

Advanced PWM and Linear Power Controller for Portable Applications

The ISL6226 provides power control and protection for two user selected output voltages required in high-performance notebook PC applications. The IC integrates a fixed, selectable frequency pulse-width-modulation (PWM) controller and a 3A LDO linear regulator controller with monitoring and protection circuitry in a single 24 lead SSOP package.

The PWM controller regulates the battery input voltage to a user selected output voltage. Synchronous converter and hysteretic operation at light loads contribute to a high efficiency over a wide range of input voltage and load variation. Efficiency can be further enhanced by using the lower MOSFET's $r_{DS(ON)}$ as the current sense element. Input voltage feed-forward ramp modulation, current-mode control, and internal feed-back compensation provide fast and stable handling of input voltage load transients encountered in advanced portable computer chip sets.

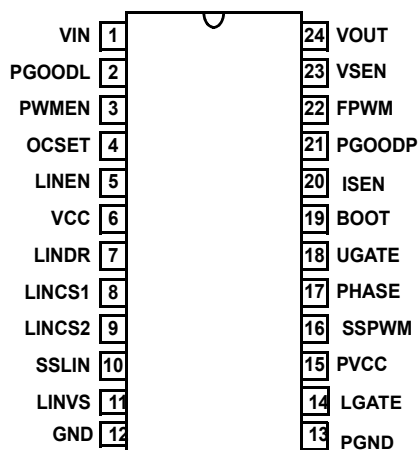
The internal linear regulator controller provides a user selected high current output from an unregulated input with user selected current limit.

Ordering Information

PART NUMBER	TEMP. RANGE (°C)	PACKAGE	PKG. NO.
ISL6226CA	-10 to 85	24 Ld SSOP	M24.15
ISL6226EVAL1	Evaluation Board		

Pinout

ISL6226 (SSOP)
TOP VIEW



Features

- Provides Two Regulated Voltages
 - Linear Regulator, User Selected, Current Limited
 - High Efficiency PWM Over Wide Line and Load Range
 - Synchronous Buck Converter on PWM Output
 - Hysteretic Operation at Light Load
- Dual Mode Operation:
 - Directly From Battery Input, V_{IN} , of 5.6V to 24V or from V_{CC} of 5V
- PWM Output Adjustable From 0.9V to $0.93 \cdot V_{IN}$
- Low Drop-Out (LDO) Linear Regulator Operates from $V_{out} + 0.5V$ to 24V
- LDO Output Adjustable
 - From 0.8V to 3.8V at 0A to 3A for NPN Transistor
 - From 0.8V to 3.0V at 0A to 3A for N-MOSFET
- 300/600kHz Switching Frequency Selectable
- Forced PWM Mode Control Option
- No Current-Sense Resistor Required on PWM Output
 - Uses MOSFET's $r_{DS(ON)}$
 - Optional Current-Sense Resistor for More Precision
- Separate Soft-Start For Both PWM and LDO
- Separate Enable Pins for both PWM and LDO for Advanced Configuration and Power Interface (ACPI) Compatibility
- Input Under voltage Lock-Out (UVLO) Protection
- Excellent Dynamic Response
 - Input Voltage Feed-Forward and Current-Mode Control
- Monitors Output Voltages and Provides PGOOD Status
- Thermal Shut-Down Protection

Applications

- Mobile PCs
- Hand-Held Portable Instruments
- Other Devices and Appliances With up to 24V Input

Related Literature

- AN1013

Absolute Maximum Ratings

Bias Voltage, Vcc	-0.3V to +7V
Input Voltage, VIN	+30.0V
Phase and Isen Pins	GND-0.3V to +30.0V
Boot and Ugate Pins	+35.0V
BOOT with respect to PHASE	+7.0V
I _{max} for LINDR Pin	70mA
All other pins	GND -0.3V to 7V
ESD Classification	Class 1

Thermal Information

Thermal Resistance (Typical, Note 1)	θ_{JA} (°C/W)
SSOP Package	90
Maximum Junction Temperature (Plastic Package)	150°C
Maximum Storage Temperature Range	-65°C to 150°C
Maximum Lead Temperature (Soldering 10s)	300°C (SSOP - Lead Tips Only)

Recommended Operating Conditions

Bias Voltage, Vcc	+5.0V ± 5 %
Input Voltage, VIN	+5.6V to +24.0V
Ambient Temperature Range	-10°C to 85°C
Junction Temperature Range	-10°C to 125°C

CAUTION: Stresses above those listed in "Absolute Maximum Ratings" may cause permanent damage to the device. This is a stress only rating and operation of the device at these or any other conditions above those indicated in the operational sections of this specification is not implied.

NOTE:

- θ_{JA} is measured with the component mounted on a high effective thermal conductivity test board in free air. See Tech Brief TB379 for details.

Electrical Specifications Operating Conditions: Vcc = 5V, T_A = -10°C to 85°C, Unless Otherwise Noted.

PARAMETER	SYMBOL	TEST CONDITIONS	MIN	TYP	MAX	UNITS
VCC SUPPLY						
Bias Current	I _{CC}	LGATE, UGATE Open, VSEN forced above regulation point (Fs = 300K Hz)	-	1100	-	μA
Bias Current	I _{CC}	LGATE, UGATE Open, VSEN forced above regulation point (Fs = 600K Hz)	-	1100	-	μA
Shut-down Current	I _{CCSN}		-	5	25	μA
VCC UVLO						
Rising Vcc Threshold			4.30	-	4.65	V
Falling Vcc Threshold			4.1	-	4.45	V
Vcc Hysterises			30		500	mV
VIN						
Input Voltage Pin Current (Sink)	I _{VIN}	VIN pin connected to the input voltage source	10	20	30	μA
Input Voltage Pin Current (Source)	I _{VIN}	VIN pin connected to ground	-7	-12	-17	μA
Shut-down Current	I _{VIN}		-	-	1	μA
OSCILLATOR						
PWM Oscillator Frequency	F _{c1}	V _{IN} =3.5V -- 24V	255	300	345	kHz
PWM Oscillator Frequency	F _{c2}	V _{IN} ≤ 0.5V	510	600	690	kHz
Ramp Amplitude, pk-pk	V _{R1}	V _{IN} = 16V	-	2	-	V
Ramp Amplitude, pk-pk	V _{R2}	V _{IN} ≤ 5V	-	1.25	-	V
Ramp Offset	V _{ROFF}		-	0.5	-	V
PWM CONTROLLER						
REFERENCE AND SOFT START						
PWM Internal Reference Voltage	V _{REF}		-	0.9	-	V
PWM Soft-Start Threshold	V _{SOFT}		-	1.5	-	V
Reference Voltage Accuracy			-2.0	-	+2.0	%

