National Semiconductor

October 17, 2011

LM5118/LM5118Q

Wide Voltage Range Buck-Boost Controller

General Description

The LM5118 wide voltage range Buck-Boost switching regulator controller features all of the functions necessary to implement a high performance, cost efficient Buck-Boost regulator using a minimum of external components. The Buck-Boost topology maintains output voltage regulation when the input voltage is either less than or greater than the output voltage making it especially suitable for automotive applications. The LM5118 operates as a buck regulator while the input voltage is sufficiently greater than the regulated output voltage and gradually transitions to the buck-boost mode as the input voltage approaches the output. This dual mode approach maintains regulation over a wide range of input voltages with optimal conversion efficiency in the buck mode and a glitch-free output during mode transitions. This easy to use controller includes drivers for the high side buck MOSFET and the low side boost MOSFET. The regulator's control method is based upon current mode control utilizing an emulated current ramp. Emulated current mode control reduces noise sensitivity of the pulse-width modulation circuit, allowing reliable control of the very small duty cycles necessary in high input voltage applications. Additional protection features include current limit, thermal shutdown and an enable input. The device is available in a power enhanced TSSOP-20 package featuring an exposed die attach pad to aid thermal dissipation.

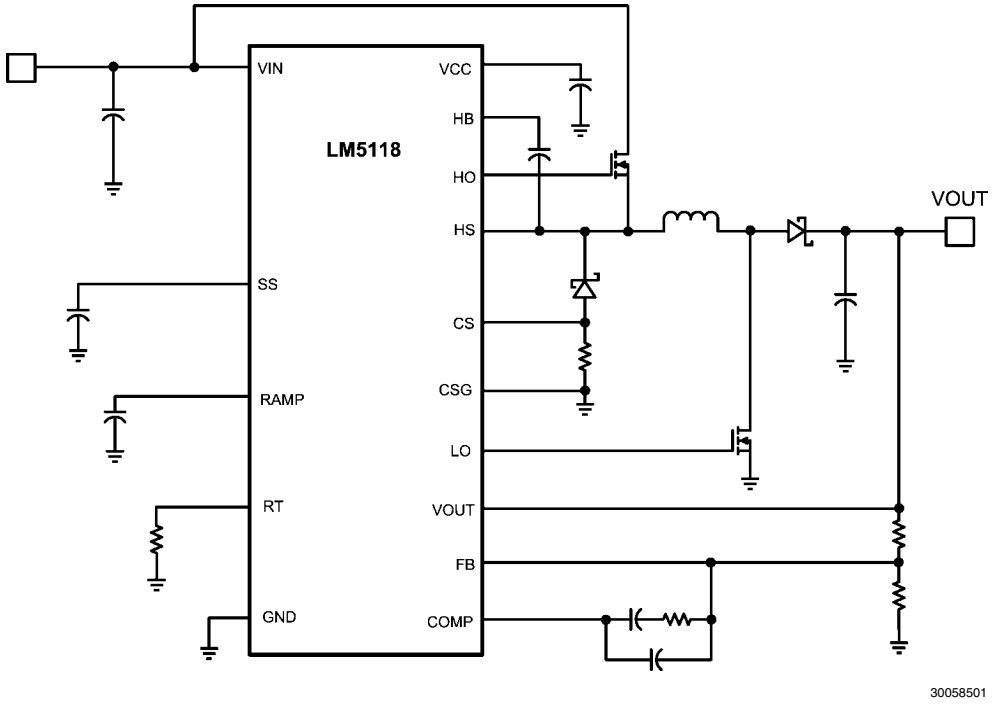
Features

- LM5118Q is an Automotive Grade product that is AEC-Q100 grade 1 qualified (-40°C to 125°C operating junction temperature)
- Ultra-wide input voltage range from 3V to 75V
- Emulated peak current mode control
- Smooth transition between step-down and step- up modes
- Switching frequency programmable to 500KHz
- Oscillator synchronization capability
- Internal high voltage bias regulator
- Integrated high and low-side gate drivers
- Programmable soft-start time
- Ultra low shutdown current
- Enable input wide bandwidth error amplifier
- 1.5% feedback reference accuracy
- Thermal shutdown

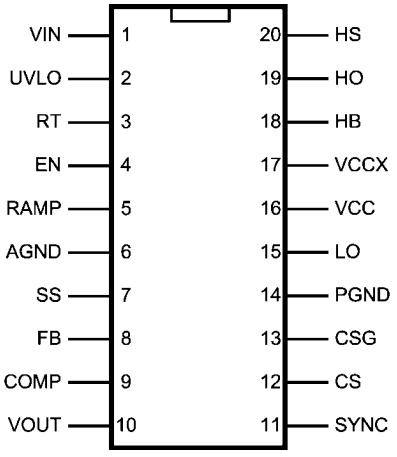
Package

TSSOP-20EP (Exposed pad)

Typical Application Circuit



Connection Diagram



Top View
See NS Package Numbers MXA20A

Ordering Information

Ordering Number	Package Type	NSC Package Drawing	Supplied As	Feature
LM5118MH	TSSOP-20EP	MXA20A	73 Units Per Anti-Static Tube	
LM5118MHX	TSSOP-20EP	MXA20A	2500 units shipped as Tape & Reel	
LM5118Q1MH	TSSOP-20EP	MXA20A	73 Units Per Anti-Static Tube	AEC-Q100 Grade 1 qualified. Automotive Grade Production Flow*
LM5118Q1MHX	TSSOP-20EP	MXA20A	2500 units shipped as Tape & Reel	

*Automotive Grade (Q) product incorporates enhanced manufacturing and support processes for the automotive market, including defect detection methodologies. Reliability qualification is compliant with the requirements and temperature grades defined in the AEC-Q100 standard. Automotive grade products are identified with the letter Q. For more information go to http://www.national.com/automotive.

Pin Descriptions

Pin	Name	Description
1	VIN	Input supply voltage.
2	UVLO	If the UVLO pin is below 1.23V, the regulator will be in standby mode (VCC regulator running, switching regulator disabled). When the UVLO pin exceeds 1.23V, the regulator enters the normal operating mode. An external voltage divider can be used to set an under-voltage shutdown threshold. A fixed 5 µA current is sourced out of the UVLO pin. If a current limit condition exists for 256 consecutive switching cycles, an internal switch pulls the UVLO pin to ground and then releases.
3	RT	The internal oscillator frequency is set with a single resistor between this pin and the AGND pin. The recommended frequency range is 50 kHz to 500 kHz.
4	EN	If the EN pin is below 0.5V, the regulator will be in a low power state drawing less than 10 µA from VIN. EN must be raised above 3V for normal operation.
5	RAMP	Ramp control signal. An external capacitor connected between this pin and the AGND pin sets the ramp slope used for emulated current mode control.
6	AGND	Analog ground.
7	SS	Soft-Start. An external capacitor and an internal 10 µA current source set the rise time of the error amp reference. The SS pin is held low when VCC is less than the VCC under-voltage threshold (< 3.7V), when the UVLO pin is low (< 1.23V), when EN is low (< 0.5V) or when thermal shutdown is active.
8	FB	Feedback signal from the regulated output. Connect to the inverting input of the internal error amplifier.
9	COMP	Output of the internal error amplifier. The loop compensation network should be connected between COMP and the FB pin.
10	VOUT	Output voltage monitor for emulated current mode control. Connect this pin directly to the regulated output.

Pin	Name	Description
11	SYNC	Sync input for switching regulator synchronization to an external clock.
12	CS	Current sense input. Connect to the diode side of the current sense resistor.
13	CSG	Current sense ground input. Connect to the ground side of the current sense resistor.
14	PGND	Power Ground.
15	LO	Boost MOSFET gate drive output. Connect to the gate of the external boost MOSFET.
16	VCC	Output of the bias regulator. Locally decouple to PGND using a low ESR/ESL capacitor located as close to the controller as possible.
17	VCCX	Optional input for an externally supplied bias supply. If the voltage at the VCCX pin is greater than 3.9V, the internal VCC regulator is disabled and the VCC pin is internally connected to VCCX pin supply. If VCCX is not used, connect to AGND.
18	HB	High side gate driver supply used in bootstrap operation. The bootstrap capacitor supplies current to charge the high side MOSFET gate. This capacitor should be placed as close to the controller as possible and connected between HB and HS.
19	HO	Buck MOSFET gate drive output. Connect to the gate of the high side buck MOSFET through a short, low inductance path.
20	HS	Buck MOSFET source pin. Connect to the source terminal of the high side buck MOSFET and the bootstrap capacitor.
	EP	Solder to the ground plane under the IC to aid in heat dissipation.

Absolute Maximum Ratings (*Note 1*)

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

VIN, EN, VOUT to GND	-0.3V to 76V
VCC, LO, VCCX, UVLO to GND(*Note 5*)	-0.3 to 15V
HB to HS	-0.3 to 15V
HO to HS	-0.3 to HB+0.3V
HS to GND	-4V to 76V
CSG, CS to GND	-0.3V to +0.3V
RAMP, SS, COMP, FB, SYNC, RT to GND	-0.3 to 7V
ESD Rating	
HBM (*Note 2*)	2 kV
Storage Temperature Range	-55°C to +150°C
Junction Temperature	+150°C

Operating Ratings (*Note 1*)

VIN (*Note 4*)	3V to 75V
VCC, VCCX	4.75V to 14V
Junction Temperature	-40°C to +125°C

Electrical Characteristics

Limits in standard type are for T_J = 25°C only; limits in **boldface type** apply over the junction temperature range of -40°C to +125°C. Unless otherwise specified, the following conditions apply: VIN = 48V, VCCX = 0V, EN = 5V, RT = 29.11 kΩ, No load on LO and HO (*Note 3*).

Symbol	Parameter	Conditions	Min	Typ	Max	Units
VIN SUPPLY						
I_{BIAS}	VIN Operating Current	VCCX = 0V		4.5	**5.5**	mA
I_{BIASX}	VIN Operating Current	VCCX = 5V		1	**1.85**	mA
I_{STDBY}	VIN Shutdown Current	EN = 0V		1	**10**	µA
VCC REGULATOR						
$V_{CC(REG)}$	VCC Regulation	VCCX = 0V	**6.8**	7	**7.2**	V
$V_{CC(REG)}$	VCC Regulation	VCCX = 0V, VIN = 6V	**5**	5.25	**5.5**	V
	VCC Sourcing Current Limit	VCC = 0	**21**	35		mA
	VCCX Switch threshold	VCCX Rising	**3.68**	3.85	**4.02**	V
	VCCX Switch hysteresis			0.2		V
	VCCX Switch RDS(ON)	ICCX = 10 mA		5	**12**	Ω
	VCCX Switch Leakage	VCCX = 0V		0.5	**1**	µA
	VCCCX Pull-down Resistance	VCCX = 3V		70		kΩ
	VCC Under-Voltage Lockout Voltage	VCC Rising	**3.52**	3.7	**3.86**	V
	VCC Under-Voltage Hysteresis			0.21		V
	HB DC Bias current	HB-HS = 15V		205	**260**	µA
	VC LDO Mode Turn-off			10		V
EN INPUT						
$V_{IL\ max}$	EN Input Low Threshold				**0.5**	V
$V_{IH\ min}$	EN Input High Threshold		**3.00**			V
	EN Input Bias Current	VEN = 3V	**-1**		**1**	µA
	EN Input Bias Current	VEN = 0.5V	**-1**		**1**	µA
	EN Input Bias Current	VEN = 75V		50		µA
UVLO THRESHOLDS						
	UVLO Standby Threshold	UVLO Rising	**1.191**	1.231	**1.271**	V
	UVLO Threshold Hysteresis			0.105		V
	UVLO Pull-up Current Source	UVLO = 0V		5		µA
	UVLO Pull-down $R_{DS(ON)}$			100	**200**	Ω
SOFT- START						
	SS Current Source	SS = 0V	**7.5**	10.5	**13.5**	µA
	SS to FB Offset	FB = 1.23V		150		mV
	SS Output Low Voltage	Sinking 100 µA, UVLO = 0V		7		mV
ERROR AMPLIFIER						
V_{REF}	FB Reference Voltage	Measured at FB pin, FB = COMP	**1.212**	1.230	**1.248**	V
	FB Input Bias Current	FB = 2V		20	**200**	nA

