NB680

28V, Low Iq, High Current, Fixed 3.3V-8A Synchronous Buck Converter with 100 mA LDO

DESCRIPTION

The NB680 is a fully integrated, high-frequency, synchronous, rectified, step-down, switch-mode converter with a fixed 3.3 V Vout. It offers a very compact solution to achieve an 8 A continuous output current and a 10 A peak output current over a wide input supply range with excellent load and line regulation.

The NB680 operates at high efficiency over a wide output current load range based on MPS proprietary switching loss reduction technology and internal low Ron power MOSFETs.

Adaptive constant-on-time (COT) control mode provides fast transient response and eases loop stabilization. The DC auto-tune loop provides good load and line regulation.

NB680 provides a fixed 3.3 V LDO, which can power the external peripheries, such as the keyboard controller in the laptop.

Also, a 250 kHz CLK is available; its output can drive an external charge pump, generating gate drive voltage for the load switches without reducing the main converter's efficiency.

Full protection features include OC limit, OVP, UVP, and thermal shutdown.

NB680 requires a minimum number of external components and is available in a QFN 2mm x 3mm package.

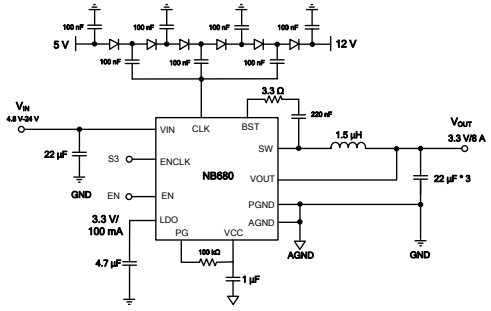
FEATURES

- Wide 4.8 V to 28 V Operating Input Range
- Fixed 3.3 V Vout
- Ultrasonic Mode with Fs over 25 kHz
- 100 µA Low Quiescent Current
- 8 A Continous Output Current
- 10 A Peak Output Current
- Adaptive COT for Fast Transient
- DC Auto-Tune Loop
- Stable with POSCAP and Ceramic Output Capacitors
- 250 kHz CLK for External Charge Pump
- Built-In 3.3 V, 100 mA LDO with Switch Over
- 1% Reference Voltage
- Internal Soft Start
- Output Discharge
- 700 kHZ Switching Frequency
- OCL, OVP, UVP, and Thermal Shutdown.
- Latch-Off Reset via EN or Power Cycle.
- QFN 2mm x 3mm Package

APPLICATIONS

- Laptop Computers
- Tablet PCs
- Networking Systems
- Servers
- Personal Video Recorders
- Flat Panel Television and Monitors
- Distributed Power Systems

TYPICAL APPLICATION



ORDERING INFORMATION

Part Number*	Package	Top Marking
NB680GD	QFN-12 (2mm x 3mm)	*See Below*

* For Tape & Reel, add suffix –Z (e.g. NB680GD–Z)

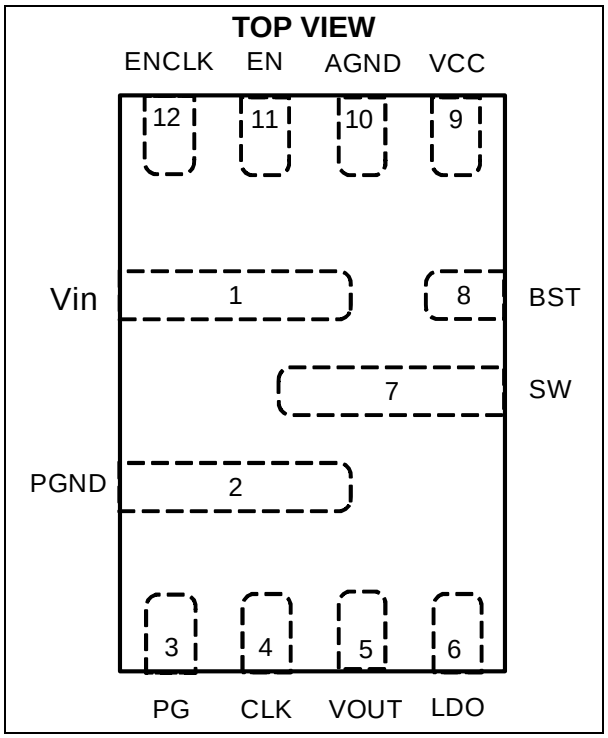
TOP MARKING

$\overline{\text{ALVY}}$

LLL

ALV: Product code of NB680GD
Y: Year code
LLL: Lot number

PACKAGE REFERENCE



ABSOLUTE MAXIMUM RATINGS (1)

Supply voltage (V_{IN}) ..... 28 V
V_{SW} (DC) ..... -1 V to 26 V
V_{SW} (25 ns) ..... -3.6 V to 28 V
V_{BST} ..... V_{SW} + 4.5 V
All other pins ..... -0.3 V to +4.5 V
Continuous power dissipation (T_A=+25°C) (2)
QFN-12 (2mm x 3mm) ..... 1.8 W
Junction temperature ..... 150°C
Lead temperature ..... 260°C
Storage temperature ..... -65°C to +150°C

Recommended Operating Conditions (3)

Supply voltage ..... 4.8 V to 24 V
Operating junction temp. (T_J). .. -40°C to +125°C

Thermal Resistance (4)

	θ_{JA}	θ_{JC}	
QFN-12 (2mm x 3mm)	70	15	°C/W

NOTES:

1) Exceeding these ratings may damage the device.
2) The maximum allowable power dissipation is a function of the maximum junction temperature T_J(MAX), the junction-to-ambient thermal resistance θ_{JA}, and the ambient temperature T_A. The maximum allowable continuous power dissipation at any ambient temperature is calculated by P_D(MAX)=(T_J(MAX)-T_A)/θ_{JA}. Exceeding the maximum allowable power dissipation produces an excessive die temperature, causing the regulator to go into thermal shutdown. Internal thermal shutdown circuitry protects the device from permanent damage.
3) The device is not guaranteed to function outside of its operating conditions.
4) Measured on JESD51-7, 4-layer PCB.

ELECTRICAL CHARACTERISTICS

V_{IN} = 12 V, T_J = 25°C, unless otherwise noted.

