$		65		mA
Switch Saturation Voltage		$I_{SW} = -0.3A$		200	300	mV
Switch Leakage Current		$V_{SW} = 0$			2	µA
Switch Minimum Bootstrap Voltage		$I_{SW} = -0.3A$		1.7	2.2	V
BST Pin Current		$I_{SW} = -0.3A$		7.1	12	mA
Freewheeling Diode Forward Voltage	V_D	$I_{SW} = -0.3A$		700		mV
Freewheeling Diode Reverse Leakage		$V_{SW} = 10V$			15	µA
Bootstrap Diode Forward Voltage		$I_{BST} = 40mA$		700		mV
Bootstrap Diode Reverse Leakage		$V_{SW} = 10V$, $V_{BIAS} = 0$			1	µA

Notes:
(1) The inductor current hysteresis is the difference between the switch current limit and the freewheeling diode valley current.

Pin Descriptions

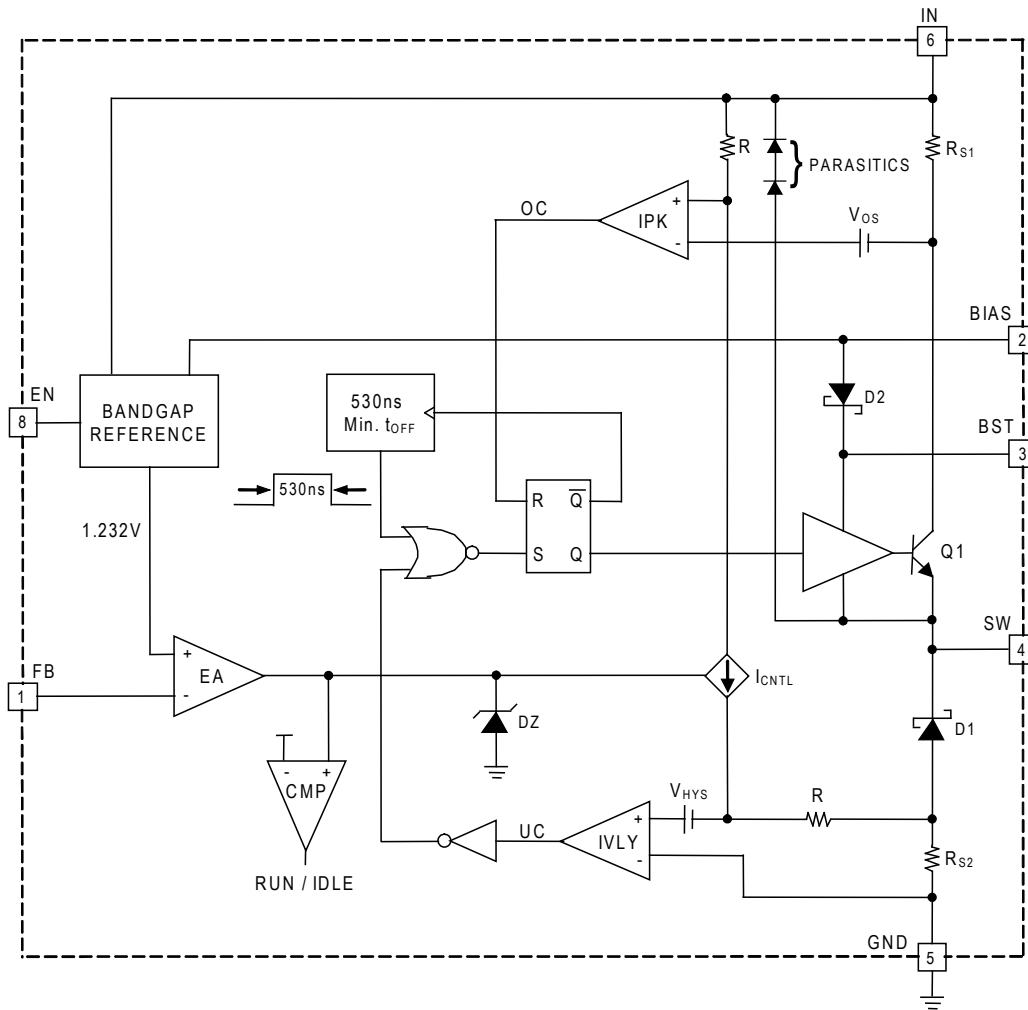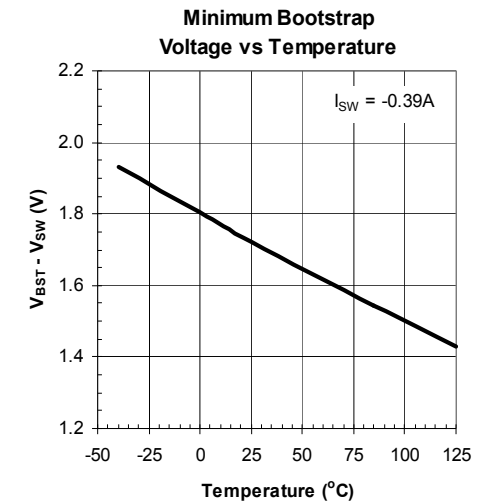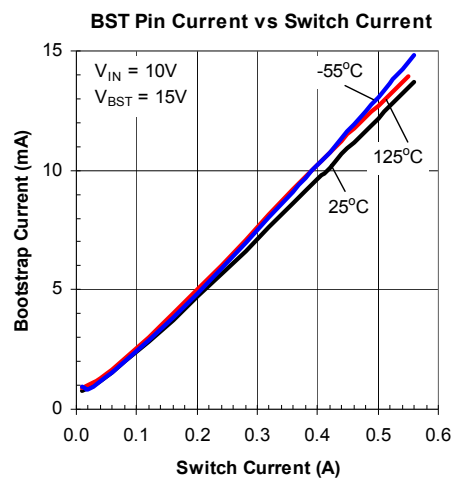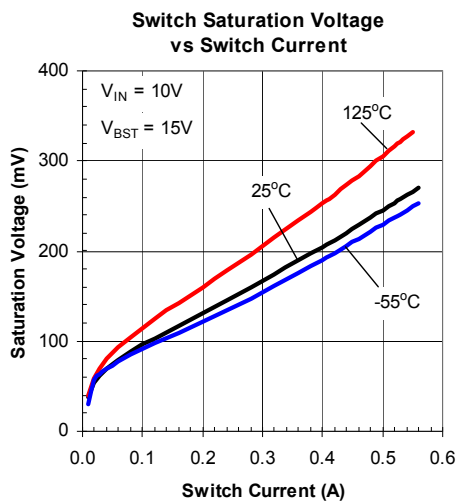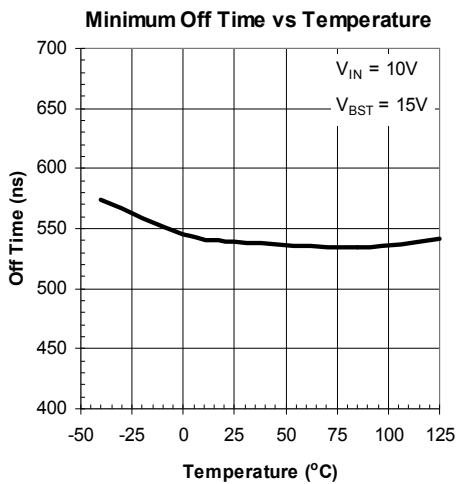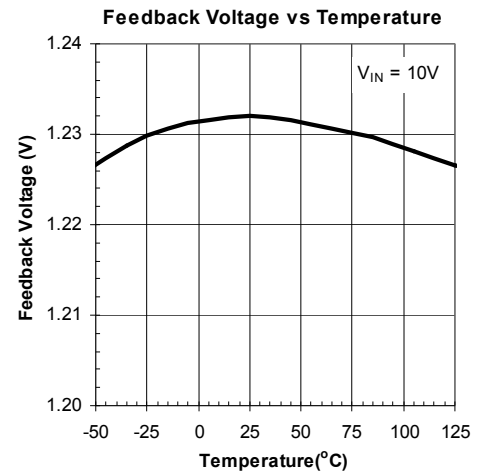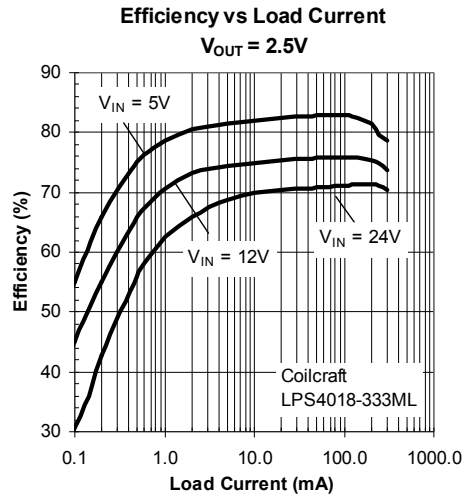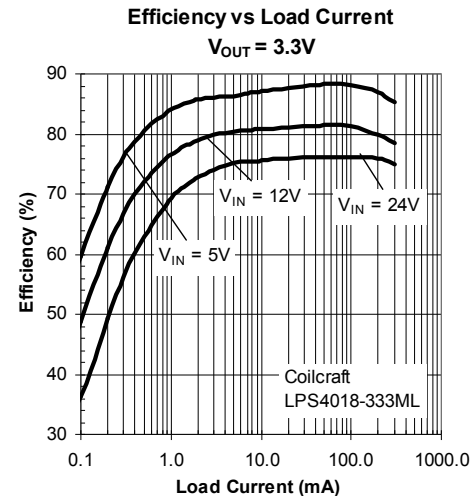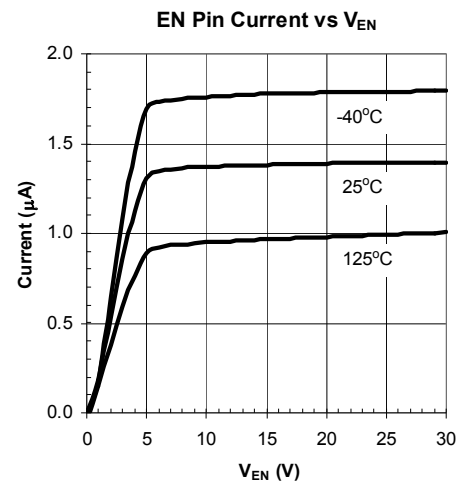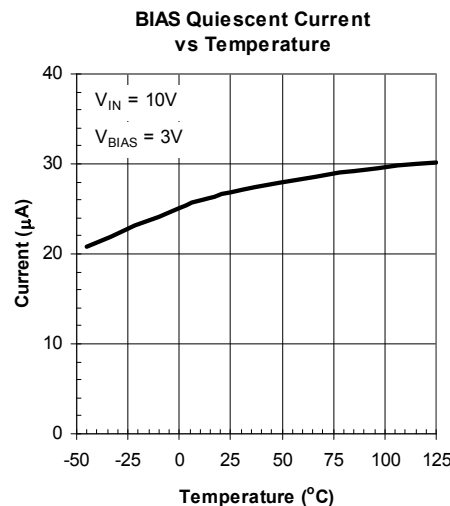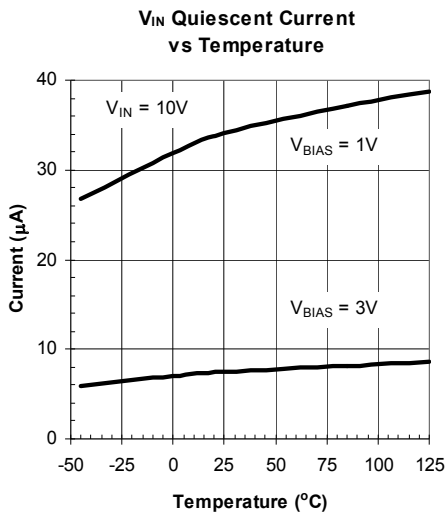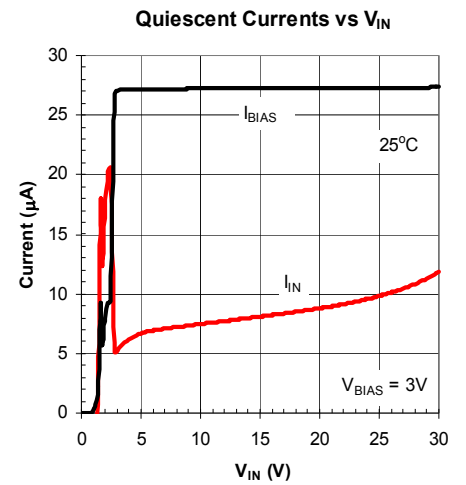| Pin #	Pin Name	Pin Function
1	FB	Inverting input of the error amplifier. The FB pin is tied to a resistive divider between the output and ground. The voltage divider sets the output voltage.
2	BIAS	Anode of the internal bootstrap diode. BIAS also powers the internal control circuit if $V_{BIAS} > 2.3V$. Tie to the output of the DC-DC converter if $V_{OUT} > 2.5V$. Tie BIAS to IN if V_{OUT} is set below 2.5V.
3	BST	Power transistor driver supply. Connect an external bootstrap capacitor from the SW pin to this pin to generate a drive voltage higher than V_{IN} to fully saturate the internal power transistor.
4	SW	The power transistor emitter and the cathode of the freewheeling diode. The SW pin is connected to an inductor and a bootstrap capacitor.
5	GND	Connect this pin to the PC board power ground plane.
6	IN	Power supply to the SC4530. It must be closely bypassed to the ground pin.
7	NC	No Connection.
8	EN	The enable pin for the SC4530. Driving this pin below 0.2V completely shuts off the SC4530. Applying more than 2V to this pin enables the SC4530. If not driven from a control circuit, tie this pin to IN. This pin cannot be floated.
9	Exposed Pad	The exposed pad at the bottom of the package serves as a thermal contact to the circuit board. It is to be soldered to the ground plane of the PC board.

