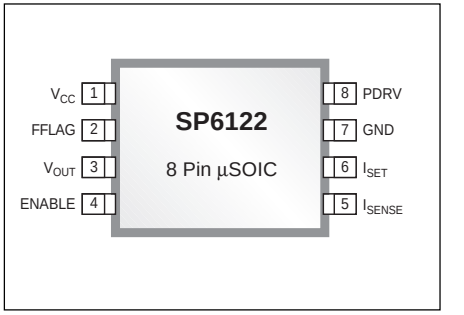
PRELIMINARY INFORMATION

SP6122

Low Voltage, Micro 8, PFET, Buck Controller

Ideal for 1A to 5A, Small Footprint, DC-DC Power Converters

- Optimized for Single Supply, 3V - 7V Applications
- High Efficiency, Greater than 90% Possible
- Small Micro 8 Package
- 20ns/1nF PFET Output Driver
- Fast Transient Response
- Open Drain Fault Output Pin
- Internal, 2ms, Soft Start Circuit (300kHz)
- Accurate 1.5% Reference
- Factory Programmable Output Voltage
- Factory Programmable Frequency, up to 600kHz
- Loss-less Adjustable Current Limit with High side $R_{DS(ON)}$ Sensing
- Hiccup or Lock-up Fault Modes
- Minimum On-Time, "Jitter & Frequency Stabilized" PFM Control: Simplifies Input and Output Filter design, provides great Light Load Efficiency and allows for Low Drop Out Regulation.
- Low 5µA Sleep Mode Quiescent Current
- Low 300µA Protected Mode Quiescent Current
- Ultra Low, 150µA Unprotected Mode Quiescent Current
- Output Over Voltage Protection

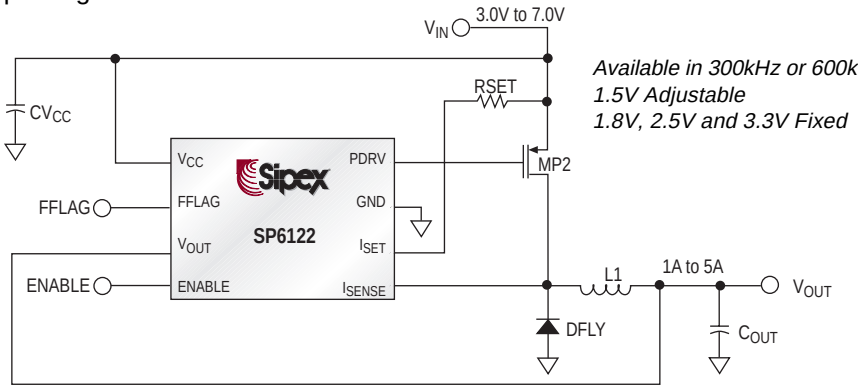


APPLICATIONS

- Video Cards
- High Power Portable
- Microcontrollers
- I/O & Logic
- Industrial Control
- Distributed Power
- Low Voltage Power

DESCRIPTION

The **SP6122** is a PWM/PFM minimum on-time controller designed to work from a single 5V or 3.3V input supply. It is engineered specifically for size and minimum components count, simplifying the transition from a linear regulator to a switcher solution. However, unlike other "micro" parts, the **SP6122** has an array of value added features like optional hiccup mode, over current protection, TTL enable, "jitter and frequency stabilization" and a fault flag pull down pin. Combined with reference and driver specifications usually found on more expensive integrated circuits, the **SP6122** delivers great performance and value in a micro 8 package.



Available in 300kHz or 600kHz
1.5V Adjustable
1.8V, 2.5V and 3.3V Fixed

ABSOLUTE MAXIMUM RATINGS

These are stress ratings only and functional operation of the device at these ratings or any other above those indicated in the operation sections of the specifications below is not implied. Exposure to absolute maximum rating conditions for extended periods of time may affect reliability.

V_{CC} 7V
All other pins ..... -0.3V to V_{CC}+0.3V
Peak Output Current < 10µs
PDRV 2A
Storage Temperature ..... -65°C to 150°C
Power Dissipation
Lead Temperature (Soldering, 10 sec) ..... 300°C
ESD Rating ..... 2kV HBM

SPECIFICATIONS

Unless otherwise specified: 0°C < T_{AMB} < 70°C, 3.0V < V_{CC} < 5.5V, C_{PDRV} = 1nF, V_{ENABLE} = V_{CC}, V_{FFLAG} = V_{CC}, I_{SET} = I_{SENSE} = V_{CC}, GND = 0V

PARAMETER	MIN	TYP	MAX	UNITS	CONDITIONS
QUIESCENT CURRENT					
V_{CC} Supply Current, OVC Enabled	-	300	400	µA	No Switching, I_{SET} = I_{SENSE} = V_{CC}
V_{CC} Supply Current, OVC Disabled	-	250	-	µA	No Switching, I_{SET} = I_{SENSE} = 0
V_{CC} Supply Current, OVC Disabled, Ultra Low IQ	-	150	-	µA	No Switching, ISET = 0, I_{SENSE}=V_{CC}
V_{CC} Supply Current, Sleep Mode	-	5	15	µA	Enable=0
REFERENCE					
Output Voltage, Initial Accuracy	VR*0.985	VR	VR*1.015	V	VR = Factory Set Voltage, see Note
Output Voltage, Over Line, Load and Temperature	VR*0.980	VR	VR*1.020	V	VR = Factory Set Voltage, see Note
PWM/PFM Reference Comparator Hysteresis	-	5	-	mV	Internal Hysteresis at Feedback Terminal
V_{OUT} Input Current	-	23	-	µA	V_{OUT} = VR
OSCILLATOR					
Oscillator Frequency	F*0.7	F	F*1.3	kHz	F = Factory Set Frequency, see Note Measured during Startup
Minimum Pulse Width during Startup (Blanking Time)	-	200	-	ns	
Soft Start					
Soft Start Ramp Time (600kHz part)	-	1	-	ms	V_{OUT} = VR – 30mV, Measure time from ENABLE = 1V to PDRV Low
Soft Start Ramp Time (300kHz part)	-	2	-	ms	V_{OUT} = VR – 30mV, Measure time from ENABLE = 1V to PDRV Low
Soft Start Voltage when PDRV Switches	-	250	-	mV	Measure VSoft Start when PDRV goes Low. (internal)

SPECIFICATIONS

Unless otherwise specified: 0°C < T_{AMB} < 70°C, 3.0V < V_{CC} < 5.5V, C_{PDRV} = 1nF, V_{ENABLE} = V_{CC}, V_{FFLAG} = V_{CC}, I_{SET} = I_{SENSE} = V_{CC}, GND = 0V

