



Monolithic Power Systems

PRELIMINARY RELEASE – SPECIFICATIONS SUBJECT TO CHANGE

# MP1570

## 3A, 23V Synchronous Rectified Step-Down Converter

### DESCRIPTION

The MP1570 is a monolithic synchronous buck regulator. The device integrates 100mΩ MOSFETS which provide 3A continuous load current over a wide operating input voltage of 4.75V to 23V. Current mode control provides fast transient response and cycle-by-cycle current limit.

Adjustable soft-start prevents inrush current at turn-on. In shutdown mode, the output is actively discharged by transferring energy in the output capacitor to the input capacitor, dropping supply current to 1μA.

This device, available in an 8-pin SOIC package, provides a very compact system solution with minimal reliance on external components.

### FEATURES

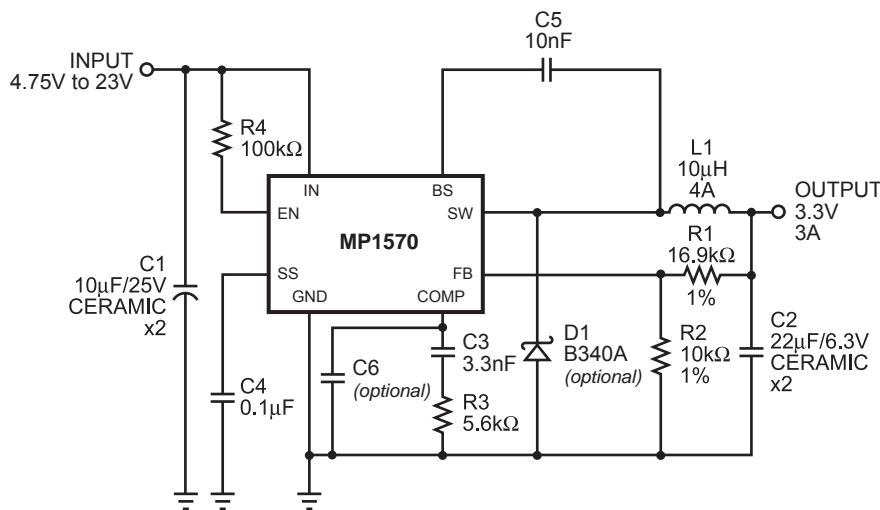
- 3A Output Current
- Wide 4.75 to 23V Operating Input Range
- Integrated 100mΩ Power MOSFET Switches
- Output Adjustable from 1.23V to 20V
- Up to 95% Efficiency
- Programmable Soft-Start
- Stable with Low ESR Ceramic Output Capacitors
- Fixed 340KHz Frequency
- Cycle-by-Cycle Over Current Protection
- Input Under Voltage Lockout
- Thermally Enhanced 8-Pin SOIC Package

### APPLICATIONS

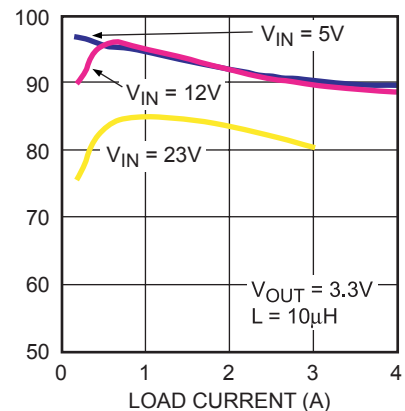
- Distributed Power Systems
- Pre-Regulator for Linear Regulators
- Notebook Computers

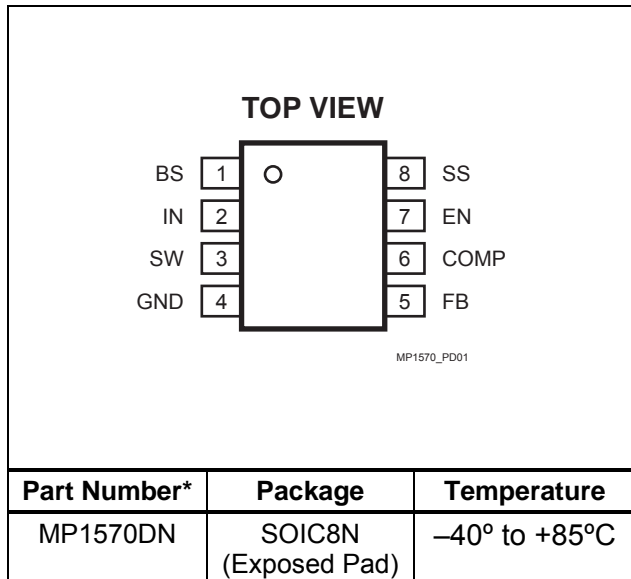
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### TYPICAL APPLICATION



