PLANAR TRANSFORMER DESIGN FOR SWITCHING MODE POWER SUPPLIES

C. L. Ebert [1], W. P. Carpes Jr [2], J. C. S. Fagundes [3]

1. GEELN - CEFET/SC – ebert@cefetsc.edu.br
2. GRUCAD - CTC - UFSC – carpes@grucad.ufsc.br
3. INEP - CTC - UFSC – fagundes@inep.ufsc.br

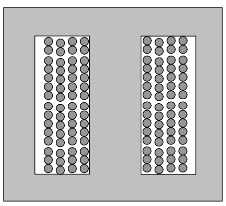
Abstract – **This paper has as main objective to present a design methodology for planar transformers used in switching mode power supplies. The magnetic elements design obtained from the presented methodology take advantages of this technology and from the appropriated technique of construction. The technology involving planar magnetic elements will be discussed in this paper, as well as the determination of the material and size of the magnetic core, types and shape of windings, core and winding losses, window utilization factor of the core, high frequency operation and integration of the magnetic element in the main plate of the converter.**

***Keywords* – planar transformer, high-frequency design, planar shape core, planar windings.**

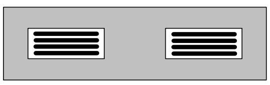
I. INTRODUCTION

The traditional techniques of transformers and other magnetic elements construction have been used for long time, but with some disadvantages compared to the techniques of planar elements [1]. The traditional construction approach requires a series of manual operations that makes high the cost design of these elements. Due to the particularities of each operator, small variations in the construction can cause great variations in the device characteristics.

The basic difference between a planar magnetic element and a conventional one is the orientation of the windings, allowing a core structure with a profile lower than the conventional (Fig. 1).



(a)



(b)

Fig. 1. (a) Conventional magnetic element and (b) planar magnetic element.

The planar magnetic elements are devices that recently are being used in energy static converters with great acceptance due to their better performance compared with a conventional magnetic element [2]. Fig. 2 shows the better efficiency of planar cores in terms of magnetic losses (Pm). The values presented in this figure were computed using Eq. 3 and Eq. 23.

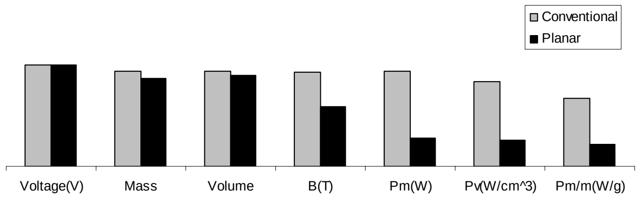


Fig. 2. Comparison between E-E conventional core and E-E planar core.

Some advantages in using planar elements are presented below:

- Structures with low profile present, for certain applications, better volumetric efficiency and high power density;
- Increasing the area of the core central leg makes possible a reduction in the number of turns. [3]
- Possibility of obtaining low leakage inductance due to simplicity in the interleaving of layers and reduced amount of turns, thus reducing the peaks of voltage and oscillations that can destroy some semiconductors devices;
- Decrease in the high-frequency winding losses, since the skin effect is minimized due to conductor thickness reduction;
- Higher conversion efficiency due to the better magnetic coupling;
- Better thermal control, because planar elements present a larger relation area/volume, increasing the available surface for the heat-transfer mechanism, thus reducing the thermal resistance;
- Easy construction with the use of Printed Circuit Board (PCB);
- Good repeatability, which is a very important characteristic in resonant topologies.

As disadvantages of the planar elements, we can quote:

- Increase of the PCB area, if the planar element is incorporated to the main plate of the converter;
- Increase of the stray capacitances;
- Low core window utilization factor, roughly 0.25 to 0.30, depending on the type of planar winding.

In addition to the advantages cited above, another factor that indicates planar magnetic elements to have a promising market is their geometric characteristic: low profile is ideal for applications where minimum height is desired, e. g., power supplies for portable computers and telecommunication system.