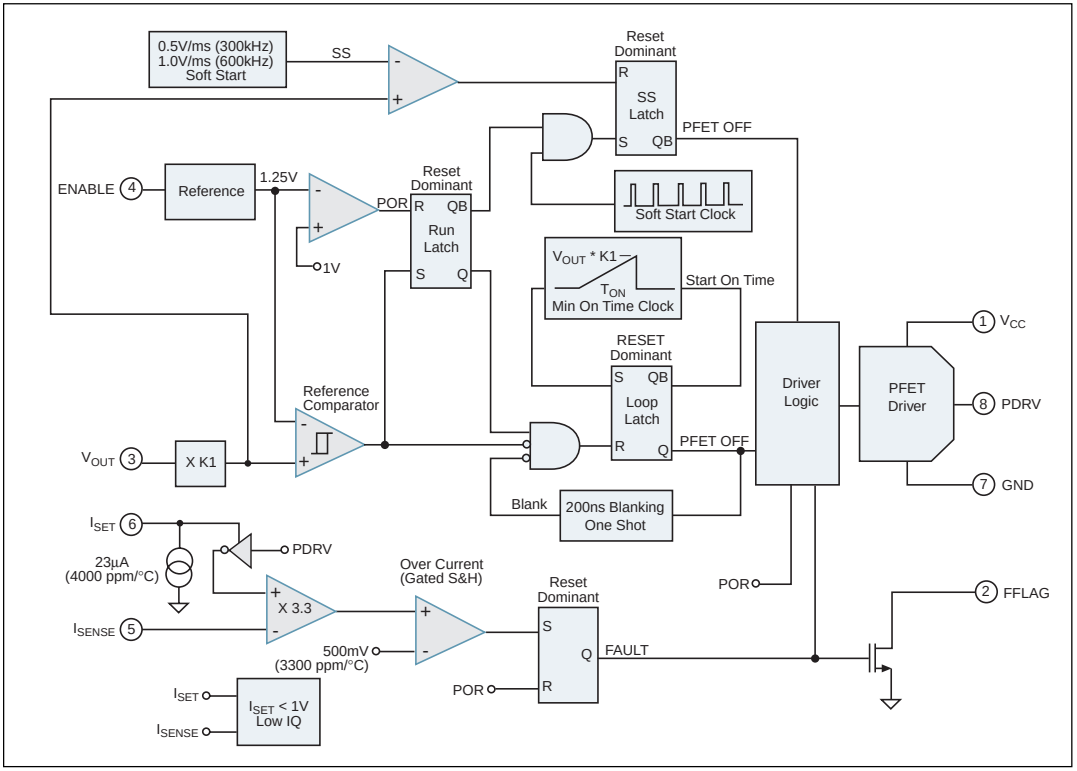
PARAMETER	MIN	TYP	MAX	UNITS	CONDITIONS
RDS OVER CURRENT COMPARATOR					
Over Current Comparator Threshold Voltage	130	150	180	mV	$V(I_{SET})$ - $V(I_{SENSE})$ 25°C only
Threshold Voltage Temperature Coefficient	-	3300	-	ppm/°C	
I_{SET} Sink Current	18	23	28	µA	Current into I_{SET} 25°C only
I_{SET} Current Temperature Coefficient	-	4000	-	ppm/°C	
I_{SENSE} Input Bias Current	-	-	100	nA	
I_{SET}, I_{SENSE} Common Mode Input Range	2.0	-	V_{CC}	V	
Over Current Peak Detection Time Constant	-	10	-	µs	
ENABLE INPUT & FFLAG OUTPUT					
ENABLE Threshold	-	1.1	-	V	
ENABLE Pin Source Current	2	5	10	µA	
FFLAG Sink Current	3	7.5		mA	V(FFLAG) = 1V
GATE DRIVER					
PDRV Rise Time		20	75	ns	0.5V to 4.5V
PDRV Fall Time		20	75	ns	4.5V to 0.5V

NOTE: Available Output Voltages: 1.5V Adj., 1.8V, 2.5V, 3.3V
Available Frequencies: 300kHz, 600kHz

PIN DESCRIPTION

PIN #	PIN NAME	DESCRIPTION
1	V_{CC}	Main Supply Pin: Decouple close to pin.
2	FFLAG	Fault Flag Pull-down Pin: Sinks current during a fault condition. Can be hooked up to ENABLE to initiate Hiccup Timing.
3	V_{OUT}	Regulated Output Voltage: This voltage is divided internally and compared to a 1.5%, 1.25V reference at the PWM/PFM comparator.
4	ENABLE	Enable Input: Floating this pin or pulling above 1.1V enables the part. Pulling this pin to less than 0.65V will disable the part. If FFLAG is hooked to ENABLE, a capacitor on ENABLE will control hiccup timing.
5	I_{SENSE}	Negative Input to the Over Current Amplifier/Comparator: This input is subtracted from the I_{SET} input and gained by a factor of 3.3. The output of this amplifier is compared with a 0.5V threshold, yielding a 150mV threshold. This threshold has a 3300 ppm/°C temperature coefficient. If the subtraction exceeds 150mV, charge is pumped into a capacitor until the capacitor hits $V_{CC}/2$. At this time, the over current fault is activated. If I_{SET} = 0V and I_{SENSE} = V_{CC}, the part enters an unprotected, 150µA quiescent current mode.
6	I_{SET}	Positive Input to the Over Current Amplifier: 23µA flows into the I_{SET} pin if it is pulled through a resistor to V_{IN}. This current has a 4000ppm/°C temperature coefficient and can be used via external resistor to raise the overcurrent trip point from 150mV to some higher value. If I_{SET} = 0V and I_{SENSE} = 0V, the part enters an unprotected, 250µA quiescent current mode.
7	GND	Power and Analog Ground: Hook directly to output ground.
8	PDRV	Drive for PFET High Side Switch: 1nF/20ns Output Driver.

BLOCK DIAGRAM



OPERATION

General Overview

The SP6122 is a minimum on-time, PFM controller for low cost DC/DC step down converters. The main control loop consists of a REFERENCE COMPARATOR, an ON-TIME CLOCK, a LOOP LATCH and a BLANKING ONESHOT. The REFERENCE comparator has 10mV of internal hysteresis and a 1.25V internal reference. Both hysteresis and reference voltage are multiplied upward by the internal feedback resistor divider, K1. This value is set by the factory and determines the output voltage of the converter. This divider is also used in the on-time algorithm for the controller. If the output voltage drops below K1*1.25V, then the DRIVER LOGIC tells the PFET switch to be "on" for a certain minimum time. The on-time is set by the Soft Start CLOCK frequency and is factory programmed to run at 300kHz or 600kHz. When the part is enabled, through V_{CC} or the ENABLE pin, the DRIVER LOGIC is configured to first look at the fixed frequency Soft Start loop. The output voltage is then controlled by a 0.5V/ms (300 kHz) internal ramp. When the output voltage reaches K1*1.25V, the Soft Start loop is switched off and the main loop takes over. In order for the main loop to appear to run at the same frequency as the fixed frequency Soft Start CLOCK, the on-time is modulated by a V_{OUT}/V_{CC} relationship. As a result, the SP6122 creates driver waveforms that look like PWM waveforms. In an effort to reduce jitter and enhance this "PWM-like" appearance, trailing edge blanking is incorporated to prevent spurious switching after the PFET switch has turned off.

Fault management is controlled either through power-on-reset (POR) or RDSon sense over current protection. Should an over current condition occur, the SP6122 will completely "lock-up" and turn the PFET switch off. The only way to recover will be to either cycle the ENABLE pin or V_{CC}. A Fault flag output (FFLAG) has been included to either signal the upstream circuitry or to engage a hiccup mode that will restart the SP6122. Tying FFLAG to ENABLE allows the controller to restart without assistance. Lastly, the SP6122 includes a powerful 4Ω PFET driver stage designed to drive a PFET associated with high speed converter designs in the 1 A – 5 A range.

Enable

Low quiescent mode or "Sleep Mode" is initiated by pulling the ENABLE pin below 650mV. The ENABLE pin has an internal 4µA pull-up current and does not require any external interface for normal operation. If the ENABLE pin is driven from a voltage source, the voltage must be above 1.1V in order to guarantee proper "awake" operation. Assuming that V_{CC} is above about 2.9V, the SP6122 transitions from "Sleep Mode" to "Awake Mode" in about 20µs – 30µs and from "Awake Mode" to "Sleep Mode" in a few microseconds. SP6122 quiescent current in sleep mode is 5µA typical. During Sleep Mode, the PFET switch is turned off, the internal SS voltage is held low and the FFLAG pin is high impedance.

