

3A 1.5MHz Synchronous Rectified Step Down Converter

General Description

The LN2431 is a monolithic synchronous high-efficiency DC/DC buck converter delivers up to 3A of output current. The device operates from an input voltage of 2.5V to 8 V and provides an output voltage from 0.8V to VIN, making the LN2431 ideal for on-board post-regulation applications. The LN2431 operate at a fixed frequency of 1.5MHz with an efficiency of up to 94%. The high operating frequency minimizes the size of external components.

Internal soft-start control circuitry reduces inrush current. Short-circuit and thermal-overload protections improve design reliability.

The LN2431 are available in a space-saving 8-pin SO/PP package.

Features

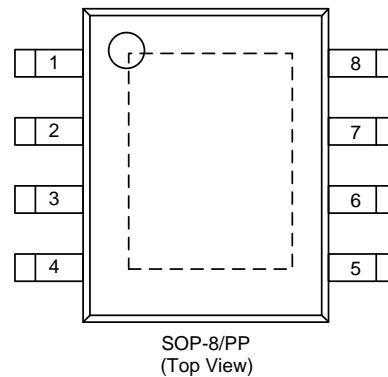
- Ceramic Input and Output Capacitors
- Efficiency Up to 94%
- Guaranteed 3A Output Current
- Operate from 2.5V to 8V Supply
- Adjustable Output from 0.8V to VIN
- Internal Soft-Start
- Short-Circuit and Thermal-Overload Protection
- ROHS Compliant

Applications

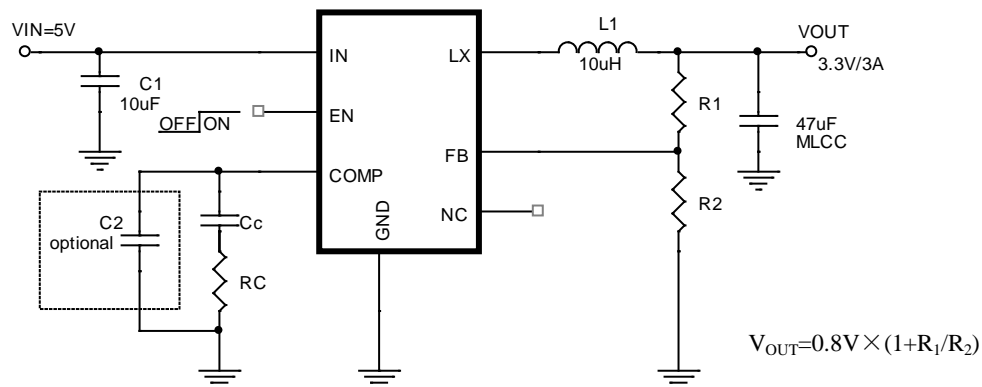
- FPGA,ASIC,DSP POWER SUPPLICES
- LCD TV
- Green Electronics/Appliances
- Notebook Computers
- Set-Top box
- Cellular Base Stations
- Networking and Telecommunications

Package

- SOP-8/PP



Typical Application Circuit



NOTE: FREQ pin is not connected :fosc=340KHz, RC=1K、CC=2.2nF ;If choose C2,C2=100PF

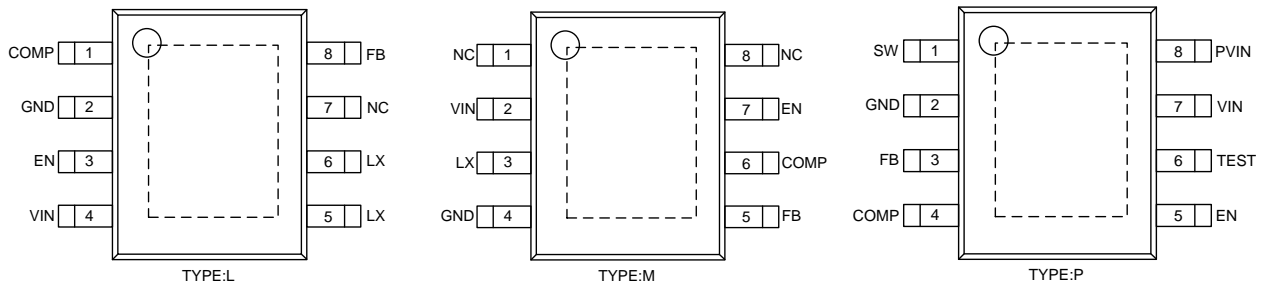
FREQ pin connect to a resistor=27KΩ :fosc=1.5MHz,RC=10K、CC=470PF;If choose C2,C2=22PF

