I-core PCB Planar Inductor Design for High Frequency and High Power Converters

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***Abstract*—The booming development of electric transportations has enlightened the need of high power converters with high power density at low cost. Conventional wire-wound inductors are limited by losses in high frequency, which explains the emergence of planar magnetics that have a better behavior in high frequency and are easier to manufacture. This paper presents the design of an I-core PCB inductor for high frequency and high power, with an inductance of at least 1 μH from 0.1 MHz to 2 MHz and total losses below 33 W. It is modeled and simulated through 3D FEA with a 50A sinusoidal current waveform, and a study on the influence of geometric parameters and frequency on inductance and losses is carried out.**

I. Introduction

The electrification of transportation requires high power converters with a high power density and a high efficiency, while keeping costs low. A common way to achieve these goals is to increase the switching frequency [1] and [2], along with the use of an appropriate switching technique. Critical soft-switching is a good case in point, since the operating frequency can be multiplied by 5 while the necessary inductance is divided by 20 [3].

Inductors usually fills most of the available volume in a converter and accounts for a significant part of the total losses [4]. PCB inductors offer an effective means to reduce the size of inductive components while remaining economical and easy to manufacture [4], [5]. Two types of designs are widely studied in literature: PCB integrated inductors [6], where the magnetic core is replaced by a magnetic sheet, within the PCB; E-core PCB inductors [7], where the windings are within the PCB but a magnetic core passes through it. While high-frequency is achieved for PCB integrated inductors, they are primarily used for low power applications [6], [8]; In comparison, E-core PCB inductors can achieve high power and high current ripple [7].

While these design have been widely studied in the literature, very little attention has been given to I-core PCB planar inductors, where the windings within the PCB are simply covered externally by planar magnetic cores. In addition to the advantages of planar inductors summarized in [9], this design offers even more flexibily for the shape of the windings and manufacturing simplicity since the core could be glued on the PCB.

This article focuses on I-core PCB planar inductor design for high power and high frequency applications. It especially aims at determining a configuration that achieves an inductance value higher than 1 μH for frequencies over 1 MHz with acceptable losses, current ripples as high as 50A for no DC current. It is fairly assumed that an inductor handling 50A of AC current can easily handle any combination of AC and DC current whose sum is lesser or equal to 50A. Several designs are simulated through 3D FEA using ANSYS Maxwell and analyzed. Three geometric parameters have been studied for their impact on inductance and core and copper losses, while the frequency is swept from 0 to 2 MHz by step of 100 kHz. This analysis allows to give a better insight into which geometric parameter change for which result.

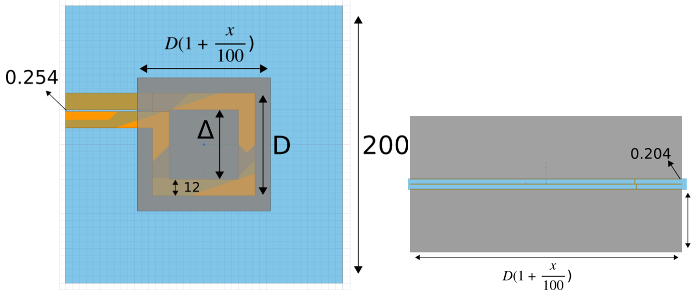
II. Design of the inductor

The I-core PCB inductor is made of windings within a PCB with one I-core on top and one on bottom of the board, centered on the windings. To handle the variations without changing its design, the FR4 epoxy PCB is a square of 200 mm side length. The design and variations are shown on figure 1. The traces have a rectangular shape with an outer side length D, a 12 mm width, and a thickness of 6 oz (0.204 mm).

The aspect ratio, defined as the ratio between the x axis length and y axis length, is equal to 1 for both the traces and the cores: they are squares. If there are several turns, they are equally spaced through the PCB and rectangular vias, of the same copper thickness as the traces, do the connections between layers. The narrow space between the two ends of a coil on a layer is 0.254 mm. The core has variable dimensions, from overlap of the windings (0%) to an increase of 100% of its side length, the aspect ratio remaining one. Its height, however, is constant to 17 mm.

