Designing Cost-Effective, Low Power Galvanic Isolation Solutions for Power Supplies

Isolated Buck Converter Topology Calculations Help Meet Isolated Data Interface Needs in Test & Measurement and Medical Applications

Bourns® Transformer with 4 kV Isolation and 6 mm Creepage for Isolated Buck Topologies

INTRODUCTION

Standard industrial interfaces such as RS-485 and RS-232 are commonly used for unidirectional and bidirectional data transfer in a range of applications that encompass everything from medical monitoring to test and measurement equipment. If the power supply used in these applications has galvanic isolation, it can deliver the required voltage level to both sides of the connecting cable without introducing switching noise or common mode voltages. The data bus and transceiver chip DC voltage supply are typically also isolated from the power supply to protect end users from the hazards of electric shock. This is particularly important in medical applications which have stringent rules on isolation of power from end users.

Furthermore, sensitive sensors in test and measurement equipment rely on highly accurate operational amplifiers (op-amps) for processing the signals with as little distortion as possible. Op-amps require very stable noise-free DC bias, which is provided by an isolation barrier. The challenge in all cases is to provide galvanic isolation and a stable power source in a cost-effective manner. Adding opto-isolated feedback for the control loop which is typically used in isolated power topologies adds complexity and cost. This solution provides high isolation in a small compact transformer and uses an isolated buck topology which removes the need for isolated feedback.

This paper presents the reasons for galvanic isolation in board designs that incorporate common RS-485 or RS-232 data interfaces or for sensors that can be sensitive to ground loops. It will also provide an optimal design methodology for the implementation of a coupled inductor with high creepage and clearance for a 2 W power supply.
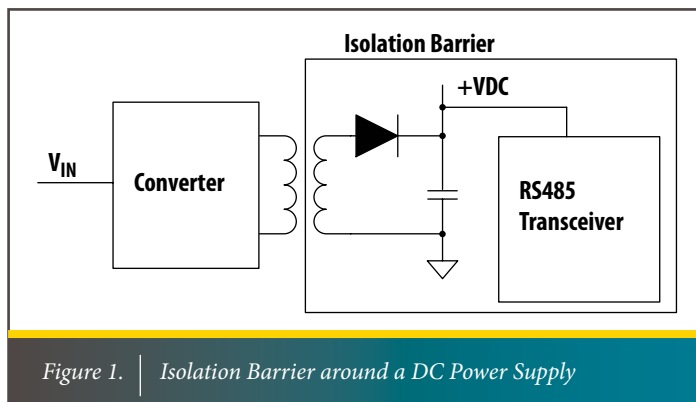

Figure 1.   *Isolation Barrier around a DC Power Supply*

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ISOLATED BUCK TECHNOLOGY

The isolated buck converter provides low-current auxiliary and bias outputs from a wide-ranging input supply.
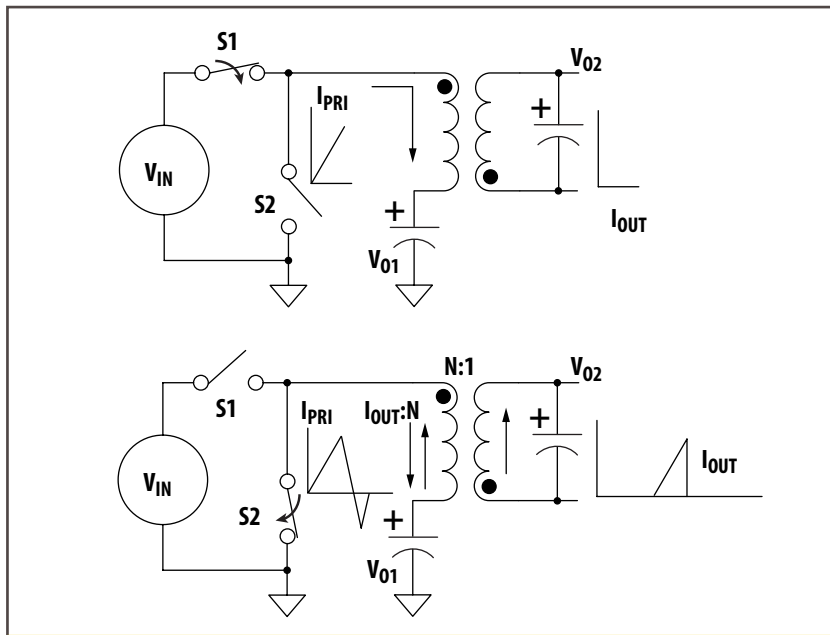

