![Texas Instruments]

LMR62014

SIMPLE SWITCHER® 20Vout, 1.4A Step-Up Voltage Regulator in SOT-23

Features

- Input voltage range of 2.7V to 14V
- Output voltage up to 20V
- Switch current up to 1.4A
- 1.6 MHz switching frequency
- Low shutdown Iq, <1 µA
- Cycle-by-cycle current limiting
- Internally compensated
- SOT23-5 packaging (2.92 x 2.84 x 1.08mm)
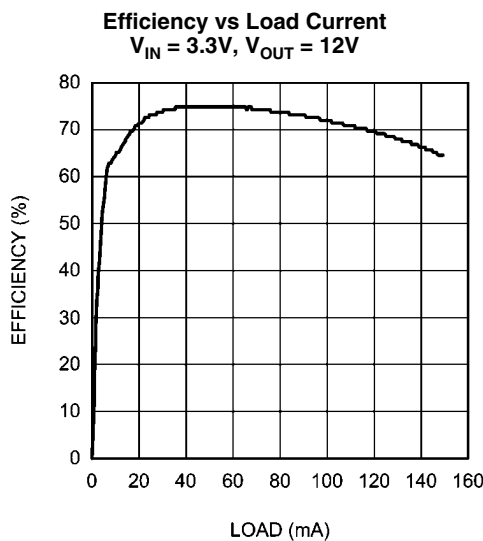- Fully enabled for WEBENCH® Power Designer

Performance Benefits

- Extremely easy to use
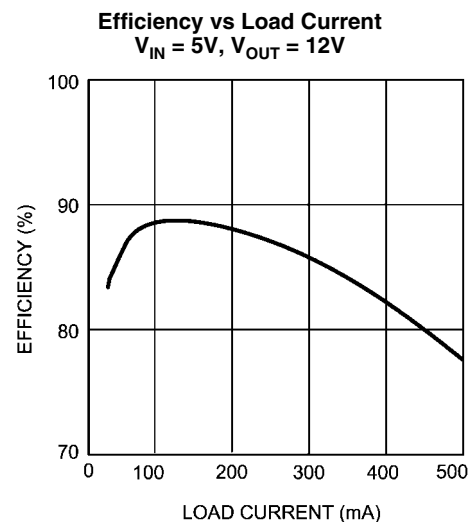- Tiny overall solution reduces system cost

Applications

- Boost Conversions from 3.3V, 5V, and 12V Rails
- Space Constrained Applications
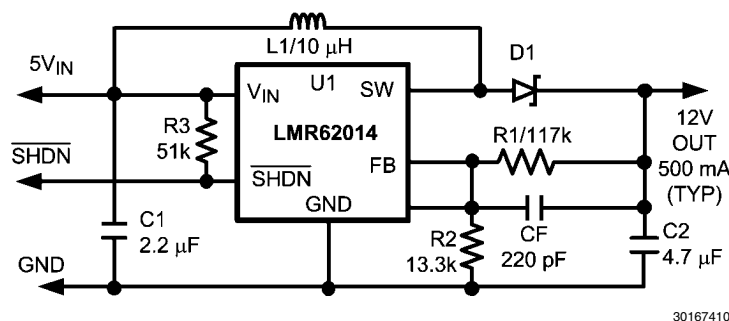- Embedded Systems
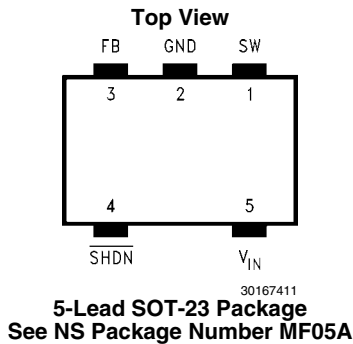- LCD Displays
- LED Applications

System Performance

Efficiency vs Load Current
V_{IN} = 3.3V, V_{OUT} = 12V

30167439

Efficiency vs Load Current
V_{IN} = 5V, V_{OUT} = 12V

30167457

30167410

Connection Diagram

Top View

30167411
5-Lead SOT-23 Package
See NS Package Number MF05A

Ordering Information

Order Number	Package Type	Package Drawing	Supplied As	Package ID
LMR62014XMFE			250 Units, Tape and Reel	SH1B
LMR62014XMF	SOT23-5	MF05A	1000 Units, Tape and Reel	
LMR62014XMFX			3000 Units, Tape and Reel	

Pin Descriptions

Pin	Name	Function
1	SW	Drain of the internal FET switch.
2	GND	Analog and power ground.
3	FB	Feedback point that connects to external resistive divider.
4	$\overline{\text{SHDN}}$	Shutdown control input. Connect to Vin if the feature is not used.
5	V_{IN}	Analog and power input.

Absolute Maximum Ratings (*Note 1*)

If Military/Aerospace specified devices are required, please contact the Texas Instruments Sales Office/ Distributors for availability and specifications.

Storage Temperature Range	−65°C to +150°C
Operating Junction Temperature Range	−40°C to +125°C
Lead Temp. (Soldering, 5 sec.)	300°C
Power Dissipation (*Note 2*)	Internally Limited
FB Pin Voltage	−0.4V to +6V

SW Pin Voltage	−0.4V to +22V
Input Supply Voltage	−0.4V to +14.5V
$\overline{\text{SHDN}}$ Pin Voltage	−0.4V to VIN + 0.3V
$\theta_{\text{J-A}}$ (SOT23-5)	265°C/W
ESD Rating (*Note 3*)	
Human Body Model	2 kV

For soldering specifications: see product folder at www.national.com and www.national.com/ms/MS/MS-SOLDERING.pdf

Electrical Characteristics

Limits in standard typeface are for T_J = 25°C, and limits in **boldface type** apply over the full operating temperature range (−40°C ≤ T_J ≤ +125°C). Unless otherwise specified: V_{IN} = 5V, V_{SHDN} = 5V, I_L = 0A.

