

A Comparative Study for Interleaved Boost Converter with Coupled/Uncoupled Inductors

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Abstract

This paper explores a comparative analysis of coupled and uncoupled inductors in context of an interleaved boost converter. The study includes the derivation of steady state variables and magnetic equations for each inductor configuration. To mitigate input current ripple issues in the conventional boost converter, interleaving techniques are employed. Additionally, the volume concerns associated with interleaving are addressed through the application of inductor coupling techniques. The utilization of coupled inductors results in a more compact, efficient, and cost-effective design when contrasted with the uncoupled configuration. Finite Element Method (FEM) is used for the analysis of the proposed inductors, and analytical loss calculations are validated. As the most suitable topology, a two-phase interleaved boost converter with an inversely coupled inductor operating at 3700 W with an efficiency of 98.17% has been designed.

1. Introduction

Interleaving is widely used in DC-DC converters to reduce current ripple within the topology and thus reduce energy losses. Interleaved Boost Converters (IBC) have found widespread use in high-power applications such as electric vehicles, battery storage systems and renewable energy sources, aiming to improve operational efficiency [1]. Although utilizing a sizable inductor reduces input current ripple in typical boost converters, interleaving eliminates the requirement for high volume inductors by minimizing current ripple and mitigates stress and losses on devices [2–4]. Through the parallel connection of boost cells and phase shifting in semiconductor devices, an increased effective circuit frequency is attainable. Studies [5,6] show that the interleaving technique also diminishes the volume of the electromagnetic interference (EMI) filter without compromising efficiency. Elevating the number of phases used in the interleaving technique yields reductions in input current ripple, output voltage ripple, and switching losses [7]. Nevertheless, the rate of this reduction decreases as the number of phases increases, necessitating careful consideration of the increased volume, cost, and complexity of the converter. Notably, the drawback of the interleaved boost converter (IBC) lies in the fixed duty cycle to achieve zero-input current ripple, tied to the phase count and devoid of arbitrary selection [8].

A coupling approach is used to limit the volume increase in the IBC. A significant increase in inductor volume can be obtained by coiling around a single core. Thanks to coupled inductors, the ripple problem can be solved without significantly increasing the inductor's size. Since greater coupling factors result in more current ripple, inductors can be loosely coupled in the cou-

pling technique [9]. Previous studies show that a loosely coupled inductor is more compatible for downsizing the inductor and efficiency enhancement compared to a single-phase or uncoupled counterpart [10, 11]. Also, in [12–14], uncoupled inductors and loosely coupled inductors were compared and downsizing of the magnetic component was achieved.

The coupling type, namely direct and inverse, is another factor affecting current ripple mitigation. Input current ripple in a boost converter with a directly coupled inductor is observed to be lower than in the uncoupled counterparts [15]. Conversely, in the case of inverse coupling, the phase current ripple is reduced compared to the uncoupled topology, while there is an increase in the input current ripple [16].

In this study, uncoupled IBC, directly coupled IBC, and inversely coupled IBC are examined in terms of volume, loss, and cost under the same operating conditions with a comprehensive approach, as compared with previous studies in the literature. Firstly, an overview of all interleaved boost converter topologies is presented, current and voltage relations and general design principles are given. After that, the magnetic design equations of uncoupled and coupled inductors are provided, and their loss equations are derived. Finally, a comparative analysis was conducted among three distinct configurations of a two-phase Interleaved Boost Converter (IBC), each designed for a 3700 W power rating with an input voltage of 200V and an output voltage of 400V. Different from previous studies, the comparison of direct and inverse coupling was made with regard to considerations related to volume, losses, and cost using a loosely coupled toroidal core.

2. Topological Overview of Coupled Structures in Interleaved Boost Converters

This section presents the steady-state equations for a two-phase interleaved boost converter with coupled and uncoupled inductors.

2.1. Conventional Interleaved Boost Converter

Fig. 1 shows a two-phase uncoupled interleaved boost converter. Each cell is equipped with its own inductor, switch, and diode, complemented by a singular output capacitor.

