MP2386

24V, 8A, Low I_Q, Synchronous Buck Converter

DESCRIPTION

The MP2386 is a fully integrated, high-frequency, synchronous, rectified, step-down, switch-mode converter. The MP2386 offers a super-compact solution that achieves 8A of continuous output current over a wide input supply range.

The MP2386 operates at high efficiency over a wide output current load range based on MPS's proprietary switching loss reduction technique and internal low $R_{DS(ON)}$ power MOSFETs.

Adaptive constant-on-time (COT) control provides fast transient response and eases loop stabilization. The DC auto-tune loop combined with the remote differential sense provides good load and line regulation.

Full protection features include over-current protection (OCP), over-voltage protection (OVP), under-voltage protection (UVP), and thermal shutdown.

The converter requires a minimal number of external components and is available in a QFN-11 (2mmx2mm) package.

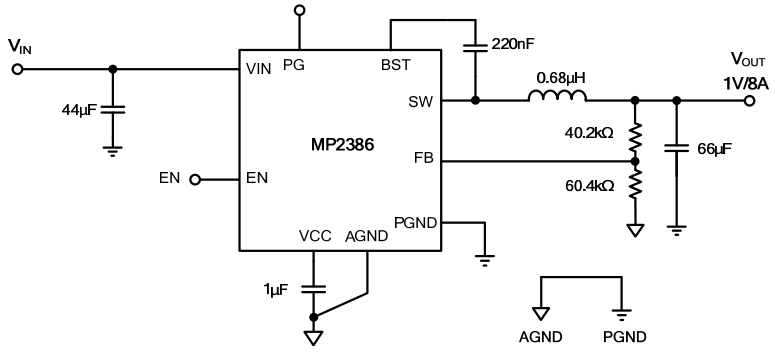
FEATURES

- Wide 4.5V to 24V Operating Input Range
- 105μA Low Quiescent Current
- 8A Continuous Output Current
- Adaptive COT for Fast Transient
- DC Auto-Tune Loop
- Low $R_{DS(ON)}$ Internal Power MOSFETs
- Proprietary Switching Loss Reduction Technique
- Power Good (PG) Indication
- Fixed 700kHz Switching Frequency
- Stable with POSCAP and Ceramic Caps
- Internal Soft Start (SS)
- Output Discharge
- OCP, OVP, UVP, and Thermal Shutdown with Auto-Retry
- Available in a QFN-11 (2mmx2mm) Package

APPLICATIONS

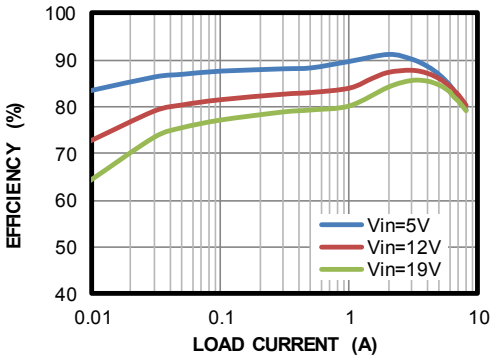
- Security Cameras
- Portable Devices, XDSL Devices
- Digital Set-Top Boxes
- Flat-Panel Televisions and Monitors
- General Purposes

TYPICAL APPLICATION



Efficiency

V_{OUT} = 1V, L = 0.68μH, DCR = 3.1mΩ



ORDERING INFORMATION

Part Number*	Package	Top Marking
MP2386GG	QFN-11 (2mmx2mm)	*See Below*

* For Tape & Reel, add suffix –Z (e.g.: MP2386GG–Z).

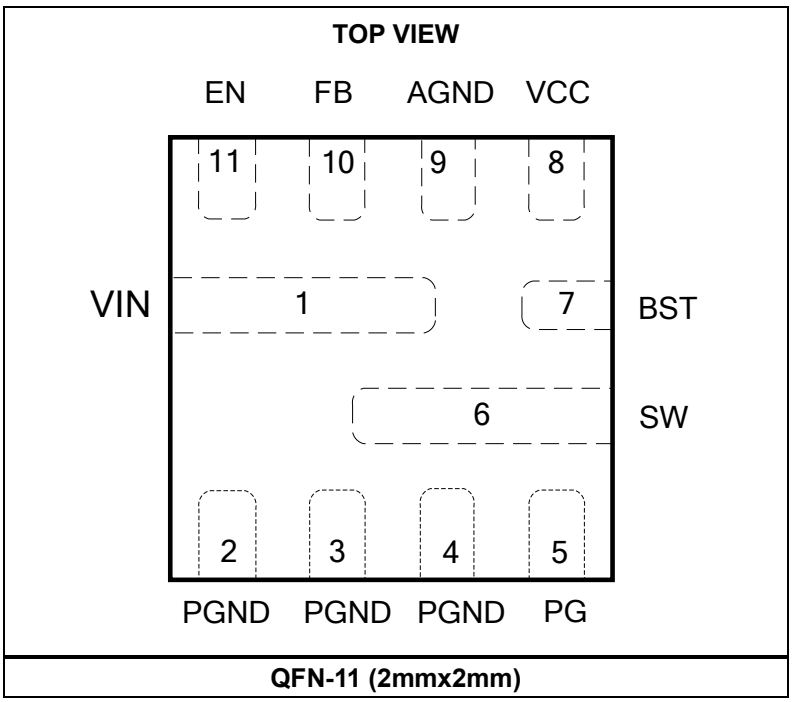
TOP MARKING

HTY

LLL

HT: Product code of MP2386GG
Y: Year code
LLL: Lot number

PACKAGE REFERENCE



QFN-11 (2mmx2mm)

PIN FUNCTIONS

PIN #	Name	Description
1	VIN	**Supply voltage.** VIN supplies power to the internal MOSFET and regulator. The MP2386 operates from a +4.5V to +24V input rail. An input capacitor is needed to decouple the input rail. Connect VIN with wide PCB traces and multiple vias. Apply at least two layers to this input trace.
2 - 4	PGND	**Power ground.** Connect PGND with wide PCB traces and multiple vias.
5	PG	**Power good output.** The output of PG is an open drain.
6	SW	**Switch output.** Connect SW to the inductor and bootstrap capacitor. SW is driven up to VIN 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. Connect SW with wide and short PCB traces.
7	BST	**Bootstrap.** Connect a capacitor between SW and BST to form a floating supply across the high-side switch driver.
8	VCC	**Internal VCC LDO output.** VCC powers the driver and control circuits. Decouple VCC with a minimum 1µF ceramic capacitor as close to VCC as possible. X7R or X5R grade dielectric ceramic capacitors are recommended for their stable temperature characteristics.
9	AGND	**Signal logic ground.** AGND is the Kelvin connection to PGND.
10	FB	**Feedback.** An external resistor divider from the output to GND tapped to FB sets the output voltage. Place the resistor divider as close to FB as possible. Avoid vias on the FB traces and V_{SEN} trace. Keep the V_{SEN} trace far away from the SW node.
11	EN	**Buck enable.** EN is a digital input that turns the buck regulator on or off. When the power supply of the control circuit is ready, drive EN high to turn on the buck regulator. Drive EN low to turn off the buck regulator. Connect EN to VIN through a resistive voltage divider for automatic start-up. Do not make the EN voltage higher than 4.5V at any time. Do not float EN.

ABSOLUTE MAXIMUM RATINGS (1)

Supply voltage (V_{IN}) 26V
V_{SW} (DC) ..... -1V to V_{IN} + 0.3V
V_{SW} (transient) ..... -9V for <2ns,
-5V to V_{IN} + 4V for <25ns (2)
V_{BST} ..... V_{SW} + 4.5V
All other pins ..... -0.3V to + 4.5V
Continuous power dissipation (T_A = +25°C) (3)(5)
QFN-11 (2mmx2mm) 3.6W
Junction temperature ..... 150°C
Lead temperature ..... 260°C
Storage temperature ..... -65°C to +150°C

Recommended Operating Conditions (4)

Supply voltage (V_{IN}) ..... 4.5V to 24V
Output voltage (V_{OUT}) ..... 0.6V to 13V
Operating junction temp. (T_J) ..... -40°C to +125°C

Thermal Resistance	θ_{JA}	θ_{JC}
QFN-11 (2mmx2mm)		
EV2386-G-00A (5)	34	9 °C/W
JESD51-7 (6)	80	16 °C/W

NOTES:

1) Exceeding these ratings may damage the device.
2) Measured by using a differential oscilloscope probe.
3) 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.
4) The device is not guaranteed to function outside of its operating conditions.
5) Measured on EV2386-G-00A, 4-layer PCB.
6) The value of θ_{JA} given in this table is only valid for comparison with other packages and cannot be used for design purposes. These values were calculated in accordance with JESD51-7 and simulated on a specified JEDEC board. They do not represent the performance obtained in an actual application.

ELECTRICAL CHARACTERISTICS

V_{IN} = 12V, T_J = -40°C to +125°C [7], typical value tested at T_J = +25°C, unless otherwise noted.

