Estimation of MnZn Ferrite Core Losses in Magnetic Components at High Frequency

Ermanno Cardelli, *Senior Member, IEEE*, Lorenzo Fiorucci, and Edward Della Torre, *Fellow, IEEE*

***Abstract*—This paper deals with a magnetic modeling for the estimation of the dynamic losses in soft ferrite cores. A time-domain MSPM based algorithm for the hysteresis losses evaluation, added to a simplified, one-dimensional modeling of the eddy current losses, are used to predict the total specific losses in cores subjected to a.c. and distorted wave-forms. Experimental analysis show satisfactory capability of prediction of the modeling proposed, used as a CAD tool.**

***Index Terms*—Eddy current losses, hysteresis losses, modeling, soft ferrites.**

I. Introduction

THE USE of Manganese–Zinc soft magnetic ferrites in transformers and inductors to power electronic and telecommunications applications is continuously increasing. In using these ferrites one of the most important determinants is the estimation of the core losses in function of the working value of the magnetic induction and for the design range of frequency. Preliminary considerations show that in the range up to 100 kHz the eddy current losses are a very small part of the total core losses and the dielectrics losses are negligible up to 1 MHz, so in the typical field of frequencies of the power electronics applications hysteresis losses are often dominant.

In this paper we present a hysteresis losses modeling technique that, taking into account a simplified dynamic analysis of the eddy current losses, has the capability to predict the amount of specific power losses in the magnetic core of the component versus frequency and amplitude of the magnetic induction. The identification of the parameters of the proposed magnetic hysteresis modeling is independent from the size and the geometry of the core and is derived only by the major magnetic loop and the virgin curve, added to the usual magnetic and electric parameters of the ferrite used.

II. Preliminary Considerations About Power Losses in Soft Ferrite Cores

A. Grain Losses

Eddy current losses inside grains are negligible compared to eddy current losses in the core. Assuming that grains are spheres and that the eddy currents inside grains don't change substantially the total value of the magnetic induction we can approximate the unitary power loss due to the eddy currents circulating in the grains by the following expression:

$$P_g = \frac{3\omega^2 \sigma_g B_{peak}^2 R_g^2}{40} \tag{1}$$

where

ω	is the angular frequency,
σ_g	is the mean electrical conductivity of the grains,
B_{peak}	is the peak value of magnetic induction, and
R_g	is the mean radius of the grains.

B. Eddy Current Losses

The dynamic power losses in the core can be estimated in closed form, by assuming that the eddy current paths in the core are circular and the current flows along the cross sections of the core, and that the presence of the eddy currents don't modify the value and the direction of the total magnetic induction, perpendicular to the cross sections of the core. If the core has O-shape and its cross section is circular; the specific eddy current power losses can be estimated by the expression:

$$P_{ec} = \frac{\omega^2 \sigma_c B_{peak}^2 R_c^2}{16} \tag{2}$$

where R_c is the external radius of the cross section of the core. Assuming $R_g \leq 10\ \mu$m, $\sigma_g \leq 1000\ \Omega^{-1}\ \mathrm{m}^{-1}$, and σ_c in the range $0.1 \div 5\ \Omega^{-1}\ \mathrm{m}^{-1}$ [2], [4] it can be proved that $P_{ec} \gg P_g$ for the usual values of working magnetic induction and for the typical dimensions of the cores.

C. Dielectric Losses

In the same hypothesis assumed in the previous case, and with the additional assumption that the displacement currents between the grains have again the same circular paths of the conduction currents, the dielectric losses can be approximated by the expression:

$$P_d = \frac{\omega^4 \varepsilon_c^2 g_c B_{peak}^2 R_c^2}{16\left(\omega^2 \varepsilon_c^2 + g_c^2\right)} \tag{3}$$

where ε_c is the mean electrical permittivity and g_c is the equivalent mean electrical conductivity in the inter-granular regions. Using the values $\varepsilon_c = 10^{-7}\ \mathrm{Fm}^{-1}$ and $g_c = 10\ \Omega^{-1}\ \mathrm{m}^{-1}$ [2], [3] it is possible to show that up to 1 MHz the hysteresis losses and eddy current losses in the core section are dominant.