Ordering Rule

LN2431 AB C D E XXXX

Designator	Represents	Symbol	Description
AB	Output Voltage	AD	Adjustable
		12	1.2V
		18	1.8V
		33	3.3V
C	Frequency	M	1.5MHZ
D	Package	S	SOP-8/PP
E	Pin assignment	L	
		M	
		P	
XXXX	Lot Number	Integer	

Pin Configuration

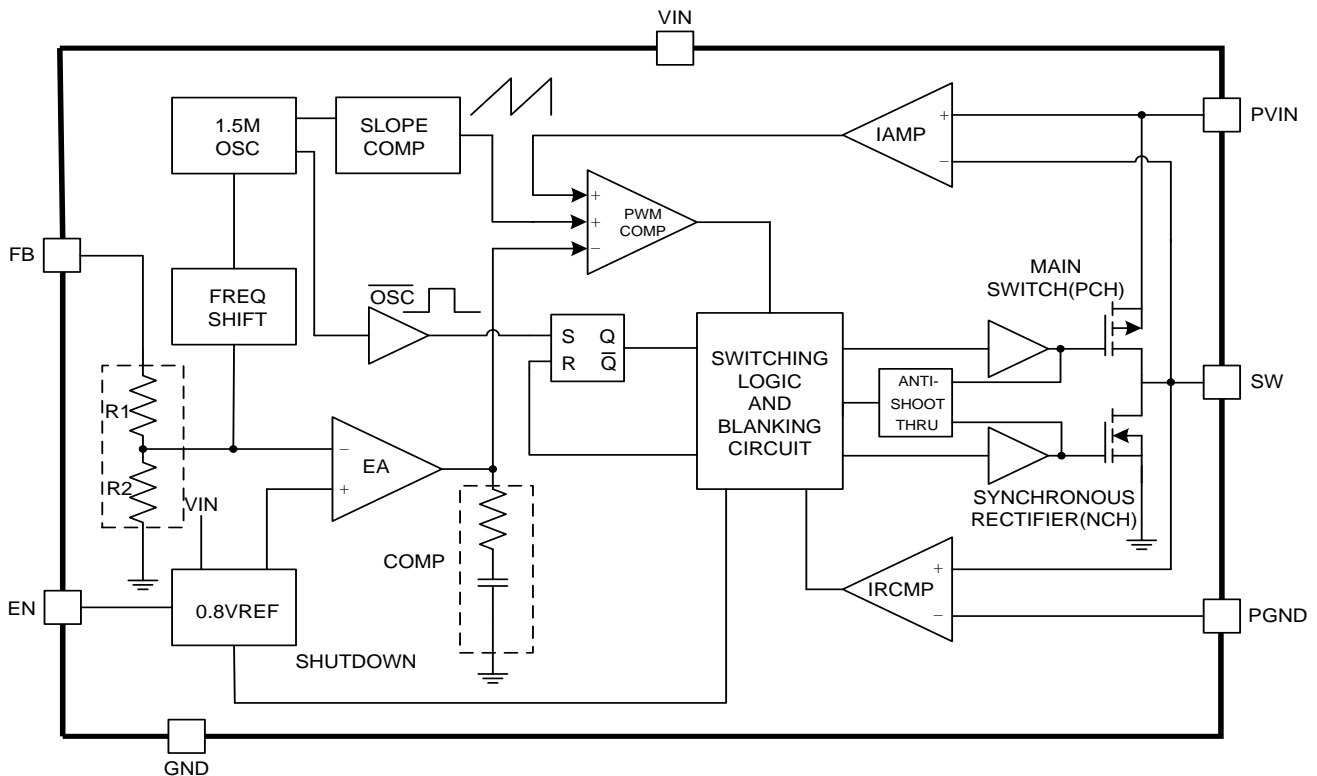


	L	M	P
Pin1	COMP	NC	SW
Pin2	GND	VIN	GND
Pin3	EN	LX	FB
Pin4	VIN	GND	COMP
Pin5	LX	FB	EN
Pin6	LX	COMP	TEST
Pin7	NC	EN	VIN
Pin8	FB	NC	PVIN

Functional Pin Description

Name	Function
COMP	Loop Compensation Pin. Add R/C network to stabilize the loop.
GND	Ground Pin.
EN	Chip Enable pin. Active high. Internal pull high for auto start up.
IN	Input Power Supply Pin.
LX	Switch Pin.
NC	Not Connect.
FB	Feedback Pin. $V_{OUT}=0.8V \times (1+R1/R2)$.
NC	Not connect.

