Total volume evaluation of high-power density non-isolated DC–DC converters with integrated magnetics for electric vehicles

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Abstract: One of the main problems in electric vehicles is the volume of their electrical systems because their bulky components carry additional mass and high cost to the total system. On this topic, interleaving-phases and magnetic coupling techniques have been reported as effective methods for increasing the power density of the DC–DC converters that work between the storage unit and the motor inverter. In that sense, a volume assessment of these topologies would provide a better understanding of the problems to be faced when an electric power train is designed. In this paper, a volume modelling methodology is introduced with the purpose of comparing four different DC–DC converters: Single-Phase Boost, Two-Phase Interleaved with non-coupled inductors, Loosely Coupled Inductor (LCI), and Integrated Winding Coupled Inductor (IWCI). The analysis considers the volume of magnetic components, power devices (conventional and next-generation), cooling devices and capacitors. The methodology can be used as a part of an optimization procedure to minimize the volume of DC–DC converters. Conclusively, LCI and IWCI were found effective to miniaturize power converters with a power density of 8.4 W/cc and 9.66 W/cc, respectively. Moreover, a maximum efficiency of 98.04% and 97.61% was obtained for a 1kW LCI and IWCI prototypes, respectively.

1 Introduction

Recently, high-power density DC–DC converters have attracted considerable attention for automotive applications because they can reduce the volume and mass of power converters in electric vehicles (EVs), hybrid electric vehicles (HEVs), and fuel cell electric vehicles [1–4]. Usually, an electric circuit with bulky and heavy components might be associated with low levels of power and autonomy in these vehicles [5, 6].

Consequently, downsizing of the electric powertrain in EVs and HEVs is required to increase the performance of these vehicles. Specifically, if the DC–DC converter that works between the storage unit and the motor inverter is downsized, the energy from the storage devices will be used in a greater proportion to supply the load instead of the power losses generated by the motion of the extra volume and mass [7, 8].

In fact, many converter topologies reported as effective for electric mobility applications have been proposed [1, 2, 4, 7–17]. Each topology offers different advantages for the vehicle performance. However, a volume comparison of these topologies is required to analyse the power density of each one and thereby raise a design criterion to downsize the electric power train of the vehicle.

In this study, the volume and efficiency of interleaving-phases and magnetic coupling techniques, usually reported as effective for EVs, are evaluated. Four different topologies were compared: single-phase boost converter (single phase) [11], two-phase interleaved boost converter with non-coupled inductors (non-coupled) [12, 13], two-phase interleaved boost converter with loosely coupled inductor (LCI) [14–16], and two-phase interleaved boost converter with an integrated winding coupled inductor (IWCI) [17–19]. Fig. 1a shows the schematic of the single-phase converter and Fig. 1b shows the schematic and the inductor arrangement of the two-phase converters: non-coupled, LCI, and IWCI.

These topologies were selected because the interleaving-phases technique can downsize the output capacitor, i.e. a higher frequency is achieved by power transmission alternation in each phase [20]. In addition, the magnetic coupling technique is effective because the inductor current ripple presents a higher frequency behaviour, and DC fluxes can be cancelled due to the mutual induction effect [15].

The analysis is conducted in several steps: (i). Geometry sizing of each magnetic component is calculated considering the inductor model. (ii). Power losses of each magnetic component are calculated with the purpose of obtaining the efficiency of each inductor. (iii). Semiconductor power loss calculation is carried out for sizing the required cooling device needed to dissipate these losses. (iv) Finally, the total volume of each converter is evaluated and compared.

For this evaluation, the specifications of the power control unit (PCU) of the Toyota Prius III were selected [4]. In addition, the circuit configuration of Fig. 2a is considered for this evaluation, where each high-voltage cell is composed of devices with a rating of 650 V (emulating the Prius Case). This model is composed of 30 cells, 1 kW each, in parallel in order to achieve the Prius' PCU power.

Moreover, considering the safety conditions of the available facilities for the tests, it was decided to evaluate the circuit configuration of Fig. 2b where each cell is a scale model of 1/30 of the Prius' converter power and 1/3 of the voltage rating. These cells can be made using a single phase or interleaved two-phase choppers as shown in Figs. 2c and d. Table 1 reports the parameters of each low-voltage cell.

The case study results will suggest the qualitative advantages and disadvantages of the four topologies (Fig. 1) by the volume evaluation of their entire structure. Finally, it is important to mention that this paper is an extended version of [3].

