TPS53355 高效 30A 同步降压 SWIFT™ 转换器，具有 Eco-mode™

1 特性

- 最高效率达 96%
- 转换输入电压范围：1.5V 至 15V
- 漏极电源电压 (VDD) 输入电压范围：4.5V 至 25V
- 输出电压范围：0.6V 至 5.5V
- 5V 低压降 (LDO) 输出
- 支持单电源轨出入
- 带有 30A 持续输出电流的集成型功率金属氧化物半导体场效应晶体管 (MOSFET)
- 自动跳跃 Eco-mode™，用于实现轻负载效率
- < 10µA 关断电流
- 针对快速瞬态响应的 D-Cap+™ 模式
- 可通过外部电阻在 250kHz 至 1MHz 范围内选择开关频率
- 可选的自动跳跃或者只支持脉宽调制 (PWM) 运行
- 内置 1% 0.6V 基准电压。
- 0.7ms，1.4ms，2.8ms 和 5.6ms 可选内部电压伺服系统软启动
- 集成升压开关
- 预充电启动能力
- 支持可调节的过流限制和温度补偿
- 过压、欠压、欠压锁定 (UVLO) 和过热保护
- 支持所有陶瓷输出电容
- 开漏电源良好指示
- 组装有 NexFET™电源块技术
- 22 引脚四方扁平无引脚 (QFN) 封装，采用 PowerPAD™
- 如需 SWIFT ™电源产品文档，请访问 http://www.ti.com.cn/ww/analog/swift
- 可选“绿色环保”（符合 RoHS 标准）

2 应用

- 服务器和存储
- 工作站和台式机
- 电信基础设施

3 说明

TPS53355 是一款集成有金属氧化物半导体场效应晶体管 (MOSFET) 的 D-CAP™模式、30A 同步开关。它的设计目标是简单易用、减少外部组件，以及适用于空间受限的电源系统。

该器件 采用 5mΩ/2.0mΩ 集成 MOSFET，精度为 1%，基准电压为 0.6V，并具备一个集成升压开关。富有竞争力的 特性 包括：1.5V 至 15V 宽转换输入电压范围、所用外部组件极少、针对超快瞬变的 D-CAP™ 模式控制、自动跳跃模式运行、内部软启动控制、可选频率并且无需补偿。

转换输入电压范围为 1.5V 至 15V，电源电压范围为 4.5V 至 25V，输出电压范围为 0.6V 至 5.5V。

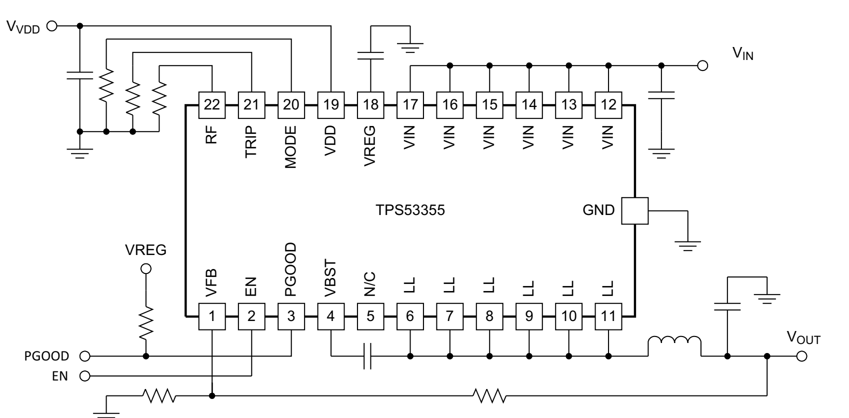
该器件采用 5mm × 6mm、22 引脚 QFN 封装，额定工作温度范围为 –40°C 至 85°C。

器件信息(1)

器件型号	封装	封装尺寸（标称值）
TPS53355	LSON-CLIP (22)	6.00mm × 5.00mm

(1) 如需了解所有可用封装，请参阅数据表末尾的可订购产品附录。

典型应用



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4 修订历史记录

注：之前版本的页码可能与当前版本有所不同。

Changes from Revision C (February 2016) to Revision D — **Page**

- 添加了*特性*：可选"绿色环保"（符合 RoHS 标准） ... 1
- Added the VQP package to the *Thermal Infomation* ... 6
- From: a SC5026-1R0 inductor is used. To: a 744355182 inductor is used. ... 8
- Changed Figure 32 and Figure 33 ... 12
- *5-V LDO and VREG Start-Up*, Changed the NOTE From: "The 5-V LDO is not controlled" To: "The 5-V LDO is controlled" ... 15
- Changed 250 µs To ~550 µs in Figure 34 ... 16

Changes from Revision B (January 2014) to Revision C — **Page**

- 将数据表标题从"具有 Eco-mode™ 的 TPS53355 高效 30A 同步降压转换器"更改为"具有 Eco-mode™ 的 TPS53355 高效 30A 同步降压 SWIFT™ 转换器" ... 1
- 添加了*特性*："有关 SWIFT™ 电源产品文档，…" ... 1

Changes from Revision A (September 2012) to Revision B — **Page**

- 已添加 *引脚配置和功能* 部分、*ESD 额定值* 表、*特性 说明* 部分、*器件功能模式*、*应用和实施* 部分、*电源相关建议* 部分、*布局* 部分、*器件和文档支持* 部分以及*机械、封装和可订购信息* 部分 ... 1

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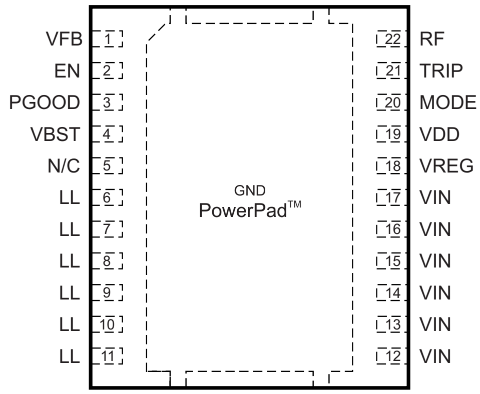
Changes from Original (August 2011) to Revision A **Page**

- 已更改 转换输入电压范围从 "3V" 改为 "15V" ... 1
- Changed VIN input voltage range minimum from "3 V" to "1.5 V" ... 4
- Changed typographical error in THERMAL INFORMATION table ... 5
- Changed VIN (main supply) input voltage range minimum from "3 V' to "1.5 V" in *Recommended Operating Conditions* ... 5
- Changed VIN pin power conversion input minimum voltage from "3 V" to "1.5 V" in ELECTRICAL CHARACTERISTICS table ... 6
- Changed conversion input voltage range from "3 V" to "1.5" in *Overview* ... 14
- Added note to the *Functional Block Diagram* ... 15
- Changed "ripple injection capacitor" to "ripple injection resistor" in *Layout Guidelines* section ... 29

5 Pin Configuration and Functions

Package With PowerPad
22-Pins (LSON-CLIP)
Top View



(1) N/C = no connection

Pin Functions

PIN		I/O/P (1)	DESCRIPTION
NAME	NO		
EN	2	I	Enable pin. Typical turn-on threshold voltage is 1.2 V. Typical turn-off threshold is 0.95 V.
GND	—	—	Ground and thermal pad of the device. Use proper number of vias to connect to ground plane.
LL	6	B	Output of converted power. Connect this pin to the output Inductor.
	7		
	8		
	9		
	10		
	11		
MODE	20	I	Soft-start and Skip/CCM selection. Connect a resistor to select soft-start time using Table 3. The soft-start time is detected and stored into internal register during start-up.
N/C	5		No connect.
PGOOD	3	O	Open drain power good flag. Provides 1-ms start-up delay after VFB falls in specified limits. When VFB goes out of the specified limits PGOOD goes low after a 2-µs delay.
RF	22	I	Switching frequency selection. Connect a resistor to GND or VREG to select switching frequency using Table 1. The switching frequency is detected and stored during the startup.
TRIP	21	I	OCL detection threshold setting pin. I_{TRIP} = 10 µA at room temperature, 4700 ppm/°C current is sourced and set the OCL trip voltage as follows: $V_{OCL} = V_{TRIP}/32$ ($V_{TRIP} \le 2.4$ V, $V_{OCL} \le 75$ mV)
VBST	4	P	Supply input for high-side FET gate driver (boost terminal). Connect capacitor from this pin to LL node. Internally connected to VREG via bootstrap MOSFET switch.
VDD	19	P	Controller power supply input. VDD input voltage range is from 4.5 V to 25 V.
VFB	1	I	Output feedback input. Connect this pin to Vout through a resistor divider.
VIN	12	P	Conversion power input. VIN input voltage range is from 1.5 V to 15 V.
	13		
	14		
	15		
	16		
	17		
VREG	18	P	5-V low drop out (LDO) output. Supplies the internal analog circuitry and driver circuitry.

(1) I=Input, O=Output, B=Bidirectional, P=Supply

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6 Specifications

6.1 Absolute Maximum Ratings(1)

			MIN	MAX	UNIT
Input voltage	VIN (main supply)		–0.3	25	V
	VDD		–0.3	28	
	VBST		–0.3	32	
	VBST (with respect to LL)		–0.3	7	
	EN, TRIP, VFB, RF, MODE		–0.3	7	
Output voltage	LL	DC	–2	25	V
		Pulse < 20ns, E=5 µJ	–7	27	
	PGOOD, VREG		–0.3	7	
	GND		–0.3	0.3	
Source/Sink current	VBST		50		mA
Operating free-air temperature, T_A			–40	85	°C
Junction temperature, T_J			–40	150	
Lead temperature 1,6 mm (1/16 inch) from case for 10 seconds				300	
Storage temperature, T_{stg}			–55	150	°C

(1) Stresses beyond those listed under Absolute Maximum Ratings may cause permanent damage to the device. These are stress ratings only and functional operation of the device at these or any other conditions beyond those indicated under *Recommended Operating Conditions* is not implied. Exposure to absolute-maximum-rated conditions for extended periods may affect device reliability.

6.2 ESD Ratings

			VALUE	UNIT
$V_{(ESD)}$	Electrostatic discharge	Human-body model (HBM), per ANSI/ESDA/JEDEC JS-001(1)	±2000	V
		Charged-device model (CDM), per JEDEC specification JESD22-C101(2)	±500	

(1) JEDEC document JEP155 states that 500-V HBM allows safe manufacturing with a standard ESD control process.
(2) JEDEC document JEP157 states that 250-V CDM allows safe manufacturing with a standard ESD control process.

6.3 Recommended Operating Conditions

over operating free-air temperature range (unless otherwise noted)

		MIN	MAX	UNIT
Input voltage range	VIN (main supply)	1.5	15	V
	VDD	4.5	25	
	VBST	4.5	28	
	VBST (with respect to LL)	4.5	6.5	
	EN, TRIP, VFB, RF, MODE	–0.1	6.5	
Output voltage range	LL	–1	22	V
	PGOOD, VREG	–0.1	6.5	
Junction temperature range, T_J		–40	125	°C

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6.4 Thermal Infomation

THERMAL METRIC(1)		TPS53355 DQP 22 PINS	TPS53355 VQP 22 PINS	UNIT
θ_{JA}	Junction-to-ambient thermal resistance	27.2	27.2	°C/W
θ_{JCtop}	Junction-to-case (top) thermal resistance	17.1	17.1	°C/W
θ_{JB}	Junction-to-board thermal resistance	5.9	5.9	°C/W
ψ_{JT}	Junction-to-top characterization parameter	0.8	0.8	°C/W
ψ_{JB}	Junction-to-board characterization parameter	5.8	5.8	°C/W
θ_{JCbot}	Junction-to-case (bottom) thermal resistance	1.2	1.2	°C/W

(1) For more information about traditional and new thermal metrics, see the *Semiconductor and IC Package Thermal Metrics* application report.

6.5 Electrical Characteristics

Over recommended free-air temperature range, V_{VDD}= 12 V (unless otherwise noted)

PARAMETER		CONDITIONS	MIN	TYP	MAX	UNIT
SUPPLY CURRENT						
V_{VIN}	VIN pin power conversion input voltage		1.5		15	V
V_{VDD}	Supply input voltage		4.5		25.0	V
$I_{VIN(leak)}$	VIN pin leakage current	V_{EN} = 0 V			1	µA
I_{VDD}	VDD supply current	T_A = 25°C, No load, V_{EN} = 5 V, V_{VFB} = 0.630 V		420	590	µA
I_{VDDSDN}	VDD shutdown current	T_A = 25°C, No load, V_{EN} = 0 V			10	µA
INTERNAL REFERENCE VOLTAGE						
V_{VFB}	VFB regulation voltage	CCM condition(1)		0.600		V
V_{VFB}	VFB regulation voltage	T_A = 25°C	0.597	0.600	0.603	V
		0°C ≤ T_A ≤ 85°C	0.5952	0.600	0.6048	
		–40°C ≤ T_A ≤ 85°C	0.594	0.600	0.606	
I_{VFB}	VFB input current	V_{VFB} = 0.630 V, T_A = 25°C		0.01	0.20	µA
LDO OUTPUT						
V_{VREG}	LDO output voltage	0 mA ≤ I_{VREG} ≤ 30 mA	4.77	5.00	5.36	V
I_{VREG}	LDO output current(1)	Maximum current allowed from LDO			30	mA
V_{DO}	Low drop out voltage	V_{VDD} = 4.5 V, I_{VREG} = 30 mA			230	mV
BOOT STRAP SWITCH						
V_{FBST}	Forward voltage	$V_{VREG-VBST}$, I_F = 10 mA, T_A = 25°C		0.1	0.2	V
I_{VBSTLK}	VBST leakage current	V_{VBST} = 23 V, V_{SW} = 17 V, T_A = 25°C		0.01	1.50	µA
DUTY AND FREQUENCY CONTROL						
$t_{OFF(min)}$	Minimum off time	T_A = 25°C	150	260	400	ns
$t_{ON(min)}$	Minimum on time	V_{IN} = 17 V, V_{OUT} = 0.6 V, R_{RF} = 39 kΩ, T_A = 25 °C(1)		35		ns
SOFT START						
t_{SS}	Internal soft-start time from V_{OUT} = 0 V to 95% of V_{OUT}	R_{MODE} = 39 kΩ		0.7		ms
		R_{MODE} = 100 kΩ		1.4		
		R_{MODE} = 200 kΩ		2.8		
		R_{MODE} = 470 kΩ		5.6		
INTERNAL MOSFETS						
$R_{DS(on)H}$	High-side MOSFET on-resistance	T_A = 25°C		5.0		mΩ
$R_{DS(on)L}$	Low-side MOSFET on-resistance	T_A = 25°C		2.0		mΩ

(1) Ensured by design. Not production tested.