Function Block Diagram



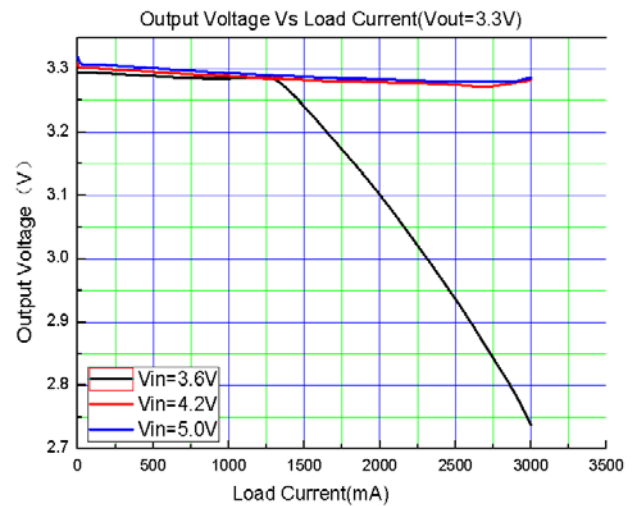
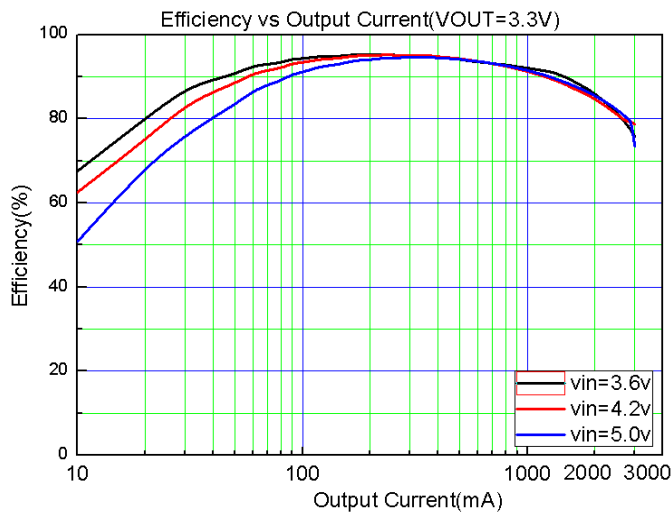
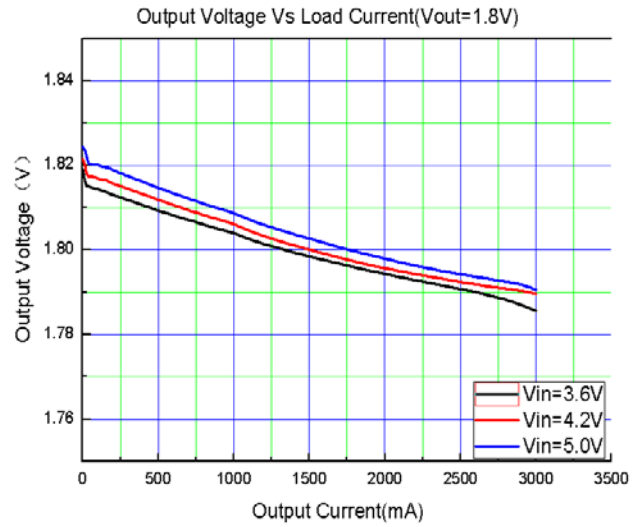