In general, a magnetic core is limited either by saturation or its losses. In high frequency, core losses due to eddy currents prevail on saturation, which is assumed not to be a problem here. The predominant core materials are MnZn and NiZn ferrites because of their low electrical conductivity and saturation flux density, which make them very efficient in reducing eddy current losses [9]. The PC200 material by TDK has been selected for the cores, as an Mn-Zn ferrite for high frequency power supplies.

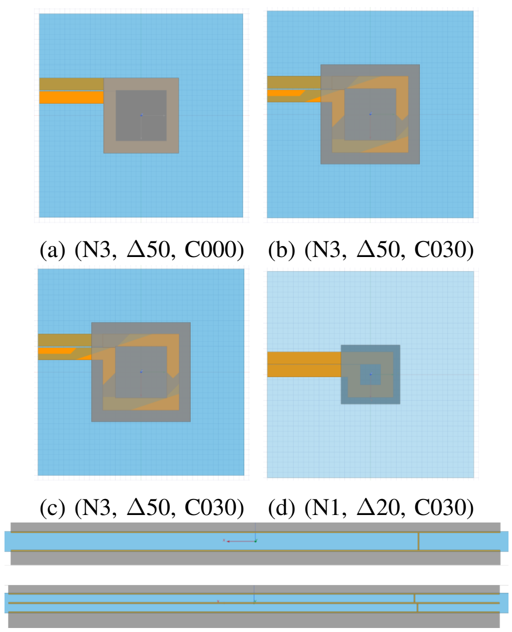
The aspect ratio being locked to 1 and the air gap being set by the thickness of the PCB, three independent geometric parameters have been varied to analyze their impact on the inductance and losses, in addition to the frequency: the inner



(a) (b)

Fig. 1: Design and dimensions (in mm) of the I-core PCB inductor.

air gap side length Δ; the core side length; the number of turns. A configuration (N, Δ, C) consists then of a number of turns, an inner air gap side length (which determines the side length of the traces) and a core side length. For example, a design with 2 turns, an inner length of 25 mm and a core length of 50% is noted: (N2, Δ25, C050). When a parameter varies, the configuration includes an 'X' instead of a number. For a number of turns that changes in the previous configuration, it comes: (NX, Δ25, C050). These variations can be seen on figure 2.



(a) (N3, Δ50, C000) (b) (N3, Δ50, C030)

(c) (N3, Δ50, C030) (d) (N1, Δ20, C030)

(e) N2 (top) and N3 (bottom)

Fig. 2: Variation of design parameters.

In this study, the number of turns is one, two or three; the inner air gap side length is modified from 5 mm to 50 mm, initially by step of 5 mm until 30 mm where the step becomes 10 mm; the core side length rises by 10% from overlap to 100%, double of its original size; the frequency is swept between 0 MHz and 2 MHz by step of 100 kHz. Since the geometric parameters are independent, their effect on the losses and inductance value are supposed independent to. In other words, the trends observed during the variation of one geometric parameter in a given configuration are assumed to be the same in other configurations. This assumption enables to study the effect of these parameters independently and to leverage these impacts to design the inductor. For instance, if one would like to increase the inductance and keep the core losses low, the number of turns could be augmented while the core size would be cut back.

A frequency of 0 MHz, while effectively simulating a DC excitation, can be considered as simulating low frequencies. When mentioning high frequencies, it is assumed any frequency greater than 0.1 MHz (100 kHz).

The inductor is subject to to a sinusoidal excitation of current, whose magnitude is 50A. The convergence criterion of the simulation is achieved when the energy error is less than 1%.

