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Synchronous Current-Mode with Constant On-Time, PWM Buck Controller

Data Sheet

ADP1872/ADP1873

FEATURES

- Power input voltage as low as 2.75 V to 20 V
- Bias supply voltage range: 2.75 V to 5.5 V
- Minimum output voltage: 0.6 V
- 0.6 V reference voltage with ±1.0% accuracy
- Supports all N-channel MOSFET power stages
- Available in 300 kHz, 600 KHz, and 1.0 MHz options
- No current-sense resistor required
- Power saving mode (PSM) for light loads (ADP1873 only)
- Resistor-programmable current-sense gain
- Thermal overload protection
- Short-circuit protection
- Precision enable input
- Integrated bootstrap diode for high-side drive
- 140 µA shutdown supply current
- Starts into a precharged load
- Small, 10-lead MSOP package

APPLICATIONS

- Telecom and networking systems
- Mid to high end servers
- Set-top boxes
- DSP core power supplies

TYPICAL APPLICATIONS CIRCUIT

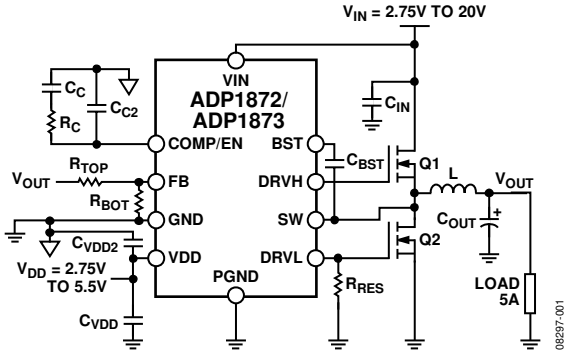

Figure 1.

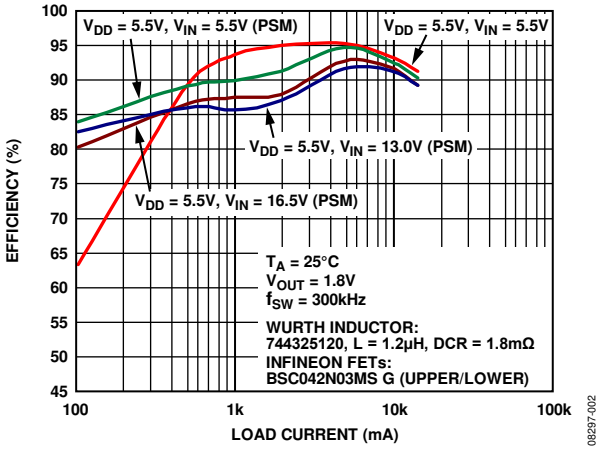

Figure 2. ADP1872 Efficiency vs. Load Current (V_{OUT} = 1.8 V, 300 kHz)

GENERAL DESCRIPTION

The ADP1872/ADP1873 are versatile current-mode, synchronous step-down controllers that provide superior transient response, optimal stability, and current limit protection by using a constant on-time, pseudo-fixed frequency with a programmable current-sense gain, current-control scheme. In addition, these devices offer optimum performance at low duty cycles by using valley current-mode control architecture. This allows the ADP1872/ADP1873 to drive all N-channel power stages to regulate output voltages as low as 0.6 V.

The ADP1873 is the power saving mode (PSM) version of the device and is capable of pulse skipping to maintain output regulation while achieving improved system efficiency at light loads (see the Power Saving Mode (PSM) Version (ADP1873) section for more information).

Available in three frequency options (300 kHz, 600 kHz, and 1.0 MHz, plus the PSM option), the ADP1872/ADP1873 are well suited for a wide range of applications. These ICs not only operate from a 2.75 V to 5.5 V bias supply, but can also accept a power input as high as 20 V.

In addition, an internally fixed, soft start period is included to limit input in-rush current from the input supply during startup and to provide reverse current protection during soft start for a pre-charged output. The low-side current-sense, current-gain scheme and integration of a boost diode, along with the PSM/forced pulse-width modulation (PWM) option, reduce the external part count and improve efficiency.

The ADP1872/ADP1873 operate over the −40°C to +125°C junction temperature range and are available in a 10-lead MSOP.

Rev. B

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3/10—Rev. 0 to Rev. A

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10/09—Revision 0: Initial Version

SPECIFICATIONS

All limits at temperature extremes are guaranteed via correlation using standard statistical quality control (SQC). VDD = 5 V, BST − SW = 5 V, VIN = 13 V. The specifications are valid for T_J = −40°C to +125°C, unless otherwise specified.

Table 1.

Parameter	Symbol	Conditions	Min	Typ	Max	Unit
POWER SUPPLY CHARACTERISTICS						
High Input Voltage Range	VIN	ADP1872ARMZ-0.3/ADP1873ARMZ-0.3 (300 kHz)	2.75	12	20	V
		ADP1872ARMZ-0.6/ADP1873ARMZ-0.6 (600 kHz)	2.75	12	20	V
		ADP1872ARMZ-1.0/ADP1873ARMZ-1.0 (1.0 MHz)	3.0	12	20	V
Low Input Voltage Range	VDD	C_{IN} = 1 µF to PGND, C_{IN} = 0.22 µF to GND				
		ADP1872ARMZ-0.3/ADP1873ARMZ-0.3 (300 kHz)	2.75	5	5.5	V
		ADP1872ARMZ-0.6/ADP1873ARMZ-0.6 (600 kHz)	2.75	5	5.5	V
		ADP1872ARMZ-1.0/ADP1873ARMZ-1.0 (1.0 MHz)	3.0	5	5.5	V
Quiescent Current	$I_{Q\_DD} + I_{Q\_BST}$	FB = 1.5 V, no switching		1.1		mA
Shutdown Current	$I_{DD, SD} + I_{BST, SD}$	COMP/EN < 285 mV		140	215	µA
Undervoltage Lockout	UVLO	Rising VDD (See Figure 34 for temperature variation)		2.65		V
UVLO Hysteresis		Falling VDD from operational state		190		mV
SOFT START						
Soft Start Period		See Figure 57		3.0		ms
ERROR AMPLIFER						
FB Regulation Voltage	V_{FB}	T_J = 25°C		600		mV
		T_J = −40°C to +85°C	595.5	600	605.4	mV
		T_J = −40°C to +125°C	594.2	600	606.5	mV
Transconductance	G_M		300	515	730	µs
FB Input Leakage Current	$I_{FB, LEAK}$	FB = 0.6 V, COMP/EN = released		1	50	nA
CURRENT-SENSE AMPLIFIER GAIN						
Programming Resistor (RES) Value from DRVL to PGND		RES = 47 kΩ ± 1%	2.7	3	3.3	V/V
		RES = 22 kΩ ± 1%	5.5	6	6.5	V/V
		RES = none	11	12	13	V/V
		RES = 100 kΩ ± 1%	22	24	26	V/V
SWITCHING FREQUENCY		Typical values measured at 50% time points with 0 nF at DRVH and DRVL; maximum values are guaranteed by bench evaluation[1]				
ADP1872ARMZ-0.3/ ADP1873ARMZ-0.3 (300 kHz)				300		kHz
On-Time		VIN = 5 V, V_{OUT} = 2 V, T_J = 25°C	1120	1200	1280	ns
Minimum On-Time		VIN = 20 V		145	190	ns
Minimum Off-Time		84% duty cycle (maximum)		320	385	ns
ADP1872ARMZ-0.6/ ADP1873ARMZ-0.6 (600 kHz)				600		kHz
On-Time		VIN = 5 V, V_{OUT} = 2 V, T_J = 25°C	500	520	580	ns
Minimum On-Time		VIN = 20 V, V_{OUT} = 0.8 V		82	110	ns
Minimum Off-Time		65% duty cycle (maximum)		320	385	ns
ADP1872ARMZ-1.0/ ADP1873ARMZ-1.0 (1.0 MHz)				1.0		MHz
On-Time		VIN = 5 V, V_{OUT} = 2 V, T_J = 25°C	285	312	340	ns
Minimum On-Time		VIN = 20 V		60	85	ns
Minimum Off-Time		45% duty cycle (maximum)		320	385	ns

Parameter	Symbol	Conditions	Min	Typ	Max	Unit
OUTPUT DRIVER CHARACTERISTICS						
High-Side Driver						
Output Source Resistance		I_{SOURCE} = 1.5 A, 100 ns, positive pulse (0 V to 5 V)		2	3.5	Ω
Output Sink Resistance		I_{SINK} = 1.5 A, 100 ns, negative pulse (5 V to 0 V)		0.8	2	Ω
Rise Time[2]	$t_{r, DRVH}$	BST − SW = 4.4 V, C_{IN} = 4.3 nF (see Figure 59)		25		ns
Fall Time[2]	$t_{f, DRVH}$	BST − SW = 4.4 V, C_{IN} = 4.3 nF (see Figure 60)		11		ns
Low-Side Driver						
Output Source Resistance		I_{SOURCE} = 1.5 A, 100 ns, positive pulse (0 V to 5 V)		1.7	3	Ω
Output Sink Resistance		I_{SINK} = 1.5 A, 100 ns, negative pulse (5 V to 0 V)		0.75	2	Ω
Rise Time[2]	$t_{r, DRVL}$	VDD = 5.0 V, C_{IN} = 4.3 nF (see Figure 60)		18		ns
Fall Time[2]	$t_{f, DRVL}$	VDD = 5.0 V, C_{IN} = 4.3 nF (see Figure 59)		16		ns
Propagation Delays						
DRVL Fall to DRVH Rise[2]	$t_{tpdh, DRVH}$	BST − SW = 4.4 V (see Figure 59)		22		ns
DRVH Fall to DRVL Rise[2]	$t_{tpdh, DRVL}$	BST − SW = 4.4 V (see Figure 60)		24		ns
SW Leakage Current	$I_{SW, LEAK}$	BST = 25 V, SW = 20 V, VDD = 5.5 V			110	µA
Integrated Rectifier						
Channel Impedance		I_{SINK} = 10 mA		22		Ω
PRECISION ENABLE THRESHOLD						
Logic High Level		VIN = 2.9 V to 20 V, VDD = 2.75 V to 5.5 V	235	285	330	mV
Enable Hysteresis		VIN = 2.9 V to 20 V, VDD = 2.75 V to 5.5 V		35		mV
COMP VOLTAGE						
COMP Clamp Low Voltage	$V_{COMP (LOW)}$	From disable state, release COMP/EN pin to enable device (2.75 V ≤ VDD ≤ 5.5 V)	0.47			V
COMP Clamp High Voltage	$V_{COMP (HIGH)}$	(2.75 V ≤ VDD ≤ 5.5 V)			2.55	V
COMP Zero Current Threshold	V_{COMP_ZCT}	(2.75 V ≤ VDD ≤ 5.5 V)		1.15		V
THERMAL SHUTDOWN	T_{TMSD}					
Thermal Shutdown Threshold		Rising temperature		155		°C
Thermal Shutdown Hysteresis				15		°C
Hiccup Current Limit Timing				6		ms

[1] The maximum specified values are with the closed loop measured at 10% to 90% time points (see Figure 59 and Figure 60), C_{GATE} = 4.3 nF and upper- and lower-side MOSFETs being Infineon BSC042N03MS G.
[2] Not automatic test equipment (ATE) tested.

ABSOLUTE MAXIMUM RATINGS

Table 2.

Parameter	Rating
VDD to GND	−0.3 V to +6 V
VIN to PGND	−0.3 V to +28 V
FB, COMP/EN to GND	−0.3 V to (VDD + 0.3 V)
DRVL to PGND	−0.3 V to (VDD + 0.3 V)
SW to PGND	−0.3 V to +28 V
SW to PGND	−2 V pulse (20 ns)
BST to SW	−0.6 V to (VDD + 0.3 V)
BST to PGND	−0.3 V to +28 V
DRVH to SW	−0.3 V to VDD
PGND to GND	±0.3 V
Operating Junction Temperature Range	−40°C to +125°C
Storage Temperature Range	−65°C to +150°C
Soldering Conditions	JEDEC J-STD-020
Maximum Soldering Lead Temperature (10 sec)	300°C

Stresses above those listed under Absolute Maximum Ratings may cause permanent damage to the device. This is a stress rating only; functional operation of the device at these or any other conditions above those indicated in the operational section of this specification is not implied. Exposure to absolute maximum rating conditions for extended periods may affect device reliability.

Absolute maximum ratings apply individually only, not in combination. Unless otherwise specified, all other voltages are referenced to PGND.

THERMAL RESISTANCE

θ_{JA} is specified for the worst-case conditions, that is, a device soldered in a circuit board for surface-mount packages.

Table 3. Thermal Resistance

Package Type	θ_{JA}	Unit
θ_{JA} (10-Lead MSOP)		
2-Layer Board	213.1	°C/W
4-Layer Board	171.7	°C/W

BOUNDARY CONDITION

In determining the values given in Table 2 and Table 3, natural convection was used to transfer heat to a 4-layer evaluation board.

ESD CAUTION

ESD (electrostatic discharge) sensitive device. Charged devices and circuit boards can discharge without detection. Although this product features patented or proprietary protection circuitry, damage may occur on devices subjected to high energy ESD. Therefore, proper ESD precautions should be taken to avoid performance degradation or loss of functionality.

PIN CONFIGURATION AND FUNCTION DESCRIPTIONS

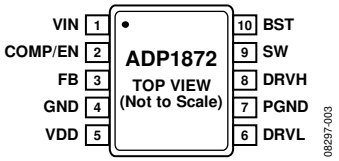

Figure 3. Pin Configuration

Table 4. Pin Function Descriptions

Pin No.	Mnemonic	Description
1	VIN	High Input Voltage. Connect VIN to the drain of the upper-side MOSFET.
2	COMP/EN	Output of the Internal Error Amplifier/IC Enable. When this pin functions as EN, applying 0 V to this pin disables the IC.
3	FB	Noninverting Input of the Internal Error Amplifier. This is the node where the feedback resistor is connected.
4	GND	Analog Ground Reference Pin of the IC. All sensitive analog components should be connected to this ground plane (see the Layout Considerations Section).
5	VDD	Bias Voltage Supply for the ADP1872/ADP1873 Controller (Includes the Output Gate Drivers). A bypass capacitor of 1 µF directly from this pin to PGND and a 0.1 µF across VDD and GND are recommended.
6	DRVL	Drive Output for the External Lower Side, N-Channel MOSFET. This pin also serves as the current-sense gain setting pin (see Figure 68).
7	PGND	Power GND. Ground for the lower side gate driver and lower side, N-channel MOSFET.
8	DRVH	Drive Output for the External Upper Side, N-Channel MOSFET.
9	SW	Switch Node Connection.
10	BST	Bootstrap for the Upper Side MOSFET Gate Drive Circuitry. An internal boot rectifier (diode) is connected between VDD and BST. A capacitor from BST to SW is required. An external Schottky diode can also be connected between VDD and BST for increased gate drive capability.

