MIC28511

60V$_{IN}$, 3A Synchronous Buck Regulator

General Description

The MIC28511 is a synchronous step-down switching regulator with internal power switches capable of providing up to 3A output current from a wide input supply range from 4.6V to 60V. The output voltage is adjustable down to 0.8V with a guaranteed accuracy of ±1%. A constant switching frequency can be programmed from 200kHz to 680kHz. The Hyper Speed Control™ and HyperLight Load® architectures of the MIC28511 allow for high V$_{IN}$ (low V$_{OUT}$) operation and ultra-fast transient response while reducing the required output capacitance and providing very good light-load efficiency.

The MIC28511 offers a full suite of protection features to ensure protections under fault conditions. These include under-voltage lockout to ensure proper operation under power sag conditions, internal soft start to reduce inrush current, foldback current limit, "hiccup" mode short-circuit protection and thermal shutdown.

Datasheets and support documentation are available on Micrel's web site at: www.micrel.com.

Features

- 4.6V to 60V operating input voltage supply
- Up to 3A output current
- Integrated high-side and low-side N-channel MOSFETs
- HyperLight Load (MIC28511-1) and Hyper Speed Control (MIC28511-2) architecture
- Enable input and power good (PGOOD) output
- Programmable current limit and foldback "hiccup" mode short-circuit protection
- Built-in 5V regulator for single-supply operation
- Adjustable 200kHz to 680kHz switching frequency
- Fixed 5ms soft-start
- Internal compensation and thermal shutdown.
- Thermally-enhanced 24-pin 3mm × 4mm FCQFN package
- Junction temperature range of –40°C to +125°C

Applications

- Industrial power supplies
- Distributed supply regulation
- Base station power supplies
- Wall transformer regulation
- High-voltage single board systems

Typical Application

Efficiency (VIN =12V) vs. Output Current MIC28511-1
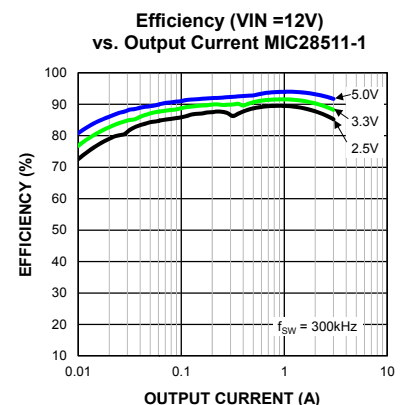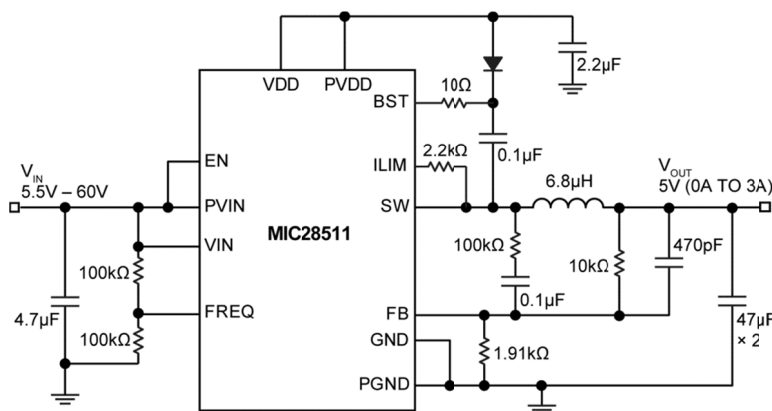
Hyper Speed Control and Ramp Control are trademarks of Micrel, Inc.
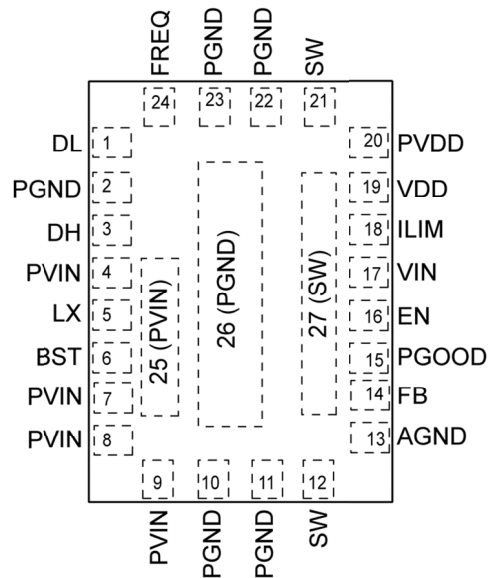HyperLight Load is a registered trademark of Micrel, Inc.

Ordering Information

Part Number	Architecture	Package[1]	Junction Temperature Range	Lead Finish
MIC28511-1YFL	HyperLight Load	24-Pin 3mm × 4mm FCQFN	–40°C to +125°C	Pb-Free
MIC28511-2YFL	Hyper Speed Control	24-Pin 3mm × 4mm FCQFN	–40°C to +125°C	Pb-Free

Note:

1. FCQFN is a lead-free package. Pb-free lead finish is Matte Tin.

Pin Configuration

24-Pin 3mm × 4mm FCQFN (FL)
(Top View)

Pin Description

Pin Number	Pin Name	Pin Description
1	DL	Low-Side Gate Drive. Internal low-side power MOSFET gate connection. This pin must be left unconnected, or floating.
2	PGND	PGND is the return path for the low-side driver circuit. Connect to the source of low-side MOSFET's (PGND, pins 10, 11 22, 23, and 26) through a low-impedance path.
3	DH	High-Side Gate Drive. Internal high-side power MOSFET gate connection. This pin must be left unconnected, or floating.
4, 7, 8, 9, 25 (25 is ePad)	PVIN	Power Input Voltage. The PVIN pins supply power to the internal power switch. Connect all PVIN pins together and bypass locally with ceramic capacitors. The positive terminal of the input capacitor should be placed as close as possible to the PVIN pins, the negative terminal of the input capacitor should be placed as close as possible to the PGND pins 10,11, 22, 23, and 26.
5	LX	The LX pin is the return path for the high-side driver circuit. Connect the negative terminal of the bootstrap capacitor directly to this pin. Also connect this pin to the SW pins 12, 21, and 27, with a low-impedance path. The controller monitors voltages on this and PGND for zero current detection.
6	BST	Bootstrap Pin. This pin provides bootstrap supply for the high-side gate driver circuit. Connect a 0.1μF capacitor and an optional resistor in series from the LX (pin 5) to the BST.

Pin Description (Continued)

Pin Number	Pin Name	Pin Description
10, 11, 22, 23, 26 (26 is ePad)	PGND	Power Ground. These pins are connected to the source of the low-side MOSFET. They are the return path for the step-down regulator power stage and should be tied together. The negative terminal of the input decoupling capacitor should be placed as close as possible to these pins.
12, 21, 27 (27 is ePad)	SW	Switch Node. The SW pins are the internal power switch outputs. These pins should be tied together and connected to the output inductor.
13	AGND	Analog Ground. The analog ground for VDD and the control circuitry. The analog ground return path should be separate from the power ground (PGND) return path.
14	FB	Feedback Inout. The FB pin sets the regulated output voltage relative to the internal reference. This pin is connected to a resistor divider from the regulated output such that the FB pin is at 0.8V when the output is at the desired voltage.
15	PGOOD	The power good output is an open drain output requiring an external pull-up resistor to external bias. This pin is a high impedance open circuit when the voltage at FB pin is higher than 90% of the feedback reference voltage (typically 0.8V).
16	EN	Enable Input. The EN pin enables the regulator. When the pin is pulled below the threshold, the regulator will shut-down to an ultra-low current state. A precise threshold voltage allows the pin to operate as an accurate UVLO. Do not tie EN to VDD
17	VIN	Supply voltage for the internal LDO. The VIN operating voltage range is from 4.6V to 60V. A ceramic capacitor from VIN to AGND is required for decoupling. The decoupling capacitor should be placed as close as possible to the supply pin.
18	ILIM	Currrent Limit Setting. Connect a resistor from this pin to the SW pin node to allow for accurate current limit sensing programming of the internal low-side power MOSFET.
19	VDD	Internal +5V Linear Regulator: VDD is the internal supply bus for the IC. Connect to an external 1µF bypass capacitor. When V_{IN} is <5.5V, this regulator operates in drop-out mode. Connect VDD to VIN.
20	PVDD	A 5V supply input for the low-side N-channel MOSFET driver circuit, which can be tied to VDD externally. A 1µF ceramic capacitor from PVDD to PGND is recommended for decoupling.
24	FREQ	Switching Frequency Adjust pin. Connect this pin to VIN to operate at 680kHz. Place a resistor divider network from VIN to the FREQ pin to program the switching frequency.

Absolute Maximum Ratings[2]

PVIN, VIN to PGND .... −0.3V to 65V
VDD, PVDD to PGND .... ……−0.3V to 6V
V_{BST} to V_{SW}, V_{LX}........ ……………….…………….−0.3V to 6V
$V_{BST\ to}$ PGND …………….…..…………−0.3V to (VIN +6V)
V_{SW}, to PGND...…………………...…..-0.3V to (VIN +0.3V)
V_{LX}, V_{FB}, V_{PG}, V_{FREQ}, V_{ILIM}, V_{EN} to AGND
………………………………… -0.3V to (VDD+ +0.3V)
PGND to AGND …………….…....................-0.3V to +0.3V
Junction Temperature (T_J) .... +150°C
Storage Temperature (T_S)......................... −65°C to +150°C
Lead Temperature (soldering, 10s)............................ 300°C
ESD HBM Rating[4].. 1.5kV
ESD MM Rating[4]...150V

Operating Ratings[3]

Supply Voltage (PVIN, VIN) .... 4.6V to 60V
Enable Input (V_{EN})...0V to VIN
V_{SW}, V_{FEQ}, V_{ILIM}, V_{EN}..0V to VIN
Junction Temperature (T_J) .... −40°C to +125°C
Junction Thermal Resistance
     3mm × 4mm FCQFN-24 ($θ_{JA}$)............................30°C/W

Electrical Characteristics[5]

V_{IN} = 12V; T_A = 25°C, unless noted. **Bold** values indicate −40°C ≤ T_J ≤ +125°C.

Parameter	Condition	Min.	Typ.	Max.	Units
Power Supply Input					
Input Voltage Range (PV$_{IN}$, V$_{IN}$)		**4.6**		**60**	**V**
Quiescent Supply Current	V$_{FB}$ = 1.5V (MIC28511-1)		0.4	**0.75**	mA
	V$_{FB}$ = 1.5V (MIC28511-2)		0.7	**1.5**	
Shutdown Supply Current	SW = unconnected, V$_{EN}$ = 0V		0.1	**10**	µA
VDD Supply					
VDD Output Voltage	V$_{IN}$ = 7V to 60V, I$_{VDD}$ = 10mA	**4.8**	5.2	**5.4**	V
VDD UVLO Threshold	V$_{VDD}$ rising	**3.8**	4.2	**4.6**	V
VDD UVLO Hysteresis			400		mV
Load Regulation @40mA		0.6	2	4.0	%
Reference					
Feedback Reference Voltage	0°C ≤ T$_J$ ≤ 85°C (±1.0%)	**0.792**	0.8	**0.808**	V
	−40°C ≤ T$_J$ ≤ 125°C (±2%)	**0.784**	0.8	**0.816**	
FB Bias Current	V$_{FB}$ = 0.8V		5	**500**	nA
Enable Control					
EN Logic Level High		**1.8**			V
EN Logic Level Low				**0.6**	V
EN Hysteresis			200		mV
EN Bias Current	V$_{EN}$ = 12V		5	**40**	µA

Notes:

2. Exceeding the absolute maximum ratings may damage the device.

3. The device is not guaranteed to function outside its operating ratings.

4. Devices are ESD sensitive. Handling precautions are recommended. Human body model, 1.5kΩ in series with 100pF.

5. Specification for packaged product only.

Electrical Characteristics[5] (Continued)

V_{IN} = 12V; T_A = 25°C, unless noted. **Bold** values indicate $-40°C \leq T_J \leq +125°C$.

| Parameter	Condition	Min.	Typ.	Max.	Units
Oscillator					
Switching Frequency	V_{FREQ} = V_{IN}	**450**	680	**800**	kHz
	V_{FREQ} = 50%V_{IN}		340		
Maximum Duty Cycle			85		%
Minimum Duty Cycle	VFB>0.8V		0		%
Minimum Off-time		110	200	270	ns
Internal MOSFETs					
High-Side NMOS On-Resistance			51		mΩ
Low-Side NMOS On-Resistance			28		mΩ
Short-Circuit Protection					
Current-Limit Threshold	V_{FB} = 0.79V	−30	−14	0	mV
Short-Circuit Threshold	V_{FB} = 0V	−24	−7	8	mV
Current-Limit Source Current	V_{FB} = 0.79V	50	70	90	µA
Short-Circuit Source Current	V_{FB} = 0V	25	36	43	µA
Leakage					
SW, BST Leakage Current				50	µA
Power Good (PGOOD)					
PGOOD Threshold Voltage	Sweep V_{FB} from low-to-high	**85**	90	**95**	%V_{OUT}
PGOOD Hysteresis	Sweep V_{FB} from low-to-high		6		%V_{OUT}
PGOOD Delay Time	Sweep V_{FB} from low-to-high		100		µs
PGOOD Low Voltage	V_{FB} < 90% × V_{NOM}, I_{PGOOD} = 1mA		70	**200**	mV
Thermal Protection					
Overtemperature Shutdown	T_J Rising		160		°C
Overtemperature Shutdown Hysteresis			**15**		°C
Soft Start					
Soft-Start Time			**5**		ms

Typical Characteristics

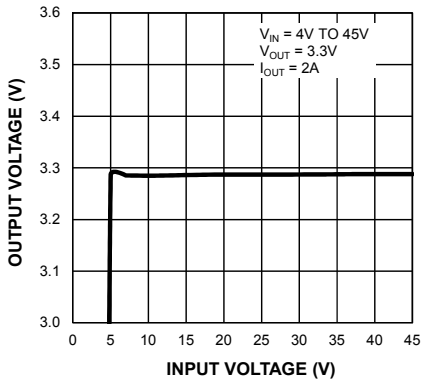**VIN Operating Supply Current
vs. Input Voltage MIC28511-1**

V_{OUT} = 5V
I_{OUT} = 0A
f_{SW} = 300kHz
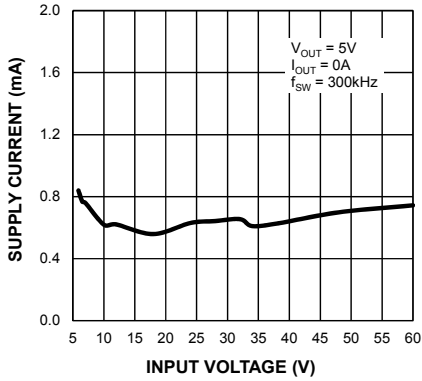
**VIN Shutdown Current
vs. Input Voltage**
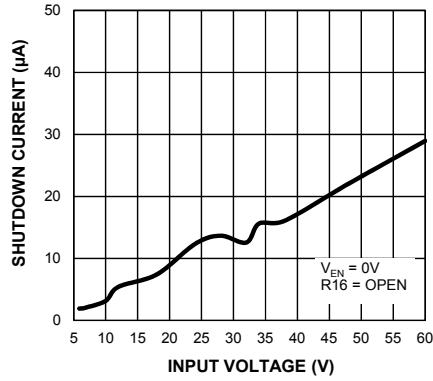
V_{EN} = 0V
R16 = OPEN