II. PLANAR TRANSFORMER DESIGN

A. Technique of construction

A planar magnetic element can be constructed either as an isolated piece (autonomous component) or integrated to the main PCB of the power supply (or another static converter).

Fig. 3 shows some ways to place a planar magnetic element in the main plate of the system.

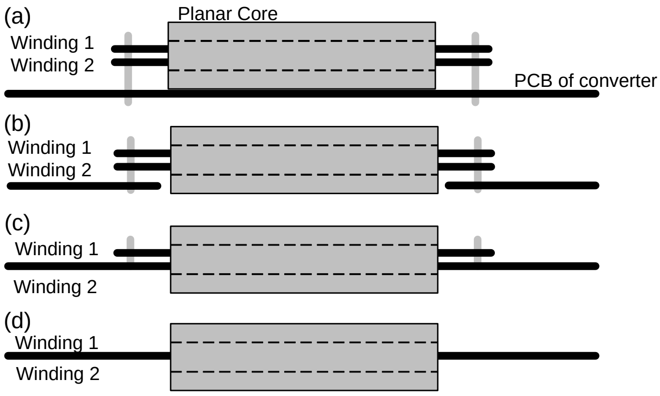


Fig. 3. Techniques of assembling the magnetic planar element in the main PCB.

B. Planar Cores

1) Geometric shape

The more studied and used planar cores are constructed by combining E-E or E-I cores, but other core shapes can be also found, as it is shown in Fig. 4.

(a) (b) (c) (d)

Fig. 4. E-E and E-I core (a), RM core (b), ER core (c) and PQ core (d).

Core shapes as the RM, ER and PQ present a larger width for the winding (bw), which permits to increase the track widths (see Fig. 5). In addition, they allow shorter turns, which reduces the copper losses and the leakage inductance. In these core shapes the windings are confined, thus reducing EMI. Because the construction form, these core have a greater cost.

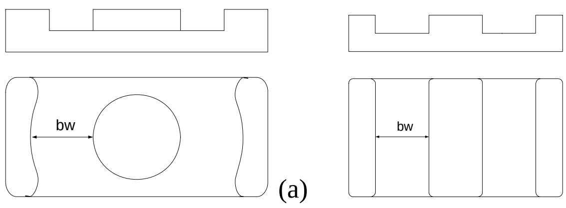


Fig. 5. Core with (a) circular central leg, and (b) core with rectangular central leg.

2) Core Material

The type of magnetic material used in fabricating planar cores is very important since this influences directly its operation and cost of production.

Some characteristics of the material must be considered before starting the project of the magnetic element. For example: electric resistance (must be high to prevent eddy currents); range of frequencies where the material can operate; the maximum allowable flux density; the magnetic permeability (must be high to facilitate flux circulation).

All these characteristics contribute to a fundamental quantity in the project of magnetic elements, namely the magnetic losses, corresponding to the power dissipated in the magnetic core. In planar magnetic elements used in high frequency ferrimagnetics materials (ferrites) are generally employed. These are ceramic magnetic materials of high resistivity, made with iron oxides (e.g., $MnFe_2O_4$ and $NiFe_2O_4$).

C. Planar Windings

Several technologies can be used for implementing a planar winding. The most used in industry (and most discussed in literature) are made with rigid PCB, flexible PCB or printed-copper foils. The use of PCB facilitates the repeatability of the windings, implying in a production profit. One of the main advantages of using PCB is that the winding of the magnetic element can be incorporated to the plate of the whole circuit, eliminating connections.

The flexible circuits, which are made of a polymer substratum covered by a thin copper layer, are better in terms of window use. This technology is more appropriate for wider copper tracks above 210µm. Windings made with printed-copper foil have lower cost when they are single turn and intended to work at higher current levels. Their disadvantage is the need for isolation between layers.

In addition to the techniques previously mentioned, hybrid windings can also be used to better exploit the geometry of the component [5]. Fig. 6 shows an example of windings made with PCB, where one of the plate sides accomodate the secondary winding and on the other side contain the primary winding.

