COMPUTER-AIDED DESIGN OPTIMIZATION OF HIGH FREQUENCY POWER TRANSFORMERS

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Abstract - **A computer-aided tool for design optimization of multiwinding high frequency power transformers has been developed. The software uses a new optimization algorithm to design the transformer with minimum overall power loss and/or volume and to determine optimal data concerning the core and windings. This algorithm takes full account of the winding current waveforms and skin and proximity effects in the process of design optimization of the windings. The developed tool can be applied to any circuit topology provided the excitation waveforms of the component are periodic. An application example is given for a transformer of a 550 W, 40 kHz active clamping forward converter.**

I. INTRODUCTION

Some procedures for designing high frequency power transformers have been presented in the literature. These procedures present, among others, the following drawbacks: do not optimize the windings [1,2,3,4]; are either described for or applicable to only two winding transformers [3,4].

The purpose of this paper is to describe in some detail the methodology that has been developed to optimize the design of multiwinding high frequency power transformers and the respective software implemented to do it. This methodology can be applied to any circuit topology provided the excitation waveforms of the component are periodic. By using the referred software two distinct design situations can occur and are identified as either a design limited by core loss or a design limited by core saturation. In the former, a transformer with minimum overall power loss and volume may be designed. In the latter, a transformer with minimum volume may be designed, but the overall power loss will not be minimum.

The significance of the paper is that the design procedure includes the optimization of the windings, while taking into account the winding current waveforms and skin and proximity effects when calculating optimal data related to number of layers, conductor data, effective resistance, current density, and loss in the respective winding. All parameters and variables in this paper are in the SI units.

II. LOSSES AND THERMAL MODELING

The design of compact and efficient magnetic components for high frequency power conversion applications, such as transformers and inductors for switched mode power supplies, requires accurate models to predict winding and core losses and the temperature of the hot spot in the component.

A. Winding Losses Modeling

Considerable mathematical modeling efforts carried out over the past few years have resulted in one- and two-dimensional models [5,6] to predict winding losses in such components. These models are referred to as thin layer model [5] and orthogonality model [6], respectively. A unified sense to these models appears in [7], where generalized expressions are derived for the effective resistance of multilayer windings in multiwinding magnetic components under arbitrary periodic current excitation. The thin layer model, in spite of its simplifying assumptions, can be used as an accurate tool for design optimization purposes [7] and, for this reason, is used in the optimization methodology described here.

Fig. 1(a) shows a cross-section of a multiwinding magnetic component where the structure of a particular multilayer winding is detailed. According to derivations in [7], based on the thin layer model, the expression for calculating the ratio, F_r, between the effective resistance, R_e, and the dc resistance, R_{dc}, of multilayer windings of foil conductors, under non-sinusoidal periodic current excitation, is given by (1). In the following equations, I_{dc} is the dc component of the current waveform, I_{ef} is the rms value of this waveform, I_{efj} is the rms value of its j'th harmonic component, *f* is the fundamental frequency, $_c$ is the conductor conductivity and $_o$ is the free space permeability. To permit calculating the winding boundary condition ratio, defined by (5), the frequency-independent magnetic field phasors between windings, caused by each harmonic component of their current waveforms, must be determined.

$$F_r = [\, I_{dc} / I_{ef} \,]^2 + \frac{1}{3} \sum_{j=1} [\, I_{efj} / I_{ef} \,]^2 [\, \underline{h}_j / |1 \quad \bar{\ }_j|^2 \,]\, x$$

$$x\{ F_1(\underline{h}_j)[(2M^2+1)(1+|\bar{\ }_j|^2)+2(M^2 \quad 1) \;(\bar{\ }_j)] +$$

$$4F_2(\underline{h}_j)[(M^2 \quad 1)(1+|\bar{\ }_j|^2)+(M^2+2) \;(\bar{\ }_j)]\} \tag{1}$$

