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Design of an Unbalanced High Power Three-Winding Planar Transformer for Electric Vehicle Application

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Abstract

This paper presents the study of a three-winding planar transformer usually required for Three-port Active Bridge converters. The transformer operates at high switching frequency under unbalanced power flow conditions and contains parallel layers to handle high current. The main difficulty is to integrate high and low voltages respecting the insulation exigencies and keeping a small number of PCB's with reduced thickness. Furthermore, the current sharing between parallel-connected layers is evaluated by using a new modified analytic method. This method is applied to different winding configurations and the results are validated by FEA (Finite Element Analysis) simulation.

1 Introduction

In order to increase converter's power density, planar transformer has become an interesting solution; due to its singularity and manufacturing process, it becomes a key component in power converter design. In addition, to reduce the number of components and increase the power density of Electric Vehicle power systems, multi-functional converters have become a frequent demand.

One of the recent designs for EV power systems is the use of a three-port converter used to integrate parts of the on-board charger (OBC) used to charge the propulsion battery; and the low/high voltage DC/DC converter (LDC) used to charge the auxiliary low voltage battery. One of the examples is presented in [1], where the power flow between the ports is controlled as in the Three-port active bridge converter. Normally, for these applications, the OBC is designed for 3.3 kW or 11 kW, while the LDC usually works with 2.2 kW or 5 kW. The focus of this paper is to use the planar transformer technology to design a three-winding transformer that integrates both structures as represented in Fig. 1. Due to the high difference between the voltage and power levels into the ports, the design of the transformer requires parallel winding to support the high current in the low voltage side.

In addition, nowadays the converters are being controlled at higher frequencies and an accurate prediction of power loss is critical for the application. In [2], the current sharing among parallel winding has been discussed and shows the importance of a balanced distribution to reduce power loss and temperature. The current balancing can be improved by using interleaved configurations, however, due to the minimum insulation thickness between low and high voltage sides required for Electric Vehicle applications, interleaved configurations associated with a high number of turns ratio could increase the difficulty in the use of planar transformers.

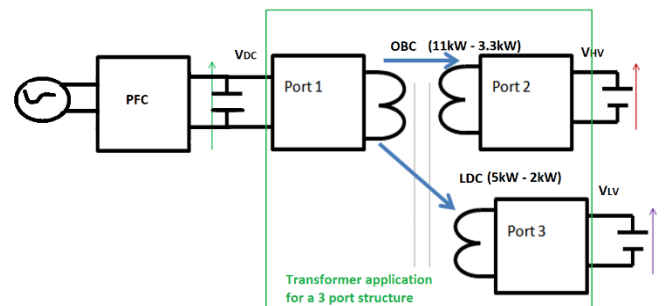


Fig. 1: Three-port converter integrating part of the OBC and the LDC converters in EV.

In this paper, three different design configurations are proposed for the three-winding planar transformer, operating for the OBC and the LDC converters. The three configurations are proposed to reduce the cost and facilitate the final manufacturing process of the unbalanced transformer.

In order to estimate a more precisely power loss on the conductor layers, a complement to the commonly used calculation method is presented in this text to help in the selection of the transformer with the best performance, taking care of the parallel winding current distribution problem. The method presented here is easily integrated into a program routine and can be applied for more than two-winding transformers or in the presence of an air gap.

This paper is organized as follows: Section 2 presents the calculation method to estimate the power loss distribution on the conductor layers of a planar structure. In section 3, is presented the design of the three-winding planar transformer for OBC and LDC for a specific power flow condition and the power losses are estimated by the method presented in section 2. In section 4, the results obtained are compared with Finite Element Analysis (FEA) results and an optimal configuration is identified.

2 The analytic approach to estimate power loss in the planar transformer

When it is necessary to use parallel-connected layers, by considering only Dowell assumptions [3], the conductor power loss is not well estimated. This occurs because the total current flowing in each conductor is considered unaltered and this is not true for parallel-connected layers. When the frequency increases the current sharing between the layers needs to be identified before the analysis of power losses. This difficulty becomes more evident in a more than two winding-transformer, as in the case of a three-winding transformer, or when the winding currents are not in phase.

