THE DATASHEET OF

LTC3866IUF#TRPBF

Current Mode Synchronous Controller for Sub Milliohm DCR Sensing

FEATURES

- **Sub Milliohm DCR Current Sensing**
- **High Efficiency: Up to 95%**
- **Selectable Current Sensing Limit**
- **Programmable DCR Temperature Compensation**
- **Die Overtemperature Thermal Shutdown**
- **± 0.5% 0.6V Output Voltage Accuracy**
- **Programmable Fixed Frequency 250kHz to 770kHz**
- **High Speed Differential Remote Sense Amplifier**
- Wide Input Voltage Range: 4.5V to 38V
- Output Voltage Range: 0.6V to 3.5V with Diffamp
- Adjustable Soft-Start or Output Voltage Tracking
- Foldback Output Current Limit
- Short-Circuit Soft Recovery
- Output Overvoltage Protection
- 24-Lead (4mm × 4mm) QFN and 24-Lead FE Packages

APPLICATIONS

- Computer Systems
- Telecom Systems
- Industrial and Medical Instruments
- DC Power Distribution Systems

DESCRIPTION

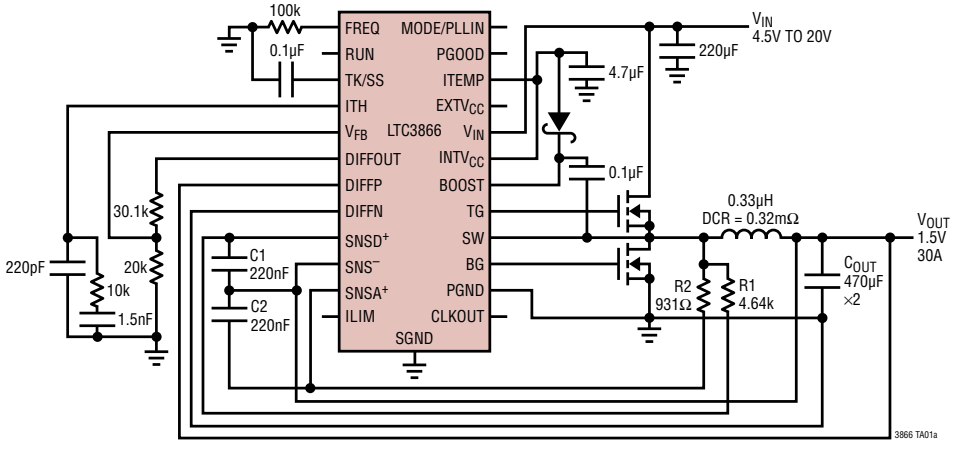
The LTC®3866 is a single phase current mode synchronous step-down switching regulator controller that drives all N-channel power MOSFET switches. It employs a unique architecture which enhances the signal-to-noise ratio of the current sense signal, allowing the use of a very low DC resistance power inductor to maximize the efficiency in high current applications. This feature also reduces the switching jitter commonly found in low DCR applications. The LTC3866 also includes a high speed remote sense differential amplifier, a programmable current sense limit that can be selected to 10mV, 15mV, 20mV, 25mV or 30mV, and DCR temperature compensation to limit the maximum output current precisely over temperature.

The LTC3866 also features a precise 0.6V reference with a guaranteed limit of ±0.5% that provides an accurate output voltage from 0.6V to 3.5V. A 4.5V to 38V input voltage range allows it to support a wide variety of bus voltages and various types of batteries.

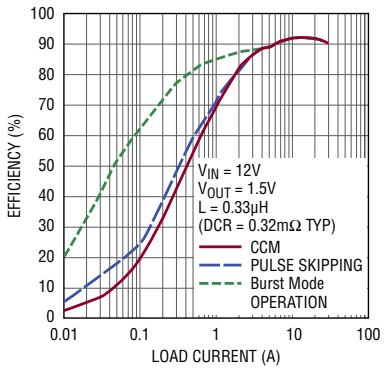
The LTC3866 is offered in a low profile 24-lead 4mm × 4mm QFN and 24-lead exposed pad FE packages.

TYPICAL APPLICATION

High Efficiency, 1.5V/30A Step-Down Converter with Very Low DCR Sensing



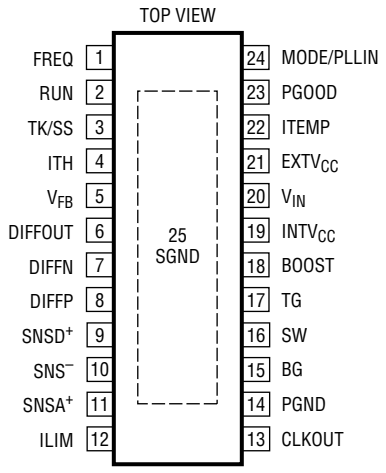
Efficiency vs Load Current and Mode



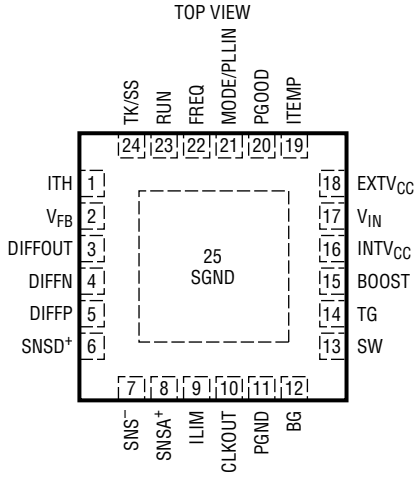
ABSOLUTE MAXIMUM RATINGS (Note 1)

Input Supply Voltage ..... –0.3V to 40V
Topside Driver Voltage (BOOST) ..... –0.3V to 46V
Switch Voltage(SW) ..... –5V to 40V
$INTV_{CC}$, $EXTV_{CC}$, RUN, PGOOD,
BOOST-SW Voltages ..... –0.3V to 6V
$SNSD^+$, $SNSA^+$, SNS^- Voltages ..... –0.3V to $INTV_{CC}$
MODE/PLLIN, ILIM, TK/SS, FREQ,
DIFFOUT Voltages ..... –0.3V to $INTV_{CC}$
DIFFP, DIFFN ..... –0.3V to $INTV_{CC}$
ITEMP, ITH, V_{FB} Voltages ..... –0.3V to $INTV_{CC}$
$INTV_{CC}$ Peak Output Current 100mA
Operating Junction Temperature Range
(Notes 2, 4) ..... –40°C to 125°C
Storage Temperature Range ..... –65°C to 125°C
Lead Temperature (Soldering, 10 sec)
FE Package ..... 300°C

PIN CONFIGURATION



FE PACKAGE
24-LEAD PLASTIC TSSOP
θ_{JA} = 33°C/W, θ_{JC} = 10°C/W
EXPOSED PAD (PIN 25) IS SGND, MUST BE SOLDERED TO PCB



UF PACKAGE
24-LEAD (4mm × 4mm) PLASTIC QFN
θ_{JA} = 47°C/W, θ_{JC} = 4.5°C/W
EXPOSED PAD (PIN 25) IS SGND, MUST BE SOLDERED TO PCB

ORDER INFORMATION http://www.linear.com/product/LTC3866#orderinfo

LEAD FREE FINISH	TAPE AND REEL	PART MARKING*	PACKAGE DESCRIPTION	TEMPERATURE RANGE
LTC3866EFE#PBF	LTC3866EFE#TRPBF	LTC3866FE	24-Lead Plastic TSSOP	–40°C to 125°C
LTC3866IFE#PBF	LTC3866IFE#TRPBF	LTC3866FE	24-Lead Plastic TSSOP	–40°C to 125°C
LTC3866EUF#PBF	LTC3866EUF#TRPBF	3866	24-Lead (4mm × 4mm) Plastic QFN	–40°C to 125°C
LTC3866IUF#PBF	LTC3866IUF#TRPBF	3866	24-Lead (4mm × 4mm) Plastic QFN	–40°C to 125°C

Consult LTC Marketing for parts specified with wider operating temperature ranges. *The temperature grade is identified by a label on the shipping container.

For more information on lead free part marking, go to: http://www.linear.com/leadfree/

For more information on tape and reel specifications, go to: http://www.linear.com/tapeandreel/. Some packages are available in 500 unit reels through designated sales channels with #TRMPBF suffix.

ELECTRICAL CHARACTERISTICS

The ● denotes the specifications which apply over the specified operating junction temperature range, otherwise specifications are at T_A = 25°C (Note 2). V_{IN} = 15V, V_{RUN} = 5V unless otherwise specified.

SYMBOL	PARAMETER	CONDITIONS		MIN	TYP	MAX	UNITS
Main Control Loops							
V_{IN}	Input Voltage Range			4.5		38	V
V_{OUT}	Output Voltage Range	with Diffamp in Loop		0.6		3.5	V
V_{FB}	Regulated Feedback Voltage	Current ITH Voltage = 1.2V (Note 5) –40°C to 85°C –40°C to 125°C	 ● ●	 0.597 0.5955	 0.6 0.6	 0.603 0.6045	 V V
I_{FB}	Feedback Current	(Note 5)			–15	–50	nA
$V_{REFLNREG}$	Reference Voltage Line Regulation	V_{IN} = 4.5V to 38V (Note 5)			0.002	0.02	%
$V_{LOADREG}$	Output Voltage Load Regulation	(Note 5) Measured in Servo Loop; ΔITH Voltage = 1.2V to 0.7V Measured in Servo Loop; ΔITH Voltage = 1.2V to 1.6V	 ● ●		 0.01 0.01	 0.1 0.1	 % %
g_m	Error Amplifier (EA) Transconductance	ITH =1.2V, Sink/Source 5µA (Note 5)			2		mmho
I_Q	Input DC Supply Current Normal Mode Shutdown	(Note 6) V_{IN} = 15V V_{IN} = 15V, V_{RUN} = 0V			 3.2 30	 50	 mA µA
UVLO	Undervoltage Lockout	V_{INTVCC} Ramping Down		3.4	3.75	4.1	V
$UVLO_{HYS}$	UVLO Hysteresis Voltage				0.5		V
V_{FBOVL}	Feedback Overvoltage Lockout	Measured at V_{FB}	●	0.64	0.66	0.68	V
I_{SNSD+}	$SNSD^+$ Pin Bias Current	V_{SNSD+} = 3.3V	●		30	100	nA
I_{SNSA+}	$SNSA^+$ Pin Bias Current	V_{SNSA+} = 3.3V	●		1	2	µA
A_{VT_SNS}	Total Sense Signal Gain to Current Comparator	(V_{SNSD+} + V_{SNSA+})/V_{SNSD+}			5		V/V
$V_{SENSE(MAX)}$	Maximum Current Sense Threshold	–40°C to 85°C $V_{SNS–}$ = 1.8V, ILIM = 0V ILIM = 1/4 V_{INTVCC} ILIM = 1/2 V_{INTVCC} or Float ILIM = 3/4 V_{INTVCC} ILIM = V_{INTVCC} –40°C to 125°C $V_{SNS–}$ = 1.8V, ILIM = 0V ILIM = 1/4V_{INTVCC} ILIM = 1/2V_{INTVCC} or Float ILIM = 3/4V_{INTVCC} ILIM = V_{INTVCC}	 ● ● ● ● ● ● ● ● ● ●	 9.2 14.2 19.2 23.5 28.5 9 14 19 23.5 28.5	 10 15 20 25 30 10 15 20 25 30	 10.8 15.8 20.8 26.5 31.5 11 16 21 26.5 31.5	 mV mV mV mV mV mV mV mV mV mV
I_{TEMP}	DCR Temperature Compensation Current	V_{ITEMP} = 0.3V	●	9	10	11	µA
$I_{TK/SS}$	Soft-Start Charge Current	$V_{TK/SS}$ = 0V	●	1.0	1.25	1.5	µA
V_{RUN}	RUN Pin On Threshold Voltage	V_{RUN} Rising	●	1.1	1.22	1.35	V
$V_{RUN(HYS)}$	RUN Pin On Hysteresis Voltage				80		mV
TG t_r t_f	Top Gate (TG) Transition Time Rise Time Fall Time	(Note 7) C_{LOAD} = 3300pF C_{LOAD} = 3300pF			 25 25		 ns ns
BG t_r t_f	Bottom Gate (BG) Transition Time Rise Time Fall Time	(Note 7) C_{LOAD} = 3300pF C_{LOAD} = 3300pF			 25 25		 ns ns

ELECTRICAL CHARACTERISTICS

The ● denotes the specifications which apply over the specified operating junction temperature range, otherwise specifications are at T_A = 25°C (Note 2). V_{IN} = 15V, V_{RUN} = 5V unless otherwise specified.

SYMBOL	PARAMETER	CONDITIONS		MIN	TYP	MAX	UNITS
TG/BG t_D	Top Gate Off to Bottom Gate On Delay, Synchronous Switch-On Delay Time	C_{LOAD} = 3300pF			30		ns
BG/TG t_D	Bottom Gate Off to Top Gate On Delay, Top Switch-On Delay Time	C_{LOAD} = 3300pF			30		ns
$t_{ON(MIN)}$	Minimum On-Time	(Note 8)			90		ns
$INTV_{CC}$ Linear Regulator							
V_{INTVCC}	Internal V_{CC} Voltage	6V < V_{IN} < 38V		5.25	5.5	5.75	V
	Load Regulation	I_{INTVCC} = 0mA to 20mA			0.5	2	%
V_{EXTVCC}	External V_{CC} Switchover Voltage	$EXTV_{CC}$ Ramping Positive		4.5	4.7		V
	$EXTV_{CC}$ Voltage Drop	I_{EXTVCC} = 20mA, V_{EXTVCC} = 5V			50	100	mV
	$EXTV_{CC}$ Hysteresis				200		mV
Oscillator and Phase-Locked Loop							
f_{NOM}	Nominal Frequency	V_{FREQ} = 1.2V		450	500	550	kHz
f_{LOW}	Lowest Frequency	V_{FREQ} = 0.4V		225	250	275	kHz
f_{HIGH}	Highest Frequency	V_{FREQ} > 2.4V		700	770	850	kHz
$R_{MODE/PLLIN}$	MODE/PLLIN Input Resistance				250		kΩ
I_{FREQ}	Frequency Setting Current			9	10	11	µA
CLKOUT	Phase Relative to the Oscillator Clock				180		Deg
$CLKOUT_{HI}$	Clock Output High Voltage	V_{INTVCC} = 5.5V		4.5	5.5		V
$CLKOUT_{LO}$	Clock Output Low Voltage				0	0.2	V
PGOOD Output							
V_{PGDLO}	PGOOD Voltage Low	I_{PGOOD} = 2mA			0.1	0.3	V
I_{PGD}	PGOOD Leakage Current	V_{PGOOD} = 5.5V				2	µA
V_{PGD}	PGOOD Trip	V_{FB} with Respect to Set Output Voltage V_{FB} Going Negative V_{FB} Going Positive			 –10 10		 % %
Differential Amplifier							
A_V	Gain	–40°C to 85°C –40°C to 125°C	● ●	0.999 0.998	1 1	1.001 1.002	V/V V/V
R_{IN}	Input Resistance	Measured at DIFFP Input			80		kΩ
V_{OS}	Input Offset Voltage	V_{DIFFP} = 1.5V, $V_{DIFFOUT}$ = 100µA				2	mV
PSRR	Power Supply Rejection Ratio	5V < V_{IN} < 38V			90		dB
I_{OUT}	Maximum Sourcing Output Current			1.5	3		mA
V_{OUT}	Maximum Output Voltage	V_{INTVCC} = 5.5V, $I_{DIFFOUT}$ = 300µA		V_{INTVCC} – 1.4	V_{INTVCC} – 1.1		V
GBW	Gain-Bandwidth Product	(Note 9)			3		MHz
SR	Slew Rate	(Note 9)			2		V/µs

ELECTRICAL CHARACTERISTICS

The ● denotes the specifications which apply over the specified operating junction temperature range, otherwise specifications are at T_A = 25°C (Note 2). V_{IN} = 15V, V_{RUN} = 5V unless otherwise specified.

SYMBOL	PARAMETER	CONDITIONS		MIN	TYP	MAX	UNITS
On-Chip Driver							
TG R_{UP}	TG Pull-Up $R_{DS(ON)}$	TG High			2.6		Ω
TG R_{DOWN}	TG Pull-Down $R_{DS(ON)}$	TG Low			1.5		Ω
BG R_{UP}	BG Pull-Up $R_{DS(ON)}$	BG High			2.4		Ω
BG R_{DOWN}	BG Pull-Down $R_{DS(ON)}$	BG Low			1.1		Ω

Note 1: Stresses beyond those listed under Absolute Maximum Ratings may cause permanent damage to the device. Exposure to any Absolute Maximum Rating condition for extended periods may affect device reliability and lifetime.

Note 2: The LTC3866 is tested under pulsed load conditions such that $T_J \approx T_A$. The LTC3866E is guaranteed to meet performance specifications from 0°C to 85°C operating junction temperature. Specifications over the –40°C to 125°C operating junction temperature range are assured by design, characterization and correlation with statistical process controls. The LTC3866I is guaranteed to meet performance specifications over the full –40°C to 125°C operating junction temperature range. The maximum ambient temperature consistent with these specifications is determined by specific operating conditions in conjunction with board layout, the package thermal impedance and other environmental factors.

Note 3: The junction temperature, T_J, is calculated from the ambient temperature, T_A, and power dissipation, P_D, according to the following formula:

LTC3866FE: $T_J = T_A + (P_D \bullet 33°C/W)$
LTC3866UF: $T_J = T_A + (P_D \bullet 47°C/W)$

Note 4: This IC includes overtemperature protection that is intended to protect the device during momentary overload conditions. The maximum rated junction temperature will be exceeded when this protection is active. Continuous operation above the absolute maximum operating junction temperature may impair device reliability or permanently damage the device.