Parameters	Symbol	Condition	Min	Typ	Max	Units
Supply current						
Supply current (quiescent)	I_{IN}	V_{EN} = V_{ENCLK} = 3.3 V,V_{OUT} = 3.5 V		120	140	μA
Supply current (standby)	I_{IN}	V_{EN} = V_{ENCLK} = 0 V, I_{LDO} = 0 A		60	80	μA
MOSFET						
High-side switch on resistance	HS_{RDS-ON}			25		mΩ
Low-side switch on resistance	LS_{RDS-ON}			12		mΩ
Switch leakage	SW_{LKG}	V_{EN} = 0 V, V_{SW} = 0 V		0	1	μA
Current limit						
Low-side valley current limit	I_{LIMIT}		10	11	12	A
Switching frequency and timer						
Switching frequency	F_S			700		kHz
Constant on timer	Ton	Vin = 6.6 V	600	710	820	ns
Minimum on time(5)	T_{ON_Min}			50		ns
Minimum off time(5)	T_{OFF_Min}			220		ns
Ultrasonic mode						
Ultrasonic mode operation period	T_{USM}		20	30	40	μs
Over-voltage and under-voltage protection						
OVP threshold	V_{OVP}		117%	122%	127%	V_{REF}
UVP-1 threshold	V_{UVP-1}		70%	75%	80%	V_{REF}
UVP-1 foldback timer	T_{UVP-1}			32		μs
UVP-2 threshold	V_{UVP-2}		45%	50%	55%	V_{REF}
Reference and soft start						
Vout REF voltage	V_{OUT_REF}		3.27	3.3	3.33	V
Soft-start time	T_{SS}			2	2.5	ms
Enable and UVLO						
Enable rising threshold	V_{EN_H}		1.18	1.28	1.38	V
Enable hysteresis	V_{EN-HYS}			150		mV
EN high limit @USM	$V_{EN\_H\_USM}$				1.8	V
EN low limit @normal	$V_{EN\_L\_Normal}$		2.6			V
Enable input current	I_{EN}	V_{EN} = 2 V		4		μA
		V_{EN} = 0 V		0		
VIN under-voltage lockout threshold rising	VIN_{VTH}			4.4	4.7	V
VIN under-voltage lockout threshold hysteresis	VIN_{HYS}			450		mV

ELECTRICAL CHARACTERISTICS *(continued)*

V_{IN} = 12 V, T_J = 25°C, unless otherwise noted.

Parameters	Symbol	Condition	Min	Typ	Max	Units
CLK output						
CLK output high-level voltage	V_{CLKH}	I_{Vclk} = -10 mA	3.1	3.3	3.5	V
CLK output low-level voltage	V_{CLKL}	I_{Vclk} = 10 mA	0	0.05	0.1	V
CLK frequency	F_{CLK}	T_J = 25°C		250		kHz
LDO regulator						
LDO regulator	V_{LDO}	V_{EN} = 0 V,	3.22	3.3	3.38	V
LDO load regulation		V_{EN} = 0 V, LDO load =100 mA		2		%
LDO current limit	I_{LDO_Limit}	V_{EN} = 0 V, V_{LDO} = 3 V		135		mA
Switch Rdson(5)	R_{Switch}	I_{LDO} = 50 mA		0.9	1.2	Ω
VCC regulator						
VCC regulator	V_{CC}		3.5	3.6	3.7	V
VCC load regulation		Icc = 5 mA		5		%
Power good						
PG when FB rising (good)	$PG_{\_Rising(Good)}$	VFB rising, percentage of VFB		95		%
PG when FB falling (fault)	$PG_{\_Falling(Fault)}$	VFB falling, percentage of VFB		85		
PG when FB rising (fault)	$PG_{\_Rising(Fault)}$	VFB rising, percentage of VFB		115		
PG when FB falling (good)	$PG_{\_Falling(Good)}$	VFB falling, percentage of VFB		105		
Power good low to high delay	PG_{Td}			750		µs
EN low to power good low delay	$PG_{Td\_EN\ low}$				5	µs
Power good sink current capability	V_{PG}	Sink 4 mA			0.4	V
Power good leakage current	I_{PG_LEAK}	V_{PG} = 3.3 V			5	µA
Thermal protection						
Thermal shutdown(5)	T_{SD}			140		°C
Thermal shutdown hysteresis(5)	T_{SD-HYS}			25		°C

NOTE:

5) Guaranteed by design.

PIN FUNCTIONS

NB680

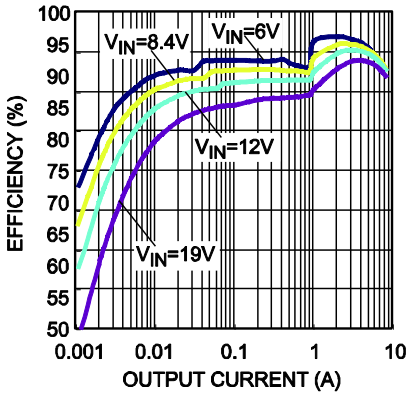
PIN #	Name	Description
1	VIN	**Supply voltage.** VIN supplies power for the internal MOSFET and regulator. The NB680 operates from a 4.8 V to 24 V input rail. An input capacitor is needed to decouple the input rail. Use wide PCB traces and multiple vias to make the connection. Apply at least two layers for this input trace.
2	PGND	**Power ground**. Use wide PCB traces and enough vias to handle the load current to make the connection. Make the PGND trace to the Vin decoupling capacitor as wide as possible.
3	PG	**Power good output**. The output of PG is an open-drain signal. It is high if the output voltage is higher than 95 percent of the nominal voltage or lower than 105 percent of the nominal voltage.
4	CLK	**250 kHZ CLK output to drive the external charge pump.**
5	VOUT	**Output voltage of the buck regulator sense.** Connect VOUT to the output capacitor of the regulator directly. Also, VOUT acts as the input of the internal LDO switch over-power input. Keep the VOUT sensing trace far away from the SW node. Avoid vias on the VOUT sensing trace. A >25 mil trace is required.
6	LDO	**Internal LDO output.** Decouple with a minimum 4.7 µF ceramic capacitor as close to LDO as possible. X7R or X5R grade dielectric ceramic capacitors are recommended for their stable temperature characteristics. Once the PG of the output voltage of the buck regulator is ready, it switches over to the LDO output to avoid power loss.
7	SW	**Switch output.** Connect SW to the inductor and bootstrap capacitor. SW is driven up to the VIN voltage by the high-side switch during the on-time of the PWM duty cycle. The inductor current drives SW negative during the off-time. The on resistance of the low-side switch and the internal diode fixes the negative voltage. Use wide and short PCB traces to make the connection. Try to minimize the area of the SW pattern.
8	BST	**Bootstrap.** A capacitor connected between SW and BS is required to form a floating supply across the high-side switch driver.
9	VCC	**Internal VCC LDO output.** The driver and control circuits are powered from this voltage. Decouple with a minimum 1 µF ceramic capacitor as close to VCC as possible. X7R or X5R grade dielectric ceramic capacitors are recommended for their stable temperature characteristics.
10	AGND	**Signal logic ground**. Make a Kelvin connection to PGND.
11	EN	**Buck enable pin.** EN is a digital input that turns the buck regulator on or off. Connect EN to 3V3 through a pull-up resistor, or connect EN with a resistive voltage divider to Vin for automatic start-up. Do NOT float EN. EN also sets USM. When EN is in the range of 1.38 V to1.8 V, it enters USM. If EN is in the range of 2.6 V to 3.6 V, it operates in normal mode.
12	ENCLK	**CLK enable pin.** The CLK Pin can control the charge pump CLK between S0 and S3/S4.

TYPICAL PERFORMANCE CHARACTERISTICS

V_{IN} = 12 V, V_{OUT} = 3.3 V, L = 1.5 µH/10 mΩ, F_S = 700 kHz, T_J=+25°C, unless otherwise noted.