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Electrical Characteristics (continued)

Over recommended free-air temperature range, V_{VDD}= 12 V (unless otherwise noted)

PARAMETER		CONDITIONS	MIN	TYP	MAX	UNIT
POWERGOOD						
V_{THPG}	PG threshold	PG in from lower	92.5%	95.0%	98.5%	
		PG in from higher	107.5%	110.0%	112.5%	
		PG hysteresis	2.5%	5.0%	7.5%	
R_{PG}	PG transistor on-resistance		15	30	55	Ω
t_{PGDEL}	PG delay	Delay for PG in	0.8	1	1.2	ms
LOGIC THRESHOLD AND SETTING CONDITIONS						
V_{EN}	EN Voltage	Enable	1.8			V
		Disable			0.6	
I_{EN}	EN Input current	V_{EN} = 5 V			1.0	µA
f_{SW}	Switching frequency	R_{RF} = 0 Ω to GND, T_A = 25°C (2)	200	250	300	kHz
		R_{RF} = 187 kΩ to GND, T_A = 25°C (2)	250	300	350	
		R_{RF} = 619 kΩ, to GND, T_A = 25°C (2)	350	400	450	
		R_{RF} = Open, T_A= 25°C (2)	450	500	550	
		R_{RF} = 866 kΩ to VREG, T_A = 25°C (2)	580	650	720	
		R_{RF} = 309 kΩ to VREG, T_A = 25°C (2)	670	750	820	
		R_{RF} = 124 kΩ to VREG, T_A = 25°C (2)	770	850	930	
		R_{RF} = 0 Ω to VREG, T_A = 25°C (2)	880	970	1070	
PROTECTION: CURRENT SENSE						
I_{TRIP}	TRIP source current	V_{TRIP} = 1 V, T_A = 25°C	9.4	10.0	10.6	µA
TC_{ITRIP}	TRIP current temperature coefficient	On the basis of 25°C (1)		4700		ppm/°C
V_{TRIP}	Current limit threshold setting range	$V_{TRIP-GND}$	0.4		2.4	V
V_{OCL}	Current limit threshold	V_{TRIP} = 2.4 V	68.5	75.0	81.5	mV
		V_{TRIP} = 0.4 V	7.5	12.5	17.5	
V_{OCLN}	Negative current limit threshold	V_{TRIP} = 2.4 V	-315	-300	-285	mV
		V_{TRIP} = 0.4 V	-58	-50	-42	
V_{AZCADJ}	Auto zero cross adjustable range	Positive	3	15		mV
		Negative		–15	–3	
PROTECTION: UVP and OVP						
V_{OVP}	OVP trip threshold	OVP detect	115%	120%	125%	
t_{OVPDEL}	OVP propagation delay	VFB delay with 50-mV overdrive		1		µs
V_{UVP}	Output UVP trip threshold	UVP detect	65%	70%	75%	
t_{UVPDEL}	Output UVP propagation delay		0.8	1.0	1.2	ms
t_{UVPEN}	Output UVP enable delay	From enable to UVP workable	1.8	2.6	3.2	ms
UVLO						
V_{UVVREG}	VREG UVLO threshold	Wake up	4.00	4.20	4.33	V
		Hysteresis		0.25		
THERMAL SHUTDOWN						
T_{SDN}	Thermal shutdown threshold	Shutdown temperature (1)		145		°C
		Hysteresis (1)		10		

(2) Not production tested. Test condition is V_{IN}= 12 V, V_{OUT}= 1.1 V, I_{OUT} = 10 A using application circuit shown in Figure 47.

6.6 Typical Characteristics

For V_{OUT} = 5 V, a 744355182 inductor is used. For 1 ≤ V_{OUT} ≤ 3.3 V, a PA0513.441 inductor is used.

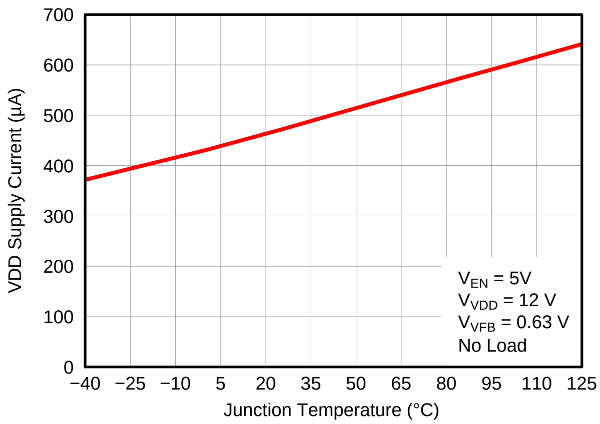

Figure 1. VDD Supply Current vs. Junction Temperature

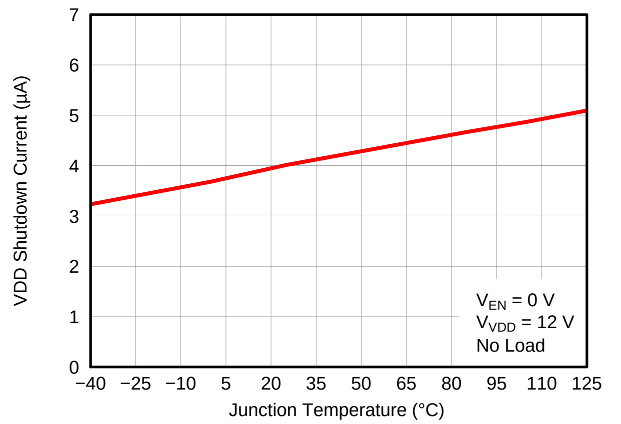

Figure 2. VDD Shutdown Current vs. Junction Temperature

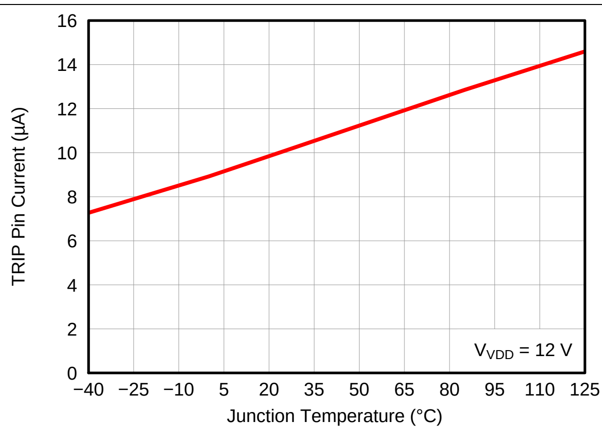

Figure 3. TRIP Pin Current vs. Junction Temperature

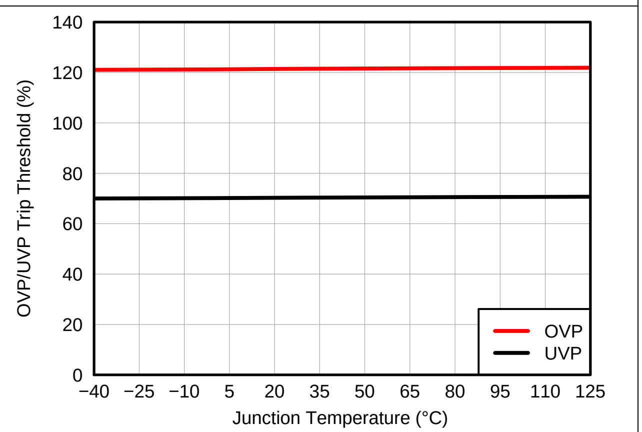

Figure 4. OVP/UVP Trip Threshold vs. Junction Temperature

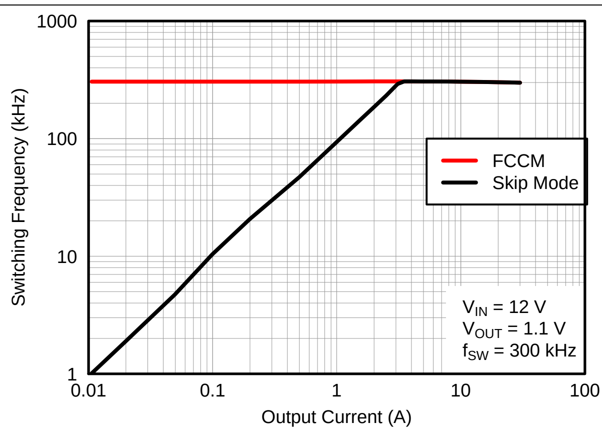

Figure 5. Switching Frequency vs. Output Current

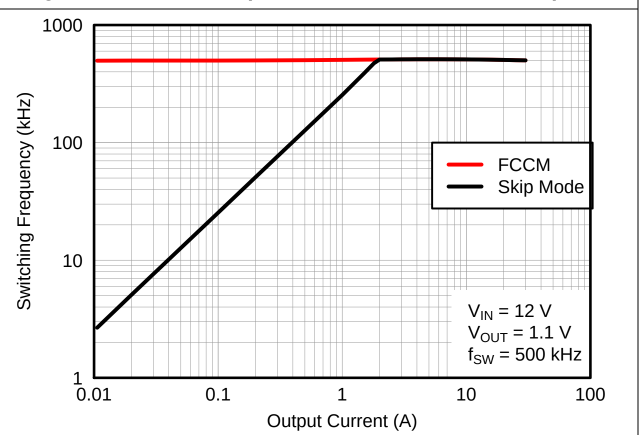

Figure 6. Switching Frequency vs. Output Current

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Typical Characteristics (continued)

For V_{OUT} = 5 V, a 744355182 inductor is used. For 1 ≤ V_{OUT} ≤ 3.3 V, a PA0513.441 inductor is used.

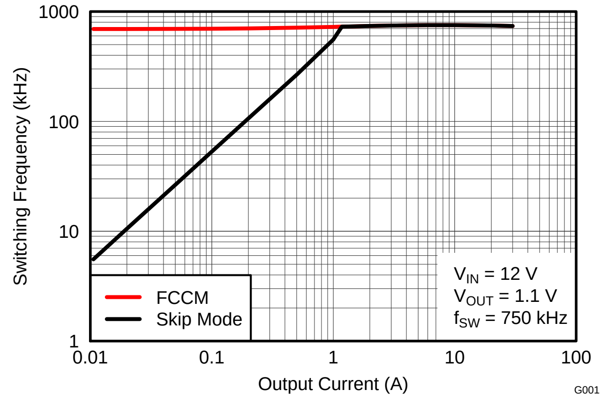

Figure 7. Switching Frequency vs. Output Current

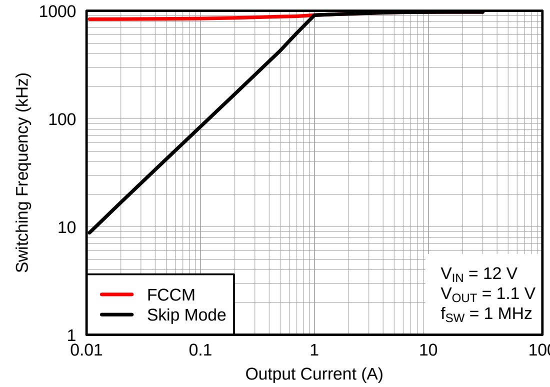

Figure 8. Switching Frequency vs. Output Current

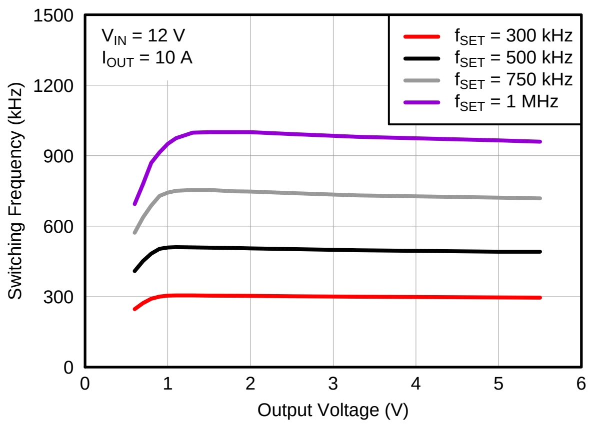

Figure 9. Switching Frequency vs. Output Voltage

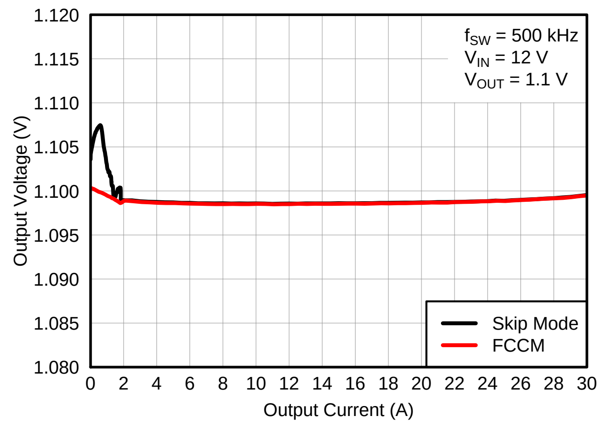

Figure 10. Output Voltage vs. Output Current

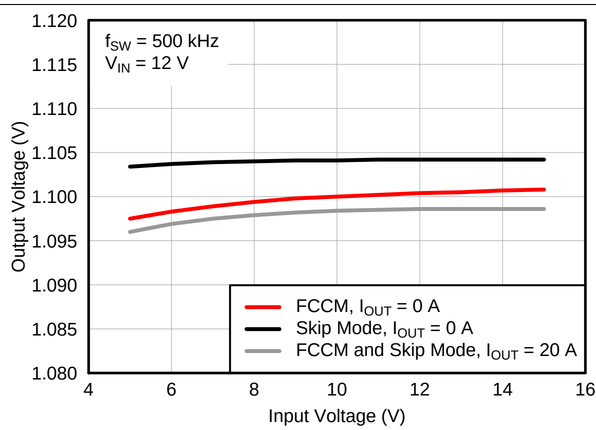

Figure 11. Output Voltage vs. Input Voltage

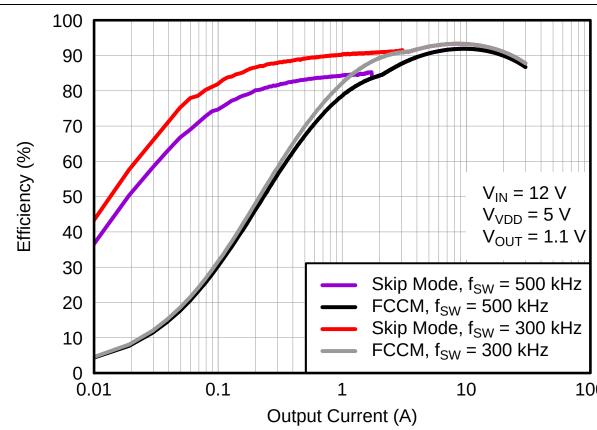

Figure 12. Efficiency vs Output Current

Typical Characteristics (continued)

For V_{OUT} = 5 V, a 744355182 inductor is used. For 1 ≤ V_{OUT} ≤ 3.3 V, a PA0513.441 inductor is used.