Block Diagram

Figure 2. SC4530 Block Diagram

Typical Characteristics

Efficiency vs Load Current
$V_{OUT} = 3.3V$

Efficiency vs Load Current
$V_{OUT} = 2.5V$

Feedback Voltage vs Temperature

Minimum Off Time vs Temperature

Maximum On Time vs Temperature

Peak and Valley Current DC Thresholds vs Temperature[1]

Switch Saturation Voltage vs Switch Current

BST Pin Current vs Switch Current

Minimum Bootstrap Voltage vs Temperature
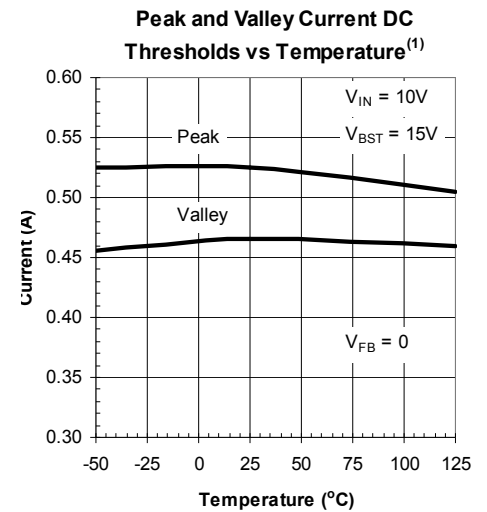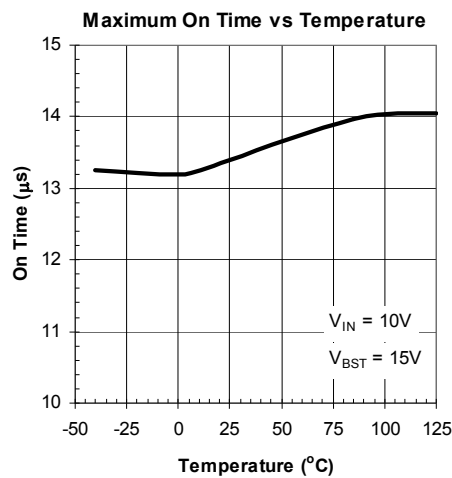
Notes:

(1) Circuit propagation delays and the error amplifier output voltage ripples may cause the actual inductor valley current to differ from its DC value.
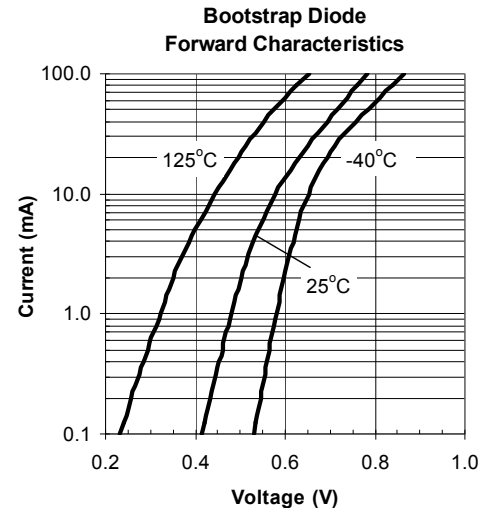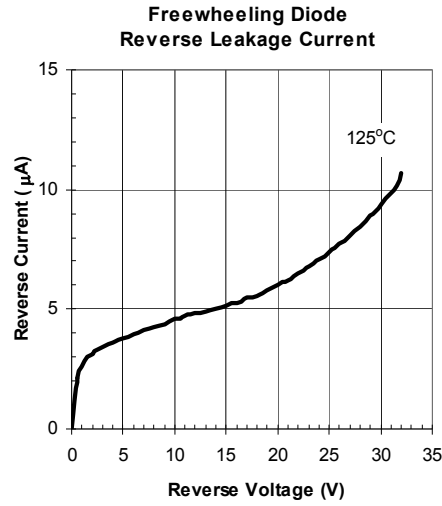
Typical Characteristics (Continued)