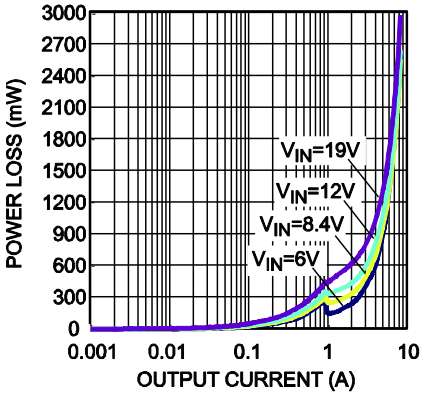
Efficiency vs. Load Current

V_{OUT}=3.3V, F_{SW}=700kHz, L=1.5µH, DCR=10mΩ, R_{BST}=3.3Ω, C_{OUT}=3x22µF



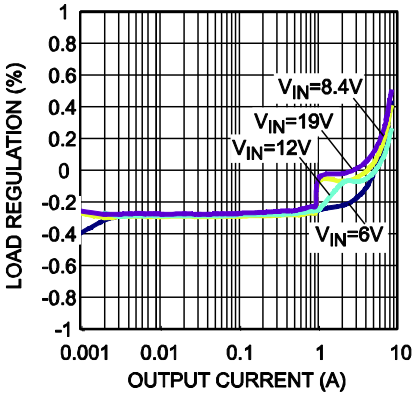
Power Loss

V_{OUT}=3.3V, F_{SW}=700kHz, L=1.5µH, DCR=10mΩ, R_{BST}=3.3Ω, C_{OUT}=3x22µF



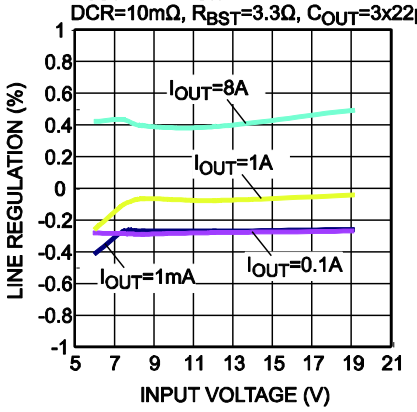
Load Regulation

V_{OUT}=3.3V, F_{SW}=700kHz, L=1.5µH, DCR=10mΩ, R_{BST}=3.3Ω, C_{OUT}=3x22µF

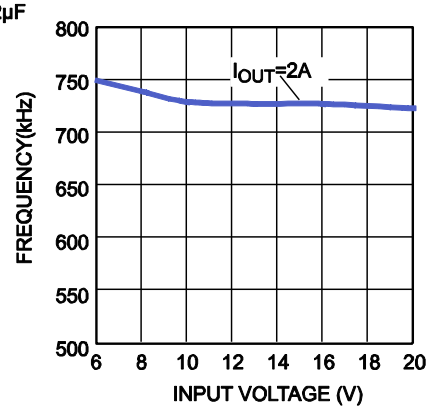


Line Regulation

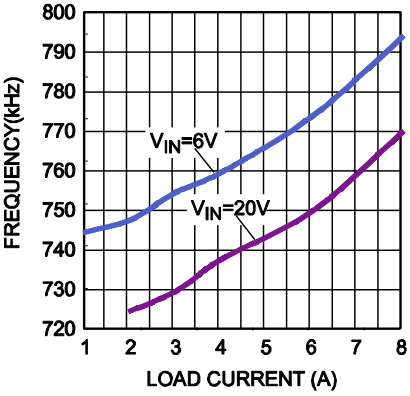
V_{OUT}=3.3V, F_{SW}=700kHz, L=1.5µH, DCR=10mΩ, R_{BST}=3.3Ω, C_{OUT}=3x22µF



Switch Frequency vs. Input Voltage

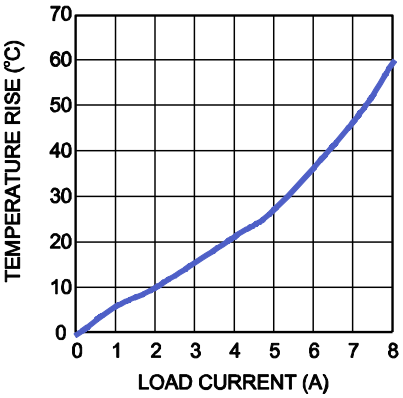


Switch Frequency vs. Load Current



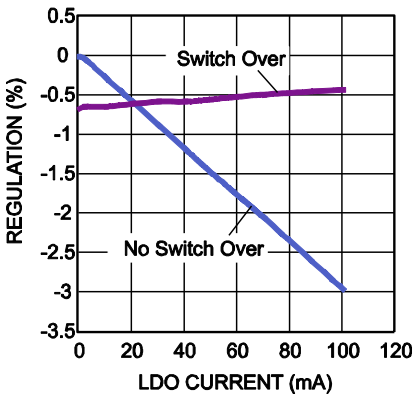
Thermal Test

V_{IN}=19V



LDO Load Regualtion

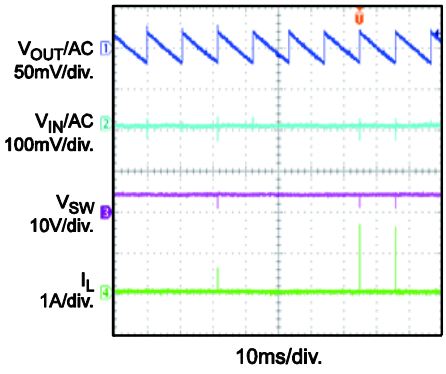
F_{SW}=700kHz



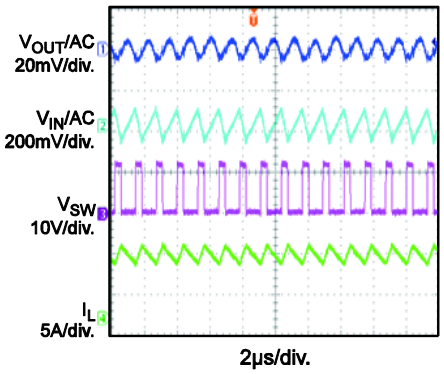
TYPICAL PERFORMANCE CHARACTERISTICS

V_{IN} = 12 V, V_{OUT} = 3.3 V, L = 1.5 µH/10 mΩ, F_S = 700 kHz, T_J=+25°C, unless otherwise noted.

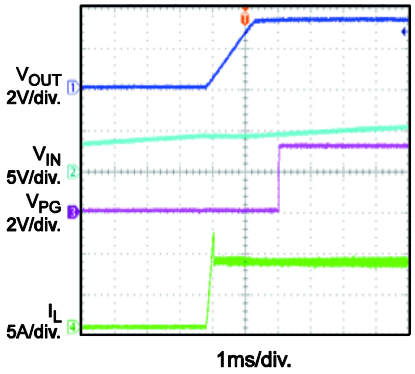
Input/Output Voltage Ripple
I_{OUT} = 0A



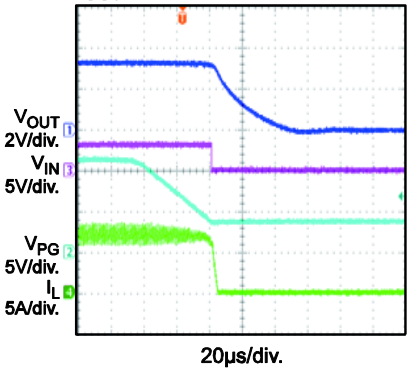
Input/Output Voltage Ripple
I_{OUT} = 8A