Symbol	Parameter	Conditions		Min (*Note 4*)	Typical (*Note 5*)	Max (*Note 4*)	Units
V_{IN}	Input Voltage			**2.7**		**14**	V
$V_{OUT\ (MIN)}$	Minimum Output Voltage Under Load	R_L = 43Ω (*Note 8*)	V_{IN} = 2.7V	**5.4**	7		V
			V_{IN} = 3.3V	**8**	10		
			V_{IN} = 5V	**13**	17		
		R_L = 15Ω (*Note 8*)	V_{IN} = 2.7V	**3.75**	5		
			V_{IN} = 3.3V	**5**	6.5		
			V_{IN} = 5V	**8.75**	11		
I_{SW}	Switch Current Limit	(*Note 6*)		1.8 **1.4**	2		A
$R_{DS}(ON)$	Switch ON Resistance	I_{SW} = 100 mA Vin = 5V			260	400 **500**	mΩ
		I_{SW} = 100 mA Vin = 3.3V			300	450 **550**	
$SHDN_{TH}$	Shutdown Threshold	Device ON		**1.5**			V
		Device OFF				**0.50**	
I_{SHDN}	Shutdown Pin Bias Current	V_{SHDN} = 0			0		µA
		V_{SHDN} = 5V			0	**2**	
V_{FB}	Feedback Pin Reference Voltage	V_{IN} = 3V		**1.205**	1.230	**1.255**	V
I_{FB}	Feedback Pin Bias Current	V_{FB} = 1.23V			60	**500**	nA
I_Q	Quiescent Current	V_{SHDN} = 5V, Switching			2	**3.0**	mA
		V_{SHDN} = 5V, Not Switching			400	**500**	µA
		V_{SHDN} = 0			0.024	1	
$\dfrac{\Delta V_{FB}}{\Delta V_{IN}}$	FB Voltage Line Regulation	2.7V ≤ V_{IN} ≤ 14V			0.02		%/V
F_{SW}	Switching Frequency (*Note 7*)			**1**	1.6	**1.85**	MHz
D_{MAX}	Maximum Duty Cycle (*Note 7*)			**86**	93		%
I_L	Switch Leakage	Not Switching V_{SW} = 5V				1	µA

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the component may occur. Electrical specifications do not apply when operating the device outside of the limits set forth under the operating ratings which specify the intended range of operating conditions.

Note 2: The maximum power dissipation which can be safely dissipated for any application is a function of the maximum junction temperature, T_J(MAX) = 125°C, the junction-to-ambient thermal resistance for the SOT-23 package, $\theta_{\text{J-A}}$ = 265°C/W, and the ambient temperature, T_A. The maximum allowable power dissipation at any ambient temperature for designs using this device can be calculated using the formula:

$$P\ (MAX)\ =\ \frac{T_J\ (MAX)\ -\ T_A}{\theta_{J-A}}\ =\ \frac{125\ -\ T_A}{265}$$

If power dissipation exceeds the maximum specified above, the internal thermal protection circuitry will protect the device by reducing the output voltage as required to maintain a safe junction temperature.

Note 3: The human body model is a 100 pF capacitor discharged through a 1.5 kΩ resistor into each pin.

Note 4: Limits are guaranteed by testing, statistical correlation, or design.

Note 5: Typical values are derived from the mean value of a large quantity of samples tested during characterization and represent the most likely expected value of the parameter at room temperature.

Note 6: Switch current limit is dependent on duty cycle (see Typical Performance Characteristics).

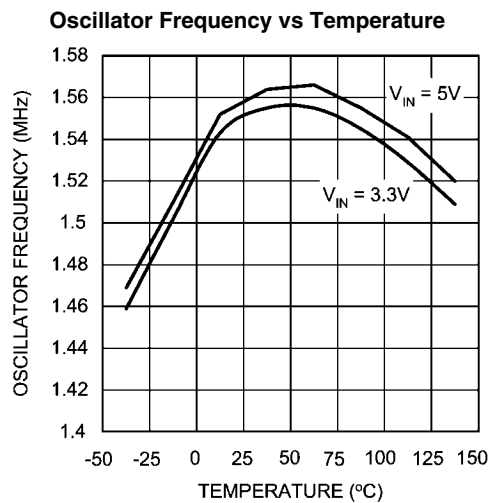**Note 7:** Guaranteed limits are the same for Vin = 3.3V input.

