Selecting IHLP Composite Inductors for Non-Isolated Converters Utilizing Vishay's Application Sheet

INTRODUCTION

This application note will provide information to assist in the specification of IHLP composite inductors based on given operating conditions utilizing Vishay's IHLP application sheets. It is assumed that the designer has a basic understanding of non-isolated dc-to-dc converters as this is not a design exercise in that family of converters. That being said, tools will be introduced to allow the designer to select an IHLP inductor and estimate its performance in their applications. These tools will work for buck, boost and buck/boost topologies. A set of application sheets will be provided containing all the relevant constants and equations needed for the selection process. This process is an estimation only and all parts should be verified in their application. It is also understood that the application sheets are a work in progress and some of the data is estimated through calculation. The estimated data will be highlighted on the application sheets.

BACKGROUND

The IHLP inductor is constructed using an "open" or "air coil" inductance coil. The two ends of the coil are connected to a lead frame that acts as a means of transport through the manufacturing operation at Vishay, and as the final termination pads when the part is singulated from the lead frame. A powdered iron core is pressed around the inductor coil after the inductor coil is welded to the lead frame. The characteristics of the powdered iron enhance the magnetic properties of the inductor and also give the inductor its final shape or footprint.

The composite inductor is essentially built backwards from a conventional inductor. In a conventional inductor the magnet wire is wound either directly on the core as in a toroid, or wound on a bobbin with the core halves inserted into it as in "E" style cores. Since each IHLP size and value has a unique coil dimension varying in outside and inside diameter and height, each inductor has different geometric parameters. This means that core constants must be calculated for each inductor size and value.

The only consistent item in a series of inductors will be the performance of the iron powder from value to value; therefore, the core loss constants for the material will remain the same. There are, however, different iron powders used in different product lines to cover a wider range of operating conditions. Within these IHLP product lines the same inductance values do not use the same air coil, which means constants will be required not only for geometry but for material as well. What it comes down to is that each inductor has its own unique parameters even within the same family size.

Composite inductors are frequently used in non-isolated dc-to-dc converters. This is not an issue, however the waveforms associated with them are not in line with conventional thinking. Core loss characterization and the resulting data are often determined using sinusoidal excitation. Dc-to-dc converters on the other hand do not operate with sine waves, instead they use a pulsed DC waveform. This means that the current waveform in the inductor determining core loss will be a triangular wave, not a sine wave. This difference will need to be compensated for in the core loss calculations.

Increasingly, dc-to-dc converters are being asked to operate at higher ambient temperatures. This in turn requires the inductor to operate at the higher temperature in addition to its own temperature rise incurred due to power losses. It is known that iron powder exhibits the effects of aging at higher temperatures in the form of increased core losses. These losses must be accounted for during the design process in order for a composite inductor to be used at temperatures in excess of 125 °C. The effects of thermal aging can be minimized by simply limiting the maximum inductor temperature to 125 °C or less.

SELECTION TOOLS

Criteria

Start the inductor selection process by establishing the selection criteria for the part. Composite inductors have a recommended maximum component temperature of 125 °C. Subtracting the ambient temperature will give us the maximum allowed temperature rise for the part. If this number should exceed 40 °C it is recommended that 40 °C be used for the allowed temperature rise. Core losses should be limited to ≤ $^1/_3$ of the total losses to mitigate any aging effects associated with the powdered iron in the core at elevated temperatures. Data sheets list a heat rated current (I_{HEAT}) as a parameter, which represents the current needed to produce a certain temperature rise indicated on the data sheet. This temperature rise is typically measured using DC current and is due to copper losses only and does not take into account core loss. However, this information is

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useful since it can be used to determine maximum power losses allowed in the inductor by multiplying the temperature corrected resistance of the inductor by the heating current squared. This will be the power loss (P_{HEAT}) to produce the temperature rise associated with the I_{HEAT} parameter.

In designing with the IHLP style of inductors core losses are characterized using the Modified Steinmetz Equation (MSE) [1] yielding P_v (core loss) in mW/cm^3,

$$P_v = K_0 \times f_e^{K_f - 1} \times B_{pk}^{K_b} \times f_o \quad (1)$$

This equation uses an "effective frequency" of the non-sinusoidal waveform along with the operational frequency of the circuit while making use of the Steinmetz parameters which are determined through curve fitting from lab testing data.

Core Constants

The core loss equation shown previously has several constants associated with it that must be determined. They are the core constant K_0, the frequency constant K_f, and the flux density constant K_b. These constants are determined by curve fitting of experimental core loss data established from lab testing at Vishay. These constants are unique to the IHLP material they are established for and do not change from inductance value to inductance value. They will, of course, change from material to material and are tabularized on the application sheets.