Note 5: The LTC3866 is tested in a feedback loop that servos V_{ITH} to a specified voltage and measures the resultant V_{FB}.

Note 6: Dynamic supply current is higher due to the gate charge being delivered at the switching frequency. See Applications Information.

Note 7: Rise and fall times are measured using 10% and 90% levels. Delay times are measured using 50% levels.

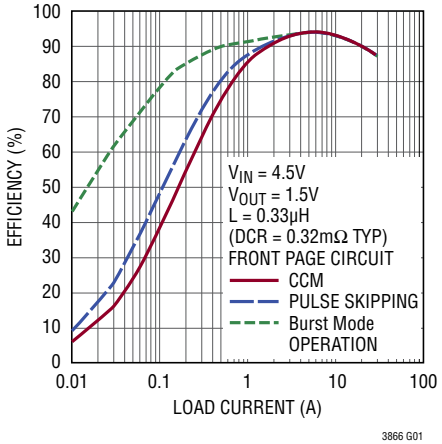
Note 8: The minimum on-time condition corresponds to the on inductor peak-to-peak ripple current ≥40% of I_{MAX} (see Minimum On-Time Considerations in the Applications Information section).

Note 9: Guaranteed by design.

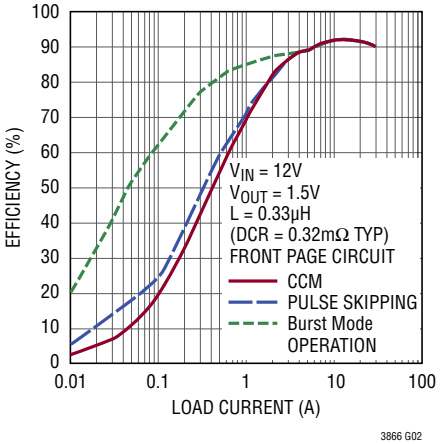
TYPICAL PERFORMANCE CHARACTERISTICS

T_A = 25°C, unless otherwise noted.

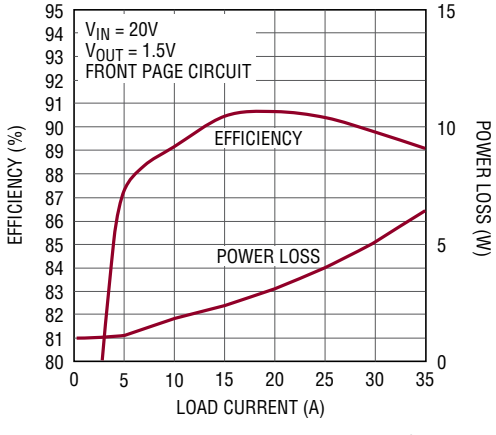
Efficiency vs Load Current and Mode



Efficiency vs Load Current and Mode

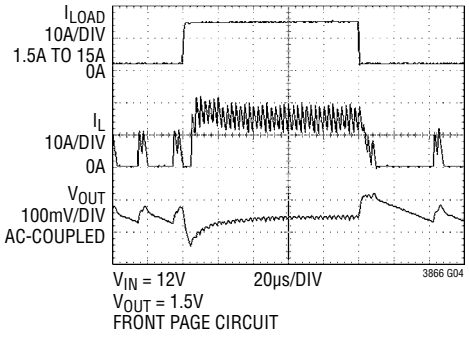


Efficiency and Power Loss vs Load Current

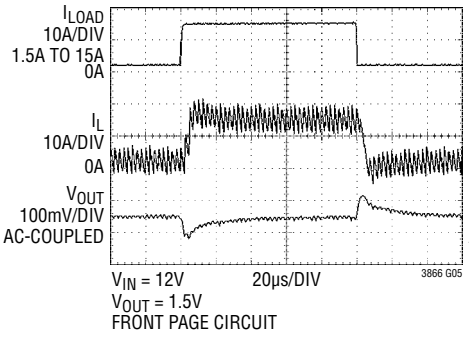


TYPICAL PERFORMANCE CHARACTERISTICS T_A = 25°C, unless otherwise noted.

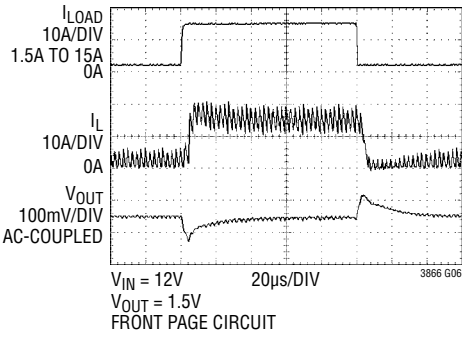
Load Step (Burst Mode® Operation)



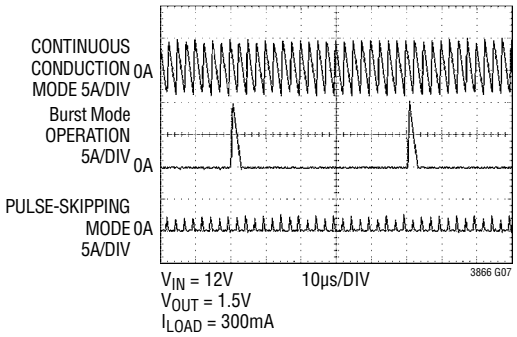
Load Step (Continuous Conduction Mode)



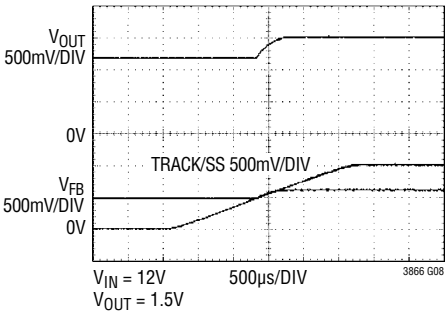
Load Step (Pulse-Skipping Mode)



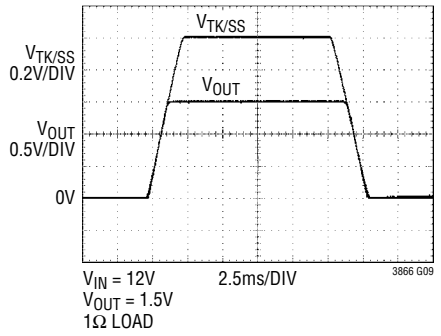
Inductor Current at Light Load



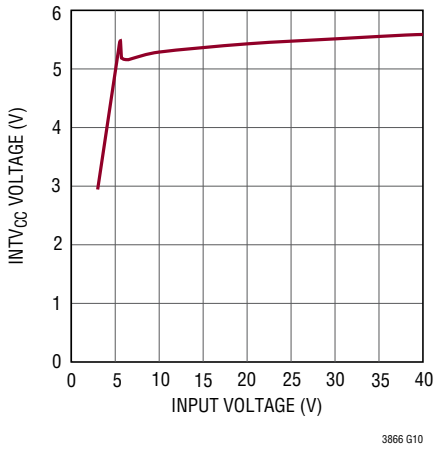
Prebiased Output at 1.2V



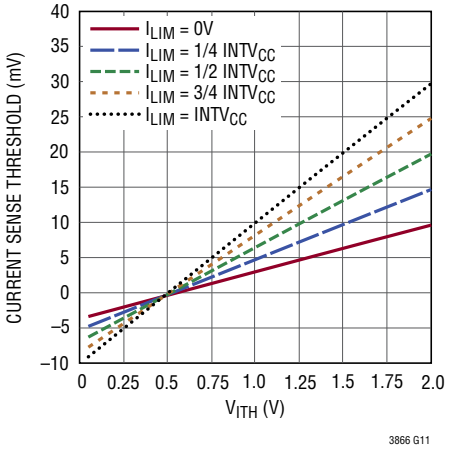
Tracking Up and Down with TK/SS External Ramp



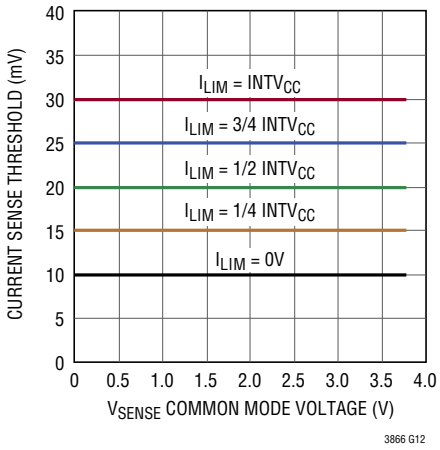
$INTV_{CC}$ Line Regulation



Current Sense Threshold vs ITH Voltage

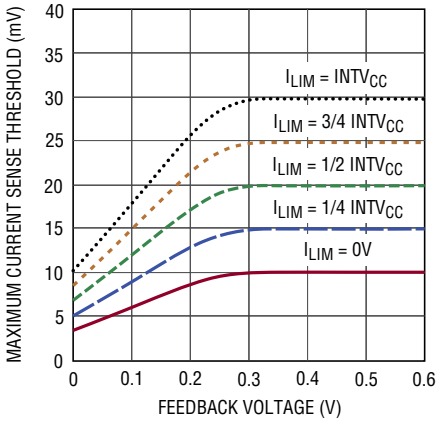


Maximum Current Sense Threshold vs Common Mode Voltage

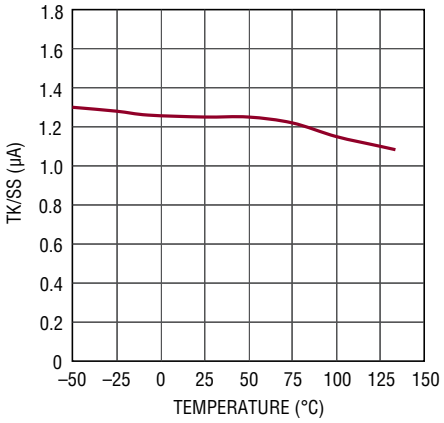


TYPICAL PERFORMANCE CHARACTERISTICS T_A = 25°C, unless otherwise noted.

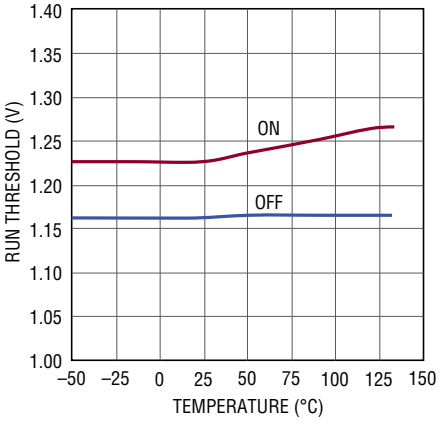
Maximum Current Sense Threshold Voltage vs Feedback Voltage (Current Foldback)



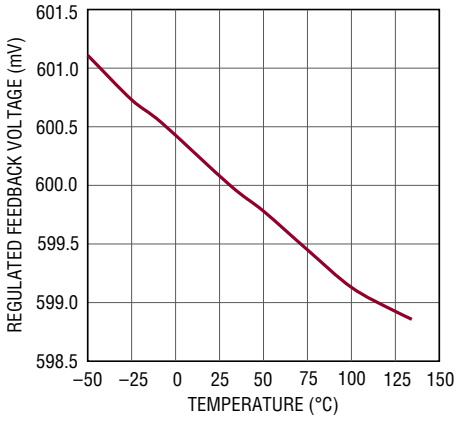
TK/SS Pull-Up Current vs Temperature



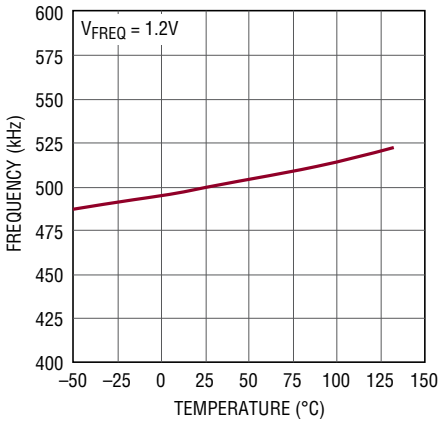
Shutdown (RUN) Threshold vs Temperature



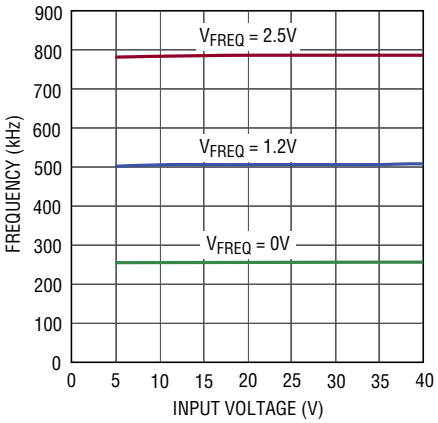
Regulated Feedback Voltage vs Temperature



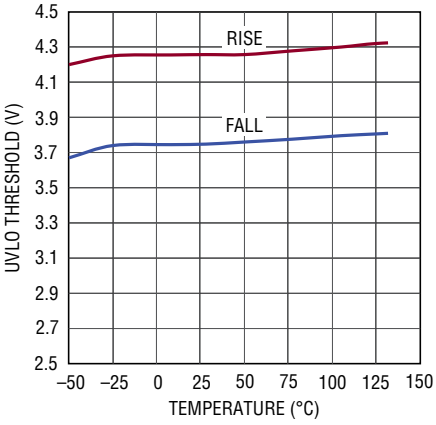
Oscillator Frequency vs Temperature



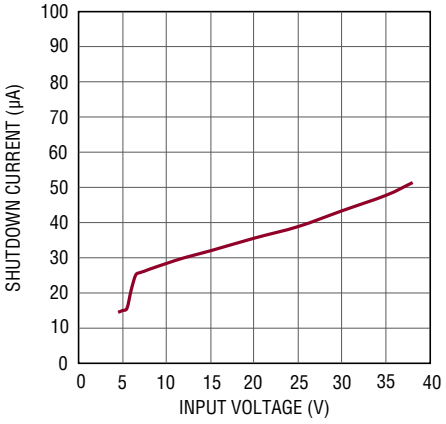
Oscillator Frequency vs Input Voltage



Undervoltage Lockout Threshold ($INTV_{CC}$) vs Temperature

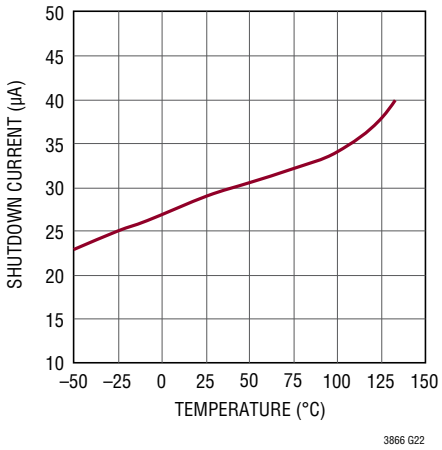


Shutdown Current vs Input Voltage

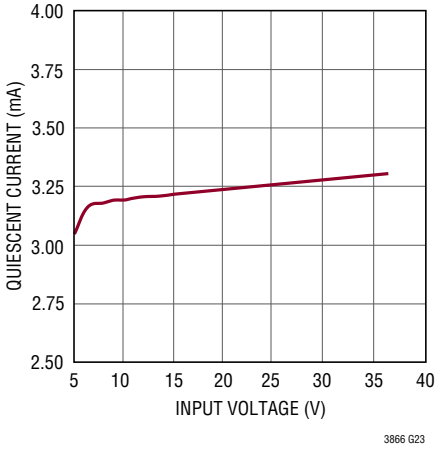


TYPICAL PERFORMANCE CHARACTERISTICS T_A = 25°C, unless otherwise noted.

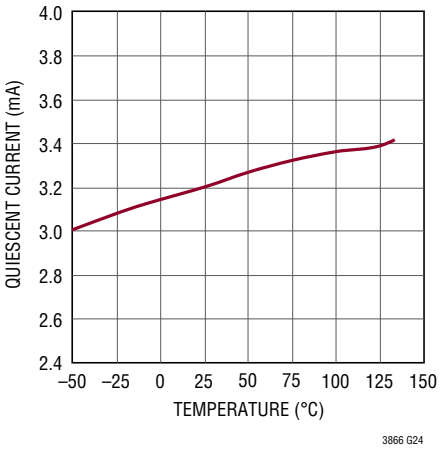
Shutdown Current vs Temperature



Input Quiescent Current vs Input Voltage without $EXTV_{CC}$



Quiescent Current vs Temperature without $EXTV_{CC}$



PIN FUNCTIONS (FE/UF)

FREQ (Pin 1/Pin 22): Oscillator Frequency Control Input. A 10µA current source flows out of this pin. Connecting a resistor between this pin and ground sets a DC voltage which in turn programs the oscillator frequency. Alternatively, this pin can be driven with a DC voltage to vary the frequency of the internal oscillator.

RUN (Pin 2/Pin 23): Run Control Input. A voltage above 1.22V turns on the IC. Pulling this pin below 1.14V causes the IC to shut down. There is a 1µA pull-up current for the pin. Once the RUN pin rises above 1.22V, an additional 4.5µA pull-up current is added to the pin.

TK/SS (Pin 3/Pin 24): Output Voltage Tracking and Soft-Start Input. An internal soft-start current of 1.25µA charges the external soft-start capacitor connected to this pin.

ITH (Pin 4/Pin 1): Current Control Threshold and Error Amplifier Compensation Pin. The current comparator tripping threshold is proportional with this voltage.

V_{FB} (Pin 5/Pin 2): Error Amplifier Feedback Input. This pin receives the remotely sensed feedback voltage to set the output voltage through an external resistive divider connected to the DIFFOUT pin or the output.

