TI Designs
150-W, 86% High-Efficiency Primary Side Regulated DCM/CCM Flyback Supply Reference Design

TI Designs

The TIDA-00367 is a 150-W flyback converter designed for powering low-voltage BLDC motor-based applications. This reference design is a primary side regulated (PSR) flyback converter based on the UCC28630 controller. It works in either discontinuous conduction mode (DCM) or continuous conduction mode (CCM), depending on the input voltage conditions, and delivers a tightly regulated 24-V output and a loosely regulated 12-V output. The hardware is designed and tested to pass the EN55011 Class B conducted emission test.

Design Resources

TIDA-00367	Design Folder
UCC28630	Product Folder
TLV1117	Product Folder

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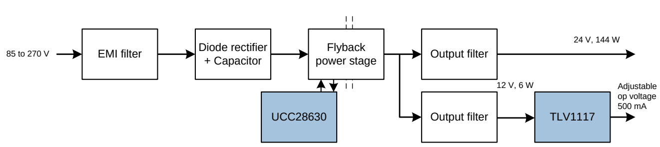
Design Features

- 150-W Flyback Designed for Operating in Wide Input Range From 85- to 270-V AC
- High Efficiency of 86% at Full Load
- Low Standby Power Consumption of 200 mW at 230-V AC
- Built-in Active X-Capacitor Discharge Function to Meet SELV Limits and Eliminates Using External Bleeder Resistor
- PSR in Both DCM and CCM Operation
- Robust Power Stage With Built-in Protection for Overcurrent, Overvoltage, Over-Temperature, AC Line Undervoltage, and Brownout
- Meets Requirements of Conducted Emissions Standard—EN55011 Class B
- Fully Tested and Validated to Deliver up to 150-W and Power a Low-Voltage BLDC Motor Using a Companion Motor Drive Board TIDA-00447

Featured Applications

- Dishwasher Pumps
- Appliances With 24-V DC Motor-Based Pumps or Fans



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1 Key System Specifications

Table 1. Key System Specifications

PARAMETER	SPECIFICATIONS
Input voltage range	85- to 270-V AC (230 V nominal)
Frequency	47 to 63 Hz
Output voltages	24-V DC and 12-V DC (center tapped winding)
Output power	144 W at 24-V DC and 6 W at 12-V DC
Efficiency	> 86% at 230 V
Protection	Output overcurrent
	Output overvoltage
	Output undervoltage
	Open-loop detection
Standby power	200 mW at 230-V AC
Line and load regulation	< ±2% at 24-V DC
Operating ambient	–10°C to 55°C
Board form factor, specs	135 × 93 mm; PCB Type: FR4, 2 layer
Conducted emissions	As per EN55011 Class B

2 System Description

Many of the home appliances uses low-power (less than 100 W) brushless motors (BLDC or PMSM) for driving pumps or fans as BLDC motors offer higher efficiency and quieter operation. For low-power brushless motors, using low-voltage motors instead of high-voltage motors brings in several advantages, such as:

- Low-voltage motors and drives can be directly used for either 110-V or 220-V line input operation using a universal input AC-DC converter.
- Low-voltage motor driver ICs typically offer higher level of integration:
  – Low R_{DS_ON} Power MOSFETs + MOSFET drivers, Current sense amplifier, programmable current regulation, overcurrent protection, over-temperature protection, auto restart, and so on.

In order to drive low-voltage motors, one needs a cost and efficiency optimized AC-DC power supply with a good output regulation. The objective of developing the TIDA-00367 reference design is to provide a low cost and robust isolated flyback power supply that delivers up to 150 W at 24 V. The TIDA-00367 can power the TI Design TIDA-00447, 24-V Dual Brushless DC Motor Pump Drives for Dishwashers.

The TIDA-00367 is a 24-V, 12-V output, 150-W isolated PSR flyback controller-based power supply, with built-in protections for overcurrent, overvoltage, feedback loop open/short detection, over-temperature, overload timer, AC line undervoltage, and brownout. The power supply is designed and tested for universal input AC range from 85- 270-V AC.

The EMI filter at the front end of the circuit is designed to meet EN55011 Class B conducted emission levels. The design achieves low standby power. Various parameters of the design like regulation, efficiency, EMI signature, output ripple, startup, and switching stresses are tested and documented.

PSR helps to eliminate using secondary side feedback and regulation components and optocoupler, enabling cost optimization, and increasing the reliability.

3 Block Diagram

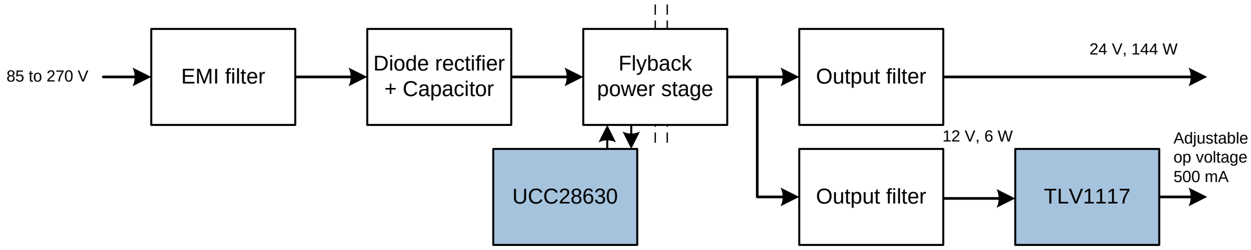

Figure 1. Block Diagram of 150-W PSR Flyback Power Supply

Figure 1 represents the block diagram of the 150-W PSR flyback power supply. The input AC mains first pass through an EMI filter and then the bridge rectifier section. The rectified DC input is fed to the flyback power stage. The flyback power stage consists of the main switching MOSFET, the flyback transformer, and the snubber. Apart from this, it also consists of the auxiliary circuit, which powers the UCC28630. The rectifier and filter stage at the output consists of the ultra-fast diodes for the 24-V and 12-V output and also the electrolytic capacitors and LC filter at the output.

3.1 Highlighted Products and Key Advantages

The following are the highlighted products used in this reference design. Key features for selecting the devices for this reference design are elucidated in the following subsections. Find the complete details of the highlighted devices in their respective product datasheets.

3.1.1 UCC28630 – Flyback Controller

The UCC28630 is a PSR flyback controller for high-power applications. With an ability to operate in both CCM and DCM, this controller is suitable for high-power flyback applications. The transformer auxiliary winding is used for sensing both the output and input voltages. Advanced sampling and processing techniques have been implemented to sense the output voltage so as to measure the output voltage accurately at a fixed instant after MOSFET turn off. This helps in avoiding the valley sensing that the typical PSR flyback controllers implement and hence gives the ability to operate in CCM. With a power full gate driving ability, it is able to keep the MOSFET switching losses low.

A few of the features influencing the choice of UCC28630 for this application are:

- Ability to operate in DCM and CCM while employing PSR
- Frequency dither to help meet EMI compliance specifications
- Built-in high-voltage startup
- Built-in active X-capacitor discharge to reduce the standby losses
- Overvoltage, overcurrent, over-temperature, overload timer, AC Line undervoltage, and brownout protections
- Peak-power mode for transient overload

The ability to operate in DCM and CCM offers the ability to increase the system efficiency at high load conditions. When operating in CCM, the RMS and peak currents through the switching MOSFET and the secondary diode are lower, which improve the efficiency at low-input voltage and high load condition. The minor drawback is an increase in the reverse recovery losses in the secondary diode.

The switching frequency of the flyback converter varies from 200 Hz to 120 kHz depending on the output power level with nominal operation at 60 kHz under full load conditions. In addition, the switching frequency is dithered by ±6.7%. Figure 2 shows the frequency dithering feature.

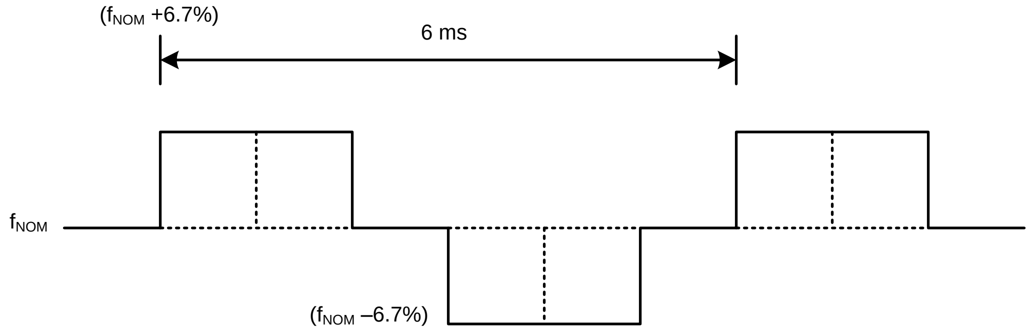

Figure 2. UCC28630 Switching Frequency Dither Feature

Since the switching frequency is dithered in a band of ±6.7% around the nominal switching frequency, the EMI spectrum becomes more broadband around the central switching frequency, thereby reducing the emission at the central frequency. This helps ease the EMI filter design to get the system to be compliant to any EMI standards.

The integrated high-voltage switched current source for startup allows fast startup during power on while at the same time reducing the static power consumption when compared to a conventional resistive startup, improving standby efficiency.

In the UCC28630, the startup resistors need to be connected on the AC side of the input bridge rectifier as the HV pin is also used for a feature called active X-capacitor discharge.

Safety standards such as EN60950 require the X-capacitors connected to the EMI filter on the AC side of the bridge rectifier quickly discharge to a safe voltage level when the AC mains is disconnected. The system should exhibit a one-second constant for discharging the input X-capacitors. Traditionally, this is achieved by having bleeder resistors in parallel to the X-capacitor. This leads to higher standby losses contributed by the steady power dissipation across the bleeder resistors. Another side effect because of this requirement is that the size of the X-capacitor will be limited, and a bigger common-mode or differential-mode inductor is required to meet the EMI specifications.

