THE DATASHEET OF

L6712QTR

L6712
L6712A

TWO-PHASE INTERLEAVED DC/DC CONTROLLER

1 Features

- 2 PHASE OPERATION WITH SYNCHRONOUS RECTIFIER CONTROL
- ULTRA FAST LOAD TRANSIENT RESPONSE
- INTEGRATED HIGH CURRENT GATE DRIVERS: UP TO 2A GATE CURRENT
- 3 BIT PROGRAMMABLE OUTPUT FROM 0.900V TO 3.300V OR WITH EXTERNAL REF.
- ±0.9% OUTPUT VOLTAGE ACCURACY
- 3mA CAPABLE AVAILABLE REFERENCE
- INTEGRATED PROGRAMMABLE REMOTE SENSE AMPLIFIER
- PROGRAMMABLE DROOP EFFECT
- 10% ACTIVE CURRENT SHARING ACCURACY
- DIGITAL 2048 STEP SOFT-START
- CROWBAR LATCHED OVERVOLTAGE PROT.
- NON-LATCHED UNDERVOLTAGE PROT.
- OVERCURRENT PROTECTION REALIZED USING THE LOWER MOSFET'S R_{dsON} OR A SENSE RESISTOR
- OSCILLATOR EXTERNALLY ADJUSTABLE AND INTERNALLY FIXED AT 150kHZ
- POWER GOOD OUTPUT AND INHIBIT FUNCTION
- PACKAGES: SO-28 & VFQFPN-36

1.1 Applications

- HIGH CURRENT DC/DC CONVERTERS
- DISTRIBUTED POWER SUPPLY

2 Description

The device implements a dual-phase step-down controller with a 180 phase-shift between each phase optimized for high current DC/DC applications.

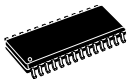
Output voltage can be programmed through the integrated DAC from 0.900V to 3.300V; programming the "111" code, an external reference from 0.800V to 3.300V is used for the regulation.

Programmable Remote Sense Amplifier avoids use of external resistor divider and recovers losses along distribution line.

The device assures a fast protection against load over current and Over / Under voltage.An internal crowbar is provided turning on the low side mosfet if Over-voltage is detected.

Output current is limited working in Constant Current mode: when Under Voltage is detected, the device resets, restarting operation.

Figure 1. Packages



SO28

VFQFPN-36 (6x6x1.0mm)

Table 1. Order Codes

Package	Tube	Tape & Reel
SO	L6712D, L6712AD	L6712DTR, L6712ADTR
VFQFPN	L6712Q, L6712AQ	L6712QTR, L6712AQTR

Figure 2. Block Diagram

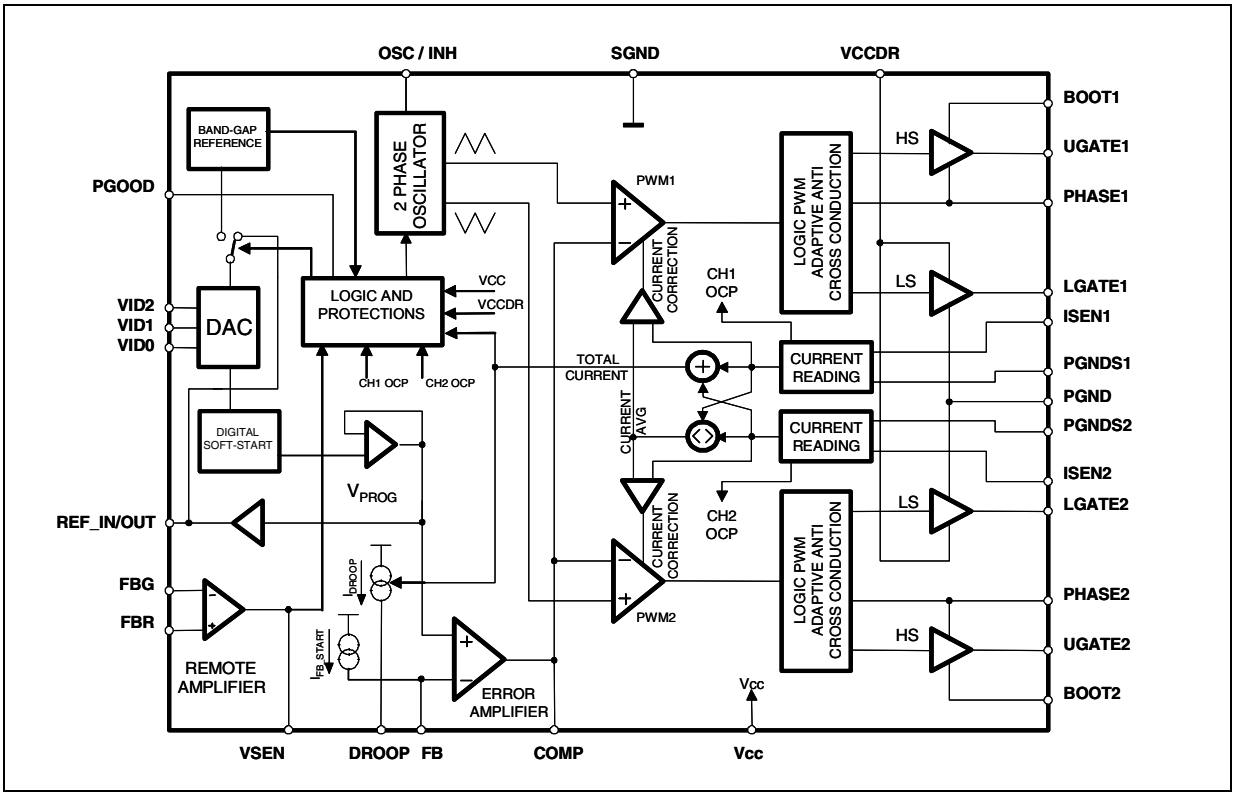


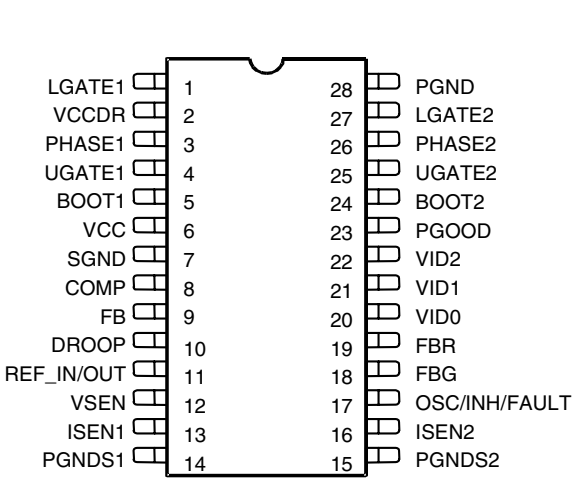
Table 2. Absolute Maximum Ratings

Symbol	Parameter	Value	Unit
V_{CC}, V_{CCDR}	To PGND	15	V
V_{BOOT}-V_{PHASE}	Boot Voltage	15	V
V_{UGATE1}-V_{PHASE1} V_{UGATE2}-V_{PHASE2}		15	V
	LGATE1, PHASE1, LGATE2, PHASE2 to PGND	-0.3 to Vcc+0.3	V
	VID0 to VID2	-0.3 to 5	V
	All other pins to PGND	-0.3 to 7	V
V_{PHASEx}	Sustainable Peak Voltage. T<20ns @ 600kHz	26	V
UGATEX Pins	Maximum Withstanding Voltage Range Test Condition: CDF-AEC-Q100-002"Human Body Model" Acceptance Criteria: "Normal Performance"	±1500	V
OTHER PINS		±2000	V

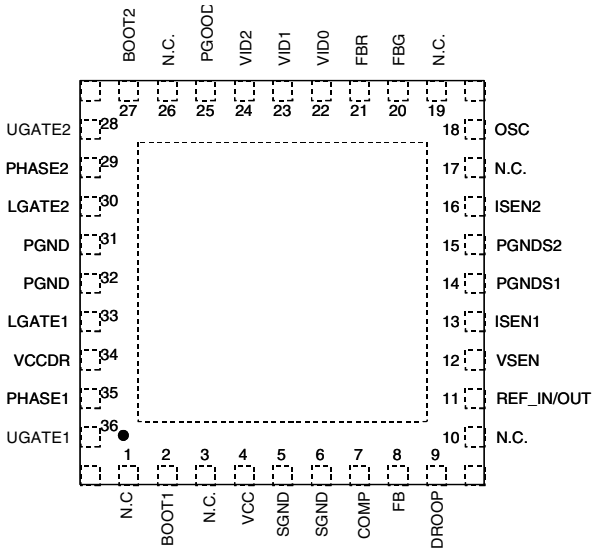
Table 3. Thermal Data

Symbol	Parameter	SO28	VFQFPN36	Unit
$R_{thj\text{-}amb}$	Thermal Resistance Junction to Ambient 4 layer PCB (2s2p)	60	30	°C/W
T_{max}	Maximum junction temperature	150	150	°C
T_{stg}	Storage temperature range	-40 to 150	-40 to 150	°C
T_j	Junction Temperature Range	-40 to 125	-40 to 125	°C
P_{MAX}	Max power dissipation at T_{amb} = 25°C	2	3.5	W

Figure 3. Pin Connection (Top view)



SO28



VFQFPN-36

Corner Pin internally connected to the Exposed Pad.

Table 4. Electrical Characteristcs

(V_{CC} = 12V±10%, T_J = 0°C to 70°C unless otherwise specified)

Symbol	Parameter	Test Condition	Min.	Typ.	Max.	Unit
Vcc SUPPLY CURRENT						
I_{CC}	VCC supply current	HGATEx and LGATEx open VCCDR=BOOTx=12V	7.5	10	12.5	mA
I_{CCDR}	VCCDR supply current	LGATEx open; VCCDR=12V	1.5	3	4	mA
I_{BOOTx}	Boot supply current	HGATEx open; PHASEx to PGND; VCC=BOOTx=12V	0.5	1	1.5	mA
POWER-ON						
	Turn-On VCC threshold	VCC Rising; VCCDR=5V	8.2	9.2	10.2	V
	Turn-Off VCC threshold	VCC Falling; VCCDR=5V	6.5	7.5	8.5	V
	Turn-On VCCDR Threshold	VCCDR Rising VCC=12V	4.2	4.4	4.6	V
	Turn-Off VCCDR Threshold	VCCDR Falling VCC=12V	4.0	4.2	4.4	V
OSCILLATOR AND INHIBIT						
f_{OSC}	Initial Accuracy	OSC = OPEN OSC = OPEN; Tj=0°C to 125°C	135 127	150	165 178	kHz kHz
INH	Inhibit threshold	I_{SINK}=5mA	0.5			V
d_{MAX}	Maximum duty cycle	**L6712**, OSC = OPEN: I_{DROOP}=0 OSC = OPEN; I_{DROOP}=70µA	72 30	80 40	- -	% %
		L6712A, OSC = OPEN	85	90		%
ΔVosc	Ramp Amplitude			3		V
FAULT	Voltage at pin OSC	OVP Active	4.75	5.0	5.25	V

Table 4. Electrical Characteristcs (continued)
(V_{CC} = 12V±10%, T_J = 0°C to 70°C unless otherwise specified)

