300kHz, Synchronous Buck Controller Featuring Adaptive On-Time Control

General Description

The Micrel MIC2174 is a fixed frequency, synchronous buck controller featuring adaptive on-time control. The MIC2174 operates over an input supply range of 3V to 40V, switches at a constant frequency of 300kHz and is capable of driving 25A of output current. The output voltage is adjustable down to 0.8V.

A unique Hyper Speed Control™ architecture allows for ultra fast transient response while reducing the output capacitance and also makes High V_{IN}/Low V_{OUT} operation possible. The MIC2174 utilizes a adaptive T_{ON} ripple controlled architecture. A UVLO is provided to ensure proper operation under power-sag conditions to prevent the external power MOSFET from overheating. A soft start is provided to reduce inrush current. Foldback current limit and "hiccup" mode short-circuit protection ensure FET and load protection.

The MIC2174 is available in a 10-pin MSOP (MAX1954A-compatible) package with a junction operating range from –40°C to +125°C.

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

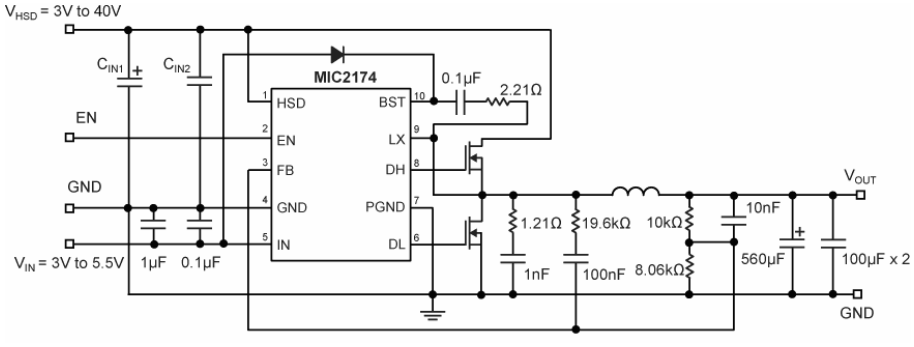
Features

- Hyper Speed Control™ architecture enables
  - High delta V operation (V_{IN}=40V and V_{OUT}=0.8V)
  - Smaller output capacitors than competitors
- 3V to 40V input voltage
- Stable with zero-ESR output capacitor
- 300kHz switching frequency
- Output down to 0.8V with ±1% FB accuracy
- Up to 94% efficiency
- Foldback current limit and "hiccup" mode short-circuit protection
- 6ms Internal soft start
- Thermal shutdown
- Pre-bias output safe
- –40°C to +125°C junction temperature range
- Available in 10-pin MSOP package

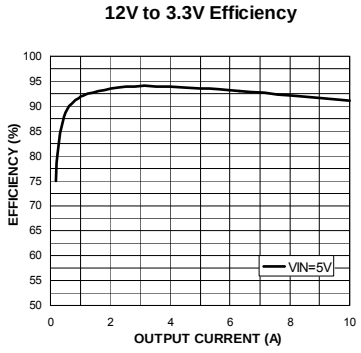
Applications

- Wide input power supply
- Industrial Equipments
- Distributed DC power systems
- Automotive applications
- PCs and servers

Typical Application



MIC2174: Synchronous Buck Controller Featuring Adaptive On-Time Control



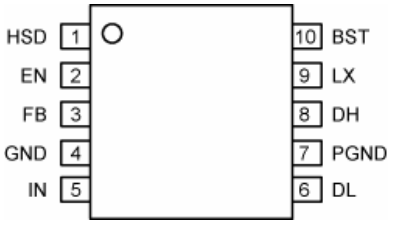
Hyper Speed Control is a trademark of Micrel. Inc.

MLF and *Micro*LeadFrame are registered trademarks of Amkor Technology, Inc.

Ordering Information

Part Number	Voltage	Switching Frequency	Junction Temp. Range	Package	Lead Finish
MIC2174-1YMM	Adj.	300kHz	–40° to +125°C	10-Pin MSOP	Pb-Free

Pin Configuration



10-Pin MSOP (MM)

Pin Description

Pin Number	Pin Name	Pin Function
1	HSD	High-Side N-MOSFET Drain Connection (input): Power to the drain of the external high-side N-channel MOSFET. The HSD operating voltage range is from 3V to 40V. Input capacitors between HSD and the power ground (PGND) are required.
2	EN	Enable (input): A logic level control of the output. The EN pin is CMOS-compatible. Logic high or floating = enable, logic low = shutdown. In the off state, supply current of the device is greatly reduced (typically 0.8mA).
3	FB	Feedback (input): Input to the transconductance amplifier of the control loop. The FB pin is regulated to 0.8V. A resistor divider connecting the feedback to the output is used to adjust the desired output voltage.
4	GND	Signal ground. GND is the ground path for the device input voltage V_{IN} and the control circuitry. The loop for the signal ground should be separate from the power ground (PGND) loop.
5	IN	Input Voltage (input): Power to the internal reference and control sections of the MIC2174. The IN operating voltage range is from 3V to 5.5V. A 1µF and 0.1µF ceramic capacitors from IN to GND are recommended for clean operation.
6	DL	Low-Side Drive (output): High-current driver output for external low-side MOSFET. The DL driving voltage swings from ground to IN.
7	PGND	Power Ground. PGND is the ground path for the MIC2174 buck converter power stage. The PGND pin connects to the sources of low-side N-Channel MOSFETs, the negative terminals of input capacitors, and the negative terminals of output capacitors. The loop for the power ground should be as small as possible and separate from the Signal ground (GND) loop.
8	DH	High-Side Drive (output): High-current driver output for external high-side MOSFET. The DH driving voltage is floating on the switch node voltage (LX). It swings from ground to V_{IN} minus the diode drop. Adding a small resistor between DH pin and the gate of the high-side N-channel MOSFETs can slow down the turn-on and turn-off time of the MOSFETs.
9	LX	Switch Node and Current Sense input: High current output driver return. The LX pin connects directly to the switch node. Due to the high speed switching on this pin, the LX pin should be routed away from sensitive nodes. LX pin also senses the current by monitoring the voltage across the low-side MOSFET during OFF time. In order to sense the current accurately, connect the low-side MOSFET drain to LX using a Kelvin connection.
10	BST	Boost (output): Bootstrapped voltage to the high-side N-channel MOSFET driver. A Schottky diode is connected between the IN pin and the BST pin. A boost capacitor of 0.1µF is connected between the BST pin and the LX pin. Adding a small resistor in series with the boost capacitor can slow down the turn-on time of high-side N-Channel MOSFETs.

Absolute Maximum Ratings[1]

IN, FB, EN to GND ..... -0.3V to +6V
BST to LX ..... -0.3V to +6V
BST to GND ..... -0.3V to +46V
DH to LX ..... -0.3V to (V_{BST} + 0.3V)
DL, COMP to GND ..... -0.3V to (V_{IN} + 0.3V)
HSD to GND ..... -0.3V to 42V
PGND to GND ..... -0.3V to +0.3V
Junction Temperature ..... +150°C
Storage Temperature (T_S) ..... -65°C to +150°C
Lead Temperature (soldering, 10sec) ..... 260°C

Operating Ratings[2]

Input Voltage (V_{IN}) ..... 3.0V to 5.5V
Supply Voltage (V_{HSD}) ..... 3.0V to 40V
Operating Temperature Range ..... -40°C to +125°C
Junction Temperature (T_J) ..... -40°C to +125°C
Junction Thermal Resistance
MSOP (θ_{JA}) ..... 130.5°C/W
Continuous Power Dissipation (T_A = 70°C) 421mW
(derate 5.6mW/°C above 70°C)

Electrical Characteristics[4]

V_{BST} - V_{LX} = 5V; T_A = 25°C, unless noted. **Bold** values indicate -40°C ≤ T_J ≤ +125°C.

Parameter	Condition	Min	Typ	Max	Units
General					
Operating Input Voltage (V_{IN}) [5]		**3.0**		**5.5**	V
HSD Voltage Range (V_{HSD})		**3.0**		**40**	V
Quiescent Supply Current	(V_{FB} = 1.5V, output switching but excluding external MOSFET gate current)		1.4	**3.0**	mA
Standby Supply Current	V_{IN} = V_{BST} = 5.5V, V_{HSD} = 40V, LX = unconnected, EN = GND [6]		0.8	**2**	mA
Under-voltage Lockout Trip Level		**2.4**	2.7	**3**	V
UVLO Hysteresis			50		mV
DC-DC Controller					
Output-Voltage Adjust Range (V_{OUT}) [7]		**0.8**			V
Error Amplifier					
FB Regulation Voltage	0°C ≤ T_J ≤ 85°C	**-1**		**1**	%
FB Regulation Voltage	-40°C ≤ T_J ≤ 125°C	**-2**		**2**	%
FB Input Leakage Current			5	**500**	nA
Current-Limit Threshold	V_{FB} = 0.8V	**103**	130	**162**	mV
	V_{FB} = 0V	**19**	48	**77**	mV
Soft-Start					
Soft-start Period			6		ms

Notes:

1. Exceeding the absolute maximum rating may damage the device.
2. The device is not guaranteed to function outside its operating rating.
3. Devices are ESD sensitive. Handling precautions recommended. Human body model, 1.5k in series with 100pF.
4. Specification for packaged product only.
5. The application is fully functional at low IN (supply of the control section) if the external MOSFETs have enough low voltage V_{TH}.
6. The current will come only from the internal 100kΩ pull-up resistor sitting on the EN Input and tied to IN.
7. The maximum V_{OUT} value is limited by the Fixed T_{ON} estimator which obtains V_{OUT} as a divided by 6 value (1/6).

