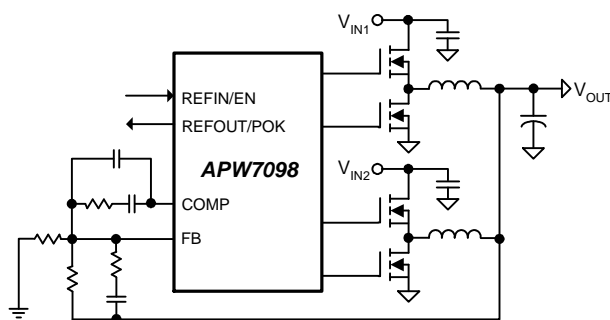


## Two-Phase Buck PWM Controller with Integrated MOSFET Drivers

### Features

- **Voltage-Mode Operation with Current Sharing**
  - Adjustable Feedback Compensation
  - Fast Load Transient Response
- **Operate with 8V~13.2V<sub>CC</sub> Supply Voltage**
- **Selectable External or Internal 0.6V Reference**
  - $\pm 1.5\%$  Accuracy Over Temperature
- **Support Single- and Two-Phase Operations**
- **5VCC and Buffered Reference Outputs**
- **8~12V Gate Drivers with Internal Bootstrap Diode**
- **Lossless Inductor DCR Current Sensing**
- **Selectable Operation Frequency**
  - 150k/300k/400kHz per Phase
- **Power-OK Indicator Output**
  - Regulated 1.5V on REFOUT/POK
- **Adjustable Over-Current Protection (OCP)**
- **Accurate Load Line (DROOP) Programming**
- **Adjustable Soft-Start**
- **Over-Voltage Protection (OVP)**
- **Under-Voltage Protection (UVP)**
- **Over-Temperature Protection (OTP)**
- **QFN4x4 24-Lead Package (QFN4x4-24A)**
- **Lead Free and Green Devices Available (RoHS Compliant)**

### Simplified Application Circuit



### General Description

The APW7098, two-phase PWM control IC, provides a precision voltage regulation system for advanced graphic microprocessors in graphics card applications. The integration of power MOSFET drivers into the controller IC and reduces the number of external parts for a cost and space saving power management solution.

The APW7098 uses a voltage-mode PWM architecture, operating with fixed-frequency, to provides excellent load transient response. The device uses the voltage across the DCRs of the inductors for current sensing. Load line voltage positioning (DROOP), channel-current balance, and over-current protection are accomplished through continuous inductor DCR current sensing.

The MODE pin programs single- or two- phase operation. When IC operates in two-phase mode normally, it can transfer two-phase mode to single-phase mode at liberty. Nevertheless, once operates in single-phase mode, the operation mode is latched. It is required to toggle SS, REFIN/EN or 5VCC pin to reset the IC. Such feature of the MODE pin makes the APW7098 ideally suitable for dual power input applications, such as PCIE interfaced graphic cards.

This control IC's protection features include a set of sophisticated over-temperature, over-voltage, under-voltage, and over-current protections. Over-voltage results in the converter turning the lower MOSFETs on to clamp the rising output voltage and protects the microprocessor. The over-current protection level is set through external resistors. The device also provides a power-on-reset function and a programmable soft-start to prevent wrong operation and limit the input surge current during power-on or start-up.


The APW7098 is available in a QFN4x4-24A package.

### Applications

- **Graphics Card GPU Core Power Supply**
- **Motherboard Chipset or DDR SDRAM Core Power Supply**
- **On-board High Power PWM Converter with Output Current up to 60A**

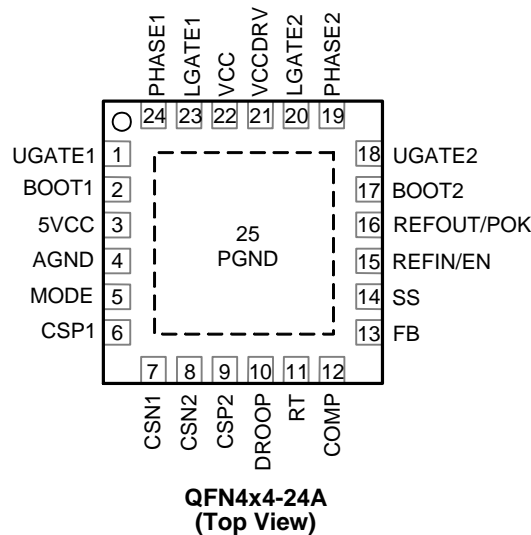
ANPEC reserves the right to make changes to improve reliability or manufacturability without notice, and advise customers to obtain the latest version of relevant information to verify before placing orders.

## Ordering and Marking Information

<p>APW7098    □□□-□□□</p> <p> <span style="margin-left: 100px;">└─</span> Assembly Material  <span style="margin-left: 80px;">└─</span> Handling Code  <span style="margin-left: 60px;">└─</span> Temperature Range  <span style="margin-left: 40px;">└─</span> Package Code         </p>	<p>Package Code QA : QFN4x4-24A</p> <p>Operating Ambient Temperature Range E : -20 to 70 °C</p> <p>Handling Code TR : Tape &amp; Reel</p> <p>Assembly Material G : Halogen and Lead Free Device</p>
<p>APW7098 QA : </p>	<p>XXXXX - Date Code</p>

Note: ANPEC lead-free products contain molding compounds/die attach materials and 100% matte tin plate termination finish; which are fully compliant with RoHS. ANPEC lead-free products meet or exceed the lead-free requirements of IPC/JEDEC J-STD-020D for MSL classification at lead-free peak reflow temperature. ANPEC defines “Green” to mean lead-free (RoHS compliant) and halogen free (Br or Cl does not exceed 900ppm by weight in homogeneous material and total of Br and Cl does not exceed 1500ppm by weight).

## Pin Configuration



## Absolute Maximum Ratings (Note 1)

Symbol	Parameter	Rating	Unit
$V_{CC}$	VCC Supply Voltage (VCC to AGND)	-0.3 ~ 15	V
$V_{BOOT1/2}$	BOOT1/2 Voltage (BOOT1/2 to PHASE1/2)	-0.3 ~ 15	V
	UGATE1/2 Voltage (UGATE1/2 to PHASE1/2)	<200ns pulse width >200ns pulse width	V
	LGATE1/2 Voltage (LGATE1/2 to PGND)	<200ns pulse width >200ns pulse width	V
	PHASE1/2 Voltage (PHASE1/2 to PGND)	<200ns pulse width >200ns pulse width	V

## Absolute Maximum Ratings (Cont.) (Note 1)

Symbol	Parameter	Rating	Unit
	BOOT1/2 to AGND Voltage <200ns pulse width >200ns pulse width	-0.3 ~ 42 -0.3 ~ 30	V
	VCCDRV to AGND Voltage	-0.3 ~ 15	V
V <sub>5VCC</sub>	5VCC Supply Voltage (5VCC to AGND, V <sub>5VCC</sub> < V <sub>CC</sub> +0.3V)	-0.3 ~ 7	V
	REFIN/EN, MODE to AGND Voltage	-0.3 ~ 7	V
	Input Voltage (REFOUT/POK, SS, FB, COMP, DROOP, RT, CSP1/2, CSN1/2 to AGND)	-0.3 ~ V <sub>5VCC</sub> +0.3	V
	PGND to AGND Voltage	-0.3 ~ +0.3	V
P <sub>DMAX</sub>	Maximum Power Dissipation	Limited Internally	W
	Maximum Junction Temperature	150	°C
T <sub>STG</sub>	Storage Temperature Range	-65 ~ 150	°C
T <sub>SDR</sub>	Maximum Soldering Temperature, 10 Seconds	260	°C

Note 1: Absolute Maximum Ratings are those values beyond which the life of a device may be impaired. Exposure to absolute maximum rating conditions for extended periods may affect device reliability.

## Thermal Characteristics

Symbol	Parameter	Typical Value	Unit
θ <sub>JA</sub>	Junction-to-Ambient Resistance <sup>(Note 2)</sup> QFN4x4-24A	45	°C/W
θ <sub>JC</sub>	Junction-to-Case Resistance <sup>(Note 3)</sup> QFN4x4-24A	7	

Note 2 : θ<sub>JA</sub> is measured with the component mounted on a high effective thermal conductivity test board in free air. The exposed pad of QFN4x4-24A is soldered directly on the PCB.

Note 3: The case temperature is measured at the center of the exposed pad on the underside of the QFN4x4-24A package.

## Recommended Operating Conditions (Note 4)

Symbol	Parameter	Range	Unit
V <sub>CC</sub>	VCC Supply Voltage	8 ~ 13.2	V
V <sub>5VCC</sub>	5VCC Supply Voltage (V <sub>5VCC</sub> < V <sub>CC</sub> +0.3V)	5 ± 5%	V
V <sub>OUT</sub>	Converter Output Voltage	0.6 ~ 2.5	V
V <sub>IN1</sub>	PWM 1 Converter Input Voltage	3.1 ~ 13.2	V
V <sub>IN2</sub>	PWM 2 Converter Input Voltage	3.1 ~ 13.2	V
I <sub>OUT</sub>	Converter Output Current	~ 60	A
V <sub>REFIN/EN</sub>	REFIN/EN Input Voltage	0 ~ 2	V
T <sub>A</sub>	Ambient Temperature	-20 ~ 70	°C
T <sub>J</sub>	Junction Temperature	-20 ~ 125	°C
C <sub>VCC</sub>	Linear Regulator Output Capacitor	0.8 ~ 15	μF
C <sub>5VCC</sub>	5VCC Linear Regulator Output Capacitor	0.8 ~ 15	μF

Note 4 : Refer to the typical application circuits.

## Electrical Characteristics

Refer to the typical application circuits. These specifications apply over  $V_{IN}=12V$ ,  $V_{OUT}=1.2V$  and  $T_A=-20 \sim 70^\circ C$ , unless otherwise specified. Typical values are at  $T_A=25^\circ C$ . The  $V_{5VCC}$  is supplied by the internal regulator.

Symbol	Parameter	Test Conditions	APW7098			Unit
			Min.	Typ.	Max.	
<b>SUPPLY CURRENT</b>						
$I_{CC}$	VCC Nominal Supply Current	UGATEx and LGATEx Open, FB forced above regulation point	-	5	10	mA
$I_{SD}$	VCC Shutdown Supply Current	SS/EN=GND	-	5	-	mA
<b>POWER-ON-RESET (POR) AND OPERATION PHASE SELECTION</b>						
$V_{5VCC\_THR}$	5VCC Rising Threshold Voltage		4.2	4.5	4.8	V
	5VCC POR Hysteresis		0.4	0.58	0.76	V
	MODE Rising Threshold Voltage	$V_{MODE}$ Rising	0.77	0.8	0.83	V
$I_{MODE}$	MODE Pin Input Current		-100	-	+100	nA
<b>VCC LINEAR CONTROLLER</b>						
$V_{RRG\_VCC}$	Regulated Voltage on VCC	$I_O=0A$ , $R_{PULL-UP}=1k\Omega$	8	8.5	9	V
	Maximum VCCDRV Sink Current	$V_{CC} = V_{REG\_VCC} + 200mV$ , $V_{VCCDRV} = 8V$	5	-	-	mA
<b>5VCC LINEAR REGULATOR</b>						
$V_{REG\_5VCC}$	Output Voltage	$I_O = 0A$ , $V_{CC} = 8V$	4.75	5	5.25	V
	Line Regulation	$I_O = 0A$ , $V_{CC} = 8V \sim 13.2V$	-20	-	20	mV
	Load Regulation	$I_O = 3mA$ , $V_{CC} > 8V$	-200	-	200	mV
	Current-Limit	5VCC = GND	20	30	-	mA
<b>REFERENCE VOLTAGE</b>						
$V_{REF}$	Regulated Voltage on FB pin	Internal reference voltage used	-	0.6	-	V
	Accuracy	$T_A=25^\circ C$	-1	-	+1	%
		Over temperature	-1.5	-	+1.5	
$I_{FB}$	FB Pin Input Current		-100	-	+100	nA
	REFIN/EN Voltage Offset	$V_{FB} - V_{REFIN/EN}$ , $V_{REFIN/EN} = 0.6V \sim 1.5V$	-5	-	5	mV
$V_{REFIN/EN\_THR}$	Device Enable Voltage Threshold	On REFIN/EN pin, $V_{REFIN/EN}$ rising	0.37	0.4	0.43	V
	Device Enable Voltage Hysteresis		-	50	-	mV
	Internal/External Reference Selection Voltage Threshold	On REFIN/EN pin	2.1	2.5	3.0	V
	Reference Selection Debounce Time	$V_{REFIN/EN}$ falling, Switching to external reference	-	20	-	$\mu s$
$I_{REFIN/EN}$	REFIN/EN Pin Input Current		-100	-	+100	nA
$V_{POK}$	REFOUT/POK Output Voltage		-	1.5	-	V
	REFOUT/POK Accuracy	$I_O = 0 \sim 3mA$ , $T_A=25^\circ C$	-2	-	+2	%
		$I_O = 0 \sim 3mA$ , Over temperature	-3	-	+3	
	REFOUT/POK Current-Limit	REFOUT/POK = GND	4	8	15	mA
	REFOUT/POK Pull-Low Resistance	$I_{REFOUT/POK} = 5mA$	-	70	100	$\Omega$

## Electrical Characteristics (Cont.)

Refer to the typical application circuits. These specifications apply over  $V_{IN}=12V$ ,  $V_{OUT}=1.2V$  and  $T_A = -20 \sim 70^\circ C$ , unless otherwise specified. Typical values are at  $T_A=25^\circ C$ . The  $V_{5VCC}$  is supplied by the internal regulator.