Low Current Operation

If over current fault protection is not needed, the SP6122 offers two options to lower its quiescent current. By grounding both I_{SET} and I_{SENSE} pins, the circuitry responsible for over current detection is turned off. This option results in a saving of about 50µA in quiescent current. Option two requires that I_{SET} is grounded and ISENSE is greater than 1.3V. This option put the SP6122 in a low performance mode that cuts the operating frequency roughly in half and slows down critical comparators in the main loop. Option two can result in additional saving of 100µA bringing the total quiescent current to only 150µA (typ).

Power On Reset (POR)

The POR command is given every time the bandgap reference is started. The internal

1.25 V reference is compared against a 1V NFET threshold. When the reference is below the threshold, FAULT and RUN latches are reset, the internal SS voltage is discharged and the PFET switch is "off". The SP6122 is allowed to begin a soft start cycle when the internal 1.25V is greater than the 1 V threshold. Note this is a "loose" threshold and should not be used to guarantee under voltage lock out with respect to V_{CC}. Care should be take to ensure that V_{CC} does not "get stuck" on the way to its regulated value.

Soft Start

Soft start is required on step-down controllers to prevent excess inrush current through the power train during start-up. On the SP6122, this is managed through turning the PFET switch on with a fixed frequency clock and then turning the switch off when divided down version of the output voltage exceeds the internal SS voltage ramp. The internal SS voltage ramp rises with a 0.5 V/ms slew rate (300 kHz part) and the internal feedback voltage follows this rate of change. The presence of the output capacitor creates extra current draw during startup. Since dV_{OUT}/dt creates an average sustained current in the output capacitor, this current must be considered while calculating peak inrush current and over current thresholds. An expression to determine the excess inrush current due to the dVOUT/dt of the output capacitor is:

$$IC_{OUT} = C_{OUT}*K1*0.5\ V/ms$$

Lock Up & Hiccup Modes

As previously stated, if the SP6122 detects an over current condition and initiates a fault, the power supply remains "locked up". That is, the FFLAG pin immediately pulls low (if loaded) and the PFET switch turns off. This condition is permanent unless the either the V_{CC} or ENABLE is cycled. However if FFLAG is tied to ENABLE, the SP6122 will restart without assistance (Hiccup Mode). Furthermore, the restart time can be controlled by the addition of a small capacitor on the ENABLE pin to ground. The restart time is equal to the amount of time it takes for the 4µA ENABLE pin current to charge the external capacitor to an NFET threshold (roughly 1V). The waveforms that describe the Hiccup Mode operation are shown below.

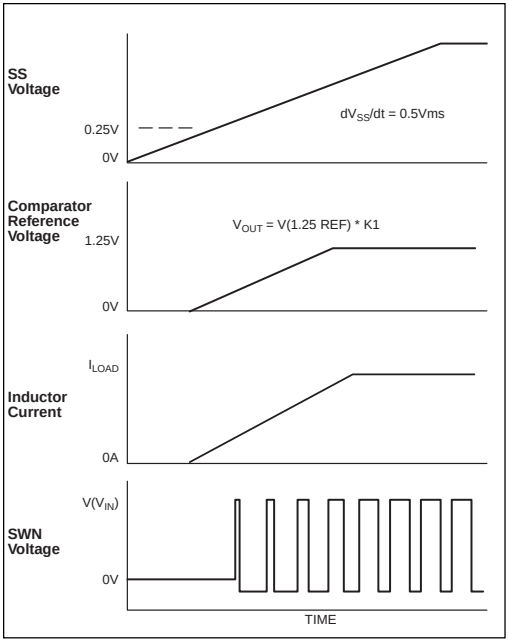


Figure 1:

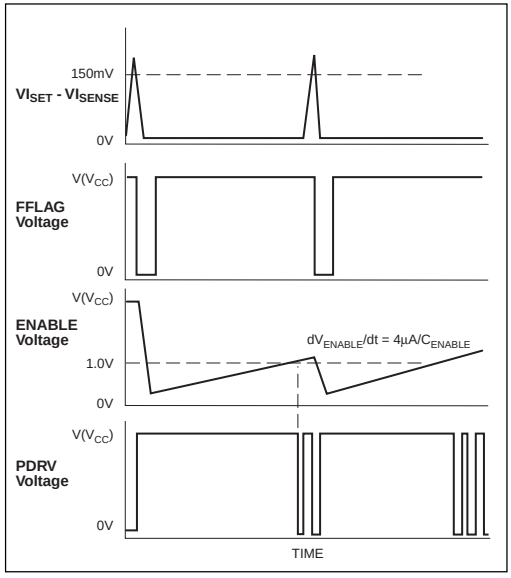


Figure 2:

Over Current Protection

Over current protection on the SP6122 is implemented through detection of an excess voltage condition across the PFET switch during conduction. This is typically referred to as high side RDSon detection. The over current comparator charges a sampling capacitor each time $V(I_{SET})$ – $V(I_{SENSE})$ exceeds 150mV (typ) and the PDRV voltage is low. The discharge current/charge current ratio on the sampling capacitor is about 2%. Therefore, provided that the over current condition persists, the capacitor voltage will be pumped up during each time PDRV switches low and this voltage will trigger an over current condition upon reaching a CMOS inverter threshold. There are many advantages to this approach. First, the filtering action of the gated S/H scheme protects against false triggering during a transient load condition or supply line noise. In addition, the total amount of time to trigger the fault depends on the on-time of the PFET switch. Ten, 1µs pulses are equivalent to twenty, 500ns pulses or one, 1µs pulse, however, depending on the period, each scenario takes a different amount of total time to trigger a fault. Therefore, the fault becomes an indicator of average power in the PFET switch. Also, because the CMOS trip threshold is dependent on V_{CC}, the over current scheme is protected against false triggering due to changes in line voltage.

Although the 150mV threshold is fixed, the overall RDSon detection voltage can be increased by placing a resistor from I_{SET} to V_{CC}. A 23µA sink current programs the additional voltage.

The 150 mV threshold and 23µA I_{SET} current have 3300 ppm/°C and 4000 ppm/°C temperature coefficients, respectively. These TC's are designed into the SP6122 in an effort to match the thermal characteristics of the PFET switch. It assumed that the SP6122 will be used in compact designs where there is a high amount of thermal coupling between the PFET and the controller.

Light Load Operation

One of the advantages of the SP6122 minimum on-time control scheme is the loop's ability to seamlessly and efficiently transition from heavy loads to light loads. In most other control schemes, the controller is notified about a light load condition and then must abruptly change control schemes in order to maintain efficiency. The SP6122 simply reduces the frequency when the average load current is less than the average inductor ripple current. As a result, switching loss decreases as the load current decreases and overall efficiency is maintained.

Output Driver

The driver stage consists of a high side, 4 ohm PFET driver. The following waveforms illustrate basic behavior of the driver.