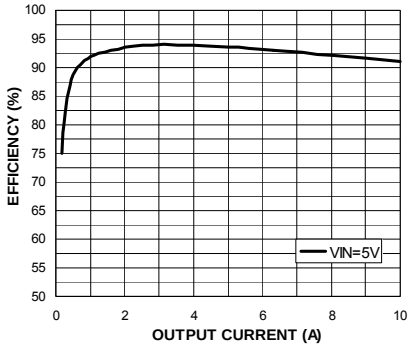
Parameter	Condition	Min	Typ	Max	Units
Oscillator					
Switching Frequency	Measured in Test Mode	**0.225**	0.3	**0.375**	MHz
Maximum Duty Cycle	Measured at DH [8]		87		%
Minimum Duty Cycle	Measured at DH		0		%
FET Drives					
DH, DL Output Low Voltage	I_{SINK} = 10mA			**0.1**	V
DH, DL Output High Voltage	I_{SOURCE} = 10mA	V_{IN}-0.1V or V_{BST}-0.1V			V
DH On-Resistance, High State			2.1	**3.3**	Ω
DH On-Resistance, Low State			1.8	**3.3**	Ω
DL On-Resistance, High State			1.8	**3.3**	Ω
DL On-Resistance, Low State			1.2	**2.3**	Ω
LX Leakage Current	V_{LX} = 40V, V_{IN} = 5.5V,V_{BST} = 45.5V			**55**	µA
HSD Leakage Current	V_{LX} = 40V, V_{IN} = 5.5V,V_{BST} = 45.5V			**21**	µA
Thermal Protection					
Over-temperature Shutdown			155		°C
Over-temperature Shutdown Hysteresis			10		°C
Shutdown Control					
En Logic Level Low	3V < V_{IN} <5.5V	**0.4**	0.8		V
En Logic Level High	3V < V_{IN} <5.5V		0.9	**1.2**	V
En Pull-up Current			50		µA

Note:

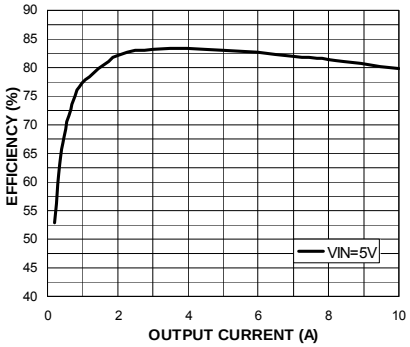
8. The maximum duty cycle is limited by the fixed mandatory off time T_{OFF} of typical 363ns.

Typical Characteristics

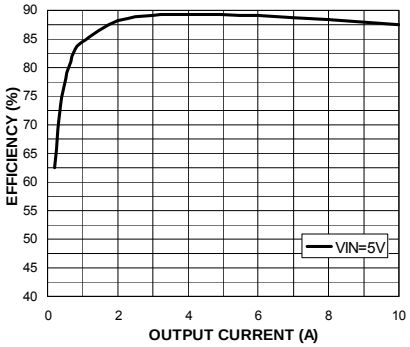
12V to 3.3V Efficiency



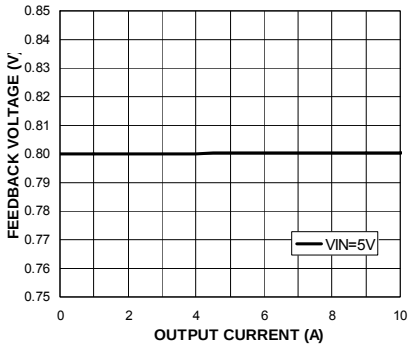
24V to 1.8V Efficiency



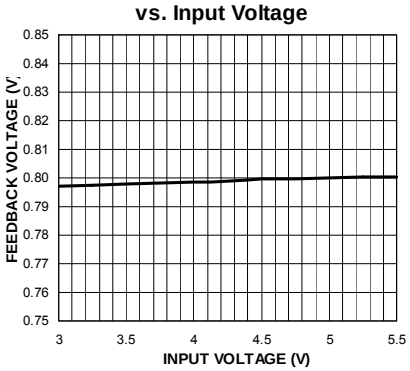
24V to 3.3V Efficiency



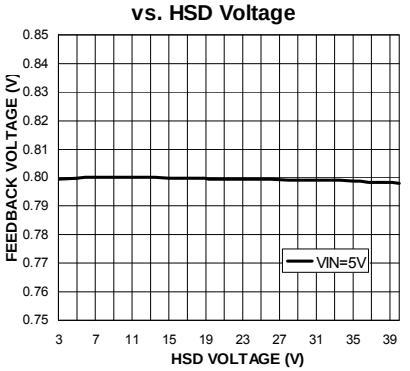
Feedback Voltage vs. Load



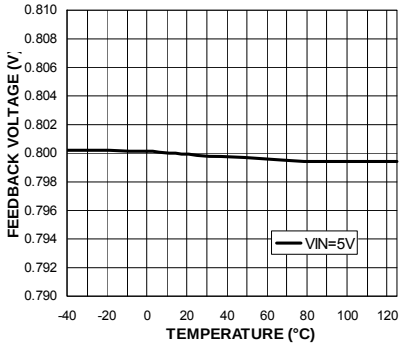
Feedback Voltage vs. Input Voltage



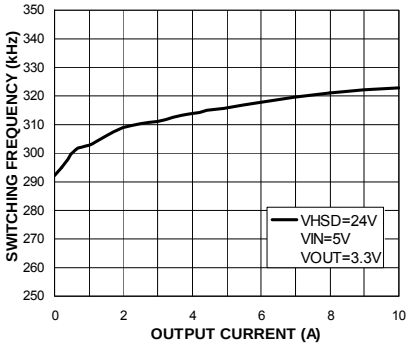
Feedback Voltage vs. HSD Voltage



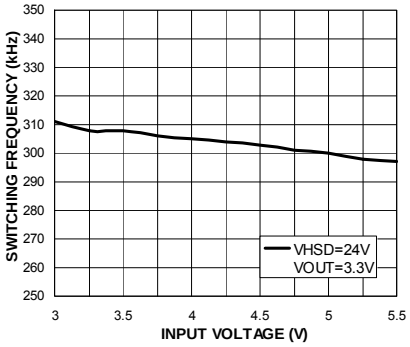
Feedback Voltage vs. Temperature



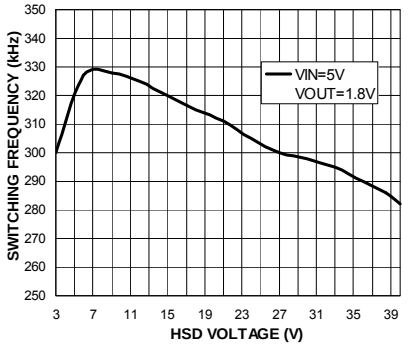
Switching Frequency vs. Load



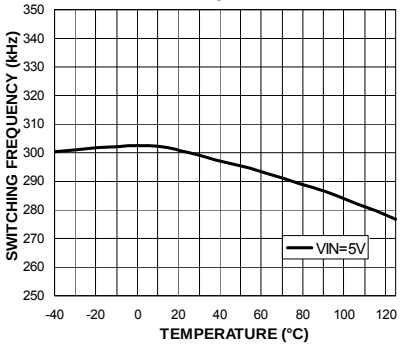
Switching Frequency vs. Input Voltage



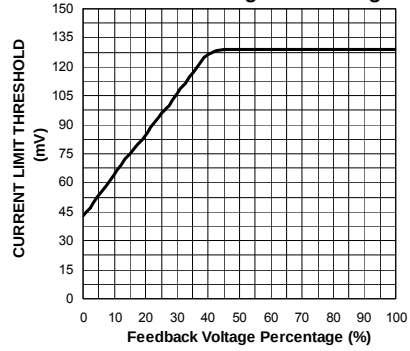
Switching Frequency vs. HSD Voltage



Switching Frequency vs. Temperature

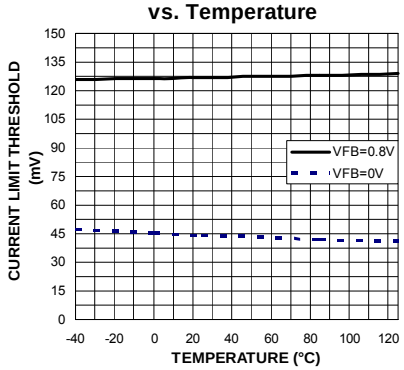


Current Limit Threshold vs. Feedback Voltage Percentage

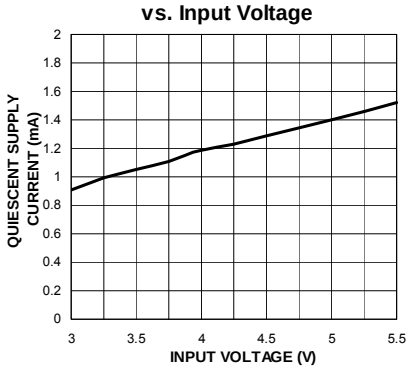


Typical Characteristics (continued)