Where:

$$F_r = R_e / R_{dc} \tag{2}$$

$$\underline{h}_j = h / \quad _j \tag{3}$$

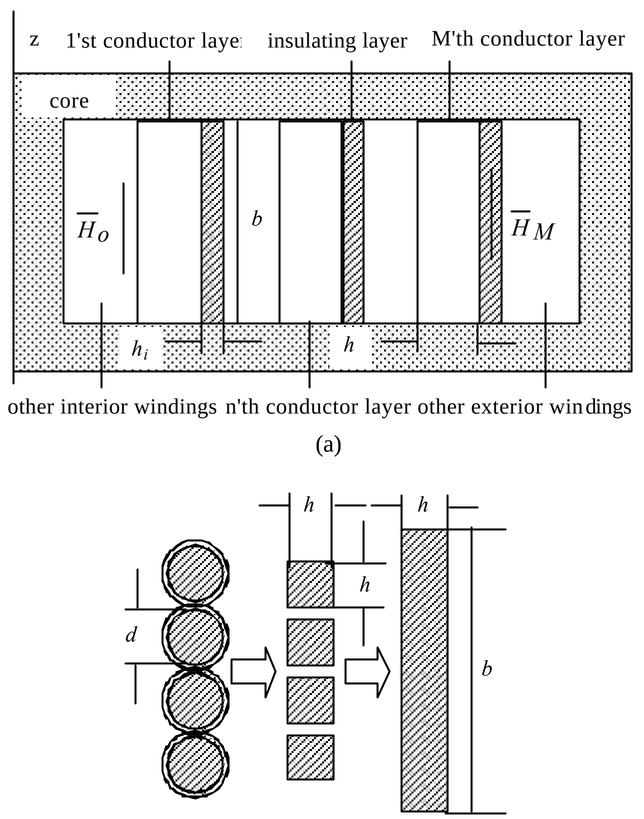
$$_j = \sqrt{2 / j2 \; f \; _o \; _c} \tag{4}$$

$$\bar{\ }_j = \overline{H}_{oj} / \overline{H}_{Mj} \tag{5}$$

$$F_1(\underline{h}_j) = \frac{senh(2\underline{h}_j) + sen(2\underline{h}_j)}{cosh(2\underline{h}_j) \quad cos(2\underline{h}_j)} \tag{6}$$

$$F_2(\underline{h}_j) = \frac{senh(\underline{h}_j)cos(\underline{h}_j) + cosh(\underline{h}_j)sen(\underline{h}_j)}{cosh(2\underline{h}_j) \quad cos(2\underline{h}_j)} \tag{7}$$

The total loss in the q'th winding is given by (8). It can be normalized by the power dissipated in it when $\underline{h}_1$ is equal to the unity, resulting in the normalized power given by (9).



(a)

(b)

Figure 1. (a) Cross-sectional view of a multiwinding magnetic component and (b) steps to transform a layer of solid round wire into an equivalent layer of foil conductor.

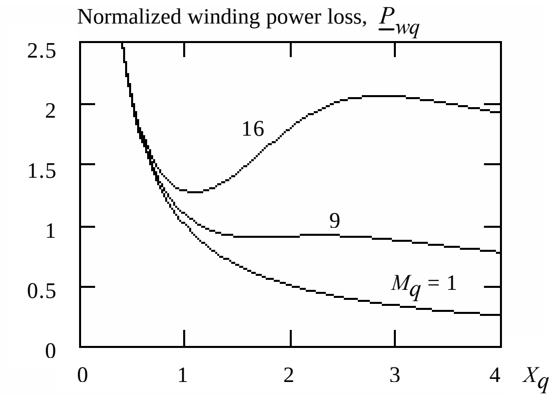


Figure 2. Normalized winding loss as a function of the X parameter.

$$P_{wq} = R_{eq} I_{efq}^2 \tag{8}$$

$$\underline{P}_{wq} = F_{rq} / \underline{h}_{1q} \tag{9}$$

The original thin layer model assumes that the winding structure is axisymmetric and constituted of foil conductors, which span the entire breadth of the core window. In the case of windings with solid round wire, bunched round wire, or litz wire, this model requires the replacement of the actual winding layers by equivalent-foil conductors, as in the original model as illustrated in Fig. 1(b) [7]. In this figure, the round conductors are replaced by square conductors of equal conducting cross-sectional area, which are brought together into a single layer foil, then stretched in order to fill the entire window breadth without changing the height of the layer. To compensate the increase in the cross-sectional area of the conducting layer, an effective conductivity $_c$ is used in the field equations for the layer. The layer porosity, , is defined as $= N_b h / b$, where N_b is the number of conductor cross-sections appearing in a cross-sectional view of the winding layer, and b is the core window breadth. Layers of bunched round wire or litz wire are transformed similarly [7]. Then, in the general case, the term $\underline{h}_j$ can be expressed as in (10), where X is given by (11).