The active X-capacitor discharge function in the UCC28630 senses the AC mains disconnect condition quickly through the HV pin and, if the design requirements are met, activates the internal HV current source to discharge the input X-capacitors to SELV of 60 V within one second.

Apart from these, the UCC28630 implements a host of protections features for overvoltage, overcurrent, over-temperature, overload timer, AC line undervoltage, and brownout.

More details on all these features can be obtained from the UCC28630 datasheet (SLUSBW3).

4 System Design Theory

The reference design is a 150-W PSR flyback converter based on the UCC28630 IC. The design is specifically developed for powering low-voltage BLDC-based systems for home appliances. It operates under wide input voltage ranges from 85- to 270-V AC. It offers a cost effective and robust solution with an overall system efficiency of above 86% in the 230-V AC input range.

The UCC28630 flyback converter can operate both in DCM and CCM. The TIDA-00367 reference design exploits this feature to operate in either DCM or CCM depending on the input voltage and output load conditions, thereby improving the system efficiency at low input voltage and high load conditions.

Low EMI, high efficiency, and low cost are main focus of this design for targeted applications.

4.1 Flyback Converter Operating Mode

Flyback converters provide a cost effective solution for AC/DC conversion needs. They are widely used for AC/DC converters up to 150 W. There are three modes of operation namely DCM, critical conduction or Quasi-resonant mode (QRM), and CCM. For lower power applications the DCM or QRM is preferred as they have reduced power losses when applied in low power applications. As the output wattage increases, CCM becomes more efficient due to the reduced peak and RMS currents. In terms of feedback, flyback converters can be classified as secondary side regulated or primary side regulated. The secondary side regulated converters are the traditional converters, which use an optocoupler to feedback the error signal to the primary side for controlling the flyback converters working. The primary side regulated converters use the auxiliary winding to sense the output voltage, thereby eliminating the use of expensive optocouplers on the secondary side. Typically, the primary side regulated converters are used in DCM or QRM of operation.

The UCC28630 controller has a unique ability to do PSR even when operating in CCM, thereby providing the possibility to develop a PSR flyback design for higher power levels.

4.2 Flyback Circuit Component Design

The UCC28630 operates in variable frequency DCM and CCM and requires minimal external component for implementing a flyback power supply. It is digitally compensated and the entire compensation is implemented inside the IC, thereby requiring minimal external components. The design process and component selection for this design are illustrated in Section 4.2.1. The design calculator for the UCC28630 section as applicable to the TIDA-00367 is attached in the "TIDA-00367_Design_Calculator".

4.2.1 Design Goal Parameters

Table 2 elucidates the design goal parameters for this design. These parameters will be used in further calculations to select components.

Table 2. Design Goal Parameters

PARAMETER		MIN	TYP	MAX	UNIT
INPUT					
V_{IN}	Input voltage	85		270	VAC
f_{LINE}	Input frequency	47		63	Hz
OUTPUT					
V_{OUT1}	Output voltage		24		VDC
P_{OUT1}	Output power		144		W
	Line regulation		5%		
	Load regulation		5%		
V_{OUT2}	Output voltage		12		VDC
P_{OUT2}	Output Power		6		W
H	Targeted efficiency		96%		

4.2.2 Current Calculations

The input fuse, bridge rectifier, and input capacitor are selected based upon the input current calculations.

At low line and rated load, the system operates in CCM. This condition is taken for calculating the input RMS current.

Calculate the maximum duty cycle D(max) based on Equation 1.

$$D(max) = \left(V_{OUT} + V_D\right) \times \frac{N_{PS}}{V_{IN(min)} + \left(V_{OUT} + V_D\right) \times N_{PS}} \tag{1}$$

$$D(max) = (24.7) \times \frac{4.91}{75.27 + (24.7 \times 4.91)}$$

$$D(max) = 0.617$$

Calculate the primary ripple current at $V_{BULK(min)}$:

$$I_{RIPPLE} = \frac{V_{BULK(min)} \times D(max)}{L_{PRI} \times F_{SW}} \tag{2}$$

$$I_{RIPPLE} = 2.58\ A$$

The primary peak current can be calculated using Equation 3:

$$I_{PRI\_PK} = \left(\frac{P_{OUT}}{V_{BULK(min)} \times D(max) \times Efficiency} + \frac{I_{RIPPLE}}{2}\right) \tag{3}$$

$$I_{PRI\_PK} = 5.135\ A$$

The primary RMS current can now be calculated using Equation 4:

$$I_{PRI\_RMS} = \sqrt{\left(D(min) \times \left(\left(\frac{P_{OUT}}{V_{IN(min)} \times D(max)}\right)^2 + \frac{I_{RIPPLE}{}^2}{3}\right)\right)} \tag{4}$$

$$I_{PRI\_RMS} = 2.79\ A$$

The maximum secondary RMS current can be calculated for the 24-V output using Equation 5.

$$I_{SEC\_RMS24} = I_{OUT\_AVG24} \times \sqrt{\left(\frac{D(max)}{1 - D(max)}\right)} \tag{5}$$

$$I_{SEC\_RMS24} = 7.61\ A$$

4.2.3 Input Capacitor

A good starting point for calculating the value of the input capacitor is to take between 1.5 to 2 µF for every watt of input nominal power. In the TIDA-00376, the input capacitor is split into two, C7 and C8, and the total input capacitance is 300 µF. The input capacitor is split into two to reduce the equivalent ESR and hence reduce the differential noise generated from the flyback power stage.

4.2.4 Bridge Rectifier

The maximum input AC voltage is 270-V AC, so the DC voltage can reach voltage levels of up to 385-V DC. Considering a safety factor of 30%, it is recommended to select a component with a voltage rating greater than 500-V DC. To optimize the power loss due to diode forward voltage drop, a higher current bridge rectifier is recommended.

For this design, the 600-V, 4-A, diode GBU4K-E3/45 was chosen for input rectification.

4.2.5 Switching MOSFET

The major parameters considered for selecting a FET for this application were the R_{DS_ON}, Q_g, and $V_{DS(max)}$. Selecting a FET that has a good R_{DS_ON} is important to limit the conduction losses on the FET at low input voltages when the RMS current is high. The Q_g of the FET also needs to be low to keep the switching losses low as the FET is hard switched in low line conditions. A FET with at least 650-V $V_{DS(max)}$ rating must be chosen to allow a sufficient snubber voltage to optimize the snubber design for losses.

Keeping this criteria in mind, the IPP65R225C7XKSA1 was chosen with an R_{DS_ON} = 0.225 Ω and Q_g = 20 nC.

The losses on the switching MOSFET $P_{TOT} = P_{COND} + P_{SW}$.

The calculations are done during the worst case when the input voltage is 85-V AC.

$$P_{COND} = I_{PRI\_RMS}^2 \times R_{DS\_ON} \quad (6)$$

$$P_{COND} = 2.3\ \text{W}$$

The switching losses are estimated using the rise time, t_r, and fall time, t_f, of the MOSFET gate, and the output capacitance losses (C_{OSS}).

$$P_{SW} = F_{SW}\left(0.5 \times C_{OSS} \times (V_{DC(min)} + V_O)^2 + \left(\left(I_{PRI\_ON} + I_{PRI\_PK}\right) \times \left(V_{DC(min)} + n \times V_O\right) t_{transiton}\right)\right) \quad (7)$$

Here, the I_{PRION} is the primary current when then MOSFET is turned on. This is calculated from the value of the I_{PRI_PK} and I_{RIPPLE}.

$$I_{PR\_ION} = I_{PRI\_PK} - I_{RIPPLE} \quad (8)$$

$$I_{PRI\_ON} = 2.55\ \text{A}$$

The $t_{transition}$ represents the time taken for the MOSFET turn on. The MOSFET turn on losses due to the voltage and current crossover during the time the miller capacitor is charged. The duration can be calculated as approximately 20 ns. Substituting the values in Equation 6, the switching loss can be calculated as P_{SW} = 1.65 W.

The total power losses comes to P_{TOT} = 3.95 W. An appropriately sized heat sink is used for MOSFET.

4.2.6 Secondary Diodes

The secondary diodes are selected keeping the following four parameters in mind: breakdown voltage (V_{BR}), forward drop (V_D), reverse recovery charge (Q_{RR}), and average current (I_{AV}). The Q_{RR} is an important parameter as the system works in CCM at high load and low-input voltage conditions.

The maximum reverse voltage applied across the secondary diode in the 24-V output is given by

$$V_{BR1} = \left(\frac{V_{IN(max)} \times \sqrt{2}}{N_{PS}}\right) + 24\ \text{V} \quad (9)$$

$$V_{BR1} = 101.5\ \text{V}$$

Since the 12-V output is taken as a center tap from the secondary winding, the maximum reverse voltage across the diode in the 12-V output can be given as V_{BR2} = 50.75 V.

The average output current through the 24-V winding is 6 A and the average current through the 12-V winding is 0.5 A. Based on the specifications above the SBR20150CT and STPS1H100U are chosen for this application.

The diode losses are estimated for the 24-V output based upon the forward voltage drop, V_F, at 125°C and the reverse recovery charge, Q_{RR}, of the diode:

$$P_{DIODE} = V_F \times I_{OUT24} + 0.5 \times F_{SW} \times V_{OUT24} \times Q_{RR} \quad (10)$$

$$P_{DIODE} = 4.5\ \text{W}$$

An appropriately sized heat sink is used for boost the diodes. For the 12-V output, since the output current is very low in the actual usage scenario, no heat sink would be required.

4.2.7 Flyback Transformer Primary Inductance Calculation

The flyback transformer's primary inductance is calculated based on the required Boundary Conduction Mode (BCM) transition point. In the current application at full output power the BCM transition point is set when the voltage across the input bulk capacitor is at 230 V., The system operates in DCM above 230 V and the system operates in CCM below 230 V.

$$L_{PRI} = \frac{1}{2 \times \left(\frac{P_{OUT}}{Efficiency}\right) \times \left(\frac{1}{V_{BULK(DCM)}} + \left(\frac{1}{N_{PS} \times (V_{OUT} + V_D)}\right)\right)^2 \times F_{SW}} \tag{11}$$

Assuming a 85% efficiency and substituting the other values, the L_{PRI} comes to 294 µH. Choosing a value of L_{PRI} = 300 µH.

