MIC2169B

500kHz PWM Synchronous Buck Control IC

General Description

The MIC2169B is a high-efficiency, simple to use 500kHz PWM synchronous buck control IC housed in a small MSOP-10 ePad package. It allows compact DC/DC solutions with a minimal external component count and cost. The device features high output driver capability to drive loads up to 30A.

The MIC2169B operates from a 3V to 14.5V input, without the need of any additional bias voltage. The output voltage can be precisely regulated down to 0.8V. The adaptive all N-Channel MOSFET drive scheme allows efficiencies over 95% across a wide load range within the smallest possible printed circuit board space area.

The MIC2169B senses current across the high-side N-Channel MOSFET, eliminating the need for an expensive and lossy current-sense resistor. Current limit accuracy is maintained by a positive temperature coefficient that tracks the increasing $R_{DS(ON)}$ of the external MOSFET. Further cost and space are saved by the internal in-rush-current limiting digital soft-start. The MIC2169B is identical to the MIC2169A except it supports pre-bias loads. Internal pre-bias circuit prevents output voltage drooping and excessive reverse inductor current when powering up with a pre-bias voltage at the output.

The MIC2169B is available in a thermally capable 10-pin ePad MSOP package, with a wide junction operating range of -40°C to +125°C.

All support documentation can be found on Micrel's web site at www.micrel.com.
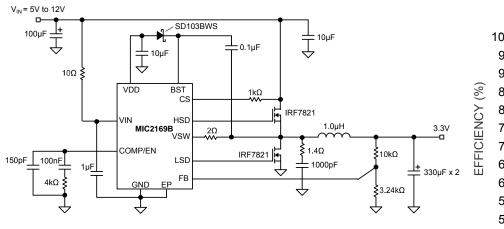
Features

- 3V to 14.5V input voltage range
- Adjustable output voltage down to 0.8V
- 500kHz PWM operation
- Up to 95% efficiency
- Output Pre-biased Protection
- Build-in 2.2Ω drivers to drive two N-channel MOSFETs
- Adaptive gate drive increases efficiency
- Simple voltage-mode PWM control with externally compensated
- Short minimum ON time of 30ns allowing very low duty cycle
- Fast transient response
- Adjustable current limit senses high-side N-Channel MOSFET current
- Hiccup mode short-circuit protection
- No external current-sense resistor
- Internal soft-start current source
- Dual function COMP and EN pin allows low-power shutdown
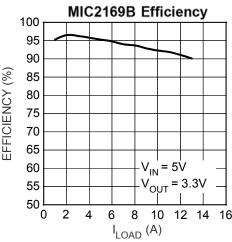- Available in a small size 10-pin MSOP ePad package

Applications

- Point-of-load DC/DC conversion
- High Current Power Supplies
- Telecom/Datacom and Networking Power Supplies
- Servers and Workstations
- Graphic cards and other PC Peripherals
- Set-top boxes
- LCD power supplies

Typical Application

V$_{IN}$ = 5V to 12V

100µF

SD103BWS

10µF

0.1µF

10µF

10Ω

VDD   BST

CS

1kΩ

IRF7821

VIN   HSD

MIC2169B   VSW

2Ω

1.0µH

3.3V

150pF   100nF

COMP/EN

1µF

LSD

IRF7821

1.4Ω

10kΩ

FB

1000pF

330µF x 2

4kΩ

GND   EP

3.24kΩ

MIC2169B Efficiency
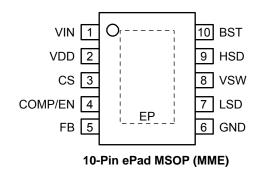
V$_{IN}$ = 5V

V$_{OUT}$ = 3.3V

EFFICIENCY (%)

I$_{LOAD}$ (A)

MIC2169B Adjustable Output 500kHz Converter

Ordering Information

Part Number	Frequency	Junction Temp. Range(1)	Package	Lead Finish
MIC2169BYMME	500kHz	-40° to +125°C	10-Pin ePad MSOP	Pb-Free