DIFFOUT (Pin 6/Pin 3): Output of Remote Sensing Differential Amplifier. Connect this pin to V_{FB} through a resistive divider to set the desired output voltage.

DIFFN (Pin 7/Pin 4): Negative Input of Remote Sensing Differential Amplifier. Connect this pin close to the ground of the output load.

DIFFP (Pin 8/Pin 5): Positive Input of Remote Sensing Differential Amplifier. Connect this pin close to the output load.

$SNSD^+$ (Pin 9/Pin 6): First Positive Current Sense Input. This pin is connected to sense the signal of the output inductor's DCR, it is to be used with a filter that matches the bandwidth, L/DCR, of the inductor.

SNS^- (Pin 10/Pin 7): Negative Current Sense Input. This negative input of the current comparator is to be connected to the output.

$SNSA^+$ (Pin 11/Pin 8): Second Positive Current Sense Input. This input is to be connected to sense the signal of the output's inductor DCR with a filter bandwidth of five times larger than L/DCR.

ILIM (Pin 12/Pin 9): Current Comparator Sense Voltage Limit. Apply a DC voltage to set the maximum current sense threshold for the current comparator.

CLKOUT (Pin 13/Pin 10): Clock Output Pin. The CLKOUT signal is 180° out of phase to the rising edge of the IC internal clock.

PIN FUNCTIONS (FE/UF)

PGND (Pin 14/Pin 11): Power Ground. Connect to the source of the bottom N-channel MOSFET and the negative terminals of the V_{IN} and $INTV_{CC}$ decoupling capacitors close to this pin.

BG (Pin 15/Pin 12): Bottom Gate Driver Output. This pin drives the gate of the bottom N-channel MOSFET and swings between $INTV_{CC}$ or $EXTV_{CC}$ and PGND.

SW (Pin 16/Pin 13): Switch Node Connection. Connect this pin to the output filter inductor, bottom N-channel MOSFET drain and top N-channel MOSFET source. Voltage swing at these pins is from a Schottky diode (external) voltage drop below ground to V_{IN}.

TG (Pin 17/Pin 14): Top Gate Driver Output. This is a floating driver to be connected to the gate of the top N-channel MOSFET. The voltage swing of this pin equals to $INTV_{CC}$ superimposed over the switch node (SW) voltage.

BOOST (Pin 18/Pin 15): Boosted Top Gate Driver Supply. The (+) terminal of the boostrap capacitor connects to this pin. This pins swings from a diode voltage drop below $INTV_{CC}$ up to V_{IN} + $INTV_{CC}$.

$INTV_{CC}$ (Pin 19/Pin 16): Internal 5.5V Regulator Output. The internal control circuits are powered from this voltage. Decouple this pin to PGND with a 4.7µF low ESR tantalum or ceramic capacitor.

V_{IN} (Pin 20/Pin 17): Main Input Supply. Decouple this pin to PGND with a capacitor (0.1µF to 1µF). For applications where the main input power is 5V, tie the V_{IN} and $INTV_{CC}$ pins together.

$EXTV_{CC}$ (Pin 21/Pin 18): External Supply Voltage Input. Whenever an external voltage supply greater than 4.7V is connected to this pin, an internal switch will close and bypass the internal low dropout regulator, and the external supply will power the IC. Do not exceed 6V on this pin and ensure $V_{IN} > V_{EXTVCC}$ at all times.

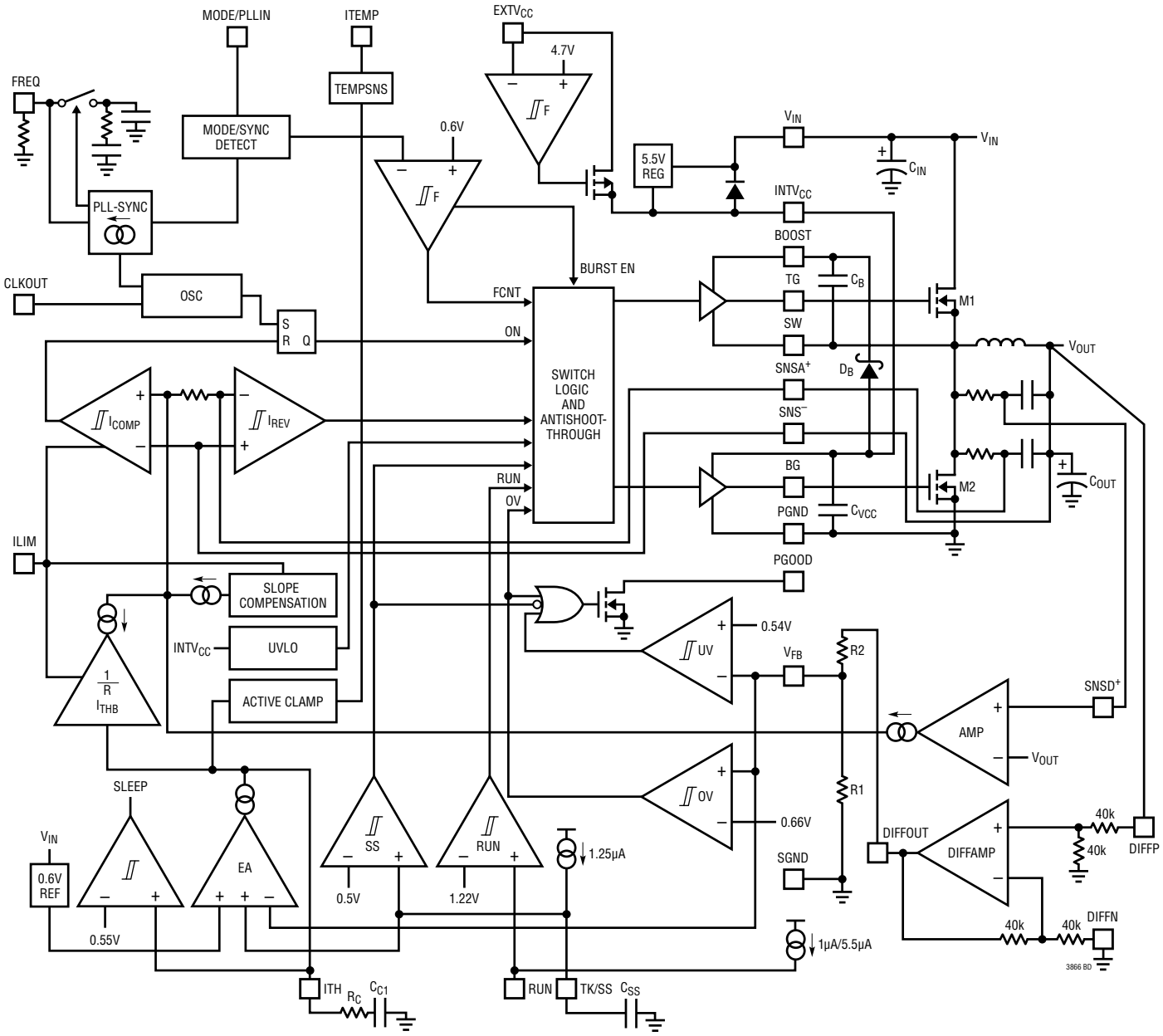
ITEMP (Pin 22/Pin 19): Temperature DCR Compensation Input. Connect to a NTC (negative tempco) resistor placed near the output inductor to compensate for its DCR change over temperature. Floating this pin or tying it to $INTV_{CC}$ disables the DCR temperature compensation function.

PGOOD (Pin 23/Pin 20): Power Good Indicator Output. Open-drain logic out that is pulled to ground when the output exceeds the 10% regulation window, after the internal 20µs power bad mask timer expires.

MODE/PLLIN (Pin24/Pin 21): Mode Operation or External Clock Synchronization. Connect this pin to SGND to set the continuous mode of operation. Connect to $INTV_{CC}$ to enable pulse-skipping mode of operation. Leaving the pin floating will enable Burst Mode operation. A clock signal applied to the pin will force the controller into continuous mode of operation and synchronizes the internal oscillator.

SGND (Exposed Pad Pin 25/ Exposed Pad Pin 25): Signal Ground. This is the ground of the controller. Connect compensation components and output setting resistors to this ground. The exposed pad must be soldered to the PCB ground plane.

FUNCTIONAL BLOCK DIAGRAM



OPERATION

Main Control Loop

The LTC3866 uses LTC proprietary current sensing, current mode step-down architecture. During normal operation, the top MOSFET is turned on every cycle when the oscillator sets the RS latch, and turned off when the main current comparator, I_{CMP}, resets the RS latch. The peak inductor current at which I_{CMP} resets the RS latch is controlled by the voltage on the ITH pin, which is the output of the error amplifier, EA. The remote sense amplifier (diffamp) produces a signal equal to the differential voltage sensed across the output capacitor divided down by the feedback divider and re-references it to the local IC ground reference. The V_{FB} pin receives this feedback signal and compares it to the internal 0.6V reference. When the load current increases, it causes a slight decrease in the V_{FB} pin voltage relative to the 0.6V reference, which in turn causes the ITH voltage to increase until the inductor's average current equals the new load current. After the top MOSFET has turned off, the bottom MOSFET is turned on until either the inductor current starts to reverse, as indicated by the reverse current comparator, I_{REV}, or the beginning of the next cycle.

The main control loop is shut down by pulling the RUN pin low. Releasing RUN allows an internal 1.0µA current source to pull up the RUN pin. When the RUN pin reaches 1.22V, the main control loop is enabled and the IC is powered up. When the RUN pin is low, all functions are kept in a controlled state.

Sensing Signal of Very Low DCR

The LTC3866 employs a unique architecture to enhance the signal-to-noise ratio that enables it to operate with a small sense signal of a very low value inductor DCR, 1mΩ or less, to improve power efficiency, and reduce jitter due to the switching noise which could corrupt the signal. The LTC3866 can sense a DCR value as low as 0.2mΩ with careful PCB layout.The LTC3866 comprises two positive sense pins, $SNSD^+$ and $SNSA^+$, to acquire signals and processes them internally to provide the response as with a DCR sense signal that has a 14dB signal-to-noise ratio improvement. In the meantime, the current limit threshold is still a function of the inductor peak current and its DCR value, and can be accurately set from 10mV to 30mV in a 5mV steps with the ILIM pin. The filter time constant, R1C1, of the $SNSD^+$ should match the L/DCR of the output inductor, while the filter at $SNSA^+$ should have a bandwidth of five times larger than $SNSD^+$, R2C2 equals R1C1/5.

$INTV_{CC}/EXTV_{CC}$ Power

Power for the top and bottom MOSFET drivers and most other internal circuitry is derived from the $INTV_{CC}$ pin. When the $EXTV_{CC}$ pin is left open or tied to a voltage less than 4.7V, an internal 5.5V linear regulator supplies $INTV_{CC}$ power from V_{IN}. If $EXTV_{CC}$ is taken above 4.7V, the 5.5V regulator is turned off and an internal switch is turned on connecting $EXTV_{CC}$ to $INTV_{CC}$. Using the $EXTV_{CC}$ pin allows the $INTV_{CC}$ power to be derived from a high efficiency external source such as a switching regulator output. The top MOSFET driver is biased from the floating bootstrap capacitor, C_B, which normally recharges during the off cycle through an external diode when the top MOSFET turns off. If the input voltage, V_{IN}, decreases to a voltage close to V_{OUT}, the loop may enter dropout and attempt to turn on the top MOSFET continuously. The dropout detector detects this and forces the top MOSFET off for about one-twelfth of the clock period plus 100ns every third cycle to allow C_B to recharge. However, it is recommended that a load be present or the IC operates at low frequency during the dropout transition to ensure C_B is recharged.

Internal Soft-Start

By default, the start-up of the output voltage is normally controlled by an internal soft-start ramp. The internal soft-start ramp connects to the noninverting input of the error amplifier. The FB pin is regulated to the lower of the error amplifier's three noninverting inputs (the internal soft-start ramp, the TK/SS pin or the internal 600mV reference). As the ramp voltage rises from 0V to 0.6V over approximately 600µs, the output voltage rises smoothly from its prebiased value to its final set value.

Certain applications can result in the start-up of the converter into a non-zero load voltage, where residual charge is stored on the output capacitor at the onset of converter switching. In order to prevent the output from discharging under these conditions, the bottom MOSFET is disabled until soft-start is greater than V_{FB}.

OPERATION

Shutdown and Start-Up (RUN and TK/SS Pins)

The LTC3866 can be shut down using the RUN pin. Pulling the RUN pin below 1.14V shuts down the main control loop for the controller and most internal circuits, including the $INTV_{CC}$ regulator. Releasing the RUN pin allows an internal 1.0µA current to pull up the pin and enable the controller. Alternatively, the RUN pin may be externally pulled up or driven directly by logic. Be careful not to exceed the absolute maximum rating of 6V on this pin. The start-up of the controller's output voltage, V_{OUT}, is controlled by the voltage on the TK/SS pin, if the internal soft-start has expired. When the voltage on the TK/SS pin is less than the 0.6V internal reference, the LTC3866 regulates the V_{FB} voltage to the TK/SS pin voltage instead of the 0.6V reference. This allows the TK/SS pin to be used to program a soft-start by connecting an external capacitor from the TK/SS pin to SGND. An internal 1.25µA pull-up current charges this capacitor, creating a voltage ramp on the TK/SS pin. As the TK/SS voltage rises linearly from 0V to 0.6V (and beyond), the output voltage, V_{OUT}, rises smoothly from zero to its final value. Alternatively, the TK/SS pin can be used to cause the start-up of V_{OUT} to *track* that of another supply. Typically, this requires connecting to the TK/SS pin an external resistor divider from the other supply to ground (see the Applications Information section). When the RUN pin is pulled low to disable the controller, or when $INTV_{CC}$ drops below its undervoltage lockout threshold of 3.75V, the TK/SS pin is pulled low by an internal MOSFET. When in undervoltage lockout, the controller is disabled and the external MOSFETs are held off.

Light Load Current Operation (Burst Mode Operation, Pulse-Skipping or Continuous Conduction)

The LTC3866 can be enabled to enter high efficiency Burst Mode operation, constant-frequency pulse-skipping mode or forced continuous conduction mode. To select forced continuous operation, tie the MODE/PLLIN pin to SGND. To select pulse-skipping mode of operation, tie the MODE/PLLIN pin to $INTV_{CC}$. To select Burst Mode operation, float the MODE/PLLIN pin. When the controller is enabled for Burst Mode operation, the peak current in the inductor is set to approximately one-third of the maximum sense voltage even though the voltage on the ITH pin indicates a lower value. If the average inductor current is higher than the load current, the error amplifier, EA, will decrease the voltage on the ITH pin. When the ITH voltage drops below 0.5V, the internal sleep signal goes high (enabling "sleep" mode) and both external MOSFETs are turned off.

In sleep mode, the load current is supplied by the output capacitor. As the output voltage decreases, the EA's output begins to rise. When the output voltage drops enough, the sleep signal goes low, and the controller resumes normal operation by turning on the top external MOSFET on the next cycle of the internal oscillator. When the controller is enabled for Burst Mode operation, the inductor current is not allowed to reverse. The reverse current comparator (I_{REV}) turns off the bottom external MOSFET just before the inductor current reaches zero, preventing it from reversing and going negative. Thus, the controller operates in discontinuous operation.

In forced continuous operation, the inductor current is allowed to reverse at light loads or under large transient conditions. The peak inductor current is determined by the voltage on the ITH pin, just as in normal operation. In this mode, the efficiency at light loads is lower than in Burst Mode operation. However, continuous mode has the advantages of lower output ripple and less interference with audio circuitry.

When the MODE/PLLIN pin is connected to $INTV_{CC}$, the LTC3866 operates in PWM pulse skipping mode at light loads. At very light loads, the current comparator, I_{CMP}, may remain tripped for several cycles and force the external top MOSFET to stay off for the same number of cycles (i.e., skipping pulses). The inductor current is not allowed to reverse (discontinuous operation). This mode, like forced continuous operation, exhibits low output ripple as well as low audio noise and reduced RF interference as compared to Burst Mode operation. It provides higher low current efficiency than forced continuous mode, but not nearly as high as Burst Mode operation.

Frequency Selection and Phase-Locked Loop (FREQ and MODE/PLLIN Pins)

The selection of switching frequency is a trade-off between efficiency and component size. Low frequency operation increases efficiency by reducing MOSFET switching losses, but requires larger inductance and/or capacitance to maintain low output ripple voltage.

OPERATION

If the MODE/PLLIN pin is not being driven by an external clock source, the FREQ pin can be used to program the controller's operating frequency from 250kHz to 770kHz. There is a precision 10µA current flowing out of the FREQ pin so that the user can program the controller's switching frequency with a single resistor to SGND. A curve is provided later in the Applications Information section showing the relationship between the voltage on the FREQ pin and switching frequency.

A phase-locked loop (PLL) is available on the LTC3866 to synchronize the internal oscillator to an external clock source that is connected to the MODE/PLLIN pin. The PLL loop filter network is integrated inside the LTC3866. The phase-locked loop is capable of locking any frequency within the range of 250kHz to 770kHz. The frequency setting resistor should always be present to set the controller's initial switching frequency before locking to the external clock. The controller operates in forced continuous mode when it is synchronized.

Sensing the Output Voltage with a Differential Amplifier

The LTC3866 includes a low offset, high input impedance, unity-gain, high bandwidth differential amplifier for applications that require true remote sensing. Sensing the load across the load capacitors directly greatly benefits regulation in high current, low voltage applications, where board interconnection losses can be a significant portion of the total error budget. Connect DIFFP to the output load, and DIFFN to the load ground. See Figure 1.