4.2.8 Sense Resistor

The value of the sense resistor R_{CS} is derived at the BCM transition point. Equation 12 is used to calculate the value of the R_{CS}.

$$R_{CS} = \frac{V_{CS(BCM)}}{\left(2 \times \frac{P_{OUT}}{Efficiency}\right) \times \left(\frac{1}{V_{BULK(DCM)}} + \frac{1}{(V_{OUT} + V_D) \times N_{PS}}\right)} \tag{12}$$

The term $V_{CS(BCM)}$ represents the peak current sense level at which the system enters BCM operation. This is equal to 640 mV.

Substituting this in Equation 12 along with the other values, the value of R_{CS} comes to 0.15 Ω.

To protect the ISENSE pin from inrush-surge current, a 1000-Ω resistor, R_{ISENSE}, is placed in series with the ISENSE pin. A 100-pF capacitor is placed close to the device to improve noise immunity on the ISENSE pin.

4.2.9 Output Capacitor

The selection of the output capacitor is based on the allowed voltage ripple $V_{OUTRIPPLE}$ and maximum ripple current I_{RIPPLE} that needs to be supported by them.

Assuming a 100-mV ripple on the 24-V output, the value of the minimum required capacitance is calculated as:

$$C_{OUT24} > \left(\frac{I_{OUT24} \times D(max)}{V_{RIPPLE24} \times F_{SW}}\right) \mu F \tag{13}$$

$$C_{OUT24} > 616\ \mu F$$

The ripple current requirement for the output capacitor can be given by Equation 14.

$$I_{OUT\_RIPPLE24} = \sqrt{I_{OUT\_RMS24}^2 - I_{OUT\_AVG24}^2} \tag{14}$$

$$I_{OUT\_RIPPLE24} = 4.68\ A$$

Two 1000-µF capacitors are used in parallel in the 24-V output to meet the ripple voltage and ripple current requirements. A 100-µF capacitor is used in the 12-V output to meet the ripple voltage requirement.

4.3 Flyback Snubber Design

The snubber design process for this application is described in this section. The snubber used in this application is a variation of a traditional RCD snubber with and additional Zener and capacitor. The reason for using this variation is also explained. The calculations for this snubber are done in the similar way to the traditional RCD snubber.

Figure 3 depicts the modified snubber.

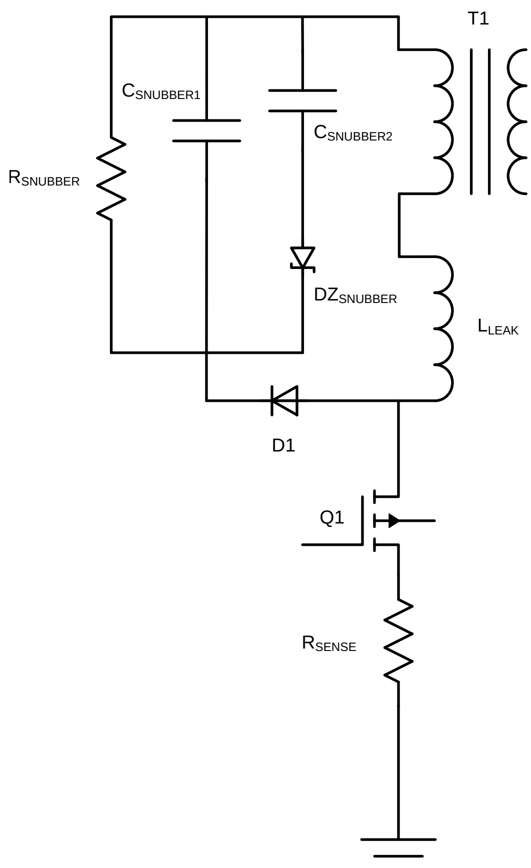

Figure 3. Modified RCD Snubber With Extra TVS Diode and Snubber Capacitor

Before beginning the calculation for the RCD snubber, the traditional RCD snubber schematic is shown in Figure 4.

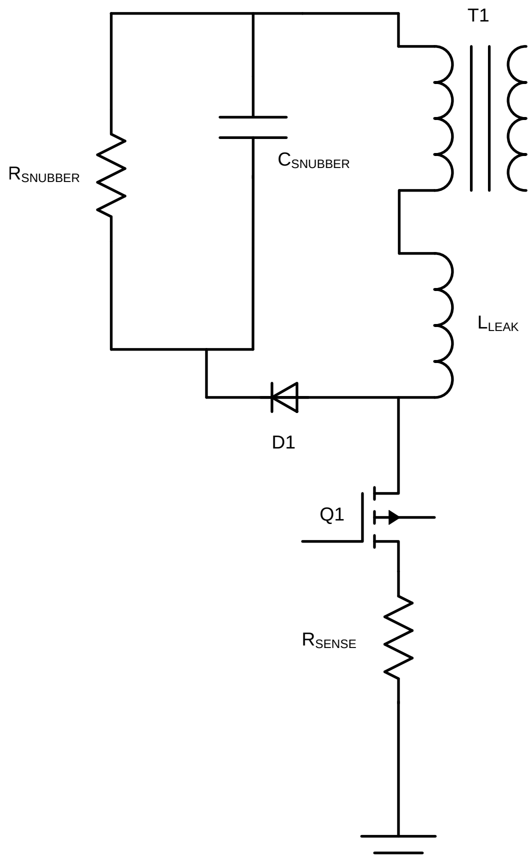

Figure 4. Traditional RCD Snubber

The snubber components of a typical RCD snubber is calculated at the maximum load under minimum input AC voltage conditions. Under these conditions, the peak current through the inductor tends to be at the maximum and this along with the leakage inductance of the flyback transformer is used to calculate the snubber components.

The leakage inductance L_{LEAK} of the flyback transformer T1 is 6 µH. The peak current has been theoretically calculated and experimentally verified to be at I_{PRI_PK} = 5.2 A.

The desired voltage across the snubber capacitor $C_{SNUBBER}$ is 220 V, with a 10% ripple. The power loss associated with the snubber can be calculated using Equation 15.

$$P_{SN} = 0.5 \times L_{LEAK} \times I_{PRI\_PK}{}^{2} \times \left(\frac{V_{SN}}{V_{SN} - nV_{O}} \right) \times F_{SW} \tag{15}$$

Substituting the values in Equation 15 results in P_{SN} = 10.7 W.

Because this entire power has to be dissipated in the snubber resistance $R_{SNUBBER}$, the value of $R_{SNUBBER}$ can be calculated from Equation 16.

$$R_{SNUBBER} = \left(\frac{V_{SN}{}^{2}}{P_{SN}} \right) \tag{16}$$

$$R_{SNUBBER} = 4.5K$$

The value of $R_{SNUBBER}$ is chosen to be slightly lower than the calculated value to increase the safety margin on the MOSFET. The actual chosen value for $R_{SNUBBER}$ is 4.2 K.

The value of the required snubber capacitance can be calculated from Equation 17.

$$C_{SNUBBER} = \frac{V_{SN}}{\Delta V_{SN} \times R_{SNUBBER} \times F_{SW}} \quad (17)$$

$$C_{SNUBBER} = 19.8 \text{ nF}$$

A higher value is chosen for $C_{SNUBBER}$ to bring down the ripple and increase safety margin. The actual value of the snubber capacitor is chosen as 22 nF.

As explained in the beginning of this section, the actual snubber used is a modification of the RCD snubber. The snubber capacitor is split into two, $C_{SNUBBER}$1 and $C_{SNUBBER}$2. A 4.7-nF capacitor is used as $C_{SNUBBER}$1 while a 22-nF capacitor is used as $C_{SNUBBER}$2. $C_{SNUBBER}$2 is in series with a TVS diode with a breakdown voltage of 120 V.

The UCC28630 samples the output voltage exactly after 1.7 µs of the DRV pin going low and switching MOSFET turning off. In order to get an accurate sampling of the actual output voltage, it is required to make sure that the leakage inductance is reset quickly and power transfer to the secondary starts before 1.7 µs off the switching MOSFET turning off. If a single high value of capacitance is used as the snubber capacitor, this condition can get violated when then output power is low. When the output power is low, the primary current and switching frequency are low. Under these conditions, if the snubber capacitance is high then the 1.7 µs timing constraint can get violated. This leads to instability in the operation of the flyback converter.

In order to avoid this condition, the snubber capacitance is split into two, and by adding a TVS in series with the 22-nF capacitor the operation of the snubber is slightly tweaked in such a way that under no-load and low output power conditions, the 4.7-nF capacitor mainly absorbs the leakage energy thereby quickly resetting the leakage energy and at higher load conditions both the 22-nF and 4.7-nF capacitors act in parallel, thereby providing a higher value of equivalent snubber capacitance to absorb the leakage energy.

An additional advantage of the modified RCD snubber is that it helps in increasing the flyback converter efficiency at low load conditions when compared to the conventional RCD snubber.

5 Getting Started Hardware

5.1 Test Conditions

For input, the power supply source (V_{IN}) must range from 85- to 270-V AC. Set the input current limit of input AC source to 2.5 A.

For output, use an electronic variable load or a variable resistive load, which must rated for ≥ 20 V and must vary the load current from 0 mA to 6.5 A.

