

Low Cost 1W Isolated Power Supply Solution with TPS61085

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ABSTRACT

This paper introduces a low cost 1 to approximately 2W isolated power supply solution with TPS61085. It is a useful solution in industry applications which need a low power isolated supply, such as the RS485 interface.

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1 Introduction

For industry applications, such as RS485 and isolated CAN transceivers, a mini power isolated power supply is required. The isolated flyback topology is suitable for these applications, due to its simple implementation and low cost.

It is possible to use a non-synchronous low voltage boost converter to build an isolated flyback power supply. This paper chooses TPS61085 as an example, which has an integrated 2A, 20V MOSFET, as shown in [Figure 1](#). The input voltage range of TPS61085 is 2.3V to approximately 6V. With the isolated solution with TPS61085, it is possible to get 5V with an isolation voltage of 2.5kV or 4kV (the isolation voltage is determined by the isolation transformer) from a 3.3V or 5V system.

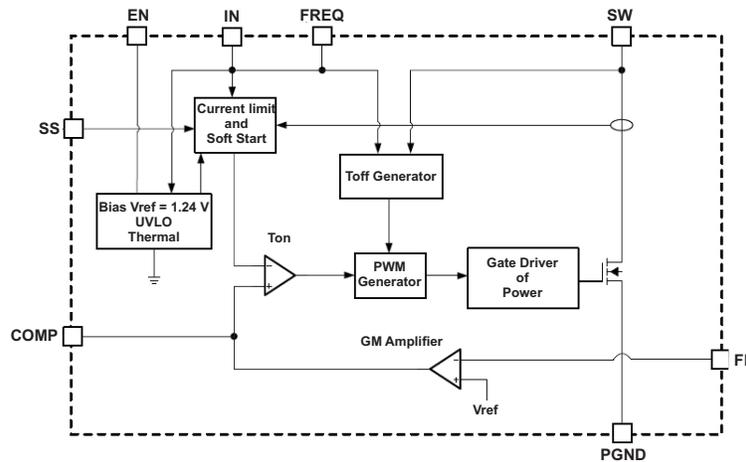


Figure 1. Internal Block Diagram of TPS61085

2 Flyback Circuit Design with TPS61085

2.1 Power Specification

This paper gives a design example for an isolated power supply for a RS485 interface with TPS61085.

The specification in detail for the isolated RS485 application follows:

- Input voltage: 4.5V to approximately 5.5V
- Output voltage: 4.5V to approximately 5.5V (The voltage range is recommended in SN65HVD3082's data sheet, which is the RS-485 transceiver)
- Output current: 200mA.

2.2 Reference Schematics

Figure 2 is the schematics for the 5V -> 5V isolated power supply.