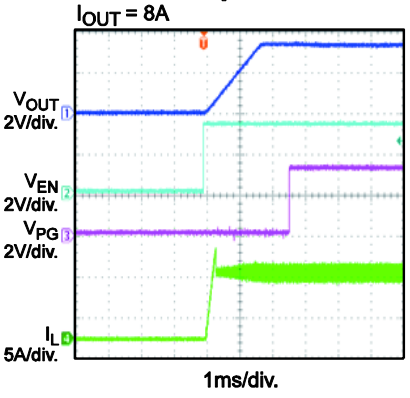
Power Good through VIN Start-Up
I_{OUT} = 8A



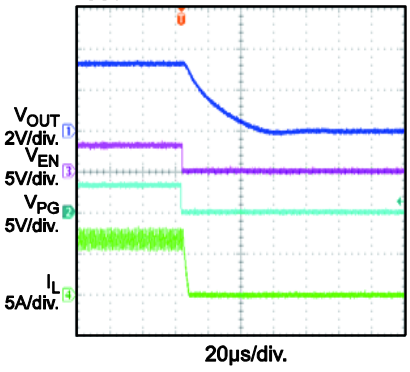
Power Good through VIN Shutdown
I_{OUT} = 8A



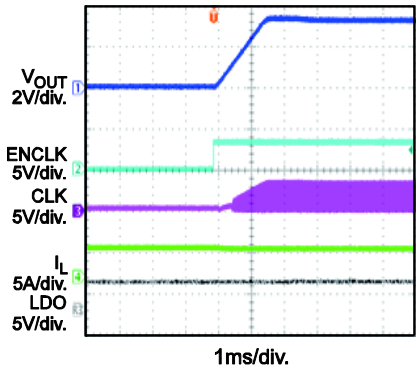
Power Good through EN Start-Up
I_{OUT} = 8A



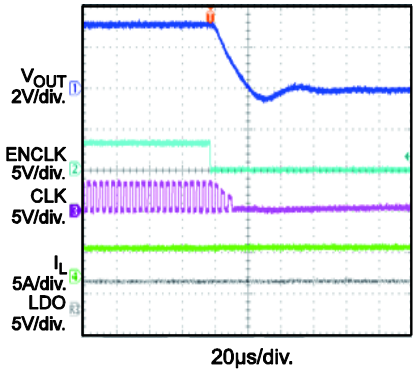
Power Good through EN Shutdown
I_{OUT} = 8A



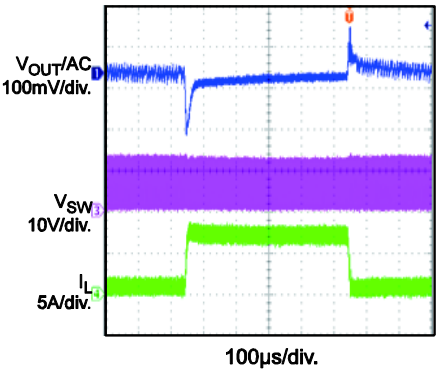
CLK with ENCLK On
I_{OUT}=8A



CLK with ENCLK Off
I_{OUT}=8A



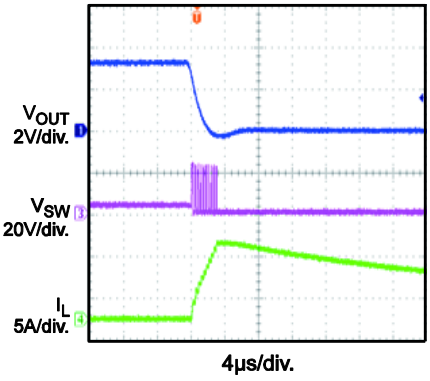
Load Transient
V_{IN}=12V, L=1.5µH, C_{OUT}=66µF
I_{OUT}=0.8-7.2A @ 1.6A/µs



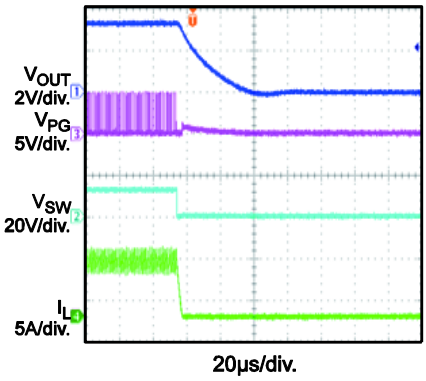
TYPICAL PERFORMANCE CHARACTERISTICS

V_{IN}=12 V, V_{OUT} =3.3 V, L=1.5 µH/10 mΩ, F_S=700 kHz, T_J=+25°C, unless otherwise noted.

