

General Description

The MIC323x are constant current boost switching controllers specifically designed to power one or more strings of high power LEDs. The MIC323x have an input voltage range from 6V to 45V and are ideal for a variety of solid state lighting applications.

The MIC323x utilizes an external power device which offers a cost conscious solution for high power LED applications. The powerful drive circuitry can deliver up to 70W to the LED system. Power consumption has been minimized through the implementation of a 250mV feedback voltage reference providing an accuracy of $\pm 3\%$. The MIC323x family is dimmable via a pulse width modulated (PWM) input signal and also features an enable pin for low power shutdown.

Multiple MIC3230 ICs can be synchronized to a common operating frequency. The clocks of these synchronized devices can be used together in order to help reduce noise and errors in a system.

An external resistor sets the adjustable switching frequency of the MIC3230/1. The switching frequency can be between 100kHz and 1MHz. Setting the switching frequency provides the mechanism by which a design can be optimized for efficiency (performance) and size of the external components (cost). The MIC323x family of LED drivers also offer the following protection features: Overvoltage protection (OVP), thermal shutdown, and undervoltage lockout (UVLO).

The MIC3231 offers a dither feature to assist in the reduction of EMI. This is particularly useful in sensitive EMI applications, and provides for a reduction of emissions by approximately 10dB.

The MIC3232 is a 400kHz fixed frequency device offered in a small 10-pin MSOP package. The MIC3230/1 are offered in both the EPAD 16-pin TSSOP package and the 12-pin 3mm $\times$ 3mm VDFN package.

Features

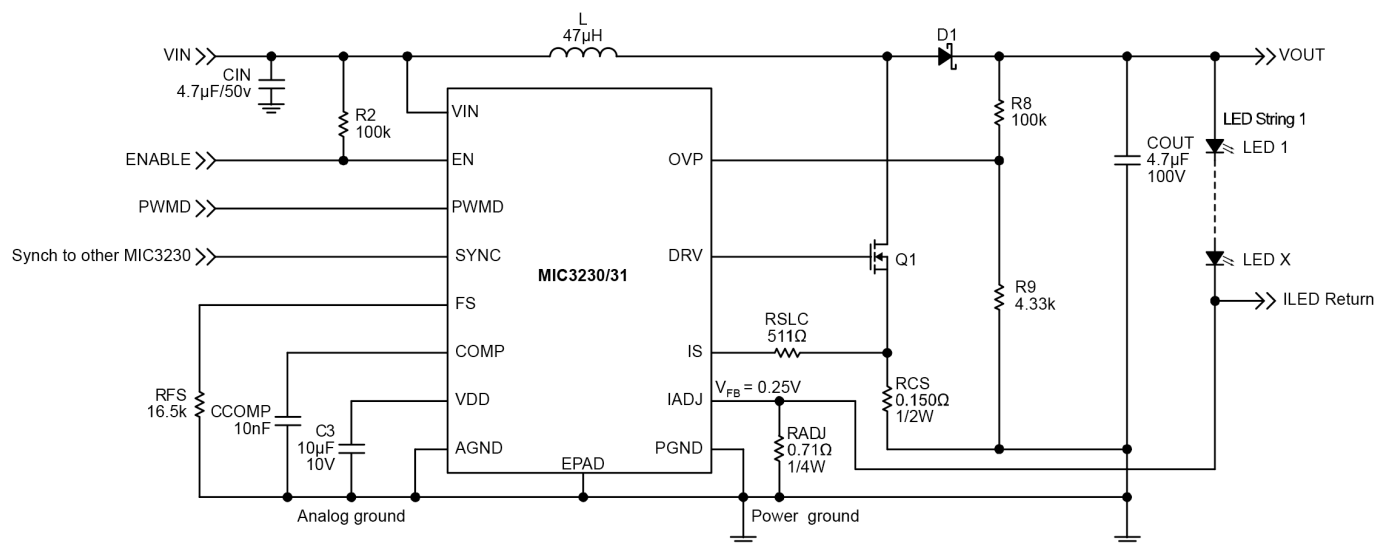
- 6V to 45V Input Supply Range
- Capable of Driving up to 70W
- Ultra-Low EMI via Dithering on the MIC3231
- Programmable LED Drive Current
- Feedback Voltage = 250mV $\pm 3\%$
- Programmable switching frequency (MIC3230/1) or 400kHz fixed frequency operation (MIC3232)
- PWM Dimming and Separate Enable Shutdown
- Frequency Synchronization with Other MIC3230s
- Protection Features:
 - Overvoltage Protection (OVP)
 - Overtemperature Protection
 - Undervoltage Lockout (UVLO)
- -40°C to $+125^{\circ}\text{C}$ Junction Temperature Range

Applications

- Street Lighting
- Solid State Lighting
- General Illumination
- Architectural Lighting
- Constant Current Power Supplies

Typical Application Circuit

Figure 1. Typical Application of the MIC3230 LED Driver



Product Option Matrix

Parameter	MIC3230	MIC3231	MIC3232
Input Voltage	6V to 45V	6V to 45V	6V to 45V
Synchronization	Yes	No	No
Dither	No	Yes	No
Frequency Range	Adjustable from 100kHz to 1MHz	Adjustable from 100kHz to 1MHz	Fixed Frequency of 400kHz
Package	16-Lead ePAD TSSOP, 12-Lead 3mm x 3mm VDFN	16-Lead ePAD TSSOP, 12-Lead 3mm x 3mm VDFN	10-Lead MSOP
Obsolete	Yes	No	Yes

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1. Electrical Characteristics

Table 1-1. Absolute Maximum Ratings

Supply Voltage (V_{IN})	+48V
Enable Pin Voltage	-0.3V to +6V
IADJ Voltage	+6V
MIC3230 ESD Rating	1.5kV HBM, 100V MM
MIC3231 ESD Rating	2kV HBM, 100V MM
MIC3232 ESD Rating	1.5kV HBM, 150V MM
Notes:	
1. Exceeding the absolute maximum rating may damage the device.	
2. Devices are ESD sensitive. Handling precautions recommended. Human body model, 1.5k Ω in series with 100pF.	

Table 1-2. Operating Ratings

Supply Voltage (V_{IN})	+6V to +45V
Notes:	
1. The device is not guaranteed to function outside its operating rating.	

Table 1-3. Electrical Characteristics

Electrical Characteristics: $V_{IN} = 12V$, $V_{EN} = 3.6V$, $L = 47\mu H$, $C = 4.7\mu F$; $T_J = +25^\circ C$; Bold values valid for $-40^\circ C \leq T_J \leq +125^\circ C$. Note 1						
Parameter	Symbol	Min.	Typ.	Max.	Units	Conditions
Supply Voltage Range	V_{IN}	6	—	45	V	—
Undervoltage Lockout	UVLO	3.5	4.9	5.5	V	—
Quiescent Current	I_{VIN}	—	3.2	10	mA	$V_{FB} > 275mV$ (to ensure device is not switching)
Shutdown Current	I_{SD}	—	30	—	μA	$V_{EN} = 0V$
Feedback Voltage at IADJ	V_{IADJ}	242.5	250	257.5	mV	Room temperature (3%)
		237.5	250	262.5		$-40^\circ C \leq T_J \leq +125^\circ C$ (5%)
Feedback Input Current	I_{ADJ}	—	1.2	3	μA	$V_{FB} = 250mV$
Line Regulation	—	—	2	—	%	$V_{IN} = 12V$ to 24V
Load Regulation	—	—	2	—	%	V_{OUT} to 2 x V_{OUT}
Maximum Duty Cycle	D_{MAX}	90	—	—	%	MIC3230 and MIC3232
		88	—	—		MIC3231
Enable Threshold	V_{EN}	1.5	1.15	—	V	Turn ON
		—	1.1	0.4	V	Turn OFF
Enable Pin Current	I_{EN}	—	17	30	μA	$V_{EN} = 3.3V$, $R_{EN} = 100k\Omega$
PWMD Threshold	V_{PWM}	1.5	0.75	—	V	Turn ON
		—	0.7	0.4	V	Turn OFF
PWMD Frequency Range	f_{PWMD}	0	—	500	Hz	Note 2, $L = 47\mu H$, $C = 4.7\mu F$
Programmable Oscillator Frequency	f_{SW}	—	109	—	kHz	$R_{FREQ} = 82.5k\Omega$
		360	400	440		$R_{FREQ} = 21k\Omega$
		—	950	—		$R_{FREQ} = 8.25k\Omega$
Fixed Frequency Option	f_{SW}	360	400	440	kHz	MIC3232YMM
Low EMI (MIC3231)	f_{DITHER}	—	± 12	—	%	Frequency dither shift from nominal

Table 1-3. Electrical Characteristics (continued)**Electrical Characteristics:** $V_{IN} = 12V$, $V_{EN} = 3.6V$, $L = 47\mu H$, $C = 4.7\mu F$; $T_J = +25^\circ C$; **Bold values valid for $-40^\circ C \leq T_J \leq +125^\circ C$.**
Note 1

Parameter	Symbol	Min.	Typ.	Max.	Units	Conditions
Current Limit Sense Voltage	V_{SENSE}	0.315	0.45	0.585	V	$R_{SENSE} = 390\Omega$
I_{SENSE} Peak Current Output	I_{SENSE}	—	250	—	μA	$R_{SENSE} = 390\Omega$
Overvoltage Protection	V_{OVP}	1.203	1.24	1.277	V	—
Driver Impedance	—	—	2.4	3.5	Ω	Sink
		—	2	—		Source
Driver Voltage High	V_{DRH}	7	9	11	V	$V_{IN} = 12V$
Overtemperature Threshold Shutdown	T_J	—	150	—	$^\circ C$	—
Thermal Shutdown	—	—	5	—	$^\circ C$	Hysteresis

Notes:

1. Specification for packaged product only.
2. Ensured by design.

Table 1-4. Temperature Specifications

Parameter	Symbol	Min.	Typ.	Max.	Units	Conditions
Temperature Ranges						
Storage Temperature Range	T_S	-65	—	+150	$^\circ C$	—
Junction Temperature Range	T_J	-40	—	+125	$^\circ C$	—
Lead Temperature	T_{LEAD}	—	—	+260	$^\circ C$	—
Package Thermal Resistances						
Thermal Resistance, MSOP-10Ld	θ_{JA}	—	130.5	—	$^\circ C/W$	—
Thermal Resistance, TSSOP-16Ld	θ_{JA}	—	36.5	—	$^\circ C/W$	—
Thermal Resistance, VDFN-12Ld	θ_{JA}	—	60.7	—	$^\circ C/W$	—

2. Typical Operating Characteristics

Note: The graphs and tables provided following this note are a statistical summary based on a limited number of samples and are provided for informational purposes only. The performance characteristics listed herein are not tested or guaranteed. In some graphs or tables, the data presented may be outside the specified operating range (e.g., outside specified power supply range) and therefore outside the warranted range.