5.2 Equipment Needed

- Isolated AC source
- Single-phase power analyzer
- Digital oscilloscope
- Multi-meters
- Electronic load or resistive load

5.3 Procedure

1. Connect input terminals (Pin-2 and Pin-3 of connector J1) of the reference board to the AC power source.
2. Connect output terminals (Pin-1 and Pin-2 of connector J2) to the electronic load, maintaining correct polarity.
3. Set and maintain a minimum load of about 10 mA.
4. Increase gradually the input voltage from 0 V to turn on voltage of 85-V AC.
5. Turn on the load to apply voltage to the output terminals of the flyback converter.
6. Observe the startup conditions for smooth switching waveforms.
7. Vary the input voltage gradually to 230-V AC to check the functional waveforms.

6 Test Results

The test results are divided in multiple sections that cover the steady state performance measurements, functional performance waveforms and test data, transient performance waveforms, thermal measurements, and conducted emission measurements.

6.1 Performance Data

6.1.1 Efficiency and Regulation With Load Variation

Table 3. 230-V AC Input, Load Variation From 10 to 150 W on 24-V Output

V_{INAC} (V)	P_{INAC} (W)	V_{OUT24} (V)	I_{OUT24} (A)	P_{OUT24} (W)	EFFICIENCY (%)
230	14.36	23.83	0.42	10	69.64
230	31.60	23.84	1.05	25	79.11
230	60.48	23.86	2.10	50	82.67
230	88.55	23.83	3.15	75	84.70
230	117.02	23.84	4.19	100	85.46
230	145.20	23.84	5.24	125	86.09
230	174.20	23.84	6.29	150	86.11

6.1.2 Efficiency and Regulation With Line Variation (AC Input)

Table 4. Input AC Voltage Varied From 85- to 270-V AC and Output Load Fixed at 150-W on 24-V Output

V_{INAC} (V)	P_{INAC} (W)	V_{OUT24} (V)	I_{OUT24} (A)	P_{OUT24} (W)	EFFICIENCY (%)
85	181.5	23.87	6.28	150	82.64
120	177.3	23.86	6.29	150	84.60
160	175.9	23.85	6.29	150	85.28
200	174.8	23.87	6.28	150	85.81
230	174.2	23.84	6.29	150	86.11
270	174.8	23.87	6.28	150	85.81

6.1.3 Standby Power

The standby power was noted at multiple AC input voltages with a 22K preload resistor across the 24-V output and a 10K preload resistor across the 12-V output.

Table 5. No Load Power Measurement With Output Voltage on 24-V Output

V_{INAC} (VAC)	P_{INAC} (mW)	V_{OUT24} (V)
85	55	23.76
120	100	23.76
160	120	23.84
200	150	23.92
230	170	24.02
270	220	24.33

6.1.4 Cross Regulation

The first test for cross regulation is conducted with 1-W load on the 24-V output and no load to full load variation on the 12-V output.

Table 6. Cross Regulation Results With Load Variation on 12-V Output

V_{OUT24} (V)	V_{OUT12} (V)	I_{OUT12} (mA)
23.72	10.24	10
23.77	9.92	50
23.77	9.71	100
23.85	9.50	200
23.94	9.32	300
23.98	9.20	400
24.13	9.10	500

The second test for cross regulation is conducted with no load on the 12-V output and no load to full load variation on the 24-V output.

Table 7. Cross Regulation Results With Load Variation on 24-V Output

V_{OUT24} (V)	V_{OUT12} (V)	P_{OUT24} (W)
24.21	10.8	0.2
23.79	11.24	10
23.84	12.53	50
23.86	13.39	100
23.89	13.6	150

6.2 Performance Curves

6.2.1 Efficiency Curves With Load and Line Variation

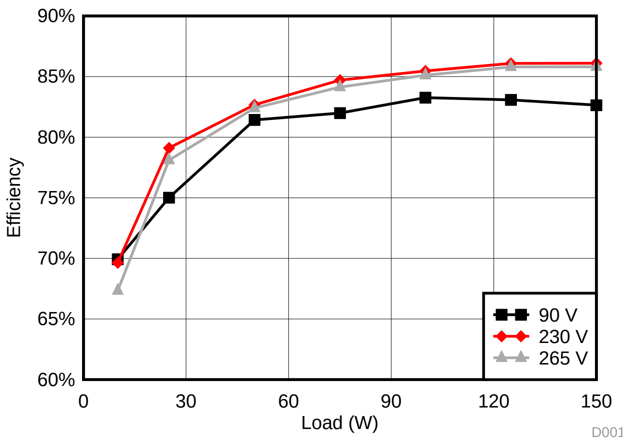

Figure 5. Efficiency versus AC Input Voltage (85- to 265-V AC)

6.2.2 Load and Line Regulation

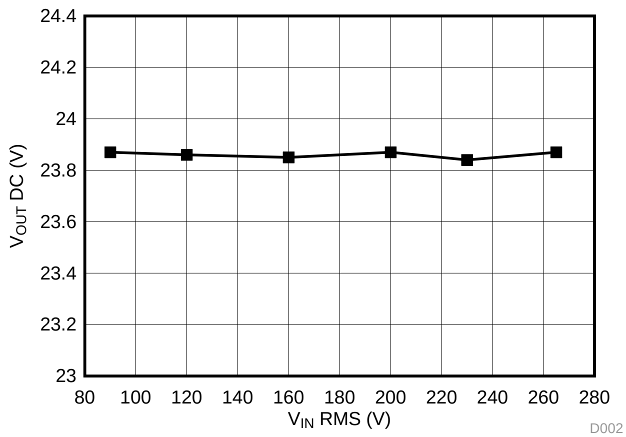

Figure 6. Output Voltage Variation With Input Voltage Variation at Full Load on 24-V Output

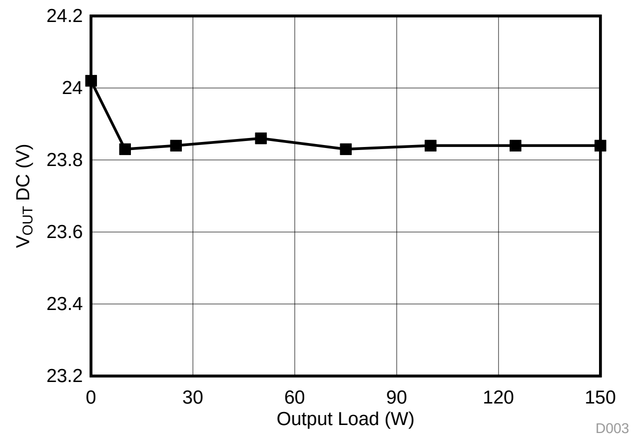

Figure 7. Output Voltage Variation With Output Load Variation at 230-V AC Input

6.2.3 Cross Regulation

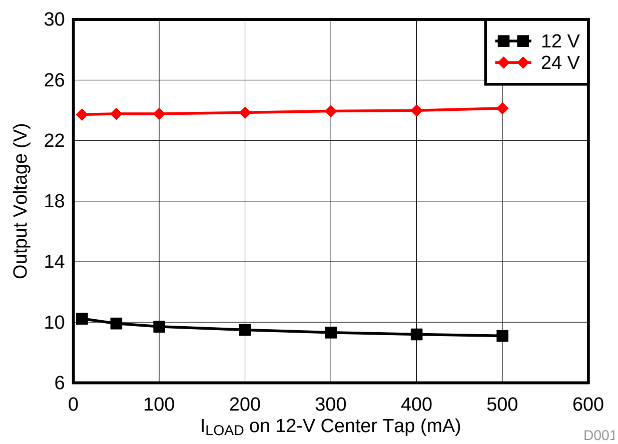

Figure 8. Cross Regulation Between 12- and 24-V Output With Load Variation on 12-V Output

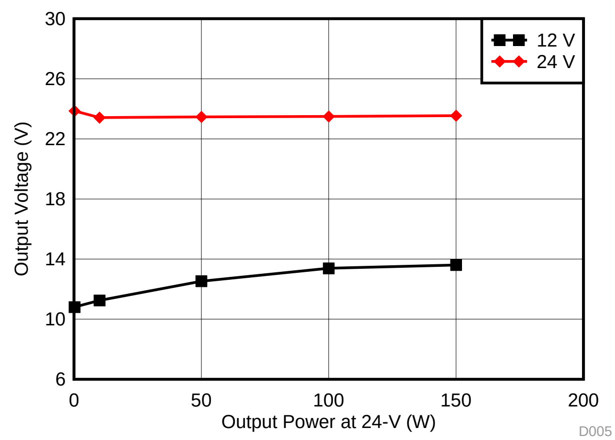

Figure 9. Cross Regulation Between 12- and 24-V Output With Load Variation on 24-V Output

6.3 Functional Waveforms

6.3.1 Switching Node Waveforms

The waveform at the MOSFET drain and the waveform across the current sense resistor were observed at full load under 85- and 270-V AC input voltage conditions.