Most of the analysis of power loss in planar transformer, considers that the frequency will only alter the current distribution across the conductors, but not the total current flowing in each conductor. Due to the difficulty in to estimate the current distribution, normally the problem is addressed to Finite Element Simulations. In [4] an analytic method is proposed to estimate power losses for

parallel winding connections and improve the transformer configuration, but it needs to use FEA at first to determine the self and mutual impedances. In [5] a method to determine static leakage inductances is presented and in [6] the frequency influence appears in the current distribution inside the copper layers, but since it is applied to a planar inductance design, it does not have the problem of parallel-connected layers or different phases.

The method presented here can be solved very quickly by a program routine and allows further improvements to an optimized version. It is a continuation of the work presented in [7].

The method can be explained by using the magnetic structure from Fig. 2. The N 's conductor layers are distributed in the schematic around the magnetic central-leg. A small air gap is added to the structure. Some assumptions are made: The magnetic flux strength is perpendicular to the central leg of the magnetic core; the core permeability is infinite and the gap is small to prevent that the field lines go too far from the central leg. At each conductor layer is considered a current i_k and a magnetic flux produced by this current, Φ_k .

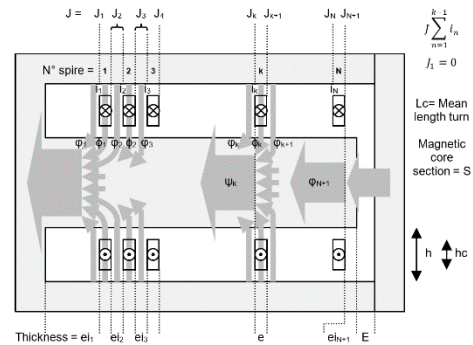


Fig. 2: The planar structure used as reference

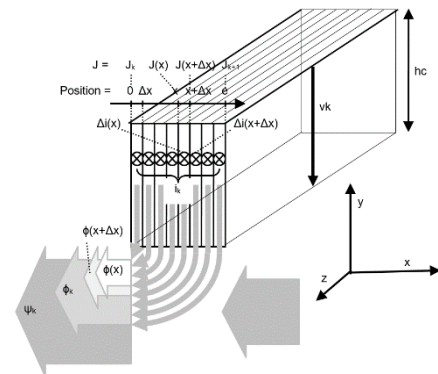


Fig. 3: A single layer divided into sub-sections

Outside the conductive region, a magneto motive force (MMF), J_k , is created and it is the image of the total current flowing into that conductor slice.

ϕ_k is the magnetic flux in the insulation layers between the conductors. The total magnetic flux just after the conductor number k is represented by ψ_k and is the contribution of all the layers placed on its left side.

In Fig. 3, we have a single layer divided into sub-sections. In each sub-section flows a current portion $\Delta i(x)$. The voltage drop on the x -axis between the portions x and $x+\Delta x$ is the equivalent variation of the magnetic flux, $\phi(x+\Delta x)$ and $\phi(x)$:

$$\Delta i(x + \Delta x) - \Delta i(x) = \frac{1}{r} \frac{d[\phi(x+\Delta x) - \phi(x)]}{dt} \quad (1)$$

$$J = \int \frac{B}{\mu_0} dy = \int \frac{\Delta \phi}{\mu_0 (\Delta L \Delta x)} dy = \frac{\Delta \phi}{\mu_0 \Delta x} \int \frac{dy}{\Delta L} \quad (2)$$

The resistance of each portion of the copper layer can be written as:

$$\frac{1}{\Delta r} = \int \frac{\Delta x}{\rho \Delta L} dy = \frac{\Delta x}{\rho} \int \frac{dy}{\Delta L} \quad (3)$$

Applying the Eq. (3) in Eq. (2) we obtain the magnetic flux that depends on the resistance:

$$\sum \Delta \phi = J \mu_0 \sum \frac{\Delta x}{\int \frac{dy}{\Delta L}} = J \frac{\mu_0 \Delta x^2}{\rho} r \quad (4)$$