$$\underline{h}_j = \sqrt{j} X \tag{10}$$

$$X = \sqrt{\ } \; h / \quad _1 \tag{11}$$

Then, (1) and (9) can be expressed, generically, in terms of X. Fig. 2 was obtained for a symmetrical triangular wave ripple current with a dc component. Without a dc current superimposed on this wave, for a certain value of M and even for $M = 1$ there is an optimum value for X, which can be determined from (12), so that the normalized power is minimum [7]. Fig. 2 shows that when a dc current is present, sufficient dc current causes the minimum for a given M to disappear. In such case, the implemented winding optimization procedure will require a greater number of foil layers. In this procedure the number of layers of equivalent-foil conductors is obtained for a given effective resistance, which allows the calculation of the number of strands in the winding cable. The respective optimum value of X allows the calculation of strands gauge.

$$\partial P_{wq} / \partial X = 0 \tag{12}$$

In the case of a multiwinding transformer with w windings the total loss in its windings is given by,

$$P_W = \sum_{q=1}^{w} R_{eq} I_{efq}^2 = \sum_{q=1}^{w} F_{rq} R_{dcq} I_{efq}^2 \tag{13}$$

B. Core Losses Modeling

Several attempts have been made to establish a reliable method suitable for measurement of core losses at high frequencies. In such measurements, the major challenge is to minimize phase errors introduced by parasitic effects in the circuit, by time delays in the coaxial cables and oscilloscope and by the digitizing process. Since a large amount of data is usually to be gathered, it is highly desirable that the measurement method used be automated in addition to being accurate and reliable. The automated measurement system for characterization of ferrite core losses described in [8] can accomplish these aims under sinusoidal voltage excitation and within specified magnetic induction and frequency ranges and at different temperatures with high accuracy, thanks to the techniques used in the acquisition and computation of the waveforms involved in loss calculation. In this system, the measurement data are used to fit the total core loss, P_c, vs. frequency, f, and peak magnetic induction, B, by applying the least chi-square plane method to (14), where V_e is the effective volume of the core. Some magnetic material manufacturers give the curve fitting parameters C_m, x, and y present in (14) for the respective frequency ranges in which the core loss was characterized. Some of them give the function that characterizes the dependence of C_m with temperature. However, practically no data have been published concerning measurements of ferrite core losses under non-sinusoidal voltage excitation and under dc biased excitation. Under these conditions, an expression different from (14) will be required in order to take into account the influence of the excitation waveform and specific magnetic effects that emerges under these conditions. However, once such expression is available, there is no apparent difficulty in took it into account in the mathematical formulation of the problem.

$$P_c = C_m f^x B^y V_e \tag{14}$$

C. Thermal Modeling

The accurate modeling of the mechanisms of heat transfer is a very complex task. Due to this complexity, the thermal modeling requires hypotheses in order to simplify the analysis. The thermal model used at the current development stage of the software is constituted of an equivalent thermal resistance of the component, R_{th}. The relationship between this thermal resistance and the maximum overall power loss allowed in the component, P_{tm}, for a given temperature rise, ΔT, is given by (15). Some ferrite core manufacturers give the value of R_{th} for each core, whereas others give an equation to calculate it as function of geometrical parameters. An example of such case is given in (16) [9], where $k = 30.5x10^{-3}$ and $n = -0.54$, which was obtained by measurements on EE, EC, EFD and ETD ferrite cores.

