MP2372

3A, 28V, 925KHz Step-Down Converter

DESCRIPTION

The MP2372 is a 925KHz step-down regulator with a built-in power MOSFET. It achieves 3A continuous output current over a wide input supply range with excellent load and line regulation.

Current mode operation provides fast transient response and eases loop stabilization.

Fault condition protection includes cycle-by-cycle current limiting and thermal shutdown. Adjustable soft-start reduces the stress on the input source at turn-on. In shutdown mode, the regulator draws 20µA of supply current.

The MP2372 is available in an 8-pin SOIC package with an exposed pad, and requires a minimum number of readily available external components to complete a 3A step-down DC to DC converter solution.

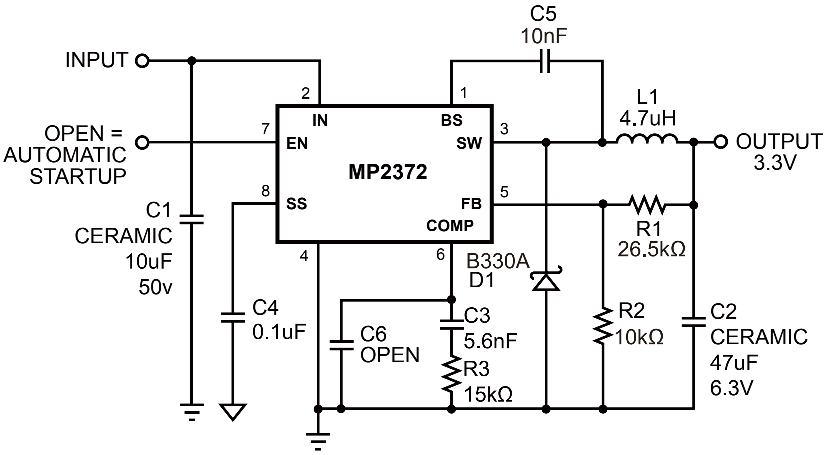
FEATURES

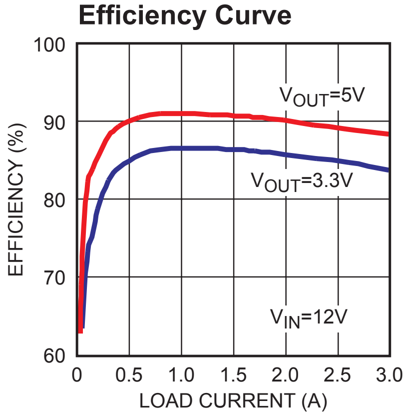
- 3A Continuous Output Current, 4A Peak Output Current
- Programmable Soft-Start
- 100mΩ Internal Power MOSFET Switch
- Stable with Low ESR Output Ceramic Capacitors
- Up to 91% Efficiency
- 20µA Shutdown Mode
- Fixed 925KHz Frequency
- Thermal Shutdown
- Cycle-by-Cycle Over Current Protection
- Wide 4.5V to 28V Operating Input Range
- Output is Adjustable From 0.92V to 21V
- Under Voltage Lockout

APPLICATIONS

- Distributed Power Systems
- Battery Chargers
- Pre-Regulator for Linear Regulators

TYPICAL APPLICATION





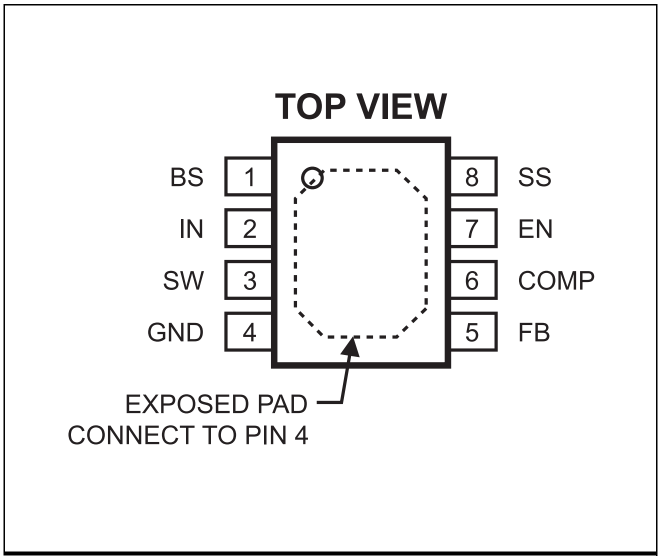
ORDERING INFORMATION

Part Number*	Package	Top Marking	Free Air Temperature (T_A)
MP2372DN	SOIC8N	MP2372DN	–40°C to +85°C

* For Tape & Reel, add suffix –Z (eg. MP2372DN–Z).