The sum of phase currents, i_{L1} and i_{L2} , constitutes the input current. Identical elements are employed in both cells to enhance input current ripple. Consequently, the self-inductances L_1 and L_2 hold the same values, thereby being denoted as L in the paper. The operational principle of the interleaved boost converter changes based on the circuit's duty cycle. When the duty cycle remains below 0.5, switches are operated sequentially, with no temporal overlap. While one phase supplies the load, in the other,

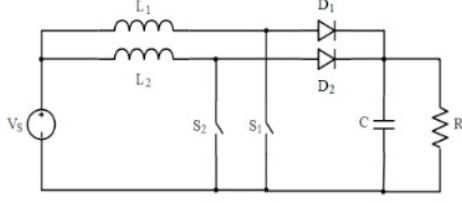


Fig. 1. Topology of the classical interleaved boost converter

the inductor restores energy. Therefore, the total frequency of the circuit, depending on the number of phases, is twice the switching frequency. Winding currents cancel each other out owing to the 180° phase shift between them, resulting in minimized input current ripple, as illustrated in Fig. 2.

Since the inductors are not coupled, the voltage and current waveforms are solely influenced by the states of the switches, leading to the same phase current ripple expression as in the conventional boost converter. The expression of phase current ripple given in (1) where V_s is the input voltage, D is the duty cycle and f_s is the switching frequency.

$$\Delta i_L = \frac{V_s \cdot D}{L \cdot f_s} \quad (1)$$

The expressions for the input current ripple can be derived as in (2).

$$\Delta i_{in} = \begin{cases} \frac{V_s}{L \cdot f_s} \cdot \frac{(1-2D) \cdot D}{(1-D)}, & 0 \leq D \leq 0.5 \\ \frac{V_s}{L \cdot f_s} \cdot (2 \cdot D - 1), & 0.5 \leq D \leq 1 \end{cases} \quad (2)$$

The input current ripple reduces as the duty cycle approaches 0.5, and in theory, the ripple reaches zero at this rate of duty cycle.

2.2. Interleaved Boost Converter with Coupled Inductors

In this section, the impacts of direct and inverse coupling on the topology will be elaborated upon, accompanied by their re-

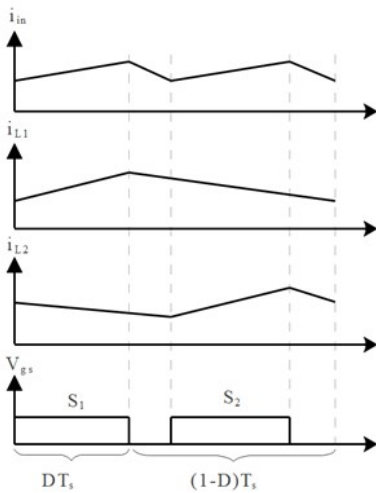


Fig. 2. Current waveforms of the uncoupled IBC according to gating triggers for $D < 0.5$

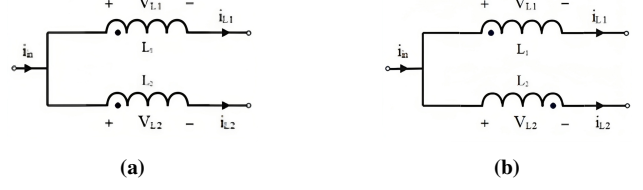


Fig. 3. Notation of coupled inductors: (a) directly coupled; (b) inversely coupled

spective equations. The coupling configurations, based on the orientations of the windings, are presented in Fig. 3.

The integration of magnetically coupled inductors into the IBC results in alterations to the voltage-current equations of the inductors as in (3) and (4) where V_1 and V_2 are the applied voltages and M is the mutual inductance.

$$V_1 = L_1 \cdot \frac{di_1}{dt} \pm M \cdot \frac{di_2}{dt} \quad (3)$$

$$V_2 = L_2 \cdot \frac{di_2}{dt} \pm M \cdot \frac{di_1}{dt} \quad (4)$$

For positive M values, in direct coupling, the sign preceding M is positive (+), while in inverse coupling, it is negative (-). Therefore, the coupling coefficient, represented below, is $k > 0$ for direct coupling, and $k < 0$ for inverse coupling.

$$k = \frac{M}{L} \quad (5)$$

This coefficient indicates the level of interaction between the coils. The influence of coupling on inductor current waveforms is depicted in Fig. 4. The variation in inductor currents is dependent not only on the associated switch but also on the other switch due to magnetic effects. The equivalent inductances are derived as given in (6)-(8).