Symbol	Parameter	Test Condition	Min.	Typ.	Max.	Unit
REFERENCE AND DAC						
V_{OUT} (1)	Output Voltage Accuracy	VIDx See Table 5, VID ≠ "11x"	-0.9	-	0.9	%
		VID = "110"	-1.0	-	1.0	%
REF_IN/OUT	Reference Accuracy	VIDx See Table 5, VID ≠ "111"	V_{OUT} -5	V_{OUT}	V_{OUT}+5	mV
	Current Capability		3			mA
	Load Regulation	I_{REF} = from 0 to 3mA			5.0	mV
V_{PROG} / REF_IN/ OUT	Accuracy with external reference	VID="111"; REF_IN/OUT = 0.8V to 3.3V	-2.0		2.0	%
REF_IN/OUT	Input impedance			400		kΩ
I_{VID}	VID pull-up Current	VIDx =SGND		5		µA
V_{VID}	VID pull-up Voltage	VIDx = OPEN		3		V
VID_{IL}	VID Input Levels	Input Low			0.4	V
VID_{IH}		Input High	1.0			V
ERROR AMPLIFIER						
V_{OS_EA}	Offset	FB = COMP	-5		5	mV
	DC Gain			80		dB
SR	Slew-Rate	COMP=10pF		15		V/µs
I_{FB_START}	Start-up Current	FB=SGND; During Soft Start…	65			µA
DIFFERENTIAL AMPLIFIER (REMOTE BUFFER)						
V_{OS_RA}	Offset	VSEN = FBG	-8		8	mV
	DC Gain			80		dB
SR	Slew Rate	VSEN = 10pF		15		V/µs
DIFFERENTIAL CURRENT SENSING						
I_{ISEN1}, I_{ISEN2}	Bias Current	I_{LOAD} = 0	45	50	55	µA
I_{PGNDSx}	Bias Current		45	50	55	µA
I_{ISEN1}, I_{ISEN2}	Bias Current at Over Current Threshold		80	85	90	µA
I_{DROOP}	Droop Current	$I_{LOAD} \leq 0$		0	1	µA
		I_{LOAD} = 100%	47.5	50	52.5	µA
GATE DRIVERS						
$t_{RISE\ HGATE}$	High Side Rise Time	BOOTx-PHASEx=10V; C_{HGATEx} to PHASEx=3.3nF		15	30	ns
I_{HGATEx}	High Side Source Current	BOOTx-PHASEx=10V		2		A
R_{HGATEx}	High Side Sink Resistance	BOOTx-PHASEx=12V;	1.5	2	2.5	Ω
$t_{RISE\ LGATE}$	Low Side Rise Time	VCCDR=10V; C_{LGATEx} to PGNDx=5.6nF		30	55	ns
I_{LGATEx}	Low Side Source Current	VCCDR=10V		1.8		A
R_{LGATEx}	Low Side Sink Resistance	VCCDR=12V	0.7	1.1	1.5	Ω

Table 4. Electrical Characteristcs (continued)
(V_{CC} = 12V±10%, T_J = 0°C to 70°C unless otherwise specified)

Symbol	Parameter	Test Condition	Min.	Typ.	Max.	Unit
PROTECTIONS						
PGOOD	Upper Threshold	VSEN Rising	108	112	115	%
	Lower Threshold	VSEN Falling	84	88	92	%
OVP	Over Voltage Threshold	VSEN Rising	115	122	130	%
UVP	Under Voltage Trip	VSEN Falling	55	60	65	%
V_{PGOODL}	PGOOD Voltage Low	I_{PGOOD} = -4mA			0.4	V
I_{PGOODH}	PGOOD Leakage	V_{PGOOD} = 5V			1	µA

Note: 1. Output voltage is specified including Error Amplifier Offset in the trimming chain. Remote Amplifier is not included.

Table 5. Voltage Identification (VID) Codes.

VID2	VID1	VID0	Output Voltage (V)
1	1	1	**Ext. Ref.**
1	1	0	**0.900**
1	0	1	**1.250**
1	0	0	**1.500**
0	1	1	**1.715**
0	1	0	**1.800**
0	0	1	**2.500**
0	0	0	**3.300**

Table 6. Pin Function

N. (*)		Name	Description
SO	VFQFPN		
1	33	LGATE1	Channel 1 LS driver output. A little series resistor helps in reducing device-dissipated power.
2	34	VCCDR	LS drivers supply: it can be varied from 5V to 12V buses. Filter locally with at least 1µF ceramic cap vs. PGND.
3	35	PHASE1	Channel 1 HS driver return path. It must be connected to the HS1 mosfet source and provides the return path for the HS driver of channel 1.
4	36	UGATE1	Channel 1 HS driver output. A little series resistor helps in reducing device-dissipated power.
5	2	BOOT1	Channel 1 HS driver supply. This pin supplies the relative high side driver. Connect through a capacitor (100nF typ.) to the PHASE1 pin and through a diode to VCC (cathode vs. boot).
6	4	VCC	Device supply voltage. The operative supply voltage is 12V ±10%. Filter with 1µF (Typ.) capacitor vs. GND.
7	5,6	SGND	All the internal references are referred to this pin. Connect it to the PCB signal ground.
8	7	COMP	This pin is connected to the error amplifier output and is used to compensate the control feedback loop.
9	8	FB	This pin is connected to the error amplifier inverting input and is used to compensate the control feedback loop.

Table 6. Pin Function (continued)

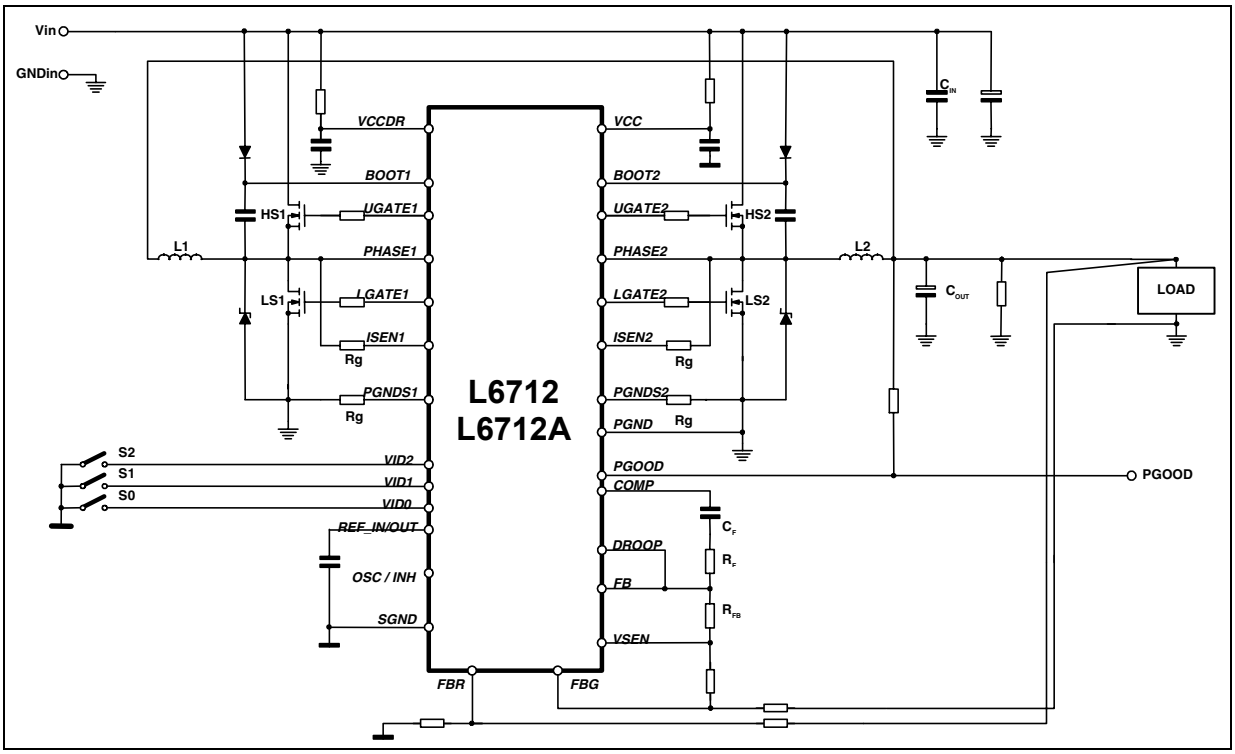
N. (*)		Name	Description
SO	VFQFPN		
10	9	DROOP	A current proportional to the sum of the current sensed in both channel is sourced from this pin (50µA at full load, 70µA at the Constant Current threshold). Short to FB to implement the Droop effect: the resistor connected between FB and VSEN (or the regulated output) allows programming the droop effect. Otherwise, connect to GND directly or through a resistor (43kΩ max) and filter with 1nF capacitor. In this last case, current information can be used for other purposes.
11	11	REF_IN / OUT	Reference input/output. Filter vs. GND with 1nF ceramic capacitor (a total of 100nF capacitor is allowed). It reproduces the reference used for the regulation following VID code: when VID=111, the reference for the regulation must be connected on this pin. References ranging from 0.800V up to 3.300V can be accepted.
12	12	VSEN	Connected to the output voltage it is able to manage Over & Under-voltage conditions and the PGOOD signal. It is internally connected with the output of the Remote Sense Amplifier for Remote Sense of the regulated voltage. Connecting 1nF capacitor max vs. GND can help in reducing noise injection at this pin. If no Remote Sense is implemented, connect it directly to the regulated voltage in order to manage OVP, UVP and PGOOD.
13	13	ISEN1	Channel 1 current sense pin. The output current may be sensed across a sense resistor or across the low-side mosfet R_{dsON}. This pin has to be connected to the low-side mosfet drain or to the sense resistor through a resistor Rg. The net connecting the pin to the sense point must be routed as close as possible to the PGNDS net in order to couple in common mode any picked-up noise.
14	14	PGNDS1	Channel 1 Power Ground sense pin. The net connecting the pin to the sense point must be routed as close as possible to the ISEN1 net in order to couple in common mode any picked-up noise.
15	15	PGNDS2	Channel 2 Power Ground sense pin. The net connecting the pin to the sense point must be routed as close as possible to the ISEN2 net in order to couple in common mode any picked-up noise.
16	16	ISEN2	Channel 2 current sense pin. The output current may be sensed across a sense resistor or across the low-side mosfet R_{dsON}. This pin has to be connected to the low-side mosfet drain or to the sense resistor through a resistor Rg. The net connecting the pin to the sense point must be routed as close as possible to the PGNDS net in order to couple in common mode any picked-up noise.
17	18	OSC/INH FAULT	Oscillator pin. It allows programming the switching frequency of each channel: the equivalent switching frequency at the load side results in being doubled. Internally fixed at 1.24V, the frequency is varied proportionally to the current sunk (forced) from (into) the pin with an internal gain of 6kHz/µA (See relevant section for details). If the pin is not connected, the switching frequency is 150kHz for each channel (300kHz on the load). The pin is forced high (5V Typ.) when an Over Voltage is detected; to recover from this condition, cycle VCC. Forcing the pin to a voltage lower than 0.6V, the device stops operation and enters the inhibit state.
18	20	FBG	Remote sense amplifier inverting input. It has to be connected to the negative side of the load to perform programmable remote sensing through apposite resistors (see relative section).
19	21	FBR	Remote sense amplifier non-inverting input. It has to be connected to the positive side of the load to perform programmable remote sensing through apposite resistors (see relative section).

Table 6. Pin Function (continued)

N. (*)		Name	Description
SO	VFQFPN		
20 to 22	22 to 24	VID0-2	Voltage IDentification pins. These input are internally pulled-up. They are used to program the output voltage as specified in Table 1 and to set the PGOOD, OVP and UVP thresholds. Connect to GND to program a '0' while leave floating to program a '1'.
23	25	PGOOD	This pin is an open collector output and is pulled low if the output voltage is not within the above specified thresholds and during soft-start. It cannot be pulled up above 5V. If not used may be left floating.
24	27	BOOT2	Channel 2 HS driver supply. This pin supplies the relative high side driver. Connect through a capacitor (100nF typ.) to the PHASE2 pin and through a diode to VCC (cathode vs. boot).
25	28	UGATE2	Channel 2 HS driver output. A little series resistor helps in reducing device-dissipated power.
26	29	PHASE2	Channel 2 HS driver return path. It must be connected to the HS2 mosfet source and provides the return path for the HS driver of channel 2.
27	30	LGATE2	Channel 2 LS driver output. A little series resistor helps in reducing device-dissipated power.
28	31, 32	PGND	LS drivers return path. This pin is common to both sections and it must be connected through the closest path to the LS mosfets source pins in order to reduce the noise injection into the device.
	PAD	THERMAL PAD	Thermal pad connects the silicon substrate and makes a good thermal contact with the PCB to dissipate the power necessary to drive the external mosfets.Connect to the GND plane with several vias to improve thermal conductivity.

(*) Pin not reported in QFN column have to be considered as Not Connected, not internally bonded.