Symbol	Parameter	Conditions	Min	Typ	Max	Units
	COMP Sink/Source Current		**3**			mA
A_{OL}	DC Gain			80		dB
f_{BW}	Unity Bain Bandwidth			3		MHz
PWM COMPARATORS						
$t_{HO(OFF)}$	Forced HO Off-time		**305**	400	**495**	ns
$T_{ON(MIN)}$	Minimum HO On-time			70		ns
	COMP to Comparator Offset			200		mV
OSCILLATOR (RT PIN)						
f_{SW1}	Frequency 1	RT = 29.11 kΩ	**178**	200	**224**	kHz
f_{SW2}	Frequency 2	RT = 9.525 kΩ	**450**	515	**575**	kHz
SYNC						
	Sync threshold falling			1.3		V
CURRENT LIMIT						
$V_{CS(TH)}$	Cycle-by-cycle Sense Voltage Threshold (CS-CSG)	RAMP = 0 Buck Mode	**-103**	-125	**-147**	mV
$V_{CS(THX)}$	Cycle-by-cycle Sense Voltage Threshold (CS-CSG)	RAMP = 0 Buck-Boost Mode	**-218**	-255	**-300**	mV
	CS Bias Current	CS = 0V		45	**60**	µA
	CSG Bias Current	CSG = 0V		45	**60**	µA
	Current Limit Fault Timer			256		cycles
RAMP GENERATOR						
I_{R1}	RAMP Current 1	VIN = 60V, VOUT = 10V	**245**	305	**365**	µA
I_{R2}	RAMP Current 2	VIN = 12V, VOUT = 12V	**95**	115	**135**	µA
I_{R3}	RAMP Current 3	VIN = 5V, VOUT = 12V	**65**	80	**95**	µA
	VOUT Bias Current	VOUT = 48V		245		µA
LOW SIDE (LO) GATE DRIVER						
V_{OLL}	LO Low-state Output Voltage	I_{LO} = 100 mA	**0.095**	0.14	**0.23**	V
V_{OHL}	LO High-state Output Voltage	I_{LO} = -100 mA $V_{OHL} = V_{CC}-V_{LO}$		0.25		V
	LO Rise Time	C-load = 1 nF, VCC = 8V		16		ns
	LO Fall Time	C-load = 1 nF, VCC = 8V		14		ns
I_{OHL}	Peak LO Source Current	V_{LO} = 0V, VCC = 8V		2.2		A
I_{OLL}	Peak LO Sink Current	V_{LO} = VCC = 8V		2.7		A
HIGH SIDE (HO) GATE DRIVER						
V_{OLH}	HO Low-state Output Voltage	I_{HO} = 100 mA	**0.1**	0.135	**0.21**	V
V_{OHH}	HO High-state Output Voltage	I_{HO} = -100 mA, $V_{OHH} = V_{HB}-V_{OH}$		0.25		V
	HO Rise Time	C-load = 1 nF, VCC = 8V		14		ns
	HO Fall Time	C-load = 1 nF, VCC = 8V		12		ns
I_{OHH}	Peak HO Source Current	V_{HO} = 0V, VCC = 8V		2.2		A
I_{OLH}	Peak HO Sink Current	V_{HO} = VCC = 8V		3.5		A
	HB-HS Under Voltage Lock-out			3		V
BUCK-BOOST CHARACTERISTICS						
	Buck-Boost Mode	Buck Duty Cycle (*Note 6*)	**69**	75	**80**	%
THERMAL						
T_{SD}	Thermal Shutdown Temp.			165		°C
	Thermal Shutdown Hysteresis			25		°C
θ_{JA}	Junction to Ambient			40		°C/W
θ_{JC}	Junction to Case			4		°C/W

Note 1: Absolute Maximum Ratings are limits beyond which damage to the device may occur. Operating Ratings indicate conditions for which the device is intended to be functional, but does not guarantee specific performance limits. For guaranteed specifications and test conditions see the Electrical Characteristics.

Note 2: The human body model is a 100pF capacitor discharged through a 1.5 kΩ resistor into each pin.

Note 3: Min and Max limits are 100% production tested at 25°C. Limits over the operating temperature range are guaranteed through correlation using Statistical Quality Control (SQC) methods. Limits are used to calculate National's Average Outgoing Quality Level (AOQL).

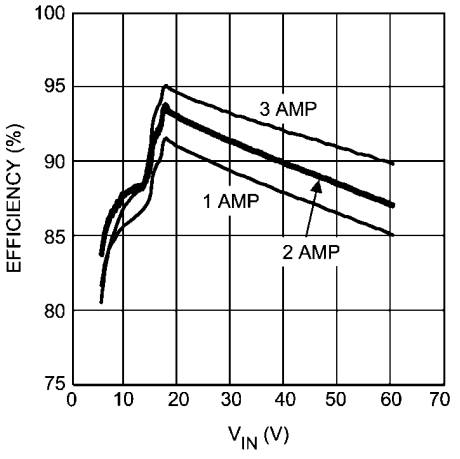
Note 4: 5V VIN is required to initially start the controller.

Note 5: These pins must not be raised above VIN.

Note 6: When the duty cycle exceeds 75%, the LM5118 controller gradually phases into the Buck-Boost mode.