Three geometric parameters are varied in addition to the frequency to study the effects on inductance and losses, assuming the trends seen in one configuration are shared in all configurations (N, Δ, C). The inductance is studied first, to rule out the designs giving too low an inductance, i.e. under 1 μH, and the influence of the 4 mentioned parameters are analyzed. The losses are then discussed in order to choose a realizable design.

III. Effect of Design on Inductance

To design an I-core PCB inductor with at least 1 μH in high frequency, the inductance variation according to the frequency, the inner air gap side length, the number of turns and the core side length is studied in a particular configuration mentioned each time.

Figure 3 shows the inductance function of the frequency. It should be noted that passed a certain frequency, the inductance becomes steady. Indeed, taking the configuration (N1, Δ20, C030) as an arbitrary reference, the value from low frequencies to high frequencies ($\geq$ 100 kHz) presents a sharp drop of 27.8%, whereas a decrease of 3.5% only is observed for the inductance between 100 kHz and 2 MHz. Consequently, the inductance can be assumed independent from the frequency when it is greater than 100 kHz. Since the frequency is independent from the geometry, it can be assumed that the trends observed for one configuration would be the same for other configurations.

The inner air gap length is a powerful parameter regarding inductance, as proves figure 4, which shows variations for the configuration (N1, ΔX, C030). Varying the length from 5 to 50 mm, with a point every 5 mm until 30 mm (10 mm after it), the inductance is multiplied by approximately 7 at 0 MHz and 14 at 2 MHz. After 100 kHz, the inductance values overlap, which confirms once again the aforementioned indenpency of the frequency. Consequently, for this configuration, we can conclude that in high frequencies, ten times the inner gap length gives 14 times the initial value of inductance and 7 times in low frequencies. The same variation has been carried out on a similar configuration but with 2 turns instead of 1. 3

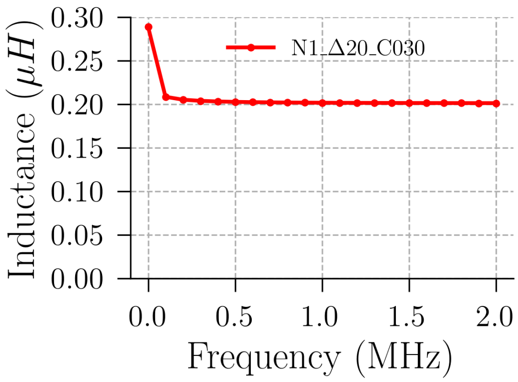


Fig. 3: Inductance value for a variation of frequency from 0 to 2 MHz, for a configuration (N1, Δ20, C030).

points have been simulated: 20 mm, 30 mm and 50 mm. The increase from 20 mm to 50 mm approaches a multiplication of 4.4 at 2 MHz, close to the one found in the one turn configuration, which is 3.8.

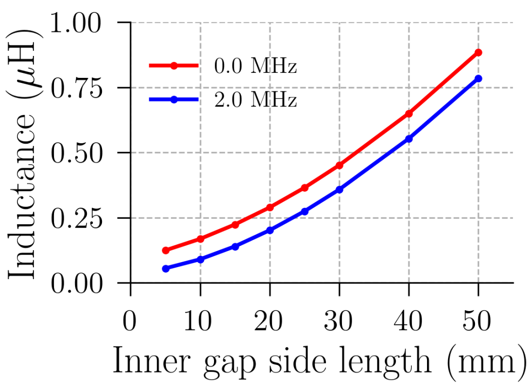


Fig. 4: Inductance depending on inner air gap length, for a configuration (N1, ΔX, C030).