Figure 2-1. Efficiency vs. PWM Dimming

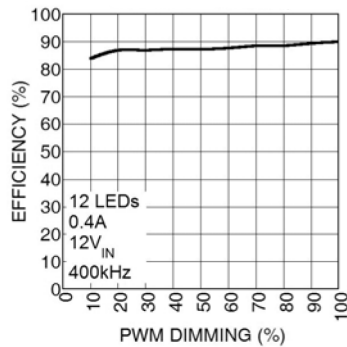


Figure 2-2. Efficiency vs. PWM Dimming

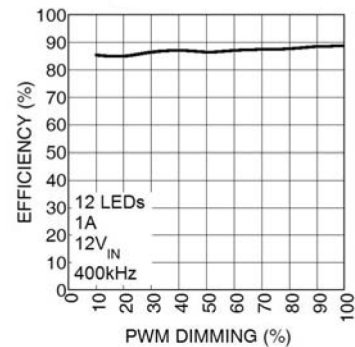


Figure 2-3. Efficiency vs. PWM Dimming

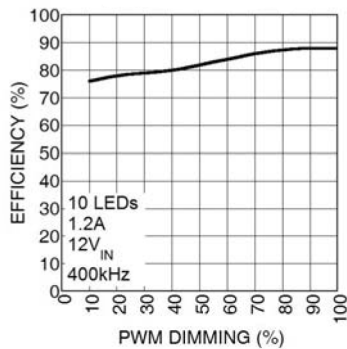


Figure 2-4. Efficiency vs. LED Current

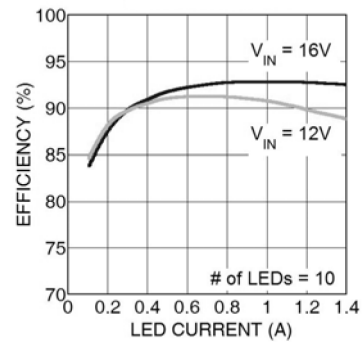


Figure 2-5. Efficiency vs. LED Current

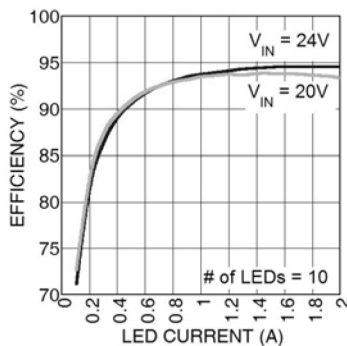


Figure 2-6. Current Limit vs. Temperature

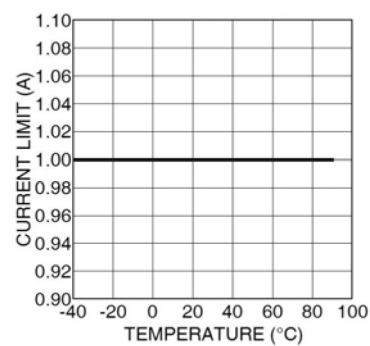


Figure 2-7. Feedback Voltage vs. Temperature

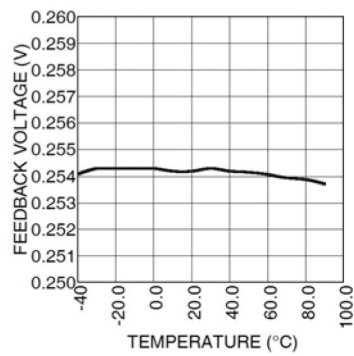


Figure 2-8. ILED vs. Temperature

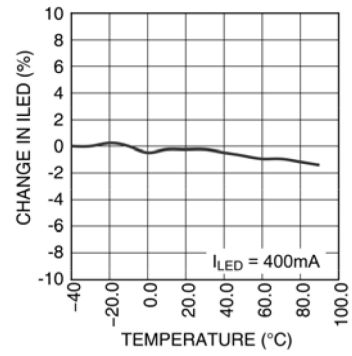


Figure 2-9. Switching Frequency vs. RFS

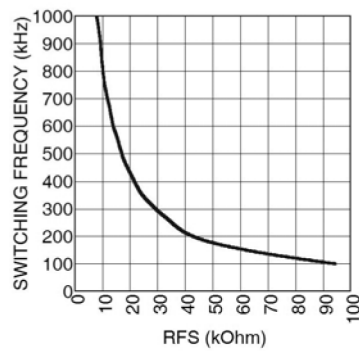


Figure 2-10. Frequency (Normalized to 400kHz) vs. Input Supply

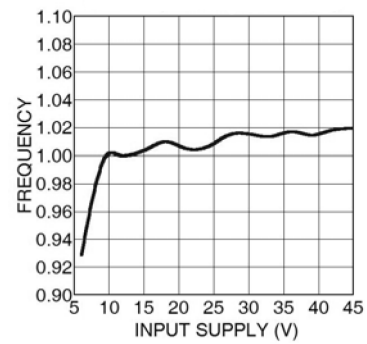


Figure 2-11. Frequency (Normalized to 400kHz) vs. Temperature

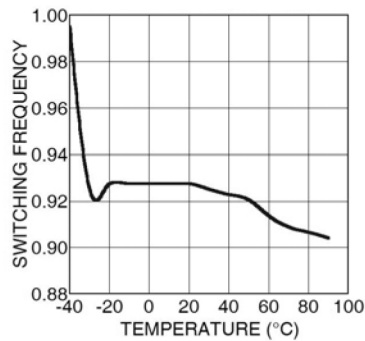


Figure 2-12. ILED vs. Input Supply

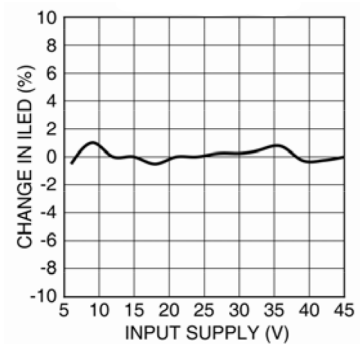


Figure 2-13. Efficiency vs. Frequency

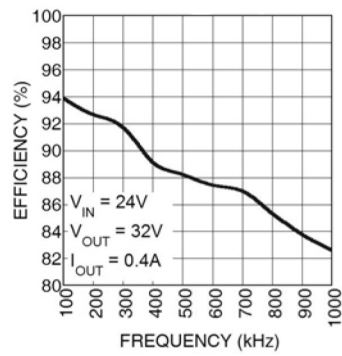


Figure 2-14. Efficiency vs. Frequency

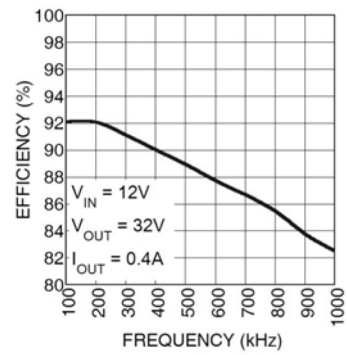
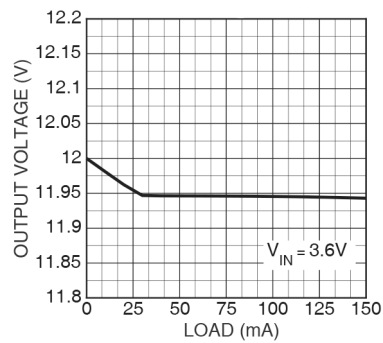


Figure 2-15. Load Regulation



3. Pin Description

VIN [1] ○ EN [2] PWMD [3] COMP [4] IADJ [5] [10] VDD [9] DRV [8] PGND [7] OVP [6] IS	VIN [1] EN [2] PWMD [3] COMP [4] IADJ [5] FS [6] [12] VDD [11] DRV [10] PGND [9] OVP [8] IS [7] SYNC/NC EPAD	N/C [1] VIN [2] EN [3] PWMD [4] COMP [5] IADJ [6] FS [7] AGND [8] [16] N/C [15] VDD [14] DRV [13] PGND [12] OVP [11] IS [10] SYNC/NC [9] N/C EPAD
MIC3232 10-Lead MSOP (MM)	MIC3230, MIC3231 12-Lead 3mm x 3mm VDFN (ML) See Product Option Matrix for selection	MIC3230, MIC3231 16-Lead TSSOP (TSE) See Product Option Matrix for selection

Table 3-1. Pin Descriptions

Pin Number VDFN	Pin Number TSSOP	Pin Number MSOP	Pin Name	Pin Function
—	1	—	NC	No connect.
1	2	1	VIN	Input Voltage (power) 6V to 45V.
2	3	2	EN	Enable Control (Input). Logic High ($\geq 1.5V$) enables the regulator. Logic Low ($\leq 0.4V$) shuts down the regulator. Connect a 100k Ω resistor from EN to VIN.
3	4	3	PWMD	PWM Dimming Input. Logic Low will disable the brightness control of the LED drivers.
4	5	4	COMP	Compensation (output): For external compensation.
5	6	5	IADJ	Feedback (input).
6	7	—	FS	Frequency Select (input). Connected to a resistor to determine the operating frequency.
—	8	—	AGND	Analog ground.
—	9	—	NC	No connect.
7	10	—	SYNC	Sync (output). Connect to another MIC3230 to synchronize multiple converters.
8	11	6	IS	Current Sense (input). Connected to external current sense resistor which in turn is connected to the source of the external FET as well as an external slope compensation resistor.
9	12	7	OVP	OVP divider connection (output). Connect the top of the divider string to the output. If the load is disconnected, the output voltage will rise until OVP reaches 1.25V and then will regulate around this point.
10	13	8	PGND	Power ground.
11	14	9	DRV	Drive Output: Connect to the gate of external FET (output).
12	15	10	VDD	VDD Filter for internal power rail. Do not connect an external load to this pin. Connect 10 μ F to GND.

Table 3-1. Pin Descriptions (continued)				
Pin Number VDFN	Pin Number TSSOP	Pin Number MSOP	Pin Name	Pin Function
—	16	—	NC	No connect.
EPAD	EPAD	—	EPAD	Connect to AGND.