Note 8: L = 10 µH, C_{OUT} = 4.7 µF, duty cycle = maximum
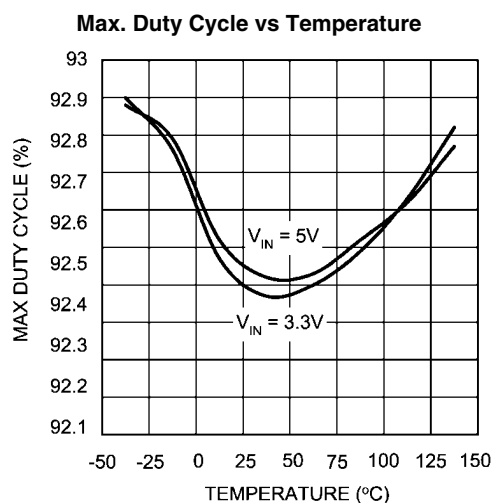
Typical Performance Characteristics Unless otherwise specified: V_{IN} = 5V, $\overline{SHDN}$ pin tied to V_{IN}.
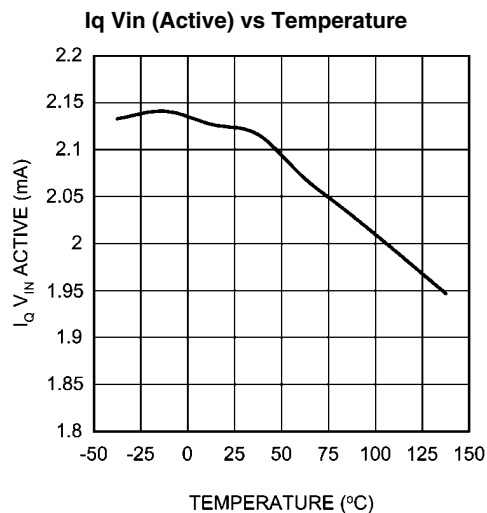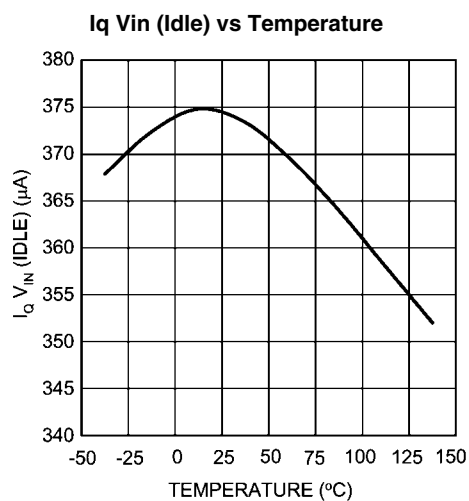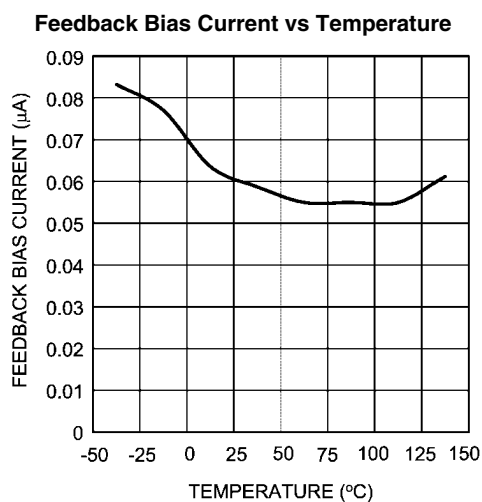
Iq Vin (Active) vs Temperature

30167402

Oscillator Frequency vs Temperature

30167405

Max. Duty Cycle vs Temperature

30167407

Iq Vin (Idle) vs Temperature

30167425

Feedback Bias Current vs Temperature

30167426

Feedback Voltage vs Temperature

30167427

R$_{DS}$(ON) vs Temperature

30167428

Current Limit vs Temperature

30167429

R$_{DS}$(ON) vs V$_{IN}$

30167452

Efficiency vs Load Current
V$_{IN}$ = 2.7V, V$_{OUT}$ = 5V

30167435

Efficiency vs Load Current
V$_{IN}$ = 3.3V, V$_{OUT}$ = 5V
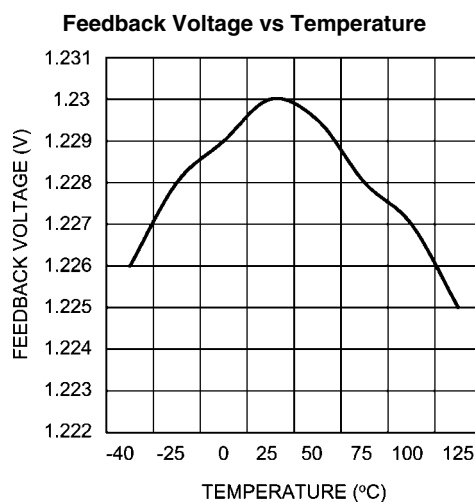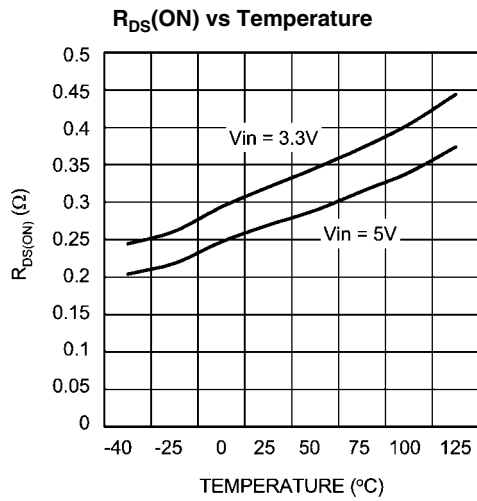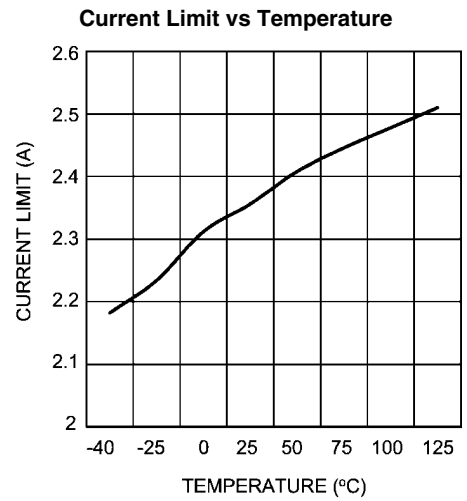
30167436
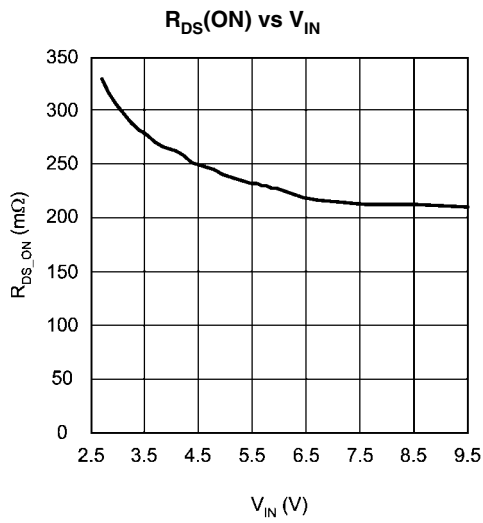
Efficiency vs Load Current
V$_{IN}$ = 4.2V, V$_{OUT}$ = 5V
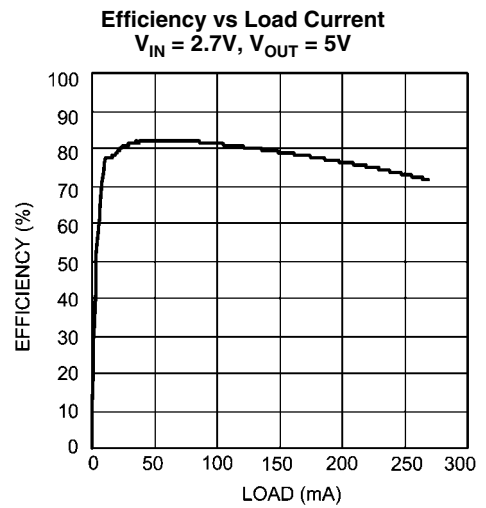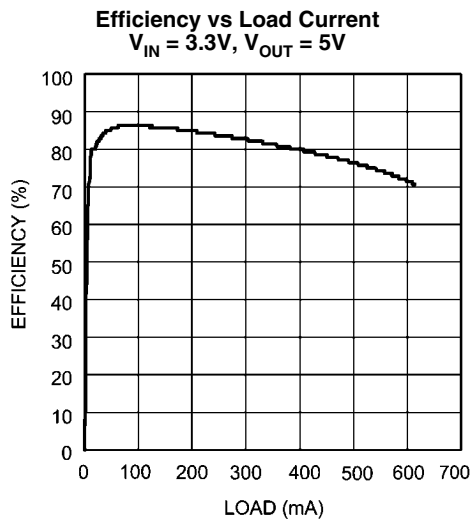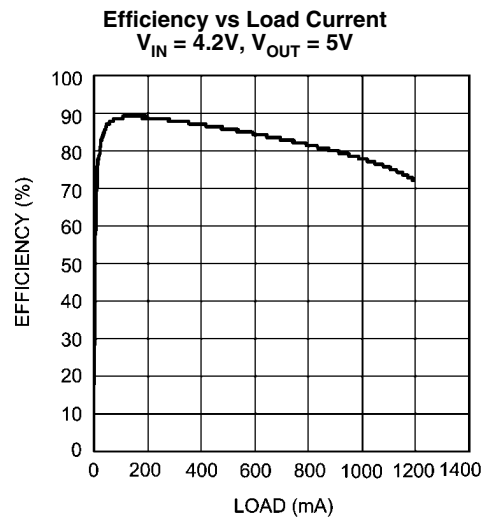
30167437