TYPICAL PERFORMANCE CHARACTERISTICS

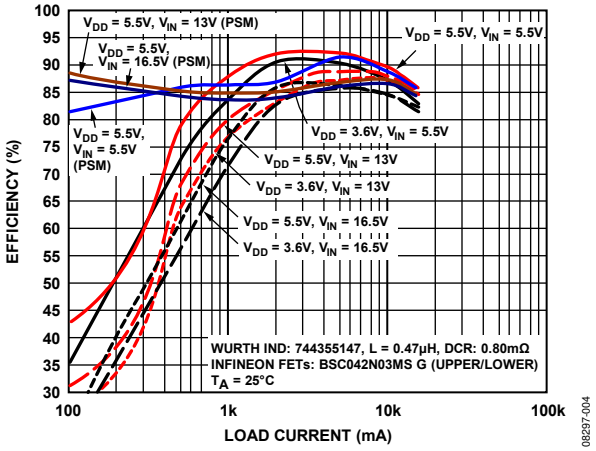

Figure 4. Efficiency—300 kHz, V_{OUT} = 0.8 V

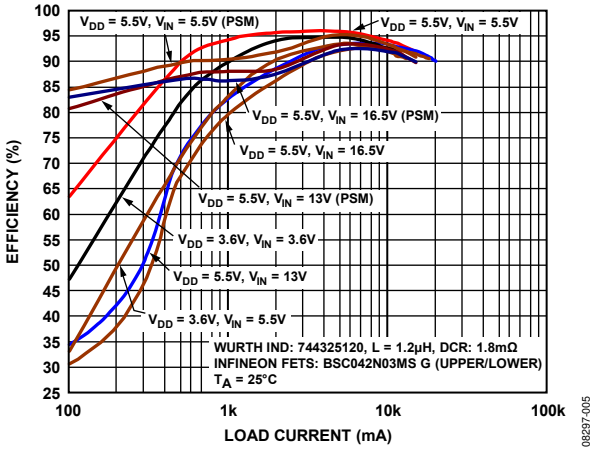

Figure 5. Efficiency—300 kHz, V_{OUT} = 1.8 V

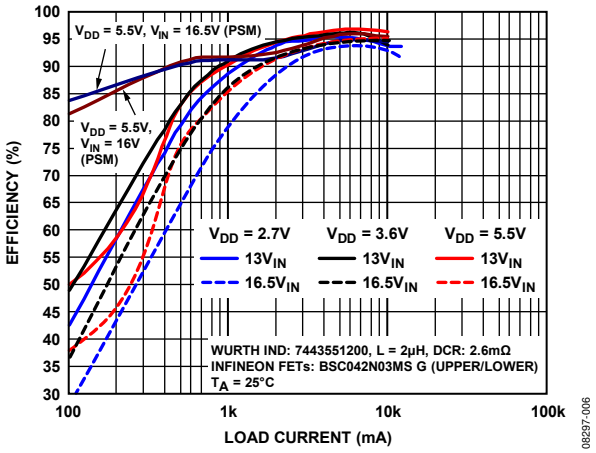

Figure 6. Efficiency—300 kHz, V_{OUT} = 7 V

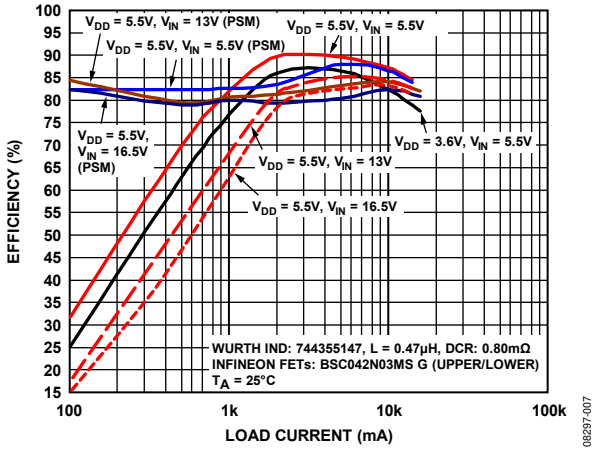

Figure 7. Efficiency—600 kHz, V_{OUT} = 0.8 V

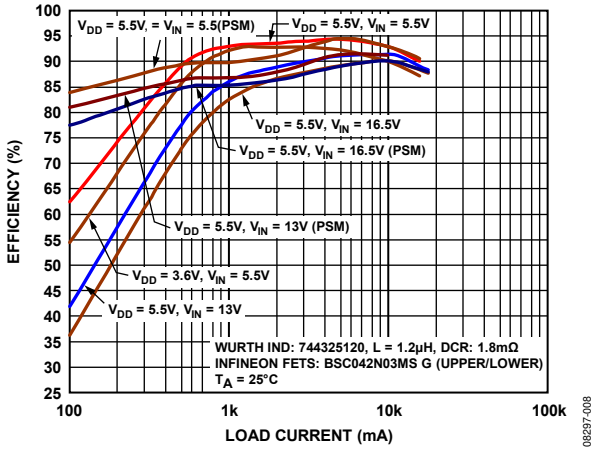

Figure 8. Efficiency—600 kHz, V_{OUT} = 1.8 V

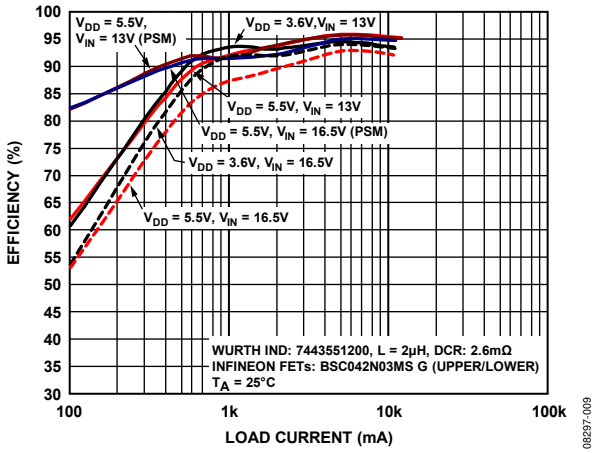

Figure 9. Efficiency—600 kHz, V_{OUT} = 5 V

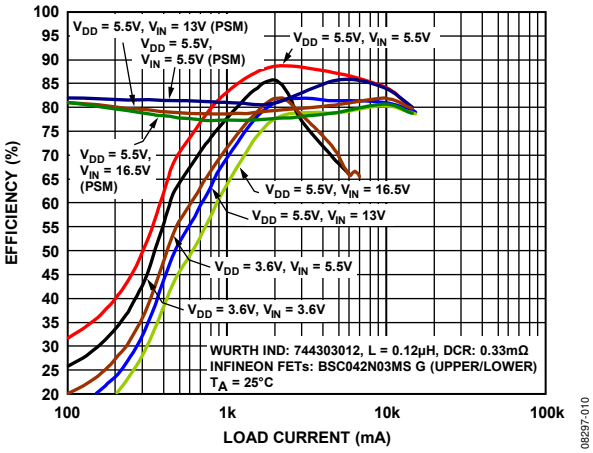

Figure 10. Efficiency—1.0 MHz, V_{OUT} = 0.8 V

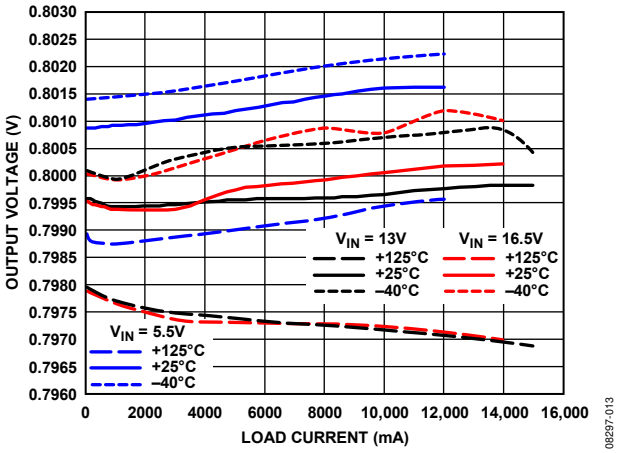

Figure 13. Output Voltage Accuracy—300 kHz, V_{OUT} = 0.8 V

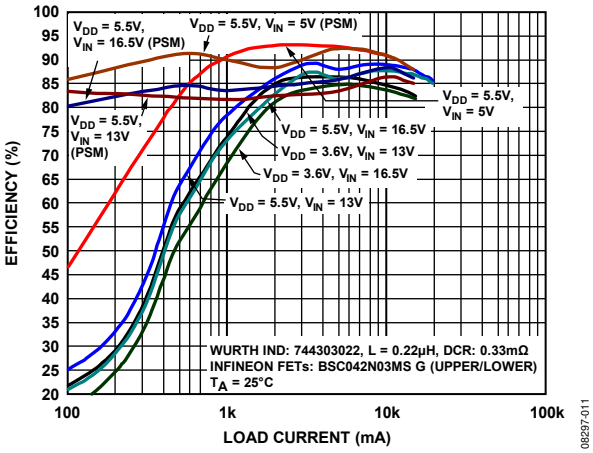

Figure 11. Efficiency—1.0 MHz, V_{OUT} = 1.8 V

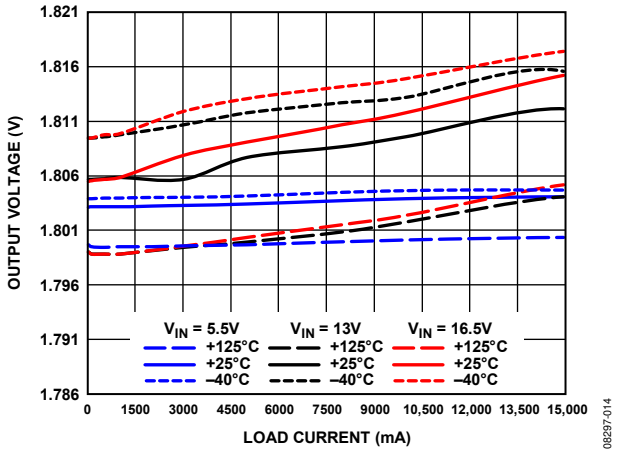

Figure 14. Output Voltage Accuracy—300 kHz, V_{OUT} = 1.8 V

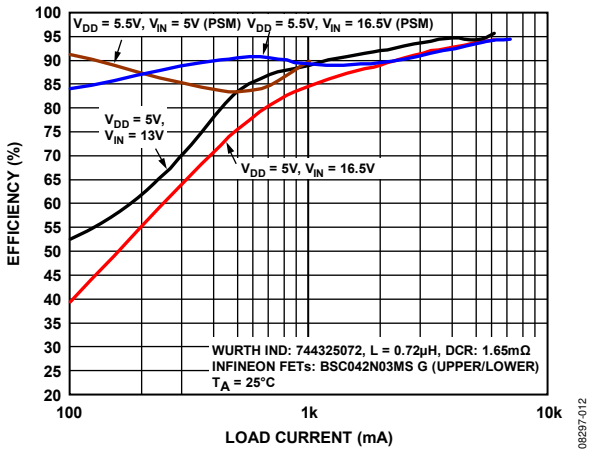

Figure 12. Efficiency—1.0 MHz, V_{OUT} = 4 V

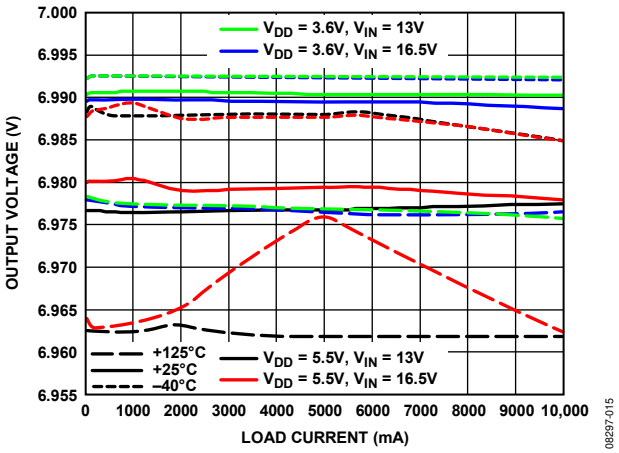

Figure 15. Output Voltage Accuracy—300 kHz, V_{OUT} = 7 V

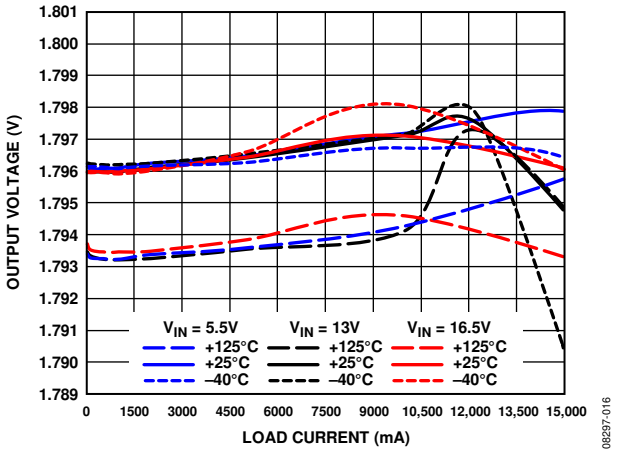

Figure 16. Output Voltage Accuracy—600 kHz, V_{OUT} = 1.8 V

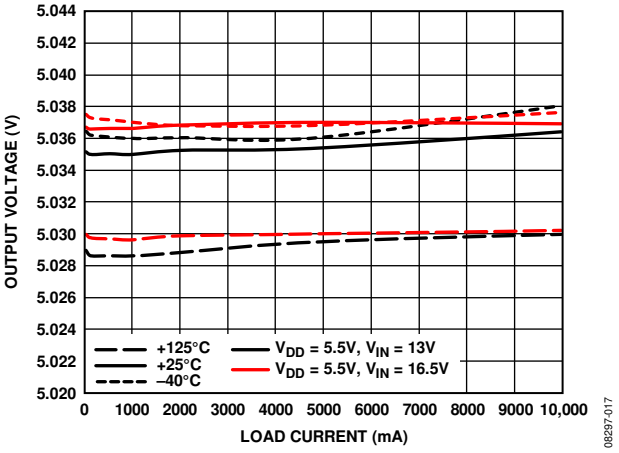

Figure 17. Output Voltage Accuracy—600 kHz, V_{OUT} = 5 V

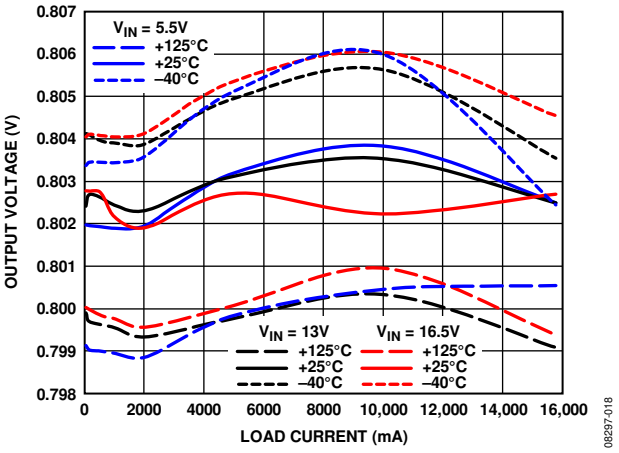

Figure 18. Output Voltage Accuracy—1 MHz, V_{OUT} = 0.8 V

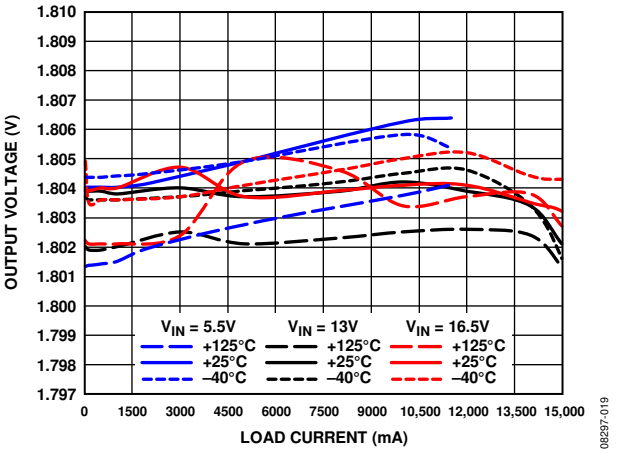

Figure 19. Output Voltage Accuracy—1.0 MHz, V_{OUT} = 1.8 V

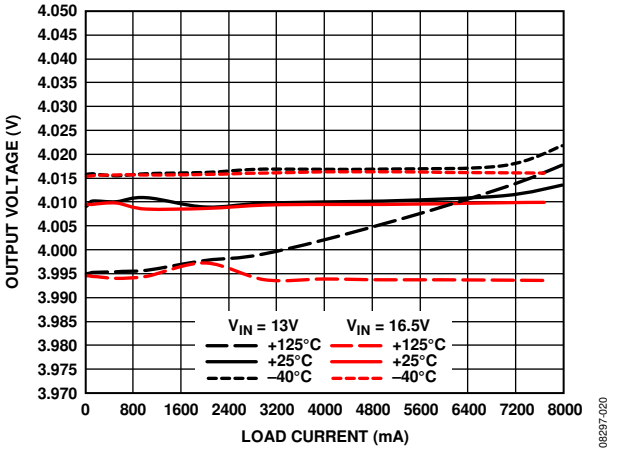

Figure 20. Output Voltage Accuracy—1.0 MHz, V_{OUT} = 4 V

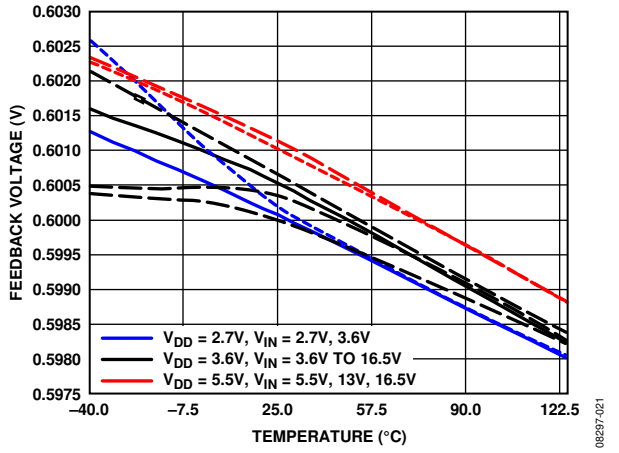

Figure 21. Feedback Voltage vs. Temperature

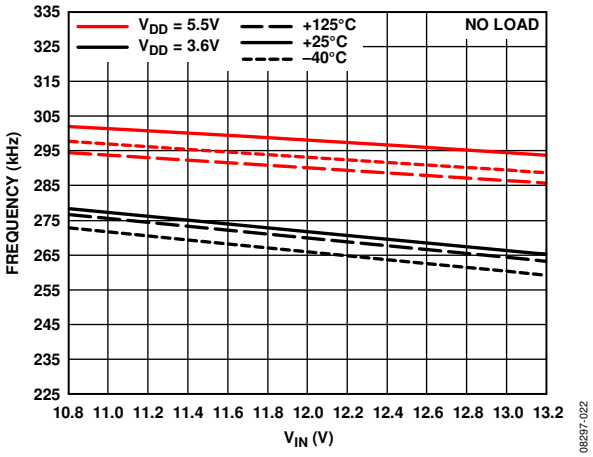

Figure 22. Switching Frequency vs. High Input Voltage, 300 kHz, ±10% of 12 V

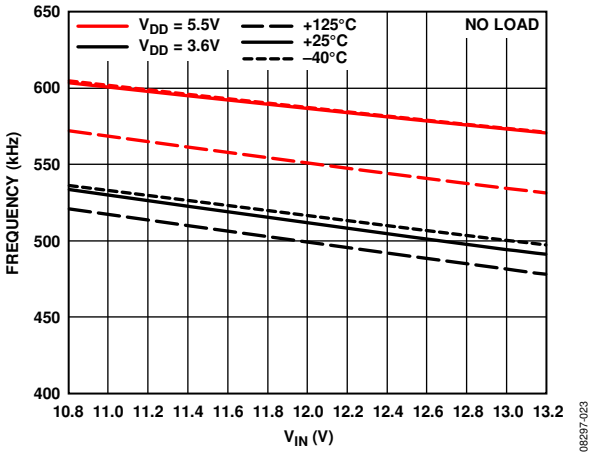

Figure 23. Switching Frequency vs. High Input Voltage, 600 kHz, V_{OUT} = 1.8 V, ±10% of 12 V