Fig. 1 Magnetic components in boost converters
 (a) Single-phase boost converter,
 (b) Interleaved boost converters: non-coupled, LCI and IWCI,
 (c) Magnetic circuit of the LCI component,
 (d) Magnetic circuit of the IWCI component

Fig. 2 Circuit configuration for the DC-DC converter of an EV
 (a) Main circuit composed of high-voltage cells,
 (b) Main circuit composed of cascaded low-voltage cells,
 (c) Chopper cell for the single-phase topology,
 (d) Chopper cell for the interleaved two-phase topologies

Table 1 Converter parameters

Parameters	Value
input voltage, V	80
output voltage, V	200
power, kW	1
switching frequency, kHz	50
input ripple current, %	20
output ripple voltage, %	0.1

2 Inductor sizing

To conduct the volume evaluation of the selected topologies, inductor volume sizing and loss calculations are required. Thus, the definition of the core and winding geometries is presented as the base of the volume and power loss analysis.

2.1 Core size

Volume analysis of the inductors of each topology is carried out based on the core modelling with the geometries presented in

Fig. 3 Inductor geometries

(a) Core geometry for single-phase and interleaved converters, (b) Core geometry for LCI and IWCI converters, (c) Winding geometry, (d) Winding resistances geometry

Fig. 3. Usually, non-coupled inductors can use two-leg cores (conventionally CC, CI or U cores) or in some cases three-leg cores. Fig. 3a shows the dimensions of the geometry for non-coupled inductors, and, Fig. 3b shows the geometry dimensions for the three-leg cores (usually EE, EI, EC or EER cores) used in the LCI and IWCI converter. These types of geometries can include a central leg in order to avoid magnetic saturation for certain cases.

With the purpose of simplifying the calculation of the dimensions and thereby the volume of the core, most of the dimensions are set according to the sectional area of the core A_e . Moreover, the sectional areas and the window areas are assumed as squares for convenience in the calculation.

The selected core material for this evaluation example is a TDK ferrite of reference PC40, with a saturation flux density of 380 mT at 100°C, a remanent flux density of 125 mT, and a relative magnetic permeability of 2300. Consequently, for these analyses, we set a maximum flux density of 250 mT. This ferrite material was chosen due to its outstanding characteristics in power losses per volume as it is presented in a previous study where amorphous, nanocrystalline and other ferrite materials are compared [9].

2.2 Core losses

Core losses are mainly dependent on the eddy currents and the hysteresis process, and they can be calculated by the well-known Steinmetz equation (SE) [20–23]. However, this calculation method is limited because it is valid only under a sinusoidal excitation condition. For this problem, the improved generalised Steinmetz equation was proposed [21]. In this context, using the SE parameters of the PC40 material, the core losses can be calculated as follows:

$$P_{cv} = \frac{1}{T_{sw}} \int_0^{T_{sw}} k_i \left| \frac{dB}{dt} \right|^\alpha (\Delta B)^\beta dt \quad (1)$$

$$k_i = \frac{k}{(2\pi)^{\alpha-1} \int_0^{2\pi} |\cos \theta|^{\alpha-2} d\theta} \quad (2)$$

where dB/dt is the slope of the flux density, ΔB is the peak-to-peak flux density, T_{sw} is the switching period, and V_c is the volume of the core. k , α , β are the Steinmetz parameters obtained from the datasheets of the PC40 core material. In the example case, $k = 4.5 \times 10^{-14}$, $\alpha = 1.55$, and $\beta = 2.5$. k_i can be calculated by applying these parameters to (2), and the core losses per volume are obtained from (1). The complete derivation of the core losses is shown in [20, 24].

2.3 Winding size

Winding volume is calculated to complete the total inductor volume. This analysis is conducted on the base of the winding geometry illustrated in Fig. 3c. As well as the core geometry description, winding geometry was set as squared for convenience in the calculation. In addition, the winding volume is calculated in accordance with the sectional area of the surrounded core, see Fig. 3.

2.4 Winding losses

Calculation of the winding losses is needed to complete the inductor efficiency analysis. These losses are generated by the DC resistance of the total winding and the AC resistance affected by the skin-effect and the proximity effect. Fig. 3d shows the geometry of each resistance. Winding losses are derived as follows:

$$P_{winding} = R_{DC} I_L^2 + R_{AC} \Delta I_{Lrms}^2 \quad (3)$$

In this context, the DC resistance is dependent on the length of the winding, the sectional area of the wire, and the wire material. R_{DC} and R_{AC} are the DC and AC (at high frequency) resistance, respectively, I_L is the DC component of the inductor current, and ΔI_{Lrms} is the effective value of the inductor current ripple. In addition, the AC resistance, where the skin effect is considered, can be calculated as

$$R_{AC} = \rho \cdot \frac{l}{\pi(d/2)^2 - \pi(d/2 - \delta)^2} = \rho \cdot \frac{l}{\pi\delta(d - \delta)} \quad (4)$$

$$\delta = \sqrt{\frac{\rho}{\pi\mu_0 f_{sw}}} \quad (5)$$

where ρ is the resistivity of the winding materials, d is the diameter of the wire, l is the winding length, δ is the skin depth, and μ_0 is the permeability of free space.