The impact of the number of turns on the inductance is well known for air core inductors and conventional inductors. The equation, built on the assumption that the inductor has an infinite length compared to its diameter and that the flux density outside the windings is negligible, highlights a squared dependency with the number of turns:

$$L = \frac{\mu_0 N^2 A}{l} \tag{1}$$

As a result, doubling or tripling the number of turns implies multiplying the inductance by 4 or 9. The same behavior can be expected for this I-core PCB inductor design, without accounting for saturation of the magnetic core. In this light, reaching an inductance of 1 μH with a minimum volume requires an inner gap length of at least 20 mm, if the number of turns is to be increased. For a maximum inner length of 50 mm, the number of turns has been varied from 1 to 3 turns, with a core length of 30% (figure 5). The increase is close to what is said in the formula: a multiplication by 4 and 8.7 when switching from one to two and one to three turns at 2 MHz, enabling to reach a value of 2.94 μH and 6.50 μH respectively; and 4 and 8 at low frequency, for values of 3.32 μH and 7.41 μH.

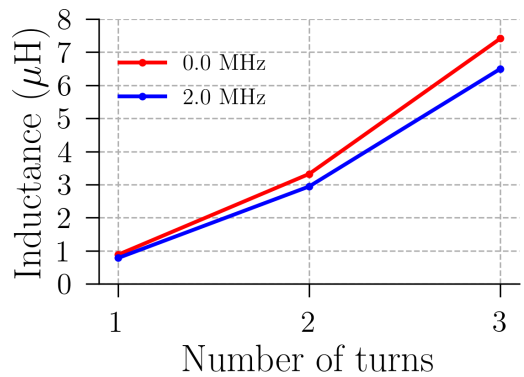


Fig. 5: Inductance relatively to the number of turns, for a configuration (NX, Δ50, C030).

A variation of the core dimensions has been carried out on a one turn inductor, with an inner length of 50 mm, since the inductance value is close to 1 μH. Compared to the inner air gap length, the effect of the core dimensions, on figure 6, are still noticeable but less striking. It seems that it reaches a saturation point where the improvement becomes less effective. The knee appears to be located around 30%. From overlap, in this configuration, an increase of 30% of the core side length leads to an increase of 66.5% and 88.0% in the inductance value for 0 and 2 MHz respectively. Whereas, the increase from 30% to 100% of core side length results in a 42.0% and 31.0% rise at 0 and 2 MHz. Nonetheless, an inductor in the following configuration achieves 1 μH in high frequencies: (N1, Δ50, C100).

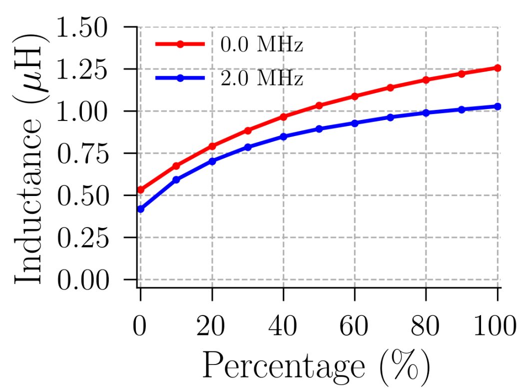


Fig. 6: Inductance according to core length percentage increase, for a configuration (N1, Δ50, CX).

The total length and surface of such an inductor is 148 mm and around 220 cm^2, which are large values, while the height is 37 mm. Moreover, losses have not been taken into account yet.

The inductance variation has been studied according to the frequency, the inner air gap length, the core side length and the number of turns. In high frequency, the value seems to be independent from the frequency. The inner length has a huged impact on the value of the inductance, which is almost proportional to the increase in length in the configuration studied. The number of turns too, as expected by the expression (1). Finally, the core variation effect is massive until around 30% increase, where its effect becomes less notable. To achieve an inductance of 1 μH, an inner air gap length bigger than 20 mm should be used.

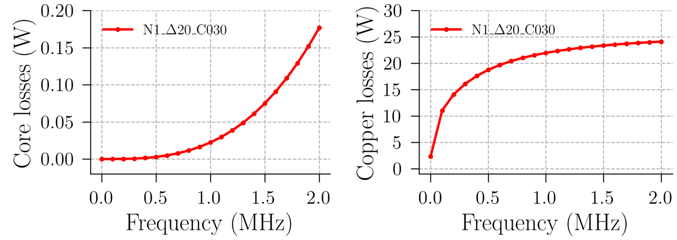
IV. Effect of design on losses

Using the same variation of parameters, i.e. the frequency, the inner length, the turn count and the core length, the effects on core and copper losses are derived.