Effective Frequency

Effective frequency (f_e) is a frequency other then the repeat frequency of the circuit used to correct for non-sinusoidal waveforms. The effective frequency for the waveforms experienced in dc-to-dc converter inductors can be characterized as follows [1]:

$$f_e = \frac{\int_0^T \left|\frac{d^2B}{dt^2}\right| dt}{2\pi \int_0^T \left|\frac{dB}{dt}\right| dt} \quad (2)$$

This method equates the frequency to the sum of the changes in the slope of the flux density divided by the changes in amplitude divided by 2π.

Method for Calculating Core Loss

To make use of the core loss equation, the effective frequency (f_e) and maximum flux density (B_{pk}) need to be established. These items are then plugged into the MSE to determine core loss per unit volume. Since the V-µs product of the circuit is known, or can be calculated, we will make use of it to determine B_{pk}. Composite inductors can be characterized in terms of their ability to handle a certain V-µs product corresponding to 100 G. This parameter is referred to as the ET_{100} constant. This constant will be different for each inductor size and value and can be used to determine B_{pk} as follows:

$$B_{pk} = \frac{ET_{ckt}}{ET_{100}} \times 100 \quad (3)$$

where ET_{ckt} is the V-µs product of the circuit and the units are in G. The V-µs product of the circuit can be taken from the IC manufacturer's data sheets and application notes or calculated from V_{OUT}, frequency and duty cycle.

It can be shown that the effective frequency depends on the duty cycle (δ) of the circuit and the operational frequency (f_0),

$$f_e = \frac{f_0}{2\pi \times (\delta - \delta^2)} \quad (4)$$

Changes in the flux density in the magnetic circuit are proportional to changes in the inductor current, assuming that the inductance remains constant. Core losses are caused by this ripple current according to the flux density levels. The inductor is subjected to the input voltage during the switch on time and a voltage of the opposite polarity during the switch off time. The resulting plot of the flux density is composed of two straight-line portions that correspond to the integration of these two voltages over time.

Once the effective frequency and the maximum flux density are known, core losses can be estimated using the MSE. The result from this equation will have the units mW/cm^3 but to simplify calculations the core volume is rolled into the K_0 constant giving us the final version of the MSE for core loss (P_{core}) in W,

$$P_{core} = K_0 \times f_e^{K_f - 1} \times B_{pk}^{K_b} \times f_0 \times 10^{-14} \quad (5)$$

The core loss constants and associated parameters for the IHLP composite inductor product line are contained in Table 1, separated by inductance value and material type.

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SELECTION PROCESS

Method

To determine core losses a designer can calculate the peak flux density, using V-μs product of the voltage waveform across the inductor during operation. Utilizing this core loss estimate, the designer can balance the combination of core and copper losses to keep the total losses and associated temperature rise less than the maximum 125 °C operating temperature of the composite inductor. Care must be taken to insure accurate copper losses by accounting for the resistance increase of copper due the change in temperature from room temperature (where DCR is specified) to the maximum operating temperature and losses due to AC components such as skin effect. Lastly, check that I_{PEAK} is $\leq I_{SAT}$ (found on the data sheet). Due to the soft saturation feature of IHLP composite inductors, I_{PEAK} can exceed I_{SAT} without a serious reduction in inductance as observed in ferrite style inductors.

DESIGN EXAMPLE

Inputs

A designer is tasked with designing an inductor for the following "buck" circuit, V_{in} = 5.0 V, V_{out} = 1.8 V, I_{out} = 20 A, Frequency = 300 kHz, ambient temperature of 50 °C. To strike a compromise between output capacitor size and cost and inductor size we will design for a ripple ratio of $0.3 \leq r \leq 0.5$. Ripple ratio is defined as the ratio of AC to DC current components in the inductor. Allowing for a diode drop of 0.5 V and estimating the switching voltage drop by,

$$V_{SW} = I_{SW} \times R_{DSon} \quad (6)$$

where I_{sw} is equal to,

$$I_{SW} = I_o \times \delta \quad (7)$$

The above conditions produce design requirements of: Duty cycle = 0.46, V-μs product of 4.14 and a required inductance of 0.54 μH. Since we wish to keep the ripple ratio around 0.4 and since ripple ratio varies with inductance choose the closest standard inductor value 0.56 μH. The circuit design equations are summarized in Appendix A.

Derating

The heating current of IHLP composite inductors is specified based on a 40 °C temperature rise due to the current only; therefore we will need to derate the DC current to allow for the heat rise due to core losses. We want to select an inductor with a higher heating current than the 20 A the circuit demands, while minimizing size. Reviewing the data sheets, we find the IHLP-4040DZ-01 series should be a good fit.