Pin Configuration

```
VIN      1 ○⌐ - - - - - ¬ 10  BST
VDD      2 |           |  9  HSD
CS       3 |           |  8  VSW
COMP/EN  4 |    EP     |  7  LSD
FB       5 |_ _ _ _ _ _|  6  GND
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10-Pin ePad MSOP (MME)

Pin Description

Pin Number	Pin Name	Pin Function
1	VIN	Supply Voltage (Input): +3V to +14.5V.
2	VDD	5V Internal Linear Regulator (Output): VDD is the external MOSFET gate drive supply voltage and an internal supply bus for the IC. When V_{IN} is <5V, short VDD to the input supply through a 10Ω resistor.
3	CS	Current Sense (Input): Current-limit comparator noninverting input. The current limit is sensed across the MOSFET during the ON time. The current can be set by the resistor in series with the CS pin.
4	COMP/EN	Compensation / Enable (Input): Dual function pin. Pin for external compensation. If this pin is pulled below 0.25V, with the reference fully up the device shuts down (50µA typical current draw).
5	FB	Feedback (Input): Input to error amplifier. Regulates error amplifier to 0.8V.
6	GND	Ground (Return).
7	LSD	Low-Side Drive (Output): High-current driver output for external synchronous MOSFET.
8	VSW	Switch (Return): High-side MOSFET driver return.
9	HSD	High-Side Drive (Output): High-current output-driver for the high-side MOSFET. When V_{IN} is between 3.0V to 5V, 2.5V threshold MOSFETs should be used. At V_{IN} > 5V, 4.5V threshold MOSFETs should be used.
10	BST	Boost (Input): Provides the drive voltage for the high-side MOSFET driver. The gate-drive voltage is higher than the source voltage by V_{DD} minus a diode drop.
ePad	EP	Connect to Ground

Absolute Maximum Ratings[1]

Supply Voltage (V_{IN}).......................................-0.3V to 15.5V
Boostrapped Voltage (V_{BST}) ....-0.3V to V_{IN} +6V
VSW ....-0.3V to 15.5V
CS 15.25V
FB ....-0.3V to 6V
Junction Temperature (T_J) ....-40°C ≤ T_J ≤ +125°C
Storage Temperature (T_S)..........................-65°C to +150°C
Peak Reflow Temperature (10 to 20 sec) ....+260°C
ESD (HBM) [3]..2kV
ESD (MM)..200V

Operating Ratings[2]

Supply Voltage (V_{IN}).....................................+3V to +14.5V
Ambient Temperature (T_A) ....-40°C to +125°C
Junction Thermal Resistance
    ePad MSOP (θ_{JA})..76.7°C/W
Output Voltage Range............................ 0.8V to V_{IN} × D_{MAX}

Electrical Characteristics[4]

T_J = 25°C, V_{IN} = 5V; bold values indicate –40°C ≤ T_J ≤ +125°C; unless otherwise specified.

Parameter	Condition	Min	Typ	Max	Units
Feedback Voltage Reference	(±1%)	0.792	0.8	0.808	V
Feedback Voltage Reference	(±2% over temp)	**0.784**	0.8	**0.816**	V
Feedback Bias Current			150	350	nA
Output Voltage Line Regulation			0.03		% / V
Output Voltage Load Regulation			0.5		%
Output Voltage Total Regulation	3V ≤ V_{IN} ≤ 14.5V; 1A ≤ I_{OUT} ≤ 10A; (V_{OUT} = 2.5V)[4]		0.6	**1.5**	%
Oscillator Section					
Oscillator Frequency		450	500	550	kHz
Maximum Duty Cycle		92			%
Minimum On-Time[5]			30	**60**	ns
Input and V_{DD} Supply					
PWM Mode Supply Current	V_{CS} = V_{IN} –0.25V; V_{FB} = 0.7V (output switching but excluding external MOSFET gate current.)		1.5	**3**	mA
Shutdown Quiescent Current	V_{COMP}/EN = 0V		50	**150**	μA
V_{COMP} Shutdown Threshold		0.1	0.25	0.35	V
V_{COMP} Shutdown Blanking Period	C_{COMP} = 100nF		675		μs
Digital Supply Voltage (V_{DD})	V_{IN} ≥ 6V	4.7	5	5.3	V
Error Amplifier					
DC Gain[5]			70		dB
Transconductance			1.1		$m\Omega^{-1}$

Parameter	Condition	Min	Typ	Max	Units
Soft-Start					
Soft-Start Current	After time out of internal timer. V_{COMP} = 0.8V	4	8.5	13	µA
Current Sense					
CS Over Current Trip Point	V_{CS} = V_{IN} –0.25V	160	200	240	µA
Temperature Coefficient			1800		ppm/°C
Gate Drivers					
Rise/Fall Time	Into 3000pF at V_{IN} > 5V		15		ns
Output Driver Impedance	Source, V_{IN} = 4.5V		2.2	3	Ω
	Sink, V_{IN} = 4.5V		1.3	3	Ω
	Source, V_{IN} = 3V		2.7	4	Ω
	Sink, V_{IN} = 3V		1.7	4	Ω
Driver Non-Overlap Time[5]			50		ns

Notes:

1. Absolute maximum ratings indicate limits beyond which damage to the component may occur. Electrical specifications do not apply when operating the device outside of its operating ratings. 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 power dissipation will result in excessive die temperature.

2. The device is not guaranteed to function outside its operating rating.

3. Devices are ESD sensitive, handling precautions required.

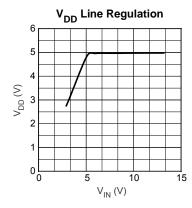4. Specification for packaged product only.
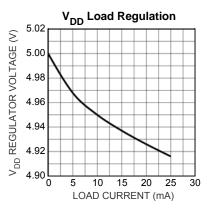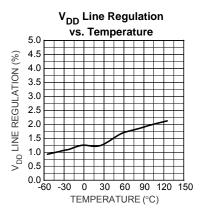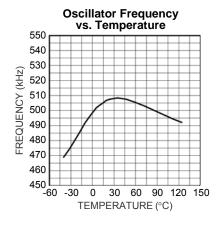
5. Guaranteed by design.

Typical Characteristics

PWM Mode Supply Current vs. Temperature

PWM Mode Supply Current vs. Supply Voltage

V$_{FB}$ Line Regulation

V$_{FB}$ vs. Temperature

V$_{DD}$ Line Regulation

V$_{DD}$ Load Regulation

V$_{DD}$ Line Regulation vs. Temperature

Oscillator Frequency vs. Temperature

Oscillator Frequency vs. Supply Voltage
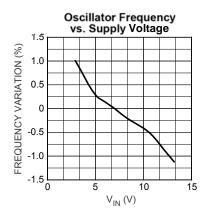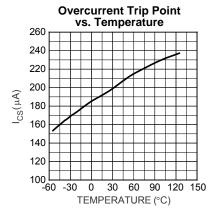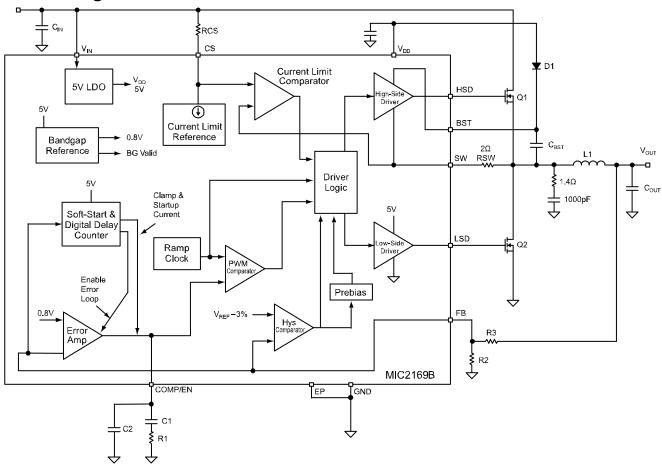
**Overcurrent Trip Point
vs. Temperature**

Functional Diagram

MIC2169B Block Diagram

Functional Description

The MIC2169B is a voltage mode, synchronous step-down switching regulator controller designed for high power. Current limit is implemented without the use of an external sense resistor. It includes an internal soft-start function which reduces the power supply input surge current at start-up by controlling the output voltage rise time, a PWM generator, a reference voltage, two MOSFET drivers, and short-circuit current limiting circuitry to form a complete 500kHz switching regulator. MIC2169B is identical to the MIC2169A except it supports pre-bias loads.