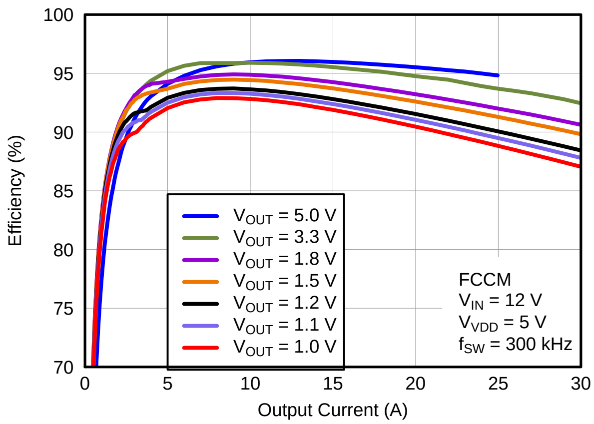

Figure 13. Efficiency vs Output Current

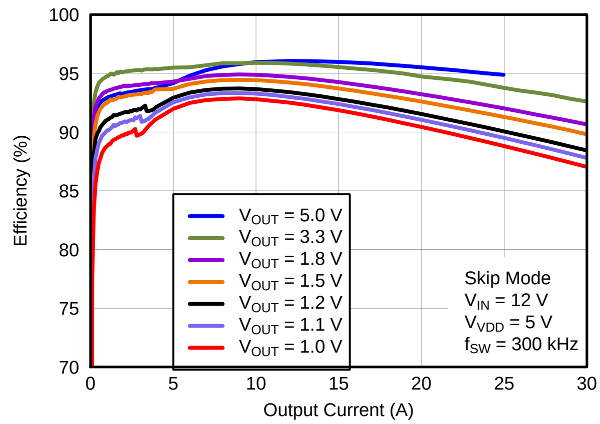

Figure 14. Efficiency vs Output Current

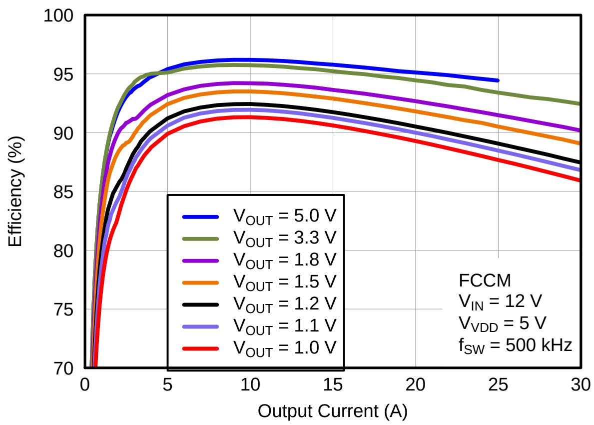

Figure 15. Efficiency vs Output Current

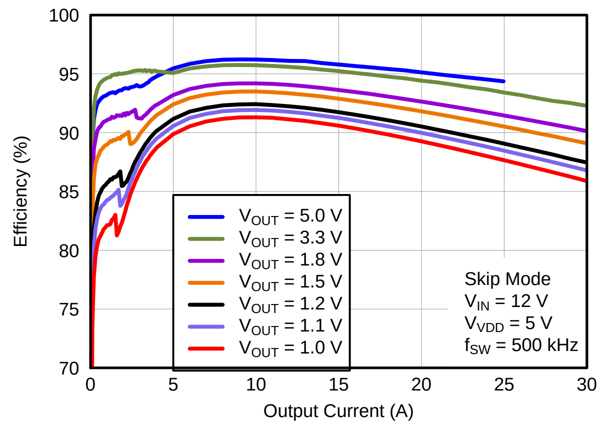

Figure 16. Efficiency vs Output Current

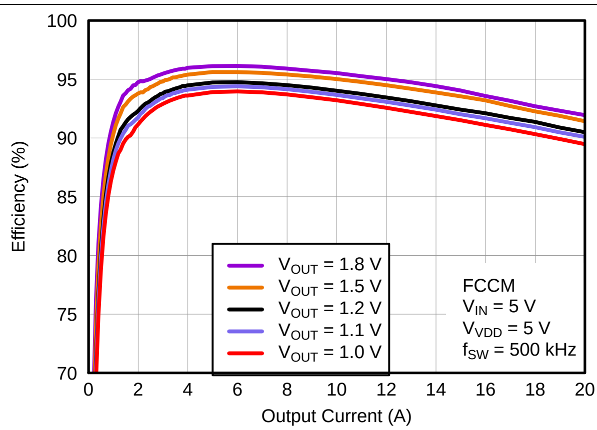

Figure 17. Efficiency vs Output Current

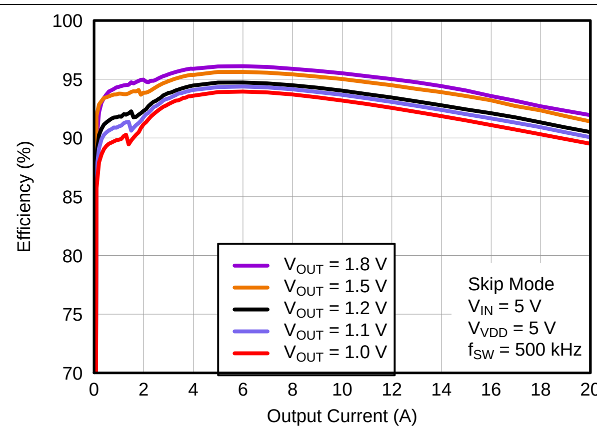

Figure 18. Efficiency vs Output Current

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Typical Characteristics (continued)

For V_{OUT} = 5 V, a 744355182 inductor is used. For 1 ≤ V_{OUT} ≤ 3.3 V, a PA0513.441 inductor is used.

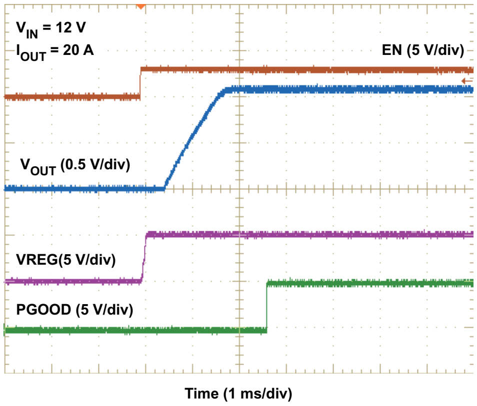

Figure 19. Start-Up Waveforms

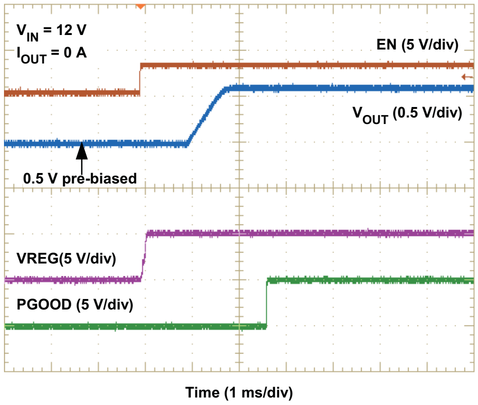

Figure 20. Pre-Bias Start-Up Waveforms

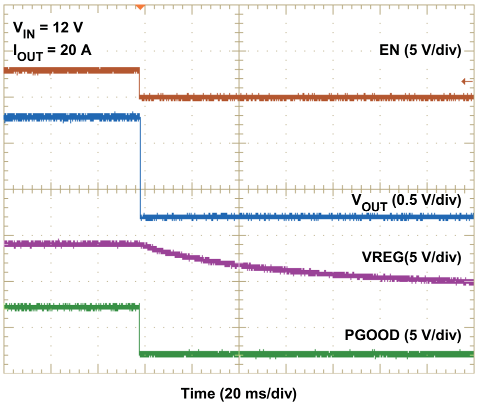

Figure 21. Shutdown Waveforms

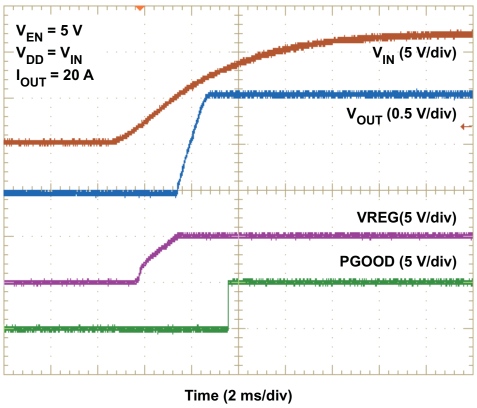

Figure 22. UVLO Start-Up Waveforms

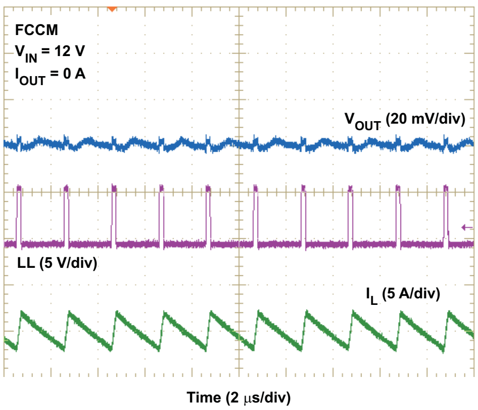

Figure 23. 1.1-V Output FCCM Mode Steady-State Operation

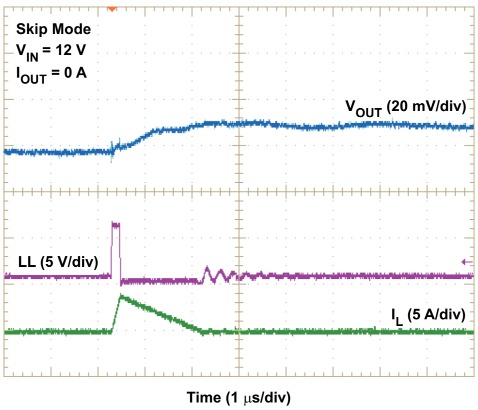

Figure 24. 1.1-V Output Skip Mode Steady-State Operation

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Typical Characteristics (continued)

For V_{OUT} = 5 V, a 744355182 inductor is used. For 1 ≤ V_{OUT} ≤ 3.3 V, a PA0513.441 inductor is used.

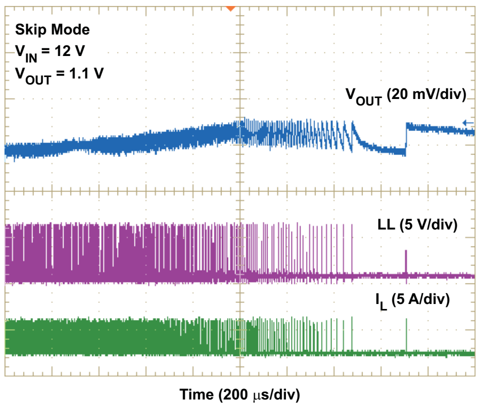

Figure 25. CCM to DCM Transition Waveforms

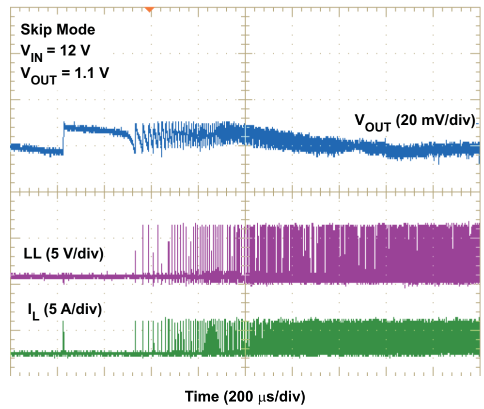

Figure 26. DCM to CCM Transition Waveforms

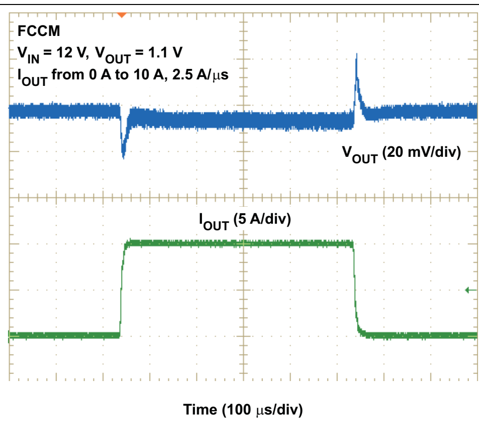

Figure 27. FCCM Load Transient

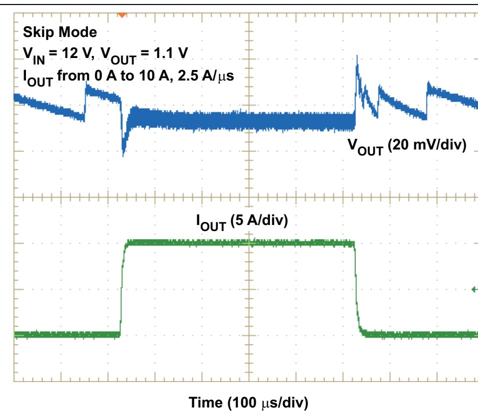

Figure 28. Skip Mode Load Transeint

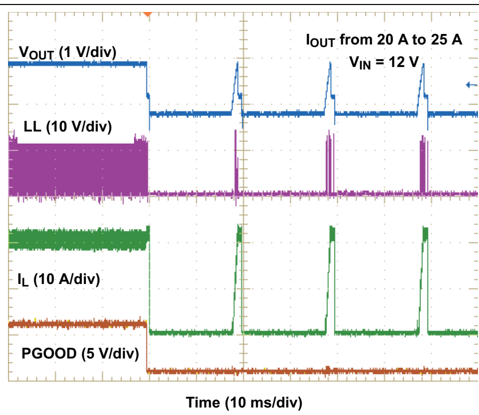

Figure 29. Overcurrent Protection Waveforms

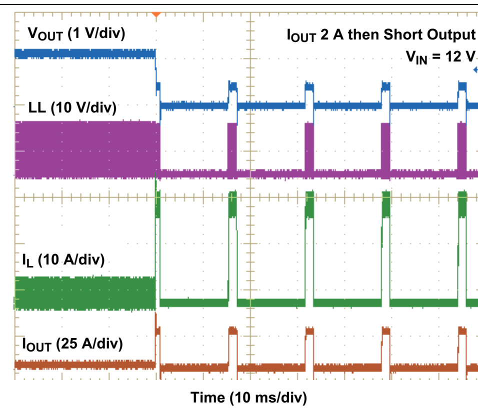

Figure 30. Output Short Circuit Protection Waveforms

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Typical Characteristics (continued)

For V_{OUT} = 5 V, a 744355182 inductor is used. For 1 ≤ V_{OUT} ≤ 3.3 V, a PA0513.441 inductor is used.

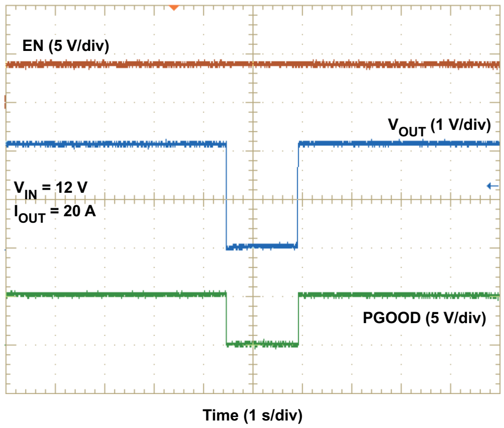
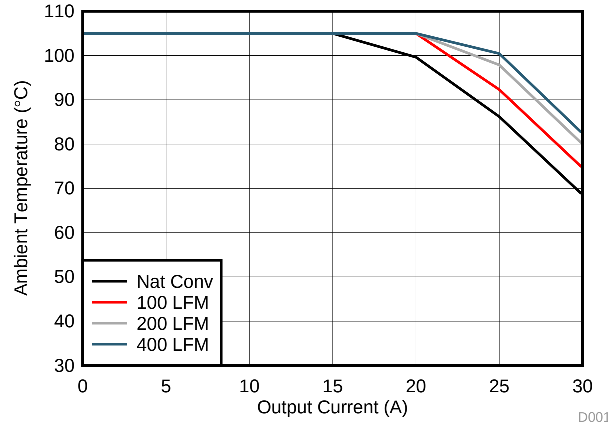
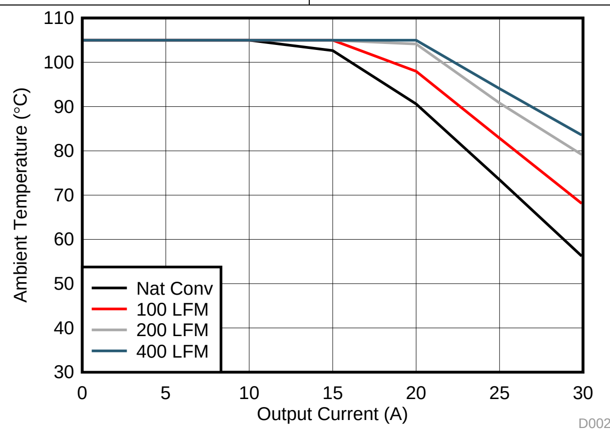


Figure 31. Over-temperature Protection Waveforms



V_{IN} = 12 V V_{OUT} = 1.2 V f_{SW} = 500 kHz

Figure 32. Safe Operating Area



V_{IN} = 12 V V_{OUT} = 5 V f_{SW} = 500 kHz

Figure 33. Safe Operating Area

7 Detailed Description

7.1 Overview

The TPS53355 is a high-efficiency, single channel, synchronous buck converter suitable for low output voltage point-of-load applications in computing and similar digital consumer applications. The device features proprietary D-CAP™ mode control combined with an adaptive on-time architecture. This combination is ideal for building modern low duty ratio, ultra-fast load step response DC-DC converters. The output voltage ranges from 0.6 V to 5.5 V. The conversion input voltage range is from 1.5 V up to 15 V and the VDD bias voltage is from 4.5 V to 25 V. The D-CAP™ mode uses the equivalent series resistance (ESR) of the output capacitor(s) to sense the device current. One advantage of this control scheme is that it does not require an external phase compensation network. This allows a simple design with a low external component count. Eight preset switching frequency values can be chosen using a resistor connected from the RF pin to ground or VREG. Adaptive on-time control tracks the preset switching frequency over a wide input and output voltage range while allowing the switching frequency to increase at the step-up of the load.