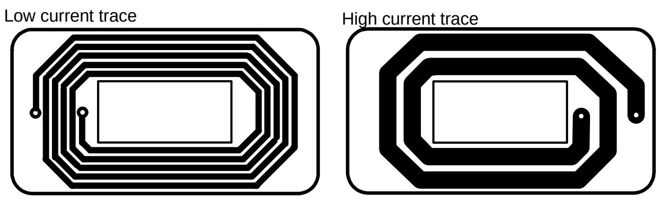


Fig. 6. Windings for E core.

D. Losses in the planar element

For the magnetic material two temperatures must be observed: the temperature Curie and the temperature for minimum losses. Besides these two temperatures, in a complete project we must take also into account the maximum allowed temperature related to the destruction of both the conductors isolation and the isolation between windings. The rise in the operation temperature is caused by the core losses as well as by the winding losses of the transformer. The core losses can be divided in hysteresis losses, and losses due to eddy current, which are significant only in high-frequency applications (range of MHz). The winding losses depend on both the DC and AC resistances.

In order to avoid an excessive temperature rise, that corresponds to operate above minimum losses temperature, the maximum permissible total loss must follow the given relation [6]:

$$P_{tot} = \frac{\Delta T}{R_{th}} = \frac{\Delta T\, V_e^{0.54}}{C_{th}} \qquad (1)$$

Where:

P_{tot} - total losses in the magnetic element = winding losses plus core losses;

ΔT - Temperature rise, i.e., the difference between the temperature at the "hot point" of magnetic element (T_p) and the ambient temperature (T_a) $\Rightarrow \Delta T = T_p - T_a$;

R_{th} - represents the thermal resistance of the transformer;

V_e - effective volume of the core (cm^3);

C_{th} - constant that depends on both the core shape and the arrangement of the winding. For a planar core and windings made in PCB (epoxy), C_{th}= 24 [7].

1) Core Losses

The volumetric magnetic losses of the material (in mW/cm3), for sinusoidal excitation, are given by [8]:

$$P_v = Cm\, f^x\, B^y (ct - ct_1\, T + ct_2\, T^2) \qquad (2)$$

For a temperature T=100^{o}C, the term in parentheses is equal to 1. Then

$$P_{mag} = V_e\; Cm\;\; f^x\, B^y\,, \qquad (3)$$

where:

B - peak of magnetic induction, given in Tesla. Parameters *Cm, x, y, ct,* ct_1, and ct_2 are specific for each type of ferrite and depend on frequency, waveform of the applied signal and the operation temperature.

2) Winding Losses

The minimization of the winding losses requires a good knowledge of skin effect and proximity effect phenomena.

The skin effect appears due to the presence of alternating current that generates an alternating circular magnetic field in the conductor. This field makes the current to flow predominantly in the conductor periphery. Consequently, the cross sectional area to current circulation is reduced, provoking an increase in the AC resistance of the conductors.

The skin effect can be reduced in the case of planar windings, when one have $h_w/\delta \le 2$, where h_w is the height of the conducting track and δ is the skin depth, given in (4).

The skin depth for a given frequency is:

$$\delta = \sqrt{\frac{\rho_c}{\pi\,\mu\, f}} \quad (m), \qquad (4)$$

where:

ρ_c - resistivity of the copper, which is temperature dependent;

μ - copper permeability, $\mu \approx \mu_o = 4\pi \times 10^{-7}$ H/m.

Thus, the thickness of the plate track for a small skin effect can be calculated by:

$$h_w \le 2\sqrt{\frac{\rho_c}{\pi\,\mu_o\,\mu_r\; f}} \quad (m) \qquad (6)$$

Where:

μ_r = 1 (non-magnetic material).

Similarly to the skin effect, the proximity effect also alters the current distribution in a conductor. But, in the case of proximity effect, the current in neighboring conductors produces the magnetic field that causes this change in current distribution. In the skin effect, on the other hand, the current in the conductor itself produces the magnetic field that alters the current distribution in it.

To analyze the proximity effect, it is necessary to verify the behavior of the magnetic field in the winding structure [9].

Fig. 7 shows an isolated conductor foil.