Figure 4. Reference Schematic



3 Device Description

The device is an integrated circuit realized in BCD technology. It provides complete control logic and protections for a high performance dual-phase step-down converter optimized for high current DC/DC applications. It is designed to drive N-Channel Mosfets in a two-phase synchronous-rectified buck topology. A 180 deg phase shift is provided between the two phases allowing reduction in the input capacitor current ripple, reducing also the size and the losses. The output voltage of the converter can be precisely regulated, programming the VID pins, from 0.900 to 3.300V with a maximum tolerance of ±0.9% over temperature and line voltage variations. The programmable Remote Sense Amplifier avoids the use of external resistor divider allowing recovering drops across distribution lines and also adjusting output voltage to different values from the available reference. The device provides an average current-mode control with fast transient response. It includes a 150kHz free-running oscillator externally adjustable through a resistor. The error amplifier features a 15V/μs slew rate that permits high converter bandwidth for fast transient performances. Current information is read across the lower mosfets R_{dsON} or across a sense resistor placed in series to the LS mos in fully differential mode. The current information corrects the PWM outputs in order to equalize the average current carried by each phase. Current sharing between the two phases is then limited at ±10% over static and dynamic conditions unless considering the sensing element spread. Droop effect can be programmed in order to minimize output filter and load transient response: the function can be disabled and the current information available on the pin can be used for other purposes. The device protects against Over-Current, with an OC threshold for each phase, entering in constant current mode. Since the current is read across the low side mosfets, the device keeps constant the bottom of the inductors current triangular waveform. When an Under Voltage is detected the device resets with all mosfets OFF and suddenly re-starts. The device also performs a crowbar Over-Voltage protection that immediately latches the operations turning ON the lower driver and driving high the FAULT pin.

3.1 OSCILLATOR

The switching frequency is internally fixed at 150kHz. Each phase works at the frequency fixed by the oscillator so that the resulting switching frequency at the load side results in being doubled.

The internal oscillator generates the triangular waveform for the PWM charging and discharging with a constant current an internal capacitor. The current delivered to the oscillator is typically 25μA (Fsw=150kHz) and may be varied using an external resistor (R_{OSC}) connected between OSC pin and SGND or Vcc. Since the OSC pin is maintained at fixed voltage (Typ. 1.237V), the frequency is varied proportionally to the current sunk (forced) from (into) the pin considering the internal gain of 6KHz/μA.

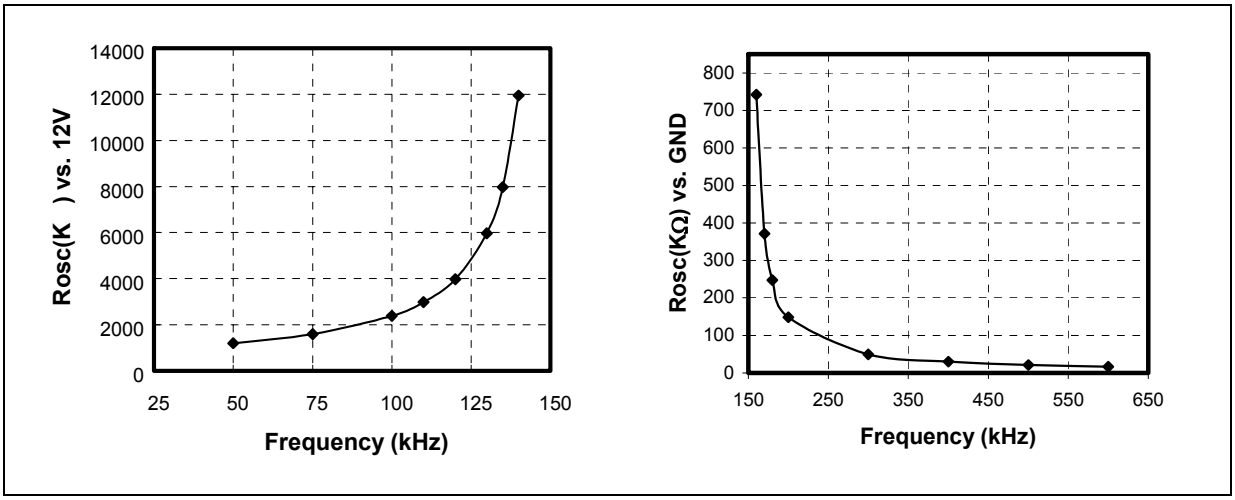
In particular connecting it to SGND the frequency is increased (current is sunk from the pin), while connecting R_{OSC} to Vcc=12V the frequency is reduced (current is forced into the pin), according to the following relationships:

$$R_{OSC} \text{ vs. GND: } F_{SW} = 150[KHz] + \frac{1.237}{R_{OSC}} \cdot 6\left[\frac{kHz}{\mu A}\right] = 150[kHz] + \frac{7.422 \cdot 10^6}{R_{OSC}[K\Omega]}[kHz]$$

$$R_{OSC} \text{ vs. 12V: } F_{SW} = 150[KHz] - \frac{12 - 1.237}{R_{OSC}} \cdot 6\left[\frac{kHz}{\mu A}\right] = 150[kHz] - \frac{6.457 \cdot 10^7}{R_{OSC}[K\Omega]}[kHz]$$

Forcing 25μA into this pin, the device stops switching because no current is delivered to the oscillator

Figure 5. R_{OSC} vs. Switching Frequency



3.2 DIGITAL TO ANALOG CONVERTER AND REFERENCE

The built-in digital to analog converter allows the adjustment of the output voltage from 0.900V to 3.300V as shown in Figure 6. Different voltages can be reached simply changing the Remote Amplifier Gain that acts as a resistor divider (See relevant section).

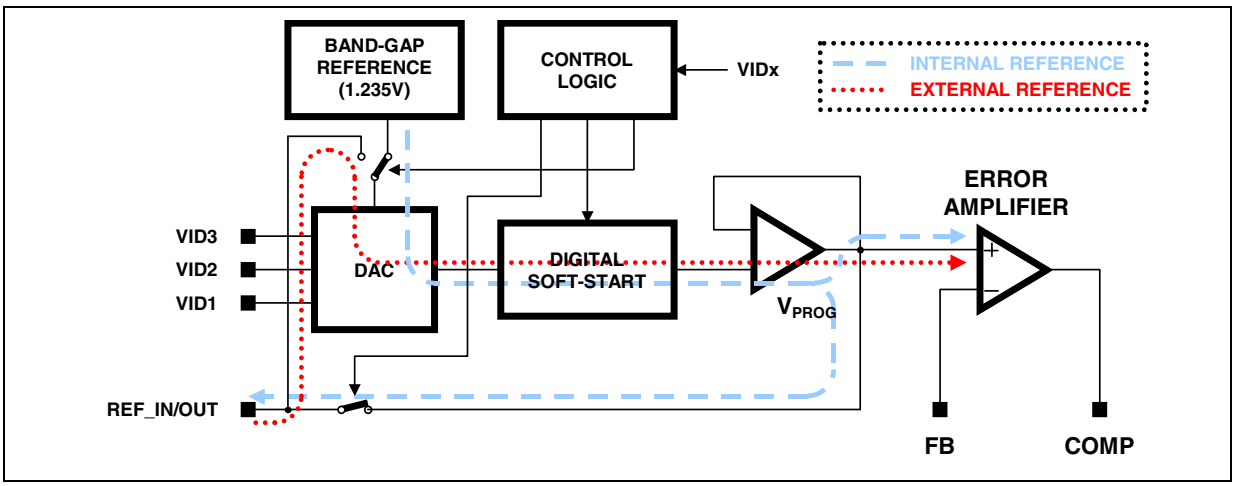
The internal reference is trimmed during production process to have an output voltage accuracy of ±0.9% and a zero temperature coefficient around 70°C including also error amplifier offset compensation. It is programmed through the voltage identification (VID) pins. These are inputs of an internal DAC that is realized by means of a series of resistors providing a partition of the internal voltage reference. The VID code drives a multiplexer that select a voltage on a precise point of the divider (see Figure 6). The DAC output is delivered to an amplifier obtaining the V_{PROG} voltage reference (i.e. the set-point of the error amplifier). Internal pull-ups are provided (realized with a 5μA current generator up to 3V typ.); in this way, to program a logic "1" it is enough to leave the pin floating, while to program a logic "0" it is enough to short the pin to SGND.

The device offers a bi-directional pin REF_IN/OUT: the internal reference used for the regulation is usually available on this pin with 3mA of maximum current capability except when VID code 111 is programmed; in this case the device accepts an external reference through the REF_IN/OUT pin and regulates on it. When external reference is used, it must range from 0.800V up to 3.300V to assure proper functionality of the device.

Figure 6 shows a block schematic of how the Reference for the regulation is managed when internal or external reference is used.

The voltage identification (VID) pin configuration or the external reference provided also sets the power-good thresholds (PGOOD) and the Over/Under voltage protection (OVP/UVP) thresholds.

Figure 6. Reference Management



The output regulated voltage accuracy can be extracted from the following relationships (worst case condition):

$$V_{OUT\_TOT\_ACC}[\%] = V_{OUT\_ACC}[\%] + K_{OS} \cdot \frac{V_{OS\_RA}}{V_{OUT}} \cdot 100 = (\pm 0.9\%) + K_{OS} \cdot \frac{(\pm 8mV)}{V_{OUT}} \cdot 100$$

(worst case with internal reference)

$$V_{OUT\_TOT\_ACC}[\%] = EXT\_REF\_Accuracy[\%] + \frac{V_{PROG}}{REF\_IN/OUT}[\%] + \left(\frac{V_{OS\_EA}}{EXT\_REF} \cdot 100\right) + K_{OS} \cdot \frac{V_{OS\_RA}}{V_{OUT}} \cdot 100 =$$

$$= EXT\_REF\_Accuracy[\%] + (\pm 2.0\%) + \left(\frac{(\pm 5mV)}{EXT\_REF} \cdot 100\right) + K_{OS} \cdot \frac{(\pm 8mV)}{V_{out}} \cdot 100$$

(worst case with external reference)

where V_{OS_RA} and V_{OS_EA} are the offsets related to the Error Amplifier and the Remote Amplifier respectively and K_{OS} = 1+1/RA_Gain reflects the impact of the Remote Amplifier Gain (RA_Gain) on the regulation (see relevant section).

A statistical analysis could consider applying the root-sum-square (RSS) method to calculate the precision since all the variables are statistically independent as follow:

$$V_{OUT\_TOT\_ACC}[\%] = \sqrt{(V_{OUT\_ACC}[\%])^2 + \left(K_{OS} \cdot \frac{V_{OS\_RA}}{V_{OUT}} \cdot 100\right)^2}$$

(with internal reference)

$$V_{OUT\_TOT\_ACC}[\%] = \sqrt{(EXT\_REF\_Accuracy[\%])^2 + \left(\frac{V_{PROG}}{REF\_IN/OUT}[\%]\right)^2 + \left(\frac{V_{OS\_EA}}{EXT\_REF} \cdot 100\right)^2 + \left(K_{OS} \cdot \frac{V_{OS\_RA}}{V_{OUT}} \cdot 100\right)^2}$$

(with external reference)

3.3 DRIVER SECTION

The integrated high-current drivers allow using different types of power MOS (also multiple MOS to reduce the R_{dsON}), maintaining fast switching transition.