For RoHS compliant packaging, add suffix –LF (eg. MP2372DN–LF–Z)

PACKAGE REFERENCE



ABSOLUTE MAXIMUM RATINGS [1]

Supply Voltage V_{IN} ....–0.3V to +30V
Switch Voltage V_{SW} .....-1V (-5V for 10ns) to 31V
Boost Voltage V_{BS} .... V_{SW} + 6V
All Other Pins.................................–0.3V to +6V
Continuous Power Dissipation (T_A = +25°C)[2]
……………………………………………….2.5W
Junction Temperature...............................150°C
Lead Temperature ....260°C
Storage Temperature .... –65°C to +150°C

Recommended Operating Conditions [3]

Input Voltage V_{IN} ....4.5V to 28V
Operating Junct. Temp (T_J)..... –40°C to +125°C

Thermal Resistance [4]	θ_{JA}	θ_{JC}	
SOIC8N..................................	50	10 ...	°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 TA. 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 will cause excessive die temperature, and the regulator will 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} = 12V, T_A = +25°C, unless otherwise noted.

Parameters	Symbol	Condition	Min	Typ	Max	Units
Shutdown Supply Current	I_{supply}	V_{EN} = 0V		20	30	μA
Quiesent Supply Current	I_q	V_{EN} = 3V, V_{FB} =1.4V		1.3	1.5	mA
Feedback Voltage	V_{FB}	$4.5V \le V_{IN} \le 28V$, V_{COMP} < 2V	0.892	0.920	0.948	V
Error Amplifier Voltage Gain	A_{VEA}			400		V/V
Error Amplifier Transconductance	G_{EA}	$\Delta I_{COMP} = \pm 10\mu A$		690		μA/V
High-Side Switch-On Resistance	$R_{DS(ON)1}$			100		mΩ
Low-Side Switch-On Resistance	$R_{DS(ON)2}$			10		Ω
High-Side Switch Leakage Current		V_{EN} = 0V, V_{SW} = 0V		0.1	10	μA
Short Circuit Current Limit				6.5		A
Current Sense to COMP Transconductance	G_{CS}			6.0		A/V
Oscillation Frequency	f_S			925		KHz
Short Circuit Oscillation Frequency		V_{FB} = 0V		110		KHz
Maximum Duty Cycle	D_{MAX}	V_{FB} = 0.8V	76	80	99	%
Minimum On Time	T_{ON}			130		ns
EN Threshold Voltage			0.9	1.2	1.5	V
Enable Pull Up Current		V_{EN} = 0V	0.9	1.6	2.3	μA
Under Voltage Lockout Threshold Rising			2.3	2.6	2.9	V
Under Voltage Lockout Threshold Hysteresis				210		mV
Soft-Start Period		C_{SS} = 0.1μF L=10μH C_O=47μF I_O=3A(CC)		10		ms
Thermal Shutdown				160		°C

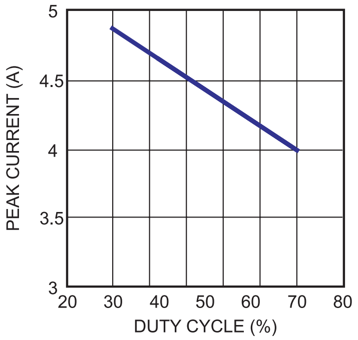
PIN FUNCTIONS

Pin #	Name	Description
1	BS	High-Side Gate Drive Boost Input. BS supplies the drive for the high-side N-Channel MOSFET switch. Connect a 10nF or greater capacitor from SW to BS to power the high side switch.
2	IN	Power Input. IN supplies the power to the IC, as well as the step-down converter switches. Drive IN with a 4.5V to 28V power source. Bypass IN to GND with a suitably large capacitor to eliminate noise on the input to the IC. See *Input Capacitor*
3	SW	Power Switching Output. SW is the switching node that supplies power to the output. Connect the output LC filter from SW to the output load. Note that a capacitor is required from SW to BS to power the high-side switch.
4	GND	Ground. Connect the exposed pad on backside to Pin 4.
5	FB	Feedback Input. FB senses the output voltage to regulate said voltage. Drive FB with a resistive voltage divider from the output voltage. The feedback threshold is 0.92V. See *Setting the Output Voltage*
6	COMP	Compensation Node. COMP is used to compensate the regulation control loop. Connect a series RC network from COMP to GND to compensate the regulation control loop. In some cases, an additional capacitor from COMP to GND is required. See *Compensation*
7	EN	Enable/UVLO. A voltage greater than 2.9V enables operation. For complete low current shutdown the EN pin voltage needs to be at less than 900mV. When the voltage on EN exceeds 1.2V, the internal regulator will be enabled and the soft-start capacitor will begin to charge. The MP2372 will start switching after the EN pin voltage reaches 2.9V.
8	SS	Soft-Start Control Input. SS controls the soft start period. Connect a capacitor from SS to GND to set the soft-start period. A 0.1µF capacitor sets the soft-start period to 10ms.