Losses are the main problem encountered in high frequency inductors, and the limiting factor for conventional wire-wound inductors to reach high frequencies. They are the consequences of two factors: AC copper losses and core losses. The first cause of losses are separated into skin effect, where the magnetic field generated by a conductor will change the current distribution to reduce it to the surface of the conductor, and proximity effect, alike skin effect but created this time by the interference of magnetic fields generated by conductors close to one another. Core losses are issued from eddy currents flowing in the core, due to the variation of the magnetic field. High current ripples significantly augment these losses.

Figure 7 shows the core and copper losses in function of the frequency for the configuration (N1, Δ20, C030). The core losses grow faster and faster when the frequency goes over 1 MHz, whereas the copper losses begins to grow fast as the frequency rises, but seem to reach an asymptote after 0.5 MHz. Indeed, an increase of 69.9% from 0.1 MHz to 0.5 MHz is observed for copper losses, compared to a 9.7% rise from 1 to 2MHz. It is noteworthy that these latter are hundred times higher than core losses. The behavior of the core material was expected because it is especially designed for low losses up to 1 MHz. For this configuration, they are particularly negligible. But the multiplication factor of 7.9 from 1 MHz to 2 MHz denoting the trend may be a real problem for other configurations if losses are higher. Since losses depends on the frequency, several frequencies are plotted for each variation of geometrical parameters.

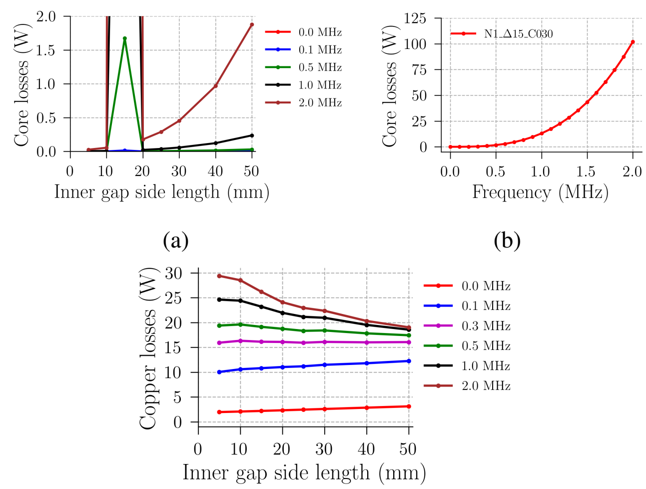
It can be noticed on figure 8 that core losses have a resonant inner air gap side length, at 15 mm. When compared to figure 7, the resonant losses figure 8 b) have the same trend but two orders of magnitude higher. For this configuration, the core losses at 2 MHz are 7.8 times those at 1 MHz. Moreover, 0.3 MHz turns out to be a frequency that demarcates two behaviors in copper losses. Before 0.3 MHz, these losses slightly rise with the inner air gap side length, but after, they decrease faster with the increase of the frequency. A 35.2% fall is noted



(a) (b)

Fig. 7: Core and copper losses according to the frequency, for a configuration (N1, Δ20, C030).

between 5 mm and 50 mm at 2 MHz in contrast to a 10.1% reduction at 1 MHz, for this configuration.



(a) (b)

(c)

Fig. 8: Core and copper losses relatively to the inner length, in (N1, ΔX, C030), together with core losses in function of the frequency for the resonant inner length (N1, Δ15, C030).