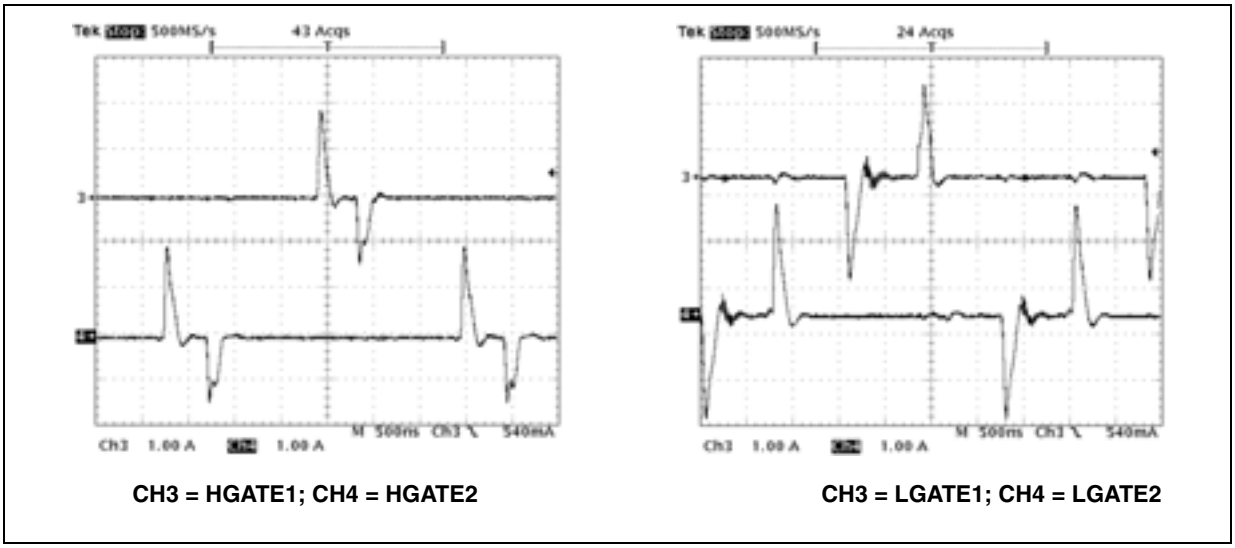
The drivers for the high-side mosfets use BOOTx pins for supply and PHASEx pins for return. The drivers for the low-side mosfets use VCCDR pin for supply and PGND pin for return. A minimum voltage of 4.6V at VCCDR pin is required to start operations of the device.

The controller embodies a sophisticated anti-shoot-through system to minimize low side body diode conduction time maintaining good efficiency saving the use of Schottky diodes in parallel to the LS mosfets. The dead time is reduced to few nanoseconds assuring that high-side and low-side mosfets are never switched on simultaneously: when the high-side mosfet turns off, the voltage on its source begins to fall; when the voltage reaches 2V, the low-side mosfet gate drive is applied with 30ns delay. When the low-side mosfet turns off, the voltage at LGATEx pin is sensed. When it drops below 1V, the high-side mosfet gate drive is applied with a delay of 30ns. If the current flowing in the inductor is negative, the source of high-side mosfet will never drop. To allow the turning on of the low-side mosfet even in this case, a watch-dog controller is enabled: if the source of the high-side mosfet don't drop for more than 240ns, the low side mosfet is switched on so allowing the negative current of the inductor to recirculate. This mechanism allows the system to regulate even if the current is negative.

The BOOTx and VCCDR pins are separated from IC's power supply (VCC pin) as well as signal ground (SGND pin) and power ground (PGND pin) in order to maximize the switching noise immunity. The separated supply for the different drivers gives high flexibility in mosfet choice, allowing the use of logic-level mosfet. Several combination of supply can be chosen to optimize performance and efficiency of the application. Power conversion is also flexible; 5V or 12V bus can be chosen freely.

The peak current is shown for both the upper and the lower driver of the two phases in Figure 7. A 10nF capacitive load has been used. For the upper drivers, the source current is 1.9A while the sink current is 1.5A with V_{BOOT} -V_{PHASE} = 12V; similarly, for the lower drivers, the source current is 2.4A while the sink current is 2A with VCCDR = 12V.

Figure 7. Drivers peak current: High Side (left) and Low Side (right)



CH3 = HGATE1; CH4 = HGATE2

CH3 = LGATE1; CH4 = LGATE2

3.4 CURRENT READING AND OVER CURRENT

The current flowing trough each phase is read using the voltage drop across the low side mosfets R_{dsON} or across a sense resistor (R_{SENSE}) in series to the LS mosfet and internally converted into a current. The transconductance ratio is issued by the external resistor Rg placed outside the chip between ISENx and PGNDSx pins toward the reading points. The differential current reading rejects noise and allows to place sensing element in different locations without affecting the measurement's accuracy. The current reading

circuitry reads the current during the time in which the low-side mosfet is on (OFF Time). During this time, the reaction keeps the pin ISENx and PGNDSx at the same voltage while during the time in which the reading circuitry is off, an internal clamp keeps these two pins at the same voltage sinking from the ISENx pin the necessary current (Needed if low-side mosfet R_{dsON} sense is implemented to avoid absolute maximum rating overcome on ISENx pin).

The proprietary current reading circuit allows a very precise and high bandwidth reading for both positive and negative current. This circuit reproduces the current flowing through the sensing element using a high speed Track & Hold transconductance amplifier. In particular, it reads the current during the second half of the OFF time reducing noise injection into the device due to the mosfet turn-on (See Figure 8-left). Track time must be at least 200ns to make proper reading of the delivered current.

This circuit sources a constant 50µA current from the PGNDSx pin: it must be connected through the Rg resistor to the ground side of the sensing element (See Figure 8-right). The two current reading circuitries use this pin as a reference keeping the ISENx pin to this voltage.

The current that flows in the ISENx pin is then given by the following equation:

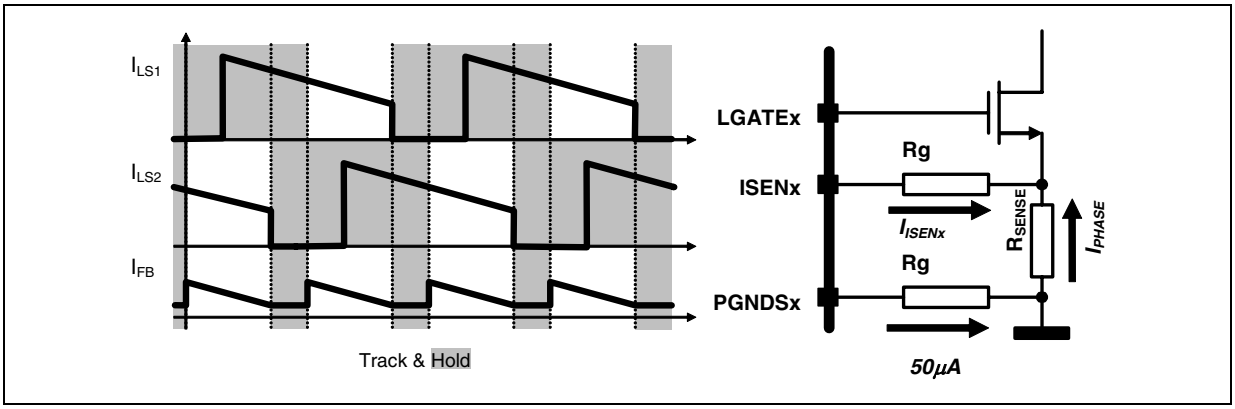
$$I_{ISENx} = 50\mu A + \frac{R_{SENSE} \cdot I_{PHASEx}}{R_g} = 50\mu A + I_{INFOx}$$

Where R_{SENSE} is an external sense resistor or the R_{dsON} of the low side mosfet and R_g is the transconductance resistor used between ISENx and PGNDSx pins toward the reading points; I_{PHASEx} is the current carried by the relative phase. The current information reproduced internally is represented by the second term of the previous equation as follow:

$$I_{INFOx} = \frac{R_{SENSE} \cdot I_{PHASEx}}{R_g}$$

Since the current is read in differential mode, also negative current information is kept; this allow the device to check for dangerous returning current between the two phases assuring the complete equalization between the phase's currents. From the current information of each phase, information about the total current delivered ($I_{FB} = I_{INFO1} + I_{INFO2}$) and the average current for each phase ($I_{AVG} = (I_{INFO1} + I_{INFO2})/2$) is taken. I_{INFOX} is then compared to I_{AVG} to give the correction to the PWM output in order to equalize the current carried by the two phases.

Figure 8. Current reading timing (left) and circuit (right)



The transconductance resistor Rg can be designed in order to have current information of 25µA per phase at full nominal load; the over current intervention threshold is set at 140% of the nominal (I_{INFOx} = 35µA).

According to the above relationship, the over current threshold (I_{OCPx}) for each phase, which has to be placed at 1/2 of the total delivered maximum current, results:

$$I_{OCPx} = \frac{35\mu A \cdot Rg}{R_{SENSE}} \qquad Rg = \frac{I_{OCPx} \cdot R_{SENSE}}{35\mu A}$$

Since the device senses the output current across the low-side mosfets (or across a sense resistors in series with them) the device limits the bottom of the inductor current triangular waveform: an over current is detected when the current flowing into the sense element is greater than I_{OCPx} ($I_{INFOx} > 35\mu A$).

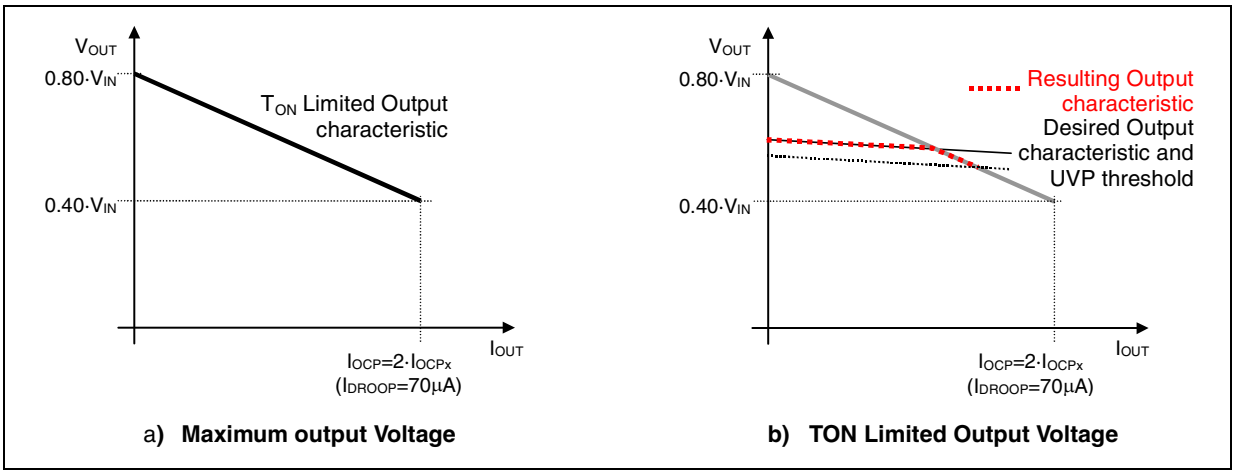
■ L6712 - Dynamic Maximum Duty Cycle Limitation

The maximum duty cycle is limited as a function of the measured current and, since the oscillator frequency is fixed once programmed, imply a maximum on-time limitation as follow (where T is the switching period $T=1/f_{SW}$ and I_{OUT} is the output current):

$$T_{ON,MAX} = (0.80 - I_{FB} \cdot 5.73k) \cdot T = \left(0.80 - \frac{R_{SENSE}}{Rg} \cdot I_{OUT} \cdot 5.73k\right) \cdot T = \begin{cases} T = 0.80 \cdot T & I_{FB} = 0\mu A \\ T = 0.40 \cdot T & I_{FB} = 70\mu A \end{cases}$$

This linear dependence has a value at zero load of 0.80·T and at maximum current of 0.40·T typical and results in two different behaviors of the device:

Figure 9. TON Limited Operation



a) **Maximum output Voltage** b) **TON Limited Output Voltage**

T_{ON} Limited Output Voltage.

This happens when the maximum ON time is reached before the current in each phase reaches I_{OCPx} ($I_{INFOx} < 35\mu A$).

Figure 9a shows the maximum output voltage that the device is able to regulate considering the T_{ON} limitation imposed by the previous relationship. If the desired output characteristic crosses the T_{ON} limited maximum output voltage, the output resulting voltage will start to drop after crossing. In this case, the device doesn't perform constant current limitation but only limits the maximum duty cycle following the previous relationship. The output voltage follows the resulting characteristic (dotted in Figure 9b) until UVP is detected or anyway until $I_{FB} = 70\mu A$.

Constant Current Operation

This happens when ON time limitation is reached after the current in each phase reaches I_{OCPx} ($I_{INFOx} > 35\mu A$).

The device enters in Quasi-Constant-Current operation: the low-side mosfets stays ON until the current read becomes lower than I_{OCPx} ($I_{INFOx} < 35\mu A$) skipping clock cycles. The high side mosfets can be turned ON with a TON imposed by the control loop at the next available clock cycle and the device works in the

usual way until another OCP event is detected.