Moreover, when an inductor is wound with two or more layers as well as when there is current flowing through an independent conductor, the proximity effect occurs and it affects the AC resistance. In this study, all inductors are designed and experimentally tested using only one layer. The AC resistance calculation considering the proximity effect generated by nearby conductors is presented in [25]. From this calculation, the proximity effect of the four magnetic components is minimal in comparison to the skin and joule effects.

In fact, as single- and two-phase topologies are evaluated, magnetic components present two different switching frequencies (50 kHz in the single-phase and 100 kHz in the two-phase converters). Therefore, due to the skin effect, AC copper losses are higher in the two-phase converters. Nonetheless, in all four converters, the DC copper losses are the biggest contributor to the total copper losses.

3 Inductor modelling

Regarding the volume comparison, the behaviour of the magnetic flux that flows in each magnetic component of the four converters is analysed. The maximum flux intensity is defined as

$$\varphi_{\max} = \varphi_{\text{DC}} + \frac{\varphi_{\text{AC}}}{2} \quad (6)$$

where Φ_{DC} and Φ_{AC} are the average and the peak-to-peak magnetic fluxes in the core. The derivation of the maximum flux in each of the four topologies is presented in [11–19].

The required sectional area can be calculated with the derived magnetic flux of each topology and the maximum magnetic flux density of the selected material ($A_e = \Phi_{\max}/B_{\max}$). In the case of the four topologies evaluated in this study, their maximum magnetic flux density has been previously derived and presented in [12–19].

3.1 Single-phase converter

As in Fig. 1a, the single-phase boost converter has only one inductor with one winding of N turns. Based on the overall modelling presented in [24], it is possible to derive the sectional area calculation as follows:

$$A_e = \frac{1}{B_{\max}} \left[\frac{(I_L + \Delta I_L/2)V_{\text{in}}D}{\Delta I_L N f_{\text{sw}}} \right] \quad (7)$$

where ΔI_L is the ripple current through the inductor, V_{in} is the input voltage, and D is the duty cycle.

3.2 Interleaved converter with non-coupled inductors

Fig. 1b shows the two-phase interleaved boost converter with non-coupled inductors. Each inductor has one winding of N turns. Consequently, as it is reported in many studies [12, 13], the operating principle of this topology is the same as the conventional single-phase boost converter except for the phase-shift in the switching process of the switches. Therefore, the sectional area calculation in function of the magnetic flux is modelled in [12] and represented as follows:

$$A_e = \frac{1}{B_{\max}} \left[\frac{(2I_{\text{ph}} + \Delta I_{\text{ph}})V_{\text{in}}D}{2\Delta I_{\text{ph}} N f_{\text{sw}}} \right] \quad (8)$$

where I_{ph} and ΔI_{ph} are the average and ripple current through each winding, respectively.

3.3 LCI converter

The modelling of coupled inductors in interleaved boost converters is more complicated than the one of conventional topologies. As in Figs. 1b and 3, the LCI is composed of one core of three legs and two windings, each one of N turns [14, 15]. In this context, the sectional area of the external legs of the core as a function of the duty cycle can be calculated from the maximum flux density reported in [14, 15] as follows:

$$A_e = \frac{1}{B_{\max}} \left[\frac{NI_{\text{ph}}}{(1 + 2\alpha)R_{\text{mo}}} + \frac{1}{2} \frac{V_{\text{in}}D}{Nf_{\text{sw}}} \right] \quad (9)$$

where R_{mo} is the magnetic reluctance of the external legs, and α is defined as the ratio of $R_{\text{mc}}/R_{\text{mo}}$, where R_{mc} is the reluctance of the

central leg. Fig. 1c shows the magnetic model of the LCI magnetic component.

Considering the behaviour of the magnetic flux in the central leg, and based on the maximum flux density derived in [14, 15], (10) shows the calculation of the sectional area when the duty cycle is lower than 0.5. In the same way, (11) shows the calculation for the case of duty cycles higher than 0.5. It is important to mention that two-phase converters have different behaviours at duty cycles lower and higher than 50%, and thereby two possible sectional areas can be calculated.

$$A_c = \frac{1}{B_{\max}} \left[\frac{2NI_{\text{ph}}}{(1 + 2\alpha)R_{\text{mo}}} + \frac{1}{2} \frac{V_{\text{in}}D}{Nf_{\text{sw}}} \frac{1 - 2D}{1 - D} \right] \quad (10)$$

$$A_c = \frac{1}{B_{\max}} \left[\frac{2NI_{\text{ph}}}{(1 + 2\alpha)R_{\text{mo}}} + \frac{1}{2} \frac{V_{\text{in}}}{Nf_{\text{sw}}} (2D - 1) \right] \quad (11)$$