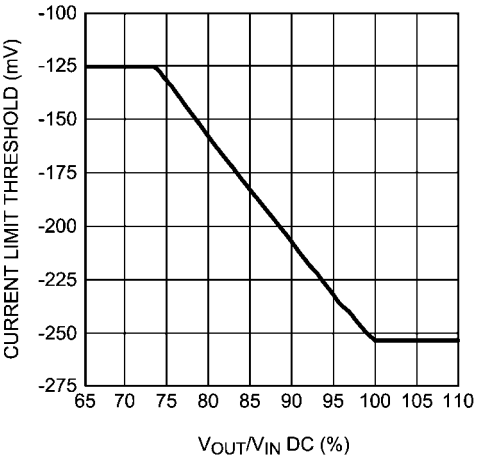
Typical Performance Characteristics

Efficiency vs VIN and IOUT
VOUT = 12V



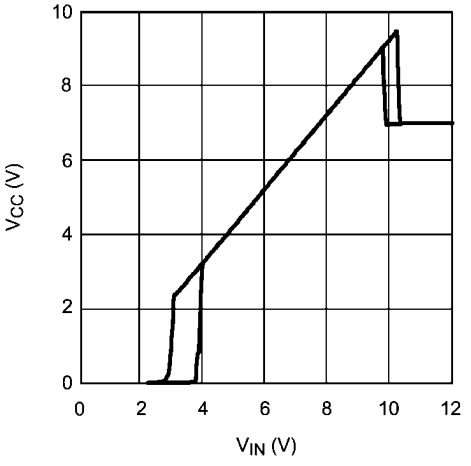
30058503

Current Limit Threshold vs VOUT/VIN
VOUT = 12V



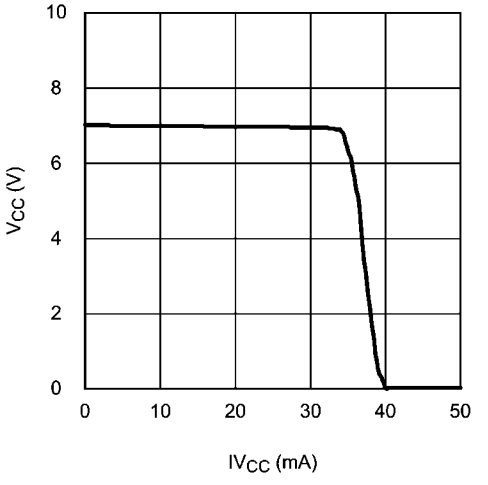
30058504

VCC vs VIN



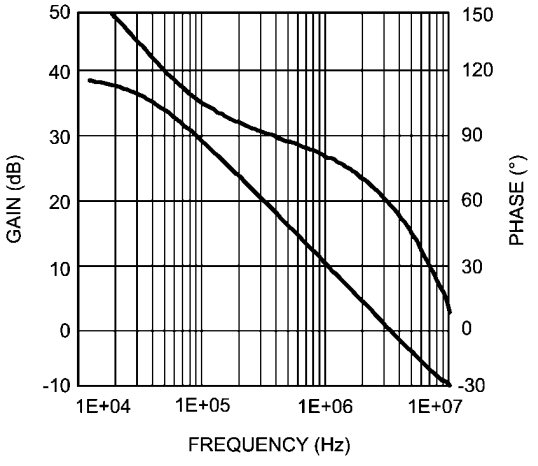
30058505

VCC vs IVCC



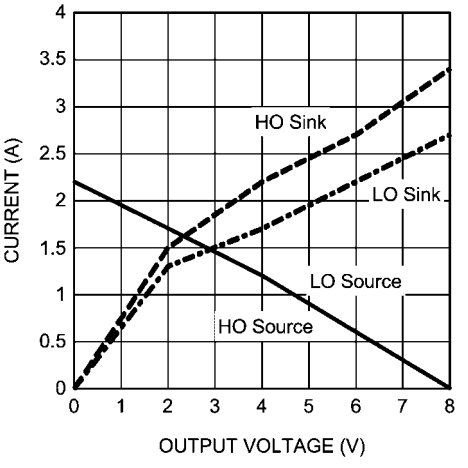
30058506

Error Amplifier Gain/Phase



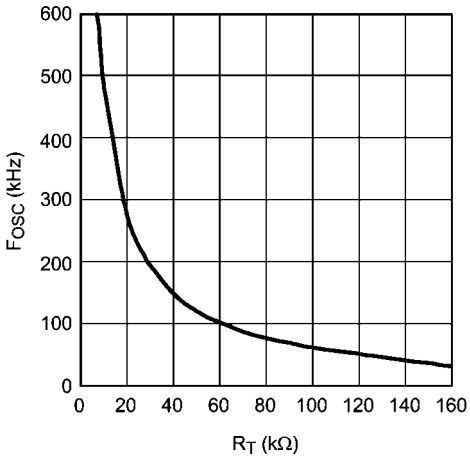
30058507

LO and HO Peak Gate Current vs Output Voltage
VCC = 8V



30058508

Oscillator Frequency vs RT



30058509

Block Diagram and Typical Application Circuit

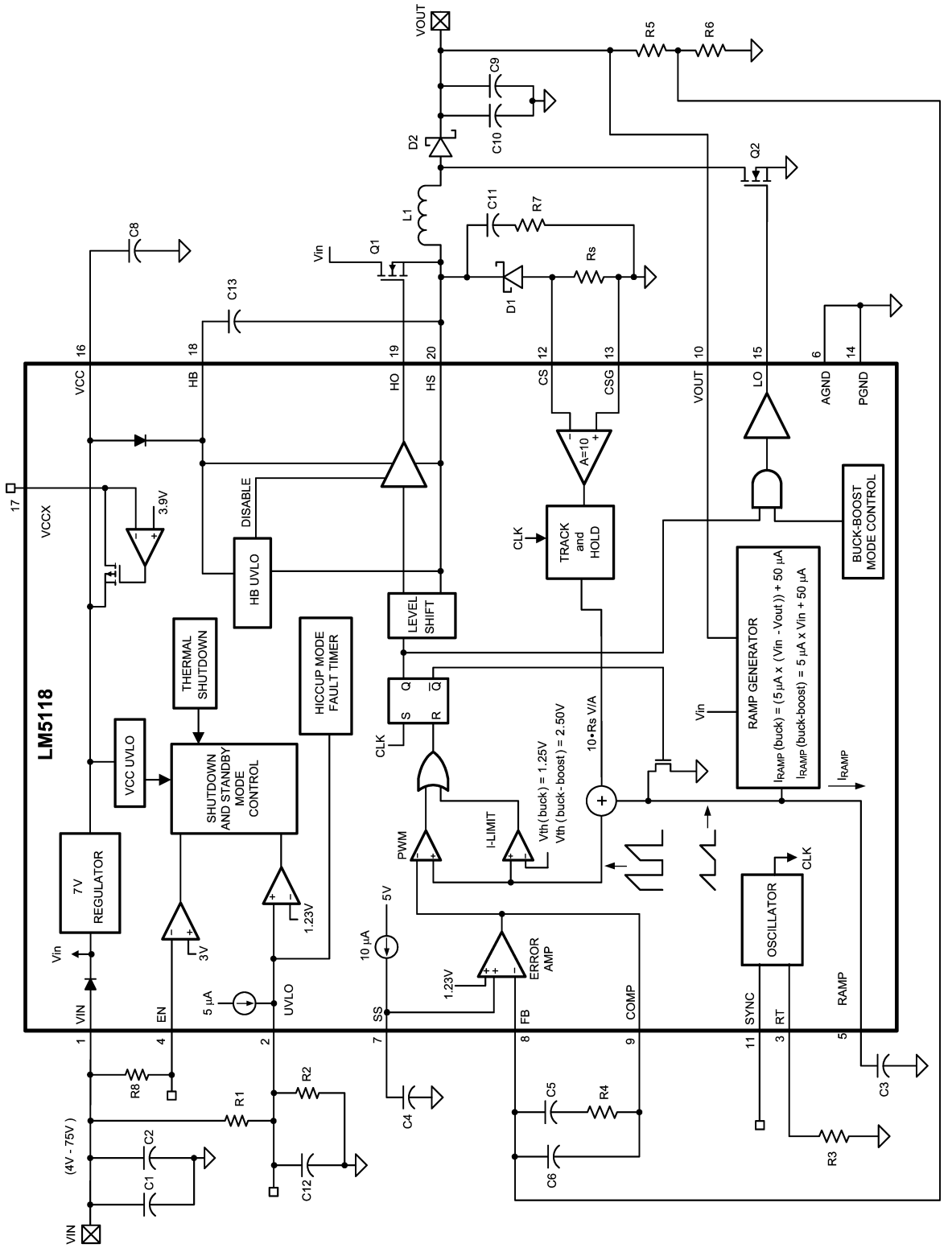


FIGURE 1.

Detailed Operating Description

The LM5118 high voltage switching regulator features all of the functions necessary to implement an efficient high voltage buck or buck-boost regulator using a minimum of external components. The regulator switches smoothly from buck to buck-boost operation as the input voltage approaches the output voltage, allowing operation with the input greater than or less than the output voltage. This easy to use regulator integrates high-side and low-side MOSFET drivers capable of supplying peak currents of 2 Amps. The regulator control method is based on current mode control utilizing an emulated current ramp. Peak current mode control provides inherent line feed-forward, cycle-by-cycle current limiting and ease of loop compensation. The use of an emulated control ramp reduces noise sensitivity of the pulse-width modulation circuit, allowing reliable processing of very small duty cycles necessary in high input voltage applications. The operating frequency is user programmable from 50 kHz to 500 kHz. An oscillator synchronization pin allows multiple LM5118 regulators to self synchronize or be synchronized to an external clock. Fault protection features include current limiting, thermal shutdown and remote shutdown capability. An undervoltage lockout input allows regulator shutdown when the input voltage is below a user selected threshold, and a low state at the enable pin will put the regulator into an extremely low current shutdown state. The device is available in the TSSOP-20EP package featuring an exposed pad to aid in thermal dissipation.

A buck-boost regulator can maintain regulation for input voltages either higher or lower than the output voltage. The challenge is that buck-boost power converters are not as efficient as buck regulators. The LM5118 has been designed as a dual mode controller whereby the power converter acts as a buck regulator while the input voltage is above the output. As the input voltage approaches the output voltage, a gradual transition to the buck-boost mode occurs. The dual mode approach maintains regulation over a wide range of input voltages, while maintaining the optimal conversion efficiency in the normal buck mode. The gradual transition between modes eliminates disturbances at the output during transitions. *Figure 2* shows the basic operation of the LM5118 regulator in the buck mode. In buck mode, transistor Q1 is active and Q2 is disabled. The inductor current ramps in proportion to the VIN - VOUT voltage difference when Q1 is active and ramps down through the recirculating diode D1 when Q1 is off. The first order buck mode transfer function is VOUT/VIN = D, where D is the duty cycle of the buck switch, Q1.

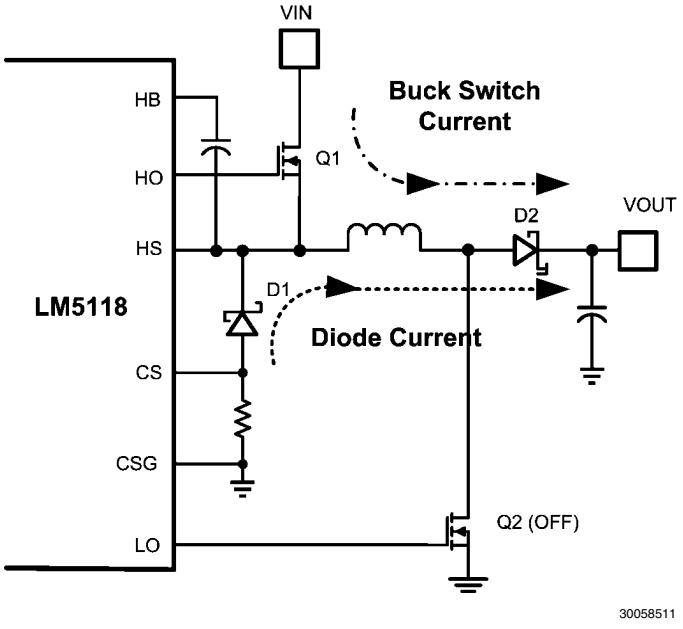


FIGURE 2. Buck Mode Operation

Figure 3 shows the basic operation of buck-boost mode. In buck-boost mode both Q1 and Q2 are active for the same time interval each cycle. The inductor current ramps up (proportional to VIN) when Q1 and Q2 are active and ramps down through the recirculating diode during the off time. The first order buck-boost transfer function is VOUT/VIN = D/(1-D), where D is the duty cycle of Q1 and Q2.

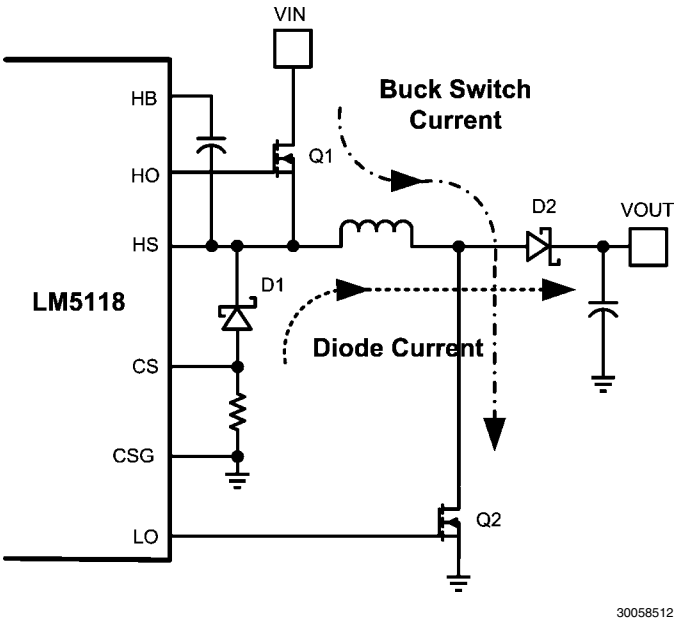


FIGURE 3. Buck-Boost Mode Operation

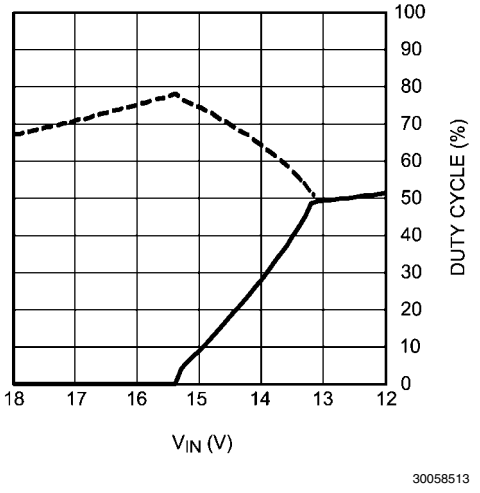


FIGURE 4. Mode Dependence on Duty Cycle (VOUT = 12V)

Operation Modes

Figure 4 illustrates how duty cycle affects the operational mode and is useful for reference in the following discussions. Initially, only the buck switch is active and the buck duty cycle increases to maintain output regulation as VIN decreases. When VIN is approximately equal to 15.5V, the boost switch begins to operate with a low duty cycle. If VIN continues to fall, the boost switch duty cycle increases and the buck switch duty cycle decreases until they become equal at VIN = 13.2V.

Buck Mode Operation: VIN > VOUT

The LM5118 buck-boost regulator operates as a conventional buck regulator with emulated current mode control while VIN is greater than VOUT and the buck mode duty cycle is less than 75%. In buck mode, the LO gate drive output to the boost switch remains low.

Buck-Boost Mode Operation: VIN ≅ VOUT

When VIN decreases relative to VOUT, the duty cycle of the buck switch will increase to maintain regulation. Once the duty cycle reaches 75%, the boost switch starts to operate with a very small duty cycle. As VIN is further decreased, the boost switch duty cycle increases until it is the same as the buck switch. As VIN is further decreased below VOUT, the buck and boost switch operate together with the same duty cycle and the regulator is in full buck-boost mode. This feature allows the regulator to transition smoothly from buck to buck-boost mode. It should be noted that the regulator can be designed to operate with VIN less than 4 volts, but VIN must be at least 5 volts during start-up *Figure 5* presents a timing illustration of the gradual transition from buck to buck-boost mode when the input voltage ramps downward over a few switching cycles.