Freewheeling Diode Forward Characteristics

Freewheeling Diode Reverse Leakage Current

Bootstrap Diode Forward Characteristics

Bootstrap Diode Reverse Leakage Current

Quiescent Currents vs BIAS Voltage

Quiescent Currents vs V$_{IN}$

V$_{IN}$ Quiescent Current vs Temperature

BIAS Quiescent Current vs Temperature

EN Pin Current vs V$_{EN}$

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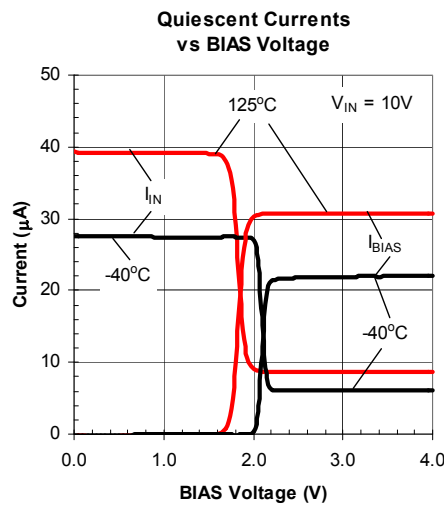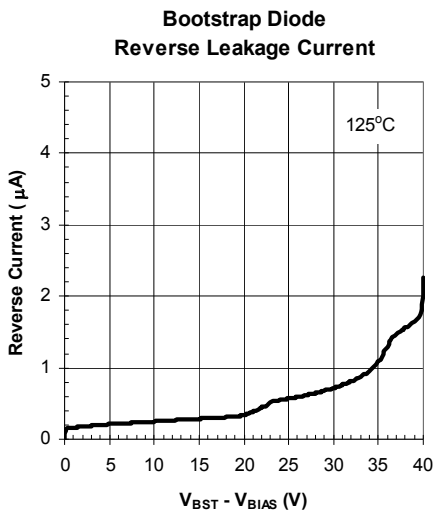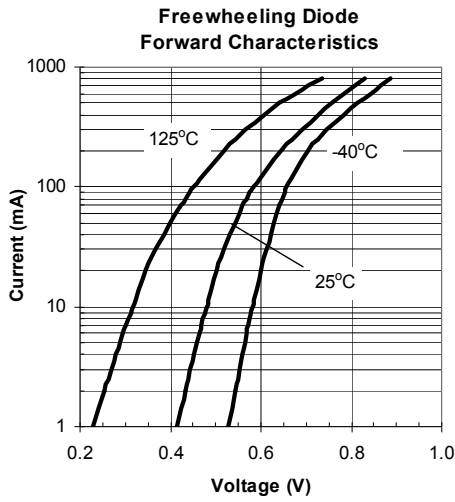
General Description and Operation

The SC4530 is a micropower, hysteretic current-mode step-down switching regulator. As shown in the block diagram in Figure 2, the converter is controlled by an error amplifier EA and two current-sensing comparators IPK and IVLY. IPK and IVLY monitor the switch (Q_1) collector current and the freewheeling diode (D_1) current respectively. The EA amplifies the differential voltage between the FB and the bandgap reference, and produces a current, I_{CNTL}, proportional to its output voltage. I_{CNTL}, in turn, adjusts the switching thresholds of both the peak and valley current comparators. The EA output voltage is high at heavy loads, as is the peak inductor current. The Zener diode D_Z clamps the amplifier output and sets the switch peak current limit.

When the switch Q_1 is turned on, the current through Q_1 ramps up until it reaches the peak threshold set by I_{CNTL}. The output of the IPK comparator, OC, goes high. This resets the latch and turns off the switch. With Q_1 off, the inductor current ramps down through the freewheeling diode D_1. When D_1 current ramps below the valley threshold established by I_{CNTL}, the output of the IVLY comparator, UC, goes high. If Q_1 has been turned off for more than 530ns, then the latch will be set and Q_1 will again turn on, starting a new cycle.

The inductor ripple current in continuous-conduction mode is independent of I_{CNTL} and is primarily determined by V_{OS} and V_{HYS}. Continuous mode switching frequency, therefore, depends on V_{IN}, V_{OUT}, the inductance L and the propagation delay times of the current comparators. If the regulator output is shorted to ground, then the amplifier output will rise to D_Z clamp voltage. Q_1 turns off as the inductor current reaches the peak current limit. With the output shorted to ground, the inductor current ramps down at a slower rate through D_1. Q_1 turns on again when the inductor current crosses the valley threshold. Therefore, short-circuiting the output merely lowers the converter switching frequency. The inductor current remains bounded by the peak switch current limit.

The RUN/IDLE comparator, CMP, monitors the output of the error amplifier. If the EA output falls below the RUN/IDLE threshold, then Q_1 and all control circuits except the reference and EA will be shut off. The output capacitor will then supply the load, causing the output voltage to fall. When the EA output rises above the RUN/IDLE threshold, the control circuit wakes up and the part starts to switch, delivering power to the output. The offset voltage V_{OS} at the input of the IPK comparator ensures that any current pulse delivered to the output has some minimum amplitude.

At very light loads, even a single minimum charge packet delivered to the output will cause the FB voltage to rise above the reference voltage. This causes the EA output voltage to fall and the part to idle. The part resumes switching when the output current discharges the FB voltage below the reference. At light loads, the part switches only as needed to keep the output in regulation. By reducing the supply current drawn when idling, high efficiency is maintained at light loads. At heavier loads, it may take a number of consecutive minimum pulses to bring the FB above the reference voltage. The part enters continuous conduction mode when the amplifier output never falls below the RUN/IDLE threshold.

Driving the base of the power transistor above the input power supply rail minimizes the power transistor turn-on voltage and maximizes efficiency. A bootstrap circuit [formed by an internal bootstrap diode D_2 (Figure 2) and an external capacitor connected between BST and SW] generates a voltage higher than V_{IN} at the BST pin. The bootstrapped voltage becomes the supply voltage of the power transistor driver.

The internal control circuit takes its power from either the input or from the BIAS pin if $V_{BIAS} > 2.3V$. For applications with output voltage higher than 2.5V, the BIAS pin should be tied to the regulator output to maximize efficiency.

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Applications Information

Setting the Output Voltage
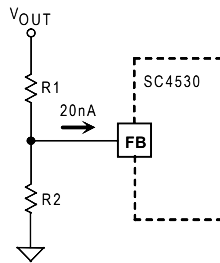
Figure 3. R_1 and R_2 Set the Output Voltage

The SC4530 output voltage is programmed using a resistive divider (Figure 3) with its center tap tied to the FB pin. For a given R_2, R_1 can be determined:

$$R_1 = R_2 \cdot \left(\frac{V_{OUT}}{1.232} - 1 \right) \tag{1}$$

The percentage error due to the input bias current of the error amplifier is:

$$\frac{\Delta V_{OUT}}{V_{OUT}} = \frac{20nA \cdot 100 \cdot (R_1 \| R_2)}{1.232V} \tag{2}$$

Example: Determine the output voltage error caused by the amplifier input bias current in a 5V output converter.

Assuming $R_2 = 200k\Omega$ and using Equations (1) and (2),

$$R_1 = 200k\Omega \cdot \left(\frac{5}{1.232} - 1 \right) \approx 619k\Omega$$

$$\frac{\Delta V_{OUT}}{V_{OUT}} = \frac{20nA \cdot 100 \cdot (200k \| 619k)}{1.232V} = 0.25\%$$

Using large R_1 and R_2 helps in maintaining light-load efficiency, since the current drawn by the feedback resistive divider is not delivered to the converter output. The simple calculation above shows that relatively large R_1 and R_2 can be used without introducing more error than that resulting from the tolerance of the standard 1% resistors.

Maximum Duty Cycle Limitation

The SC4530 is a non-synchronous, step-down switching regulator. Its duty cycle in continuous-conduction mode (CCM) primarily depends on the input and output voltages:

$$D = \frac{V_{OUT} + V_D}{V_{IN} + V_D - V_{CESAT}} \tag{3}$$

where $V_{CESAT} = 0.25V$ is the switch saturation voltage and $V_D = 0.6V$ is the forward voltage drop of the freewheeling diode.

Whenever the power switch is turned off, it is kept off for at least 530ns. Moreover, the control circuit prevents the power transistor from turning on for more than 13.5µs. The inductor current pulls the SW node low as the power switch turns off, allowing the inductor current to charge the bootstrap capacitor. The maximum on-time ensures that the bootstrap capacitor gets replenished after a long switch-on interval. The minimum off-time, together with the maximum on-time, put an upper limit on the achievable duty cycle (≈ 0.96). From Equation (3), the minimum V_{IN} to avoid dropout is:

$$V_{IN(MIN)} = \frac{V_{OUT} + V_D}{0.96} + V_{CESAT} - V_D \tag{4}$$

If V_{IN} falls below this minimum, then the regulator will not be able to attain its set output voltage regardless load. Using Equation (4), the input supply voltage must be at least 5.5V in order to generate a 5V output.