Current Limit Threshold vs. Temperature

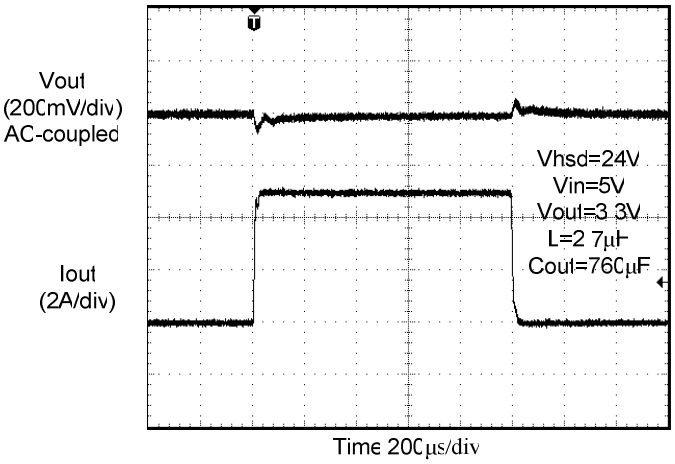


Quiescent Supply Current vs. Input Voltage

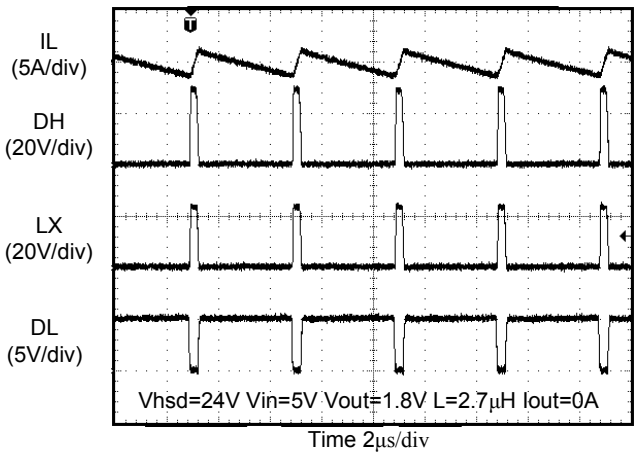


Functional Characteristics

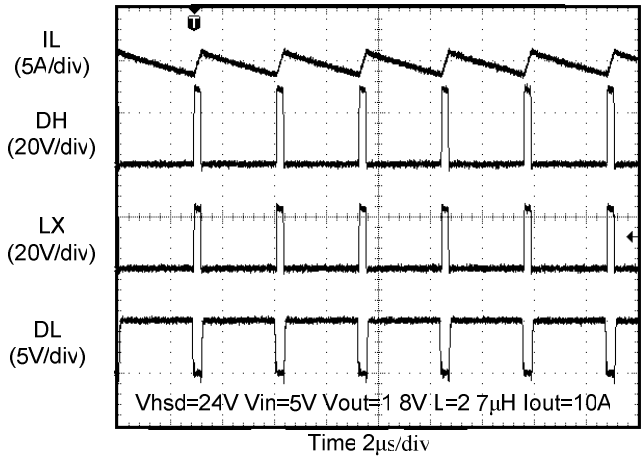
Load Transient



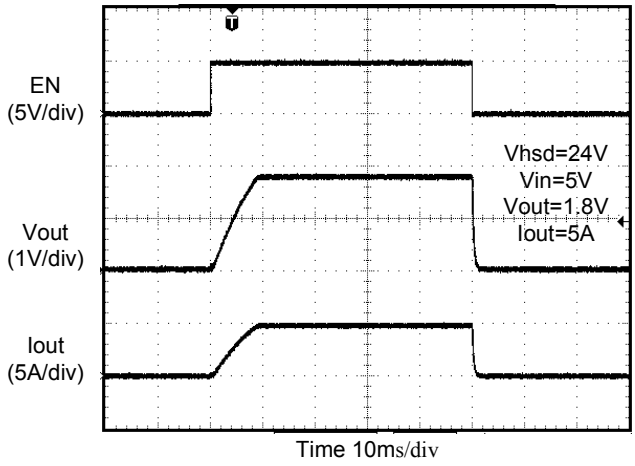
Switching Waveform (No Load)



Switching Waveform (Heavy Load)

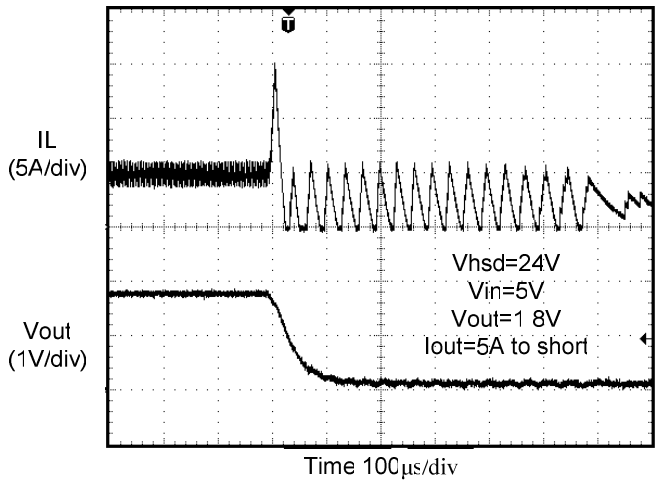


Power-Up/Power-Down

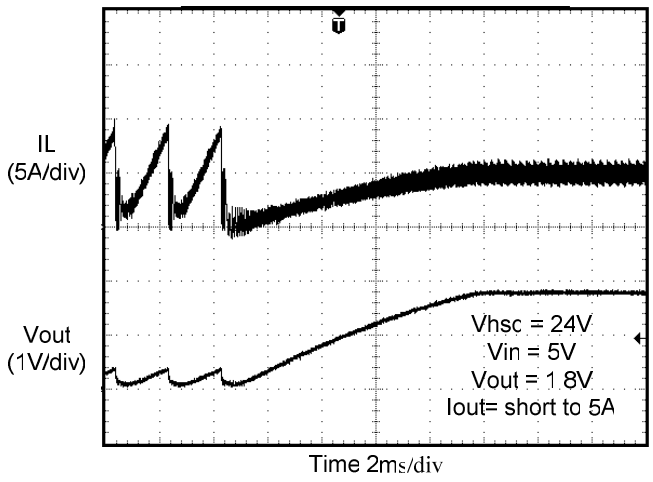


Functional Characteristics (continued)