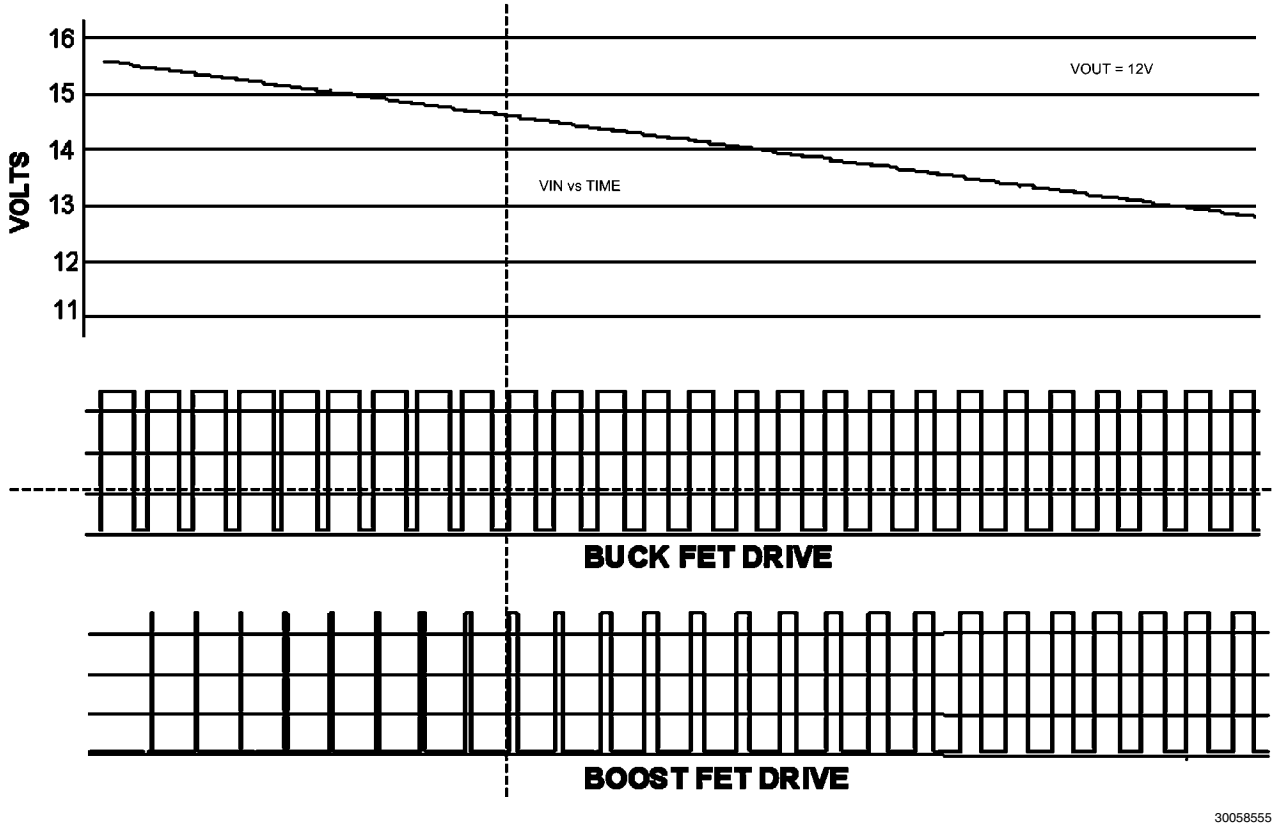


FIGURE 5. Buck (HO) and Boost (LO) Switch Duty Cycle vs. Time, Illustrating Gradual Mode Change with Decreasing Input Voltage

High Voltage Start-Up Regulator

The LM5118 contains a dual mode, high voltage linear regulator that provides the VCC bias supply for the PWM controller and the MOSFET gate driver. The VIN input pin can be connected directly to input voltages as high as 75V. For input voltages below 10V, an internal low dropout switch connects VCC directly to VIN. In this supply range, VCC is approximately equal to VIN. For VIN voltages greater than 10V, the low dropout switch is disabled and the VCC regulator is enabled to maintain VCC at approximately 7V. A wide operating range of 4V to 75V (with a startup requirement of at least 5 volts) is achieved through the use of this dual mode regulator.

The output of the VCC regulator is current limited to 35 mA, typical. Upon power up, the regulator sources current into the capacitor connected to the VCC pin. When the voltage at the VCC pin exceeds the VCC under-voltage threshold of 3.7V and the UVLO input pin voltage is greater than 1.23V, the gate driver outputs are enabled and a soft-start sequence begins. The gate driver outputs remain enabled until VCC falls below 3.5V or the voltage at the UVLO pin falls below 1.13V.

In many applications the regulated output voltage or an auxiliary supply voltage can be applied to the VCCX pin to reduce the IC power dissipation. For output voltages between 4V and 15V, VOUT can be connected directly to VCCX. When the voltage at the VCCX pin is greater than 3.85V, the internal VCC regulator is disabled and an internal switch connects VCCX to VCC, reducing the internal power dissipation.

In high voltage applications extra care should be taken to ensure the VIN pin voltage does not exceed the absolute maximum voltage rating of 76V. During line or load transients, voltage ringing on the VIN line that exceeds the absolute maximum rating can damage the IC. Both careful PC board layout and the use of quality bypass capacitors located close to the VIN and GND pins are essential.

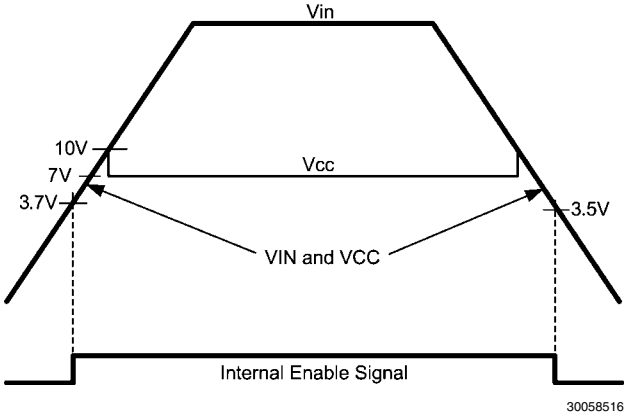


FIGURE 6. VIN and VCC Sequencing

Enable

The LM5118 contains an enable function which provides a very low input current shutdown mode. If the EN pin is pulled below 0.5V, the regulator enters shutdown mode, drawing less than 10 µA from the VIN pin. Raising the EN input above

3V returns the regulator to normal operation. The EN pin can be tied directly to the VIN pin if this function is not needed. It must not be left floating. A 1MΩ pull-up resistor to VIN can be used to interface with an open collector or open drain control signal.

UVLO

An under-voltage lockout pin is provided to disable the regulator when the input is below the desired operating range. If the UVLO pin is below 1.13V, the regulator enters a standby mode with the outputs disabled, but with VCC regulator operating. If the UVLO input exceeds 1.23V, the regulator will resume normal operation. A voltage divider from the input to ground can be used to set a VIN threshold to disable the regulator in brown-out conditions or for low input faults.

If a current limit fault exists for more than 256 clock cycles, the regulator will enter a "hiccup" mode of current limiting and the UVLO pin will be pulled low by an internal switch. This switch turns off when the UVLO pin approaches ground potential allowing the UVLO pin to rise. A capacitor connected to the UVLO pin will delay the return to a normal operating level and thereby set the off-time of the hiccup mode fault protection. An internal 5 µA pull-up current pulls the UVLO pin to a high state to ensure normal operation when the VIN UVLO function is not required and the pin is left floating.

Oscillator and Sync Capability

The LM5118 oscillator frequency is set by a single external resistor connected between the RT pin and the AGND pin. The R_T resistor should be located very close to the device and connected directly to the pins of the IC. To set a desired oscillator frequency (f), the necessary value for the R_T resistor can be calculated from the following equation:

$$R_T = \frac{6.4 \times 10^9}{f} - 3.02 \times 10^3$$

The SYNC pin can be used to synchronize the internal oscillator to an external clock. The external clock must be of higher frequency than the free-running frequency set by the R_T resistor. A clock circuit with an open drain output is the recommended interface from the external clock to the SYNC pin. The clock pulse duration should be greater than 15 ns.

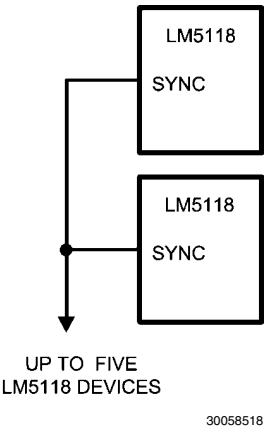


FIGURE 7. Sync from Multiple Devices

Multiple LM5118 devices can be synchronized together simply by connecting the SYNC pins together. In this configuration all of the devices will be synchronized to the highest frequency device. The diagram in *Figure 7* illustrates the SYNC input/output features of the LM5118. The internal oscillator circuit drives the SYNC pin with a strong pull-down/weak pull-up inverter. When the SYNC pin is pulled low, either by the internal oscillator or an external clock, the ramp cycle of the oscillator is terminated and forced 400 ns off-time is initiated before a new oscillator cycle begins. If the SYNC pins of several LM5118 IC's are connected together, the IC with the highest internal clock frequency will pull all the connected SYNC pins low and terminate the oscillator ramp cycles of the other IC's. The LM5118 with the highest programmed clock frequency will serve as the master and control the switching frequency of all the devices with lower oscillator frequencies.

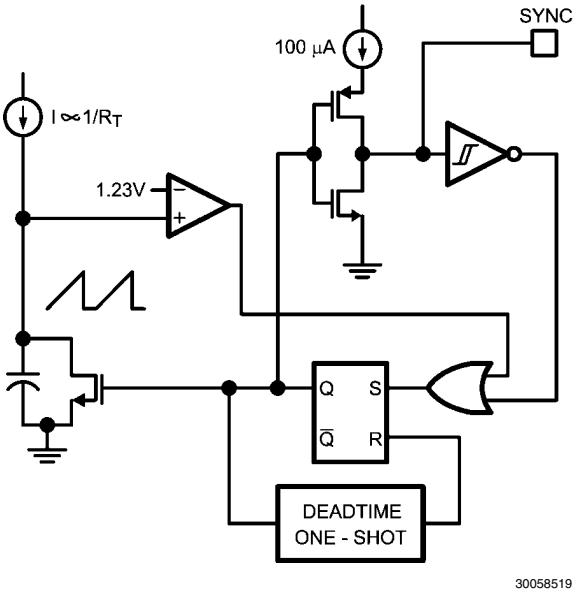


FIGURE 8. Simplified Oscillator and Block Diagram with Sync I/O Circuit

Error Amplifier and PWM Comparator

The internal high gain error amplifier generates an error signal proportional to the difference between the regulated output voltage and an internal precision reference (1.23V). The output of the error amplifier is connected to the COMP pin. Loop compensation components, typically a type II network illustrated in *Figure 1* are connected between the COMP and FB pins. This network creates a low frequency pole, a zero, and a noise reducing high frequency pole. The PWM comparator compares the emulated current sense signal from the RAMP generator to the error amplifier output voltage at the COMP pin. The same error amplifier is used for operation in buck and buck-boost mode.