**VDD Voltage
vs. Input Voltage MIC28511-1**

I_{DD} = 10mA
I_{DD} = 40mA
V_{OUT} = 5.0V

**Output Voltage
vs. Input Voltage MIC28511-1**

V_{IN} = 4V TO 45V
V_{OUT} = 3.3V
I_{OUT} = 2A
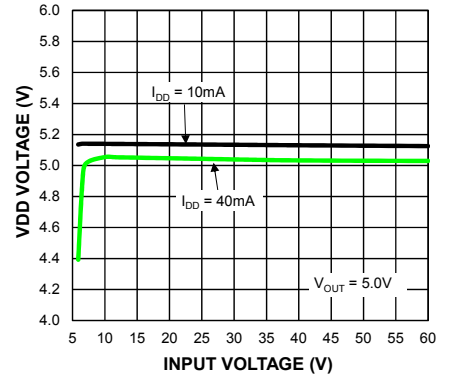
**Enable Threshold
vs. Input Voltage MIC28511-1**
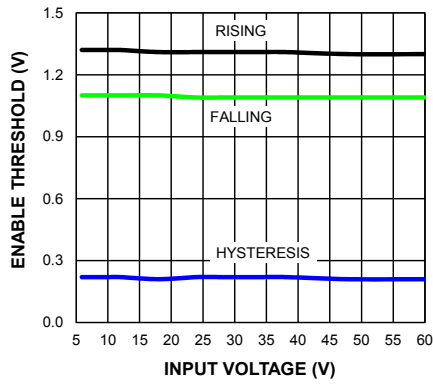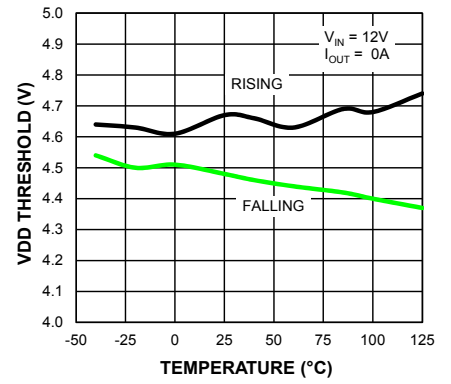
RISING
FALLING
HYSTERESIS

**VDD UVLO Threshold
vs. Temperature MIC28511-1**

V_{IN} = 12V
I_{OUT} = 0A
RISING
FALLING

**Output Peak Current Limit
vs. Temperature MIC28511-1**

V_{IN} = 12V
V_{OUT} = 5.0V
f_{SW} = 300kHz

**Feedback Voltage
vs. Temperature MIC28511-1**
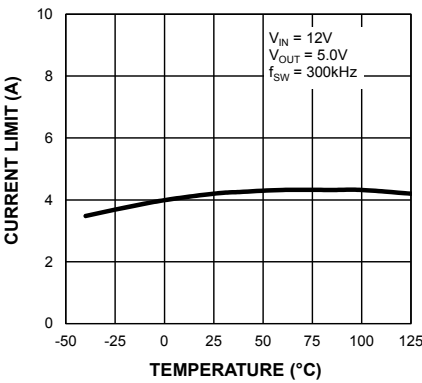
V_{IN} = 12V
V_{OUT} = 5.0V
I_{OUT} = 0A

**Enable Threshold
vs. Temperature MIC28511-1**
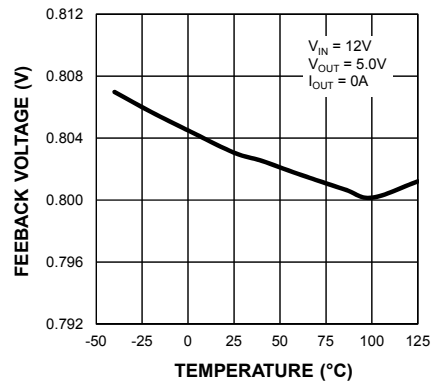
V_{IN} = 12V
V_{DD} = 5V
RISING
FALLING
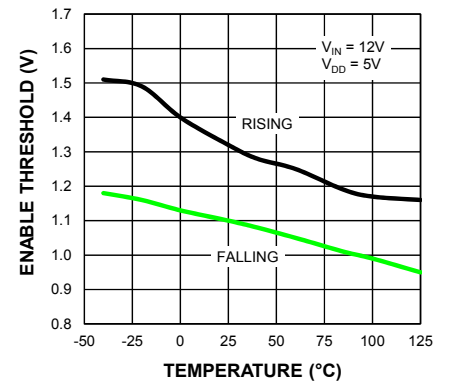
Typical Characteristics (Continued)

**Output Voltage
vs. Output Current MIC28511-1**
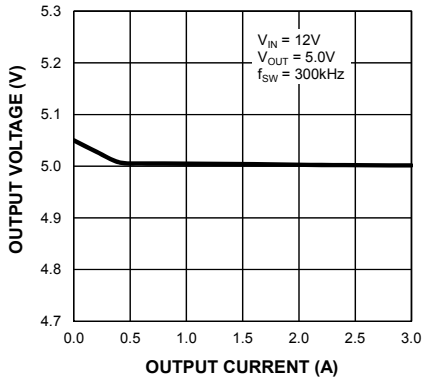
V_{IN} = 12V
V_{OUT} = 5.0V
f_{SW} = 300kHz

**Efficiency (VIN =12V)
vs. Output Current MIC28511-1**

5.0V
3.3V
2.5V

f_{SW} = 300kHz

**Efficiency (VIN = 24V)
vs. Output Current MIC28511-1**

5.0V
3.3V
2.5V

f_{SW} = 300kHz

**Efficiency (VIN = 48V)
vs. Output Current MIC28511-1**

5.0V
3.3V
2.5V

f_{SW} = 300kHz

**Switching Frequency
vs. Output Current MIC28511-1**

V_{IN} = 12V
V_{OUT} = 5.0V

**IC Power Dissipation
vs. Output Current MIC28511-1**

V_{IN} = 12V
f_{SW} = 300kHz

5.0V
3.3V
2.5V

**12V Input Thermal Derating
MIC28511-1**

2.5V
5.0V
3.3V

V_{IN} =12V
f_{SW} = 300kHz
T_{JMAX} =125°C
θ_{JA} = 30°C/W

**IC Power Dissipation
vs. Output Current MIC28511-1**

V_{IN} = 24V
f_{SW} = 300kHz

5.0V
3.3V
2.5V

**IC Power Dissipation
vs. Output Current MIC28511-1**

V_{IN} = 48V
f_{SW} = 300kHz

5.0V
3.3V
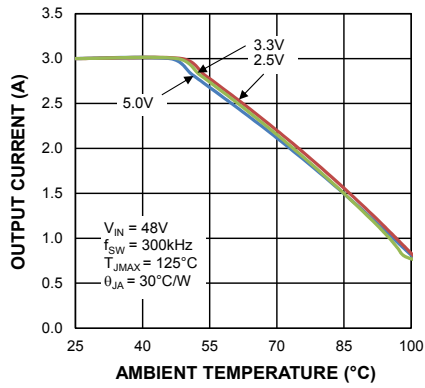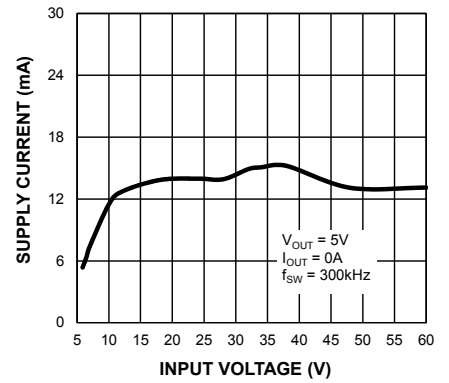2.5V
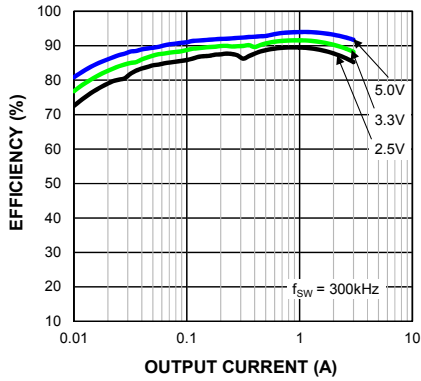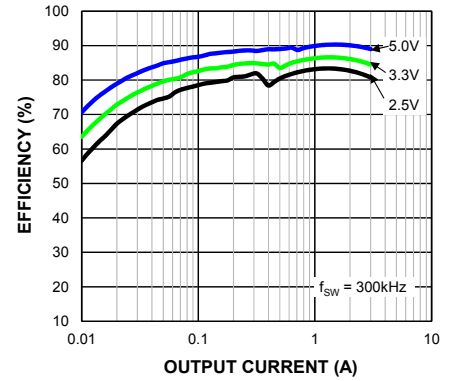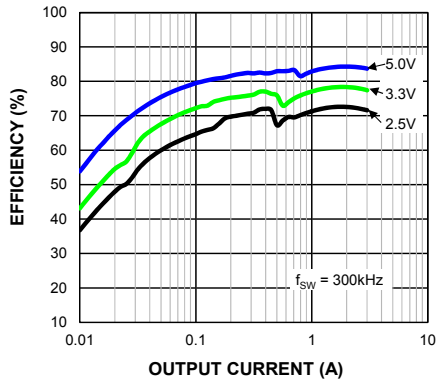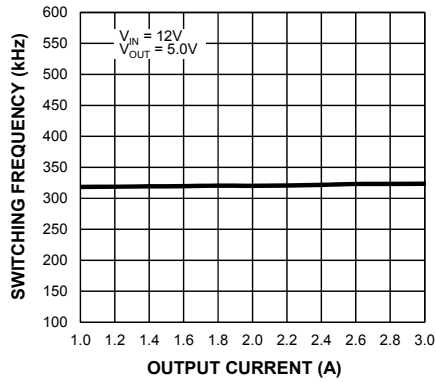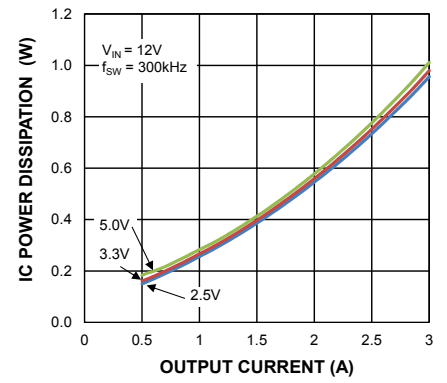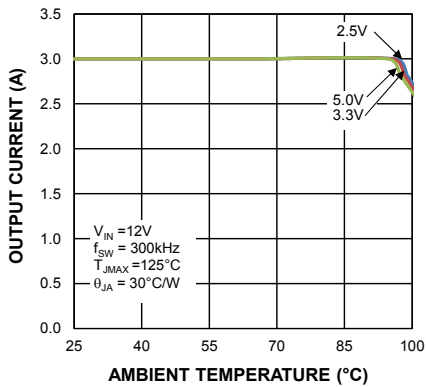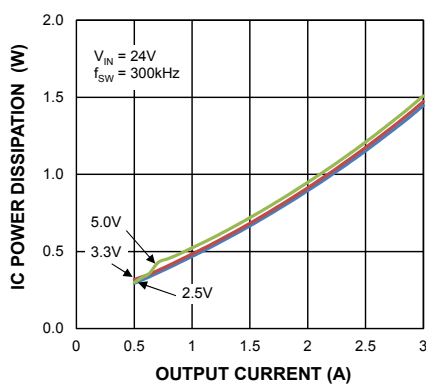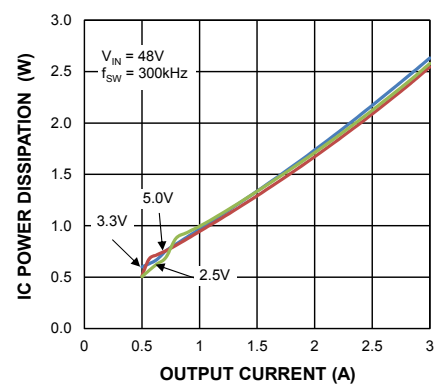
Typical Characteristics (Continued)

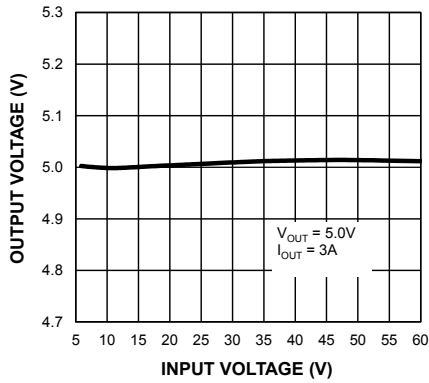**24V Input Thermal Derating MIC28511-1**
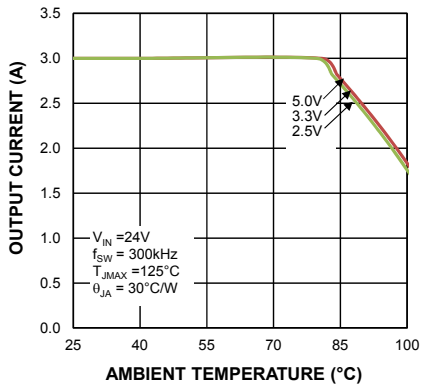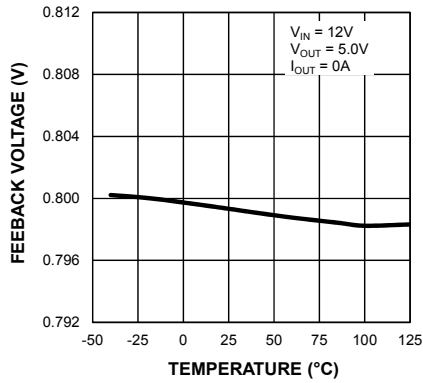

48V Input Thermal Derating MIC28511-1

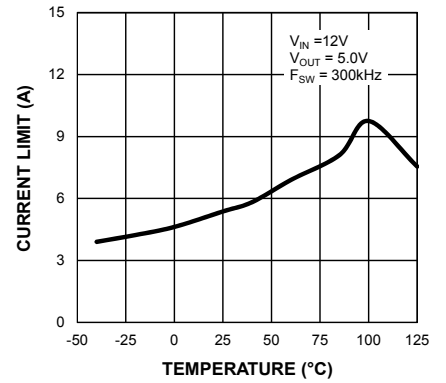**VIN Operating Supply Current vs. Input Voltage MIC28511-2**