The TPS53355 has a MODE pin to select between auto-skip mode and forced continuous conduction mode (FCCM) for light load conditions. The MODE pin also sets the selectable soft-start time ranging from 0.7 ms to 5.6 ms as shown in Table 3.

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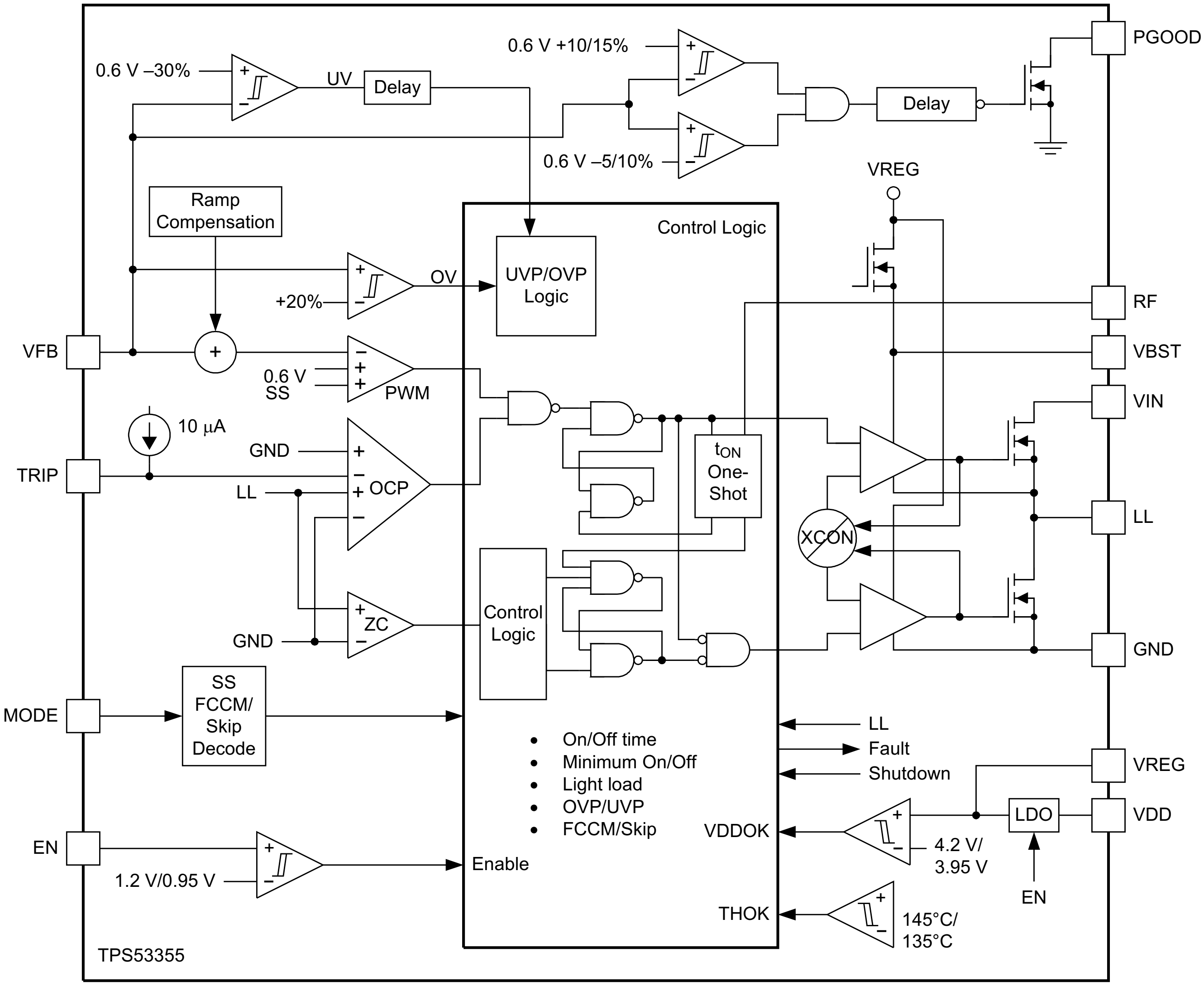
7.2 Functional Block Diagram



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NOTE

The thresholds in this block diagram are typical values. Refer to the *Electrical Characteristics* table for threshold limits.

7.3 Feature Description

7.3.1 5-V LDO and VREG Start-Up

TPS53355 provides an internal 5-V LDO function using input from VDD and output to VREG. When the VDD voltage rises above 2 V, the internal LDO is enabled and outputs voltage to the VREG pin. The VREG voltage provides the bias voltage for the internal analog circuitry and also provides the supply voltage for the gate drives.

NOTE

The 5-V LDO is controlled by the EN pin. The LDO starts-up any time VDD rises to approximately 2 V. Figure 34

Feature Description (continued)

7.3.2 Adaptive On-Time D-CAP Control and Frequency Selection

The TPS53355 does not have a dedicated oscillator to determine switching frequency. However, the device operates with pseudo-constant frequency by feed-forwarding the input and output voltages into the on-time one-shot timer. The adaptive on-time control adjusts the on-time to be inversely proportional to the input voltage and proportional to the output voltage ($t_{ON} \propto V_{OUT}/V_{IN}$).

This makes the switching frequency fairly constant in steady state conditions over a wide input voltage range. The switching frequency is selectable from eight preset values by a resistor connected between the RF pin and GND or between the RF pin and the VREG pin as shown in Table 1. (Maintaining open resistance sets the switching frequency to 500 kHz.)

Table 1. Resistor and Switching Frequency

RESISTOR (R_{RF}) CONNECTIONS		SWITCHING FREQUENCY (f_{SW}) (kHz)
VALUE (kΩ)	CONNECT TO	
0	GND	250
187	GND	300
619	GND	400
OPEN	n/a	500
866	VREG	650
309	VREG	750
124	VREG	850
0	VREG	970

The off-time is modulated by a PWM comparator. The VFB node voltage (the mid-point of resistor divider) is compared to the internal 0.6-V reference voltage added with a ramp signal. When both signals match, the PWM comparator asserts a set signal to terminate the off time (turn off the low-side MOSFET and turn on high-side MOSFET). The set signal is valid if the inductor current level is below the OCP threshold, otherwise the off time is extended until the current level falls below the threshold.

Figure 35 and Figure 36 show two on-time control schemes.

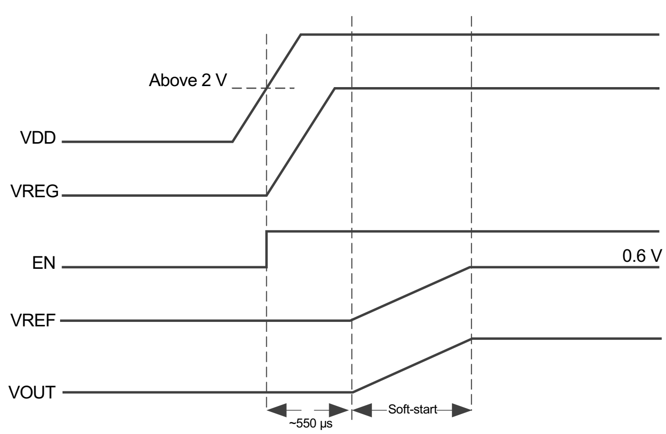

Figure 34. Power Up Sequence

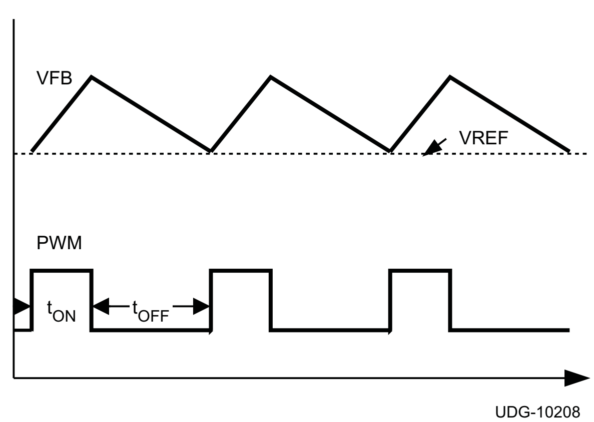

Figure 35. On-Time Control Without Ramp Compensation

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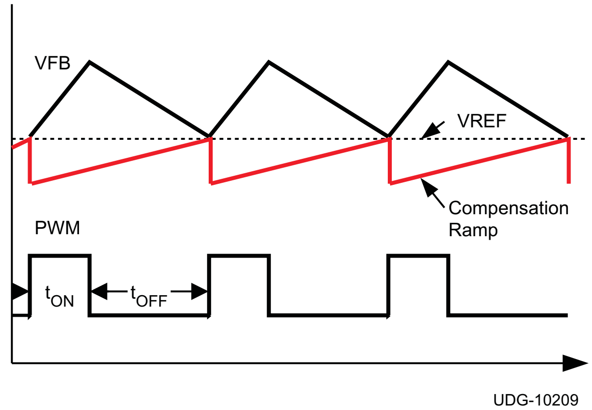

Figure 36. On-Time Control With Ramp Compensation

7.3.3 Ramp Signal

The TPS53355 adds a ramp signal to the 0.6-V reference in order to improve jitter performance. As described in the previous section, the feedback voltage is compared with the reference information to keep the output voltage in regulation. By adding a small ramp signal to the reference, the signal-to-noise ratio at the onset of a new switching cycle is improved. Therefore the operation becomes less jittery and more stable. The ramp signal is controlled to start with –7 mV at the beginning of an on-cycle and becomes 0 mV at the end of an off-cycle in steady state.

During skip mode operation, under discontinuous conduction mode (DCM), the switching frequency is lower than the nominal frequency and the off-time is longer than the off-time in CCM. Because of the longer off-time, the ramp signal extends after crossing 0 mV. However, it is clamped at 3 mV to minimize the DC offset.

7.3.4 Adaptive Zero Crossing

The TPS53355 has an adaptive zero crossing circuit which performs optimization of the zero inductor current detection at skip mode operation. This function pursues ideal low-side MOSFET turning off timing and compensates inherent offset voltage of the Z-C comparator and delay time of the Z-C detection circuit. It prevents SW-node swing-up caused by too late detection and minimizes diode conduction period caused by too early detection. As a result, better light load efficiency is delivered.

7.3.5 Power-Good

The TPS53355 has power-good output that indicates high when switcher output is within the target. The power-good function is activated after soft-start has finished. If the output voltage becomes within +10% and –5% of the target value, internal comparators detect power-good state and the power-good signal becomes high after a 1-ms internal delay. If the output voltage goes outside of +15% or –10% of the target value, the power-good signal becomes low after two microsecond (2-µs) internal delay. The power-good output is an open drain output and must be pulled up externally.

The power-good MOSFET is powered through the VDD pin. V_{VDD} must be >1 V in order to have a valid power-good logic. It is recommended to pull PGOOD up to VREG (or a voltage divided from VREG) so that the power-good logic is still valid even without VDD supply.

7.3.6 Current Sense, Overcurrent and Short Circuit Protection

TPS53355 has cycle-by-cycle overcurrent limiting control. The inductor current is monitored during the *OFF* state and the controller maintains the *OFF* state during the period in that the inductor current is larger than the overcurrent trip level. In order to provide both good accuracy and cost effective solution, TPS53355 supports temperature compensated MOSFET $R_{DS(on)}$ sensing. The TRIP pin should be connected to GND through the trip voltage setting resistor, R_{TRIP}. The TRIP terminal sources current (I_{TRIP}) which is 10 µA typically at room temperature, and the trip level is set to the OCL trip voltage V_{TRIP} as shown in Equation 1.

$$V_{TRIP}\left(mV\right)=R_{TRIP}\left(k\Omega\right)\times I_{TRIP}\left(\mu A\right) \quad (1)$$

The inductor current is monitored by the LL pin. The GND pin is used as the positive current sensing node and the LL pin is used as the negative current sense node. The trip current, I_{TRIP} has 4700ppm/°C temperature slope to compensate the temperature dependency of the $R_{DS(on)}$.

As the comparison is made during the *OFF* state, V_{TRIP} sets the valley level of the inductor current. Thus, the load current at the overcurrent threshold, I_{OCP}, can be calculated as shown in Equation 2.

$$I_{OCP} = \frac{V_{TRIP}}{\left(32 \times R_{DS(on)}\right)} + \frac{I_{IND(ripple)}}{2} = \frac{V_{TRIP}}{\left(32 \times R_{DS(on)}\right)} + \frac{1}{2 \times L \times f_{SW}} \times \frac{\left(V_{IN} - V_{OUT}\right) \times V_{OUT}}{V_{IN}} \quad (2)$$

In an overcurrent or short circuit condition, the current to the load exceeds the current to the output capacitor thus the output voltage tends to decrease. Eventually, it crosses the undervoltage protection threshold and shuts down. After a hiccup delay (16 ms with 0.7 ms sort-start), the controller restarts. If the overcurrent condition remains, the procedure is repeated and the device enters hiccup mode.

Hiccup time calculation:

$t_{HIC(wait)} = (2^n + 257) \times 4\ \mu s$

where

- n = 8, 9, 10, or 11 depending on soft start time selection (3)

$t_{HIC(dly)} = 7 \times (2^n + 257) \times 4\ \mu s$ (4)

Table 2. Hiccup Delay

SELECTED SOFT-START TIME (t_{SS}) (ms)	n	HICCUP WAIT TIME ($t_{HIC(wait)}$) (ms)	HICCUP DELAY TIME ($t_{HIC(dly)}$) (ms)
0.7	8	2.052	14.364
1.4	9	3.076	21.532
2.8	10	5.124	35.868
5.6	11	9.220	64.540

7.3.7 Overvoltage and Undervoltage Protection

TPS53355 monitors a resistor divided feedback voltage to detect over and under voltage. When the feedback voltage becomes lower than 70% of the target voltage, the UVP comparator output goes high and an internal UVP delay counter begins counting. After 1ms, TPS53355 latches OFF both high-side and low-side MOSFETs drivers. The controller restarts after a hiccup delay (16 ms with 0.7 ms soft-start). This function is enabled 1.5-ms after the soft-start is completed.