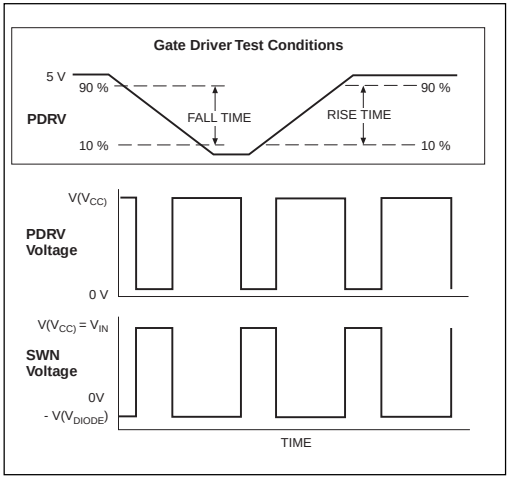


Figure 2:

APPLICATION INFORMATION

As an SP6122 application example, we will use the circuit from the SP6122 Evaluation Board Manual. This evaluation board uses the Sipex SP6122CU-A2, 1.5V adjustable, 300kHz PFET controller to realize a 3.3V to 1.9V step down converter. The board is optimized for 1A – 4A operation and has an RDSon over current trip threshold of about 7A. The body of the applications section contains:

- Data for the Evaluation Board
- Guidelines for Component Selection
- Features and Protection
- Layout Guidelines
- Introduction to the "Buck Cad Calculator"

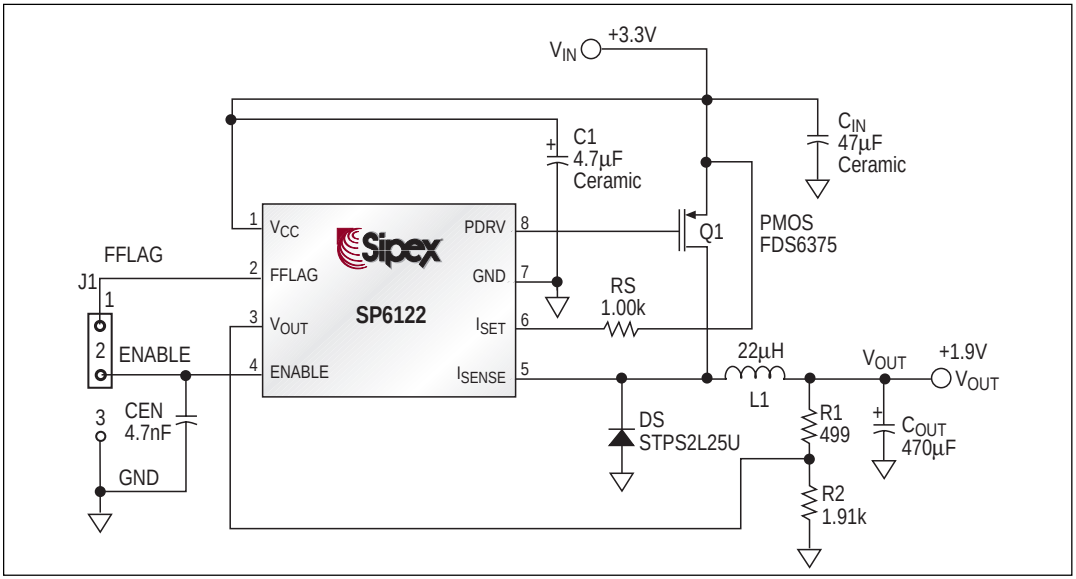


Figure 1. SP6122 Evaluation Board Application Schematic

Spreadsheet

Data For Evaluation Board

The SP6122 is engineered for size and minimum pin count, yet has a very accurate 2.0% reference over line, load and temperature. Figure 2 data shows a typical SP6122 Evaluation Board Efficiency plot, with efficiencies to 88% and output currents to 4A. Load Regulation plot in Figure 3 shows an essentially flat response of only 3mV change for up to 4A load. Figure 4 Line Regulation illustrates a 1.90V output that varies only 4mV or 0.2% for an input voltage change from 3.0V to 5.5V. While data on individual power supply boards may vary, the capability of the SP6122 of achieving high accuracy over load and line shown here is quite impressive and desirable for accurate power supply design.

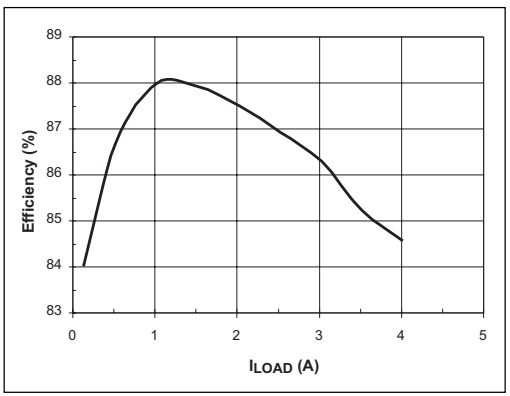


Figure 2. SP6122 Efficiency with V_{IN} = 3.3V, V_{OUT} = 1.9V.

Data For Evaluation Board: *continued*

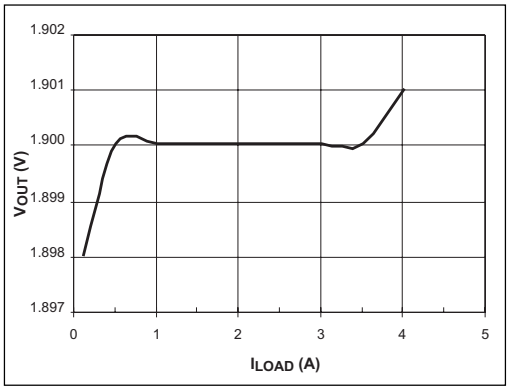


Figure 3. SP6122 Load Regulation with Input Voltage = 3.3V.

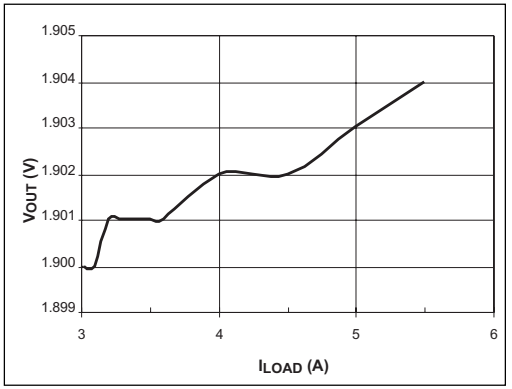


Figure 4. SP6122 Line Regulation with I_{LOAD} = 2A.

Guidelines for Component Selection

GENERAL

The SP6122 is a minimum on-time PFM controller. This means there is no error amp controlling the loop. Although an internal algorithm adjusts the on-time approximate the performance of a fixed frequency controller, the loop control is generated by looking at OUTPUT RIPPLE. The peak to peak value of this output ripple must be no less than 2% of the DC output voltage in order to maintain reasonable fixed frequency operation. In addition, as with all PFM controllers, board layout is critical and careful attention must be paid to minimize paths that can generate noise. Fortunately, the SP6122 is designed for simplicity and minimal external components, making it easy to design small, quiet power converters up to 12W.

INDUCTOR SELECTION

In a SP6122 application, the main factors for choosing an inductor are likely to be cost, size, saturation current and efficiency. If you use low inductor values, you get the smallest size, but you may cause larger ripple currents and poor efficiency and require more output capacitance to smooth the output ripple. Increasing the inductor value will decrease the output voltage ripple but degrade the transient response. For a good compromise between size, losses and cost, set the inductor ripple current between 20% to 40% of the maximum output current.

The inductor operating point and switching frequency determine the inductor value as seen in the following expression:

$$L = (V_{OUT} + V_{DIODE})*(V_{IN} - V_{OUT})/((V_{IN} + V_{DIODE})*(F_S\ K_R\ I_{OUT(max)}))$$

Where F_S = switching frequency (see Soft Start Frequency Specification)

K_R = ratio of the ac inductor ripple current to the maximum output current

V_{DIODE} = forward Schottky diode voltage

For an application with 1.9V out, 4A maximum I_{OUT}, 3.3V input supply, 400 mV typical forward diode voltage, 300kHz clock frequency and a 30% inductor ripple current, a 2.2µH inductor was selected (see Table 1 SP6122 Component Selection).