4. Functional Description

A constant output current converter is the preferred method for driving LEDs. Small variations in current have a minimal effect on the light output, whereas small variations in voltage have a significant impact on light output. The MIC323x family of LED drivers are specifically designed to operate as constant current LED Drivers. See the [Typical Application Circuit](#) for more detail.

The MIC323x family is designed to operate as a boost controller, where the output voltage is greater than the input voltage. This configuration allows for the design of multiple LEDs in series to help maintain color and brightness. The MIC323x family can also be configured as a SEPIC controller, where the output voltage can be either above or below the input voltage.

The MIC323x have a very wide input voltage range, between 6V and 45V, to help accommodate a diverse range of input voltage applications. In addition, the LED current can be programmed to a wide range of values through the use of an external resistor. This provides design flexibility in adjusting the current for a particular application's need.

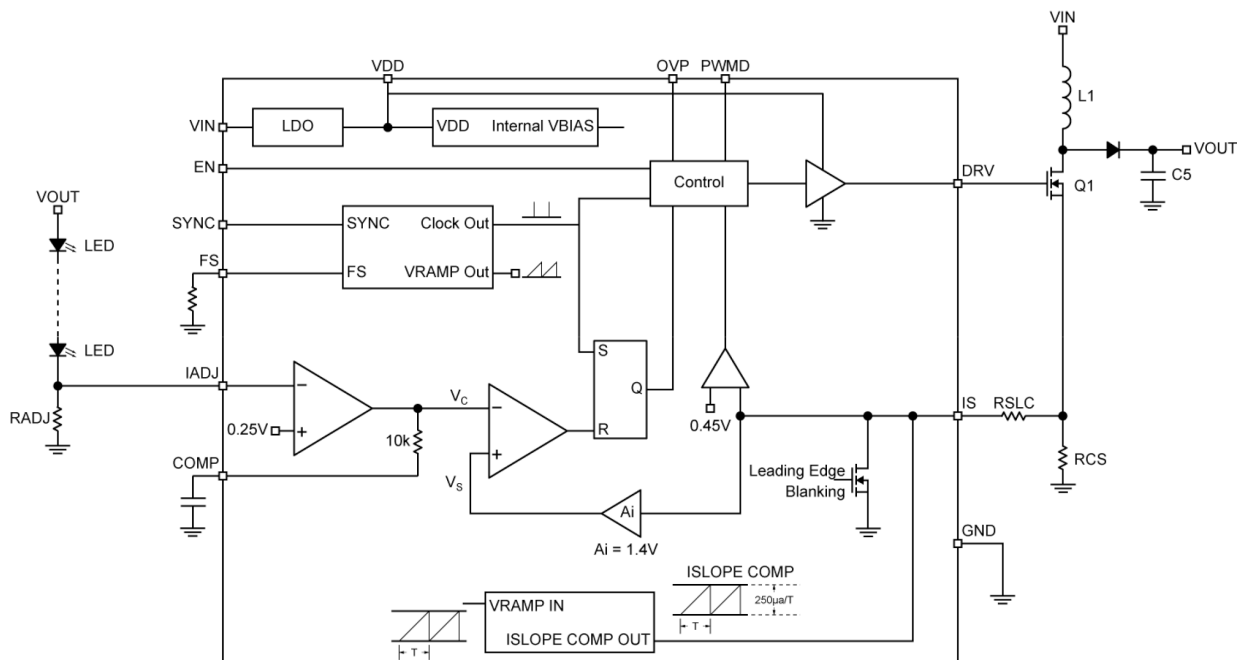
The MIC323x features a low impedance gate driver capable of switching large MOSFETs. This low impedance helps provide higher operating efficiency.

The MIC323x family can control the brightness of the LEDs via its PWM dimming capability. Applying a PWM signal (up to 500Hz) to the PWMD pin allows for control of the brightness of the LED.

Each member of the MIC323x family employs peak current mode control. Peak current mode control offers advantages over voltage mode control in the following manner: Current mode control can achieve a superior line transient performance compared to voltage mode control and through small signal analysis (not shown here), current mode control is easier to compensate than voltage mode control, thus allowing for a less complex control loop stability design. The functional block diagram is shown below.

4.1. Functional Block Diagram

Figure 4-1. MIC3230 Functional Block Diagram



5. Power Topology

5.1. Constant Output Current Controller

The MIC323x family is peak current mode boost controllers designed to drive high power LEDs. Unlike a standard constant output voltage controller, the MIC323x family has been designed to provide a constant output current. The MIC323x family is designed for a wide input voltage range, from 6V to 45V. In the boost configuration, the output can be set from V_{IN} up to 100V.

As a peak current mode controller, the MIC323x family provides the benefits of superior line transient response as well as an easier to design compensation.

This family of LED drivers features a built-in soft-start circuitry in order to prevent start-up surges. Other protection features include:

- Current Limit (I_{LIMIT}) - Current sensing for overcurrent and overload protection.
- Overvoltage Protection (OVP) - Output overvoltage protection to prevent operation above a safe upper limit.
- Undervoltage Lockout (UVLO) - UVLO is designed to prevent operation at very low input voltages.

5.2. Setting the LED Current

The current through the LED string is set via the value chosen for the current sense resistor, R_{ADJ} . This value can be calculated using the following equation:

Equation 5-1.

$$I_{LED} = \frac{0.25V}{R_{ADJ}}$$

Another important parameter to be aware of in the boost controller design is the ripple current. The amount of ripple current through the LED string is equal to the output ripple voltage divided by the LED AC resistance (R_{LED} , provided by the LED manufacturer) plus the current sense resistor (R_{ADJ}). The amount of allowable ripple through the LED string is dependent upon the application and is left to the designer's discretion. This equation is shown below:

Equation 5-2.

$$\Delta I_{LED} \approx \frac{V_{OUTRIPPLE}}{(R_{LED} + R_{ADJ})}$$

Where:

Equation 5-3.

$$V_{OUTRIPPLE} = \frac{I_{LED} \times D \times T}{C_{OUT}}$$

5.3. Reference Voltage

The voltage feedback loop of the MIC323x uses an internal reference voltage of 0.25V with an accuracy of $\pm 3\%$. The feedback voltage is the voltage drop across the current setting resistor (R_{ADJ}) as shown in the [Typical Application Circuit](#). When in regulation, the voltage at IADJ will equal 0.25V.

5.4. Output Overvoltage Protection

The MIC323x provide an OVP circuitry in order to help protect the system from an overvoltage fault condition. This OVP point can be programmed through the use of external resistors (R_8 and R_9 in the [Typical Application Circuit](#)). A reference value of 1.245V is used for the OVP. The following equation can be used to calculate the resistor value for R_9 to set the OVP point.

Equation 5-4.

$$R9 = \frac{R8}{\left(\frac{V_{OVP}}{1.245}\right) - 1}$$

5.5. LED Dimming

The MIC323x family of LED drivers can control the brightness of the LED string via the use of pulse width modulated (PWM) dimming. A PWM input signal of up to 500Hz can be applied to the PWM DIM pin (see the [Typical Application Circuit](#)) to pulse the LED string ON and OFF. It is recommended to use PWM dimming signals above 120Hz to avoid any recognizable flicker by the human eye. PWM dimming is the preferred way to dim a LED in order to prevent color/wavelength shifting, as occurs with analog dimming. The output current level remains constant during each PWMD pulse.

5.6. Oscillator and Switching Frequency Selection

The MIC323x family features an internal oscillator that synchronizes all of the switching circuits internal to the IC. This frequency is adjustable on the MIC3230 and MIC3231 and fixed at 400kHz in the MIC3232. In the MIC3230/1, the switching frequency can be set by choosing the appropriate value for the resistor, R1 according to the following equation:

Equation 5-5.

$$R_{FS}(k\Omega) = \left(\frac{7526}{f_{SW}(kHz)}\right)^{1.035}$$

5.7. SYNC (MIC3230 Only)

Multiple MIC3230 ICs can be synchronized by connecting their SYNC pins together. When synchronized, the MIC3230 with the highest frequency (host) will override the other MIC3230s (clients). The internal oscillator of the host IC will override the oscillator of the client part(s) and all MIC3230 will be synchronized to the same host switching frequency.

The SYNC pin is designed to be used only by other MIC3230s and is available on the MIC3230 only. If the SYNC pin is being unused, it is to be left floating (open). In the MIC3231, the SYNC pin is to be left floating (open).

5.8. Dithering (MIC3231 Only)

The MIC3231 has a feature which dithers the switching frequency by $\pm 12\%$. The purpose of this dithering is to help achieve a spread spectrum of the conducted EMI noise. This can allow for an overall reduction in noise emission by approximately 10dB.

5.9. Internal Gate Driver

External FETs are driven by the MIC323x's internal low impedance gate drivers. These drivers are biased from the V_{DD} and have a source resistance of 2Ω and a sink resistance of 3.5Ω .

5.10. VDD

VDD is an internal linear regulator powered by VIN and VDD is the bias supply for the internal circuitry of the MIC323x. A $10\mu F$ ceramic bypass capacitor is required at the VDD pin for proper operation. This pin is for filtering only and should not be utilized for operation.

5.11. Current Limit

The MIC323x family features a current limit protection feature to prevent any current runaway conditions. The current limit circuitry monitors current on a pulse by pulse basis. It limits the current through the inductor by sensing the voltage across R_{CS} . When 0.45V is present at the IS pin, the pulse is truncated. The next pulse continues as normally until the IS pin reaches 0.45V and it is truncated once again. This will continue until the output load is decreased.

Select R_{CS} using the following equation:

Equation 5-6.

$$R_{CS} = \frac{0.45}{\frac{(V_{OUT(MAX)} - V_{IN(MIN)}) \times D}{L \times f_{SW}}} + I_{LPKLIMIT}$$

5.12. Slope Compensation

The MIC323x is a peak current mode controller and requires slope compensation. Slope compensation is required to maintain internal stability across all duty cycles and prevent any unstable oscillations. The MIC323x uses slope compensation that is set by an external resistor, R_{SLC} . The ability to set the proper slope compensation through the use of a single external component results in design flexibility. This slope compensation resistor, R_{SLC} , can be calculated using the following equation:

Equation 5-7.