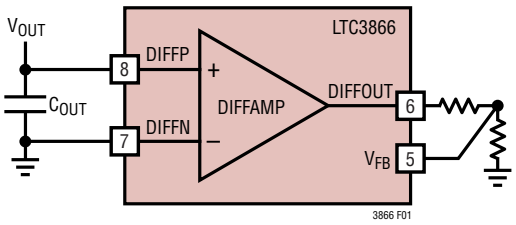


Figure 1. Differential Amplifier Connection

The LTC3866 differential amplifier has a typical output slew rate of 2V/µs. The amplifier is configured for unity gain, meaning that the difference between DIFFP and DIFFN is translated to DIFFOUT, relative to SGND.

Care should be taken to route the DIFFP and DIFFN PCB traces parallel to each other all the way to the remote sensing points on the board. In addition, avoid routing these sensitive traces near any high speed switching nodes in the circuit. Ideally, the DIFFP and DIFFN traces should be shielded by a low impedance ground plane to maintain signal integrity.

Power Good (PGOOD Pin)

The PGOOD pin is connected to the open drain of an internal N-channel MOSFET. The MOSFET turns on and pulls the PGOOD pin low when the V_{FB} pin voltage is not within ±10% of the 0.6V reference voltage. The PGOOD pin is also pulled low when the RUN pin is below 1.14V or when the LTC3866 is in the soft-start or tracking up phase. When the V_{FB} pin voltage is within the ±10% regulation window, the MOSFET is turned off and the pin is allowed to be pulled up by an external resistor to a source of up to 6V. The PGOOD pin will flag power good immediately when the V_{FB} pin is within the regulation window. However, there is an internal 20µs power-bad mask when the V_{FB} goes out of the window.

Output Overvoltage Protection

An overvoltage comparator, OV, guards against transient overshoots (>10%) as well as other more serious conditions that may overvoltage the output. In such cases, the top MOSFET is turned off and the bottom MOSFET is turned on until the overvoltage condition is cleared.

Undervoltage Lockout

The LTC3866 has two functions that help protect the controller in case of undervoltage conditions. A precision UVLO comparator constantly monitors the $INTV_{CC}$ voltage to ensure that an adequate gate-drive voltage is present. It locks out the switching action when $INTV_{CC}$ is below 3.75V. To prevent oscillation when there is a disturbance on the $INTV_{CC}$, the UVLO comparator has 600mV of precision hysteresis.

OPERATION

Another way to detect an undervoltage condition is to monitor the V_{IN} supply. Because the RUN pin has a precision turn-on reference of 1.22V, one can use a resistor divider to V_{IN} to turn on the IC when V_{IN} is high enough. An extra 4.5µA of current flows out of the RUN pin once the RUN pin voltage passes 1.22V. The RUN comparator itself has about 80mV of hysteresis. One can program additional hysteresis for the RUN comparator by adjusting the values of the resistive divider. For accurate V_{IN} undervoltage detection, V_{IN} needs to be higher than 4.75V.

APPLICATIONS INFORMATION

The Typical Application on the first page of this data sheet is a basic LTC3866 application circuit. The LTC3866 is designed and optimized for use with a very low DCR value by utilizing a novel approach to reduce the noise sensitivity of the sensing signal by a factor of 14dB. DCR sensing is becoming popular because it saves expensive current sensing resistors and is more power efficient, especially in high current applications. However, as the DCR value drops below 1mΩ, the signal-to-noise ratio is low and current sensing is difficult. LTC3866 uses an LTC proprietary technique to solve this issue. In general, external component selection is driven by the load requirement, and begins with the DCR and inductor value. Next, power MOSFETs are selected. Finally, input and output capacitors are selected.

Current Limit Programming

The ILIM pin is a 5-level logic input which sets the maximum current limit of the controller. When ILIM is either grounded, floated or tied to $INTV_{CC}$, the typical value for the maximum current sense threshold will be 10mV, 20mV or 30mV, respectively. Setting ILIM to one-fourth $INTV_{CC}$ and three-fourths $INTV_{CC}$ for maximum current sense thresholds of 15mV and 25mV.

Which setting should be used? For the best current limit accuracy, use the highest setting that is applicable to the output requirements.

SNSD+, SNSA+ and SNS– Pins

The SNSA+ and SNS– pins are the inputs to the current comparators, while the SNSD+ pin is the input of an internal amplifier. The operating input voltage range of 0V to 3.5V is for SNSA+, SNS– and SNSD+ when the internal differential amplifier is used to remotely sense the output. All the positive sense pins that are connected to the current comparator or the amplifier are high impedance with input bias currents of less than 1µA, but there is also a resistance of about 300k from the SNS– pin to ground. The SNS– should be connected directly to V_{OUT}. The SNSD+ pin connects to the filter that has a R1C1 time constant matched to L/DCR of the inductor. The SNSA+ pin is connected to the second filter with the time constant one-fifth that of R1C1. Care must be taken not to float these pins during normal operation. Filter components, especially capacitors, must be placed close to the LTC3866, and the sense lines should run close together to a Kelvin connection underneath the current sense element (Figure 2). Because the LTC3866 is designed to be used with a very low DCR value to sense inductor current, without proper care, the parasitic resistance, capacitance and inductance will degrade the current sense signal integrity, making the programmed current limit unpredictable. As shown in Figure 3, resistors R1 and R2 are placed close to the output inductor and capacitors C1 and C2 are close to the IC pins to prevent noise coupling to the sense signal.

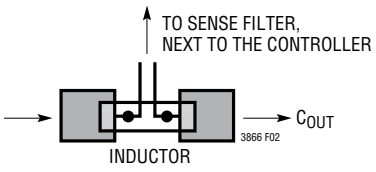


Figure 2. Sense Lines Placement with Inductor DCR

APPLICATIONS INFORMATION

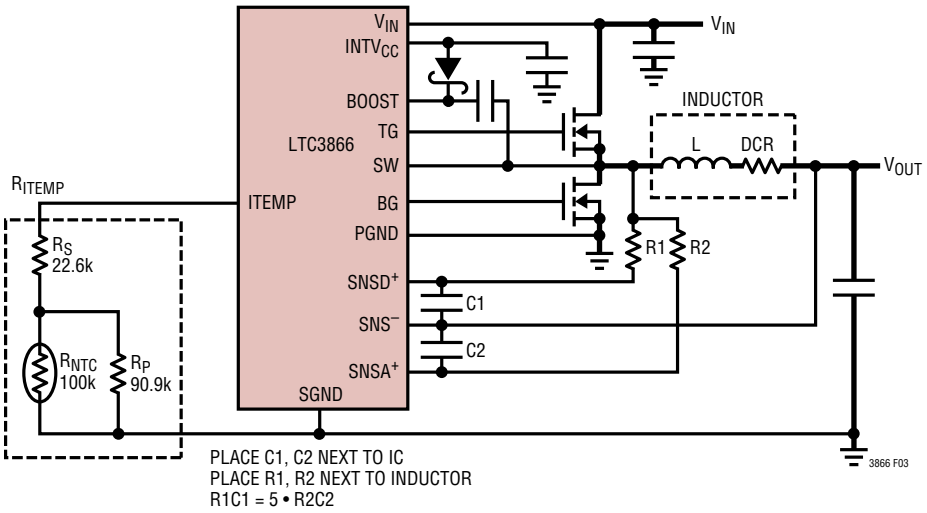


Figure 3. Inductor DCR Current Sensing

The LTC3866 could also be used like any typical current mode controller by disabling the SNSD+ pin, shorting it to ground. An R_{SENSE} resistor or a RC filter can be used to sense the output inductor signal and connects to the SNSA+ pin. If the RC filter is used, its time constant, R • C, is equaled to L/DCR of the output inductor. In these applications, the current limit, $V_{SENSE\ (MAX)}$, will be five times larger for the specified ILIM, and the operating voltage range of SNSA+ and SNS– is from 0V to 5.25V. Without using the internal differential amplifier, the output voltage of 5V can be generated as shown in the Typical Applications section.

Inductor DCR Sensing

The LTC3866 is specifically designed for high load current applications requiring the highest possible efficiency; it is capable of sensing the signal of an inductor DCR in the sub milliohm range (Figure 3). The DCR is the DC winding resistance of the inductor's copper, which is often less than 1mΩ for high current inductors. In high current and low output voltage applications, a conduction loss of a high DCR or a sense resistor will cause a significant reduction in power efficiency. For a specific output requirement, chose the inductor with the DCR that satisfies the maximum desirable sense voltage, and uses the relationship of the sense pin filters to output inductor characteristics as depicted below.

$$DCR = \frac{V_{SENSE(MAX)}}{I_{MAX} + \frac{\Delta I_L}{2}}$$

$$L/DCR = R1 \bullet C1 = 5 \bullet R2 \bullet C2$$

where:

$V_{SENSE(MAX)}$: Maximum sense voltage for a given ILIM threshold

I_{MAX}: Maximum load current

ΔI_L: Inductor ripple current

L, DCR: Output inductor characteristics

R1, C1: Filter time constant of the SNSD+ pin

R2, C2: Filter time constant of the SNSA+ pin

To ensure that the load current will be delivered over the full operating temperature range, the temperature coefficient of DCR resistance, approximately 0.4%/°C, should be taken into account. The LTC3866 features a DCR temperature compensation circuit that uses an NTC temperature sensing resistor for this purpose. See the Inductor DCR Sensing Temperature Compensation section for details.

APPLICATIONS INFORMATION

Typically, C1 and C2 are selected in the range of 0.047µF to 0.47µF. If C1 and C2 are chosen to be 220nF, and an inductor of 330nH with 0.32mΩ DCR is selected, R1 and R2 will be 4.7k and 942Ω respectively. The bias current at $SNSD^+$ and $SNSA^+$ is about 30nA and 500nA respectively, and it causes some small error to the sense signal.

There will be some power loss in R1 and R2 that relates to the duty cycle, and will be the most in continuous mode at the maximum input voltage:

$$P_{LOSS}(R) = \frac{\left(V_{IN(MAX)} - V_{OUT}\right) \bullet V_{OUT}}{R}$$

Ensure that R1 and R2 have a power rating higher than this value. However, DCR sensing eliminates the conduction loss of a sense resistor; it will provide a better efficiency at heavy loads. To maintain a good signal-to-noise ratio for the current sense signal, using a minimum ΔV_{SENSE} of 2mV for duty cycles less than 40% is desirable. The actual ripple voltage will be determined by the following equation:

$$\Delta V_{SENSE} = \frac{V_{OUT}}{V_{IN}} \bullet \frac{V_{IN} - V_{OUT}}{R1C1 \bullet f_{OSC}}$$

Inductor DCR Sensing Temperature Compensation with NTC Thermistor

For DCR sensing applications, the temperature coefficient of the inductor winding resistance should be taken into account when the accuracy of the current limit is critical over a wide range of temperature. The main element used in inductors is Copper; that has a positive tempco of approximately 4000ppm/°C. The LTC3866 provides a feature to correct for this variation through the use of the ITEMP pin. There is a 10µA precision current source flowing out of the ITEMP pin. A thermistor with a NTC (negative temperature coefficient) resistance can be used in a network, R_{ITEMP} (Figure 3) connected to maintain the current limit threshold constant over a wide operating temperature. The ITEMP voltage range that activates the correction is from 0.7V or less. If floating this pin, its voltage will be at $INTV_{CC}$ potential, about 5.5V. When the ITEMP voltage is higher than 0.7V, the temperature compensation is inactive.

The following guideline will help to choose components for temperature correction. The initial compensation is for 25°C ambient temperature:

$ITEMP \bullet R_{ITEMP} = 0.7V$ for 25°C

R_{ITEMP} is a thermistor resistance network connected to ITEMP pin.

Since ITEMP = 10µA, choose R_{ITEMP} network = 70kΩ at 25°C

$TC_{RITEMP} = -(1.5/0.7) \bullet TC_{DCR}$

Typically TC_{DCR} = 4000ppm/°C, tempco of DCR which is usually Copper. For ideal compensation, the tempco of the R_{ITEMP} should be:

$TC_{RITEMP} = -(1.5/0.7) \bullet 4000$ ppm/°C = –8570 ppm/°C

For example, a Murata NTC thermistor of 100k with B = 4334 that has a nonlinear temperature characteristic as described in R[T] = R[T0] • EXP B (1/T – 1/T0) where T0 is the temperature at 300°K. Resistors R_S and R_P of 22.6k and 90.9k respectively are used to linearize the network as shown in Figure 4.The current limit threshold will be compensated from 25°C to over 100°C of the inductor temperature, Figure 5. Once the temperature compensation is done, it will remain valid for all programmable current sense limit scales.

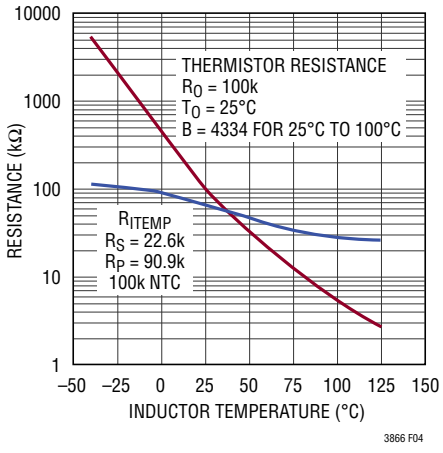


Figure 4. Resistance Versus Temperature for the ITEMP Pin Network and the 100k NTC

APPLICATIONS INFORMATION

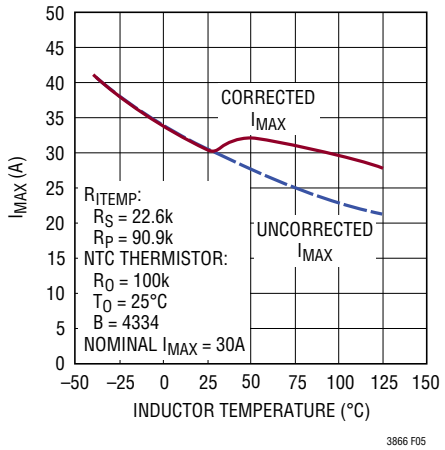


Figure 5. Worst-Case I_{MAX} Versus Inductor Temperature Curve with and without NTC Temperature Compensation

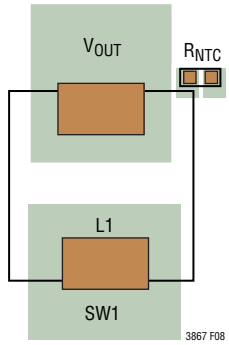


Figure 6. Thermistor Location. Place the Thermistor Next to the Inductor for Accurate Sensing of the Inductor Temperature, But Keep the ITEMP Pin Away from the Switch Nodes and Gate Drive Traces

For the most accurate temperature detection, place the thermistor next to the output inductor as shown in Figure 6. Care should be taken to keep the ITEMP sense line away from switch nodes.

Pre-Biased Output Start-Up

There may be situations that require the power supply to start up with a pre-bias on the output capacitors. In this case, it is desirable to start up without discharging that output pre-bias. The LTC3866 can safely power up into a pre-biased output without discharging it.

The LTC3866 accomplishes this by disabling both TG and BG until the TK/SS pin voltage and the internal soft-start voltage are above the V_{FB} pin voltage. When V_{FB} is higher than TK/SS or the internal soft-start voltage, the error amp output is railed low. The control loop would like to turn BG on, which would discharge the output. Disabling BG and TG prevents the pre-biased output voltage from being discharged. When TK/SS and the internal soft-start both cross 500mV or V_{FB}, whichever is lower, TG and BG are enabled. If the pre-bias is higher than the OV threshold, the bottom gate is turned on immediately to pull the output back into the regulation window.

Overcurrent Fault Recovery

When the output of the power supply is loaded beyond its preset current limit, the regulated output voltage will collapse depending on the load. The output may be shorted to ground through a very low impedance path or it may be a resistive short, in which case the output will collapse partially, until the load current equals the preset current limit. The controller will continue to source current into the short. The amount of current sourced depends on the ILIM pin setting and the V_{FB} voltage as shown in the Current Foldback graph in the Typical Performance Characteristics section.

Upon removal of the short, the output soft starts using the internal soft-start, thus reducing output overshoot. In the absence of this feature, the output capacitors would have been charged at current limit, and in applications with minimal output capacitance this may have resulted in output overshoot. Current limit foldback is not disabled during an overcurrent recovery. The load must step below the folded back current limit threshold in order to restart from a hard short.

Thermal Protection

Excessive ambient temperatures, loads and inadequate airflow or heat sinking can subject the chip, inductor, FETs etc. to high temperatures. This thermal stress reduces component life and if severe enough, can result in immediate catastrophic failure (Note 4). To protect the power supply from undue thermal stress, the LTC3866 has a fixed chip temperature-based thermal shutdown. The internal thermal shutdown is set for approximately 160°C with 10°C of hysteresis. When the chip reaches 160°C, both TG and BG are disabled until the chip cools down below 150°C.

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Inductor Value Calculation

Given the desired input and output voltages, the inductor value and operating frequency, f_{OSC}, directly determine the inductor's peak-to-peak ripple current:

$$I_{RIPPLE} = \frac{V_{OUT}}{V_{IN}}\left(\frac{V_{IN} - V_{OUT}}{f_{OSC} \bullet L}\right)$$

Lower ripple current reduces core losses in the inductor, ESR losses in the output capacitors, and output voltage ripple. Thus, highest efficiency operation is obtained at low frequency with a small ripple current. Achieving this, however, requires a large inductor.

A reasonable starting point is to choose a ripple current that is about 40% of $I_{OUT(MAX)}$. Note that the largest ripple current occurs at the highest input voltage. To guarantee that ripple current does not exceed a specified maximum, the inductor should be chosen according to:

$$L \geq \frac{V_{IN} - V_{OUT}}{f_{OSC} \bullet I_{RIPPLE}} \bullet \frac{V_{OUT}}{V_{IN}}$$

Inductor Core Selection

Once the inductance value is determined, the type of inductor must be selected. Core loss is independent of core size for a fixed inductor value, but it is very dependent on inductance selected. As inductance increases, core losses go down. Unfortunately, increased inductance requires more turns of wire and therefore copper losses will increase.