Efficiency vs Load Current



**PACKAGE REFERENCE**


\* For Tape & Reel, add suffix -Z (eg. MP1570DN-Z)  
For Lead Free, add suffix -LF (eg. MP1570DN-LF-Z)

**ABSOLUTE MAXIMUM RATINGS <sup>(1)</sup>**

Supply Voltage  $V_{IN}$ ..... -0.3V to 26V  
 Switch Voltage  $V_{SW}$ ..... -1V to  $V_{IN} + 0.3V$   
 Boost Voltage  $V_{BS}$ .....  $V_{SW} - 0.3V$  to  $V_{SW} + 6V$   
 All Other Pins..... -0.3V to 6V  
 Junction Temperature..... 150°C  
 Lead Temperature ..... 260°C  
 Storage Temperature ..... -65°C to 150°C

**Recommended Operating Conditions <sup>(2)</sup>**

Input Voltage  $V_{IN}$ ..... 4.75V to 23V  
 Output Voltage  $V_{OUT}$  ..... 1.23V to 20V  
 Ambient Operating Temperature . -40°C to +85°C

**Thermal Resistance <sup>(3)</sup>**

$\Theta_{JA}$  ..... 50°C/W  
 $\Theta_{JC}$  ..... 10°C/W

**Notes:**

- 1) Exceeding these ratings may damage the device.
- 2) The device is not guaranteed to function outside of its operating conditions.
- 3) Measured on approximately 1" square of 1 oz copper.

**ELECTRICAL CHARACTERISTICS**

$V_{IN} = 12V$ ,  $T_A = 25^\circ C$ , unless otherwise noted.

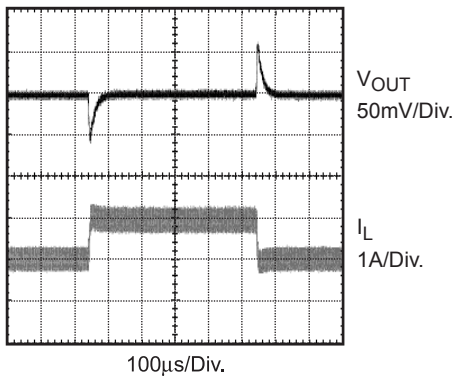
Parameter	Symbol	Condition	Min	Typ	Max	Units
Shutdown Supply Current		$V_{EN} = 0V$		1		$\mu A$
Supply Current		$V_{EN} = 2.7V$ ; $V_{FB} = 1.4V$		1.3	1.5	mA
Feedback Voltage	$V_{FB}$	$4.75V \leq V_{IN} \leq 23V$ ; $V_{COMP} < 2V$	1.202	1.230	1.258	V
Feedback Overvoltage Threshold				1.5		V
Error Amplifier Voltage Gain				400		V/V
Error Amplifier Transconductance		$\Delta I_C = \pm 10\mu A$		800		$\mu A/V$
High Side Switch On Resistance				100		m $\Omega$
Low Side Switch On Resistance				100		m $\Omega$
High Side Switch Leakage Current		$V_{EN} = 0V$ ; $V_{SW} = 0V$		0	10	$\mu A$
Upper Switch Current Limit				5.8		A
Lower Switch Current Limit		From Drain to Source		0.9		A
COMP to Current Sense Transconductance				5.4		A/V
Oscillation Frequency	$f_{SW}$			340		KHz
Short Circuit Oscillation Frequency		$V_{FB} = 0V$		100		KHz
Maximum Duty Cycle		$V_{FB} = 1.0V$		90		%
Minimum On Time				220		nS
EN Shutdown Threshold Voltage		$V_{EN}$ Rising	1.1	1.3	1.5	V

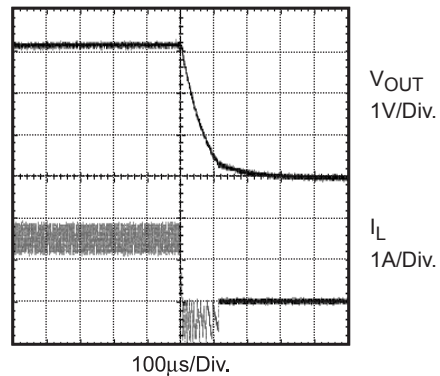
**ELECTRICAL CHARACTERISTICS (continued)**
 $V_{IN} = 12V$ ,  $T_A = 25^\circ C$ , unless otherwise noted.