Inductor Selection

The SC4530 uses a hysteretic current-mode control topology. The peak-to-peak inductor ripple current, ΔI_L, is theoretically constant. However, propagation delays of the current comparators (IPK and IVLY in Figure 2), as well as the error amplifier (EA) output ripples, will cause the actual inductor ripple current to vary depending on the input voltage and the duty cycle. The inductor should be chosen so that the valley current comparator, not the minimum off-time, determines the switch turn-on instant. To simplify inductance calculation, we will assume that ΔI_L is constant and equal to 150mA. Furthermore, we will use 1.5 times the typical $t_{OFF(MIN)}$, to allow for tolerance and temperature variation.

$$L_{MIN} = \frac{(V_{OUT} + V_D) \cdot 1.5 \, t_{OFF(MIN)}}{\Delta I_L} \tag{5}$$

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Applications Information (Continued)

For a given V_{IN} and inductance L, the continuous-conduction switching frequency is:

$$f = \frac{(V_{OUT} + V_D)(1-D)}{L \cdot \Delta I_L} = \frac{(V_{OUT} + V_D)(V_{IN} - V_{OUT} - V_{CESAT})}{L(V_{IN} + V_D - V_{CESAT})\Delta I_L} \quad (6)$$

The minimum inductance is first found using Equation (5). Next the switching frequencies are estimated at V_{IN} extremes using Equation (6). The inductance is then adjusted for achieving desired switching frequency. The resulting switch on-time at the maximum V_{IN} must exceed the minimum controllable switch on-time, which can be as high as 180ns. This prevents the inductor current from running away when the output is shorted to ground.

Example: Select the inductor for a 3.3V output regulator, with input voltage ranging from 10V to 26V. The desired switching frequency is about 600kHz.

The minimum inductance is found using Equation (5):

$$L_{MIN} = \frac{(3.3 + 0.6) \times 0.53 \times 1.5}{0.15} \approx 22\mu H$$

Duty cycles and switching frequencies at the input voltage extremes can be found using Equations (3) and (6) respectively. The results for 22µH and 33µH are tabulated (Table 1).

Table 1. Estimated Switching Frequencies for 3.3V Output

Input Voltage V_{IN} (V)	Duty Cycle D (%)	Switching Frequency f (kHz)	
		L = 22µH	L = 33µH
10	37.7	740	490
26	14.8	1000	670

The 33µH inductance will be chosen, as it gives the desired switching frequency range.

The resulting switch on-time is checked against the minimum controllable switch on-time. The switch on-time can be calculated using Equation (7) below:

$$t_{ON} = \frac{L \cdot \Delta I_L}{V_{IN} - V_{CESAT} - V_{OUT}} \quad (7)$$

With L = 33µH, the switch on-time at 10V and 26V V_{IN} are 770ns and 220ns respectively, above the 180ns minimum controllable on-time.

Table 2 lists some recommended inductor values for various output voltages.

Table 2. Recommended Inductor Values

Output Voltage V_{OUT} (V)	Inductor Value (µH)	
	V_{IN} = 16 V	V_{IN} = 30 V
1.8	22	–
2.5	22	33
3.3	22	33
5.0	33	33
12	68	68
18	–	100

Low-cost inductors with powder iron cores are not suitable for high-frequency switching due to their high core losses. Inductors with ferrite cores are recommended for high efficiency.

It should be noted that the inductor saturation current should be designed based on the inductor peak current in output short circuit and startup instead of the nominal output current. If output short circuit protection or fast dVIN/dt (>100V/ms) at startup(EN = VIN) is required at high VIN (up to 30V), at least a 33µH inductor with a minimum 900mA saturation current is needed. If output short circuit does not exist and the dVIN/dt at power on (EN = VIN) is less than 100V/ms, inductors with 600mA or more saturation current may be used. Please check with the inductor manufacturers for the saturation current at the maximum inductor temperature in the real application.

Input Capacitor Selection

A step-down regulator draws pulse current from the input power supply. A capacitor placed between the supply and the converter filters the AC current and keeps the current drawn from the supply to a DC constant. The input capacitance should be high enough to filter the pulse input current. Its equivalent series resistance (ESR) should be low so that power dissipated in the capacitor does not result in significant temperature rise and degrade reliability.

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Applications Information (Continued)

Multi-layer ceramic capacitors, which have very low ESR (a few mΩ) and can easily handle high RMS ripple current, are the ideal choice. A single 4.7µF (X5R or X7R) ceramic capacitor should be enough for most applications. Using a larger capacitor (for example, 10µF) will reduce SW node jitters if the minimum input voltage is less than 0.7V above the output voltage. For applications with high input voltage, a small (1µF ~ 2.2µF) ceramic capacitor can be placed in parallel with a low ESR electrolytic capacitor to satisfy both the ESR and bulk capacitance requirements.

Output Capacitor Selection

The output ripple voltage ΔV_{OUT} of a step-down regulator in continuous conduction can be expressed as:

$$\Delta V_{OUT} = \Delta I_L \left(ESR + \frac{t_{ON} + t_{OFF}}{8 C_{OUT}} \right) \quad (8)$$

where C_{OUT} is the output capacitance.

The first term in Equation (8) results from the equivalent series resistance (ESR) of the output capacitor while the rest is due to the charging and discharging of C_{OUT} by the inductor ripple current.

Substituting $\Delta I_L = 150mA$, $t_{ON} + t_{OFF} = 2\mu s$, $C_{OUT} = 22\mu F$ and $ESR = 3m\Omega$ in Equation (8), we get:

$$\Delta V_{OUT} = 0.15A \cdot (3m\Omega + 11.4m\Omega)$$
$$= 0.45 + 1.7 = 2.2mV$$

Depending on the switching period and the type of the capacitor used, the output voltage ripple resulting from charging/discharging of C_{OUT} may be higher than the ripple due to the ESR. The example above also shows that the output voltage ripple in continuous mode is very low.

The SC4530 relies on fast amplifier response to reduce the output voltage overshoot during power-up. Neither the error amplifier output nor the reference is ramped during start-up. The Zener diode D_Z (refer to page 5) clamps the amplifier output, while the regulator output voltage ramps up. As a result, the switch Q_1 is turned off every cycle at the switch current limit, I_{LIM} (typically 0.5A). The regulator thus delivers about 0.5A to its output until V_{OUT} rises to its set value. If the load is light, then the amplifier output voltage will fall below the RUN/IDLE threshold

following regulation. This causes the regulator to idle. However the energy previously stored in the inductor still flows to the output, causing the output voltage to rise above its regulation level. The minimum output capacitance required to keep the overshoot to less than 1% of the nominal output voltage is:

$$C_{OUT} > \frac{50\,L\,I_{LIM}^2}{V_{OUT}\left(V_{OUT} + V_D\right)} \quad (9)$$

The minimum output capacitance for various output voltages can be estimated from Equation (9) using the inductances given in Table 2. The results are shown in Table 3. Smaller output capacitors may also be used if higher output voltage overshoot is acceptable.

Table 3. Calculated Minimum Output Capacitance for 1% V_{OUT} Overshoot during Start-up

V_{OUT} (V)	Minimum C_{OUT} (µF)	
	$V_{IN} = 16V$	$V_{IN} = 30V$
1.8	64	-
2.5	36	53
3.3	22	33
5.0	15	15
12	5.6	5.6
18	-	3.7

Ceramic capacitors are the best choice for most applications. Sanyo TPE series polymer capacitors in B-case, which offer large capacitors (>100µF) with slightly higher ESR, are also good alternatives. Ripple current in the output capacitor is not a concern because the inductor current of a step-down converter directly feeds C_{OUT}, resulting in very low ripple current. Avoid using Z5U or Y5V ceramic capacitors because these types of capacitors have high temperature and high voltage coefficients.

Bootstrapping the Power Transistor

To reduce the switch on-state voltage and maximize efficiency, the base of the power transistor should be driven from a power supply higher in voltage than V_{IN}. The required driver supply voltage (at least 2.2V higher than the SW) is generated with a bootstrap capacitor C_3 connected between the BST and the SW nodes (Figure

Applications Information (Continued)

1) and the bootstrap diode D_2 (Figure 2). The D_2 anode is connected to the BIAS pin.

During startup, the power transistor in the SC4530 is first switched on so the current flows through to the inductor. When the transistor is switched off, the inductor current pulls the SW voltage low, allowing C_3 to be charged through the internal bootstrap diode D_2. When the power switch is turned on again, the SW voltage goes high. This brings the BST voltage to $V_{SW} + V_{C3}$, thus back-biasing D_2. The C_3 voltage increases with each subsequent switching cycle, as does the bootstrapped voltage at the BST pin. After a number of switching cycles, C_3 will be fully charged to a voltage approximately equal to that applied to the anode of D_2. The minimum BST to SW voltage required to fully saturate the power transistor is shown in the Typical Characteristics (pages 6-7). This difference voltage must be at least 1.72V at room temperature. This is also specified in the Electrical Characteristics (pages 3-4) as the Minimum Bootstrap Voltage. The minimum required V_{C3} increases as temperature decreases. The bootstrap

circuit reaches equilibrium when the base charge drawn from C_3 during transistor on-time is equal to the charge replenished during the off interval.