Output Voltage vs. Input Voltage
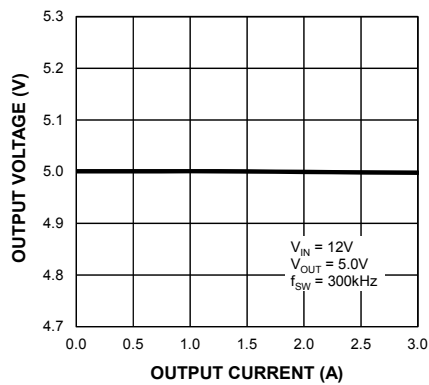

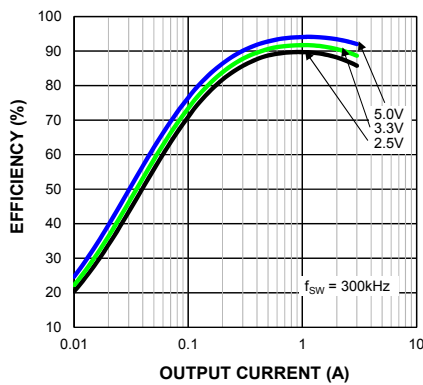**Feedback Voltage vs.Temperature MIC28511-2**

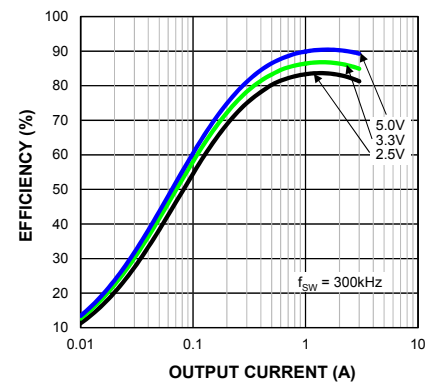**Output Peak Current Limit vs. Temperature MIC28511-2**

Output Voltage vs. Output Current MIC28511-2

Efficiency (VIN = 12V) vs. Output Current MIC28511-2

Efficiency (VIN = 24V) vs. Output Current MIC28511-2

Typical Characteristics (Continued)

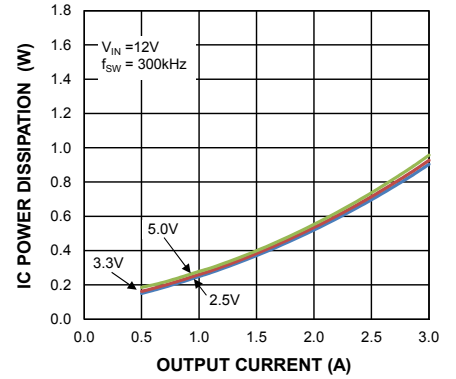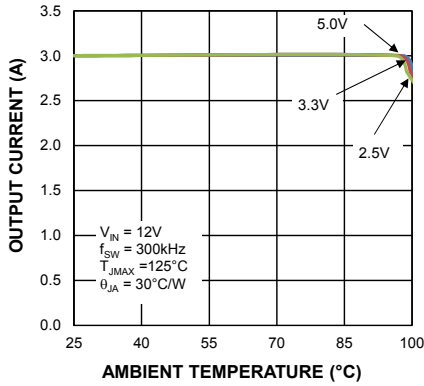**Efficiency (VIN = 48V)
vs. Output Current MIC28511-2**
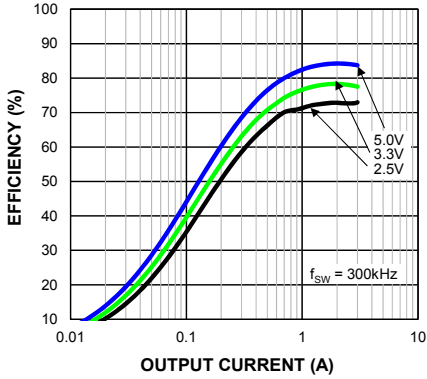
**Switching Frequency
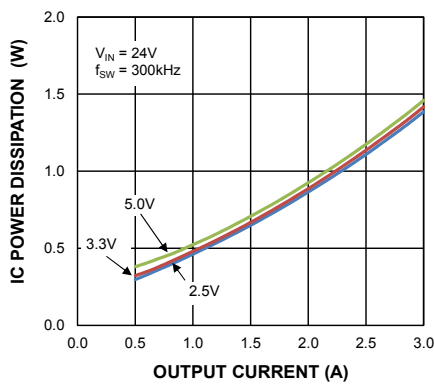vs. Output Current MIC28511-2**
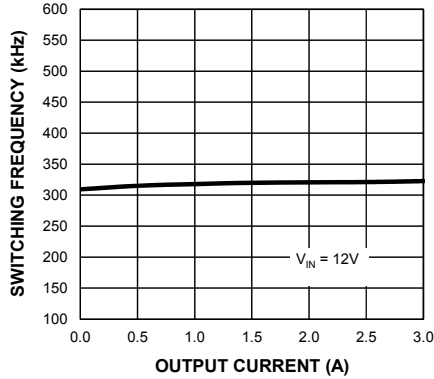
**IC Power Dissipation
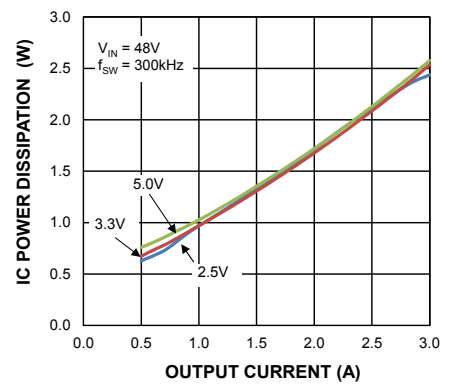vs. Output Current MIC28511-2**

**12V Input Thermal Derating
MIC28511-2**

**IC Power Dissipation
vs. Output Current MIC28511-2**

**IC Power Dissipation
vs. Output Current MIC28511-2**

**24V Input Thermal Derating
MIC28511-2**
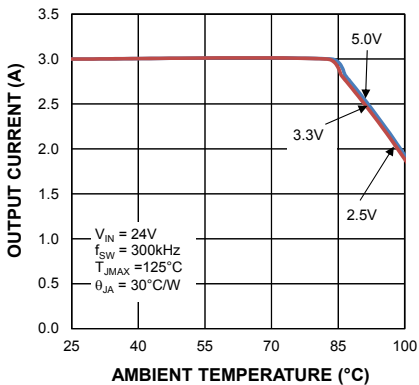
**48V Input Thermal Derating
MIC28511-2**
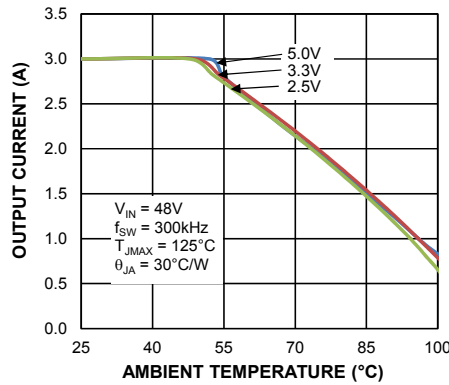
Functional Characteristics

Enable Turn-On Delay Time
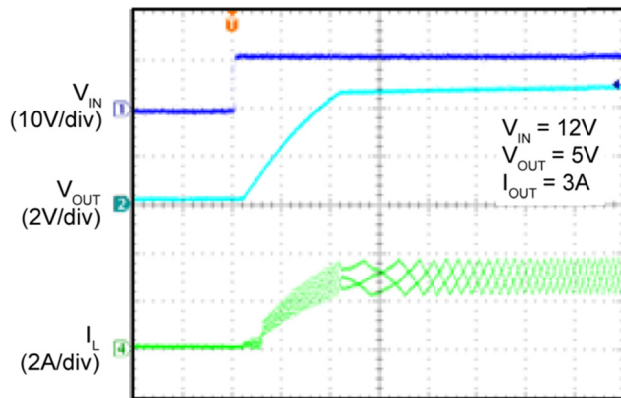
$V_{IN} = 12V$
$V_{OUT} = 5V$
$I_{OUT} = 3A$

V_{IN} (10V/div)
V_{OUT} (2V/div)
I_L (2A/div)

Time (2.0ms/div)

Start-Up with Pre-Biased Output at 1V

$V_{IN} = 12V$
$V_{OUT} = 5V$
$I_{OUT} = 0A$
$V_{PRE-BIAS} = 1V$

V_{IN} (10V/div)
LX (10V/div)
V_{OUT} (2V/div)

Time (2.0ms/div)

Start-Up with Pre-Biased Output at 2V

$V_{IN} = 12V$
$V_{OUT} = 5V$
$I_{OUT} = 0A$
$V_{PRE-BIAS} = 2V$
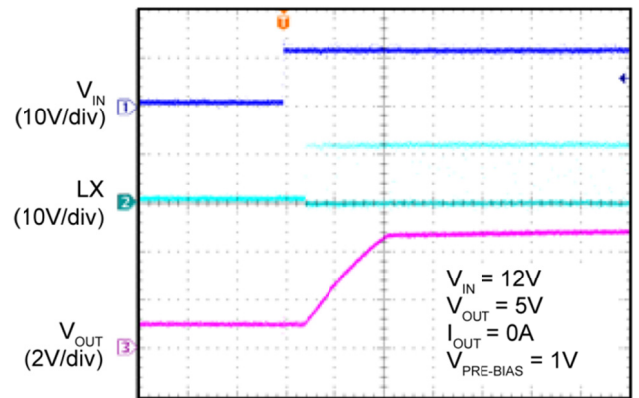
V_{IN} (10V/div)
LX (10V/div)
V_{OUT} (2V/div)
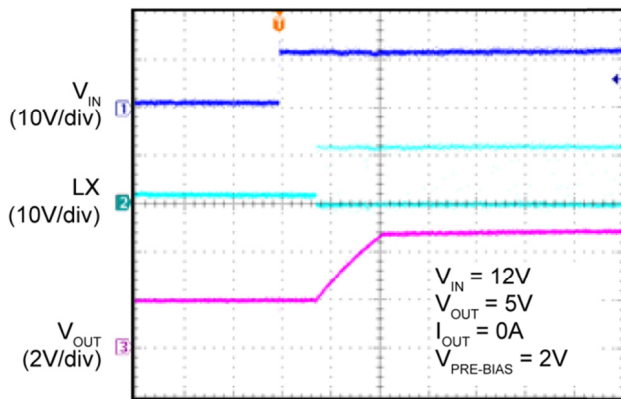
Time (2.0ms/div)

Enable Turn-Off Delay and Fall Time

$V_{IN} = 12V$
$V_{OUT} = 5V$
$I_{OUT} = 3A$

V_{EN} (2V/div)
V_{OUT} (2V/div)
I_L (2A/div)

Time (2.0ms/div)

V_{IN} Turn-On

$V_{IN} = 12V$
$V_{OUT} = 5V$
$I_{OUT} = 3A$

V_{IN} (10V/div)
LX (10V/div)
V_{OUT} (5V/div)
I_{OUT} (2A/div)
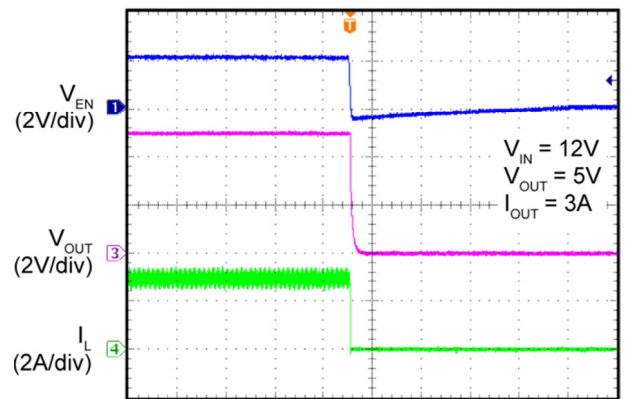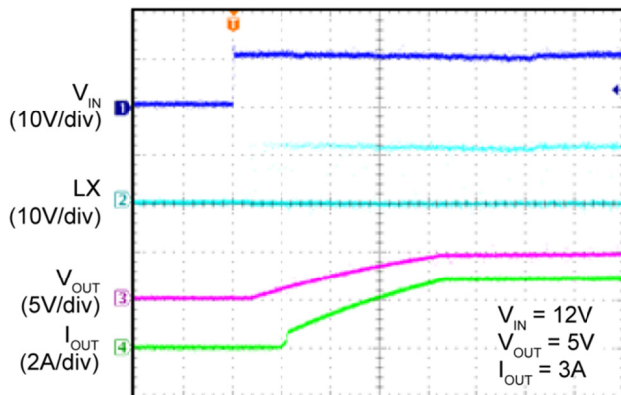
Time (1.0ms/div)

V_{IN} Turn-Off

$V_{IN} = 12V$
$V_{OUT} = 5V$
$I_{OUT} = 3A$

V_{IN} (10V/div)
LX (10V/div)
V_{OUT} (5V/div)
I_L (2A/div)
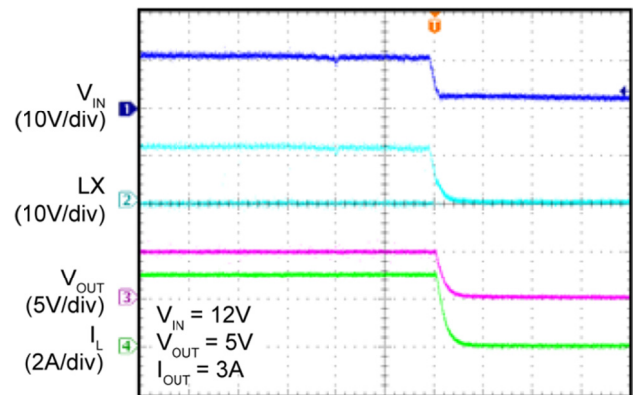
Time (1.0ms/div)

Functional Characteristics (Continued)

Enable Turn-On/Turn-Off

V_{EN} (2V/div)

V_{IN} = 12V
V_{OUT} = 5V
I_{OUT} = 3A

V_{OUT} (2V/div)

Time (10ms/div)

V_{IN} UVLO Thresholds

V_{IN} (2V/div)

V_{OUT} = 3.3V
I_{OUT} = 0.5A
V_{EN} = 5V

V_{OUT} (2V/div)