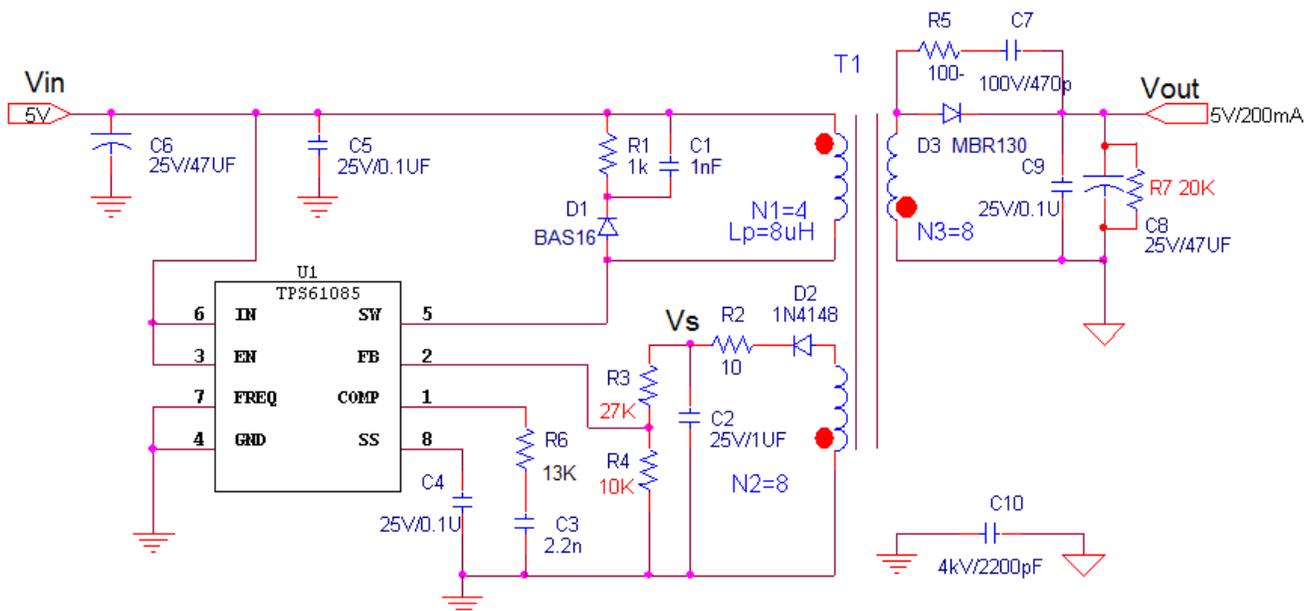


Figure 2. Schematic for 5V Isolated Power Supply with TPS61085

2.3 Frequency Selection

The TPS61085's frequency select pin, **FREQ**, allows to set the switching frequency of the device to 650 kHz (**FREQ** = low) or 1.2 MHz (**FREQ** = high). For this isolated power supply solution, the low frequency is preferred because higher switching frequencies cause more switching losses, EMI issues, and so on.

2.4 Transformer Design

There are two ways to design the transformer for the flyback converter. One is using the maximum duty cycle for the PWM controller; another is using the reflected voltage from the secondary-side.

TPS61085 has an integrated 20V MOSFET, thus the maximum drain-source voltage, V_{DS} , of the MOSFET is fixed and; therefore, the second method is used in this paper. Using 70% derating for the internal MOSFET, approximate the maximum allowed reflected secondary-side voltage of the MOSFET using [Equation 1](#):

$$V_{\text{reflect}} = 0.7 \times V_{DS} - V_L - V_{IN(\text{max})} = 0.7 \times 20 \text{ V} - 5.5 \text{ V} - 5.5 \text{ V} = 3 \text{ V} \quad (1)$$

where V_{DS} is the required drain to source voltage rating of the internal MOSFET. V_L is the voltage spike due to the leakage inductance of the transformer, estimated to be equal to $V_{IN(\text{max})}$.

The ratio of the transformer can be calculated by using [Equation 2](#):

$$N = \frac{N_P}{N_S} = \frac{V_{\text{reflect}}}{V_{\text{out}} + V_F} = \frac{3 \text{ V}}{5 \text{ V} + 0.6 \text{ V}} = 0.54 \approx 0.5 \quad (2)$$

V_F is the forward voltage drop across the rectifier schottky diode, which is assumed to be 0.6V. To simplify the transformer ratio, we used N_P/N_S of 0.5 instead of 0.54.

The maximum duty cycle of the flyback converter is:

$$D_{\text{max}} = \frac{N \times (V_{\text{out}} + V_F)}{V_{\text{inmin}} + N \times (V_{\text{out}} + V_F)} = \frac{0.5 \times (5 \text{ V} + 0.6 \text{ V})}{4.5 \text{ V} + 0.5 \times (5 \text{ V} + 0.6 \text{ V})} = 0.36 \quad (3)$$

To calculate the primary inductance, ΔI_L is set to half the primary current as an acceptable ripple current. For a CCM flyback design, the peak primary current is then calculated as:

$$I_{\text{peak}} = \frac{I_{O\text{max}}}{N(1 - D_{\text{max}})} + \frac{\Delta I_L}{2} \quad (4)$$

By replacing ΔI_L with $\frac{1}{2}(I_{\text{Peak}})$, $I_{O\text{max}}$ with 0.2A, D_{max} with 0.36, and N with 0.5 which was calculated earlier, the peak primary current is calculated to be 0.625A and ΔI_L calculates to 0.313A.

The inductance of the transformer's primary-side is calculated as follows:

$$L_P = V_{\text{in}(\text{min})} \times \frac{D_{\text{max}}}{f \times \Delta I_L} = 4.5 \text{ V} \times \frac{0.36}{650 \text{ kHz} \times 0.313 \text{ A}} = 8 \mu\text{H} \quad (5)$$

For the mini-power application, we can choose a small magnetic core, such as EE8. The A_e of the EE8 core is 7mm^2 .

$$N_P = \frac{V_{\text{inmin}} \times D_{\text{max}}}{A_e \times B_{\text{max}} \times f_{\text{min}}} = \frac{4.5 \text{ V} \times 0.36}{7\text{mm}^2 \times 0.15\text{T} \times 480 \text{ kHz}} = 3.2 \approx 4 \quad (6)$$

$$N_S = N_P / N = 4 / 0.5 = 8 \quad (7)$$

2.5 Output Voltage Setting

To reduce the cost, the circuit uses an auxiliary winding on the primary-side of the transformer, N2, which can save the cost for the optocoupler and a reference like TLV431 as the secondary-side control. As a result, the output load transition is not as good as with secondary-side control, but it is adequate for the RS485's power supply and some other applications.