The Eq. (1) and Eq. (4) are equivalents:

$$\frac{d\phi}{dt} = \frac{dJ}{dt} \frac{\mu_0 \Delta x^2 r}{\rho} = r [\Delta i(x + \Delta x) - \Delta i(x)] \quad (5)$$

Supposing sinusoidal current waveform, the solution for the MMF, J_k , is obtained from Eq. (6):

$$j\omega J \frac{\mu_0}{\rho} = \frac{[\Delta i(x + \Delta x) - \Delta i(x)]}{\Delta x^2} = \frac{d^2 J}{dx^2} \quad (6)$$

$$J = J_{Ak} \cdot e^{\frac{(1+j)x}{x_0}} + J_{Bk} \cdot e^{-\frac{(1+j)x}{x_0}}, \quad x_0 = \sqrt{\frac{2\rho}{\mu_0 \omega}} \quad (7)$$

The coefficients, J_{Ak} and J_{Bk} in Eq. (7), are determined by the conditions at $x=0$ and $x=e$, J_k and J_{k+1} , respectively, where $\alpha=(1+j)/x_0$.

$$J_{Ak} = \frac{J_{k+1} - J_k \cdot e^{-\alpha e}}{e^{\alpha e} - e^{-\alpha e}} \quad J_{Bk} = \frac{-J_k e^{\alpha e} - J_{k+1}}{e^{\alpha e} - e^{-\alpha e}} \quad (8)$$

The voltage drop in each portion of the conductor layer can now be calculated:

$$v_k = j\omega \psi_k + r \frac{dJ}{dx} \bigg|_{x=J_k-J_{k+1}} \quad (9)$$

$$v_k = j\omega \psi_k + r \Delta x \alpha (J_{Ak} - J_{Bk}) \quad (10)$$

The total magnetic flux in the center of the magnetic core will be the contribution of the magnetic flux flowing between the conductors, ϕ_k , and flowing in the conductors, ϕ_k .

$$\psi_k = \sum_{n=k}^N \phi_n + \sum_{n=k+1}^{N+1} \phi_n \quad (11)$$

$$\phi_k = \frac{\mu_0 L_c}{\alpha h} \cdot \frac{e^{\alpha e} + e^{-\alpha e} - 2}{e^{\alpha e} - e^{-\alpha e}} (J_k + J_{k+1}) \quad (12)$$

$$\phi_k = \frac{\mu_0 L_c}{\alpha h} \cdot e i_k J_k, \quad (13)$$

Where $e i_k$ is the insulation thickness.

In Eq. (9) we see that the voltage drop in a single layer is the contribution of the magnetic fluxes ϕ_k and ϕ_k , flowing in the other conductors and in the insulating surfaces, plus the contribution that becomes of its own current distribution, which is also the contribution of the skin effect.

The expressions in Eq. (7), (11), (12) and (13) show the linear dependence of the voltage difference created in a single layer, v_k , and the MMF, J_k , generated by each conductor portion.

Now we only have to find the solution of the system with the linear equations. The system has N voltages v_k linked to the $(N+1)$ MMF, J_k . By using the information about the winding connections and the winding electrical conditions, as voltage or current, we can find the solution associated to the linear system. There are necessary (N_w) conditions to describe the winding connections, such as voltage and current, where N_w is the number of windings. The indication of voltage and

current for a same winding can also contain a phase between them different from zero.

For turns that are series connected, the total current does not change from one layer to another. For winding where is necessary to include parallel connections, the voltage drop between $x=[0: e]$ is the same for all the layers.

The last condition is that the MMF=0 at the end of the layout arrangement is zero, $J_1=0$.

A matrix $(2N+1) \times (2N+1)$ is then obtained relating J_k and ψ_k , using Eq. (9) and Eq. (11).

$$\begin{bmatrix} \psi_1 \\ \psi_k \\ \dots \end{bmatrix} = \begin{bmatrix} \frac{\mu_o L_c}{\alpha h} \cdot e i_1 \dots \\ \dots \end{bmatrix} \begin{bmatrix} J_1 \\ J_k \\ \dots \end{bmatrix} \quad (14)$$

A routine program in Matlab consisting in to inform the characteristics of the layer conductors, the space between conductors, dimensions of the magnetic material, frequency, the arrangement of the layers and the electrical conditions was used to verify the power loss of the transformers proposed in section 3.