$$P_{tm} = \Delta T / R_{th} \tag{15}$$

$$R_{th} = k V_e^n \tag{16}$$

III. OPTIMIZATION METHODOLOGY

A. The Basic Equations for Transformer Design

The following are the basic equations for transformer design:

$$S_c = K_v A_e A_J f B \sum_{q=1}^{w} K_q J_{efq} \tag{17}$$

$$S_r = \sum_{q=1}^{w} S_q \tag{18}$$

$$R_{dcq} = N_q^2 l_t / (\rho_c K_q A_J) \tag{19}$$

Where:

$$K_v = 4 V_{efq} / (f \int_0^T |v_q(t)| dt) \tag{20}$$

$$S_q = V_{efq} I_{efq} \tag{21}$$

$$V_{efq} = K_v N_q A_e f B \tag{22}$$

$$I_{efq} = K_q A_J J_{efq} / N_q \tag{23}$$

In the above equations, S_c is an apparent power that express the transformer power transfer capacity, S_r is total required apparent power, K_v, is the voltage form factor, A_e is the effective cross sectional area of the core, A_J is core window area, l_t is the mean turn length, and the subscript "q" is used to describe the following variables related to the q'th winding: the dc resistance, R_{dcq}, the window utilization factor, K_q, the rms current density, J_{efq}, the rms voltage, V_{efq}, the rms current, I_{efq}, and the number of turns, N_q. The window utilization factor of the q'th winding, K_q, is defined by the ratio of the total conductive cross sectional area of this winding to the core window area. The window utilization factor, K_u, is defined by the sum of the window utilization factor of all windings and must be lower than 1.

The basic input data for designing a transformer are the following: core data (geometry, material grade and data), winding data (conductor types, turns ratios), current and voltage waveform data (Fourier components), bobbin data, insulation data, maximum temperature rise and maximum allowed magnetic induction. The design consists basically in determining the specific magnetic core, all the windings number of turns and respective conductor data suitable to build the transformer. Suitable means that the temperature rise will not be greater than its allowed maximum value, the winding will fit into the core window, the core will not saturate and will be capable to transfer the total required apparent power. The latter condition is mathematically expressed by (24).

$$K_v A_e A_J f\, B_d \sum_{q=1}^{w} K_{qd} J_{efqd} \quad \sum_{q=1}^{w} V_{efq} I_{efq} \qquad (24)$$

How to calculate K_{qd}, J_{efqd} and B_d in order to optimize the transformer design? The answer to this question is given in the following item.

B. Optimizing the Total Loss

By ignoring the dielectric losses, the transformer total loss is given by:

$$P_t = P_w + P_c \qquad (25)$$

Taking into account (13), (19), (21) and (22) the total winding loss can be expressed by (26).

$$P_c = \frac{l_t}{c A_J (K_v A_e f B)^2} \sum_{q=1}^{w} \frac{F_{rq} S_q^2}{K_q} \qquad (26)$$

From (26) it should be noted that the winding loss presents an inverse dependence on the peak magnetic induction. This originates from (22), which expresses the fact that increasing the peak magnetic induction there is a tendency to reduce the number of turns, and hence to reduce the winding dc resistance. On the other hand, from (14) it should be noted that the core loss increases with increasing peak magnetic induction. Therefore, it may be possible to optimize the transformer total loss with respect to peak magnetic induction. Defining B_o to be the value at which the derivative of P_t with respect to B is zero and P_{wo} and P_{co} the resultant values for winding and core losses yields (27), which gives the optimum distribution of winding and core losses for minimum transformer total loss.

$$P_{wo} = y P_{co} / 2 \qquad (27)$$

Combining (15), (25) and (27) yields

$$P_{wo} = (y / y+2)(\ T / R_{th}) \qquad (28)$$

$$P_{co} = (2 / y+2)(\ T / R_{th}) \qquad (29)$$

Therefore, from (14) and (27) the following expression for the optimum peak magnetic induction yields

$$B_o = \left[2\ T \Big/ (y+2) R_T C_m f^x V_e \right]^{1/y} \qquad (30)$$

However, as mentioned before, two distinct design conditions can occur. The design condition can be identified from the analysis of the dependence of the optimum peak magnetic induction on frequency. From (30) it should be noted that the peak magnetic induction increases with decreasing frequency. Therefore, there is a frequency at which the optimum peak magnetic induction is equal to the respective maximum value allowed by the designer, given by $B_m = g B_s$, where B_s is the saturation magnetic induction. An expression for this frequency, named transition frequency, is given by (31). If the transition frequency is lower than the voltage fundamental frequency, then the design is limited by core loss and the peak magnetic induction is B_o, which is lower than B_m. In this case, a transformer with minimum overall power loss and volume may be designed. On the other hand, if the transition frequency is greater than the voltage fundamental frequency, then the design is limited by core saturation and the design peak magnetic induction is B_m. In this case, a transformer with minimum volume may be designed, but the overall power loss will not be minimum. The first step in a design procedure is to establish the design condition. The value of B_d in (24) depends on this condition. In the design limited by core loss, one has $B_d = B_o$ and the total winding loss is limited according to (28). In the design limited by core saturation, one has $B_d = B_m$ and the total winding loss is limited according to (32). As can be seen from (23), the value of J_{efd} depends on K_q, which are obtained in the implemented procedure for designing the windings.