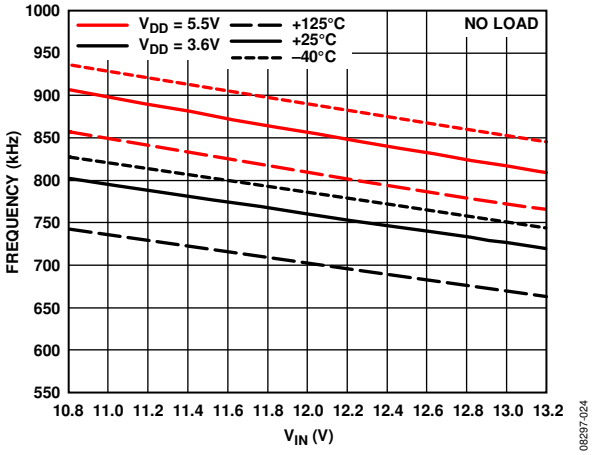

Figure 24. Switching Frequency vs. High Input Voltage, 1.0 MHz, ±10% of 12 V

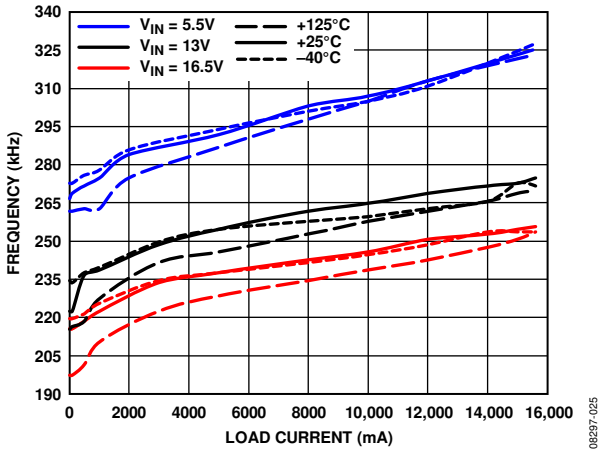

Figure 25. Frequency vs. Load Current, 300 kHz, V_{OUT} = 0.8 V

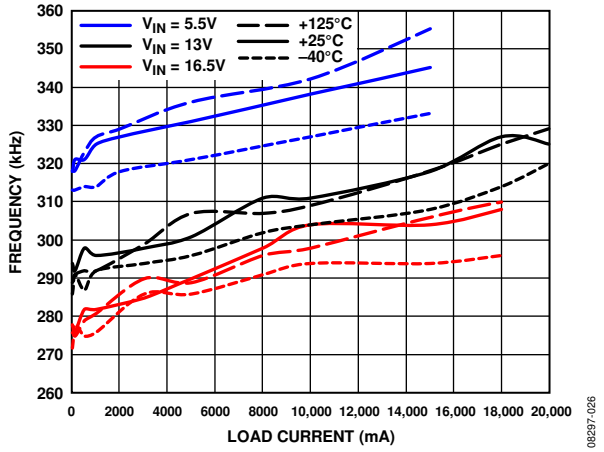

Figure 26. Frequency vs. Load Current, 300 kHz, V_{OUT} = 1.8 V

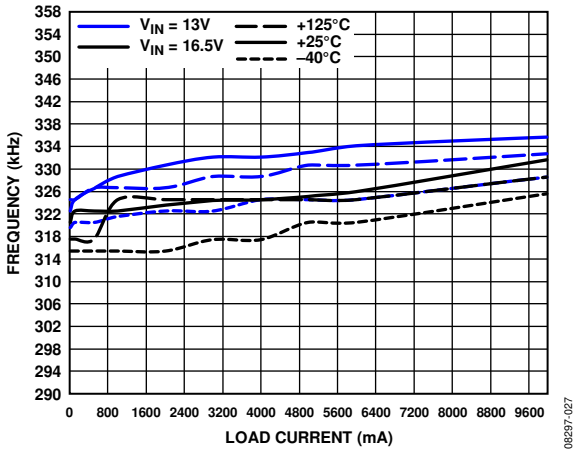

Figure 27. Frequency vs. Load Current, 300 kHz, V_{OUT} = 7 V

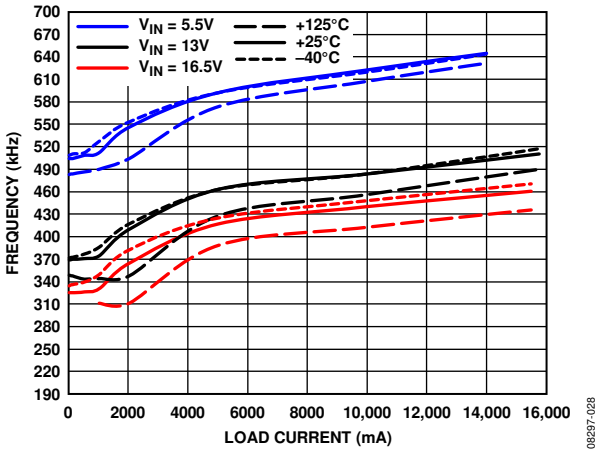

Figure 28. Frequency vs. Load Current, 600 kHz, V_{OUT} = 0.8 V

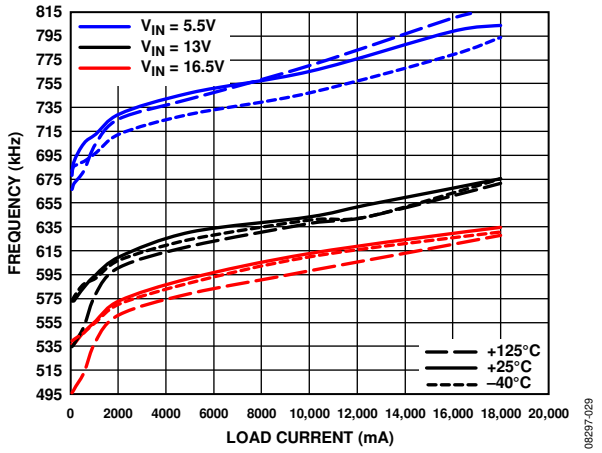

Figure 29. Frequency vs. Load Current, 600 kHz, V_{OUT} = 1.8 V

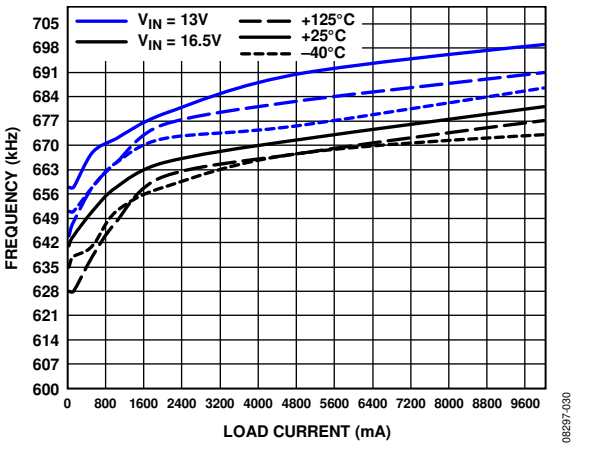

Figure 30. Frequency vs. Load Current, 600 kHz, V_{OUT} =5 V

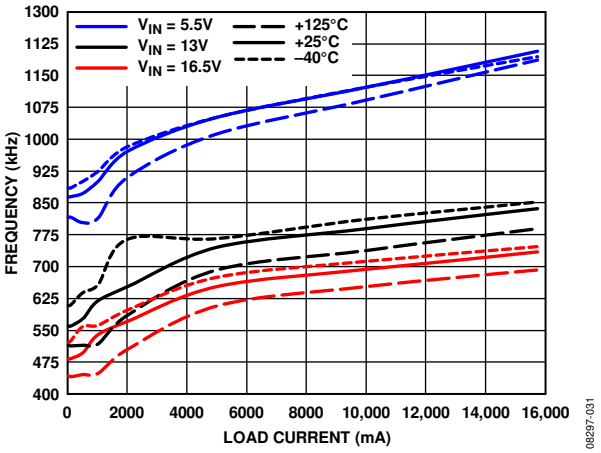

Figure 31. Frequency vs. Load Current, V_{OUT} = 1.0 MHz, 0.8 V

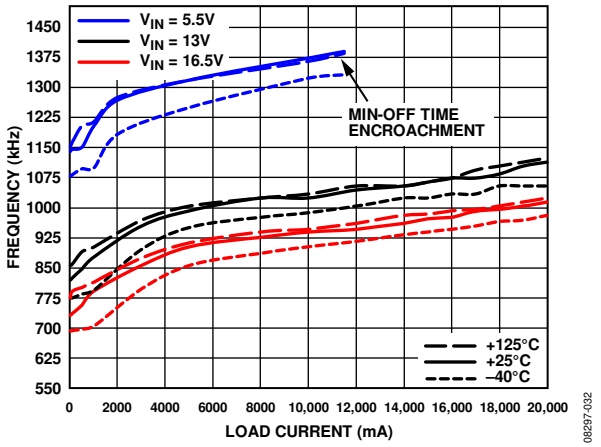

Figure 32. Frequency vs. Load Current, 1.0 MHz, V_{OUT} = 1.8 V

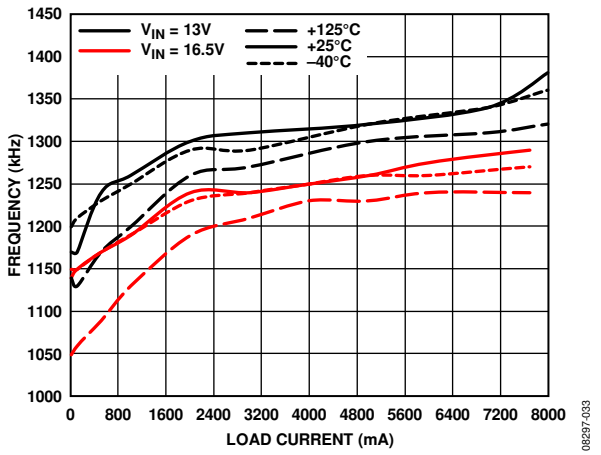

Figure 33. Frequency vs. Load Current, 1.0 MHz, V_{OUT} = 4 V

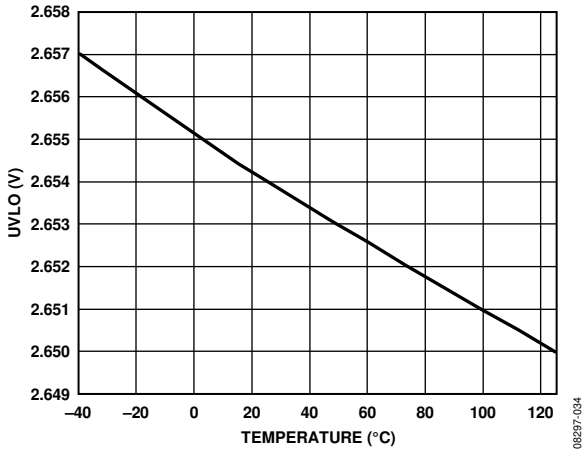

Figure 34. UVLO vs. Temperature