This means that the average current delivered can slightly increase also in Over Current condition since the current ripple increases. In fact, the ON time increases due to the OFF time rise because of the current has to reach the I_{OCPx} bottom. The worst-case condition is when the ON time reaches its maximum value.

When this happens, the device works in Constant Current and the output voltage decrease as the load increase. Crossing the UVP threshold causes the device to reset.

Figure 10 shows this working condition.

It can be observed that the peak current (I_{peak}) is greater than the I_{OCPx} but it can be determined as follow:

$$I_{peak} = I_{OCPx} + \frac{V_{IN} - Vout_{min}}{L} \cdot Ton_{MAX} = I_{OCPx} + \frac{V_{IN} - Vout_{MIN}}{L} \cdot 0.40 \cdot T$$

Where V_{outMIN} is the minimum output voltage (VID-40% as follow).

The device works in Constant-Current, and the output voltage decreases as the load increase, until the output voltage reaches the under-voltage threshold ($Vout_{MIN}$).

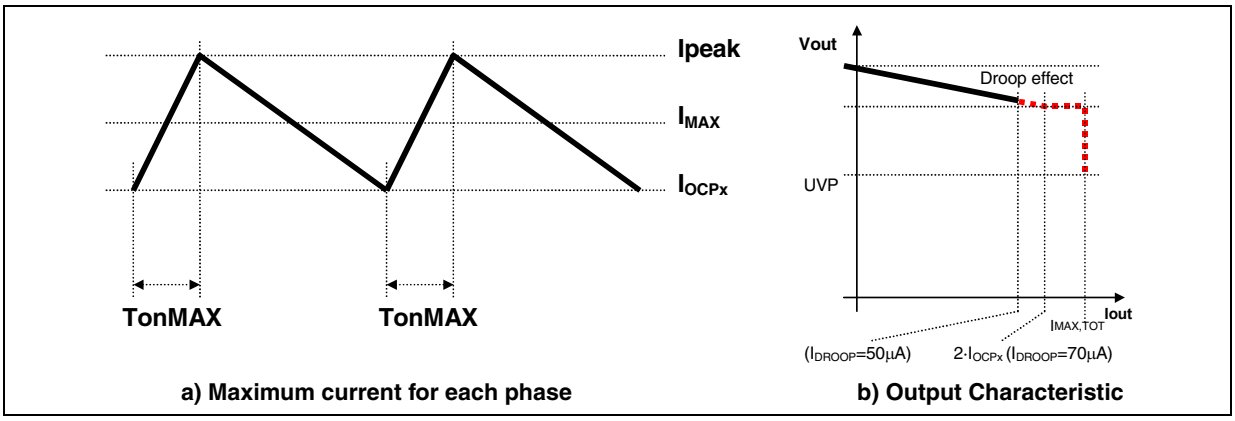
The maximum average current during the Constant-Current behavior results:

$$I_{MAX,TOT} = 2 \cdot I_{MAX} = 2 \cdot \left(I_{OCPx} + \frac{Ipeak - I_{OCPx}}{2} \right)$$

In this particular situation, the switching frequency results reduced. The ON time is the maximum allowed (T_{onMAX}) while the OFF time depends on the application:

$$T_{OFF} = L \cdot \frac{Ipeak - I_{OCPx}}{V_{OUt}} \qquad f = \frac{1}{T_{ONmax} + T_{OFF}}$$

Figure 10. Constant Current operation



a) Maximum current for each phase **b) Output Characteristic**

Over current is set anyway when I_{INFOx} reaches 35µA (I_{FB}=70µA). The full load value is only a convention to work with convenient values for I_{FB}. Since the OCP intervention threshold is fixed, to modify the percentage with respect to the load value, it can be simply considered that, for example, to have on OCP threshold of 200%, this will correspond to I_{INFOx} = 35µA (I_{FB} = 70µA). The full load current will then correspond to I_{INFOx} = 17.5µA (IFB = 35µA).

Once the UVP threshold has been intercepted, the device resets with all power mosfets turned OFF. Another soft start is then performed allowing the device to recover from OCP once the over load cause has been removed.

Crossing the UVP threshold causes the device to reset: all mosfets are turned off and a new soft start is

then implemented allowing the device to recover if the over load cause has been removed.

■ **L6712A - Fixed Maximum Duty Cycle Limitation**

The maximum duty cycle is fixed and constant with the delivered current. The device works in constant current operation once the OCP threshold has overcome. Refer to the above Constant Current section in which only the different value in the maximum duty has to be considered as follow:

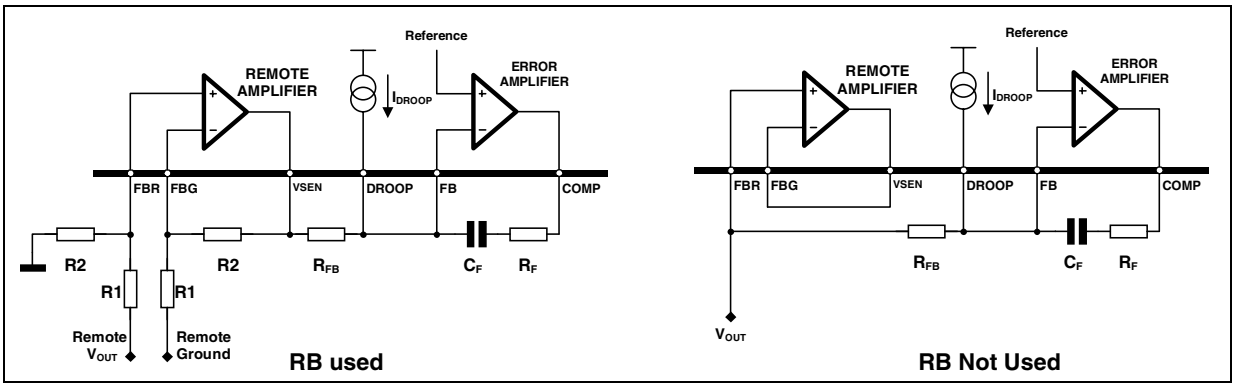
$$I_{peak} = I_{OCPx} + \frac{V_{IN} - Vout_{min}}{L} \cdot Ton_{MAX} = I_{OCPx} + \frac{V_{IN} - Vout_{MIN}}{L} \cdot 0.85 \cdot T$$

All the above reported relationships about the deliverable current once in quasi-constant current and constant current are still valid in this case.

3.5 REMOTE SENSE AMPLIFIER

Remote Sense Amplifier is integrated in order to recover from losses across PCB traces and wiring in high current DC/DC converter remote sense of the regulated voltage is required to maintain precision in the regulation. The integrated amplifier is a low-offset error amplifier; external resistors are needed as shown in Figure 11 to implement a differential remote sense amplifier.

Figure 11. Remote Sense Amplifier Connections



Equal resistors give to the resulting amplifier a unity gain: the programmed reference will be regulated across the remote load.

To regulate output voltages different from the available references, the Remote Amplifier gain can be adjusted simply changing the value of the external resistors as follow (see Figure 11):

$$RA\_Gain = \frac{V_{VSEN}}{Remote\_V_{OUT} - Remote\_GND} = \frac{R2}{R1}$$

to regulate a voltage double of the reference, the above reported gain must be equal to ½.

Modifying the Remote Amplifier Gain (in particular with values higher than 1) allows also to regulate voltages lower than the programmed reference.

Since this Amplifier is connected as a differential amplifier, when calculating the offset introduced in the regulated output voltage, the "native" offset of the amplifier must be multiplied by the term $K_{OS} = [1+(1/RA\_Gain)]$ because a voltage generator insisting on the non-inverting input represents the offset.

If remote sense is not required, it is enough connecting R_{FB} directly to the regulated voltage: VSEN becomes not connected and still senses the output voltage through the remote amplifier. In this case the use of the external resistors R1 and R2 becomes optional and the Remote Sense Amplifier can simply be connected as a "buffer" to keep VSEN at the regulated voltage (See Figure 11). Avoiding use of Remote Amplifier saves its offset in the accuracy calculation but doesn't allow remote sensing.

3.6 INTEGRATED DROOP FUNCTION (Optional)

Droop function realizes dependence between the regulated voltage and the delivered current (Load Regulation). In this way, a part of the drop due to the output capacitor ESR in the load transient is recovered. As shown in Figure 12, the ESR drop is present in any case, but using the droop function the total deviation of the output voltage is minimized.

Connecting DROOP pin and FB pin together, forces a current I_{DROOP}, proportional to the output current, into the feedback resistor R_{FB} implementing the load regulation dependence. If RA_Gain is the Remote Amplifier gain, the Output Characteristic is then given by the following relationship (when droop enabled):

$$V_{OUT} = \frac{1}{RA\_Gain} \cdot (VID - R_{FB} \cdot I_{DROOP}) = \frac{1}{RA\_Gain} \cdot \left(VID - R_{FB} \cdot \frac{R_{SENSE}}{Rg} \cdot I_{OUT}\right)$$

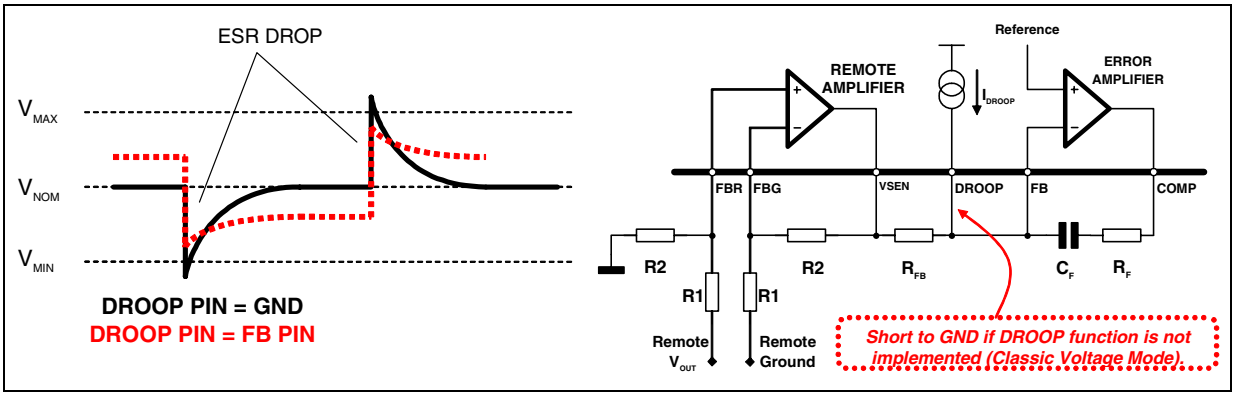
with a remote amplifier gain of 1/2, the regulated output voltage results in being doubled.
The Droop current is equal to 50µA at nominal full load and 70µA at the OC intervention threshold, so the maximum output voltage deviation is equal to:

$$\Delta V_{FULL-POSITIVE-LOAD} = -\frac{1}{RA\_Gain} \cdot R_{FB} \cdot 50\mu A \qquad \Delta V_{OC-INTERVENTION} = -\frac{1}{RA\_Gain} \cdot R_{FB} \cdot 70\mu A$$

Droop function is provided only for positive load; if negative load is applied, and then $I_{INFOx}<0$, no current is sunk from the FB pin. The device regulates at the voltage programmed by the VID.

If this effect is not desired, shorting DROOP pin to SGND, the device regulates as a Voltage Mode Buck converter.

Figure 12. Load Transient response (Left) and DROOP pin connection (Right).



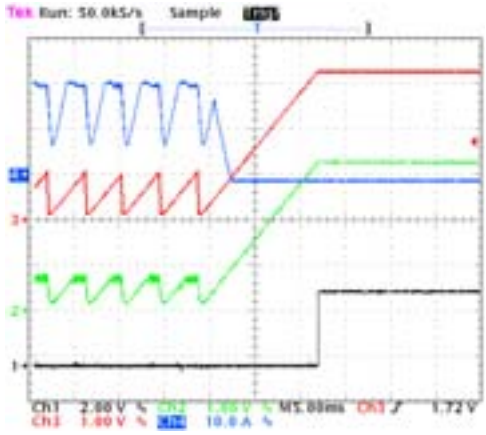
3.7 MONITOR AND PROTECTIONS

The device monitors through pin VSEN the regulated voltage in order to build the PGOOD signal and manage the OVP / UVP conditions.