When the feedback voltage becomes higher than 120% of the target voltage, the OVP comparator output goes high and the circuit latches OFF the high-side MOSFET driver and latches ON the low-side MOSFET driver. The output voltage decreases. If the output voltage reaches UV threshold, then both high-side MOSFET and low-side MOSFET driver will be OFF and the device restarts after a hiccup delay. If the OV condition remains, both high-side MOSFET and low-side MOSFET driver remains OFF until the OV condition is removed.

7.3.8 UVLO Protection

The TPS53355 uses VREG undervoltage lockout protection (UVLO). When the VREG voltage is lower than 3.95 V, the device shuts off. When the VREG voltage is higher than 4.2 V, the device restarts. This is a non-latch protection.

7.3.9 Thermal Shutdown

TPS53355 monitors the temperature of itself. If the temperature exceeds the threshold value (typically 145°C), TPS53355 is shut off. When the temperature falls about 10°C below the threshold value, the device will turn back on. This is a non-latch protection.

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7.4 Device Functional Modes

7.4.1 Enable, Soft Start, and Mode Selection

When the EN pin voltage rises above the enable threshold voltage (typically 1.2 V), the controller enters its start-up sequence. The internal LDO regulator starts immediately and regulates to 5 V at the VREG pin. The controller then uses the first 250 μs to calibrate the switching frequency setting resistance attached to the RF pin and stores the switching frequency code in internal registers. During this period, the MODE pin also senses the resistance attached to this pin and determines the soft-start time. Switching is inhibited during this phase. In the second phase, an internal DAC starts ramping up the reference voltage from 0 V to 0.6 V. Depending on the MODE pin setting, the ramping up time varies from 0.7 ms to 5.6 ms. Smooth and constant ramp-up of the output voltage is maintained during start-up regardless of load current.

Table 3. Soft-Start and MODE Settings

MODE SELECTION	ACTION	SOFT-START TIME (ms)	R_{MODE} (kΩ)
Auto Skip	Pull down to GND	0.7	39
		1.4	100
		2.8	200
		5.6	475
Forced CCM(1)	Connect to PGOOD	0.7	39
		1.4	100
		2.8	200
		5.6	475

(1) Device enters FCCM after the PGOOD pin goes high when MODE is connected to PGOOD through the resistor R_{MODE}.

After soft-start begins, the MODE pin becomes the input of an internal comparator which determines auto skip or FCCM mode operation. If MODE voltage is higher than 1.3 V, the converter enters into FCCM mode. Otherwise it will be in auto skip mode at light load condition. Typically, when FCCM mode is selected, the MODE pin is connected to PGOOD through the R_{MODE} resistor, so that before PGOOD goes high the converter remains in auto skip mode.

7.4.2 Auto-Skip Eco-mode™ Light Load Operation

While the MODE pin is pulled low via R_{MODE}, TPS53355 automatically reduces the switching frequency at light load conditions to maintain high efficiency. Detailed operation is described as follows. As the output current decreases from heavy load condition, the inductor current is also reduced and eventually comes to the point that its rippled valley touches zero level, which is the boundary between continuous conduction and discontinuous conduction modes. The synchronous MOSFET is turned off when this zero inductor current is detected. As the load current further decreases, the converter runs into discontinuous conduction mode (DCM). The on-time is kept almost the same as it was in the continuous conduction mode so that it takes longer time to discharge the output capacitor with smaller load current to the level of the reference voltage. The transition point to the light-load operation $I_{OUT(LL)}$ (i.e., the threshold between continuous and discontinuous conduction mode) can be calculated as shown in Equation 5.

$$I_{OUT(LL)} = \frac{1}{2 \times L \times f_{SW}} \times \frac{(V_{IN} - V_{OUT}) \times V_{OUT}}{V_{IN}} \quad (5)$$

where

- f_{SW} is the PWM switching frequency

Switching frequency versus output current in the light load condition is a function of L, V_{IN} and V_{OUT}, but it decreases almost proportionally to the output current from the $I_{OUT(LL)}$ given in Equation 5. For example, it is 60 kHz at $I_{OUT(LL)}/5$ if the frequency setting is 300 kHz.

7.4.3 Forced Continuous Conduction Mode

When the MODE pin is tied to PGOOD through a resistor, the controller keeps continuous conduction mode (CCM) in light load condition. In this mode, switching frequency is kept almost constant over the entire load range which is suitable for applications that need tight control of the switching frequency at a cost of lower efficiency.

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8 Application and Implementation

NOTE

Information in the following applications sections is not part of the TI component specification, and TI does not warrant its accuracy or completeness. TI's customers are responsible for determining suitability of components for their purposes. Customers should validate and test their design implementation to confirm system functionality.

8.1 Application Information

The TPS53355 is a high-efficiency, single channel, synchronous buck converter suitable for low output voltage point-of-load applications in computing and similar digital consumer applications. The device features proprietary D-CAP mode control combined with an adaptive on-time architecture. This combination is ideal for building modern low duty ratio, ultra-fast load step response DC-DC converters. The output voltage ranges from 0.6 V to 5.5 V. The conversion input voltage range is from 1.5 V up to 15 V and the VDD bias voltage is from 4.5 V to 25 V. The D-CAP mode uses the equivalent series resistance (ESR) of the output capacitor(s) to sense the device current . One advantage of this control scheme is that it does not require an external phase compensation network. This allows a simple design with a low external component count. Eight preset switching frequency values can be chosen using a resistor connected from the RF pin to ground or VREG. Adaptive on-time control tracks the preset switching frequency over a wide input and output voltage range while allowing the switching frequency to increase at the step-up of the load.

8.1.1 Small Signal Model

From small-signal loop analysis, a buck converter using D-CAP™ mode can be simplified as shown in Figure 37.

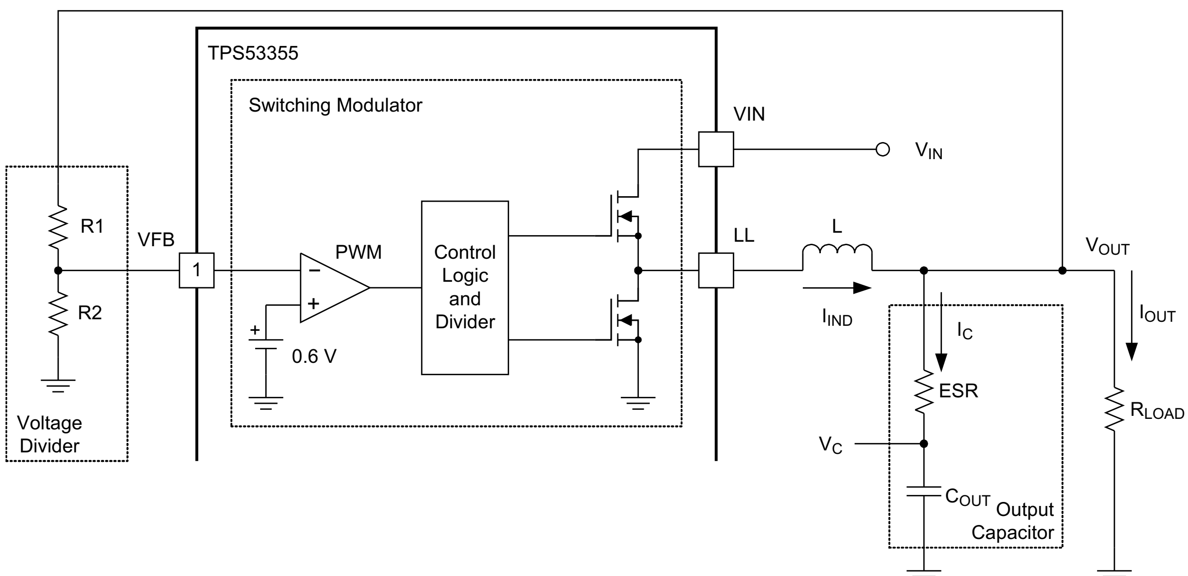

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Figure 37. Simplified Modulator Model

The output voltage is compared with the internal reference voltage (ramp signal is ignored here for simplicity). The PWM comparator determines the timing to turn on the high-side MOSFET. The gain and speed of the comparator can be assumed high enough to keep the voltage at the beginning of each on cycle substantially constant.

$$H(s) = \frac{1}{s \times ESR \times C_{OUT}} \tag{6}$$

For loop stability, the 0-dB frequency, f_0, defined below need to be lower than 1/4 of the switching frequency.

Application Information (continued)

$$f_0 = \frac{1}{2\pi \times ESR \times C_{OUT}} \leq \frac{f_{SW}}{4} \quad (7)$$

According to the equation above, the loop stability of D-CAP™ mode modulator is mainly determined by the capacitor's chemistry. For example, specialty polymer capacitors (SP-CAP) have an output capacitance in the order of several 100 µF and ESR in range of 10 mΩ. These makes f_0 on the order of 100 kHz or less, creating a stable loop. However, ceramic capacitors have an f_0 at more than 700 kHz, and need special care when used with this modulator. An application circuit for ceramic capacitor is described in *External Component Selection Using All Ceramic Output Capacitors*.

8.2 Typical Applications

8.2.1 Typical Application Circuit Diagram with Ceramic Output Capacitors

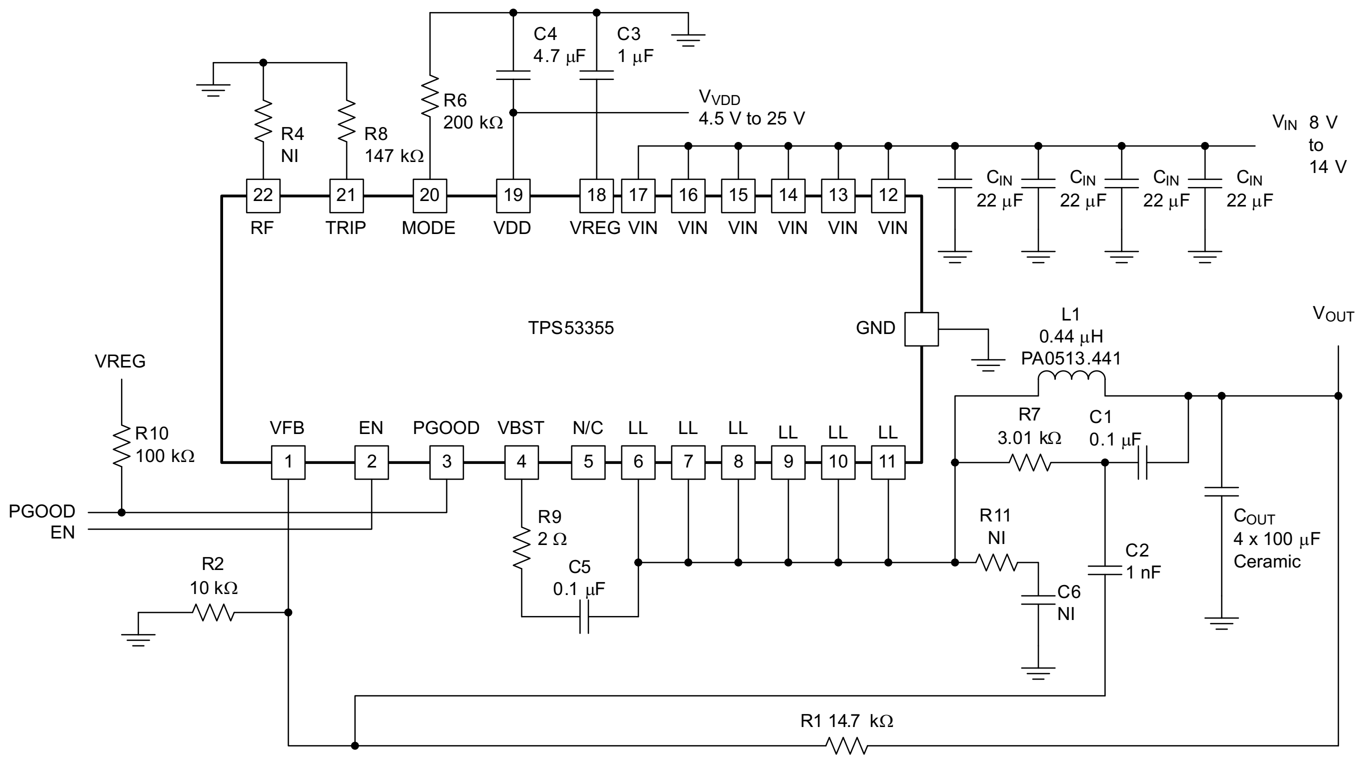

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Figure 38. Typical Application Circuit Diagram with Ceramic Output Capacitors Schematic

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Typical Applications (continued)

8.2.1.1 Design Requirements

Table 4. Design Parameters

PARAMETER		TEST CONDITION	MIN	TYP	MAX	UNIT
INPUT CHARACTERISTICS						
V_{IN}	Voltage range		8	12	14	V
I_{MAX}	Maximum input current	V_{IN} = 8 V, I_{OUT} = 30 A		6.3		A
	No load input current	V_{IN} = 14 V, I_{OUT} = 0 A with auto-skip mode		1		mA
OUTPUT CHARACTERISTICS						
V_{OUT}	Output voltage			1.5		
	Output voltage regulation	Line regulation, 8 V ≤ V_{IN} ≤ 15 V		0.1%		
		Load regulation, V_{IN} = 12 V, 0 A ≤ I_{OUT} ≤ 30 A with FCCM		0.2%		
V_{RIPPLE}	Output voltage ripple	V_{IN} = 12 V, I_{OUT} = 30 A with FCCM		20		mV_{PP}
I_{LOAD}	Output load current		0		30	A
I_{OCP}	Output overcurrent threshold			34		A
t_{SS}	Soft-start time			1.4		ms
SYSTEMS CHARACTERISTICS						
f_{SW}	Switching frequency			500		kHz
η	Peak efficiency	V_{IN} = 12 V, V_{OUT} = 1.1 V, I_{OUT} = 10 A		91.87%		
	Full load efficiency	V_{IN} = 12 V, V_{OUT} = 1.1 V, I_{OUT} = 30 A		89.46%		
T_A	Operating temperature			25		°C

8.2.1.2 Detailed Design Procedure

8.2.1.2.1 External Component Selection

The external components selection is a simple process when using organic semiconductors or special polymer output capacitors.

1. **Select operation mode and soft-start time**

   Select operation mode and soft-start time using Table 3.

2. **Select switching frequency**

   Select the switching frequency from 250 kHz to 1 MHz using Table 1.