Parameters	Symbol	Condition	Min	Typ	Max	Units
Supply Current						
Supply current (quiescent)	I_{IN}	V_{EN} = 3.3V, V_{FB} = 0.62V		105	145	μA
Supply current (Shutdown)	I_{IN}	V_{EN} = 0V			2	μA
MOSFET						
High-side switch on resistance	HS_{RDS-ON}			34		mΩ
Low-side switch on resistance	LS_{RDS-ON}			10		mΩ
Switch leakage	SW_{LKG}	V_{EN} = 0V, V_{SW} = 0V		0	5	μA
Current Limit						
Low-side valley current limit	I_{LIMIT_LS}		7	9	11	A
Zero crossing current	I_{ZCD}	V_{OUT} = 3.3V, Lo = 2.2μH	0	150	400	mA
Switching Frequency and Minimum Off Timer						
Switching frequency	F_S		600	700	800	kHz
Minimum on time [8]	T_{ON_Min}			50		ns
Minimum off time [8]	T_{OFF_Min}			200		ns
Over-Voltage and Under-Voltage Protection (OVP, UVP)						
OVP threshold	V_{OVP}	V_{FB}	125%	130%	135%	V_{REF}
UVP-1 threshold	V_{UVP}	V_{FB}	70%	75%	80%	V_{REF}
UVP-1 hold off timer [8]	T_{OC}	V_{OUT} = 60% V_{REF}		32		μs
UVP-2 threshold	V_{UVP}	V_{FB}	45%	50%	55%	V_{REF}
Reference and Soft Start (REF, SS)						
REF voltage	V_{REF}		590	600	610	mV
Soft-start time [8]	T_{SS}		1.1	1.7	2.3	ms
Enable and Under-Voltage Lockout (EN, UVLO)						
Enable rising threshold	V_{EN_Rising}		1.15	1.25	1.35	V
Enable hysteresis	V_{EN_HYS}			150		mV
Enable Input Current	I_{EN}	V_{EN} = 3.3V		3.3		μA
VCC UVLO threshold rising	VCC_{Vth_R}		3.1	3.3	3.5	V
VCC UVLO threshold hysteresis	VCC_{HYS}			420		mV
VIN UVLO threshold rising	VIN_{VTH_R}		4.2	4.35	4.48	V
VIN UVLO threshold hysteresis	VIN_{HYS}			550		V
VCC Regulator						
VCC voltage	V_{CC}		3.45	3.65	3.85	V
VCC load regulation	V_{CC_Reg}	I_{VCC} = 5mA		5		%
Thermal Protection						
Thermal shutdown [8]	T_{SD}			150		°C
Thermal shutdown hysteresis [8]	T_{SD_HYS}			25		°C

NOTES:

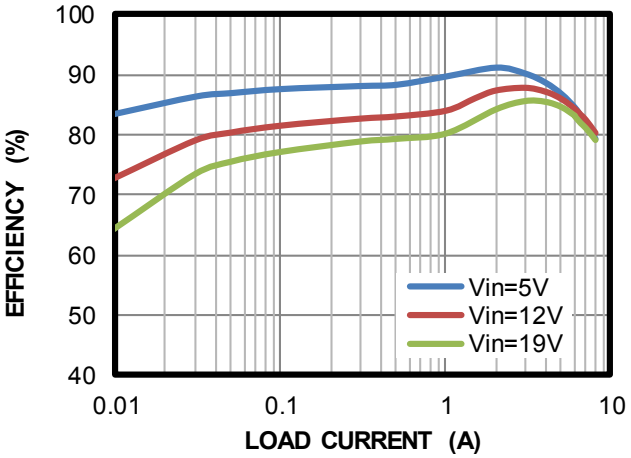
7) Not tested in production. Guaranteed by over-temperature correlation.
8) Guaranteed by engineering sample characterization.

TYPICAL CHARACTERISTICS

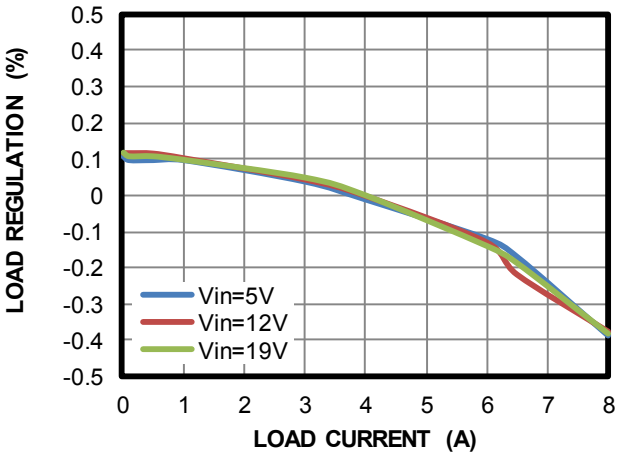
V_{IN} = 12V, V_{OUT} = 1V, L = 0.68µH, T_A = +25°C, unless otherwise noted.

Efficiency

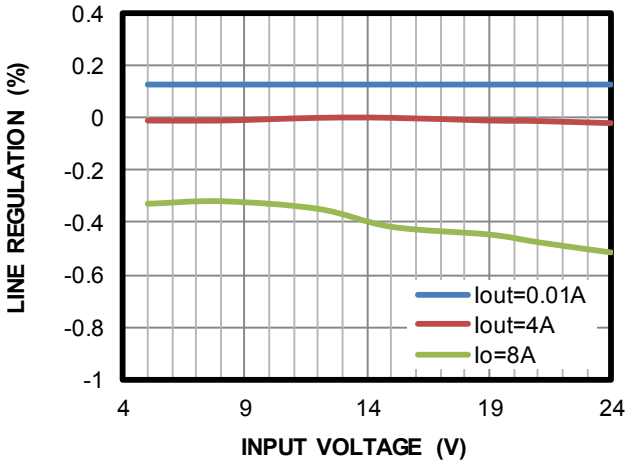
V_{OUT} = 1V, L = 0.68µH, DCR = 3.1mΩ



Load Regulation

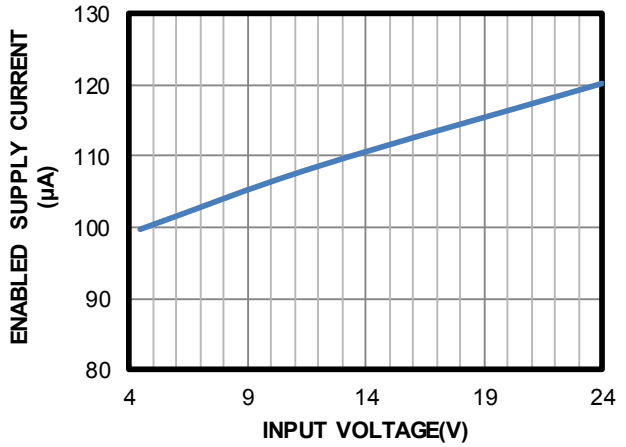


Load Regulation



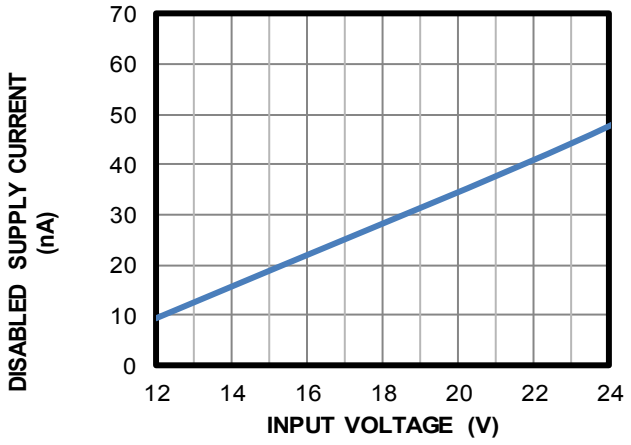
Enabled Supply Current vs. Input Voltage

V_{EN} = 3.3V, V_{FB} = 0.62V



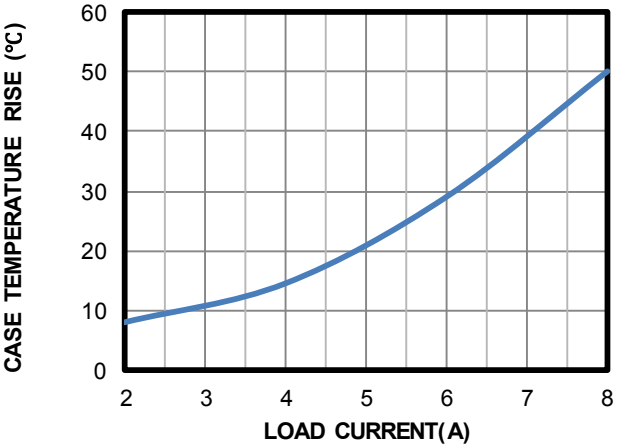
Disabled Supply Current vs. Input Voltage

V_{EN} = 0V



Case Temperature Rise vs. Load Current

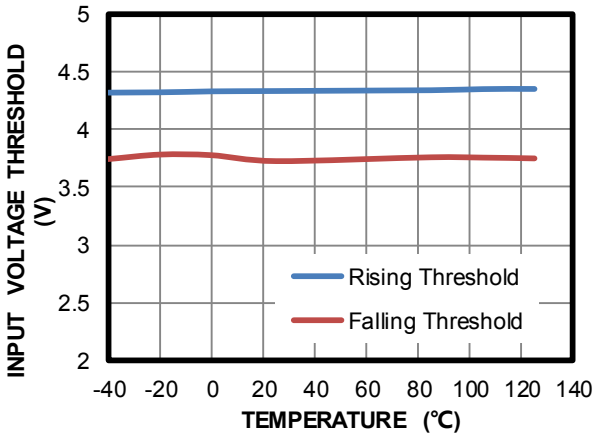
V_{IN} = 12V, V_{OUT} = 1V



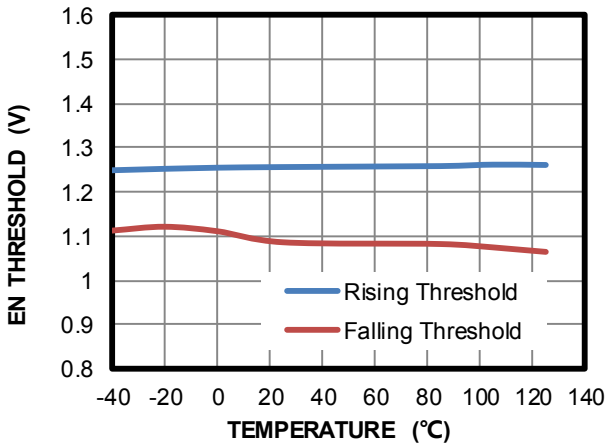
TYPICAL CHARACTERISTICS *(continued)*

V_{IN} = 12V, V_{OUT} = 1V, L = 0.68µH, T_A = +25°C, unless otherwise noted.