Efficiency vs Load Current
V_{IN} = 2.7V, V_{OUT} = 12V

Efficiency vs Load Current
V_{IN} = 2.7V, V_{OUT} = 12V

Efficiency vs Load Current
V_{IN} = 3.3V, V_{OUT} = 12V
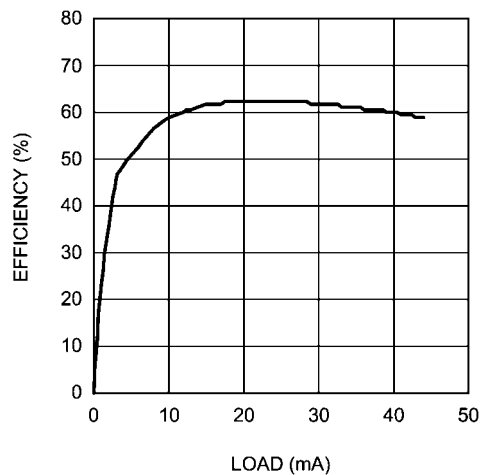
30167438
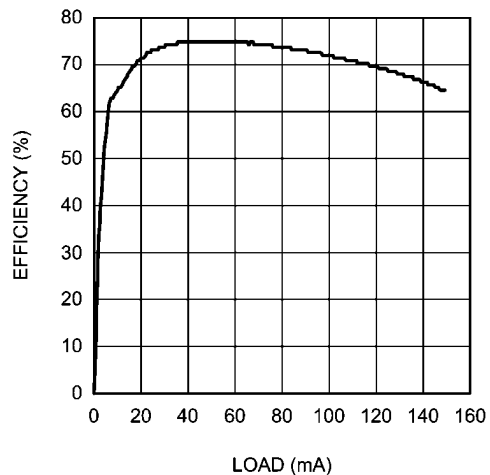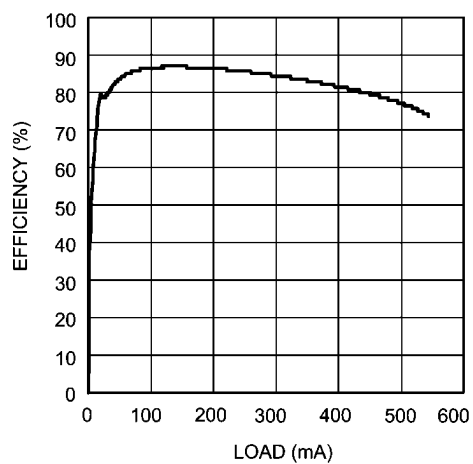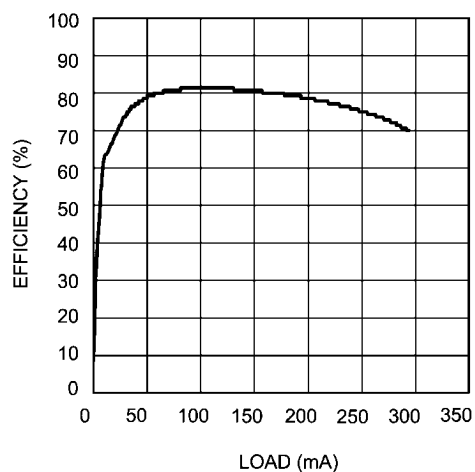
30167439