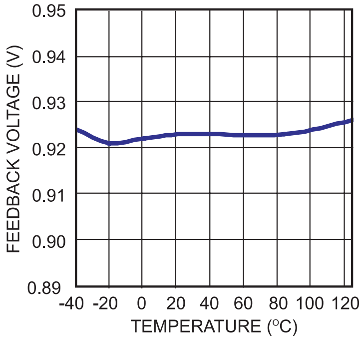
ELECTRICAL CHARACTERISTICS

V_{IN} = 12V, C1 = 10μF, C2 = 47μF, L = 10μH and T_A = +25°C, unless otherwise noted.

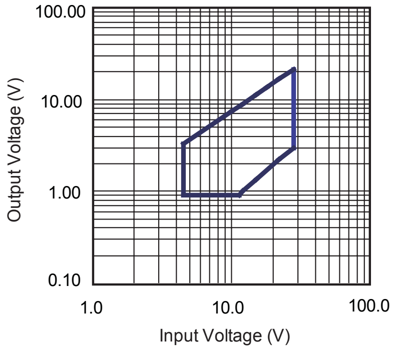
Peak Current vs Duty Cycle



Feedback Voltage vs Die Temperature

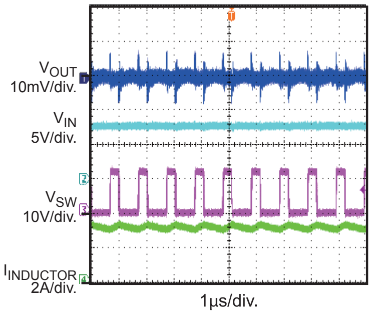


Operating Range



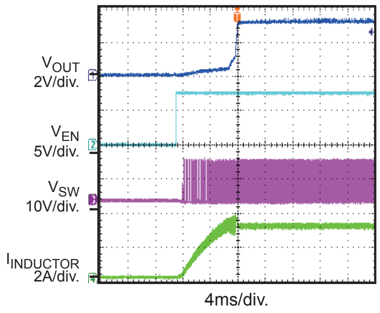
Steady State Operation

I_{OUT} = 3A, V_{OUT}=3.3V



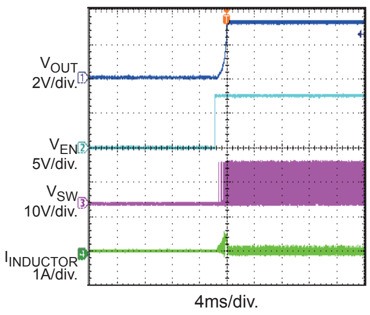
Startup Through Enable

I_{OUT} = 3A, V_{OUT}=3.3V



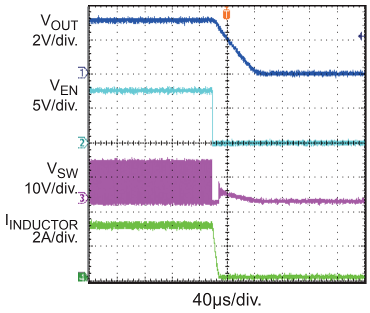
Startup Through Enable

I_{OUT} = 0A, V_{OUT}=3.3V



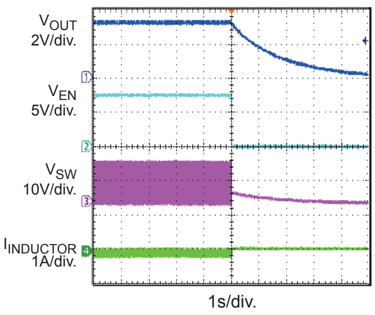
Shutdown Through Enable

I_{OUT} = 3A, V_{OUT}=3.3V

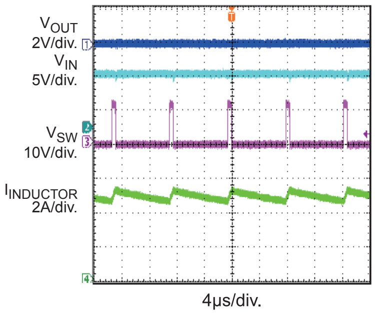


Shutdown Through Enable

I_{OUT} = 0A, V_{OUT}=3.3V



SHORT



OPERATION

The MP2372 is a current-mode step-down regulator. It regulates input voltages from 4.5V to 28V down to an output voltage as low as 0.92V, and is able to supply up to 3A of load current.