3 Design of the three-winding planar transformer

The application addressed in this section concerns the design of a three-winding planar transformer as an integrated part of a three-port converter that integrates the OBC and LDC converters, as it has been discussed in section 1.

In this context, the nominal condition of the transformer operation is the case when the electrical grid through the Power Factor Corrector (PFC) charges both batteries from Fig.1, V_{HV} and V_{LV} . V_{DC} provides 5 kW for V_{HV} and 2 kW for V_{LV} as can be seen in Tab. 1. The converter operation is not detailed in this text, but it is considered that the switching frequency is kept constant at 300 kHz for all the ports.

	N1	N2	N3
Vnom (V)	480	480	20
Pmax (kW)	7.0	5.0	2.0

Tab. 1: Design requirements for the application example. Switching frequency required at 300 kHz.

The E planar core shape is a plausible choice since it allows higher power density integration. In addition, at $F_{sw}=300$ kHz, the 3F36 and the 3F4 from Ferroxcube are the candidates pre-selected. The closest magnetic core is selected for a maximum current density of $J_i=20$ A/mm² and a window-filling factor of $F=0.25$. The total number of layers is determined with the information about voltage level, turns ratio and the maximum current density.

It is considered that the larger PCB has 105 μ m of copper thickness and a maximum of 8 layers. No more than 3 PCB's are used to reduce the complexity of the fabrication process.

As it discussed in section 1, the current balancing is improved by using interleaved configurations in a planar transformer. However, due to the minimum insulation thickness between low and high voltage sides, required in EV (European standard IEC60950), interleaved configurations associated with a large number of turns ratio can increase the PCB thickness or the number of PCB's, making it difficult the use of planar transformers.

Parameters	Values
Number of layers Port 1	8 (Series) – 3 turns by layer
Number of layers Port 2	8 (Series) – 3 turns by layer
Number of layers Port 3	Between 5 or 8 (Parallel)
Copper thickness	105 μ m (1,2 and 3)
Initial permeability core	2400
Mean turn length	112.4 mm
Core section	518 mm ²
Max. N° Layers by PCB	8
Max. PCB total thickness	3.2 mm

Tab. 2: Transformer parameters.

Three proposed arrangements are presented in Fig. 4. The three configurations have the same conductor layout for both high voltage sides. The maximum PCB thickness allowed in this configuration is 3.2 mm. The transformer parameters are presented in Tab. 2.

The design consists of 3 turns by layer connected in series for port 1 (P), then 8 layers also series connected to attend the turns ratio. For port 2 (S), the same configuration is used. The port 3 (T) can have between 5 or 8 layers connected in parallel to handle the high current.

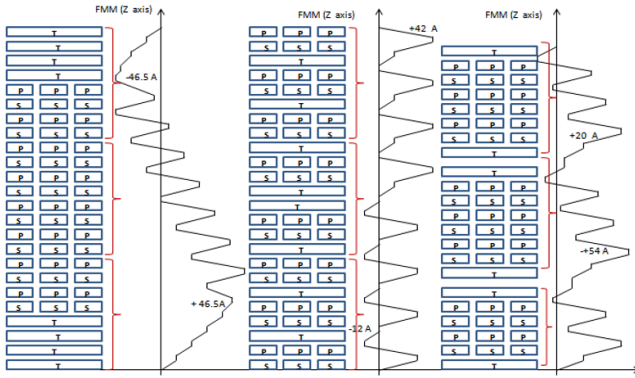


Fig. 4: Transformer arrangement propositions and magneto-motive force estimated with approximations for the current distribution and magnetic field. A) Non-interleaved. B) Full-interleaved. C) Partial-interleaved.