Time (20ms/div)
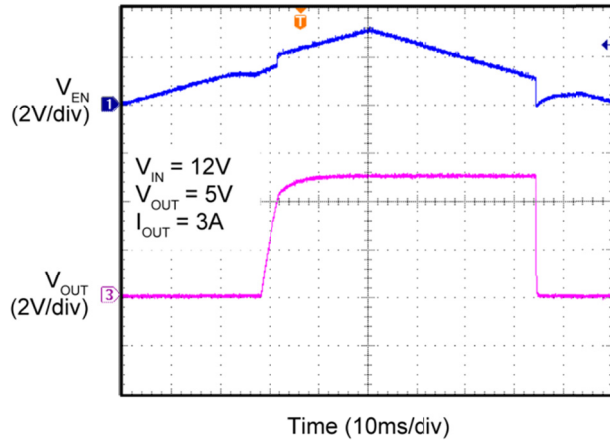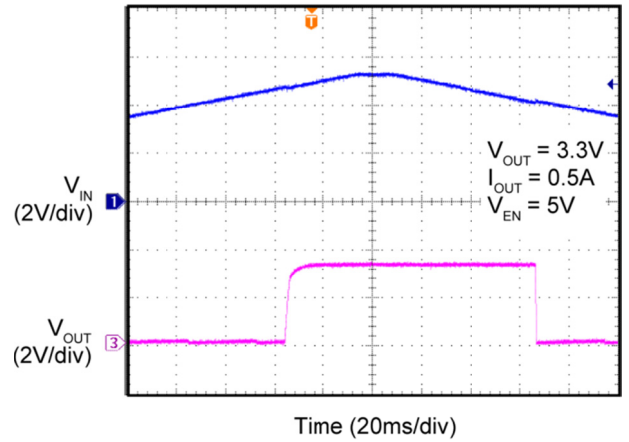
Power-Up into Short Circuit
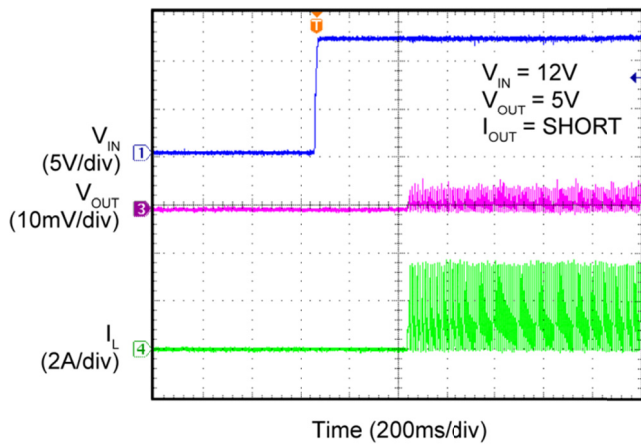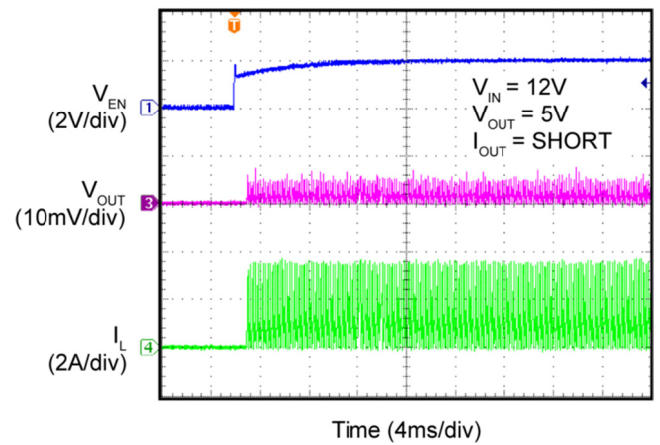
V_{IN} = 12V
V_{OUT} = 5V
I_{OUT} = SHORT

V_{IN} (5V/div)
V_{OUT} (10mV/div)

I_L (2A/div)

Time (200ms/div)

Enabled into Short

V_{EN} (2V/div)

V_{IN} = 12V
V_{OUT} = 5V
I_{OUT} = SHORT

V_{OUT} (10mV/div)

I_L (2A/div)

Time (4ms/div)

Output Peak Curent-Limit Threshold

V_{OUT} (5V/div)

I_L (5A/div)

V_{IN} = 12V
V_{OUT} = 5V

Time (40ms/div)

Short Circuit

V_{IN} = 12V
V_{OUT} = 5V

V_{OUT} (2V/div)

I_L (2A/div)

Time (20μs/div)
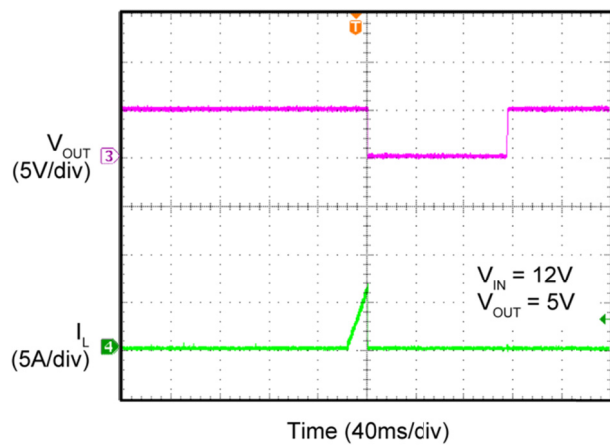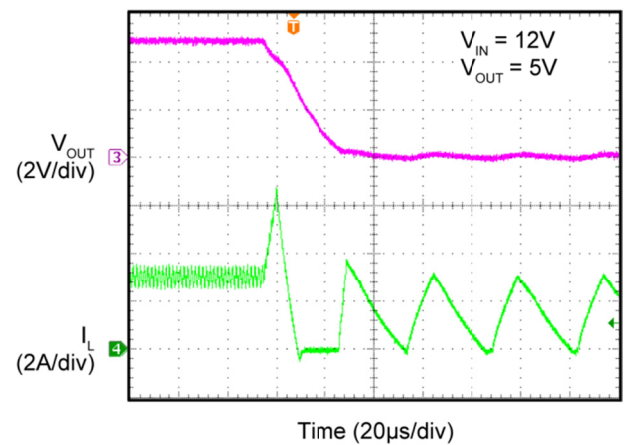
Functional Characteristics (Continued)

Output Recovery from Short Circuit

V_{IN} = 12V
V_{OUT} = 5V
I_{OUT} = 3A TO SHORT

V_{OUT} (2V/div)

I_L (2A/div)

Time (200µs/div)
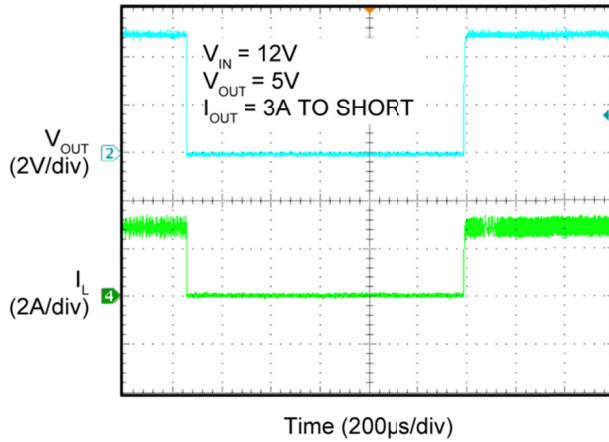
Output Recovery from Short Circuit (Zoomed-In Version)

V_{IN} = 12V
V_{OUT} = 5V
I_{OUT} = 3A TO SHORT

V_{OUT} (2V/div)

I_L (2A/div)

Time (100µs/div)
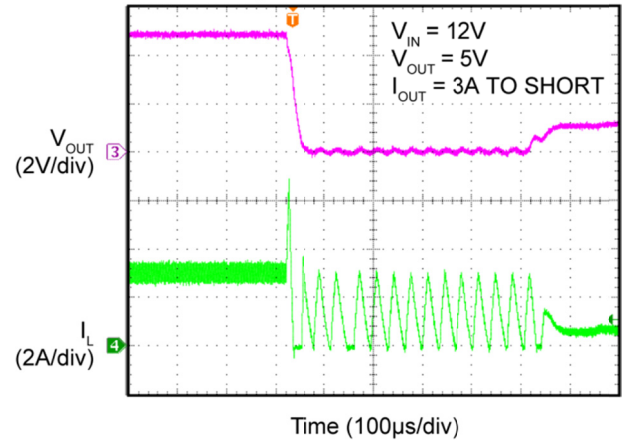
Output Recovery from Thermal Shutdown
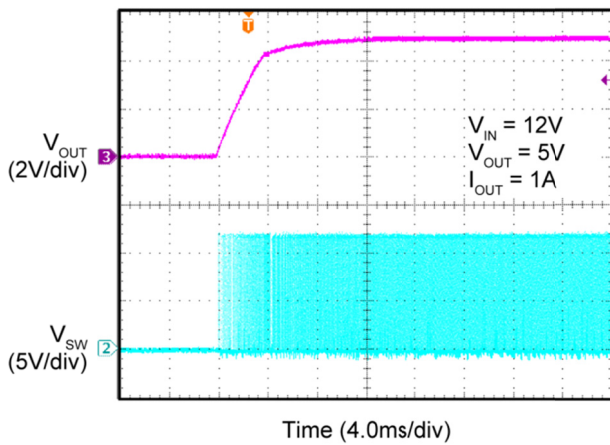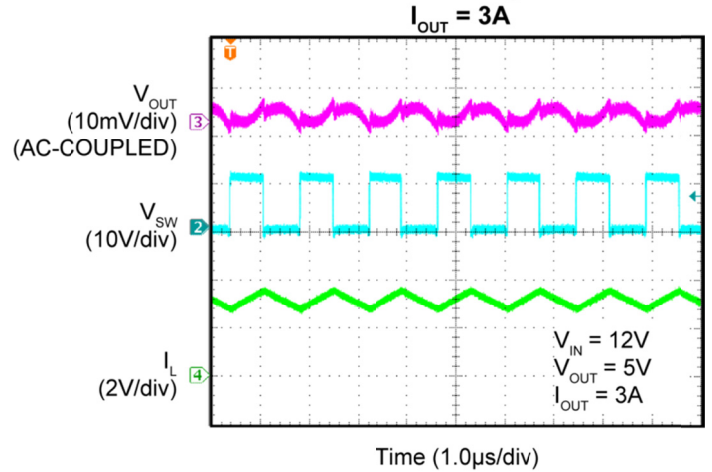
V_{IN} = 12V
V_{OUT} = 5V
I_{OUT} = 1A

V_{OUT} (2V/div)

V_{SW} (5V/div)

Time (4.0ms/div)

MIC28511-2 Switching Waveforms I_{OUT} = 3A

V_{OUT} (10mV/div) (AC-COUPLED)

V_{SW} (10V/div)

I_L (2V/div)

V_{IN} = 12V
V_{OUT} = 5V
I_{OUT} = 3A

Time (1.0µs/div)

MIC28511-2 Switching Waveforms I_{OUT} = 0A

V_{IN} = 12V, V_{OUT} = 5V, I_{OUT} = 0A

V_{OUT} (10mV/div) (AC-COUPLED)

V_{SW} (10V/div)

I_L (2V/div)

Time (1.0µs/div)

MIC28511-1 Switching Waveforms I_{OUT} = 0A

V_{IN} = 12V, V_{OUT} = 5V, I_{OUT} = 0A

V_{OUT} (50mV/div) (AC-COUPLED)

V_{SW} (10V/div)

I_L (1V/div)

Time (400µs/div)
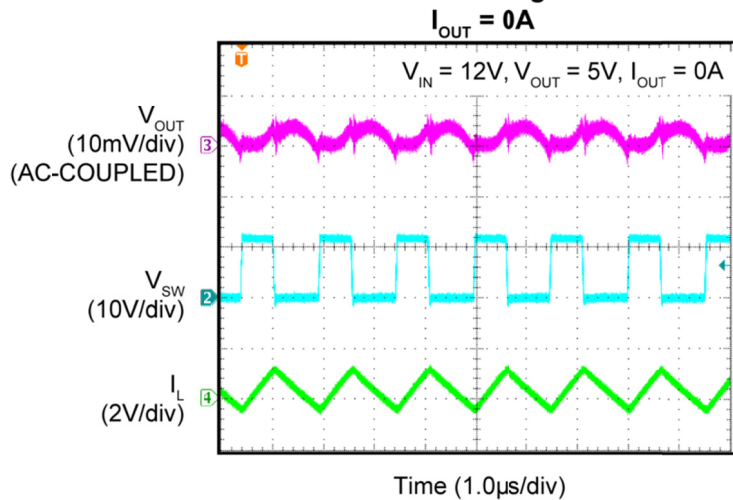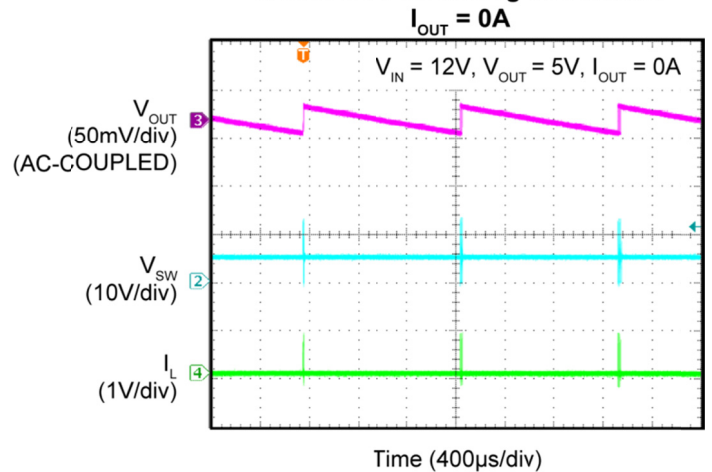
Functional Characteristics (Continued)

MIC28511-1 Switching Waveforms
I_{OUT} = 3A

V_{OUT}
(20mV/div)
(AC-COUPLED)

V_{SW}
(10V/div)

I_L
(2V/div)

V_{IN} = 12V
V_{OUT} = 5V
I_{OUT} = 3A

Time (1µs/div)
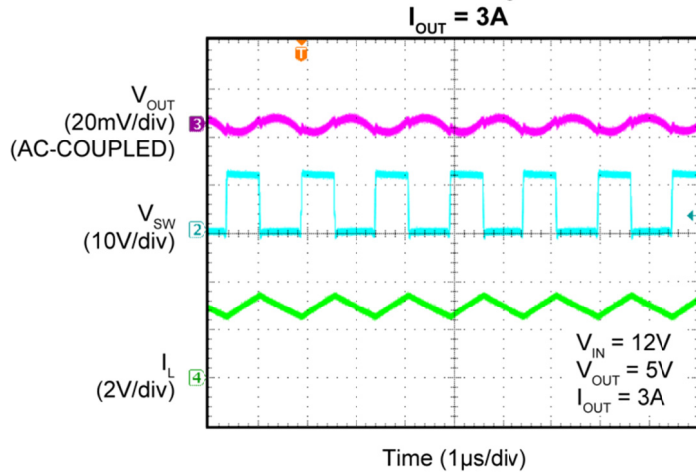
MIC28511-2 Transient Response

V_{OUT}
(500mV/div)
(AC-COUPLED)

V_{IN} = 12V
V_{OUT} = 5V
I_{OUT} = 0A TO 3A

I_{OUT}
(2A/div)

Time (100µs/div)
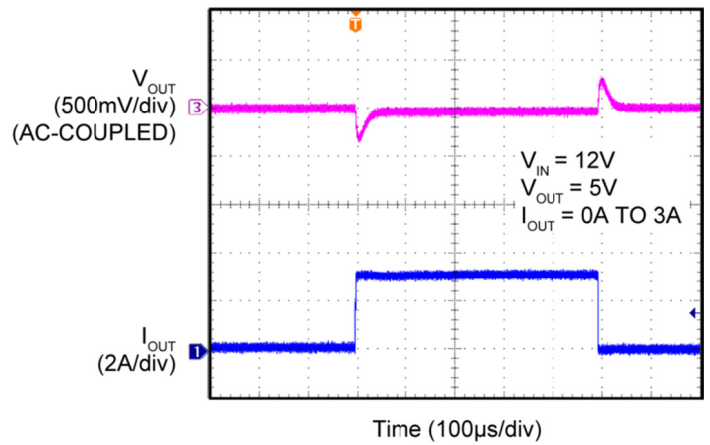
MIC28511-1 Transient Response