Figure 2. | *Operation of Isolated Buck Converter as the Switches S1 and S2 are Activated*

Isolated buck converters are proven solutions for applications of up to 10 W, while flyback power supplies (isolated buck-boost) are also used in applications of up to 250 W.

Figure 2 shows the operation of the isolated buck toplogy. During time D x Ts when the switch S1 is on, the primary winding conducts current I_{pri} while the secondary winding is open due to the reverse biased diode D1. During (1-D)Ts, the diode is forward biased while S2 is now on, and both windings conduct with the current in the primary winding decaying to zero and then changing direction as it continues to carry current coupled from the secondary. The output current I_{out} will be the average current over one complete switching cycle and this is calculated by integrating the peak current in the primary over that time. The maximum current that the switch can conduct ($I_{sw,peak}$) is equivalent to the sum of the output current and the ripple currents in both windings ($\frac{N_{sec}}{N_{pri}}I_{out} + \frac{\Delta I_{pri} + \Delta I_{sec}}{2}$). This is the same as the peak current during D x Ts. However, no current flows in the secondary winding during DTs so ΔI_{sec} is actually equal to zero. Therefore:

$\Delta I_{pri} = 2 (I_{sw,peak} - \frac{N_{sec}}{N_{pri}}I_{out})$: *Equation 1*

Once ΔI_{pri} (peak current ripple) is known, the inductance value of the primary winding L_{pri} can be calculated using the equation $V = L \frac{\Delta I}{\Delta t}$: *Equation 2.*

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BOURNS SOLUTION FOR ISOLATED POWER FOR A TEST AND MEASUREMENT APPLICATION

A typical isolated power solution for the RS-485 interface on many test and measurement equipment is shown in figure 3. The specification for the power supply is as follows:
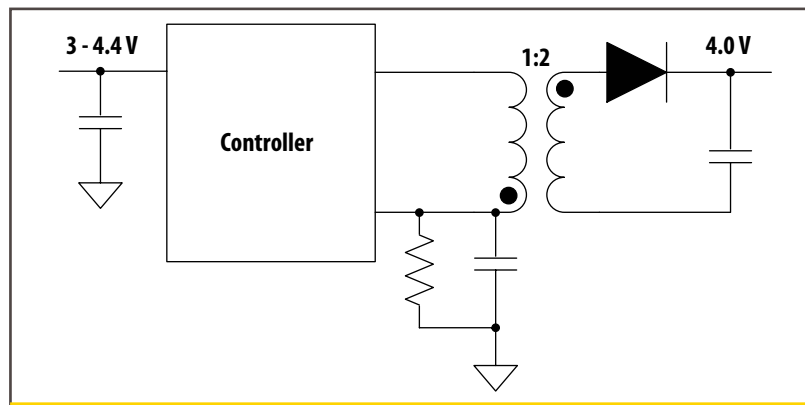

Figure 3. | *Circuit Diagram of Isolated Power Supply*

Specification

    A. Input Voltage (V_{in}): 3.0 V – 4.4 VDC

    B. Output Voltage (V02): 4.0 V

    C. Power: 2 W

    D. Frequency: 200 kHz

    E. Dielectric Strength: 4 kV

    F. Creepage and Clearance: 6 mm

The peak current in this calculation depends on the maximum switch current. The controller has a current limit of 2 A in the high side switch. Using Equation 1, the ripple current calculation in the primary winding ΔIpri is equal to 2 A. From there, the calculation of the primary inductance $L_{pri} = V01 \frac{\Delta t}{\Delta I_{pri}}$ during the time (1-D)Ts and that V01 is equal to D x V_{in} (like a buck converter).