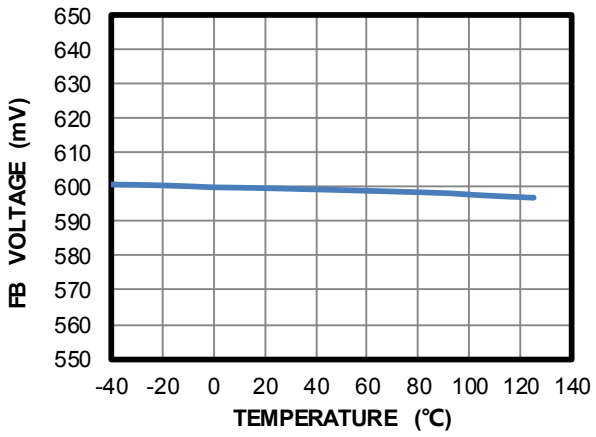
Input Voltage Threshold vs. Temperature



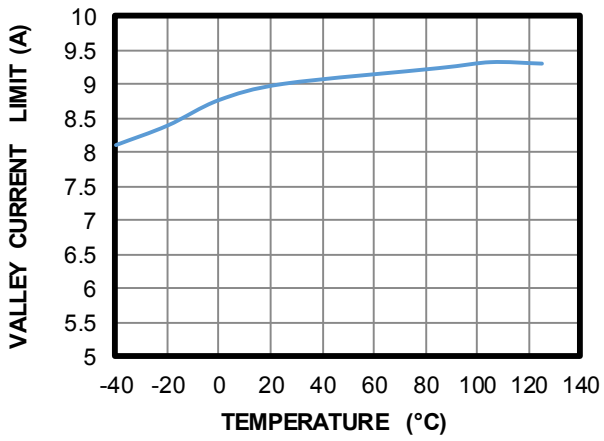
EN Threshold vs. Temperature



FB Voltage vs. Temperature



Valley Current Limit vs. Temperature

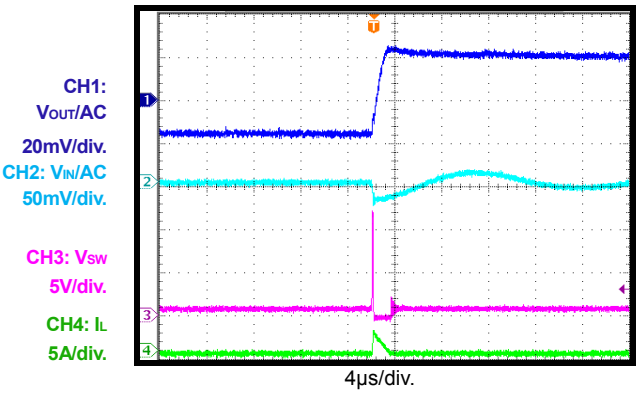


TYPICAL PERFORMANCE CHARACTERISTICS

V_{IN} = 12V, V_{OUT} = 1V, L = 0.68µH, T_A = +25°C, unless otherwise noted.

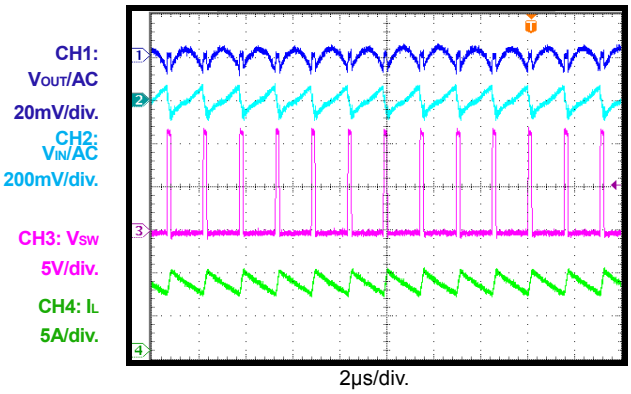
Input/Output Ripple

I_{OUT} = 0A



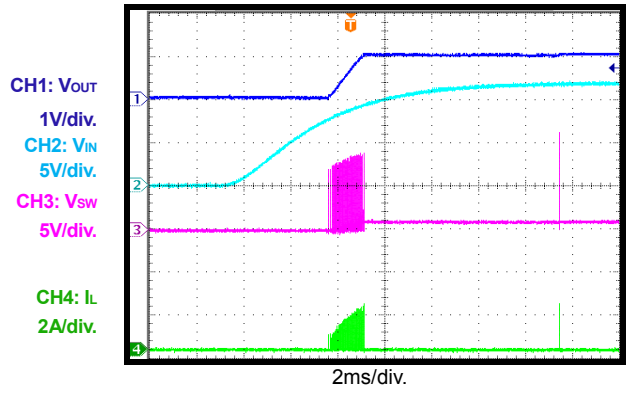
Input/Output Ripple

I_{OUT} = 8A



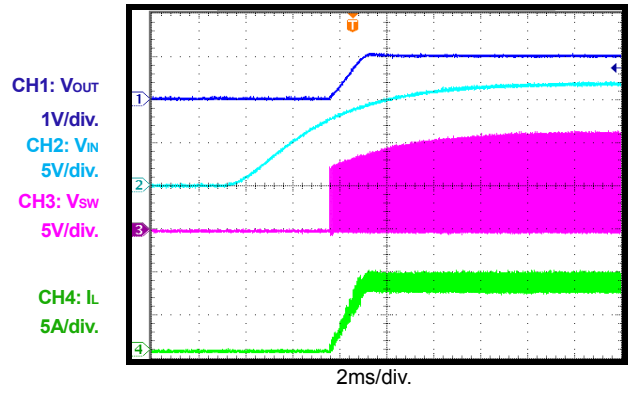
Start-Up through Input Voltage

I_{OUT} = 0A



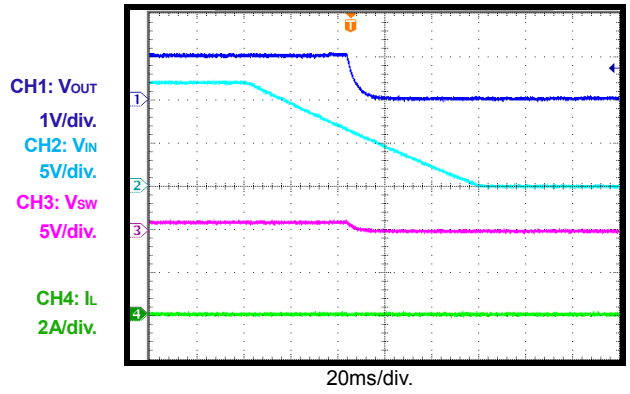
Start-Up through Input Voltage

I_{OUT} = 8A



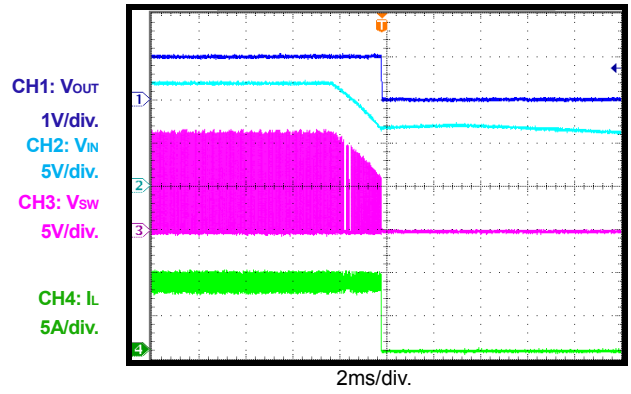
Shutdown through Input Voltage

I_{OUT} = 0A



Shutdown through Input Voltage

I_{OUT} = 8A

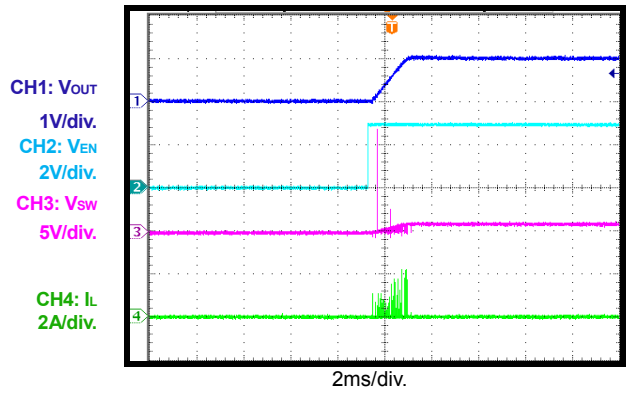


TYPICAL PERFORMANCE CHARACTERISTICS *(continued)*

V_{IN} = 12V, V_{OUT} = 1V, L = 0.68µH, T_A = +25°C, unless otherwise noted.