NOTE: For Figure 10: Yellow trace: Drain voltage, 100 V/div; Blue trace: Current sense voltage, 500 mV/div

For Figure 11: Yellow trace: Drain voltage, 200 V/div; Blue trace: Current sense voltage, 500 mV/div

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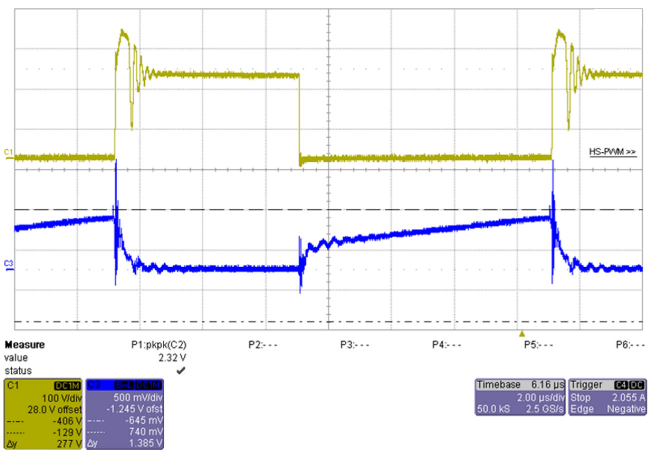


Figure 10. MOSFET Drain Voltage Waveform and Voltage Waveform Across Current Sense Resistor at V_{INAC} = 85-V AC, Full Load

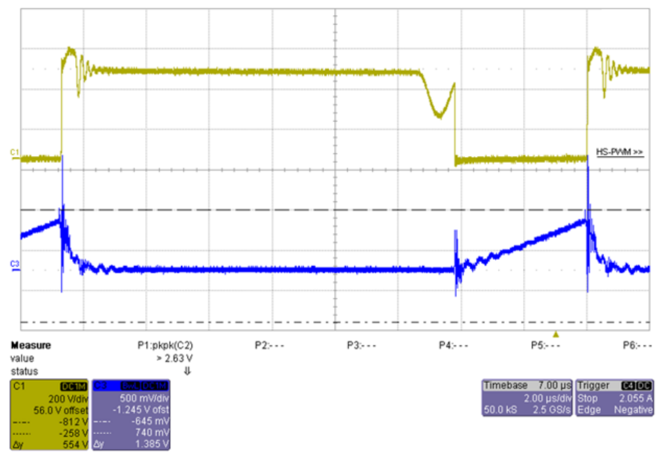


Figure 11. MOSFET Drain Voltage Waveform and Voltage Waveform Across Current Sense Resistor at V_{INAC} = 270-V AC, Full Load

From Figure 10 and Figure 11, the system operates in CCM under low-input AC voltage conditions and in DCM under high-input AC voltage conditions.

In Figure 12 and Figure 13, the voltage across the secondary rectifier diode across the 24-V output are shown. These are captured at full load under 85- and 270-V AC input conditions.

NOTE: For Figure 12: Blue trace: Voltage across the rectifier diode on the 24-V output, 20 V/div

For Figure 13: Blue trace: Voltage across the secondary rectifier diode on the 24-V output, 20 V/div

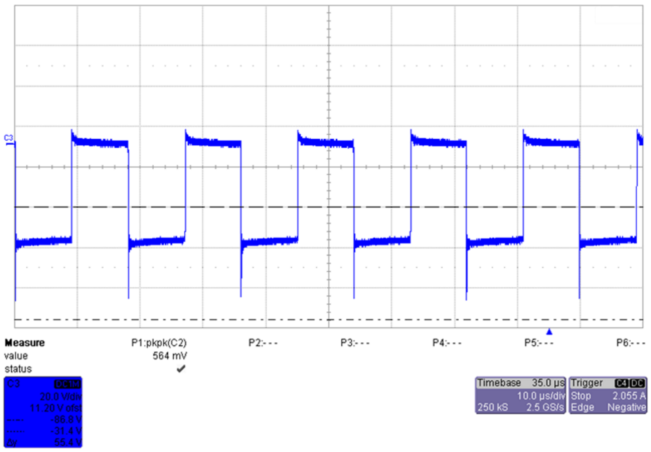


Figure 12. 24-V Output Secondary Rectifier Diode Voltage Waveform at V_{INAC} = 85-V AC, Full Load

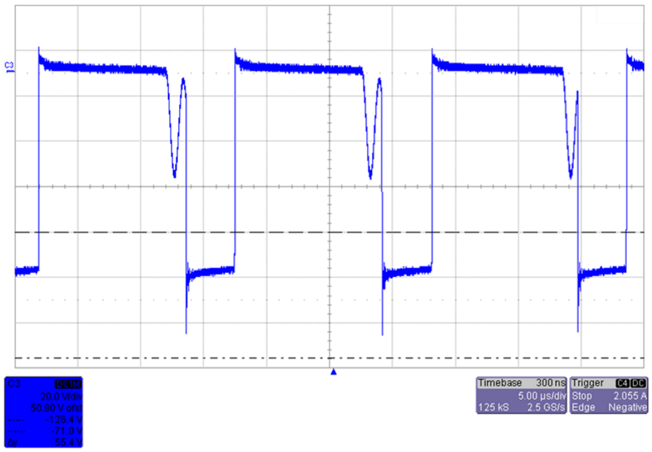


Figure 13. 24-V Output Secondary Rectifier Diode Voltage Waveform at V_{INAC} = 270-V AC, Full Load

6.3.2 Output Ripple

The voltage ripple across the 24-V output at no load and full load 85- and 230-V AC input conditions. A 50-Hz ripple corresponding to the input AC mains frequency can be seen at the output.

NOTE: Blue trace: Ripple at the 24-V output, 500 mV/div

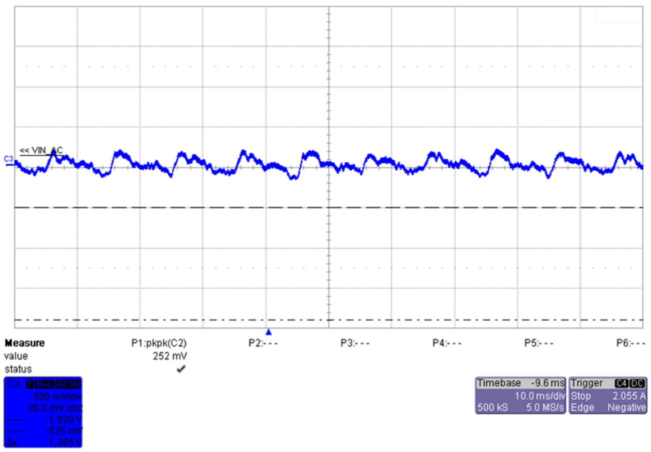


Figure 14. Output Voltage Ripple on 24-V Output at Input 85-V AC, Full Load, Only 50-Hz Component

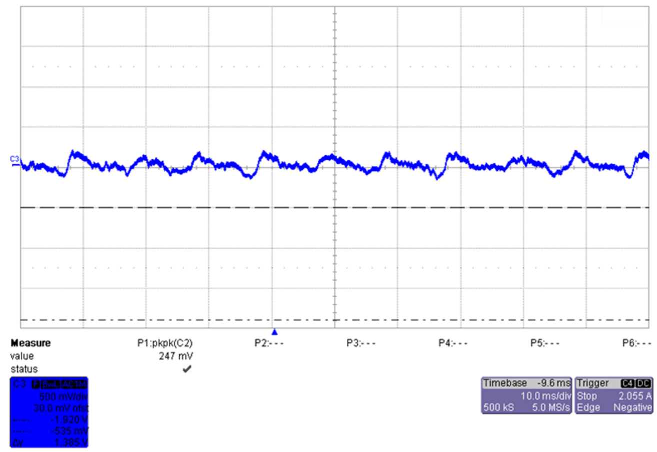


Figure 15. Output Voltage Ripple on 24-V Output at Input 230-V AC, Full Load, Only 50-Hz Component

6.3.3 Turn-On Characteristics

The turn-on characteristics of the 24-V output were recorded at no load and full load at 230-V AC input conditions.

NOTE: Blue trace: 24-V output, 5 V/div.

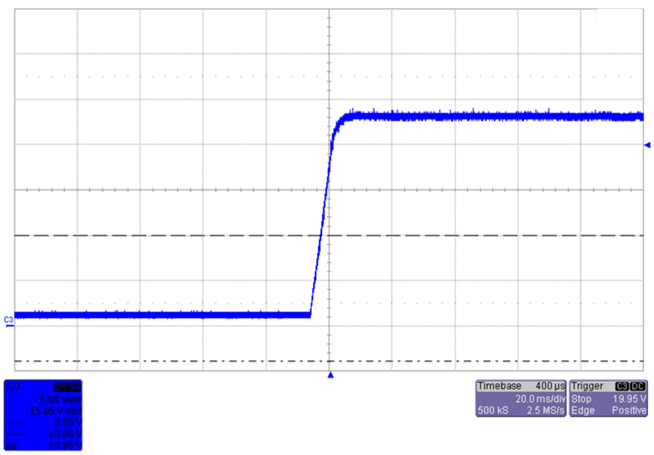


Figure 16. Output Turn ON Waveform at 230-V AC Input With No Load

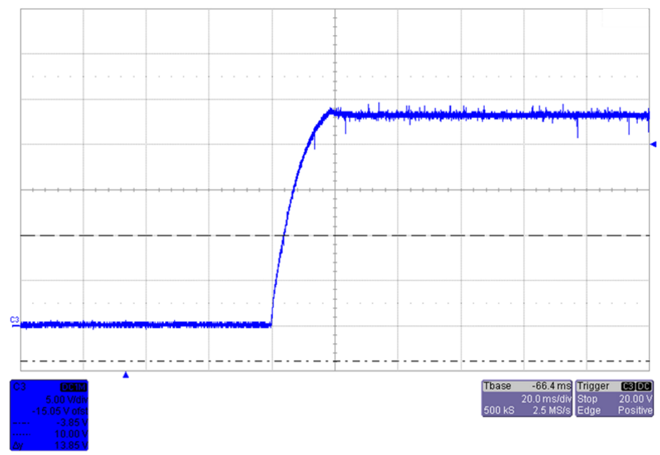


Figure 17. Output Turn ON Waveform at 230-V AC Input With Full Load

6.4 Transient Waveforms

6.4.1 Transient Load Response

Load transient performance is observed with load switched at 0.2-m wire length. The output load is switched using electronic load. The input voltage is at 85-V AC to simulate the rapid change of system operating from DCM to CCM and then back to DCM.

The load transient is applied on the 24-V output varying from 0.25 to 6.25 A and back to 0.25 A. The slope of the transient is set at 100 mA/µs.

NOTE: Red trace: Output voltage, 5 V/div; Yellow trace: Output current, 2 A/div.