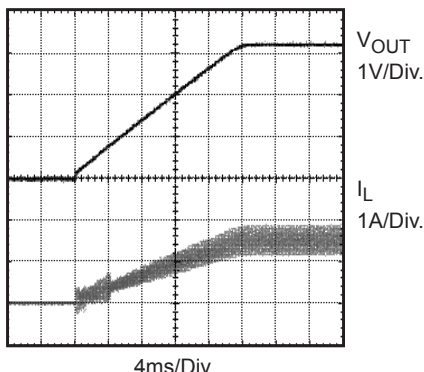
Parameter	Symbol	Condition	Min	Typ	Max	Units
EN Lockout Threshold Voltage				2.5		V
EN Lockout Hysteresis				210		mV
Input Under Voltage Lockout Threshold		$V_{IN}$ Rising	3.80	4.05	4.30	V
Input Under Voltage Lockout Threshold Hysteresis				210		mV
Soft Start Current		$V_{SS} = 0V$		6		$\mu A$
Soft Start Period		$C_{SS} = 0.1\mu F$		20		ms
Thermal Shutdown				160		$^\circ C$

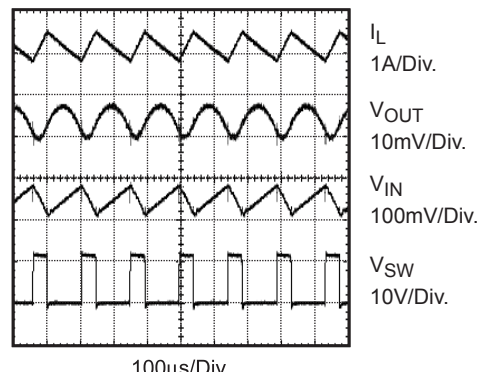
**TYPICAL PERFORMANCE CHARACTERISTICS**

Refer to Typical Application Schematic on Page 1

**Load Transient Waveforms**

 $V_{IN} = 12V$ ,  $V_{OUT} = 3.3V$ , 1A - 2A STEP  
 MP1570-WF01

**Turn Off Waveforms**

 $V_{IN} = 12V$ ,  $V_{OUT} = 3.3V$ ,  $R_{LOAD} = 2\Omega$   
 MP1570-WF02

**Soft-Start Waveforms**

 $V_{IN} = 12V$ ,  $V_{OUT} = 3.3V$ ,  $R_{LOAD} = 2\Omega$   
 MP1570-WF03

**Switching Waveforms**

 $V_{IN} = 12V$ ,  $V_{OUT} = 3.3V$ ,  $R_{LOAD} = 2\Omega$   
 MP1570-WF04



## OPERATION

### FUNCTIONAL DESCRIPTION

The MP1570 is a synchronous rectified, current-mode, step-down regulator. It regulates input voltages from 4.75V to 23V down to an output voltage as low as 1.230V, and supplies up to 3A of load current.

The MP1570 uses current-mode control to regulate the output voltage. The output voltage is measured at FB through a resistive voltage divider and amplified through the internal transconductance error amplifier. The voltage at COMP pin is compared to the switch current measured internally to control the output voltage.

The converter uses internal N-Channel MOSFET switches to step-down the input voltage to the regulated output voltage. Since the high side MOSFET requires a gate voltage greater than the input voltage, a boost capacitor connected between SW and BS is needed to drive the high side gate. The boost capacitor is charged from the internal 5V rail when SW is low.

When the MP1570 FB pin exceeds 20% of the nominal regulation voltage of 1.230V, the overvoltage comparator is tripped and latched, the COMP pin and the SS pin are discharged to GND, forcing the high-side switch off. Latch cannot be cleared unless the EN or IN pin is reset.