3. **Choose the inductor**

   The inductance value should be determined to give the ripple current of approximately 1/4 to 1/2 of maximum output current. Larger ripple current increases output ripple voltage and improves signal-to-noise ratio and helps ensure stable operation, but increases inductor core loss. Using 1/3 ripple current to maximum output current ratio, the inductance can be determined by Equation 8.

$$L = \frac{1}{I_{IND(ripple)} \times f_{SW}} \times \frac{\left(V_{IN(max)} - V_{OUT}\right) \times V_{OUT}}{V_{IN(max)}} = \frac{3}{I_{OUT(max)} \times f_{SW}} \times \frac{\left(V_{IN(max)} - V_{OUT}\right) \times V_{OUT}}{V_{IN(max)}} \quad (8)$$

   The inductor requires a low DCR to achieve good efficiency. It also requires enough room above peak inductor current before saturation. The peak inductor current can be estimated in Equation 9.

$$I_{IND(peak)} = \frac{V_{TRIP}}{32 \times R_{DS(on)}} + \frac{1}{L \times f_{SW}} \times \frac{\left(V_{IN(max)} - V_{OUT}\right) \times V_{OUT}}{V_{IN(max)}} \quad (9)$$

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4. **External component selection with all ceramic output capacitors**

   Refer to *External Component Selection Using All Ceramic Output Capacitors* to select external components because ceramic output capacitors are used in this design.

5. **Choose the overcurrent setting resistor**

   The overcurrent setting resistor, R_{TRIP}, can be determined by Equation 10.

$$R_{TRIP}(k\Omega) = \frac{\left(I_{OCP} - \left(\frac{1}{2 \times L \times f_{SW}}\right) \times \frac{(V_{IN} - V_{OUT}) \times V_{OUT}}{V_{IN}}\right) \times 32 \times R_{DS(on)}(m\Omega)}{I_{TRIP}(\mu A)}$$

where

- I_{TRIP} is the TRIP pin sourcing current (10 µA)
- $R_{DS(on)}$ is the thermally compensated on-time resistance value of the low-side MOSFET (10)

Use an $R_{DS(on)}$ value of 1.5 mΩ for an overcurrent level of approximately 30 A. Use an $R_{DS(on)}$ value of 1.7 mΩ for overcurrent level of approximately 10 A.

8.2.1.2.2 External Component Selection Using All Ceramic Output Capacitors

When a ceramic output capacitor is used, the stability criteria in Equation 7 cannot be satisfied. The ripple injection approach as shown in Figure 38 is implemented to increase the ripple on the VFB pin and make the system stable. In addition to the selections made using steps 1 through step 6 in *External Component Selection*, the ripple injection components must be selected. The C2 value can be fixed at 1 nF. The value of C1 can be selected between 10 nF to 200 nF.

$$\frac{L \times C_{OUT}}{R7 \times C1} > N \times \frac{t_{ON}}{2}$$

where

- N is the coefficient to account for L and C_{OUT} variation (11)

N is also used to provide enough margin for stability. It is recommended N=2 for V_{OUT} ≤ 1.8 V and N=4 for V_{OUT} ≥ 3.3 V or when L ≤ 250 nH. The higher V_{OUT} needs a higher N value because the effective output capacitance is reduced significantly with higher DC bias. For example, a 6.3-V, 22-µF ceramic capacitor may have only 8 µF of effective capacitance when biased at 5 V.

Because the VFB pin voltage is regulated at the valley, the increased ripple on the VFB pin causes the increase of the VFB DC value. The AC ripple coupled to the VFB pin has two components, one coupled from SW node and the other coupled from the VOUT pin and they can be calculated using Equation 12 and Equation 13 when neglecting the output voltage ripple caused by equivalent series inductance (ESL).

$$V_{INJ\_SW} = \frac{V_{IN} - V_{OUT}}{R7 \times C1} \times \frac{D}{f_{SW}} \qquad (12)$$

$$V_{INJ\_OUT} = ESR \times I_{IND(ripple)} + \frac{I_{IND(ripple)}}{8 \times C_{OUT} \times f_{SW}} \qquad (13)$$

It is recommended that V_{INJ_SW} to be less than 50 mV. If the calculated V_{INJ_SW} is higher than 50 mV, then other parameters need to be adjusted to reduce it. For example, C_{OUT} can be increased to satisfy Equation 11 with a higher R7 value, thereby reducing V_{INJ_SW}.

The DC voltage at the VFB pin can be calculated by Equation 14:

$$V_{VFB} = 0.6 + \frac{V_{INJ\_SW} + V_{INJ\_OUT}}{2} \qquad (14)$$

And the resistor divider value can be determined by Equation 15:

$$R1 = \frac{V_{OUT} - V_{VFB}}{V_{VFB}} \times R2 \qquad (15)$$

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8.2.1.3 Application Curves

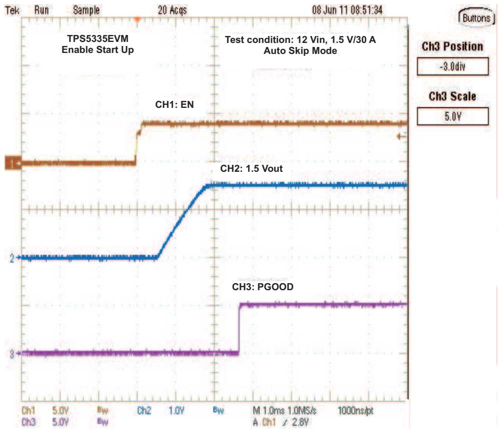

Figure 39. Enable Turn-on

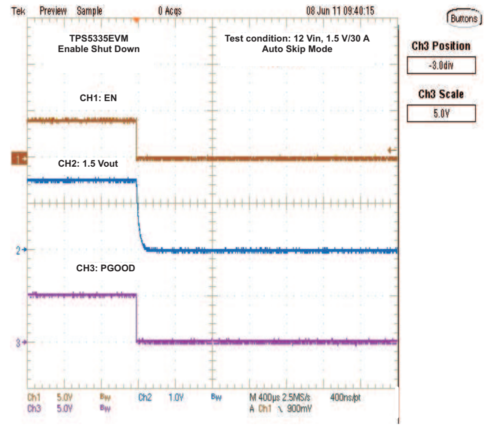

Figure 40. Enable Turn-off

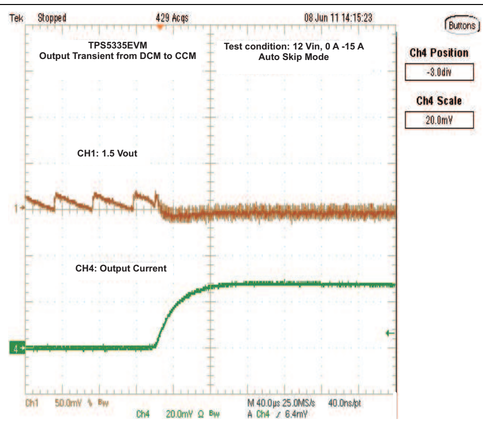

Figure 41. Output Transient From DCM to CCM

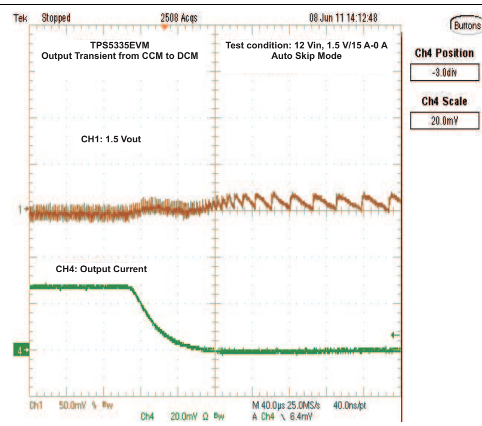

Figure 42. Output Transient From CCM to DCM

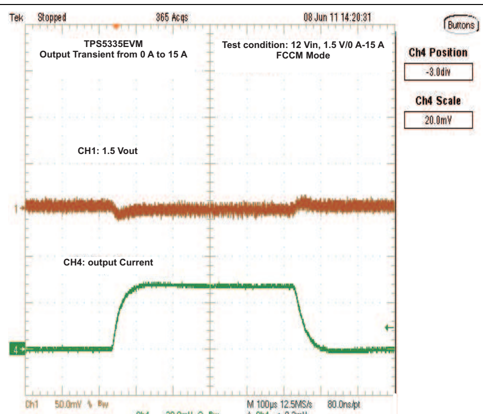

Figure 43. Output Transient With FCCM Mode

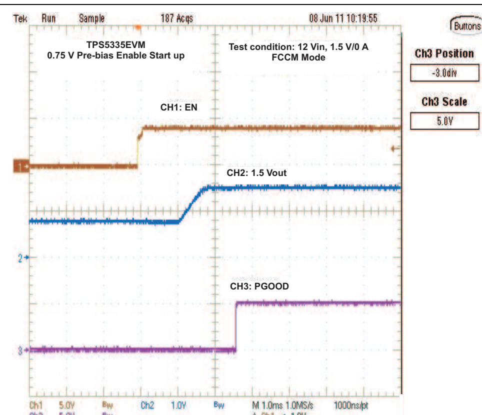

Figure 44. Output 0.75-V Prebias Turn-on

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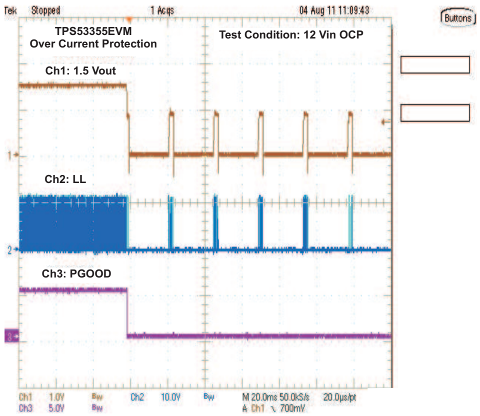

Figure 45. Output Overcurrent Protection

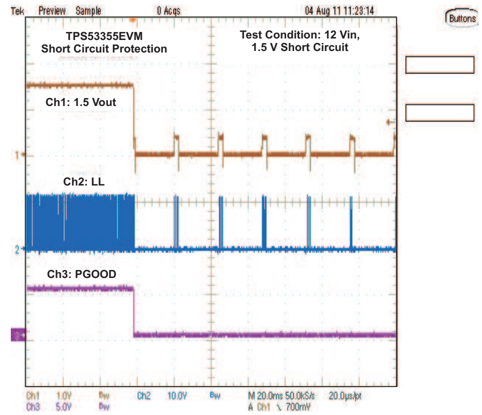

Figure 46. Output Short Circuit Protection

8.2.2 Typical Application Circuit

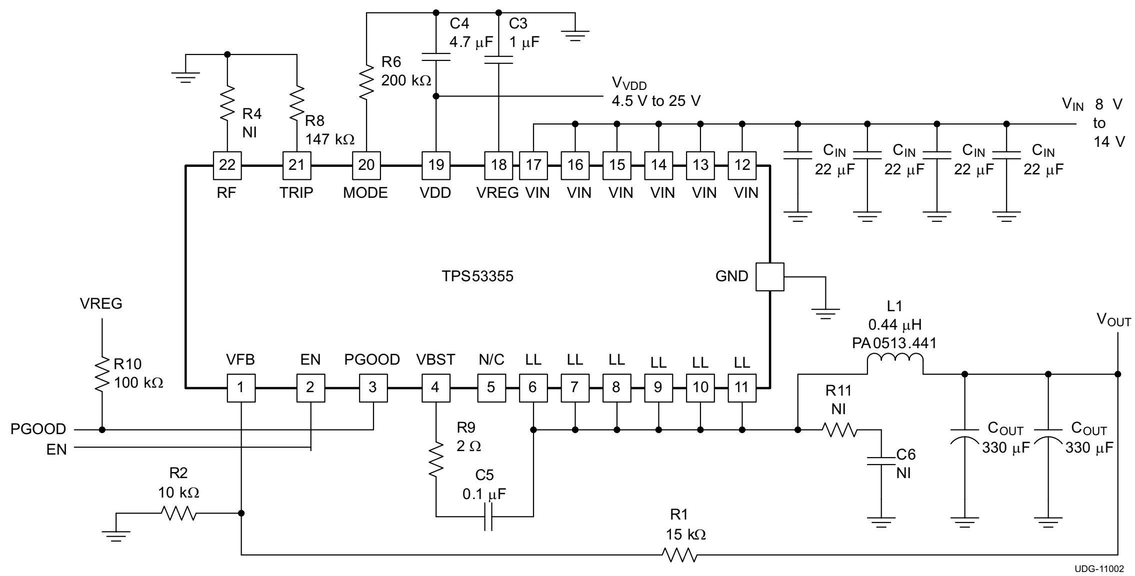

Figure 47. Typical Application Circuit Diagram

8.2.2.1 Design Requirements

Table 5. Design Parameters

PARAMETER		TEST CONDITION	MIN	TYP	MAX	UNIT
INPUT CHARACTERISTICS						
V_{IN}	Voltage range		8	12	14	V
I_{MAX}	Maximum input current	V_{IN} = 8 V, I_{OUT} = 30 A		6.3		A
	No load input current	V_{IN} = 14 V, I_{OUT} = 0 A with auto-skip mode		1		mA
OUTPUT CHARACTERISTICS						
V_{OUT}	Output voltage			1.5		
	Output voltage regulation	Line regulation, 8 V ≤ V_{IN} ≤ 15 V		0.1%		
		Load regulation, V_{IN} = 12 V, 0 A ≤ I_{OUT} ≤ 30 A with FCCM		0.2%		
V_{RIPPLE}	Output voltage ripple	V_{IN} = 12 V, I_{OUT} = 30 A with FCCM		20		mV_{PP}
I_{LOAD}	Output load current		0		30	A
I_{OCP}	Output overcurrent threshold			34		A

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Table 5. Design Parameters (continued)

PARAMETER		TEST CONDITION	MIN	TYP	MAX	UNIT
t_{SS}	Soft-start time			1.4		ms
SYSTEMS CHARACTERISTICS						
f_{SW}	Switching frequency			500		kHz
η	Peak efficiency	V_{IN} = 12 V, V_{OUT} = 1.1 V, I_{OUT} = 10 A		91.87%		
	Full load efficiency	V_{IN} = 12 V, V_{OUT} = 1.1 V, I_{OUT} = 30 A		89.46%		
T_A	Operating temperature			25		°C

8.2.2.2 Detailed Design Procedure

8.2.2.2.1 External Component Selection

Refer to *External Component Selection Using All Ceramic Output Capacitors* for guidelines for this design with all ceramic output capacitors.

The external components selection is a simple process when using organic semiconductors or special polymer output capacitors.

1. **Select operation mode and soft-start time**

   Select operation mode and soft-start time using Table 3.

2. **Select switching frequency**

   Select the switching frequency from 250 kHz to 1 MHz using Table 1.