Symbol	Parameter	Test Conditions	APW7098			Unit	
			Min.	Typ.	Max.		
<b>ERROR AMPLIFIER</b>							
	DC Gain	$R_L = 10k\Omega$ to the ground	-	85	-	dB	
	Gain-Bandwidth Product	$C_L = 100pF$ , $R_L = 10k\Omega$ to the ground	-	20	-	MHz	
	Slew Rate	$C_L = 100pF$ , $I_O = \pm 400\mu A$	-	8	-	V/ $\mu s$	
	Upper Clamp Voltage	$I_O = 1mA$	2.7	3.0	-	V	
	Lower Clamp Voltage	$I_O = -1mA$	-	-	0.1	V	
	COMP Pull-Low Resistance	In fault or shutdown condition	-	2	-	k $\Omega$	
<b>OSCILLATOR</b>							
$F_{OSC}$	Oscillator Frequency	RT = GND	135	150	165	kHz	
		RT = Floating	270	300	330		
		RT = 5VCC	360	400	440		
$\Delta V_{OSC1/2}$	Oscillator Sawtooth Amplitude		-	1.5	-	V	
$I_{RT}$	RT Input Current	RT = GND/5VCC(5V)	-100	-	+100	$\mu A$	
	RT 5VCC Level	For $F_{OSC} = 150kHz$	$V_{5VCC}-0.5$	-	-	V	
	RT Floating Voltage	For $F_{OSC} = 300kHz$	1.2	3.6	$V_{5VCC}-1.2$	V	
	RT GND Level	For $F_{OSC} = 400kHz$	-	-	0.3	V	
	Maximum Duty Cycle		85	88	-	%	
<b>MOSFET GATE DRIVERS</b>							
	UGATE1/2 Source Current	$V_{BOOT} = 12V$ , $V_{UGATE}-V_{PHASE} = 2V$	-	2.6	-	A	
	UGATE1/2 Sink Current	$V_{BOOT} = 12V$ , $V_{UGATE}-V_{PHASE} = 2V$	-	1	-	A	
	LGATE1/2 Source Current	$V_{CC} = 12V$ , $V_{LGATE} = 2V$	-	2.6	-	A	
	LGATE1/2 Sink Current	$V_{CC} = 12V$ , $V_{LGATE} = 2V$	-	1.4	-	A	
	UGATE1/2 Source Resistance	$V_{BOOT} = 12V$ , 100mA Source Current	-	2.5	3.75	$\Omega$	
	UGATE1/2 Sink Resistance	$V_{BOOT} = 12V$ , 100mA Sink Current	-	2	3	$\Omega$	
	LGATE1/2 Source Resistance	$V_{CC} = 12V$ , 100mA Source Current	-	2	3	$\Omega$	
	LGATE1/2 Sink Resistance	$V_{CC} = 12V$ , 100mA Sink Current	-	1.4	2.1	$\Omega$	
$T_D$	Dead-Time		-	30	-	ns	
<b>CURRENT SENSE AND DROOP FUNCTION</b>							
$I_{CSP}$	CSP1/2 Pin Input Current		-100	-	+100	nA	
$I_{CSN}$	CSN1/2 Maximum Output Current	$R_{CSN1/2} = 2k\Omega$ ,	Sourcing current	80	-	-	$\mu A$
			Sinking current	15	-	-	
	Current Sense Amplifier Bandwidth		-	3	-	MHz	
	DROOP Output Current Accuracy	$R_{DROOP} = 2k\Omega$ , $V_{DROOP} = 0.005V$	-	50	-	$\mu A$	
	DROOP Accuracy	$\Delta V_{FB} = V_{DROOP}/20$ , $V_{DROOP} = 1V$	-5	-	+5	mV	
	Current Difference Between Channel1/2 and Average Current		-10	-	+10	%	

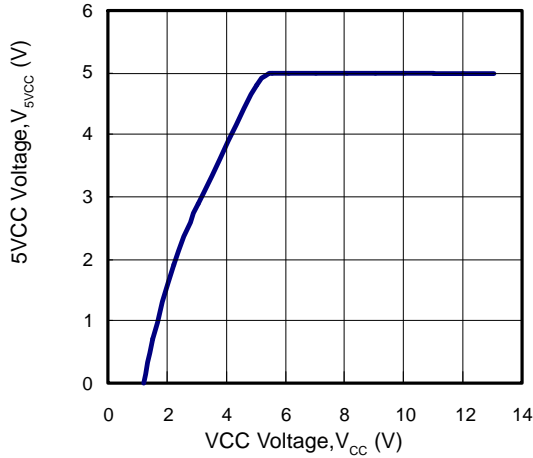
## Electrical Characteristics (Cont.)

Refer to the typical application circuits. These specifications apply over  $V_{IN}=12V$ ,  $V_{OUT}=1.2V$  and  $T_A = -20 \sim 70^\circ C$ , unless otherwise specified. Typical values are at  $T_A=25^\circ C$ . The  $V_{SVCC}$  is supplied by the internal regulator.

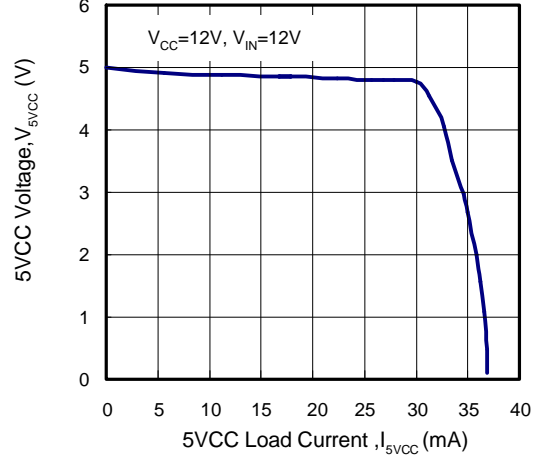
Symbol	Parameter	Test Conditions	APW7098			Unit
			Min.	Typ.	Max.	
<b>SOFT-START AND ENABLE</b>						
$I_{SS}$	Soft-Start Current Source	Flowing out of SS pin	8	10	12	$\mu A$
	Soft-Start Complete Threshold		-	3.2	-	V
	SS Pull-low Resistance		-	10	18	k $\Omega$
<b>POWER-OK AND PROTECTIONS</b>						
	Over-Current Trip Level	$I_{CS1} + I_{CS2}$	110	120	140	$\mu A$
$V_{UV}$	FB Under-Voltage Threshold	$\sim 2\mu s$ noise filter, $V_{FB}$ falling, Percentage of $V_R$ at Error Amplifier	40	50	60	%
$V_{POK\_L}$	POK Lower Threshold		-	87.5	-	%
$V_{OV}$ , $V_{POK\_H}$	FB Over-Voltage Threshold and POK Upper Threshold	$\sim 2\mu s$ noise filter, $V_{FB}$ rising Percentage of $V_R$ at Error Amplifier	115	125	135	%
	FB Over-Voltage Hysteresis		-	60	80	mV
$T_{OTR}$	Over-Temperature Trip Level	$T_J$ rising	-	150	-	$^\circ C$
	Over-Temperature Hysteresis		-	50	-	$^\circ C$