V_{OUT}
(500mV/div)
(AC-COUPLED)

V_{IN} = 12V
V_{OUT} = 5V
I_{OUT} = 0A TO 3A

I_{OUT}
(2A/div)

Time (100µs/div)

Power Good at V_{IN} Soft Turn-On

V_{IN}
(10V/div)

V_{OUT}
(5V/div)

V_{IN} = 12V
V_{OUT} = 5V
I_{OUT} = 0A

V_{PG}
(5V/div)

Time (400ms/div)

Power Good at V_{IN} Soft Turn-Off

V_{IN}
(10V/div)

V_{OUT}
(5V/div)

V_{IN} = 12V
V_{OUT} = 5V
I_{OUT} = 0A

V_{PG}
(5V/div)

Time (20.0ms/div)
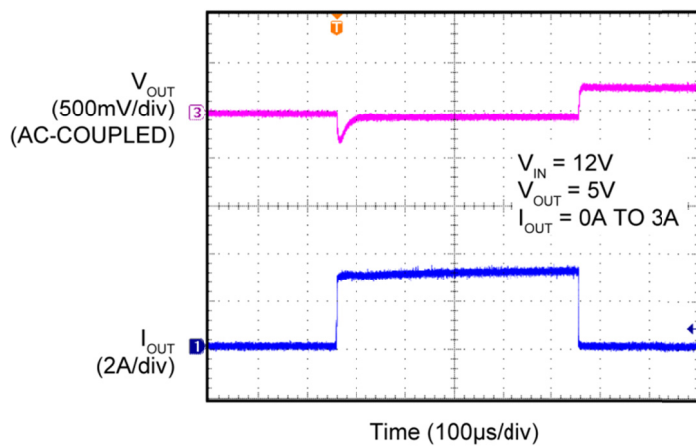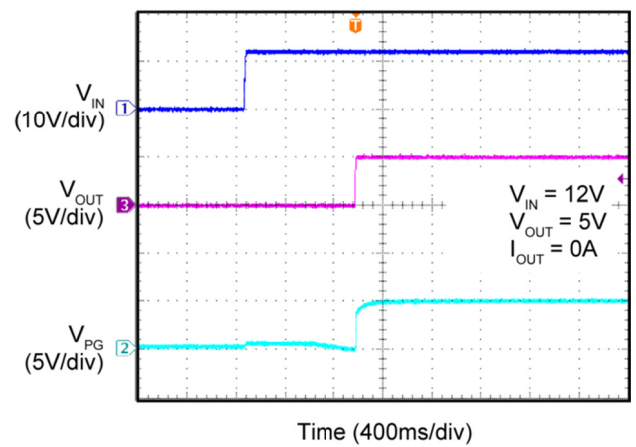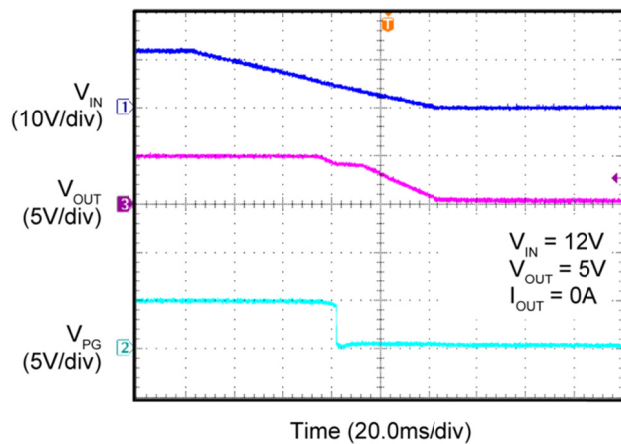
Functional Diagram
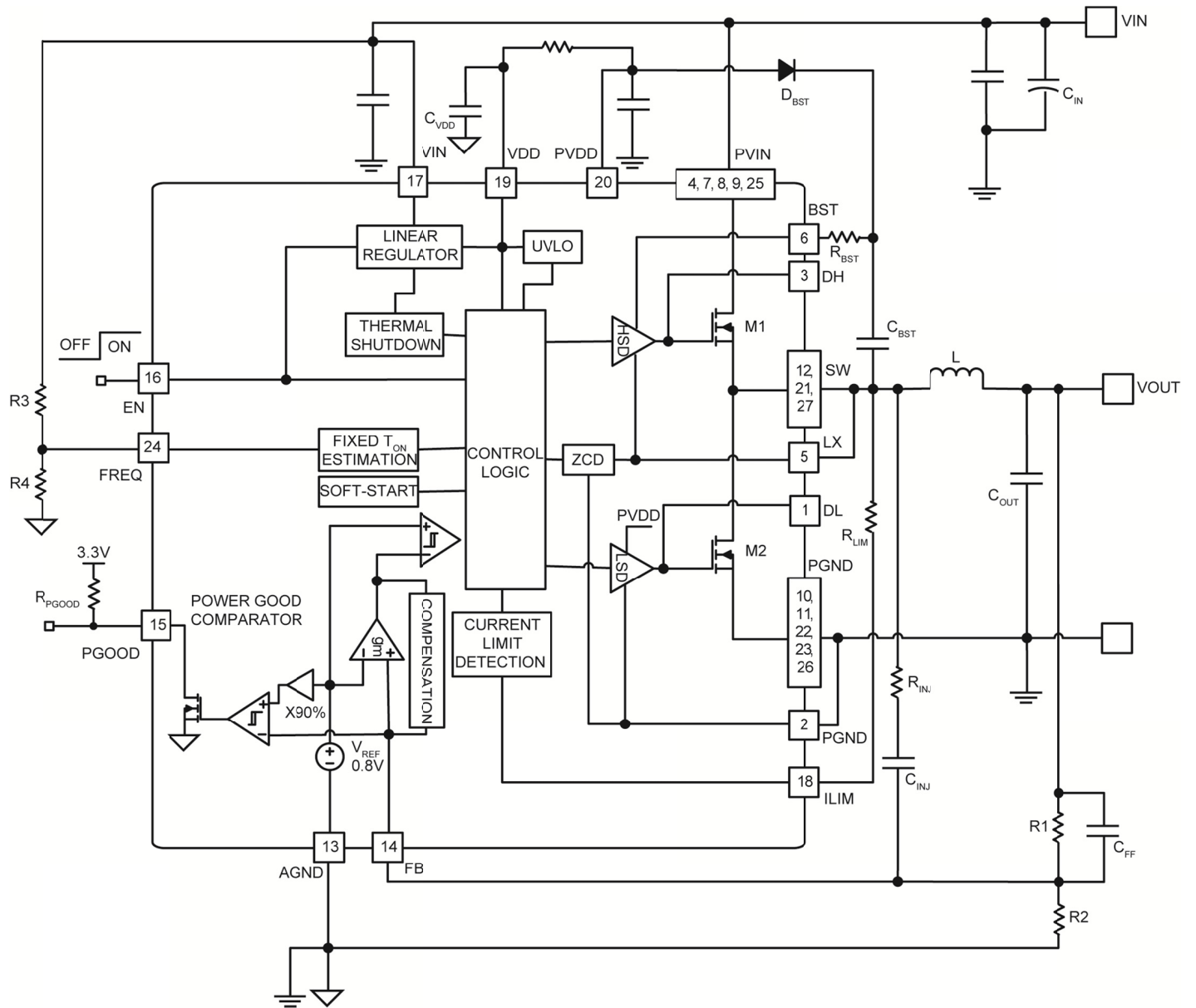
Functional Description

The MIC28511 is an adaptive on-time synchronous buck regulator with integrated high-side and low-side MOSFETs suitable for high-input voltage to low-output voltage conversion applications. It is designed to operate over a wide input voltage range (4.6V to 60V) which is suitable for automotive and industrial application. The output is adjustable with an external resistive divider. An adaptive on-time control scheme is employed to produce a constant switching frequency in continuous-conduction mode and reduced switching frequency in discontinuous-operation mode, improving light-load efficiency. Overcurrent protection is implemented by sensing low-side MOSFET's $R_{DS(ON)}$. The device features internal soft-start, enable, UVLO, and thermal shutdown.

Theory of Operation

As illustrated in the *Functional Diagram*, the output voltage of the MIC28511 is sensed by the feedback (FB) pin via voltage dividers R1 and R2, and compared to a 0.8V reference voltage V_{REF} at the error comparator through a low-gain transconductance (g_M) amplifier. If the feedback voltage decreases and the amplifier output is below 0.8V, then the error comparator will trigger the control logic and generate an ON-time period. The ON-time period length is predetermined by the fixed t_{ON} estimator circuitry:

$$t_{ON(ESTIMATED)} = \frac{V_{OUT}}{V_{IN} \times f_{SW}} \quad \text{Eq. 1}$$

where V_{OUT} is the output voltage, V_{IN} is the power stage input voltage, and f_{SW} is the switching frequency.

At the end of the ON-time period, the internal high-side driver turns off the high-side MOSFET and the low-side driver turns on the low-side MOSFET. The OFF-time period length depends upon the feedback voltage in most cases. When the feedback voltage decreases and the output of the g_M amplifier is below 0.8V, then the ON-time period is triggered and the OFF-time period ends. If the OFF-time period determined by the feedback voltage is less than the minimum OFF-time $t_{OFF(MIN)}$, which is about 200ns (typical), the MIC28511 control logic will apply the $t_{OFF(MIN)}$ instead. The $t_{OFF(MIN)}$ is required to maintain enough energy in the boost capacitor (C_{BST}) to drive the high-side MOSFET.

The maximum duty cycle is obtained from:

$$D_{MAX} = 1 - t_{OFF(MIN)} \times f_{SW} \quad \text{Eq. 2}$$

It is not recommended to use MIC28511 with an OFF-time close to $t_{OFF(MIN)}$ during steady-state operation.

The adaptive ON-time control scheme results in a constant switching frequency in the MIC28511. The actual ON-time and resulting switching frequency will vary with the different rising and falling times of the external MOSFETs. Also, the minimum t_{ON} results in a lower switching frequency in high V_{IN} to V_{OUT} applications. During load transients, the switching frequency is changed due to the varying OFF-time.

Figure 1 shows the allowable range of the output voltage versus the input voltage. The minimum output voltage is 0.8V which is limited by the reference voltage. The maximum output voltage is 24V which is limited by the internal circuitry.
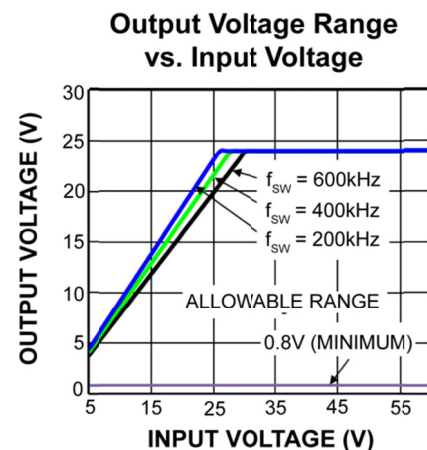
Figure 1. Allowable Output Voltage Range vs. Input Voltage

To illustrate the control loop operation, both the steady-state and load transient scenarios will be analyzed.

Figure 2 shows the MIC28511 control loop timing during steady-state operation. During steady-state, the g_M amplifier senses the feedback voltage ripple, which is proportional to the output voltage ripple and the inductor current ripple, to trigger the ON-time period. The ON-time is predetermined by the t_{ON} estimator. The termination of the OFF-time is controlled by the feedback voltage. At the valley of the feedback voltage ripple, which occurs when V_{FB} falls below V_{REF}, the OFF period ends and the next ON-time period is triggered through the control logic circuitry.

Figure 2. MIC28511 Control Loop Timing

Figure 3 shows the operation of the MIC28511 during a load transient. The output voltage drops due to the sudden load increase, which causes the V_{FB} to be less than V_{REF}. This will cause the error comparator to trigger an ON-time period. At the end of the ON-time period, a minimum OFF-time $t_{OFF(MIN)}$ is generated to charge C_{BST} since the feedback voltage is still below V_{REF}. Then, t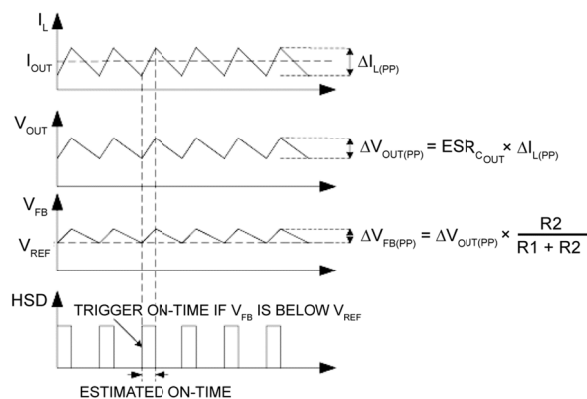he next ON-time period is triggered due to the low feedback voltage. Therefore, the switching frequency changes during the load transient, but returns to the nominal fixed frequency once the output has stabilized at the new load current level. With the varying duty cycle and switching frequency, the output recovery time is fast and the output voltage deviation is small in MIC28511 converter.
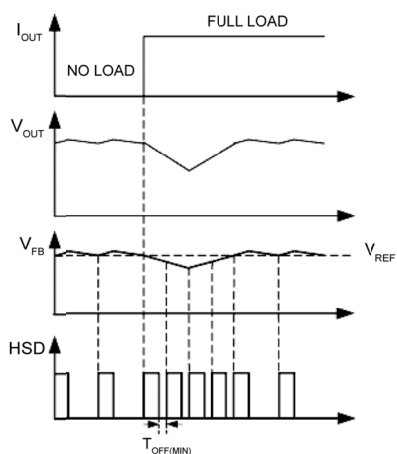

Figure 3. MIC28511 Load Transient Response

Unlike true current-mode control, the MIC28511 uses the output voltage ripple to trigger an ON-time period. The output voltage ripple is proportional to the inductor current ripple if the ESR of the output capacitor is large enough. The MIC28511 control loop has the advantage of eliminating the need for slope compensation.

In order to meet the stability requirements, the MIC28511 feedback voltage ripple should be in phase with the inductor current ripple and large enough to be sensed by the g_M amplifier and the error comparator. The recommended feedback voltage ripple is 20mV~100mV.

If a low-ESR output capacitor is selected, then the feedback voltage ripple may be too small to be sensed by the g_m amplifier and the error comparator. Also, if the ESR of the output capacitor is very low, the output voltage ripple and the feedback voltage ripple are not necessarily in phase with the inductor current ripple. In these cases, ripple injection is required to ensure proper operation. Please refer to "Ripple Injection" subsection in *Application Information* for more details about the ripple injection technique.

Discontinuous Mode (MIC28511-1 Only)

In continuous mode, the inductor current is always greater than zero; however, at light loads the MIC28511-1 is able to force the inductor current to operate in discontinuous mode. Discontinuous mode occurs when the inductor current falls to zero, as indicated by trace (I_L) shown in Figure 4. During this period, the efficiency is optimized by shutting down all the non-essential circuits and minimizing the supply current. The MIC28511-1 wakes up and turns on the high-side MOSFET when the feedback voltage V_{FB} drops below 0.8V.