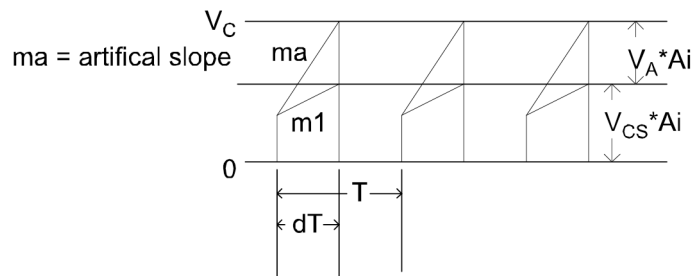
$$R_{SLC} = \frac{(V_{OUT(MAX)} - V_{IN(MIN)}) \times R_{CS}}{L \times 250 \mu A \times f_{SW}}$$

Where $V_{IN(MIN)}$ and $V_{OUT(MAX)}$ can be selected to system specifications.

5.13. Current Sense

The IS pin monitors the rising slope of the inductor current ($m1$ in Current Waveforms) and also sources a ramp current ($250\mu A/T$) that flows through R_{SLC} that is used for slope compensation. This ramp of $250\mu A$ per period, T , generates a ramped voltage across R_{SLC} and is labeled V_A in the following figure. The signal at the IS pin is the sum of $V_{CS} + V_A$ (as shown below). The current sense circuitry and block diagram is displayed in Current Sense Circuit. The IS pin is also used as the current limit (see the previous section on [Current Limit](#)).

Figure 5-1. Slope Compensation Waveforms



5.14. Soft-Start

The boost switching converter features a soft-start in order to power-up in a controlled manner, thereby limiting the inrush current from the line supply. Without this soft-start, the inrush current could be too high for the supply. To prevent this, a soft-start delay can be set using the compensation capacitor (C_{COMP} in the [Typical Application Circuit](#)). For switching to begin, the voltage on the compensation capacitor must reach about 0.7V. Switching starts with the minimum duty cycle and increases to the final duty cycle. As the duty cycle increases, V_{OUT} will increase from V_{IN} to its final value. A $6\mu A$ current source charges the compensation capacitor and the soft-start time can be calculated in the following equation:

Equation 5-8.

$$t_{SOFTSTART} \approx \frac{C_{COMP} \times V_{COMP(STEADYSTATE)}}{6 \mu A}$$

$V_{COMP(STEADYSTATE)}$ is usually between 0.7V and 3V, but can be as high as 5V.

Equation 5-9.

$$V_{COMP(STEADYSTATE)} = A_i \times (V_{A(PK)} + V_{CS(PK)})$$

Where:

$$V_{A(PK)} = I_{RAMP}/t \times R_{SLC} \times D \times T$$

$$V_{CS(PK)} = I_{L(PK)} \times R_{CS}$$

$$A_i = 1.4V/V$$

D = Duty cycle (0 to 1)

T = Period

A 10nF ceramic capacitor will make this system stable at all operating conditions.

5.15. Leading Edge Blanking

Large transient spikes due to the reverse recovery of the diode may be present at the leading edge of the current sense signal. Note that drive current can also cause such spikes. For this reason, a switch is employed to blank the first 100ns of the current sense signal. See [Figure 5-4](#).

Equation 5-10.

$$I_{IN(RMS)} = \frac{V_{OUT} \times I_{OUT}}{eff \times V_{IN}}$$

Equation 5-11.

$$I_{IN(AVE)} = \sqrt{(I_{IN(RMS)})^2 - \frac{(I_{IN(PP)})^2}{12}}$$

Equation 5-12.

$$I_{IN(PEAK)} = I_{IN(AVE)} + \frac{I_{IN(PP)}}{2}$$

Figure 5-2. Current Sense Circuit (An Explanation of the IS Pin)

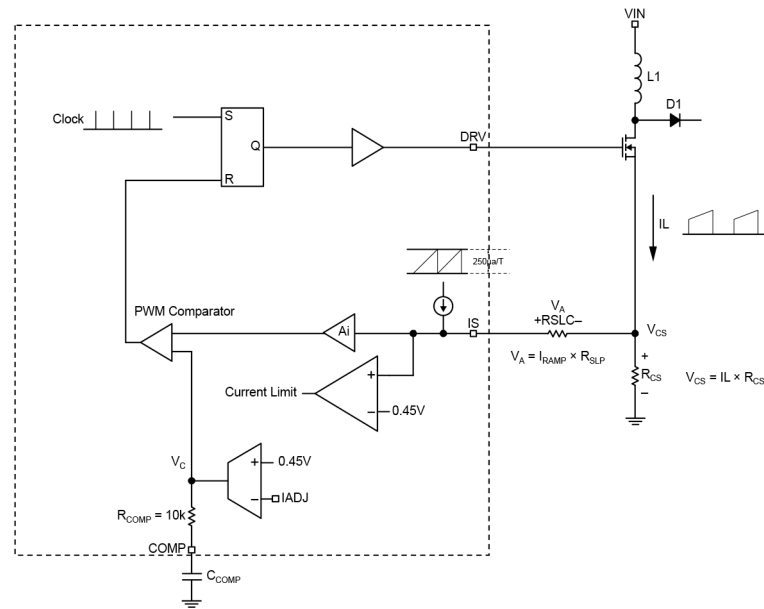


Figure 5-3. Current Waveforms

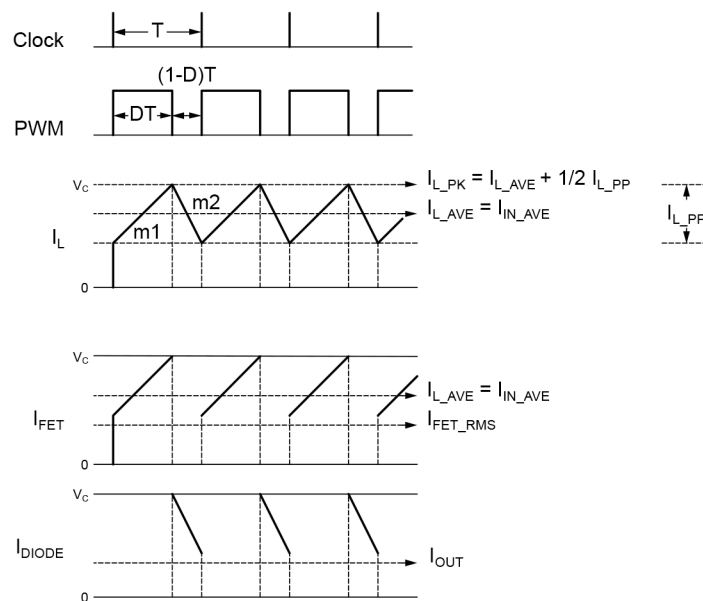
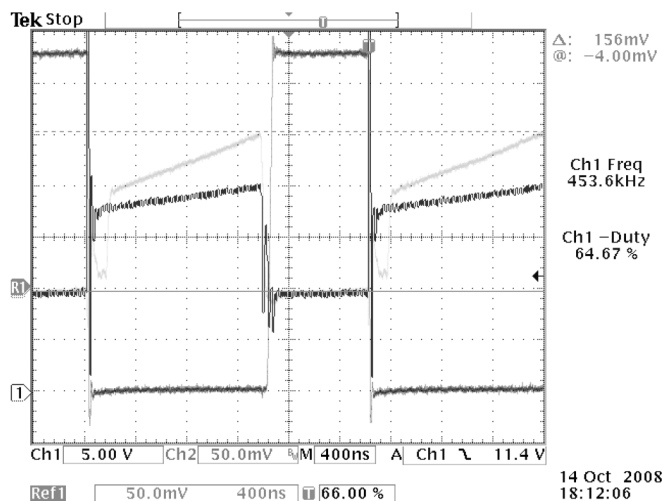


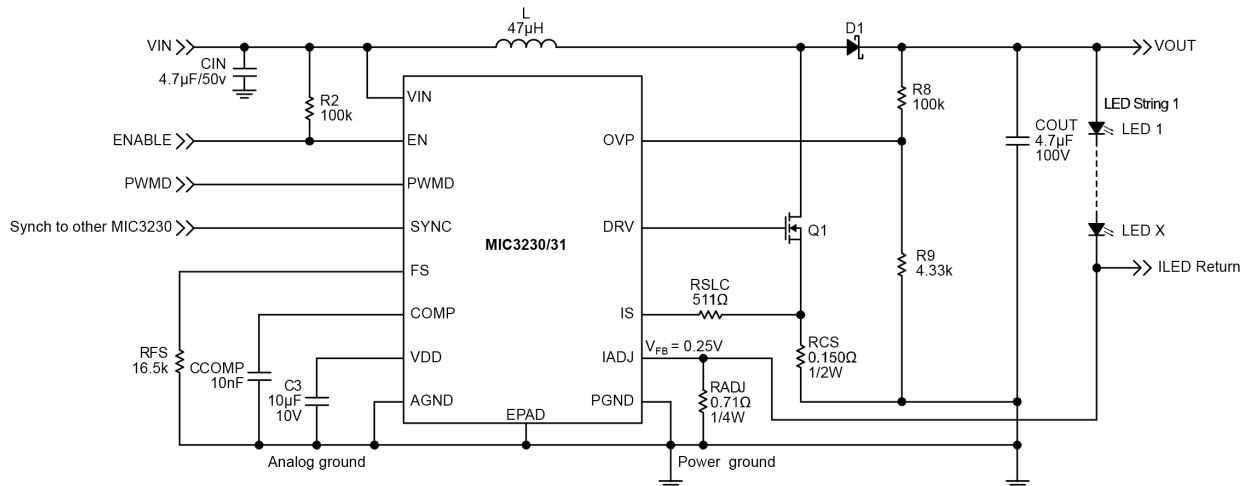
Figure 5-4. IS Pin and V_{RCS} (Ch1 = Switch Node, Ch2 = IS Pin, Ref1 = V_{CS})

5.16. Design Procedure for an LED Driver

Table 5-1. Design Example Parameters

Symbol	Parameter	Min.	Nom.	Max.	Units
Input					
V_{IN}	Input Voltage	8	12	14	V
I_{IN}	Input Current	—	—	2	A
Output					
LEDs	Number of LEDs	5	6	7	—
V_F	Forward Voltage of LED	3.2	3.5	4.0	V
V_{OUT}	Output Voltage	16	21	28	V
I_{LED}	LED Current	0.33	0.35	0.37	A
I_{PP}	Required I Ripple	—	40	—	mA
PWMD	PWM Dimming	0	—	100	%
OVP	Output Overvoltage Protection	—	30	—	V
System					
f_{SW}	Switching Frequency	—	500	—	kHz
eff	Efficiency	—	80	—	%
V_{DIODE}	Forward Drop of Schottky Diode	—	0.6	—	V

Figure 5-5. Design Example Schematic



5.17. Design Example

In this example, we will be designing a boost LED driver operating off a 12V input. This design has been created to drive six LEDs at 350mA with a ripple of about 12%. We are designing for 80% efficiency at a switching frequency of 500kHz.