Typical Operating Characteristics

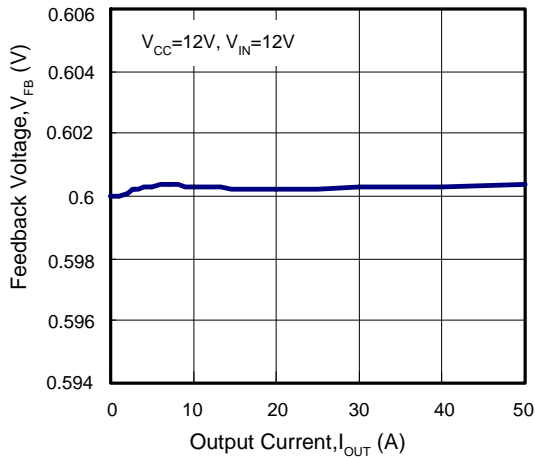
5VCC Line Regulation



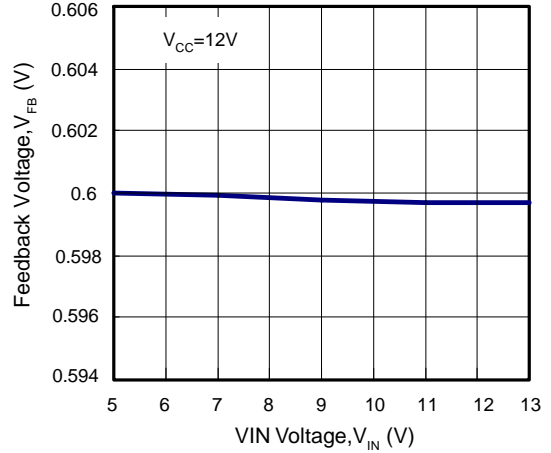
5VCC Load Regulation



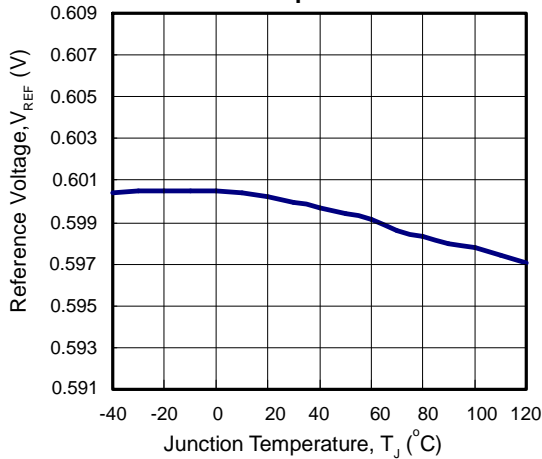
Output Voltage Load Regulation



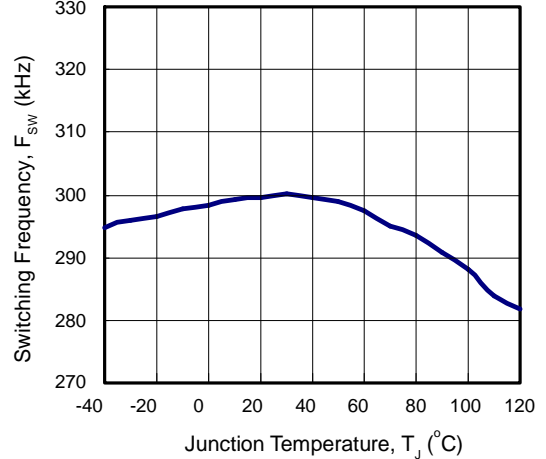
Output Voltage Line Regulation



Reference Voltage Accuracy Over Temperature

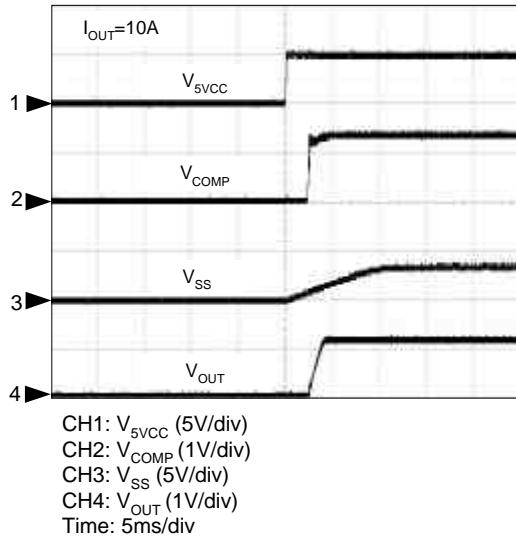


Switching Frequency Over Temperature

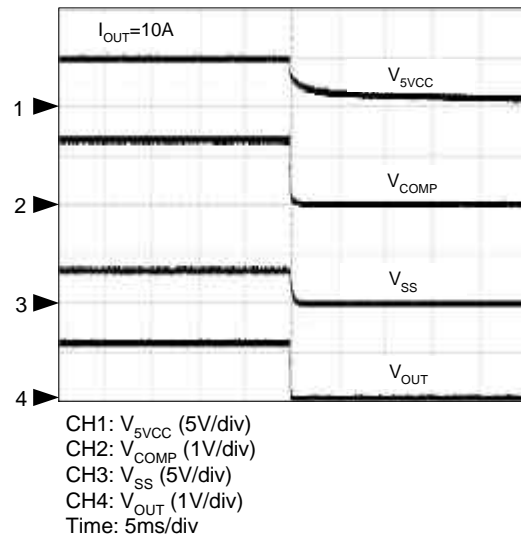


Operating Waveforms

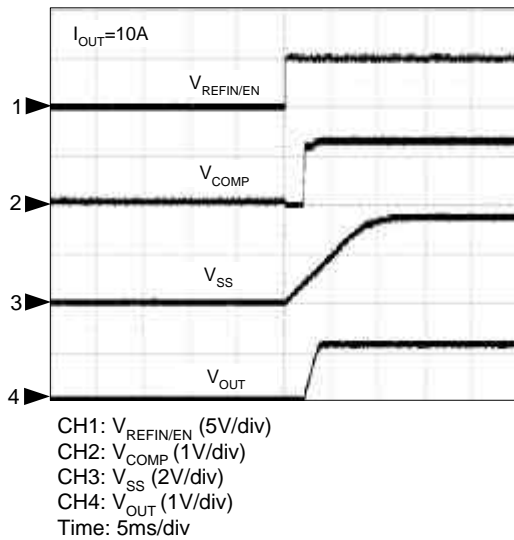
Power On



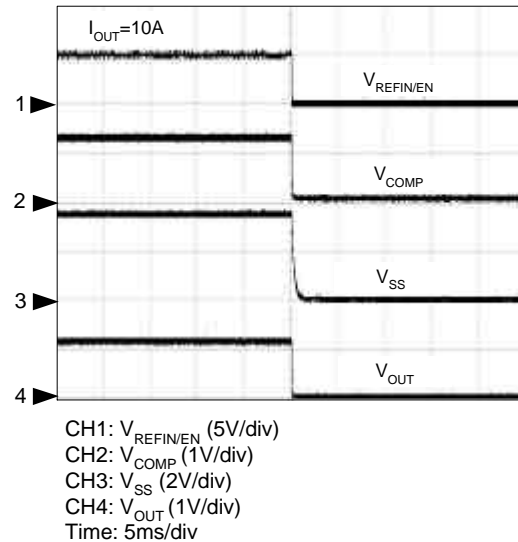
Power Off



Enable by REFIN/EN Pin



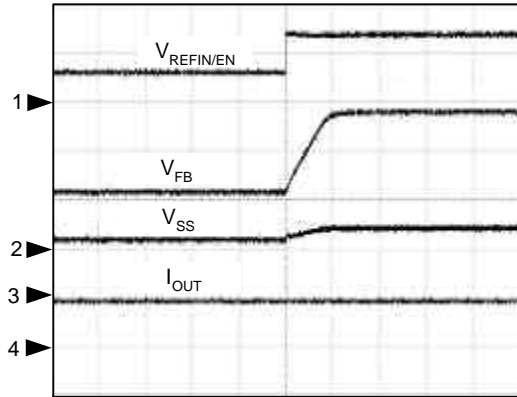
Shutdown by REFIN/EN Pin





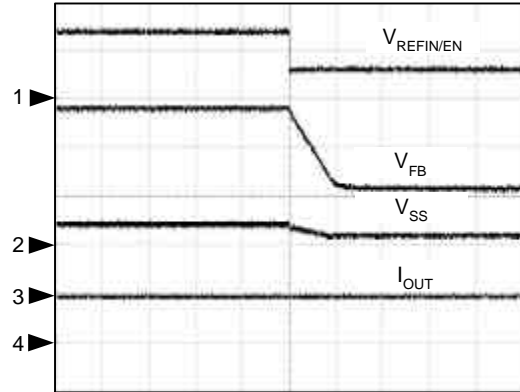
Operating Waveforms (Cont.)

External Step-Up Reference by  $V_{REFIN/EN}$



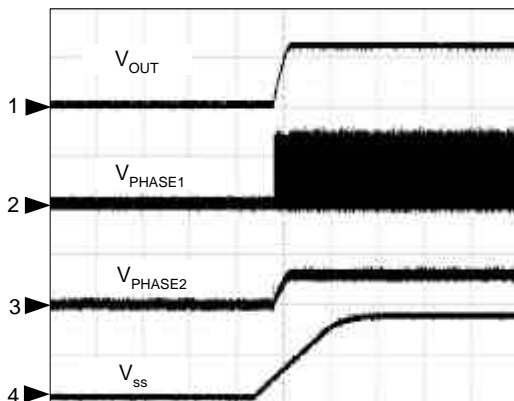
CH1:  $V_{REFIN/EN}$  (1V/div)  
 CH2:  $V_{FB}$  (500mV/div)  
 CH3:  $V_{SS}$  (1V/div)  
 CH4:  $I_{OUT}$  (10A/div)  
 Time: 200 $\mu$ s/div

External Step-Down Reference by  $V_{REFIN/EN}$



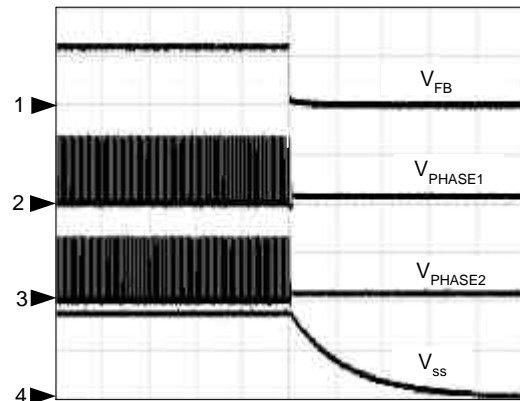
CH1:  $V_{REFIN/EN}$  (1V/div)  
 CH2:  $V_{FB}$  (500mV/div)  
 CH3:  $V_{SS}$  (1V/div)  
 CH4:  $I_{OUT}$  (10A/div)  
 Time: 200 $\mu$ s/div

Power On Without  $V_{IN2}$  Voltage



CH1:  $V_{OUT}$  (1V/div)  
 CH2:  $V_{PHASE1}$  (10V/div)  
 CH3:  $V_{PHASE2}$  (2V/div)  
 CH4:  $V_{SS}$  (2V/div)  
 Time: 5ms/div

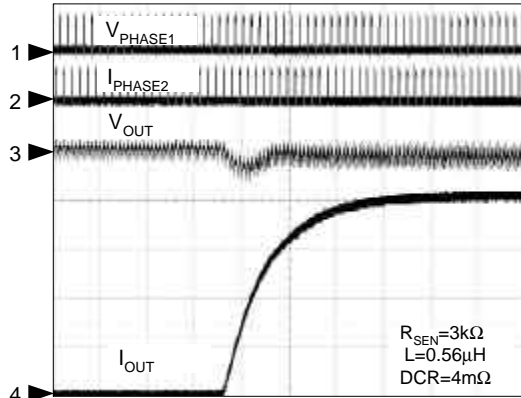
Under-Voltage Protection (UVP)



CH1:  $V_{FB}$  (500mV/div)  
 CH2:  $V_{PHASE1}$  (10V/div)  
 CH3:  $V_{PHASE2}$  (10V/div)  
 CH4:  $V_{SS}$  (2V/div)  
 Time: 200 $\mu$ s/div

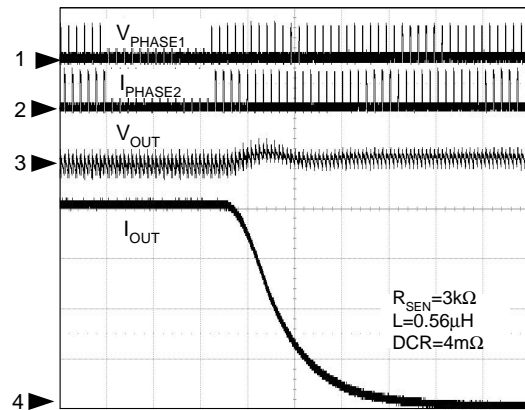
Operating Waveforms (Cont.)

Load Transient , 0A==>40A



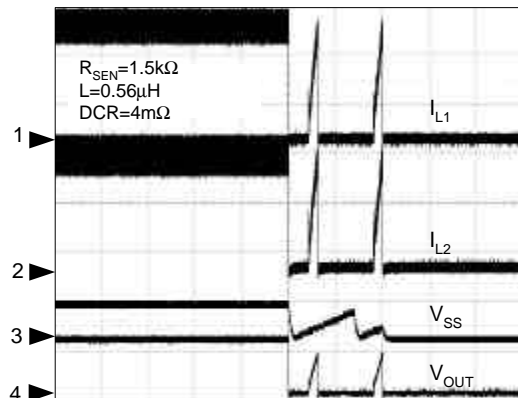
CH1:  $V_{PHASE1}$  (20V/div)  
 CH2:  $I_{PHASE2}$  (20A/div)  
 CH3:  $V_{OUT}$  (AC, 200mV/div)  
 CH4:  $I_{OUT}$  (10A/div)  
 Time: 20μs/div

Load Transient , 40A==>0A



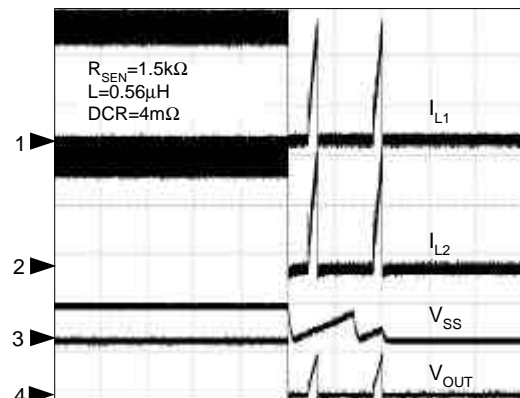
CH1:  $V_{PHASE1}$  (20V/div)  
 CH2:  $I_{PHASE2}$  (20A/div)  
 CH3:  $V_{OUT}$  (AC, 200mV/div)  
 CH4:  $I_{OUT}$  (10A/div)  
 Time: 20μs/div

OCP at Slow Slew  $I_{OUT}$



CH1:  $I_{L1}$  (10A/div)  
 CH2:  $I_{L2}$  (10A/div)  
 CH3:  $V_{SS}$  (5V/div)  
 CH4:  $V_{OUT}$  (1V/div)  
 Time: 5ms/div

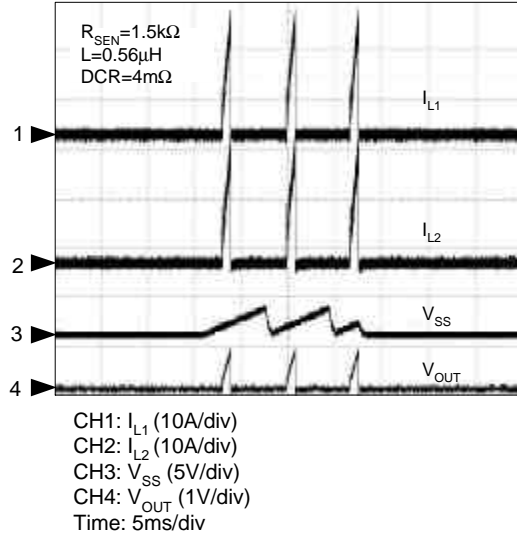
Short-Circuit Test After Power On



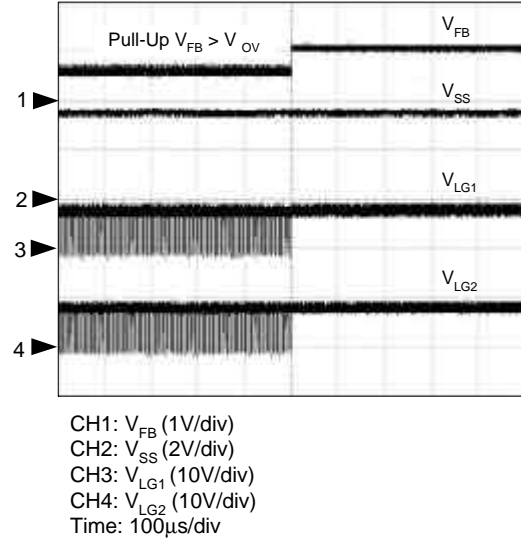
CH1:  $I_{L1}$  (10A/div)  
 CH2:  $I_{L2}$  (10A/div)  
 CH3:  $V_{SS}$  (5V/div)  
 CH4:  $V_{OUT}$  (1V/div)  
 Time: 5ms/div

Operating Waveforms (Cont.)