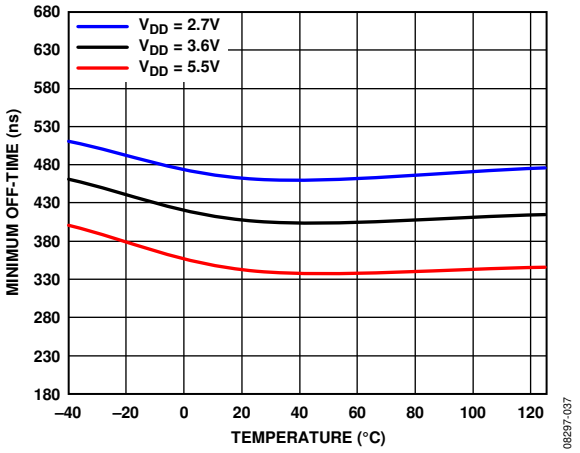

Figure 37. Minimum Off-Time vs. Temperature

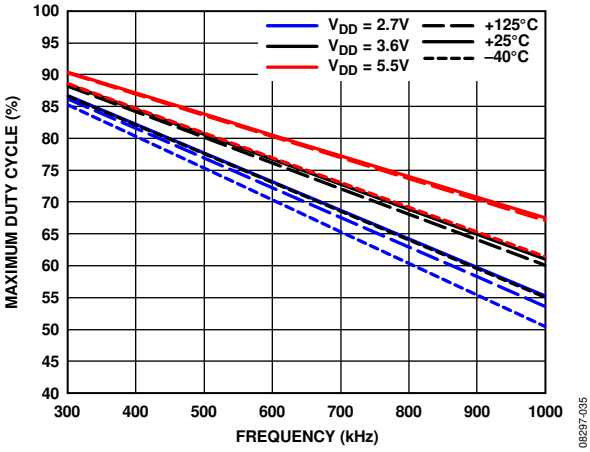

Figure 35. Maximum Duty Cycle vs. Frequency

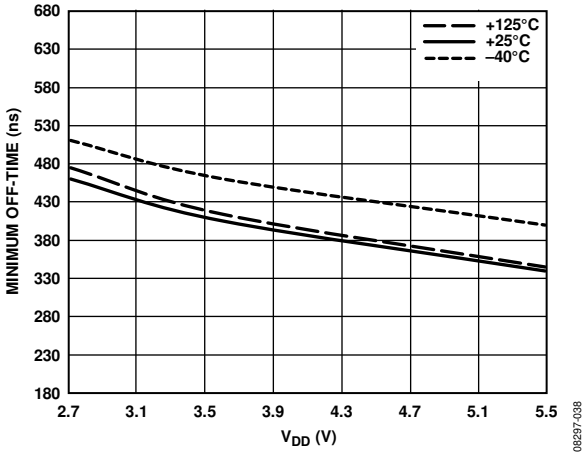

Figure 38. Minimum Off-Time vs. V_{DD} (Low Input Voltage)

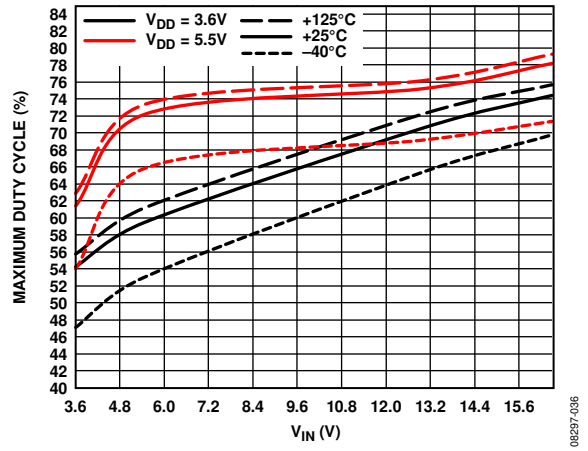

Figure 36. Maximum Duty Cycle vs. High Voltage Input (V_{IN})

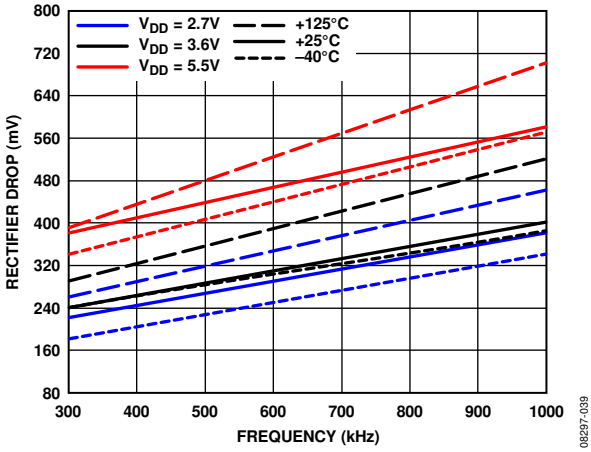

Figure 39. Internal Rectifier Drop vs. Frequency

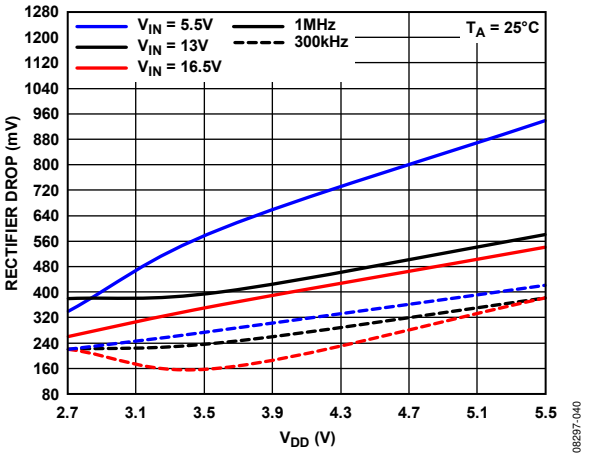

Figure 40. Internal Boost Rectifier Drop vs. V_{DD} (Low Input Voltage) over VIN Variation

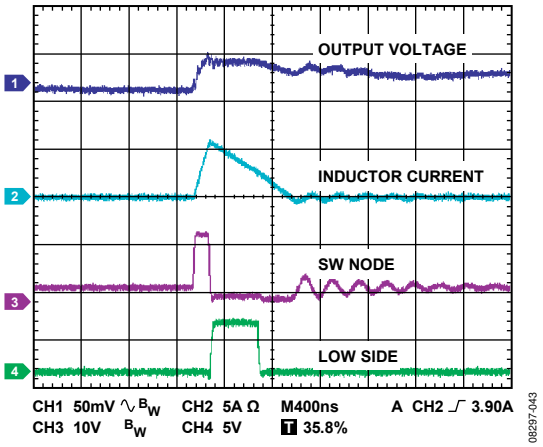

Figure 43. Power Saving Mode (PSM) Operational Waveform, 100 mA

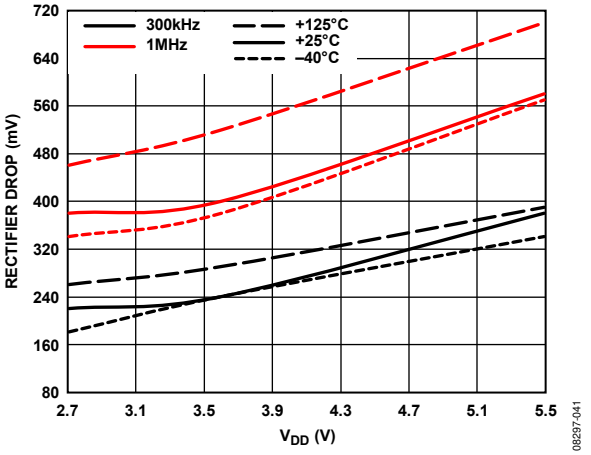

Figure 41. Internal Boost Rectifier Drop vs. V_{DD}

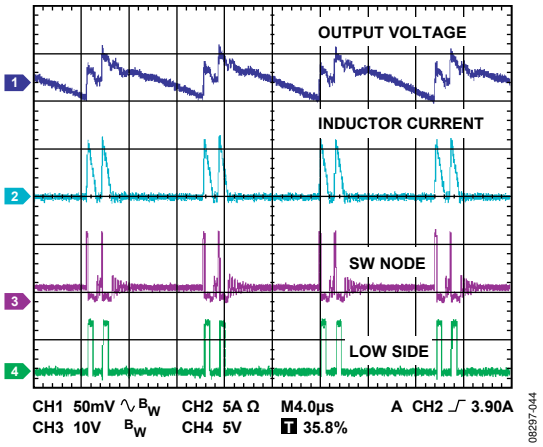

Figure 44. PSM Waveform at Light Load, 500 mA

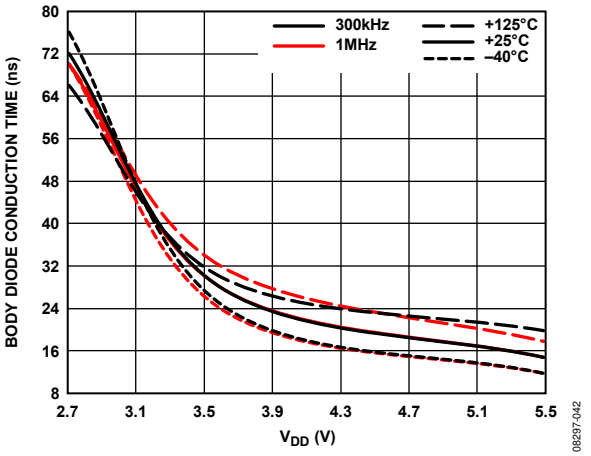

Figure 42. Lower Side MOSFET Body Conduction Time vs. V_{DD} (Low Input Voltage)

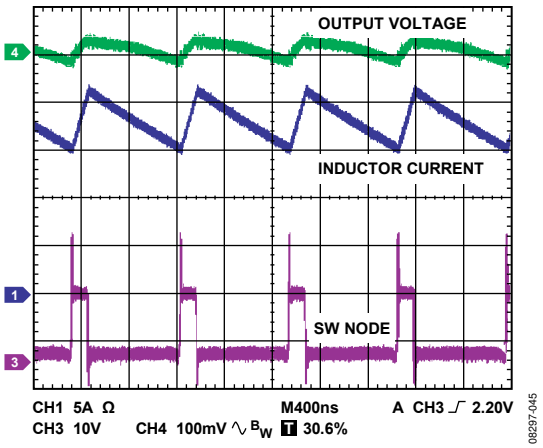

Figure 45. CCM Operation at Heavy Load, 18 A (See Figure 91 for Application Circuit)