5.17.1. Select R_{FS}

To operate at a switching frequency of 500kHz, the R_{FS} resistor must be chosen using the following equation.

Equation 5-13.

$$R_{FS}(k\Omega) = \frac{7526^{1.035}}{500} = 20.6k\Omega$$

Use the closest standard resistor value of 20.6kΩ.

5.17.2. Select R_{ADJ}

Having chosen the LED drive current to be 350mA in this example, the current can be set by choosing the R_{ADJ} resistor using the following equation:

Equation 5-14.

$$R_{ADJ} = \frac{0.25V}{0.35A} = 0.71\Omega$$

The power dissipation in this resistor is:

Equation 5-15.

$$P(R_{ADJ}) = I^2 \times R_{ADJ} = 87mW$$

Use a resistor rated at ¼ watt or higher. Choose the closest value from a resistor manufacturer.

5.17.3. Operating Duty Cycle

The operating duty cycle can be calculated using the equation below:

Equation 5-16.

$$D = \frac{V_{OUT} - eff \times V_{IN} + V_{DIODE}}{V_{OUT} + V_{DIODE}}$$

These can be calculated for the nominal (typical) operating conditions, but should also be understood for the minimum and maximum system conditions as listed below.

Equation 5-17.

$$D_{NOM} = \frac{V_{OUT(NOM)} - eff \times V_{IN(NOM)} + V_{SCHOTTKY}}{V_{OUT(NOM)} + V_{SCHOTTKY}}$$

$$D_{MAX} = \frac{V_{OUT(MAX)} - eff \times V_{IN(MIN)} + V_{SCHOTTKY}}{V_{OUT(MAX)} + V_{SCHOTTKY}}$$

$$D_{MIN} = \frac{V_{OUT(MIN)} - eff \times V_{IN(MAX)} + V_{SCHOTTKY}}{V_{OUT(MIN)} + V_{SCHOTTKY}}$$

Therefore, $D_{NOM} = 56\%$, $D_{MAX} = 78\%$, and $D_{MIN} = 33\%$.

5.17.4. Inductor Selection

First, it is necessary to calculate the RMS input current (nominal, min and max) for the system given the operating conditions listed in [Table 5-1](#). This minimum value of the RMS input current is necessary to ensure proper operation. The following values can then be calculated:

Equation 5-18.

$$I_{IN_RMS(MAX)} = \frac{V_{OUT(MAX)} \times I_{OUT(MAX)}}{eff \times V_{IN(MIN)}} = 1.64A_{RMS}$$

$$I_{IN_RMS(NOM)} = \frac{V_{OUT(NOM)} \times I_{OUT(NOM)}}{eff \times V_{IN(NOM)}} = 0.78A_{RMS}$$

$$I_{IN_RMS(MIN)} = \frac{V_{OUT(MIN)} \times I_{OUT(MIN)}}{eff \times V_{IN(MAX)}} = 0.48A_{RMS}$$

I_{OUT} is the same as I_{LED} .

Selecting the inductor current (peak-to-peak), I_{L_PP} , to be between 20% to 50% of $I_{IN_RMS(NOM)}$, in this case 40%, we obtain:

Equation 5-19.

$$I_{IN_PP(NOM)} = 0.4 \times I_{IN_RMS(NOM)} = 0.4 \times 0.78 = 0.31A_{PP}$$

See the [Current Waveforms](#).

It can be difficult to find large inductor values with high saturation currents in a surface mount package. Due to this, the percentage of the ripple current may be limited by the available inductor. It is recommended to operate in the continuous conduction mode. The selection of L described here is for continuous conduction mode.

Equation 5-20.

$$L = \frac{V_{IN} \times D \times T}{I_{IN_PP}}$$

Using the nominal values, we get:

Equation 5-21.

$$L = \frac{12V \times 0.56 \times 2 \mu s}{0.31A} = 43 \mu H$$

Select the next higher standard inductor value of 47μH.

Going back and calculating the actual ripple current gives:

Equation 5-22.

$$I_{IN_PP} = \frac{V_{IN(NOM)} \times D_{NOM} \times T}{L} = \frac{12V \times 0.56 \times 2 \mu s}{47 \mu H} = 0.29A_{PP}$$

The average input current is different than the RMS input current because of the ripple current. If the ripple current is low, then the average input current nearly equals the RMS input current. In the case where the average input current is different than the RMS, the average input current is calculated thusly:

Equation 5-23.

$$I_{IN_AVE(MAX)} = \sqrt{(I_{IN_RMS(MAX)})^2 - \frac{I_{IN_PP}^2}{12}}$$

$$I_{IN_AVE(MAX)} = \sqrt{1.64^2 - \frac{0.29^2}{12}} \approx 1.64A$$

The Maximum Peak input current, I_{L_PK} , can found using the following equation:

Equation 5-24.

$$I_{L_PK(MAX)} = I_{IN_AVE(MAX)} + 0.5 \times I_{L_PP(MAX)} = 1.78A$$

The saturation current (I_{SAT}) at the highest operating temperature of the inductor must be rated higher than this.

The power dissipated in the inductor is:

Equation 5-25.

$$P_{INDUCTOR} = I_{IN_RMS(MAX)}^2 \times DCR$$

5.17.5. Current Limit and Slope Compensation

Having calculated the I_{L_PK} above, we can set the current limit 20% above this maximum value:

Equation 5-26.

$$I_{L_PK} = 1.2 \times 1.6A = 1.9A$$

The internal current limit comparator reference is set at 0.45V. Therefore, when $V_{IS_PIN} = 0.45$, the IC enters current limit.

Equation 5-27.

$$0.45 = V_{A(PK)} + V_{CS(PK)}$$

Where:

$V_{A(PK)}$ = The peak of the V_A waveform.

$V_{CS(PK)}$ = The peak of the V_{CS} waveform.

Equation 5-28.

$$0.45 = I_{RAMP} \times R_{SLC} \times D \times I_{L_PK(LIMIT)} \times R_{CS}$$

To calculate the value of the slope compensation resistance, R_{SLC} , we can use the following equation:

Equation 5-29.

$$R_{SLC} = \frac{(V_{OUT(MAX)} - V_{IN(MIN)}) \times R_{CS}}{L \times 250 \mu A \times f_{SW}}$$

First, we must calculate R_{CS} , which is given below:

Equation 5-30.

$$R_{CS} = \frac{0.45}{\frac{(V_{OUT(MAX)} - V_{IN(MIN)}) \times D_{MAX}}{L \times f_{SW}} + I_{L_PK(LIMIT)}}$$

Therefore:

Equation 5-31.

$$R_{CS} = \frac{0.45}{\frac{(28V - 8V) \times 0.50}{47 \mu H \times 500kHz} + 1.9A} = 179m\Omega$$

Using a standard value 150mΩ resistor for R_{CS} , we obtain the following for R_{SLC} :

Equation 5-32.

$$R_{SLC} = \frac{(28 - 8) \times 150m\Omega}{47 \mu H \times 250 \mu A \times 500kHz} = 511\Omega$$

Use the next higher standard value if this not a standard value. In this example, 511Ω is a standard value.

Check: Because we must use a standard value for R_{CS} and R_{SLC} ; $I_{L_PK(LIMIT)}$ may be set at a different level (if the calculated value isn't a standard value) and we must calculate the actual $I_{L_PK(LIMIT)}$ value (remember, $I_{L_PK(LIMIT)}$ is the same as $I_{IN_PK(LIMIT)}$). Now we solve for $I_{L_PK(LIMIT)}$:

Equation 5-33.

$$I_{IN_PK(LIMIT)} = \frac{0.45 - I_{RAMP} \times R_{SLC} \times D}{R_{CS}}$$

$$I_{IN_ACTUAL(LIMIT)} = \frac{0.45 - 250 \mu A \times 511 \times 0.75}{0.15} = 2.34A$$

This is higher than the initial $1.2 \times I_{L_PK(MAX)} = 1.9A$ limit because we have to use standard values for R_{CS} and for R_{SLC} . If $I_{IN_ACTUAL(LIMIT)}$ is too high, then use a higher value for R_{CS} . The calculated value of R_{CS} for a 1.9A current limit is 179mΩ. In this example, we have chosen a lower value for R_{CS} , which results in a higher current limit. If we use a higher standard value, the current limit will have a lower value. The designer does not have the same choices for small valued resistors as with larger valued resistors. The choices differ from resistor manufacturers. If too large a current sense resistor is selected, the maximum output power may not be able to be achieved at low input line voltage

levels. Make sure the inductor will not saturate at the actual current limit $I_{IN_ACTUAL(LIMIT)}$. Perform a check at $I_{IN} = 2.34A_{PK}$.

Equation 5-34.

$$V_{IS_PIN} = 250 \mu A \times 0.78 \times 511\Omega + 2.34A \times 150m\Omega = 0.45V$$

Maximum power dissipated in R_{CS} is:

Equation 5-35.

$$P_{RCS} = I_{RCS(RMS)}^2 \times R_{CS}$$

Equation 5-36.

$$I_{RCS_RMS(MAX)} = I_{FET_RMS(MAX)} - \sqrt{D \times \left(I_{IN_AVE(MAX)}^2 + \frac{I_{L(PP)}^2}{12} \right)}$$

$$I_{RCS(RMS)} = \sqrt{0.78 \times \left(1.64^2 + \frac{0.26^2}{12} \right)} = 1.44A_{RMS}$$

$$P_{RCS} = 1.25^2 \times .15 = 0.31W$$

Use a ½ Watt resistor for R_{CS} .

5.17.6. Output Capacitor

In this LED driver application, the ILED ripple current is a more important factor compared to that of the output ripple voltage (although the two are directly related). To find the C_{OUT} for a required ILED ripple use the following calculation:

For an output ripple $I_{LED_RIPPLE} = 20\%$ of I_{LED_NOM} :

Equation 5-37.

$$I_{LED_RIPPLE} = 0.2 \times 0.35 = 70mA$$

$$C_{OUT} = \frac{I_{LED_NOM} \times D_{NOM} \times T}{I_{LED_RIPPLE} \times (R_{ADJ} + R_{LED(TOTAL)})}$$

Find the equivalent AC resistance, R_{LED_AC} , from the data sheet of the LED. This is the inverse slope of the ILED vs. V_F curve, i.e.:

Equation 5-38.

$$R_{LED_AC} = \frac{\Delta V_F}{\Delta I_{LED}}$$

In this example, use $R_{LED_AC} = 0.1\Omega$ for each LED.