3.4 IWCI converter

Finally, the IWCI is composed of one magnetic core with three different windings installed in each leg. Fig. 1d shows the magnetic model of the IWCI component. The numbers of turns are N_1 for the central winding and N_2 for the external windings (see Figs. 1b and 3). Consequently, and based on the overall modelling of this magnetic component and the maximum magnetic flux density derivation presented in [17–19], (12) shows the sectional area of the external legs for the cases of duty cycles lower than 0.5, and (13) shows the sectional area calculation for the external legs when the duty cycle is higher than 0.5. In this converter, the operating states are different for duty cycles lower or higher than 0.5, and thereby the sectional area calculation is different for both cases

$$A_e = \frac{1}{B_{\max}} \left[\frac{(1 + 2\beta)N_2I_{\text{ph}}}{(1 + 2\alpha)R_{\text{mo}}} + \frac{1}{2} \frac{V_{\text{in}}D}{(1 + 2\beta)N_2f_{\text{sw}}} \left(1 + \frac{\beta}{1 - D}\right) \right] \quad (12)$$

$$A_e = \frac{1}{B_{\max}} \left[\frac{(1 + 2\beta)N_2I_{\text{ph}}}{(1 + 2\alpha)R_{\text{mo}}} + \frac{1}{2} \frac{V_{\text{in}}D}{(1 + 2\beta)N_2f_{\text{sw}}} \left(1 + \frac{\beta}{D}\right) \right] \quad (13)$$

where β is defined as the ratio of N_1/N_2 and, for this volume study, it is set as $\beta = 1$. This value of β is chosen due to not only the convenience of having the same turn number because the height of the central and external legs is the same in regular EE cores, but also the opportunity of optimising the volume of the core, because each leg will have the same turn number and therefore the window utilisation is improved. Furthermore, (14) shows the calculation of the sectional area of the central leg when the duty cycle is lower than 0.5, and (15) shows the case for a duty cycle higher than 0.5

$$A_c = \frac{1}{B_{\max}} \left[\frac{2(1 + 2\beta)N_2I_{\text{ph}}}{(1 + 2\alpha)R_{\text{mo}}} + \frac{1}{2} \frac{V_{\text{in}}D}{(1 + 2\beta)N_2f_{\text{sw}}} \frac{1 - 2D}{1 - D} \right] \quad (14)$$

$$A_c = \frac{1}{B_{\max}} \left[\frac{2(1 + 2\beta)N_2I_{\text{ph}}}{(1 + 2\alpha)R_{\text{mo}}} + \frac{1}{2} \frac{V_{\text{in}}}{(1 + 2\beta)N_2f_{\text{sw}}} (2D - 1) \right] \quad (15)$$

3.5 Volume and losses comparison

A comparison among the selected topologies was performed considering the geometric models of cores and windings, their loss models, and the magnetic flux modelling of each topology. This was possible by solving (7)–(15) with the evaluation of different turn number, the calculation of each variable with the parameters defined in Table 1, and having in mind the temperature limitation of 100°C for the selected ferrite material.

Fig. 4 shows the comparison between the volume of each inductor (or pair of inductors in the case of the interleaved converter) and their power losses. This comparison was made considering a varying turn number in each inductor, because the turn number influences the core size and the inductor losses (both core and copper losses). The values of volume presented in these figures correspond to the net value of windings and cores, i.e. the volume of the core windows is not considered.

Fig. 4 Inductor volume versus inductor losses

Table 2 Power semiconductor characteristics

Transistors	Mosfet	Super junction	Diodes	Si diode	SiC diode
$R_{DS(on)}$, m Ω^a	46.2	27.5	V_F , V ^a	0.9	1.35
$R_{DS(on)}$, m Ω^b	45.9	27	V_F , V ^b	0.8	1.18
C_{iss} pF @200 V	9600	9900			
C_{oss} pF @200 V	350	190	Q_r , nC	65	—
t_{rise} , ns	52	27	t_{rr} , ns	50	—
t_{fall} , ns	81	5			

^a $I_D = 12.5$ A, $T_J = 100^\circ\text{C}$.

^b $I_D = 6.25$ A, $T_J = 100^\circ\text{C}$.

The dots on each line correspond to the value of the turn number. In addition, the increase of the turn number produces a reduction in the total inductor volume in each converter. This behaviour is generated by the winding-core dependency, where the lower the turn number, the larger the core size to accomplish the filtering requirements. Nevertheless, there is an observable limitation in the figure, which is more evident in the IWCI case. This limitation corresponds to the case where the inductor cannot be downsized anymore even if there is space in the window for more turns. This point of minimum volume corresponds to the case where the magnetic core reaches its maximum power allowance.

Every dot in each line of Fig. 4 (that corresponds to a turn number) is assumed with a constant diameter of the winding wire that is suitable for the current ratings of each inductor. On the other hand, every dot in each line requires a different sectional area of the core. When the turn number is increased, the sectional area is reduced in order to comply with (7)–(15).