The MIC28511-1 has a zero crossing comparator that monitors the inductor current by sensing the voltage drop across the low-side MOSFET during its ON-time. If the V_{FB} > 0.8V and the inductor current goes slightly negative, then the MIC28511-1 automatically powers down most of the IC circuitry and goes into a low-power mode.

Once the MIC28511-1 goes into discontinuous mode, both DH and DL are low, which turns off the high-side and low-side MOSFETs. The load current is supplied by the output capacitors and V_{OUT} drops. If the drop of V_{OUT} causes V_{FB} to go below V_{REF}, then all the circuits will wake up into normal continuous mode. First, the bias currents of most circuits reduced during the discontinuous mode are restored, and then a t_{ON} pulse is triggered before the drivers are turned on to avoid any possible glitches. Finally, the high-side driver is turned on. Figure 4 shows the control loop timing in discontinuous mode.
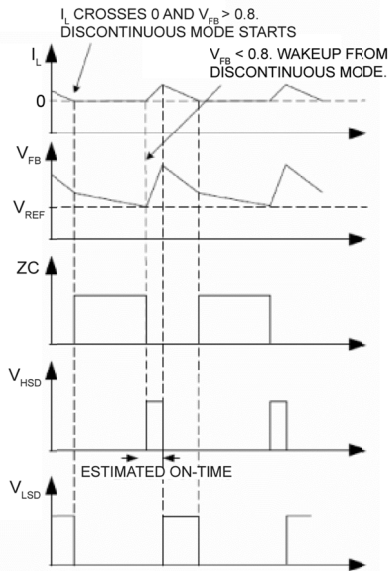
Figure 4. MIC28511-1 Control Loop Timing (Discontinuous Mode)

During discontinuous mode, the bias current of most circuits are reduced. As a result, the total power supply current during discontinuous mode is only about 450μA, allowing the MIC28511-1 to achieve high efficiency in light load applications.

V_{DD} Regulator

The MIC28511 provides a 5V regulated V_{DD} to bias internal circuitry for V_{IN} ranging from 5.5V to 60V. When V_{IN} is less than 5.5V, V_{DD} should be tied to V_{IN} pins to bypass the internal linear regulator.

Soft-Start

Soft-start reduces the power supply inrush current at startup by controlling the output voltage rise time while the output capacitor charges.

The MIC28511 implements an internal digital soft-start by ramping up the 0.8V reference voltage (V_{REF}) from 0 to 100% in about 5ms with 9.7mV steps. This controls the output voltage rate of rise at turn on, minimizing inrush current and eliminating output voltage overshoot. Once the soft-start cycle ends, the related circuitry is disabled to reduce current consumption.

Current Limit

The MIC28511 uses the $R_{DS(ON)}$ of the internal low-side power MOSFET to sense over-current conditions. In each switching cycle, the inductor current is sensed by monitoring the low-side MOSFET during its ON period. The sensed voltage, $V_{(ILIM)}$, is compared with the power ground (PGND) after a blanking time of 150ns.

The voltage drop of the resistor R_{ILIM} is compared with the low-side MOSFET voltage drop to set the over-current trip level. The small capacitor connected from ILIM pin to PGND can be added to filter the switching node ringing allowing a better short limit measurement. The time constant created by R_{LIM} and the filter capacitor should be much less than the minimum off time.

The overcurrent limit can be programmed by using Equation 3:

$$R_{ILIM} = \frac{\left(I_{CLIM} - 0.5 \times \Delta I_{L(PP)}\right) \times R_{DS(ON)} + V_{CL}}{I_{CL}} \qquad \text{Eq. 3}$$

Where:

I_{CLIM} = Desired current limit.

$R_{DS(ON)}$ = On-resistance of low-side power MOSFET 40mΩ (typical).

V_{CL} = Current-limit threshold 14mV (typical absolute value). See the *Electrical Characteristics*[5] table.

I_{CL} = Current-limit source current 70μA (typical). See the *Electrical Characteristics*[5] table.

$\Delta I_{L(PP)}$ = Inductor current peak-to-peak (use Equation 4 to calculate the inductor ripple current).

The peak-to-peak inductor current ripple is:

$$\Delta I_{L(PP)} = \frac{V_{OUT} \times \left(V_{IN(MAX)} - V_{OUT}\right)}{V_{IN(MAX)} \times f_{SW} \times L} \qquad \text{Eq. 4}$$

The MOSFET $R_{DS(ON)}$ varies 30% to 40% with temperature; therefore, it is recommended to use the $R_{DS(ON)}$ at max junction temperature with 20% margin to calculate R_{ILIM} in Equation 3.

In case of hard short, the current limit threshold is folded down to allow an indefinite hard short on the output without any destructive effect. It is mandatory to make sure that the inductor current used to charge the output capacitor during soft start is under the folded short limit; otherwise the supply will go in hiccup mode and may not be finishing the soft start successfully.

Power Good (PGOOD)

The power good (PGOOD) pin is an open drain output which indicates logic high when the output is nominally 90% of its steady state voltage.

MOSFET Gate Drive

The *Functional Diagram* shows a bootstrap circuit, consisting of D_{BST}, C_{BST} and R_{BST}. This circuit supplies energy to the high-side drive circuit. Capacitor C_{BST} is charged, while the low-side MOSFET is on, and the voltage on the SW pin is approximately 0V. When the high-side MOSFET driver is turned on, energy from C_{BST} is used to turn the MOSFET on. As the high-side MOSFET turns on, the voltage on the SW pin increases to approximately V_{IN}. Diode D_{BST} is reverse biased and C_{BST} floats high while continuing to bias the high-side gate driver. The bias current of the high-side driver is less than 10mA so a 0.1µF to 1µF is sufficient to hold the gate voltage with minimal droop for the power stroke (high-side switching) cycle, i.e. $\Delta BST = 10mA \times 1.25\mu s/0.1\mu F = 125mV$. When the low-side MOSFET is turned back on, C_{BST} is then recharged through the boost diode. A 30Ω resistor R_{BST}, which is in series with BST pin, is required to slow down the turn-on time of the high-side N-channel MOSFET.

Application Information

Output Voltage Setting Components

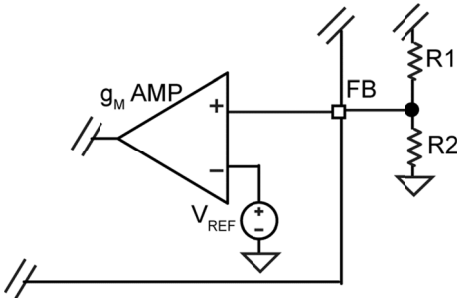The MIC28511 requires two resistors to set the output voltage as shown in Figure 5:

Figure 5. Voltage-Divider Configuration

The output voltage is determined by Equation 5:

$$V_{OUT} = V_{FB} \times \left(1 + \frac{R1}{R2}\right)$$ Eq. 5

Where: V_{FB} = 0.8V

A typical value of R1 used on the standard evaluation board is 10kΩ. If R1 is too large, it may allow noise to be introduced into the voltage feedback loop. If R1 is too small in value, it will decrease the efficiency of the power supply, especially at light loads. Once R1 is selected, R2 can be calculated using Equation 6:

$$R2 = \frac{V_{FB} \times R1}{V_{OUT} - V_{FB}}$$ Eq. 6

Setting the Switching Frequency

The MIC28511 switching frequency can be adjusted by changing the resistor divider network from VIN.
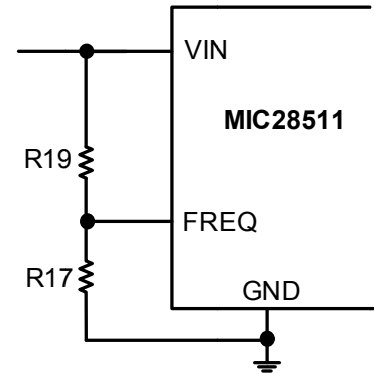

Figure 6. Switching Frequency Adjustment

Equation 7 gives the estimated switching frequency:

$$F_{SW} = F_0 \times \left(\frac{R17}{R17 + R19}\right)$$ Eq. 7

Where:

f_O = Switching frequency when R17 is open, 680kHz typically

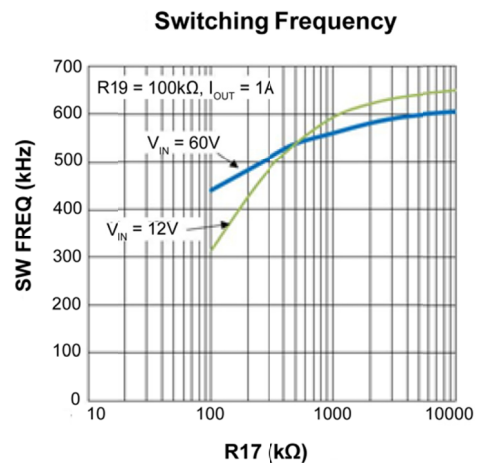Figure 7 shows the switching frequency versus the resistor R17 when R19 = 100kΩ:

Figure 7. Switching Frequency vs. R17

Inductor Selection

Values for inductance, peak, and RMS currents are required to select the output inductor. The input and output voltages and the inductance value determine the peak-to-peak inductor ripple current. Generally, higher inductance values are used with higher input voltages. Larger peak-to-peak ripple currents will increase the power dissipation in the inductor and MOSFETs. Larger output ripple currents will also require more output capacitance to smooth out the larger ripple current. Smaller peak-to-peak ripple currents require a larger inductance value and therefore a larger and more expensive inductor. A good compromise between size, loss and cost is to set the inductor ripple current to be equal to 20% of the maximum output current. The inductance value is calculated by:

$$L = \frac{V_{OUT} \times (V_{IN(MAX)} - V_{OUT})}{V_{IN(MAX)} \times \Delta I_{L(PP)} \times f_{SW}} \qquad \text{Eq. 8}$$

Where:

f_{SW} = Switching frequency.

$\Delta I_{L(PP)}$ = The peak-to-peak inductor current ripple, typically 20% of the maximum output current.

In the continuous conduction mode, the peak inductor current is equal to the average output current plus one half of the peak-to-peak inductor current ripple.

$$I_{L(PK)} = I_{OUT} + 0.5 \times \Delta I_{L(PP)} \qquad \text{Eq. 9}$$

The RMS inductor current is used to calculate the I^2R losses in the inductor.

$$I_{L(RMS)} = \sqrt{I^2_{OUT(MAX)} + \frac{\Delta I^2_{L(PP)}}{I^2}} \qquad \text{Eq. 10}$$

Maximizing efficiency requires the proper selection of core material and minimizing the winding resistance. The high frequency operation of the MIC28511 requires the use of ferrite materials for all but the most cost sensitive applications. Lower cost iron powder cores may be used but the increase in core loss will reduce the efficiency of the power supply. This is especially noticeable at low output power. The winding resistance decreases efficiency at the higher output current levels.

The winding resistance must be minimized although this usually comes at the expense of a larger inductor. The power dissipated in the inductor is equal to the sum of the core and copper losses. At higher output loads, the core losses are usually insignificant and can be ignored. At lower output currents, the core losses can be a significant contributor. Core loss information is usually available from the magnetics vendor. Copper loss in the inductor is calculated by Equation 11:

$$P_{L(Cu)} = I_{L(RMS)}{}^2 \times DCR \qquad \text{Eq. 11}$$

The resistance of the copper wire, DCR, increases with the temperature. The value of the winding resistance used should be at the operating temperature.

$$DCR_{(HT)} = DCR_{20°C} \times (1 + 0.0042 \times (T_H - T_{20°C})) \qquad \text{Eq. 12}$$

Where:

T_H = Temperature of wire under full load.

$T_{20°C}$ = Ambient temperature.

$DCR_{(20°C)}$ = Room temperature winding resistance (usually specified by the manufacturer).

Output Capacitor Selection

The type of the output capacitor is usually determined by its equivalent series resistance (ESR). Voltage and RMS current capability are also important factors in selecting an output capacitor. Recommended capacitor types are ceramic, tantalum, low-ESR aluminum electrolytic, OS-CON and POSCAP. For high ESR electrolytic capacitors, ESR is the main cause of the output ripple. The output capacitor ESR also affects the control loop from a stability point of view. For a low ESR ceramic output capacitor, ripple is dominated by the reactive impedance.

The maximum value of ESR is calculated:

$$ESR_{COUT} \leq \frac{\Delta V_{OUT(PP)}}{\Delta I_{L(PP)}} \qquad \text{Eq. 13}$$

Where:

$\Delta V_{OUT(pp)}$ = peak-to-peak output voltage ripple

$\Delta I_{L(PP)}$ = peak-to-peak inductor current ripple

The total output ripple is a combination of the ESR and output capacitance. The total ripple is calculated by Equation 14:

$$\Delta V_{OUT(PP)} = \sqrt{\left(\frac{2 \times \Delta I_{L(PP)}}{C_{OUT} \times f_{SW} \times 8}\right) + \left(\Delta I_{L(PP)} \times ESR_{COUT}\right)^2}$$

<div align="right">Eq. 14</div>

Where:

D = Duty cycle.

C_{OUT} = Output capacitance value.

f_{SW} = Switching frequency.

As described in the "Theory of Operation" section in the *Functional Characteristics* section, the MIC28511 requires at least 20mV peak-to-peak ripple at the FB pin for the g_m amplifier and the error comparator to operate properly. Also, the ripple on FB pin should be in phase with the inductor current. Therefore, the output voltage ripple caused by the output capacitors value should be much smaller than the ripple caused by the output capacitor ESR. If low-ESR capacitors, such as ceramic capacitors, are selected as the output capacitors, a ripple injection method should be applied to provide the enough feedback voltage ripple. Please refer to the "Ripple Injection" section for more details.

The voltage rating of the capacitor should be twice the output voltage for a tantalum and 20% greater for aluminum electrolytic or OS-CON. The output capacitor RMS current is calculated by Equation 15:

$$I_{COUT(RMS)} = \frac{\Delta I_{L(PP)}}{\sqrt{12}}$$

<div align="right">Eq. 15</div>

The power dissipated in the output capacitor is:

$$P_{DISS(COUT)} = I^2_{COUT(RMS)} \times ESR_{COUT}$$

<div align="right">Eq. 16</div>

Input Capacitor Selection

The input capacitor for the power stage input V_{IN} should be selected for ripple current rating and voltage rating. Tantalum input capacitors may fail when subjected to high inrush currents, caused by turning the input supply on. A tantalum input capacitor's voltage rating should be at least two times the maximum input voltage to maximize reliability. Aluminum electrolytic, OS-CON, and multilayer polymer film capacitors can handle the higher inrush currents without voltage de-rating. The input voltage ripple will primarily depend on the input capacitor's ESR. The peak input current is equal to the peak inductor current, so:

$$\Delta V_{IN} = I_{L(PK)} \times ESR_{CIN}$$

<div align="right">Eq. 17</div>

The input capacitor must be rated for the input current ripple. The RMS value of input capacitor current is determined at the maximum output current. Assuming the peak-to-peak inductor current ripple is low:

$$I_{CIN(RMS)} \approx I_{OUT(MAX)} \times \sqrt{D \times (1-D)}$$

<div align="right">Eq. 18</div>

The power dissipated in the input capacitor is:

$$P_{DISS(CIN)} = I^2_{CIN(RMS)} \times ESR_{CIN}$$

<div align="right">Eq. 19</div>

Ripple Injection

The V_{FB} ripple required for proper operation of the MIC28511's g_m amplifier and error comparator is 20mV to 100mV. However, the output voltage ripple is generally designed as 1% to 2% of the output voltage. If the feedback voltage ripple is so small that the g_m amplifier and error comparator can't sense it, then the MIC28511 will lose control and the output voltage is not regulated. In order to have some amount of V_{FB} ripple, a ripple injection method is applied for low output voltage ripple applications.