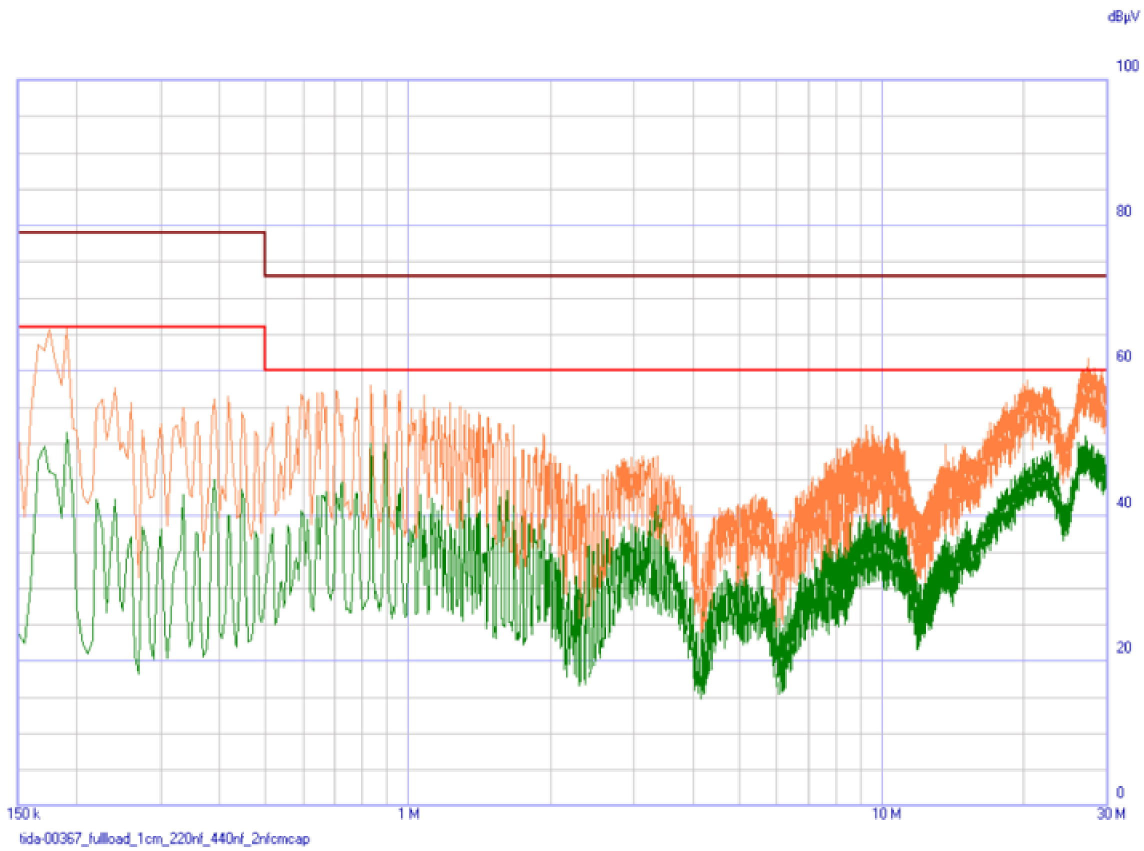
Figure 18. Output Voltage Waveform at 230-V AC Input Voltage, Load Transient From 0.25 to 6.25 A

6.5 Conducted Emissions

Generally, conducted emissions will be more at full load. Therefore, this operating point is chosen for measuring conducted EMI.

6.5.1 With Resistive Load at Output

The 230-V AC input, 2.31-A resistive load is connected to PSU with short leads. The conducted emissions in a pre-compliance test setup were compared against EN55011 class A and EN55011 class B limits and found to meet them comfortably. To meet the EN55011 class A limits, results are obtained with common-mode choke L1 unmounted. This result is shown in Figure 19.



	Start [MHz]	Stop [MHz]	Step	Detector	Hold Time	RBW	Min Att	Pre Amp	Pre Sel	Prompt start	Ancillary
1	0.15	30	AUTO (5 kHz)	P A 55022bav 55022bqp	10 ms	9 kHz	10	OFF	OFF	...	L1, L2

Pulse Limiter ON
Ancillary = L2 7010

Peak
Avg

Figure 19. Conducted Emissions as per EN55011 Class A

To meet the EN55011 class B limits, the common-mode choke L1 is also mounted. The result is shown in Figure 20.

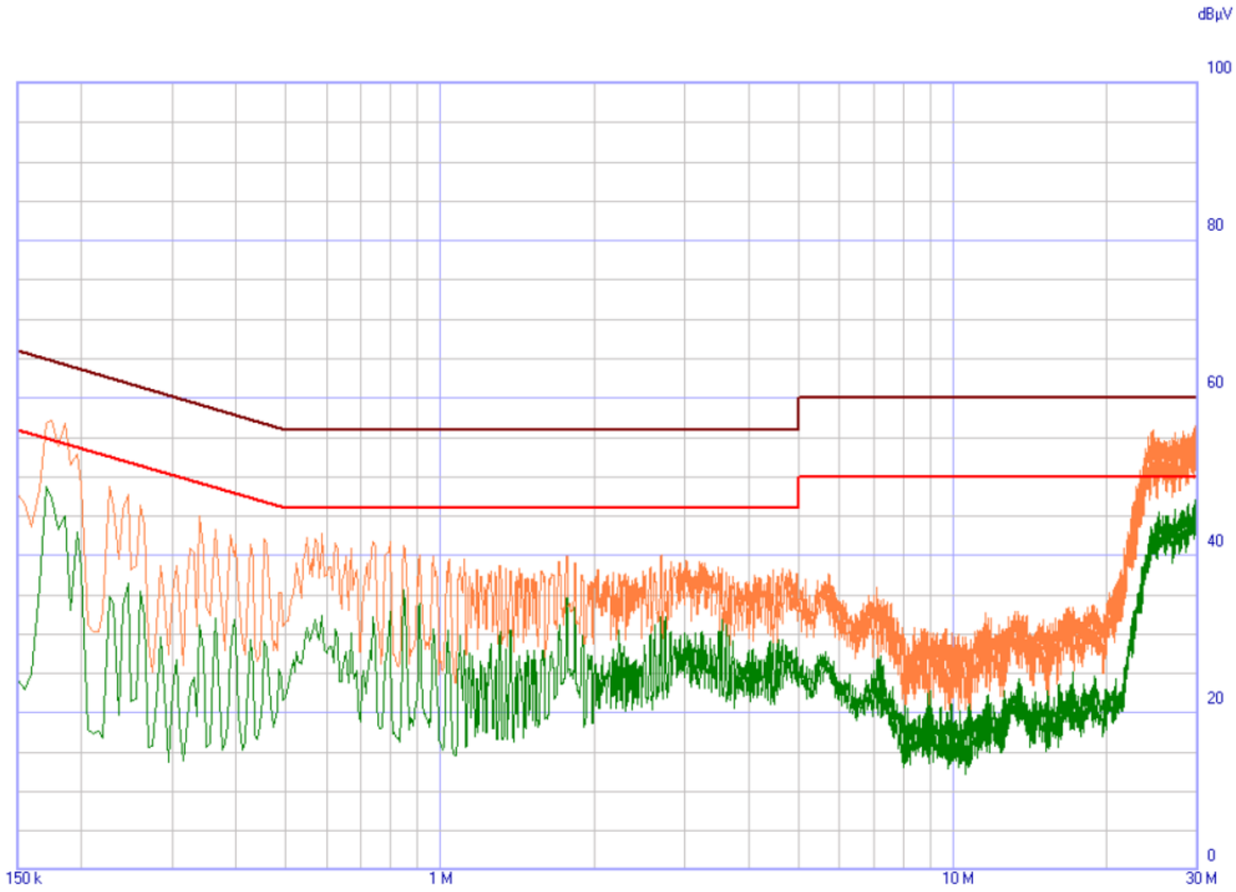


Figure 20. Conducted Emissions as per EN55011 Class B

6.6 Thermal Measurements

Thermal images are plotted at room temperature (25°C) within an enclosure, with no airflow, and at full load conditions. The board runs for 30 minutes before capturing the thermal image.

Figure 22 was taken after powering the board with a 230-V AC input and running it at 150-W load at the 24-V output.