Efficiency vs Load Current
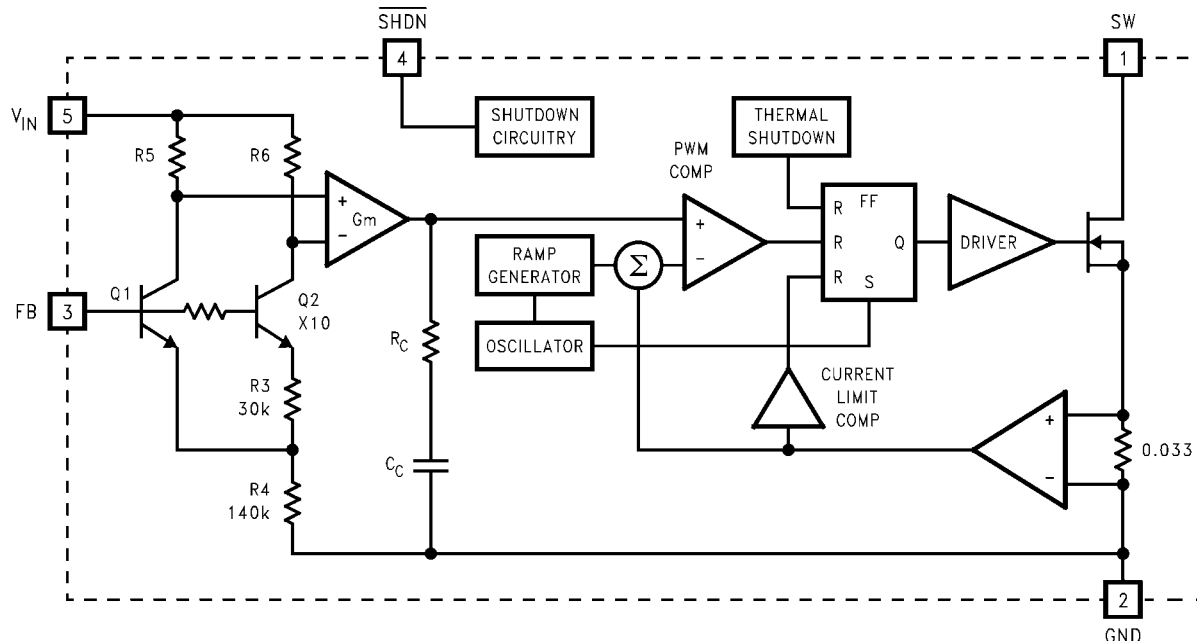V_{IN} = 5V, V_{OUT} = 12V

Efficiency vs Load Current
V_{IN} = 5V, V_{OUT} = 18V

30167440

30167441

Block Diagram

30167412

General Description

The LMR62014 switching regulator is a current-mode boost converter operating at fixed frequency of 1.6 MHz.

The use of SOT-23 package, made possible by the minimal power loss of the internal 1.4A switch, and use of small inductors and capacitors result in the industry's highest power density. The LMR62014 is capable of greater than 90% duty cycle, making it ideal for boosting to voltages up to 20V.

These parts have a logic-level shutdown pin that can be used to reduce quiescent current and extend battery life.

Protection is provided through cycle-by-cycle current limiting and thermal shutdown. Internal compensation simplifies design and reduces component count.

Theory of Operation

The LMR62014 is a switching converter IC that operates at a fixed frequency (1.6 MHz) for fast transient response over a wide input voltage range and incorporates pulse-by-pulse current limiting protection. Because this is current mode control, a 33 mΩ sense resistor in series with the switch FET is used to provide a voltage (which is proportional to the FET current) to both the input of the pulse width modulation (PWM) comparator and the current limit amplifier.

At the beginning of each cycle, the S-R latch turns on the FET. As the current through the FET increases, a voltage (proportional to this current) is summed with the ramp coming from the ramp generator and then fed into the input of the PWM comparator. When this voltage exceeds the voltage on the other input (coming from the Gm amplifier), the latch resets and turns the FET off. Since the signal coming from the Gm amplifier is derived from the feedback (which samples the voltage at the output), the action of the PWM comparator constantly sets the correct peak current through the FET to keep the output voltage in regulation.

Q1 and Q2 along with R3 - R6 form a bandgap voltage reference used by the IC to hold the output in regulation. The currents flowing through Q1 and Q2 will be equal, and the

feedback loop will adjust the regulated output to maintain this. Because of this, the regulated output is always maintained at a voltage level equal to the voltage at the FB node "multiplied up" by the ratio of the output resistive divider.

The current limit comparator feeds directly into the flip-flop that drives the switch FET. If the FET current reaches the limit threshold, the FET is turned off and the cycle terminated until the next clock pulse. The current limit input terminates the pulse regardless of the status of the output of the PWM comparator.

Application Hints

SELECTING THE EXTERNAL CAPACITORS

The best capacitors for use with the LMR62014 are multi-layer ceramic capacitors. They have the lowest ESR (equivalent series resistance) and highest resonance frequency which makes them optimum for use with high frequency switching converters.

When selecting a ceramic capacitor, only X5R and X7R dielectric types should be used. Other types such as Z5U and Y5F have such severe loss of capacitance due to effects of temperature variation and applied voltage, they may provide as little as 20% of rated capacitance in many typical applications. Always consult capacitor manufacturer's data curves before selecting a capacitor.

SELECTING THE OUTPUT CAPACITOR

A single ceramic capacitor of value 4.7 µF to 10 µF will provide sufficient output capacitance for most applications. If larger amounts of capacitance are desired for improved line support and transient response, tantalum capacitors can be used. Aluminum electrolytics with ultra low ESR such as Sanyo Oscon can be used, but are usually prohibitively expensive. Typical Al electrolytic capacitors are not suitable for switching frequencies above 500 kHz due to significant ringing and temperature rise due to self-heating from ripple current. An output capacitor with excessive ESR can also reduce phase margin and cause instability.

In general, if electrolytics are used, it is recommended that they be paralleled with ceramic capacitors to reduce ringing, switching losses, and output voltage ripple.