Following discharge, the MP1570 actively recycles the energy stored in the output capacitor. Initially the low-side synchronous rectifier turns on. Once the internal, negative 900mA current limit is reached, the low-side switch turns off, forcing inductor current to flow through the high-side switch body diode. The inductor current is recycled back into the input as an energy saving feature. This cycle continues until the output voltage is discharged below 10% of the initial regulation voltage (0.123V at FB), at which time the low-side switch turns off.

## APPLICATIONS INFORMATION

### COMPONENT SELECTION

#### Setting the Output Voltage

The output voltage is set using a resistive voltage divider from the output voltage to FB pin. The voltage divider divides the output voltage down to the feedback voltage by the ratio:

$$V_{FB} = V_{OUT} \frac{R2}{R1 + R2}$$

Thus the output voltage is:

$$V_{OUT} = 1.23 \times \frac{R1 + R2}{R2}$$

A typical value for R2 can be as high as 100kΩ, but a typical value is 10kΩ. Using that value, R1 is determined by:

$$R1 = 8.18 \times (V_{OUT} - 1.23)(k\Omega)$$

For example, for a 3.3V output voltage, R2 is 10kΩ, and R1 is 17kΩ.

#### Inductor

The inductor is required to supply constant current to the output load while being driven by the switched input voltage. A larger value inductor will result in less ripple current that will result in lower output ripple voltage. However, the larger value inductor will have a larger physical size, higher series resistance, and/or lower saturation current. A good rule for determining the inductance to use is to allow the peak-to-peak ripple current in the inductor to be approximately 30% of the maximum switch current limit. Also, make sure that the peak inductor current is below the maximum switch current limit. The inductance value can be calculated by:

$$L = \frac{V_{OUT}}{f_s \times \Delta I_L} \times \left( 1 - \frac{V_{OUT}}{V_{IN}} \right)$$

Where,  $V_{OUT}$  is the output voltage,  $V_{IN}$  is the input voltage,  $f_s$  is the 340KHz switching

frequency, and  $\Delta I_L$  is the peak-to-peak inductor ripple current.

Choose an inductor that will not saturate under the maximum inductor peak current. The peak inductor current can be calculated by:

$$I_{LP} = I_{LOAD} + \frac{V_{OUT}}{2 \times f_S \times L} \times \left(1 - \frac{V_{OUT}}{V_{IN}}\right)$$

Where,  $I_{LOAD}$  is the load current,  $V_{OUT}$  is the output voltage,  $V_{IN}$  is the input voltage, and  $f_S$  is the 340KHz switching frequency.

Table 1 lists a number of suitable inductors from various manufacturers. The choice of which style inductor to use mainly depends on the price vs. size requirements and any EMI requirement.

**Table 1—Inductor Selection Guide**

Vendor/ Model	Core Type	Core Material	Package Dimensions (mm)		
			W	L	H
<b>Sumida</b>					
CR75	Open	Ferrite	7.0	7.8	5.5
CDH74	Open	Ferrite	7.3	8.0	5.2
CDRH5D28	Shielded	Ferrite	5.5	5.7	5.5
CDRH5D28	Shielded	Ferrite	5.5	5.7	5.5
CDRH6D28	Shielded	Ferrite	6.7	6.7	3.0
CDRH104R	Shielded	Ferrite	10.1	10.0	3.0
<b>Toko</b>					
D53LC Type A	Shielded	Ferrite	5.0	5.0	3.0
D75C	Shielded	Ferrite	7.6	7.6	5.1
D104C	Shielded	Ferrite	10.0	10.0	4.3
D10FL	Open	Ferrite	9.7	1.5	4.0
<b>Coilcraft</b>					
DO3308	Open	Ferrite	9.4	13.0	3.0
DO3316	Open	Ferrite	9.4	13.0	5.1

Table 2 defines optional Schottky diodes that can be used to improve overall efficiency.

**Table 2—Diode Selection Guide**

Vendor	Part Number
Diode, Inc.	B340A
ON Semi	MBRS340

**Input Capacitor**

The input current to the step-down converter is discontinuous, therefore a capacitor is required to supply the AC current to the step-down converter while maintaining the DC input voltage. Use low ESR capacitors for the best performance. Ceramic capacitors are preferred, but tantalum or low-ESR electrolytic capacitors may also suffice. Choose X5R or X7R dielectrics when using ceramic capacitors.