If the LEDs are connected in series, multiply $R_{LED_AC} = 0.1\Omega$ by the total number of LEDs. In this example of 6 LEDs, we obtain the following:

Equation 5-39.

$$R_{LED_TOTAL} = 6 \times 0.1\Omega = 0.6\Omega$$

$$C_{OUT} = \frac{I_{LED_NOM} \times D_{NOM} \times T}{I_{LED_RIPPLE} \times (R_{ADJ} + R_{LED(TOTAL)})} = 4.1 \mu F$$

Use the next highest standard value, which is 4.7μF.

There is a trade off between the output ripple and the rising edge of the PWM pulse. This is because between PWM dimming pulses, the converter stops pulsing and C_{OUT} will start to discharge. The amount that C_{OUT} will discharge depends on the time between PWM dimming pulses. At the next PWM pulse, C_{OUT} has to be charged up to the full output voltage V_{OUT} before the desired LED current flows.

5.17.7. Input Capacitor

The input current is shown in Current Waveforms. For superior performance, ceramic capacitors should be used because of their low equivalent series resistance (ESR). The input ripple current is equal to the ripple in the inductor plus the ripple voltage across the input capacitor, which is the ESR of C_{IN} times the inductor ripple. The input capacitor will also bypass the EMI generated by the converter as well as any voltage spikes generated by the inductance of the input line.

For a required V_{IN_RIPPLE} :

Equation 5-40.

$$C_{IN} = \frac{I_{IN_PP}}{8 \times V_{IN_RIPPLE} \times f_{SW}} = \frac{0.28A}{8 \times 50mV \times 500kHz} = 1.4 \mu F$$

This is the minimum value that should be used. The input capacitor should also be rated for the maximum RMS input current. To protect the IC from inductive spikes or any overshoot, a larger value of input capacitance may be required and it is recommended that ceramic capacitors be used. In this design example, a value of 4.7μF ceramic capacitor was selected.

5.17.8. MOSFET Selection

In this design example, the FET has to hold off an output voltage maximum of 30V. It is recommended to use an 80% derating value on switching FETs, so a minimum of a 38V FET should be selected. In this design example, a 75V FET has been selected.

The switching FET power losses are the sum of the conduction loss and the switching loss:

Equation 5-41.

$$P_{FET} = P_{FET_COND} + P_{FET_SWITCH}$$

The conduction loss of the FET is when the FET is turned on. The conduction power loss of the FET is found by the following equation:

Equation 5-42.

$$P_{FET_COND} = I_{FET_RMS}^2 \times R_{DS(ON)}$$

Where:

Equation 5-43.

$$I_{FET_RMS} = \sqrt{D \times \left(I_{IN_AVE}^2 + \frac{I_{L_PP}^2}{12} \right)}$$

The switching losses occur during the switching transitions of the FET. The transition times, $t_{TRANSITION}$, are the times when the FET is turning off and on. There are two transition times per period, T . It is important not to confuse T (the period) with the transition time, $t_{TRANSITION}$.

Equation 5-44.

$$T = \frac{1}{f_{SW}}$$

Equation 5-45.

$$P_{FET_SWITCH(MAX)} = I_{FET_AVE(MAX)} \times V_{OUT(MAX)} \times t_{TRANSITION(MAX)} \times f_{SW}$$

To find $t_{TRANSITION(MAX)}$:

Equation 5-46.

$$t_{TRANSITION(MAX)} \approx \frac{Q_G}{I_{GATEDRV}}$$

Where:

Q_G = The total gate charge of the external MOSFET provided by the MOSFET manufacturer and the Q_G should be chosen at a $V_{GS} \approx 10V$. This is not an exact value, but is more of an estimate of $t_{TRANSITION(MAX)}$. The FET manufacturers' provide a gate charge at a specified V_{GS} voltage:

Equation 5-47.

$$C_{IN_FET} = \frac{Q_G}{@V_{GS}}$$

This is the FET's input capacitance. Select a FET with $R_{DS(ON)}$ and Q_G such that the external power is below about 0.7W for a SO-8 or about 1W for a PowerPak (FET package). The Vishay Siliconix Si7148DP in a PowerPak SO-8 package is one good choice. The internal gate driver in the MIC3230/1/2 is 2A. From the Si7148DP data sheet:

- $R_{DS(ON),25^{\circ}C} = 0.0145\Omega$
- Total Gate Charge = 68nC (typical)

The $R_{DS(ON)}(temp)$ is a function of temperature. As the temperature in the FET increases so does the $R_{DS(ON)}$. To find $R_{DS(ON)}(temp)$ use the following equation or simply double the $R_{DS(ON)}(25^{\circ}C)$ for $R_{DS(ON)}(125^{\circ}C)$.

Equation 5-48.

$$R_{DS(ON)}(temp) = R_{DS(ON)}(25^{\circ}C) \times 1.007^{(TEMP-25^{\circ}C)}$$

The $R_{DS(ON)}(temp)$ at $125^{\circ}C$ is:

Equation 5-49.

$$R_{DS(ON)}(temp) = 0.0145 \times (1.007^{(125^{\circ}C - 25^{\circ}C)}) \approx 30m\Omega$$

$$P_{FET_COND} = 1.64^2 \times 30m\Omega = 62mW$$

$$t_{TRANSITION} \approx \frac{Q_G}{I_{GATEDRV}} = \frac{68nC}{2A} = 34ns$$

$$I_{FET_AVE(MAX)} = 1.64A$$

$$V_{OUT(MAX)} = 28V$$

$$P_{FET_SWITCH(MAX)} = 1.64A \times 28V \times 34ns \times 500kHz = 0.78W$$

$$P_{FET} = 62mW + 0.78W = 0.84W$$

This is about the limit for a part on a circuit board without having to use any additional heat sinks.

5.17.9. Rectifier Diode

A Schottky Diode is best used here because of the lower forward voltage and the low reverse recovery time. The voltage stress on the diode is the maximum V_{OUT} and, therefore, a diode with a higher rating than the maximum V_{OUT} should be used. An 80% de-rating is recommended here as well.

Equation 5-50.

$$P_{DIODE} \approx V_{SCHOTTKY} \times I_{OUT(MAX)}$$

Equation 5-51.

$$P_{DIODE} \approx V_{SCHOTTKY} \times I_{LED(MAX)} \approx 0.25W$$

5.17.10. MIC3230 Power Losses

The power losses in MIC3230 are:

Equation 5-52.

$$P_{MIC3230} = Q_{GATE} \times V_{GATE} \times f_{SW} + I_Q \times V_{IN}$$

Where:

Q_{GATE} = The total gate charge of the external MOSFET.

V_{GATE} = The gate drive voltage of the MIC3230.

f_{SW} = The switching frequency.

I_Q = The quiescent current of the MIC3230, found in the Electrical Characteristics table (3.2mA).

V_{IN} = The voltage at the VIN pin of the MIC3230.

So, using that equation above:

Equation 5-53.

$$P_{MIC3230} = 68nf \times 12 \times 500kHz + 3.2mA \times 14 = 0.45W$$

5.17.11. Overvoltage Protection

Set OVP higher than the maximum output voltage by at least one volt. To find the resistor divider values for OVP use the following equation and set the OVP = 30V and R8 = 100kΩ:

Equation 5-54.

$$R9 = \frac{100k\Omega \times 1.245}{30 - 1.245} = 4.33k\Omega$$

5.18. PCB Layout

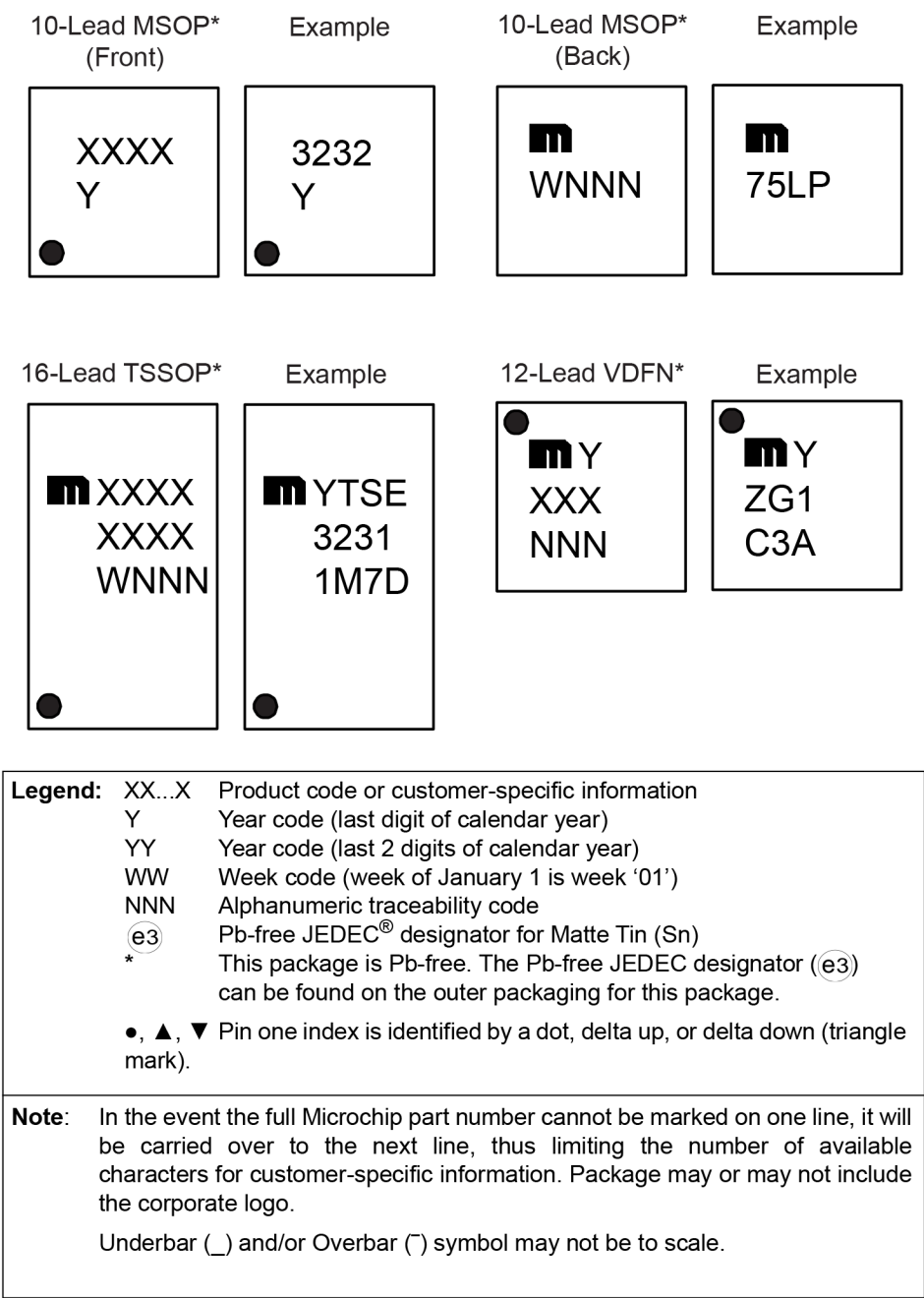
1. All typologies of DC-to-DC converters have a reverse recovery current (RRC) of the flyback or (freewheeling) diode. Even a Schottky diode, which is advertised as having zero RRC, it really is not zero. The RRC of the freewheeling diode in a boost converter is even greater than in the Buck converter. This is because the output voltage is higher than the input voltage and the diode has to charge up to $-V_{OUT}$ during each on-time pulse and then discharge to V_F during the off-time.