SELECTING THE INPUT CAPACITOR

An input capacitor is required to serve as an energy reservoir for the current which must flow into the coil each time the switch turns ON. This capacitor must have extremely low ESR, so ceramic is the best choice. We recommend a nominal value of 2.2 µF, but larger values can be used. Since this capacitor reduces the amount of voltage ripple seen at the input pin, it also reduces the amount of EMI passed back along that line to other circuitry.

FEED-FORWARD COMPENSATION

Although internally compensated, the feed-forward capacitor Cf is required for stability (see *Basic Application Circuit*). Adding this capacitor puts a zero in the loop response of the converter. The recommended frequency for the zero fz should be approximately 6 kHz. Cf can be calculated using the formula:

$$Cf = 1 / (2 \times \pi \times R1 \times fz)$$

SELECTING DIODES

The external diode used in the typical application should be a Schottky diode. The diode must be rated to handle the maximum output voltage and load current. A 20V diode such as the MBR0520 is recommended.

The MBR05XX series of diodes are designed to handle a maximum average current of 0.5A. For applications exceeding 0.5A average, a Toshiba CRS08 can be used.

LAYOUT HINTS

High frequency switching regulators require very careful layout of components in order to get stable operation and low noise. All components must be as close as possible to the LMR62014 device. It is recommended that a 4-layer PCB be used so that internal ground planes are available.

As an example, a recommended layout of components is shown:
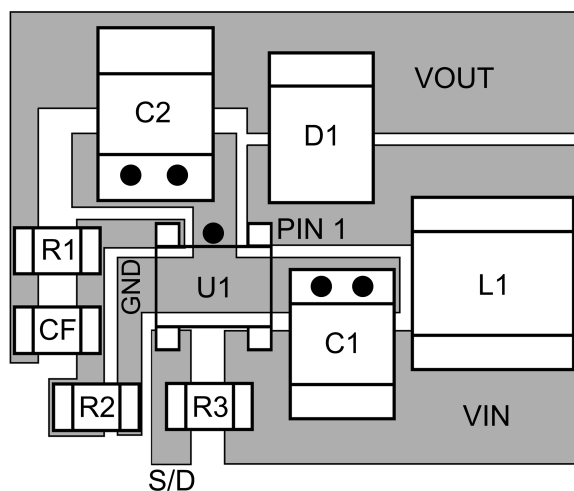
● VIAS TO GROUND PLANE

**Recommended PCB Component Layout**   30167416

Some additional guidelines to be observed:

1. Keep the path between L1, D1, and C2 extremely short. Parasitic trace inductance in series with D1 and C2 will increase noise and ringing.

2. The feedback components R1, R2 and CF must be kept close to the FB pin of U1 to prevent noise injection on the FB pin trace.

3. If internal ground planes are available (recommended) use vias to connect directly to ground at pin 2 of U1, as well as the negative sides of capacitors C1 and C2.

SETTING THE OUTPUT VOLTAGE

The output voltage is set using the external resistors R1 and R2 (see *Basic Application Circuit*). A value of approximately 13.3 kΩ is recommended for R2 to establish a divider current of approximately 92 µA. R1 is calculated using the formula:

$$R1 = R2 \times (V_{OUT}/1.23 - 1)$$
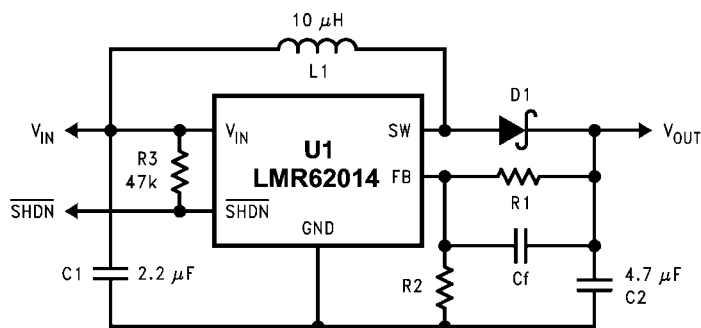
Basic Application Circuit

30167417

DUTY CYCLE

The maximum duty cycle of the switching regulator determines the maximum boost ratio of output-to-input voltage that the converter can attain in continuous mode of operation. The duty cycle for a given boost application is defined as:

$$\text{Duty Cycle} = \frac{V_{OUT} + V_{DIODE} - V_{IN}}{V_{OUT} + V_{DIODE} - V_{SW}}$$

This applies for continuous mode operation.

INDUCTANCE VALUE

The first question we are usually asked is: "How small can I make the inductor?" (because they are the largest sized component and usually the most costly). The answer is not simple and involves trade-offs in performance. Larger inductors mean less inductor ripple current, which typically means less output voltage ripple (for a given size of output capacitor). Larger inductors also mean more load power can be delivered because the energy stored during each switching cycle is:

$$E = L/2 \times (Ip)^2$$

Where "Ip" is the peak inductor current. An important point to observe is that the LMR62014 will limit its switch current based on peak current. This means that since Ip(max) is fixed, increasing L will increase the maximum amount of power available to the load. Conversely, using too little inductance may limit the amount of load current which can be drawn from the output.

Best performance is usually obtained when the converter is operated in "continuous" mode at the load current range of interest, typically giving better load regulation and less output ripple. Continuous operation is defined as not allowing the inductor current to drop to zero during the cycle. It should be noted that all boost converters shift over to discontinuous operation as the output load is reduced far enough, but a larger inductor stays "continuous" over a wider load current range.

To better understand these trade-offs, a typical application circuit (5V to 12V boost with a 10 μH inductor) will be analyzed. We will assume:

$V_{IN} = 5V$, $V_{OUT} = 12V$, $V_{DIODE} = 0.5V$, $V_{SW} = 0.5V$

Since the frequency is 1.6 MHz (nominal), the period is approximately 0.625 μs. The duty cycle will be 62.5%, which means the ON time of the switch is 0.390 μs. It should be noted that when the switch is ON, the voltage across the inductor is approximately 4.5V.