Short-Circuit Test Before Power On



OVP After Power On



Pin Description

PIN		FUNCTION
NO.	NAME	
1	UGATE1	High-side Gate Driver Output for channel 1. Connect this pin to the gate of high-side MOSFET. This pin is monitored by the adaptive shoot-through protection circuitry to determine when the high-side MOSFET has turned off.
2	BOOT1	Bootstrap Supply for the floating high-side gate driver of channel 1. Connect the Bootstrap capacitor between the BOOT1 pin and the PHASE1 pin to form a bootstrap circuit. The bootstrap capacitor provides the charge to turn on the high-side MOSFET. Typical values for $C_{BOOT}$ ranged from 0.1μF to 1μF. Ensure that $C_{BOOT}$ is placed near the IC.
3	5VCC	Internal Regulator Output. This is the output pin of the linear regulator, which is converting power from VCC and provides output current up to 20mA minimums for internal bias and external usage.
4	AGND	Signal Ground for the IC. All voltage levels are measured with respect to this pin. Tie this pin to the ground island/plane through the lowest impedance connection available.
5	MODE	Operation Phase Selection Input. Pulling this pin lower than 0.64V sets two-phase operation with both channels enabled. Pulling this pin higher than 0.8V sets single-phase operation with the channel 2 disabled. Once operating in single-phase mode, the operation mode is latched. It is required to toggle SS, REFIN/EN, or 5VCC pin to reset the IC.
6	CSP1	Positive Input of current sensing Amplifier for channel 1. This pin combined with CSN1 senses the inductor current through an RC network.
7	CSN1	Negative Input of current sensing amplifier for channel 1. This pin combined with CSP1 senses the inductor current through an RC network.
8	CSN2	Negative Input of current sensing amplifier for channel 2. This pin combined with CSP2 senses the inductor current through an RC network.
9	CSP2	Positive Input of current sensing Amplifier for Channel 2. This pin combined with CSN2 senses the inductor current through an RC network.

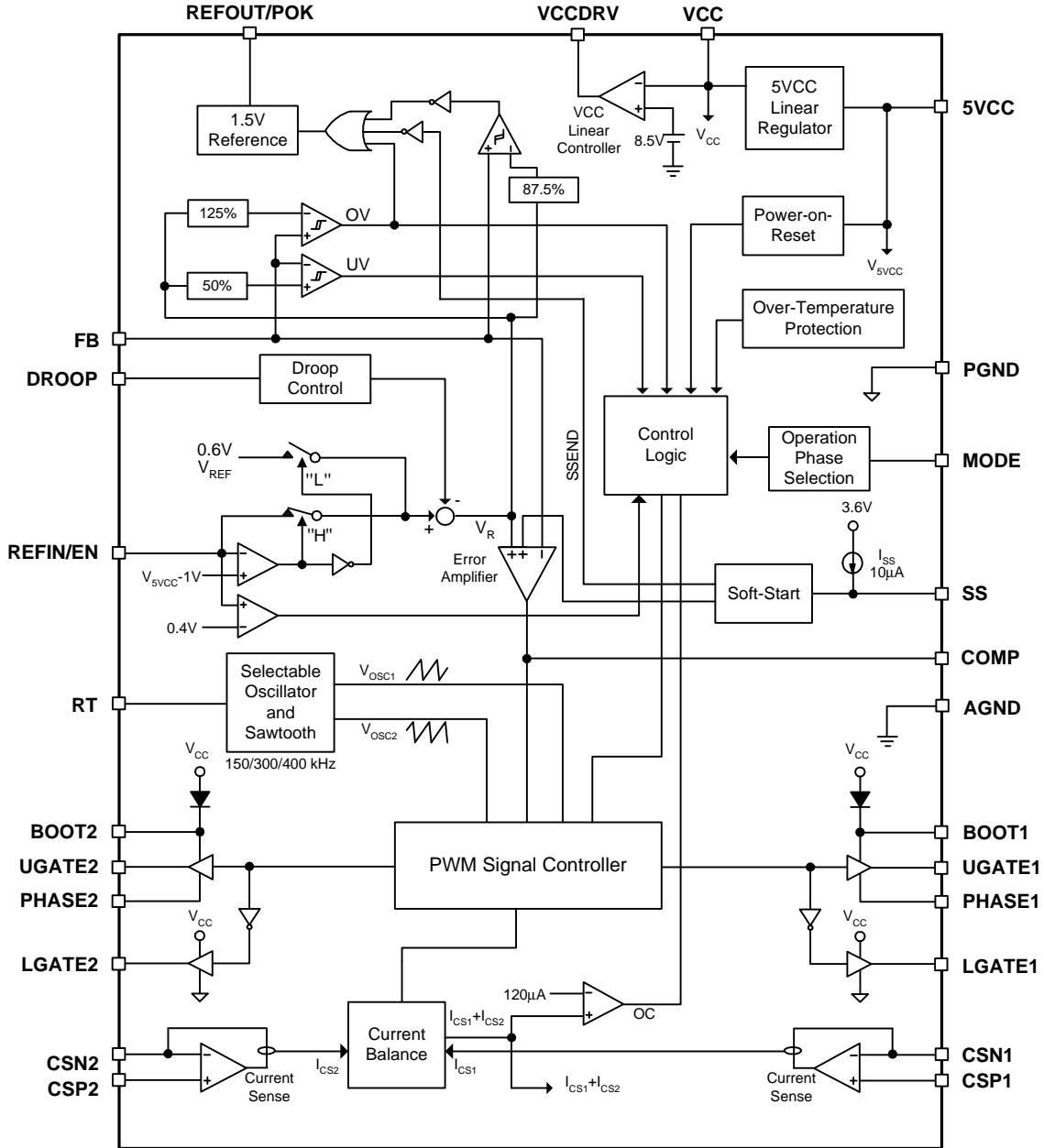
Pin Description (Cont.)

PIN		FUNCTION	
NO.	NAME		
10	DROOP	Load Line (droop) Setting. Connect a resistor between this pin and AGND to set the droop. A sourcing current, proportional to output current is present on the DROOP pin. The droop scale factor is set by the resistors (connected with CSP1, CSP2, and DROOP), resistance of the output inductors, and the internal voltage divider with the ratio of 5%.	
11	RT	Operating Frequency Setting. The three-level input pin sets the operating frequency for each channel.	
		RT	Operating Frequency (kHz)
		GND	150
		Floating	300
	5VCC	400	
12	COMP	Error Amplifier Output. Connect the compensation network between COMP, FB, and V <sub>OUT</sub> for Type 2 or Type 3 feedback compensation.	
13	FB	Feedback Voltage. This pin is the inverting input to the error comparator. A resistor divider from the output to the AGND is used to set the regulation voltage.	
14	SS	Soft-start Current Output. Connect a capacitor from this pin to the AGND to set the soft-start interval. Pulling the voltage on this pin below 0.5V causes COMP to pull low and then shuts off the output.	
15	REFIN/EN	External Reference and Enable Input. The IC uses the voltage (V <sub>REFIN/EN</sub> ) as reference voltage of the converter with soft-start control. If this pin is driven by an external voltage ranged from 0.4V to 2V. The IC is disabled if the voltage is below 0.4V (typical). If external reference is not available, then connect this pin to 5VCC for internal 0.6V reference.	
16	REFOUT/POK	Power-OK and 1.5V Reference Output. This pin is a reference output used to indicate the status of the voltages on SS pin and FB pin. REFOUT/POK provides 1.5V reference if V <sub>FB</sub> > 87.5% of reference (V <sub>R</sub> ).	
17	BOOT2	Bootstrap Supply for the floating high-side gate driver of channel 2. Connect the Bootstrap capacitor between the BOOT2 pin and the PHASE2 pin to form a bootstrap circuit. The bootstrap capacitor provides the charge to turn on the high-side MOSFET. Typical values for C <sub>BOOT</sub> range from 0.1μF to 1μF. Ensure that C <sub>BOOT</sub> is placed near the IC.	
18	UGATE2	High-side Gate Driver Output for Channel 2. Connect this pin to the gate of high-side MOSFET. This pin is monitored by the adaptive shoot-through protection circuitry to determine when the high-side MOSFET has turned off.	
19	PHASE2	Switch Node for Channel 2. Connect this pin to the source of high-side MOSFET and the drain of the low-side MOSFET. This pin is used as sink for UGATE2 driver. This pin is also monitored by the adaptive shoot-through protection circuitry to determine when the high-side MOSFET has turned off. An Schottky diode between this pin and the ground is recommended to reduce negative transient voltage that is common in a power supply system.	
20	LGATE2	Low-side Gate Driver Output for Channel 2. Connect this pin to the gate of low-side MOSFET. This pin is monitored by the adaptive shoot-through protection circuitry to determine when the low-side MOSFET has turned off.	
21	VCCDRV	Drive for External Linear Regulator. This pin is the drive output for the external linear regulator. Connect this pin to base/gate of NPN/NMOS transistor as the pass element.	
22	VCC	Supply Voltage. This pin along with VCCDRV pin and external pass element provides 8.5V regulated bias supply, low-side gate drivers, and the bootstrap circuit for high-side drivers. This pin can receive a well-decoupled 8V~13.2V supply voltage alone if the VCCDRV is left open. Ensure that this pin is bypassed by a ceramic capacitor next to the pin.	
23	LGATE1	Low-side Gate Driver Output for Channel 1. Connect this pin to the gate of low-side MOSFET. This pin is monitored by the adaptive shoot-through protection circuitry to determine when the low-side MOSFET has turned off.	

## Pin Description (Cont.)

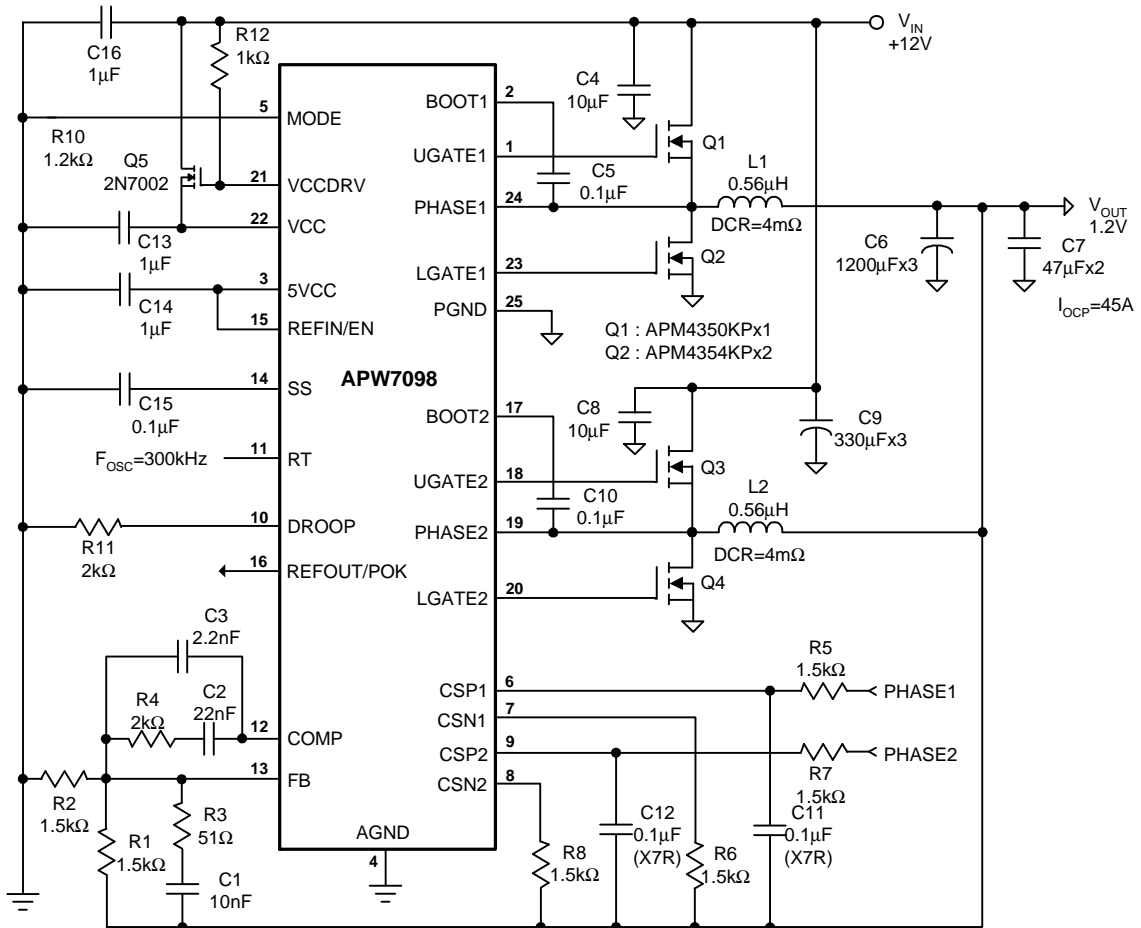
PIN		FUNCTION
NO.	NAME	
24	PHASE1	Switch Node for Channel 1. Connect this pin to the source of high-side MOSFET and the drain of the low-side MOSFET. This pin is used as sink for UGATT1 driver. This pin is also monitored by the adaptive shoot-through protection circuitry to determine when the high-side MOSFET has turned off. An Schottky diode between this pin and the ground is recommended to reduce negative transient voltage, which is common in a power supply system.
25	PGND	Power Ground for the low-side gate drivers. Connect this pin to the source of low-side MOSFETs. This pin is used as sink for LGATE1 and LGATE2 drivers.