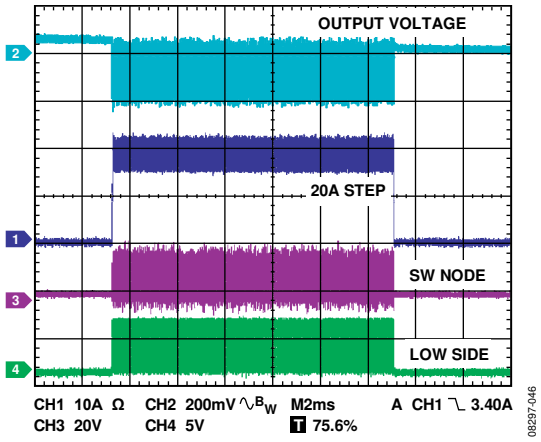

Figure 46. Load Transient Step—PSM Enabled, 20 A (See Figure 91 Application Circuit)

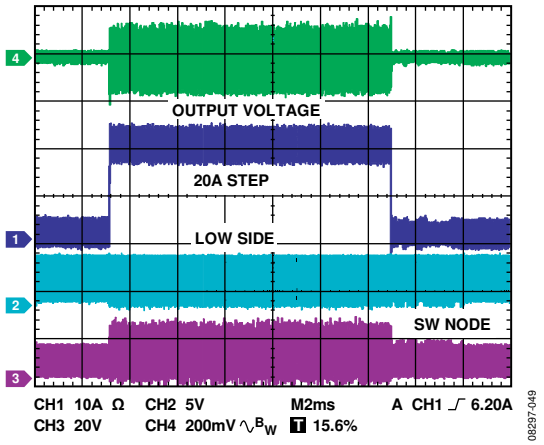

Figure 49. Load Transient Step—Forced PWM at Light Load, 20 A (See Figure 91 Application Circuit)

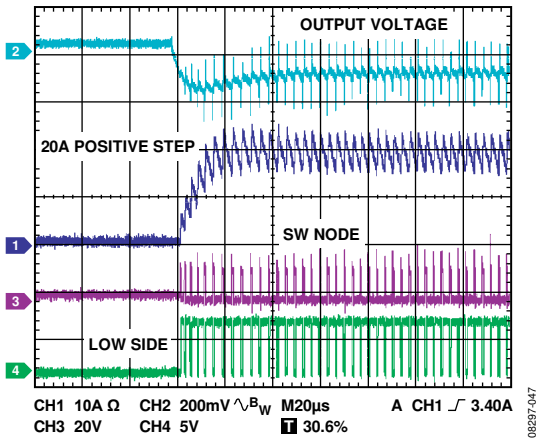

Figure 47. Positive Step During Heavy Load Transient Behavior—PSM Enabled, 20 A, V_{OUT} = 1.8 V (See Figure 91 Application Circuit)

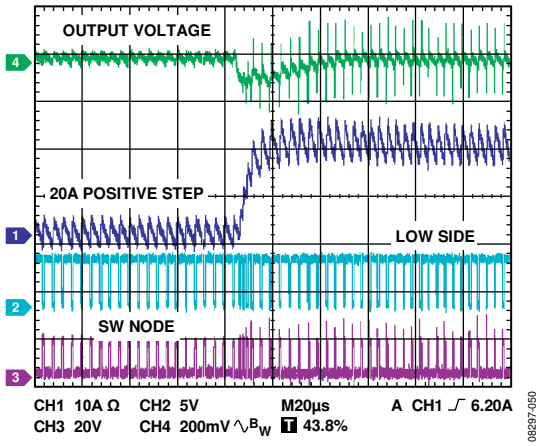

Figure 50. Positive Step During Heavy Load Transient Behavior—Forced PWM at Light Load, 20 A, V_{OUT} = 1.8 V (See Figure 91 Application Circuit)

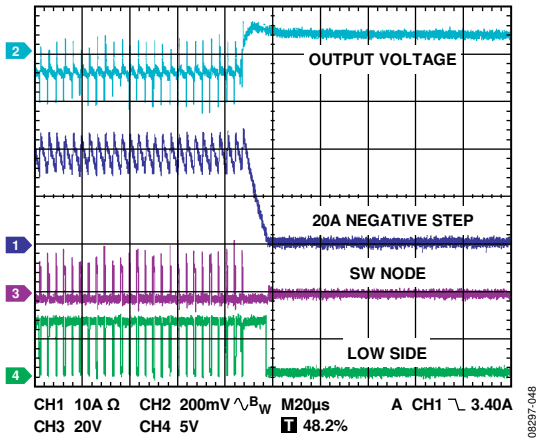

Figure 48. Negative Step During Heavy Load Transient Behavior—PSM Enabled, 20 A (See Figure 91 Application Circuit)

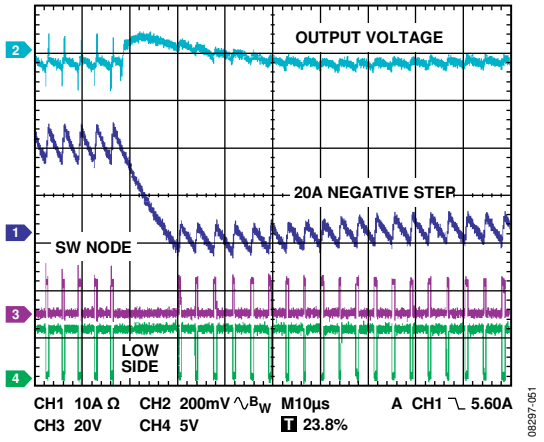

Figure 51. Negative Step During Heavy Load Transient Behavior—Forced PWM at Light Load, 20 A (See Figure 91 Application Circuit)

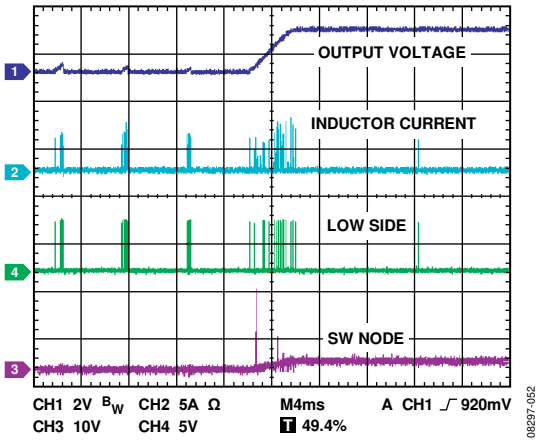

Figure 52. Output Short-Circuit Behavior Leading to Hiccup Mode

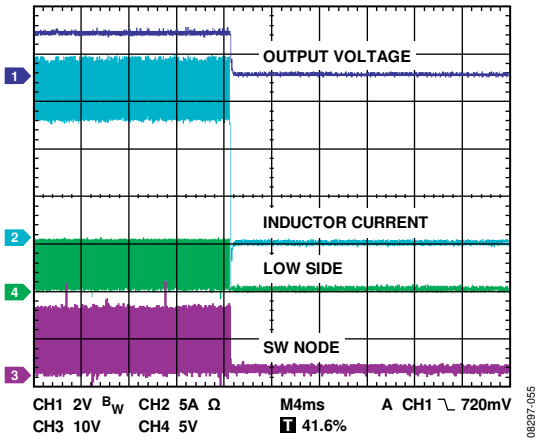

Figure 55. Power-Down Waveform During Heavy Load

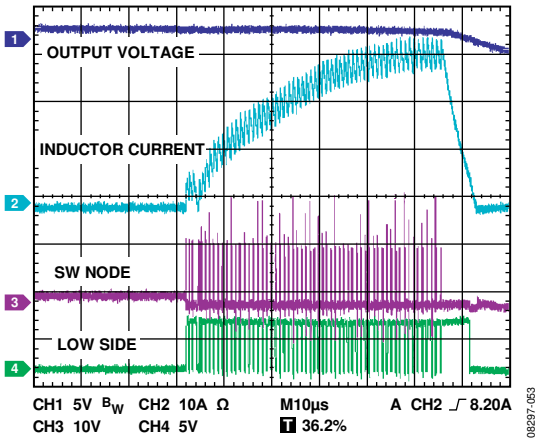

Figure 53. Magnified Waveform During Hiccup Mode

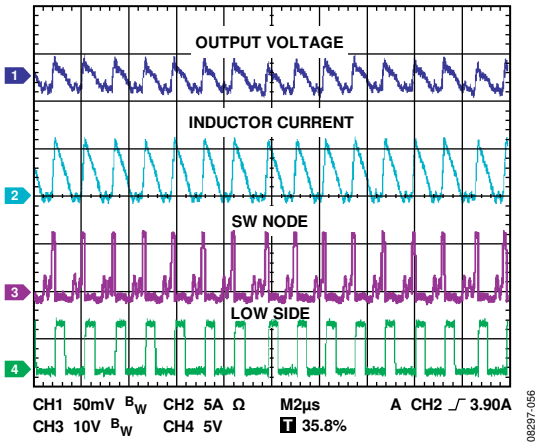

Figure 56. Output Voltage Ripple Waveform During PSM Operation at Light Load, 2 A

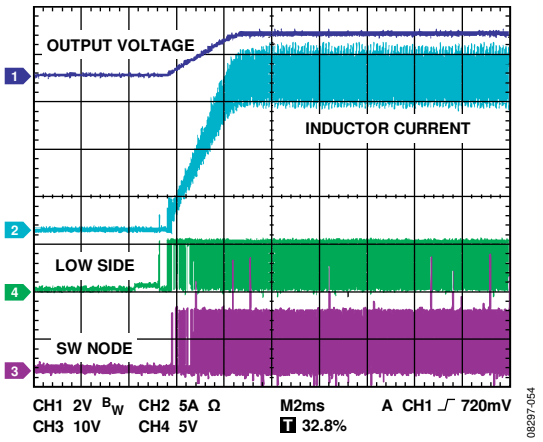

Figure 54. Start-Up Behavior at Heavy Load, 18 A, 300 kHz (See Figure 91 Application Circuit)

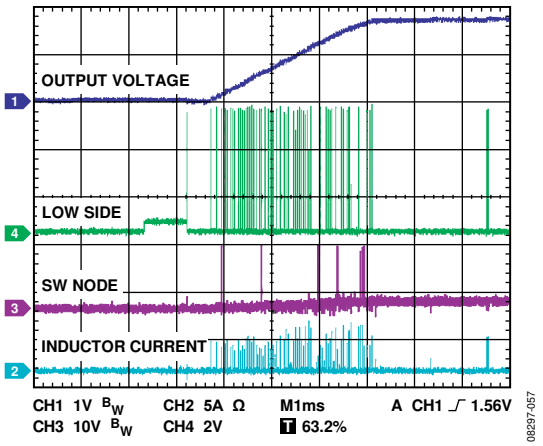

Figure 57. Soft Start and RES Detect Waveform

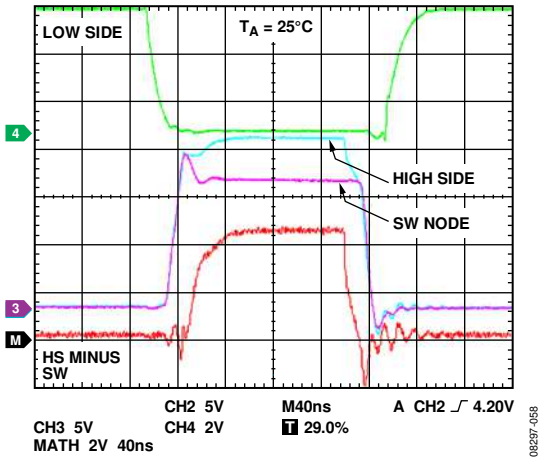

Figure 58. Output Drivers and SW Node Waveforms

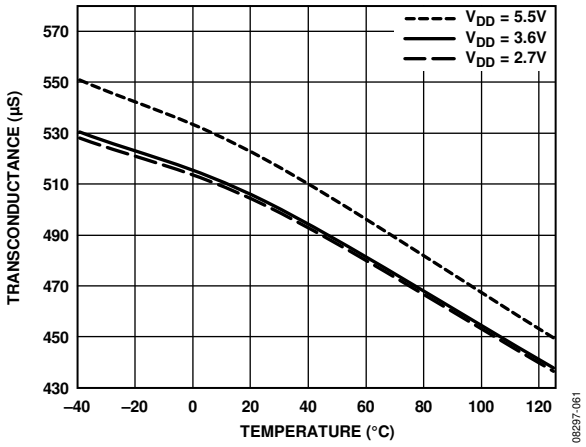

Figure 61. Transconductance (G_M) vs. Temperature

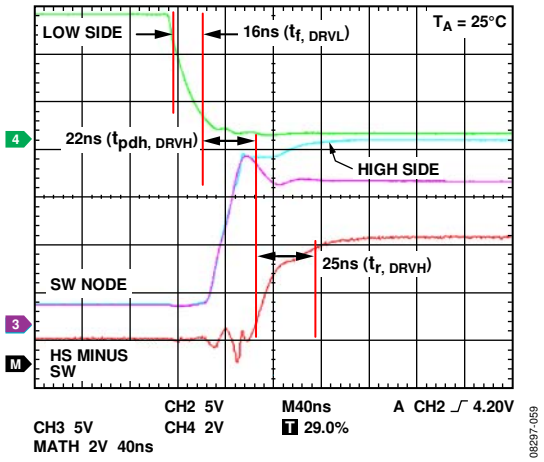

Figure 59. Upper Side Driver Rising and Lower Side Falling Edge Waveforms
(C_{GATE} = 4.3 nF (Upper/Lower Side MOSFET),
Q_{TOTAL} = 27 nC (V_{GS} = 4.4 V (Q1), V_{GS} = 5 V (Q3))