Therefore, $L_{pri} = \frac{V_{in}.D.(D - 1)\Delta I_{pri}}{F_{sw}} = 1.8$ µH

Reducing the ripple current will help reduce power lost due to proximity of the windings. Increasing the inductance to 10 µH will reduce the ripple current to 0.375 A.

The turns ratio $\frac{N_{sec}}{N_{pri}}$ will have to be equal to 2 in order to keep V02 closer to the target of 4.0 V. V02 actually tracks V01. Therefore, there is no need for opto-isolation and an auxiliary winding for regulation of the isolated output. It follows that the cost of the coupled inductor can be greatly reduced.

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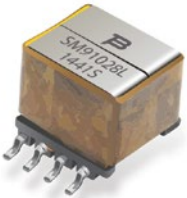
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WINDING DESIGN

The core chosen for this design is an EP13 core with a surface mount 8 pin bobbin. The minimum number of turns will depend on the area of the core and the inductance value. For this, the equation $B_{s}A_{x}N > L_{prix} I_{peak}$ can be used to calculate the minimum number of turns needed without saturating the core. The following values are used to calculate the number of turns:

    A. B_s: 0.4 T
    B. A: 0.149 cm² (Core Area of Ep13)
    C. L_{pri}: 10 µH
    D. I_{peak}: 2 A. This is the current limit of the controller chip.

The minimum number of turns is 3.8 and can be rounded up to 4. The flux density B_{ac} is calculated using the following equation $B = \frac{E.t}{N.A}$; where E is the applied voltage to the primary and t is the time that the voltage is applied to the primary. This determines that the flux density B at N turns is 0.16 T. ΔB is as a rule calculated to be 50 % of the total flux density or 0.08 T. By consulting the core manufacturer's graph of flux density (figure 4), it is possible to calculate the power loss in the core with a volume of 0.4 cm³ at 200 kHz to be 0.12 watts. Increasing the turns to 8 reduces the flux density by half, from 0.16 T to 0.08 T and the core loss to 8 mW.
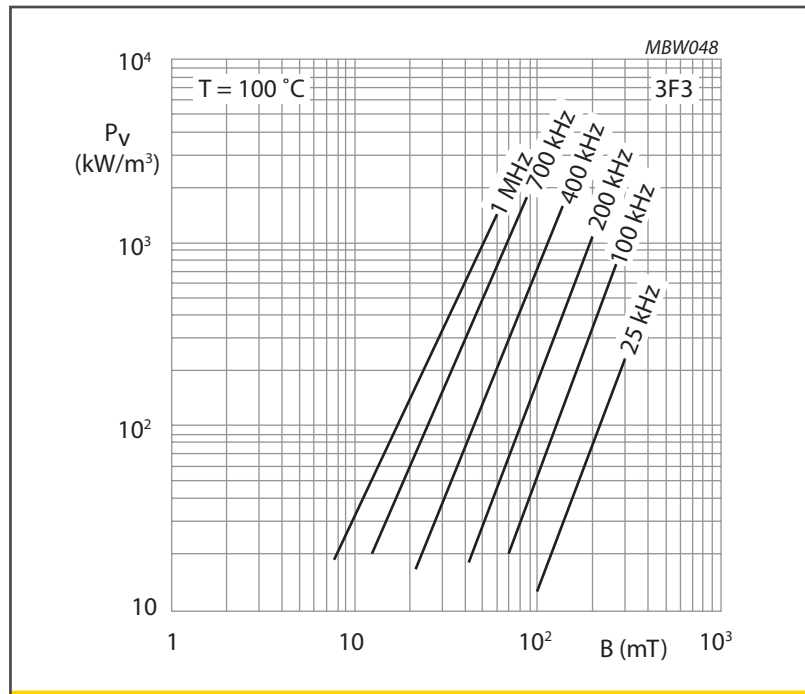