Ferrite designs have very low core loss and are preferred at high switching frequencies, so design goals can concentrate on copper loss and preventing saturation. Ferrite core material saturates "hard," which means that inductance collapses abruptly when the peak design current is exceeded. This results in an abrupt increase in inductor ripple current and consequent output voltage ripple. Do not allow the core to saturate!

Power MOSFET and Schottky Diode (Optional) Selection

At least two external power MOSFETs need to be selected: One N-channel MOSFET for the top (main) switch and one or more N-channel MOSFET(s) for the bottom (synchronous) switch. The number, type and on-resistance of all MOSFETs selected take into account the voltage step-down ratio as well as the actual position (main or synchronous) in which the MOSFET will be used. A much smaller and much lower input capacitance MOSFET should be used for the top MOSFET in applications that have an output voltage that is less than one-third of the input voltage. In applications where $V_{IN} >> V_{OUT}$, the top MOSFETs' on-resistance is normally less important for overall efficiency than its input capacitance at operating frequencies above 300kHz. MOSFET manufacturers have designed special purpose devices that provide reasonably low on-resistance with significantly reduced input capacitance for the main switch application in switching regulators.

The peak-to-peak MOSFET gate drive levels are set by the internal regulator voltage, V_{INTVCC}, requiring the use of logic-level threshold MOSFETs in most applications. Pay close attention to the BV_{DSS} specification for the MOSFETs as well; many of the logic-level MOSFETs are limited to 30V or less. Selection criteria for the power MOSFETs include the on-resistance, $R_{DS(ON)}$, input capacitance, input voltage and maximum output current. MOSFET input capacitance is a combination of several components but can be taken from the typical *gate charge* curve included on most data sheets (Figure 7). The curve is generated by forcing a constant input current into the gate of a common source, current source loaded stage and then plotting the gate voltage versus time.

The initial slope is the effect of the gate-to-source and the gate-to-drain capacitance. The flat portion of the curve is the result of the Miller multiplication effect of the drain-to-gate capacitance as the drain drops the voltage across the current source load. The upper sloping line is

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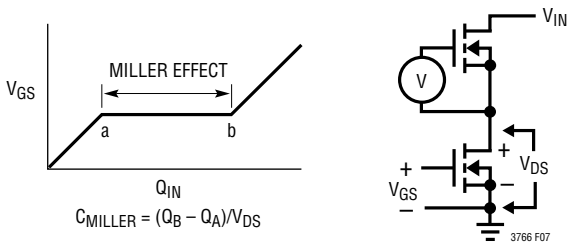

Figure 7. Gate Charge Characteristic

due to the drain-to-gate accumulation capacitance and the gate-to-source capacitance. The Miller charge (the increase in coulombs on the horizontal axis from a to b while the curve is flat) is specified for a given V_{DS} drain voltage, but can be adjusted for different V_{DS} voltages by multiplying the ratio of the application V_{DS} to the curve specified V_{DS} values. A way to estimate the C_{MILLER} term is to take the change in gate charge from points a and b on a manufacturer's data sheet and divide by the stated V_{DS} voltage specified. C_{MILLER} is the most important selection criteria for determining the transition loss term in the top MOSFET but is not directly specified on MOSFET data sheets. C_{RSS} and C_{OS} are specified sometimes but definitions of these parameters are not included. When the controller is operating in continuous mode the duty cycles for the top and bottom MOSFETs are given by:

$$\text{Main Switch Duty Cycle} = \frac{V_{OUT}}{V_{IN}}$$

$$\text{Synchronous Switch Duty Cycle} = \left(\frac{V_{IN} - V_{OUT}}{V_{IN}}\right)$$

The power dissipation for the main and synchronous MOSFETs at maximum output current are given by:

$$P_{MAIN} = \frac{V_{OUT}}{V_{IN}}\left(I_{MAX}\right)^2(1+\delta)R_{DS(ON)} + (V_{IN})^2\left(\frac{I_{MAX}}{2}\right)(R_{DR})(C_{MILLER})\bullet\left[\frac{1}{V_{INTVCC} - V_{TH(MIN)}} + \frac{1}{V_{TH(MIN)}}\right]\bullet f$$

$$P_{SYNC} = \frac{V_{IN} - V_{OUT}}{V_{IN}}\left(I_{MAX}\right)^2(1+\delta)R_{DS(ON)}$$

where δ is the temperature dependency of $R_{DS(ON)}$, R_{DR} is the effective top driver resistance (approximately 2Ω at $V_{GS} = V_{MILLER}$), V_{IN} is the drain potential and the change in drain potential in the particular application. $V_{TH(MIN)}$ is the data sheet specified typical gate threshold voltage specified in the power MOSFET data sheet at the specified drain current. C_{MILLER} is the calculated capacitance using the gate charge curve from the MOSFET data sheet and the technique described above.

Both MOSFETs have I^2R losses while the topside N-channel equation includes an additional term for transition losses, which peak at the highest input voltage. For $V_{IN} < 20V$, the high current efficiency generally improves with larger MOSFETs, while for $V_{IN} > 20V$, the transition losses rapidly increase to the point that the use of a higher $R_{DS(ON)}$ device with lower C_{MILLER} actually provides higher efficiency. The synchronous MOSFET losses are greatest at high input voltage when the top switch duty factor is low or during a short-circuit when the synchronous switch is on close to 100% of the period.

The term $(1 + \delta)$ is generally given for a MOSFET in the form of a normalized $R_{DS(ON)}$ vs temperature curve, but $\delta = 0.005/°C$ can be used as an approximation for low voltage MOSFETs.

An optional Schottky diode across the synchronous MOSFET conducts during the dead time between the conduction of the two large power MOSFETs. This prevents the body diode of the bottom MOSFET from turning on, storing charge during the dead time and requiring a reverse-recovery period which could cost as much as several percent in efficiency. A 2A to 8A Schottky is generally a good compromise for both regions of operation due to the relatively small average current. Larger diodes result in additional transition loss due to their larger junction capacitance.

C_{IN} and C_{OUT} Selection

In continuous mode, the source current of the top MOSFET is a square wave of duty cycle $(V_{OUT})/(V_{IN})$. To prevent large voltage transients, a low ESR capacitor sized for the maximum RMS current of one channel must be used. The maximum RMS capacitor current is given by:

$$C_{IN}\ \text{Required}\ I_{RMS} \approx \frac{I_{MAX}}{V_{IN}}\left[(V_{OUT})(V_{IN} - V_{OUT})\right]^{1/2}$$

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This formula has a maximum at $V_{IN} = 2V_{OUT}$, where $I_{RMS} = I_{OUT}/2$. This simple worst-case condition is commonly used for design because even significant deviations do not offer much relief. Note that capacitor manufacturers' ripple current ratings are often based on only 2000 hours of life. This makes it advisable to further derate the capacitor, or to choose a capacitor rated at a higher temperature than required. Several capacitors may be paralleled to meet size or height requirements in the design. Due to the high operating frequency of the LTC3866, ceramic capacitors can also be used for C_{IN}. Always consult the manufacturer if there is any question.

Ceramic capacitors are becoming very popular for small designs but several cautions should be observed. X7R, X5R and Y5V are examples of a few of the ceramic materials used as the dielectric layer, and these different dielectrics have very different effect on the capacitance value due to the voltage and temperature conditions applied. Physically, if the capacitance value changes due to applied voltage change, there is a concomitant piezo effect which results in radiating sound! A load that draws varying current at an audible rate may cause an attendant varying input voltage on a ceramic capacitor, resulting in an audible signal. A secondary issue relates to the energy flowing back into a ceramic capacitor whose capacitance value is being reduced by the increasing charge. The voltage can increase at a considerably higher rate than the constant current being supplied because the capacitance value is decreasing as the voltage is increasing! Nevertheless, ceramic capacitors, when properly selected and used, can provide the lowest overall loss due to their extremely low ESR.

A small (0.1µF to 1µF) bypass capacitor, C_{IN}, between the chip V_{IN} pin and ground, placed close to the LTC3866, is also suggested. A 2.2Ω to 10Ω resistor placed between C_{IN} and V_{IN} pin provides further isolation.

The selection of C_{OUT} is driven by the required effective series resistance (ESR). Typically once the ESR requirement is satisfied the capacitance is adequate for filtering. The steady-state output ripple (ΔV_{OUT}) is determined by:

$$\Delta V_{OUT} \approx \Delta I_{RIPPLE}\left(ESR + \frac{1}{8fC_{OUT}}\right)$$

where f = operating frequency, C_{OUT} = output capacitance and ΔI_{RIPPLE} = ripple current in the inductor. The output ripple is highest at maximum input voltage since ΔI_{RIPPLE} increases with input voltage. The output ripple will be less than 50mV at maximum V_{IN} with $\Delta I_{RIPPLE} = 0.4I_{OUT(MAX)}$ assuming:

C_{OUT} required ESR < N • R_{SENSE}

and

$$C_{OUT} > \frac{1}{(8f)(R_{SENSE})}$$

The emergence of very low ESR capacitors in small, surface mount packages makes very small physical implementations possible. The ability to externally compensate the switching regulator loop using the ITH pin allows a much wider selection of output capacitor types. The impedance characteristic of each capacitor type is significantly different than an ideal capacitor and therefore requires accurate modeling or bench evaluation during design. Manufacturers such as Nichicon, Nippon Chemi-Con and Sanyo should be considered for high performance through-hole capacitors. The OS-CON semiconductor dielectric capacitors available from Sanyo and the Panasonic SP surface mount types have a good (ESR)(size) product.

Once the ESR requirement for C_{OUT} has been met, the RMS current rating generally far exceeds the $I_{RIPPLE(P-P)}$ requirement. Ceramic capacitors from AVX, Taiyo Yuden, Murata and TDK offer high capacitance value and very low ESR, especially applicable for low output voltage applications.

In surface mount applications, multiple capacitors may have to be paralleled to meet the ESR or RMS current handling requirements of the application. Aluminum electrolytic and dry tantalum capacitors are both available in surface mount configurations. New special polymer surface mount capacitors offer very low ESR also but have much lower capacitive density per unit volume. In the case of tantalum, it is critical that the capacitors are surge tested for use in switching power supplies. Several excellent choices are the AVX TPS, AVX TPSV, the KEMET T510 series of surface mount tantalums or the Panasonic SP series of surface mount special polymer capacitors

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available in case heights ranging from 2mm to 4mm. Other capacitor types include Sanyo POSCAP, Sanyo OS-CON, Nichicon PL series and Sprague 595D series. Consult the manufacturers for other specific recommendations.

Differential Amplifier

The LTC3866 has true remote voltage sense capability. The sense connections should be returned from the load, back to the differential amplifier's inputs through a common, tightly coupled pair of PC traces. The differential amplifier rejects common mode signals capacitively or inductively radiated into the feedback PC traces as well as ground loop disturbances. The LTC3866 diffamp has 80kΩ input impedance on DIFFP. It is designed to be connected directly to the output. The output of the diffamp connects to the V_{FB} pin through a voltage divider, setting the output voltage.

External Soft-Start and Tracking

The LTC3866 has the ability to either soft-start by itself or track the output of another channel or external supply. When the controller is configured to soft-start by itself, a capacitor may be connected to its TK/SS pin or the internal soft-start may be used. The controller is in the shutdown state if its RUN pin voltage is below 1.14V and its TK/SS pin is actively pulled to ground in this shutdown state. If the RUN pin voltage is above 1.22V, the controller powers up. A soft-start current of 1.25µA then starts to charge the TK/SS soft-start capacitor. Note that soft-start or tracking is achieved not by limiting the maximum output current of the controller but by controlling the output ramp voltage according to the ramp rate on the TK/SS pin. Current foldback is disabled during this phase to ensure smooth soft-start or tracking. The soft-start or tracking range is defined to be the voltage range from 0V to 0.6V on the TK/SS pin. The total soft-start time can be calculated as:

$$t_{SOFTSTART} = 0.6 \bullet \frac{C_{SS}}{1.25\mu A}$$

Regardless of the mode selected by the MODE/PLLIN pin, the controller always starts in discontinuous mode up to TK/SS = 0.5V. Between TK/SS = 0.5V and 0.54V, it will operate in forced continuous mode and revert to the selected mode once TK/SS > 0.54V. The output ripple is minimized during the 40mV forced continuous mode window, ensuring a clean PGOOD signal. When the channel is configured to track another supply, the feedback voltage of the other supply is duplicated by a resistor divider and applied to the TK/SS pin. Therefore, the voltage ramp rate on this pin is determined by the ramp rate of the other supply's voltage. It is only possible to track another supply that is slower than the internal soft-start ramp. Note that the small soft-start capacitor charging current is always flowing, producing a small offset error. To minimize this error, select the tracking resistive divider value to be small enough to make this error negligible. In order to track down another channel or supply after the soft-start phase expires, the LTC3866 is forced into continuous mode of operation as soon as V_{FB} is below the undervoltage threshold of 0.54V regardless of the setting on the MODE/PLLIN pin. However, the LTC3866 should always be set in forced continuous mode tracking down when there is no load. After TK/SS drops below 0.1V, the controller operates in discontinuous mode.

The LTC3866 allows the user to program how its output ramps up and down by means of the TK/SS pin. Through these pins, the output can be set up to either coincidentally or ratiometrically track another supply's output, as shown in Figure 8. In the following discussions, V_{OUT2} refers to the LTC3866's output as a slave and V_{OUT1} refers to another supply output as a master. To implement the coincident tracking in Figure 8a, connect an additional resistive divider to V_{OUT1} and connect its mid-point to the TK/SS pin of the slave controller. The ratio of this divider should be the same as that of the slave controller's feedback divider shown in Figure 9a. In this tracking mode, V_{OUT1} must be set higher than V_{OUT2}. To implement the ratiometric tracking in Figure 8b, the ratio of the V_{OUT2} divider should be exactly the same as the master controller's feedback divider shown in Figure 9b . By selecting different resistors, the LTC3866 can achieve different modes of tracking including the two in Figure 8.

So which mode should be programmed? While either mode in Figure 8 satisfies most practical applications,

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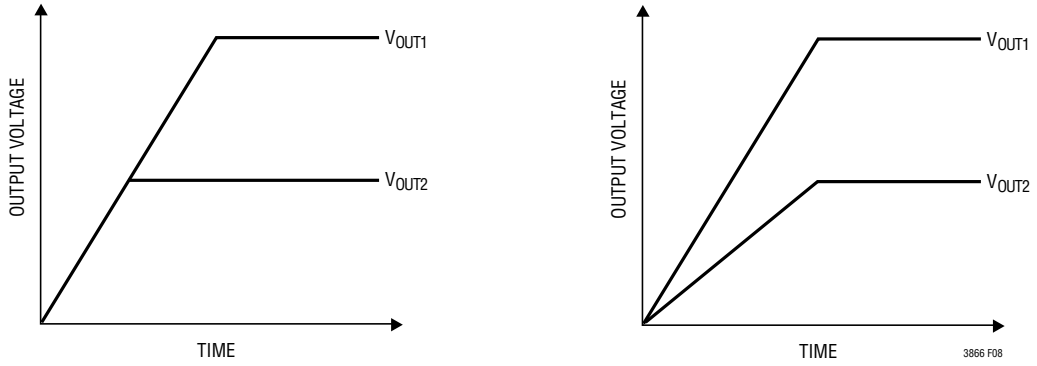
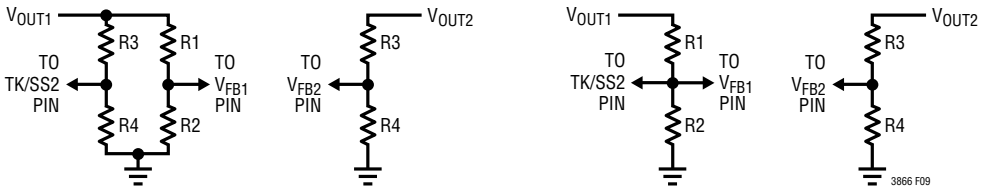


(8a) Coincident Tracking (8b) Ratiometric Tracking

Figure 8. Two Different Modes of Output Voltage Tracking



(9a) Coincident Tracking Setup (9b) Ratiometric Tracking Setup

Figure 9. Setup and Coincident and Ratiometric Tracking

some trade-offs exist. The ratiometric mode saves a pair of resistors, but the coincident mode offers better output regulation. Under ratiometric tracking, when the master controller's output experiences dynamic excursion (under load transient, for example), the slave controller output will be affected as well. For better output regulation, use the coincident tracking mode instead of ratiometric.