Theory of Operation

The MIC2169B is a voltage mode step-down regulator. The figure above illustrates the block diagram for the voltage control loop. The output voltage variation due to

load or line changes will be sensed by the inverting input of the transconductance error amplifier via the feedback resistors R3, and R2 and compared to a reference voltage at the non-inverting input. This will cause a small change in the DC voltage level at the output of the error amplifier which is the input to the PWM comparator. The other input to the comparator is a 0.95V to 1.45V triangular waveform. The comparator generates a rectangular waveform whose width t_{ON} is equal to the time from the start of the clock cycle t_0 until t_1, the time the triangle crosses the output waveform of the error amplifier. To illustrate the control loop, let us assume the output voltage drops due to sudden load turn-on, this would cause the inverting input of the error amplifier, which is divided down version of V_{OUT}, to be slightly less than the reference voltage, causing the output voltage of the error amplifier to go high. This will cause the PWM comparator to increase t_{ON} time of the top side

MOSFET, causing the output voltage to go up and bringing V_{OUT} back in regulation.

Soft-Start

The COMP/EN pin on the MIC2169B is used for the following three functions:

1. Disables the part by grounding this pin
2. External compensation to stabilize the voltage control loop
3. Soft-start

For better understanding of the soft-start feature, assume V_{IN} = 12V, and the MIC2169B is allowed to power-up by un-grounding the COMP/EN pin. The COMP pin has an internal 8.5µA current source that charges the external compensation capacitor. As soon as this voltage rises to 250mV (t = Cap_COMP × 0.25V/8.5µA) and V_{IN} crosses the 2.6V UVLO threshold, the MIC2169B allows the internal V_{DD} linear regulator to power up, and the chip's internal oscillator starts switching. At this point in time, the COMP pin current source increases to 40µA and an internal 11-bit counter starts counting which takes approximately 2ms to complete. During counting, the COMP voltage is clamped at 0.65V. After this counting cycle the COMP current source is reduced to 8.5µA and the COMP pin voltage rises from 0.65V to 0.95V, the bottom edge of the saw-tooth oscillator. This is the beginning of 0% duty cycle and it increases slowly causing the output voltage to rise slowly. The MIC2169B has one hysteretic comparator whose output is asserted high when V_{OUT} is within -3% of steady state. When the output voltage reaches 97% of programmed output voltage then the g_m error amplifier is enabled along with the hysteretic comparator output is asserted high. This point onwards, the voltage control loop (g_m error amplifier) is fully in control and will regulate the output voltage.

Soft-start time can be calculated approximately by adding the following four time frames:

t1 = Cap_COMP × 0.25V/8.5µA

t2 = 12 bit counter, approx 2ms

t3 = Cap_COMP × 0.3V/8.5µA

$$t4 = \left(\frac{V_{OUT}}{V_{IN}}\right) \times 0.5 \times \frac{Cap\_COMP}{8.5\mu A}$$

Soft-Start Time(Cap_COMP=100nF) = t1 + t2 + t3 + t4 = 2.9ms + 2ms + 3.5ms + 1.6ms = 10ms

Current Limit

The MIC2169B uses the $R_{DS(ON)}$ of the top power MOSFET to measure output current. Since it uses the drain to source resistance of the power MOSFET, it is

not very accurate. This scheme is adequate to protect the power supply and external components during a fault condition by cutting back the time the top MOSFET is on if the feedback voltage is greater than 0.67V. In case of a hard short when feedback voltage is less than 0.67V, the MIC2169B discharges the COMP capacitor to 0.65V, resets the digital counter and automatically shuts off the top gate drive, the g_m error amplifier is completely disabled, the –3% hysteretic comparators is asserted low, and the soft-start cycles restart from t_2 to t_4. This mode of operation is called the "hiccup mode" and its purpose is to protect the down stream load in case of a hard short. The circuit in Figure 1 illustrates the MIC2169B current limiting circuit.
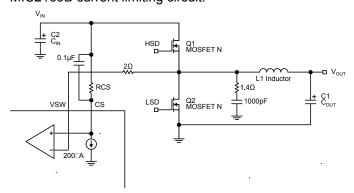

Figure 1. The MIC2169B Current Limiting Circuit

The current limiting resistor R_{CS} is calculated by the following equation:

$$R_{CS} = \frac{R_{DS(ON)Q1} \times I_L}{200\mu A}$$

where:

$$I_L = I_{LOAD} + \frac{Inductor\ Ripple\ Current}{2}$$

$$Inductor\ Ripple\ Current = V_{OUT} \times \frac{(V_{IN} - V_{OUT})}{V_{IN} \times F_S \times L}$$

F_S = 500kHz

200µA is the internal sink current to program the MIC2169B current limit.

The MOSFET $R_{DS(ON)}$ varies 30% to 40% with temperature; therefore, it is recommended to add a 50% margin to the load current (I_{LOAD}) in the above equation to avoid false current limiting due to increased MOSFET junction temperature rise. It is also recommended to connect R_{CS} resistor directly to the drain of the top MOSFET Q1, and the R_{SW} resistor to the source of Q1 to accurately sense the MOSFETs $R_{DS(ON)}$. To make the MIC2169B insensitive to board layout and noise generated by the switch node, a 1.4Ω resistor and a

1000pF capacitor is recommended between the switch node and GND.

Internal V$_{DD}$ Supply

The MIC2169B controller internally generates V$_{DD}$ for self biasing and to provide power to the gate drives. This V$_{DD}$ supply is generated through a low-dropout regulator and generates 5V from V$_{IN}$ supply greater than 5V. For supply voltage less than 5V, the V$_{DD}$ linear regulator is approximately 200mV in dropout. Therefore, it is recommended to short the V$_{DD}$ supply to the input supply through a 10Ω resistor for input supplies between 3.0V to 5V.