2. Even though the RRC is very short (tens of nanoseconds), the peak currents are high (multiple amperes). The high RRC causes a voltage drop on the ground trace of the PCB and, if the converter control IC is referenced to this voltage drop, the output regulation will suffer.
3. It is important to connect the IC's reference to the same point as the output capacitors to avoid the voltage drop caused by RRC. This is also called a star connection or single point grounding.
4. Feedback trace: The high impedance traces of the FB should be short.

6. Package Information

6.1. Package Marking Drawing

Figure 6-1. Package Marking Drawing



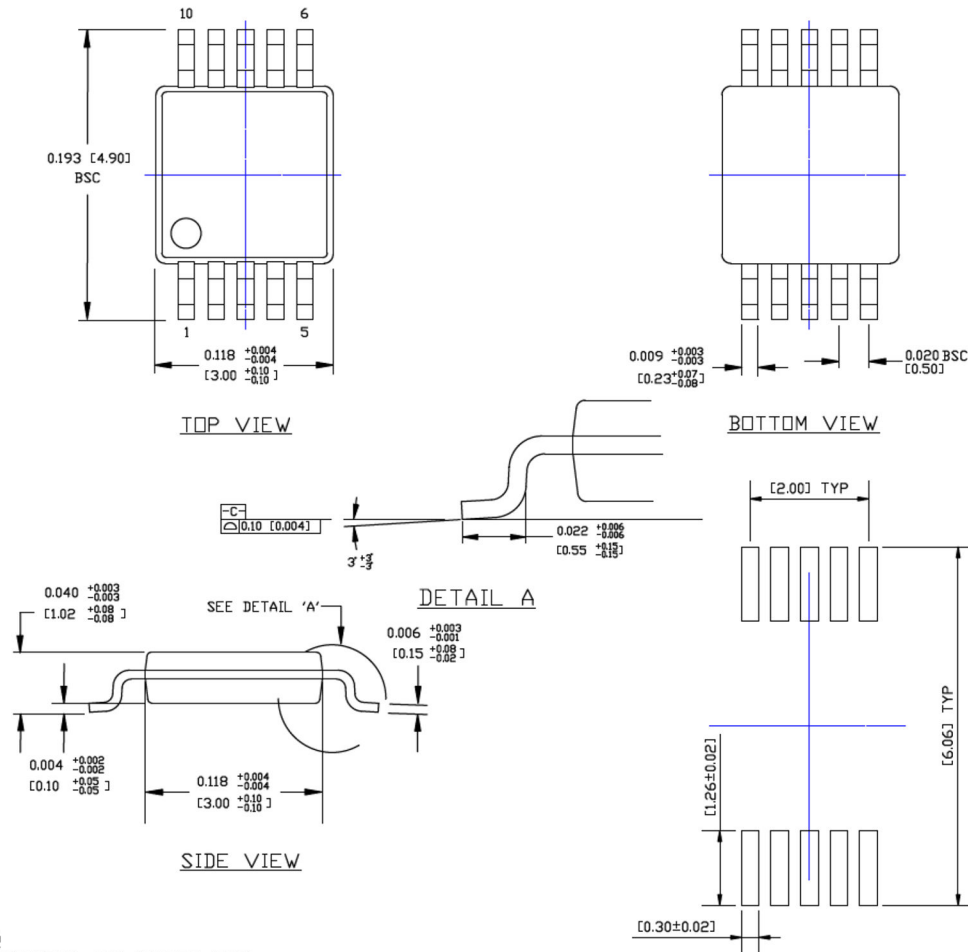
Note: If the full seven-character YYWWNNN code cannot fit on the package, the following truncated codes are used based on the available marking space: 6 Characters = YWWNNN; 5 Characters = WWNNN; 4 Characters = WNNN; 3 Characters = NNN; 2 Characters = NN; 1 Character = N

6.2. Package Outline Drawings

TITLE

10 LEAD MSOP PACKAGE OUTLINE & RECOMMENDED LAND PATTERN

DRAWING #	MSOP-10LD-PL-1	UNIT	INCH [MM]
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NOTES:

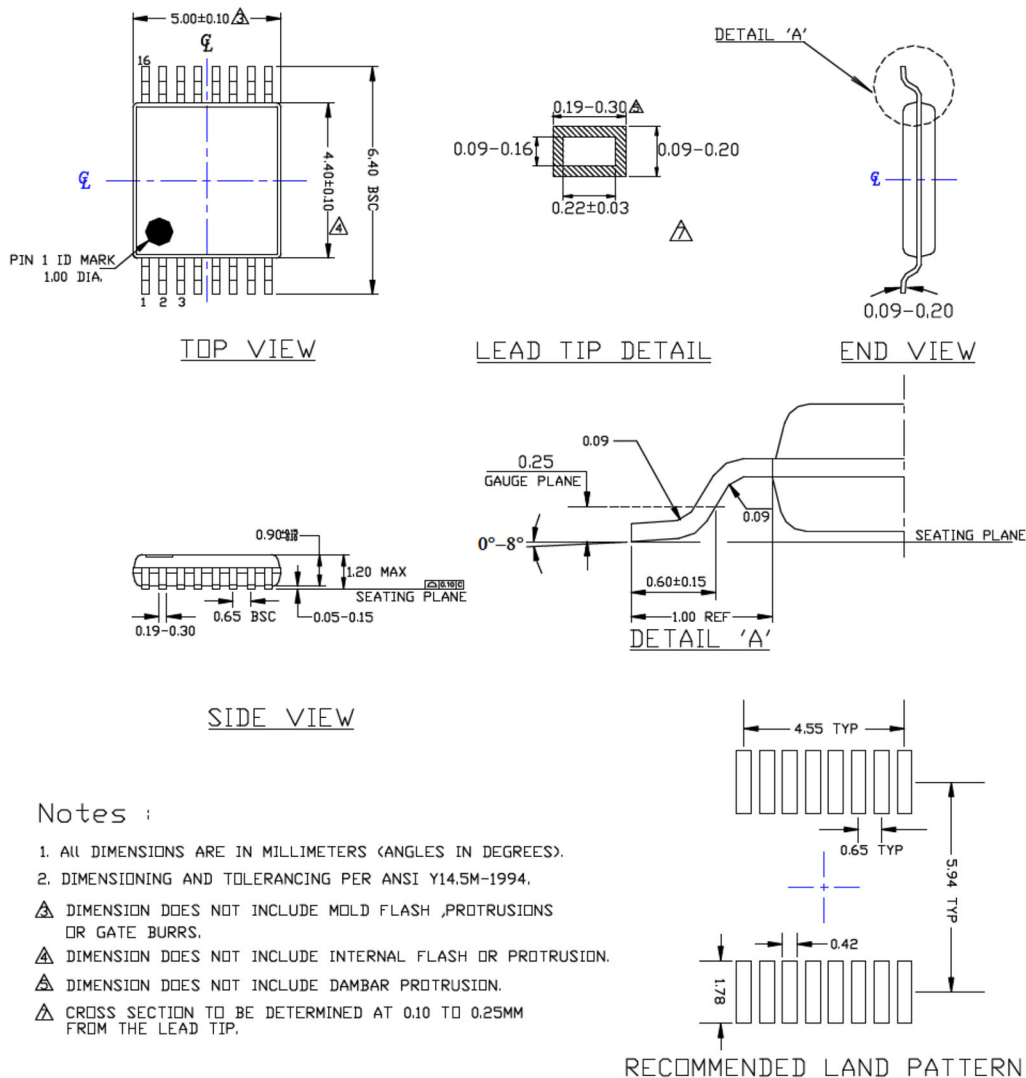
1. DIMENSIONS ARE INCHES [MM].
2. CONTROLLING DIMENSION: MM
3. DIMENSION DOES NOT INCLUDE MOLD FLASH OR PROTRUSIONS, EITHER OF WHICH SHALL NOT EXCEED 0.008 [0.20] PER SIDE.

Note: For the most current package drawings, please see the Microchip Packaging Specification located at <http://www.microchip.com/packaging>.