$INTV_{CC}$ (LDO) and $EXTV_{CC}$

The LTC3866 features a true PMOS LDO that supplies power to $INTV_{CC}$ from the V_{IN} supply. $INTV_{CC}$ powers the gate drivers and much of the LTC3866's internal circuitry. The LDO regulates the voltage at the $INTV_{CC}$ pin to 5.5V when V_{IN} is greater than 6V. $EXTV_{CC}$ connects to $INTV_{CC}$ through a P-channel MOSFET and can supply the needed power when its voltage is higher than 4.7V. Either of these can supply a peak current of 100mA and must be bypassed to ground with a minimum of 4.7µF ceramic capacitor or low ESR electrolytic capacitor. No matter what type of bulk capacitor is used, an additional 0.1µF ceramic capacitor placed directly adjacent to the $INTV_{CC}$ and PGND pins is highly recommended. Good bypassing is needed to supply the high transient currents required by the MOSFET gate drivers. High input voltage applications in which large MOSFETs are being driven at high frequencies may cause the maximum junction temperature rating for the LTC3866 to be exceeded. The $INTV_{CC}$ current, which is dominated by the gate charge current, may be supplied by either the 5.5V LDO or $EXTV_{CC}$. When the voltage on the $EXTV_{CC}$ pin is less than 4.5V, the LDO is enabled. Power dissipation for the IC in this case is highest and is equal to $V_{IN} \bullet I_{INTVCC}$. The gate charge current is dependent on operating frequency as discussed in the Efficiency Considerations section. The junction temperature can be estimated by using the equations given in Note 2 of the Electrical Characteristics tables. For example, the LTC3866 $INTV_{CC}$ current is limited to less than 39mA from a 38V supply in the UF package and not using the $EXTV_{CC}$ supply with a 70°C ambient temperature:

$$T_J = 70°C + (39mA)(38V)(37°C/W) \cong 125°C$$

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To prevent the maximum junction temperature from being exceeded, the input supply current must be checked while operating in continuous conduction mode (MODE/PLLIN = SGND) at maximum V_{IN}. When the voltage applied to $EXTV_{CC}$ rises above 4.7V, the $INTV_{CC}$ LDO is turned off and the $EXTV_{CC}$ is connected to the $INTV_{CC}$. The $EXTV_{CC}$ remains on as long as the voltage applied to $EXTV_{CC}$ remains above 4.5V. Using the $EXTV_{CC}$ allows the MOSFET driver and control power to be derived from an efficient switching regulator output during normal operation. If more current is required through the $EXTV_{CC}$ than is specified, an external Schottky diode can be added between the $EXTV_{CC}$ and $INTV_{CC}$ pins. Do not apply more than 6V to the $EXTV_{CC}$ pin and make sure that $EXTV_{CC} < V_{IN}$.

Significant efficiency and thermal gains can be realized by powering $INTV_{CC}$ from $EXTV_{CC}$, since the V_{IN} current resulting from the driver and control currents will be scaled by a factor of (duty cycle)/(switcher efficiency). Tying the $EXTV_{CC}$ pin to a 5V supply reduces the junction temperature in the previous example from 125°C to:

$$T_J = 70°C + (39mA)(5V)(37°C/W) = 77°C$$

However, for low voltage outputs, additional circuitry is required to derive $INTV_{CC}$ power from the output.

The following list summarizes the three possible connections for $EXTV_{CC}$:

1. $EXTV_{CC}$ left open (or grounded). This will cause $INTV_{CC}$ to be powered from the internal LDO resulting in an efficiency penalty of up to 10% at high input voltages.
2. $EXTV_{CC}$ connected to an external supply. If a 5V external supply is available, it may be used to power $EXTV_{CC}$ providing it is compatible with the MOSFET gate drive requirements.
3. $EXTV_{CC}$ connected to an output-derived boost network. For 3.3V and other low voltage regulators, efficiency gains can still be realized by connecting $EXTV_{CC}$ to an output-derived voltage that has been boosted to greater than 4.7V.

For applications where the main input power is 5V, tie the V_{IN} and $INTV_{CC}$ pins together and tie the combined pins to the 5V input with a 1Ω or 2.2Ω resistor as shown in Figure 10 to minimize the voltage drop caused by the gate charge current. This will override the $INTV_{CC}$ linear regulator and will prevent $INTV_{CC}$ from dropping too low due to the dropout voltage. Make sure the $INTV_{CC}$ voltage is at or exceeds the $R_{DS(ON)}$ test voltage for the MOSFET which is typically 4.5V for logic-level devices.

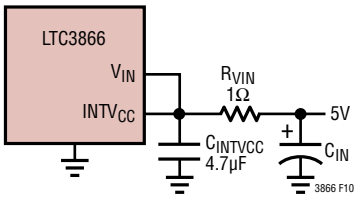

Figure 10. Setup for a 5V Input

Topside MOSFET Driver Supply (C_B, D_B)

External bootstrap capacitor, C_B, connected to the BOOST pin supplies the gate drive voltages for the topside MOSFET. Capacitor C_B in the Functional Diagram is charged though external diode D_B from $INTV_{CC}$ when the SW pin is low. When the topside MOSFET is to be turned on, the driver places the C_B voltage across the gate source of the MOSFET. This enhances the MOSFET and turns on the topside switch. The switch node voltage, SW, rises to V_{IN} and the BOOST pin follows. With the topside MOSFET on, the boost voltage is above the input supply:

$$V_{BOOST} = V_{IN} + V_{INTVCC} – V_{DB}$$

The value of the boost capacitor, C_B, needs to be 100 times that of the total input capacitance of the topside MOSFET(s). The reverse breakdown of the external Schottky diode must be greater than $V_{IN(MAX)}$. When adjusting the gate drive level, the final arbiter is the total input current for the regulator. If a change is made and the input current decreases, then the efficiency has improved. If there is no change in input current, then there is no change in efficiency.

Setting Output Voltage

The LTC3866 output voltage is set by an external feedback resistive divider carefully placed across the DIFFOUT pin,

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as shown in Figure 11. The regulated output voltage is determined by:

$$V_{OUT} = 0.6V \bullet \left(1 + \frac{R_B}{R_A}\right)$$

To improve the frequency response, a feedforward capacitor, C_{FF}, may be used. Great care should be taken to route the V_{FB} line away from noise sources, such as the inductor or the SW line.

To minimize the effect of the voltage drop caused by high current flowing through board conductance; connect DIFFN and DIFFP sense lines close to the ground and the load output respectively.

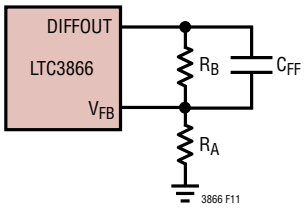

Figure 11. Setting Output Voltage

Fault Conditions: Current Limit and Current Foldback

The LTC3866 includes current foldback to help limit load current when the output is shorted to ground. If the output falls below 50% of its nominal output level, then the maximum sense voltage is progressively lowered from its maximum programmed value to one-third of the maximum value. Foldback current limiting is disabled during the soft-start or tracking up using the TK/SS pin. It is not disabled for internal soft-start. Under short-circuit conditions with very low duty cycles, the LTC3866 will begin cycle skipping in order to limit the short-circuit current. In this situation the bottom MOSFET will be dissipating most of the power but less than in normal operation. The short circuit ripple current is determined by the minimum on-time $t_{ON(MIN)}$ of the LTC3866 (≈90ns), the input voltage and inductor value:

$$\Delta I_{L(SC)} = t_{ON(MIN)} \bullet \frac{V_{IN}}{L}$$

The resulting short-circuit current is:

$$I_{SC} = \left(\frac{1/3\, V_{SENSE(MAX)}}{R_{SENSE}} - \frac{1}{2}\Delta I_{L(SC)}\right)$$

After a short, or while starting with internal soft-start, make sure that the load current takes the folded-back current limit into account. The output may not restart if the load current exceeds the fold-back current limit. Therefore, it is not recommended to set ILIM to ground due to its small fold-back current limits.

Phase-Locked Loop and Frequency Synchronization

The LTC3866 has a phase-locked loop (PLL) comprised of an internal voltage-controlled oscillator (VCO) and a phase detector. This allows the turn-on of the top MOSFET to be locked to the rising edge of an external clock signal applied to the MODE/PLLIN pin. The phase detector is an edge sensitive digital type that provides zero degrees phase shift between the external and internal oscillators. This type of phase detector does not exhibit false lock to harmonics of the external clock.

The output of the phase detector is a pair of complementary current sources that charge or discharge the internal filter network. There is a precision 10µA current flowing out of the FREQ pin. This allows the user to use a single resistor to SGND to set the switching frequency when no external clock is applied to the MODE/PLLIN pin. The internal switch between the FREQ pin and the integrated PLL filter network is on, allowing the filter network to be pre-charged to the same voltage as the FREQ pin. The relationship between the voltage on the FREQ pin and operating frequency is shown in Figure 12 and specified in the Electrical Characteristics table. If an external clock is detected on the MODE/PLLIN pin, the internal switch mentioned above turns off and isolates the influence of the FREQ pin. Note that the LTC3866 can only be synchronized to an external clock whose frequency is within range of the LTC3866's internal VCO. This is guaranteed to be between 250kHz and 770kHz. A simplified block diagram is shown in Figure 13.

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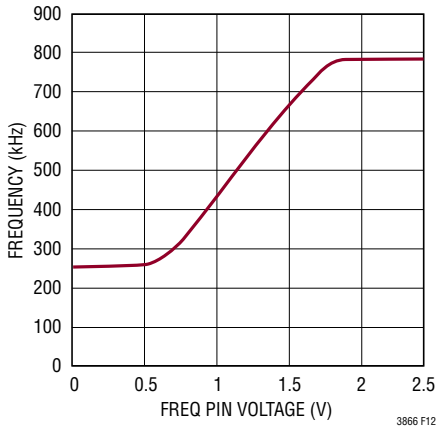


Figure 12. Relationship Between Oscillator Frequency and Voltage at the FREQ Pin

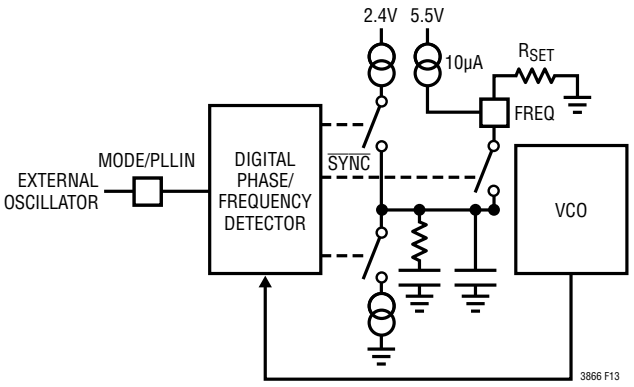


Figure 13. Phase-Locked Loop Block Diagram

If the external clock frequency is greater than the internal oscillator's frequency, f_{OSC}, then current is sourced continuously from the phase detector output, pulling up the filter network. When the external clock frequency is less than f_{OSC}, current is sunk continuously, pulling down the filter network. If the external and internal frequencies are the same but exhibit a phase difference, the current sources turn on for an amount of time corresponding to the phase difference. The voltage on the filter network is adjusted until the phase and frequency of the internal and external oscillators are identical. At the stable operating point, the phase detector output is high impedance and the filter capacitor C_{LP} holds the voltage.

Typically, the external clock (on the MODE/PLLIN pin) input high threshold is 1.6V, while the input low threshold is 1V.

Minimum On-Time Considerations

Minimum on-time, $t_{ON(MIN)}$, is the smallest time duration that the LTC3866 is capable of turning on the top MOSFET. It is determined by internal timing delays and the gate charge required to turn on the top MOSFET. Low duty cycle applications may approach this minimum on-time limit and care should be taken to ensure that:

$$t_{ON(MIN)} < \frac{V_{OUT}}{V_{IN}(f)}$$

If the duty cycle falls below what can be accommodated by the minimum on-time, the controller will begin to skip cycles. The output voltage will continue to be regulated, but the voltage ripple and current ripple will increase. The minimum on-time for the LTC3866 is approximately 90ns, with good PCB layout, minimum 30% inductor current ripple and at least 2mV ripple on the current sense signal. The minimum on-time can be affected by PCB switching noise in the voltage and current loop. As the peak sense voltage decreases the minimum on-time gradually increases to about 110ns. This is of particular concern in forced continuous applications with low ripple current at light loads. If the duty cycle drops below the minimum on-time limit in this situation, a significant amount of cycle skipping can occur with correspondingly larger current and voltage ripple.

Efficiency Considerations

The percent efficiency of a switching regulator is equal to the output power divided by the input power times 100%. It is often useful to analyze individual losses to determine what is limiting the efficiency and which change would produce the most improvement. Percent efficiency can be expressed as:

$$\%Efficiency = 100\% - (L1 + L2 + L3 + ...)$$

where L1, L2, etc. are the individual losses as a percentage of input power.

Although all dissipative elements in the circuit produce losses, four main sources usually account for most of the losses in LTC3866 circuits: 1) IC V_{IN} current, 2) $INTV_{CC}$ regulator current, 3) I^2R losses, 4) topside MOSFET transition losses.

APPLICATIONS INFORMATION

1. The V_{IN} current is the DC supply current given in the Electrical Characteristics table, which excludes MOSFET driver and control currents. V_{IN} current typically results in a small (<0.1%) loss.

2. $INTV_{CC}$ current is the sum of the MOSFET driver and control currents. The MOSFET driver current results from switching the gate capacitance of the power MOSFETs. Each time a MOSFET gate is switched from low to high to low again, a packet of charge dQ moves from $INTV_{CC}$ to ground. The resulting dQ/dt is a current out of $INTV_{CC}$ that is typically much larger than the control circuit current. In continuous mode, $I_{GATECHG} = f(Q_T + Q_B)$, where Q_T and Q_B are the gate charges of the topside and bottom side MOSFETs. Supplying $INTV_{CC}$ power through $EXTV_{CC}$ from an output-derived source will scale the V_{IN} current required for the driver and control circuits by a factor of (duty cycle)/(efficiency). For example, in a 20V to 5V application, 10mA of $INTV_{CC}$ current results in approximately 2.5mA of V_{IN} current. This reduces the mid-current loss from 10% or more (if the driver was powered directly from V_{IN}) to only a few percent.

3. I^2R losses are predicted from the DC resistances of the fuse (if used), MOSFET, inductor and current sense resistor (if used). In continuous mode, the average output current flows through L and R_{SENSE}, but is *chopped* between the topside MOSFET and the synchronous MOSFET. If the two MOSFETs have approximately the same $R_{DS(ON)}$, then the resistance of one MOSFET can simply be summed with the resistances of L and R_{SENSE} to obtain I^2R losses. For example, if each $R_{DS(ON)} = 10m\Omega$, $R_L = 10m\Omega$, $R_{SENSE} = 5m\Omega$, then the total resistance is $25m\Omega$. This results in losses ranging from 2% to 8% as the output current increases from 3A to 15A for a 5V output, or a 3% to 12% loss for a 3.3V output.

   Efficiency varies as the inverse square of V_{OUT} for the same external components and output power level. The combined effects of increasingly lower output voltages and higher currents required by high performance digital systems is not doubling but quadrupling the importance of loss terms in the switching regulator system!

4. Transition losses apply only to the topside MOSFET(s), and become significant only when operating at high input voltages (typically 15V or greater). Transition losses can be estimated from:

   $$\text{Transition Loss} = (1.7)\ V_{IN}^2 \bullet I_{O(MAX)} \bullet C_{RSS} \bullet f$$

   Other *hidden* losses such as copper trace and internal battery resistances can account for an additional 5% to 10% efficiency degradation in portable systems. It is very important to include these *system* level losses during the design phase. The internal battery and fuse resistance losses can be minimized by making sure that C_{IN} has adequate charge storage and very low ESR at the switching frequency. A 25W supply will typically require a minimum of 20µF to 40µF of capacitance having a maximum of $20m\Omega$ to $50m\Omega$ of ESR. Other losses, including Schottky conduction losses during dead time and inductor core losses, generally account for less than 2% total additional loss.

Checking Transient Response

The regulator loop response can be checked by looking at the load current transient response. Switching regulators take several cycles to respond to a step in DC (resistive) load current. When a load step occurs, V_{OUT} shifts by an amount equal to $\Delta I_{LOAD} \bullet ESR$, where ESR is the effective series resistance of C_{OUT}. ΔI_{LOAD} also begins to charge or discharge C_{OUT}, generating the feedback error signal that forces the regulator to adapt to the current change and return V_{OUT} to its steady-state value. During this recovery time V_{OUT} can be monitored for excessive overshoot or ringing, which would indicate a stability problem. The availability of the ITH pin not only allows optimization of control loop behavior but also provides a DC-coupled and AC-filtered closed-loop response test point. The DC step, rise time and settling at this test point truly reflects the closed-loop response. Assuming a predominantly second order system, phase margin and/or damping factor can be estimated using the percentage of overshoot seen at this pin. The bandwidth can also be estimated by examining the rise time at the pin. The ITH external components shown in the Typical Application circuit will provide an

APPLICATIONS INFORMATION

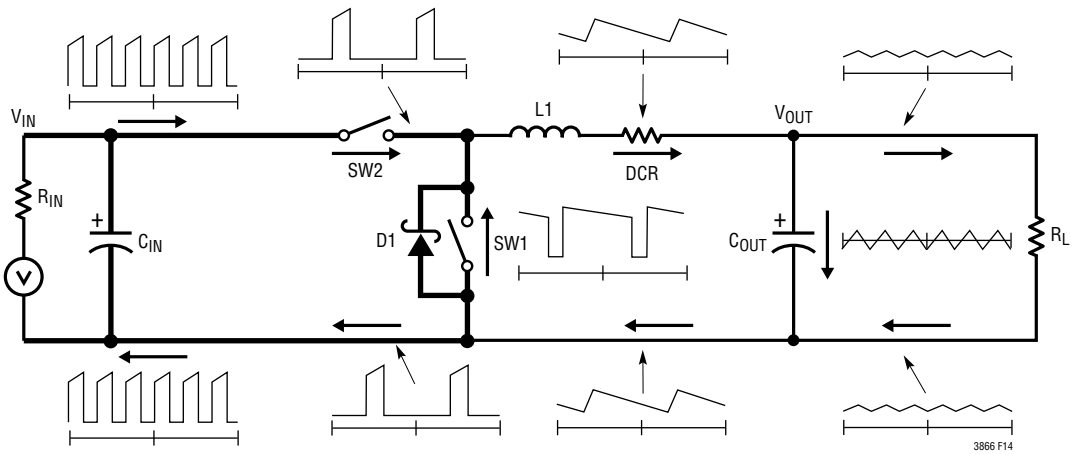
adequate starting point for most applications. The ITH series R_C-C_C filter sets the dominant pole-zero loop compensation. The values can be modified slightly (from 0.5 to 2 times their suggested values) to optimize transient response once the final PC layout is done and the particular output capacitor type and value have been determined. The output capacitors need to be selected because the various types and values determine the loop gain and phase. An output current pulse of 20% to 80% of full-load current having a rise time of 1µs to 10µs will produce output voltage and ITH pin waveforms that will give a sense of the overall loop stability without breaking the feedback loop. Placing a power MOSFET directly across the output capacitor and driving the gate with an appropriate signal generator is a practical way to produce a realistic load step condition. The initial output voltage step resulting from the step change in output current may not be within the bandwidth of the feedback loop, so this signal cannot be used to determine phase margin. This is why it is better to look at the ITH pin signal which is in the feedback loop and is the filtered and compensated control loop response. The gain of the loop will be increased by increasing R_C and the bandwidth of the loop will be increased by decreasing C_C. If R_C is increased by the same factor that C_C is decreased, the zero frequency will be kept the same, thereby keeping the phase shift the same in the most critical frequency range of the feedback loop. The output voltage settling behavior is related to the stability of the closed-loop system and will demonstrate the actual overall supply performance.