MOSFET Gate Drive

The MIC2169B high-side drive circuit is designed to switch an N-Channel MOSFET. The Functional Block Diagram on page 8 shows a bootstrap circuit, consisting of D1 and C$_{BST}$, supplies energy to the high-side drive circuit. Capacitor C$_{BST}$ is charged while the low-side MOSFET is on and the voltage on the VSW pin is approximately 0V. When the high-side MOSFET driver is turned on, energy from C$_{BST}$ is used to turn the MOSFET on. As the MOSFET turns on, the voltage on the VSW pin increases to approximately V$_{IN}$. Diode D1 is reversed biased and C$_{BST}$ floats high while continuing to keep the high-side MOSFET on. When the low-side switch is turned back on, C$_{BST}$ is recharged through D1. The drive voltage is derived from the internal 5V V$_{DD}$ bias supply. The nominal low-side gate drive voltage is 5V and the nominal high-side gate drive voltage is approximately 4.5V due the voltage drop across D1. An approximate 50ns delay between the high-side and low-side driver transitions is used to prevent current from simultaneously flowing unimpeded through both MOSFETs (shoot-through).

Adaptive gate drive is implemented on the high-side (off) to low-side (on) driver transition to reduce losses in the flywheel diode and to prevent shoot-through. This is operated by detecting the VSW pin; once this pin is detected to reach 1.5V, the high-side MOSFET can be assumed to be off and the low side driver is enabled.

Application Information

MOSFET Selection

The MIC2169B controller works from input voltages of 3V to 14.5V and has an internal 5V regulator to provide power to turn the external N-Channel power MOSFETs for high- and low-side switches. For applications where V_{IN} < 5V, the internal V_{DD} regulator operates in dropout mode, and it is necessary that the power MOSFETs used are sub-logic level and are in full conduction mode for V_{GS} of 2.5V. For applications when V_{IN} > 5V; logic-level MOSFETs, whose operation is specified at V_{GS} = 4.5V must be used. For the lower (<5V) applications, the V_{DD} supply can be connected directly to V_{IN} to help increase the driver voltage to the MOSFET.

It is important to note the on-resistance of a MOSFET increases with increasing temperature. A 75°C rise in junction temperature will increase the channel resistance of the MOSFET by 50% to 75% of the resistance specified at 25°C. This change in resistance must be accounted for when calculating MOSFET power dissipation and in calculating the value of current-sense (CS) resistor. Total gate charge is the charge required to turn the MOSFET on and off under specified operating conditions (V_{DS} and V_{GS}). The gate charge is supplied by the MIC2169B gate-drive circuit. At 500kHz switching frequency and above, the gate charge can be a significant source of power dissipation in the MIC2169B. At low output load, this power dissipation is noticeable as a reduction in efficiency. The average current required to drive the high-side MOSFET is:

$$I_{G[high-side](avg)} = Q_G \times f_S$$

where:

$I_{G[high-side](avg)}$ = average high-side MOSFET gate current.

Q_G = total gate charge for the high-side MOSFET taken from manufacturer's data sheet for V_{GS} = 5V.

The low-side MOSFET is turned on and off at V_{DS} = 0 because the freewheeling diode is conducting during this time. The switching loss for the low-side MOSFET is usually negligible. Also, the gate-drive current for the low-side MOSFET is more accurately calculated using CISS at V_{DS} = 0 instead of gate charge.

For the low-side MOSFET:

$$I_{G[low-side](avg)} = C_{ISS} \times V_{GS} \times f_S$$

Since the current from the gate drive comes from the input voltage, the power dissipated in the MIC2169B due to gate drive is:

$$P_{GATEDRIVE} = V_{IN} \times \left(I_{G[high-side](avg)} + I_{G[low-side](avg)} \right)$$

A convenient figure of merit for switching MOSFETs is the on resistance times the total gate charge $R_{DS(ON)} \times Q_G$. Lower numbers translate into higher efficiency. Low gate-charge logic-level MOSFETs are a good choice for use with the MIC2169B.

Parameters that are important to MOSFET switch selection are:

- Voltage rating
- On-resistance
- Total gate charge

The voltage ratings for the top and bottom MOSFET are essentially equal to the input voltage. A safety factor of 20% should be added to the V_{DS}(max) of the MOSFETs to account for voltage spikes due to circuit parasitics.

The power dissipated in the switching transistor is the sum of the conduction losses during the on-time ($P_{CONDUCTION}$) and the switching losses that occur during the period of time when the MOSFETs turn on and off (P_{AC}).

$$P_{SW} = P_{CONDUCTION} + P_{AC}$$

where:

$$P_{CONDUCTION} = I_{SW(rms)}^2 \times R_{SW}$$

$$P_{AC} = P_{AC(off)} + P_{AC(on)}$$

R_{SW} = on-resistance of the MOSFET switch

$$D = \text{duty cycle} = \left(\frac{V_O}{V_{IN}} \right)$$

Making the assumption the turn-on and turn-off transition times are equal; the transition times can be approximated by:

$$t_T = \frac{C_{ISS} \times V_{GS} + C_{OSS} \times V_{IN}}{I_G}$$

where:

C_{ISS} and C_{OSS} are measured at V_{DS} = 0

I_G = gate-drive current (1.4A for the MIC2169B)

The total high-side MOSFET switching loss is:

$$P_{AC} = (V_{IN} + V_D) \times I_{PK} \times t_T \times f_S$$

where:

t_T = switching transition time (typically 20ns to 50ns)

V_D = freewheeling diode drop, typically 0.5V

f_S it the switching frequency, nominally 500kHz