Figure 4 summarizes various ways of bootstrapping the SC4530. In Figure 4(a) the BIAS pin is connected to the converter output. The bootstrap charge is obtained from the output of the step-down converter. The input-referred charge is reduced by the step-down ratio. This is the most efficient configuration and it also results in the least voltage stress at the BST pin. The maximum BST pin voltage is about $V_{IN} + V_{OUT}$. If the output voltage is between 2.5V and 3V, then a 0.33-0.47µF bootstrap capacitor may be needed to reduce droop. In most other cases, a 0.22µF ceramic capacitor is adequate.

Figure 4(b) shows the SC4530 can also be bootstrapped from the input. This way it is not as efficient as the configuration shown in Figure 4(a). However this may be only option if the output voltage is less than 2.5V and there is no other supply with voltage higher than 2.5V.

(a)

(b)

(c)
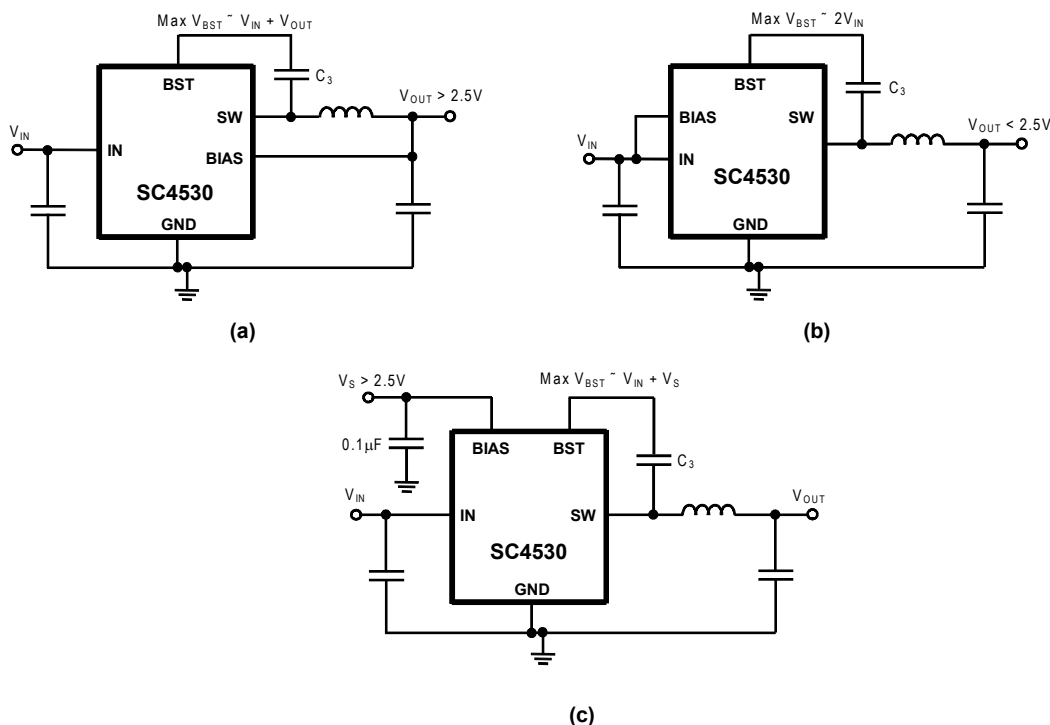
Figure 4. Methods of Bootstrapping the SC4530

Applications Information (Continued)

Voltage stress at the BST pin can be somewhat higher than $2V_{IN}$. The BST pin voltage should not exceed its absolute maximum rating of 42V. Figure 4(c) shows how to bootstrap the SC4530 from an independent power supply V_S with its voltage > 2.5V.

(a)

(b)
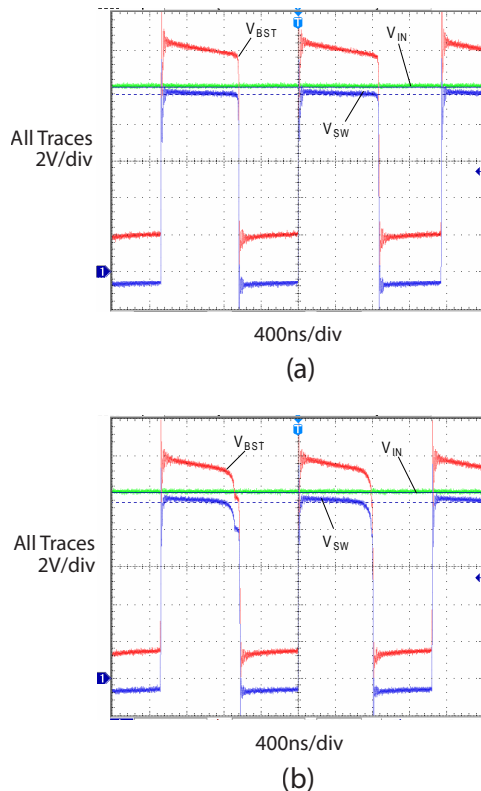
Figure 5. **Switching Waveforms of a 10V to 5V Regulator**

(a) Sufficient Bootstrap Voltage Drives the Power Transistor into Saturation, Minimizing Power Loss.
(b) Excessive Droop in Bootstrap Capacitor Voltage fails to keep the Power Transistor Saturated near the End of its Conduction Cycles, Causing Jitters and Low Efficiency.

To demonstrate the effect of an under-sized bootstrap capacitor, C_3 (Figure 1, page 1) is deliberately reduced to 10nF. The BIAS pin is tied to an external power supply similar to Figure 4(c). By adjusting the external supply voltage V_S, the bootstrap voltage can be varied. Figure 5(a) shows the switching waveforms of a correctly bootstrapped 10V to 5V regulator with $V_S = 2.5V$. All three traces share the

same ground level. When the power transistor is turned on, V_{SW} should come within a few hundred millivolts of V_{IN} and V_{BST} should have at least 2.2V of headroom above V_{SW}. As V_S is reduced to 1.9V, excessive V_{BST} droop decreases transistor driver headroom, as shown in Figure 5(b). The power transistor can no longer be fully saturated (as evidenced by the round V_{SW} turn-off corners), resulting in high power dissipation. When bootstrapping from a low-voltage output or supply, checking the bootstrap voltage is a good precaution.

Since the inductor current charges C_3, the bootstrap circuit requires some minimum load current to function. Figures 6(a) and 6(b) show the minimum input voltage required to saturate the power transistor and to produce a regulated output as a function of the load current. Once started, the bootstrap circuit is able to sustain itself down to zero load.

Feed-Forward Compensation

A feed-forward capacitor C_4 (connected across the upper feedback resistor R_1) is needed for stability. An initial estimate of C_4 can be found using Equation (10) below:

$$C_4 = \frac{6.8 \times 10^{-6}}{R_1} \qquad (10)$$

The value of C_4 can be optimized empirically by observing the inductor current and the output voltage during load transient. Starting with the initial estimate, C_4 is tuned until there is no excessive ringing or overshoot in the inductor current or the output voltage during load transient.

(a)

Applications Information (Continued)

Minimum Input Voltage

(b)
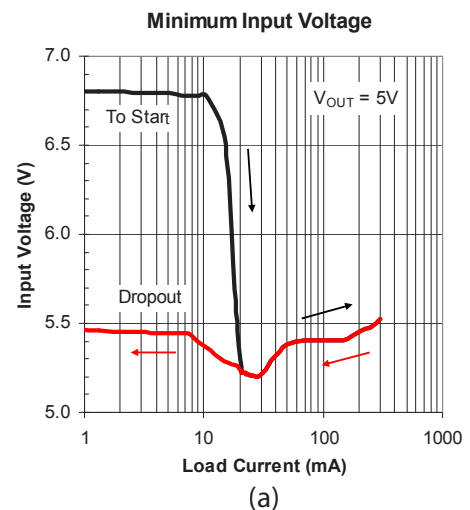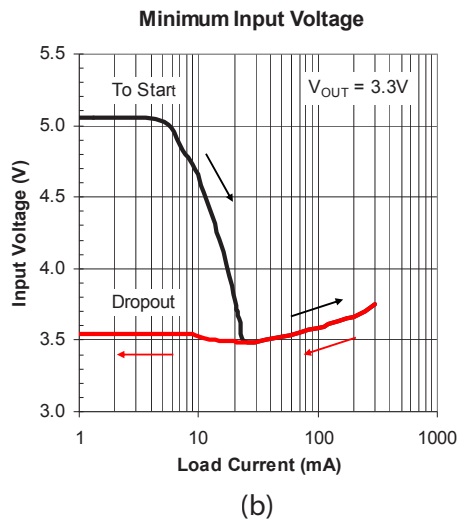
Figure 6. The Minimum Input Voltage Required to Start and to Operate Before Dropout
(a) V_{OUT} =5V
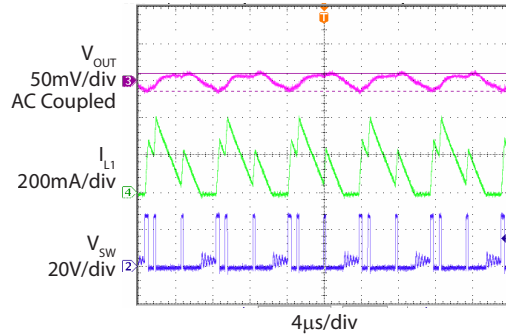**(b) V_{OUT} = 3.3V**