Block Diagram



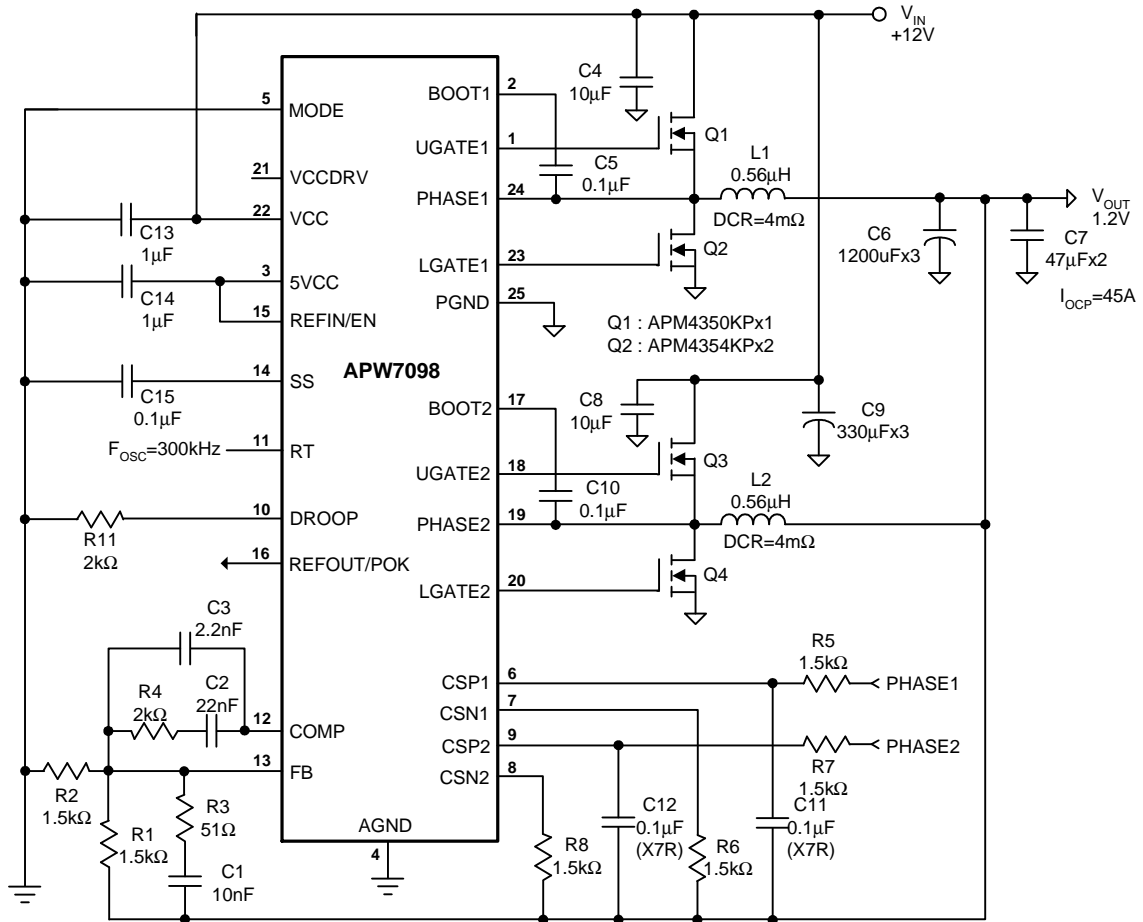
## Typical Application Circuits

1. APW7098 PWM Converter With 8V Gate Drive



## Typical Application Circuits (Cont.)

2. APW7098 PWM Converter With 12V Gate Drive





## Function Description

### VCC Linear Controller

The VCC linear-regulator controller is an analog gain block with an open-drain n-channel output. It drives an external NPN or N-channel MOSFET pass transistor with a 1kΩ (typical) pull-up resistor and senses the feedback voltage via VCC pin. The regulator uses a 1μF (minimum) ceramic output capacitor and is designed to deliver 100mA (at 8.5V) for VCC.

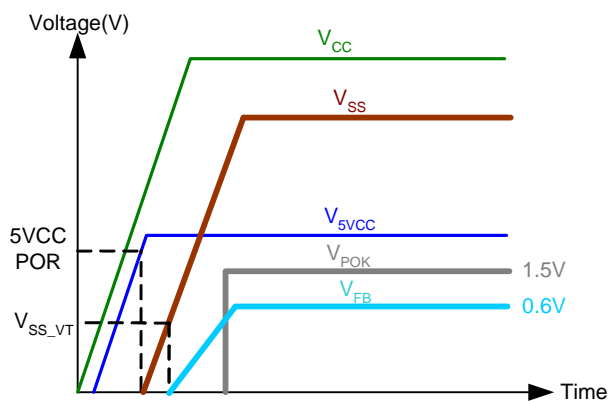
### 5VCC Linear Regulator

5VCC is the output terminal of the internal 5V linear regulator which regulates a 5V voltage on 5VCC by controlling an internal bypass transistor between VCC and 5VCC. The linear regulator powers the internal control circuitry and is stable with a low-ESR ceramic output capacitor. Bypass 5VCC to GND with a ceramic capacitor of at least 1μF. Place the capacitor physically close to the IC to provide good noise decoupling. The linear regulator can also provide output current up to 20mA for external loads. The linear regulator with current-limit protection can protect itself during over-load or short-circuit conditions on 5VCC pin.

The 5VCC linear regulator stops regulating in Over-Temperature Protection. When the junction temperature is cooled by 50°C, the 5VCC linear regulator starts to regulate the output voltage again.

### 5VCC Power-On-Reset (POR) and REFIN/EN (External Reference and Enable Input)

Figure 1 shows the power sequence. The APW7098 keeps monitoring the voltage on 5VCC pin to prevent wrong logic operations which may occur when 5VCC voltage is not high enough for the internal control circuitry to operate. The 5VCC POR has a rising threshold of 4.6V (typical) with 0.58V of hysteresis. After the 5VCC voltage exceeds its rising Power-On-Reset (POR) voltage threshold, the IC starts a start-up process and then ramps up the output voltage to the setting of output voltage. The 5VCC POR signal resets the fault latch, set by the under-voltage or over-current event, when the signal is at a low level.



When soft-start is initiated, the internal 10μA current source starts to charge the capacitor. When the soft-start voltage across the soft-start capacitor reaches the enabled threshold about 0.8V ( $V_{SS\_VT}$ ), the internal reference starts to rise and follows the soft-start voltage with converter operating at 150k/300k/400kHz PWM switching frequency. When output voltage rises to 87.5% of the regulation voltage, the power-ok is enabled. The soft-start time (from the moment of enabling the IC to the moment when  $V_{POK}$  goes high) can be expressed as the following equation:

$$T_{SS} = \frac{C_{SS} \times (V_{SS\_VT} + V_{REF} \times 0.875)}{I_{SS}}$$

where

$C_{SS}$  = external soft-start capacitor

$V_{SS\_VT}$  = internal soft-start threshold voltage, is about 0.8V

$V_{REF}$  = 0.6V or the voltage on the REFIN/EN pin

$I_{SS}$  = soft-start current = 10μA

During soft-start stage, the under-voltage protection is inhibited; however, the over-voltage and over-current protection functions are enabled. If the output capacitor has residue voltage before start-up, both lower and upper MOSFETs are in off-state until the internal soft-start voltage equals to the FB pin voltage. This will ensure the output voltage starts from its existing voltage level.

### Reference Voltage Selection and Shutdown Control

The APW7098 features a reference selection function to use either internal 0.6V or external reference voltage. During the beginning of soft-start, the voltage on

## Function Description (Cont.)

### Reference Voltage Selection and Shutdown Control (Cont.)

REFIN/EN pin determines which reference voltage is used. If this REFIN/EN pin is driven by an external voltage ranged from 0.4V to 2V, the IC uses the  $V_{REFIN/EN}$  voltage as reference voltage of the converter with soft-start control. If external reference is not available, connect this pin to 5VCC for internal 0.6V reference used. Once the internal or external reference is selected, the reference source is latched. Cycling the POR signal resets the latch.

The other function of REFIN/EN pin is used to enable or shut off the IC. Pulling the  $V_{REFIN/EN}$  voltage below 0.4V (typical) shuts down the two-phase PWM controller. In the shutdown mode, the two-phase UGATE and LGATE signals are pulled to PHASE and PGND respectively, the output is floating.

### Operation Phase Selection

The MODE pin programs single- or two- phase operation. It has a typical value for rising threshold of 0.8V,  $V_{MODE\_THR}$ , with 0.16V of hysteresis (0.64V),  $V_{MODE\_THF}$ . When the MODE pin voltage is higher than  $V_{MODE\_THR}$ , the device operates in single-phase; when the MODE pin voltage is lower than  $V_{MODE\_THF}$  and  $V_{IN2}$  supply voltage is above approximate 4V, the device operates in two-phase operation. This function makes the APW7098 ideally suitable for dual power input applications like PCIE interfaced graphic cards.

The figure 2 shows the power sources of the two channels. The input power of PWM1 converter is supplied by PCIE bus power and the input power of PWM2 converter is supplied by an external power. If the input power connector of PWM2 converter is not plugged into the socket before start-up, the internal  $V_{IN2}$  sensing circuit can sense the absence of  $V_{IN2}$  and set the IC to operate in single-phase mode with PWM2 disabled. When the IC operates in two-phase mode, it can switch the operating mode from two-phase to single-phase operation. Once operating in single-phase mode, the operation mode is latched. It is required to toggle SS, REFIN/EN, or 5VCC pin to reset the IC.

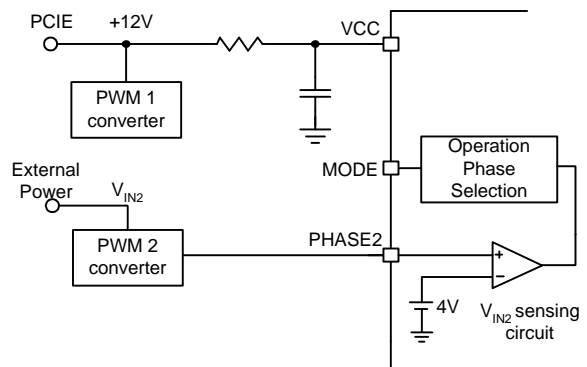


Figure 2.  $V_{IN2}$  Sensing Circuit

### Over-Voltage Protection (OVP)

The over-voltage protection function monitors the output voltage through the FB pin. When the FB voltage increases over 125% of the reference voltage ( $V_R$ ) due to the high-side MOSFET failure or other reasons, the over-voltage protection comparator designed with a 2 $\mu$ s noise filter will force the low-side MOSFET gate drivers high. This action actively pulls down the output voltage and eventually attempts to trigger the over-current shutdown of an ATX power supply. As soon as the output voltage is within regulation, the OVP comparator is disengaged. The chip will restore its normal operation. When the OVP occurs, the REFOUT/POK will drop to low as well.