Since the input capacitor absorbs the input switching current it requires an adequate ripple current rating. The RMS current in the input capacitor can be estimated by:

$$I_{CIN} = I_{LOAD} \times \sqrt{\frac{V_{OUT}}{V_{IN}} \times \left(1 - \frac{V_{OUT}}{V_{IN}}\right)}$$

Where,  $I_{LOAD}$  is the load current,  $V_{OUT}$  is the output voltage, and  $V_{IN}$  is the input voltage. The worst-case condition occurs at  $V_{IN} = 2V_{OUT}$ , where  $I_{CIN} = I_{LOAD}/2$ . For simplification, choose the input capacitor whose RMS current rating greater than half of the maximum load current.

The input capacitor can be electrolytic, tantalum or ceramic. When using electrolytic or tantalum capacitors, a small, high quality ceramic capacitor, i.e. 0.1µF, should be placed as close to the IC as possible. When using ceramic capacitors, make sure that they have enough capacitance to provide sufficient charge to prevent excessive voltage ripple at input. The input voltage ripple caused by capacitance can be estimated by:

$$\Delta V_{IN} = \frac{I_{LOAD}}{C_{IN}} \times \frac{V_{OUT}}{V_{IN}} \times \left(1 - \frac{V_{OUT}}{V_{IN}}\right)$$

Where,  $C_{IN}$  is the input capacitance value,  $I_{LOAD}$  is the output current,  $V_{OUT}$  is the output voltage, and  $V_{IN}$  is the input voltage.

**Output Capacitor**

The output capacitor is required to maintain the DC output voltage. Ceramic, tantalum, or low ESR electrolytic capacitors are recommended. Low ESR capacitors are preferred to keep the output voltage ripple low. The output voltage ripple can be estimated by:

$$\Delta V_{OUT} = \frac{V_{OUT}}{f_S \times L} \times \left(1 - \frac{V_{OUT}}{V_{IN}}\right) \times \left(R_{ESR} + \frac{1}{8 \times f_S \times C_O}\right)$$

Where,  $V_{OUT}$  is the output voltage,  $V_{IN}$  is the input voltage,  $f_s$  is the 340KHz switching frequency,  $L$  is the inductor value,  $C_O$  is the output capacitance value, and  $R_{ESR}$  is the equivalent series resistance (ESR) value of the output capacitor.

In the case of ceramic capacitors, the impedance at the switching frequency is dominated by the capacitance. The output voltage ripple is mainly caused by the capacitance. For simplification, the output voltage ripple can be estimated by:

$$\Delta V_{OUT} = \frac{V_{OUT}}{8 \times f_s^2 \times L \times C_O} \times \left( 1 - \frac{V_{OUT}}{V_{IN}} \right)$$

Where,  $V_{OUT}$  is the output voltage,  $V_{IN}$  is the input voltage,  $f_s$  is the 340KHz switching frequency,  $L$  is the inductor value, and  $R_{ESR}$  is the equivalent series resistance (ESR) value of the output capacitor.

In the case of tantalum or electrolytic capacitors, the ESR dominates the impedance at the switching frequency. For simplification, the output ripple can be approximated to:

$$\Delta V_{OUT} = \frac{V_{OUT}}{f_s \times L} \times \left( 1 - \frac{V_{OUT}}{V_{IN}} \right) \times R_{ESR}$$

Where,  $V_{OUT}$  is the output voltage,  $V_{IN}$  is the input voltage,  $f_s$  is the 340KHz switching frequency,  $L$  is the inductor value, and  $C_O$  is the output capacitance value.

The characteristics of the output capacitor also affect the stability of the regulation system. The MP1570 can be optimized for a wide range of capacitance and ESR values.

### Compensation Components

MP1570 employs current mode control for easy compensation and fast transient response. The system stability and transient response are controlled through the COMP pin. COMP pin is the output of the internal transconductance error amplifier. A series capacitor-resistor combination sets a pole-zero combination to control the characteristics of the control system.

The DC gain of the voltage feedback loop is given by:

$$A_{VDC} = R_{LOAD} \times G_{CS} \times A_{VEA} \times \frac{V_{FB}}{V_{OUT}}$$

Where,  $V_{FB}$  is the feedback voltage, 1.23V;  $V_{OUT}$  is the output voltage;  $A_{VEA}$  is the error amplifier voltage gain, 400V/V;  $G_{CS}$  is the current sense transconductance, 5.4A/V;  $R_{LOAD}$  is the load resistor value.