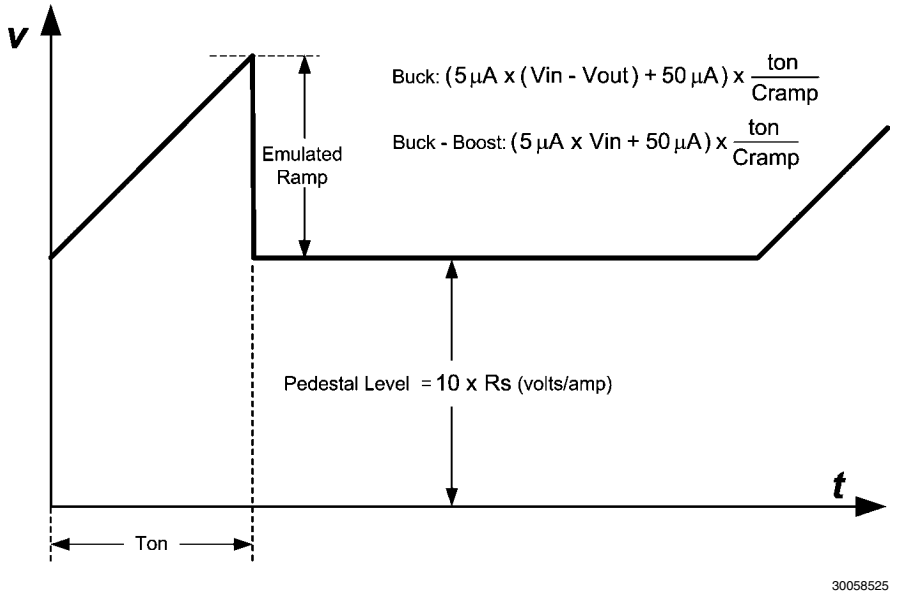


FIGURE 9. Composition of Emulated Current Signal

Ramp Generator

The ramp signal of a pulse-width modulator with current mode control is typically derived directly from the buck switch drain current. This switch current corresponds to the positive slope portion of the inductor current signal. Using this signal for the PWM ramp simplifies the control loop transfer function to a single pole response and provides inherent input voltage feed-forward compensation. The disadvantage of using the buck switch current signal for PWM control is the large leading edge spike due to circuit parasitics. The leading edge spike must be filtered or blanked to avoid early termination of the PWM pulse. Also, the current measurement may introduce significant propagation delays. The filtering, blanking time and propagation delay limit the minimal achievable pulse width. In applications where the input voltage may be relatively large in comparison to the output voltage, controlling a small pulse width is necessary for regulation. The LM5118 utilizes a unique ramp generator which does not actually measure the buck switch current but instead creates a signal representing or emulating the inductor current. The emulated ramp provides signal to the PWM comparator that is free of leading edge spikes and measurement or filtering delays. The current reconstruction is comprised of two elements, a sample-and-hold pedestal level and a ramp capacitor which is charged by a controlled current source. Refer to *Figure 9* for details.

The sample-and-hold pedestal level is derived from a measurement of the recirculating current through a current sense resistor in series with the recirculating diode of the buck regulator stage. A small value current sensing resistor is required between the recirculating diode anode and ground. The CS and CSG pins should be Kelvin connected directly to the sense resistor. The voltage level across the sense resistor is sampled and held just prior to the onset of the next conduction interval of the buck switch. The current sensing and sample-and-hold provide the DC level of the reconstructed current signal. The sample and hold of the recirculating diode current is valid for both buck and buck-boost modes. The positive slope inductor current ramp is emulated by an external capacitor connected from the RAMP pin to the AGND and an internal voltage controlled current source. In buck mode, the ramp current source that emulates the inductor current is a function of the VIN and VOUT voltages per the following equation:

$$I_{RAMP}\ (buck) = \frac{5\ \mu A}{V} \times (VIN - VOUT) + 50\ \mu A$$

In buck-boost mode, the ramp current source is a function of the input voltage VIN, per the following equation:

$$I_{RAMP}\ (buck - boost) = \frac{5\ \mu A}{V} \times VIN + 50\ \mu A$$

Proper selection of the RAMP capacitor (C_{RAMP}) depends upon the value of the output inductor (L) and the current sense resistor (R_S). For proper current emulation, the sample and hold pedestal value and the ramp amplitude must have the same relative relationship to the actual inductor current. That is:

$$R_S \times A = \frac{g_m \times L}{C_{RAMP}}$$

$$C_{RAMP} = \frac{g_m \times L}{A \times R_S}$$

Where g_m is the ramp generator transconductance (5 µA/V) and A is the current sense amplifier gain (10V/V). The ramp capacitor should be located very close to the device and connected directly to the RAMP and AGND pins.

The relationship between the average inductor current and the pedestal value of the sampled inductor current can cause instability in certain operating conditions. This instability is known as sub-harmonic oscillation, which occurs when the inductor ripple current does not return to its initial value by the start of the next switching cycle. Sub-harmonic oscillation is normally characterized by observing alternating wide and narrow pulses at the switch node. Adding a fixed slope voltage ramp (slope compensation) to the current sense signal prevents this oscillation. The 50µA of offset current provided from the emulated current source adds enough slope compensation to the ramp signal for output voltages less than or equal to 12V. For higher output voltages, additional slope compensation may be required. In such applications, the ramp capacitor can be decreased from the nominal calculated value to increase the ramp slope compensation.

The pedestal current sample is obtained from the current sense resistor (Rs) connected to the CS and CSG pins. It is sometimes helpful to adjust the internal current sense amplifier gain (A) to a lower value in order to obtain the higher current limit threshold. Adding a pair of external resistors RG in a series with CS and CSG as shown in *Figure 10* reduces the current sense amplifier gain A according to the following equation:

$$A = \frac{10k}{1k + R_G}$$

Current Limit

In the buck mode the average inductor current is equal to the output current (IOUT). In buck-boost mode the average inductor current is approximately equal to:

$$Iout \times \left(1 + \frac{VOUT}{VIN}\right)$$

Consequently, the inductor current in buck-boost mode is much larger especially when VOUT is large relative to VIN. The LM5118 provides a current monitoring scheme to protect the circuit from possible over-current conditions. When set correctly, the emulated current sense signal is proportional to the buck switch current with a scale factor determined by the current sense resistor. The emulated ramp signal is applied to the current limit comparator. If the peak of the emulated ramp signal exceeds 1.25V when operating in the buck mode, the PWM cycle is immediately terminated (cycle-by-cycle current limiting). In buck-boost mode the current limit threshold is increased to 2.50V to allow higher peak inductor current. To further protect the external switches during prolonged overload conditions, an internal counter detects consecutive cycles of current limiting. If the counter detects 256 consecutive current limited PWM cycles, the LM5118 enters a low power dissipation hiccup mode. In the hiccup mode, the output drivers are disabled, the UVLO pin is momentarily pulled low, and the soft-start capacitor is discharged. The regulator is restarted with a normal soft-start sequence once the UVLO pin charges back to 1.23V. The hiccup mode off-time can be programmed by an external capacitor connected from UVLO pin to ground. This hiccup cycle will repeat until the output overload condition is removed.

In applications with low output inductance and high input voltage, the switch current may overshoot due to the propagation delay of the current limit comparator and control circuitry. If an overshoot should occur, the sample-and-hold circuit will detect the excess recirculating diode current. If the sample-and-hold pedestal level exceeds the internal current limit threshold, the buck switch will be disabled and will skip PWM cycles until the inductor current has decayed below the current limit threshold. This approach prevents current runaway conditions due to propagation delays or inductor saturation since the inductor current is forced to decay before the buck switch is turned on again.

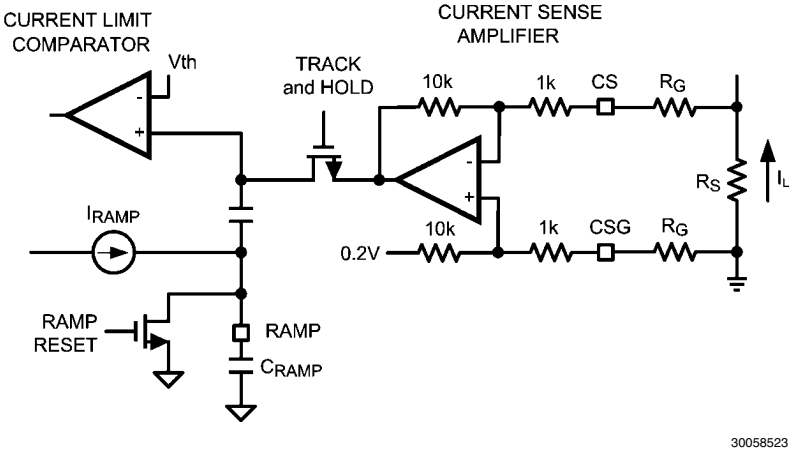


FIGURE 10. Current Limit and Ramp Circuit

Maximum Duty Cycle

Each conduction cycle of the buck switch is followed by a forced minimum off-time of 400ns to allow sufficient time for the recirculating diode current to be sampled. This forced off-time limits the maximum duty cycle of the controller. The actual maximum duty cycle will vary with the operating frequency as follows:

$$D_{MAX} = 1 - f \times 400 \times 10^{-9}$$

where f is the oscillator frequency in Hz

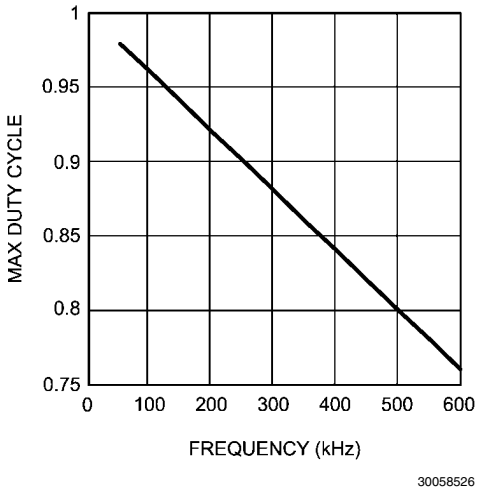


FIGURE 11. Maximum Duty Cycle vs Frequency

Limiting the maximum duty cycle will limit the maximum boost ratio (VOUT/VIN) while operating in buck-boost mode. For example, from *Figure 11*, at an operating frequency of 500 kHz, D_{MAX} is 80%. Using the buck-boost transfer function.

$$D = \frac{Vout}{Vin + Vout}$$

With D= 80%, solving for VOUT results in,

VOUT = 4 x VIN

With a minimum input voltage of 5 volts, the maximum possible output voltage is 20 volts at f = 500 kHz. The buck-boost step-up ratio can be increased by reducing the operating frequency which increases the maximum duty cycle.

Soft-Start

The soft-start feature allows the regulator to gradually reach the initial steady state operating point, thus reducing start-up stresses and surges. The internal 10 µA soft-start current source gradually charges an external soft-start capacitor connected to the SS pin. The SS pin is connected to the positive input of the internal error amplifier. The error amplifier controls the pulse-width modulator such that the FB pin approximately equals the SS pin as the SS capacitor is charged. Once the SS pin voltage exceeds the internal 1.23V reference voltage, the error amp is controlled by the reference instead of the SS pin. The SS pin voltage is clamped by an internal amplifier at a level of 150 mV above the FB pin voltage. This feature provides a soft-start controlled recovery in the event a severe overload pulls the output voltage (and FB pin) well below normal regulation but doesn't persist for 256 clock cycles.

Various sequencing and tracking schemes can be implemented using external circuits that limit or clamp the voltage level of the SS pin. The SS pin acts as a non-inverting input to the error amplifier anytime SS voltage is less than the 1.23V reference. In the event a fault is detected (over-temperature, VCC under-voltage, hiccup current limit), the soft-start capacitor will be discharged. When the fault condition is no longer present, a new soft-start sequence will begin.

HO Output

The LM5118 contains a high side, high current gate driver and associated high voltage level shift. This gate driver circuit works in conjunction with an internal diode and an external bootstrap capacitor. A 0.1 µF ceramic capacitor, connected with short traces between the HB pin and HS pin is recommended for most circuit configurations. The size of the bootstrap capacitor depends on the gate charge of the external FET. During the off time of the buck switch, the HS pin voltage is approximately -0.5V and the bootstrap capacitor is charged from VCC through the internal bootstrap diode. When operating with a high PWM duty cycle, the buck switch will be forced off each cycle for 400ns to ensure that the bootstrap capacitor is recharged.

Thermal Protection

Internal Thermal Shutdown circuitry is provided to protect the integrated circuit in the event the maximum junction temperature is exceeded. When activated, typically at 165°C, the controller is forced into a low power reset state, disabling the output driver and the bias regulator. This protection is provided to prevent catastrophic failures from accidental device overheating.