$$L_{eq1} = \frac{L^2 - M^2}{L + \frac{D}{1-D} \cdot M} \quad (6)$$

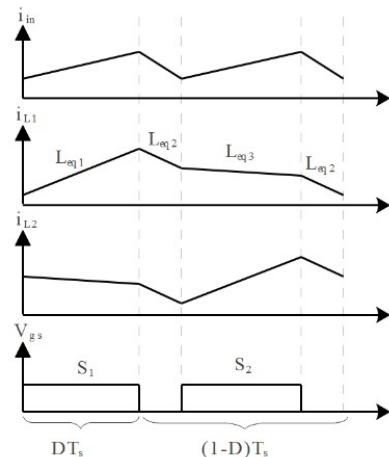


Fig. 4. Current waveforms of the coupled IBC according to gating triggers for $D < 0.5$

$$L_{eq2} = L + M \quad (7)$$

$$L_{eq3} = \frac{L^2 - M^2}{L + \frac{1-D}{D} \cdot M} \quad (8)$$

To obtain inductance equation for inversely coupled inductors, all M values have to be replaced by $-M$. The phase current ripple and input current ripple are given in (9) and (10), respectively.

$$\Delta i_{L_{cp}} = \frac{V_S \cdot D \cdot (k \cdot D - D + 1)}{L \cdot f_s \cdot (1 - D) \cdot (1 - k^2)} \quad (9)$$

$$\Delta i_{in_{cp}} = \begin{cases} \frac{V_S \cdot (1-2 \cdot D) \cdot D}{L \cdot f_s \cdot (1-D) \cdot (1+k)}, & 0 \leq D \leq 0.5 \\ \frac{V_S \cdot (2 \cdot D - 1)}{L \cdot f_s \cdot (1+k)}, & 0.5 \leq D \leq 1 \end{cases} \quad (10)$$

As evidenced from these equations, it is observed that depending on the sign of k , the input current ripple is lower in direct coupling, whereas the phase current ripple is lower in inverse coupling.

3. Magnetic Analysis and Loss Model of Coupled/Uncoupled Inductors

In this section, the magnetic design equations of uncoupled and coupled inductors are presented, and their loss formula are derived. Uncoupled inductors are shown in Fig. 5.

In order to calculate core losses, maximum and minimum magnetic field strengths for uncoupled inductors can be calculated by (11) and (12) where N is the number of turns, I_{max} is the maximum phase current, I_{min} is the minimum phase current and l_e is the magnetic path length. The coupled inductor cores and flux distributions are illustrated in Fig. 6.

$$H_{max} = 4 \cdot \pi \cdot \frac{N \cdot I_{max}}{l_e} \quad (11)$$

$$H_{min} = 4 \cdot \pi \cdot \frac{N \cdot I_{min}}{l_e} \quad (12)$$

The magnetic flux equations are given in (13) and (14) where ϕ_{lk1} and ϕ_{lk2} are the leakage fluxes, ϕ_{21} and ϕ_{12} are the flux linkages between windings due to i_{L1} and i_{L2} , respectively.

$$\Phi_{11} = \Phi_{21} + \Phi_{lk1} \quad (13)$$

$$\Phi_{22} = \Phi_{12} + \Phi_{lk2} \quad (14)$$

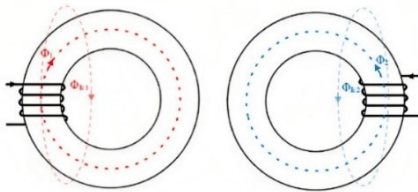


Fig. 5. Magnetic flux distribution of the uncoupled inductors

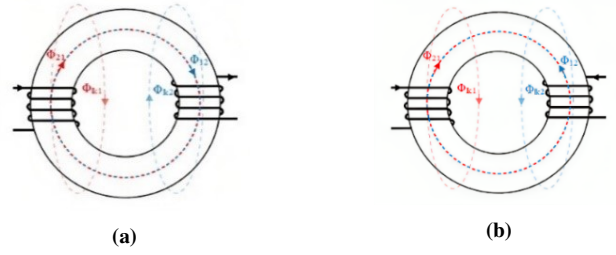


Fig. 6. Magnetic flux distribution (a) directly coupled; (b) inversely coupled

The total flux in the magnetic core is

$$\Phi_m = \Phi_{21} \pm \Phi_{12} \quad (15)$$

The sign is positive (+) in direct coupling and negative (-) in inverse coupling. This difference in flux can be seen in Fig. 7.