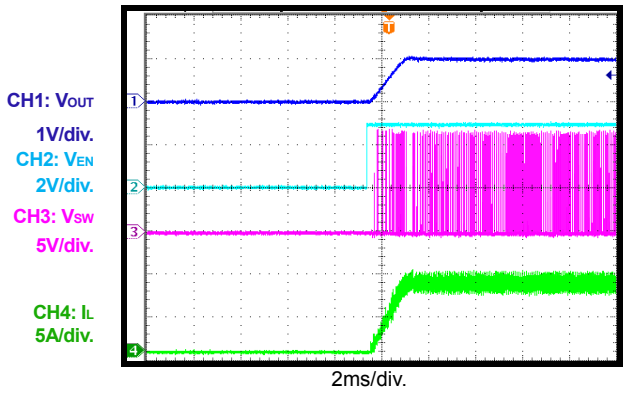
Start-Up through Enable

I_{OUT} = 0A



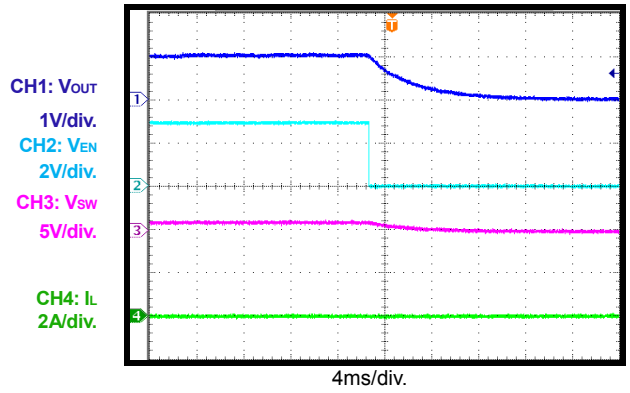
Start-Up through Enable

I_{OUT} = 8A



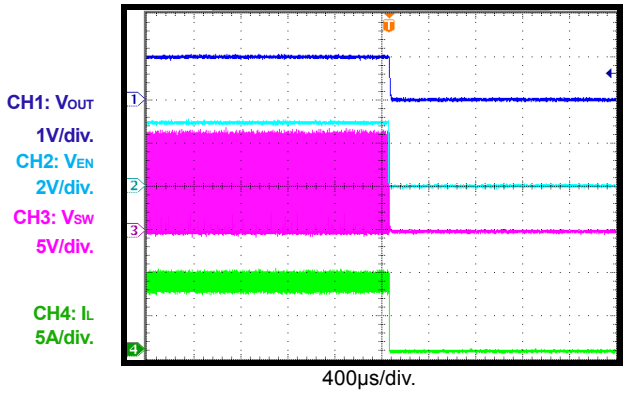
Shutdown through Enable

I_{OUT} = 0A

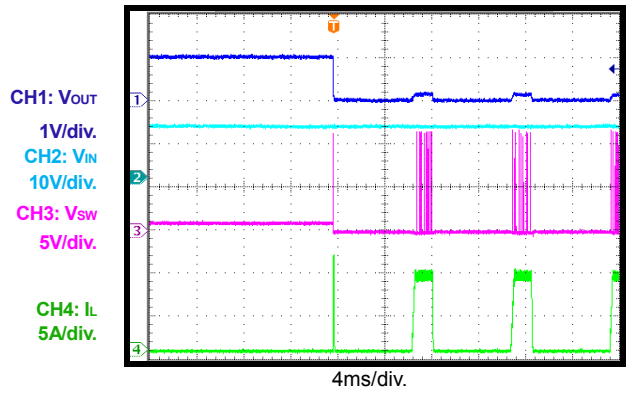


Shutdown through Enable

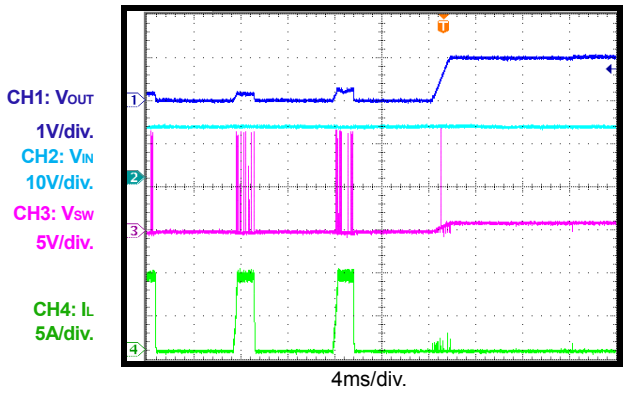
I_{OUT} = 8A



Short-Circuit Entry



Short-Circuit Recovery

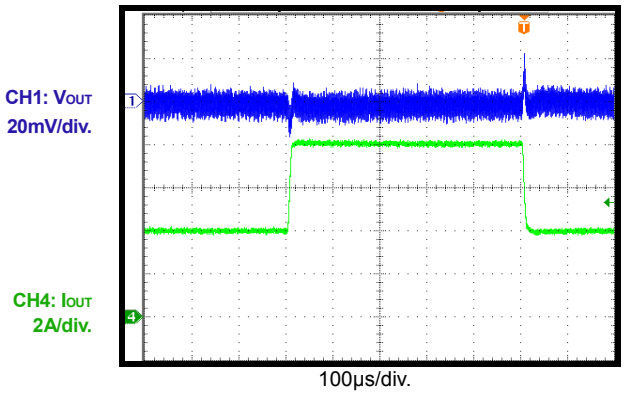


TYPICAL PERFORMANCE CHARACTERISTICS *(continued)*

V_{IN} = 12V, V_{OUT} = 1V, L = 0.68µH, T_A = +25°C, unless otherwise noted.

Load Transient

I_{OUT} = 4 - 8A



BLOCK DIAGRAM

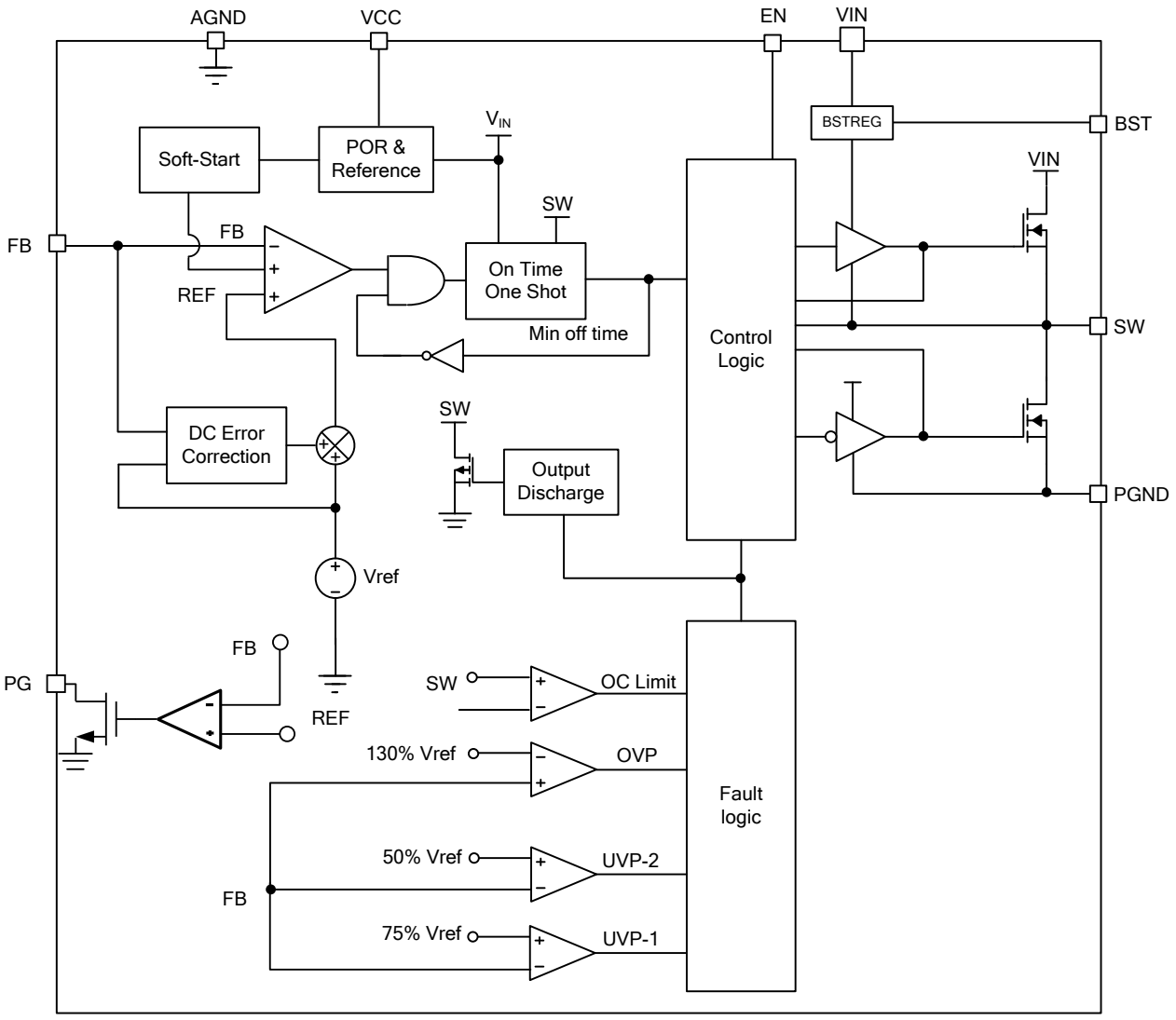


Figure 1: Functional Block Diagram

OPERATION

The MP2386 is a fully integrated, synchronous, rectified, step-down, switch-mode converter. Constant-on-time (COT) control is employed to provide fast transient response and easy 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 an insufficient output voltage. The on period is determined by the output voltage and 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. It is turned on again when V_{FB} drops below V_{REF}. By repeating operation in 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 conduction loss. There is a dead short between the input and GND if both the HS-FET and LS-FET are turned on at the same time. This is called shoot-through. To avoid shoot-through, a dead time (DT) is generated internally between the HS-FET off and LS-FET on period or the LS-FET off and HS-FET on period.