In the 1° configuration, the insulation thickness can be reduced because the low voltage side does not participate in the interleaving layout. This only occurs because the highest insulation thickness is always required between the layers P and T or between the layers S and T. Considering first, a linear behavior for the magnetic field inside the conductive layers and second, the current equally distributed between the parallel-connected layers, the magneto-motive force is drawn in Fig. 4.

We see in the first configuration, at the low voltage side, that the parallel layers seem to behave as a conductor block with a big surface area with a substantial influence of the skin effect. The unbalanced magnetic field profile generates a very poor current distribution on the parallel-connected layers.

In the 2° configuration, to improve the magnetic field distribution, the layers T are interleaved between P and the S type layers, but its fabrication process and mechanical assembling are more difficult. Due to the difficulty in the mechanical assembling, the 3° configuration is proposed, where the T layers are set to be the external layers of each PCB to make easier the assembling process. The last proposition is supposed to represent a midterm between the first and second arrangements considering efficiency and fabrication complexity.

The estimated MMF is illustrated on the left side of each configuration for the conditions in Tab. 1. Note that the first configuration seems to have the poorest performance and the last one seems to be acceptable.

3.1 Using the analytic method to estimate power losses

Now we are going to use the analytic procedure described in section 2.

Using the analytic approach, first we need to define the connection conditions. Only three conditions sets the three-winding transformer, voltage or currents, v_k or J_k and these conditions do not need to be defined for all the windings.

For example, we can define the voltage and current for the first winding and the current for the second winding only, as in Tab. 3.

The current in the first winding and in the secondary winding does not change from one layer to another since they are series-connected. However, for the third winding, the voltage is the one kept the same for all the layers. With that information, 25 equations relate the voltage and the MMF, Eq. (9) and Eq. (11).

Parameters	Values
V1	480 V / ref
I1	20 A / 0°
I2	14 A / 180°

Tab. 3: Winding electrical conditions.

The magnetic flux density and the current density of each conductor layer in the first arrangement of Fig. 4 is presented in Fig. 5.

The magnetic flux density contains real and imaginary parts and the x-axis represents the layer order, from the bottom to the top layer. Note that the current in the parallel layers T is concentrated in the internal layers.

Figures 6 and 7 present the results for the full-interleaved and partial-interleaved arrangements.

As we can see from the results, the partial-interleaved arrangement has higher power loss than the first configuration.

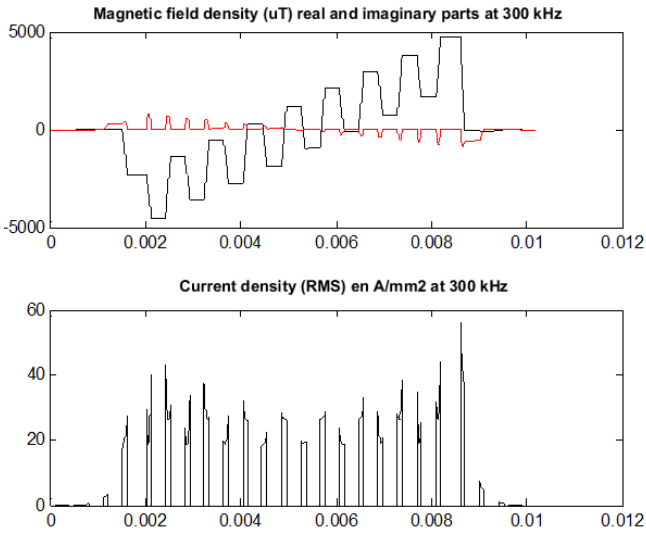


Fig. 5: Magnetic flux density and current density for the non-interleaved arrangement. Red: imaginary parts.