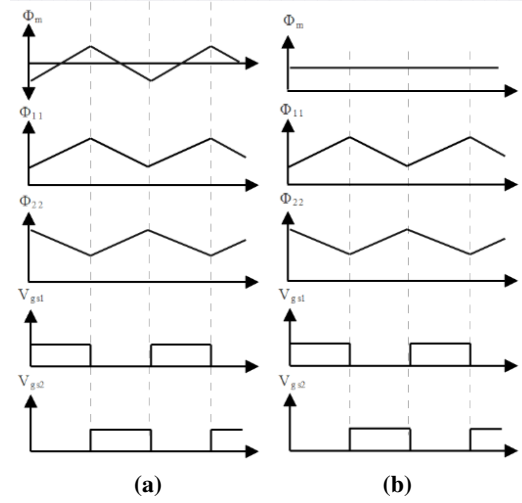


Fig. 7. Flux waveforms of (a) inversely coupled inductor; (b) directly coupled inductor at $D = 0.5$

In direct coupling, the AC magnetic field strength is absent since the ripple flux is cancelled at 0.5 duty cycle. Therefore, the generated magnetic field strength is as follows

$$H_{dc} = 4 \cdot \pi \cdot \frac{2 \cdot N \cdot I_L}{l_e} \quad (16)$$

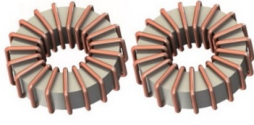
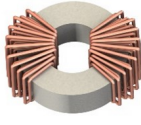
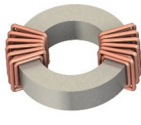
Due to the zero ΔB , there should theoretically be no core loss in a directly coupled inductor. Due to the negative flux alternans, the DC component of H does not exist in inverse coupling. Hence, the peak value of H points in inverse coupling are as follows

$$H_{max} = 2 \cdot \pi \cdot \frac{N \cdot \Delta i_L}{l_e} \quad (17)$$

$$H_{min} = -2 \cdot \pi \cdot \frac{N \cdot \Delta i_L}{l_e} \quad (18)$$

The corresponding B_{max} and B_{min} values are calculated using the equation provided by the core manufacturer. Core loss is determined by (19), where V_e is the volume of the core, ΔB is AC flux density and a, b, c are coefficients which are given by the manu-

Table 1. Specifications for uncoupled and loosely coupled inductors

Items	Uncoupled Inductor	Directly Coupled Inductor	Inversely Coupled Inductor
Schematic View	 (19 turns)	 (28 turns)	 (13 turns)
Core Part Number	Magnetics 79074	Magnetics 78072	Magnetics 79110
Inductance (μH)	21.62	Self-L 108.10 Mutual-L 86.48	Self-L 12.01 Mutual-L 9.61
Coupling Coefficient	–	0.8	-0.8
Volume (liter)	0.284	0.117	0.065

facturer.

$$P_{\text{core}} = a \cdot \left(\frac{\Delta B}{2} \right)^b \cdot f_s^c \cdot V_e \quad (19)$$

DC resistance of the winding is calculated in (20), where ρ_w is the resistivity of the conductor, MLT is the mean length turn and A_w is the cross-section of the winding.

$$R_{dc} = \frac{N \cdot \rho_w \cdot MLT}{A_w} \quad (20)$$

The skin effect, which leads to a rise in the resistivity of a conductor at high frequencies, can be calculated by (21). This value is used to determine the AC resistance in (22) [17].

$$\delta = \frac{1}{\sqrt{\pi \cdot f_s \cdot \mu \cdot \sigma}} \quad (21)$$

$$R_{ac} = R_{dc} \cdot \left[1 + \frac{(r/\delta)^4}{48 + 0.8 \cdot (r/\delta)^4} \right] \quad (22)$$

After obtaining AC resistance, the copper loss can be calculated in (23) where $I_{L_{rms}}$ is RMS phase current. The total loss is revised, according to the number of winding and cores.

$$P_{cu} = R_{ac} \cdot (I_{L_{rms}})^2 \quad (23)$$

To limit the temperature-rise in topology, thermal resistance is calculated using the empirical formula in (24) [17]. As a design specification, the maximum temperature rise is restricted to 90°C. Then, the maximum dissipation caused by the temperature rise is determined in (25).