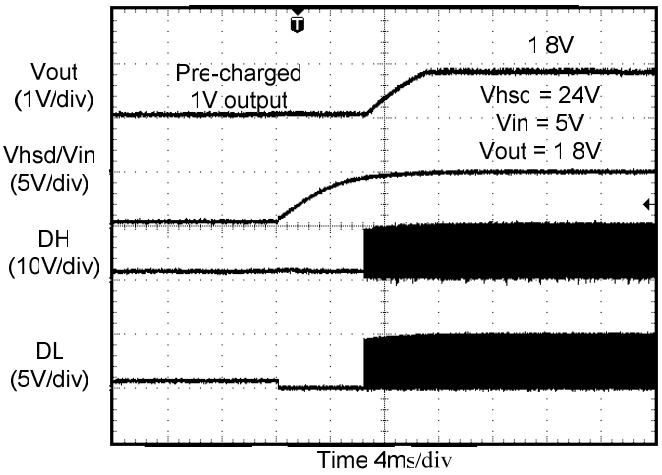
Short Circuit



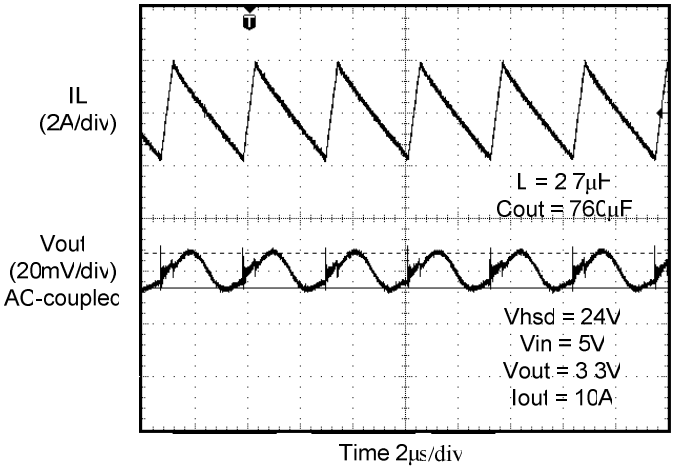
Short Circuit Recovery



Start-up with Prebias Vout



Output Voltage Ripple



Functional Diagram

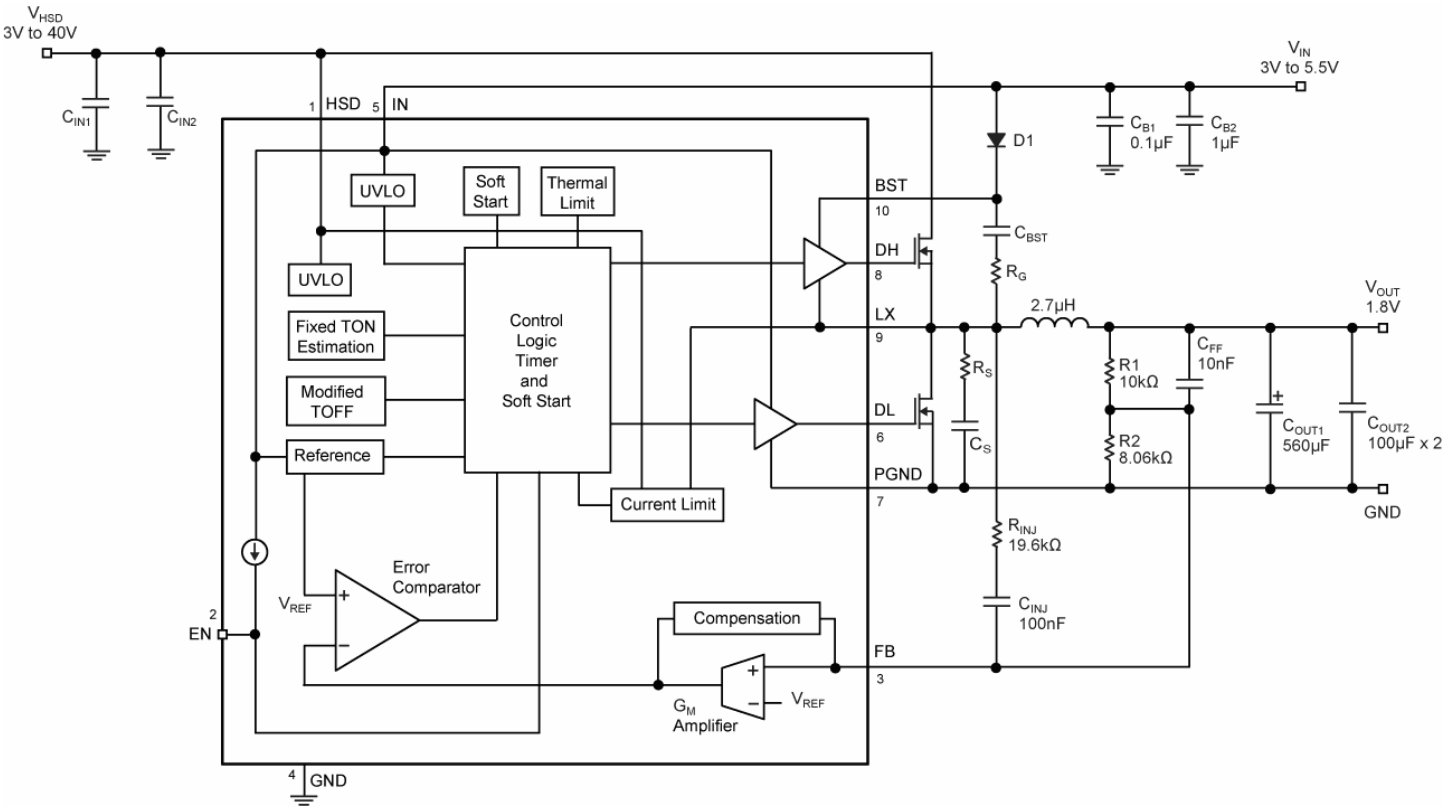


Figure 1. MIC2174 Block Diagram

Functional Description

The MIC2174 is an adaptive on-time synchronous buck controller built for low cost and high performance. It is designed for wide input voltage range from 3V to 40V and for high output power buck converters. An estimated-ON-time method is applied in MIC2174 to obtain a constant switching frequency and to simplify the control compensation. The over-current protection 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.

Theory of Operation

The MIC2174 is an adaptive on-time buck controller. Figure 1 illustrates the block diagram for the control loop. The output voltage variation will be sensed by the MIC2174 feedback pin FB via the voltage divider R1 and R2, and compared to a 0.8V reference voltage V_{REF} at the error comparator through a low gain transconductance (gm) amplifier, which improves the MIC2174 converter output voltage regulation. If the FB voltage decreases and the output of the gm amplifier is below 0.8V, the error comparator will trigger the control logic and generate an ON-time period, in which DH pin is logic high and DL pin is logic low. The ON-time period length is predetermined by the "FIXED T_{ON} ESTIMATION" circuitry:

$$T_{ON(estimated)} = \frac{V_{OUT}}{V_{HSD} \times 300kHz} \quad (1)$$

where V_{OUT} is the output voltage, V_{HSD} is the power stage input voltage.

After ON-time period, the MIC2174 goes into the OFF-time period. In which DH pin is logic low and DL pin is logic high. The OFF-time period length depends upon the FB voltage in most cases. When the FB voltage decreases and the output of the gm amplifier is below 0.8V, the ON-time period is trigger and the OFF-time period ends. If the OFF-time period decided by the FB voltage is less than the minimum OFF time $T_{OFF(min)}$, which is about 363ns typical, the MIC2174 control logic will apply the $T_{OFF(min)}$ instead. $T_{OFF(min)}$ is required by the BST charging. The maximum duty cycle is obtained from the 363ns $T_{OFF(min)}$:

$$Dmax = \frac{T_S - T_{OFF(min)}}{T_S} = 1 - \frac{363ns}{T_S}$$

where Ts = 1/300kHz = 3.33µs. It is not recommended to use MIC2174 with a OFF time close to $T_{OFF(min)}$ at the steady state.

The power stage input voltage V_{HSD} is fed into the Fixed Ton Estimation block through a 6:1 divider and 5V voltage clamper. Therefore, if the V_{HSD} is higher than 30V, the Fixed T_{ON} Estimation block uses 30V to estimate Ton instead of the real V_{HSD}. As a result, the switching frequency will be less than 300kHz:

$$f_{SW(VHDS>30V)} = \frac{30V}{V_{HSD}} \times 300kHz \quad (2)$$

The estimated-ON-time method results in a constant 300kHz switching frequency up to 30V V_{HSD}. The actual ON time is varied with the different rising and falling time of the external MOSFETs. Therefore, the type of the external MOSFETs, the output load current, and the control circuitry power supply V_{IN} will modify the actual ON time and the switching frequency. Also, the minimum Ton results in a lower switching frequency in the high V_{HSD} and low V_{OUT} applications, such as 36V to 1.0V application. The minimum Ton measured on the MIC2174 evaluation board with Si7148DP MOSFETs is about 184ns. During the load transient, the switching frequency is changed due to the varying OFF time.

To illustrate the control loop, the steady-state scenario and the load transient scenario are analyzed. For easy analysis, the gain of the gm amplifier is assumed to be 1. With this assumption, the inverting input of the error comparator is the same as the FB voltage. Figure 2 shows the MIC2174 control loop timing during the steady-state. During the steady-state, the gm amplifier senses the FB voltage ripple, which is proportional to the output voltage ripple and the inductor current ripple, to trigger the ON-time period. The ON time is predetermined by the estimation. The ending of OFF time is controlled by the FB voltage. At the valley of the FB voltage ripple, which is below than V_{REF}, OFF period ends and the next ON-time period is triggered through the control logic circuitry.

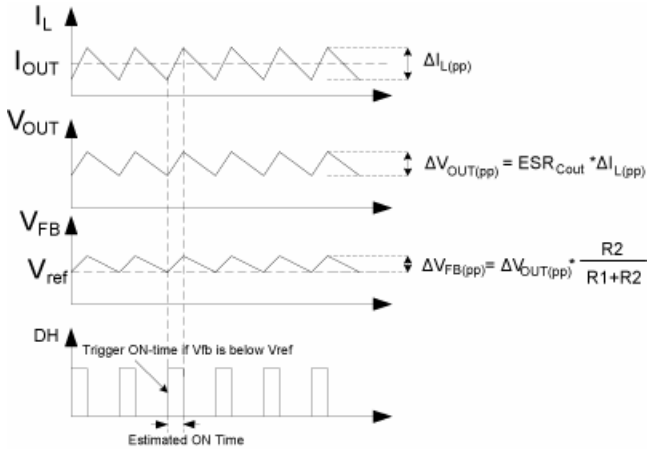


Figure 2. MIC2174 Control Loop Timing

Figure 3 shows the load transient scenario of the MIC2174 converter. The output voltage drops due to the sudden load increasing, which would cause the FB voltage to be less than V_{REF}. This will cause the error

comparator to trigger ON-time period. At the end of the ON-time period, a minimum OFF time $T_{OFF(min)}$ is generated to charge BST since the FB voltage is still below the V_{REF}. Then, the next ON-time period is triggered due to the low FB voltage. Therefore, the switching frequency changes during the load transient. With the varying duty cycle and switching frequency, the output recovery time is fast and the output voltage deviation is small in MIC2174 converter.

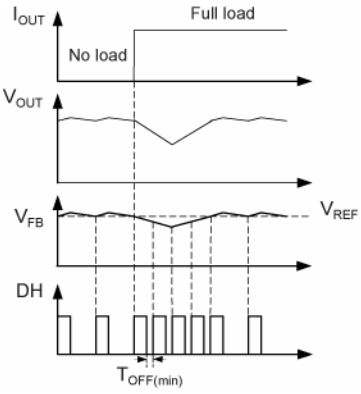


Figure 3. MIC2174 Load-Transient Response

Unlike the current-mode control, MIC2174 uses the output voltage ripple, which is proportional to the inductor current ripple if the ESR of the output capacitor is large enough, to trigger an ON-time period. The predetermined ON time makes MIC2174 control loop has the advantage as the adaptive on-time mode control. Therefore, the slope compensation, which is necessary for the current-mode control, is not required in the MIC2174.

The MIC2174 has its own stability concern: the FB voltage ripple should be in phase with the inductor current ripple and large enough to be sensed by the gm amplifier and the error comparator. The recommended minimum FB voltage ripple is 20mV. If a low ESR output capacitor is selected, the FB voltage ripple may be too small to be sensed by the gm amplifier and the error comparator. Also, the output voltage ripple and the FB voltage ripple are not in phase with the inductor current ripple if the ESR of the output capacitor is very low. Therefore, the ripple injection is required for a low ESR output capacitor. Please refer to "Ripple Injection" subsection in "Application Information" for more details about the ripple injection.

Soft-Start

Soft-start reduces the power supply input surge current at startup by controlling the output voltage rise time. The input surge appears while the output capacitor is charged up. A slower output rise time will draw a lower input surge current.