III. Numerical Modeling of the Hysteresis Losses

Usually the data sheets given by the manufacturers of soft ferrites report only the data about the major loop and the first

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E. Cardelli and L. Fiorucci are with the Department of Industrial Engineering, University of Perugia (e-mail: ermanno2@isten.ing.unipg.it).

E. Della Torre is with the George Washington University, Department of Electrical Engineering and Computer Science, Washington, D.C. 20052.

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magnetization (virgin) curve. In this case a useful tool to calculate the hysteresis losses can be the Modified Scalar Preisach Model (MSPM) [1]. The MSPM uses as input the Preisach Probability function and the amount of the reversible part of the magnetization and gives in output the minor loops in the plane H, M. The key points of MSPM are:

- Stochastic independence between switching probabilities in order to express the probability function is a product of two single-value functions.
- Introduction of a reversible part of magnetization added to the Preisach probability function.
- Introduction of a product function in order to simulate the experimental behavior of "noncongruency" of minor loops.
- Possibility to treat multi-frequency waveforms and d-c components added to a.c. components.

We have applied the MSPM to the problem of the estimation of the hysteresis losses, and we have derived the relation:

$$A = \int_{H1}^{H2} \Gamma(H_1, U, -V)\, dH - \int_{H2}^{H1} \Gamma(H_2, -V, U)\, dH \tag{4}$$

where A is the area of the hysteresis loop, and:

$$\Gamma(X, Y, Z) = \int_X^H K(|M(Y)|) \cdot \left[W(Y) + P_S(Y) \int_X^Y P_S(-Z)\, dZ\right] dY. \tag{5}$$

IV. Verification and Discussion

We have applied the magnetic modeling technique described in the previous section to commercial O-Shape, C-Shape, and M-Shape Mn–Zn ferrite Cores. Assuming the hypothesis of local linearity of the ferrite we have calculated in 1-D (cylindrical coordinates) the distribution of magnetic induction and current density into the core, and we have used these results to evaluate the hysteresis losses by the:

$$P_h = \frac{\int_{\Omega c} A d\, \Omega c}{\int_{\Omega c} d\, \Omega c} f \tag{6}$$

and the eddy current losses by the:

$$P_{ec} = \int_{\Omega c} \frac{J^2}{\sigma_c}\, d\, \Omega c \tag{7}$$

where Ωc is the volume of the magnetic core. Under the hypotheses made in Section II-B (7) became (2). Typical computational time for the modeling parameters identification is 20–40 sec, using a 400 MHz Pentium III INTEL© processor and a dedicated code developed in MATLAB© environment. In Figs. 1–4 are reported the specific power losses either calculated using this approach, or measured. Measurements have been performed by the manufacturer [5], [6] at 25 and 100 Celsius degrees. We have seen (Figs. 1 and 2) that up to 100 kHz the

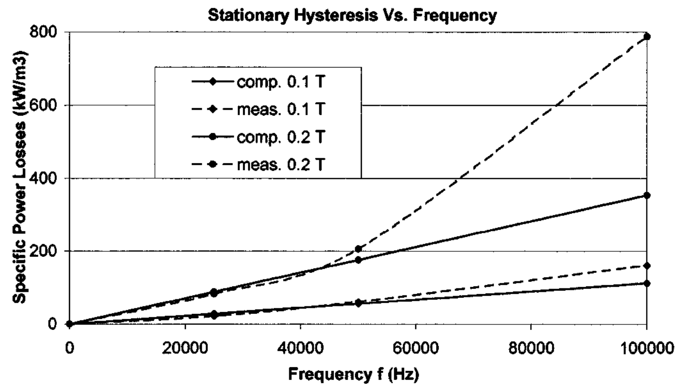

Fig. 1. Low frequency losses vs. frequency in linear zone (0.1 T), and in saturated region (0, 2 T). $MnZnFe_2O_4$O-shape core. Length 300 mm, height 220 mm. Cross section (square) 40×40 mm. Temperature 25 degrees Celsius.