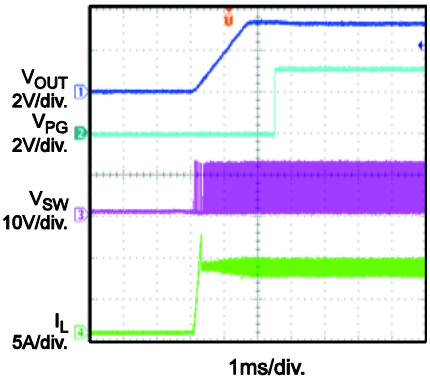
Short-Circuit Protection
V_{IN} = 22V



Thermal Shutdown
V_{IN} = 19V, I_{OUT} = 8A



Thermal Recovery
V_{IN} = 19V, I_{OUT} = 8A



FUNCTIONAL BLOCK DIAGRAM

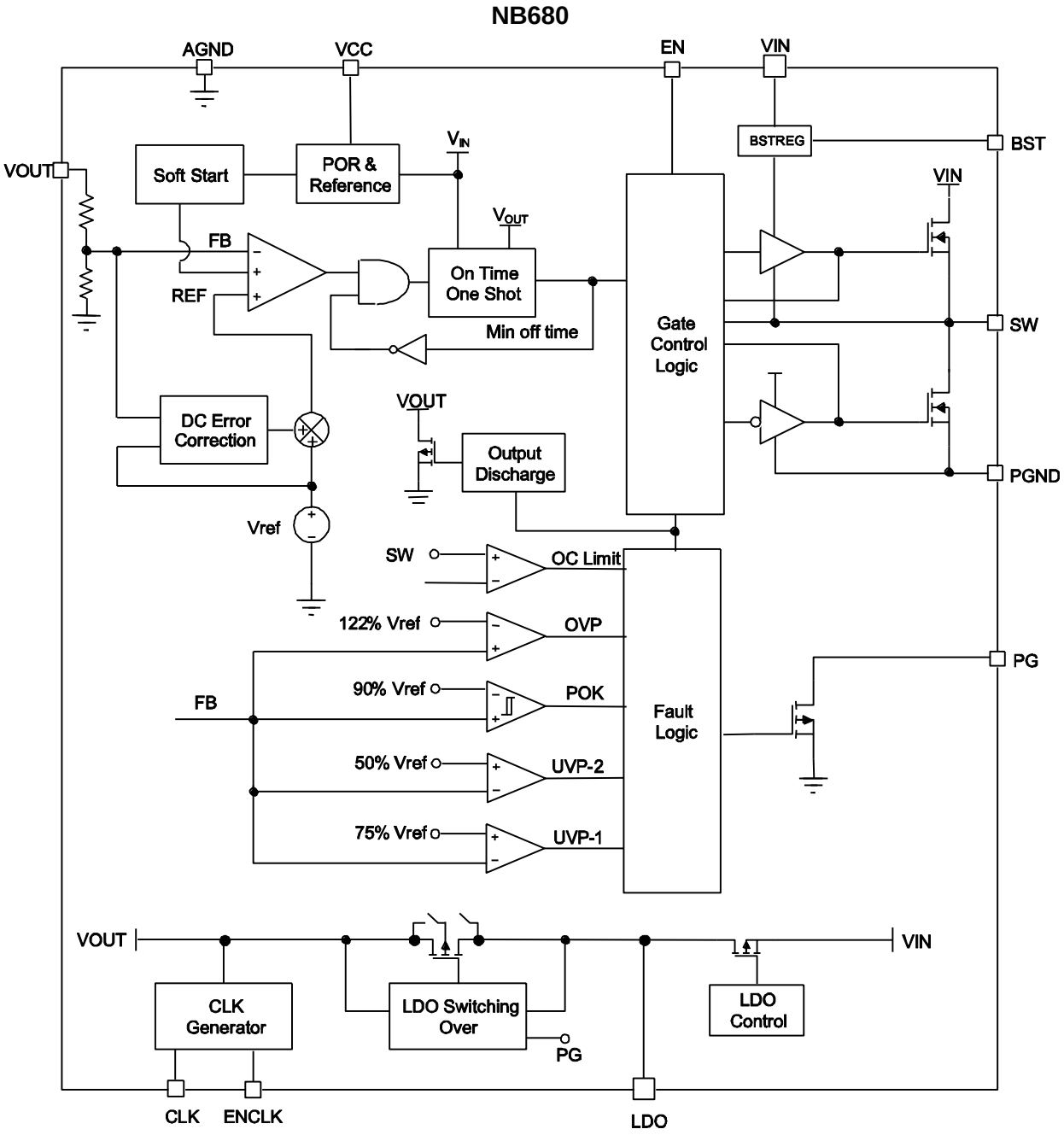


Figure 1—Functional block diagram

OPERATION

PWM Operation

The NB680 is a fully integrated, synchronous, rectified, step-down, switch-mode converter with a fixed 3.3 V output. Constant-on-time (COT) control provides fast transient response and eases loop stabilization. At the beginning of each cycle, the high-side MOSFET (HS-FET) is turned on when the feedback voltage (V_{FB}) is below the reference voltage (V_{REF}), which indicates insufficient output voltage. The on period is determined by the output voltage and the input voltage to make the switching frequency fairly constant over the input voltage range.

After the on period elapses, the HS-FET is turned off or enters an off state. It is turned on again when V_{FB} drops below V_{REF}. By repeating operation this way, the converter regulates the output voltage. The integrated low-side MOSFET (LS-FET) is turned on when the HS-FET is in its off state to minimize the conduction loss. There is a dead short between the input and GND if both the HS-FET and the LS-FET are turned on at the same time (shoot-through). In order to avoid shoot-through, a dead time (DT) is generated internally between the HS-FET off and the LS-FET on period or the LS-FET off and the HS-FET on period.

Internal compensation is applied for COT control for stable operation even when ceramic capacitors are used as output capacitors. This internal compensation improves the jitter performance without affecting the line or load regulation.

CCM Operation

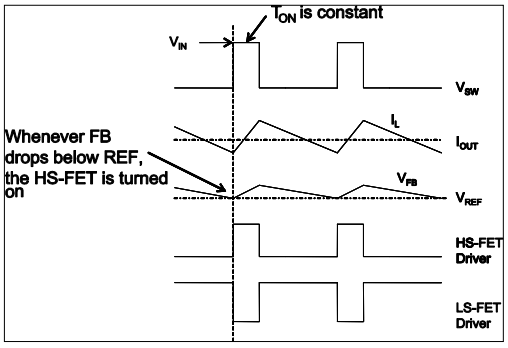


Figure 2—CCM operation

Continuous conduction mode (CCM) occurs when the output current is high, and the inductor current is always above zero amps (see Figure 2). When V_{FB} is below V_{REF}, the HS-FET is turned on for a fixed interval. When the HS-FET is turned off, the LS-FET is turned on until the next period.

In CCM operation, the switching frequency is fairly constant (PWM mode).

DCM Operation

With the load decreases, the inductor current will decrease as well. Once the inductor current reaches zero, the device transitions from CCM to discontinuous conduction mode (DCM).

DCM operation is shown in Figure 3. When V_{FB} is below V_{REF}, the HS-FET is turned on for a fixed interval, which is determined by the one-shot on timer. See Equation (1). When the HS-FET is turned off, the LS-FET is turned on until the inductor current reaches zero. In DCM operation, the V_{FB} does not reach V_{REF} when the inductor current is approaching zero. The LS-FET driver turns into tri-state (high Z) whenever the inductor current reaches zero. A current modulator takes over the control of the LS-FET and limits the inductor current to less than -1 mA. Hence, the output capacitors discharge slowly to GND through the LS-FET. As a result, the efficiency during a light-load condition is improved greatly. The HS-FET is not turned on as frequently during a light-load condition as it is during a heavy-load condition (skip mode).

At a light-load or no-load condition, the output drops very slowly, and the NB680 reduces the switching frequency naturally, achieving high efficiency at light load.

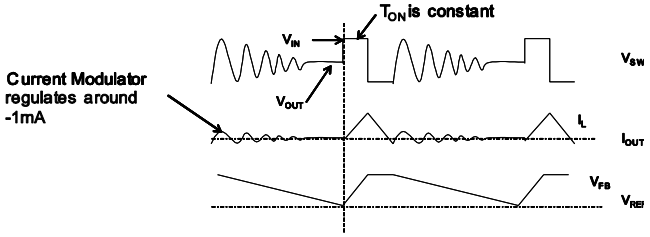


Figure 3—DCM Operation

As the output current increases from the light-load condition, the time period within which the current modulator regulates becomes shorter. The HS-FET is turned on more frequently. Hence, the switching frequency increases accordingly. The output current reaches the critical level when the current modulator time is zero. The critical level of the output current is determined with Equation (1):

$$I_{OUT} = \frac{(V_{IN} - V_{OUT}) \times V_{OUT}}{2 \times L \times F_{SW} \times V_{IN}} \quad (1)$$

The part enters PWM mode once the output current exceeds the critical level. After that, the switching frequency stays fairly constant over the output current range.

DC Auto-Tune Loop

The NB680 applies a DC auto-tune loop to balance the DC error between V_{FB} and V_{REF} by adjusting the comparator input REF to make V_{FB} always follow V_{REF}. This loop is quite slow, so it improves the load and line regulation without affecting the transient performance. The relationship between V_{FB}, V_{REF}, and REF is shown in Figure 4.

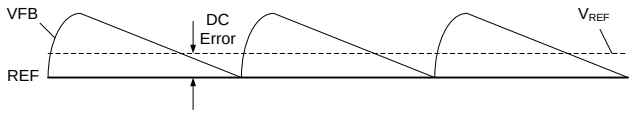


Figure 4—DC auto-tune loop operation

Ultrasonic Mode (USM)

Ultrasonic mode (USM) keeps the switching frequency above an audible frequency area during light-load or no-load conditions. Once the part detects that both the HS-FET and the LS-FET are off (for about 32 µs), it shrinks the Ton to keep Vout under regulation with optimal efficiency. If the load continues to decrease, the part discharges Vout to make sure FB is less than 102 percent of the internal reference. The HS-FET turns on again once the internal FB reaches VREF and then stops switching.