MIC2174 implements an internal digital soft-start by making the 0.8V reference voltage V_{REF} ramp from 0 to 100% in about 6ms with a 9.7mV step. Therefore, the output voltage is controlled to increase slowly by a stair-case V_{REF} ramp. Once the soft-start ends, the related circuitry is disabled to reduce the current consumption. V_{IN} should be powered up no earlier than V_{HSD} to make the soft-start function behavior correctly.

Current Limit

The MIC2174 uses the $R_{DS(ON)}$ of the low-side power MOSFET to sense over-current conditions. The lower-side MOSFET is used because it displays much lower parasitic oscillations during switching then the high-side MOSFET. Using the low-side MOSFET $R_{DS(ON)}$ as a current sense is an excellent method for circuit protection. This method will avoid adding cost, board space and power losses taken by discrete current sense resistors.

In each switching cycle of the MIC2174 converter, the inductor current is sensed by monitoring the low-side MOSFET in the OFF period. The sensed voltage is compared with a current-limit threshold voltage V_{CL} after a blanking time of 150ns. If the sensed voltage is over V_{CL}, which is 130mV typical at 0.8V feedback voltage, the MIC2174 turns off the high-side MOSFET and a soft-start sequence is trigged. 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 current limit threshold V_{CL} has a fold back characteristics related to the FB voltage. Please refer to the "Typical Characteristics" for the curve of V_{CL} vs. FB voltage. The circuit in Figure 4 illustrates the MIC2174 current limiting circuit.

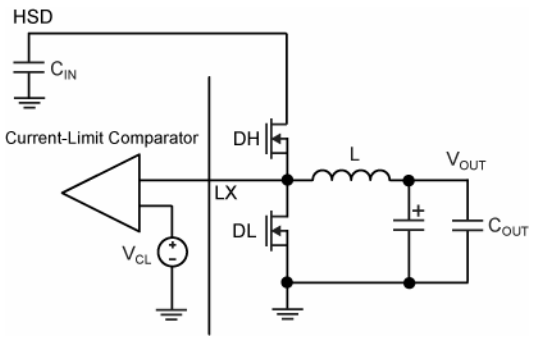


Figure 4. MIC2174 Current Limiting Circuit

Using the typical V_{CL} value of 130mV, the current limit value is roughly estimated as:

$$I_{CL} \approx \frac{130mV}{R_{DS(ON)}}$$

For designs where the current ripple is significant compared to the load current I_{OUT}, or for low duty cycle operation, calculating the current limit I_{CL} should take into account that one is sensing the peak inductor

current and that there is a blanking delay of approximately 150ns.

$$I_{CL} = \frac{130mV}{R_{DS(ON)}} + \frac{V_{OUT} \times T_{DLY}}{L} - \frac{\Delta I_{L(pp)}}{2} \quad (3)$$

$$\Delta I_{L(pp)} = \frac{V_{OUT} \times (1-D)}{f_{SW} \times L} \quad (4)$$

where:

V_{OUT} = The output voltage

T_{DLY} = Current limit blanking time, 150ns typical

$\Delta I_{L(pp)}$ = Inductor current ripple peak-to-peak value

D = Duty Cycle

f_{SW} = Switching frequency

The MOSFET $R_{DS(ON)}$ varies 30% to 40% with temperature; therefore, it is recommended to add a 50% margin to I_{CL} in the above equation to avoid false current limiting due to increased MOSFET junction temperature rise. It is also recommended to connect LX pin directly to the drain of the low-side MOSFET to accurately sense the MOSFETs $R_{DS(ON)}$.

MOSFET Gate Drive

The MIC2174 high-side drive circuit is designed to switch an N-Channel MOSFET. The Block Diagram of Figure 1 shows a bootstrap circuit, consisting of D1 (a Schottky diode is recommended) and C_{BST}. This circuit 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 LX 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 high-side MOSFET turns on, the voltage on the LX pin increases to approximately V_{HSD}. Diode D1 is reversed biased and C_{BST} floats high while continuing to keep the high-side MOSFET on. The bias current of the high-side driver is less than 10mA so a 0.1µF to 1µF is sufficient to hold the gate voltage with minimal droop for the power stroke (high-side switching) cycle, i.e. ΔBST = 10mA x 3.33µs/0.1µF = 333mV. When the low-side MOSFET is turned back on, C_{BST} is recharged through D1. A small resistor R_G, which is in series with C_{BST}, can slow down the turn-on time of the high-side N-channel MOSFET.

The drive voltage is derived from the supply voltage VIN. The nominal low-side gate drive voltage is V_{IN} and the nominal high-side gate drive voltage is approximately $V_{IN} – V_{DIODE}$, where V_{DIODE} is the voltage drop across D1. An approximate 30ns delay between the high-side and low-side driver transitions is used to prevent current from simultaneously flowing unimpeded through both MOSFETs.

Application Information

MOSFET Selection

The MIC2174 controller works from power stage input voltages of 3V to 40V and has an external 3V to 5.5V V_{IN} to provide power to turn the external N-Channel power MOSFETs for the high- and low-side switches. For applications where V_{IN} < 5V, 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.

There are different criteria for choosing the high-side and low-side MOSFETs. These differences are more significant at lower duty cycles such as a 12V to 1.8V conversion. In such an application, the high-side MOSFET is required to switch as quickly as possible to minimize transition losses, whereas the low-side MOSFET can switch slower, but must handle larger RMS currents. When the duty cycle approaches 50%, the current carrying capability of the high-side MOSFET starts to become critical.

It is important to note that 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 limit. 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 MIC2174 gate-drive circuit. At 300kHz switching frequency and above, the gate charge can be a significant source of power dissipation in the MIC2174. 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:

$$I_{G[high\text{-}side]}(avg) = Q_G \times f_{SW} \quad (5)$$

where:

$I_{G[high\text{-}side]}(avg)$ = Average high-side MOSFET gate current

Q_G = Total gate charge for the high-side MOSFET taken from the manufacturer's data sheet for $V_{GS} = V_{IN}$.

f_{SW} = Switching Frequency (300kHz)

The low-side MOSFET is turned on and off at V_{DS} = 0 because an internal body diode or external 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 C_{ISS} at V_{DS} = 0 instead of gate charge.

For the low-side MOSFET:

$$I_{G[low\text{-}side]}(avg) = C_{ISS} \times V_{GS} \times f_{SW} \quad (6)$$

Since the current from the gate drive comes from the V_{IN}, the power dissipated in the MIC2174 due to gate drive is:

$$P_{GATEDRIVE} = V_{IN} \times (I_{G[high\text{-}side]}(avg) + I_{G[low\text{-}side]}(avg)) \quad (7)$$

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 MIC2174. Also, the $R_{DS(ON)}$ of the low-side MOSFET will determine the current limit value. Please refer to "Current Limit" subsection in "Functional Description" for more details.

Parameters that are important to MOSFET switch selection are:

- Voltage rating
- On-resistance
- Total gate charge

The voltage ratings for the high-side and low-side MOSFETs are essentially equal to the power stage input voltage V_{HSD}. A safety factor of 20% should be added to the V_{DS}(max) of the MOSFETs to account for voltage spikes due to circuit parasitic elements.

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

$$P_{SW} = P_{CONDUCTION} + P_{AC} \quad (8)$$

$$P_{CONDUCTION} = I_{SW(RMS)}^2 \times R_{DS(ON)} \quad (9)$$

$$P_{AC} = P_{AC(off)} + P_{AC(on)} \quad (10)$$

where:

$R_{DS(ON)}$ = on-resistance of the MOSFET switch

D = Duty Cycle = V_{OUT} / V_{HSD}

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

$$t_T = \frac{C_{ISS} \times V_{IN} + C_{OSS} \times V_{HSD}}{I_G} \quad (11)$$

where:

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

I_G = gate-drive current

The total high-side MOSFET switching loss is:

$$P_{AC} = (V_{HSD} + V_D) \times I_{PK} \times t_T \times f_{SW} \quad (12)$$

where:

t_T = Switching transition time

V_D = Body diode drop (0.5v)

f_{SW} = Switching Frequency (300kHz)

The high-side MOSFET switching losses increase with the input voltage V_{HSD} due to the longer turn-on time and turn-off time. 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.

$$L = \frac{V_{OUT} \times (V_{HSD(max)} - V_{OUT})}{V_{HSD(max)} \times f_{sw} \times 20\% \times I_{OUT(max)}} \quad (13)$$

where:

f_{SW} = switching frequency, 300 kHz

20% = ratio of AC ripple current to DC output current

$V_{HSD(max)}$ = maximum power stage input voltage

The peak-to-peak inductor current ripple is:

$$\Delta I_{L(pp)} = \frac{V_{OUT} \times (V_{HSD(max)} - V_{OUT})}{V_{HSD(max)} \times f_{sw} \times L} \quad (14)$$

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

$$I_{L(pk)} = I_{OUT(max)} + 0.5 \times \Delta I_{L(pp)} \quad (15)$$

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

$$I_{L(RMS)} = \sqrt{I_{OUT(max)}^2 + \frac{\Delta I_{L(PP)}^2}{12}} \quad (16)$$

Maximizing efficiency requires the proper selection of core material and minimizing the winding resistance. The high frequency operation of the MIC2174 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:

$$P_{INDUCTOR(Cu)} = I_{L(RMS)}^2 \times R_{WINDING} \quad (17)$$

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

$$P_{WINDING(Ht)} = R_{WINDING(20°C)} \times (1 + 0.0042 \times (T_H - T_{20°C})) \quad (18)$$

where:

T_H = temperature of wire under full load

$T_{20°C}$ = ambient temperature

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

Output Capacitor Selection

The type of the output capacitor is usually determined by its ESR (equivalent series resistance). Voltage and RMS current capability are two other important factors for selecting the output capacitor. Recommended capacitors are tantalum, low-ESR aluminum electrolytic, OS-CON and POSCAPS. The output capacitor's ESR is usually the main cause of the output ripple. The output capacitor ESR also affects the control loop from a stability point of view. The maximum value of ESR is calculated:

$$ESR_{C_{OUT}} \leq \frac{\Delta V_{OUT(pp)}}{\Delta I_{L(PP)}} \quad (19)$$

where:

$\Delta V_{OUT(pp)}$ = peak-to-peak output voltage ripple

$\Delta I_{L(PP)}$ = peak-to-peak inductor current ripple

The total output ripple is a combination of the ESR and output capacitance. The total ripple is calculated below:

$$\Delta V_{OUT(pp)} = \sqrt{\left(\frac{\Delta I_{L(PP)}}{C_{OUT} \times f_{SW} \times 8}\right)^2 + \left(\Delta I_{L(PP)} \times ESR_{C_{OUT}}\right)^2} \quad (20)$$

where:

D = duty cycle

C_{OUT} = output capacitance value

f_{SW} = switching frequency

As described in the "Theory of Operation" subsection in "Functional Description", the MIC2174 requires at least 20mV peak-to-peak ripple at the FB pin to make the gm amplifier and the error comparator to behavior properly. Also, the output voltage ripple should be in phase with the inductor current. Therefore the output voltage ripple caused by the output capacitor C_{OUT} should be much smaller than the ripple caused by the output capacitor ESR. If low ESR capacitors are selected as the output capacitors, such as ceramic capacitors, a ripple injection method is applied to provide the enough FB voltage ripples. Please refer to the "Ripple Injection" subsection for more details.

The voltage rating of the capacitor should be twice the output voltage for a tantalum and 20% greater for aluminum electrolytic or OS-CON. The output capacitor RMS current is calculated below:

$$I_{C_{OUT}(RMS)} = \frac{\Delta I_{L(PP)}}{\sqrt{12}} \quad (21)$$

The power dissipated in the output capacitor is:

$$P_{DISS(C_{OUT})} = I_{C_{OUT}(RMS)}^{2} \times ESR_{C_{OUT}} \quad (22)$$

Input Capacitor Selection

The input capacitor for the power stage input V_{HSD} 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 two 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 de-rating. 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:

$$\Delta V_{IN} = I_{L(pk)} \times ESR_{CIN} \quad (23)$$

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 current ripple is low:

$$I_{CIN(RMS)} \approx I_{OUT(max)} \times \sqrt{D \times (1-D)} \quad (24)$$

The power dissipated in the input capacitor is:

$$P_{DISS(CIN)} = I_{CIN(RMS)}^{2} \times ESR_{CIN} \quad (25)$$

Voltage Setting Components

The MIC2174 requires two resistors to set the output voltage as shown in Figure 5.

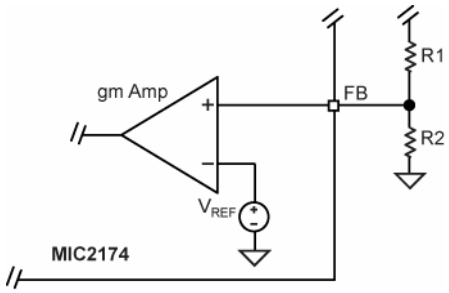


Figure 5. Voltage-Divider Configuration

The output voltage is determined by the equation:

$$V_{OUT} = V_{REF} \times (1 + \frac{R1}{R2}) \quad (26)$$

where, V_{REF} = 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:

$$R2 = \frac{V_{REF} \times R1}{V_{OUT} - V_{REF}} \quad (27)$$

External Schottky Diode (Optional)

An external freewheeling diode, which is not necessary, 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 30ns. 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.

$$I_{D(avg)} = I_{OUT} \times 2 \times 30ns \times f_{SW} \quad (28)$$

The reverse voltage requirement of the diode is:

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

The power dissipated by the Schottky diode is:

$$P_{DIODE} = I_{D(avg)} \times V_F \quad (29)$$

where, V_F = forward voltage at the peak diode current.

The external Schottky diode is not necessary for the circuit operation since the low-side MOSFET contains a parasitic body diode. The external diode will improve efficiency and decrease the 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.

Ripple Injection

The minimum FB voltage ripple requested by the MIC2174 gm amplifier and error comparator is 20mV. However, the output voltage ripple is generally designed as 1% to 2% of the output voltage. For a low output voltage, such as 1V output, the output voltage ripple is only 10mV to 20mV, and the FB voltage ripple is less than 20mV. If the FB voltage ripple is so small that the gm amplifier and error comparator could not sense it, the MIC2174 will lose control and the output voltage is not regulated. In order to have some amount of FB voltage ripple, the ripple injection method is applied for low output voltage ripple applications.

The applications are divided into three situations according to the amount of the FB voltage ripple:

1) Enough ripple at the FB voltage due to the large ESR of the output capacitors.

As shown in Figure 6a, the converter is stable without any adding in this situation. The FB voltage ripple is:

$$\Delta V_{FB(pp)} = \frac{R2}{R1+R2} \times ESR_{C_{OUT}} \times \Delta I_{L(pp)} \quad (30)$$

where $\Delta I_{L(pp)}$ is the peak-to-peak value of the inductor current ripple.

2) Inadequate ripple at the FB voltage due to the small ESR of the output capacitors.

The output voltage ripple is fed into the FB pin through a feedforward capacitor Cff in this situation, as shown in Figure 6b. The typical Cff value is between 1nF to 100nF. With the feedforward capacitor, the FB voltage ripple is very close to the output voltage ripple:

$$\Delta V_{FB(pp)} \approx ESR \times \Delta I_{L(pp)} \quad (31)$$

3) Invisible ripple at the FB voltage due to the very low ESR of the output capacitors.

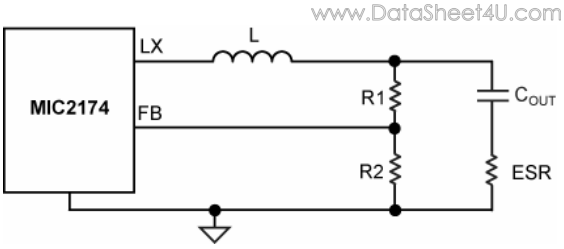


Figure 6a. Enough Ripple at FB

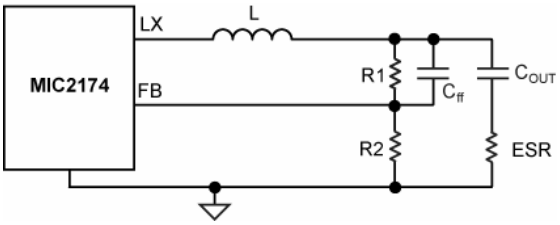


Figure 6b. Inadequate Ripple at FB

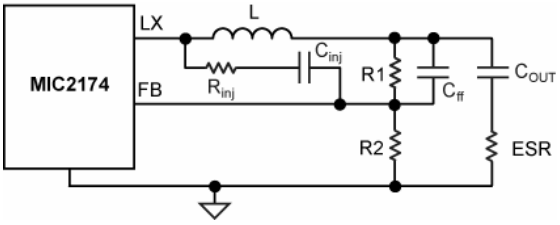


Figure 6c. Invisible Ripple at FB

In this situation, the output voltage ripple is less than 20mV. Therefore, additional ripple is injected into the FB pin from the switching node LX via a resistor Rinj and a capacitor Cinj, as shown in Figure 6c. The injected ripple is:

$$\Delta V_{FB(pp)} = V_{HSD} \times K_{div} \times D \times (1\text{-}D) \times \frac{1}{f_{SW} \times \tau} \quad (32)$$

$$K_{div} = \frac{R1//R2}{Rinj + R1//R2} \quad (33)$$

where

V_{HSD} = Power stage input voltage at HSD pin

D = Duty Cycle

f_{SW} = switching frequency

τ = (R1//R2//Rinj) × Cff

In the formula (32) and (33), it is assumed that the time constant associated with Cff must be much greater than the switching period:

$$\frac{1}{f_{SW} \times \tau} = \frac{T}{\tau} << 1$$

If the voltage divider resistors R1 and R2 are in the kΩ range, a Cff of 1nF to 100nF can easily satisfy the large time constant consumption. Also, a 100nF injection capacitor Cinj is used in order to be considered as short for a wide range of the frequencies.

The process of sizing the ripple injection resistor and capacitors is:

Step 1. Select Cff to feed all output ripples into the feedback pin and make sure the large time constant assumption is satisfied. Typical choice of Cff is 1nF to 100nF if R1 and R2 are in kΩ range.

Step 2. Select Rinj according to the expected feedback voltage ripple. According to the equation (33),

$$K_{div} = \frac{\Delta V_{FB(pp)}}{V_{HSD}} \times \frac{f_{SW} \times \tau}{D \times (1 - D)} \quad (34)$$

Then the value of Rinj is obtained as:

$$R_{inj} = (R1 // R2) \times \left(\frac{1}{K_{div}} - 1\right) \quad (35)$$

Step 3. Select Cinj as 100nF, which could be considered as short for a wide range of the frequencies.