The q'th winding effective resistance for a design limited by core loss or for a design limited by core saturation can be obtained from (28) or (32), respectively, combined with (8). To do this, it is assumed that total loss in the windings is equally distributed among them. Therefore, taking into account (2), (9), (10), and (19) yields the expression for K_q given by (33).

$$f_t = [P_{co} / (C_m B_m^y V_e)]^{1/x} \tag{31}$$

$$P_{wm} = P_{tm} \quad P_c \tag{32}$$

$$K_q = l_t N_q^2 X_q \underline{P}_{wq} / (\quad {}_c A_J R_{eq}) \tag{33}$$

Taking into account (16), (29) and (31), the following expression for the minimum effective volume that a core must have to be candidate to a design limited by core loss yields

$$V_{e\,min} = \left[2 \quad T / k(y+2) C_m f^x B_m^y \right]^{1/(1 \quad n)} \tag{34}$$

This completes the fundamental expressions and restrictions used in the implemented procedure for design optimization of high frequency power transformers.

IV. Design Example and Software Features

The main features of the implemented software will be described through a design example. The application example which follows is based on a 550 W, 48 V, 40 kHz zero-voltage switching pulse-width modulated active clamping forward converter (ACF) [10]. Fig. 3 shows the basic converter power stage diagram, which was simulated through PSpice . The converter operates at a duty cycle equal to 0.32. The transformer magnetization inductance was made equal to 1.6 mH, so that the resonant inductor current ripple is 2 A.

The transformer optimization design algorithm was implemented by using DELPHI . Fig. 4 (a) shows the main screen of the implemented software, which was named MAGNO. This software includes databases concerning conductors and ferrite cores and accessories, as can be seen from Figs. 4 (a), (b) and (c). The former file includes data about solid and bunched enamelled round copper wires and litz wires. There is also an option that allows the software to design the bunch of round wires by using solid round wires (Fig. 4 (b)). Fig. 4 (c) shows that the latter file includes all core and bobbin geometrical data, the parameters C_m, x, y for all frequency ranges in which the core loss was characterized, and the saturation magnetic induction (some data are hidden in these figures). Wherever possible the parameter C_m is calculated as a function of temperature. The software allows the user to add new data on databases from your preferred manufacturers. Figs. 4 (a), (b) and (c) shows that the input data include: fundamental frequency, ambient temperature, temperature rise, fraction of saturation magnetic induction allowed for transformer operation, core data (geometry, material grade, type of bobbin terminals, manufacturer), winding data (conductor type and its manufacturer, turns ratios, position from the inside out), current and voltage waveform data (number of Fourier components to be considered and their dc, peak and phase angle values), and permissivities of the insulating materials among others. The program allows the design of transformers with until four primary windings and twelve secondary windings and allows the user to move windings around from one position to another. Fig. 4 (d) shows that the output data include: core data (manufacturer code, peak magnetic induction, magnetic loss), winding data (copper loss, number of turns, wire gauge (AWG and ABNT metric), number of strands per conductor, number of conductors per cable, number of layers, effective and dc resistances and the ratio between them, rms current density, effective leakage inductance), equivalent parasitic capacitance, temperature rise, window utilization factor, power transfer capacity, required apparent power, and efficiency among others.