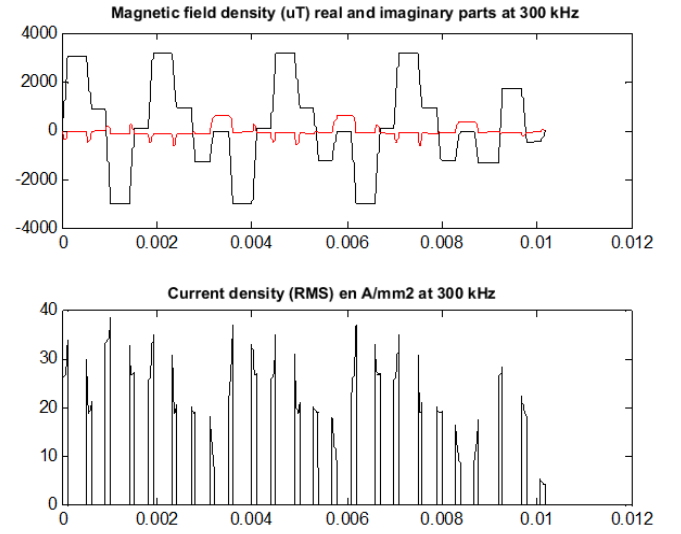


Fig. 7: Magnetic flux density and current density for the partial-interleaved arrangement.

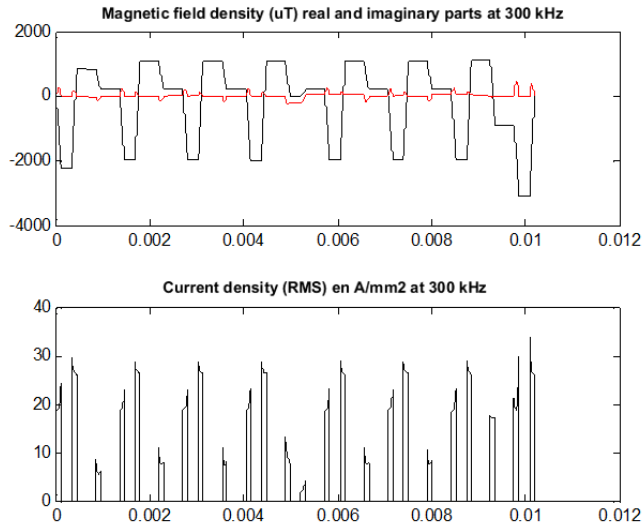


Fig. 6: Magnetic flux density and current density for the full-interleaved arrangement.

3.2 Finite-elements analysis simulations

In order to validate the results obtained, Finite Elements Analysis is now used to confirm the profile obtained with the analytic method.

Since in the analytic study, the magnetic field intensity is considered as being perpendicular to the central-leg of the magnetic core, a 2D-FEA analysis will be enough to verify the profile of the frequency dependent magnetic flux and current density inside the conductor layers. The software used is the FEMM.

The same layout configurations described before and the electrical conditions are indicated in the analysis. The distances between the conductor layers are configured and the frequency is set at 300 kHz.

Figure 8 shows the magnetic field intensity and current density for the non-interleaved arrangement. Figures 9 and 10 present the results for the full interleaved and the partial-interleaved arrangements, respectively. We can see that the magnetic field profile obtained with the simulation is very similar to the results obtained with the analytic approach.

In the three evaluated configurations, we see that the second arrangement has the best profile for current distribution and reduced power losses. The third configuration, however, has the poorest performance. When the method for power loss calculation considers a linear magnetic field variation inside the conductor layers and that the total current in the conductor is unaltered for parallel-connected layers, the third configuration seems to have an acceptable performance. Nevertheless, in reality, if we consider the frequency dependency for the magnetic lines inside the conductors and that the current is not homogeneous distributed between the parallel layers, the third configuration has higher power losses, as can be seen in Fig. 11.