Using the equation:

$$V = L \ (di/dt)$$

We can then calculate the di/dt rate of the inductor which is found to be 0.45 A/μs during the ON time. Using these facts, we can then show what the inductor current will look like during operation:
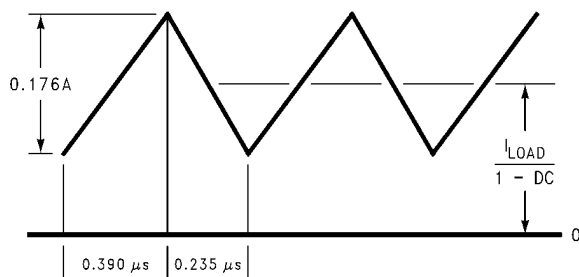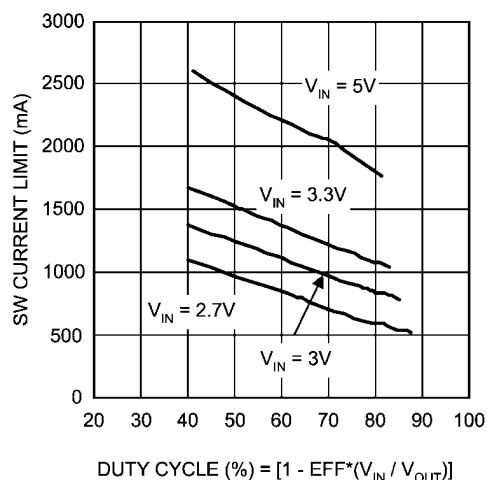

**10 μH Inductor Current,
5V–12V Boost (LMR62014X)**

30167418

During the 0.390 μs ON time, the inductor current ramps up 0.176A and ramps down an equal amount during the OFF time. This is defined as the inductor "ripple current". It can also be seen that if the load current drops to about 33 mA, the inductor current will begin touching the zero axis which means it will be in discontinuous mode. A similar analysis can be performed on any boost converter, to make sure the ripple current is reasonable and continuous operation will be maintained at the typical load current values.

MAXIMUM SWITCH CURRENT

The maximum FET switch current available before the current limiter cuts in is dependent on duty cycle of the application. This is illustrated in the graphs below which show typical values of switch current as a function of effective (actual) duty cycle:

DUTY CYCLE (%) = [1 - EFF*(V$_{IN}$ / V$_{OUT}$)]

30167450

Switch Current Limit vs Duty Cycle

V$_{IN}$ (V)

30167448

Max. Load Current (typ) vs V$_{IN}$

CALCULATING LOAD CURRENT

As shown in the figure which depicts inductor current, the load current is related to the average inductor current by the relation:

$$I_{LOAD} = I_{IND}(AVG) \times (1 - DC)$$

Where "DC" is the duty cycle of the application. The switch current can be found by:

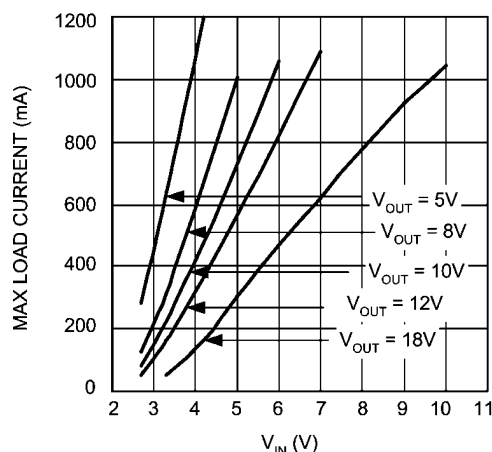$$I_{SW} = I_{IND}(AVG) + \tfrac{1}{2}(I_{RIPPLE})$$

Inductor ripple current is dependent on inductance, duty cycle, input voltage and frequency:

$$I_{RIPPLE} = DC \times (V_{IN} - V_{SW}) / (f \times L)$$

combining all terms, we can develop an expression which allows the maximum available load current to be calculated:

$$I_{LOAD}(max) = (1 - DC) \times \left(I_{SW}(max) - \frac{DC(V_{IN} - V_{SW})}{2fL} \right)$$

The equation shown to calculate maximum load current takes into account the losses in the inductor or turn-OFF switching losses of the FET and diode. For actual load current in typical applications, we took bench data for various input and output voltages that displayed the maximum load current available for a typical device in graph form:

DESIGN PARAMETERS V$_{SW}$ AND I$_{SW}$

The value of the FET "ON" voltage (referred to as V$_{SW}$ in the equations) is dependent on load current. A good approximation can be obtained by multiplying the "ON Resistance" of the FET times the average inductor current.

FET on resistance increases at V$_{IN}$ values below 5V, since the internal N-FET has less gate voltage in this input voltage range (see Typical performance Characteristics curves). Above V$_{IN}$ = 5V, the FET gate voltage is internally clamped to 5V.

The maximum peak switch current the device can deliver is dependent on duty cycle. For higher duty cycles, see Typical performance Characteristics curves.

THERMAL CONSIDERATIONS

At higher duty cycles, the increased ON time of the FET means the maximum output current will be determined by power dissipation within the LMR62014 FET switch. The switch power dissipation from ON-state conduction is calculated by:

$$P_{(SW)} = DC \times I_{IND}(AVE)^2 \times R_{DS}(ON)$$

There will be some switching losses as well, so some derating needs to be applied when calculating IC power dissipation.

INDUCTOR SUPPLIERS

Recommended suppliers of inductors for this product include, but are not limited to Sumida, Coilcraft, Panasonic, TDK and Murata. When selecting an inductor, make certain that the continuous current rating is high enough to avoid saturation at peak currents. A suitable core type must be used to minimize core (switching) losses, and wire power losses must be considered when selecting the current rating.