Fig. 4 also shows that LCI and IWCI topologies offer low inductor volume. IWCI can obtain higher downsizing in comparison with the LCI, but this miniaturisation can lead to the increase in the inductor losses, this is because of the trade-off between the core size and the inductor losses. Additionally, it is possible to see the effectiveness of the magnetic coupling technique compared with the non-coupled inductors. Also, the performance of the LCI converter is remarkable because it presents the lowest power losses with small volume for this case study. However, IWCI inductor presents high power losses because of its topology; its central winding conducts all the input current, carrying larger copper losses than the other windings where the current is divided.

Comparing the blue and red (upper) lines with the green and magenta (lower) lines, it is possible to infer that magnetic coupling technique is an effective technique for downsizing and, in some cases, for increasing the efficiency of magnetic components.

4 Volume of cooling devices

To calculate the volume of the cooling devices, the estimation of the losses in semiconductor devices and the heat sink modelling are required.

4.1 Losses in semiconductor devices

Power losses in semiconductor devices can be classified into: switching and conduction losses that are dependent on the equivalent series resistance (ESR), voltage drops, parasitic capacitances, and parasitic inductances, among others [25–27].

In fact, losses in switches are produced by the static drain-source on-state resistance $R_{DS(on)}$, the transistor input capacitance, the output capacitance, and the switching transition process. In addition, losses in diodes are produced by the diode voltage drop, the diode resistance, and the reverse recovery when the converter operates in continuous conduction mode. The overall power loss model is explained in detail in [24]. Conventional silicon and next-generation devices (super junction and SiC devices) were chosen for evaluating their losses and thereby the volume of the required cooling devices. Table 2 shows the specifications of the selected semiconductors, considering the voltage and current stresses of each topology. All the selected devices have a TO-247 package and a voltage rating of 650 V for the super-junction devices and 600 V for the other devices.

Based on the power loss model of [24], the parameters of Table 1, and the power semiconductors of Table 2, the individual power losses of the transistors and diodes of each converter are displayed in Table 3.

4.2 Heat sink modelling

Heat sinks were selected for this study because they are conventionally used for low-power dissipation. Therefore, to model the semiconductor cooling device, it is necessary to calculate the required thermal resistance from the cooling device to the air [28]. This resistance can be calculated from the thermal circuit presented in Fig. 5a.

Table 3 Power semiconductors losses

Single-phase		Interleaved		LCI		IWCI	
Transistor losses, W							
Si	S-Jun	Si	S-Jun	Si	S-Jun	Si	S-Jun
13.11	4.85	5.7	1.91	6.04	2.11	6.04	2.11
Diode losses, W							
Si	SiC	Si	SiC	Si	SiC	Si	SiC
13.28	7.75	5.78	3.2	5.78	3.2	5.78	3.2

Fig. 5 Thermal circuit and heat sink

(a) Circuit, (b) Heat sink geometry

Table 4 Heat sink parameters and dimensions

Parameters	Value	Dimensions, mm	Value
h , W/(m ² °C)	25	s	4
λ_{HS} , W/(m°°C)	237	L	25
V , m ³ /s	0.006	t	1
ρ_{AIR} , kg/m ³	0.99	d	2
$c_{p,AIR}$, J/(kg°°C)	1010	c	10

The junction temperature T_J (defined by the manufacturer) can be calculated using (16).

$$T_J = T_{AMB} + (R_{\phi JC} + R_{\phi CH} + R_{\phi HA})P_{Loss} \quad (16)$$

where T_{AMB} is the ambient temperature (usually 50°C for the ambient within the converter [29]), $R_{\phi JC}$ is the thermal resistance from the junction to the semiconductor's case, $R_{\phi CH}$ is the thermal resistance from the case to the heat sink (usually neglected due to its very small value), $R_{\phi HA}$ is the thermal resistance from the heat sink to the air, and P_{Loss} is the dissipated power in each power device. Thus, using (16), it is possible to calculate the required heat sink thermal resistance.

All the selected power devices have a maximum junction temperature of 175°C; however, the heat sink volume calculation is conducted assuming a maximum junction temperature of 100°C with the purpose of protecting the power devices and preventing high ambient temperature rises.

Once the heat sink's thermal resistance from the base plate surface to the ambient is calculated, the next step is to model the size of the heat sink. Fig. 5b shows the definitions of the heat sink geometry. Based on [30–32], it is possible to derive the thermal resistance of the heat sink in relation to its geometry as follows:

$$R_{\phi HA} = \frac{1}{n} \left[R_{th,d} + \frac{1}{2} (R_{th,FIN} + R_{th,A}) \right] + \frac{0.5}{\rho_{air} c_{p,air} V} \quad (17)$$

where n is the number of the channels, ρ_{air} is the air density, $c_{p,air}$ is the specific thermal capacitance of air, V is the air volume flow, and $R_{th,d}$ is the thermal resistance of the heat sink base of height d . $R_{th,d}$ is calculated as follows:

$$R_{th,d} = \frac{nd}{A_{HS}\lambda_{HS}} \quad (18)$$

where A_{HS} is the size of the heat sink plate, and λ_{HS} is the thermal conductivity of the heat sink material (generally, heat sinks are manufactured with aluminium alloys). Additionally, $R_{th,FIN}$ is defined as the thermal resistance of the fins and is expressed as:

$$R_{th,FIN} = \frac{c}{t\lambda_{HS}} \quad (19)$$

where c , t , and L are the dimensions of the defined heat sink geometry. Finally, $R_{th,A}$ is the thermal resistance between the fin surface and the air channel:

$$R_{th,A} = \frac{1}{hLc} \quad (20)$$

where h is the convective heat transfer coefficient.