ISL6226

PARAMETER	SYMBOL	TEST CONDITIONS	MIN	TYP	MAX	UNITS
Soft-Start Current During Start-up	I _{SOFT}		-	5	-	μA
PWM CONVERTER						
Load Regulation		0.0mA < I _{VOUT1} < 5.0A; 5.0V < V _{IN} < 24.0V	-2.0	-	+2.0	%
VSEN pin bias current	I _{VSEN}		-	80	-	nA
VOUT pin input impedance	I _{VOUT}		135	150	198	kOhm
Under-Voltage Shut-Down Level	V _{UV1}	Fraction of the set point; ~3μs noise filter	70	-	76	%
Under Voltage Delay	UV_DLY		3	-	5.5	μS
Over-Voltage Protection	V _{OVP1}	Fraction of the set point; ~1μs noise filter	113.5	-	118	%
Over Voltage Delay	OVP_DLY		0.5	-	2.7	μS
PWM CONTROLLER GATE DRIVER						
Upper Drive Pull-Up Resistance	R _{2UGPUP}		6.2	8.5	13.5	Ω
Upper Drive Pull-Down Resistance	R _{2UGPDN}		2.5	3.0	4.7	Ω
Lower Drive Pull-Up Resistance	R _{2LGPUP}		6.2	8.5	13.4	Ω
Lower Drive Pull-Down Resistance	R _{2LGPDN}		1.3	1.6	2.7	Ω
POWER GOOD AND CONTROL FUNCTIONS						
Power Good Lower Threshold	V _{PG-}	Fraction of the set point; ~3μs noise filter	-13.5	-	-7.5	%
Power Good Higher Threshold	V _{PG+}	Fraction of the set point; ~3μs noise filter	7.4	-	13.5	%
PGOOD Leakage Current	I _{PGLKG}	V _{PULLUP} = 5.5V	-	-	1	μA
PGOOD Voltage Low		I _{PGOOD} = -4mA	-	-	0.5	V
EN- Low (Off)			-	-	0.8	V
EN - High (On)			2.0	-	-	V
FCCM -Hysteretic Operation Enabled			-	V _{CC} /2	-	V
LOW DROP OUT CONTROLLER FUNCTIONS						
Linear Input Voltage	V _{in}	No External Passive Device	-	V _{CC}	-	V
Linear Input Voltage	V _{in}	With External Passive Device	Device	Rating		
Linear Output Voltage	LDO _{out}	No External Passive Device	0.8	-	4.5	V
Output Drive Current for NPN	I(DRV) out		50	-	-	mA
OverCurrent Feedback Voltage	V(CS)			100		mV
Under-Voltage Shut-Down Level	V _{UV}	Fraction of the set point; ~3μs noise filter	71	74	77	%
Under-Voltage Delay	UV _{DLY}		5.5	-	8	μS
REFERENCE AND SOFT START						
LDO Internal Reference Voltage	V _{REF}		-	0.8	-	V
LDO Soft-Start Threshold	V _{SOFT}			1.5		V
Reference Voltage Accuracy			-2.0	-	+2.0	%
Soft-Start Current During Start-up	I _{SOFT}		-	5	-	μA
POWER GOOD AND CONTROL FUNCTIONS						
Power Good Lower Threshold	V _{PG-}	V _O < V _{REF}	-15.5	-12	-9	%
Power Good Higher Threshold	V _{PG+}	V _O > V _{REF}	10.5	13.5	16.5	%
PGOOD Leakage Current	I _{PGLKG}	V _{PULLUP} = 5.5V	-	-	1.0	μA
PGOOD Voltage Low	V _{PGOOD}	I _{PGOOD} = -4mA, V _{CC} = 4.75V	-	-	0.5	V
LINEN-Low (OFF)			-	-	0.8	V

PARAMETER	SYMBOL	TEST CONDITIONS	MIN	TYP	MAX	UNITS
LINEN-High (ON)			2.0	-	-	V
Input Bias Current LINCS1 and LINCS2	Ibias1	Vlincs = 0.8 V	-	-80	-	nA
Input Bias Current LINVS	Ibias2	Vlinvs = 0.8 V	-	-80	-	nA

Functional Pin Descriptions

VIN (Pin 1)

Provides battery voltage to the oscillator for feed-forward rejection of the input voltage variation. Also, this pin programs frequency of the internal clock and gain of the ramp generator.

PGOODL (Pin 2)

PGOODL is an open drain output used to indicate the status of the LDO output voltages. This pin is pulled low when the output voltage is not within of its respective nominal voltage or over 120mV between pin 8 and 9.

PWMEN (Pin 3)

This pin provides enable/disable function the PWM output. The output is enabled when this pin is high. The PWM output is held off when this pin is pulled to ground. Intersil recommends to use hysteresis mode when input voltage higher than 18V and output voltage is 1.25V or less.

OCSET (Pin 4)

A resistor on this pin to ground sets the over current threshold for the PWM controller.

LINEN (Pin 5)

This pin provides enable/disable function and soft-start for the LDO. The output is enabled when this pin is high. The LDO is held off when this pin is pulled to ground.

VCC (Pin 6)

Input power for the controller and the upper MOSFET gate drive. The IC starts to operate when the voltage on this pin exceeds 4.3V and stops operating when the voltage on this pin drops below approximately 4.45V.

LINDR (Pin 7)

Current output to drive the NPN transistor.

LINCS1 (Pin 8)

High side of current sense resistor.

LINCS2 (Pin 9)

Low side of current sense resistor. Current limit for the linear regulator is initiated when the voltage difference between LINCS1 and LINCS2 is 120mV.

SSLIN (Pin 10)

This pin provides soft start of the LDO controller. When the EN pin is pulled high, the voltage on the capacitor connected to the soft start pin is rising linearly due to the 5 μ A pull-up current. The output voltage follows the voltage on the

capacitor till it reaches the value of 0.8V. At this moment the output voltage starts to regulate and soft start continues to rise to 1.5V. At this time, the soft start is complete and the PGOODL will be high to indicate the output voltage within its respective nominal voltage. The further rise of soft start capacitor does not affect the output voltage. The soft-start time can be obtained from the following equation.

$$T_{sslin} = \frac{1.5V \times C_{ss}}{5\mu A}$$

LINVS (Pin 11)

Voltage regulation point (0.8V) for the linear output.

GND (Pin 12)

Signal ground for the IC. All voltage levels are measured with respect to this pin.

PGND (Pin 13)

This is the lower MOSFET gate drive return connection for PWM converter. Tie the lower MOSFET source directly to this pin.

LGATE (Pin 14)

This pin provides the gate drive for the lower MOSFET. Connect the lower MOSFET gate to this pin.

PVCC (Pin 15)

This pin powers the lower MOSFET gate driver.

SSPWM (Pin 16)

This pin provides soft start of the PWM controller. This pin has the same function as SSLIN pin. When the EN pin is pulled high, the voltage on the capacitor connected to the soft start pin is rising linearly due to the 5 μ A pull-up current. The output voltage follows the voltage on the capacitor till it reaches the value of 0.9V. At this moment the output voltage starts to regulate and soft start continues to rise to 1.5V. At this time, the soft start is complete and the PGOODPWM will be high to indicate the output voltage within its respective nominal voltage. The further rise of soft start capacitor does not affect the output voltage. The soft-start time can be obtained from the following equation.

$$T_{sspwm} = \frac{1.5V \times C_{ss}}{5\mu A}$$

PHASE (Pin17)

The phase node is the junctions of the upper MOSFET source, output filter inductor, and lower MOSFET drain. Connect the PHASE pin directly to the PWM converter's lower MOSFET drain.

UGATE (Pin18)

This pin provides the gate drive for the upper MOSFET. Connect UGATE pin to the PWM converter's upper MOSFET gate.

BOOT (Pin 19)

Power is supplied to the upper MOSFET driver of the PWM converter via the BOOT pin. Connect this pin to the junction the bootstrap capacitor with the cathode of the bootstrap diode. The anode of the bootstrap diode is connected to pin 6, VCC.

ISEN (Pin 20)

This pin is used to monitor the voltage drop across the lower MOSFET for current feedback control. For more precise current detection, this input can be connected to optional current sense resistors placed in series with the source of the lower MOSFET.

PGOODP (Pin 21)

PGOODP is an open drain output used to indicate the status of the PWM converter's output voltages. This pin is pulled low when the output is not within of their respective nominal voltage.

FPWM (Pin 22)

This pin when pulled to VCC restrains hysteretic operation in light loads.

VSEN (Pin 23)

This pin is connected to the PWM voltage divider to provide the voltage feedback signal for the PWM controller. The PGOODP, overvoltage protection (OVP) and undervoltage shutdown circuits use this signal to determine output voltage status and/or to initiate undervoltage shut down.

VOUT (Pin 24)

The output voltage is sensed on this pin to provide feedback during the PWM to Hysteretic mode transition.