In this example, the first ten harmonic of the waveforms were taken into account and EE core geometry of F grade ferrite from Magnetics Inc. was chosen. The option for designing "bunched round wire" was selected and the percentage of saturation magnetic induction was 90 % ($g = 0{,}9$). In this case, it was possible to design the transformer with minimum overall power loss and volume. In fact, it can be noted from the results that there is a "practical" optimum distribution of winding and core losses and the transformer efficiency is 98,82 %. From the output data, it can also be noted that: the current densities in the windings are lower than the respective maximum commonly used value, i.e. 5 MA/m^2, although not always this happens, the magnetic induction is much lower than its saturation value, i.e. 0.36 T for the F grade ferrite at 100°C, the temperature rise is lower than the specified maximum value, i.e. 45°C, and that the power transfer capacity is higher than the required apparent power. Unlike classical design procedures, the window utilization factor is an output data. Besides, it is particularly noteworthy that the number of layers results integer, which results from the original purpose to design a transformer while respecting theoretical models and practical aspects. As a final note, the volume of the selected core (EE 5530) is smaller than that specified in [10] (EE 6526).

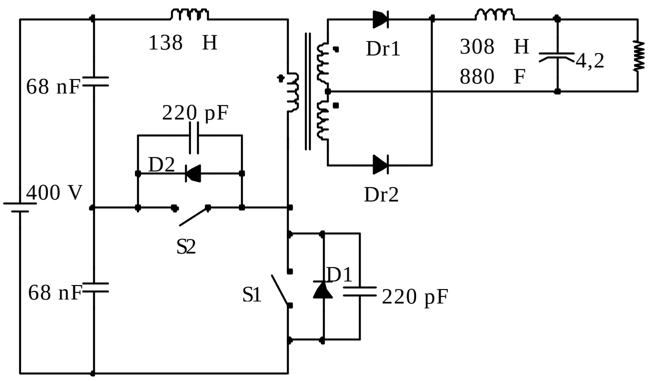
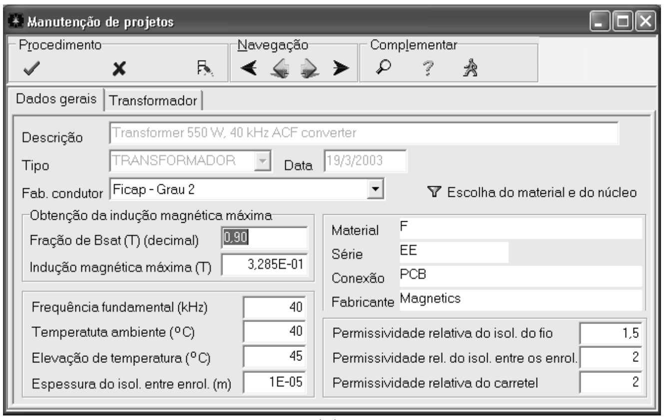
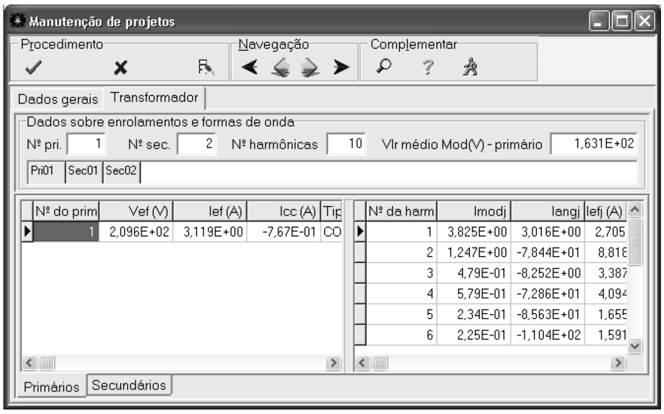


Figure 3. ACF converter power stage diagram.



(a)



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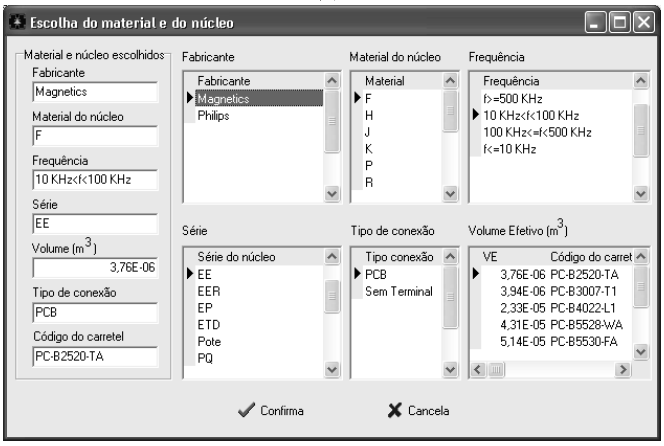