The system has 2 poles of importance. One is due to the compensation capacitor ( $C_3$ ) and the output resistor of error amplifier, and the other is due to the output capacitor and the load resistor. These poles are located at:

$$f_{P1} = \frac{G_{EA}}{2\pi \times C_3 \times A_{VEA}}$$

$$f_{P2} = \frac{1}{2\pi \times C_O \times R_{LOAD}}$$

Where,  $G_{EA}$  is the error amplifier transconductance, 800 $\mu$ A/V;  $A_{VEA}$  is the error amplifier voltage gain, 400V/V;  $R_{LOAD}$  is the load resistor value,  $C_O$  is the output capacitance value.

The system has one zero of importance, due to the compensation capacitor ( $C_3$ ) and the compensation resistor ( $R_3$ ). This zero is located at:

$$f_{Z1} = \frac{1}{2\pi \times C_3 \times R_3}$$

The system may have another zero of importance, if the output capacitor has a large capacitance and/or a high ESR value. The zero, due to the ESR and capacitance of the output capacitor, is located at:

$$f_{ESR} = \frac{1}{2\pi \times C_O \times R_{ESR}}$$

Where,  $C_O$  is the output capacitance value, and  $R_{ESR}$  is the ESR value of the output capacitor. In this case (as shown in Figure 2), a third pole set by the compensation capacitor ( $C_6$ ) and the compensation resistor ( $R_3$ ) is used to compensate the effect of the ESR zero on the loop gain. This pole is located at:

$$f_{P3} = \frac{1}{2\pi \times C_6 \times R_3}$$

The goal of compensation design is to shape the converter transfer function to get a desired loop gain. The system crossover frequency where the feedback loop has the unity gain is important.

Lower crossover frequencies result in slower line and load transient responses, while higher crossover frequencies could cause system unstable. A good rule of thumb is to set the crossover frequency to approximately one-tenth of the switching frequency. Switching frequency for the MP1570 is 340KHz, so the desired crossover frequency is ~34KHz.

Table 3 lists the typical values of compensation components for some standard output voltages with various output capacitors and inductors. The values of the compensation components have been optimized for fast transient responses and good stability at given conditions.

**Table 3—Compensation Values for Typical Output Voltage/Capacitor Combinations**

V <sub>OUT</sub>	L	C <sub>O</sub>	R3	C3	C6
1.8V	4.7μH	100μF Ceramic	5.6kΩ	3.3nF	None
2.5V	4.7-6.8μH	47μF Ceramic	4.7kΩ	4.7nF	None
3.3V	6.8-10μH	22μF×2 Ceramic	5.6kΩ	3.3nF	None
5V	10-15μH	22μF×2 Ceramic	7.5kΩ	3.3nF	None
12V	15-22μH	22μF×2 Ceramic	10kΩ	1.2nF	None
1.8	4.7μH	100μF SP-CAP	10kΩ	2.2nF	100pF
2.5V	4.7-6.8μH	47μF SP-CAP	5.6kΩ	3.3nF	None
3.3V	6.8-10μH	47μF SP-CAP	6.8kΩ	2.2nF	None
5V	10-15μH	47μF SP CAP	10kΩ	2.2nF	None
2.5V	4.7-6.8μH	560μF Al. 30mΩ ESR	10kΩ	7.5nF	1.5nF
3.3V	6.8-10μH	560μF Al 30mΩ ESR	10kΩ	10nF	1.5nF
5V	10-15μH	470μF Al. 30mΩ ESR	15kΩ	7.5nF	1nF
12V	15-22μH	220μF Al. 30mΩ ESR	15kΩ	10nF	390pF

To optimize the compensation components for conditions not listed in Table 2, the following procedure can be used.

1. Choose the compensation resistor (R3) to set the desired crossover frequency. Determine the R3 value by the following equation:

$$R3 = \frac{2\pi \times C_O \times f_C}{G_{EA} \times G_{CS}} \times \frac{V_{FB}}{V_{OUT}}$$

Where, V<sub>FB</sub> is the feedback voltage, 1.23V; V<sub>OUT</sub> is the output voltage; G<sub>EA</sub> is the error amplifier transconductance, 800μA/V; G<sub>CS</sub> is the current sense transconductance, 5.4 A/V; C<sub>O</sub> is the output capacitance value; and f<sub>C</sub> is the desired crossover frequency, 34KHz.