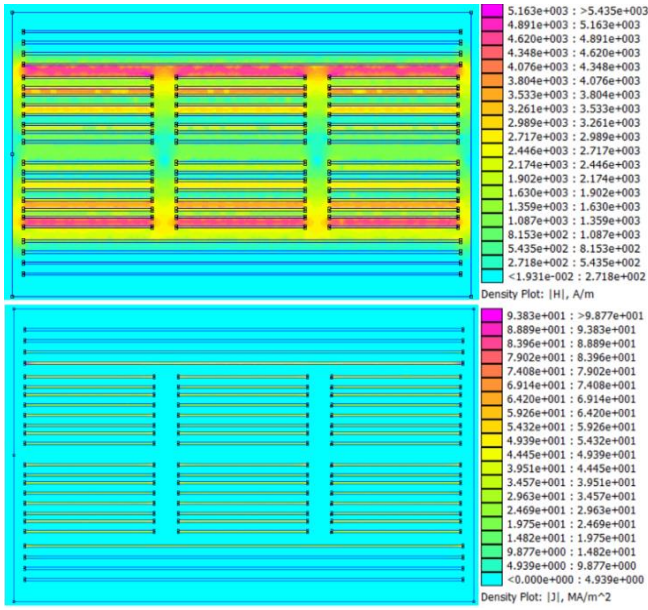


Fig. 8: Magnetic flux density and current density for the non-interleaved arrangement obtained with FEA.



Fig. 10: Magnetic flux density and current density for the partial interleaved arrangement obtained with FEA.

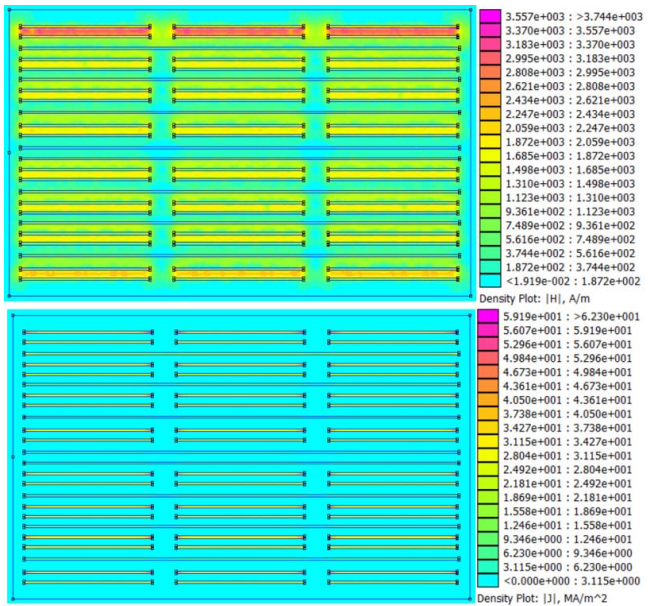


Fig. 9: Magnetic flux density and current density for the full-interleaved arrangement obtained with FEA.

These results are correctly estimated with the proposed modified calculation and by FEA.

The analytic method offers a simple, precise and fast solution to the problem and can be used in a pre-design phase for example to compare different arrangements as it has been made in this paper. The method also allows the determination of a physical equivalent model for a transformer with N windings. For example, as it was made in section 3, the application of some conditions in the windings allows determine an impedance matrix or the leakage flux between them by powering up each of the windings while keeping the others open.

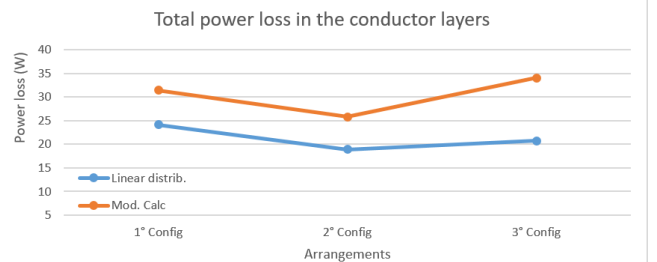


Fig. 11: Total power loss obtained with the analytic method and with the simplification of considering a linear distribution of the magnetic flux on the conductor layers.

4 Conclusion

In this paper, the design of a three-winding planar transformer for EV application was presented. The transformer is studied as an important part of the integrated system for OBC and LDC converters, responsible for the power transfer between high and low voltage batteries in the EV. Three different configurations were proposed to make easier its fabrication. Owing to the fact that is necessary parallel-connected layers to handle the high current on the low voltage side, a modified analytic method was studied to determine the current and the power loss distribution. The method allows to quickly highlight the drawbacks of the planar magnetic design. In the application discussed here, it helped in the selection of the configuration with the best performance. The results obtained with the analytic methodology and compared with results from FEA show that the second configuration presents a better power loss distribution for this specific application.

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