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 MIC2174 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 HSD input capacitor next.
- Place the HSD input capacitors on the same side of the board and as close to the MOSFETs as possible.
- Keep both the HSD and PGND connections short.
- Place several vias to the ground plane close to the HSD 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.
- The 1µF and 0.1µF capacitors, which connect to the V_{IN} terminal, must be located right at the IC. The V_{IN} 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 (LX) short.
- Do not route any digital lines underneath or close to the inductor.
- Keep the switch node (LX) away from the feedback (FB) pin.
- The LX pin should be connected directly to the drain of the low-side MOSFET to accurate sense the voltage across the low-side MOSFET.
- 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.

Schottky Diode (Optional)

- Place the Schottky diode on the same side of the board as the MOSFETs and HSD 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 (LX) must be keep as short as possible.

RC Snubber

- Place the RC snubber on the same side of the board and as close to the MOSFETs as possible.

Evaluation Board Schematics

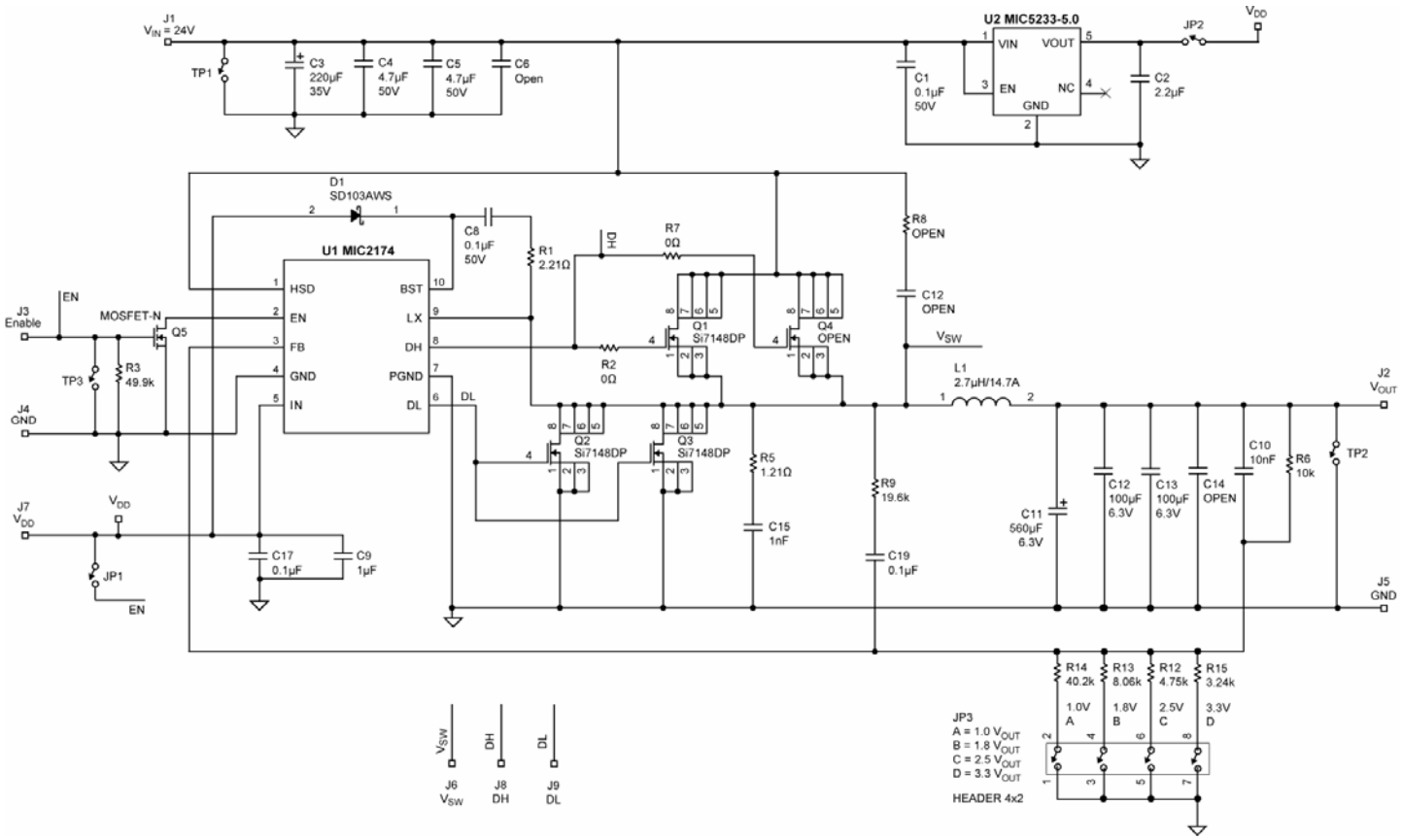


Figure 7. Schematic of MIC2174 Evaluation Board

Bill of Materials

Item	Part Number	Manufacturer	Description	Qty
C1, C8, C17, C19	06035D104MAT	AVX[1]	0.1µF Ceramic Capacitor, X5R, Size 0603, 50V	4
C2	0805ZD225MAT	AVX	2.2µF Ceramic Capacitor, X5R, Size 0805, 10V	1
	GRM216R61A225ME24D	MuRata[2]		
	C2012X5R1A225K/0.85	TDK[3]		
C3	222215095001	Vishay[4]	220µF Aluminum Capacitor, SMD, 35V	1
C4, C5	12105C475KAT2A	AVX	4.7µF Ceramic Capacitor, X7R, Size 1210, 50V	2
	GRM32ER71H475KA88L	MuRata		
C9	0805ZD105MAT	AVX	1µF Ceramic Capacitor, X5R, Size 0805, 10V	1
	GRM219R61A105MA01D	MuRata		
C10	06035C103KAT2A	AVX	10nF Ceramic Capacitor, X7R, Size 0603, 50V	1
	GRM188R71H103KA01D	MuRata		
C11	6SEPC560MX	Sanyo[5]	560µF OSCON Capacitor, 6.3V	1
C12, C13	12106D107MAT	AVX	100µF Ceramic Capacitor, X5R, Size 1210, 6.3V	2
	GRM32ER60J107ME20L	muRata		
	C3225X5R0J107M	TDK		
C15	06035D102MAT	AVX	1nF Ceramic Capacitor, X5R, 0603, 50V	1
D1	SD103AWS	Vishay	Small Signal Schottky Diode	1
L1	CDEP134-2R7MC-H	Sumida[6]	2.7µH Inductor, 14.7A Saturation Current	1
Q1, Q2, Q3	Si7148DP	Vishay	75V N-Channel TrenchFET Power MOSFET, 14.5mΩ Rds(on) @ 4.5V	3
Q5	CMPDM7002A	Central Semiconductor[7]	Signal MOSFET, 60V	1
	2N7002E-T1-E3	Vishay		
R1	CRCW06032R21FKEY3	Vishay-Dale[4]	2.21Ω Resistor, Size 0603, 1%	1
R2, R7	CRCW06030000FKEY3	Vishay-Dale	0Ω Resistor, Size 0603, 1%	2
R3	CRCW06034992FKEY3	Vishay-Dale	49.9kΩ Resistor, Size 0603, 1%	1
R5	CRCW06031R21FKEY3	Vishay-Dale	1.21Ω Resistor, Size 0603, 1%	1
R6	CRCW06031002FKEY3	Vishay-Dale	10kΩ Resistor, Size 0603, 1%	1
R9	CRCW06031962FKEY3	Vishay-Dale	19.6kΩ Resistor, Size 0603, 1%	1
R12	CRCW06034751FKEY3	Vishay-Dale	4.75kΩ Resistor, Size 0603, 1%	1
R13	CRCW06038061FKEY3	Vishay-Dale	8.06kΩ Resistor, Size 0603, 1%	1
R14	CRCW06034022FKEY3	Vishay-Dale	40.2kΩ Resistor, Size 0603, 1%	1
R15	CRCW06033241FKEY3	Vishay-Dale	3.24kΩ Resistor, Size 0603, 1%	1
U1	**MIC2174-1YMM**	**Micrel, Inc.**[8]	300kHz Buck Controller	1
U2	**MIC5233-5.0YM5**	**Micrel, Inc.**	LDO	1

Notes:

1. AVX: www.avx.com
2. MuRata: www.murata.com
3. TDK: www.tdk.com
4. Vishay: www.vishay.com
5. Sanyo: www.sanyo.com
6. Sumida: www.sumida.com
7. Central Semiconductor: www.centralsemi.com
8. **Micrel, Inc: www.micrel.com**