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

Heavy-Load Operation

Continuous conduction mode (CCM) occurs when the output current is high and the inductor current is always above zero amps. CCM operation is shown in Figure 2.

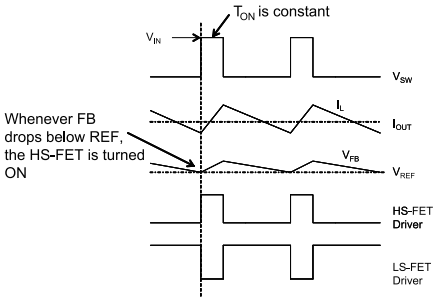


Figure 2: Heavy-Load Operation

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 begins.

In CCM operation, the switching frequency is fairly constant. This is called pulse-width modulation (PWM) mode.

Light-Load Operation

The inductor current decreases with the load. Once the inductor current reaches zero, the operation transitions from CCM to discontinuous conduction mode (DCM).

Light-load operation is shown in Figure 3. 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 inductor current reaches zero. In DCM operation, the V_{FB} does not reach V_{REF} when the inductor current approaches zero. The LS-FET driver enters tri-state (Hi-Z) whenever the inductor current reaches zero. As a result, the efficiency at light-load is improved greatly. At light-load condition, the HS-FET is not turned on as frequently as it is in heavy-load condition. This is called skip mode.

At light-load or no-load condition, the output drops very slowly, and the MP2386 reduces the switching frequency naturally. High efficiency is then achieved at light load.

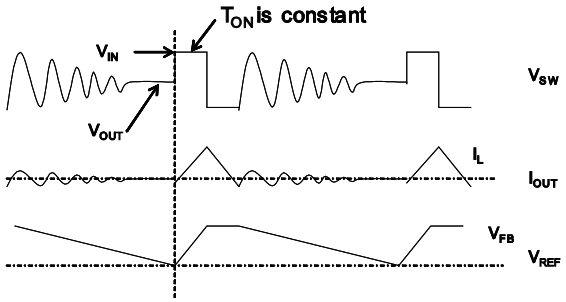


Figure 3: Light-Load Operation

As the output current increases from light-load condition, the current modulator regulation time period becomes shorter. The HS-FET is turned on more frequently, and the switching frequency increases. 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):

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

The device enters PWM mode once the output current exceeds the critical level. Afterward, the switching frequency remains fairly constant over the output current range.

Large Duty Cycle Operation

When V_{IN} is below 7V and V_{OUT} is above 4.2V, the MP2386 reduces the switching frequency to about 280kHz to support large-duty operation. If V_{OUT} is below 3.9V, the MP2386 returns to the normal switching frequency.

Jitter and FB Ramp

Jitter occurs in both PWM and skip mode when noise in the V_{FB} ripple propagates a delay to the HS-FET driver (see Figure 4 and Figure 5). Jitter can affect system stability with noise immunity proportional to the steepness of V_{FB}'s downward slope. Therefore, the jitter in DCM is usually larger than that in CCM. However, V_{FB} ripple does not affect noise immunity directly.

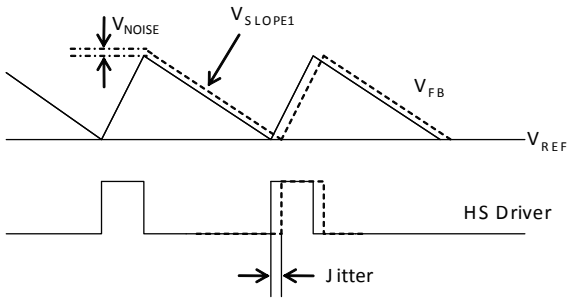


Figure 4: Jitter in PWM Mode

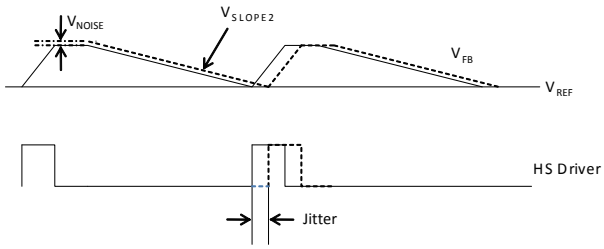


Figure 5: Jitter in Skip Mode

Operating with External Ramp Compensation

The MP2386 is able to support ceramic output capacitors without an external ramp, typically. However, in some cases, the internal ramp may not be enough to stabilize the system, or the jitter is too large. In these cases, external ramp compensation is needed. Refer to the Setting the Output Voltage with External Compensation section on page 15 for design steps using external ramp compensation.

Configuring the EN Control

The enable pin (EN) is used to enable or disable the entire chip. Pull EN high to turn on the regulator. Pull EN low to turn off the regulator. Do not float EN.

For automatic start-up, EN can be pulled up to the input voltage through a resistive voltage divider. There is an internal 1MΩ resistor from EN to GND. Choose the values for the pull-up resistor (R_{UP} from VIN to EN) and the pull-down resistor (R_{DOWN} from EN to GND) to determine the automatic start-up voltage with Equation (2):

$$V_{IN-START} = 1.25 \times \frac{R_{UP} + R_{DOWN}//1000K}{R_{DOWN}//1000K} (V) \quad (2)$$

For example, when R_{UP} = 150kΩ and R_{DOWN} = 51kΩ, $V_{IN-START}$ is 5.11V.

The EN voltage must not exceed the 4.5V maximum value to avoid damaging the internal circuit.

Power Good (PG)

The power good pin (PG) indicates whether the output voltage is in the normal range compared to the internal reference voltage. PG is an open-drain structure and requires an external pull-up supply. During power-up, the PG output is pulled low. This indicates to the system to remain off and keep the load on the output to a minimum. This helps reduce inrush current during start-up.

When the output voltage is higher than 95% and lower than 115% of the internal reference voltage and the soft start is finished, the PG signal is pulled high. When the output voltage is lower than 90% after the soft start finishes, the PG signal remains low. When the output voltage is higher than 115% of the internal reference, PG is switched low. The PG signal rises high again after the output voltage drops below 105% of the internal V_{REF}. The PG output is pulled low when either the EN under-voltage

lockout (UVLO), input UVLO, over-current protection (OCP), or over-temperature protection (OTP) is triggered.

Soft Start (SS)

The MP2386 employs a soft start (SS) mechanism to ensure a smooth output during power-up. When the part starts up, the internal V_{REF} ramps up gradually, and the output voltage ramps up smoothly as well. Once V_{REF} reaches the target value, the soft start finishes, and the device 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 low-side switches until the voltage on the internal reference exceeds the sensed output voltage at the FB node.

Over-Current Limit

The MP2386 has a cycle-by-cycle over-current limiting control. The current-limit circuit employs a valley current-sensing algorithm. The MP2386 uses the $R_{DS(ON)}$ 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, even if FB is lower than REF. Figure 6 shows the detailed operation of the valley-current limit.

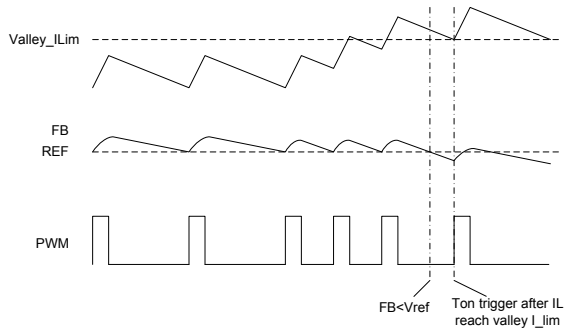


Figure 6: Valley Current-Limit Control

Since the comparison is done during the low-side on state, the OC trip level sets the valley level of the inductor current.

The maximum load current at the over-current threshold (I_{OC}) can be calculated with Equation (3):

$$I_{OC} = I\_limit + \frac{\Delta I_{inductor}}{2} \quad (3)$$

The OCL itself limits the inductor current and does not latch off. In an over-current condition, the current to the load exceeds the current to the output capacitor, making the output voltage fall off. Eventually, the current ends up crossing the under-voltage protection (UVP) threshold, and MP2386 enters hiccup protection mode.

Over-/Under-Voltage Protection (OVP/UVP)

The MP2386 monitors a resistor-divided feedback voltage to detect over- and under-voltage conditions. When V_{FB} rises higher than 130% of the target voltage, the over-voltage protection (OVP) comparator output goes high, and the circuit latches as the HS-FET driver turns off. A discharge MOSFET on SW turns on to discharge the output cap voltage.