The applications are divided into three situations according to the amount of the feedback voltage ripple:

1. Enough ripple at the feedback voltage due to the large ESR of the output capacitors.

   As shown in Figure 8, the converter is stable without any ripple injection. The feedback voltage ripple is:

$$\Delta V_{FB(PP)} = \frac{R2}{R1+R2} \times ESR_{COUT} \times \Delta I_{L(PP)} \quad \text{Eq. 20}$$

where $\Delta I_{L(pp)}$ is the peak-to-peak value of the inductor current ripple.

2. Inadequate ripple at the feedback voltage due to the small ESR of the output capacitors.

   The output voltage ripple is fed into the FB pin through a feed forward capacitor C_{FF} in this situation, as shown in Figure 9. The typical C_{FF} value is selected by:

$$R1 \times C_{FF} \approx \frac{10}{f_{SW}} \quad \text{Eq. 21}$$

With the feed forward capacitor, the feedback voltage ripple is very close to the output voltage ripple:

$$\Delta V_{FB(PP)} \approx ESR_{COUT} \times \Delta I_{L(PP)} \quad \text{Eq. 22}$$

3. Virtually no ripple at the FB pin voltage due to the very-low ESR of the output capacitors.
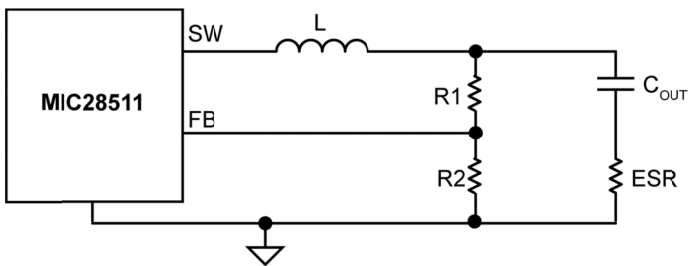
Figure 8. Enough Ripple
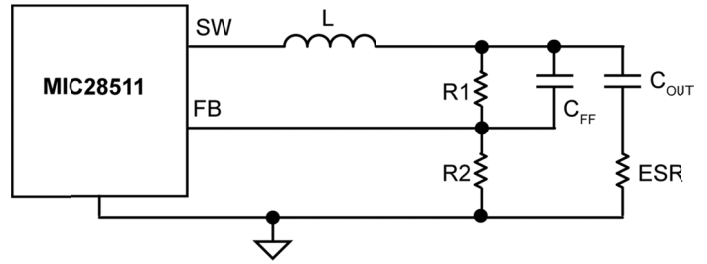
Figure 9. Inadequate Ripple
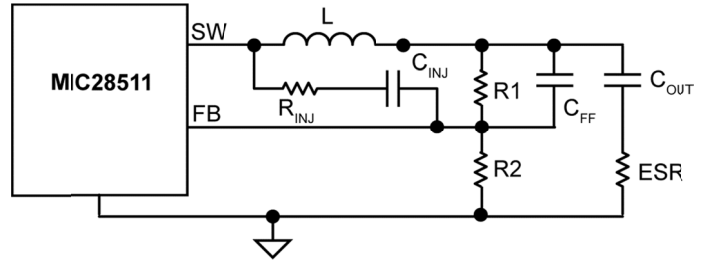
Figure 10. Invisible Ripple

In this situation, the output voltage ripple is less than 20mV. Therefore, additional ripple is injected into the FB pin from the switching node SW via a resistor R_{INJ} and a capacitor C_{INJ}, as shown in Figure 10. The injected ripple is:

$$\Delta V_{FB(pp)} = V_{IN} \times K_{div} \times D \times (1-D) \times \frac{1}{f_{SW} \times \tau} \quad \text{Eq. 23}$$

$$K_{DIV} = \frac{R1//R2}{R_{INJ} + R1//R2} \quad \text{Eq. 24}$$

Where:

V_{IN} = Power stage input voltage

D = Duty cycle

f_{SW} = Switching frequency

τ = (R1//R2//R_{INJ}) × C_{FF}

In Equations 23 and 25, it is assumed that the time constant associated with C_{FF} must be much greater than the switching period:

$$\frac{1}{f_{SW} \times \tau} = \frac{T}{\tau} << 1 \qquad \text{Eq. 25}$$

If the voltage divider resistors R1 and R2 are in the kΩ range, a C_{FF} of 1nF to 100nF can easily satisfy the large time constant requirements. Also, a 100nF injection capacitor C_{INJ} is used in order to be considered as short for a wide range of the frequencies.

The process of sizing the ripple injection resistor and capacitors is:

Step 1. Select C_{FF} to feed all output ripples into the feedback pin and make sure the large time constant assumption is satisfied. Typical choice of C_{FF} is 1nF to 100nF if R1 and R2 are in kΩ range.

Step 2. Select R_{INJ} according to the expected feedback voltage ripple using Equation 26:

$$K_{DIV} = \frac{\Delta V_{FB(pp)}}{V_{IN}} \times \frac{f_{SW} \times \tau}{D \times (1-D)} \qquad \text{Eq. 26}$$

Then the value of R_{INJ} is obtained as:

$$R_{INJ} = (R1//R2) \times (\frac{1}{K_{DIV}} - 1) \qquad \text{Eq. 27}$$

Step 3. Select C_{INJ} as 100nF, which could be considered as short for a wide range of the frequencies.

PCB Layout Guidelines

Warning: To minimize EMI and output noise, follow these layout recommendations.

PCB Layout is critical to achieve reliable, stable and efficient performance. A ground plane is required to control EMI and minimize the inductance in power, signal and return paths.

Figure 11 is optimized from small form factor point of view shows top and bottom layer of a four-layer PCB. It is recommended to use Mid-Layer 1 as a continuous ground plane.
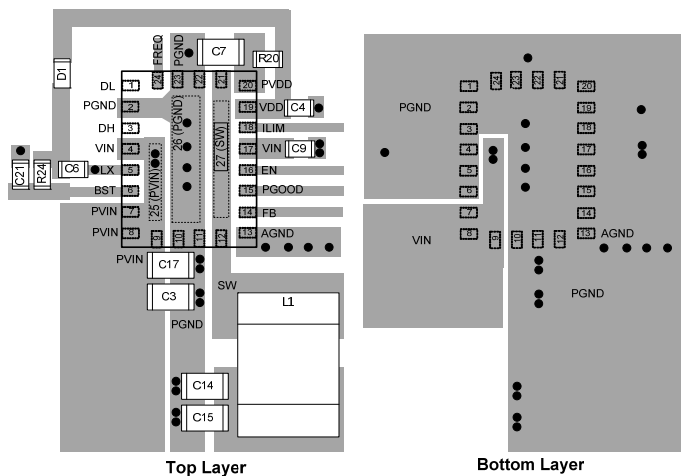
Figure 11. Top and Bottom Layer of a Four-Layer Board

The following guidelines should be followed to insure proper operation of the MIC28511 converter:

IC

- The analog ground pin (AGND) must be connected directly to the ground planes. Do not route the AGND pin to the PGND pin on the top layer.

- Place the IC close to the point of load (POL).

- Use copper planes to route the input and output power lines.

- Analog and power grounds should be kept separate and connected at only one location.

Input Capacitor

- Place the input capacitors on the same side of the board and as close to the PVIN and PGND pins as possible.

- Place several vias to the ground plane close to the input capacitor ground terminal.

- Use either X7R or X5R dielectric input capacitors. Do not use Y5V or Z5U type capacitors.

- Do not replace the ceramic input capacitor with any other type of capacitor. Any type of capacitor can be placed in parallel with the input capacitor.

- If a Tantalum input capacitor is placed in parallel with the input capacitor, it must be recommended for switching regulator applications and the operating voltage must be derated by 50%.

- In "Hot-Plug" applications, a Tantalum or Electrolytic bypass capacitor must be used to limit the over-voltage spike seen on the input supply with power is suddenly applied.

SW Node

- Do not route any digital lines underneath or close to the SW node.

- Keep the switch node (SW) away from the feedback (FB) pin.

Output Capacitor

- Use a copper island to connect the output capacitor ground terminal to the input capacitor ground terminal.

- Phase margin will change as the output capacitor value and ESR changes. Contact the factory if the output capacitor is different from what is shown in the BOM.

- The feedback trace should be separate from the power trace and connected as close as possible to the output capacitor. Sensing a long high-current load trace can degrade the DC load regulation.

Thermal Measurements

Measuring the IC's case temperature is recommended to insure it is within its operating limits. Although this might seem like a very elementary task, it is easy to get erroneous results. The most common mistake is to use the standard thermal couple that comes with a thermal meter. This thermal couple wire gauge is large, typically 22 gauge, and behaves like a heatsink, resulting in a lower case measurement.

Two methods of temperature measurement are using a smaller thermal couple wire or an infrared thermometer. If a thermal couple wire is used, it must be constructed of 36 gauge wire or higher then (smaller wire size) to minimize the wire heat-sinking effect. In addition, the thermal couple tip must be covered in either thermal grease or thermal glue to make sure that the thermal couple junction is making good contact with the case of the IC. Omega brand thermal couple (5SC-TT-K-36-36) is adequate for most applications.

Wherever possible, an infrared thermometer is recommended. The measurement spot size of most infrared thermometers is too large for an accurate reading on a small form factor ICs. However, a IR thermometer from Optris has a 1mm spot size, which makes it a good choice for measuring the hottest point on the case. An optional stand makes it easy to hold the beam on the IC for long periods of time.

MIC2851X Evaluation Board Schematic
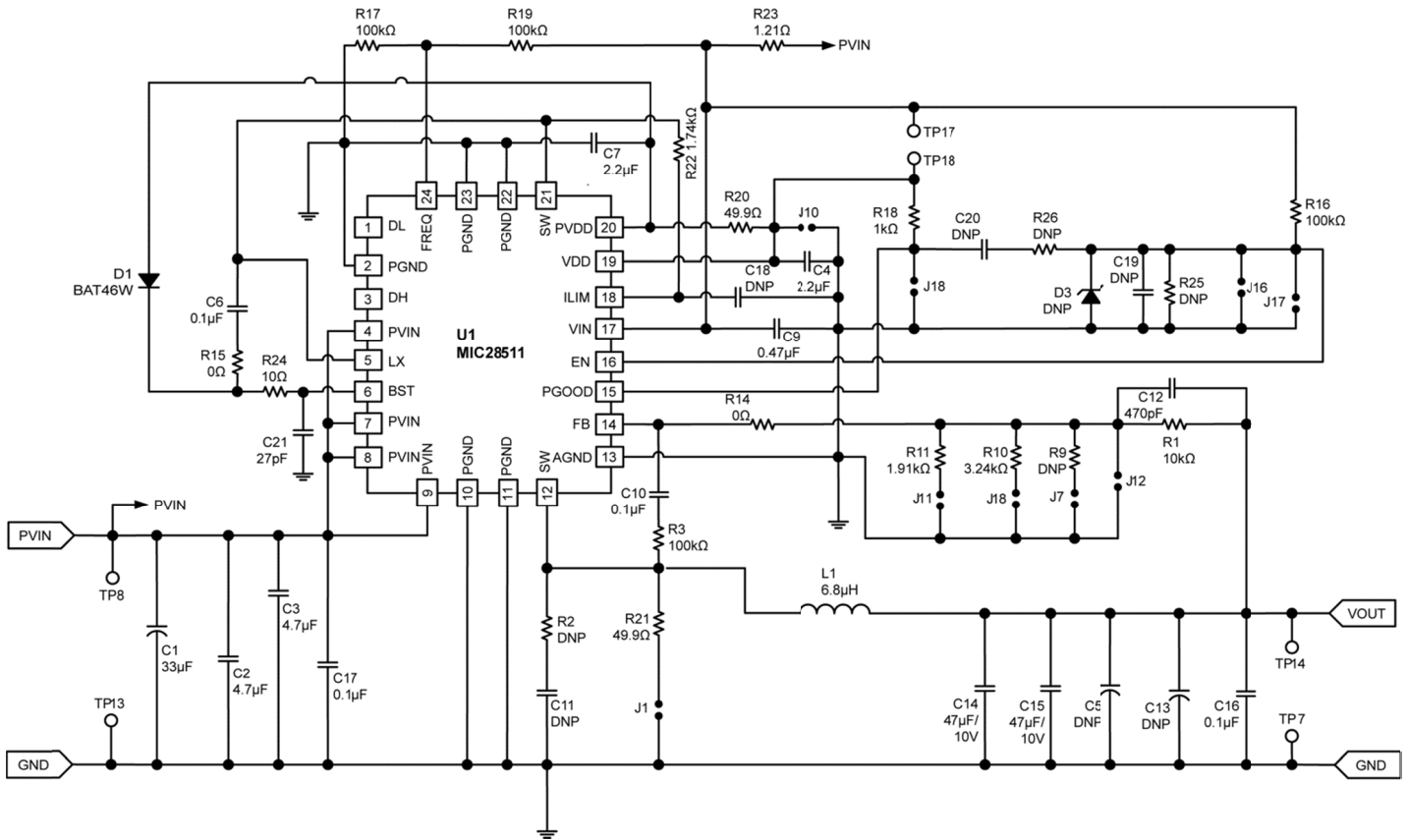
Bill of Materials

Item	Part Number	Manufacturer	Description	Qty.
C1	UVZ2A330MPD	Nichicon[6]	33µF/100V 20% Radial Aluminum Capacitor	1
C2, C3	12061Z475KAT2A	AVX[7]	4.7µF/100V, X7S, Size 1206 Ceramic Capacitor	2
C4, C7	C1608X7R1A225K080AC	TDK[8]	2.2µF/10V, X7R, Size 0603 Ceramic Capacitor	2
C5, C13			OPEN	NA
C6, C16	C0603C104K8RACTU	Kemet[9]	0.1µF/10V, X7R, Size 0603 Ceramic Capacitor	2
C9	GRM21BR72A474KA73	Murata[9]	0.47µF/100V, X7R, Size 0805 Ceramic Capacitor	1
	08051C474KAT2A	AVX		
C10, C17	GRM188R72A104KA35D	Murata	0.1µF/100V, X7R, Size 0603 Ceramic Capacitor	2
C11			OPEN	NA
C12	CGA3E2X7R1H471K	TDK	470pF/50V, X7R, Size 0603 Ceramic Capacitor	1

Notes:

6. Nichicon: www.nichicon.co.jp/english.
7. AVX: www.avx.com.
8. TDK: www.tdk.com.
9. Kemet: www.kemet.com.
10. Murata: www.murata.com.