Figure 9 highlights that core losses and coppper losses skyrocket with the increasing number of turns in high frequency. Consequently, a maximum of 3 turns has been simulated. If the rise is still sharp in core losses, it remains manageable for a frequency of 1 MHz with a value of 12W, for this configuration. They become prohibitive from around 1.5 MHz and higher, reaching at least 40W. Similarly, copper losses increase with the number of turns, but the threshhold seen in figure 7 appears again here. At 2 MHz, the copper losses for 3 turns are 4 times those for 1 turn.

Core losses increase with the area of the core, with a linear trend for frequencies under 1 MHz, and an asymptote seems to be reached for higher frequencies, as shown in Figure 10. More points should be simulated to confirm the asymptote hypothesis. Meanwhile, an initial reduction (from 0 to 10% increase in core length) in copper losses becomes sharper as the frequency augments: 3.9W at 0.1 MHz and 21.5W at 2 MHz in this configuration, so a division by 1.3 and 2.2

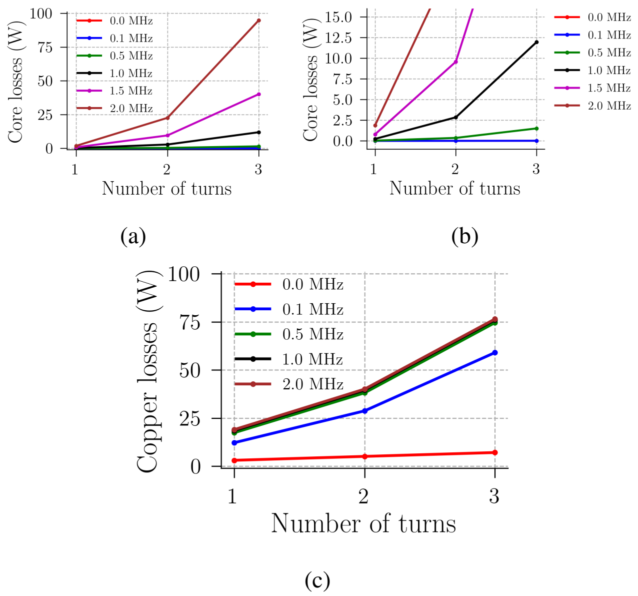


Fig. 9: Core and copper losses according to the number of turns in (NX, Δ50, C030).

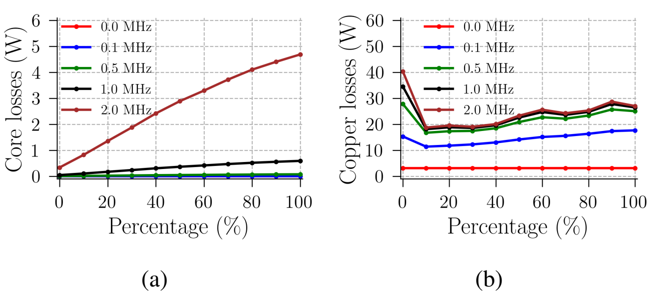


Fig. 10: Core and copper losses depending on core dimensions, for a configuration (N1, Δ50, CX).

respectively. However, they remain steady until 30% of core variation and go up again, slightly.

Consequently, copper losses are initially impacted by the frequency but seems to reach an asymptote after 1 MHz, whereas core losses keep increasing faster as the frequency goes up. For a given frequency, to decrease copper losses, the inner air gap length can be increased if the frequency is greater or equal to 0.3 MHz; the number of turns can be reduced, and the core side length should be between 10% and 40%. Considering the reduction of core losses, the inner air gap side length should be reduced and the resonant value of 15 mm, avoided. The number of turns should be lessened and the size of the core, reduced.

V. 1 μH INDUCTANCE DESIGN IN HIGH FREQUENCY

The results of section III and IV shed light on the effect of geometrical parameters on inductance and losses. It allows to thoughtfully decide which parameter to configure to increase the inductance and decrease the losses. Obviously, a tradeoff has to be made: an increase in the inductance leads to an increase in losses and volume. For a given value of inductance, an optimization could be done to reach the best compromise between losses and volume. This is beyond the scope of this article. Instead, an I-core PCB inductor design is proposed that achieves 1μH in high frequency, reasonable losses and size.