When V_{FB} is between 50% and 75% of V_{REF}, the UVP-1 comparator output goes high, and the MP2386 enters hiccup mode if V_{FB} remains in this range for about 32µs. During this period, the valley current limit helps control the inductor current.

When V_{FB} drops below 50% of V_{REF}, the UVP-2 comparator output goes high, and the MP2386 enters hiccup mode directly after the comparator and logic delay.

UVLO Protection

The MP2386 has two types of UVLO protection: VCC UVLO and V_{IN} UVLO. The MP2386 starts up only when both VCC and V_{IN} exceed their respective UVLO thresholds. The MP2386 shuts down when either VCC is lower than the VCC falling threshold voltage or VIN is lower than the V_{IN} falling threshold. These are both non-latch off protections.

If an application requires a higher UVLO, use EN as shown in Figure 7 to adjust the input voltage UVLO using two external resistors.

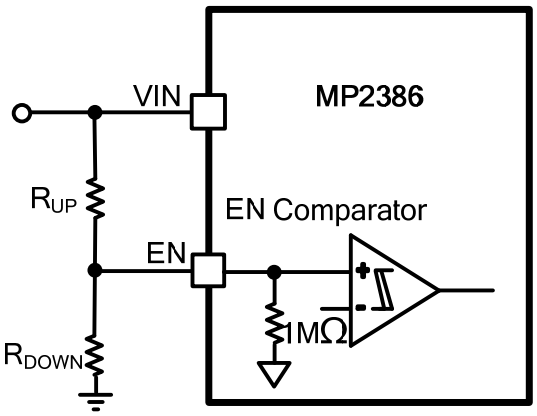


Figure 7: Adjustable UVLO

Thermal Shutdown

The MP2386 has a thermal shutdown function. The junction temperature of the IC is monitored internally. If the junction temperature exceeds the threshold value (typically 150°C), the converter shuts off. This is a non-latch protection. There is a hysteresis of about 25°C. Once the junction temperature drops to about 125°C, a soft start is initiated.

Output Discharge

The MP2386 discharges the output when the controller is turned off by a protection function (UVP, OCP, OVP, UVLO, thermal shutdown). The discharge resistor on the output is 40Ω, typically.

APPLICATION INFORMATION

Setting the Output Voltage without External Compensation

The MP2386 has an internal ramp. When the internal compensation is sufficient for stable operation with ceramic output capacitors, the MP2386 does not require external ramp compensation. The output voltage is then set by feedback resistors R1 and R2 (see Figure 8).

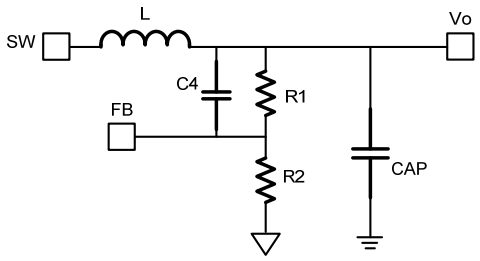


Figure 8: Simplified Circuit without an External Ramp

First, choose a value for R2. R2 should be chosen reasonably, since a small R2 leads to considerable quiescent current loss, but a large R2 makes FB noise-sensitive. Set R2 to be between 5 - 100kΩ. Considering the output ripple, R1 can be determined with Equation (4):

$$R_1 = \frac{V_{OUT} - V_{REF}}{V_{REF}} \cdot R_2 \quad (4)$$

C4 acts as a feed-forward cap to improve the transient. A larger C4 lead to better transient but more noise sensitivity.

Table 1 lists the recommended resistor values for common output voltages.

Table 1: Parameters Selection for Common Output Voltages [9]

V_{OUT} (V)	R1 (kΩ)	R2 (kΩ)	C4 (pF)	L (μH)
5	40.2	5.49	33	2.2
3.3	40.2	8.87	33	2.2
2.5	40.2	12.7	33	1.5
1.8	40.2	20	33	1.5
1.5	40.2	26.7	33	1
1.2	40.2	40.2	33	1
1	40.2	60.4	33	0.68

NOTE:

9) For additional component parameters, refer to the Typical Application Circuits on page 18 to page 20.

Setting the Output Voltage with External Compensation

If the system is not stable enough or the jitter is too large when ceramic capacitors are used in the output, an external voltage ramp should be added to FB through resistor R4 and capacitor C4. Since an internal ramp has already been added in the system, a 1MΩ (R4), 220pF (C4) ramp is sufficient for the ramp, typically.

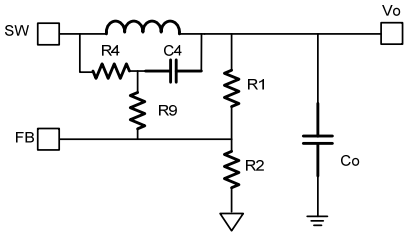


Figure 9: Simplified Circuit with External Ramp

The output voltage is influenced by R4 in addition to the R1 and R2 divider shown in Figure 9. R2 should be chosen reasonably, since a small R2 leads to considerable quiescent current loss while a large R2 makes FB noise-sensitive. Set R2 to be between 5 - 100kΩ. R1 can then be determined with Equation (5):

$$R_1 = \frac{1}{\frac{V_{REF}}{V_{OUT} - V_{REF}} - \frac{R2}{R4}} \cdot R_2 \quad (5)$$

Usually, R9 is set to 0Ω. To get a pole for better noise immunity, R9 can also be set using Equation (6):

$$R_9 = \frac{1}{2\pi \times C_4 \times 2F_{SW}} \quad (6)$$

R9 should be set in the range of 100Ω to 1kΩ to reduce its influence on the ramp.

Selecting the Input Capacitor

The input current to the step-down converter is discontinuous and therefore requires a capacitor to supply AC current to the step-down converter while maintaining the DC input voltage. Ceramic capacitors are recommended for the best performance and should be placed as close to VIN as possible. Capacitors with X5R and X7R ceramic dielectrics are

recommended since they are fairly stable with temperature fluctuations.

The capacitors must also have a ripple current rating greater than the maximum input ripple current of the converter. The input ripple current can be estimated with Equation (7):

$$I_{CIN} = I_{OUT} \times \sqrt{\frac{V_{OUT}}{V_{IN}} \times (1 - \frac{V_{OUT}}{V_{IN}})} \quad (7)$$

The worst-case condition occurs at $V_{IN} = 2V_{OUT}$, shown in Equation (8):

$$I_{CIN} = \frac{I_{OUT}}{2} \quad (8)$$

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 (9):

$$\Delta V_{IN} = \frac{I_{OUT}}{F_{SW} \times C_{IN}} \times \frac{V_{OUT}}{V_{IN}} \times (1 - \frac{V_{OUT}}{V_{IN}}) \quad (9)$$

The worst-case condition occurs at $V_{IN} = 2V_{OUT}$, shown in Equation (10):

$$\Delta V_{IN} = \frac{1}{4} \times \frac{I_{OUT}}{F_{SW} \times C_{IN}} \quad (10)$$

Selecting the Output Capacitor

An 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 (11):

$$\Delta V_{OUT} = \frac{V_{OUT}}{F_{SW} \times L} \times (1 - \frac{V_{OUT}}{V_{IN}}) \times (R_{ESR} + \frac{1}{8 \times F_{SW} \times C_{OUT}}) \quad (11)$$

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 with Equation (12):

$$\Delta V_{OUT} = \frac{V_{OUT}}{8 \times F_{SW}^2 \times L \times C_{OUT}} \times (1 - \frac{V_{OUT}}{V_{IN}}) \quad (12)$$

In the case of POSCAP capacitors, the ESR dominates the impedance at the switching frequency. The output ripple can be approximated with Equation (13):

$$\Delta V_{OUT} = \frac{V_{OUT}}{F_{SW} \times L} \times (1 - \frac{V_{OUT}}{V_{IN}}) \times R_{ESR} \quad (13)$$

The maximum output capacitor limitation should also be considered in the design application. The MP2386 has a soft-start time period of around 1.7ms. If the output capacitor value is too high, then the output voltage cannot reach the design value during the soft-start time and will fail to regulate. The maximum output capacitor value (C_{O_MAX}) can be limited approximately with Equation (14):

$$C_{O\_MAX} = (I_{LIM\_AVG} - I_{OUT}) \times T_{ss} / V_{OUT} \quad (14)$$

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

Selecting the Inductor

An inductor is necessary for supplying constant current to the output load while being driven by the switched input voltage. A larger-value inductor results in less ripple current and a lower output ripple voltage but also has a larger physical footprint, higher series resistance, and 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 - 50% of the maximum output current and ensure that the peak inductor current is below the maximum switch current limit. The inductance value can be calculated with Equation (15):

$$L = \frac{V_{OUT}}{F_{SW} \times \Delta I_L} \times (1 - \frac{V_{OUT}}{V_{IN}}) \quad (15)$$

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

The inductor should not saturate under the maximum inductor peak current, including a short current.

PCB Layout Guidelines

Efficient PCB layout is critical for stable operation. For best results, refer to Figure 10 and follow the guidelines below. A four-layer layout is recommended for better thermal performance.