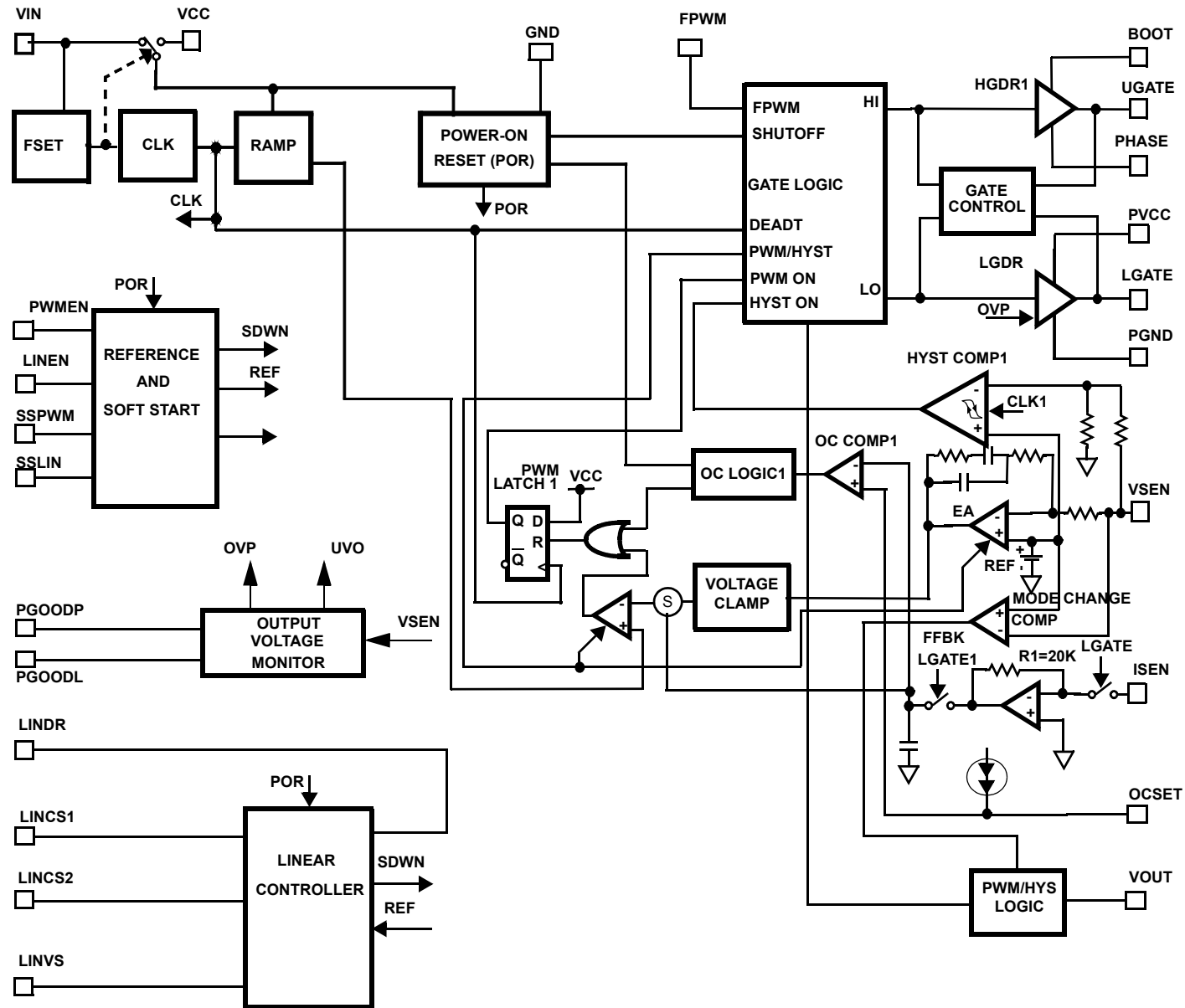


FIGURE 1. FUNCTIONAL BLOCK DIAGRAM

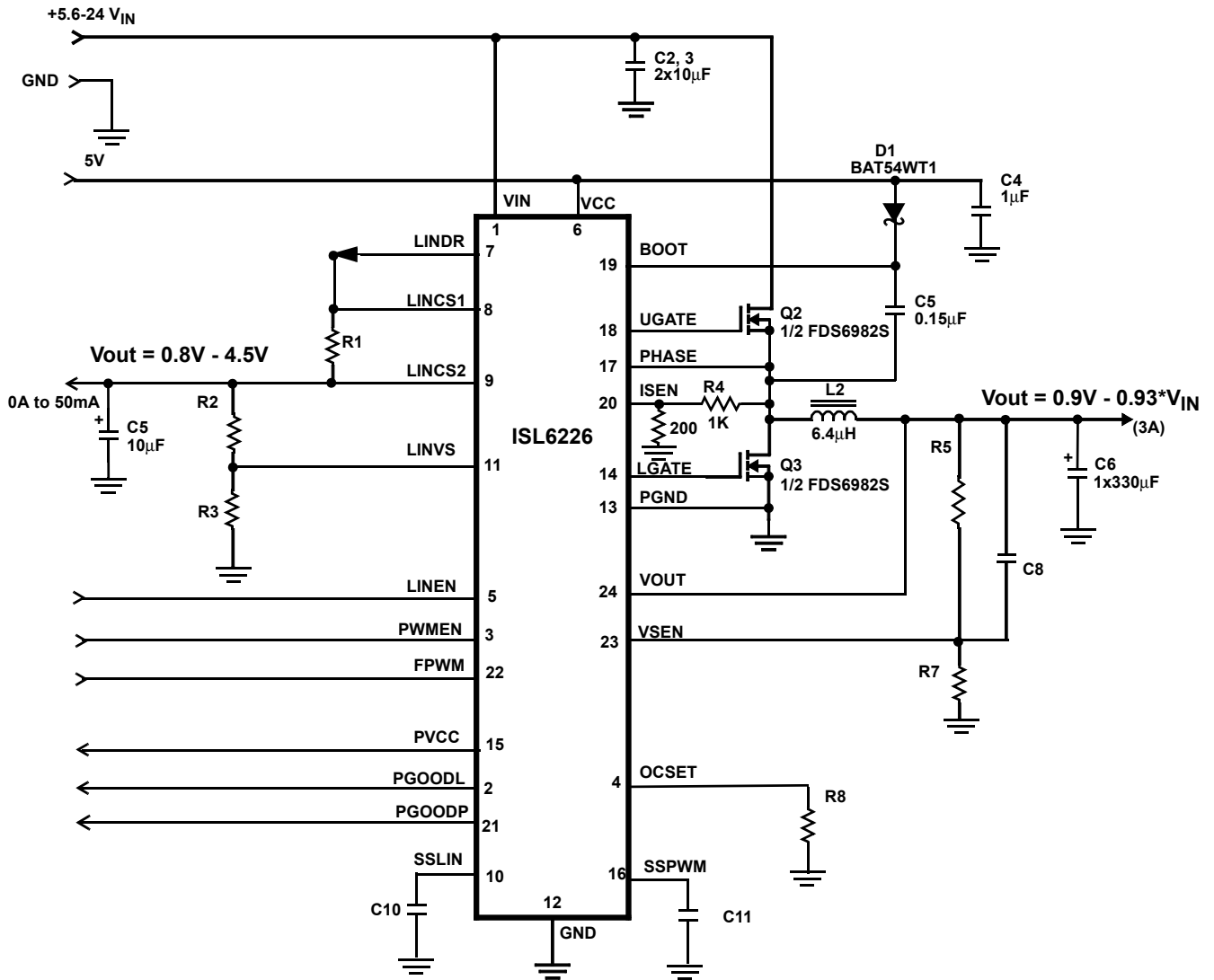


FIGURE 2. 0A to 50mA OUTPUT CURRENT LDO AND SYNCHORNOUS BUCK CONVERTER (F_s=300KHz)