The required heat sink dimensions, and thereby its volume, can be calculated by solving (17) using (18)–(20). This calculation is made using the heat sink parameters shown in Table 4. Note that (17) presents several variables: s , L , t , d , c , and n . Based on several heat sinks, suitable for the TO-247 package and available in the market, the dimensions s , L , t , d , and c were selected as it is shown in Table 4. Using the calculated thermal resistance for the heat sinks, it is possible to derive the number of channels n that the heat sink needs, and therefore its volume is estimated.

The dimensions of Table 4 (columns 3 and 4) were set as constants with the purpose of varying only the number of channels (i.e. the thick) and the height of the heat sink. Therefore, it will be possible to see how the heat sink dimensions influence the comparison.

5 Volume comparison

5.1 Power devices

When next-generation devices are used, instead of conventional silicon semiconductors, a reduction in power losses and heat sinks volume is produced. It is important to mention that all power devices, either silicon or next-generation, have the same die area for a fair comparison purpose. Table 5 shows the volume of the heat sinks set (pair or single) needed to dissipate the losses of each individual semiconductor. This heat sinks set corresponds to one device in the case of the single phase, and two devices in the case of the other three topologies. Thus, the use of next-generation power devices can produce a reduction in the power losses and thereby in the heat sink volume up to 60% in comparison with the

Table 5 Heat sink volume

Single-phase		Interleaved		LCI		IWCI	
Transistor heat sink volume, cc							
Si	S-Jun	Si*	S-Jun*	Si*	S-Jun*	Si*	S-Jun*
7.18	2.79	6.46	2.54	6.81	2.75	6.81	2.75
CSPI, °C/(W L)							
18.74	17.59	17.86	15.09	17.9	15.44	17.9	15.44
Diode heat sink volume, cc							
Si	SiC	Si*	SiC*	Si*	SiC*	Si*	SiC*
6.36	3.7	5.52	3.27	5.52	3.27	5.52	3.27
CSPI, °C/(W L)							
18.66	18.03	17.56	16.13	17.56	16.13	17.56	16.13

Table 6 Total volume comparison

Component	Boost	Interleaved	LCI	IWCI
total volume, cc	564.5	687.6	76.7	54.6
inductor (or pair of), cc	509.0	654.2	43.1	21.0
heat sinks, cc	6.5	5.8	6.0	6.0
semiconductors, cc	6.0	6.0	6.0	6.0
capacitors, cc	43.0	21.5	21.5	21.5

Fig. 6 Core volume versus inductor losses

conventional silicon semiconductors, for the case of the defined 1 kW prototype.

In addition, to have a better understanding of the calculated volume, Table 5 also reports the Cooling System Performance Index (CSPI), defined as the power density capability of the cooling system, described in detail in [30]. From this comparison and the derivation of (17)–(20), it was possible to notice that heat sinks with larger effective area may further reduce the volume in comparison to those who have longer height.

5.2 Total volume

Based on the volume estimations of the inductors and the heat-sinks for power semiconductors described above, the total volume of the selected topologies under the defined parameters was calculated. This calculation was set in the operating point where the inductors of each converter present their lowest power losses (see Table 6). This comparison was calculated using the values of the super-junction Mosfet and the SiC diode, as well as their corresponding heat sinks. Moreover, conventional electrolytic capacitors were used for the comparison.

Fig. 4 and Table 6 show the opposition between inductor losses and inductor volume, i.e. first, the topology that offers the lowest power losses is the LCI converter, and second, LCI has a bigger volume in certain numbers of turns in comparison with IWCI. However, IWCI has much larger inductor loss for most of all numbers of turns in comparison with the other three topologies.

Conclusively, the IWCI converter exhibits the smallest volume, but for this case study, it presents the highest power losses.