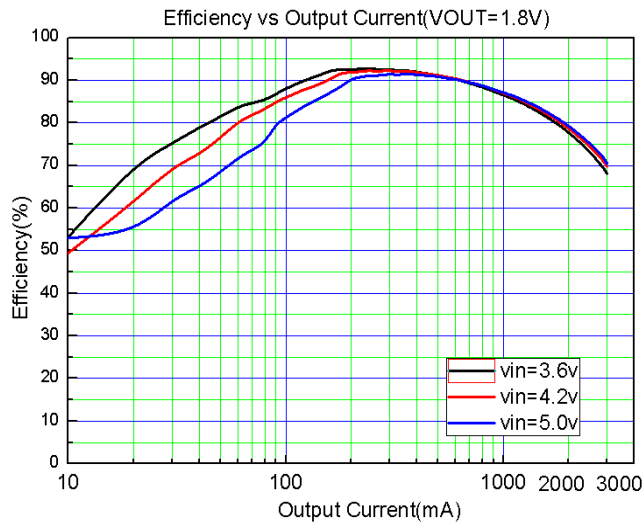
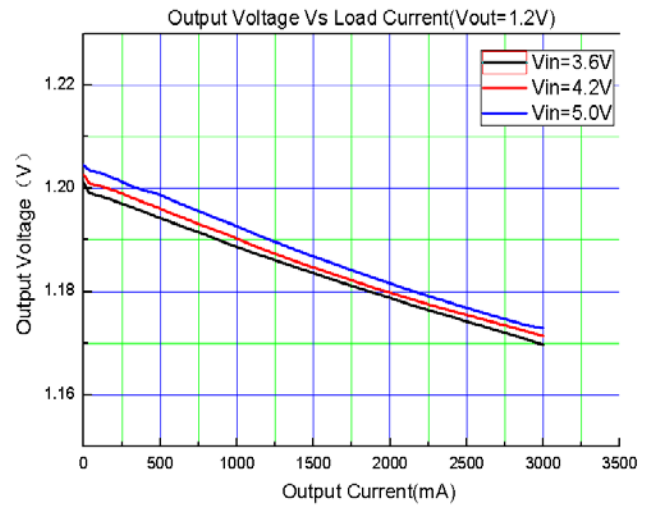
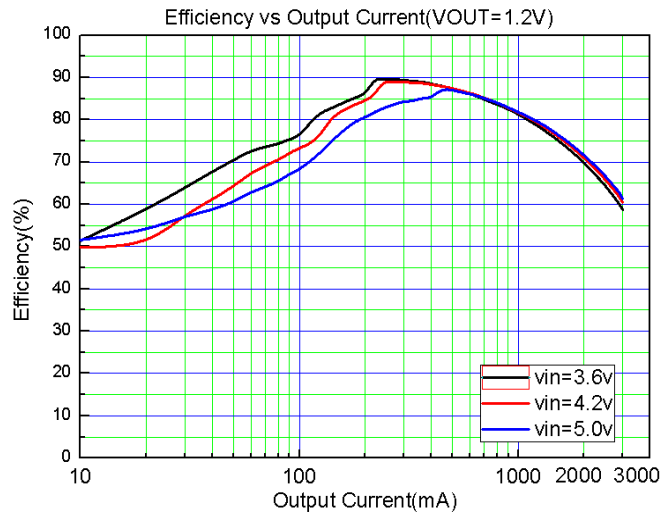
Absolute Maximum Ratings