Process

Data and application sheet specifications for the IHLP-4040DZ-01 0.56 μH inductor are summarized in Table 1.

	IHLP-4040DZ-01
Inductance	0.56 μH
$DCR_{MAX.}$	0.0017 Ω
I_{HEAT}	27.5 A
I_{SAT}	49.0 A
R_{TH}	26.96 °C/W
P_{HEAT}	1.48 W
ET_{100}	0.88
K_0	18.31
K_1	0.00340
K_f	1.188
K_b	2.118

Table 1

Start by determining B_{pk}, to do this use (3). Plugging in the circuit ET product and the ET_{100} parameter of the inductor gives us a B_{pk} of 470.5 G. The effective frequency from (4) will be 192 216 kHz. The resulting core losses using (5) are:

$$P_{core} = 18.31 \times 192\,216^{1.188-1} \times 470.5^{2.118} \times 300\,000 \times 10^{-14} = 0.248 \text{ W}$$

To continue with the selection process we need to determine total losses which must include copper losses. To calculate copper losses we must determine operational resistance (R_{OPER}), circuit ripple current and the inductor DC current. R_{OPER} is the temperature corrected resistance of the inductor and can be found, assuming a 40 °C rise in temperature with

$$R_{OPER} = R_{MAX.} \times \frac{234.5 + T_{AMB} + T_{RISE}}{259.5} \quad (8)$$

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Delta I (ΔI), or ripple current, for a buck inductor is related to the output voltage, inductance in μH, frequency in Hz and duty cycle (δ) according to,

$$\Delta I = \frac{V_{out} + V_{DIODE}}{L \times f_0} \times (1 - \delta) \times 10^6 \quad (9)$$

The resulting ΔI for this circuit is 7.39 A, the power loss asoociated with AC effects can be determined using:

$$P_{ac} = K_1 \times \Delta I^2 \times \sqrt{f_{OP}} \times R_{OPER} \quad (10)$$

Multiplying the square of the DC current in the inductor (equal to the output current in a buck inductor) by the temperature adjusted operational resistance from (9) results in DC copper losses of 0.852 W. The power losses associated with AC effects, from (10), are 0.217 W. Adding these results to the core losses produces total losses of 1.317 W. The temperature rise can be found by multiplying the total losses by the thermal resistance (R_{TH}) found in the inductor specification sheet. The temperature rise for the inductor will then be,

$$\Delta T_{0.56} = 1.917 \times 26.96 = 35.51\ °C$$

It can be seen that by adding the temperature rise to the ambient temperature the inductor will not exceed the maximum recommended component temperature of 125 °C.

Based on this it appears the IHLP-4040DZ-01 0.56 μH is good choice for this circuit. The last item to check is to see if the peak inductor current (I_{PEAK}) is less then the saturation current of the inductor. The peak current can be determined by,

$$I_{PEAK} = I_{DC} + \frac{\Delta I}{2} \quad (11)$$

I_{PEAK} for this inductor is 23.70 A. Checking the data sheet information the 0.56 μH part has a saturation current rating of 49 A, therefore we can safely use this inductor for this application.

References

[1] Jieli Li, T. Abdallah, and C.R. Sullivan, "Improved calculation of core loss with non-sinusoidal waveforms", IEEE Industry Application Society annual meeting, 2001, pp. 2203-2210

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APPENDIX A

Buck Circuit Equations

$$\delta = \frac{V_o + V_d}{V_{in} - V_{SW} + V_d}$$

$$ET = \frac{V_o + V_d}{f} \times (1 - \delta) \times 10^6$$

$$L_{REQ} = \frac{V_o + V_d}{I_o \times r \times f} \times (1 - \delta) \times 10^6$$

$$\Delta I = \frac{V_o + V_d}{L \times f} \times (1 - \delta) \times 10^6$$

$$I_{DC} = I_o$$

$$I_{SW} = I_o \times \delta$$

$$V_{SW} = I_{SW} \times R_{DSon}$$

Definitions

δ = Circuit duty cycle

V_o = Circuit output voltage in V

V_d = Diode voltage drop in V

V_{in} = Circuit input voltage in V

V_{SW} = Switch (MOSFET) voltage drop in V

ET = Circuit V-μs product

f = Circuit switching frequency in Hz

L_{REQ} = Required inductance in μH

I_o = Circuit output current in A

r = Ripple ratio

ΔI = AC current component in the inductor in A

I_{DC} = DC component in the inductor in A

I_{SW} = Current in the MOSFET switch in A

R_{DSon} = The "on resistence" of the MOSFET switch in W