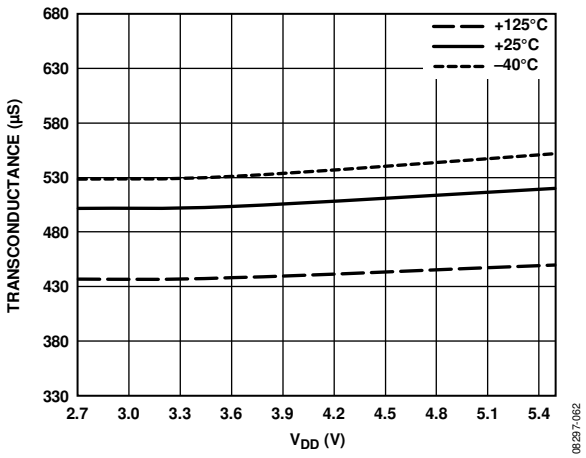

Figure 62. Transconductance (G_M) vs. V_{DD}

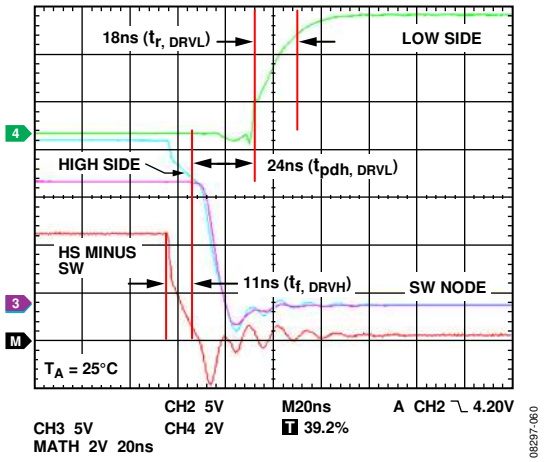

Figure 60. Upper Side Driver Falling and Lower Side Rising Edge Waveforms
(C_{GATE} = 4.3 nF (Upper/Lower Side MOSFET),
Q_{TOTAL} = 27 nC (V_{GS} = 4.4 V (Q1), V_{GS} = 5 V (Q3))

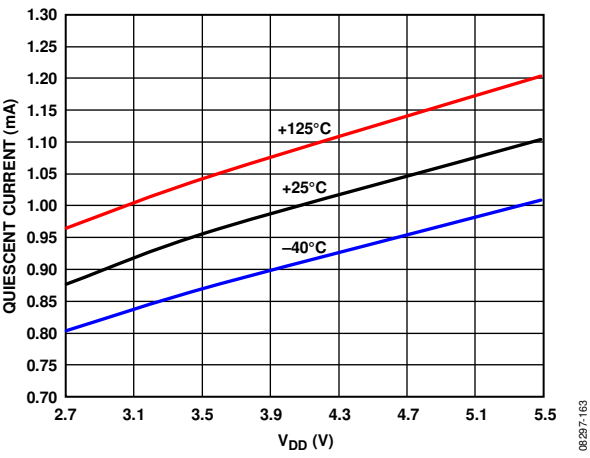

Figure 63. Quiescent Current vs. V_{DD} (VIN = 13 V)

ADP1872/ADP1873 BLOCK DIGRAM

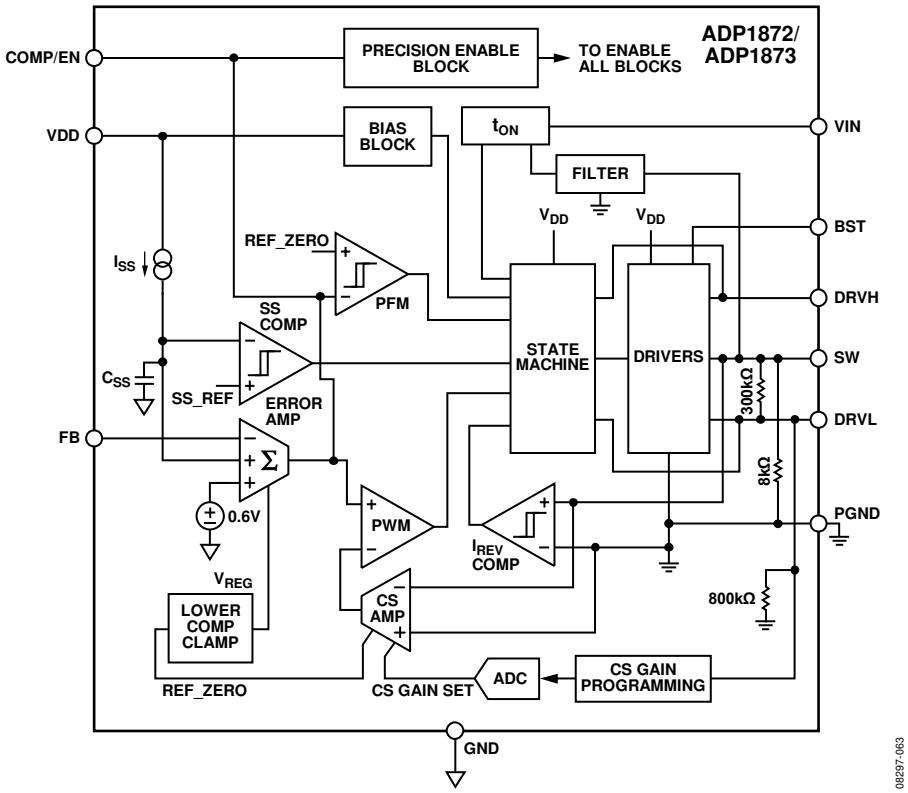

Figure 64. ADP1872/ADP1873 Block Diagram

THEORY OF OPERATION

The ADP1872/ADP1873 are versatile current-mode, synchronous step-down controllers that provide superior transient response, optimal stability, and current limit protection by using a constant on-time, pseudo-fixed frequency with a programmable current-sense gain, current-control scheme. In addition, these devices offer optimum performance at low duty cycles by using valley current-mode control architecture. This allows the ADP1872/ADP1873 to drive all N-channel power stages to regulate output voltages as low as 0.6 V.

STARTUP

The ADP1872/ADP1873 have an input low voltage pin (VDD) for biasing and supplying power for the integrated MOSFET drivers. A bypass capacitor should be located directly across the VDD (Pin 5) and PGND (Pin 7) pins. Included in the power-up sequence is the biasing of the current-sense amplifier, the current-sense gain circuit (see the Programming Resistor (RES) Detect Circuit section), the soft start circuit, and the error amplifier.

The current-sense blocks provide valley current information (see the Programming Resistor (RES) Detect Circuit section) and are a variable of the compensation equation for loop stability (see the Compensation Network section). The valley current information is extracted by forcing 0.4 V across the DRVL output and the PGND pin, which generates a current depending on the resistor across DRVL and PGND in a process performed by the RES detect circuit. The current through the resistor is used to set the current-sense amplifier gain. This process takes approximately 800 µs, after which the drive signal pulses appear at the DRVL and DRVH pins synchronously and the output voltage begins to rise in a controlled manner through the soft start sequence.

The rise time of the output voltage is determined by the soft start and error amplifier blocks (see the Soft Start section). At the beginning of a soft start, the error amplifier charges the external compensation capacitor, causing the COMP/EN pin to rise above the enable threshold of 285 mV, thus enabling the ADP1872/ADP1873.

SOFT START

The ADP1872/ADP1873 have digital soft start circuitry, which involves a counter that initiates an incremental increase in current, by 1 µA, via a current source on every cycle through a fixed internal capacitor. The output tracks the ramping voltage by producing PWM output pulses to the upper side MOSFET. The purpose is to limit the in-rush current from the high voltage input supply (VIN) to the output (V_{OUT}).

PRECISION ENABLE CIRCUITRY

The ADP1872/ADP1873 employ precision enable circuitry. The enable threshold is 285 mV typical with 35 mV of hysteresis. The devices are enabled when the COMP/EN pin is released, allowing the error amplifier output to rise above the enable threshold (see Figure 65). Grounding this pin disables the ADP1872/ADP1873, reducing the supply current of the devices to approximately 140 µA. For more information, see Figure 66.

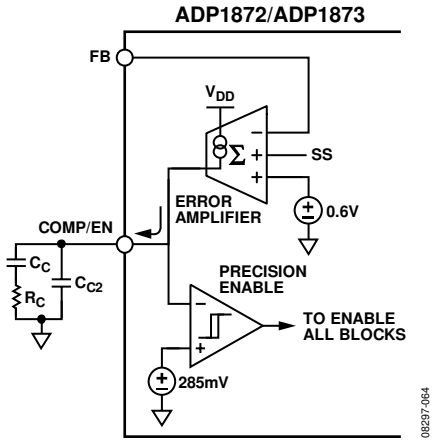

Figure 65. Release COMP/EN Pin to Enable the ADP1872/ADP1873

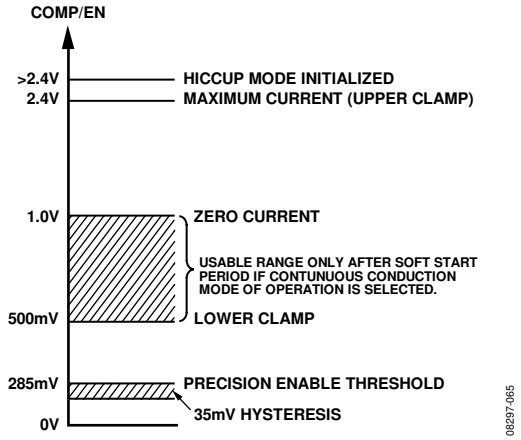

Figure 66. COMP/EN Voltage Range

UNDERVOLTAGE LOCKOUT

The undervoltage lockout (UVLO) feature prevents the part from operating both the upper side and lower side MOSFETs at extremely low or undefined input voltage (VDD) ranges. Operation at an undefined bias voltage may result in the incorrect propagation of signals to the high-side power switches. This, in turn, results in invalid output behavior that can cause damage to the output devices, ultimately destroying the device tied at the output. The UVLO level has been set at 2.65 V (nominal).

THERMAL SHUTDOWN

The thermal shutdown is a self-protection feature to prevent the IC from damage due to a very high operating junction temperature. If the junction temperature of the device exceeds 155°C, the part enters the thermal shutdown state. In this state, the device shuts off both the upper side and lower side MOSFETs and disables the entire controller immediately, thus reducing the power consumption of the IC. The part resumes operation after the junction temperature of the part cools to less than 140°C.

PROGRAMMING RESISTOR (RES) DETECT CIRCUIT

Upon startup, one of the first blocks to become active is the RES detect circuit. This block powers up before soft start begins. It forces a 0.4 V reference value at the DRVL output (see Figure 67) and is programmed to identify four possible resistor values: 47 kΩ, 22 kΩ, open, and 100 kΩ.

The RES detect circuit digitizes the value of the resistor at the DRVL pin (Pin 6). An internal ADC outputs a 2-bit digital code that is used to program four separate gain configurations in the current-sense amplifier (see Figure 68). Each configuration corresponds to a current-sense gain (A_{CS}) of 3 V/V, 6 V/V, 12 V/V, 24 V/V, respectively (see Table 5 and Table 6). This variable is used for the valley current-limit setting, which sets up the appropriate current-sense gain for a given application and sets the compensation necessary to achieve loop stability (see the Valley Current-Limit Setting and Compensation Network sections).

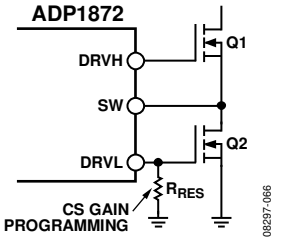

Figure 67. Programming Resistor Location

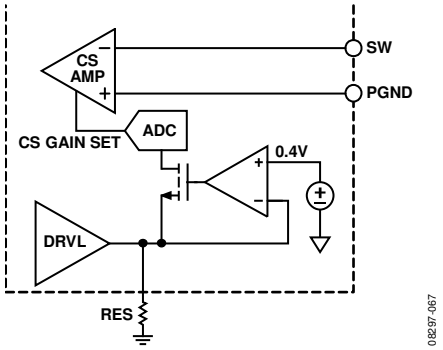

Figure 68. RES Detect Circuit for Current-Sense Gain Programming

Table 5. Current-Sense Gain Programming

Resistor	A_{CS} (V/V)
47 kΩ	3
22 kΩ	6
Open	12
100 kΩ	24

VALLEY CURRENT-LIMIT SETTING

The architecture of the ADP1872/ADP1873 is based on valley current-mode control. The current limit is determined by three components: the R_{ON} of the lower side MOSFET, the error amplifier output voltage swing (COMP), and the current-sense gain. The COMP range is internally fixed at 1.4 V. The current-sense gain is programmable via an external resistor at the DRVL pin (see the Programming Resistor (RES) Detect Circuit section). The R_{ON} of the lower side MOSFET can vary over temperature and usually has a positive T_C (meaning that it increases with temperature); therefore, it is recommended to program the current-sense gain resistor based on the rated R_{ON} of the MOSFET at 125°C.

Because the ADP1872/ADP1873 are based on valley current control, the relationship between I_{CLIM} and I_{LOAD} is

$$I_{CLIM} = I_{LOAD} \times \left(1 - \frac{K_I}{2}\right)$$

where:
I_{CLIM} is the desired valley current limit.
I_{LOAD} is the current load.
K_I is the ratio between the inductor ripple current and the desired average load current (see Figure 10). Establishing K_I helps to determine the inductor value (see the Inductor Selection section), but in most cases, K_I = 0.33.

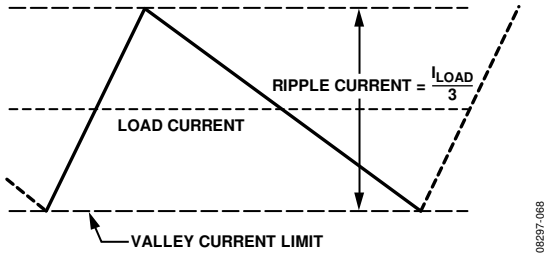

Figure 69. Valley Current Limit to Average Current Relation

When the desired valley current limit (I_{CLIM}) has been determined, the current-sense gain can be calculated by

$$I_{CLIM} = \frac{1.4\ \text{V}}{A_{CS} \times R_{ON}}$$

where:
A_{CS} is the current-sense gain multiplier (see Table 5 and Table 6).
R_{ON} is the channel impedance of the lower side MOSFET.