3. **Choose the inductor**

   The inductance value should be determined to give the ripple current of approximately 1/4 to 1/2 of maximum output current. Larger ripple current increases output ripple voltage and improves signal-to-noise ratio and helps ensure stable operation, but increases inductor core loss. Using 1/3 ripple current to maximum output current ratio, the inductance can be determined by Equation 16.

$$L = \frac{1}{I_{IND(ripple)} \times f_{SW}} \times \frac{\left(V_{IN(max)} - V_{OUT}\right) \times V_{OUT}}{V_{IN(max)}} = \frac{3}{I_{OUT(max)} \times f_{SW}} \times \frac{\left(V_{IN(max)} - V_{OUT}\right) \times V_{OUT}}{V_{IN(max)}} \quad (16)$$

   The inductor requires a low DCR to achieve good efficiency. It also requires enough room above peak inductor current before saturation. The peak inductor current can be estimated in Equation 9.

$$I_{IND(peak)} = \frac{V_{TRIP}}{32 \times R_{DS(on)}} + \frac{1}{L \times f_{SW}} \times \frac{\left(V_{IN(max)} - V_{OUT}\right) \times V_{OUT}}{V_{IN(max)}} \quad (17)$$

4. **Choose the output capacitors**

   When organic semiconductor capacitor(s) or specialty polymer capacitor(s) are used, for loop stability, capacitance and ESR should satisfy Equation 7. For jitter performance, Equation 18 is a good starting point to determine ESR.

$$ESR = \frac{V_{OUT} \times 10\,mV \times (1 - D)}{0.6\,V \times I_{IND(ripple)}} = \frac{10\,mV \times L \times f_{SW}}{0.6\,V} = \frac{L \times f_{SW}}{60}\,(\Omega)$$

   where

   - D is the duty factor.
   - The required output ripple slope is approximately 10 mV per t_{SW} (switching period) in terms of VFB terminal voltage. (18)

5. **Determine the value of R1 and R2**

The output voltage is programmed by the voltage-divider resistor, R1 and R2 shown in Figure 37. R1 is connected between VFB pin and the output, and R2 is connected between the VFB pin and GND. Recommended R2 value is from 1 kΩ to 20 kΩ. Determine R1 using Equation 19.

$$R1 = \frac{V_{OUT} - \frac{I_{IND(ripple)} \times ESR}{2} - 0.6}{0.6} \times R2 \tag{19}$$

6. **Choose the overcurrent setting resistor**

The overcurrent setting resistor, R_{TRIP}, can be determined by Equation 10.

$$R_{TRIP}(k\Omega) = \frac{\left(I_{OCP} - \left(\frac{1}{2 \times L \times f_{SW}}\right) \times \frac{(V_{IN} - V_{OUT}) \times V_{OUT}}{V_{IN}}\right) \times 32 \times R_{DS(on)}(m\Omega)}{I_{TRIP}(\mu A)} \tag{20}$$

where

- I_{TRIP} is the TRIP pin sourcing current (10 µA)
- $R_{DS(on)}$ is the thermally compensated on-time resistance value of the low-side MOSFET

Use an $R_{DS(on)}$ value of 1.5 mΩ for an overcurrent level of approximately 30 A. Use an $R_{DS(on)}$ value of 1.7 mΩ for overcurrent level of approximately 10 A.

8.2.2.3 Application Curves

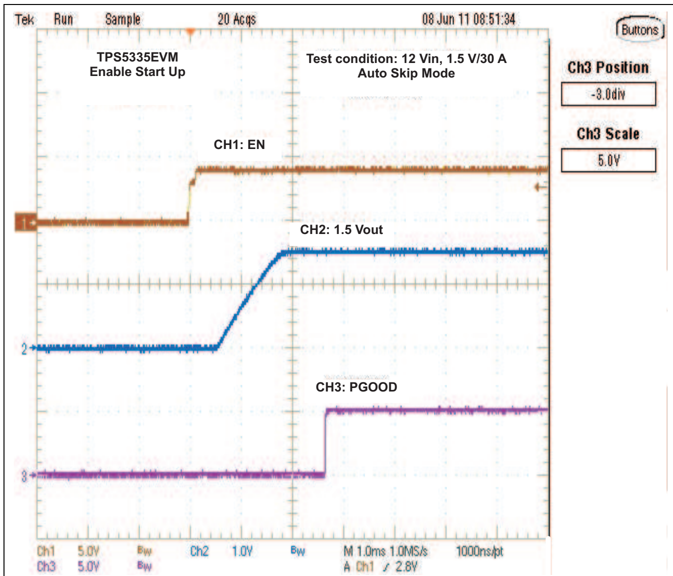

Figure 48. Enable Turn-on

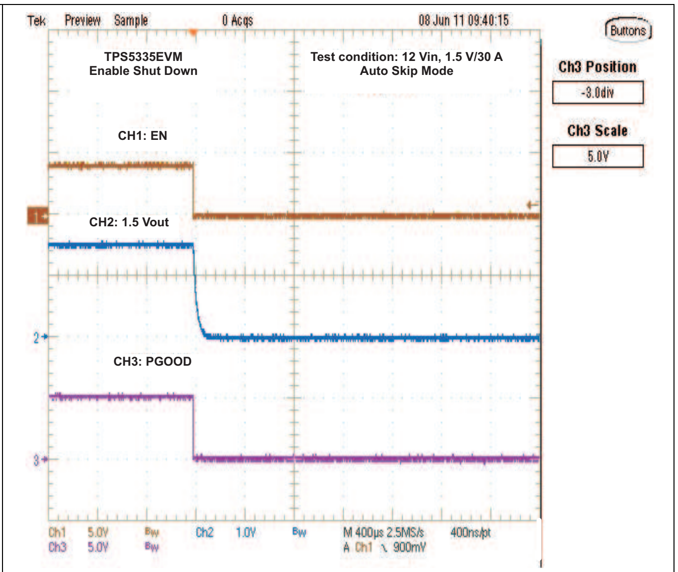

Figure 49. Enable Turn-off

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9 Power Supply Recommendations

The device is designed to operate from an input voltage supply range between 1.5 V and 22 V (4.5-V to 25-V biased). This input supply must be well regulated. Proper bypassing of input supplies and internal regulators is also critical for noise performance, as is PCB layout and grounding scheme. See the recommendations in *Layout*.

10 Layout

10.1 Layout Guidelines

Certain points must be considered before starting a layout work using the TPS53355.

- The power components (including input/output capacitors, inductor and TPS53355) should be placed on one side of the PCB (solder side). At least one inner plane should be inserted, connected to ground, in order to shield and isolate the small signal traces from noisy power lines.
- All sensitive analog traces and components such as VFB, PGOOD, TRIP, MODE and RF should be placed away from high-voltage switching nodes such as LL, VBST to avoid coupling. Use internal layer(s) as ground plane(s) and shield feedback trace from power traces and components.
- Place the VIN decoupling capacitors as close to the VIN and PGND pins as possible to minimize the input AC current loop.
- Because the TPS53355 controls output voltage referring to voltage across VOUT capacitor, the top-side resistor of the voltage divider should be connected to the positive node of the VOUT capacitor. The GND of the bottom side resistor should be connected to the GND pad of the device. The trace from these resistors to the VFB pin should be short and thin.
- Place the frequency setting resistor (R_F), OCP setting resistor (R_{TRIP}) and mode setting resistor (R_{MODE}) as close to the device as possible. Use the common GND via to connect them to GND plane if applicable.
- Place the VDD and VREG decoupling capacitors as close to the device as possible. Make sure GND vias are provided for each decoupling capacitor and make the loop as small as possible.
- The PCB trace defined as switch node, which connects the LL pins and high-voltage side of the inductor, should be as short and wide as possible.
- Connect the ripple injection V_{OUT} signal (V_{OUT} side of the C1 capacitor in Figure 38) from the terminal of ceramic output capacitor. The AC coupling capacitor (C2 in Figure 38) should be placed near the device, and R7 and C1 can be placed near the power stage.
- Use separate vias or trace to connect LL node to snubber, boot strap capacitor and ripple injection resistor. Do not combine these connections.

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10.2 Layout Example

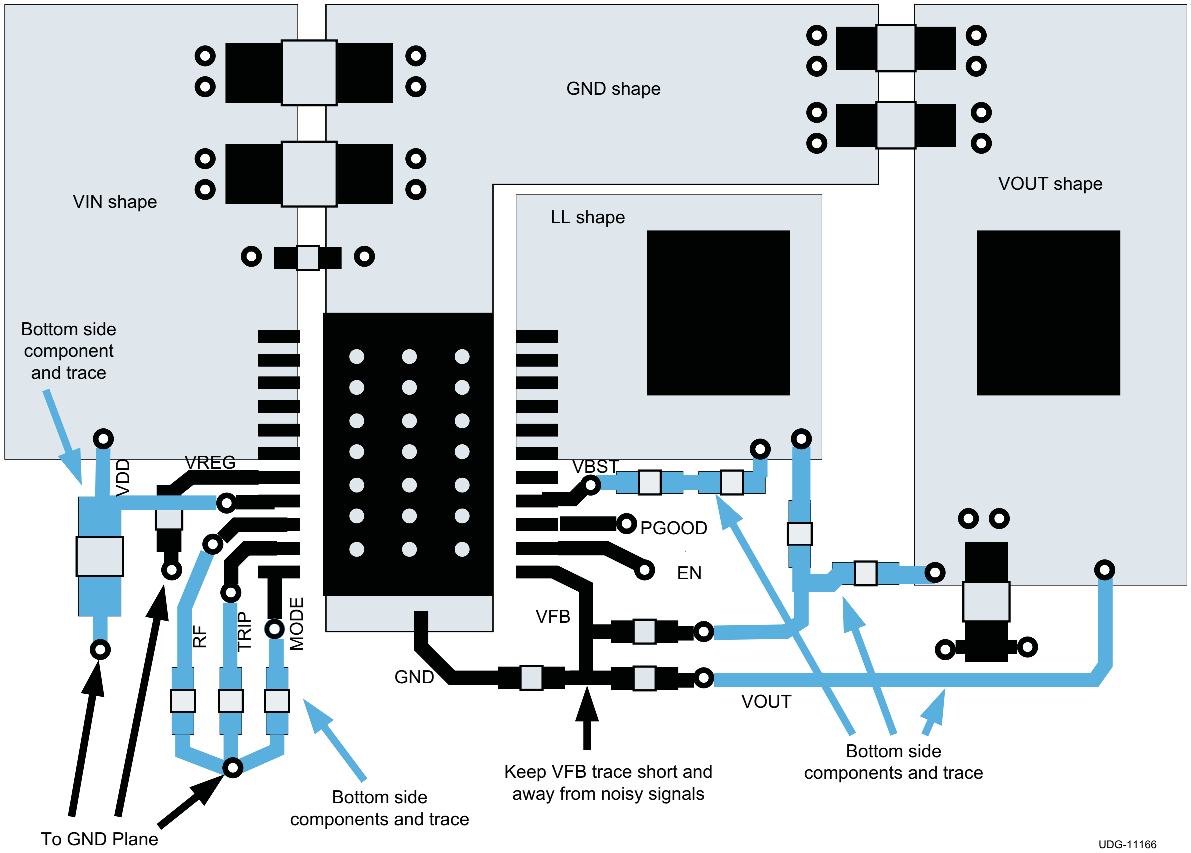

Figure 50. Layout Recommendation

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11 器件和文档支持

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要接收文档更新通知，请导航至 ti.com 上的器件产品文件夹。请单击右上角的*通知我* 进行注册，即可收到任意产品信息更改每周摘要。有关更改的详细信息，请查看任意已修订文档中包含的修订历史记录。

11.2 社区资源

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Design Support ***TI's Design Support*** Quickly find helpful E2E forums along with design support tools and contact information for technical support.

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11.4 静电放电警告

这些装置包含有限的内置 ESD 保护。 存储或装卸时，应将导线一起截短或将装置放置于导电泡棉中，以防止 MOS 门极遭受静电损伤。

11.5 Glossary

SLYZ022 — *TI Glossary*.

This glossary lists and explains terms, acronyms, and definitions.

12 机械、封装和可订购信息

以下页面包括机械、封装和可订购信息。这些信息是指定器件的最新可用数据。这些数据发生变化时，我们可能不会另行通知或修订此文档。如欲获取此产品说明书的浏览器版本，请参见左侧的导航栏。

PACKAGING INFORMATION

Orderable Device	Status (1)	Package Type	Package Drawing	Pins	Package Qty	Eco Plan (2)	Lead/Ball Finish (6)	MSL Peak Temp (3)	Op Temp (°C)	Device Marking (4/5)	Samples
HPA01111DQPR	ACTIVE	LSON-CLIP	DQP	22	2500	Pb-Free (RoHS Exempt)	CU NIPDAU	Level-2-260C-1 YEAR	-40 to 85	53355DQP	Samples
TPS53355DQPR	ACTIVE	LSON-CLIP	DQP	22	2500	Pb-Free (RoHS Exempt)	CU NIPDAU | CU SN	Level-2-260C-1 YEAR	-40 to 85	53355DQP	Samples
TPS53355DQPT	ACTIVE	LSON-CLIP	DQP	22	250	Pb-Free (RoHS Exempt)	CU NIPDAU | CU SN	Level-2-260C-1 YEAR	-40 to 85	53355DQP	Samples

(1) The marketing status values are defined as follows:
ACTIVE: Product device recommended for new designs.
LIFEBUY: TI has announced that the device will be discontinued, and a lifetime-buy period is in effect.
NRND: Not recommended for new designs. Device is in production to support existing customers, but TI does not recommend using this part in a new design.
PREVIEW: Device has been announced but is not in production. Samples may or may not be available.
OBSOLETE: TI has discontinued the production of the device.

(2) **RoHS:** TI defines "RoHS" to mean semiconductor products that are compliant with the current EU RoHS requirements for all 10 RoHS substances, including the requirement that RoHS substance do not exceed 0.1% by weight in homogeneous materials. Where designed to be soldered at high temperatures, "RoHS" products are suitable for use in specified lead-free processes. TI may reference these types of products as "Pb-Free".
RoHS Exempt: TI defines "RoHS Exempt" to mean products that contain lead but are compliant with EU RoHS pursuant to a specific EU RoHS exemption.
Green: TI defines "Green" to mean the content of Chlorine (Cl) and Bromine (Br) based flame retardants meet JS709B low halogen requirements of <=1000ppm threshold. Antimony trioxide based flame retardants must also meet the <=1000ppm threshold requirement.

(3) MSL, Peak Temp. - The Moisture Sensitivity Level rating according to the JEDEC industry standard classifications, and peak solder temperature.

(4) There may be additional marking, which relates to the logo, the lot trace code information, or the environmental category on the device.

(5) Multiple Device Markings will be inside parentheses. Only one Device Marking contained in parentheses and separated by a "~" will appear on a device. If a line is indented then it is a continuation of the previous line and the two combined represent the entire Device Marking for that device.

(6) Lead/Ball Finish - Orderable Devices may have multiple material finish options. Finish options are separated by a vertical ruled line. Lead/Ball Finish values may wrap to two lines if the finish value exceeds the maximum column width.