A second, more severe transient is caused by switching in loads with large (>1µF) supply bypass capacitors. The discharged bypass capacitors are effectively put in parallel with C_{OUT}, causing a rapid drop in V_{OUT}. No regulator can alter its delivery of current quickly enough to prevent this sudden step change in output voltage if the load switch resistance is low and it is driven quickly. If the ratio of C_{LOAD} to C_{OUT} is greater than 1:50, the switch rise time should be controlled so that the load rise time is limited to approximately 25 • C_{LOAD}. Thus a 10µF capacitor would require a 250µs rise time, limiting the charging current to about 200mA.

PC Board Layout Checklist

When laying out the printed circuit board, the following checklist should be used to ensure proper operation of the IC. These items are also illustrated graphically in the layout diagram of Figure 14. Check the following in the PC layout:

1. The $INTV_{CC}$ decoupling capacitor should be placed immediately adjacent to the IC between the $INTV_{CC}$ pin and PGND plane. A 1µF ceramic capacitor of the X7R or X5R type is small enough to fit very close to the IC to minimize the ill effects of the large current pulses drawn to drive the bottom MOSFETs. An additional 4.7µF to 10µF of ceramic, tantalum or other very low ESR capacitance is recommended in order to keep the internal IC supply quiet.



BOLD LINES INDICATE HIGH, SWITCHING CURRENTS. KEEP LINES TO A MINIMUM LENGTH

Figure 14. Branch Current Waveforms

APPLICATIONS INFORMATION

2. Place the feedback divider between the + and – terminals of C_{OUT}. Route DIFFP and DIFFN with minimum PC trace spacing from the IC to the feedback divider.
3. Are the $SNSD^+$, $SNSA^+$ and SNS^- printed circuit traces routed together with minimum PC trace spacing? The filter capacitors between $SNSD^+$, $SNSA^+$ and SNS^- should be as close as possible to the pins of the IC. Connect the $SNSD^+$ and $SNSA^+$ pins to the filter resistors as illustrated in Figure 3.
4. Do the (+) plates of C_{IN} connect to the drain of the topside MOSFET as closely as possible? This capacitor provides the pulsed current to the MOSFET.
5. Keep the switching nodes, SW, BOOST and TG away from sensitive small-signal nodes ($SNSD^+$, $SNSA^+$, SNS^-, DIFFP, DIFFN, V_{FB}). Ideally the SW, BOOST and TG printed circuit traces should be routed away and separated from the IC and especially the *quiet* side of the IC. Separate the high dv/dt traces from sensitive small-signal nodes with ground traces or ground planes.
6. Use a low impedance source such as a logic gate to drive the MODE/PLLIN pin and keep the lead as short as possible.
7. The 47pF to 330pF ceramic capacitor between the I_{TH} pin and signal ground should be placed as close as possible to the IC. Figure 14 illustrates all branch currents in a switching regulator. It becomes very clear after studying the current waveforms why it is critical to keep the high switching current paths to a small physical size. High electric and magnetic fields will radiate from these *loops* just as radio stations transmit signals. The output capacitor ground should return to the negative terminal of the input capacitor and not share a common ground path with any switched current paths. The left half of the circuit gives rise to the *noise* generated by a switching regulator. The ground terminations of the synchronous MOSFET and Schottky diode should return to the bottom plate(s) of the input capacitor(s) with a short isolated PC trace since very high switched currents are present. External OPTI-LOOP® compensation allows overcompensation for PC layouts which are not optimized but this is not the recommended design procedure.
8. Are the signal and power grounds kept separate? The combined IC signal ground pin and the ground return of C_{INTVCC} must return to the combined C_{OUT} (–) terminals. The V_{FB} and ITH traces should be as short as possible. The path formed by the top N-channel MOSFET, Schottky diode and the C_{IN} capacitor should have short leads and PC trace lengths. The output capacitor (–) terminals should be connected as close as possible to the (–) terminals of the input capacitor by placing the capacitors next to each other and away from the Schottky loop described above.
9. Use a modified "star ground" technique: a low impedance, large copper area central grounding point on the same side of the PC board as the input and output capacitors with tie-ins for the bottom of the $INTV_{CC}$ decoupling capacitor, the bottom of the voltage feedback resistive divider and the SGND pin of the IC.

Design Example

As a design example of the front page circuit for a single channel high current regulator, assume V_{IN} = 12V(nominal), V_{IN} = 20V(maximum), V_{OUT} = 1.5V, I_{MAX} = 30A, and f = 400kHz (see front page schematic).

The regulated output voltage is determined by:

$$V_{OUT} = 0.6V \bullet \left(1 + \frac{R_B}{R_A}\right)$$

Using a 20k 1% resistor from the V_{FB} node to ground, the top feedback resistor is (to the nearest 1% standard value) 30.1k.

The frequency is set by biasing the FREQ pin to 1V (see Figure 12).

The inductance value is based on a 35% maximum ripple current assumption (10.5A). The highest value of ripple current occurs at the maximum input voltage:

$$L = \frac{V_{OUT}}{f \bullet \Delta I_{L(MAX)}}\left(1 - \frac{V_{OUT}}{V_{IN(MAX)}}\right)$$

APPLICATIONS INFORMATION

This design will require 0.33µH. The Würth 744301033, 0.32µH inductor is chosen. At the nominal input voltage (12V), the ripple current will be:

$$\Delta I_{L(NOM)} = \frac{V_{OUT}}{f \bullet L}\left(1 - \frac{V_{OUT}}{V_{IN(NOM)}}\right)$$

It will have 10A (33%) ripple. The peak inductor current will be the maximum DC value plus one-half the ripple current, or 35A.

The minimum on-time occurs at the maximum V_{IN}, and should not be less than 90ns:

$$t_{ON(MIN)} = \frac{V_{OUT}}{V_{IN(MAX)}f} = \frac{1.5V}{20V(400kHz)} = 187ns$$

DCR sensing is used in this circuit. If C1 and C2 are chosen to be 220nF, based on the chosen 0.33µH inductor with 0.32mΩ DCR, R1 and R2 can be calculated as:

$$R1 = \frac{L}{DCR \bullet C1} = 4.69k$$

$$R2 = \frac{L}{DCR \bullet C2 \bullet 5} = 937\Omega$$

Choose R1 = 4.64k and R2 = 931Ω.

The maximum DCR of the inductor is 0.34Ω. The $V_{SENSE(MAX)}$ is calculated as:

$$V_{SENSE(MAX)} = I_{PEAK} \bullet DCR_{MAX} = 12mV$$

The current limit is chosen to be 15mV. If temperature variation is considered, please refer to Inductor DCR Sensing Temperature Compensation with NTC Thermistor.

The power dissipation on the topside MOSFET can be easily estimated. Choosing an Infineon BSC050NE2LS MOSFET results in: $R_{DS(ON)}$ = 7.1mΩ (max), V_{MILLER} = 2.8V, $C_{MILLER} \cong$ 35pF. At maximum input voltage with T_J (estimated) = 75°C:

$$P_{MAIN} = \frac{1.5V}{20V}(30A)^2\left[1+(0.005)(75°C-25°C)\right] \bullet$$

$$(0.0071\Omega) + (20V)^2\left(\frac{30A}{2}\right)(2\Omega)(35pF) \bullet$$

$$\left[\frac{1}{5.5V-2.8V} + \frac{1}{2.8V}\right](400kHz)$$

$$= 599mW + 122mW$$

$$= 721mW$$

For a 0.32mΩ DCR, a short-circuit to ground will result in a folded back current of:

$$I_{SC} = \frac{(1/3)15mV}{0.0032\Omega} - \frac{1}{2}\left(\frac{90ns(20V)}{0.33\mu H}\right) = 12.9A$$

An Infineon BSC010NE2LS, $R_{DS(ON)}$ = 1.1mΩ, is chosen for the bottom FET. The resulting power loss is:

$$P_{SYNC} = \frac{20V-1.5V}{20V}(30A)^2 \bullet$$

$$\left[1+(0.005)\bullet(75°C-25°C)\right]\bullet 0.0011\Omega$$

$$P_{SYNC} = 1.14W$$

C_{IN} is chosen for an equivalent RMS current rating of at least 13.7A. C_{OUT} is chosen with an equivalent ESR of 4.5mΩ for low output ripple. The output ripple in continuous mode will be highest at the maximum input voltage. The output voltage ripple due to ESR is approximately:

$$V_{ORIPPLE} = R_{ESR}(\Delta I_L) = 0.0045\Omega \bullet 10A = 45mV_{P-P}$$

Further reductions in output voltage ripple can be made by placing a 100µF ceramic capacitor across C_{OUT}.

TYPICAL APPLICATIONS

Very Low Output Ripple Converter

The LTC3866 can work with very low DCR inductors because it can operate with only a small peak-to-peak sense voltage. Two inductor characteristics can diminish this signal: lower DC resistance and higher inductance. While lower DCR improves efficiency, higher inductance reduces output ripple. Because the LTC3866 only requires a ripple signal about a quarter of the sense signal of the next best current mode converters, output ripple can be drastically reduced by increasing the inductance and capacitance of the output filter. The very small output voltage ripple is critical for low noise applications such as audio systems and noise sensitive systems.

The schematic as shown Figure 15 is similar to that of the front page circuit, except that three times the inductance and double the output capacitance are used. The compensation components are changed to maintain the same crossover frequency and phase margin. Figure 16 shows the transient response of 15A load step, and Figure 17 demonstrates that the output voltage ripple is a factor of six smaller than that of typical current mode converters.

Increasing the inductance, while maintaining the same physical size inductor, will invariably increase conduction losses due to higher DC resistance. However, reduced ripple current will decrease the core loss and the AC resistance loss often enough to negate the extra DC conduction losses. Figure 18 shows a high efficiency converter with the benefit of low output ripple current.

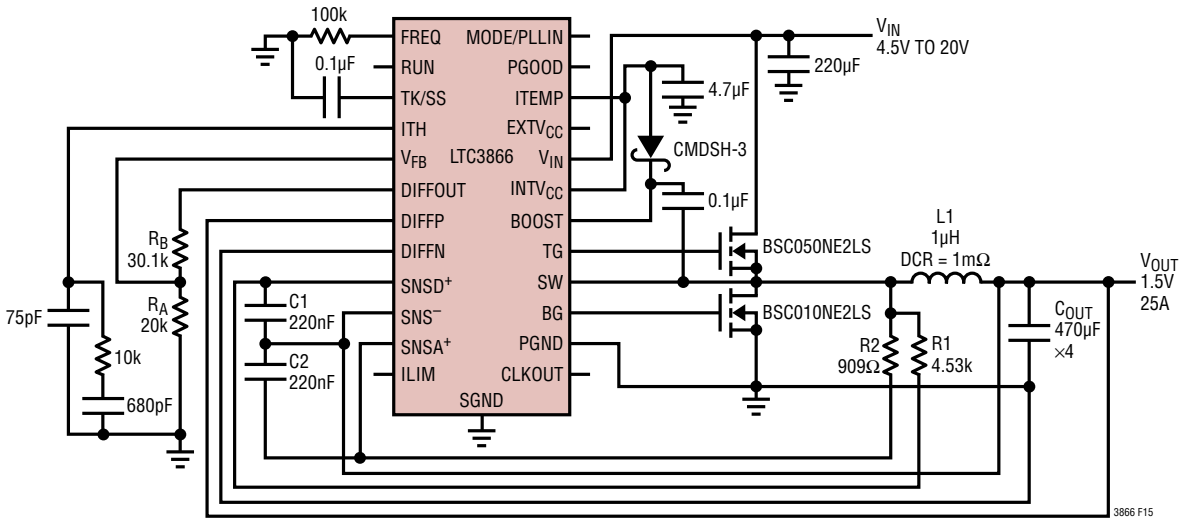


Figure 15. High Efficiency, 1.5V/25A Step-Down Converter with Very Low Output Ripple

TYPICAL APPLICATIONS

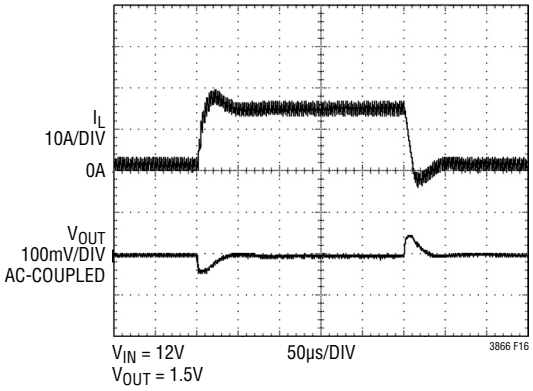


Figure 16. Load Step Transient Response

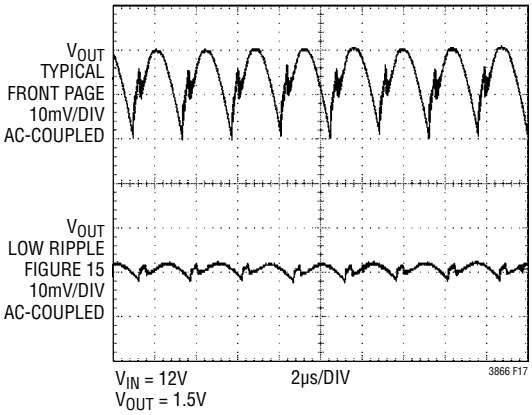


Figure 17. Very Low Output Voltage Ripple

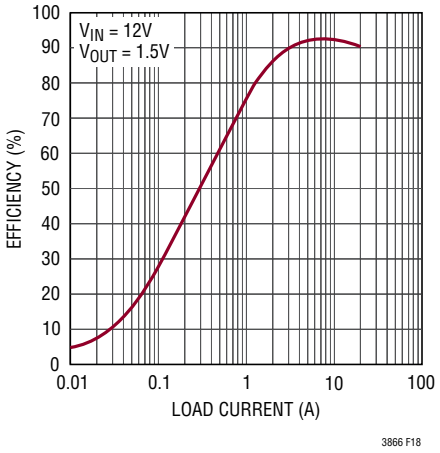
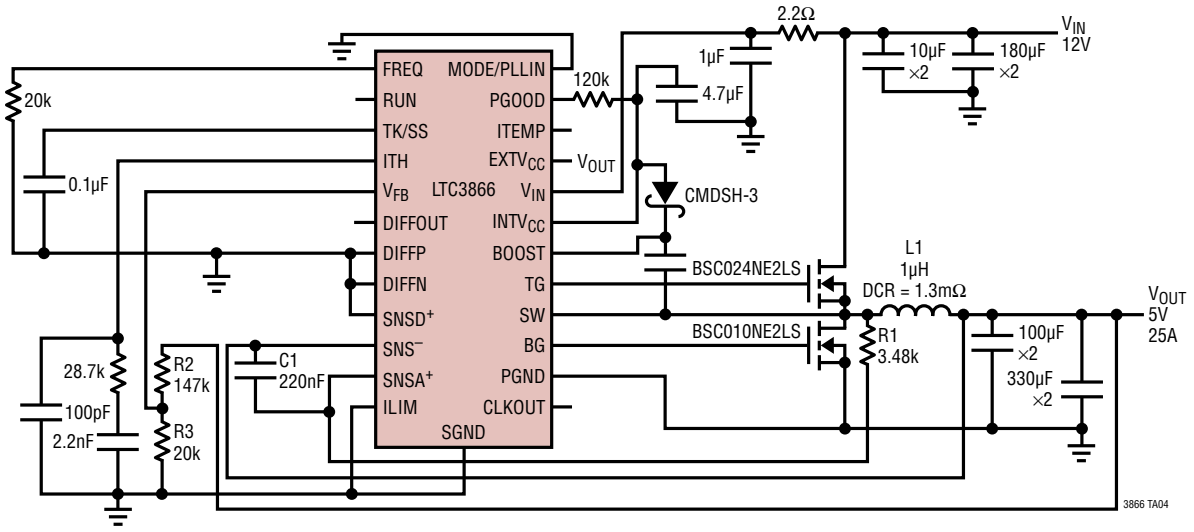