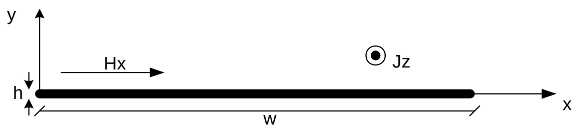


Fig. 7. Conductor foil.

Considering that a current flows through the conductor, and that its thickness h_w is very smaller than the width w, the border effects can be neglected. It is also assumed that the magnetic field is perpendicular to the current (tangent to the conductor). The losses in conduction per unit length of the conductor are given by [9]:

$$P = (\Lambda_F + 1)\, I_{rms}{}^2\, R_{dc} + \Lambda_G\, w^2\, Hx^2\, R_{dc} \qquad (7)$$

The factors Λ_F and Λ_G are related to the skin and proximity losses, respectively, and are given by:

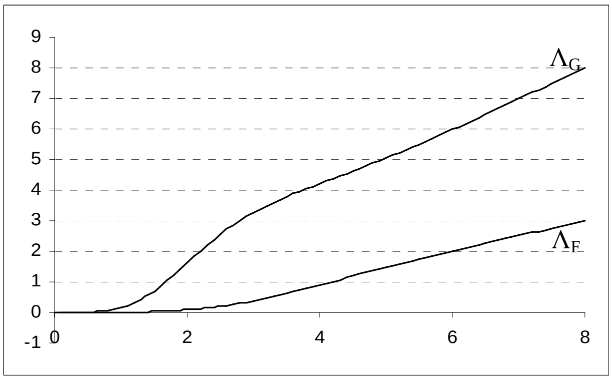
$$\Lambda_F = \frac{\alpha}{2}\frac{\sinh\alpha + \sin\alpha}{\cosh\alpha - \cos\alpha} - 1 \qquad (8)$$

$$\Lambda_G = \alpha\,\frac{\sinh\alpha - \sin\alpha}{\cosh\alpha + \cos\alpha} \qquad (9)$$

R_{dc} is the DC resistance per unit length, $R_{dc} = 1/(\sigma\, h\, w)$, where σ is the conductivity and $\alpha = h/\delta$.

In low frequencies, where δ is large, or in foils with small thickness h_w, the skin effect can be neglected, because $\alpha \cong 0$ ($h << \delta$).

The factor Λ_G is greater than the factor Λ_F for the same values of α, and both increase with increasing frequency. This can be seen in Fig. 8.



α

Fig. 8. Factors Λ_F and Λ_G as a function of α.

This also implies that in high frequencies the losses caused by the proximity effect will be much larger than that caused by the skin effect.

Equation 7 shows that the proximity effect losses are proportional to the square of the magnetic field. This means that at the points where the field is more intense the losses will be larger, and that is dominant in multilayer windings.

E. Parasitic effects in planar transformers

1) Leakage inductance

The analysis of the magnetic field properties shows that this field is related to stored energy. In fact, the spatial distribution of the magnetic field represents the distribution of this energy. Hence, knowing the magnetic field properties shows not only where this energy is stored but also as this energy is coupled with to other circuit elements.

By applying the physical principles of magnetic fields in planar magnetic structures it is possible to optimize the project and to calculate the magnitude of the parasitic elements, such as leakage inductances. A high stray flux inside the window of the core cause excessive losses. Furthermore, if this stray flux is external to the window, it will cause problems of noise in external components. Figure 9 shows the flux lines in a planar transformer. It can be observed that there are some lines that make the magnetic coupling of the windings (magnetizing flux), but there are also other flux lines that are between the windings, which are responsible for the leakage inductance.

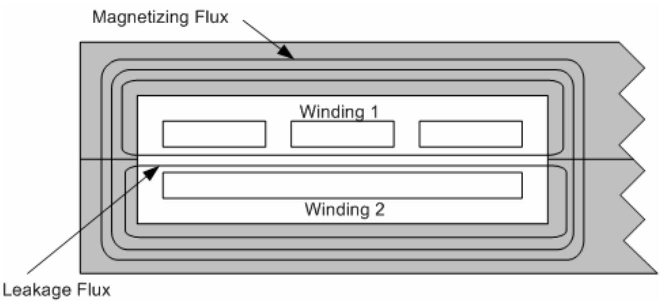


Fig. 9. Flux lines in a planar transformer.