Application Information

The procedure for calculating the external components is illustrated with the following design example. The designations used in the design example correlate to the final schematic shown in *Figure 18*. The design specifications are:

- VOUT = 12V
- VIN = 5V to 75V
- F = 300 kHz
- Minimum load current (CCM operation) = 600 mA
- Maximum load current = 3A

R7 = R_T

RT sets the oscillator switching frequency. Generally speaking, higher operating frequency applications will use smaller components, but have higher switching losses. An operating frequency of 300 kHz was selected for this example as a reasonable compromise for both component size and efficiency. The value of R_T can be calculated as follows:

$$R_T = \frac{6.4 \times 10^9}{f} - 3.02 \times 10^3$$

therefore, R7 = 18.3 kΩ

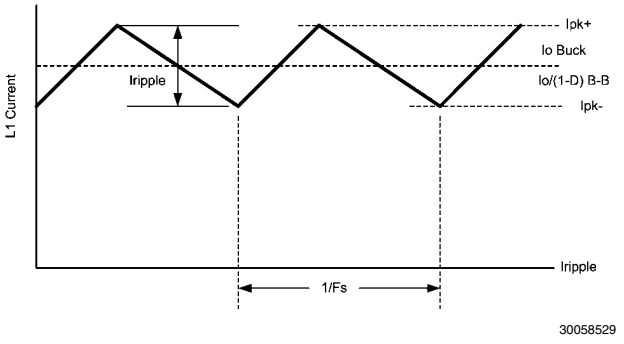


FIGURE 12. Inductor Current Waveform

INDUCTOR SELECTION

L1

The inductor value is determined based upon the operating frequency, load current, ripple current and the input and output voltages. Refer to *Figure 12* for details.

To keep the circuit in continuous conduction mode (CCM), the maximum ripple current IRIPPLE should be less than twice the minimum load current. For the specified minimum load of 0.6A, The maximum ripple current is 1.2A p-p. Also, the minimum value of L must be calculated both for a buck and buck-boost configurations. The final value of inductance will generally be a compromise between the two modes. It is desirable to have a larger value inductor for buck mode, but the saturation current rating for the inductor must be large for buck-boost mode, resulting in a physically large inductor. Additionally, large value inductors present buck-boost mode loop compensation challenges which will be discussed in error amplifier configuration section. For the design example, the inductor values in both modes are calculated as:

$$L1 = \frac{V_{OUT}\,(V_{IN1} - V_{OUT})}{V_{IN1} \times f \times I_{RIPPLE}} \quad \text{Buck Mode}$$

$$L1 = \frac{V_{IN2}\,(V_{OUT})}{(V_{OUT} + V_{IN2}) \times f \times I_{RIPPLE}} \quad \text{Buck-Boost Mode}$$

Where:

V_{OUT} is the output voltage

V_{IN1} is the maximum input voltage

f is the switching frequency

I_{RIPPLE} is the selected inductor peak to peak ripple current (1.2 A selected for this example)

V_{IN2} is the minimum input voltage

The resulting inductor values are:

L1 = 28 µH, Buck Mode

L1 = 9.8 µH Buck-Boost mode

A 10 µH inductor was selected which is a compromise between these values, while favoring the buck-boost mode. As will be illustrated in the compensation section below, the inductor value should be as low as possible to move the buck-boost right-half-plane zero to a higher frequency. The ripple current is then rechecked with the selected inductor value using the equations above,

$I_{RIPPLE(BUCK)}$ = 3.36A

$I_{RIPPLE(BUCK-BOOST)}$ = 1.17A

Because the inductor selected is lower than calculated for the Buck mode, the minimum load current for CCM in buck mode is 1.68 A at maximum VIN.

With a 10 µH inductor, the worst case peak inductor currents can be estimated for each case, assuming a 20% inductor value tolerance.

$$I_{1(PEAK)} = \frac{I_{OUT}}{0.8} + \frac{I_{RIPPLE}}{2} \quad \text{Buck Mode}$$

$$I_{2(PEAK)} = \frac{(V_{OUT} + Vin)I_{OUT}}{0.8 \times V_{IN2}} + \frac{I_{RIPPLE}}{2} \quad \text{Buck-Boost Mode}$$

For this example, the two equations yield:

$I_{1(PEAK)}$ = 5.43A

$I_{2(PEAK)}$ = 13.34A

An acceptable current limit setting would be 6.7A for buck mode since the LM5118 automatically doubles the current limit threshold in buck-boost mode. The selected inductor must have a saturation current rating at least as high as the buck-boost mode cycle-by-cycle current limit threshold, in this case at least 13.5A. A 10 µH 15 amp inductor was chosen for this application.

R13 = R_{SENSE}

To select the current sense resistor value, begin by calculating the value of R_{SENSE} for both modes of operation.

$$R13_{(BUCK)} = \frac{1.25V}{10 \times I_{PEAK}}$$

$R13_{(BUCK)}$ = 23 mΩ

For the buck-boost mode, R_{SENSE} is given by:

$$R13_{(BUCK\text{-}BOOST)} = \frac{2.5V}{10 \times I_{PEAK}}$$

$R13_{(BUCK\text{-}BOOST)}$ = 18.7 mΩ

A R_{SENSE} value of no more than 18.7 mΩ must be used to guarantee the required maximum output current in the buck-boost mode. A value of 15 mΩ was selected for component tolerances and is a standard value.

R13 = 15 mΩ

C15 = C_{RAMP}

With the inductor value selected, the value of C3 necessary for the emulation ramp circuit is:

$$C15 = C_{RAMP} = \frac{L \times 10^{-6}}{2 \times R_{SENSE}}$$

With the inductance value (L1) selected as 10 µH, the calculated value for C_{RAMP} is 333 pF. A standard value of 330 pF was selected.

C9 - C12 = OUTPUT CAPACITORS

In buck-boost mode, the output capacitors C9 - C12 must supply the entire output current during the switch on-time. For this reason, the output capacitors are chosen for operation in buck-boost mode, the demands being much less in buck operation. Both bulk capacitance and ESR must be considered

to guarantee a given output ripple voltage. Buck-boost mode capacitance can be estimated from:

$$C_{MIN} = \frac{I_{OUT} \times D_{MAX}}{f \times \Delta V_{OUT}} \quad \text{With } D_{MAX} = \frac{V_{OUT}}{V_{IN2} + V_{OUT}}$$

ESR requirements can be estimated from:

$$ESR_{MAX} = \frac{\Delta V_{OUT}}{I_{PEAK}}$$

For our example, with a ΔVOUT (output ripple) of 50 mV,

C_{MIN} = 141 µF

ESR_{MAX} = 3.8 mΩ

If hold-up times are a consideration, the values of input/output capacitors must be increased appropriately. Note that it is usually advantageous to use multiple capacitors in parallel to achieve the ESR value required. Also, it is good practice to put a .1 µF - .47 µF ceramic capacitor directly on the output pins of the supply to reduce high frequency noise. Ceramic capacitors have good ESR characteristics, and are a good choice for input and output capacitors. It should be noted that the effective capacitance of ceramic capacitors decreases with dc bias. For larger bulk values of capacitance, a low ESR electrolytic is usually used. However, electrolytic capacitors have poor tolerance, especially over temperature, and the selected value should be selected larger than the calculated value to allow for temperature variation. Allowing for component tolerances, the following values of COUT were chosen for this design example:

Two 180 µF Oscon electrolytic capacitors for bulk capacitance

Two 47 µF ceramic capacitors to reduce ESR

Two 0.47 µF ceramic capacitors to reduce spikes at the output .

D1

Reverse recovery currents degrade performance and decrease efficiency. For these reasons, a Schottky diode of appropriate ratings should be used for D1. The voltage rating of the boost diode should be equal to VOUT plus some margin. Since D1 only conducts during the buck switch off time in either mode, the current rating required is:

$I_{DIODE} = I_{OUT} \times (1\text{-}D)$ Buck Mode

$I_{DIODE} = I_{OUT}$ Buck-Boost Mode

D4

A Schottky type recirculating diode is required for all LM5118 applications. The near ideal reverse recovery characteristics and low forward voltage drop are particularly important diode characteristics for high input voltage and low output voltage applications. The reverse recovery characteristic determines how long the current surge lasts each cycle when the buck switch is turned on. The reverse recovery characteristics of Schottky diodes minimize the peak instantaneous power in the buck switch during the turn-on transition. The reverse breakdown rating of the diode should be selected for the maximum VIN plus some safety margin.

The forward voltage drop has a significant impact on the conversion efficiency, especially for applications with a low output voltage. “Rated” current for diodes vary widely from various manufacturers. For the LM5118 this current is user selectable through the current sense resistor value. Assuming a worst case 0.6V drop across the diode, the maximum diode power dissipation can be high. The diode should have a voltage rating of VIN and a current rating of IOUT. A conservative design would at least double the advertised diode rating since specifications between manufacturers vary. For the reference design a 100V, 10A Schottky in a D2PAK package was selected.

C1 - C5 = INPUT CAPACITORS

A typical regulator supply voltage has a large source impedance at the switching frequency. Good quality input capacitors are necessary to limit the ripple voltage at the VIN pin while supplying most of the switch current during the buck switch on-time. When the buck switch turns on, the current into the buck switch steps from zero to the lower peak of the inductor current waveform, then ramps up to the peak value, and then drops to the zero at turn-off. The RMS current rating of the input capacitors depends on which mode of operation is most critical.

$$I_{RMS(BUCK)} = I_{OUT} \sqrt{D(1\text{-}D)}$$

This value is a maximum at 50% duty cycle which corresponds to VIN = 24 volts.

$$I_{RMS(BUCK\text{-}BOOST)} = \frac{I_{OUT}}{1 - D} \sqrt{D(1\text{-}D)}$$

Checking both modes of operation we find:

$I_{RMS(BUCK)}$ = 1.5 Amps

$I_{RMS(BUCK\text{-}BOOST)}$ = 4.7 Amps

Therefore C1-C5 should be sized to handle 4.7A of ripple current. Quality ceramic capacitors with a low ESR should be selected. To allow for capacitor tolerances, four 2.2 µF, 100V ceramic capacitors will be used. If step input voltage transients are expected near the maximum rating of the LM5118, a careful evaluation of the ringing and possible spikes at the device VIN pin should be completed. An additional damping network or input voltage clamp may be required in these cases.

C20

The capacitor at the VCC pin provides noise filtering and stability for the VCC regulator. The recommended value of C20 should be no smaller than 0.1 µF, and should be a good quality, low ESR, ceramic capacitor. A value of 1 µF was selected for this design. C20 should be 10 x C8.

If operating without VCCX, then

$$f_{OSC} \times (Q_C \text{Buck + Boost}) + I_{LOAD(INTERNAL)}$$

must be less than the VCC current limit.

C8

The bootstrap capacitor between the HB and HS pins supplies the gate current to charge the buck switch gate at turn-on. The recommended value of C8 is 0.1 µF to 0.47 µF, and should be a good quality, low ESR, ceramic capacitor. A value of 0.1 µF was chosen for this design.

C16 = C_{SS}

The capacitor at the SS pin determines the soft-start time, i.e. the time for the reference voltage and the output voltage, to reach the final regulated value. The time is determined from:

$$t_{SS} = \frac{C16 \times 1.23V}{10\ \mu A}$$

and assumes a current limit>Iload + ICout

For this application, a C16 value of 0.1 µF was chosen which corresponds to a soft-start time of about 12 ms.

R8, R9

R8 and R9 set the output voltage level, the ratio of these resistors is calculated from:

$$\frac{R8}{R9} = \frac{V_{OUT}}{1.23V} - 1$$

For a 12V output, the R8/R9 ratio calculates to 9.76. The resistors should be chosen from standard value resistors and a good starting point is to select resistors within power ratings appropriate for the output voltage. Values of 309Ω for R9 and 2.67 kΩ for R8 were selected.