1. Place the high-current paths (PGND, VIN, SW) very close to the device with short, direct, and wide traces.
2. Place the input capacitors as close to VIN and PGND as possible.
3. Place the decoupling capacitor as close to VCC and AGND as possible.
4. Keep the switching node SW short and away from the feedback network.
5. Keep the BST voltage path as short as possible.
6. Keep the VIN and PGND pads connected with large coppers to achieve better thermal performance.
7. Add several vias close to the VIN and PGND pads to help with thermal dissipation.

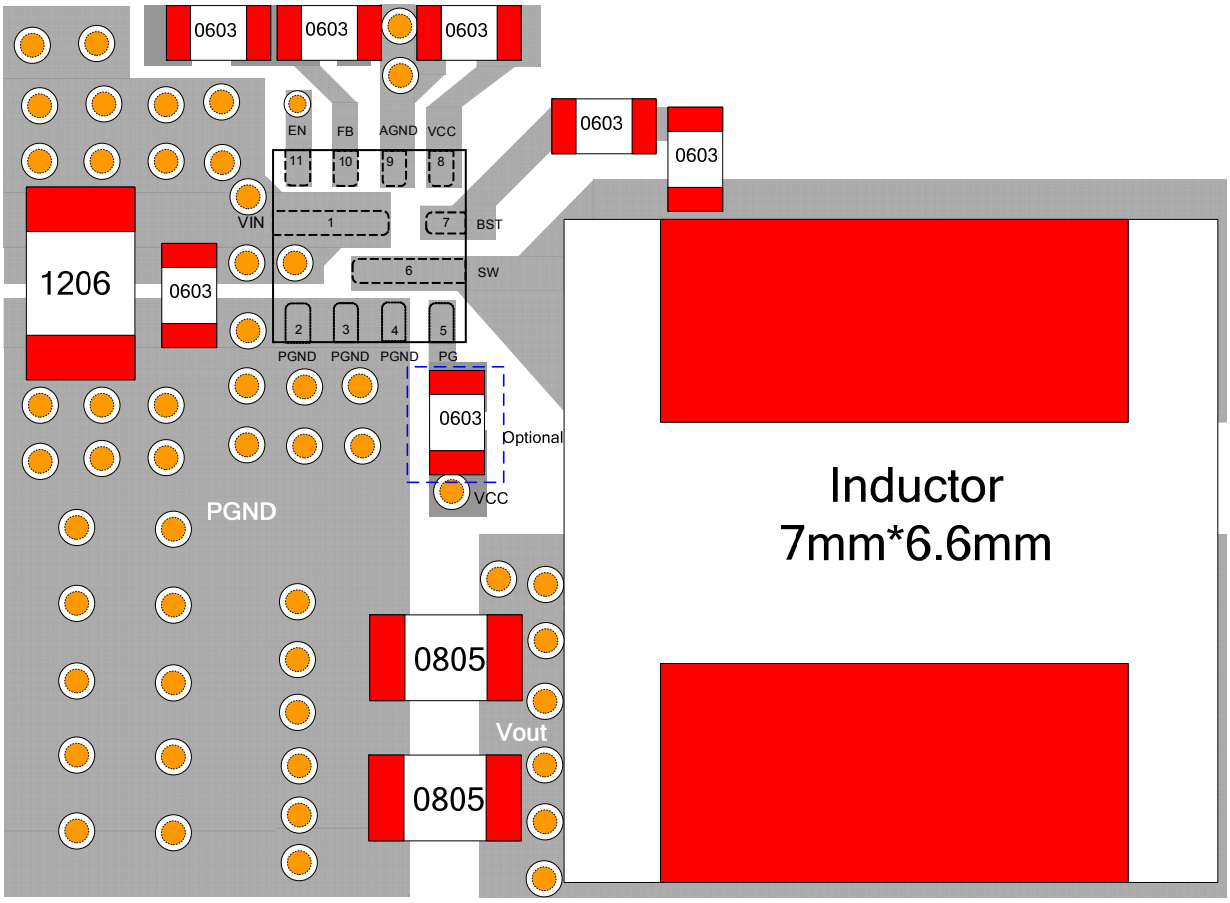


Figure 10: Recommended Layout

Design Example

Table 2 shows a design example when ceramic capacitors are applied.

Table 2: Design Example

V_{IN}	8V to 24V
V_{OUT}	1V
I_{OUT}	8A

The detailed application schematics are shown in Figure 11 through Figure 17. The typical performance and waveforms are shown in the Typical Characteristics section. For more devices applications, please refer to the related evaluation board datasheet.

TYPICAL APPLICATION CIRCUITS (10)