2. Choose the compensation capacitor (C3) to achieve the desired phase margin. For applications with typical inductor values, setting the compensation zero, f<sub>Z1</sub>, to one forth of the crossover frequency provides sufficient phase margin. Determine the C3 value by the following equation:

$$C3 = \frac{4}{2\pi \times R3 \times f_C}$$

Where, R3 is the compensation resistor value and f<sub>C</sub> is the desired crossover frequency, 34KHz.

3. Determine if the second compensation capacitor (C6) is required. It is required if the ESR zero of the output capacitor is located at less than half of the 340KHz switching frequency, or the following relationship is valid:

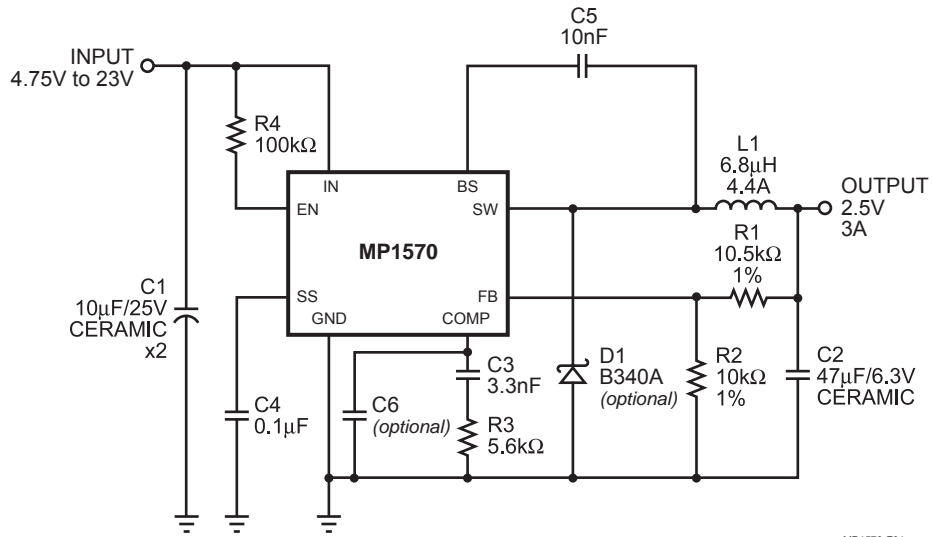
$$\frac{1}{2\pi \times C_O \times R_{ESR}} \leq \frac{f_S}{2}$$

Where, C<sub>O</sub> is the output capacitance value, R<sub>ESR</sub> is the ESR value of the output capacitor, and f<sub>S</sub> is the 340KHz switching frequency. If this is the case, then add the second compensation capacitor (C6) to set the pole f<sub>P3</sub> at the location of the ESR zero. Determine the C6 value by the equation:

$$C6 = \frac{C_O \times R_{ESR}}{R3}$$

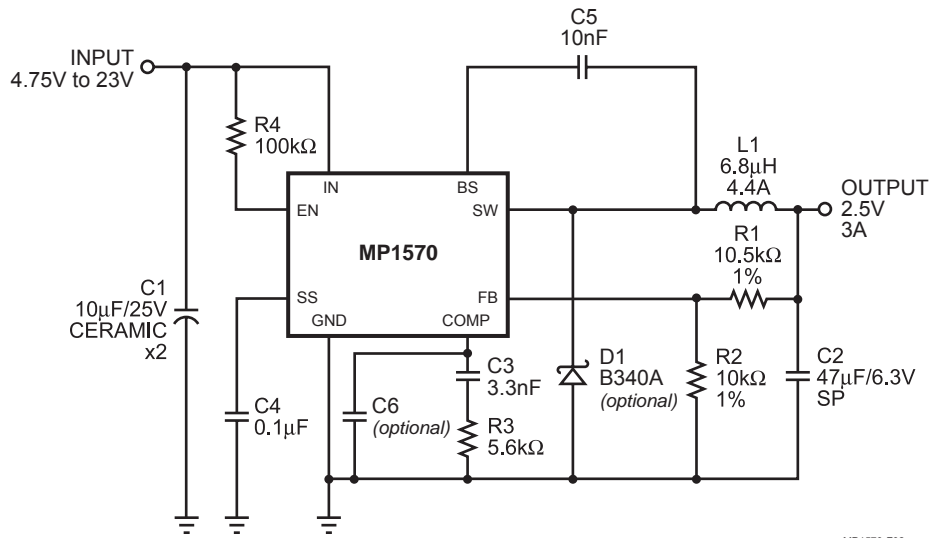
Where, C<sub>O</sub> is the output capacitance value, R<sub>ESR</sub> is the ESR value of the output capacitor, and R3 is the compensation resistor.