Generally, the leakage inductance is small compared with the magnetizing inductance, and it can be calculated from the windings dimensions.

According to the Ampere's Law, the total magnetomotive force in the region between the windings (leakage inductance) is equal to the load-related Ampère-turns flowing in the windings. If the distance between the windings is uniform along all the windings width, the field intensity H is constant in this entire region and it is given by:

$$H = \frac{F}{b_w} = \frac{NI}{b_w} \text{ (A/m)}, \tag{10}$$

where:

b_w: available width for the winding. Considering that practically all this width is occupied by the tracks, $b_w \cong w_t.N$, where w_t is the width of tracks.

The permeability in the region between the windings, which is a non magnetic material, is equal to the permeability of air, $\mu_0 = 4\pi \times 10^{-7}$ H/m. As H is constant, the flux density (B) will be also constant, since B = μH.

As the energy density per unit of volume is equal to $\frac{1}{2}B.H$, the energy density between the windings can be considered constant:

$$\frac{E}{v} = \frac{1}{2}BH = \frac{1}{2}\mu_o H^2 = \frac{1}{2}\mu_o\left(\frac{NI}{b_w}\right)^2 \text{ (Joules/m}^3\text{)} \tag{11}$$

As a consequence, the total energy can be calculated by multiplying the energy density by the total volume of the region between the windings. This energy is related to the leakage inductance, according to:

$$E = \frac{1}{2}LI^2 \tag{12}$$

Then, the leakage inductance for each winding can be calculated from:

$$L_L = \frac{2E}{I^2}, \tag{13}$$

where I is the current in each winding. After some algebraic manipulations, we obtain:

$$L_L = \frac{\mu_o N^2 h_L MLT}{b_w}, \tag{14}$$

where:

h_L: distance between windings;
MLT: mean length of turns.

Equation (14) can be rewritten as:

$$L_L = \frac{\mu_o N^2 A_L}{b_w}, \tag{15}$$

where:

A_L: perpendicular area to the lines of the leakage flux.

Therefore, it can be seen that the leakage inductance does not depend on the load current, but only on the winding geometry.

2) Capacitances between windings

The interleaving of windings presents a significant negative effect: the growth of the parasite capacitances due to the increase in the surface of the conducting area. These capacitances can cause undesirable consequences in applications with electrically isolated structures, since AC coupling is always present, increasing the possibility of conduction noise. The planar transformers are reputed to have great capacitances between windings, compared to conventional transformers. This comes from the fact that the interleaving is much more used in planar transformer (due to easy implementation) than in conventional transformers. If in both types of transformers were used the same core and the same relation between width of winding and intercalation (which would gives the same AC resistance and leakage inductance), the capacitance between windings would be similar in both structures [3].

Equation (16) allows the calculation of the capacitance between windings and, in this way, to verify the factors influencing this capacitance [10]:

$$C_w = \frac{0.0085\,\varepsilon_r\, MLT\, b_w\, N}{h_L} \quad \text{(pF)}, \tag{16}$$

where:

- Relative dielectric constant (ε): the use of isolating materials with low dielectric constant (e.g., Mylar or Teflon) reduces the capacitance between windings. The dielectric constant does not have effect on both the leakage inductance and the AC resistance;
- Thickness of the isolation (h_L): a denser insulation reduces the capacitance, but increases proportionally the leakage inductance. The thickness of the insulation does not affect AC resistance;
- Number of turns: using a single-layer winding and minimizing the number of turns (the core operating with maximum flux density) reduces the area of the winding, which reduces the capacitances between windings. Reduction of turn number also minimizes the leakage inductance and the AC resistance;
- Width of the winding: increasing the width of the winding, directly or for interleaving, also increases the capacitance, but reduces the leakage inductance and the AC resistance;
- Length of turns: decreasing the length of turns reduces their area and, as a consequence, also reduces their capacitance;
- Faraday Protectors: when correctly placed, they help to reduce the capacitive effects, but they make the structure of the interleaving most complex. In addition, they enlarge the volume between windings, causing a proportional increase of the leakage inductance.