The MP2372 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 error amplifier. The output current of the transconductance error amplifier is presented at COMP where a network compensates the regulation control system. The voltage at COMP is compared to the switch current measured internally to control the output voltage.

The converter uses an internal N-Channel MOSFET switch to step-down the input voltage to the regulated output voltage. Since the MOSFET requires a gate voltage greater than the input voltage, a boost capacitor connected between SW and BS drives the gate. The capacitor is internally charged while SW is low.

An internal 10Ω switch from SW to GND is used to insure that SW is pulled to GND when SW is low to fully charge the BS capacitor.

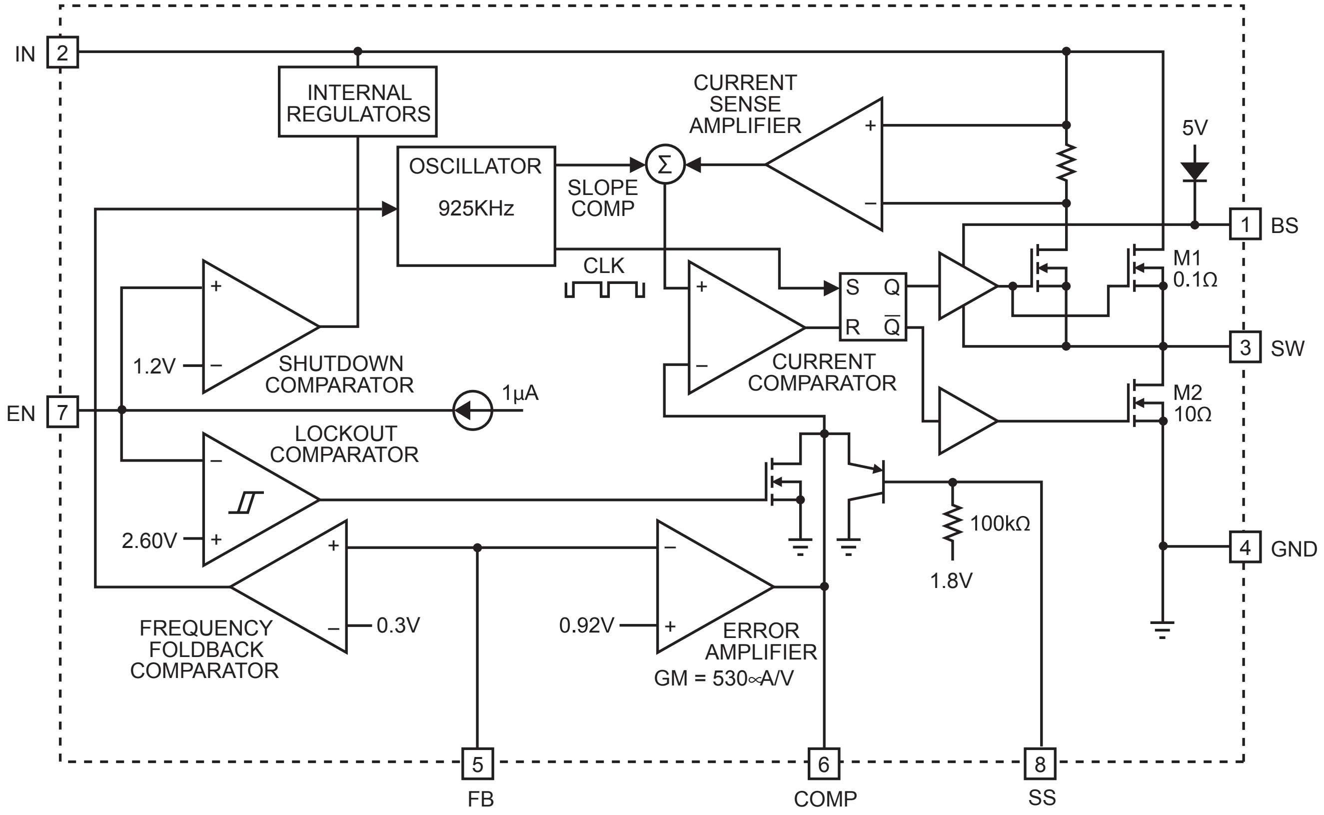


Figure 1—Functional Block Diagram

APPLICATION INFORMATION

COMPONENT SELECTION

(Refer to Figure 5)