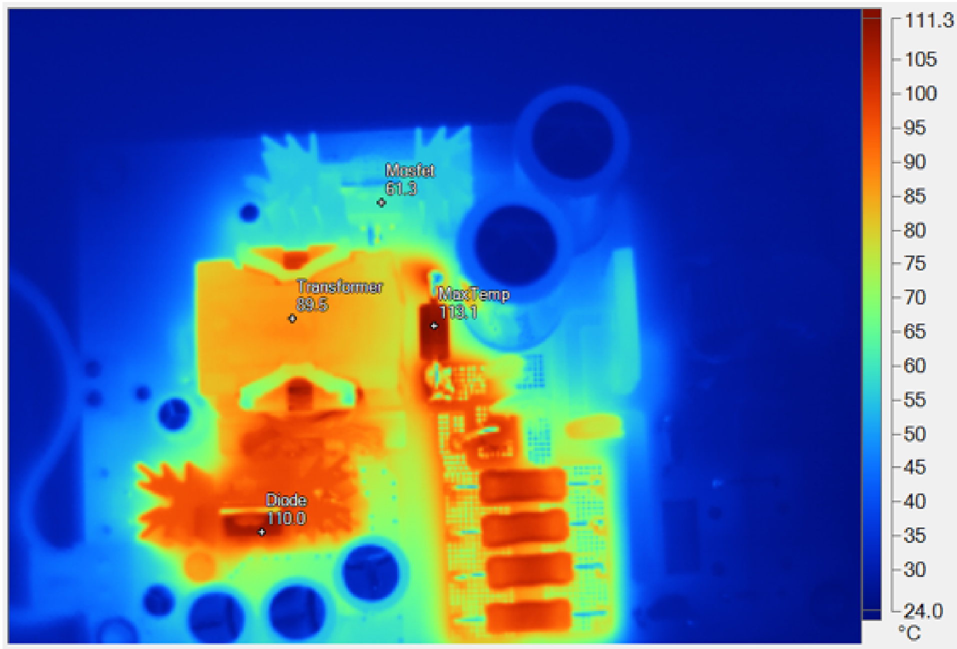


Figure 21. Thermal Image of Top Side of TIDA-00367 Board

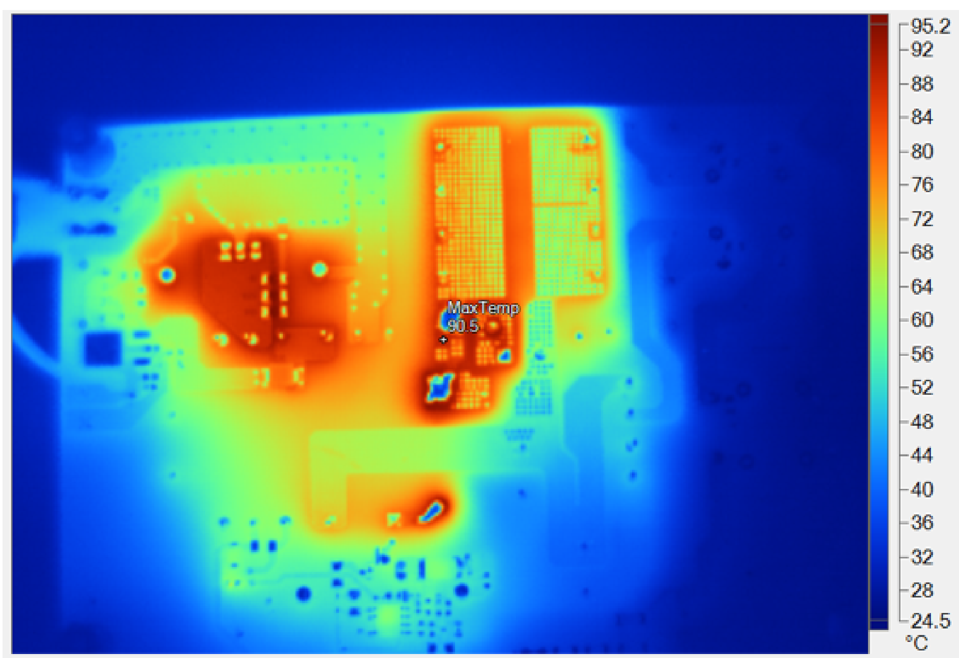


Figure 22. Thermal Image of Bottom Side of TIDA-00367 Board

7 Design Files

7.1 Schematics

To download the schematics, see the design files (TIDRIU0).

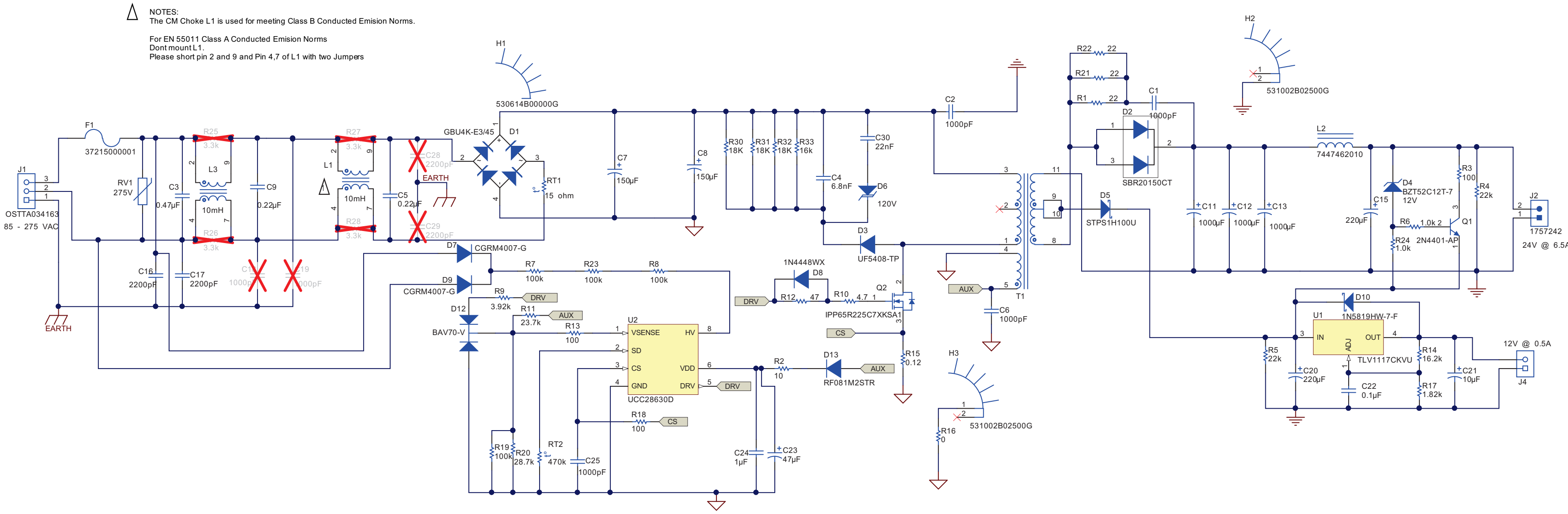


Figure 23. TIDA-00367 Schematics

7.2 Bill of Materials

To download the bill of materials (BOM), see the design files (TIDRIU1).

Table 8. TIDA-00367 BOM

QTY	REFERENCE	PART DESCRIPTION	MANUFACTURER	MANUFACTURER PARTNUMBER	NOTE
1	!PCB1	Printed Circuit Board	Any	TIDA-00367	Fitted
1	C1	CAP, CERM, 1000 pF, 250 V, +/- 10%, X7R, 0805	Taiyo Yuden	QMK212B7102KD-T	Fitted
1	C2	CAP, CERM, 1000 pF, 250 V, +/- 20%, E, Disc, 8x12mm	MuRata	DE1E3KX102MA5BA01	Fitted
1	C3	CAP FILM 0.47UF 20% 630VDC RAD	TDK	B32922C3474M	Fitted
1	C4	CAP CER 6800PF 1KV X7R RADIAL	MuRata	RDER73A682K3K1H03B	Fitted
1	C5	CAP, Film, 0.22 µF, 1000 V, +/- 5%, TH	TDK	B32653A224J	Fitted
1	C6	CAP, CERM, 1000 pF, 1000 V, +/- 5%, C0G/NP0, 1206	Vishay-Vitramon	VJ1206A102JXGAT5Z	Fitted
2	C7, C8	CAP, AL, 150 µF, 400 V, +/- 20%, TH	Chemi-Con	450BXW150MEFC18X45	Fitted
1	C9	CAP, Film, 0.22 µF, 630 V, +/- 20%, TH	TDK	B32922C3224M	Fitted
3	C11, C12, C13	CAP, AL, 1000 µF, 35 V, +/- 20%, 0.018 ohm, TH	Panasonic	EEU-FR1V102B	Fitted
2	C15, C20	CAP, AL, 220 µF, 35 V, +/- 20%, 0.087 ohm, TH	Chemi-Con	EKY-350ELL221MH15D	Fitted
2	C16, C17	CAP, CERM, 2200 pF, 250 V, +/- 20%, E, Cap, 7x12x9 mm	MuRata	DE1E3KX222MA4BL01	Fitted
1	C21	CAP, AL, 10 µF, 16 V, +/- 20%, ohm, TH	Panasonic	ECE-A1CKS100	Fitted
1	C22	CAP, CERM, 0.1uF, 50V, +/-10%, X7R, 0603	AVX	06035C104KAT2A	Fitted
1	C23	CAP, AL, 47 µF, 35 V, +/- 20%, 0.4 ohm, TH	Rubycon	35ZLJ47MTA5X11	Fitted
1	C24	CAP, CERM, 1 µF, 50 V, +/- 10%, X7R, 0805	AVX	08055C105KAT2A	Fitted
1	C25	CAP, CERM, 1000 pF, 50 V, +/- 10%, X7R, 0805	AVX	08055C102KAT2A	Fitted
1	C30	CAP CER 0.022UF 1KV X7R RADIAL	MuRata	RDER73A223K3M1H03A	Fitted
1	D1	Diode, Switching-Bridge, 800 V, 4 A, TH	Vishay-Semiconductor	GBU4K-E3/45	Fitted
1	D2	Diode, Super Barrier Rectifier, 60 V, 10 A, TO-220AB	Diodes Inc.	SBR20150CT	Fitted
1	D3	DIODE GEN PURP 1KV 3A DO201AD	Vishay-Semiconductor	UF5408-TP	Fitted
1	D4	Diode, Zener, 12 V, 300 mW, SOD-523	Diodes Inc.	BZT52C12T-7	Fitted
1	D5	Diode, Schottky, 100 V, 1 A, SMB	Diodes Inc.	STPS1H100U	Fitted
1	D6	Diode, TVS, Uni, 120 V, 1500 W, SMC	Littelfuse	1.5KE120A-E3/54	Fitted
2	D7, D9	Diode, P-N, 1000 V, 1 A, 3.9x1.7x1.8mm	Comchip Technology	CGRM4007-G	Fitted
1	D8	Diode, Switching, 75 V, 0.25 A, SOD-323	Micro Commercial Components	1N4448WX	Fitted
1	D10	Diode, Schottky, 40 V, 1 A, SOD-123	Diodes Inc.	1N5819HW-7-F	Fitted
1	D12	Diode, Switching, 70 V, 0.25 A, SOT-23	Vishay-Semiconductor	BAV70-V	Fitted
1	D13	Diode, Fast Rectifier, 200 V, 0.8 A, SOD-123	Rohm	RF081M2STR	Fitted
1	F1	Fuse, 5 A, 250 V, TH	Littelfuse	37215000001	Fitted

Table 8. TIDA-00367 BOM (continued)