TITLE

16 LEAD TSSOP PACKAGE OUTLINE & RECOMMENDED LAND PATTERN

DRAWING #	TSSOP-16LD-PL-1	UNIT	MM
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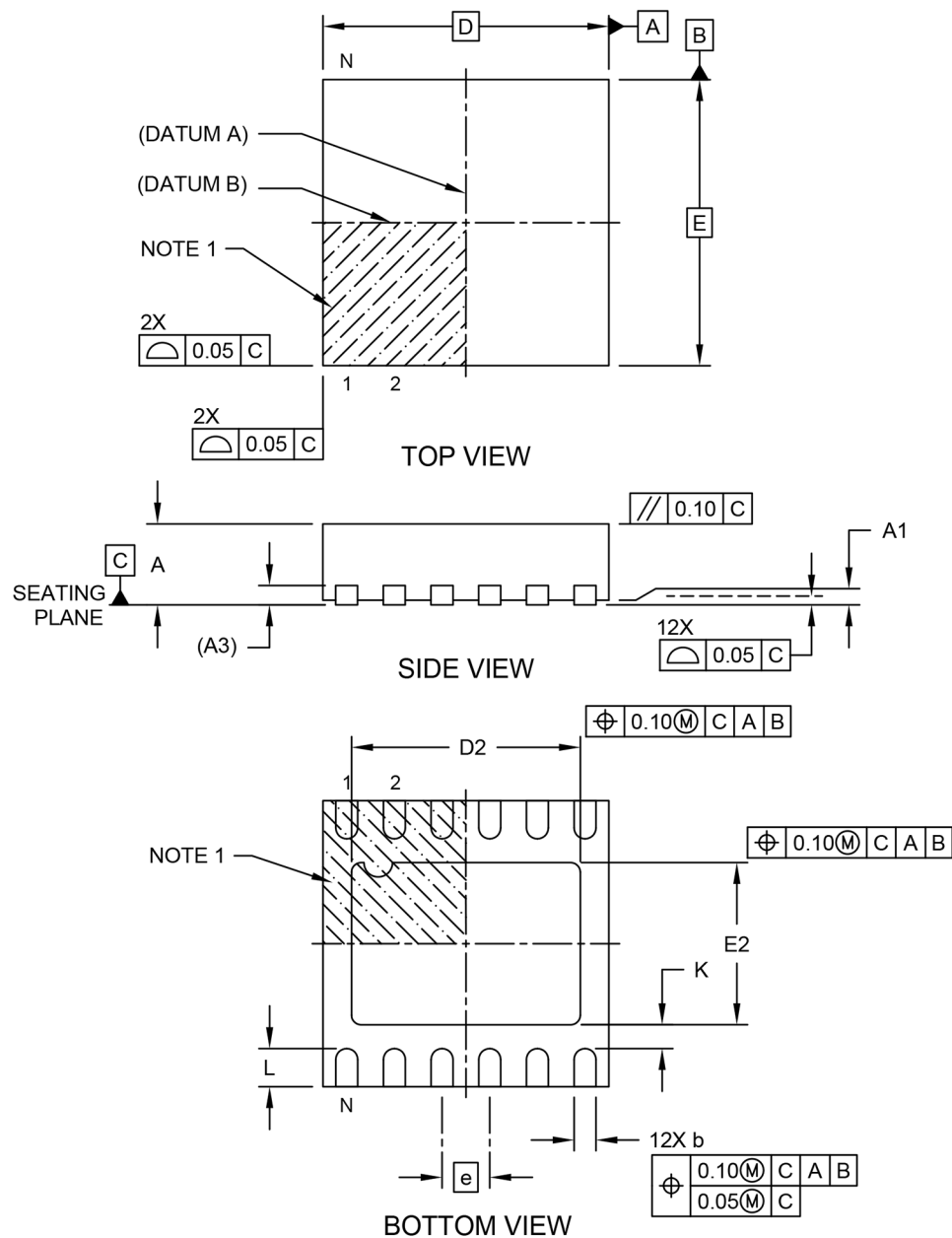
Notes :

1. ALL DIMENSIONS ARE IN MILLIMETERS (ANGLES IN DEGREES).
2. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M-1994.
3. DIMENSION DOES NOT INCLUDE MOLD FLASH, PROTRUSIONS OR GATE BURRS.
4. DIMENSION DOES NOT INCLUDE INTERNAL FLASH OR PROTRUSION.
5. DIMENSION DOES NOT INCLUDE DAMBAR PROTRUSION.
6. CROSS SECTION TO BE DETERMINED AT 0.10 TO 0.25MM FROM THE LEAD TIP.

Note: For the most current package drawings, please see the Microchip Packaging Specification located at <http://www.microchip.com/packaging>.

**12-Lead Very Thin Plastic Dual Flat, No Lead Package (JQA) - 3x3 mm Body [VDFN]
With 2.4x1.7 mm Exposed Pad; Micrel Legacy Package**

Note: For the most current package drawings, please see the Microchip Packaging Specification located at <http://www.microchip.com/packaging>

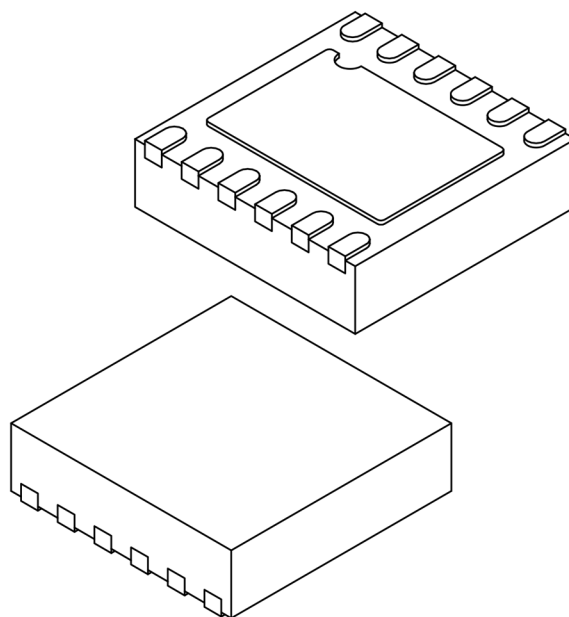


Microchip Technology Drawing C04-1020 Rev. A Sheet 1 of 2

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12-Lead Very Thin Plastic Dual Flat, No Lead Package (JQA) - 3x3 mm Body [VDFN] With 2.4x1.7 mm Exposed Pad; Micrel Legacy Package

Note: For the most current package drawings, please see the Microchip Packaging Specification located at <http://www.microchip.com/packaging>



Units		MILLIMETERS		
Dimension Limits		MIN	NOM	MAX
Number of Terminals	N	12		
Pitch	e	0.50 BSC		
Overall Height	A	0.80	0.85	0.90
Standoff	A1	0.00	0.02	0.05
Terminal Thickness	A3	0.203 REF		
Overall Length	D	3.00 BSC		
Exposed Pad Length	D2	2.35	2.40	2.45
Overall Width	E	3.00 BSC		
Exposed Pad Width	E2	1.65	1.70	1.75
Terminal Width	b	0.18	0.23	0.28
Terminal Length	L	0.35	0.40	0.45
Terminal-to-Exposed-Pad	K	0.20	-	-

Notes:

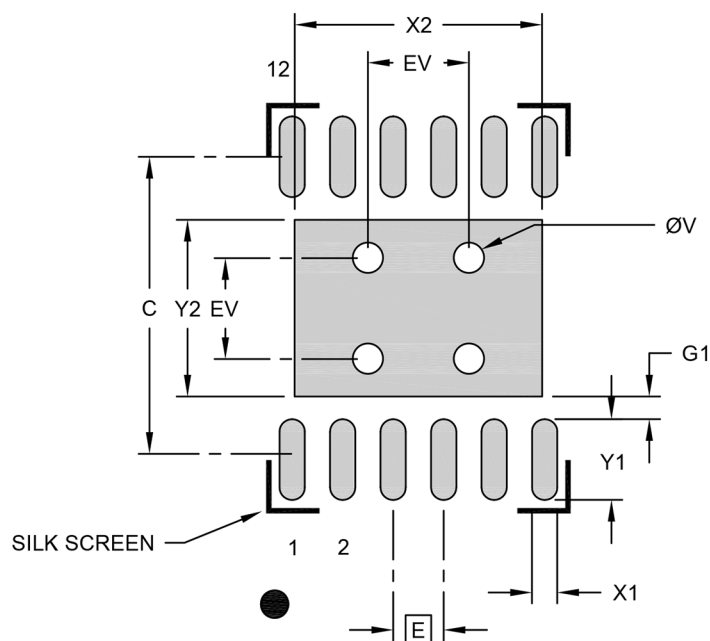
- Pin 1 visual index feature may vary, but must be located within the hatched area.
- Package is saw singulated
- Dimensioning and tolerancing per ASME Y14.5M
 - BSC: Basic Dimension. Theoretically exact value shown without tolerances.
 - REF: Reference Dimension, usually without tolerance, for information purposes only.

Microchip Technology Drawing C04-1020 Rev. A Sheet 2 of 2

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12-Lead Very Thin Plastic Dual Flat, No Lead Package (JQA) - 3x3 mm Body [VDFN] With 2.4x1.7 mm Exposed Pad; Micrel Legacy Package

Note: For the most current package drawings, please see the Microchip Packaging Specification located at <http://www.microchip.com/packaging>



RECOMMENDED LAND PATTERN

Dimension Limits	Units	MILLIMETERS		
		MIN	NOM	MAX
Contact Pitch	E	0.50 BSC		
Optional Center Pad Width	X2			2.45
Optional Center Pad Length	Y2			1.75
Contact Pad Spacing	C		3.00	
Contact Pad Width (X12)	X1			0.25
Contact Pad Length (X12)	Y1			0.80
Contact Pad to Center Pad (X12)	G1	0.20		
Thermal Via Diameter	V		0.30	
Thermal Via Pitch	EV		1.00	

Notes:

1. Dimensioning and tolerancing per ASME Y14.5M

BSC: Basic Dimension. Theoretically exact value shown without tolerances.

2. For best soldering results, thermal vias, if used, should be filled or tented to avoid solder loss during reflow process

Microchip Technology Drawing C04-3020 Rev. A

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7. Revision History

The revision history describes the changes that were implemented in the document. The changes are listed by revision, starting with the most current publication.

Revision	Date	Section	Description
DS20007173A	08/2026	—	Converted Micrel document MIC3230/1/2 to Microchip data sheet DS20007173A. Minor text changes throughout.

8. Product Identification System

PART NO.	X	XX	-XX
Device	Junction Temperature Range	Package	Media Type
Device:	MIC3231: Constant Current Boost Controller for Driving High Power LEDs		
Junction Temperature Range:	Y = -40°C to +125°C		
Package:	ML = 12-Lead VDFN TSE = 16-Lead TSSOP		
Media Type:	<blank> = 94/Tube (MIC3231 TSSOP only) TR = 2,500/Reel (MIC3231 TSSOP only) TR = 5,000/Reel (MIC3231 VDFN only)		

Examples:

- MIC3231YML-TR: MIC3231, -40°C to +125°C Temperature Range, 12-Lead VDFN, 5,000/Reel
- MIC3231YTSE: MIC3231, -40°C to +125°C Temperature Range, 16-Lead TSSOP, 94/Tube
- MIC3231YTSE-TR: MIC3231, -40°C to +125°C Temperature Range, 16-Lead TSSOP, 2,500/Reel

Note: Tape and Reel identifier only appears in the catalog part number description. This identifier is used for ordering purposes and is not printed on the device package. Check with your Microchip Sales Office for package availability with the Tape and Reel option.

Obsolete Part Variants:

- MIC3230YTSE
- MIC3230YTSE-TR
- MIC3230YML-TR
- MIC3232YMM
- MIC3232YMM-TR

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