Mode Transition and the FB Pin

If the upper feedback resistor R_1 (Figure 3, page 9) is large and is about the same magnitude as R_2, then fast switching transients may couple into the FB pin, disturbing or delaying the transition from light-load operating mode to continuous-conduction mode (CCM). As described previously, the output ripple voltage is very low in continuous-conduction mode. Delayed CCM transition extends the load range in which the converter produces larger output voltage ripples. This disturbance becomes more pronounced when V_{IN} is increased above 21V and when large feedback resistors are used. The regulator becomes insensitive to switching disturbances after it enters continuous-conduction mode.
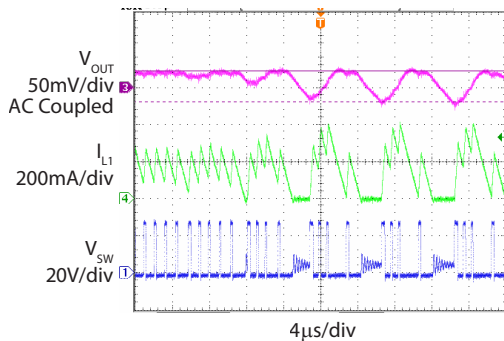
(a) C_4 = 10pF, C_5 not Placed
V_{OUT} = 5V, I_{OUT} = 174mA
CCM Onset I_{OUT} = 176mA

(a) C_4 = 22pF, C_5 not Placed
V_{OUT} = 3.3V, I_{OUT} = 145mA
CCM Onset I_{OUT} = 150mA

(b) C_4 = 10pF, C_5 = 33pF
V_{OUT} = 5V, I_{OUT} = 110mA
CCM Onset I_{OUT} = 126mA
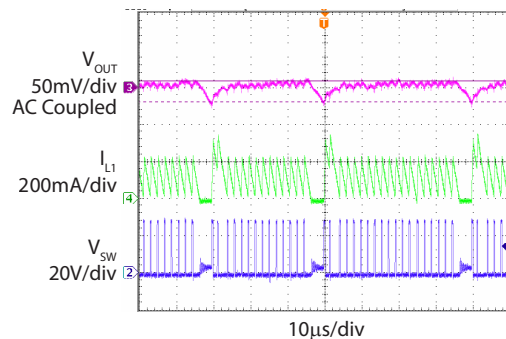
(b) C_4 = 22pF, C_5 = 47pF
V_{OUT} = 3.3V, I_{OUT} = 118mA
CCM Onset I_{OUT} = 121mA

Figure 7. Switching Waveforms of a 28V to 5V Converter Just Before It Enters Continuous-Conduction Mode

Figure 8. Switching Waveforms of a 28V to 3.3V Converter Just Before It Enters Continuous-Conduction Mode

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Applications Information (Continued)

The operating mode transition can be significantly smoothed by filtering the FB node. A capacitor between FB pin and ground (Capacitor C_5, as shown in Figures 13(a), page 18) serves this purpose. It should be chosen so that it improves mode transition without significantly slowing down load transients. Switching waveforms of a 5V output regulator (Figure 13(a), page 18) immediately before it enters continuous-conduction mode are shown in Figure 7. The inductor current waveform appears to be more jagged without filtering. Moreover, transition to CCM occurs at an output current of 176mA, instead of 126mA with FB filtering. Figure 8 compares the corresponding switching waveforms of an output 3.3V (Figure 14(a), page 19) regulator.

If the converter output voltage is 1.8V or less, or if R_2 is reduced to below 2kΩ, then C_5 will not be necessary. C_5 is also optional in Figures 13(a) and 14(a) if the maximum V_{IN} never exceeds 21V. Bench testing shows that removing C_5

from these converters still results in acceptable transitional behavior, provided that $V_{IN} < 21V$.

C_5 can be estimated using the following empirical equation:

$$C_5 = \frac{8 \times 10^{-6}}{R_1 \| R_2} - C_4 \qquad (11)$$

FB filtering has no significant impact on the output ripple voltage. However, it improves the converter efficiency by 0.25% to 0.5% around the mode transition point (Figure 9). Regulator efficiencies are slightly lower (< 0.25%) at light loads when filtering the FB voltage. Positive values in Figure 9 imply that FB filtering improves efficiency compared to no filtering.

Reverse Input Protection

Consider a circuit board where the input power source supplies several DC-DC converters, including an SC4530
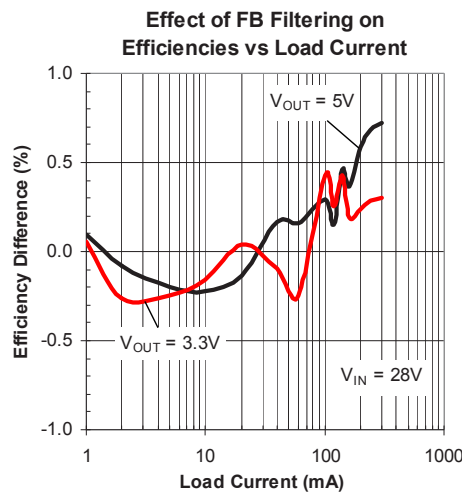
Figure 9. Effect of FB Filtering on Converter Efficiency (V_{IN} = 28V)
Plotted Efficiency = the Efficiency of a FB-Filtered Converter -
the Efficiency of the Same Converter without FB-Filtering

Applications Information (Continued)

regulator with a large output capacitor. During power-off, the SC4530 regulator output may be held high by its output capacitor, while V_{IN} is discharged rapidly by other DC-DC converters. If V_{IN} falls to two diode voltages below V_{OUT}, then the parasitic junction diodes inside the SC4530 (see Figure 2, page 5) will draw current from the output through the SW pin to the input. If the load is light and the output capacitor is large, then high reverse current will flow, or even damage the internal circuits.

Figure 10 shows two protection schemes. In Figure 10(a), a Schottky diode D_4 placed at the input blocks the reverse

current. This method has the disadvantage that it lowers the converter efficiency. A PN junction diode placed from the converter output to the input [(as shown in Figure 10(b)] shunts the reverse current away from the part, thus protecting the part. This scheme is not suitable in a power supply system where a backup battery is diode OR-ed with the SC4530 regulator output and with the SC4530 input grounded.

Board Layout Considerations

In a step-down switching regulator, the input bypass capacitor, the main power switch and the freewheeling power diode carry pulse current with high di/dt (Figure 11). To minimize jittering, the size of the loop formed by these components must be minimized. Since the main power switch and the freewheeling diode are already integrated inside the part, connecting the input bypass capacitor close to the ground pin minimizes size of the switched current loop.

(a)

(b)

Figure 10. Reversed Input Protection Schemes

(a) D_4 Blocks the Reverse Current
(b) D_4 Shunts the Reverse Current from the Part During Power-off.
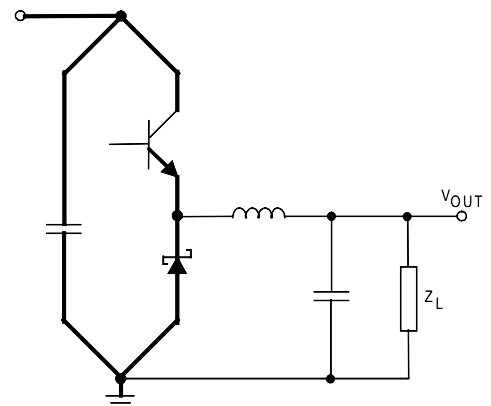
Figure 11. Heavy Lines Show the Fast Switching Current Paths in a Step-down Converter. The Input Capacitor Should be Placed Close to the Part for Improved Switching Performance.

Applications Information (Continued)

Shortening the traces at the SW and BST nodes reduces the parasitic trace inductance at these nodes. This not only reduces EMI, but also decreases switching voltage spikes at these nodes. Shielding the FB trace from the SW and the BST nodes with ground traces is a good precaution in mitigating switching transient disturbance.