In section III, it has been shown that to reach an inductance value greater than 1 μH, the inner air gap length should be of at least 20 mm. A one-turn inductor would lead to a too large volume and a three-turn, too high losses in high frequency. As a result, a 2-turn inductor turns out to be a good tradeoff. A core length between 10% and 30% increase from overlap would keep core and copper losses low enough. The configuration (N2, Δ30, C010) matches our requirements.

In high frequency, the inductance reaches 1.05 μH. The core, copper and total losses depending on the frequency can be seen on figure 11. A peak of 32.3 W of losses needs to be dissipated at 2 MHz. At 1 MHz, the losses in the core are of 0.5 W while the copper losses are of 27.7 W. The total length of the inductor is 54 mm and its height, 37 mm. The total volume is hence: 107,892 mm^3 or 108 cm^3. The final design can be seen on figure 12.

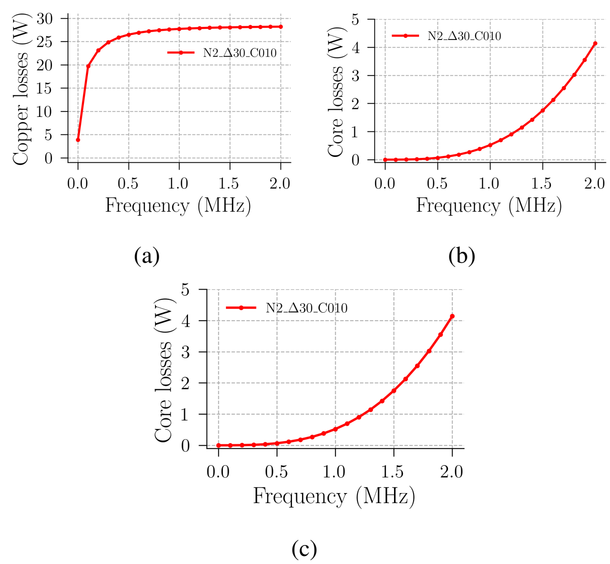


Fig. 11: Core, copper and total losses for the design (N2, Δ30, C010).

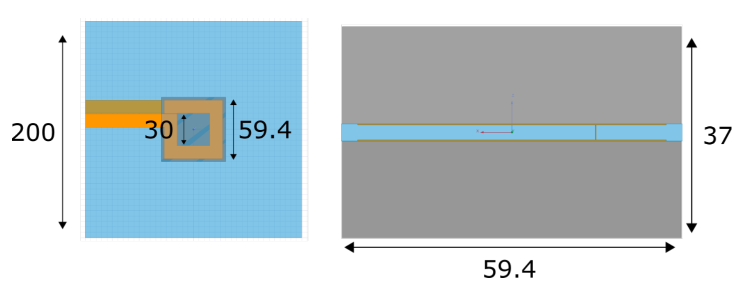


Fig. 12: Design and dimensions (in mm) of the inductor (N2, Δ30, C010).

VI. Conclusion

This paper introduces the design of an I-core PCB planar inductor for high frequency and high power. Through 3D FEA, using Ansys Maxwell, the effects of geometric parameters along with the frequency on the inductance and losses are analyzed to give an insight into which parameter to use to build such an inductor. The aspect ratio was locked to 1 and the inner air gap length, the number of turns and the core side length have been varied. Based on these results, an inductor has been designed achieving 1 μH over a range of 0.1 MHz to 2 MHz, with losses kept below 33 W at 2 MHz.

An optimization of volume and losses for a given inductance could be carried out in order to reach the best tradeoff possible for the design of the inductor. Further perspectives include building prototypes to validate the trends and values derived from the simulations, as well as determining the relative dependency of geometric parameters on inductance and losses.

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