- **PGOOD.** Power good output is forced low if the voltage sensed by VSEN is not within ±12% (Typ.) of the programmed value (RA_Gain=1). It is an open drain output and it is enabled only after the soft start is finished (2048 clock cycles after start-up). During Soft-Start this pin is forced low.
- **UVP.** If the output voltage monitored by VSEN drops below the 60% of the reference voltage for more than one clock period, the device turns off all mosfets and resets restarting operations with a new soft-start phase (hiccup mode, see Figure 13).
- **OVP.** Enabled once VCC crosses the turn-ON threshold: when the voltage monitored by VSEN reaches 115% (min) of the programmed voltage (or the external reference) the controller permanently switches on both the low-side mosfets and switches off both the high-side mosfets in order to protect the load. The OSC/ FAULT pin is driven high (5V) and power supply (VCC) turn off and on is required to restart operations.
  Both Over Voltage and Under Voltage are active also during soft start (Under Voltage after than the

reference voltage reaches 0.6V). The reference used in this case to determine the UV thresholds is the increasing voltage driven by the 2048 soft start digital counter while the reference used for the OV threshold is the final reference programmed by the VID pins or available on the REF_IN/OUT pin.

Figure 13. UVP Protection & Hiccup Mode.



CH1=PGOOD; CH2=Vout; CH3=REF_OUT; CH4=Iout

3.8 SOFT START AND INHIBIT

At start-up a ramp is generated increasing the loop reference from 0V to the final value programmed by VID in 2048 clock periods as shown in Figure 14.

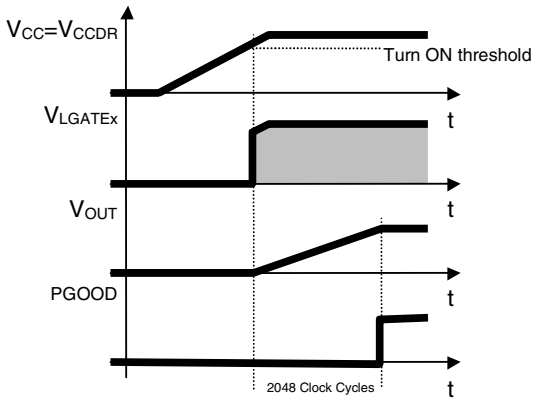
Once the soft start begins, the reference is increased: upper and lower Mosfets begin to switch and the output voltage starts to increase with closed loop regulation. At the end of the digital soft start, the Power Good comparator is enabled and the PGOOD signal is then driven high (See Figure 14).

The Under Voltage comparator is enabled when the increasing reference voltage reaches 0.6V while OVP comparator is always active with a threshold equal to the +15%_min of the final reference.

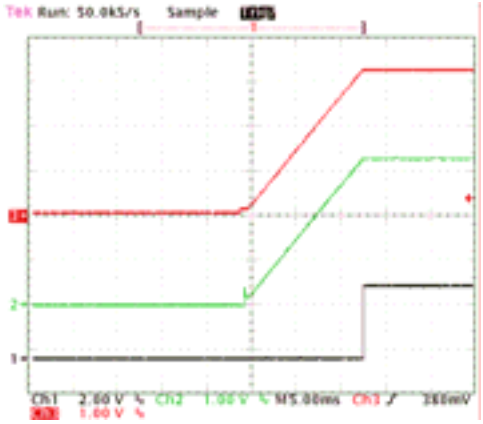
The Soft-Start will not take place, if both VCC and VCCDR pins are not above their own turn-on thresholds.

During normal operation, if any under-voltage is detected on one of the two supplies the device shuts down. Forcing the OSC/INH pin to a voltage lower than 0.5V (Typ.) disables the device: all the power mosfets and protections are turned off until the condition is removed.

Figure 14. Soft Start.



Timing Diagram



Acquisition:
CH1=PGOOD; CH2=VOUT; CH3=REF_OUT

3.9 INPUT CAPACITOR

The input capacitor is designed considering mainly the input RMS current that depends on the duty cycle as reported in Figure 15. Considering the two-phase topology, the input RMS current is highly reduced comparing with a single-phase operation.

It can be observed that the input RMS value is one half of the single-phase equivalent input current in the worst case condition that happens for D=0.25 and D=0.75.

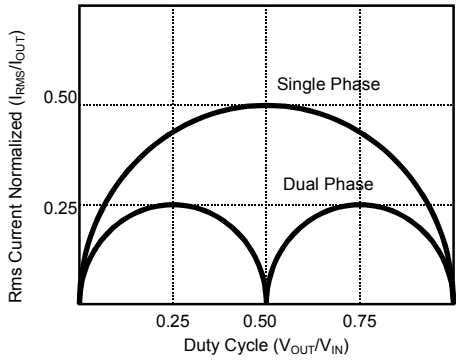
The power dissipated by the input capacitance is then equal to:

$$P_{RMS} = ESR \cdot (I_{RMS})^2$$

Input capacitor is designed in order to sustain the ripple relative to the maximum load duty cycle. To reach the RMS value needed and also to minimize components cost, the input capacitance is realized by more than one physical capacitor. The equivalent RMS current is simply the sum of the single capacitor's RMS current.

Input bulk capacitor must be equally divided between high-side drain mosfets and placed as close as possible to reduce switching noise above all during load transient. Ceramic capacitor can also introduce benefits in high frequency noise de coupling, noise generated by parasitic components along power path.

Figure 15. Input RMS Current vs. Duty Cycle (D) and Driving Relationships.



$$I_{rms} = \begin{cases} \frac{I_{OUT}}{2} \cdot \sqrt{2D \cdot (1-2D)} & \text{if} \quad D < 0.5 \\ \frac{I_{OUT}}{2} \cdot \sqrt{(2D-1) \cdot (2-2D)} & \text{if} \quad D > 0.5 \end{cases}$$

Where $D = V_{OUT}/V_{IN}$

3.10 OUTPUT CAPACITOR

The output capacitor is a basic component for the fast response of the power supply.

Two-phase topology reduces the amount of output capacitance needed because of faster load transient response (switching frequency is doubled at the load connections). Current ripple cancellation due to the 180° phase shift between the two phases also reduces requirements on the output ESR to sustain a specified voltage ripple.

Moreover, if DROOP function is enabled, bigger ESR can be used still keeping the same transient tolerances. In fact, when a load transient is applied to the converter's output, for first few microseconds the current to the load is supplied by the output capacitors. The controller recognizes immediately the load transient and increases the duty cycle, but the current slope is limited by the inductor value.

The output voltage has a first drop due to the current variation inside the capacitor (neglecting the effect of the ESL):

$$\Delta V_{OUT} = \Delta I_{OUT} \cdot ESR$$

A minimum capacitor value is required to sustain the current during the load transient without discharge

it. The voltage drop due to the output capacitor discharge is given by the following equation:

$$\Delta V_{OUT} = \frac{\Delta I^2_{OUT} \cdot L}{4 \cdot C_{OUT} \cdot (V_{In} \cdot d_{max} - V_{OUT})}$$

Where D_{MAX} is the maximum duty cycle value. The lower is the ESR, the lower is the output drop during load transient and the lower is the output voltage static ripple.

3.11 INDUCTOR DESIGN

The inductance value is defined by a compromise between the transient response time, the efficiency, the cost and the size. The inductor has to be calculated to sustain the output and the input voltage variation to maintain the ripple current ΔI_L between 20% and 30% of the maximum output current. The inductance value can be calculated with this relationship:

$$L = \frac{V_{IN} - V_{OUT}}{fs \cdot \Delta I_L} \cdot \frac{V_{OUT}}{V_{IN}}$$

Where F_{SW} is the switching frequency, V_{IN} is the input voltage and V_{OUT} is the output voltage.

Increasing the value of the inductance reduces the ripple current but, at the same time, reduces the converter response time to a load transient. The response time is the time required by the inductor to change its current from initial to final value. Since the inductor has not finished its charging time, the output current is supplied by the output capacitors. Minimizing the response time can minimize the output capacitance required.

The response time to a load transient is different for the application or the removal of the load: if during the application of the load the inductor is charged by a voltage equal to the difference between the input and the output voltage, during the removal it is discharged only by the output voltage. The following expressions give approximate response time for ΔI load transient in case of enough fast compensation network response:

$$t_{application} = \frac{L \cdot \Delta I}{V_{IN} - V_{OUT}} \qquad t_{removal} = \frac{L \cdot \Delta I}{V_{OUT}}$$

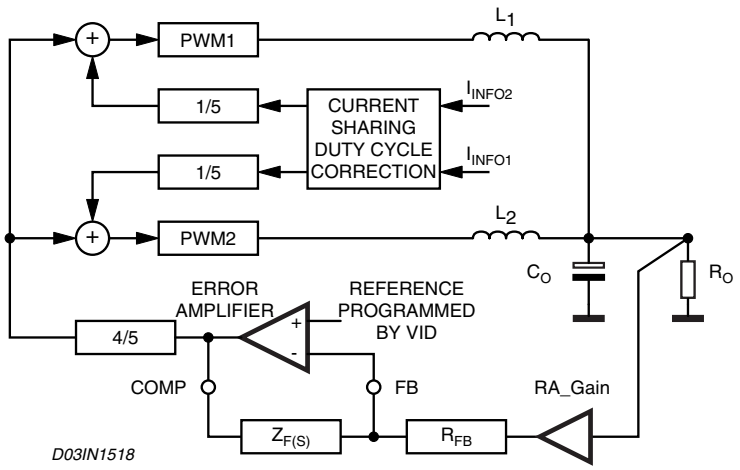
The worst condition depends on the input voltage available and the output voltage selected. Anyway the worst case is the response time after removal of the load with the minimum output voltage programmed and the maximum input voltage available.

3.12 MAIN CONTROL LOOP

The system control loop topology depends on the DROOP pin connection: if connected to FB (droop function active) an Average Current Mode topology must be considered while, if connected to GND (droop function not active) a Voltage Mode topology must be considered instead.

Anyway, the system control loop encloses the Current Sharing control loop to allow proper sharing to the inductor' currents. Each loop gives, with a proper gain, the correction to the PWMs in order to minimize the error in its regulation: the Current Sharing control loop equalize the currents in the inductors while the output voltage control loop fixes the output voltage equal to the reference programmed by VID (with or without the droop effect and with or without considering the Remote Amplifier Gain). Figure 16 reports the block diagram of the main control loop.

Figure 16. Main Control Loop Diagram



3.12.1 Current Sharing (CS) Control Loop

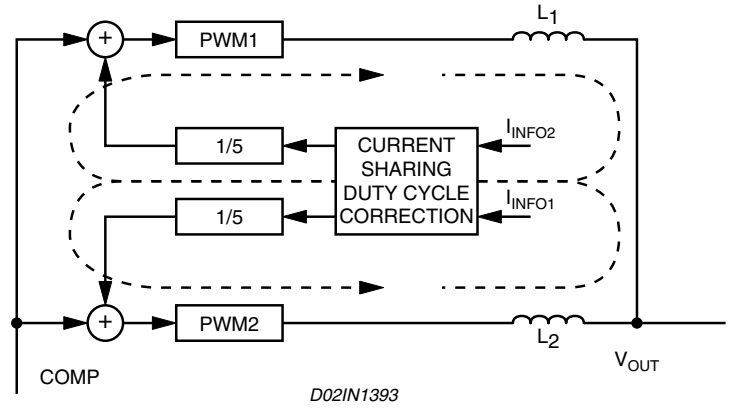
Active current sharing is implemented using the information from Trans conductance differential amplifier. A current reference equal to the average of the read current (I_{AVG}) is internally built; the error between the read current and this reference is converted to a voltage with a proper gain and it is used to adjust the duty cycle whose dominant value is set by the error amplifier at COMP pin (See Figure 17).

The current sharing control is a high bandwidth control loop allowing current sharing even during load transients.