Figure 12 shows an example of external component placement around the SC4530. The exposed pad should be soldered to a large power ground plane as the ground copper acts as a heat sink for the device.
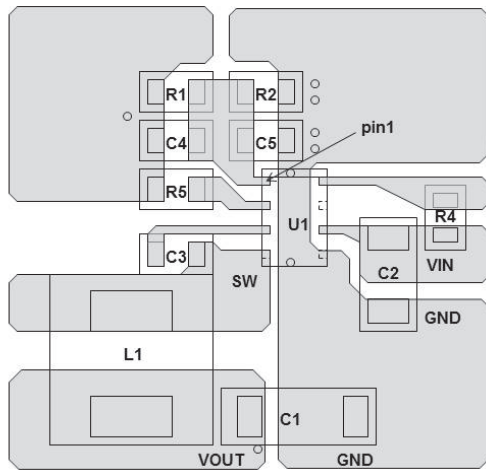
Figure 12. Suggested PCB Layout for the SC4530
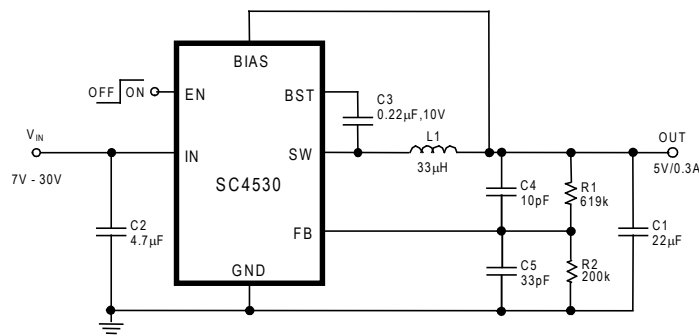
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Typical Application Circuits

(a)

(b)

(c)

(d)

(e)
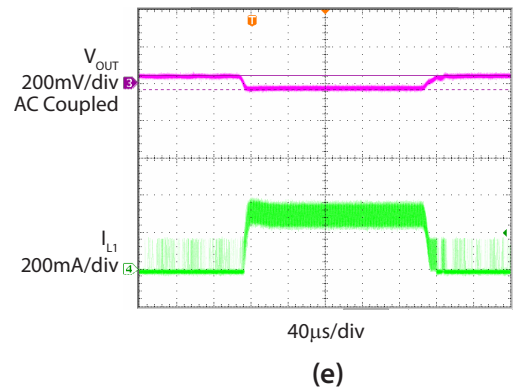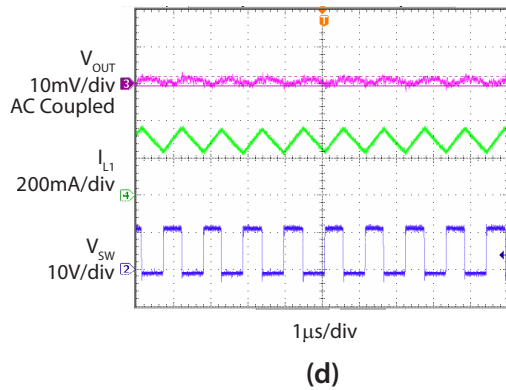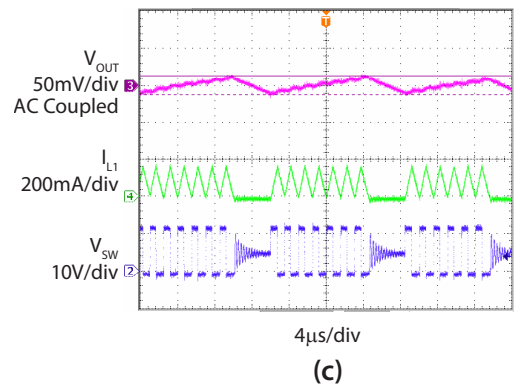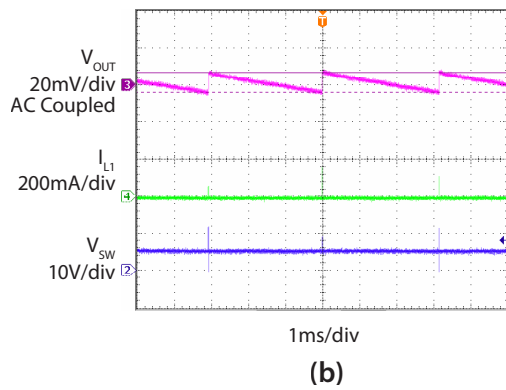
Figure 13. (a) 5V Output Step-Down Converter
(b) Switching Waveforms of the Figure 13(a) Circuit. V_{IN} = 12V, I_{OUT} = 10µA
(c) V_{IN} = 12V, I_{OUT} = 60mA
(d) V_{IN} = 12V, I_{OUT} = 300mA
(e) Load Transient. V_{IN} = 12V, I_{OUT} is Switched Between 0 and 300mA

Typical Application Circuits (Continued)
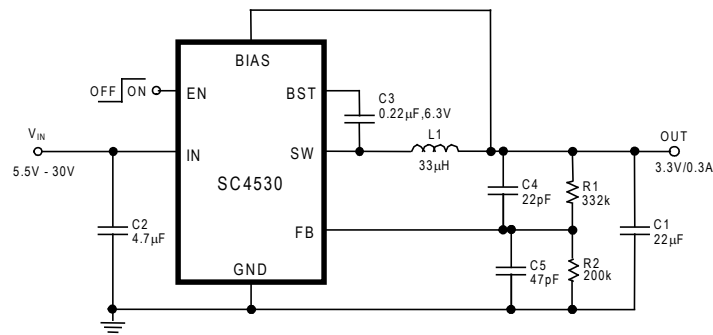
L1: Coilcraft LPS6225

C1: Murata GRM31CR60J226K
C2: Murata GRM31CR71H475K

(a)

400μs/div

(b)

4μs/div

(c)

1μs/div

(d)

40μs/div
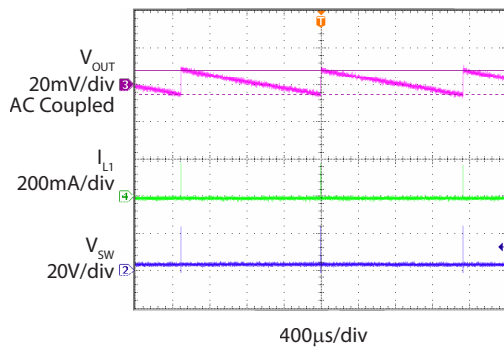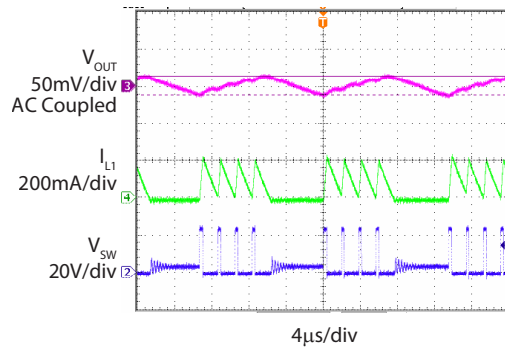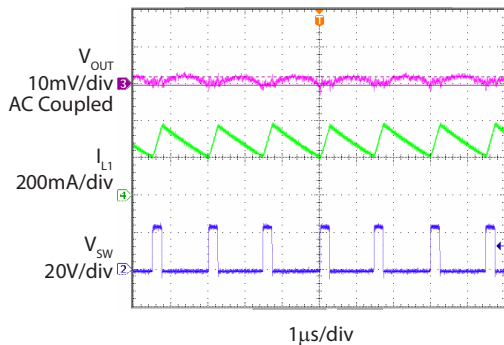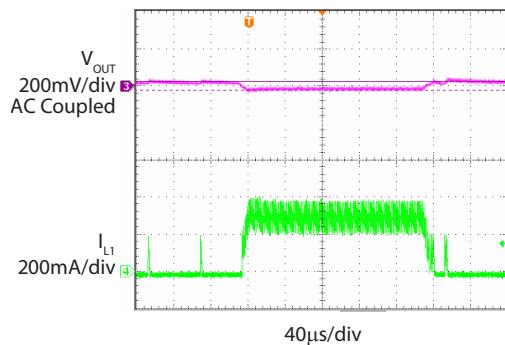
(e)

Figure 14. **(a) 3.3V Output Step-Down Converter**
(b) Switching Waveforms of the Figure 14(a) Circuit. V_{IN} = 24V, I_{OUT} = 100μA
(c) V_{IN} = 24V, I_{OUT} = 60mA
(d) V_{IN} = 24V, I_{OUT} = 300mA
(e) Load Transient. V_{IN} = 24V, I_{OUT} is Switched Between 0 and 300mA

Typical Application Circuits (Continued)
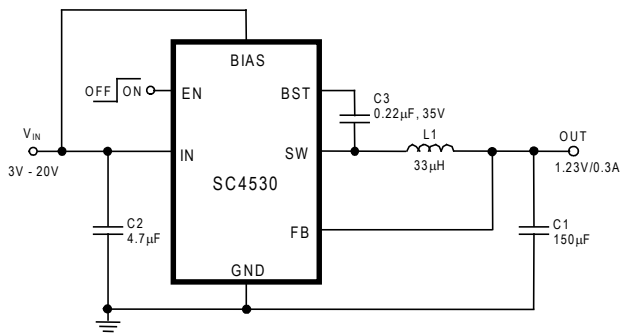
L1: Coilcraft LPS6225   C1: Sanyo POSCAP 4TPE150MAUB
C2: Murata GRM31CR71H475K