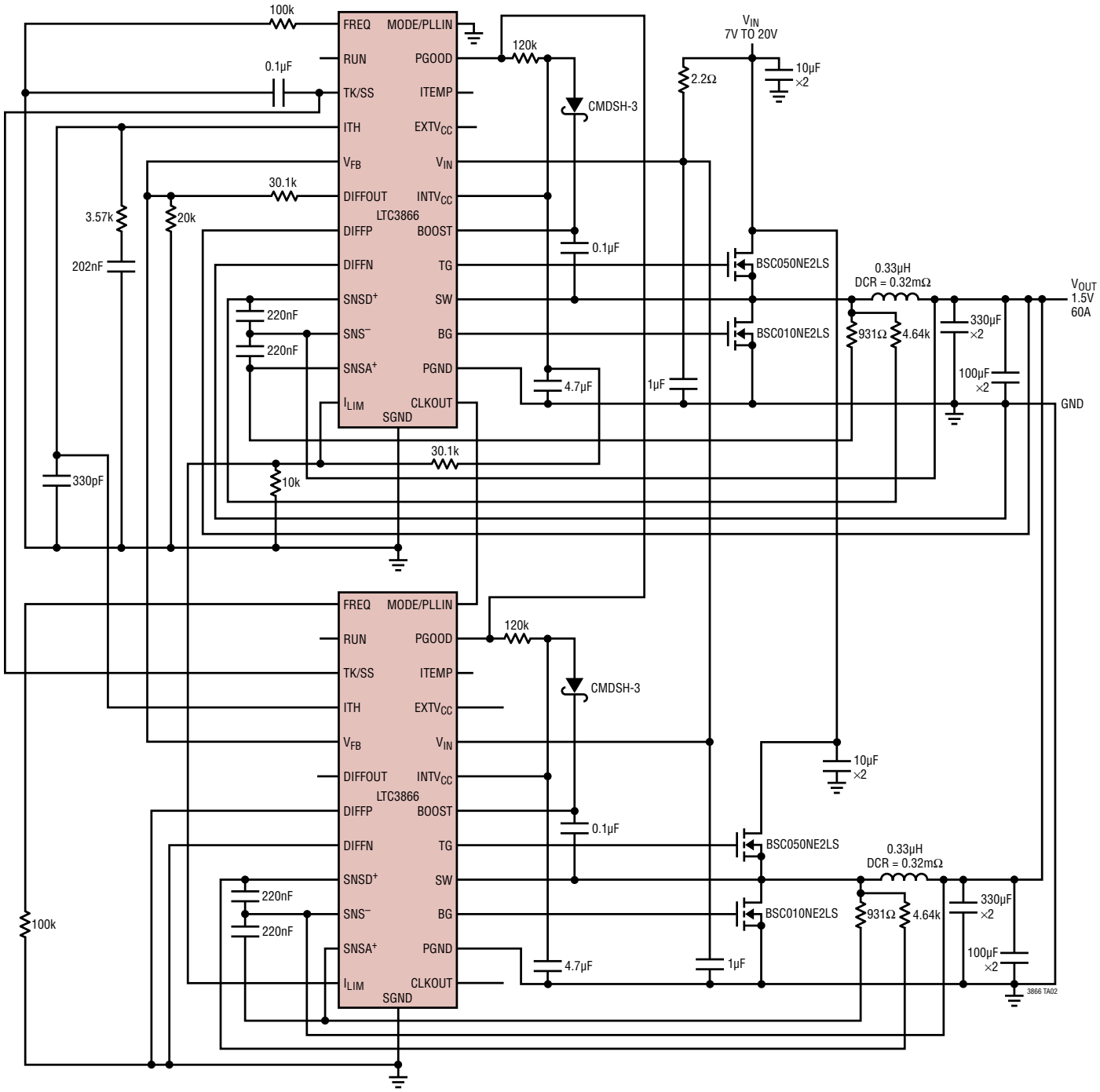
Figure 18. Power Efficiency vs Load Current

5V/25A Step-Down Converter



TYPICAL APPLICATIONS

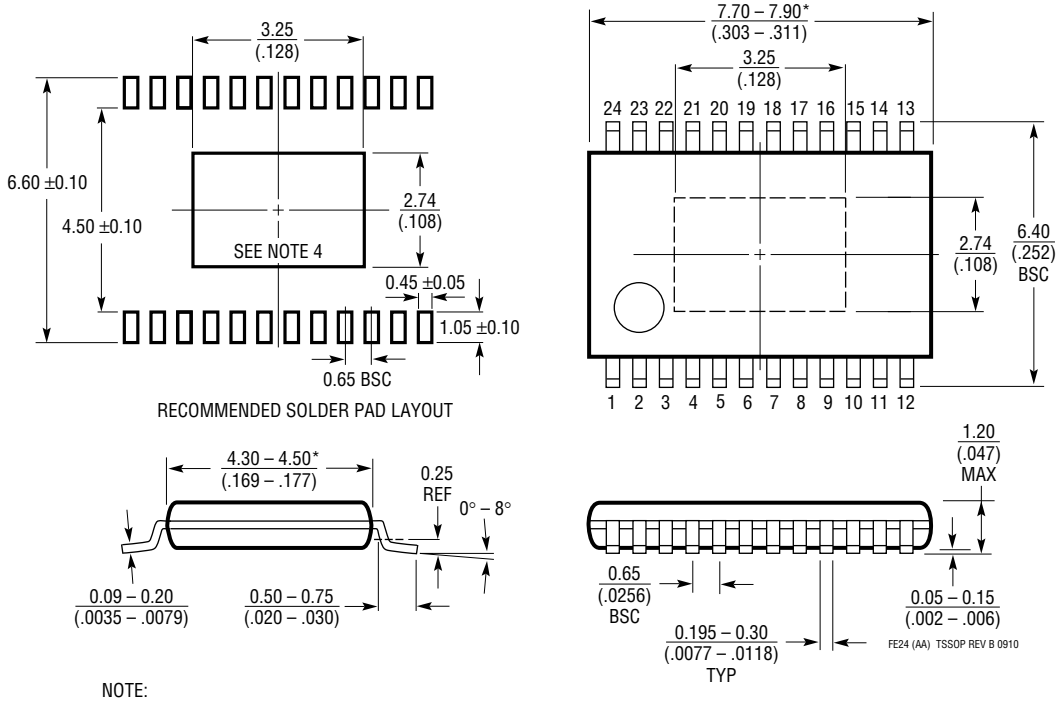
High Efficiency, Dual Phase Very Low DCR Sensing 1.5V/60A Step-Down Supply



PACKAGE DESCRIPTION

Please refer to http://www.linear.com/product/LTC3866#packaging for the most recent package drawings.

FE Package
24-Lead Plastic TSSOP (4.4mm)
(Reference LTC DWG # 05-08-1771 Rev B)
Exposed Pad Variation AA



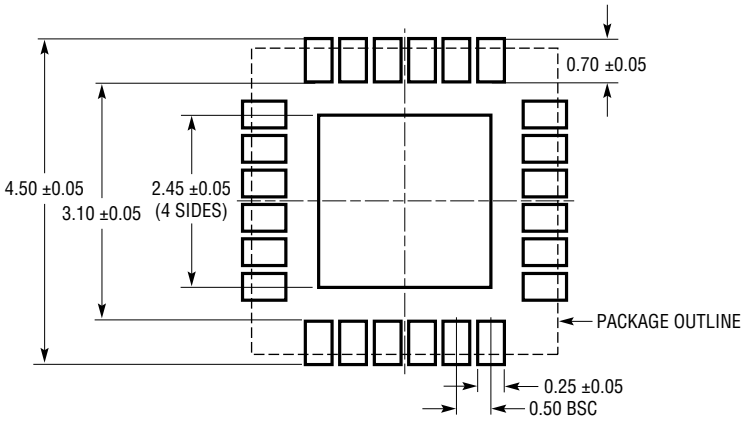
NOTE:
1. CONTROLLING DIMENSION: MILLIMETERS
2. DIMENSIONS ARE IN $\frac{\text{MILLIMETERS}}{(\text{INCHES})}$
3. DRAWING NOT TO SCALE
4. RECOMMENDED MINIMUM PCB METAL SIZE FOR EXPOSED PAD ATTACHMENT

*DIMENSIONS DO NOT INCLUDE MOLD FLASH. MOLD FLASH SHALL NOT EXCEED 0.150mm (.006") PER SIDE

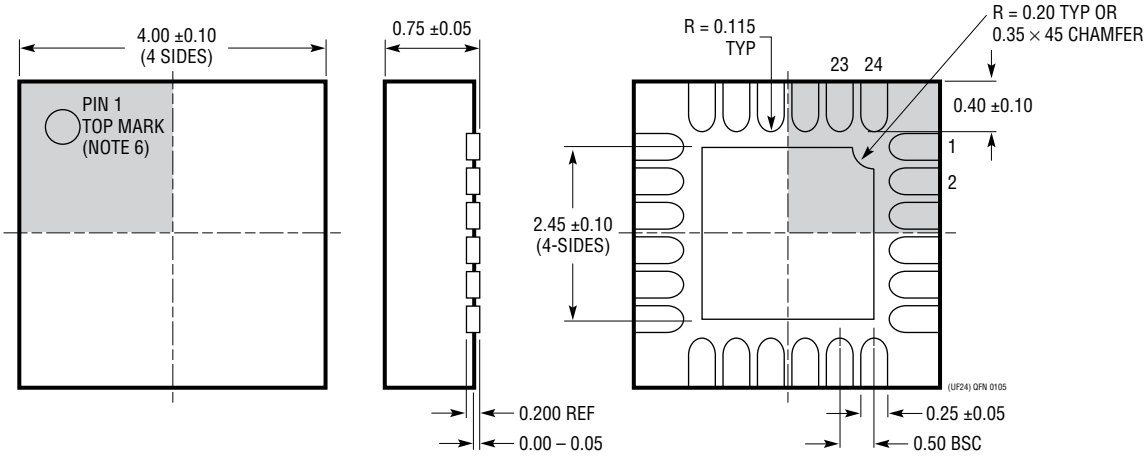
PACKAGE DESCRIPTION

Please refer to http://www.linear.com/product/LTC3866#packaging for the most recent package drawings.

UF Package
24-Lead Plastic QFN (4mm × 4mm)
(Reference LTC DWG # 05-08-1697)



RECOMMENDED SOLDER PAD PITCH AND DIMENSIONS



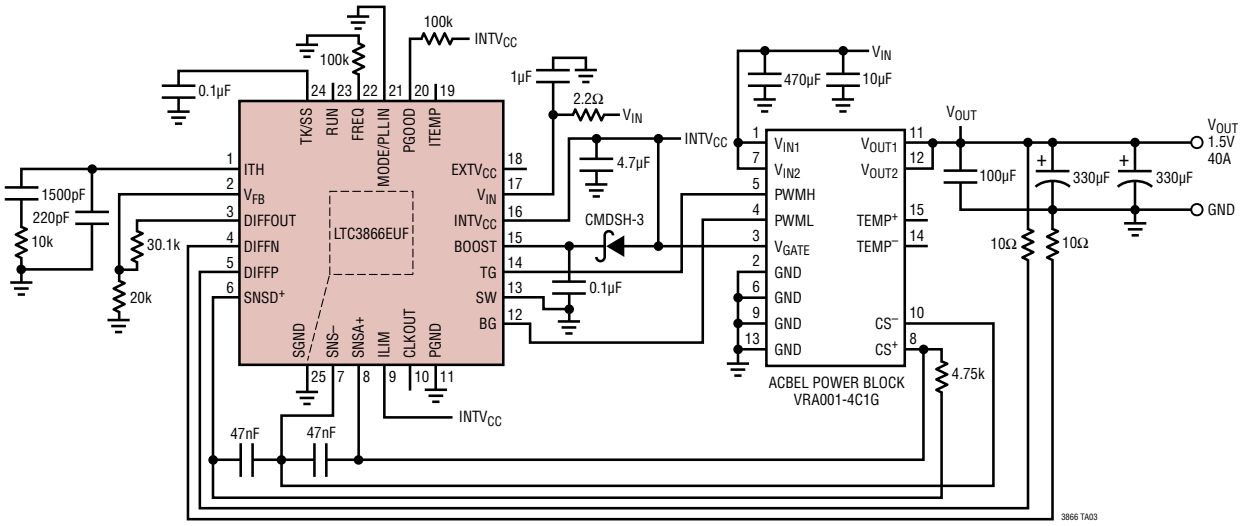
NOTE:
1. DRAWING PROPOSED TO BE MADE A JEDEC PACKAGE OUTLINE MO-220 VARIATION (WGGD-X)—TO BE APPROVED
2. DRAWING NOT TO SCALE
3. ALL DIMENSIONS ARE IN MILLIMETERS
4. DIMENSIONS OF EXPOSED PAD ON BOTTOM OF PACKAGE DO NOT INCLUDE
   MOLD FLASH. MOLD FLASH, IF PRESENT, SHALL NOT EXCEED 0.15mm ON ANY SIDE, IF PRESENT
5. EXPOSED PAD SHALL BE SOLDER PLATED
6. SHADED AREA IS ONLY A REFERENCE FOR PIN 1 LOCATION
   ON THE TOP AND BOTTOM OF PACKAGE

REVISION HISTORY

REV	DATE	DESCRIPTION	PAGE NUMBER
A	08/12	Clarified operating temperatures.	2-5
		Modified the P_D equation thermal resistance value.	5
		Modified the Block Diagram sense amplifier.	10
		Clarified the SNS section values.	14
		Modified the ripple value in the Soft-Start section.	21
		Modified values in the $INTV_{CC}$ and $EXTV_{CC}$ section.	22-23
		Modified the 5V/25A Step-Down Converter circuit schematic.	31
B	10/12	Added "junction" to clarify operating temperature range.	3, 4, 5
C	10/17	Clarified Fault Conditions paragraph	24

TYPICAL APPLICATION

High Efficiency Step-Down Converter with Power Block



RELATED PARTS

PART NUMBER	DESCRIPTION	COMMENTS
LTC3833	Fast Accurate Step-Down DC/DC Controller with Differential Output Sensing	Very Fast Transient Response, $t_{ON(MIN)}$ = 20ns, 4.5V ≤ V_{IN} ≤ 38V, 0.6V ≤ V_{OUT} ≤ 5.5V, TSSOP-20E, 3mm × 4mm QFN-20
LTC3878/LTC3879	No R_{SENSE}™ Constant On-Time Synchronous Step-Down DC/DC Controllers	Very Fast Transient Response, $t_{ON(MIN)}$ = 43ns, 4V ≤ V_{IN} ≤ 38V, 0.6V ≤ V_{OUT} ≤ 0.9V_{IN}, SSOP-16, MSOP-16E, 3mm × 3mm QFN-16
LTC3775	High Frequency Synchronous Voltage Mode Step-Down DC/DC Controller	Very Fast Transient Response, $t_{ON(MIN)}$ = 30ns, 4V ≤ V_{IN} ≤ 38V, 0.6V ≤ V_{OUT} ≤ 0.8V_{IN}, MSOP-16E, 3mm × 3mm QFN-16
LTC3854	Small Footprint Synchronous Step-Down DC/DC Controller	Fixed 400kHz Operating Frequency, 4.5V ≤ V_{IN} ≤ 38V, 0.8V ≤ V_{OUT} ≤ 5.25V, 2mm × 3mm QFN-12
LTC3851A/LTC3851A-1	No R_{SENSE} Wide V_{IN} Range Synchronous Step-Down DC/DC Controllers	PLL Fixed Frequency 250kHz to 750kHz, 4V ≤ V_{IN} ≤ 38V, 0.8V ≤ V_{OUT} ≤ 5.25V, MSOP-16E, 3mm × 3mm QFN-16, SSOP-16
LTC3891	60V, Low I_Q Synchronous Step-Down DC/DC Controller	PLL Capable Fixed Frequency 50kHz to 900kHz, 4V ≤ V_{IN} ≤ 60V, 0.8V ≤ V_{OUT} ≤ 24V, I_Q = 50µA
LTC3856	2-Phase, Single Output Synchronous Step-Down DC/DC Controller with Diff Amp and DCR Temp Compensation	PLL Fixed 250kHz to 770kHz Frequency, 4.5V ≤ V_{IN} ≤ 38V, 0.6V ≤ V_{OUT} ≤ 5.25V
LTC3829	3-Phase, Single Output Synchronous Step-Down DC/DC Controller with Diff Amp and DCR Temp Compensation	PLL Fixed 250kHz to 770kHz Frequency, 4.5V ≤ V_{IN} ≤ 38V, 0.6V ≤ V_{OUT} ≤ 5.25V
LTC3855	2-Phase, Dual Output Synchronous Step-Down DC/DC Controller with Differential Remote Sense	PLL Fixed Frequency 250kHz to 770kHz, 4.5V ≤ V_{IN} ≤ 38V, 0.6V ≤ V_{OUT} ≤ 12.5V
LTC3860	Dual, Multiphase, Synchronous Step-Down DC/DC Controller with Diff Amp and Three-State Output Drive	Operates with Power Blocks, DRMOS Devices or External Drivers/ MOSFETs, 3V ≤ V_{IN} ≤ 24V, $t_{ON(MIN)}$ = 20ns
LTC3869/LTC3869-2	2-Phase, Dual Output Synchronous Step-Down DC/DC Controllers, with Accurate Multiphase Current Matching	PLL Fixed Frequency 250kHz to 780kHz, 4V ≤ V_{IN} ≤ 30V, 0.6V ≤ V_{OUT} ≤ 12.5V, 4mm × 5mm QFN-28, SSOP-28
LTC3867	Synchronous Step-Down DC/DC Controller with Nonlinear Control and Remote Sense	Fast Transient Response, $t_{ON(MIN)}$ = 65ns, 4V ≤ V_{IN} ≤ 38V, 0.6V ≤ V_{OUT} ≤ 14V, 4mm × 4mm QFN-24

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- Analog Devices Inc. Information

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- Cost Control Management
- Shortage Management
- Alternative Solution
- Excess Inventory Management