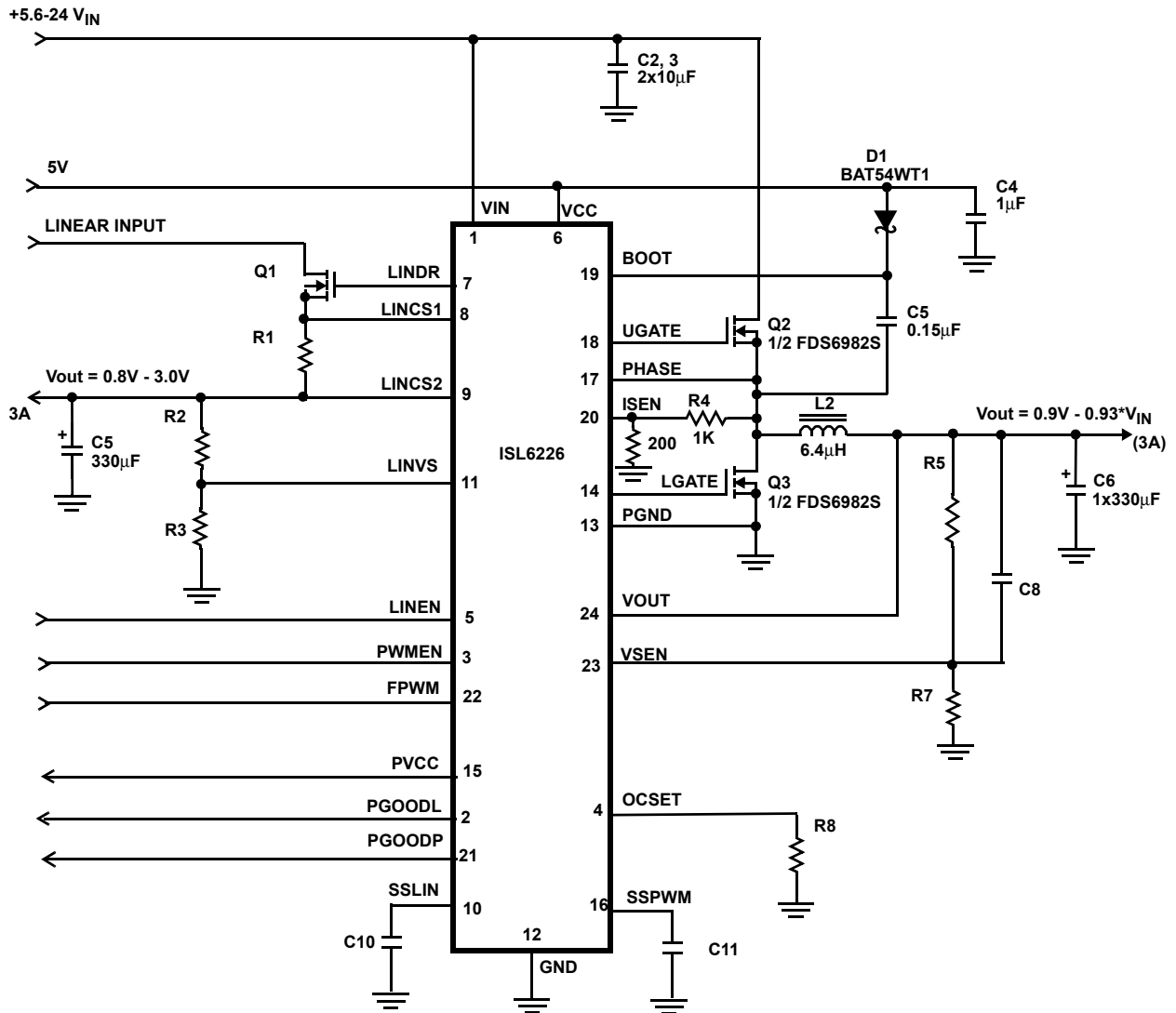


Figure 3. LDO WITH EXTERNAL N-MOSFET AND SYNCHORNOUS BUCK CONVERTER (Fs = 300KHz)

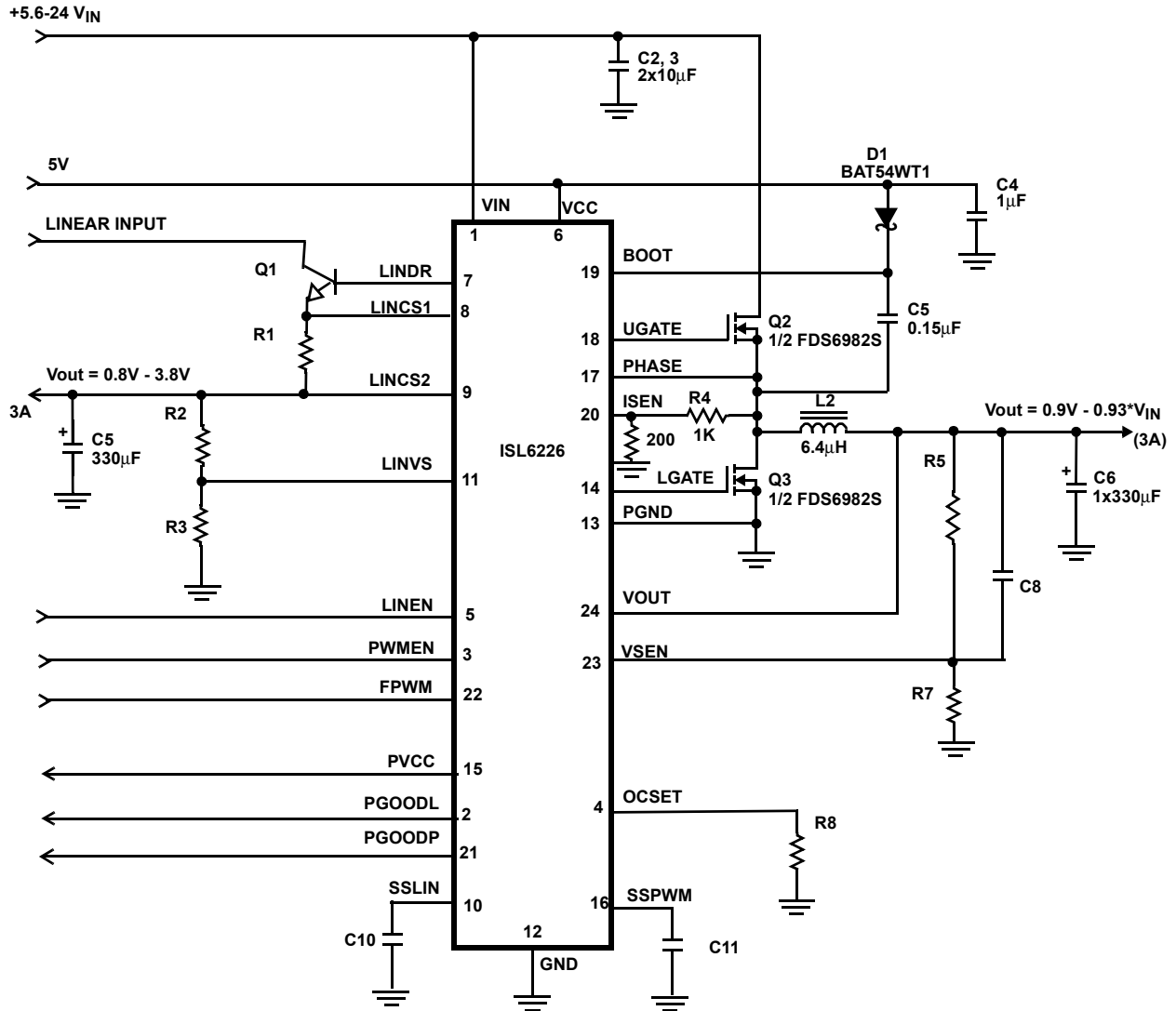


FIGURE 4. LDO with External NPN and Synchronous Buck Converter (Fs = 300KHz)