TYPICAL APPLICATIONS / TEST CIRCUITS



MP1570-F01

Figure 1— MP1570 with AVX 47µF, 6.3V Ceramic Output Capacitor

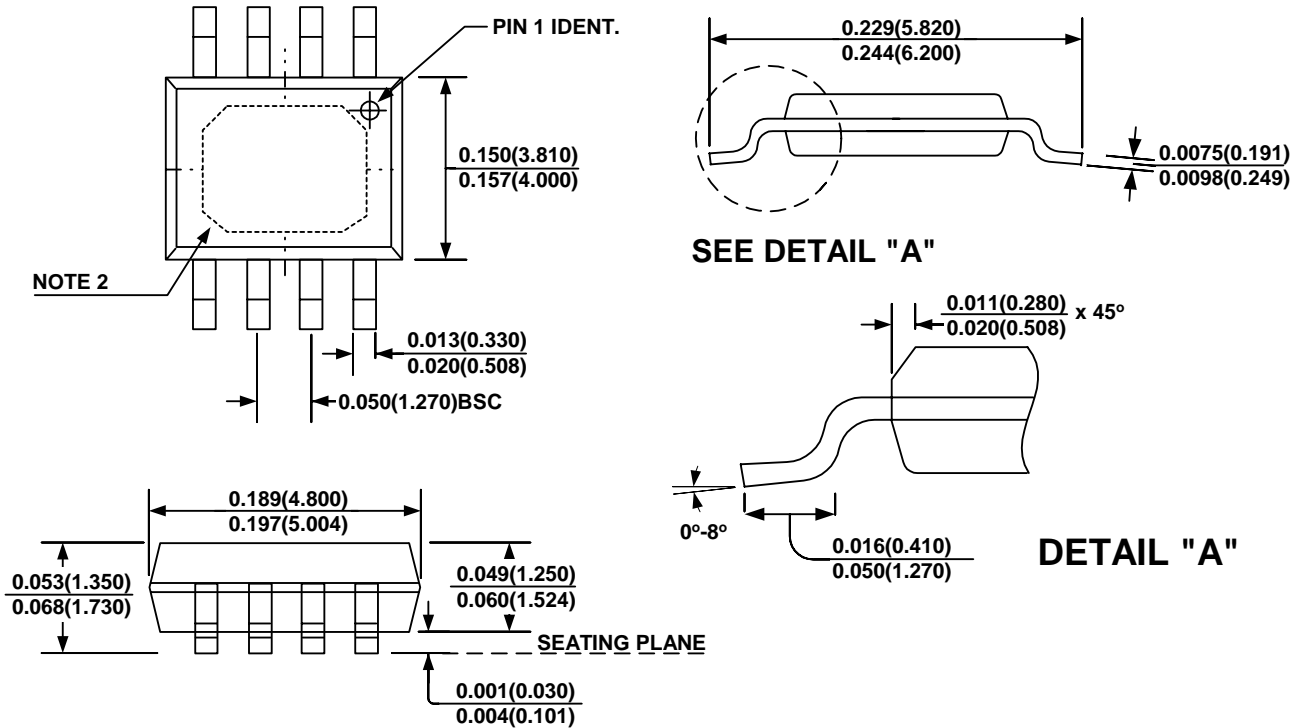


MP1570-F02

Figure 2— MP1570 with Panasonic 47µF, 6.3V Solid Polymer Output Capacitor

PACKAGE INFORMATION

SOIC8N (EXPOSED PAD)



NOTE:

- 1) Control dimension is in inches. Dimension in bracket is millimeters.
- 2) Exposed Pad Option Only (N-Package) ; 2.55+/- 0.25mm x 3.38 +/- 0.44mm.  
Recommended Solder Board Area: 2.80mm x 3.82mm = 10.7mm<sup>2</sup> (16.6mil<sup>2</sup>)

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