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:

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

Where V_{FB} is the feedback voltage and V_{OUT} is the output voltage.

Thus the output voltage is:

$$V_{OUT} = 0.92 \times \frac{R1+R2}{R2}$$

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

$$R1 = 10.87 \times (V_{OUT} - 0.92)(k\Omega)$$

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:

$$L1 = \frac{V_{OUT}}{f_S \times \Delta I_L} \times \left(1 - \frac{V_{OUT}}{V_{IN}}\right)$$

Where V_{IN} is the input voltage, f_S is the 925KHz switching frequency and ΔI_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.

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
Sumida					
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
Toko					
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
Coilcraft					
DO3308	Open	Ferrite	9.4	13.0	3.0
DO3316	Open	Ferrite	9.4	13.0	5.1

Output Rectifier Diode

The output rectifier diode supplies the current to the inductor when the high-side switch is off. To reduce losses due to the diode forward voltage and recovery times, use a Schottky diode.

Choose a diode whose maximum reverse voltage rating is greater than the maximum input voltage, and whose current rating is greater than the maximum load current. Table 2 lists example Schottky diodes and manufacturers.

Input Capacitor

The input current to the step-down converter is discontinuous, therefore a capacitor is required

Table 2—Diode Selection Guide

Diode	Voltage/Current Rating	Manufacture
SK33	30V, 3A	Diodes Inc.
SK34	40V, 3A	Diodes Inc.
B330	30V, 3A	Diodes Inc.
B340	40V, 3A	Diodes Inc.
MBRS330	30V, 3A	On Semiconductor
MBRS340	40V, 3A	On Semiconductor

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.

Since the input capacitor (C1) absorbs the input switching current, it requires an adequate ripple current rating. The RMS current in the input capacitor can be estimated by:

$$I_{C1} = I_{LOAD} \times \sqrt{\frac{V_{OUT}}{V_{IN}} \times \left(1 - \frac{V_{OUT}}{V_{IN}}\right)}$$

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

$$I_{C1} = \frac{I_{LOAD}}{2}$$

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:

$$\Delta V_{IN} = \frac{I_{LOAD}}{f_S \times C1} \times \frac{V_{OUT}}{V_{IN}} \times \left(1 - \frac{V_{OUT}}{V_{IN}}\right)$$

Output Capacitor

The output capacitor (C2) 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:

$$\Delta V_{OUT} = \frac{V_{OUT}}{f_S \times L} \times \left(1 - \frac{V_{OUT}}{V_{IN}}\right) \times \left(R_{ESR} + \frac{1}{8 \times f_S \times C2}\right)$$

Where L is the inductor 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:

$$\Delta V_{OUT} = \frac{V_{OUT}}{8 \times f_S^{\,2} \times L \times C2} \times \left(1 - \frac{V_{OUT}}{V_{IN}}\right)$$

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:

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

The characteristics of the output capacitor also affect the stability of the regulation system. The MP2372 can be optimized for a wide range of capacitance and ESR values.

Compensation Components

MP2372 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 A_{VEA} is the error amplifier voltage gain, G_{CS} is the current sense transconductance and R_{LOAD} is the load resistor value.

The system has two poles of importance. One is due to the compensation capacitor (C3) 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:

$$f_{P1} = \frac{G_{EA}}{2\pi \times C3 \times A_{VEA}}$$

$$f_{P2} = \frac{1}{2\pi \times C2 \times R_{LOAD}}$$

Where G_{EA} is the error amplifier transconductance, 530μA/V.