Figure 4. | *Typical Core Loss Curve for Ferrite Material*

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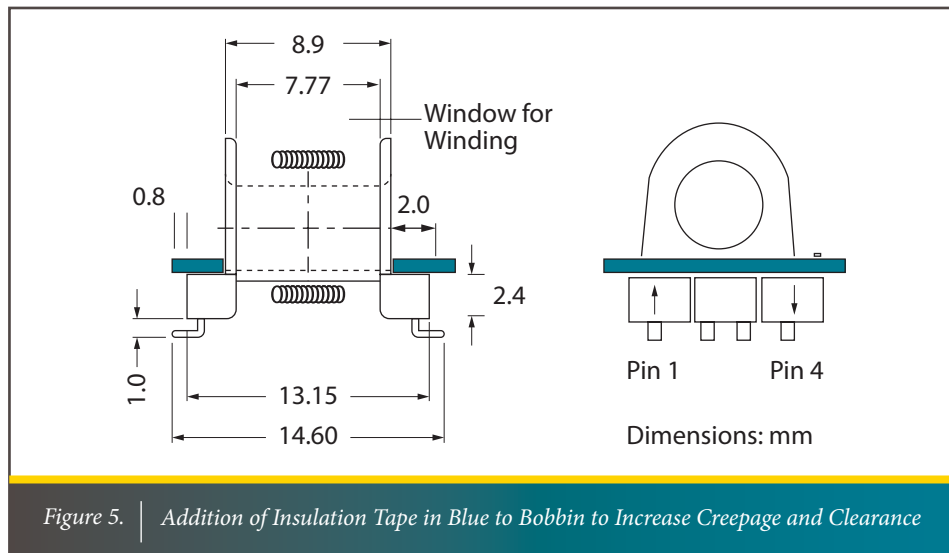
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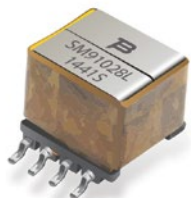
WINDING DESIGN *(Continued)*

"Creepage" refers to the distance over a surface between conductors of the primary and secondary windings. This can be over multiple surfaces and through conductive elements, of which the ferrite core is one. Clearance is another aspect to be considered, and is described as the line of sight distance between primary and secondary, although, quite often this is the same as the creepage. The distances for both are defined in various UL standards: UL 60950 (IT equipment and transformers) and IEC 60601 (medical equipment). In addition, UL distance definitions are: UL 60950 (IT equipment), IEC 61558-1 (transformers) and IEC 60601 (medical equipment).

The core is considered conductive and, therefore, any flash over between one pin of the core will track across the core. Consequently, the shortest path between two pins of a winding may actually not be the distance over the circuit board but could be through the core itself. To overcome this, adding insulation tape to the bobbin where the core normally sits is recommended. Insulation tape to the core can also be added where it meets the bobbin. This allows designers to effectively increase the distance between the pin and the core by up to 2.0 mm. The shield also must extend 0.8 mm around the outside of the core, on all sides, to maintain creepage and clearance. The distance between two pins of a winding will now be twice the sum of the tape width and the height above the board of the core.

Figure 5. | *Addition of Insulation Tape in Blue to Bobbin to Increase Creepage and Clearance*

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WINDING DESIGN *(Continued)*

Creepage between the primary and secondary on such a small bobbin is achieved by using a triple insulated wire on the primary. This negates the need to use margin tape with ordinary enamel wire which is not considered to have insulation. The secondary winding must also contain at least one layer of insulation. Otherwise, the secondary will be considered to be at risk of flashing across to the core which would halve the creepage distance from 8 mm to 4 mm.

Using the available window length of 7.7 mm of the EP13 bobbin, consulting winding tables and the wire manufacturers' data sheet enables designers to calculate the resistances of the primary and secondary windings. Using this information and the power supply specification enables the measurement of DC power loss. Due to the high frequency there will also be power loss due to the proximity effect. The proximity effect changes the distribution of the ripple current in the winding effectively increasing the resistance seen by the ripple current. This resistance is known as the AC resistance and can be calculated once the thickness of the wire in the winding and the frequency are known.