Although the ADP1872/ADP1873 have only four discrete current-sense gain settings for a given R_{ON} variable, Table 6 and Figure 70 outline several available options for the valley current setpoint based on various R_{ON} values.

Table 6. Valley Current Limit Program[1]

R_{ON} (mΩ)	Valley Current Level			
	47 kΩ	22 kΩ	Open	100 kΩ
	A_{CS} = 3 V/V	A_{CS} = 6 V/V	A_{CS} = 12 V/V	A_{CS} = 24 V/V
1.5				38.9
2				29.2
2.5				23.3
3			39.0	19.5
3.5			33.4	16.7
4.5			26.0	13
5			23.4	11.7
5.5			21.25	10.6
10		23.3	11.7	5.83
15	31.0	15.5	7.75	3.87
18	26.0	13.0	6.5	3.25

[1] Refer to Figure 70 for more information and a graphical representation.

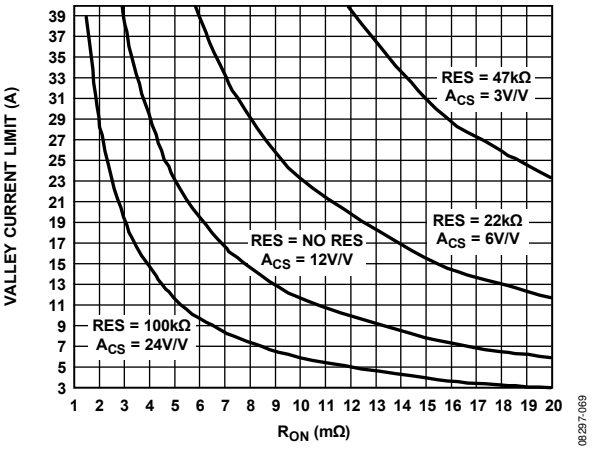

Figure 70. Valley Current-Limit Value vs. R_{ON} of the Lower Side MOSFET for Each Programming Resistor (RES)

The valley current limit is programmed as outlined in Table 6 and Figure 70. The inductor chosen must be rated to handle the peak current, which is equal to the valley current from Table 6 plus the peak-to-peak inductor ripple current (see the Inductor Selection section). In addition, the peak current value must be used to compute the worst-case power dissipation in the MOSFETs (see Figure 71).

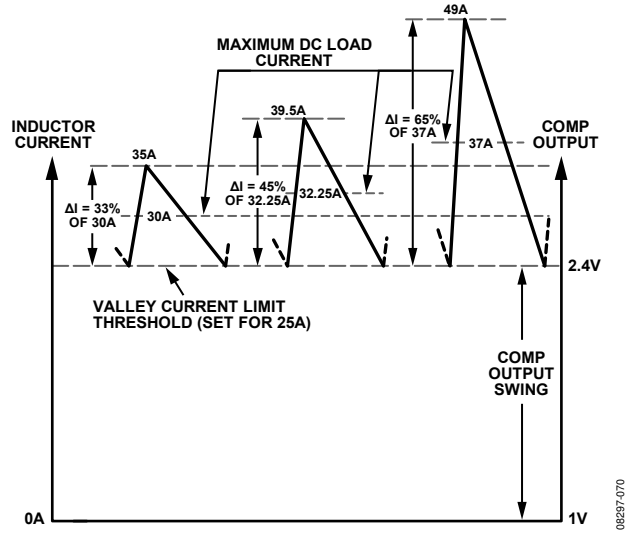

Figure 71. Valley Current-Limit Threshold in Relation to Inductor Ripple Current

HICCUP MODE DURING SHORT CIRCUIT

A current-limit violation occurs when the current across the source and drain of the lower side MOSFET exceeds the current-limit setpoint. When 32 current-limit violations are detected, the controller enters idle mode and turns off the MOSFETs for 6 ms, allowing the converter to cool down. Then, the controller re-establishes soft start and begins to cause the output to ramp up again (see Figure 72). While the output ramps up, COMP is monitored to determine if the violation is still present. If it is still present, the idle event occurs again, followed by the full-chip power-down sequence. This cycle continues until the violation no longer exists. If the violation disappears, the converter is allowed to switch normally, maintaining regulation.

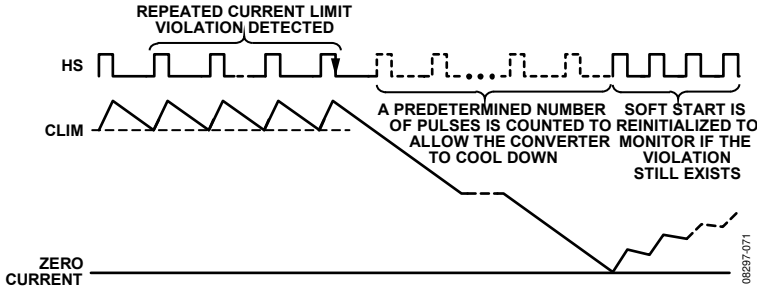

Figure 72. Idle Mode Entry Sequence Due to Current-Limit Violations

SYNCHRONOUS RECTIFIER

The ADP1872/ADP1873 employ an internal lower side MOSFET driver to drive the external upper side and lower side MOSFETs. The synchronous rectifier not only improves overall conduction efficiency but also ensures proper charging to the bootstrap capacitor located at the upper side driver input. This is beneficial during startup to provide sufficient drive signal to the external upper side MOSFET and attain fast turn-on response, which is essential for minimizing switching losses. The integrated upper and lower side MOSFET drivers operate in complementary fashion with built-in anticross conduction circuitry to prevent unwanted shoot-through current that may potentially damage the MOSFETs or reduce efficiency as a result of excessive power loss.

POWER SAVING MODE (PSM) VERSION (ADP1873)

The power saving mode version of the ADP1872 is the ADP1873. The ADP1873 operates in the discontinuous conduction mode (DCM) and pulse skips at light load to midload currents. It outputs pulses as necessary to maintain output regulation. Unlike the continuous conduction mode (CCM), DCM operation prevents negative current, thus allowing improved system efficiency at light loads. Current in the reverse direction through this pathway, however, results in power dissipation and therefore a decrease in efficiency.

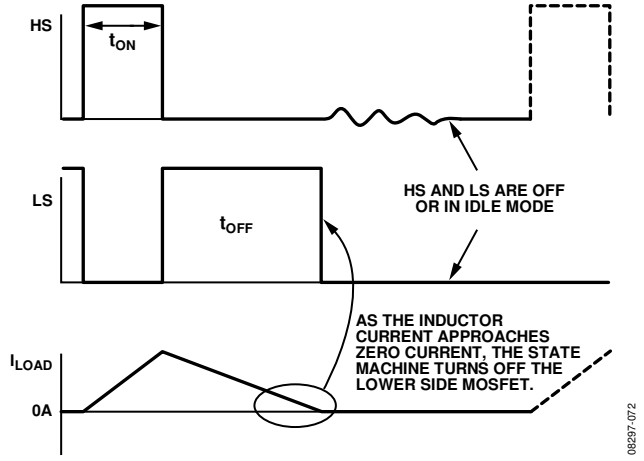

Figure 73. Discontinuous Mode of Operation (DCM)

To minimize the chance of negative inductor current buildup, an on-board, zero-cross comparator turns off all upper side and lower side switching activities when the inductor current approaches the zero current line, causing the system to enter idle mode, where the upper side and lower side MOSFETs are turned off. To ensure idle mode entry, a 10 mV offset, connected in series at the SW node, is implemented (see Figure 74).

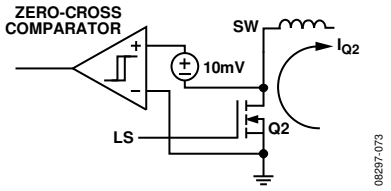

Figure 74. Zero-Cross Comparator with 10 mV of Offset

As soon as the forward current through the lower side MOSFET decreases to a level where

$$10\ \text{mV} = I_{Q2} \times R_{ON(Q2)}$$

the zero-cross comparator (or I_{REV} comparator) emits a signal to turn off the lower side MOSFET. From this point, the slope of the inductor current ramping down becomes steeper (see Figure 75) as the body diode of the lower side MOSFET begins to conduct current and continues conducting current until the remaining energy stored in the inductor has been depleted.

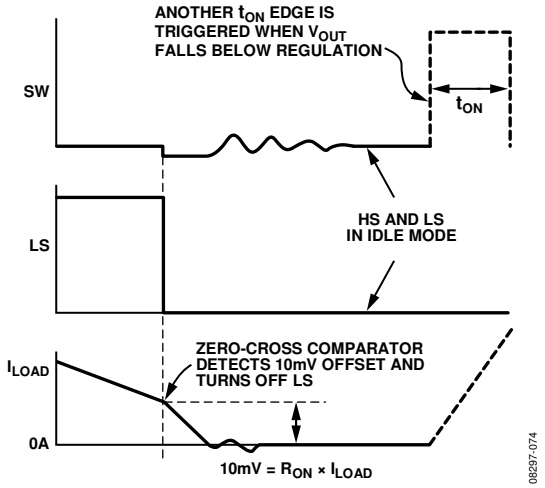

Figure 75. 10 mV Offset to Ensure Prevention of Negative Inductor Current

The system remains in idle mode until the output voltage drops below regulation. A PWM pulse is then produced, turning on the upper side MOSFET to maintain system regulation. The ADP1873 does not have an internal clock; therefore, it switches purely as a hysteretic controller, as described in this section.

TIMER OPERATION

The ADP1872/ADP1873 employ a constant on-time architecture, which provides a variety of benefits, including improved load and line transient response when compared with a constant (fixed) frequency current-mode control loop of comparable loop design. The constant on-time timer, or t_{ON} timer, senses the high input voltage (VIN) and the output voltage (V_{OUT}) using SW waveform information to produce an adjustable one-shot PWM pulse that varies the on-time of the upper side MOSFET in response to dynamic changes in input voltage, output voltage, and load current conditions to maintain regulation. It then generates an on-time (t_{ON}) pulse that is inversely proportional to V_{IN}.

$$t_{ON} = K \times \frac{V_{OUT}}{VIN}$$

where K is a constant that is trimmed using an RC timer product for the 300 kHz, 600 kHz, and 1.0 MHz frequency options.

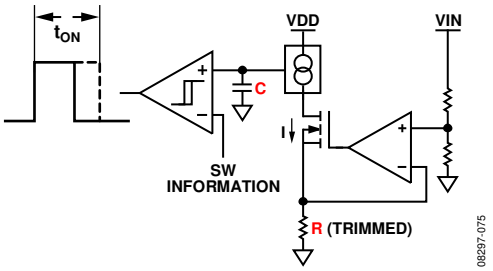

Figure 76. Constant On-Time Timer

The constant on-time (t_{ON}) is not strictly constant because it varies with VIN and V_{OUT}. However, this variation occurs in such a way as to keep the switching frequency virtually independent of VIN and V_{OUT}.

The t_{ON} timer uses a feedforward technique, applied to the constant on-time control loop, making it pseudo-fixed frequency to a first order. Second-order effects, such as dc losses in the external power MOSFETs (see the Efficiency Consideration section), cause some variation in frequency vs. load current and line voltage. These effects are shown in Figure 22 to Figure 33. The variations in frequency are much reduced compared with the variations generated when the feedforward technique is not used.

The feedforward technique establishes the following relationship:

$$f_{SW} = 1/K$$

where f_{SW} is the controller switching frequency (300 kHz, 600 kHz, and 1.0 MHz).

The t_{ON} timer senses VIN and V_{OUT} to minimize frequency variation with VIN and V_{OUT} as previously explained. This provides a pseudo-fixed frequency, see the Pseudo-Fixed Frequency section for additional information. To allow headroom for VIN/V_{OUT} sensing, the following two equations must be adhered to. For typical applications where V_{DD} is 5 V, these equations are not relevant; however, for lower V_{DD}, care may be required.

$$V_{DD} \geq VIN/8 + 1.5$$

$$V_{DD} \geq V_{OUT}/4$$

PSEUDO-FIXED FREQUENCY

The ADP1872/ADP1873 employ a constant on-time control scheme. During steady state operation, the switching frequency stays relatively constant, or pseudo-fixed. This is due to the one-shot t_{ON} timer that produces a high-side PWM pulse with a fixed duration, given that external conditions such as input voltage, output voltage, and load current are also at steady state. During load transients, the frequency momentarily changes for the duration of the transient event so that the output comes back within regulation quicker than if the frequency were fixed or if it were to remain unchanged. After the transient event is complete, the frequency returns to a pseudo-fixed value to a first-order.

To illustrate this feature more clearly, this section describes one such load transient event—a positive load step—in detail. During load transient events, the high-side driver output pulse width stays relatively consistent from cycle to cycle; however, the off-time (DRVL on-time) dynamically adjusts according to the instantaneous changes in the external conditions mentioned.

When a positive load step occurs, the error amplifier (out of phase of the output, V_{OUT}) produces new voltage information at its output (COMP). In addition, the current-sense amplifier senses new inductor current information during this positive load transient event. The error amplifier's output voltage reaction is compared to the new inductor current information that sets the start of the next switching cycle. Because current information is produced from valley current sensing, it is sensed at the down ramp of the inductor current, whereas the voltage loop information is sensed through the counter action upswing of the error amplifier's output (COMP).

The result is a convergence of these two signals (see Figure 77), which allows an instantaneous increase in switching frequency during the positive load transient event. In summary, a positive load step causes V_{OUT} to transient down, which causes COMP to transient up and therefore shortens the off time. This resulting increase in frequency during a positive load transient helps to quickly bring V_{OUT} back up in value and within the regulation window.

Similarly, a negative load step causes the off time to lengthen in response to V_{OUT} rising. This effectively increases the inductor demagnetizing phase, helping to bring V_{OUT} to within regulation. In this case, the switching frequency decreases, or experiences a foldback, to help facilitate output voltage recovery.

Because the ADP1872/ADP1873 has the ability to respond rapidly to sudden changes in load demand, the recovery period in which the output voltage settles back to its original steady state operating point is much quicker than it would be for a fixed-frequency equivalent. Therefore, using a pseudo-fixed frequency, results in significantly better load transient performance than using a fixed frequency.