The system has one zero of importance, due to the compensation capacitor (C3) and the compensation resistor (R3). This zero is located at:

$$f_{Z1} = \frac{1}{2\pi \times C3 \times R3}$$

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:

$$f_{ESR} = \frac{1}{2\pi \times C2 \times R_{ESR}}$$

In this case, a third pole set by the compensation capacitor (C6) and the compensation resistor (R3) is used to compensate the effect of the ESR zero on the loop gain. This pole is located at:

$$f_{P3} = \frac{1}{2\pi \times C6 \times R3}$$

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 or lower. The switching frequency for the MP2372 is 925KHz, so the desired crossover frequency is equal to or less than 92.5KHz.

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_{OUT} (V)	L (μH)	C2 (μF, Ceramic)	R3 (kΩ)	C3 (nF)	C6
1	1	47	3	10	None
1.2	1	47	5.1	6.8	None
1.8	2.2	47	7.5	3.3	None
2.5	2.2 - 4.7	47	10	4.7	None
3.3	2.2 - 4.7	47	15	5.6	None
5	4.7 – 6.8	2 x 22	20	4.7	None
12	6.8 - 10	2 x 22	44.2	2.2	None

To optimize the compensation components for conditions not listed in Table 3, 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:

$$R3 = \frac{2\pi \times C2 \times f_C}{G_{EA} \times G_{CS}} \times \frac{V_{OUT}}{V_{FB}}$$

Where f_C is the desired crossover frequency.

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

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

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 925KHz switching frequency, or the following relationship is valid:

$$\frac{1}{2\pi \times C2 \times R_{ESR}} < \frac{f_S}{2}$$

If this is the case, then add the second compensation capacitor (C6) to set the pole f_{P3} at the location of the ESR zero. Determine the C6 value by the equation:

$$C6 = \frac{C2 \times R_{ESR}}{R3}$$

PCB Layout Guide

PCB layout is very important to achieve stable operation. Please follow these guidelines and take Figure2 and 3 for references.

1) Keep the path of switching current short and minimize the loop area formed by Input cap, high-side MOSFET and schottky diode.
2) Keep the connection of schottky diode between SW pin and input power ground as short and wide as possible.
3) Ensure all feedback connections are short and direct. Place the feedback resistors and compensation components as close to the chip as possible.
4) Route SW away from sensitive analog areas such as FB.
5) Connect IN, SW, and especially GND respectively to a large copper area to cool the chip to improve thermal performance and long-term reliability. For single layer, do not solder exposed pad of the IC.

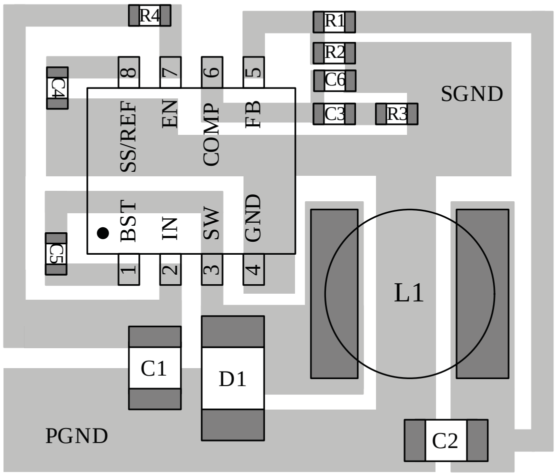
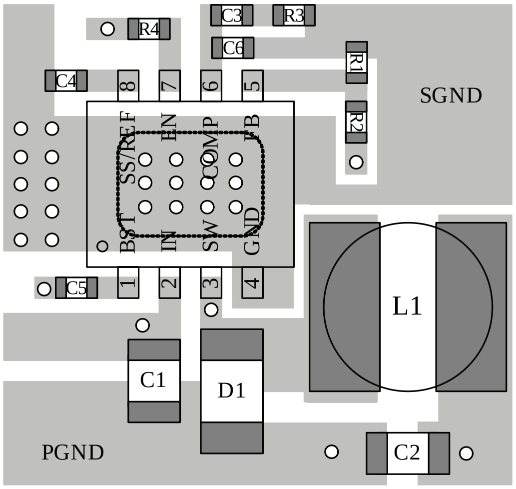
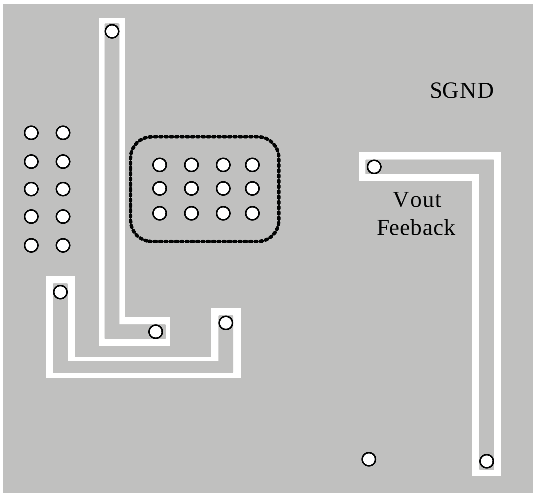