The low-side MOSFET switching losses are negligible and can be ignored for these calculations.

Inductor Selection

Values for inductance, peak, and RMS currents are required to select the output inductor. The input and output voltages and the inductance value determine the peak-to-peak inductor ripple current. Generally, higher inductance values are used with higher input voltages. Larger peak-to-peak ripple currents will increase the power dissipation in the inductor and MOSFETs. Larger output ripple currents will also require more output capacitance to smooth out the larger ripple current. Smaller peak-to-peak ripple currents require a larger inductance value and therefore a larger and more expensive inductor. A good compromise between size, loss and cost is to set the inductor ripple current to be equal to 20% of the maximum output current. The inductance value is calculated by the equation below.

$$L = \frac{V_{OUT} \times \left(V_{IN(max)} - V_{OUT} \right)}{V_{IN(max)} \times f_S \times 0.2 \times I_{OUT(max)}}$$

where:

f_S = switching frequency, 500kHz

0.2 = ratio of AC ripple current to DC output current

$V_{IN(max)}$ = maximum input voltage

The peak-to-peak inductor current (AC ripple current) is:

$$I_{PP} = \frac{V_{OUT} \times \left(V_{IN(max)} - V_{OUT} \right)}{V_{IN(max)} \times f_S \times L}$$

The peak inductor current is equal to the average output current plus one half of the peak-to-peak inductor ripple current.

$$I_{PK} = I_{OUT(max)} + 0.5 \times I_{PP}$$

The RMS inductor current is used to calculate the $I^2 \times R$ losses in the inductor.

$$I_{INDUCTOR} = \sqrt{\left(I_{OUT\_MAX} \right)^2 + \frac{I_{PP}^2}{12}}$$

Maximizing efficiency requires the proper selection of core material and minimizing the winding resistance. The high frequency operation of the MIC2169B requires the use of ferrite materials for all but the most cost sensitive applications.

Lower cost iron powder cores may be used but the increase in core loss will reduce the efficiency of the power supply. This is especially noticeable at low output power. The winding resistance decreases efficiency at the higher output current levels. The winding resistance must be minimized although this usually comes at the expense of a larger inductor. The power dissipated in the inductor is equal to the sum of the core and copper losses. At higher output loads, the core losses are

usually insignificant and can be ignored. At lower output currents, the core losses can be a significant contributor. Core loss information is usually available from the magnetics vendor. Copper loss in the inductor is calculated by the equation below:

$$P_{INDUCTORCu} = I_{INDUCTOR(rms)}{}^2 \times R_{WINDING}$$

The resistance of the copper wire, $R_{WINDING}$, increases with temperature. The value of the winding resistance used should be at the operating temperature.

$$R_{WINDING(hot)} = R_{WINDING(20°C)} \times (1 + 0.0042 \times (T_{HOT} - T_{20°C}))$$

where:

T_{HOT} = temperature of the wire under operating load

$T_{20°C}$ = ambient temperature

$R_{WINDING(20°C)}$ = room temperature winding resistance (usually specified by the manufacturer)

Output Capacitor Selection

The output capacitor values are usually determined by the capacitors ESR (equivalent series resistance). Voltage and RMS current capability are two other important factors selecting the output capacitor. Recommended capacitors are tantalum, low-ESR aluminum electrolytics, and POSCAPS. The output capacitor's ESR is usually the main cause of output ripple. The output capacitor ESR also affects the overall voltage feedback loop from stability point of view. See "Feedback Loop Compensation" section for more information. The maximum value of ESR is calculated:

$$R_{ESR} \le \frac{\Delta V_{OUT}}{I_{PP}}$$

where:

V_{OUT} = peak-to-peak output voltage ripple

I_{PP} = peak-to-peak inductor ripple current

The total output ripple is a combination of the ripple due to the output capacitors' ESR and the ripple due to the output capacitor. The total ripple is calculated below:

$$\Delta V_{OUT} = \sqrt{\left(\frac{I_{PP} \times (1-D)}{C_{OUT} \times f_S} \right)^2 + \left(I_{PP} \times R_{ESR} \right)^2}$$

where:

D = duty cycle

C_{OUT} = output capacitance value

f_S = switching frequency

The voltage rating of capacitor should be twice the voltage for a tantalum and 20% greater for aluminum electrolytic.

The output capacitor RMS current is calculated below:

$$I_{C_{OUT(rms)}} = \frac{I_{PP}}{\sqrt{12}}$$

The power dissipated in the output capacitor is:

$$P_{DISS(C_{OUT})} = \left(I_{C_{OUT(rms)}}\right)^2 \times R_{ESR(C_{OUT})}$$

Input Capacitor Selection

The input capacitor should be selected for ripple current rating and voltage rating. Tantalum input capacitors may fail when subjected to high inrush currents, caused by turning the input supply on. A tantalum input capacitor's voltage rating should be at least 2 times the maximum input voltage to maximize reliability. Aluminum electrolytic, OS-CON, and multilayer polymer film capacitors can handle the higher inrush currents without voltage derating. The input voltage ripple will primarily depend on the input capacitor's ESR. The peak input current is equal to the peak inductor current, so:

$$\Delta V_{IN} = I_{INDUCTOR(peak)} \times R_{ESR(C_{IN})}$$

The input capacitor must be rated for the input current ripple. The RMS value of input capacitor current is determined at the maximum output current. Assuming the peak-to-peak inductor ripple current is low:

$$I_{C_{IN(rms)}} \approx I_{OUT(max)} \times \sqrt{D \times (1-D)}$$

The power dissipated in the input capacitor is:

$$P_{DISS(C_{IN})} = \left(I_{CIN(rms)}\right)^2 \times R_{ESR(C_{IN})}$$

Voltage Setting Components

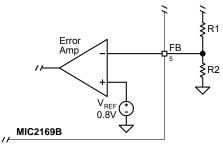The MIC2169B requires two resistors to set the output voltage as shown in Figure 2.

Figure 2. Voltage-Divider Configuration

The output voltage is determined by the equation:

$$V_O = V_{REF} \times \left(1 + \frac{R1}{R2}\right)$$

where: V_{REF} for the MIC2169B is typically 0.8V

A typical value of R1 can be between 3kΩ and 10kΩ. If R1 is too large, it may allow noise to be introduced into the voltage feedback loop. If R1 is too small, in value, it will decrease the efficiency of the power supply, especially at light loads. Once R1 is selected, R2 can be calculated using:

$$R2 = \frac{V_{REF} \times R1}{V_O - V_{REF}}$$

External Schottky Diode

An external freewheeling diode is used to keep the inductor current flow continuous while both MOSFETs are turned off. This dead time prevents current from flowing unimpeded through both MOSFETs and is typically 50ns. The diode conducts twice during each switching cycle. Although the average current through this diode is small, the diode must be able to handle the peak current.

$$I_{D(avg)} = I_{OUT} \times 2 \times 50ns \times f_S$$

The reverse voltage requirement of the diode is:

$$V_{DIODE(rrm)} = V_{IN}$$

The power dissipated by the Schottky diode is:

$$P_{DIODE} = I_{D(avg)} \times V_F$$

where:

V_F = forward voltage at the peak diode current

The external Schottky diode, D1, is not necessary for circuit operation since the low-side MOSFET contains a parasitic body diode. The external diode will improve efficiency and decrease high frequency noise. If the MOSFET body diode is used, it must be rated to handle the peak and average current. The body diode has a relatively slow reverse recovery time and a relatively high forward voltage drop. The power lost in the diode is proportional to the forward voltage drop of the diode. As the high-side MOSFET starts to turn on, the body diode becomes a short circuit for the reverse recovery period, dissipating additional power. The diode recovery and the circuit inductance will cause ringing during the high-side MOSFET turn-on. An external Schottky diode conducts at a lower forward voltage preventing the body diode in the MOSFET from turning on. The lower forward voltage drop dissipates less power than the body diode. The lack of a reverse recovery mechanism in a Schottky diode causes less ringing and less power loss. Depending on the circuit components and operating conditions, an external Schottky diode will give a $^1/_2$% to 1% improvement in efficiency.

Feedback Loop Compensation

The MIC2169B controller comes with an internal transconductance error amplifier used for compensating the voltage feedback loop by placing a capacitor (C1) in series with a resistor (R1) and another capacitor C2 in parallel from the COMP pin to ground. See "Functional

Block Diagram."

Power Stage

The power stage of a voltage mode controller has an inductor, L1, with its winding resistance (DCR) connected to the output capacitor, C_{OUT}, with its electrical series resistance (ESR) as shown in Figure 3. The transfer function G(s), for such a system is:
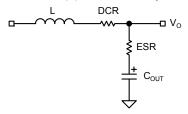

Figure 3.   The Output LC Filter in a Voltage Mode Buck Converter

$$G(s) = \left(\frac{(1 + ESR \times s \times C)}{DCR \times s \times C + s^2 \times L \times C + 1 + ESR \times s \times C} \right)$$

Plotting this transfer function with the following assumed values (L=1μH, DCR=0.009Ω, C_{OUT}=660μF, ESR=0.025Ω) gives lot of insight as to why one needs to compensate the loop by adding resistor and capacitors on the COMP pin. Figures 4 and 5 show the gain curve and phase curve for the above transfer function.
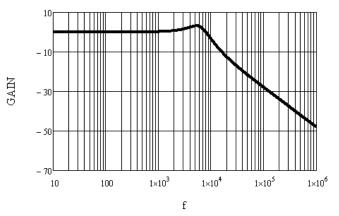

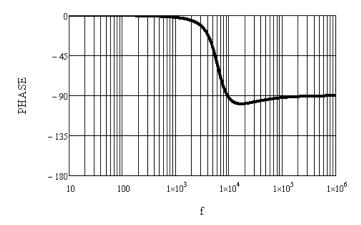**Figure 4.   The Gain Curve for G(s)**

Figure 5.   Phase Curve for G(s)

It can be seen from the transfer function G(s) and the gain curve that the output inductor and capacitor create a two pole system with a break frequency at:

$$f_{LC} = \frac{1}{2 \times \pi \sqrt{L \times C_{OUT}}}$$

Therefore, f_{LC} = 6.2kHz

By looking at the phase curve, it can be seen that the output capacitor ESR (0.025Ω) cancels one of the two poles (LC_{OUT}) system by introducing a zero at:

$$f_{ZERO} = \frac{1}{2 \times \pi \times ESR \times C_{OUT}}$$

Therefore, F_{ZERO} = 9.6kHz.

From the point of view of compensating the voltage loop, it is recommended to use higher ESR output capacitors since they provide a 90° phase gain in the power path. For comparison purposes, Figure 6, shows the same phase curve with an ESR value of 0.002Ω.
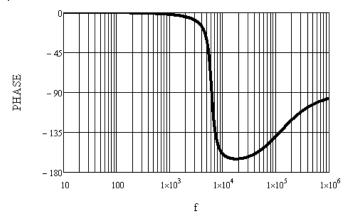

Figure 6.   The Phase Curve with ESR = 0.002Ω

It can be seen from Figure 5 that at 50kHz, the phase is approximately –90° versus Figure 6 where the number is –150°. This means that the transconductance error amplifier has to provide a phase boost of about 45° to achieve a closed loop phase margin of 45° at a crossover frequency of 50kHz for Figure 5, versus 105° for Figure 6. The simple RC and C2 compensation scheme allows a maximum error amplifier phase boost of about 90°. Therefore, it is easier to stabilize the MIC2169B voltage control loop by using high ESR value output capacitors.

g_m Error Amplifier

It is undesirable to have high error amplifier gain at high frequencies because high frequency noise spikes would be picked up and transmitted at large amplitude to the output, thus, gain should be permitted to fall off at high frequencies. At low frequency, it is desired to have high open-loop gain to attenuate the power line ripple. Thus, the error amplifier gain should be allowed to increase rapidly at low frequencies.