USM is selected by the EN voltage level. When EN is in the range of 1.38 V to 1.8 V, it enters USM. If EN is in the range of 2.6 V to 3.6 V, it enters normal mode.

Configuring the EN Control

The NB680 has two enable pins to control the on/off of the internal regulators and CLK.

For NB680, the 3V3 LDO is always on when Vin passes UVLO. EN controls both the buck and the CLK. Once EN is on, the ENCLK is able to control the CLK on/off. See Table1 for the NB680 EN logic control.

Table 1—ENCLK/EN control

State	ENCLK	EN	VCC	VOUT	CLK	3V3 LDO
S0	1	1	ON	ON	ON	ON
S3	0	1	ON	ON	OFF	ON
S3/S5	0	0	ON	OFF	OFF	ON
Others	1	0	ON	OFF	OFF	ON

For automatic start-up, EN can be pulled up to the input voltage through a resistive voltage divider. Refer to the “UVLO Protection” section for more details.

Soft Start (SS)

The NB680 employs a soft-start (SS) mechanism to ensure smooth output during power-up. When EN goes high, the internal reference voltage ramps up gradually; hence, the output voltage ramps up smoothly as well. Once the reference voltage reaches the target value, the soft start finishes, and the part enters steady-state operation.

If the output is pre-biased to a certain voltage during start-up, the IC disables the switching of both the high-side and the low-side switches until the voltage on the internal reference exceeds the sensed output voltage at the internal FB node.

3.3 V Linear Regulator

There is a built-in 100 mA standby linear regulator with a fixed output at 3.3 V, controlled by VIN UVLO. Once Vin passes its UVLO, it is on. The 3.3 V LDO is not controlled by EN or ENCLK. This LDO is intended mainly for an auxiliary 3.3 V supply for the notebook system in standby mode.

Add a ceramic capacitor with a value between 4.7 µF and 22 uF placed close to the LDO pins to stabilize the LDOs.

LDO Switch Over

When the output voltage becomes higher than 3.15 V and the power good (PG) is ok, the internal LDO regulator is shut off, and the LDO output is connected to VOUT by the internal switch-over MOSFET, reducing power loss from the LDO.

CLK for Charge Pump

The 250 kHz CLK signal drives an external charge pump circuit to generate approximately 10 V-12 V DC voltage. The CLK voltage becomes available once Vin is higher than the UVLO threshold, and ENCLK is pulled high (see Figure 5).

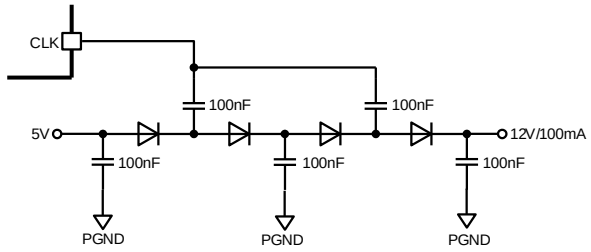


Figure 5—Charge pump circuit

Power Good (PG)

The NB680 has power-good (PG) output used to indicate whether the output voltage of the buck regulator is ready. PG is the open drain of a MOSFET. It should be connected to V_{CC} or another voltage source through a resistor (e.g. 100k). After the input voltage is applied, the MOSFET is turned on so that PG is pulled to GND before SS is ready. Once FB voltage rises to 95 percent of the REF voltage, PG is pulled high after 750 µs.

When the FB voltage drops to 85 percent of the REF voltage, PG is pulled low.

Over-Current Protection (OCP)

NB680 has cycle-by-cycle over-current limiting control. The current-limit circuit employs a "valley" current-sensing algorithm. The part uses the Rds(on) of the LS-FET as a current-sensing element. If the magnitude of the current-sense signal is above the current-limit threshold, the PWM is not allowed to initiate a new cycle.

The trip level is fixed internally. The inductor current is monitored by the voltage between GND and SW. GND is used as the positive current sensing node, so GND should be connected to the source terminal of the bottom MOSFET.

Since the comparison is done during the HS-FET off state and the LS-FET on state, the OC trip level sets the valley level of the inductor current. Thus, the load current at the over-current threshold (I_{OC}) is calculated with Equation (2):

$$I_{OC} = I\_limit + \frac{\Delta I_{inductor}}{2} \quad (2)$$

In an over-current condition, the current to the load exceeds the current to the output capacitor; thus, the output voltage tends to fall off. Eventually, it ends up crossing the under-voltage protection threshold and shuts down. Fault latching can be reset by EN going low or the power cycling of VIN.

Over/Under-Voltage Protection (OVP/UVP)

NB680 monitors the output voltage to detect over and under voltage. Once the feedback voltage becomes higher than 122 percent of the target voltage, the OVP comparator output goes high, and the circuit latches as the HS-FET driver turns off, and the LS-FET driver turns on, acting as an -1.8 A current source.

To protect the part from damage, there is an absolute OVP on VOUT (usually set at 6.2 V). Once Vout > 6.2 V, the controller turns off both the HS-FET and the LS-FET. This protection is not latched off and will keep switching once the Vout returns to its normal value.

When the feedback voltage drops below 75 percent of the Vref but remains higher than 50 percent of the Vref, the UVP-1 comparator output goes high, and the part latches if the FB voltage remains in this range for about 32 µs (latching the HS-FET off and the LS-FET on). The LS-FET remains on until the inductor current hits zero. During this period, the valley current limit helps control the inductor current.

When the feedback voltage drops below 50 percent of the Vref, the UVP-2 comparator output goes high, and the part latches off directly after the comparator and logic delay (latching the HS-FET off and the LS-FET on). The LS-FET remains on until the inductor current hits zero. Fault latching can be re-set by EN going low or the power cycling of VIN.

UVLO Protection

The part starts up only when the Vin voltage is higher than the UVLO rising threshold voltage. The part shuts down when the VIN is lower than the Vin falling threshold. The UVLO protection is non-latch off. Fault latching can be re-set by EN going low or the power cycling of VIN.
If an application requires a higher under-voltage lockout (UVLO), use EN to adjust the input voltage UVLO by using two external resistors (see Figure 6).

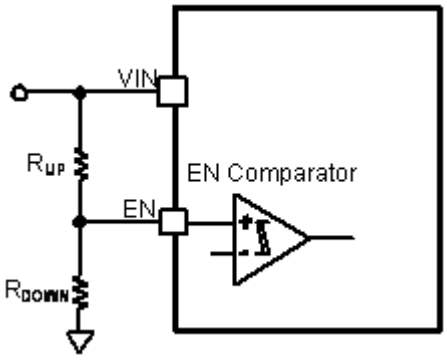


Figure 6—Adjustable UVLO

To avoid too much sink current on EN, the EN resistor (Rup) is usually in the range of 1 M-2 MΩ. A typical pull-up resistor is 2 MΩ.

Thermal Shutdown

Thermal shutdown is employed in the NB680. The junction temperature of the IC is monitored internally. If the junction temperature exceeds the threshold value (140ºC, typically), the converter shuts off. This is a non-latch protection. There is about 25ºC hysteresis. Once the junction temperature drops to about 115ºC, it initiates a SS.