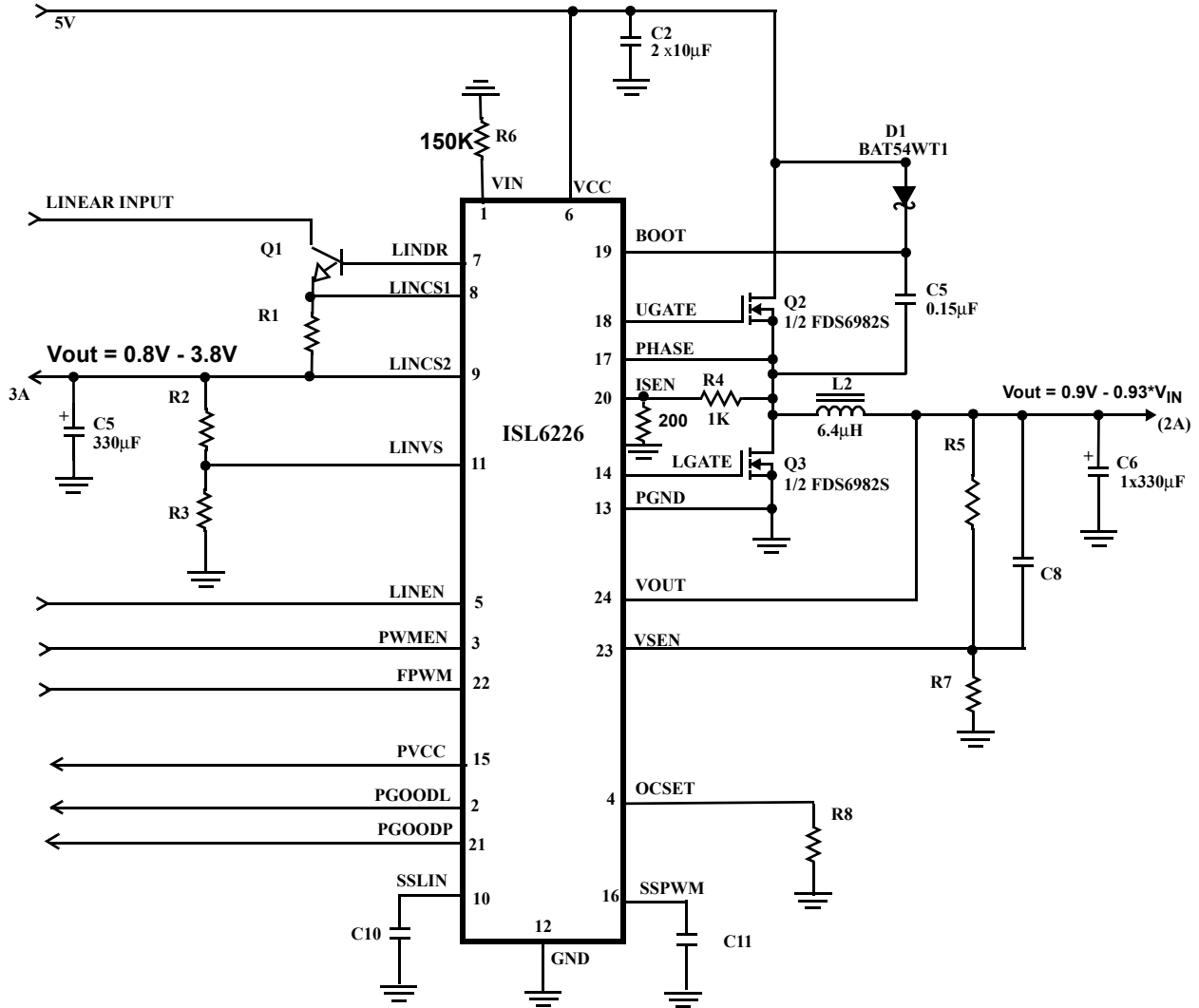


FIGURE 5. APPLICATION CIRCUIT FOR TWO STEP 300 kHz CONVERSION (MODE 2)

NOTE: Q1 can be either N-MOSFET, NPN or not use for 50 mA

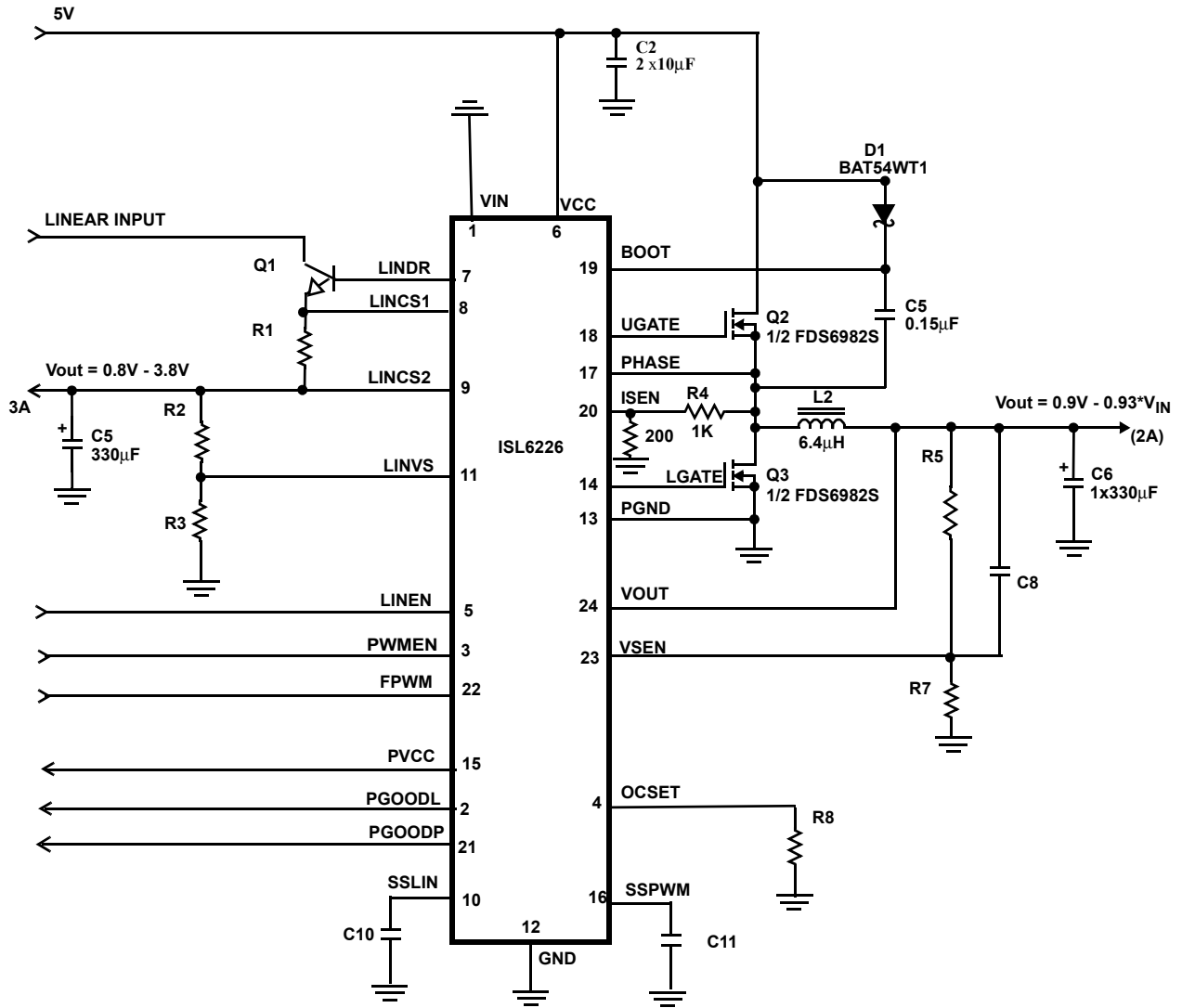


FIGURE 6. APPLICATION CIRCUIT FOR TWO STEP 600 kHz CONVERSION (MODE 3)

NOTE: Q1 can be either N-MOSFET, NPN or not used for 50 mA

Description Of Operation

The ISL6226 is a single channel PWM controller and linear regulator controller intended for chipset, DRAM, audio or other low voltage power needs of modern notebook and sub-notebook PCs. The IC integrates control circuits for a single synchronous buck converter whose output voltage is set in the range of 0.9V to 0.93*Vin by an external resistive divider and a linear regulator controller which drives an external NMOSFET or NPN pass transistor whose output is set in the range of 0.8V to 3.8V by an external resistor divider. The synchronous buck converter can be operated in several modes to optimize the application. The 300kHz frequency operation is recommended when powered from the battery, which voltage varies from 5.6V to 24V. In this case the VIN pin is connected to the battery terminal and the ramp generator gain is dynamically adjusted according to the input voltage.