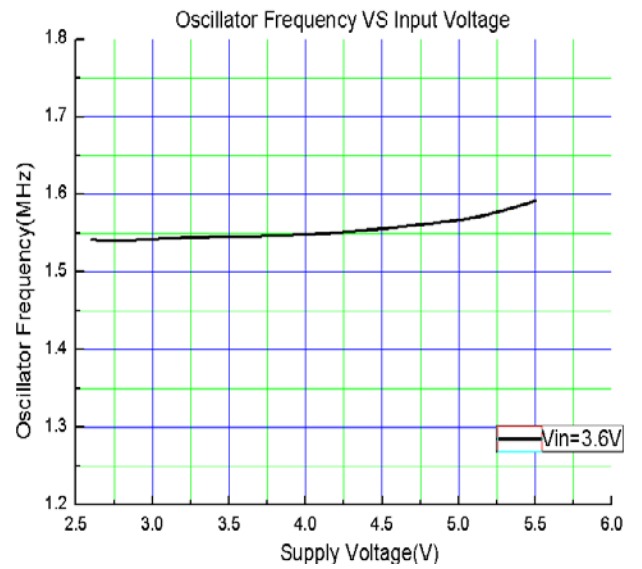
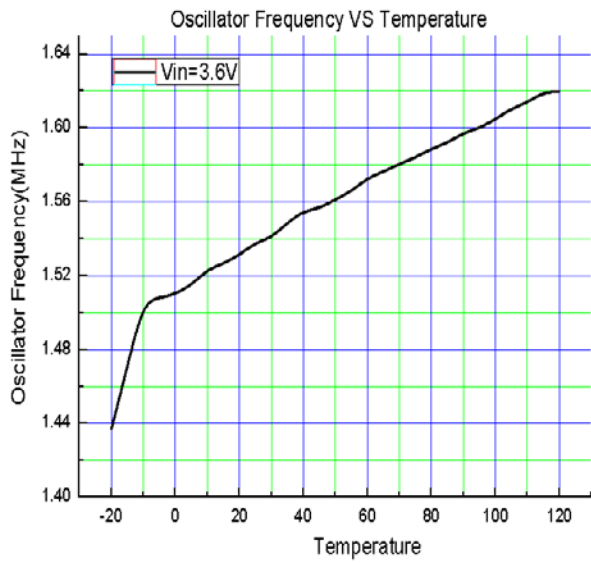
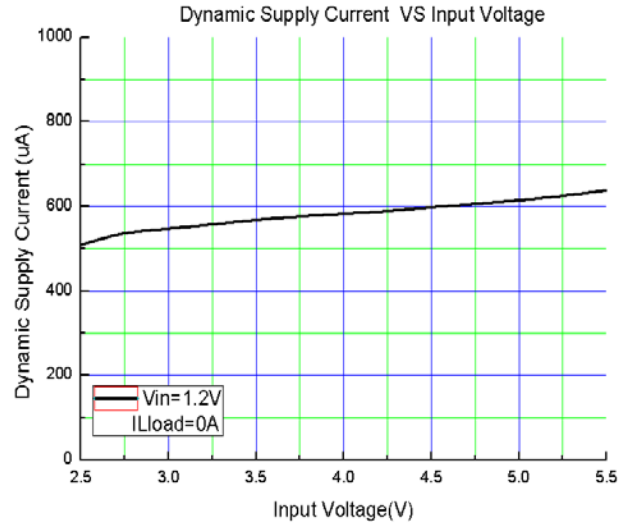
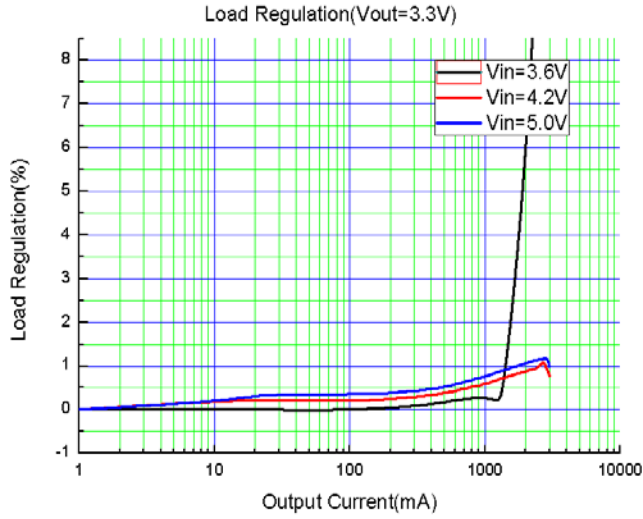
Parameter	Symbol	Maximum Rating	Unit
Input Voltage	V_{IN}	$V_{SS}-0.3 \sim V_{SS}+8$	V
	V_{LX}	$V_{SS}-0.3 \sim V_{IN}+0.7$	
	$V_{EN,FB,COMP}$	$V_{SS}-0.3 \sim V_{IN}+0.3$	
Power Dissipation	P_D	SOP8/PP 1400	mW
Operating Ambient Temperature	T_{opr}	-40~+85	°C
Storage Temperature	T_{stg}	-40~+125	
Reflow Temperature(soldeing,10s)	T_{refl}	250	