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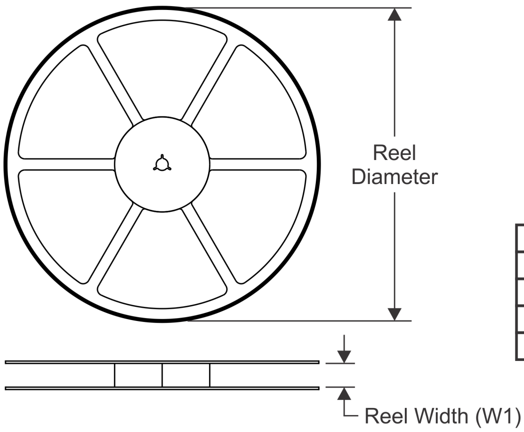
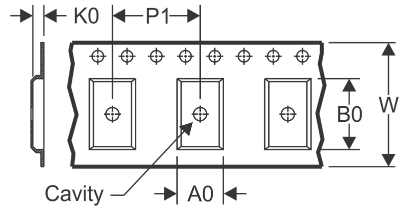
TAPE AND REEL INFORMATION

REEL DIMENSIONS

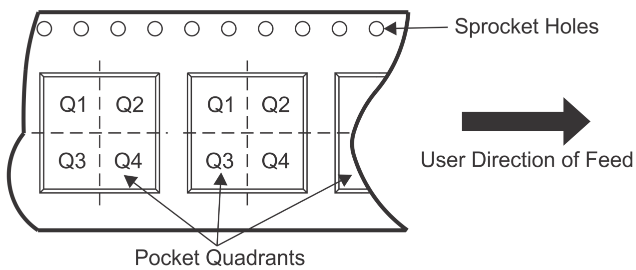


TAPE DIMENSIONS



A0	Dimension designed to accommodate the component width
B0	Dimension designed to accommodate the component length
K0	Dimension designed to accommodate the component thickness
W	Overall width of the carrier tape
P1	Pitch between successive cavity centers

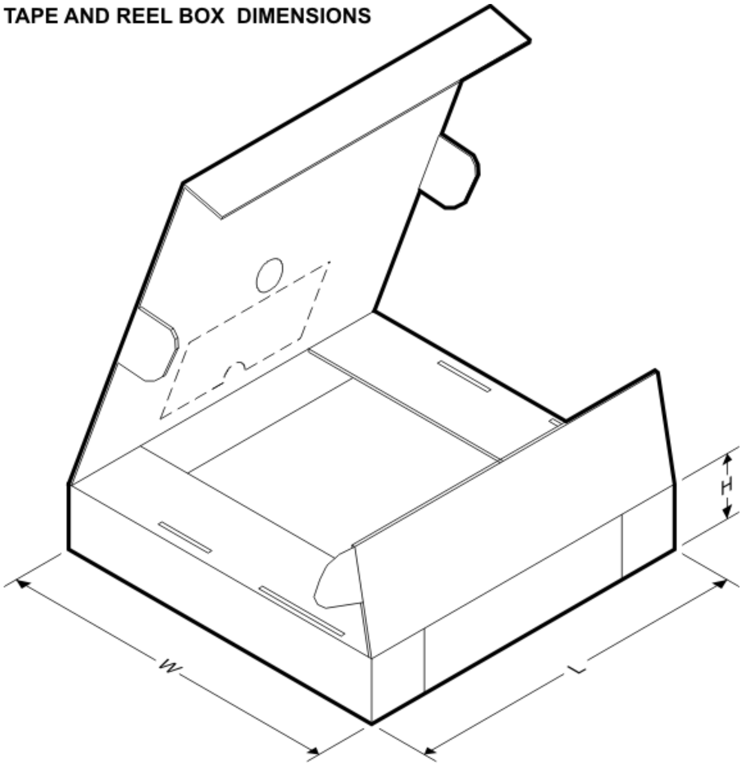
QUADRANT ASSIGNMENTS FOR PIN 1 ORIENTATION IN TAPE



*All dimensions are nominal

Device	Package Type	Package Drawing	Pins	SPQ	Reel Diameter (mm)	Reel Width W1 (mm)	A0 (mm)	B0 (mm)	K0 (mm)	P1 (mm)	W (mm)	Pin1 Quadrant
TPS53355DQPR	LSON-CLIP	DQP	22	2500	330.0	15.4	5.3	6.3	1.8	8.0	12.0	Q1
TPS53355DQPT	LSON-CLIP	DQP	22	250	330.0	15.4	5.3	6.3	1.8	8.0	12.0	Q1

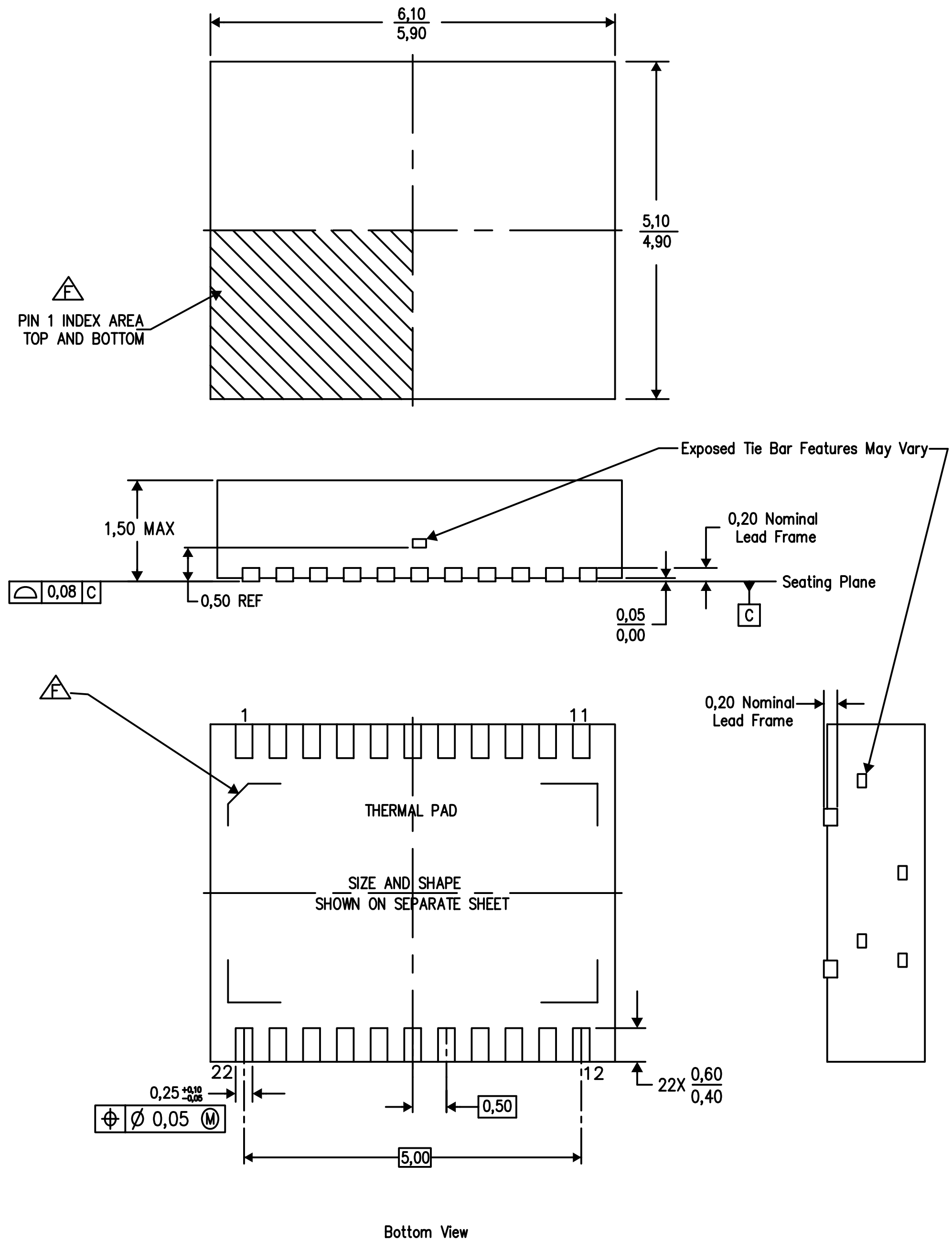
TAPE AND REEL BOX DIMENSIONS



*All dimensions are nominal

Device	Package Type	Package Drawing	Pins	SPQ	Length (mm)	Width (mm)	Height (mm)
TPS53355DQPR	LSON-CLIP	DQP	22	2500	336.6	336.6	41.3
TPS53355DQPT	LSON-CLIP	DQP	22	250	336.6	336.6	41.3

DQP (R–PSON–N22) PLASTIC SMALL OUTLINE NO–LEAD



4210472–3/E 09/11

NOTES:
- A. All linear dimensions are in millimeters. Dimensioning and tolerancing per ASME Y14.5M–1994.
- B. This drawing is subject to change without notice.
- C. Small Outline No–Lead (SON) package configuration.
- D. The package thermal pad must be soldered to the board for thermal and mechanical performance.
- E. See the additional figure in the Product Data Sheet for details regarding the exposed thermal pad features and dimensions.
- F. Pin 1 identifiers are located on both top and bottom of the package and within the zone indicated. The Pin 1 identifiers are either a molded, marked, or metal feature.

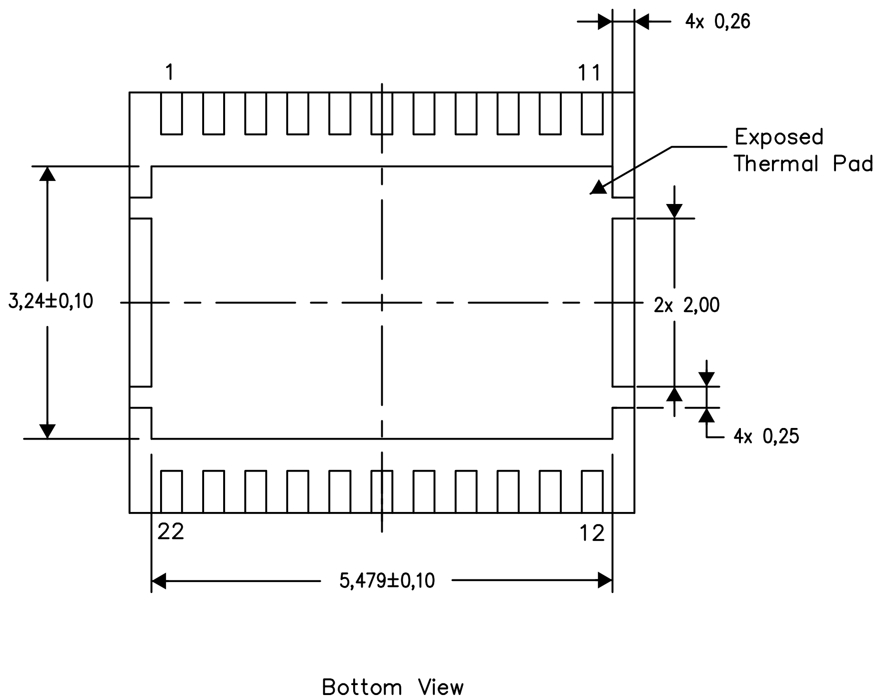
DQP (R–PSON–N22) PLASTIC SMALL OUTLINE NO–LEAD

THERMAL INFORMATION

This package incorporates an exposed thermal pad that is designed to be attached directly to an external heatsink. The thermal pad must be soldered directly to the printed circuit board (PCB). After soldering, the PCB can be used as a heatsink. In addition, through the use of thermal vias, the thermal pad can be attached directly to the appropriate copper plane shown in the electrical schematic for the device, or alternatively, can be attached to a special heatsink structure designed into the PCB. This design optimizes the heat transfer from the integrated circuit (IC).

For information on the Quad Flatpack No–Lead (QFN) package and its advantages, refer to Application Report, QFN/SON PCB Attachment, Texas Instruments Literature No. SLUA271. This document is available at www.ti.com.

The exposed thermal pad dimensions for this package are shown in the following illustration.

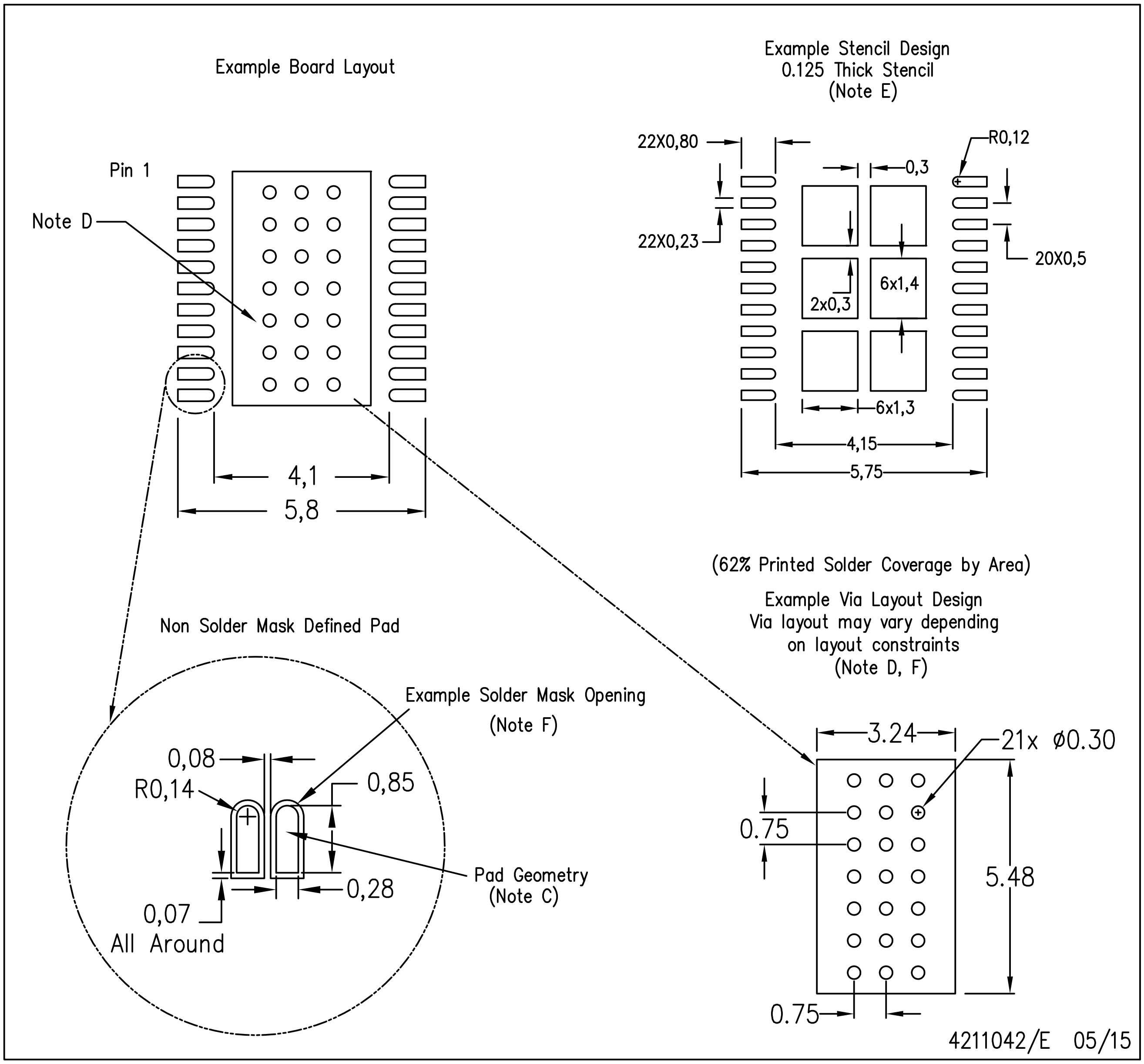


Exposed Thermal Pad Dimensions

4211024–3/H 08/15

NOTE: All linear dimensions are in millimeters

DQP (R-PSON-N22) PLASTIC SMALL OUTLINE NO-LEAD



NOTES:
A. All linear dimensions are in millimeters.
B. This drawing is subject to change without notice.
C. Publication IPC-7351 is recommended for alternate designs.
D. This package is designed to be soldered to a thermal pad on the board. Refer to Application Note, QFN/SON PCB Attachment, Texas Instruments Literature No. SLUA271, and also the Product Data Sheets for specific thermal information, via requirements, and recommended board layout. These documents are available at www.ti.com <http: //www.ti.com>.
E. Laser cutting apertures with trapezoidal walls and also rounding corners will offer better paste release. Customers should contact their board assembly site for stencil design recommendations. Refer to IPC 7525 for stencil design considerations.
F. Customers should contact their board fabrication site for recommended solder mask tolerances and via tenting recommendations for vias placed in the thermal pad.

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