The IC also can be optimized for a two stage power conversion configuration where the output voltage is derived from a 5V input in lieu of the raw battery. Two operating frequencies of 300kHz or 600kHz are available in this mode. If 300kHz operation is desired while operating from the 5V system bus, the VIN pin should be connected to ground via a 150K ohm resistor. The ramp generator gain is corrected for the lower input voltage.

When powered from the system 5V rail, the converter can be configured to operate on increased frequency of 600kHz to minimize the output filter components. In this case, the VIN pin is connected to the ground. Table 1 summarizes the various modes of operation. Figure 7 below shows plots of the ramp speed compensation.

Table 1. Configuration for Modes of Operation

Operation	Pin 1 connection	Pin 1 potential
One stage 300kHz	VIN	V1>5.6V
Two stage 300kHz	150K-GND	1V<V1<2V
Two stage 600kHz	GND	V1<0.5V

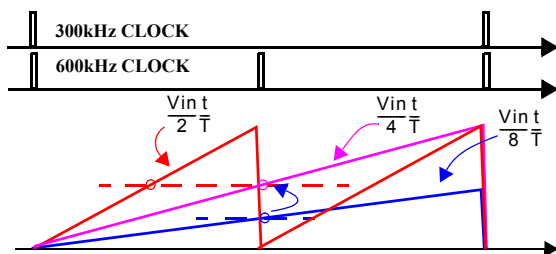


Figure 7. Ramp Speed Compensation Vo=2.5V

Current Sensing and Current Limit Protection

The PWM converter uses the lower MOSFET on-state resistance, $r_{DS(ON)}$, as the current-sensing element. This technique eliminates the need for a current sense resistor and the associated power losses. If more accurate current protection is desired, current sense resistors may be used in series with the lower MOSFETs' source.

A current proportional signal is used to provide average current mode control and over-current protection. The gain in the current sense circuit is set by the resistor connected from ISEN pin to the switching node of buck converter. The value of this resistor can be estimated by the following expression:

$$R_{isen} = \frac{I_{omax} \cdot R_{dson}}{75\mu A} - 100$$

where I_{omax} is the maximum inductor current. The value of R_{isen} should be specified for the expected maximum MOSFET operating temperature.

The resulting current out of the ISEN pin through R_{isen} , is used for current feedback and current limit protection. An over-current protection threshold is set by an external resistor connected from OCSET (pin 4) to ground. The value of this resistor can be obtained from the following expression:

$$R_{ocset} = \frac{9.6 \cdot (R_{isen} - 100)}{I_{oc} \cdot R_{dson}}$$

where I_{oc} is the value of over current. This is compared with an internal current limit threshold. When a sampled value of the output current is determined to be above the current limit threshold, the PWM drive is terminated and a counter is initiated. This limits the inductor current build-up and essentially switches the converter into current-limit mode. If an overcurrent is detected between 26 μ s to 53 μ s later, an overcurrent shutdown is initiated. If during the 26 μ s to 53 μ s period, an overcurrent is not detected, the counter is reset and sampling continues as normal.

This current limit scheme has proven to be very robust in applications like portable computers where fast inductor current build-up is common due to a large difference between input and output voltages and a low value of the inductor.

Light-Load (Hysteretic) Operation

In the light-load (hysteretic) mode the output voltage is regulated by the hysteretic comparator which regulates the output voltage by maintaining the output voltage ripple as shown in Figure 8. In Hysteretic mode, the inductor current flows only when the output voltage reaches the lower limit of the hysteretic comparator and turns off at the upper limit. Hysteretic mode saves converter energy at light loads by supplying energy only at the time when the output voltage

requires it. This mode conserves energy by reducing the power dissipation associated with continuous switching.

During the time between inductor current pulses, both the upper and lower MOSFETs are turned off. This is referred to as 'diode emulation mode' because the lower MOSFET performs the function of a diode. This diode emulation mode prevents the output capacitor from discharging through the lower MOSFET when the upper MOSFET is not conducting.

NOTE: the FPWM operation can intentionally be forced by tying pin 22, FPWM, to V_{cc} .

Operation-Mode Control

The mode-control circuit changes the converter's mode of operation based on the voltage polarity of the phase node when the lower MOSFET is conducting and just before the upper MOSFET turns on. For continuous inductor current, the phase node is negative when the lower MOSFET is conducting and the converter operates in fixed-frequency PWM mode as shown in Figure 9. When the load current decreases to the point where the inductor current flows through the lower MOSFET in the 'reverse' direction, the phase node becomes positive, and the mode is changed to hysteretic.

A phase comparator handles the timing of the phase node voltage sensing. A low level on the phase comparator output indicates a negative phase voltage during the conduction time of the lower MOSFET. A high level on the phase comparator output indicates a positive phase voltage.

When the phase node is positive (phase comparator high), at the end of the lower MOSFET conduction time, for eight consecutive clock cycles, the mode is changed to hysteretic as shown in Figure 9. The dashed lines indicate when the phase node goes positive and the phase comparator output goes high. The solid vertical lines at 1,2,...8 indicate the sampling time, of the phase comparator, to determine the polarity (sign) of the phase node. At the transition between PWM and hysteretic mode both the upper and lower MOSFETs are turned off. The phase node will 'ring' based on the output inductor and the parasitic capacitance on the phase node and settle out at the value of the output voltage.

The mode change from hysteretic to PWM can be caused by one of two events. One event is the same mechanism that causes a PWM to hysteretic transition. But instead of looking for eight consecutive positive occurrences on the phase node, it is looking for eight consecutive negative occurrences on the phase node. The operation mode will be changed from hysteretic to PWM when these eight consecutive pulses occur. This transition technique prevents jitter of the operation mode at load levels close to boundary.

The other mechanism for changing from hysteretic to PWM is due to a sudden increase in the output current. This step load causes an instantaneous decrease in the output voltage due to the voltage drop on the output capacitor ESR and ESL. If the decrease causes the output voltage to drop below the hysteretic regulation level, the mode is changed to

PWM on the next clock cycle. This insures the full power required by the increase in output current.

Gate Control Logic

The gate control logic translates generated PWM control signals into the MOSFET gate drive signals providing necessary amplification, level shifting and shoot-through protection. Also, it has functions that help optimize the IC performance over a wide range of operational conditions. Since MOSFET switching time can vary dramatically from type to type and with the input voltage, the gate control logic provides adaptive dead time by monitoring the gate-to-source voltages of both upper and lower MOSFETs. The lower MOSFET is not turned on until the gate-to-source voltage of the upper MOSFET has decreased to less than approximately 1 volt. Similarly, the upper MOSFET is not turned on until the gate-to-source voltage of the lower MOSFET has decreased to less than approximately 1 volt. This allows a wide variety of upper and lower MOSFETs to be used without a concern for simultaneous conduction, or shoot-through.

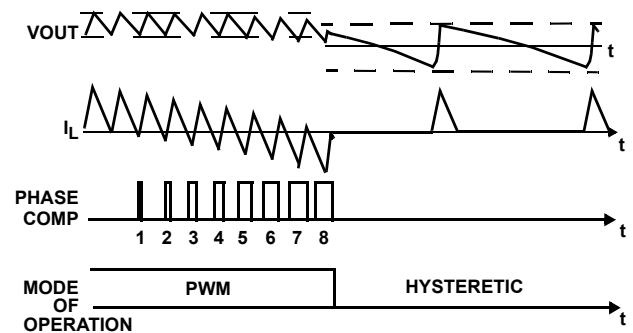


FIGURE 8. REGULATION IN HYSTERETIC MODE

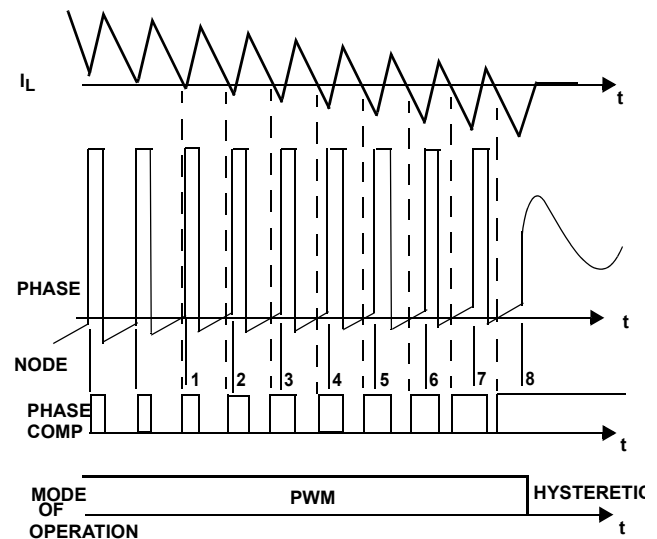


FIGURE 9. MODE CONTROL WAVEFORMS

Feedback Loop Compensation

To reduce the number of external components and remove the burden of determining compensation components from a system designer, PWM controller has internally compensated error amplifiers. To make internal compensation possible several design measures were taken.