This OVP scheme only clamps the voltage overshoot, and does not invert the output voltage when otherwise activated with a continuously high output from low-side MOSFETs driver, which is a common problem for OVP schemes with a latch.

### Under-Voltage Protection (UVP)

In the process of operation, when a short-circuit occurs, the output voltage will drop quickly. Before the over-current protection responds, the output voltage will fall out of the required regulation range. The under-voltage protection continually monitors the  $V_{FB}$  voltage after soft-start is completed. If a load step is strong enough to pull the output voltage lower than the under-voltage threshold, the IC shuts down converter's output. Cycling the 5VCC POR or REFIN/EN signal resets the fault latch and starts a start-up process. The under-voltage threshold is 50% of the nominal output voltage. The under-voltage comparator has a built-in 2 $\mu$ s noise filter to prevent the chips from wrong UVP shutdown being caused by noise.

## Function Description (Cont.)

### Over-Current Protection (OCP)

Figure 3 shows the circuit of sensing inductor current. Connecting a series resistor ( $R_s$ ) and a capacitor ( $C_s$ ) network in parallel with the inductor and measuring the voltage ( $V_c$ ) across the capacitor can sense the inductor current.

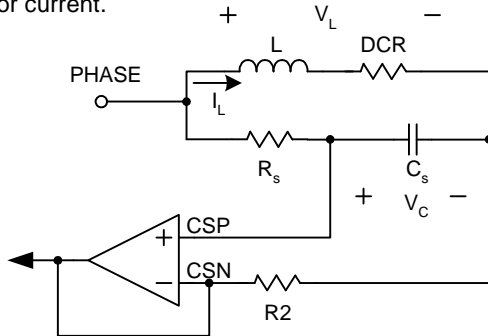


Figure 3. Illustration of Inductor Current Sensing Circuit

The equations of the sensing network are:

$$V_L(s) = I_L(s) \times (SL + DCR)$$

$$V_c(s) = V_L(s) \times \frac{1}{1 + SRsCs} = \frac{I_L(s) \times (SL + DCR)}{1 + SRsCs}$$

Take

$$R_sCs = \frac{L}{DCR}$$

for example, if the above equation is true, the voltage across the capacitor  $C_s$  is equal to voltage drop across the inductor DCR, and the voltage  $V_c$  is proportional to the current  $I_L$ . The sensing current through the resistor  $R2$  can be expressed as the following equation:

$$I_{CS} = \frac{I_L \times DCR}{R2}$$

where

- $I_{CS}$  is the sensed current
- $I_L$  is the inductor current
- DCR is the inductor resistance
- $R2$  is the sense resistor

The APW7098 is a two-phase PWM controller; therefore, the IC has two sensed current parts,  $I_{CS1}$  and  $I_{CS2}$ . When  $I_{CS1}$  plus  $I_{CS2}$  is greater than  $120\mu A$ , the over current occurs. In over-current protection, the IC shuts off the converter and then initials a new soft-start process. After 3 over-current events are counted, the device turns off both high-side and low-side MOSFETs and the converter's output is latched to be floating.

### Current Sharing

The APW7098 uses inductor's DCRs and external networks to sense the both currents flowing through the inductors of the PWM1 and PWM2 channels. The current sharing circuit, with closed-loop control, uses the sensed currents to adjust the two-phase inductor currents. For example, if the sensed current of PWM1 is bigger than PWM2, the duty of PWM1 will decrease and the duty of PWM2 will increase. Then, the device will reduce  $I_{L1}$  current and increase  $I_{L2}$  current for current sharing.

### DROOP

In some high current applications, a requirement on precisely controlled output impedance is imposed. This dependence of output voltage on load current is often termed droop regulation.

As shown in figure 4, the droop control block generates a voltage through external resistor  $R_{DROOP}$  and then set the droop voltage. The droop voltage,  $V_{DROOP}$ , is proportional to the total current in two channels. As shown in the following equation:

$$V_{DROOP} = 0.05 \times [(I_{CS1} + I_{CS2}) \times R_{DROOP}]$$

The  $V_{DROOP}$  voltage is used the regulator to adjust the output voltage, therefore, it is equal to the reference voltage minus the droop voltage.

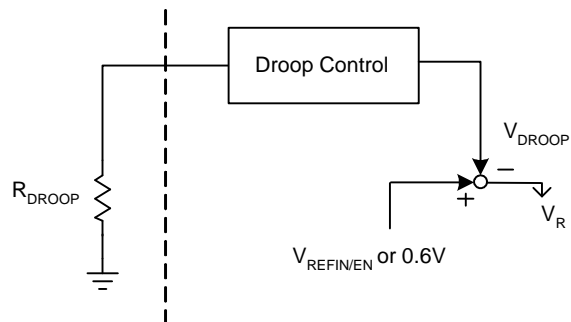


Figure 4. Illustration of Droop Setting Function

## Function Description (Cont.)

### Over-Temperature Protection (OTP)

When the junction temperature increases above the rising threshold temperature  $T_{OTR}$ , the IC will enter the over-temperature protection state that suspends the PWM, which forces the LGATE and UGATE gate drivers to output low voltages and turns off the 5VCC linear regulator output. The thermal sensor allows the converters to start a start-up process and regulate the output voltage again after the junction temperature cools by 50°C. The OTP is designed with a 50°C hysteresis to lower the average  $T_J$  during continuous thermal overload conditions, which increases lifetime of the APW7098.

## Application Information

### Output Voltage Setting

The output voltage is adjustable from 0.6V to 2.5V with a resistor-divider connected with FB, AGND, and converter's output. Using 1% or better resistors for the resistor-divider is recommended. The output voltage is determined by:

$$V_{OUT} = 0.6 \times \left( 1 + \frac{R_{TOP}}{R_{GND}} \right)$$

Where 0.6 is the reference voltage,  $R_{TOP}$  is the resistor connected from converter's output to FB, and  $R_{GND}$  is the resistor connected from FB to the the AGND. Suggested  $R_{GND}$  is in the range from 1K to 20kΩ. To prevent stray pickup, locate resistors  $R_{TOP}$  and  $R_{GND}$  close to the APW7098.

### PWM Compensation

The output LC filter of a step down converter introduces a double pole, which contributes with -40dB/decade gain slope and 180 degrees phase shift in the control loop. A compensation network among COMP, FB, and  $V_{OUT}$  should be added. The compensation network is shown in Figure 8. The output LC filters consists of the output inductors and output capacitors. For two-phase convertor, when assuming  $V_{IN1}=V_{IN2}=V_{IN}$ ,  $L1=L2=L$ , the transfer function of the LC filter is given by:

$$GAIN_{LC} = \frac{1 + s \times ESR \times C_{OUT}}{s^2 \times \frac{1}{2} L \times C_{OUT} + s \times ESR \times C_{OUT} + 1}$$

The poles and zero of this transfer functions are:

$$F_{LC} = \frac{1}{2 \times \pi \times \sqrt{\frac{1}{2} L \times C_{OUT}}}$$

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

The  $F_{LC}$  is the double-pole frequency of the two-phase LC filters, and  $F_{ESR}$  is the frequency of the zero introduced by the ESR of the output capacitors.

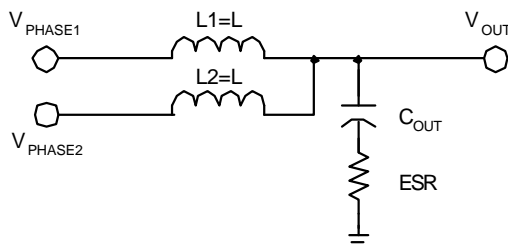


Figure 5. The Output LC Filter

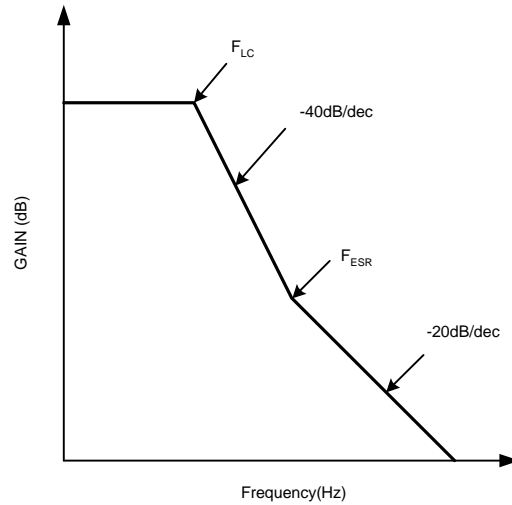


Figure 6. Frequency Resopnse of the LC filters

The PWM modulator is shown in figure 7. The input is the output of the error amplifier and the output is the PHASE node. The transfer function of the PWM modulator is given by :

$$GAIN_{PWM} = \frac{V_{IN}}{\Delta V_{OSC}}$$

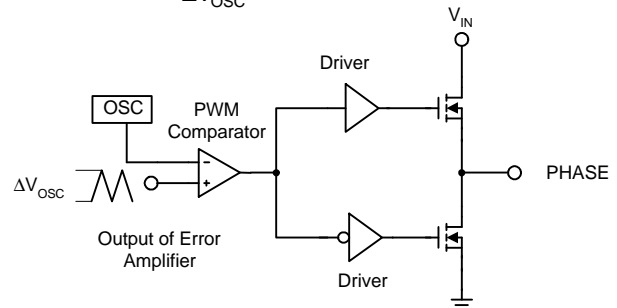


Figure 7. The PWM Modulator

The compensation network is shown in figure 8. It provides a close loop transfer function with the highest zero crossover frequency and sufficient phase margin.

The transfer function of error amplifier is given by :

$$GAIN_{AMP} = \frac{V_{COMP}}{V_{OUT}} = \frac{\frac{1}{sC1} // \left( R2 + \frac{1}{sC2} \right)}{R1 // \left( R3 + \frac{1}{sC3} \right)}$$

$$= \frac{R1+R3}{R1 \times R3 \times C1} \times \left( s + \frac{1}{R2 \times C2} \right) \times \left( s + \frac{1}{(R1+R3) \times C3} \right)$$

$$= \frac{R1+R3}{R1 \times R3 \times C1} \times \left( s + \frac{1}{R2 \times C2} \right) \times \left( s + \frac{1}{R3 \times C3} \right)$$

## Application Information (Cont.)

### PWM Compensation (Cont.)

The pole and zero frequencies of the transfer function are:

$$F_{Z1} = \frac{1}{2 \times \pi \times R2 \times C2}$$

$$F_{Z2} = \frac{1}{2 \times \pi \times (R1 + R3) \times C3}$$

$$F_{P1} = \frac{1}{2 \times \pi \times R2 \times \left( \frac{C1 \times C2}{C1 + C2} \right)}$$

$$F_{P2} = \frac{1}{2 \times \pi \times R3 \times C3}$$

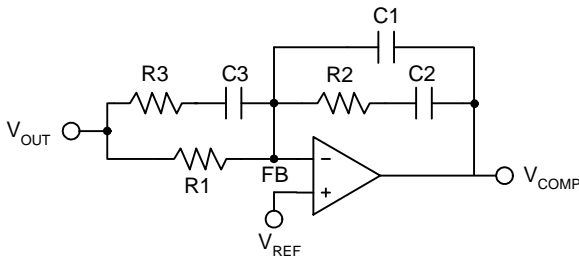


Figure 8. Compensation Network

The closed loop gain of the converter can be written as:

$$GAIN_{LC} \times GAIN_{PWM} \times GAIN_{AMP}$$

Figure 9. shows the asymptotic plot of the closed loop converter gain, and the following guidelines will help to design the compensation network. Using the below guidelines should give a compensation similar to the curve plotted. A stable closed loop has a -20dB/decade slope and a phase margin greater than 45 degree.