Electrical Characteristics

Parameter	Conditions	Min	Typ	Max	Units
Input Voltage Range		3		8	V
Input UVLO		2	2.3	2.6	V
Quiescent Current	$V_{FB} = 1V$ (no switching)	—	460	550	μA
	$V_{EN} = 0V$	—	0	+1	μA
FB Pin Voltage		0.784	0.8	0.816	V
FB Pin Current		-50	0	+50	nA
Load Regulation	$0A < I_{OUT} < 3A$		0.3		%
Line Regulation	$3.3V < V_{IN} < 5V$		0.17		%
EN Pin Voltage High		0.9			V
EN Pin Voltage Low				0.75	V
EN Pin Leakage Current	$V_{EN}=3V$		0.1	1	μA
Switching Frequency		1275	1500	1725	KHz
Short-Circuit Frequency			1/4		FSW
Current Limit		3.5	4.0	4.5	A
Maximum Duty				100	%
Minimum Duty		0			%
Minimum On Time			180		nS
Error Amp Transconductance		300	400	500	μmho
P-Switch Leakage Current	$V_{LX} = 0V, V_{EN} = 0V$		0.1	20	μA
P-Switch RDS(ON)	$I_{LX} = 150mA$		130	200	m Ω
N-Switch RDS(ON)	$I_{LX} = 150mA$		110	180	m Ω
Thermal Shutdown Protection	Rising		160		$^{\circ}C$
	Hysteresis		-20		$^{\circ}C$

Typical Performance Characteristic





Application Information

Inductor Selection

For most applications, the value of the inductor will fall in the range of 4.7μH to 22μH. Its value is chosen based on the desired ripple current. Large value inductors lower ripple current and small value inductors result in higher ripple currents. Higher VIN or VOUT also increase the ripple current ΔIL:

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

$$C_{IN} \text{ requires } I_{RMS} \cong I_{QMAX} \frac{\sqrt{V_{OUT}(V_{IN} - V_{OUT})}}{V_{IN}}$$

$$\Delta V_{OUT} \cong \Delta I_L \left(ESR + \frac{1}{8fC_{OUT}} \right)$$

$$V_{OUT} = 0.8 \times \left[1 + \frac{R1}{R2} \right] \text{ Volt}$$

where f=switching frequency, L=inductance. A reasonable inductor current ripple is usually set as 1/3 to 1/10 of maximum out current.

The DC current rating of the inductor should be at least equal to the maximum load current plus half the ripple current to prevent core saturation. For better efficiency, choose a low DCR inductor.

Capacitor Selection

In continuous mode, the source current of the top MOSFET is a square wave of duty cycle VOUT/VIN. To prevent large voltage transients, a low ESR input capacitor sized for maximum RMS current must be used. The maximum RMS capacitor current is given by:

$$C_{IN} \text{ requires } I_{RMS} \cong I_{QMAX} \frac{\sqrt{V_{OUT}(V_{IN} - V_{OUT})}}{V_{IN}}$$

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

This formula has a maximum at VIN=2VOUT, where IRMS=IOUT/2. This simple worst case condition is commonly used for design because even significant deviations do not offer much relief.

The selection of COUT is driven by the required effective series resistance (ESR). Typically, once the ESR requirement for COUT has been met, the RMS current rating generally far exceeds the IRIPPLE(P-P) requirement.

The output ripple ΔVOUT is determined by:

$$\Delta V_{OUT} \cong \Delta I_L \left(ESR + \frac{1}{8fC_{OUT}} \right)$$

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

For a fixed output voltage, the output ripple is highest at maximum input voltage since ΔIL increases with input voltage.

Nowadays, higher value, lower cost ceramic capacitors are becoming available in smaller case sizes. Their high ripple current, high voltage rating and low ESR make them ideal for switching regulator applications. Because the LN2431's control loop does not depend on the output capacitor's ESR for stable operation, ceramic capacitors can be used freely to achieve very low output ripple and small circuit size.

When choosing the input and output ceramic capacitors, choose the X5R or X7R dielectric formulations. These dielectrics have the best temperature and voltage characteristics of all the ceramics for given value and size.

Output Voltage Programming

The output voltage of the LN2431 is set by a resistive divider according to the following formula:

$$V_{OUT} = 0.8 \times \left[1 + \frac{R1}{R2} \right] \text{ Volt}$$

Some standard value of R1, R2 for most commonly used output voltage values are listed in Table 1.

V _{OUT} (V)	R ₁ (kΩ)	R ₂ (kΩ)
1.1	7.5	20
1.2	10	20
1.5	13	15
1.8	30	24
2.5	51	24
3.3	75	24

Loop Compensation

The LN2431 employs peak current mode control for easy use and fast transient response. Peak current mode control eliminates the double pole effect of the output L-C filter. It greatly simplifies the compensation loop design.