Choose for N2 the same turns as for N3, the secondary side winding. The voltage at the point V_S (see [Figure 2](#)) is similar to V_{OUT} , because the turn ratio of N2:N3 = 1. R3 is chosen to be 10 k Ω (on the basis of experience), R4 can be calculated with [Equation 8](#):

$$R_4 = \frac{(V_S - V_{FB}) \times R_3}{V_{FB}} \quad (8)$$

V_{FB} is given as 1.238V (the feedback voltage of TPS61085), and V_S is supposed to be 5V, which leads to a value of R4 of 30k Ω . For the actual circuit, V_S varies with the actual turns ratio of N2:N3 and the leakage inductance of the transformer. So R4 needs to be adopted in the actual circuit. In this case, R4 is finally set to 27 k Ω .

R7 is a dummy load, which can hold the output voltage in the range when there is no load at the output. Reducing the value of R7, can improve the load regulation, but, as a trade off, will increase the no load power dissipation.

2.6 Compensation Circuit

For C_{COMP} (C3), a value of 2.2nF or 3.3nF is sufficient for most applications. R_{COMP} (R6) is set as 13k Ω as recommended in the datasheet.

2.7 Input and Output Capacitors

For the input and output capacitors, it is recommended to use the low ESR ceramic capacitor with 10 μ F. For low cost applications, it is possible to use the capacitor combination of one 47 μ F electrolytic capacitor and one 0.1 μ F ceramic capacitor.

3 Test Results for the Flyback Solution with TPS61085

The efficiency for the TPS61085 isolated power solution with an input voltage of 5V is shown above in [Figure 3](#). The efficiency at full load is around 75%, the main power loss of the circuit are the conduction and switching losses of the MOSFET integrated into TPS61085, the conduction loss in the output diode D3, and the power dissipation in the transformer. For light load efficiency, the power dissipation on the dummy load can not be ignored.

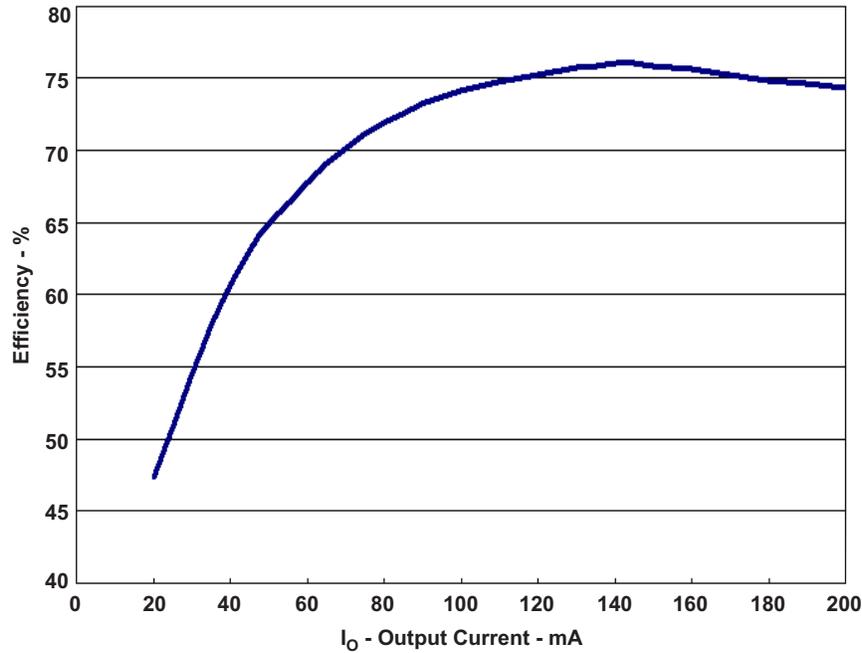


Figure 3. Efficiency with 5V Input

[Figure 4](#) is the load regulation of this power solution at an input voltage of 5V. The total load regulation is limited to $\pm 5\%$, which meets the requirement of the RS485 application.