1. Choose a value for R1, usually between 1K and 5K.

2. Select the desired zero crossover frequency

$$F_o = (1/5 \sim 1/10) \times F_{SW}$$

Use the following equation to calculate R2:

$$R2 = \frac{\Delta V_{OSC}}{V_{IN}} \times \frac{F_o}{F_{LC}} \times R1$$

3. Place the first zero  $F_{Z1}$  before the output LC filter double pole frequency  $F_{LC}$ .

$$F_{Z1} = 0.75 \times F_{LC}$$

Calculate the C2 by the equation:

$$C2 = \frac{1}{2 \times \pi \times R2 \times F_{LC} \times 0.75}$$

4. Set the pole at the ESR zero frequency  $F_{ESR}$ :

$$F_{P1} = F_{ESR}$$

Calculate the C1 by the following equation:

$$C1 = \frac{C2}{2 \times \pi \times R2 \times C2 \times F_{ESR} - 1}$$

5. Set the second pole  $F_{P2}$  at the half of the switching frequency and also set the second zero  $F_{Z2}$  at the output LC filter double pole  $F_{LC}$ . The compensation gain should not exceed the error amplifier open loop gain, check the compensation gain at  $F_{P2}$  with the capabilities of the error amplifier.

$$F_{P2} = 0.5 \times F_{SW}$$

$$F_{Z2} = F_{LC}$$

Combine the two equations will get the following component calculations:

$$R3 = \frac{R1}{\frac{F_{SW}}{2 \times F_{LC}} - 1}$$

$$C3 = \frac{1}{\pi \times R3 \times F_{SW}}$$

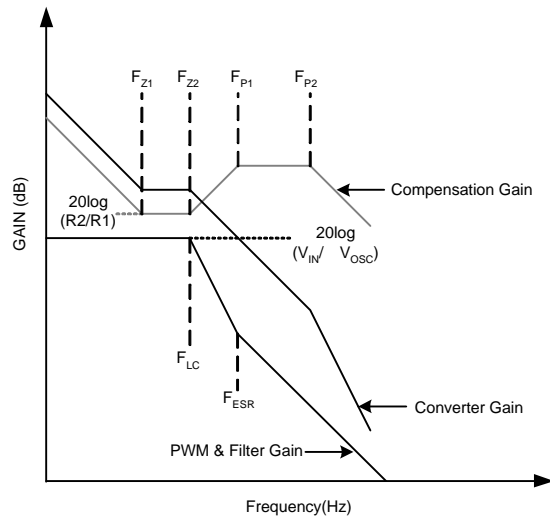


Figure 9. Converter Gain and Frequency

### Output Inductor Selection

The duty cycle (D) of a buck converter is the function of the input voltage and output voltage. Once an output voltage is fixed, it can be written as:

## Application Information (Cont.)

### Output Inductor Selection (Cont.)

$$D = \frac{V_{OUT}}{V_{IN}}$$

For two-phase converter, the inductor value (L) determines the sum of the two inductor ripple currents,  $\Delta I_{P-P}$ , and affects the load transient response. Higher inductor value reduces the output capacitors' ripple current and induces lower output ripple voltage. The ripple current can be approximated by:

$$\Delta I_{P-P} = \frac{V_{IN} - 2V_{OUT}}{F_{SW} \times L} \times \frac{V_{OUT}}{V_{IN}}$$

Where  $F_{SW}$  is the switching frequency of the regulator. Although the inductor value and frequency are increased and the ripple current and voltage are reduced, a tradeoff exists between the inductor's ripple current and the regulator load transient response time.

A smaller inductor will give the regulator a faster load transient response at the expense of higher ripple current. Increasing the switching frequency ( $F_{SW}$ ) also reduces the ripple current and voltage, but it will increase the switching loss of the MOSFETs and the power dissipation of the converter. The maximum ripple current occurs at the maximum input voltage. A good starting point is to choose the ripple current to be approximately 30% of the maximum output current. Once the inductance value has been chosen, select an inductor that is capable of carrying the required peak current without going into saturation. In some types of inductors, especially core that is made of ferrite, the ripple current will increase abruptly when it saturates. This results in a larger output ripple voltage.

### Output Capacitor Selection

Output voltage ripple and the transient voltage deviation are factors that have to be taken into consideration when selecting output capacitors. Higher capacitor value and lower ESR reduce the output ripple and the load transient drop. Therefore, selecting high performance low ESR capacitors is recommended for switching regulator applications. In addition to high frequency noise related to MOSFET turn-on and turn-off, the output voltage ripple includes the capacitance voltage drop  $\Delta V_{C_{OUT}}$  and ESR voltage drop  $\Delta V_{ESR}$

caused by the AC peak-to-peak sum of the inductor's current. The ripple voltage of output capacitors can be represented by:

$$\Delta V_{C_{OUT}} = \frac{\Delta I_{P-P}}{8 \times C_{OUT} \times F_{SW}}$$

$$\Delta V_{ESR} = \Delta I_{P-P} \times R_{ESR}$$

These two components constitute a large portion of the total output voltage ripple. In some applications, multiple capacitors have to be paralleled to achieve the desired ESR value. If the output of the converter has to support another load with high pulsating current, more capacitors are needed in order to reduce the equivalent ESR and suppress the voltage ripple to a tolerable level. A small decoupling capacitor in parallel for bypassing the noise is also recommended, and the voltage rating of the output capacitors are also must be considered.

To support a load transient that is faster than the switching frequency, more capacitors are needed for reducing the voltage excursion during load step change. For getting same load transient response, the output capacitance of two-phase converter only needs around half of output capacitance of single-phase converter.

Another aspect of the capacitor selection is that the total AC current going through the capacitors has to be less than the rated RMS current specified on the capacitors in order to prevent the capacitor from overheating.

### Input Capacitor Selection

Use small ceramic capacitors for high frequency decoupling and bulk capacitors to supply the surge current needed each time high-side MOSFET turns on. Place the small ceramic capacitors physically close to the MOSFETs and between the drain of high-side MOSFET and the source of low-side MOSFET.

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

## Application Information (Cont.)

### Input Capacitor Selection (Cont.)

RMS current of the bulk input capacitor is roughly calculated as the following equation :

$$I_{RMS} = \frac{I_{OUT}}{2} \times \sqrt{2D \cdot (1 - 2D)}$$

For a through hole design, several electrolytic capacitors may be needed. For surface mount design, solid tantalum capacitors can be used, but caution must be exercised with regard to the capacitor surge current rating.

### MOSFET Selection

The APW7098 requires two N-Channel power MOSFETs on each phase. These should be selected based upon  $R_{DS(ON)}$ , gate supply requirements, and thermal management requirements.

In high-current applications, the MOSFET power dissipation, package selection, and heatsink are the dominant design factors. The power dissipation includes two loss components, conduction loss, and switching loss. The conduction losses are the largest component of power dissipation for both the high-side and the low-side MOSFETs. These losses are distributed between the two MOSFETs according to duty factor (see the equations below). Only the high-side MOSFET has switching losses since the low-side MOSFETs body diode or an external Schottky rectifier across the lower MOSFET clamps the switching node before the synchronous rectifier turns on. These equations assume linear voltage-current transitions and do not adequately model power loss due the reverse-recovery of the low-side MOSFET body diode. The gate-charge losses are dissipated by the APW7098 and don't heat the MOSFETs. However, large gate-charge increases the switching interval,  $t_{sw}$  which increases the high-side MOSFET switching losses. Ensure that all MOSFETs are within their maximum junction temperature at high ambient temperature by calculating the temperature rise according to package thermal-resistance specifications. A separate heatsink may be necessary depending upon MOSFET power, package type, ambient temperature and air flow.

For the high-side and low-side MOSFETs, the losses are approximately given by the following equations:

$$P_{high-side} = I_{OUT}^2 (1 + TC)(R_{DS(ON)})D + (0.5)(I_{OUT})(V_{IN})(t_{sw})F_{sw}$$

$$P_{low-side} = I_{OUT}^2 (1 + TC)(R_{DS(ON)})(1 - D)$$

where

$I_{OUT}$  is the load current

TC is the temperature dependency of  $R_{DS(ON)}$

$F_{sw}$  is the switching frequency

$t_{sw}$  is the switching interval

D is the duty cycle

Note that both MOSFETs have conduction losses while the high-side MOSFET includes an additional transition loss. The switching interval,  $t_{sw}$ , is the function of the reverse transfer capacitance  $C_{RSS}$ . The (1+TC) term is a factor in the temperature dependency of the  $R_{DS(ON)}$  and can be extracted from the " $R_{DS(ON)}$  vs. Temperature" curve of the power MOSFET.

### Layout Consideration

In any high switching frequency converter, a correct layout is important to ensure proper operation of the regulator. With power devices switching at higher frequency, the resulting current transient will cause voltage spike across the interconnecting impedance and parasitic circuit elements. As an example, consider the turn-off transition of the PWM MOSFET. Before turn-off condition, the MOSFET is carrying the full load current. During turn-off, current stops flowing in the MOSFET and is freewheeling by the low side MOSFET and parasitic diode. Any parasitic inductance of the circuit generates a large voltage spike during the switching interval. In general, using short and wide printed circuit traces should minimize interconnecting impedances and the magnitude of voltage spike. Besides, signal and power grounds are to be kept separating and finally combined using ground plane construction or single point grounding. The best tie-point between the signal ground and the power ground is at the negative side of the output capacitor on each channel, where there is less noise. Noisy traces beneath the IC are not recommended. Figure 10. illustrates the layout, with bold lines indicating high current paths; these traces must be short and wide. Components along the bold lines should be placed close together. Below is a checklist for your layout:



## Application Information (Cont.)

### Layout Consideration (Cont.)

- Keep the switching nodes (UGATEx, LGATEx, BOOTx, and PHASEx) away from sensitive small signal nodes since these nodes are fast moving signals. Therefore, keep traces to these nodes as short as possible and there should be no other weak signal traces in parallel with these traces on any layer.
- The signals going through these traces have both high dv/dt and high di/dt with high peak charging and discharging current. The traces from the gate drivers to the MOSFETs (UGATEx and LGATEx) should be short and wide.
- Place the source of the high-side MOSFET and the drain of the low-side MOSFET as close as possible. Minimizing the impedance with wide layout plane between the two pads reduces the voltage bounce of the node. In addition, the large layout plane between the drain of the MOSFETs ( $V_{IN}$  and PHASEx nodes) can get better heat sinking.
- For experiment result of accurate current sensing, the current sensing components are suggested to place close to the inductor part. To avoid the noise interference, the current sensing trace should be away from the noisy switching nodes.
- Decoupling capacitors, the resistor-divider, and boot capacitor should be close to their pins. (For example, place the decoupling ceramic capacitor close to the drain of the high-side MOSFET as close as possible).
- The input bulk capacitors should be close to the drain of the high-side MOSFET, and the output bulk capacitors should be close to the loads. The input capacitor's ground should be close to the grounds of the output capacitors and low-side MOSFET.
- Locate the resistor-divider close to the FB pin to minimize the high impedance trace. In addition, FB pin traces can't be close to the switching signal traces (UGATEx, LGATEx, BOOTx, and PHASEx).

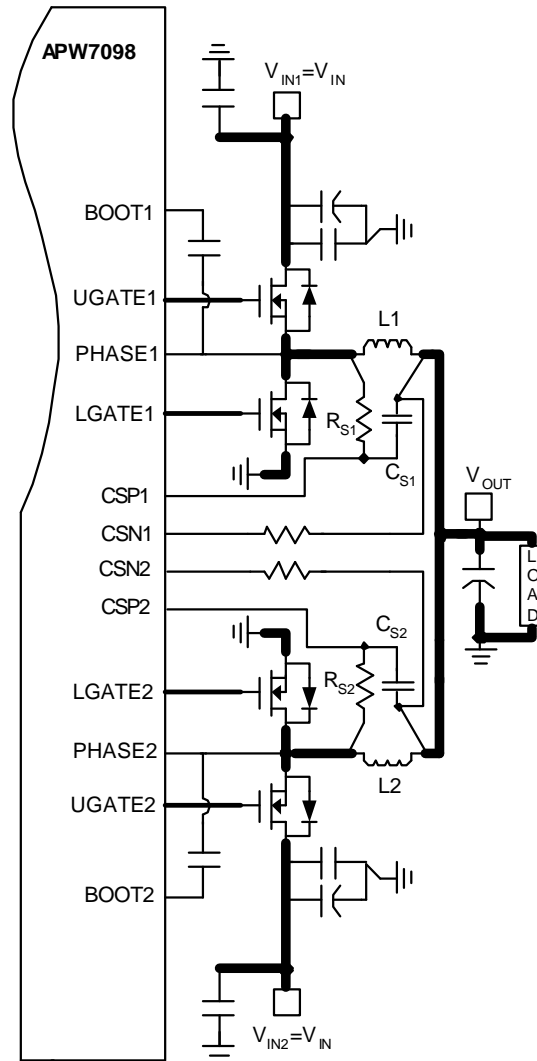
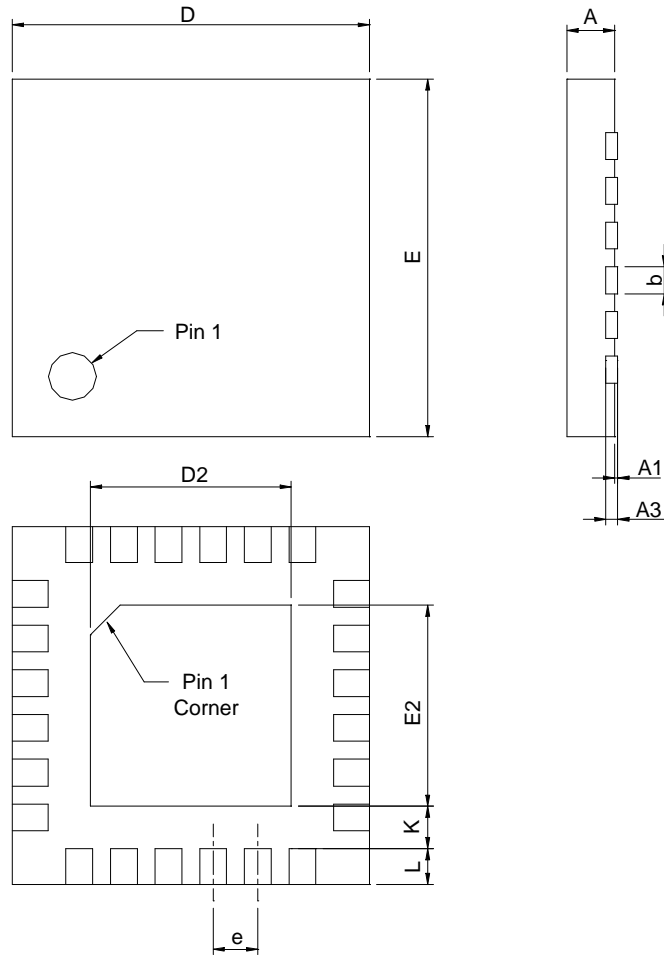