With peak current mode control, the buck power stage can be simplified to be a one-pole and one-zero system in frequency domain. The pole can be calculated by:

$$f_{p1} = \frac{1}{2\pi \times C_{OUT} \times R_L}$$

$$f_{z1} = \frac{1}{2\pi \times C_{OUT} \times ESR_{COUT}}$$

$$f_{p2} = \frac{1}{2\pi \times C_C \times (R_C + \frac{A_{EA}}{G_{EA}})} \cong \frac{G_{EA}}{2\pi \times C_C \times A_{EA}}$$

$$f_{z2} = \frac{1}{2\pi \times C_C \times R_C}$$

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

$$C_C = \frac{C_{OUT} \times R_L}{R_C}$$

$$R_{LX} = (R_{DS(ON)})D + (R_D(F))(1-D)$$

The zero is a ESR zero due to output capacitor and its ESR. It can be calculated by:

$$f_{z1} = \frac{1}{2\pi \times C_{OUT} \times ESR_{COUT}}$$

Where COUT is the output capacitor; RL is load resistance; ESR_{COUT} is the equivalent series resistance of output capacitor.

The compensation design is to shape the converter close loop transfer function to get desired gain and phase. For most cases, a series capacitor and resistor network connected to the COMP pin sets the pole-zero and is adequate for a stable high-bandwidth control loop.

In the LN2431, FB pin and COMP pin are the inverting input and the output of internal transconductance error amplifier (EA). A series RC and CC compensation network connected to COMP pin provides one pole and one zero: for $R_C \ll A_{EA}/G_{EA}$

$$f_{p2} = \frac{1}{2\pi \times C_C \times (R_C + \frac{A_{EA}}{G_{EA}})} \cong \frac{G_{EA}}{2\pi \times C_C \times A_{EA}}$$

$$f_{z2} = \frac{1}{2\pi \times C_C \times R_C}$$

where GEA is the error amplifier transconductance
AEA is the error amplifier voltage gain
RC is the compensation resistor
CC is the compensation capacitor

The desired crossover frequency f_C of the system is defined to be the frequency where the control loop has unity gain. It is also called the bandwidth of the converter. In general, a higher bandwidth means faster response to load transient. However, the bandwidth should not be too high because of system stability concern. When designing the compensation loop, converter stability under all line and load condition must be considered. Usually, it is recommended to set the bandwidth to be less than 1/10 of switching frequency.

Using selected crossover frequency, f_C , to calculate RC:

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

where $G_{CS} = 2 A/V$ is the current sense circuit transconductance.

The compensation capacitor CC and resistor RC together make zero. This zero is put somewhere close to the pole f_{P1} of selected frequency. CC is selected by:

$$C_C = \frac{C_{OUT} \times R_L}{R_C}$$

Checking Transient Response

The regulator loop response can be checked by looking at the load transient response. Switching regulators take several cycles to respond to a step in load current. When a load step occurs, VOUT immediately shifts by an amount equal to $(\Delta I_{LOAD} \times ESR)$, where ESR is the effective series resistance of COUT. ΔI_{LOAD} also begins to charge or discharge COUT, which generates a feedback error signal. The regulator loop then acts to return VOUT to its steady-state value. During this recovery time VOUT can be monitored for overshoot or ringing that would indicate a stability problem.

Efficiency Considerations

Although all dissipative elements in the circuit produce losses, one major source usually account for most of the losses in LN2431 circuits: I^2R losses. The I^2R loss dominates the efficiency loss at medium to high load currents. The I^2R losses are calculated from the resistances of the internal switches, RLX, and external inductor RL. In continuous mode, the average output current flowing through inductor L is "chopped" between the main switch and the external diode. Thus the series resistance looking into the LX pin is a function of internal high-side switch's $R_{DS(ON)}$, external low-side diode's forward resistance $R_D(F)$ and the duty cycle (D) as follows:

$$R_{LX} = (R_{DS(ON)})D + (R_D(F))(1-D)$$

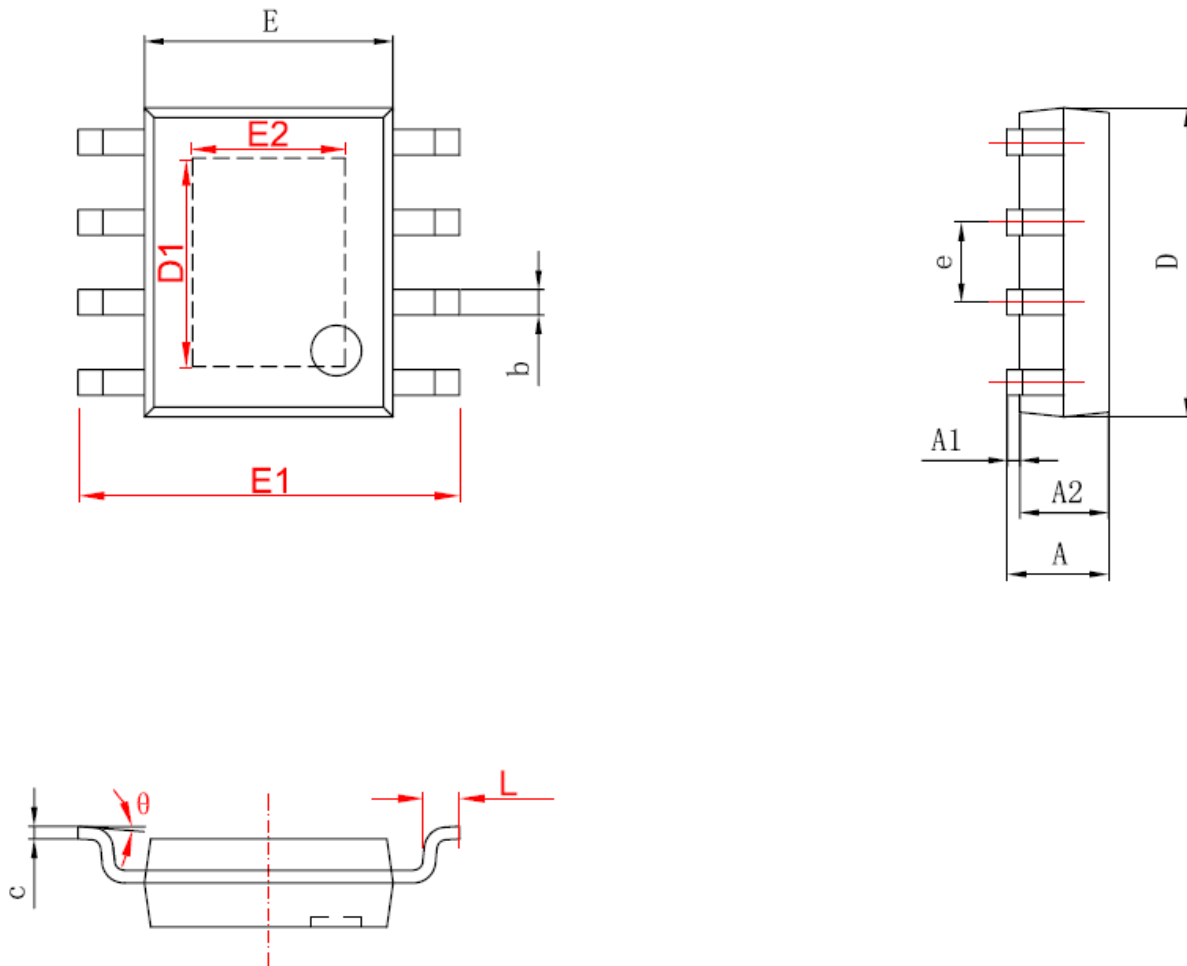
Thus, to obtain I^2R losses, simply add RLX to RL and multiply the result by the square of the average output current. Other losses including CIN and COUT ESR dissipative losses and inductor core losses generally account for less than 2% total additional loss.

Thermal considerations

In most application the LN2431 does not dissipate much heat due to its high efficiency. But, in applications where the LN2431 is running at high ambient temperature with low supply voltage and high duty cycles, such as in dropout, the heat dissipated may exceed the maximum junction temperature of the part. If the junction temperature reaches approximately 140°C, both power switches will be turned off and the LX node will become high impedance.

■ Package Information

- SOP-8/PP



字符	Dimensions In Millimeters		Dimensions In Inches	
	Min	Max	Min	Max
A	1.350	1.750	0.053	0.069
A1	0.050	0.150	0.004	0.010
A2	1.350	1.550	0.053	0.061
b	0.330	0.510	0.013	0.020
c	0.170	0.250	0.006	0.010
D	4.700	5.100	0.185	0.200
D1	3.202	3.402	0.126	0.134
E	3.800	4.000	0.150	0.157
E1	5.800	6.200	0.228	0.244
E2	2.313	2.513	0.091	0.099
e	1.270 (BSC)		0.050 (BSC)	
L	0.400	1.270	0.016	0.050
θ	0°	8°	0°	8°