The transfer function with R1, C1, and C2 for the internal g_m error amplifier can be approximated by the following equation:

$$\text{Error Amplifier}(z) = gm \times \left[\frac{1 + s \times R1 \times C1}{s \times (C1 + C2) \times \left(1 + s \times R1 \times \frac{C1 \times C2}{C1 + C2}\right)} \right]$$

The above equation can be simplified by assuming C2<<C1,

$$\text{Error Amplifier}(z) = gm \times \left[\frac{1 + s \times R1 \times C1}{s \times C1 \times (1 + s \times R1 \times C2)} \right]$$

From the above transfer function, one can see that R1 and C1 introduce a zero and R1 and C2 a pole at the following frequencies:

$F_{ZERO} = 1/2 \pi \times R1 \times C1$

$F_{POLE} = 1/2 \pi \times C2 \times R1$

$F_{POLE@origin} = 1/2 \pi \times C1$

Figures 7 and 8 show the gain and phase curves for the above transfer function with R1 = 4.02k, C1 = 100nF, C2 = 150pF, and gm = 1.1mΩ⁻¹.
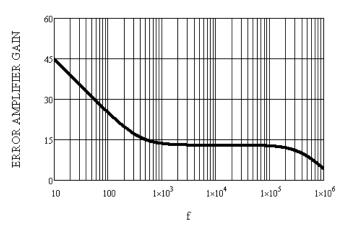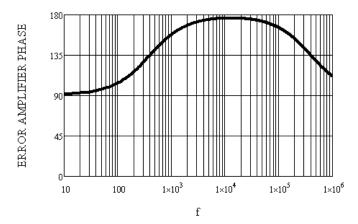

Figure 7. Error Amplifier Gain Curve

Figure 8. Error Amplifier Phase Curve

Total Open-Loop Response

The open-loop response for the MIC2169B controller is easily obtained by adding the power path and the error amplifier gains together, since they already are in Log scale. It is desirable to have the gain curve intersect zero dB at tens of kilohertz, this is commonly called crossover frequency; the phase margin at crossover frequency should be at least 45°. Phase margins of 30° or less cause the power supply to have substantial ringing when subjected to transients, and have little tolerance for component or environmental variations.

Figures 9 and 10 show the open-loop gain and phase margin for the 5V input and 1.8V output application, and it can be seen from Figure 9 that the gain curve intersects the 0dB at approximately 50kHz, and from Figure 10 that at 50kHz, the phase shows approximately 74° of margin.
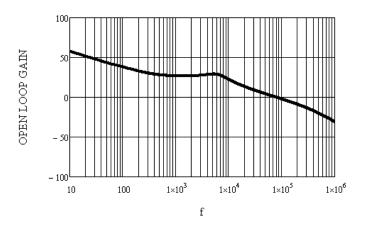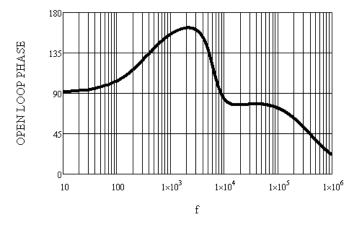
Figure 9.   Open-Loop Gain Margin

Figure 10.   Open-Loop Phase Margin

Pre-Biased Loads

The MIC2169B supports pre-biased loads. Some applications have a pre-existing voltage on the output. This pre-existing or pre-biased load is generated by an external supply (other than the MIC2169B). During startup without pre-bias support, MIC2169A will pull the output voltage to ground through the inductor and low side FET (see figure 11).

The MIC2169B prevents the current sinking of any pre-existing voltage source at the output (see figure 12). It does this by keeping the low side FET off during the soft start period. In some applications this pre-bias current sink is not a problem, and the MIC2169A may be used. In some applications the pre-bias current sink may cause a problem, and the MIC2169B should be used. The MIC2169B can support up to 90% of a pre-bias condition (up to 90% of the final regulated output voltage) see figure 13.
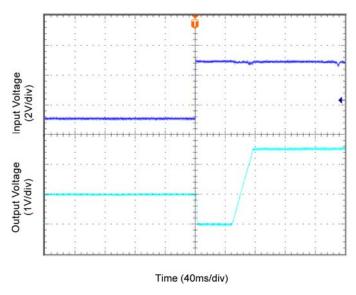

Figure 11. MIC2169A Startup without Pre-Bias Support

Figure 11 shows MIC2169A startup with a pre-bias of 1V on the output, in which the pre-existing output voltage discharges during soft start.

Figure 12. MIC2169B startup with pre-bias support
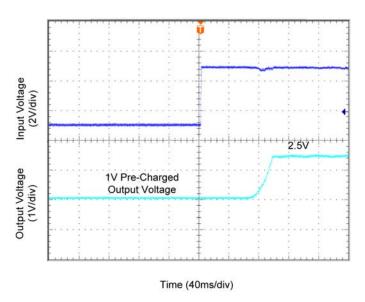
Figure 12 shows MIC2169B startup with a pre-bias of 1V on the output, in which the pre-existing output voltage has no discharge.
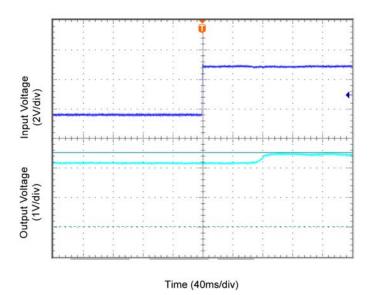
Time (40ms/div)

Fig 13 MIC2169B startup with pre-bias support, pre-bias at 90% of Vout_final

Figure 13 shows MIC2169B startup with a pre-bias of 2.2V on the output (90% of V_{OUT}) without the pre-existing output voltage discharge.

Design and PCB Layout Guideline

Warning!!! To minimize EMI and output noise, follow these layout recommendations.

PCB Layout is critical to achieve reliable, stable and efficient performance. A ground plane is required to control EMI and minimize the inductance in power, signal and return paths.

The following guidelines should be followed to insure proper operation of the MIC2169B converter.

IC

- Place the IC and MOSFETs close to the point of load (POL).
- Use fat traces to route the input and output power lines.
- Signal and power grounds should be kept separate and connected at only one location.

Input Capacitor

- Place the V_{IN} input capacitor next.
- Place the V_{IN} input capacitors on the same side of the board and as close to the MOSFETs as possible.
- Keep both the V_{IN} and power GND connections short.
- Place several vias to the ground plane close to the V_{IN} input capacitor ground terminal.
- Use either X7R or X5R dielectric input capacitors. Do not use Y5V or Z5U type capacitors.
- Do not replace the ceramic input capacitor with any other type of capacitor. Any type of capacitor can be placed in parallel with the input capacitor.
- If a Tantalum input capacitor is placed in parallel with the input capacitor, it must be recommended for switching regulator applications and the operating voltage must be derated by 50%.
- In "Hot-Plug" applications, a Tantalum or Electrolytic bypass capacitor must be used to limit the over-voltage spike seen on the input supply with power is suddenly applied.
- An additional Tantalum or Electrolytic bypass input capacitor of 22µF or higher is required at the input power connection.
- Use a 5Ω resistor from the input supply to the VDD pin on the MIC2169B. Also, place a 1µF ceramic capacitor from this pin to GND, preferably not through a via. The capacitor must be located right at the IC. The V_{DD} terminal is very noise sensitive and placement of the capacitor is very critical. Connections must be made with wide trace.