6 Inductor size evaluation

In the previous sections, ideal cores with defined geometries (Fig. 3) have been modelled. These geometries were defined using squares for calculation convenience. However, in practice, it is difficult to find the exact core that fills the design parameters. In this context, there are two possibilities: to use a customised core that fulfils all the design requirements, resulting in an overcost due to the personalised core; or to use a core available in the market that can fulfil the requirements. Consequently, to validate the modelling presented so far and to compare the changes of inductor losses and volume in the defined geometries with cores available in the market, four different cores were selected to be compared with the results exhibited in Fig. 4. These cores were selected because their volume and effective sectional area fit into the calculated values of Fig. 4, they are fabricated with the selected core material (TDK ferrite of reference PC40), and they offer a convenient trade-off between efficiency and volume based on Fig. 4. In consequence, Fig. 6 shows the core volume of the selected cores for each inductor or pair of inductors in the case of the non-coupled interleaved converter.

Fig. 6 shows that non-coupled inductors (single-phase and interleaved) require large cores to obtain the required filtering. Therefore, the region of considerable large turn number (where the

Fig. 7 FEM results in Teslas
(a) Single phase, (b) Interleaved, (c) LCI, (d) IWCI

points represent a volume smaller than 200 cc) is suitable for this study because cores with volumes larger than 200 cc are not affordable (the region of low turn number, the upper side of Fig. 6). In this context, EC90 and EE90 cores (manufactured by TDK with reference PC40) are selected for the single-phase and the interleaved inductors, respectively. These cores were selected considering the analysis of Fig. 6, where their volume matches with the region of suitable core sizes. It is important to mention that the interleaved converter with non-coupled inductors (blue line in Fig. 6) uses two cores, obtaining a total core volume of 118.1 cc for the case of two EE90 cores. Additionally, for the case of magnetic coupled inductors (LCI and IWCI), they can be made with smaller cores. EE60 core was selected for the case of the LCI, and EE50 for the IWCI.

To validate this modelling procedure, a finite-element method (FEM) was conducted for each inductor to check the magnetic flux density of each core and corroborate the saturation absence. Fig. 7 shows the results of the FEM presenting the normal magnetic flux density in the surface of the cores, and the FEM results using slices of the cores to display the inner magnetic flux density. All the FEM results are presented in Teslas.

Based on these results, the inductor modelling is validated because none of the models exceed the defined as the maximum magnetic flux density, 250 mT. Also, taking into account the relationship between the flux density and the internal temperature in a magnetic core, this FEM validation shows the areas where possibly a temperature increase will occur.

Moreover, having in mind the concept of the low profile design presented in [33–35], it is possible to state that under a low thermal resistance condition, the temperature of the magnetic components does not exhibit a rise performance. Therefore, low profile coupled inductors are one of the effective way to reduce temperature rising. In this context, it is possible to see the difference between the flux of Figs. 7c and a.

7 Experimental validation

To validate the volume comparison presented above, an experimental verification was conducted. This validation was carried out considering the results presented in Fig. 6.

7.1 Inductors

As it was explained before, the volume comparison conducted in Section 5 was made using custom core geometries; however, only specific cores could be used for the experimental validation due to access restriction of geometries available in the market. In this context, the experimental tests were performed using the cores

evaluated in the previous section: EC90, EE90, EE60 and EE50 (Ferrites of reference PC40 manufactured by TDK). The prototypes were designed using the method illustrated in Sections 2 and 3.

7.2 Power devices

The power devices used for the prototypes were the super-junction Mosfet and the SiC diode presented and described in Table 2. This selection was made based on the higher efficiency performance of these components.

7.3 Heat sinks

As it is shown in Section 4, the heat sink modelling was made with custom geometries. Although the heat sink parameters and dimensions of Table 4 were chosen based on real parameters of regular heat sinks for TO-247 packages, the total estimated volume is slightly different from the heat sinks available in the market. Therefore, the selected heat sinks are described as follows: single-phase's Mosfet: 7.9 cc, 16.5°C/W; single-phase's diode: 9.4 cc, 14.2°C/W; all devices of the two-phase topologies (interleaved, LCI and IWCI): 4.9 cc, 22.2°C/W.

All the selected heat sinks present a thermal resistance slightly lower than the one calculated in Section 4. This means that the junction temperature will be lower than the designed one (100°C). In addition, the selected heat sinks present larger volumes than the designed ones because only the net volume (without window volumes) was considered in the analytical design of Section 4.

7.4 Capacitors

To select a suitable capacitor for the prototypes, multilayer ceramic capacitors (MLCC), metallised polypropylene film, and electrolytic capacitors were compared. The required capacitance is approximately 300 and 150 µF for the single-phase and the two-phase converters, respectively. Therefore, capacitors with a capacitance nearby to 50 µF were compared. It is important to mention that the frequency of each converter was considered, i.e. the capacitor of the single-phase converter operates at 50 kHz while the capacitors of the other converters operate at 100 kHz.

Table 7 shows the specifications of the selected capacitors. It is possible to highlight the low ESR of the film and the MLCC capacitors, the small volume of the MLCC and electrolytic capacitors, and the large ESR of the electrolytic capacitor.