The peak to peak inductor ripple current is:

$$I_{PP} = (V_{OUT} + V_{DIODE})*(V_{IN} - V_{OUT})/((V_{IN} + V_{DIODE})*(F_S\ L))$$

For that same 2.2µH inductor application, the I_{PP} = 1.32A.

The inductor must be selected to not saturate the core at the peak inductor current:

$$I_{PEAK} = I_{OUT(max)} + I_{PP}/2$$

Guidelines for Component Selection: *continued*

Again, for that same 2.2µH application, I_{PEAK} = 4.6A. Therefore, a 2.2µH inductor with at least a 5A rating would be desired.

The type of core material to use must also be determined. For low cost, powdered iron cores can be used, and they have a gradual saturation characteristic, but they can cause ac core loss when the inductor value is low and ripple current is high. Ferrite cores, on the other hand, have an abrupt saturation characteristic and the inductor value drops sharply when the peak design current is exceeded. But, ferrites are preferred for high switching frequencies because they have low core losses as long as the saturation current is avoided.

Table 1 lists examples of both shielded and unshielded ferrite core inductors for applications appropriate for SP6122 applications from 2A to 5A output current. The inductors listed are both shielded and unshielded, the customer can decide what is needed for their application. For the SP6122 Evaluation Board, the unshielded ferrite inductor 2.2µH Coilcraft DO3316P-222 was selected for its cost/performance features.

INDUCTORS - SURFACE MOUNT *Note: Components highlighted in **bold** are those used on the SP6122 Evaluation Board.*

		INDUCTOR SPECIFICATION				
Inductance (µH)	Manufacturer/ Part No.	Series R (Ω)	Isat (A)	Size LxWxH (mm)	Inductor Type	Manufacturer Website
1.5	Coilcraft DO3316P-152	0.010	8.0	12.9x9.4x5.0	Unshielded Ferrite Core	www.coilcraft.com
2.2	**Coilcraft DO3316P-222**	**0.012**	**7.0**	**12.9x9.4x5.0**	**Unshielded Ferrite Core**	www.coilcraft.com
3.3	Coilcraft DO3316P-332	0.015	6.4	12.9x9.4x5.0	Unshielded Ferrite Core	www.coilcraft.com
1.5	Sumida CDRH104R-1R5	0.006	10.0	10x10x3.8	Shielded Ferrite Core	www.sumida.com
2.5	Sumida CDRH104R-2R5	0.008	7.5	10x10x3.8	Shielded Ferrite Core	www.sumida.com
3.8	Sumida CDRH104R-3R8	0.010	6.0	10x10x3.8	Shielded Ferrite Core	www.sumida.com
1.5	Murata LQN6C1R5M04	0.019	3.7	5.0x5.7x4.7	Unshielded Ferrite Core	www.murata.com
2.2	Murata LQN6C2R2M04	0.024	3.2	5.0x5.7x4.7	Unshielded Ferrite Core	www.murata.com
3.3	Murata LQN6C3R3M04	0.029	2.7	5.0x5.7x4.7	Unshielded Ferrite Core	www.murata.com

CAPACITORS - SURFACE MOUNT & THROUGH HOLE *Note: Components highlighted in **bold** are those used on the SP6122 Evaluation Board.*

		CAPACITOR SPECIFICATION					
Capacitance (µF)	Manufacturer/ Part No.	ESR Ω (max)	Ripple Current (A) @ 25°C	Size LxWxH (mm)	Voltage (V)	Capacitor Type	Manufacturer Website
470	**SANYO 6TPB470M**	**0.035**	**3.0**	**7343H**	**10.0**	**SMT Tant.**	www.sanyovideo.com
47	**TDK C4532X5R0J476M**	**0.005**	**4.0**	**1812**	**6.3**	**SMT X5R Cer.**	www.tdk.com
4.7	**TDK C3216X5R1C475M**	**0.020**	**4.0**	**1206**	**10.0**	**SMT X5R Cer.**	www.tdk.com
100	SANYO 16SA100M	0.030	2.7	8Dx10L	16.0	Thru-hole OS-CON	www.sanyovideo.com

PMOS SWITCH - SURFACE MOUNT *Note: Components highlighted in **bold** are those used on the SP6122 Evaluation Board.*

	PMOS SPECIFICATION					
Manufacturer/Part No.	$R_{DS(ON)}$ Ω @ 3.3V	Gate Charge nc @ 3.3V	Crss (pF)	Id (max) (A)	Package Type	Manufacturer Website
Fairchild FDS6375	**0.022**	**15**	**300**	**8**	**SO-8**	www.fairchildsemi.com
Siliconix SI4463DY	0.015	34	800	10	SO-8	www.siliconix.com
Intersil ITF86172SK8T	0.023	17	400	8	SO-8	www.intersil.com

SCHOTTKY DIODE - SURFACE MOUNT *Note: Components highlighted in **bold** are those used on the SP6122 Evaluation Board.*

	DIODE SPECIFICATION					
Manufacturer/Part No.	$V_{F\ @\ IF}$ (V)	$I_{F(AV)}$ (A)	Size LxWxH (mm)	Reverse V (V)	Package Type	Manufacturer Website
STMicro STPS2L25U	**0.50**	**4.0**	**5.5x3.9x2.5**	**25**	**SMB**	www.st.com
On-Semi MBRD835L	0.50	8.0	9.4x6.7x2.3	35	DPAK	www.onsemi.com

Table 1: SP6122 Component Selection

Guidelines for Component Selection: *continued*

The copper loss in the inductor can be calculated from the equation:

$P_{L(Cu)} = I_{L(RMS)}^2 R_{WINDING} \cong I_{OUT(max)}^2 R_{WINDING}$

For the 2.2µH example with 0.012Ω ESR in the winding, 4A load and 1.9V output, the copper loss in the inductor is 190mW.

OUTPUT CAPACITOR SELECTION

The output capacitor is typically selected based on its ability to maintain the output within specified tolerance limits during load transients. During an output load transient, the output capacitor must supply all the additional current demanded by the load until the SP6122 adjusts the inductor current to the new value. Therefore the capacitance must be large enough so that the output voltage is held up while the inductor current ramps up or down to the value corresponding to the new load current. For power converters delivering greater than 1A at less than 1MHz switching frequency, the output capacitor is typically greater than 100µF. Typically, tantalum and OSCON capacitors are used to get high output capacitance in a small space. These capacitors have a high Equivalent Series Resistance (ESR) when compared to ceramic capacitors and this ESR is both a curse and a blessing. Unfortunately, the ESR (Equivalent Series Resistance) in the output capacitor causes a step in the output voltage equal to the ESR value multiplied by the change in load current. As a result, in a power supply using a tantalum, aluminum electrolytics or OSCON output capacitor, the value of output capacitance (or number of output capacitors) is typically chosen to minimize the output variation due to the load step imposed on this ESR. However, the SP6122 takes advantage of the natural presence of this ESR to control the loop. Because the inductor ripple current also flows through this ESR, and output ripple voltage is created and the waveform is resembles a miniature current-mode timing waveform. For a 1.9V output voltage, the required ripple is a reasonable 38 mV. The designer must chose all other trade-offs wisely to maintain this ripple

$$0.02 * V_{OUT} < I_{PP} * R_{ESR}$$

and

$$\Delta I_{LOAD} * R_{ESR} < \Delta V_{TOL}$$

where:

V_{OUT} = DC output voltage

R_{ESR} = ESR of the output capacitor

DI_{LOAD} = change in current due to load step

DV_{TOL} = tolerable deviation due to load transient

I_{PP} = peak to peak inductor ripple current

Output ripple is due primarily to the output ripple current and the output capacitor ESR value as seen in the following equation:

$$\Delta V_{OUT} \cong I_{PP} R_{ESR}$$

For our SP6122 evaluation board example with ESR = 35mΩ and I_{PP} = 1.32A, ΔV_{OUT} = 46mV. Note that a 4A step creates a 140mV deviation. If this is unacceptable, ESR and I_{PP} must be reconsidered in order to improve step response and maintain output ripple.