Figure2—PCB Layout for Single Layer





Top Layer

Bottom Layer

Figure3—PCB Layout for Double Layer

External Bootstrap Diode

Enough voltage of the bootstrap capacitor can enhance the efficiency of the regulator and increase the output capacity. The MP2372 can't output 3.3V/3A when the input voltage is only about 5V. An External charge circuit shown in Figure 4 can be a solution, and the applicable conditions of the circuit are:

- V_{IN} is only about 5V, and V_{OUT} is 3.3V or less
- The Duty cycle is high: $D=\frac{V_O}{V_{IN}}>60\%$
- The external bootstrap diode circuit can't be applied when $V_{IN}>6V$

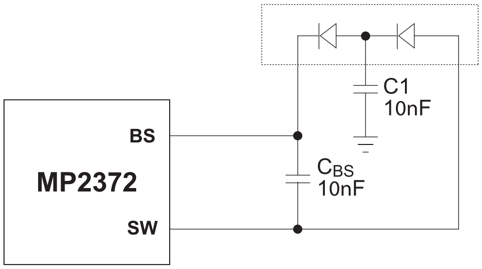


Figure 4—External Bootstrap Diode Circuit

D1 integrates two diodes in SOT-523 package such as BAV199T which will not increase the PCB size and the cost. The additional capacitor C1 can be only 10nF.