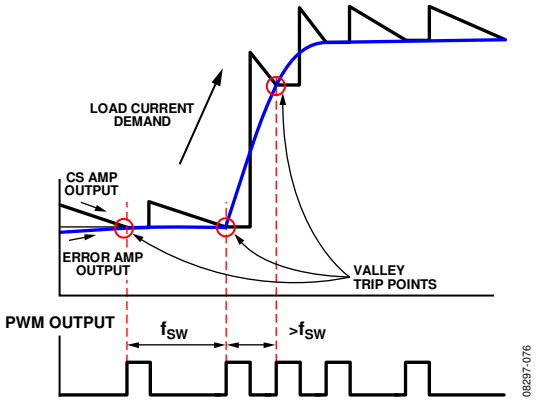

Figure 77. Load Transient Response Operation

APPLICATIONS INFORMATION

FEEDBACK RESISTOR DIVIDER

The required resistor divider network can be determine for a given V_{OUT} value because the internal band gap reference (V_{REF}) is fixed at 0.6 V. Selecting values for R_T and R_B determines the minimum output load current of the converter. Therefore, for a given value of R_B, the R_T value can be determined by

$$R_T = R_B \times \frac{(V_{OUT} - 0.6\text{ V})}{0.6\text{ V}}$$

INDUCTOR SELECTION

The inductor value is inversely proportional to the inductor ripple current. The peak-to-peak ripple current is given by

$$\Delta I_L = K_I \times I_{LOAD} \approx \frac{I_{LOAD}}{3}$$

where K_I is typically 0.33.

The equation for the inductor value is given by

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

where:
VIN is the high voltage input.
V_{OUT} is the desired output voltage.
f_{SW} is the controller switching frequency (300 kHz, 600 kHz, and 1.0 MHz).

When selecting the inductor, choose an inductor saturation rating that is above the peak current level and then calculate the inductor current ripple (see the Valley Current-Limit Setting section and Figure 78).

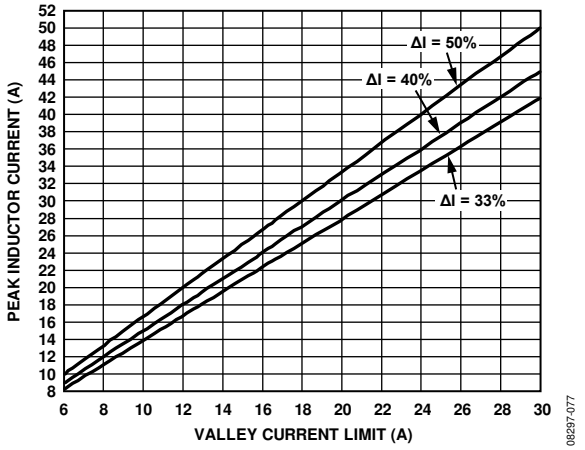

Figure 78. Peak Current vs. Valley Current Threshold for 33%, 40%, and 50% of Inductor Ripple Current

Table 7. Recommended Inductors

L (µH)	DCR (mΩ)	I_{SAT} (A)	Dimensions (mm)	Manufacturer	Model No.
0.12	0.33	55	10.2 × 7	Würth Elektronic	744303012
0.22	0.33	30	10.2 × 7	Würth Elektronic	744303022
0.47	0.8	50	14.2 × 12.8	Würth Elektronic	744355147
0.72	1.65	35	10.5 × 10.2	Würth Elektronic	744325072
0.9	1.6	28	13 × 12.8	Würth Elektronic	744355090
1.2	1.8	25	10.5 × 10.2	Würth Elektronic	744325120
1.0	3.3	20	10.5 × 10.2	Würth Elektronic	7443552100
1.4	3.2	24	14 × 12.8	Würth Elektronic	744318180
2.0	2.6	22	13.2 × 12.8	Würth Elektronic	7443551200
0.8		27.5		Sumida	CEP125U-0R8

OUTPUT RIPPLE VOLTAGE (ΔV_{RR})

The output ripple voltage is the ac component of the dc output voltage during steady state. For a ripple error of 1.0%, the output capacitor value needed to achieve this tolerance can be determined using the following equation. (Note that an accuracy of 1.0% is only possible during steady state conditions, not during load transients.)

$$\Delta V_{RR} = (0.01) \times V_{OUT}$$

OUTPUT CAPACITOR SELECTION

The primary objective of the output capacitor is to facilitate the reduction of the output voltage ripple; however, the output capacitor also assists in the output voltage recovery during load transient events. For a given load current step, the output voltage ripple generated during this step event is inversely proportional to the value chosen for the output capacitor. The speed at which the output voltage settles during this recovery period depends on where the crossover frequency (loop bandwidth) is set. This crossover frequency is determined by the output capacitor, the equivalent series resistance (ESR) of the capacitor, and the compensation network.

To calculate the small signal voltage ripple (output ripple voltage) at the steady state operating point, use the following equation:

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

where *ESR* is the equivalent series resistance of the output capacitors.

To calculate the output load step, use the following equation:

$$C_{OUT} = 2 \times \frac{\Delta I_{LOAD}}{f_{SW} \times (\Delta V_{DROOP} - (\Delta I_{LOAD} \times ESR))}$$

where ΔV_{DROOP} is the amount that V_{OUT} is allowed to deviate for a given positive load current step (ΔI_{LOAD}).

Ceramic capacitors are known to have low ESR. However, the trade-off of using X5R technology is that up to 80% of its capacitance may be lost due to derating because the voltage applied across the capacitor is increased (see Figure 79). Although X7R series capacitors can also be used, the available selection is limited to only up to 22 µF.

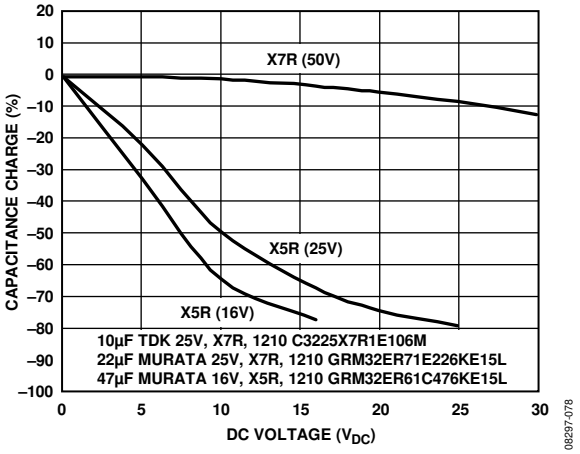

Figure 79. Capacitance vs. DC Voltage Characteristics for Ceramic Capacitors

Electrolytic capacitors satisfy the bulk capacitance requirements for most high current applications. Because the ESR of electrolytic capacitors is much higher than that of ceramic capacitors, when using electrolytic capacitors, several MLCCs should be mounted in parallel to reduce the overall series resistance.

COMPENSATION NETWORK

Due to its current-mode architecture, the ADP1872/ADP1873 require Type II compensation. To determine the component values needed for compensation (resistance and capacitance values), it is necessary to examine the converter's overall loop gain (H) at the unity gain frequency ($f_{SW}/10$) when H = 1 V/V.

$$H = 1\ \text{V/V} = G_M \times G_{CS} \times \frac{V_{OUT}}{V_{REF}} \times Z_{COMP} \times Z_{FILT}$$

Examining each variable at high frequency enables the unity gain transfer function to be simplified to provide expressions for the R_{COMP} and C_{COMP} component values.

Output Filter Impedance (Z_{FILT})

Examining the filter's transfer function at high frequencies simplifies to

$$Z_{FILTER} = \frac{1}{sC_{OUT}}$$

at the crossover frequency ($s = 2\pi f_{CROSS}$).

Error Amplifier Output Impedance (Z_{COMP})

Assuming C_{C2} is significantly smaller than C_{COMP}, C_{C2} can be omitted from the output impedance equation of the error amplifier. The transfer function simplifies to

$$Z_{COMP} = \frac{R_{COMP}(f_{CROSS} + f_{ZERO})}{f_{CROSS}}$$

and

$$f_{CROSS} = \frac{1}{12} \times f_{SW}$$

where f_{ZERO}, the zero frequency, is set to be 1/4th of the crossover frequency for the ADP1872.

Error Amplifier Gain (G_M)

The error amplifier gain (transconductance) is

G_M = 500 µA/V

Current-Sense Loop Gain (G_{CS})

The current-sense loop gain is

$$G_{CS} = \frac{1}{A_{CS} \times R_{ON}}\ \text{(A/V)}$$

where:
A_{CS} (V/V) is programmable for 3 V/V, 6 V/V, 12 V/V, and 24 V/V (see the Programming Resistor (RES) Detect Circuit and Valley Current-Limit Setting sections).
R_{ON} is the channel impedance of the lower side MOSFET.

Crossover Frequency

The crossover frequency is the frequency at which the overall loop (system) gain is 0 dB (H = 1 V/V). It is recommended for current-mode converters, such as the ADP1872, that the user set the crossover frequency between 1/10th and 1/15th of the switching frequency.

$$f_{CROSS} = \frac{1}{12} f_{SW}$$

The relationship between C_{COMP} and f_{ZERO} (zero frequency) is

$$f_{ZERO} = \frac{1}{2\pi \times R_{COMP} \times C_{COMP}}$$

The zero frequency is set to 1/4th of the crossover frequency.

Combining all of the above parameters results in

$$R_{COMP} = \frac{f_{CROSS}}{f_{CROSS} + f_{ZERO}} \times \frac{2\pi f_{CROSS} C_{OUT}}{G_M G_{CS}} \times \frac{V_{OUT}}{V_{REF}}$$

$$C_{COMP} = \frac{1}{2 \times \pi \times R_{COMP} \times f_{ZERO}}$$

EFFICIENCY CONSIDERATION

One of the important criteria to consider in constructing a dc-to-dc converter is efficiency. By definition, efficiency is the ratio of the output power to the input power. For high power applications at load currents up to 20 A, the following are important MOSFET parameters that aid in the selection process:

- $V_{GS\ (TH)}$: the MOSFET support voltage applied between the gate and the source.
- $R_{DS\ (ON)}$: the MOSFET on resistance during channel conduction.
- Q_G: the total gate charge
- C_{N1}: the input capacitance of the upper side switch
- C_{N2}: the input capacitance of the lower side switch

The following are the losses experienced through the external component during normal switching operation:

- Channel conduction loss (both the MOSFETs)
- MOSFET driver loss
- MOSFET switching loss
- Body diode conduction loss (lower side MOSFET)
- Inductor loss (copper and core loss)

Channel Conduction Loss

During normal operation, the bulk of the loss in efficiency is due to the power dissipated through MOSFET channel conduction. Power loss through the upper side MOSFET is directly proportional to the duty cycle (D) for each switching period, and the power loss through the lower side MOSFET is directly proportional to 1 − D for each switching period. The selection of MOSFETs is governed by the amount of maximum dc load current that the converter is expected to deliver. In particular, the selection of the lower side MOSFET is dictated by the maximum load current because a typical high current application employs duty cycles of less than 50%. Therefore, the lower side MOSFET is in the on state for most of the switching period.

$$P_{N1,\ N2\ (CL)} = [D \times R_{N1\ (ON)} + (1 - D) \times R_{N2\ (ON)}] \times I_{LOAD}^2$$

MOSFET Driver Loss

Other dissipative elements are the MOSFET drivers. The contributing factors are the dc current flowing through the driver during operation and the Q_{GATE} parameter of the external MOSFETs.

$$P_{DR(LOSS)} = \left[V_{DR} \times \left(f_{SW} C_{upperFET} V_{DR} + I_{BIAS}\right)\right] + \left[V_{DD} \times \left(f_{SW} C_{lowerFET} V_{DD} + I_{BIAS}\right)\right]$$

where:
$C_{upperFET}$ is the input gate capacitance of the upper-side MOSFET.
$C_{lowerFET}$ is the input gate capacitance of the lower-side MOSFET.
V_{DR} is the driver bias voltage (that is, the low input voltage (V_{DD}) minus the rectifier drop (see Figure 80)).
I_{BIAS} is the dc current flowing into the upper- and lower-side drivers.
V_{DD} is the bias voltage.

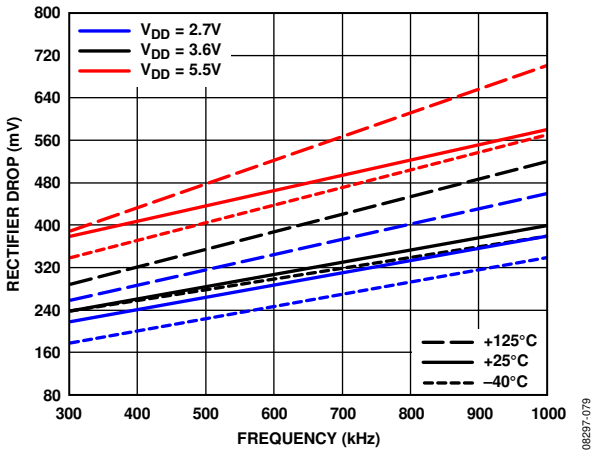

Figure 80. Internal Rectifier Voltage Drop vs. Switching Frequency

MOSFET Switching Loss

The SW node transitions due to the switching activities of the upper side and lower side MOSFETs. This causes removal and replenishing of charge to and from the gate oxide layer of the MOSFET, as well as to and from the parasitic capacitance associated with the gate oxide edge overlap and the drain and source terminals. The current that enters and exits these charge paths presents additional loss during these transition times. This can be approximately quantified by using the following equation, which represents the time in which charge enters and exits these capacitive regions.

$$t_{SW\text{-}TRANS} = R_{GATE} \times C_{TOTAL}$$

where:
R_{GATE} is the gate input resistance of the MOSFET.
C_{TOTAL} is the C_{GD} + C_{GS} of the external MOSFET used.

The ratio of this time constant to the period of one switching cycle is the multiplying factor to be used in the following expression:

$$P_{SW(LOSS)} = \frac{t_{SW\text{-}TRANS}}{t_{SW}} \times I_{LOAD} \times VIN \times 2$$

or

$$P_{SW\ (LOSS)} = f_{SW} \times R_{GATE} \times C_{TOTAL} \times I_{LOAD} \times VIN \times 2$$