A way to reduce the capacitance between winding consists in place the tracks of a winding between the tracks of the adjacent winding, as Fig. 10 shows.

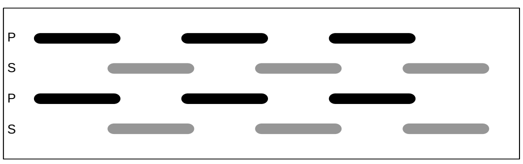


Fig. 10. Winding design to reduce the capacitances.

F. Design methodology of planar transformers

When initiating the project of a planar magnetic element, we must clearly know the objective to be reached in constructing this device. The present work has as main objective to present a design methodology that contemplates the minimum losses requirements. As the total losses in a magnetic element are composed by the magnetic losses (core losses) and by the joule losses (winding losses), we must determine each contribution so that the total losses are minimized [11].

The design where the objective is minimal losses is different to design where maximum power capacity is desired.

The optimum value of B corresponding to the minimum losses (B_{opt}) can be obtained from:

$$\left(\frac{dP_{tot}}{dB}\right)_{B=B_{opt}} = 0, \tag{17}$$

which results in:

$$P_{mag} = \frac{2}{y+2} P_{tot} = \frac{2}{y+2}\frac{\Delta T}{R_{th}} \tag{18}$$

$$P_j = \frac{y}{y+2} P_{tot} = \frac{y}{y+2}\frac{\Delta T}{R_{th}} \tag{19}$$

From the above equations, we can determine the optimum values of B and J for each material and shape core.

$$B_{opt} = \left[\frac{2}{y+2}\left(\frac{\Delta T}{R_{th}}\right)\frac{1}{Cm\ f^{x}\ V_e}\right]^{1/y} \tag{20}$$

$$J_{opt} = \left[\frac{y}{y+2}\left(\frac{\Delta T}{R_{th}}\right)\frac{1}{\rho\, MLT\, k_u\, W_a}\right]^{1/2} \tag{21}$$

where:

k_u: utilization factor of windows;
W_a: cross-sectional area of winding window.

It is also possible to express the relation between Pmag and Pj by:

$$\frac{Pmag}{Pj} = \frac{2}{y} \tag{22}$$

The choice of the core must be made through the minimization of losses, for a given frequency and a given flux density necessary to obtain the desired voltage across the transformer windings. For this choice, we can use the graphs presented in figure 11, obtained from Eq. 23, which shows the voltage across the transformer windings as a function of the magnetic induction B for a given frequency and for several FERROXCUBE cores made with material 3F3.

$$V_{rms} = B_{\max}\, f\, AeKvN \tag{23}$$

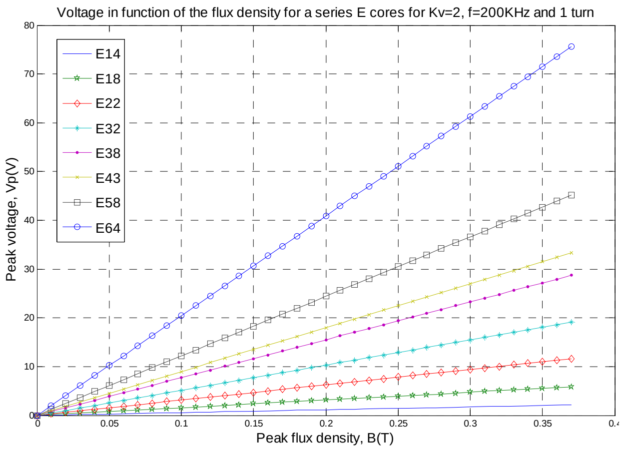


Fig. 11. Voltage for several E cores

The procedure starts by considering only a single turn in the winding with smaller voltage, which gives a smaller electric resistance and a smaller leakage inductance. From the graphs presented in Fig. 11 one can determine the smaller core and the flux density corresponding to a given operation voltage. The core chosen, the optimum flux density, B_{opt}, can be determined.