Output Discharge

NB680 discharges the output when EN is low, or the controller is turned off by the protection functions UVP, OCP, OCP, OVP, UVLO, and thermal shutdown. The part discharges outputs using an internal MOSFET.

APPLICATION INFORMATION

Input Capacitor

The input current to the step-down converter is discontinuous, and therefore requires a capacitor to supply the AC current to the step-down converter while maintaining the DC input voltage. Ceramic capacitors are recommended for best performance and should be placed as close to the V_{IN} pin as possible. Capacitors with X5R and X7R ceramic dielectrics are recommended because they are fairly stable with temperature fluctuations.

The capacitors must have a ripple-current rating greater than the maximum input ripple current of the converter. The input ripple current can be estimated with Equation (3) and Equation (4):

$$I_{CIN} = I_{OUT} \times \sqrt{\frac{V_{OUT}}{V_{IN}} \times (1 - \frac{V_{OUT}}{V_{IN}})} \quad (3)$$

The worst-case condition occurs at $V_{IN} = 2V_{OUT}$, where:

$$I_{CIN} = \frac{I_{OUT}}{2} \quad (4)$$

For simplification, choose an input capacitor with an RMS current rating greater than half of the maximum load current.

The input capacitance value determines the input voltage ripple of the converter. If there is an input voltage ripple requirement in the system, choose an input capacitor that meets the specification.

The input voltage ripple can be estimated with Equation (5) and Equation (6):

$$\Delta V_{IN} = \frac{I_{OUT}}{F_{SW} \times C_{IN}} \times \frac{V_{OUT}}{V_{IN}} \times (1 - \frac{V_{OUT}}{V_{IN}}) \quad (5)$$

The worst-case condition occurs at $V_{IN} = 2V_{OUT}$, where:

$$\Delta V_{IN} = \frac{1}{4} \times \frac{I_{OUT}}{F_{SW} \times C_{IN}} \quad (6)$$

Output Capacitor

The output capacitor is required to maintain the DC output voltage. Ceramic or POSCAP capacitors are recommended. The output voltage ripple can be estimated with Equation (7):

$$\Delta V_{OUT} = \frac{V_{OUT}}{F_{SW} \times L} \times (1 - \frac{V_{OUT}}{V_{IN}}) \times (R_{ESR} + \frac{1}{8 \times F_{SW} \times C_{OUT}}) \quad (7)$$

When using ceramic capacitors, the impedance at the switching frequency is dominated by the capacitance. The output voltage ripple is caused mainly by the capacitance. For simplification, the output voltage ripple can be estimated using Equation (8):

$$\Delta V_{OUT} = \frac{V_{OUT}}{8 \times F_{SW}^2 \times L \times C_{OUT}} \times (1 - \frac{V_{OUT}}{V_{IN}}) \quad (8)$$

When using POSCAP capacitors, the ESR dominates the impedance at the switching frequency. The output ripple can be approximated using Equation (9):

$$\Delta V_{OUT} = \frac{V_{OUT}}{F_{SW} \times L} \times (1 - \frac{V_{OUT}}{V_{IN}}) \times R_{ESR} \quad (9)$$

The maximum output capacitor limitation should be considered in design application. For a small soft-start time period (if the output capacitor value is too high), the output voltage cannot reach the design value during the soft-start time, causing it to fail to regulate. The maximum output capacitor value (C_{o_max}) can be limited approximately with Equation (10):

$$C_{O\_MAX} = (I_{LIM\_AVG} - I_{OUT}) \times T_{ss} / V_{OUT} \quad (10)$$

Where I_{LIM_AVG} is the average start-up current during the soft-start period, and T_{ss} is the soft-start time.

Inductor

The inductor is necessary to supply constant current to the output load while being driven by the switched input voltage. A larger value inductor results in less ripple current, resulting in a lower output ripple voltage. However, a larger value inductor has a larger physical footprint, a higher series resistance, and/or a lower saturation current. A good rule for determining the inductance value is to design the peak-to-peak ripple current in the inductor to be in the range of 30 percent to 50 percent of the maximum output current, with the peak inductor current below the maximum switch current limit. The inductance value can be calculated with Equation (11):

$$L = \frac{V_{OUT}}{F_{SW} \times \Delta I_L} \times (1 - \frac{V_{OUT}}{V_{IN}}) \quad (11)$$

Where ΔI_L is the peak-to-peak inductor ripple current.

The inductor should not saturate under the maximum inductor peak current (including short current), so it is suggested to choose Isat > 10 A.

PCB Layout Guidelines

Efficient PCB layout is critical for optimum IC performance. For best results, refer to Figure 7 and follow the guidelines below:

1. Place the high-current paths (GND, IN, and SW) very close to the device with short, direct, and wide traces. **The PGND trace should be as wide as possible (This should be the number one priority).**
2. Place the input capacitors as close to IN and GND as possible on the same layer as the IC.
3. Place the decoupling capacitor as close to VCC and GND as possible. Keep the switching node (SW) short and away from the feedback network.
4. Keep the BST voltage path as short as possible with a >50 mil trace.
5. Keep the IN and GND pads connected with a large copper plane to achieve better thermal performance. Add several vias with 8 mil drill/16 mil copper width close to the IN and GND pads to help thermal dissipation.
6. A 4-layer layout is strongly recommended to achieve better thermal performance.