R1, R3, C21

A voltage divider can be connected to the UVLO pin to set a minimum operating voltage $VIN_{(UVLO)}$ for the regulator. If this feature is required, the easiest approach to select the divider resistor values is to choose a value for R1 between 10 kΩ and 100 kΩ, while observing the minimum value of R1 necessary to allow the UVLO switch to pull the UVLO pin low. This value is:

$R1 \geq 1000 \times V_{IN(MAX)}$

R1 ≥ 75k in our example

R3 is then calculated from

$$R3 = 1.23 \times \left[\frac{R1}{V_{IN(MIN)} + 5\ \mu A \times R1 - 1.23} \right]$$

Since $VIN_{(MIN)}$ for our example is 5V, set $VIN_{(UVLO)}$ to 4.0V for some margin in component tolerances and input ripple.

R1 = 75k is chosen since it is a standard value

R3 = 29.332k is calculated from the equation above. 29.4k was used since it is a standard value.

Capacitor C21 provides filtering for the divider and the off time of the "hiccup" duty cycle during current limit. The voltage at the UVLO pin should never exceed 15V when using an external set-point divider. It may be necessary to clamp the UVLO pin at high input voltages.

Knowing the desired off time during "hiccup" current limit, the value of C21 is given by:

$$t_{OFF} = -\left(\frac{R1 \times R3}{R1 + R3}\right) \times C21 \times \ln\left[1 - \frac{1.23 \times (R1 + R3)}{V_{IN} \times R1}\right]$$

Notice that t_{OFF} varies with V_{IN}

In this example, C21 was chosen to be 0.1 µF. This will set the t_{OFF} time to 956 µs with VIN = 12V.

R2

A 1M pull-up resistor connected from the EN pin to the VIN pin is sufficient to keep enable in a high state if on-off control is not used.

SNUBBER

A snubber network across the buck recirculating diode reduces ringing and spikes at the switching node. Excessive ringing and spikes can cause erratic operation and increase noise at the regulator output. In the limit, spikes beyond the maximum voltage rating of the LM5118 or the recirculating diode can damage these devices. Selecting the values for the snubber is best accomplished through empirical methods. First, make sure the lead lengths for the snubber connections are very short. Start with a resistor value between 5 and 20 Ohms. Increasing the value of the snubber capacitor results in more damping, however the snubber losses increase. Select a minimum value of the capacitor that provides adequate clamping of the diode waveform at maximum load. A snubber may be required for the boost diode as well. The same empirical procedure applies. Snubbers were not necessary in this example.

Error Amplifier Configuration

R4, C18, C17

These components configure the error amplifier gain characteristics to accomplish a stable overall loop gain. One advantage of current mode control is the ability to close the loop with only three feedback components, R4, C18 and C17. The overall loop gain is the product of the modulator gain and the error amplifier gain. The DC modulator gain of the LM5118 is as follows:

$$DCGain_{(MOD)} = \frac{R_{LOAD} \times V_{IN}}{10R_S(V_{IN} + 2V_{OUT})}$$

The dominant, low frequency pole of the modulator is determined by the load resistance (R_{LOAD}) and output capacitance (C_{OUT}). The corner frequency of this pole is:

$$f_{P(MOD)} = \frac{1 + D_{MIN}}{2\pi \times R_{LOAD} \times C_{OUT}}$$

For this example, R_{LOAD} = 4Ω, D_{MIN} = 0.294, and C_{OUT} = 454 µF, therefore:

$f_{P(MOD)}$ = 149 Hz

DC $Gain_{(MOD)}$ =3.63 = 11.2 dB

Additionally, there is a right-half plane (RHP) zero associated with the modulator. The frequency of the RHP zero is:

$$f_{RHPzero} = \frac{R_{LOAD}\ (1 - D)^2}{2\pi \times L \times D}$$

$f_{RHPzero}$ = 7.8 kHz

The output capacitor ESR produces a zero given by:

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

ESR_{ZERO} = 70 kHz

The RHP zero complicates compensation. The best design approach is to reduce the loop gain to cross zero at about 30% of the calculated RHP zero frequency. The Type II error amplifier compensation provided by R4, C18 and C17 places one pole at the origin for high DC gain. The 2nd pole should

be located close to the RHP zero. The error amplifier zero (see below) should be placed near the dominate modulator pole. This is a good starting point for compensation. Refer to the on-line LM5118 Quick-Start calculator for ready to use equations and more details.

Components R4 and C18 configure the error amplifier as a type II configuration which has a DC pole and a zero at

$$f_z = \frac{1}{2 \times \pi \times R4 \times C18}$$

C17 introduces an additional pole used to cancel high frequency switching noise. The error amplifier zero cancels the modulator pole leaving a single pose response at the crossover frequency of the loop gain if the crossover frequency is much lower than the right half plane zero frequency. A single pole response at the crossover frequency yields a very stable loop with 90 degrees of phase margin.

For the design example, a target loop bandwidth (crossover frequency) of 2.0 kHz was selected (about 30% of the right-half-plane zero frequency). The error amplifier zero (fz) should be selected at a frequency near that of the modulator pole and much less than the target crossover frequency. This constrains the product of R4 and C18 for a desired compensation network zero to be less than 2 kHz. Increasing R4, while proportionally decreasing C18 increases the error amp gain. Conversely, decreasing R4 while proportionally increasing C18 decreases the error amp gain. For the design example C18 was selected for 4.7 nF and R4 was selected to be 10 kΩ. These values set the compensation network zero at 149 Hz. The overall loop gain can be predicted as the sum (in dB) of the modulator gain and the error amp gain.

If a network analyzer is available, the modulator gain can be measured and the error amplifier gain can be configured for the desired loop transfer function. If a network analyzer is not available, the error amplifier compensation components can be designed with the guidelines given. Step load transient tests can be performed to verify acceptable performance. The step load goal is minimal overshoot with a damped response.

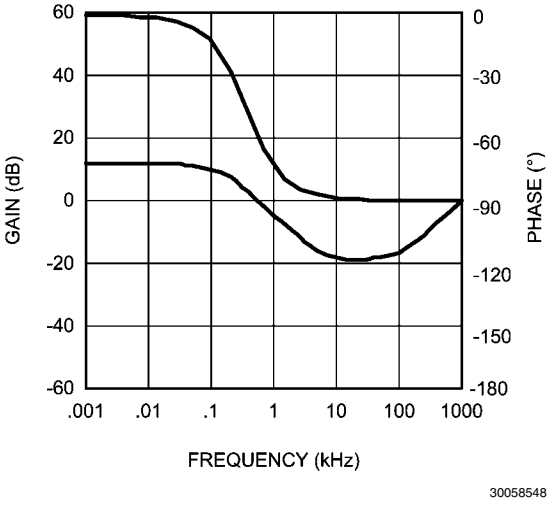


FIGURE 13. Modulator Gain and Phase

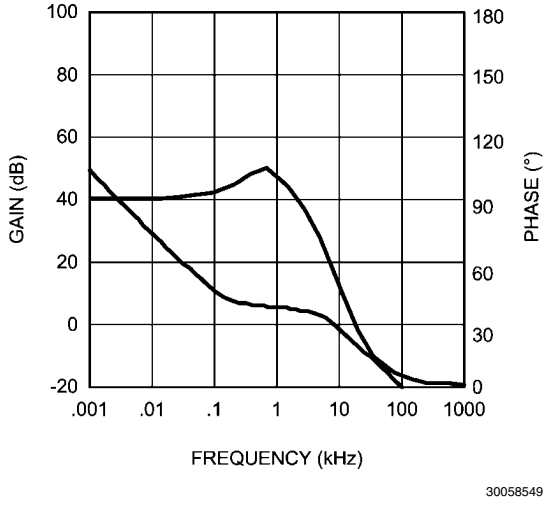


FIGURE 14. Error Amplifier Gain and Phase

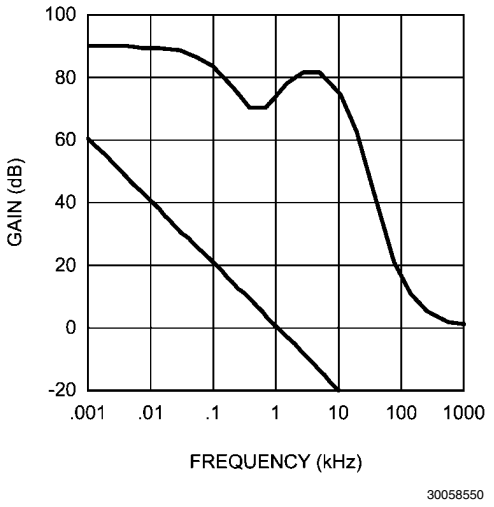


FIGURE 15. Overall Loop Gain and Phase

The plots shown in Figures 13, 14 and 15 illustrate the gain and phase diagrams of the design example. The overall bandwidth is lower in a buck-boost application due the compensation challenges associated with the right-half-plane zero. For a pure buck application, the bandwidth could be much higher. The LM5116 datasheet is a good reference for compensation design of a pure buck mode regulator.

Bias Power Dissipation Reduction

Buck or Buck-boost regulators operating with high input voltage can dissipate an appreciable amount of power while supplying the required bias current of the IC. The VCC regulator must step-down the input voltage VIN to a nominal VCC level of 7V. The large voltage drop across the VCC regulator translates into high power dissipation in the VCC regulator. There are several techniques that can significantly reduce this bias regulator power dissipation. Figures 16 and 17 depict two methods to bias the IC, one from the output voltage and one from a separate bias supply. In the first case, the internal VCC regulator is used to initially bias the VCC pin. After the output voltage is established, the VCC pin bias current is supplied through the VCCX pin, which effectively disables the internal VCC regulator. Any voltage greater than 4.0V can supply VCC bias through the VCCX pin. However, the voltage applied to the VCCX pin should never exceed 15V. The voltage supplied through VCCX must be large enough to drive the switching MOSFETs into full saturation.