Inductor

- Keep the inductor connection to the switch node (SW) short.
- Do not route any digital lines underneath or close to the inductor.
- Keep the switch node (SW) away from the feedback (FB) pin.
- To minimize noise, place a ground plane underneath the inductor.

Output Capacitor

- Use a wide trace to connect the output capacitor ground terminal to the input capacitor ground terminal.
- Phase margin will change as the output capacitor value and ESR changes. Contact the factory if the output capacitor is different from what is shown in the BOM.
- The feedback trace should be separate from the power trace and connected as close as possible to the output capacitor. Sensing a long high current load trace can degrade the DC load regulation.

MOSFETs

- Low gate charge MOSFETs should be used to maximize efficiency, such as Si4800, Si4804BDY, IRF7821, IRF8910, FDS6680A and FDS6912A, etc.

RC Snubber

- Add a RC snubber of 1.4Ω resistor and a 1000pF capacitor from the switch node to ground pin. Place the snubber on the same side of the board and as close to the MOSFETs as possible. See page 8, Current Limiting section for more detail.

Schottky Diode (Optional)

- Place the Schottky diode on the same side of the board as the MOSFETs and V_{IN} input capacitor.
- The connection from the Schottky diode's Anode to the input capacitors ground terminal must be as short as possible.
- The diode's Cathode connection to the switch node (SW) must be keep as short as possible.
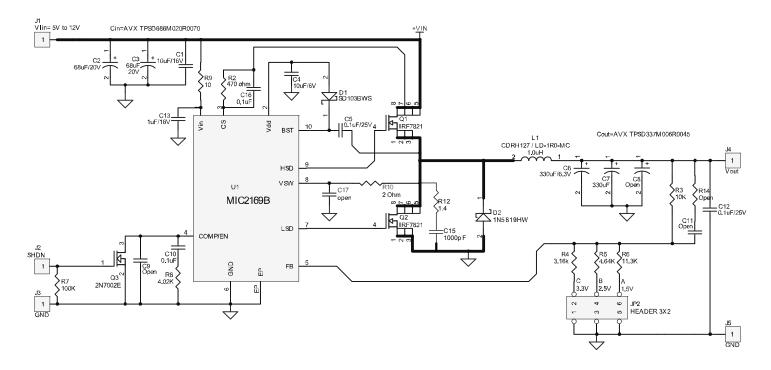
Others

- Connect the current limiting (R2) resistor directly to the drain of top MOSFET Q1.
- The feedback resistors R3 and R4/R5/R6 should be placed close to the FB pin. The top side of R3 should connect directly to the output node. Run this trace away from the switch node (junction of Q1, Q2, and L1). The bottom side of R3 should connect to the GND pin on the MIC2169B.

- The compensation resistor and capacitors should be placed right next to the COMP pin and the other side should connect directly to the GND pin on the MIC2169B rather than going to the plane.

- Add place holders for gate resistors on the top and bottom MOSFET gate drives. If necessary, gate resistors of 10Ω or less should be used.

Evaluation Board Schematics

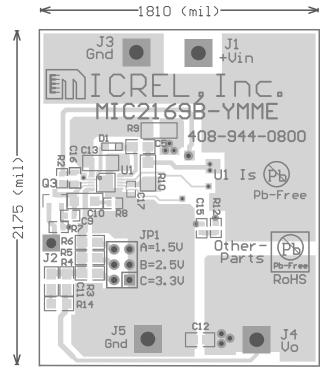**MIC2169B Evaluation Board Schematic**

Bill of Materials

Item	Part Number	Manufacturer	Description	Qty.
U1	**MIC2169B-YMME**	**Micrel, Inc.**	**Buck controller**	**1**
Q1, Q2	IRF7821-TR SI4390DY	IR Vishay	30V, N channel HEXFET , Power MOSFET	2 0
Q3	2N7002E	On Semiconductor	60V, N channel MOSFET	0
D1	SD103BWS	Vishay	30V, Schottky Diode	1
D2	1N5819HW SL04 CMMSH1-40	Diodes Inc. Vishay Central Semi	40V, Schottky Diode	1 0 0
L1	CDRH127LDNP-1R0NC HC5-1R0 SER1360-1R0	Sumida Cooper Electronic Coilcraft	1.0uH, 10A inductor	1 0 0
C1	C3225X7R1C106M	TDK	10uF/16V, X7R Ceramic cap.	1
C2, C3	TPSD686M020R0070 594D686X0020D2T	AVX Vishay/Sprague	68uF, 20V Tantalum	2 0
C4	C2012X5R0J106M	TDK	10uF/6.3V, 0805 Ceramic cap.	1
C5, C10, C12	VJ1206Y104KXXAT	Vishay Victramon	0.1uF/25V Ceramic cap.	3
C6, C7	TPSD337M006R0045	AVX	330uF, 6.3V, Tantalum	2
C8		Vishay/Sprague	Open	0
C9, C11		Vishay Dale	open	0
C13	C2012X7R1C105K GRM21BR71C105KA01B. VJ1206S105KXJAT	TDK muRata Vishay Victramon	1uF/16V, 0805 Ceramic cap.	1 0 0
C15	VJ0603A102KXXAT	Vishay Victramon	1000pF /25V, 0603 , NPO	1
C16	VJ0603Y104KXXAT	Vishay Victramon	0.1uF/25V Ceramic cap.	1
C17			open	1
R2	CRCW06034700JRT1	Vishay	470Ω, 0603, 1/16W, 5%.	1
R3	CRCW08051002FRT1	Vishay	10kΩ, 0805, 1/10W, 1%	1
R4	CRCW08053161FRT1	Vishay	3.16kΩ, 0805, 1/10W , 1%	1
R5	CRCW08054641FRT1	Vishay	4.64kΩ, 0805, 1/10W , 1%	1
R6	CRCW08051132FRT1	Vishay	11.3kΩ, 0805, 1/10W, 1%	1
R7	CRCW08051003FRT1	Vishay	100kΩ, 0805, 1/10W, 1%	
R8	CRCW06034021FRT1	Vishay	4.02kΩ, 0603 ,1/16W, 1%	1
R9	CRCW120610R0FRT1	Vishay	10Ω, 1/8W, 1206, 1%	1
R10	CRCW12062R00FRT1	Vishay	2Ω, 1/8 W, 1206, 1%	1
R12	CRCW12061R40FRT1	Vishay	1.4Ω, 1/8 W, 1206, 1%	
R14			Open	0
J1, J3, J4, J5	2551-2-00-01-00-00-07-0	MilMax	Turret Pins	4

Notes:

1. Micrel.Inc   408-944-0800
2. Vishay corp   206-452-5664
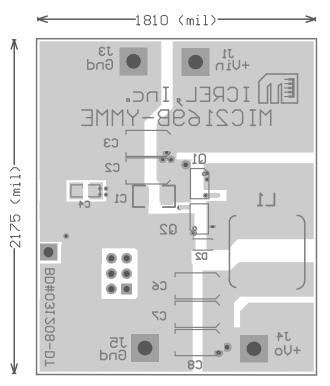3. Diodes. Inc   805-446-4800

4. Sumida   408-321-9660
5. TDK   847-803-6100
6. muRata   800-831-9172
7. AVX   843-448-9411
8. International Rectifier   847-803-6100
9. Fairchild Semiconductor   207-775-8100
10. Cooper Electronic   561-752-5000
11. Coilcraft   1-800-322-2645
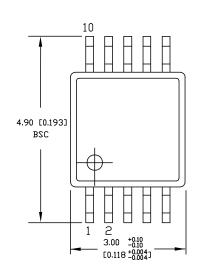12. Central Semi   631-435-1110
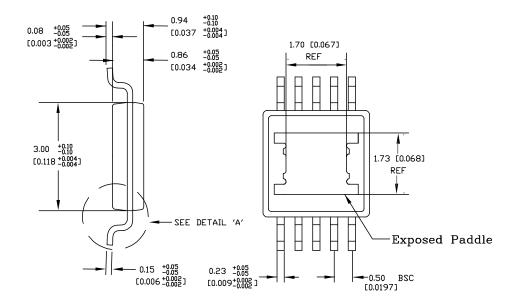
MIC2169B PCB Layout

MIC2169B Top Layer

MIC2169B Bottom Layer

Package Information

0.94 +0.10 −0.10 [0.037 +0.004 −0.004]

0.08 +0.05 −0.05 [0.003 +0.002 −0.002]

0.86 +0.05 −0.05 [0.034 +0.002 −0.002]

1.70 [0.067] REF

10

4.90 [0.193] BSC

3.00 +0.10 −0.10 [0.118 +0.004 −0.004]

1.73 [0.068] REF

SEE DETAIL 'A'

Exposed Paddle

1 2

3.00 +0.10 −0.10 [0.118 +0.004 −0.004]

0.15 +0.05 −0.05 [0.006 +0.002 −0.002]
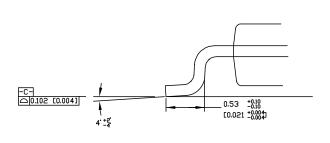
0.23 +0.05 −0.05 [0.009 +0.002 −0.002]

0.50 BSC [0.0197]

TOP VIEW

END VIEW

BOTTOM VIEW

NOTES:
1. DIMENSIONS ARE IN MM [INCHES].
2. CONTROLLING DIMENSION: MM
3. DIMENSION DOES NOT INCLUDE MOLD FLASH OR PROTRUSIONS, EITHER OF WHICH SHALL NOT EXCEED 0.20 [0.008] PER SIDE.

−C−
0.102 [0.004]

4° +2° −4°

0.53 +0.10 −0.10 [0.021 +0.004 −0.004]

DETAIL A

10-Pin ePad MSOP (MME)

MICREL, INC. 2180 FORTUNE DRIVE SAN JOSE, CA 95131 USA
TEL +1 (408) 944-0800 FAX +1 (408) 474-1000 WEB http://www.micrel.com

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