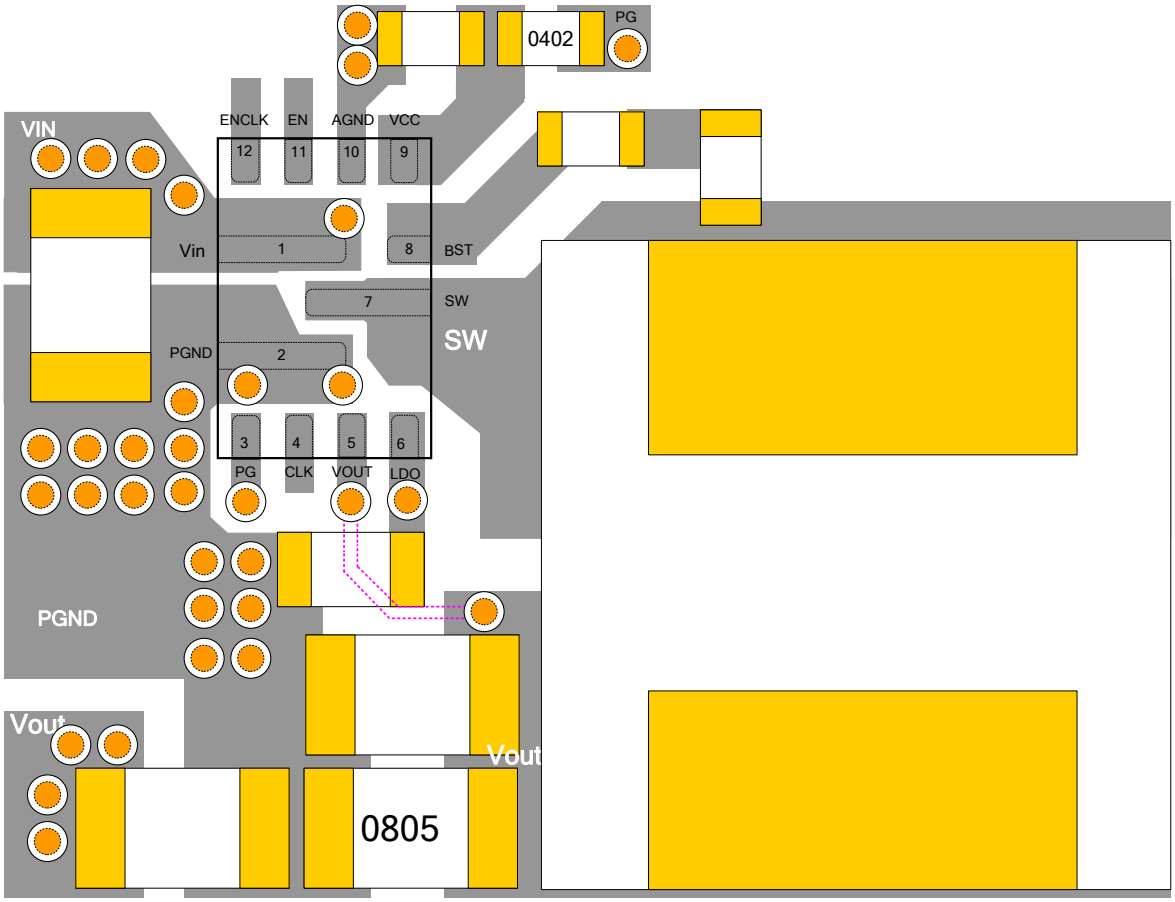
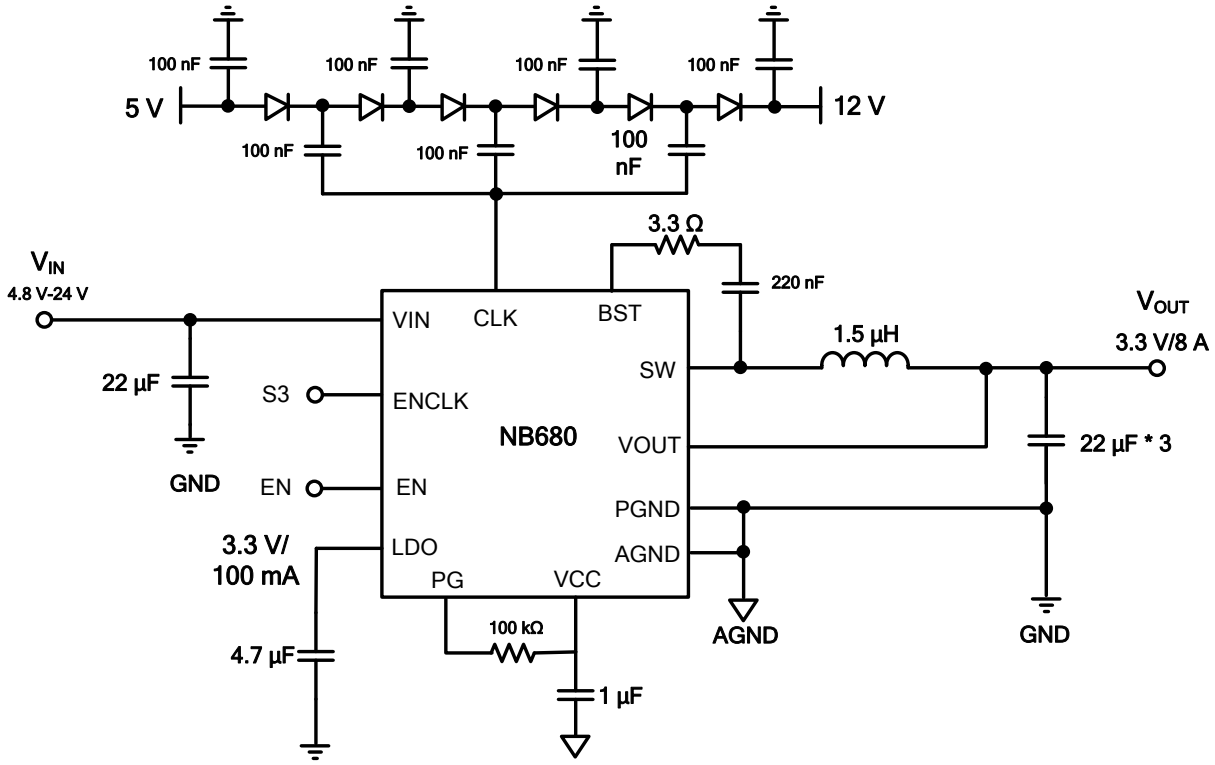


Figure 7— Recommend PBC layout

TYPICAL APPLICATION

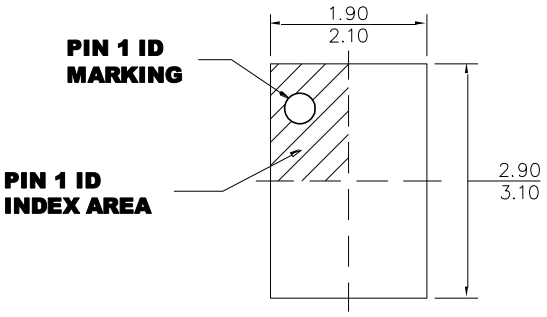


NOTE: If the charge pump function is not used, leave CLK open.

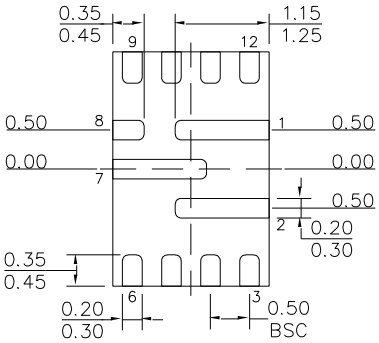
Figure 8—Typical application schematic with ceramic output capacitors

PACKAGE INFORMATION

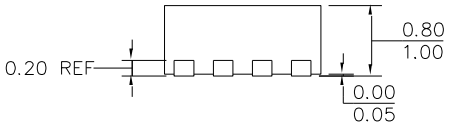
QFN-12 (2mm x 3mm)



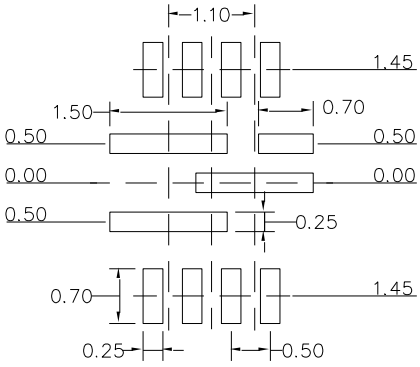
TOP VIEW



BOTTOM VIEW



SIDE VIEW



RECOMMENDED LAND PATTERN

NOTE:

1) ALL DIMENSIONS ARE IN MILLIMETERS.
2) EXPOSED PADDLE SIZE DOES NOT INCLUDE MOLD FLASH.
3) LEAD COPLANARITY SHALL BE 0.10 MILLIMETERS MAX.
4) JEDEC REFERENCE IS MO-220.
5) DRAWING IS NOT TO SCALE.