Although the volume of the electrolytic and the MLCC capacitors is similar (4.02 versus 4.74 cc), the required PCB area of both capacitors is different (2.01 versus 11.84 cm²). Therefore, electrolytic capacitors were chosen because they require much less

Table 7 Capacitor comparison

Specification	Film	Electrolytic	MLCC $\times 2$
capacitance, μF	50	47	30×2
rated voltage, V	500	400	400
ESR, m Ω datasheet	4	—	1 each
ESR, m Ω measured	5.65	416	2.8 each
volume, cc	56.7	4.02	2.37×2
PCB area, cm^2	12.6	2.56 (square)	7.82×2

Fig. 8 Prototypes of the four converters
(a) Single phase, (b) Interleaved, (c) LCI, (d) IWCI

PCB area leading to a more compact prototype in comparison to the case of the MLCC.

7.5 Volume evaluation

Fig. 8 shows the prototypes of the four converters made with the selected components described before. The gate drivers were made using surface mount components with the purpose of reducing the volume. Fig. 8 shows the volume difference between the prototypes. Each figure presents a pie chart where the measured volume of each component group is presented. As in Table 6, it is confirmed that the inductor volume of the single-phase and the interleaved two-phase converters represents $>75\%$ of the total prototype volume.

As a matter of fact, the largest volume of the four prototypes was presented by the single-phase topology (335.3 cc). Note that in Table 6, the largest calculated volume was exhibited by the interleaved two-phase converter. Nevertheless, the analytical comparison presented in Section 5 was made using net values without considering dead spaces between the heat sinks or inside the inductors. In practice, the prototype of the single-phase converter presents the largest volume because the window volume (dead space) of the EC90 is much bigger than the one of the two EE90 cores (82 versus 19 cc, respectively). Conclusively, the interleaved two-phase topology is better than the single-phase converter, in volume terms because it is more compact.

Finally, having the measured volume of each topology (Fig. 8) and knowing that the prototypes were designed for 1 kW, it is possible to calculate the power density of each prototype as follows: Single-Phase: 2.98 W/cc; Interleaved: 3.36 W/cc; 8.4 W/cc; and IWCI: 9.66 W/cc.

In conclusion, the volume comparison and the sizing modelling, presented in Sections 3, 4 and 5, are validated. It was confirmed that IWCI offers the highest power density because of the interleaving phases and the magnetic coupling techniques.

7.6 Experimental results

Fig. 9a shows the experimental waveforms of the LCI converter tested with a 1 kW load. Blue and yellow signals correspond to the drain-source voltages of both phases (200 V) and the purple signal corresponds to the input current (2.5 A of ripple). It is important to highlight that Fig. 9a shows the current ripple without a great difference between their peaks, therefore it is validated the current balancing between the phases. Also, it was confirmed that magnetic saturation in the magnetic components did not occur and a stable operation was realised. Therefore, the accuracy of the inductor design is validated from the experimental results.

The efficiency of the four converters was measured with the conditions presented in Table 1, and a 98.05% of efficiency was observed at 1 kW for the LCI converter. It is important to mention that the single-phase converter presents a higher efficiency at low-load condition than the other converters. In the same way, IWCI

Fig. 9 Experimental results

(a) Experimental waveforms of the LCI prototype, (b) Efficiency measurement of the four converters, (c) Temperature rise in the Mosfets of the four converters

converter exhibited a maximum efficiency of 97.61%. Fig. 9b shows the efficiency of these converters with the output power scanning from 200 W until 1 kW.

In addition, Fig. 9c shows the temperature rise of the heat sinks attached to the Mosfets of each topology, where a maximum temperature of 74°C was measured after 10 min of testing the LCI converter and 85 in the case of the single phase. Although, the single-phase converter has bigger heat sinks, due to the higher current that each device flows, the steady state temperature is higher.

8 Conclusion

A volume modelling methodology of four DC–DC converter topologies, combining geometry sizing, inductor modelling, power loss evaluation, and heat sinks modelling of conventional and next-generation devices, was presented in this paper. For the 1 kW case study presented in this article, the IWCI converter, with a large turn number, offered the smallest volume in comparison with the other studied topologies. However, although the IWCI can reduce the size, it can lead to a magnetic loss increase. On the other hand, this research suggests that the LCI converter is effective for reducing the size and improving the efficiency.

It was confirmed that the use of next-generation power devices can reduce the power losses and thereby the heat sink volume up to 60% in comparison with the conventional silicon devices. Moreover, based on the experimental results, a 98.05% of efficiency and a power density of 8.4 W/cc was measured in the LCI prototype.

The proposed methodology can be used for a designer of DC–DC converters, intended to be applied in modular and cascaded converters for electric mobility applications, because it gives an overall understanding of the effect of each component characteristics of the entire converter volume and efficiency. Moreover, the methodology can be used as a part of an optimisation procedure of the converter, e.g. a multi-objective optimisation of the volume, efficiency, and temperature of the converter.

9 References

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