First, the ramp signal applied to the PWM comparator is proportional to the input voltage provided via the VIN pin as shown in Figure 7. Second, the load current proportional signal is derived from the voltage drop across the lower MOSFET during on time of lower MOSFET. This signal is added to the amplifier error signal on the comparator input. This effectively creates an internal current control loop. Because of current loop, a compensation network with one zero and two poles is usually sufficient. One pole is placed at origin which has high DC gain. The zero is placed around 9kHz, and another pole is around 137kHz. The zero-pole pair causes a flat gain region at frequencies in between the zero and the pole. This region is also made to make the phase bump or reduce phase shift. The amount of phase shift reduction depends on how wide the region of flat gain and has a maximum value of 90°.

Low Drop Out Linear Regulator

The linear has separate Enable, Pgood and Soft Start pins from the PWM. The linear also has over current and under voltage protection features.

The over current protection will shut down the LDO controller when the voltage between pin 8 and 9 is more than 100mV. The LDO controller will automatically restart again when the voltage is less than 100mV. Intersil recommends to use this feature to protect the ISL6226 IC otherwise the IC will be damaged when over current happens.

The under voltage protection will start when the output voltage is 26% lower than the nominal output voltage.

The ISL6226 has an internal P-channel MOSFET pass transistor. This provides several advantages over PNP pass transistor. PNP regulator wastes considerable amounts of current in dropout when the pass transistor saturates. The P-channel MOSFET requires no base drive current which reduces quiescent current.

Output Voltage Calculation

Figure 10 describes the output voltage which is relative to the bias voltage. Due to the over current protection circuitry working properly, the internal P-channel MOSFET has an approximate 0.5V drop across it. Therefore, the maximum voltage at the LINDR pin will be.

$$V_{LINDR} = V_{CC} - 0.5V$$

The maximum output voltage equals to the voltage at LINDR

pin minus the diode drop voltage of between the base and emitter of NPN transistor device and minus 100mV across pin 8 and 9. In the case of using MOSFET instead of NPN transistor, the maximum output voltage equals to the voltage at LINDR pin minus the gate threshold voltage of MOSFET and minus 100mV across pin 8 and 9.

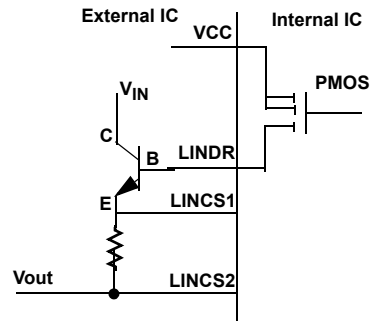


FIGURE 10. LDO

Output Current

The LDO controller is capable of sourcing 50mA out of the LINDR pin. This amount of current is used to drive the NPN transistor or as load current up to 50mA. The output current of the system equals the current out of the LINDR pin times the current gain of the NPN transistor (β , Beta).

Component Selection Guidelines

LDO Device Considerations

The NPN transistor or MOSFET should be chosen to carry the load current and power dissipation required by the conditions of the load current and voltage drop across the device.

Output Capacitor Selection-LDO

For stable operation, the output current up to 50mA or without the external device, a minimum output capacitor value of 4.7 μ F is recommended. The design simulation results indicated that the phase margins vary from 70 degrees to 90 degrees while the output currents vary from 5 μ A to 50mA, respectively. The higher output capacitance indicates the higher phase margin.

For stable operation with the external device, a minimum output capacitor value of 220 μ F is recommended. Larger capacitor values provide better transient response. A capacitor value of 330 μ F, such as SANYO POSCAP 6TPB330M, will provide good phase margin and better transient response.

Output Capacitor Selection-PWM

The output capacitors have unique requirements. In general, the output capacitors should be selected to meet the dynamic regulation requirements including ripple voltage and load transients.

Selection of the output capacitors is also dependent on the output inductor so some inductor analysis is required to select the output capacitors.

One of the parameters limiting the converter's response to a load transient is the time required for the inductor current to slew to its new level. Given a sufficiently fast control loop design, the ISL6226 will provide either 0% or 93% duty cycle in response to a load transient. The response time is the time interval required to slew the inductor current from an initial current value to the load current level. During this interval the difference between the inductor current and the transient current level must be supplied by the output capacitor(s). Minimizing the response time can minimize the output capacitance required. Also, if the load transient rise time is slower than the inductor response time, as in a hard drive or CD drive, this reduces the requirement on the output capacitor.

The maximum capacitor value required to provide the full, rising step, transient load current during the response time of the inductor is:

$$C_{OUT} = \frac{L_O \times I_{TRAN}}{(V_{IN} - V_{OUT}) \times 2} \times \frac{I_{TRAN}}{DV_{OUT}}$$

Where: C_{OUT} is the output capacitor(s) required, L_O is the output inductor, I_{TRAN} is the transient load current step, V_{IN} is the input voltage, V_{OUT} is output voltage, and DV_{OUT} is the drop in output voltage allowed during the load transient.

High frequency capacitors initially supply the transient current and slow the load rate-of-change seen by the bulk capacitors. The bulk filter capacitor values are generally determined by the ESR (Equivalent Series Resistance) and voltage rating requirements as well as actual capacitance requirements. The output voltage ripple is due to the inductor ripple current and the ESR of the output capacitors as defined by:

$$V_{RIPPLE} = \Delta I_L \times ESR$$

where, ΔI_L is calculated in the Inductor Selection section.

High frequency decoupling capacitors should be placed as close to the power pins of the load as physically possible. Be careful not to add inductance in the circuit board wiring that could cancel the usefulness of these low inductance components. Consult with the manufacturer of the load circuitry for specific decoupling requirements.

Use only specialized low-ESR capacitors intended for switching-regulator applications, at 300kHz, for the bulk capacitors. In most cases, multiple electrolytic capacitors of small case size perform better than a single large case capacitor.

The stability requirement on the selection of the output capacitor is that the 'ESR zero', f_Z , be between 1.8kHz and 45kHz. This range is set by an internal, single compensation

zero at 9kHz. The ESR zero can be a factor of five on either side of the internal zero and still contribute to increased phase margin of the control loop. Therefore:

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

In conclusion, the output capacitors must meet three criteria:

1. They must have sufficient bulk capacitance to sustain the output voltage during a load transient while the output inductor current is slewing to the value of the load transient
2. The ESR must be sufficiently low to meet the desired output voltage ripple due to the output inductor current, and
3. The ESR zero should be placed, in a rather large range, to provide additional phase margin.

Output Inductor Selection

The output inductor is selected to meet the output voltage ripple requirements. The inductor value determines the converter's ripple current and the ripple voltage is a function of the ripple current and output capacitor(s) ESR. The ripple voltage expression is given in the capacitor selection section and the ripple current is approximated by the following equation:

$$\Delta I_L = \frac{V_{IN} - V_{OUT}}{F_S \times L} \times \frac{V_{OUT}}{V_{IN}}$$

where F_S is the switching frequency.