Recommended capacitors that can be used effectively in SP6122 applications are: low-ESR aluminum electrolytic capacitors, OSCON capacitors that provide a very high performance/size ratio for electrolytic capacitors and low-ESR tantalum capacitors. AVX TPS series and Kemet T510 surface mount capacitors are popular tantalum capacitors that work well in SP6122 applications. POSCAP from Sanyo is a solid electrolytic chip capacitor that has low ESR and high capacitance. For the same ESR value, POSCAP has lower profile compared with a tantalum capacitor.

INPUT CAPACITOR SELECTION

The input capacitor should be selected for ripple current rating, capacitance and voltage rating. The input capacitor must meet the ripple current requirement imposed by the switching current. In continuous conduction mode, the source current of the high-side MOSFET is approximately a square wave of duty cycle V_{OUT}/V_{IN}. Most of this current is supplied by the input bypass capacitors. The RMS value of input capacitor current is determined at the maximum output current and under the assumption that the peak to peak inductor ripple current is low, it is given by:

$$I_{CIN(RMS)} = I_{OUT(MAX)}SQRT(D(1-D))$$

The worse case occurs when the duty cycle D is 50% and gives an RMS current value equal to $I_{OUT}/2$. Select input capacitors with adequate ripple current rating to ensure reliable operation. The power dissipated in the input capacitor is:

$$P_{CIN} = I_{CIN}^2{}_{(RMS)} R_{ESR(CIN)}$$

This can become a significant part of power losses in a converter and hurt the overall energy transfer efficiency. The input voltage ripple primarily depends on

the input capacitor ESR and capacitance. Ignoring the inductor ripple current, the input voltage ripple can be determined by:

$$\Delta V_{IN} = I_{OUT\ (MAX)} R_{ESR(CIN)} + I_{OUT(MAX)}V_{OUT}(V_{IN} - V_{OUT})/(F_S C_{IN} V_{IN}^2)$$

The capacitor type suitable for the output capacitors can also be used for the input capacitors. However, exercise extra caution when tantalum capacitors are considered. Tantalum capacitors are known for catastrophic failure when exposed to surge current, and input capacitors are prone to such surge current when power supplies are connected 'live' to low impedance power sources. Certain tantalum capacitors, such as AVX TPS series, are surge tested. For generic tantalum capacitors, use 2:1 voltage derating to protect the input capacitors from surge fall-out.

For accurate control it is important to keep ripple voltages on Vin to a minimum. Vin powers the SP6122 and its internal reference used to maintain output regulation, so proper input bypassing is critical to reduce reference noise. With a reference comparator internal hysteresis of 5mV, and a 1.25V reference voltage, noise on the V_{CC} of the I_{CC} should be kept to about 20mV or less to reduce reference noise effect on output regulation.

The use of very low ESR capacitors is recommended for Vin bypassing, through the use of parallel combinations of tantalum capacitors or even better using some of the new large valued multi-layer ceramic capacitors. ESR values as low as 0.005Ω can be obtained with a 47µF ceramic (see table 1 capacitor selection) and these ceramic capacitors will reduce the power loss in the input capacitance greatly by their reduced ESR values.

For the SP6122 example using the 47µF ceramic input capacitor, the P_{CIN} = 20mW, which is very efficient, and the input ripple voltage at the V_{IN} post (not the V_{CC} pin of the IC) is about 90mV.

MOSFET SELECTION

A SP6122 design uses a PMOS switch on the high side, without the need for a high side charge pump, simplifying the application circuit. The losses associated with the PMOS can be divided into conduction and switching losses. Conduction losses are related to the on resistance of the PMOS, and increase with the load current. Switching losses occur on each on/off transition when the PMOS experiences both high current and voltage. The switching losses are difficult to quantify due to all the variables affecting turnon/turnoff time. However, the following equation provides an approximation on the switching losses associated with the PMOS driven by SP6122.

Guidelines for Component Selection: *continued*

$P_{SH(MAX)} \cong 1/2\ I_{OUT(MAX)} V_{IN(MAX)} (t_{RISE} + t_{FALL}) F_S$

where t_{RISE} (SP6122) for 8A PMOS is typically 20ns and t_{FALL} (SP6122) for 8A PMOS is typically 40ns.

Switching losses need to be taken into account for high switching frequency, since they are directly proportional to switching frequency. The conduction losses associated with the PMOS is determined by:

$$P_{CH(MAX)} = I_{OUT\ (MAX)}{}^2\ R_{DS(ON)}\ D$$

Where $R_{DS(ON)}$ = drain to source on resistance.

The total power losses of the PMOS are the sum of switching and conduction losses. For input voltages of 3.3V and 5V, conduction losses often dominate switching losses. Therefore, lowering the $R_{DS(ON)}$ of the PMOS always improves efficiency even though it gives rise to higher switching losses due to increased C_{RSS}.

For the SP6122 design example, the Fairchild PMOS FDS6375 was selected for its low $R_{DS(ON)}$ and good switching characteristics including low gate charge at the 3.3V input. Using table 1 values for $R_{DS(ON)}$ and t_{RISE} and t_{FALL} for the SP6122, we calculate;

$P_{SH(MAX)}$ = 119mW and $P_{CH(MAX)}$ = 203mW.

$R_{DS(ON)}$ varies greatly with the gate driver voltage. The MOSFET vendors often specify $R_{DS(ON)}$ on multiple gate to source voltages (VGS), as well as provide typical curve of $R_{DS(ON)}$ versus VGS. For 5V input, use the $R_{DS(ON)}$ specified at 4.5V VGS. At the time of this publication, vendors, such as Fairchild, Siliconix and International Rectifier, have started to specify $R_{DS(ON)}$ at VGS less than 3V. This has provided necessary data for designs in which these MOSFETs are driven with 3.3V and made it possible to use SP6122 in 3.3V only applications.

Thermal calculation must be conducted to ensure the MOSFET can handle the maximum load current. The junction temperature of the MOSFET, determined as follows, must stay below the maximum rating.

$$T_{J(MAX)} = T_{A\ (MAX)} + P_{MOSFET(MAX)}\ R\theta_{JA}$$

where

$T_{A\ (MAX)}$ = maximum ambient temperature

$P_{MOSFET(MAX)}$ = maximum power dissipation of the MOSFET, including both switching and conduction losses

$R\theta_{JA}$ = junction to ambient thermal resistance.

$R\theta_{JA}$ of the device depends greatly on the board layout, as well as device package. Significant thermal improvement can be achieved in the maximum power dissipation through the proper design of copper mounting pads on the circuit board. For example, in a SO-8 package, placing two 0.04 square inches copper pad directly under the package, without occupying additional board space, can increase the maximum power from approximately 1 to 1.2W.

For the PMOS FDS6375, assuming $T_{A\ (MAX)}$ = 20°C, $P_{MOSFET(MAX)} = P_{SH(MAX)} + P_{CH(MAX)}$ = 321mW, and assuming per FDS6375 data sheet, $R\theta_{JA}$ = 50°C/W if using 0.5 in^2 pad of 2oz Cu,

$$T_{J(MAX)} = 36°C$$

which is only a 16°C rise from ambient.