The current sharing error is affected by the choice of external components; choose precise Rg resistor (±1% is necessary) to sense the current. The current sharing error is internally dominated by the voltage offset of Trans conductance differential amplifier; considering a voltage offset equal to 2mV across the sense resistor, the current reading error is given by the following equation:

$$\frac{\Delta I_{READ}}{I_{MAX}} = \frac{2mV}{R_{SENSE} \cdot I_{MAX}}$$

Figure 17. Current Sharing Control Loop.



Where ΔI_{READ} is the difference between one phase current and the ideal current ($I_{MAX}/2$).

For R_{SENSE} = 4mΩ and I_{MAX} = 40A the current sharing error is equal to 2.5%, neglecting errors due to Rg and R_{SENSE} mismatches.

3.12.2 Average Current Mode (ACM) Control Loop (DROOP=FB)

The average current mode control loop is reported in Figure 18. The current information I_{DROOP} sourced by the DROOP pin flows into R_{FB} implementing the dependence of the output voltage from the read current.

The ACM control loop gain results (obtained opening the loop after the COMP pin):

$$G_{LOOP}(s) = -\frac{PWM \cdot Z_F(s) \cdot (R_{DROOP} + RA\_Gain \cdot Z_P(s))}{(Z_P(s) + Z_L(s)) \cdot \left[\frac{Z_F(s)}{A(s)} + \left(1 + \frac{1}{A(s)}\right) \cdot R_{FB}\right]}$$

Where:

- $R_{DROOP} = \frac{Rsense}{Rg} \cdot R_{FB}$ is the equivalent output resistance determined by the droop function;
- $Z_P(s)$ is the impedance resulting by the parallel of the output capacitor (and its ESR) and the applied load Ro;
- $Z_F(s)$ is the compensation network impedance;
- $Z_L(s)$ is the parallel of the two inductor impedance;
- A(s) is the error amplifier gain;
- $PWM = \frac{4}{5} \cdot \frac{V_{IN}}{\Delta V_{OSC}}$ is the ACM PWM transfer function where ΔV_{OSC} is the oscillator ramp amplitude and has a typical value of 3V
- RA_Gain is the Remote Amplifier Gain.

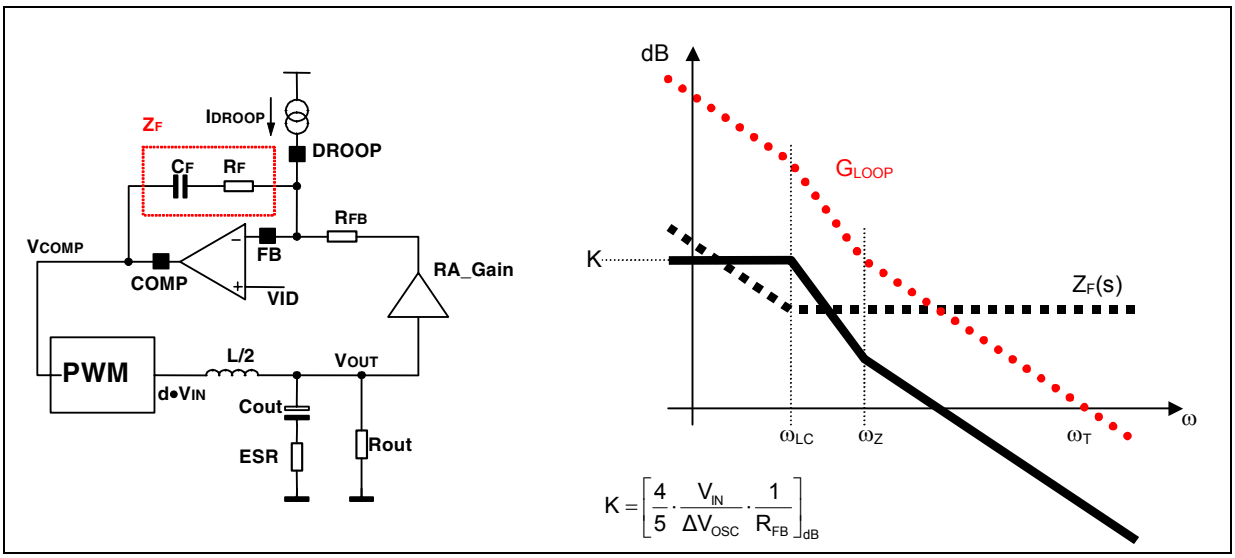
Removing the dependence from the Error Amplifier gain, so assuming this gain high enough, the control loop gain results:

$$G_{LOOP}(s) = -\frac{4}{5} \cdot \frac{V_{IN}}{\Delta V_{OSC}} \cdot \frac{Z_F(s)}{Z_P(s) + Z_L(s)} \cdot \left(\frac{Rs}{Rg} + \frac{RA\_Gain \cdot Z_P(s)}{R_{FB}}\right)$$

Considering now that in the application of interest it can be assumed that $R_o >> R_L$; $ESR << R_o$ and $R_{DROOP} << R_o$, it results:

$$G_{LOOP}(s) = -\frac{4}{5} \cdot \frac{V_{IN}}{\Delta V_{OSC}} \cdot \frac{Z_F(s)}{R_{FB}} \cdot \frac{1 + s \cdot Co \cdot \left(\frac{R_{DROOP}}{RA\_Gain} + ESR\right)}{s^2 \cdot Co \cdot \frac{L}{2} + s \cdot \left[\frac{L}{2 \cdot Ro} + Co \cdot ESR + Co \cdot \frac{L}{2 \cdot Ro}\right] + 1} \cdot RA\_Gain$$

Figure 18. ACM Control Loop Gain Block Diagram (left) and Bode Diagram (right).



The ACM control loop gain is designed to obtain a high DC gain to minimize static error and cross the 0dB axes with a constant -20dB/dec slope with the desired crossover frequency ω_T. Neglecting the effect of ZF(s), the transfer function has one zero and two poles. Both the poles are fixed once the output filter is designed and the zero is fixed by ESR and the Droop resistance.

To obtain the desired shape an R_F-C_F series network is considered for the $Z_F(s)$ implementation. A zero at $\omega_F=1/R_FC_F$ is then introduced together with an integrator. This integrator minimizes the static error while placing the zero in correspondence with the L-C resonance a simple -20dB/dec shape of the gain is assured (See Figure 18). In fact, considering the usual value for the output filter, the LC resonance results to be at frequency lower than the above reported zero.Compensation network can be simply designed placing $\omega_Z = \omega_{LC}$ and imposing the cross-over frequency ω_T as desired obtaining:

$$R_F = \frac{R_{FB} \cdot \Delta V_{OSC}}{V_{IN}} \cdot \frac{5}{4} \cdot \omega_T \cdot \frac{L}{2 \cdot \left(\frac{R_{DROOP}}{RA\_Gain} + ESR\right)} \qquad C_F = \frac{\sqrt{Co \cdot \frac{L}{2}}}{R_F}$$

3.12.3 Voltage Mode (VM) Control Loop (DROOP = SGND)

Disconnecting the DROOP pin from the Control Loop, the system topology becomes a Voltage Mode. The simplest way to compensate this loop still keeping the same compensation network consists in placing the RF-CF zero in correspondence with the L-C filter resonance.

The loop gain becomes now:

$$G_{LOOP}(s) = -\frac{V_{IN}}{\Delta V_{OSC}} \cdot \frac{Z_F(s)}{R_{FB}} \cdot \frac{Z_P(s)}{Z_P(s) + Z_L(s)} \cdot RA\_Gain$$

3.13 LAYOUT GUIDELINES

Since the device manages control functions and high-current drivers, layout is one of the most important things to consider when designing such high current applications.

A good layout solution can generate a benefit in lowering power dissipation on the power paths, reducing radiation and a proper connection between signal and power ground can optimize the performance of the control loops.

Integrated power drivers reduce components count and interconnections between control functions and drivers, reducing the board space.

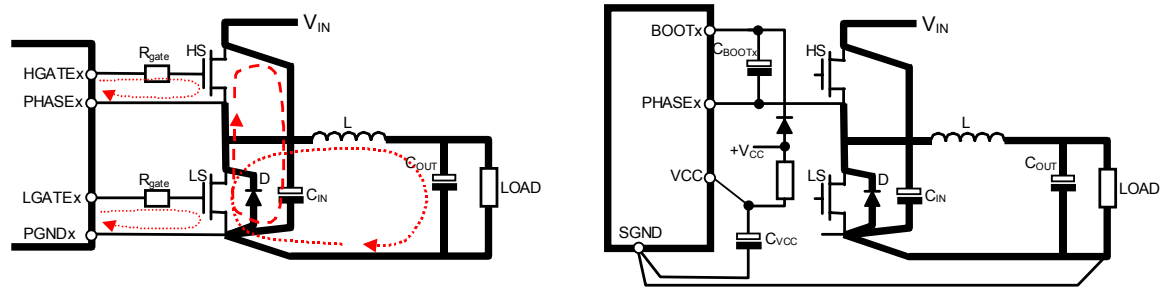
Here below are listed the main points to focus on when starting a new layout and rules are suggested for a correct implementation.

■ Power Connections.

These are the connections where switching and continuous current flows from the input supply towards the load. The first priority when placing components has to be reserved to this power section, minimizing the length of each connection as much as possible.

To minimize noise and voltage spikes (EMI and losses) these interconnections must be a part of a power plane and anyway realized by wide and thick copper traces.

Figure 19. Power connections and related connections layout guidelines (same for both phases).



a. PCB power and ground planes areas

b. PCB small signal components placement

The critical components, i.e. the power transistors, must be located as close as possible, together and to the controller. Considering that the "electrical" components reported in figure are composed by more than one "physical" component, a ground plane or "star" grounding connection is suggested to minimize effects due to multiple connections.

Figure 19a shows the details of the power connections involved and the current loops. The input capacitance (C_{IN}), or at least a portion of the total capacitance needed, has to be placed close to the power section in order to eliminate the stray inductance generated by the copper traces. Low ESR and ESL capacitors are required.

■ Power Connections Related.

Figure 19b shows some small signal components placement, and how and where to mix signal and power ground planes. The distance from drivers and mosfet gates should be reduced as much as possible. Propagation delay times as well as for the voltage spikes generated by the distributed inductance along the copper traces are so minimized.

In fact, the further the mosfet is from the device, the longer is the interconnecting gate trace and as a consequence, the higher are the voltage spikes corresponding to the gate PWM rising and falling signals. Even if these spikes are clamped by inherent internal diodes, propagation delays, noise and potential causes of instabilities are introduced jeopardizing good system behavior. One important consequence is that the switching losses for the high side mosfet are significantly increased.

For this reason, it is suggested to have the device oriented with the driver side towards the mosfets and the GATEx and PHASEx traces walking together toward the high side mosfet in order to minimize distance (see Figure 20). In addition, since the PHASEx pin is the return path for the high side driver, this pin must be connected directly to the High Side mosfet Source pin to have a proper driving for this mosfet. For the LS mosfets, the return path is the PGND pin: it can be connected directly to the power ground plane (if

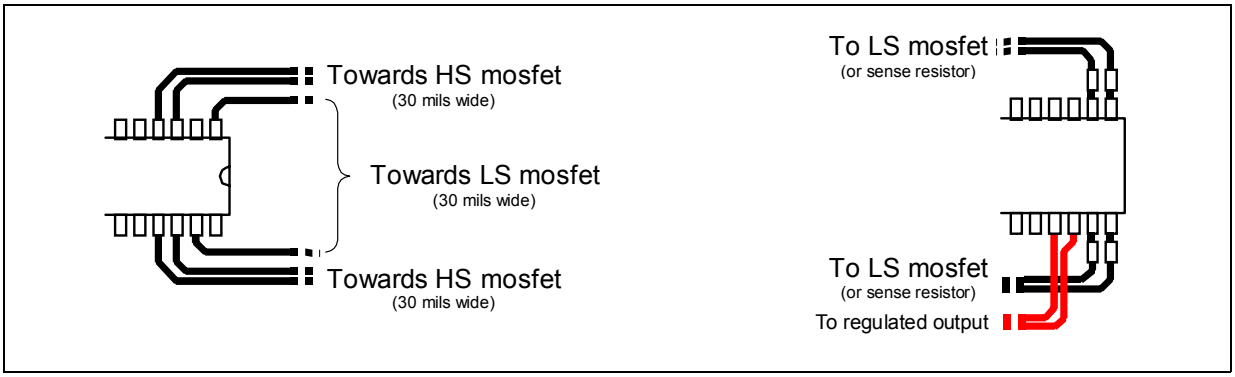
implemented) or in the same way to the LS mosfets Source pin. GATEx and PHASEx connections (and also PGND when no power ground plane is implemented) must also be designed to handle current peaks in excess of 2A (30 mils wide is suggested).