$$R_{\theta} = \frac{0.06}{\sqrt{V_c}} \quad (24)$$

$$P_{\text{max}} = \frac{\Delta T}{R_{\theta}} \quad (25)$$

4. Results

Three 3700 W two-phase IBC converters with an input voltage of 200V and an output voltage of 400V have been compared in terms of volume, loss, and cost. All three topologies have a 1 MHz switching frequency and 50% phase ripple current. In Table 1, specifications for uncoupled, and loosely coupled inductors are provided. Since coupling has an effect on ripple current, different inductance values are employed to achieve a 4.63 A phase ripple current. In directly coupled inductor, a larger inductance value is used compared to the uncoupled inductor, as direct coupling increases the phase ripple current. On the contrary, a smaller inductance value is used when inverse coupling is chosen, as it reduces the phase ripple current. It is proved theoretically that leakage inductance should be low in direct coupling in order to decrease input ripple current. However, low leakage inductance also leads to an increase in phase ripple current. Therefore, the choice of a coupling coefficient should be made based on the specific topology requirements. Core loss does not occur due to AC flux cancellation in a directly coupled inductor. Accordingly, despite using a larger inductance value, smaller cores can be chosen compared to uncoupled cores. The only factor affecting efficiency difference in the topologies is inductor loss since the current ripple is the same. As shown in Fig. 8, the directly coupled inductor is the most efficient, while the inversely coupled inductor is the smallest inductor.

In Table 2, the presented data encompasses the outcomes of analytical core loss computations alongside the findings obtained through FEM analysis. The observed disparities in the outcomes can be attributed to the modestly heightened magnetic flux ripple within the FEM analysis, arising as a consequence of the relaxed coupling conditions.

Table 2. Comparison of core losses of uncoupled and loosely coupled inductors

Items	Uncoupled inductor	Directly coupled inductor	Inversely coupled inductor
	P_{core} (W)	P_{core} (W)	P_{core} (W)
Analytical	6.940	0	5.410
FEM Analysis	6.960	1.330	6.114

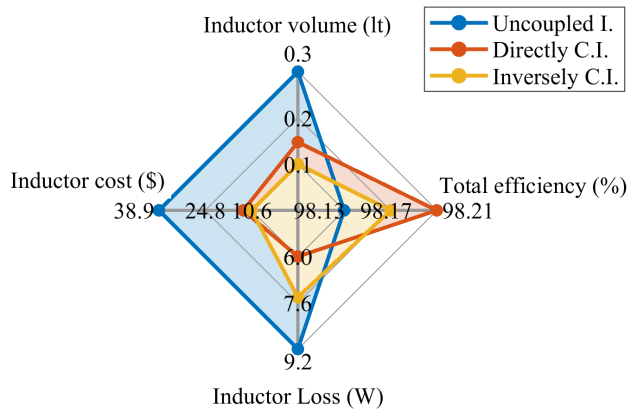


Fig. 8. Radar chart for cost, loss and efficiency and inductor volume of IBC topologies

5. Conclusions

In the present study, a comprehensive comparative analysis was undertaken to evaluate the performance of a two-phase interleaved boost converter employing coupled and uncoupled inductors under identical operational conditions. Specifically, the investigation was conducted at a switching frequency of 1 MHz with a 50% phase ripple current. The inductors were classified into three categories: uncoupled, directly coupled, and inversely coupled, and their attributes in terms of volume, efficiency, and cost were scrutinized. The outcomes of this study reveal that the highest level of efficiency, quantified at 98.21%, was attained by the directly coupled inductor. Analytically, it was discerned that both coupling methods, whether direct or inverse, offered significant advantages over the uncoupled configuration in terms of losses, volume reduction, and cost savings. Furthermore, the inversely coupled magnetic core exhibited a diminished DC flux, rendering it particularly suitable for high-power applications where concerns regarding core saturation are prevalent. However, in terms of efficiency, direct coupling demonstrated superiority over inverse coupling due to its lower core loss attributable to reduced AC flux. Significantly, using directly coupled inductors resulted in a 58.8% reduction in volume, while using inversely coupled inductor resulted in a 77% reduction in volume.

6. References

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