PCB Layout

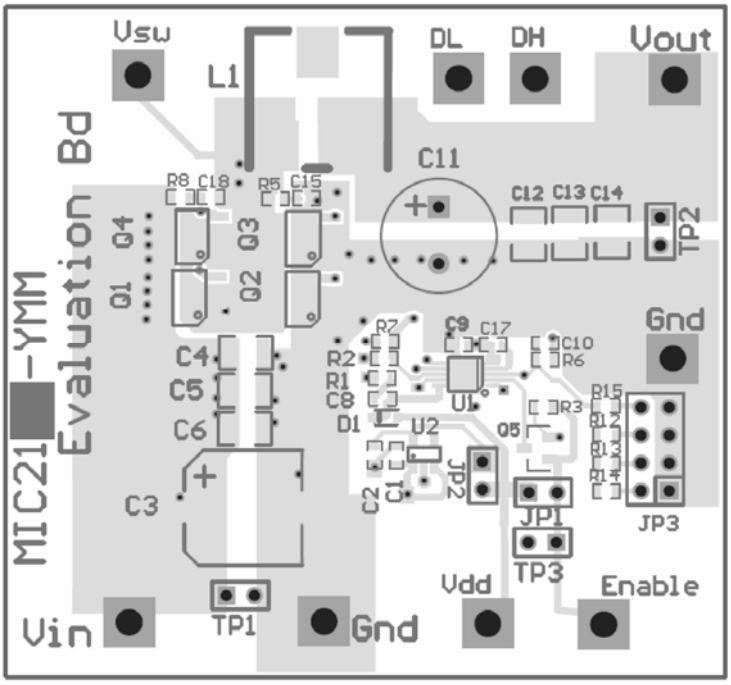


Figure 8. MIC2174 Evaluation Board Top Layer

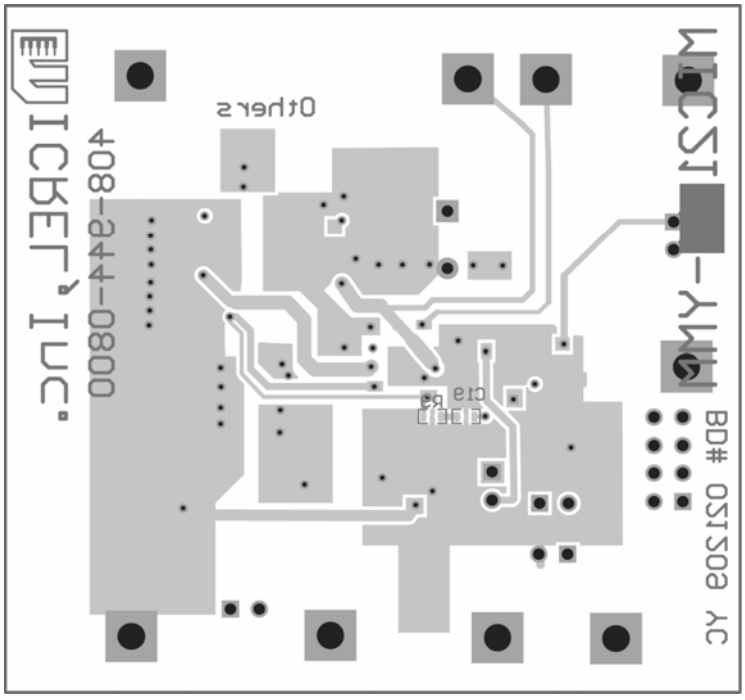


Figure 9. MIC2174 Evaluation Board Bottom Layer

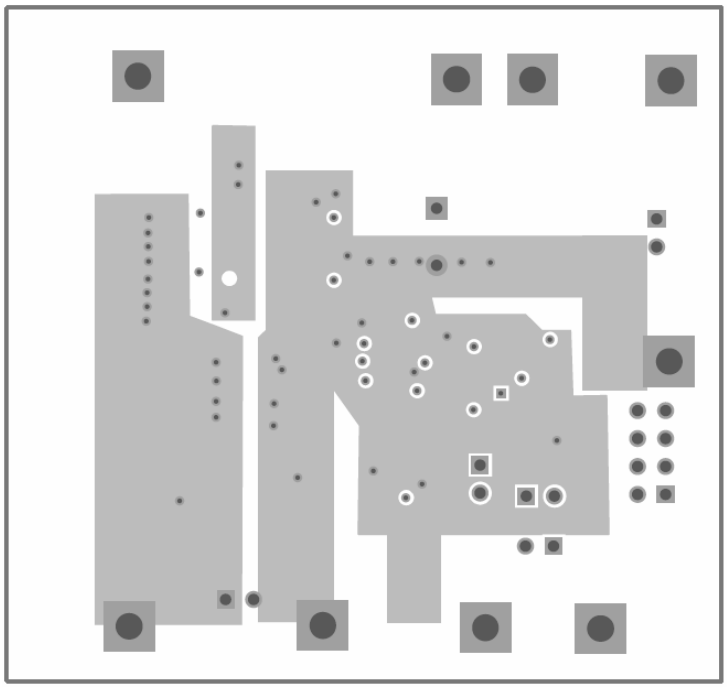


Figure 10. MIC2174 Evaluation Board Mid-Layer 1

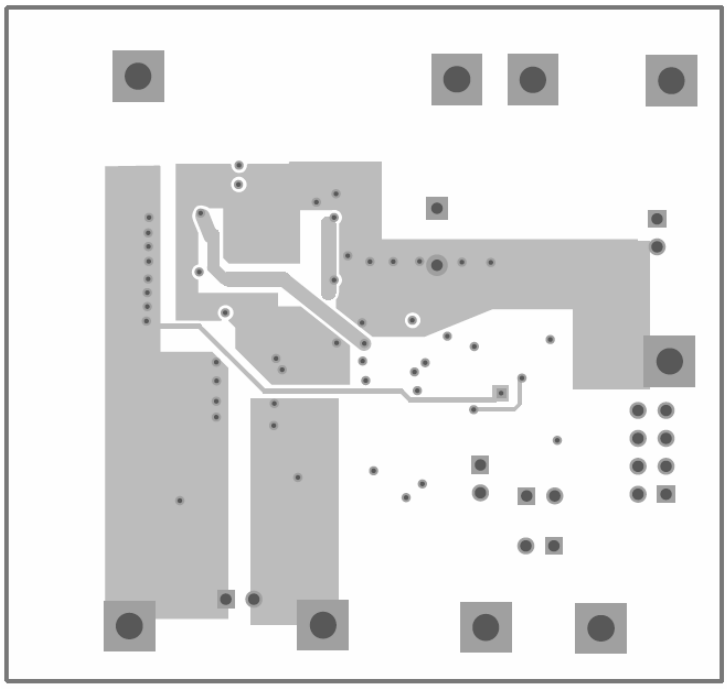
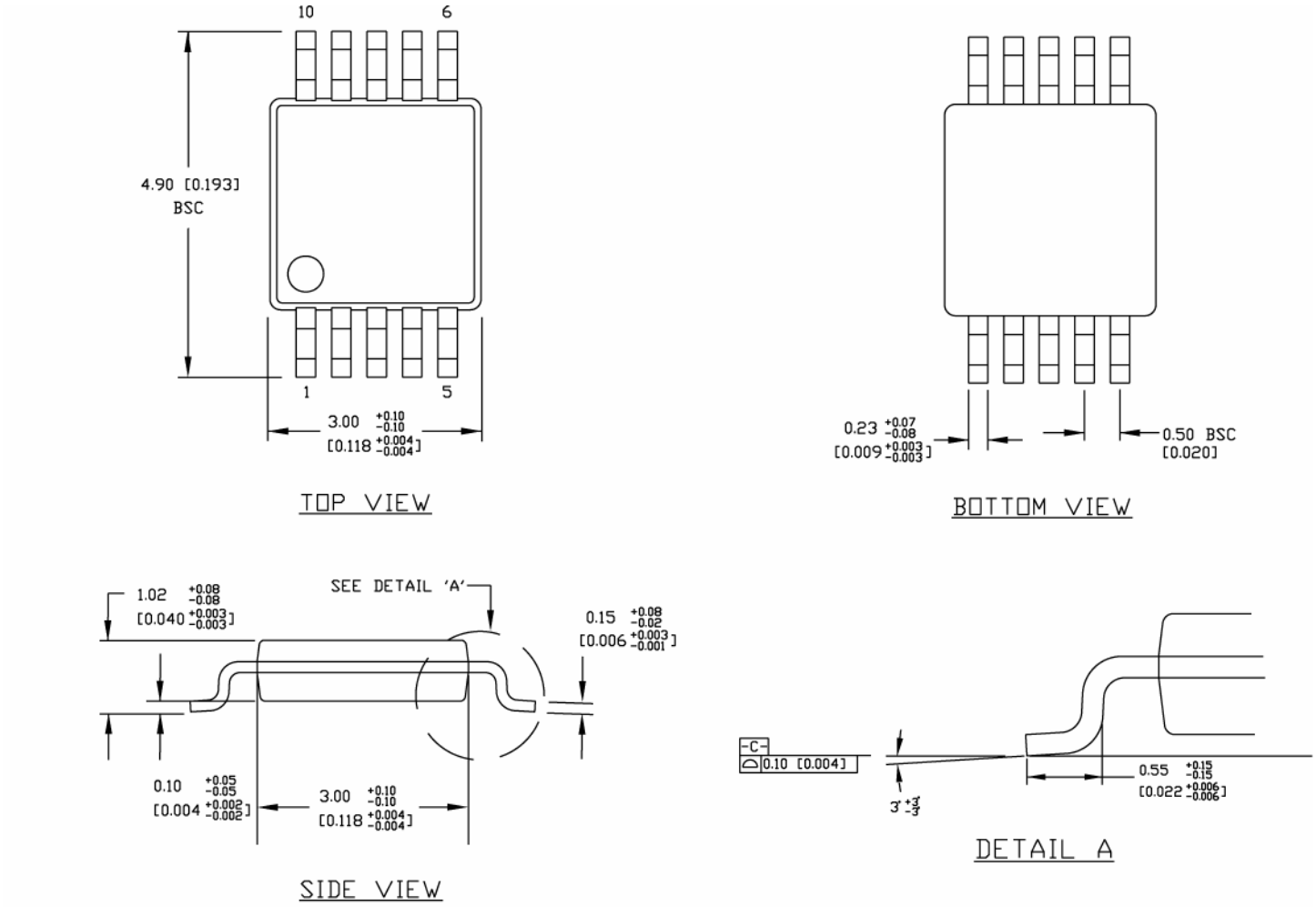


Figure 11. MIC2174 Evaluation Board Mid-Layer 2

Package Information



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.

10-Pin MSOP (MM)

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