Bill of Materials (Continued)

Item	Part Number	Manufacturer	Description	Qty.
C14, C15	GRM32ER71A476KE15L	Murata	47µF/10V, X7R, Size 1210 Ceramic Capacitor	2
C18			Open	NA
C19			Open	NA
C20			Open	NA
C21	C1608NP02A270J080AA	TDK	27pF 100V, NPO, Size 0603 Ceramic Capacitor	1
D1	BAT46W-TP	MCC[11]	100V Small Signal Schottky Diode, SOD123	1
D3			Open	NA
J1, J7, J8, J10, J11, J12, J16, J17, J18	77311-118-02LF	FCI[12]	CONN HEADER 2POS VERT T/H	9
L1	XAL7030-682MED	Coilcraft[13]	6.8µH, 10.7A sat current	1
R1	CRCW060310K0FKEA	Vishay Dale[14]	10.0kΩ, Size 0603, 1% Resistor	1
R2			Open	NA
R9			Open	NA
R10	CRCW06033K24FKEA	Vishay Dale	3.24kΩ, Size 0603, 1% Resistor	1
R11	CRCW06031K91FKEA	Vishay Dale	1.91kΩ, Size 0603, 1% Resistor	1
R14, R15	CRCW06030000FKEA	Vishay Dale	0.0 Ω, Size 0603, Resistor Jumper	2
R26			Open	NA
R16, R17, R19, R3	CRCW0603100K0FKEA	Vishay Dale	100kΩ, Size 0603, 1% Resistor	4
R25			Open	NA
R18	CRCW06031K00JNEA	Vishay Dale	1.0kΩ, Size 0603, 5% Resistor	1
R20, R21	CRCW060349R9FKEA	Vishay Dale	49.9Ω, Size 0603, 1% Resistor	2
R22	CRCW06031K74FKEA	Vishay Dale	1.74kΩ, Size 0603, 1% Resistor	1
R23	CRCW08051R21FKEA	Vishay Dale	1.21Ω, Size 0805, 1% Resistor	1
R24	CRCW060310R0FKEA	Vishay Dale	10.0Ω, Size 0603, 1% Resistor	1
TP1 – TP2			Open	
TP7 – TP14	77311-118-02LF	FCI	CONN HEADER 2POS VERT T/H	1
TP8 – TP13	77311-118-02LF	FCI	CONN HEADER 2POS VERT T/H	1
TP17 – TP18	77311-118-02LF	FCI	CONN HEADER 2POS VERT T/H	1
TP9, TP10, TP11, TP12	1502	Keystone Electronics[15]	Testpoint Turret, .090	4
U1	**MIC28511-1YFL**	**Micrel. Inc.**[16]	**60V$_{IN}$, 3A Synchronous Buck Regulator**	**1**
	MIC28511-2YFL			

Notes:

11. MCC: www.mccsemi.com.

12. FCI: www.fciconnect.com.

13. Coilcraft: www.coilcraft.com.

14. Vishay Dale: www.vishay.com.

15. Keystone Electronics: www.keystone.com.

16. Micrel Inc.: www.micrel.com.

MIC2851X Evaluation Board Layout
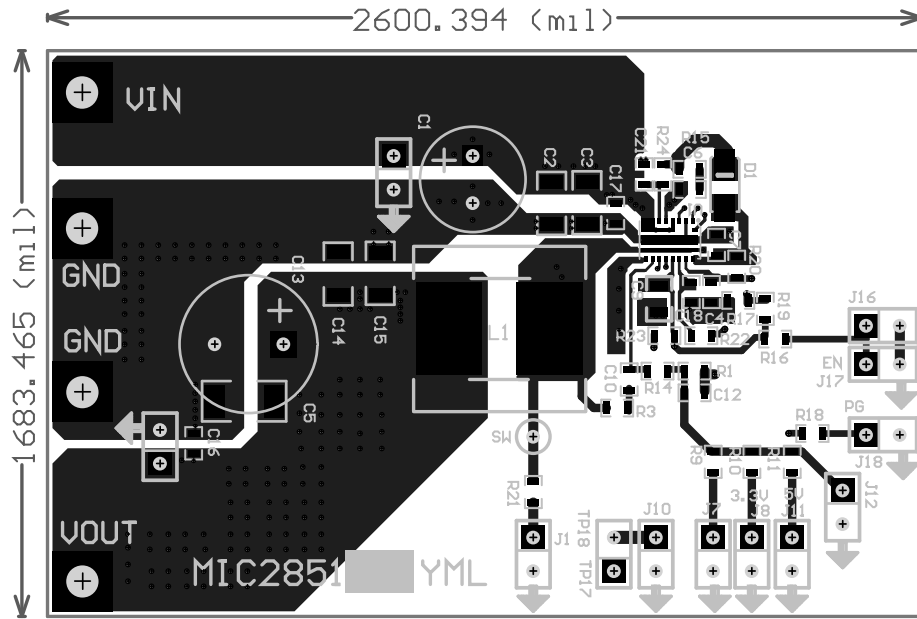
Top Layer
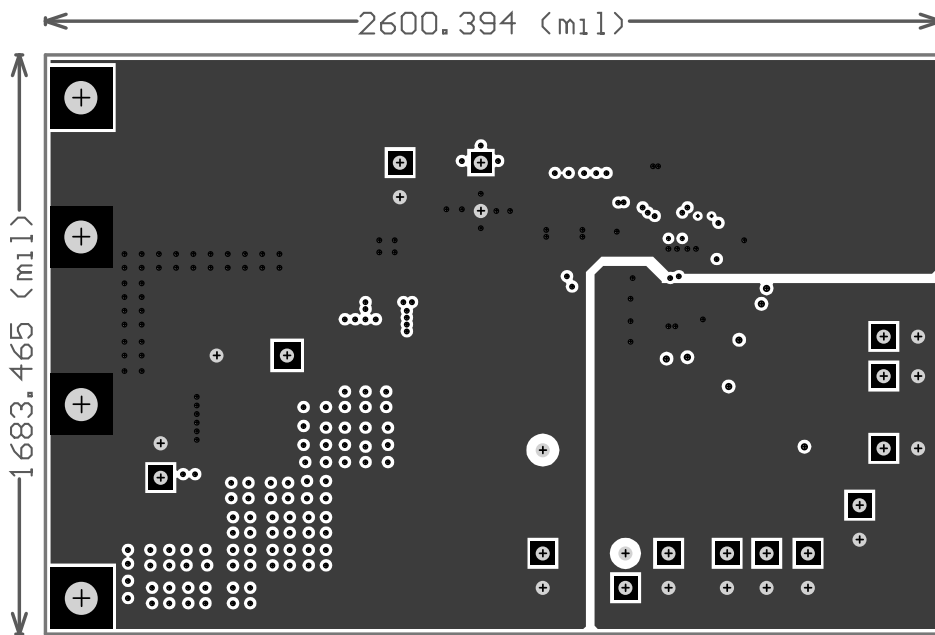
Mid Layer 1

MIC2851X Evaluation Board Layout (Continued)
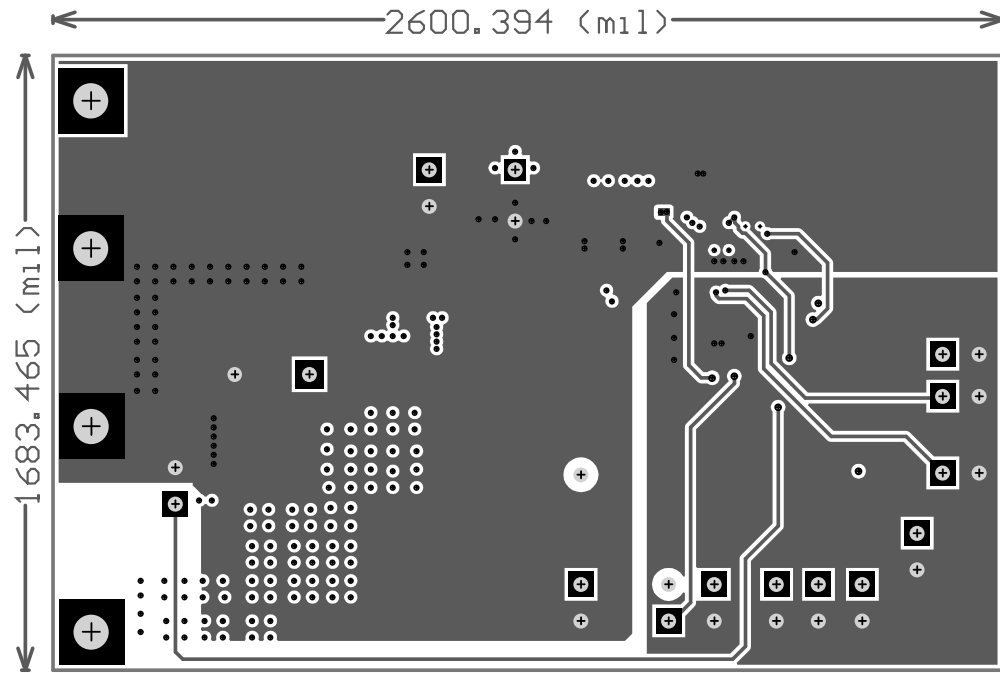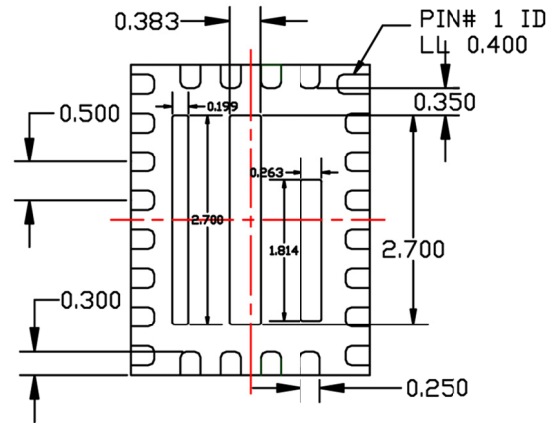

Mid Layer 2
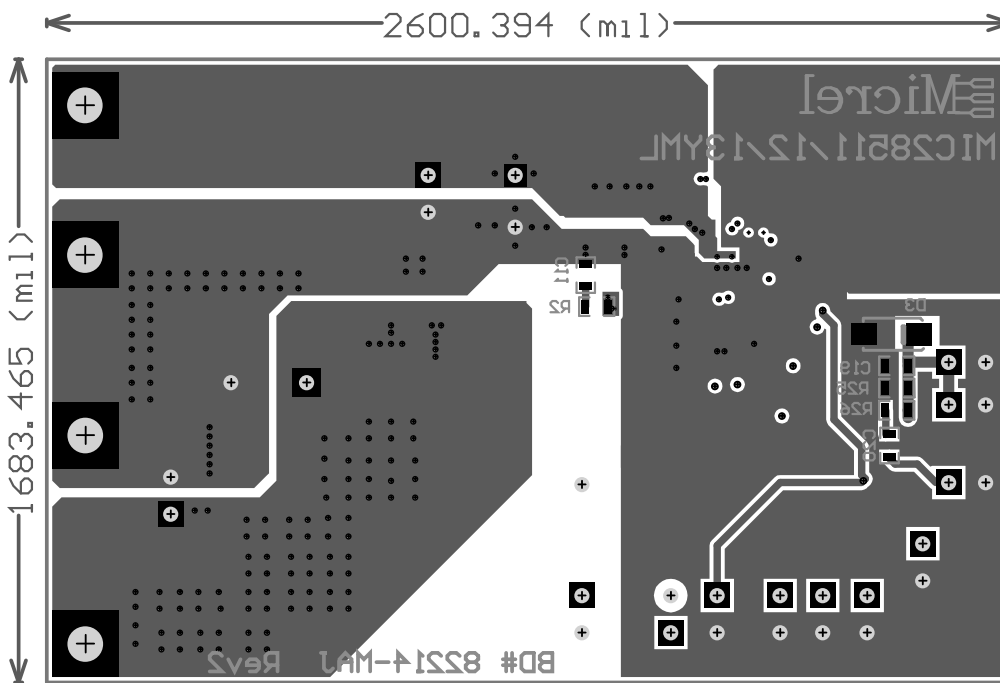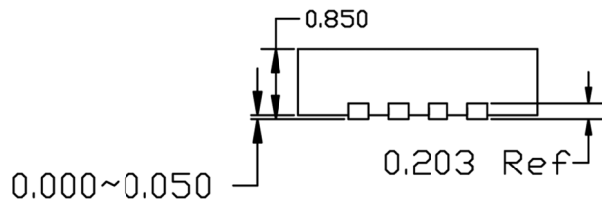

Bottom Layer

Package Information and Recommended Land Pattern[17]
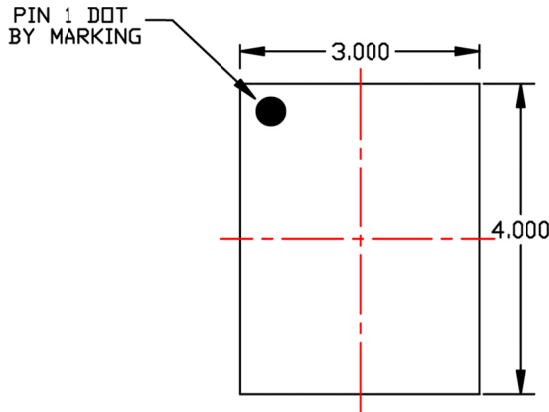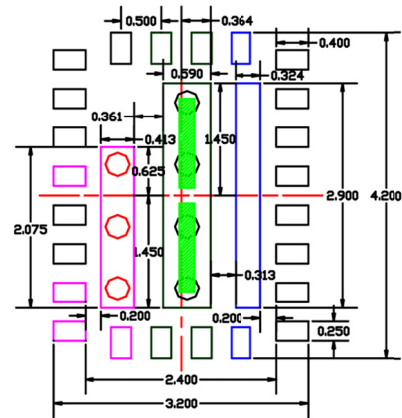

TOP VIEW
NOTE : 1,2,3

BOTTOM VIEW
NOTE : 1,2,3

SIDE VIEW
NOTE : 1,2,3

RECOMMENDED
LAND PATTERN
NOTE : 4,5,6

NOTE:
1. MAX PACKAGE WARPAGE IS 0.05 MM
2. MAX ALLOWABLE BURR IS 0.076MM   IN ALL DIRECTIONS
3. PIN #1 IS ON TOP WILL BE LASER MARKED
4. BLACK CIRCLES IN LAND PATTERN REPRESENT THERMAL VIA & SHOULD BE CONNECTED
TO GND FOR MAXIMUM PERFORMANCE, SIZE IS 0.30~0.35mm, 0.80mm PITCH.
5. GREEN RECTANGLES (SHADED AREA) REPRESENT SOLDER STENCIL OPENING ON EXPOSED
PAD AREA, SIZE IS 0.190x1.15mm, PITCH IS 1.350mm.
6. MAGENTA & BLUE COLORED PADS REPRESENT DIFFERENT POTENTIAL, DO NOT CONNECT
TO GND.

24-Pin 3mm × 4mm FQFN Package Type (FL)

MICREL, INC. 2180 FORTUNE DRIVE SAN JOSE, CA 95131 USA
TEL +1 (408) 944-0800 FAX +1 (408) 474-1000 WEB http://www.micrel.com

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