TYPICAL APPLICATION CIRCUITS

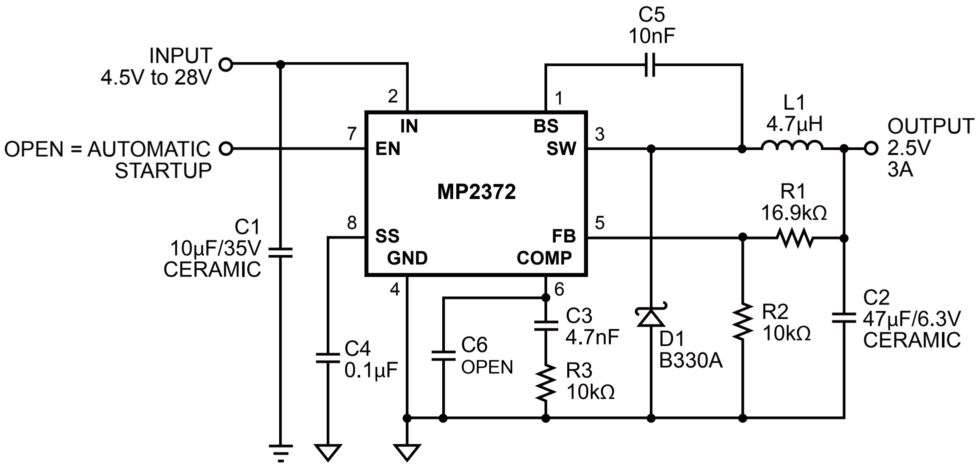


Figure 5—2.5V Output Typical Application Schematic

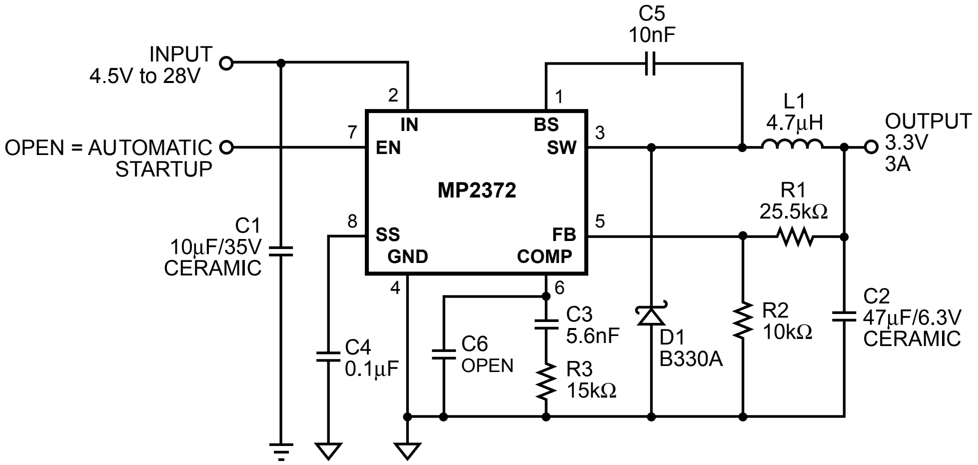
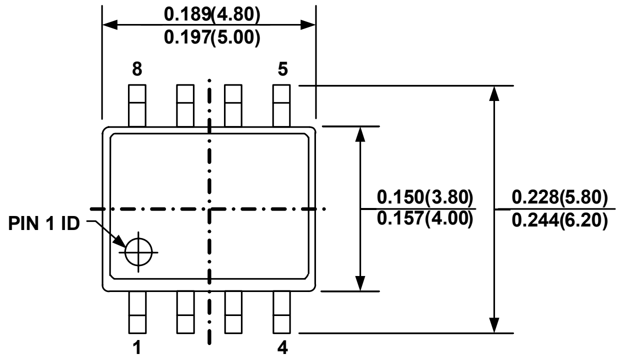


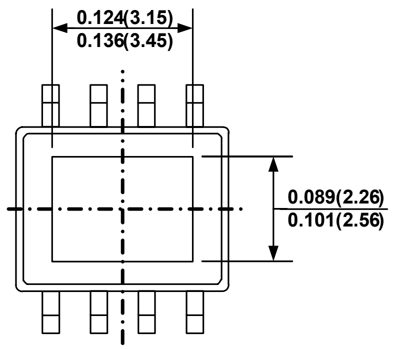
Figure 6—3.3V Output Typical Application Schematic

PACKAGE INFORMATION

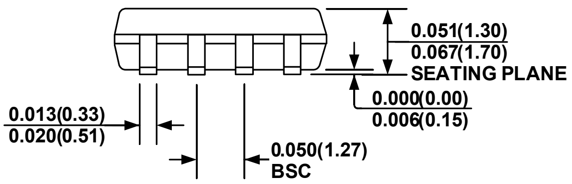
SOIC8N (EXPOSED PAD)



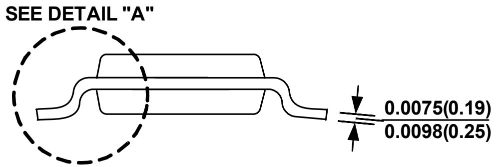
TOP VIEW



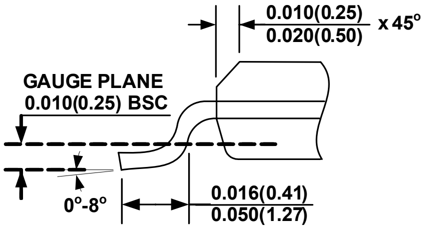
BOTTOM VIEW



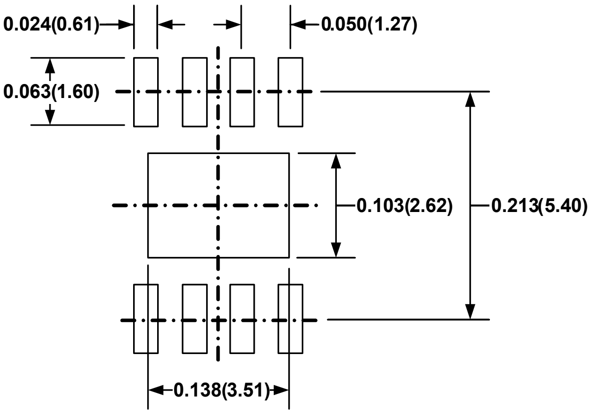
FRONT VIEW



SIDE VIEW



DETAIL "A"



RECOMMENDED LAND PATTERN

NOTE:

1) CONTROL DIMENSION IS IN INCHES DIMENSION IN BRACKET IS IN MILLIMETERS
2) PACKAGE LENGTH DOES NOT INCLUDE MOLD FLASH, PROTRUSIONS OR GATE BURRS
3) PACKAGE WIDTH DOES NOT INCLUDE INTERLEAD FLASH OR PROTRUSIONS.
4) LEAD COPLANARITY(BOTTOM OF LEADS AFTER FORMING) SHALL BE 0.004" INCHES MAX.
5) DRAWING CONFORMS TO JEDEC MS-012, VARIATION BA.
6) DRAWING IS NOT TO SCALE