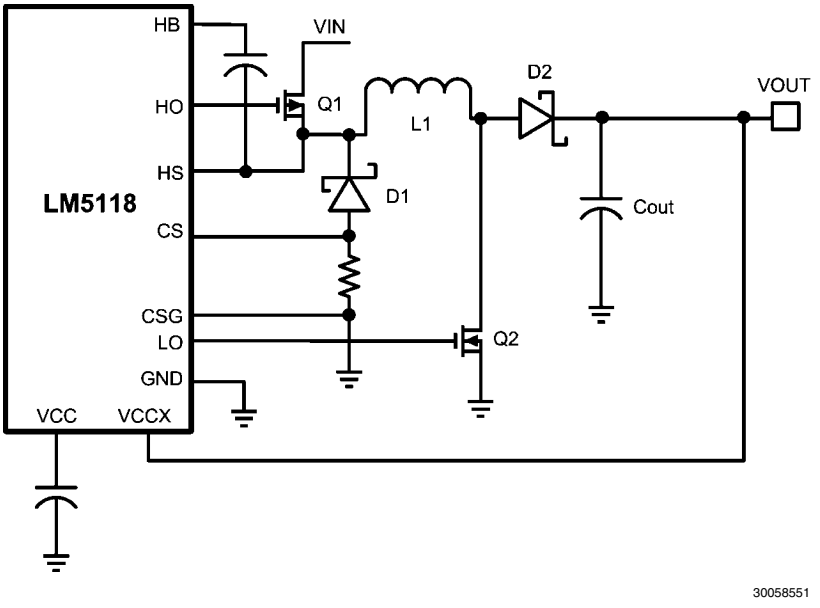


FIGURE 16. VCC Bias from VOUT 4V < VOUT < 15V

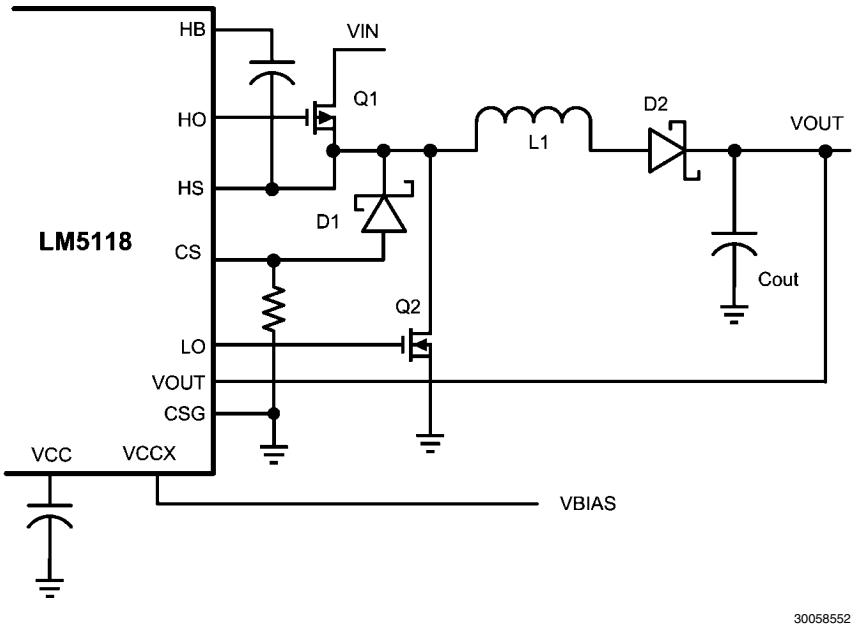


FIGURE 17. VCC Bias with Additional Bias Supply

PCB Layout and Thermal Considerations

In a buck-boost regulator, there are two loops where currents are switched very fast. The first loop starts from the input capacitors, and then to the buck switch, the inductor, the boost switch then back to the input capacitor. The second loop starts from the inductor, and then to the output diode, the output capacitor, the recirculating diode, and back to the inductor. Minimizing the PC board area of these two loops reduces the stray inductance and minimizes noise and the possibility of erratic operation. A ground plane in the PC board is recommended as a means to connect the input filter capacitors to the output filter capacitors and the PGND pins of the LM5118. Connect all of the low current ground connections (C_{SS}, R_T, C_{RAMP}) directly to the regulator AGND pin. Connect the AGND and PGND pins together through topside copper area covering the entire underside of the device. Place several vias in this underside copper area to the ground plane of the input capacitors.

The highest power dissipating components are the two power MOSFETs, the recirculating diode, and the output diode. The easiest way to determine the power dissipated in the MOSFETs is to measure the total conversion losses (P_{IN} - P_{OUT}), then subtract the power losses in the Schottky diodes, output inductor and any snubber resistors. An approximation for the recirculating Schottky diode loss is:

$P = (1\text{-}D) \times I_{OUT} \times V_{FWD}$.

The boost diode loss is

$P = I_{OUT} \times V_{FWD}$.

If a snubber is used, the power loss can be estimated with an oscilloscope by observation of the resistor voltage drop at both turn-on and turn-off transitions. The LM5118 package has an exposed thermal pad to aid power dissipation. Selecting diodes with exposed pads will aid the power dissipation of the diodes as well. When selecting the MOSFETs, pay careful attention to $R_{DS(ON)}$ at high temperature. Also, selecting MOSFETs with low gate charge will result in lower switching losses.

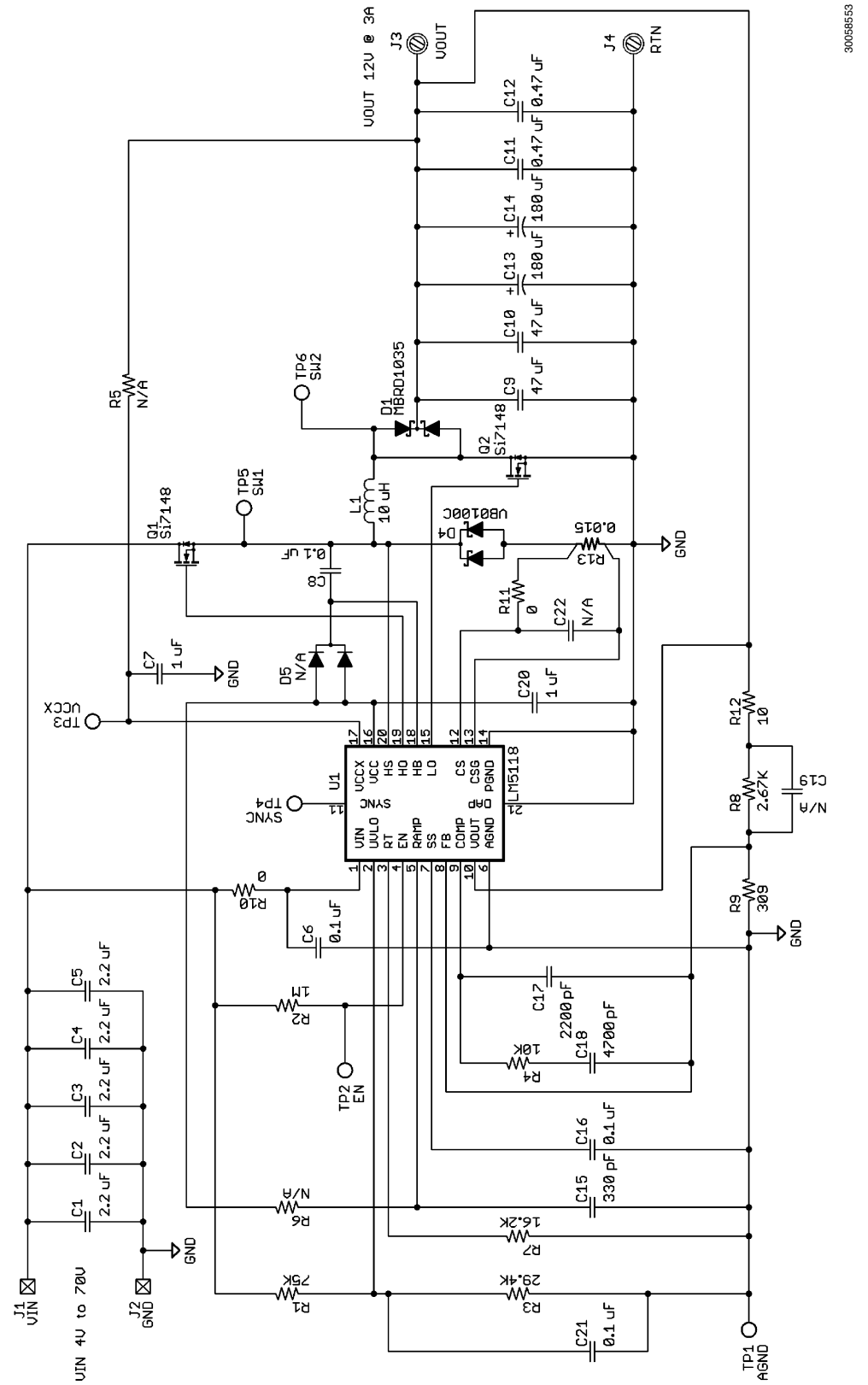
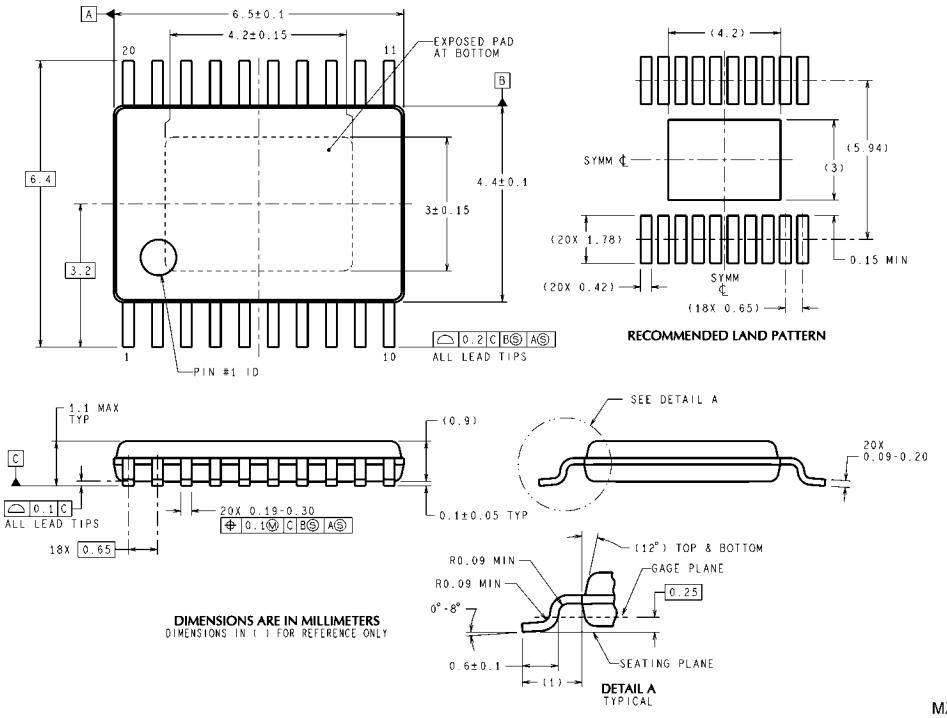


FIGURE 18. 12V, 3A Typical Application Schematic

Physical Dimensions inches (millimeters) unless otherwise noted



MXA20A (Rev C)

TSSOP-20EP Outline Drawing
NS Package Number MXA20A

Notes

Notes

For more National Semiconductor product information and proven design tools, visit the following Web sites at: www.national.com

Products		Design Support	
Amplifiers	www.national.com/amplifiers	WEBENCH® Tools	www.national.com/webench
Audio	www.national.com/audio	App Notes	www.national.com/appnotes
Clock and Timing	www.national.com/timing	Reference Designs	www.national.com/refdesigns
Data Converters	www.national.com/adc	Samples	www.national.com/samples
Interface	www.national.com/interface	Eval Boards	www.national.com/evalboards
LVDS	www.national.com/lvds	Packaging	www.national.com/packaging
Power Management	www.national.com/power	Green Compliance	www.national.com/quality/green
Switching Regulators	www.national.com/switchers	Distributors	www.national.com/contacts
LDOs	www.national.com/ldo	Quality and Reliability	www.national.com/quality
LED Lighting	www.national.com/led	Feedback/Support	www.national.com/feedback
Voltage References	www.national.com/vref	Design Made Easy	www.national.com/easy
PowerWise® Solutions	www.national.com/powerwise	Applications & Markets	www.national.com/solutions
Serial Digital Interface (SDI)	www.national.com/sdi	Mil/Aero	www.national.com/milaero
Temperature Sensors	www.national.com/tempsensors	SolarMagic™	www.national.com/solarmagic
PLL/VCO	www.national.com/wireless	PowerWise® Design University	www.national.com/training

National Semiconductor Americas Technical Support Center
Email: support@nsc.com
Tel: 1-800-272-9959

National Semiconductor Europe Technical Support Center
Email: europe.support@nsc.com

National Semiconductor Asia Pacific Technical Support Center
Email: ap.support@nsc.com

National Semiconductor Japan Technical Support Center
Email: jpn.feedback@nsc.com