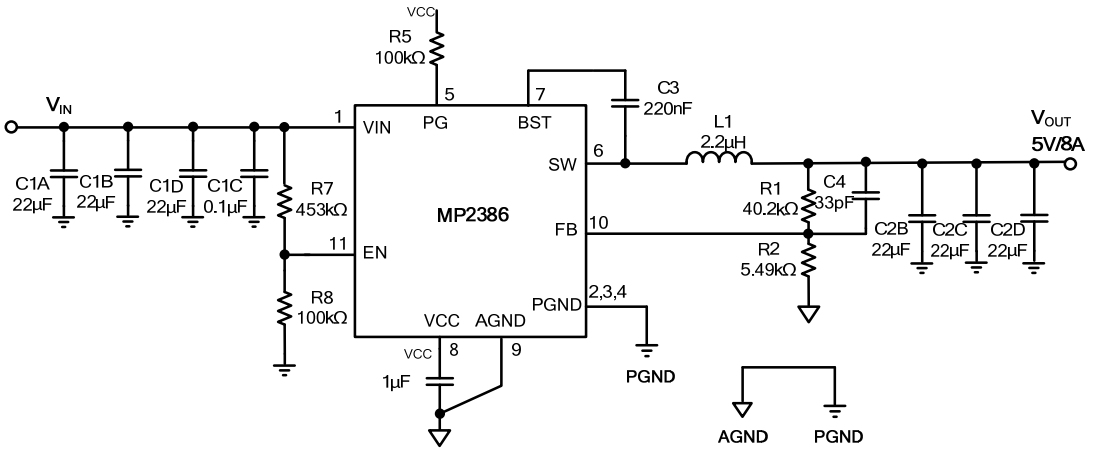


Figure 11: V_{IN} = 19V, V_{OUT} = 5V/8A

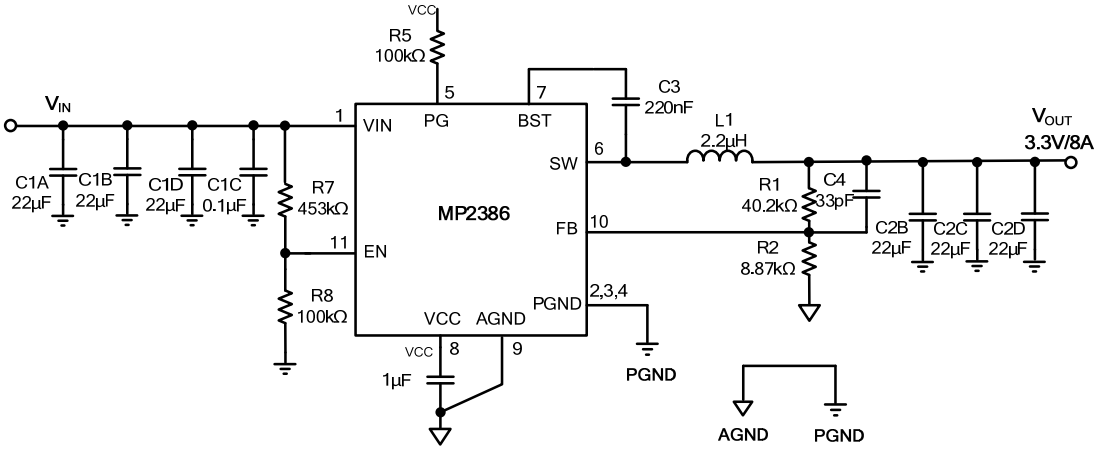


Figure 12: V_{IN} = 19V, V_{OUT} = 3.3V/8A

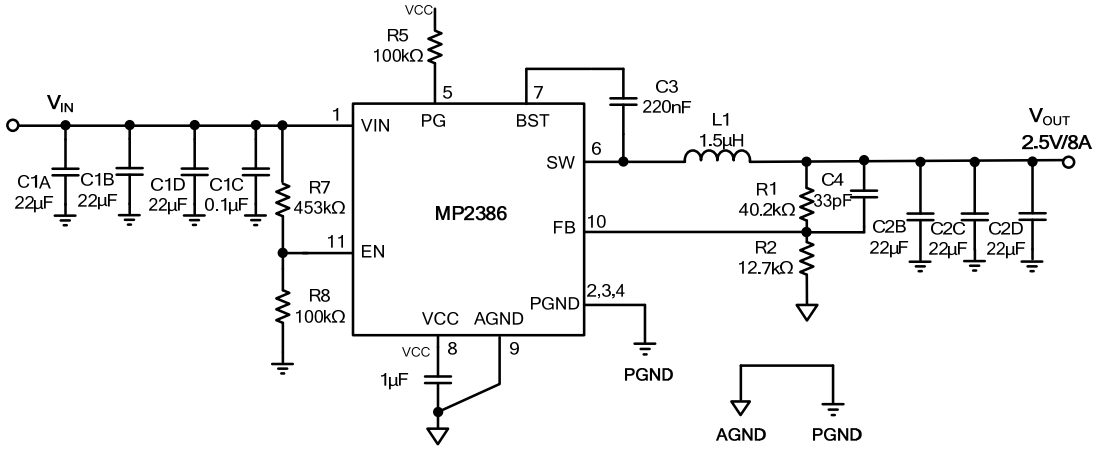


Figure 13: V_{IN} = 19V, V_{OUT} = 2.5V/8A

TYPICAL APPLICATION CIRCUITS (10) *(continued)*

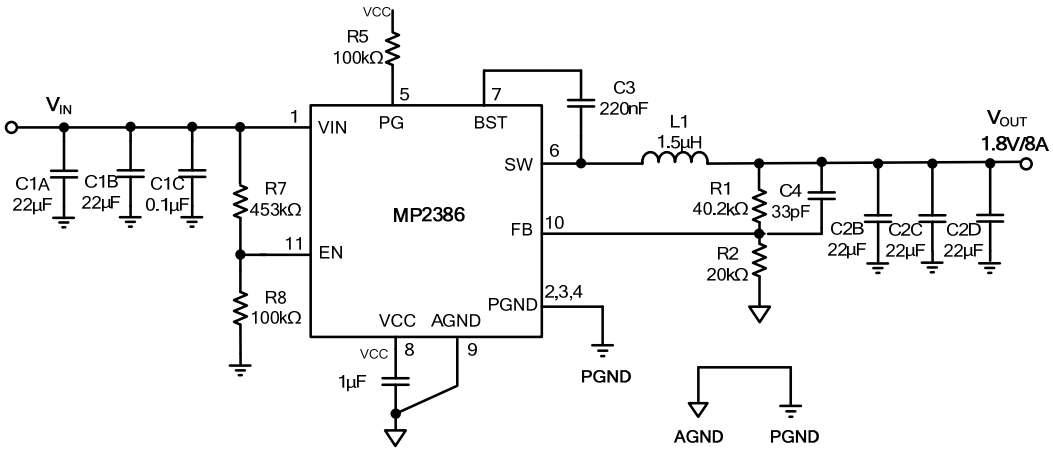


Figure 14: V_{IN} = 19V, V_{OUT} = 1.8V/8A

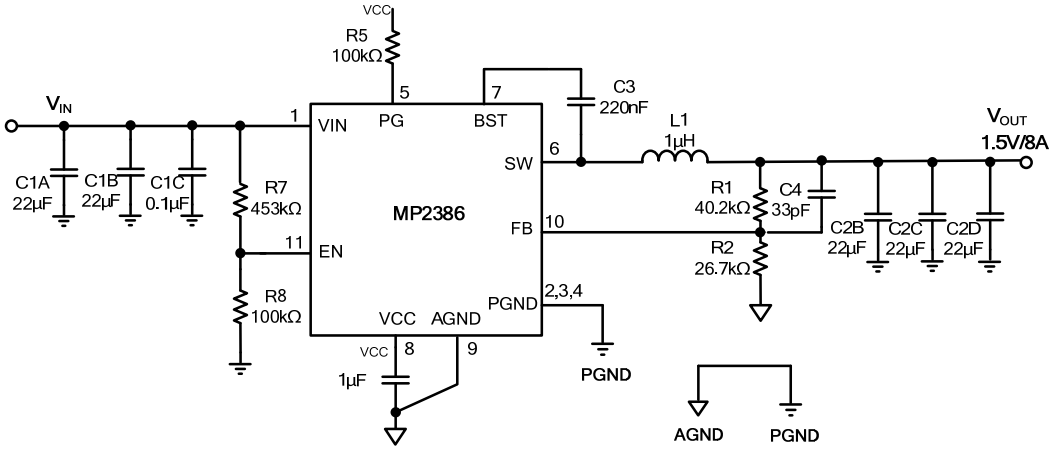


Figure 15: V_{IN} = 19V, V_{OUT} = 1.5V/8A

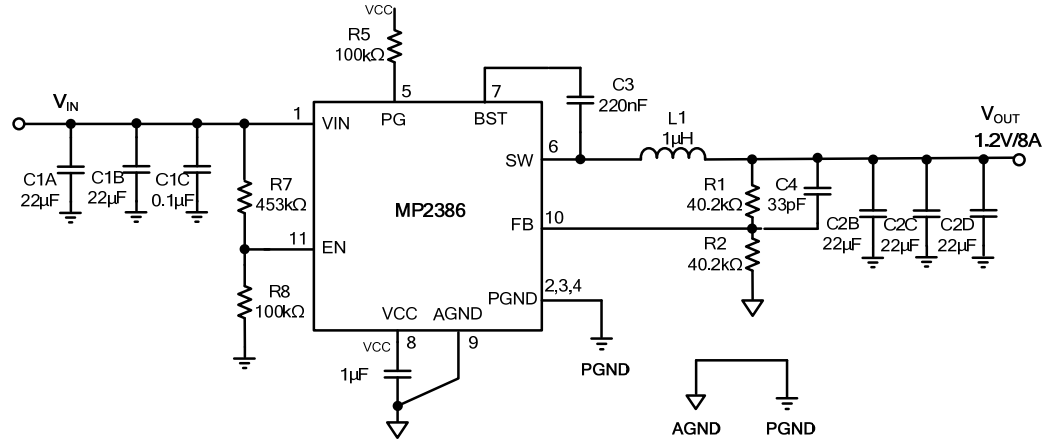


Figure 16: V_{IN} = 19V, V_{OUT} = 1.2V/8A

TYPICAL APPLICATION CIRCUITS (10) *(continued)*

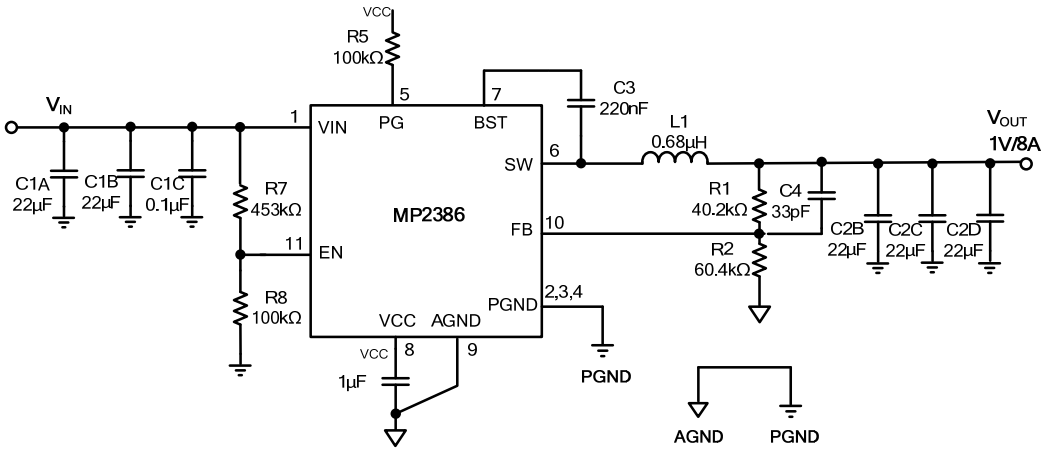


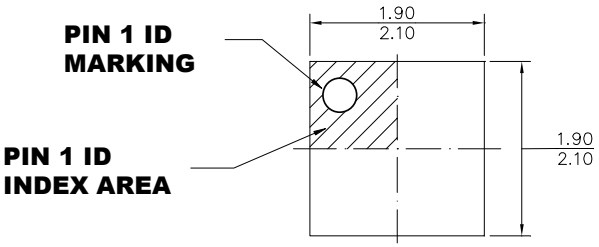
Figure 17: V_{IN} = 19V, V_{OUT} = 1V/8A

NOTE:

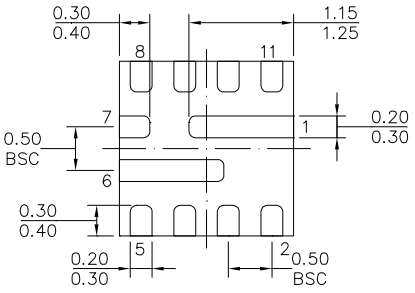
10) An EN resistor divider sets the VIN threshold to 7.5V. For 5V input applications, change the EN resistor accordingly.

PACKAGE INFORMATION

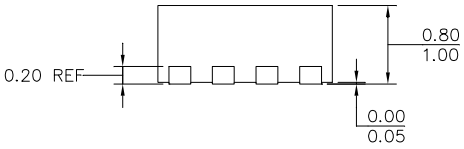
QFN-11 (2mmx2mm)



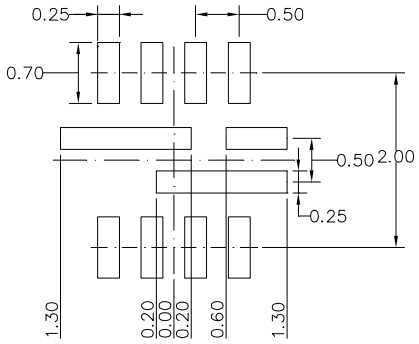
TOP VIEW



BOTTOM VIEW



SIDE VIEW



RECOMMENDED LAND PATTERN

NOTE:

1) LAND PATTERNS OF PIN1 AND PIN6 HAVE THE SAME LENGTH AND WIDTH
2) ALL DIMENSIONS ARE IN MILLIMETERS.
3) LEAD COPLANARITY SHALL BE 0.10 MILLIMETERS MAX.
4) JEDEC REFERENCE IS MO-220, VARIATION VCCD.
5) DRAWING IS NOT TO SCALE.

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