## SHUTDOWN PIN OPERATION

The device is turned off by pulling the shutdown pin low. If this function is not going to be used, the pin should be tied directly to V$_{IN}$. If the SHDN function will be needed, a pull-up resistor must be used to V$_{IN}$ (approximately 50k-100kΩ recommended). The SHDN pin must not be left unterminated.

Efficiency vs Load Current
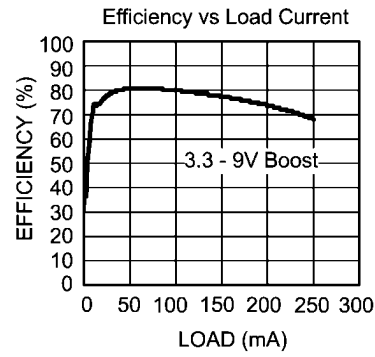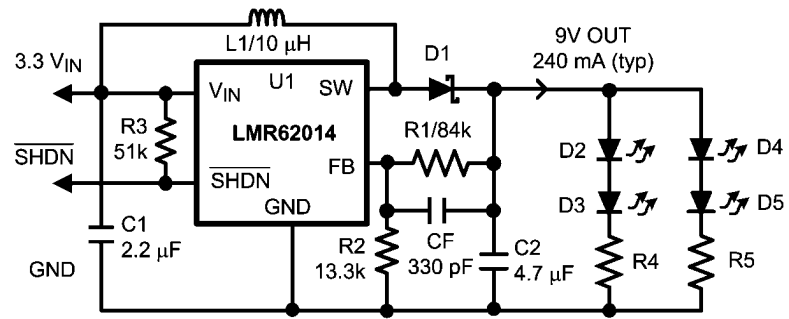
FIGURE 1. Flash LED Application

30167456

Physical Dimensions inches (millimeters) unless otherwise noted

LAND PATTERN RECOMMENDATION
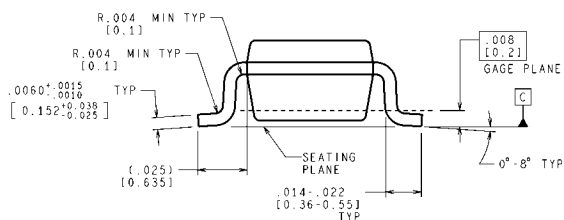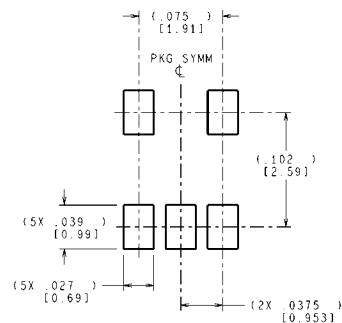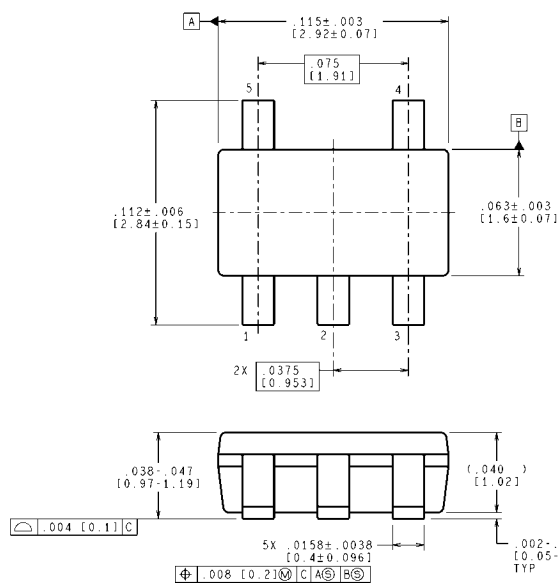
CONTROLLING DIMENSION IS INCH
VALUES IN [ ] ARE MILLIMETERS
DIMENSIONS IN ( ) FOR REFERENCE ONLY

MF05A (Rev D)

5-Lead SOT-23 Package
Order Number LMR62014XMF, LMR62014XMFX
NS Package Number MF05A

Notes

TI/NATIONAL INTERIM IMPORTANT NOTICE

Texas Instruments has purchased National Semiconductor. As of Monday, September 26th, and until further notice, products sold or advertised under the National Semiconductor name or logo, and information, support and interactions concerning such products, remain subject to the preexisting National Semiconductor standard terms and conditions of sale, terms of use of website, and Notices (and/or terms previously agreed in writing with National Semiconductor, where applicable) and are not subject to any differing terms and notices applicable to other TI components, sales or websites. To the extent information on official TI and National websites and business social networking media, etc., pertains to both TI and National-branded products, both companies' instructions, warnings and limitations in the above-referenced terms of use apply.

Following are URLs where you can obtain information on other Texas Instruments products and application solutions:

Products		Applications	
Audio	www.ti.com/audio	Communications and Telecom	www.ti.com/communications
Amplifiers	amplifier.ti.com	Computers and Peripherals	www.ti.com/computers
Data Converters	dataconverter.ti.com	Consumer Electronics	www.ti.com/consumer-apps
DLP® Products	www.dlp.com	Energy and Lighting	www.ti.com/energy
DSP	dsp.ti.com	Industrial	www.ti.com/industrial
Clocks and Timers	www.ti.com/clocks	Medical	www.ti.com/medical
Interface	interface.ti.com	Security	www.ti.com/security
Logic	logic.ti.com	Space, Avionics and Defense	www.ti.com/space-avionics-defense
Power Mgmt	power.ti.com	Transportation and Automotive	www.ti.com/automotive
Microcontrollers	microcontroller.ti.com	Video and Imaging	www.ti.com/video
RFID	www.ti-rfid.com	Wireless	www.ti.com/wireless-apps
RF/IF and ZigBee® Solutions	www.ti.com/lprf	**TI E2E Community Home Page**	e2e.ti.com

Mailing Address: Texas Instruments, Post Office Box 655303, Dallas, Texas 75265

Copyright© 2011 Texas Instruments Incorporated