Figure 15. 1.23V Output Step-Down Converter
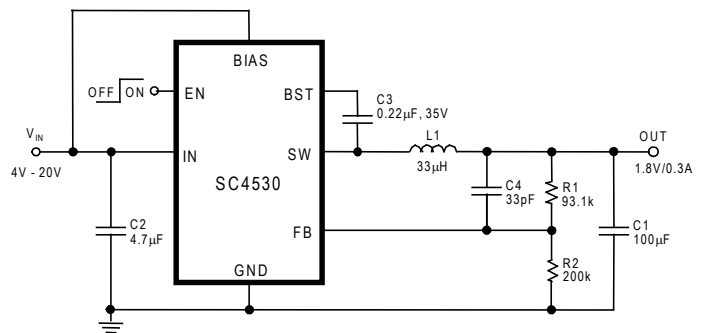
L1: Coilcraft LPS6225   C1: Murata GRM31CR60G107M
C2: Murata GRM31CR71H475K

Figure 16. 1.8V Output Step-Down Converter
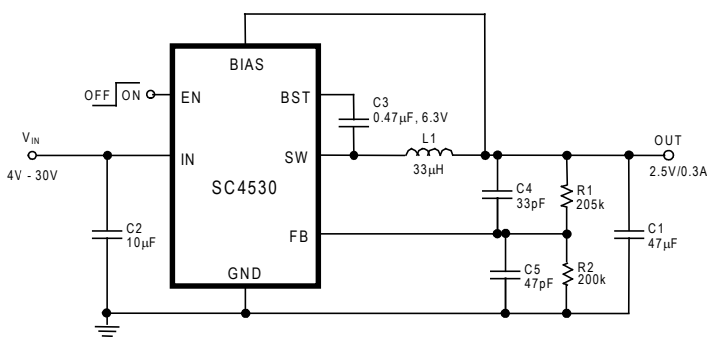
L1: Coilcraft LPS6225   C1: Murata GRM21BR60J476M
C2: Murata GRM32ER71H106K

Figure 17. 2.5V Output Step-Down Converter
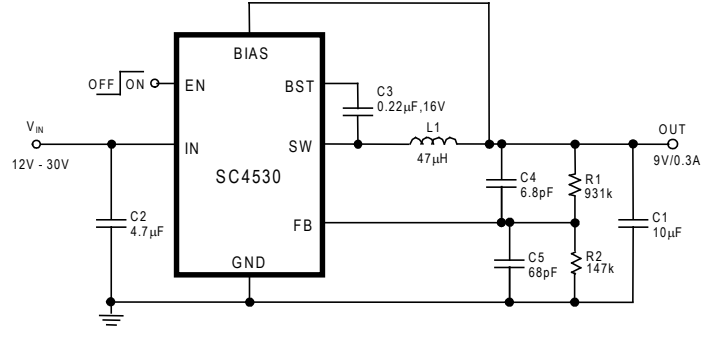
L1: Coilcraft LPS6225   C1: Murata GRM31CR61C106K
C2: Murata GRM31CR71H475K

Figure 18. 9V Output Step-Down Converter
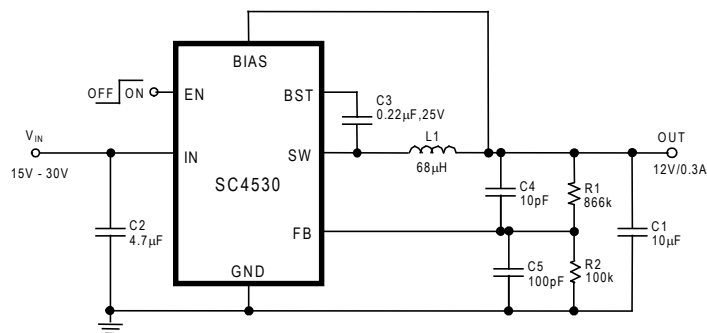
L1: Coilcraft LPS6235   C1: Murata GRM31CR61C106K
C2: Murata GRM31CR71H475K

Figure 19. 12V Output Step-Down Converter

Outline Drawing – MLPD-W-8 3x2
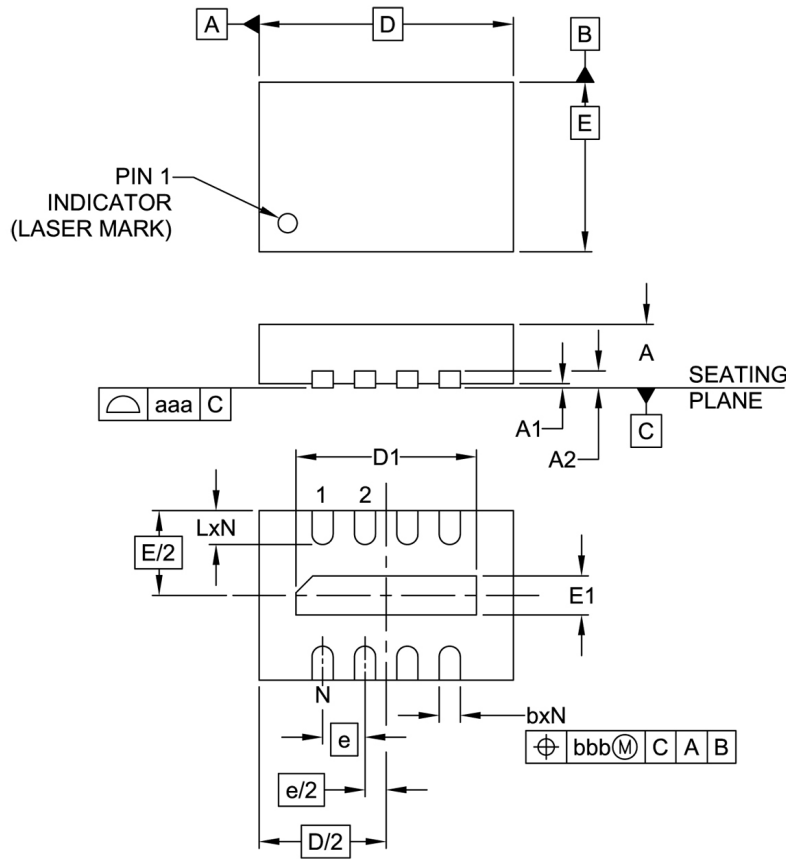
DIMENSIONS			
DIM	MILLIMETERS		
	MIN	NOM	MAX
A	0.70	0.75	0.80
A1	0.00	0.02	0.05
A2	-	(0.20)	-
b	0.18	0.25	0.30
D	2.90	3.00	3.10
D1	1.984	2.134	2.234
E	1.90	2.00	2.10
E1	0.409	0.559	0.659
e	0.50 BSC		
L	0.30	0.40	0.50
N	8		
aaa	0.08		
bbb	0.10		

NOTES:
1. CONTROLLING DIMENSIONS ARE IN MILLIMETERS (ANGLES IN DEGREES).
2. COPLANARITY APPLIES TO THE EXPOSED PAD AS WELL AS TERMINALS.

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Land Pattern – MLPD-W-8 3x2
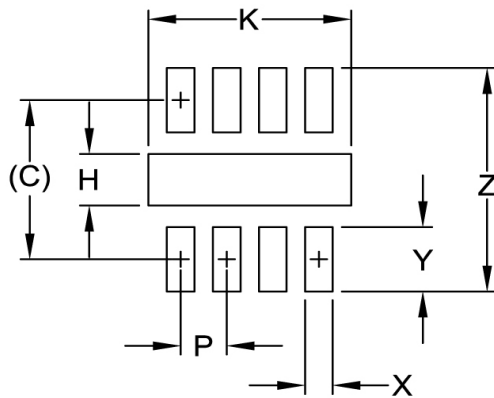
DIMENSIONS	
DIM	MILLIMETERS
C	(1.85)
G	1.10
H	0.60
K	2.20
P	0.50
X	0.30
Y	0.75
Z	2.60

NOTES:

1. CONTROLLING DIMENSIONS ARE IN MILLIMETERS (ANGLES IN DEGREES).

2. THIS LAND PATTERN IS FOR REFERENCE PURPOSES ONLY.
    CONSULT YOUR MANUFACTURING GROUP TO ENSURE YOUR
    COMPANY'S MANUFACTURING GUIDELINES ARE MET.

3. THERMAL VIAS IN THE LAND PATTERN OF THE EXPOSED PAD
    SHALL BE CONNECTED TO A SYSTEM GROUND PLANE.
    FAILURE TO DO SO MAY COMPROMISE THE THERMAL AND/OR
    FUNCTIONAL PERFORMANCE OF THE DEVICE.

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