Figure 10. Layout Guidelines

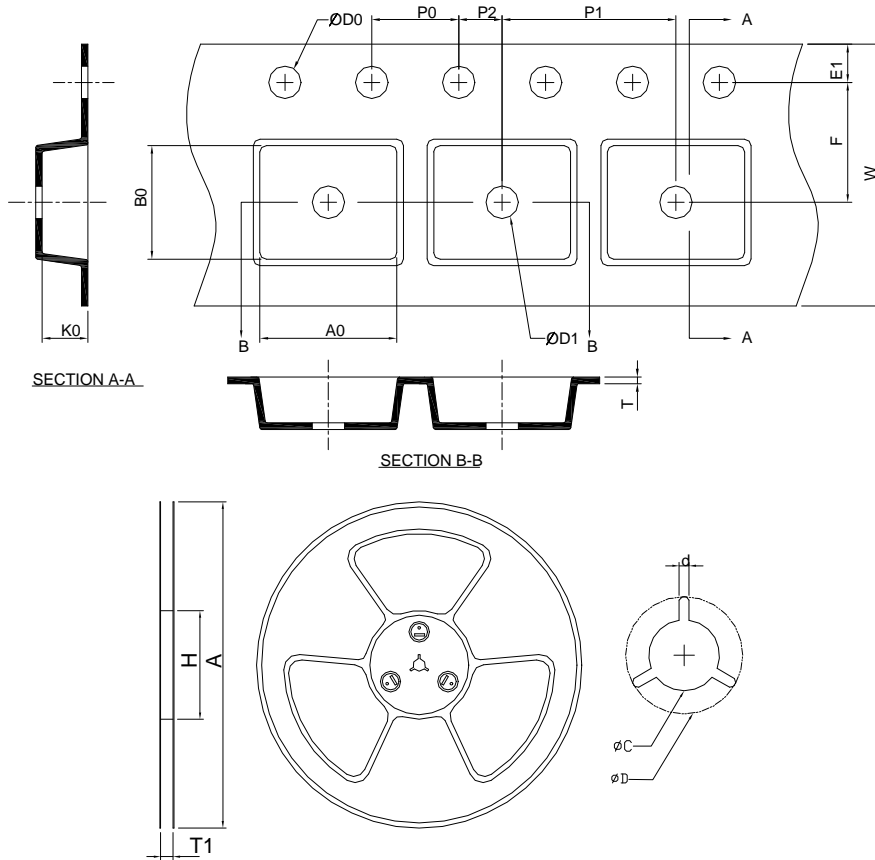
Package Information

QFN4x4-24A



DIMENSIONS	QFN4x4-24A			
	MILLIMETERS		INCHES	
	MIN.	MAX.	MIN.	MAX.
A	0.80	1.00	0.031	0.039
A1	0.00	0.05	0.000	0.002
A3	0.20 REF		0.008 REF	
b	0.18	0.30	0.008	0.012
D	3.90	4.10	0.154	0.161
D2	2.00	2.50	0.079	0.098
E	3.90	4.10	0.154	0.161
E2	2.00	2.50	0.079	0.098
e	0.50 BSC		0.020 BSC	
L	0.35	0.45	0.014	0.018
K	0.20		0.008	

Carrier Tape & Reel Dimensions



Application	A	H	T1	C	d	D	W	E1	F
QFN4x4-24A	330.0 ±0.00	50 MIN.	12.4+2.00 -0.00	13.0+0.50 -0.20	1.5 MIN.	20.2 MIN.	12.0 ±0.30	1.75 ±0.10	5.5 ±0.05
	P0	P1	P2	D0	D1	T	A0	B0	K0
	4.0 ±0.10	8.0 ±0.10	2.0 ±0.05	1.5+0.10 -0.00	1.5 MIN.	0.6+0.00 -0.40	4.30 ±0.20	4.30 ±0.20	1.30 ±0.20

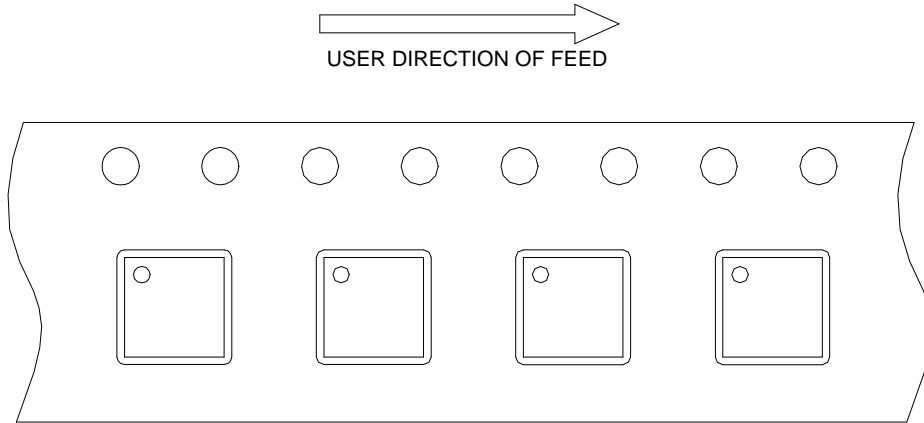
(mm)

Devices Per Unit

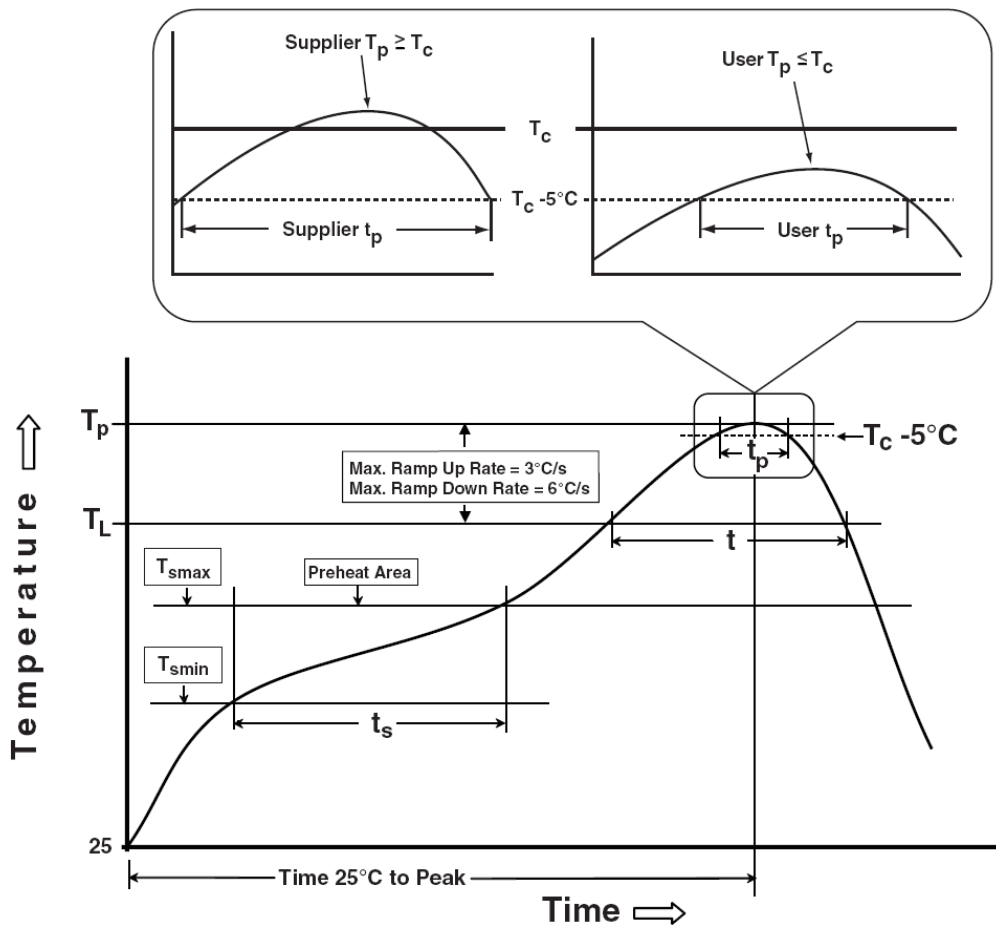
Package Type	Unit	Quantity
QFN4x4-24A	Tape & Reel	3000

### Taping Direction Information

QFN4x4-24A



### Classification Profile



### Classification Reflow Profiles

Profile Feature	Sn-Pb Eutectic Assembly	Pb-Free Assembly
<b>Preheat &amp; Soak</b>		
Temperature min ( $T_{smin}$ )	100 °C	150 °C
Temperature max ( $T_{smax}$ )	150 °C	200 °C
Time ( $T_{smin}$ to $T_{smax}$ ) ( $t_s$ )	60-120 seconds	60-120 seconds
Average ramp-up rate ( $T_{smax}$ to $T_p$ )	3 °C/second max.	3°C/second max.
Liquidous temperature ( $T_L$ )	183 °C	217 °C
Time at liquidous ( $t_L$ )	60-150 seconds	60-150 seconds
Peak package body Temperature ( $T_p$ )*	See Classification Temp in table 1	See Classification Temp in table 2
Time ( $t_p$ )** within 5°C of the specified classification temperature ( $T_c$ )	20** seconds	30** seconds
Average ramp-down rate ( $T_p$ to $T_{smax}$ )	6 °C/second max.	6 °C/second max.
Time 25°C to peak temperature	6 minutes max.	8 minutes max.
* Tolerance for peak profile Temperature ( $T_p$ ) is defined as a supplier minimum and a user maximum.		
** Tolerance for time at peak profile temperature ( $t_p$ ) is defined as a supplier minimum and a user maximum.		

Table 1. SnPb Eutectic Process – Classification Temperatures ( $T_c$ )

Package Thickness	Volume mm <sup>3</sup> <350	Volume mm <sup>3</sup> ≥350
<2.5 mm	235 °C	220 °C
≥2.5 mm	220 °C	220 °C

Table 2. Pb-free Process – Classification Temperatures ( $T_c$ )

Package Thickness	Volume mm <sup>3</sup> <350	Volume mm <sup>3</sup> 350-2000	Volume mm <sup>3</sup> >2000
<1.6 mm	260 °C	260 °C	260 °C
1.6 mm – 2.5 mm	260 °C	250 °C	245 °C
≥2.5 mm	250 °C	245 °C	245 °C

### Reliability Test Program

Test item	Method	Description
SOLDERABILITY	JESD-22, B102	5 Sec, 245°C
HOLT	JESD-22, A108	1000 Hrs, Bias @ 125°C
PCT	JESD-22, A102	168 Hrs, 100%RH, 2atm, 121°C
TCT	JESD-22, A104	500 Cycles, -65°C~150°C
HBM	MIL-STD-883-3015.7	VHBM 2KV
MM	JESD-22, A115	VMM 200V
Latch-Up	JESD 78	10ms, 1 <sub>tr</sub> 100mA

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