If the value of B_{opt} is inferior to the required B for the core and the desired voltage, three options appear:

- increase the number of turns, which increases the electric resistance and the leakage inductance;
- select a larger core, but with a corresponding increase in volume, weight and cost; or
- operate with a flux density higher than the optimum, which however increases the magnetic losses.

In order to solve this problem, we can use an algorithm that compares the three options and verify which presents the smaller losses.

After that, the core will be already chosen as well as the flux density, the current density and the number of turns in the smaller-voltage winding can be determined.

The number of turns in the other winding must respect the transformation ratio, Np = Ns.Vp/Vs.

To finish the design, it only remains to calculate the conductor cross section area for both windings. As the height of the PCB conducting track is known, the track width is obtained from the calculated area. Care must be taken in verifying if the conducting track height is greater than, at least, two skin depths, otherwise there will be an increase in joule losses.

With all the parameters defined, at the end of the project the magnetic and joule losses must be computed, taking in account both the skin and proximity effects, in order to verify the increase in temperature.

The leakage inductance, as well as the capacitance between the windings, can also be calculated and verified to be in accordance with the static converter design.

III. CONCLUSION

When a core is underutilized, the designed element becomes more expensive and bigger than it should be. On the other hand, if it is used beyond the nominal ratings, there is the risk of malfunctioning or temperature rise above the tolerated limits. A core is well used when it operates with the maximum allowed flux density and the winding works with the maximum allowed current density, limited by magnetic losses and joule losses, respectively.

It is important to stress that, in order to take advantage of the planar elements, their design must be carried out differently from that of conventional cores. Moreover, the assembly of the planar transformer presents particular characteristics at the final stage of its construction, in which the designer has control over the transformer parameters.

This work presented a construction technology of magnetic elements still little used in industry. Constructive aspects, calculation of parameters and a design methodology had been discussed, showing some advantages of this type of element.

Currently, we are working on new parameters and correction factors in order to predicted core and winding losses for non-sinusoidal waveform. Methods of optimization using multiobjective genetic algorithms, study of the design sensitivity on the involved elements, as well as experimental work for adjustments in the design methodology, are also in development.

REFERENCES

[1] R. L. Nagel, M. E. Partridge, “Planar Magnetic Component Development”, *Sandia National Laboratories*, Albuquerque, New Mexico, 1994.

[2] S. Wang, “Modeling and Design of Planar Integrated Magnetic Component”, *Thesis (Master of Science*), Virginia Polytechnic Institute at State University, Virginia, 2003.

[3] L. Dixon, “Designing Planar Magnetics”, *Power Supply Design Seminar*, Texas Instruments. 2006-2007.

[4] Ferroxcube Components Inc. “Design of planar power transformers”, *Application Note*, 2001.

[5] T. R. Gunewardena, “Manufacturing Design Considerations for Planar Magnetics”, Electrical Insulation Conference, 1997

[6] S. A. Mulder, “Application note on the design of low profile high frequency transformers”, *Philips Components*, May 1990.

[7] Ferroxcube, “Soft ferrite design tools”, *Help*. 2002.

[8] S. A. Mulder, “Loss formulas for power ferrites and their use in transformer design”, *Philips Components*, February 1994.

[9] S. Wang, M.A. de Rooij, W. G. Odendaal, J. D. Van Wyk, D. Boroyevich, “Reduction of High-Frequency Conduction Losses Using a Planar Litz Structure”, IEEE Transactions on Power Electronics, vol. 20, nº. 2, March 2005.

[10] C. W. T. McLyman, “Transformer and Inductor Design Handbook”. Third Edition.California, USA, 2004.

[11] P. Viarouge, “Eléments de dimensionement des composants magnétiques en électronique de puissance”, *Raport RHAE/CNPq LAMEP*, Florianópolis/SC - Brasil, 1994.