Using the following Hurley equation, the ratio of the AC resistance to the DC resistance is as follows:

$$\frac{R_{ac}}{R_{dc}} = \Delta \left| \frac{Sinh2\Delta + Sin2\Delta}{Cosh2\Delta + Cos2\Delta} + 2 \frac{(P^2 - 1)}{3} \frac{Sinh2\Delta + Sin2\Delta}{Cosh2\Delta + Cos2\Delta} \right| \quad \text{equation 3}$$

where Δ is the ratio of the wire diameter to the skin depth ($\sqrt{\frac{60}{F_{sw}}}$) and P is the number of layers in the winding. Once the DC resistance is known for the windings, the AC resistance can be calculated:

Primary Side
- DC Resistance: 0.015 Ω
- AC Resistance: 0.07 Ω

Secondary Side
- DC Resistance: 0.067 Ω
- AC Resistance: 0.22 Ω

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WINDING DESIGN *(Continued)*

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The ripple current in the primary was already calculated as being 0.375 A. The secondary output current has to be recalculated given the higher inductance in the primary. Replacing the value L_{pri} in equation 1 and solving for I_{out} gives us an output current of 0.9 A. The peak current is calculated from the formula $I_{rms} = I_{pk}\sqrt{\frac{D}{3}}$ which works out as $I_{pk} = 2.25$ A. The DC component of that is taken from the formula $I_{dc} = I_{peak}\sqrt{\frac{D}{2}}$ or 0.56 A. Subtracting this from the peak current will give us the true peak of the AC current from which we can calculate its rms value of 0.676 A.

The input current can be calculated from the minimum input voltage taking into account the power and target efficiency. Allowing for 95 % efficiency, the input current is calculated as being 0.6 A.

From the calculations above, the following is a summary of the power lost in the transformer:

Primary (Watts)	DC Power	0.0054
	AC Power	0.0098
Secondary (Watts)	DC Power	0.0210
	AC Power	0.1000
Core (Watts)		0.0080
Total (Watts)		0.145

The sum of the total power lost in the transformer (taking DC, core and AC losses into account) is calculated as 0.145 watts or 7.2 % of the specification. The power lost can be diminished by reducing AC losses in the secondary. This can be achieved using multi-stranded wire instead of a single gauge wire. This has the effect of reducing the overall wire thickness and, hence, the AC resistance. However, multi-stranded wire is more expensive than single gauge and is more difficult to assemble. A trade-off between efficiency and cost will have to be made to arrive at an optimum solution.

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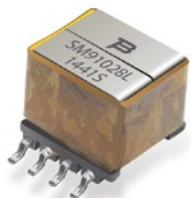
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SUMMARY

Isolated power for data interface drivers such as RS485 or for analog circuits processing sensitive sensor signals is often a requirement in test & measurement, medical and industrial applications. The efficiency of the power supply depends on the design of the magnetic component as this has an effect on the DC, AC and core power dissipation. Using the right magnetic component materials is the key to achieving high creepage rules in such small surface mount components. Transformers for low-power isolated buck applications can be cost-effective due to the simplicity of the design as there is no need for isolated feedback and extra auxiliary windings. However, achieving the right efficiency may involve a compromise on cost as often more expensive multi-strand wires are needed to reduce AC power losses.

Bourns offers in-house capabilities to help customers design and prototype inductors for low power circuits. The company's extensive inductor experience allows it to meet stringent international standards from its comprehensive portfolio of cores and bobbins in stock to meet fast time-to-market needs.

REFERENCES

Bourns: "Designing a Low Power Flyback Power Supply":
http://www.bourns.com/data/global/pdfs/Bourns_Consumer_Grade_Flyback_Power_Supply_Appnote.pdf

ADDITIONAL RESOURCES

For more information about Bourns' complete line of circuit protection products, please visit:

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EMEA: *Tel* +36 88 520 390
Fax +36 88 520 211

Asia-Pacific: *Tel* +886-2 256 241 17
Fax +886-2 256 241 16