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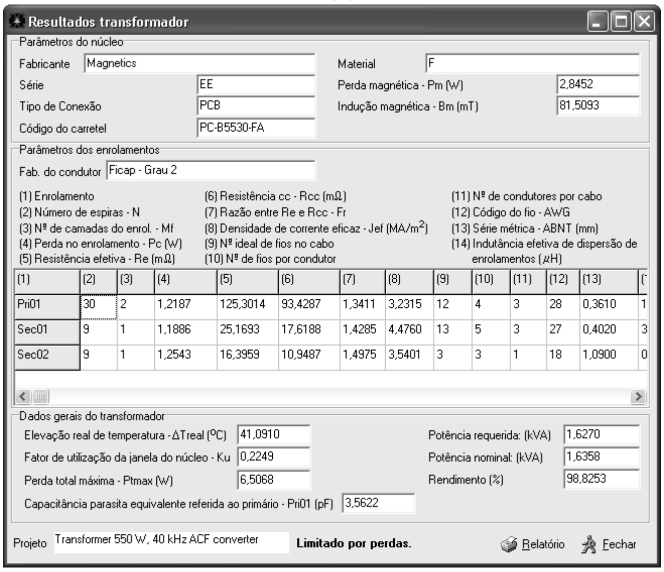

(d)

Figure 4. Software screens: (a) main screen, (b) windings input data, (c) core and material input data, (d) output data.

V. CONCLUSIONS

A new methodology and the respective software to optimize the design of high frequency power transformers have been described. The program searches for the smallest core that meets the design specifications and concurrently, wherever possible, it minimizes the total transformer loss. In the process of winding design optimization, it takes full account of the current waveform and skin and proximity effects and searches for the cables that better fits to the design. It can be applied to any circuit topology provided the voltage and current waveforms are periodic. By using the developed software, the designer can explore many design possibilities in order to choose the most adequate one for its application. An application example is given for a transformer of a 550 W, 40 kHz active clamping forward converter (ACF).

VI. ACKNOWLEDGEMENTS

This work was supported by the Campanhia Energética de Goiás – CELG in the context of its R&D policy.

VII. REFERENCES

[1] Intusoft. Magnetics Designer Manual, 252 p., 1997.

[2] P. M. Gradzki, M. M. Jovanovic, F. C. Lee, "Computer-aided design for high-frequency power transformers", *in Applied Power Electronics Conference,* pp. 336-43, 1990.

[3] R. Petkov, "Optimum Design of a High Power, High Frequency Transformer", *IEEE Trans. on Power Electronics*, v. 11, n. 1, p. 33-42, 1996.

[4] W. G. Hurley, W. H. Wölfle, J. G. Breslin, "*Optimized Transformer Design: Inclusive of High-Frequency Effects*", *IEEE Trans. on Power Electronics*, v. 13, n. 4, p. 651-659, 1998.

[5] J. P. Vandelac and P. Ziogas, "A Novel Approach for Minimizing High Frequency Transformer Copper Losses", *IEEE Trans. on Power Electronics*, v. 3, n. 3, pp. 266-277, 1988.

[6] J. A. Ferreira, "Improved Analytical Modeling of Conductive Losses in Magnetic Components", *IEEE Trans. on Power Electronics*, v. 9, n. 1, pp. 127-131, 1994.

[7] A. J. Batista, J. C. S. Fagundes, P. Viarouge, "Effective Resistance and Leakage Inductance in Multiwinding Transformers: Modeling, Measurement and Comparison with Numerical Solution", *in Congresso Brasileiro de Eletrônica de Potência*, pp. 731-736, 1999.

[8] A. J. Batista, J. C. S. Fagundes, P. Viarouge, "An Automated Measurement System for Core Loss Characterization", *IEEE Trans. on Instrumentation and Measurement*, v.48, n.2, pp. 663-667, 1999.

[9] S. A. Mulder, "On the Design of Low Profile High Frequency Transformers", in *Proc. of the HFPC*, pp. 141-159, 1990.

[10] G. J. Maass, A. F. Souza, I. Barbi, "Analysis and experimentation of a forward converter with active clamping and output rectifier using a center tapped transformer", in *Power Electronics Specialists Conference*, v. 4, pp. 1767-1772, 2001.