QTY	REFERENCE	PART DESCRIPTION	MANUFACTURER	MANUFACTURER PARTNUMBER	NOTE
1	H1	Heat Sink, TO-220, TH	Aavid	530614B00000G	Fitted
2	H2, H3	HEATSINK TO-220 W/PINS 1" TALL	Aavid	531002B02500G	Fitted
4	H4, H5, H6, H7	Machine Screw, Round, #4-40 x 1/4, Nylon, Philips panhead	B&F Fastener Supply	NY PMS 440 0025 PH	Fitted
4	H8, H9, H10, H11	Standoff, Hex, 0.5"L #4-40 Nylon	Keystone	1902C	Fitted
1	J1	Terminal Block, 3x1, 5.08 mm, TH	On-Shore Technology	OSTTA034163	Fitted
1	J2	Header (Shrouded), 5.08 mm, 2x1, Tin, R/A, TH	Phoenix Contact	1757242	Fitted
1	J4	Terminal Block, 2-pole, 200mil, TH	On-Shore Technology	OSTTC022162	Fitted
1	L3	Coupled inductor, 10 mH, 3 A, 0.040 ohm, +/- 30%, TH	Abracon	ALFT-04-9	Fitted
1	L2	Inductor, Unshielded Drum Core, Ferrite, 1 µH, 8 A, 0.006 ohm, TH	Wurth Elektronik	7447462010	Fitted
1	Q1	Transistor, NPN, 40 V, 0.5 A, TO-92AP	Micro Commercial Components	2N4401-AP	Fitted
1	Q2	MOSFET, N-CH, 650 V, 16 A, TO-220AB	Infineon Technologies	IPP65R225C7XKSA1	Fitted
3	R1, R21, R22	RES, 22, 5%, 0.25 W, 1206	Vishay-Dale	CRCW120622R0JNEA	Fitted
1	R2	RES, 10, 5%, 0.1 W, 0603	Vishay-Dale	CRCW060310R0JNEA	Fitted
1	R3	RES, 100, 1%, 0.25 W, TH	Ohmite	OF101JE	Fitted
2	R4, R5	RES, 22 k, 5%, 0.1 W, 0603	Vishay-Dale	CRCW060322K0JNEA	Fitted
2	R6, R24	RES, 1.0 k, 5%, 0.125 W, 0805	Vishay-Dale	CRCW08051K00JNEA	Fitted
3	R7, R8, R23	RES, 100 k, 5%, 0.25 W, 1206	Vishay-Dale	CRCW1206100KJNEA	Fitted
1	R9	RES, 3.92 k, 1%, 0.125 W, 0805	Panasonic	ERJ-6ENF3921V	Fitted
1	R10	RES, 4.7, 5%, 0.125 W, 0805	Panasonic	ERJ-6GEYJ4R7V	Fitted
1	R11	RES, 23.7 k, 1%, 0.063 W, 0402	Vishay-Dale	CRCW040223K7FKED	Fitted
1	R12	RES, 47, 5%, 0.125 W, 0805	Panasonic	ERJ-6GEYJ470V	Fitted
2	R13, R18	RES, 100, 1%, 0.125 W, 0805	Vishay-Dale	CRCW0805100RFKEA	Fitted
1	R14	RES, 16.2 k, 1%, 0.063 W, 0402	Vishay-Dale	CRCW040216K2FKED	Fitted
1	R15	RES, 0.12, 1%, 2 W, 2512	TE Connectivity	2-2176057-7	Fitted
1	R16	RES, 0, 5%, 0.1 W, 0603	Vishay-Dale	CRCW06030000Z0EA	Fitted
1	R17	RES, 1.82 k, 1%, 0.063 W, 0402	Vishay-Dale	CRCW04021K82FKED	Fitted
1	R19	RES, 100 k, 0.1%, 0.063 W, 0603	TE Connectivity	CPF0603B100KE	Fitted
1	R20	RES, 28.7 k, 1%, 0.063 W, 0402	Vishay-Dale	CRCW040228K7FKED	Fitted
3	R30, R31, R32	RES, 18 k, 5%, 3W, TH	Vishay-Dale	ERG-3SJ183	Fitted
1	R33	RES, 16k, 5%, 3W, TH	Vishay-Dale	ERG-3SJ163	Fitted
1	RT1	Thermistor NTC, 15 ohm, 20%, 8.5mm Disc	EPCOS Inc	B57153S150M	Fitted
1	RT2	Thermistor NTC, 470k, 5%, Disc, 5.5x5mm	TDK	B57164K474J	Fitted

Table 8. TIDA-00367 BOM (continued)

QTY	REFERENCE	PART DESCRIPTION	MANUFACTURER	MANUFACTURER PARTNUMBER	NOTE
1	RV1	Varistor, 430V, 4.5KA, TH	EPCOS Inc	B72214S0271K101	Fitted
1	T1	Transformer, 300 uH, TH	Wurth Elektronik	750342946	Fitted
1	U1	ADJUSTABLE AND FIXED LDO VOLTAGE REGULATOR, KVU0003A	Texas Instruments	TLV1117CKVU	Fitted
1	U2	High-Power Flyback Controller with Primary-Side Regulation and Peak-Power Mode, D0007A	Texas Instruments	UCC28630D	Fitted
0	C18, C19	CAP, CERM, 1000 pF, 250 V, +/- 20%, E, Disc, 8x12mm	MuRata	DE1E3KX102MA5BA01	Not Fitted
0	C28, C29	CAP, CERM, 2200 pF, 250 V, +/- 20%, E, Cap, 7x12x9 mm	MuRata	DE1E3KX222MA4BL01	Not Fitted
0	FID1, FID2, FID3	Fiducial mark. There is nothing to buy or mount.	N/A	N/A	Not Fitted
0	R25, R26, R27, R28	RES, 3.3 k, 5%, 0.125 W, 0805	Vishay-Dale	CRCW08053K30JNEA	Not Fitted
0	L1	CHOKE TOROID 3.3MH 4A VERTICAL	Wurth Elektronik	744824433	Fitted only for Class B

7.3 Layout Guidelines

A careful PCB layout is critical and extremely important in a high-current fast-switching circuit to provide appropriate device operation and design robustness. As with all switching power supplies, attention to detail in the layout can save much time in troubleshooting later on.

7.3.1 Power Stage Specific Guidelines

The following are key guidelines for routing power stage components:

- Minimize the loop area and trace length of the power path circuits, which contain high frequency switching currents. This will help in reducing EMI and improve converter overall performance.
- Keep the switch node as short as possible. A short and optimal trace width helps to reduce induced ringing caused by parasitic inductance.
- The snubber capacitors and TVS diode should be close to the switching MOSFET and secondary rectifier diode.
- Keep traces with high dV/dt potential and high di/dt capability away from or shielded from sensitive signal traces with adequate clearance and ground shielding.
- When multiple capacitors are used in parallel for current sharing, layout should be symmetrical across both leads of the capacitors. If the layout is not identical, the capacitor with the lower series trace impedance will see higher peak currents and become hotter (i^2R).
- Tie the heat-sinks of all the power switching components to their respective power grounds.
- Place protection devices such as TVS, snubbers, capacitors, or diodes physically close to the device they are intended to protect, and route with short traces to reduce inductance.
- Choose the width of PCB traces based on acceptable temperature rise at the rated current as per IPC2152 as well as acceptable DC and AC impedances. Also, the traces should withstand the fault currents (such as short circuit current) before the activation of electronic protection such as a fuse or circuit breaker.
- Determine the distances between various traces of the circuit according to the requirements of applicable standards. For this design, UL 60950-1 safety standard is followed to maintain the creepage and clearance from live line to neutral line and to safety ground, as defined in the Tables 2K through 2N of this standard.
- Adapt thermal management to fit the end-equipment requirements

7.3.2 Controller Specific Guidelines

The following are key guidelines for routing of controller components and signal circuits:

- The optimum placement of decoupling capacitor is closest to the VCC and GND terminals of the device. Minimize the loop area formed by the bypass-capacitor connection and the GND terminal of the IC.
- Use a copper plane or island as the reference ground for the control devices.
- Make the trace routing for the current sensing circuit components to the device as short as possible to reduce parasitic effects on the current limit and current monitoring accuracy. These traces should not have any coupling to switching signals on the board.
- Resistors R11, R13, R19, and R20 form the voltage sensing network and need to be placed near the IC VSENSE pin. No decoupling capacitor should be placed at this pin.
- The DRV trace should be short as this is a switching signal.

7.3.2.1 Layout Prints

To download the layout prints, see the design files (TIDRIU3).

7.4 Altium Project

To download the Altium project files, see the design files at TIDA-00367.

7.5 Gerber Files

To download the Gerber files, see the design files at TIDA-00367.

7.6 Assembly Drawings

To download the assembly drawings, see the design files at TIDA-00367.

7.7 Design Calculator Spreadsheet

To download the design calculator spreadsheet, see the design files at TIDA-00367.

8 References

1. Texas Instruments, *UCC28630, UCC28631, UCC28632 and UCC28633, High-Power Flyback Controller with Primary-Side Regulation and Peak-Power Mode*, UCC28630 Datasheet (SLUSBW3)
2. Texas Instruments, *PMP10912 Rev A Test Results* (TIDUBC3)
3. Texas Instruments, UCC28630 Excel Design Calculator (SLUC537)

9 Terminology

The TI Glossary lists and explains terms, acronyms, and definitions (SLYZ022).

10 About the Author

RAMKUMAR S is a Systems Engineer at Texas Instruments, where he is responsible for developing reference design solutions for the industrial segment. Ramkumar brings to this role his diverse experience in analog and digital power supplies design. Ramkumar earned his master of technology (M.Tech) from the Indian Institute of Technology, Delhi.

Revision History

NOTE: Page numbers for previous revisions may differ from page numbers in the current version.

Changes from Original (December 2015) to B Revision	Page
• Changed title from *150-W High-Efficiency Primary Side Regulated DCM/CCM Flyback Supply for Low-Voltage Motor Drives* to *150-W, 86% High-Efficiency Primary Side Regulated DCM/CCM Flyback Supply Reference Design*	1

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Mailing Address: Texas Instruments, Post Office Box 655303, Dallas, Texas 75265