Gate resistors of few ohms help in reducing the power dissipated by the IC without compromising the system efficiency.

The placement of other components is also important:

- The bootstrap capacitor must be placed as close as possible to the BOOTx and PHASEx pins to minimize the loop that is created.
- Decoupling capacitor from VCC and SGND placed as close as possible to the involved pins.
- Decoupling capacitor from VCCDR and PGND placed as close as possible to those pins. This capacitor sustains the peak currents requested by the low-side mosfet drivers.
- Refer to SGND all the sensible components such as frequency set-up resistor (when present) and Remote Amplifier Divider.
- Connect SGND to PGND plane on a single point to improve noise immunity. Connect at the load side (output capacitor) if Remote Sense is not implemented to avoid undesirable load regulation effect.
- An additional 100nF ceramic capacitor is suggested to place near HS mosfet drain. This helps in reducing noise.
- PHASE pin spikes. Since the HS mosfet switches in hard mode, heavy voltage spikes can be observed on the PHASE pins. If these voltage spikes overcome the max breakdown voltage of the pin, the device can absorb energy and it can cause damages. The voltage spikes must be limited by proper layout, the use of gate resistors, Schottky diodes in parallel to the low side mosfets and/or snubber network on the low side mosfets, to a value lower than 26V, for 20ns, at F_{SW} of 600kHz max.
- Boot Capacitor Extra Charge. Systems that do not use Schottky diodes in parallel to the LS mosfet might show big negative spikes on the phase pin. This spike can be limited as well as the positive spike but has an additional consequence: it causes the bootstrap capacitor to be over-charged. This extra-charge can cause, in the worst case condition of maximum input voltage and during particular transients, that boot-to-phase voltage overcomes the abs. max. ratings also causing device failures. It is then suggested in this cases to limit this extra-charge by:
  - adding a small resistor in series to the boot diode (one resistor can be enough for all the diodes if placed upstream the diode anode)
  - using low capacitance diodes.

Figure 20. Device orientation (left) and sense nets routing (right).



■ Sense Connections.

Remote Amplifier: Place the external resistors near the device to minimize noise injection and refer to SGND. The connections for these resistors (from the remote load) must be routed as parallel nets in order to compensate losses along the output power traces and also to avoid the pick-up of any noise. Connecting these pins in points far from the load will cause a non-optimum load regulation, increasing output tol-

erance.

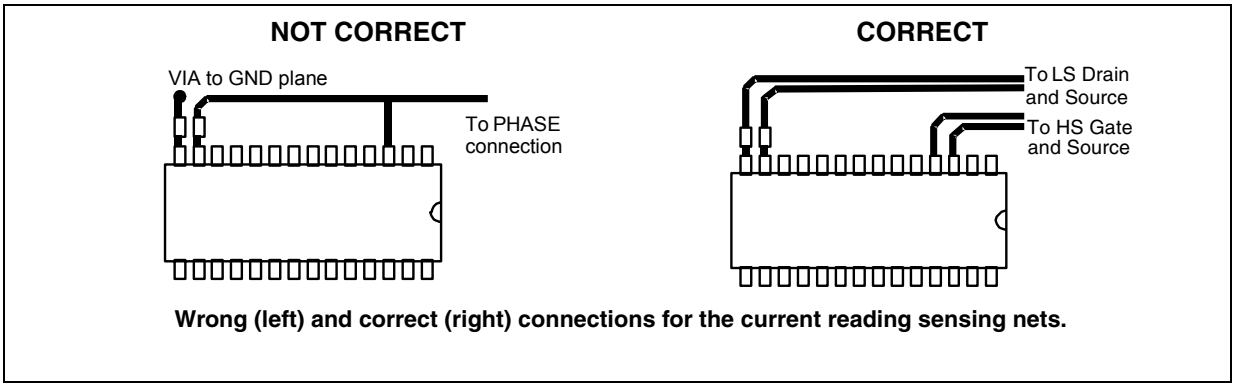
Current Reading: The Rg resistor has to be placed as close as possible to the ISENx and PGNDSx pins in order to limit the noise injection into the device. The PCB traces connecting these resistors to the reading point must be routed as parallel traces in order to avoid the pick-up of any noise. It's also important to avoid any offset in the measurement and to get a better precision, to connect the traces as close as possible to the sensing elements, dedicated current sense resistor or low side mosfet R_{dsON}.

Moreover, when using the low side mosfet R_{dsON} as current sense element, the ISENx pin is practically connected to the PHASEx pin. DO NOT CONNECT THE PINS TOGETHER AND THEN TO THE HS SOURCE! The device won't work properly because of the noise generated by the return of the high side driver. In this case route two separate nets: connect the PHASEx pin to the HS Source (route together with HGATEx) with a wide net (30 mils) and the ISENx pin to the LS Drain (route together with PGNDSx). Moreover, the PGNDSx pin is always connected, through the Rg resistor, to the PGND: DO NOT CONNECT DIRECTLY TO THE PGND! In this case the device won't work properly. Route anyway to the LS mosfet source (together with ISENx net).

Right and wrong connections are reported in Figure 21.

Symmetrical layout is also suggested to avoid any unbalance between the two phases of the converter.

Figure 21. PCB layout connections for sense nets.



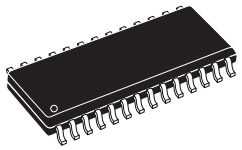
Wrong (left) and correct (right) connections for the current reading sensing nets.

4 Package informations

Figure 22. SO-28 Mechanical Data & Package Dimensions

DIM.	mm			inch		
	MIN.	TYP.	MAX.	MIN.	TYP.	MAX.
A			2.65			0.104
a1	0.1		0.3	0.004		0.012
b	0.35		0.49	0.014		0.019
b1	0.23		0.32	0.009		0.013
C		0.5			0.020	
c1	45° (typ.)					
D	17.7		18.1	0.697		0.713
E	10		10.65	0.394		0.419
e		1.27			0.050	
e3		16.51			0.65	
F	7.4		7.6	0.291		0.299
L	0.4		1.27	0.016		0.050
S	8 ° (max.)					

OUTLINE AND MECHANICAL DATA



SO-28

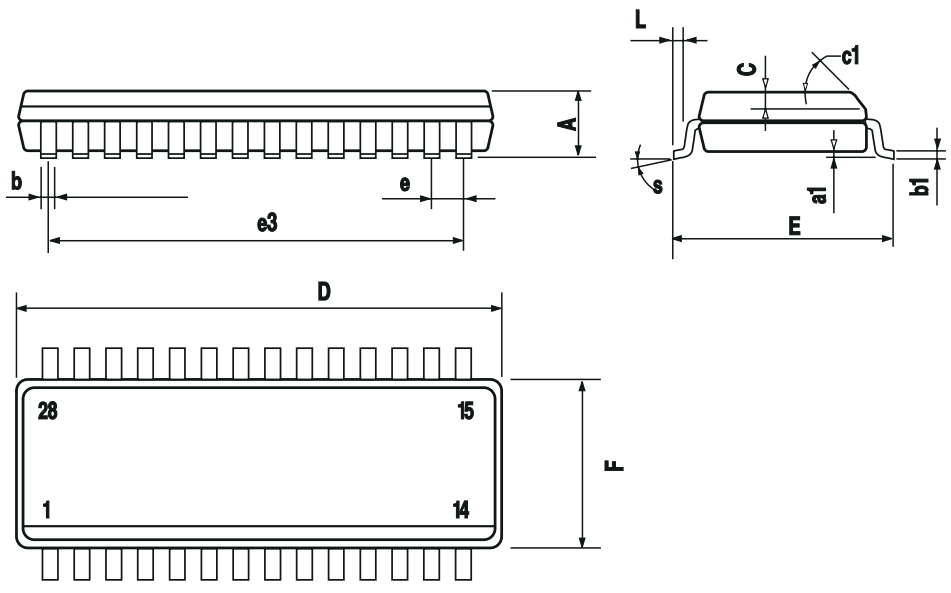
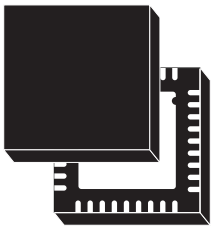


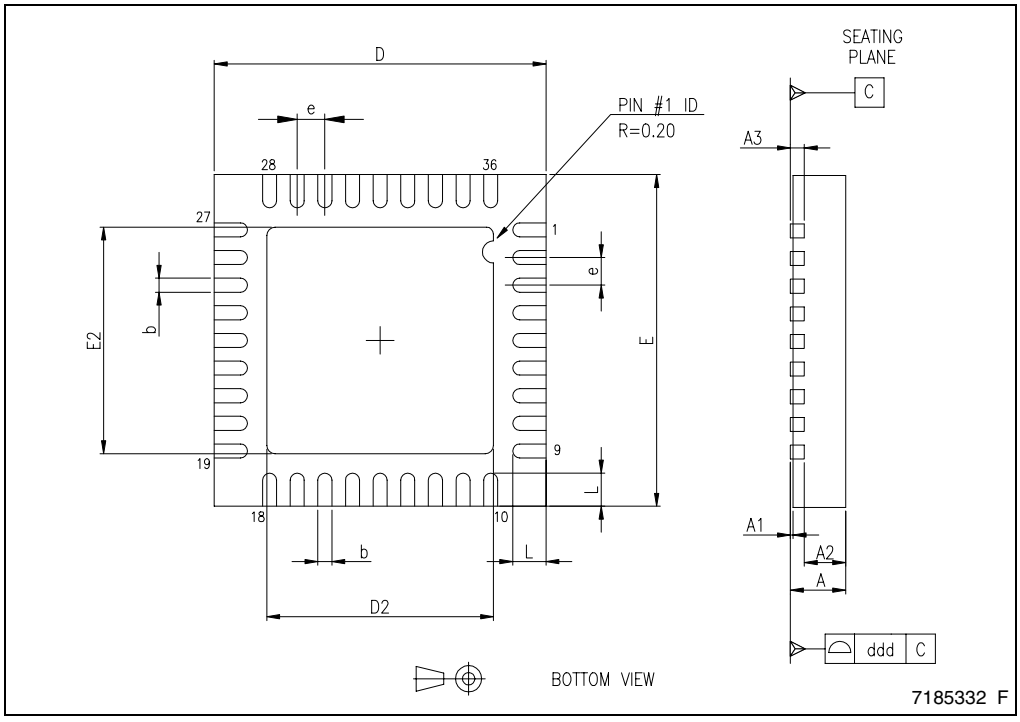
Figure 23. VFQFPN-36 Mechanical Data & Package Dimensions

DIM.	mm			inch		
	MIN.	TYP.	MAX.	MIN.	TYP.	MAX.
A	0.800	0.900	1.000	0.031	0.035	0.039
A1		0.020	0.050		0.0008	0.0019
A2		0.650	1.000		0.025	0.039
A3		0.250			0.01	
b	0.180	0.230	0.300	0.007	0.009	0.012
D	5.875	6.000	6.125	0.231	0.236	0.241
D2	1.750	3.700	4.250	0.069	0.146	0.167
E	5.875	6.000	6.125	0.231	0.236	0.241
E2	1.750	3.700	4.250	0.069	0.146	0.167
e	0.450	0.500	0.550	0.018	0.020	0.022
L	0.350	0.550	0.750	0.014	0.022	0.029
ddd			0.080			0.003

OUTLINE AND MECHANICAL DATA



VFQFPN-36 (6x6x1.0mm)
Very **F**ine **Q**uad **F**lat **P**ackage **N**o lead



5 Revision History

Table 7. Revision History

Date	Revision	Description of Changes
March 2004	2	First Issue in EDOCS.
June 2005	3	Changed look and feel. Inserted “Boot Capacitor Extra Charge” paragraph to page 27.

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