SCHOTTKY DIODE SELECTION

The schottky diode is selected for low forward voltage, current capability and fast switching speed. The average power dissipation of the schottky diode is determined by

$$P_{DIODE} = V_F\ I_{OUT}\ (1- D)$$

Where V_F is the forward voltage of the schottky diode at I_{OUT}.

For the SP6122 example, the schottky STPS2L25U has V_F = 0.5V for I_{OUT} of 4A, the power loss in the schottky P_{DIODE} = 848mW.

Note that this power dissipation is 2.5 times greater than the MOSFET. If we assume the same thermal conductivity as the MOSFET (according to the data sheets, this is close) we should get a 40°C rise due to the Schottky diode alone. It is apparent that due to the proximity of all the components involved that the board temperature is higher than ambient and this temperature rise must be considered when attempting to protect the power converter.

Features and Protection

PROGRAMMING THE SP6122 OUTPUT VOLTAGE

As you can see by the schematic in Figure 5, the SP6122 uses an internal feedback divider to initially trim the output voltage using $R_{IN}1$ and $R_{IN}2$, where $R_{IN}2$ is approximately 62.5K. To accommodate the user who wants to externally program the SP6122 output voltage, the SP6122 Evaluation Board has 2 external divider resistors, R1 and R2, which can be used to program the output voltage above, but not below the voltage set by the internal resistor divider. The relationships for the external divider resistors are derived below:

$$i1 = V_A/R_{IN}2 = 1.25/62.5k = 20\mu A$$

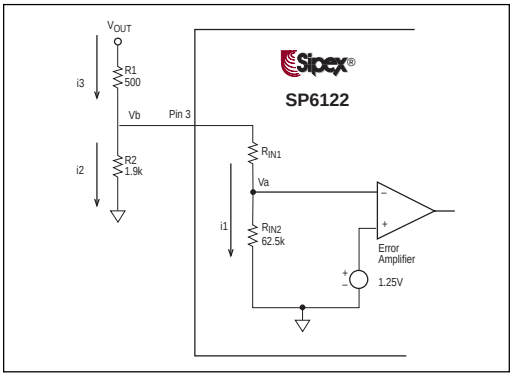


Figure 5: Schematic: Output Voltage Divider Resistors

Since $R_{IN}2$ has ±50% tolerance, ib change is +20µA/-7µA. If the SP6122 is trimmed to 1.5V output, Vb will be fixed at 1.5V in a closed loop. As a result, i2 is not affected by the variation of the internal resistors, and the +20uA/-7uA current variation will be passed on to i3.

$$V_{OUT} = Vb + i3*R1,$$

Therefore,

$$DV_{OUT} = Di3*R1 = +20\mu A/-7\mu A*R1.$$

That is, the additional variation on the output voltage is caused by the internal voltage divider. For example, with R1 selected to be 500Ω, the variation on the output voltage will be +10mV/-3.5mV.

SP6122 Evaluation Board Divider Resistors: *The values of R1 and R2 are selected to program 1.9V output, with 1.5V trim voltage Vb, and R1 = 500Ω:*

$$i3 = (V_{OUT}-Vb)/R1=(1.9-1.5)/500 = 800\mu A.$$

The external voltage divider will add only 800µA to the load. Divider Resistor R2 is:

$$R2 = Vb/(i3-i1) =1.5/(800\mu A-20\mu A) = 1.9k$$

The user of the SP6122 Evaluation Board can use the above equations for i3 and R2 to modify R2 and change the output voltage to be any voltage from Vb (1.5V) to as high as the input voltage. And if you want the output voltage to be the preset voltage Vb (1.5V), just short a wire across R1.

SOFT START

The SP6122 has a built-in soft start feature that automatically limits the inrush currents to reasonable levels for most power supplies. For our 300kHz, 1.9V example, the soft start time is 2 ms. The inrush current on start up is:

$$I_{INRUSH} = 470\mu F * 1.9V/2ms = 447mA$$

This extra current must be factored in when calculating over current margins.

Features and Protection: *continued*

LOCK-UP AND HICCUP MODES

Basically, when the SP6122 sees an over current fault, the part can react in two ways. If the FFLAG is not tied to ENABLE, the part will put the driver into a low impedance state to the high rail during a fault. The ENABLE pin must be manually cycled to remove the fault. This mode is useful when power supply sequencing and system fault management is complex. If the FFLAG pin is tied to ENABLE, then a 'hiccup' time can be designed by adding a capacitor from ENABLE to ground. The 4µA ENABLE pin charge current acts as a timer. The driver will be put into a low impedance state to the high rail for a certain amount of time.

$$T_{OFF} = C_{ENABLE} * 1.1V/4\mu A$$

For C_{ENABLE} = 4.7nF, this time equals 1.3ms. This represents a 'cool off' time required for the power supply to cycle and see if the fault has been removed. This mode is useful for short term faults or in single supply systems.

$R_{DS(ON)}$ OVER CURRENT PROTECTION

Fault conditions are detected via an over voltage condition across the PMOS switch during conduction. This is commonly known as RDSon sensing. RDSon sensing is inaccurate but efficient and is used where an indicator of over current behavior is required for protection. Two advanced features are incorporated in the SP6122 RDSon sensing scheme. The sensing environment is very noisy. Typical schemes require some external filtering in order to avoid spurious faults due to noise or load transients, often compromising the protection and performance at low duty ratios. The SP6122 incorporates a 10µs internal sample and hold filter after the main sense comparator. In this fashion, small pulse widths can be detected while maintaining adequate filtering against false glitches. In addition, temperature compensation is added such that the over current detection threshold at any temperature can be calculated with reasonable accuracy at room temperature. For our evaluation board example:

$$I_{TRIP} = (150mV + I_{SET}R_{SET})/R_{DS(ON)} =$$
$$(150mV + 23\mu A * 1k\Omega)/25m\Omega = 6.92A$$

This is the about the same trip threshold at room, hot or cold because a temperature coefficient has been added to both the 150mV and the 23µA set currents. This temperature coefficient tracks the 25mΩ RDSon of the external FET. Due to the small size of these power supplies, thermal coupling exists between the PFET and the SP6122, making this thermal compensation reasonable, but not perfect. Notice there is about a 50% pad between the maximum usable current (5A) and the over current trip threshold (7A) in order to accommodate PFET and overall system variation.

Layout Guidelines

PCB layout plays a critical role in proper function of the converters and EMI control. In switch mode power supplies, loops carrying high di/dt give rise to EMI and ground bounce. The goal of layout optimization is to identify these loops and minimize them. It is also crucial on how to connect the controller ground such that its operation is not affected by noise. The following guidelines should be followed to ensure proper operation.

1. A ground plane is recommended for minimizing noises, copper losses and maximizing heat dissipation.
2. Connect the ground of the feedback divider to the GND pin of the IC. Then connect this pin as close as possible to the ground of the output capacitor.
3. The Vcc bypass capacitor should be right next to the Vcc and GND pins.
4. The traces connecting to the feedback resistors and current sense components should be short and far away from the switch node and switching components.
5. Minimize the trace length/maximize the trace width between the PDRV pin and the gate of the PMOS.
6. Minimize the loop composed of input capacitors, PMOS and Schottky diode, as this loop carries high di/dt current. Also increase the trace width to reduce copper losses.
7. Maximize the trace width of the loop connecting the inductor, output capacitors, and Schottky diode.
8. For an layout example of an SP6122 power supply (3.3Vin and 1.9Vout at 4A) see the SP6122 Evaluation Board Manual.