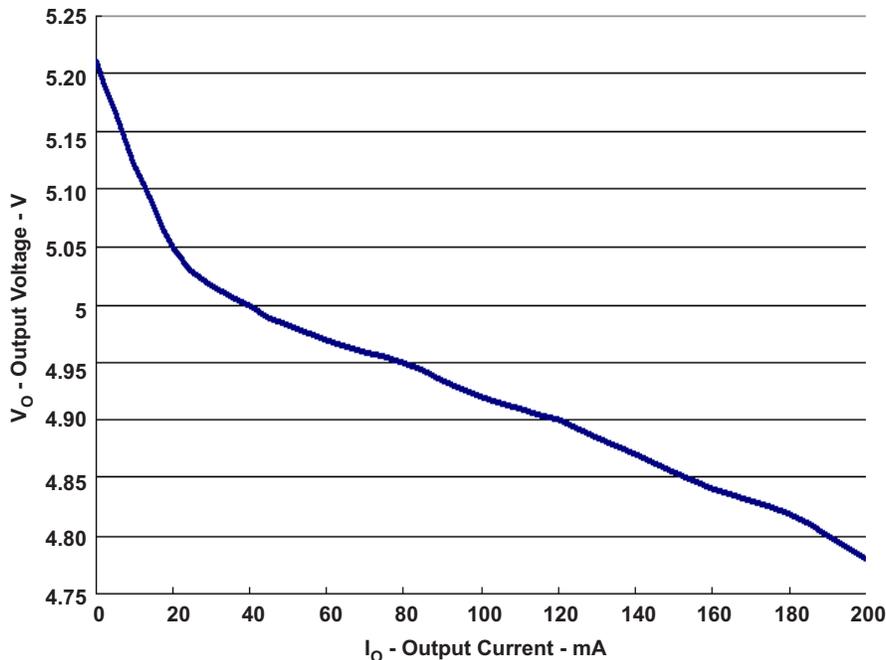


Figure 4. Load Regulation with 5V Input

4 Conclusion

The application report has introduced the isolated mini-power supply solution with TPS61085 and gives the design process for a 5V to 5V 1W isolated power supply.

In some applications, if a tighter load regulation is required, [Figure 5](#) provides another regulation possibility. It uses secondary side control with an optocoupler and the TL431. [Figure 5](#) demonstrates a solution with a tighter load regulation for a 12V output application. This is a paper design for the 12V output, and some components may require changes in an actual design.

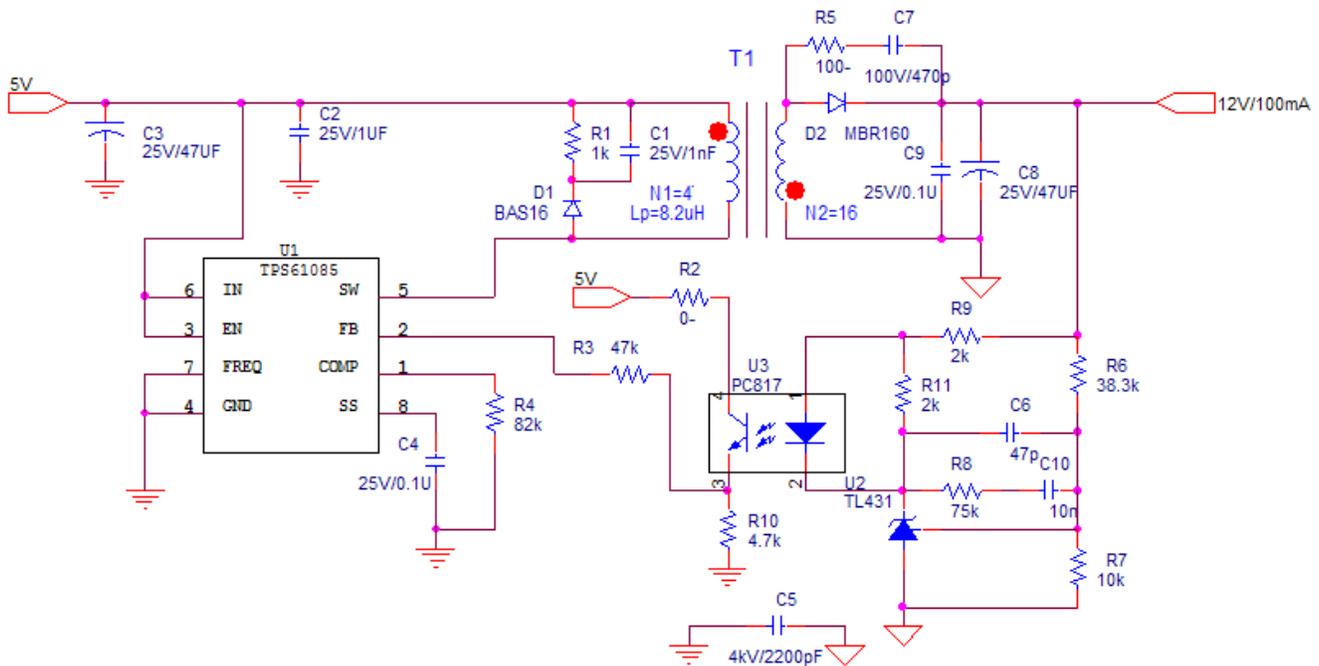


Figure 5. TPS61085 12V Isolated Solution with TL431 (Paper Design)

5 Reference

U-165 ([SLUU096](#)) Reference Design: Isolated 50 Watt Flyback Converter Using the UCC3809 Primary Side Controller

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