Input Capacitor Selection

The important parameters for the bulk input capacitor(s) are the voltage rating and the RMS current rating. For reliable operation, select bulk input capacitors with voltage and current ratings above the maximum input voltage and largest RMS current required by the circuit. The capacitor voltage rating should be at least 1.25 times greater than the maximum input voltage and 1.5 times is a conservative guideline.

The AC RMS input current varies with load. Depending on the specifics of the input power and its impedance, most (or all) of this current is supplied by the input capacitor(s).

Use a mix of input bypass capacitors to control the voltage ripple across the MOSFETs. Use ceramic capacitors for the high frequency decoupling and bulk capacitors to supply the RMS current. Small ceramic capacitors can be placed very close to the upper MOSFET to suppress the voltage induced in the parasitic circuit impedances.

For board designs that allow through-hole components, the Sanyo OS-CON® series offer low ESR and good temperature performance.

For surface mount designs, solid tantalum capacitors can be used, but caution must be exercised with regard to the

capacitor surge current rating. These capacitors must be capable of handling the surge-current at power-up. The TPS series available from AVX is surge current tested.

MOSFET Considerations

The logic level MOSFETs are chosen for optimum efficiency given the potentially wide input voltage range and output power requirements. One dual N-Channel or Two N-channel MOSFETs are used in each of the synchronous-rectified buck converters for the outputs. These MOSFETs should be selected based upon $r_{DS(ON)}$, gate supply requirements, and thermal management considerations.

The power dissipation includes two loss components; conduction loss and switching loss. These losses are distributed between the upper and lower MOSFETs according to duty cycle (see the following equations). The conduction losses are the main component of power dissipation for the lower MOSFETs. Only the upper MOSFET has significant switching losses, since the lower device turns on and off into near zero voltage.

$$P_{UPPER} = \frac{I_O^2 \times r_{DS(ON)} \times V_{OUT}}{V_{IN}} + \frac{I_O \times V_{IN} \times t_{SW} \times F_S}{2}$$

$$P_{LOWER} = \frac{I_O^2 \times r_{DS(ON)} \times (V_{IN} - V_{OUT})}{V_{IN}}$$

The equations assume linear voltage-current transitions and do not model power loss due to the reverse-recovery of the lower MOSFET's body diode. The gate-charge losses are dissipated by the ISL6226 and do not heat the MOSFETs. However, a large gate-charge increases the switching time, t_{SW} which increases the upper MOSFET switching losses. Ensure that both MOSFETs are within their maximum junction temperature at high ambient temperature by calculating the temperature rise according to package thermal-resistance specifications.

Layout Considerations

MOSFETs switch very fast and efficiently. The speed with which the current transitions from one device to another causes voltage spikes across the interconnecting impedances and parasitic circuit elements. The voltage spikes can degrade efficiency, radiate noise into the circuit, and lead to device over voltage stress. Careful component layout and printed circuit design minimizes the voltage spikes in the converter. Consider, as an example, the turn-off transition of one of the upper PWM MOSFETs. Prior to turn-off, the upper MOSFET is carrying the full load current. During the turn-off, current stops flowing in the upper MOSFET and is picked up by the lower MOSFET. Any inductance in the switched current path generates a voltage spike during the switching interval. Careful component selection, tight layout of the critical components, and short, wide circuit traces minimize the magnitude of voltage spikes. See the Application Note

AN1013 for the evaluation board component placement and the printed circuit board layout details.

There are two sets of critical components in a DC-DC converter using an ISL6226 controller. The switching power components are the most critical because they switch large amounts of energy, and as such, they tend to generate equally large amounts of noise. The critical small signal components are those connected to sensitive nodes or those supplying critical bias currents.

Power Components Layout Considerations

The power components and the controller IC should be placed first. Locate the input capacitors, especially the high-frequency ceramic decoupling capacitors, close to the power MOSFETs. Locate the output inductor and output capacitors between the MOSFETs and the load. Locate the PWM controller close to the MOSFETs.

Insure the current paths from the input capacitors to the MOSFETs, to the output inductors and output capacitors are as short as possible with maximum allowable trace widths.

A multi-layer printed circuit board is recommended. Dedicate one solid layer for a ground plane and make all critical component ground connections with via to this layer. Dedicate another solid layer as a power plane and break this plane into smaller islands of common voltage levels. The power plane should support the input power and output power nodes. Use copper filled polygons on the top and bottom circuit layers for the phase nodes, but do not unnecessarily oversize these particular islands. Since the phase nodes are subjected to very high dV/dt voltages, the stray capacitor formed between these islands and the surrounding circuitry will tend to couple switching noise. Use the remaining printed circuit layers for small signal wiring. The wiring traces from the control IC to the MOSFET gate and source should be sized to carry 2A peak currents.

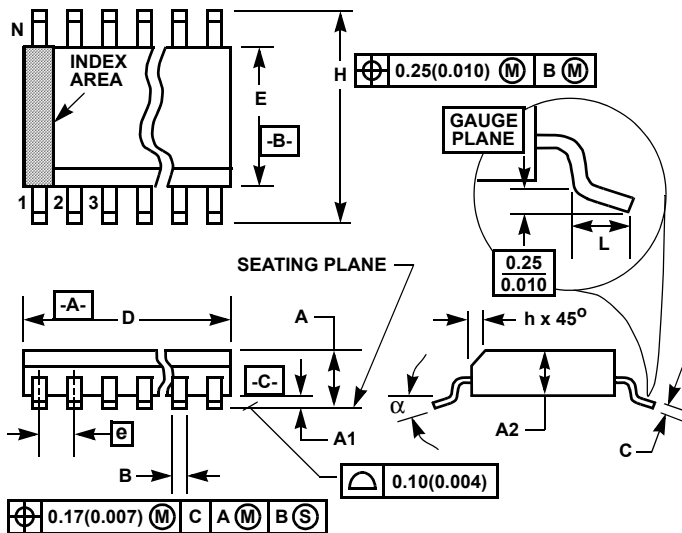
Small Components Signal Layout Considerations

The Vin pin 1 input should be bypassed with a 1.0uF capacitor. The bypass capacitors for Vin and the soft-start capacitor, should be located close to their connecting pins on the control IC.

Refer to the Application Note AN1013 for a recommended component placement and interconnections.

Plastic Packages for Integrated Circuits

Shrink Small Outline Plastic Packages (SSOP)



NOTES:

1. Symbols are defined in the "MO Series Symbol List" in Section 2.2 of Publication Number 95.
2. Dimensioning and tolerancing per ANSI Y14.5M-1982.
3. Dimension "D" does not include mold flash, protrusions or gate burrs. Mold flash, protrusion and gate burrs shall not exceed 0.15mm (0.006 inch) per side.
4. Dimension "E" does not include interlead flash or protrusions. Interlead flash and protrusions shall not exceed 0.25mm (0.010 inch) per side.
5. The chamfer on the body is optional. If it is not present, a visual index feature must be located within the crosshatched area.
6. "L" is the length of terminal for soldering to a substrate.
7. "N" is the number of terminal positions.
8. Terminal numbers are shown for reference only.
9. Dimension "B" does not include dambar protrusion. Allowable dambar protrusion shall be 0.10mm (0.004 inch) total in excess of "B" dimension at maximum material condition.
10. Controlling dimension: INCHES. Converted millimeter dimensions are not necessarily exact.

M24.15

24 LEAD THIN SHRINK NARROW BODY SMALL OUTLINE PLASTIC PACKAGE

SYMBOL	INCHES		MILLIMETERS		NOTES
	MIN	MAX	MIN	MAX	
A	0.053	0.069	1.35	1.75	-
A1	0.004	0.010	0.10	0.25	-
A2	-	0.061	-	1.54	-
B	0.008	0.012	0.20	0.30	9
C	0.007	0.010	0.18	0.25	-
D	0.337	0.344	8.55	8.74	3
E	0.150	0.157	3.81	3.98	4
e	0.025 BSC		0.635 BSC		-
H	0.228	0.244	5.80	6.19	-
h	0.0099	0.0196	0.26	0.49	5
L	0.016	0.050	0.41	1.27	6
N	24		24		7
α	0°	8°	0°	8°	-

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