SP6122 Design Calculator Example: Evaluation Board with 3.3V_{IN}, 1.9V_{OUT}

Table 2, SP6122 Design Calculator, illustrates the calculations and formulas contained in the Sipex Non-Synchronous Buck Cad Calculator spreadsheet, (available in the applications section of the Sipex website at www.sipex.com). The example shown is the same SP6122 Evaluation Board used previously with V_{IN} = 3.3V, V_{OUT} = 1.9V at 4A. As you can see, the SP6122 efficiency at 4A output is calculated to be 84.3%. Compare this with the Typical Performance Characteristics curve of 84.5%, which is very close considering the tolerances of various components, and you see how useful this easy design calculator is to evaluate your SP6122 designs.

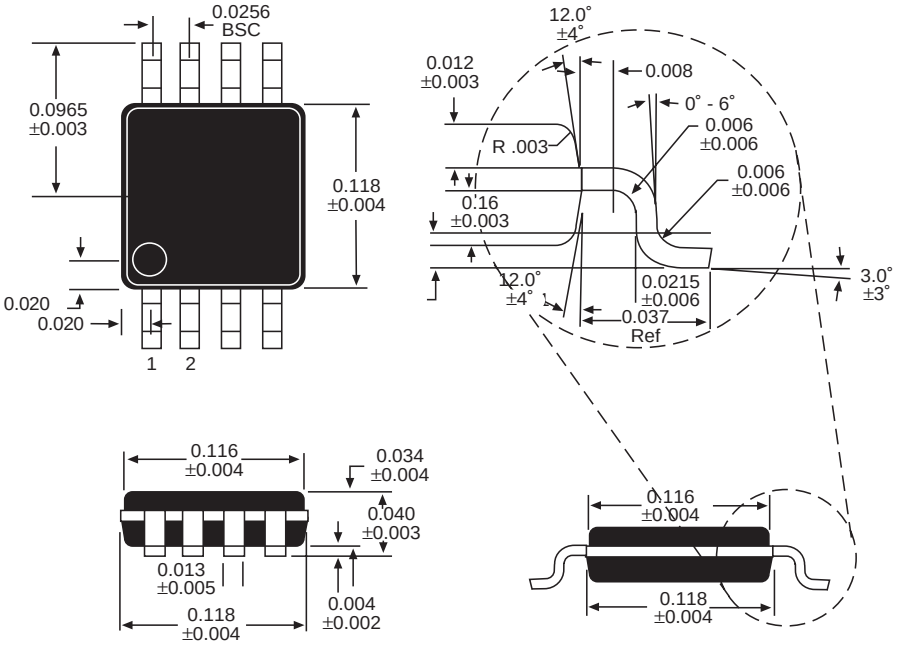
SP6122 Non-Synchronous Buck Design Calculator

STEADY STATE CALCULATION				
Enter Values		**Calculation Results**		**Formula**
V_{IN} = Input Voltage (V)	**3.3**	D = Duty Cycle	**0.58**	= V_{OUT}/V_{IN}
V_{OUT} = Output Voltage (V)	**1.9**	Iripple = Ripple Current (A)	**1.22**	= $(V_{IN}-V_{OUT})*V_{OUT}/(Fs*1000*L*0.000001*V_{IN})$
Fs = Switching Frequency (kHz)	**300**	Ipeak = Peak Inductor Current (A)	**4.61**	= I_{OUT}+Iripple/2
I_{OUT} = Load Current (A)	**4**	Output Ripple (mV)	**42.75**	= Iripple*ESRout
L = Inductance (µH)	**2.2**	Iin = Max Input Current (A)	**2.56**	= I_{OUT}*D/0.9
ESRin = Input Capacitor ESR (mΩ)	**5**	Max Input Ripple (mV)	**96.99**	= I_{OUT}*ESRin+Iin*(1-D)/(Fs*C_{IN}*0.000001)
C_{IN} = Input Capacitance (µF)	**47**	Iin_rms = Input Cap RMS Current (A)	**1.98**	= I_{OUT}*SQRT(D*(1-D))
ESR_{OUT} = Output Capacitor ESR (Ω)	**35**			

EFFICIENCY CALCULATION				
Enter Values		**Calculation Results**		**Formula**
RGH = GH Impedance (Ω)	4	Pic = IC Power (switching) (mW)	**31.35**	= Icc*V_{IN}+Chs*V_{IN}*Fs*0.001
PMOS				
T_{RISE} = SP6122 typ. PMOS rise time (ns)	20	Psch = Schottky Conducting Loss (mw)	**848.48**	= Vf*I_{OUT}*(1-D)*1000
T_{FALL} = SP6122 typ. PMOS rise time (ns)	40			
Chs = PMOS Gate Charge @ V_{IN} (nc)	**15**	Pch = PMOS Conducting Loss (mW)	202.67	= I_{OUT}*I_{OUT}*D*Rhs
Rhs = $R_{DS(ON)}$ @ V_{IN} (mΩ)	**22**	Psh = PMOS Switching Loss (mW)	118.80	= 1/2*I_{OUT}*V_{IN}*(T_{RISE}+T_{FALL})*Fs*0.001
		Phs = Total PMOS Loss (mW)	**321.47**	= Pch + Psh
Vf = Schottky Forward Voltage	**0.5**	Pl = Inductor loss (mW)	**192.00**	= I_{OUT}*I_{OUT}*ESR_L
I_{CC} = Supply Current (no switch) (mA)	5	Pc_{IN} = Input Capacitor Loss(mW)	**19.54**	= ESR_{IN}*Iin_rms*Iin_rms
ESR_L = Inductor ESR (mΩ)	**12**	Pltot = Total Power Losses (mW)	**1412.84**	= Pic+Pls+Phs+Pl+Psch
		Efficiency (%)	**84.32**	= V_{OUT}*I_{OUT}/(V_{OUT}*I_{OUT} - Pltot/1000)*100

Table 2: Design Calculator

PACKAGE: 8 PIN PLASTIC MICRO SMALL OUTLINE (µSOIC)



All package dimensions in inches

ORDERING INFORMATION

Part Number	Operating Temperature Range	Package Type
(300kHz)		
SP6122CUA-1.5	0°C to 70°C	8 Pin µSOIC
SP6122CUA-1.5/TR	0°C to 70°C *(Tape & Reel)*	8 Pin µSOIC
SP6122CUA-1.8	0°C to 70°C	8 Pin µSOIC
SP6122CUA-1.8/TR	0°C to 70°C *(Tape & Reel)*	8 Pin µSOIC
SP6122CUA-2.5	0°C to 70°C	8 Pin µSOIC
SP6122CUA-2.5/TR	0°C to 70°C *(Tape & Reel)*	8 Pin µSOIC
SP6122CUA-3.3	0°C to 70°C	8 Pin µSOIC
SP6122CUA-3.3/TR	0°C to 70°C *(Tape & Reel)*	8 Pin µSOIC
(600kHz)		
SP6122CUB-1.5	0°C to 70°C	8 Pin µSOIC
SP6122CUB-1.5/TR	0°C to 70°C *(Tape & Reel)*	8 Pin µSOIC
SP6122CUB-1.8	0°C to 70°C	8 Pin µSOIC
SP6122CUB-1.8/TR	0°C to 70°C *(Tape & Reel)*	8 Pin µSOIC
SP6122CUB-2.5	0°C to 70°C	8 Pin µSOIC
SP6122CUB-2.5/TR	0°C to 70°C *(Tape & Reel)*	8 Pin µSOIC
SP6122CUB-3.3	0°C to 70°C	8 Pin µSOIC
SP6122CUB-3.3/TR	0°C to 70°C *(Tape & Reel)*	8 Pin µSOIC

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