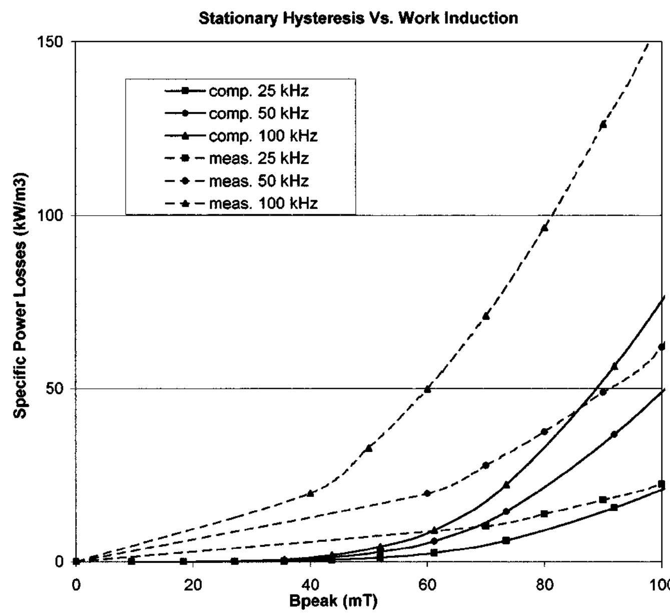

Fig. 2. Low frequency losses vs. the rate of magnetic induction. Same core and temperature as Fig. 1.

hysteresis losses dominate the total losses. Moreover when the magnetic induction is limited within the linear zone of the behavior of the material (far from the saturation level), i.e., if the hypothesis of local linearity is in fact verified, and taking into account that up to these frequencies the magnetic skin depth is larger than R_c, the core losses can be approximated using only the d.c. data. In Figs. 3 and 4 the hysteresis losses are calculated using the MSPM and the eddy current losses amount is added. Dielectric losses were calculated using (3) and resulted negligible up to 1 MHz. Comparison with the measured data show agreement. The predicted curves fit the experimental ones up to 1 MHz and for all the values of the rate of magnetic induction. The predictive capability of the proposed model can be used also in the case of nonsinusoidal waveforms. This is an important aspect, taking into account the soft ferrite magnetic components that work to power electronic devices in which exists a consistent d.c. level. Figs. 5 and 6 show the calculated dynamic specific power losses when a consistent d.c. component is superimposed on the a.c. one. In

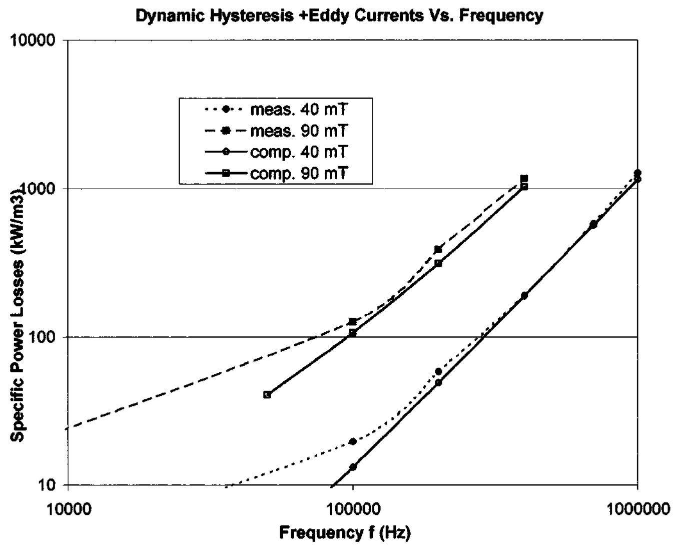

Fig. 3. Dynamic losses vs. frequency. Same core as Fig. 1. Temperature 100 degrees Celsius.

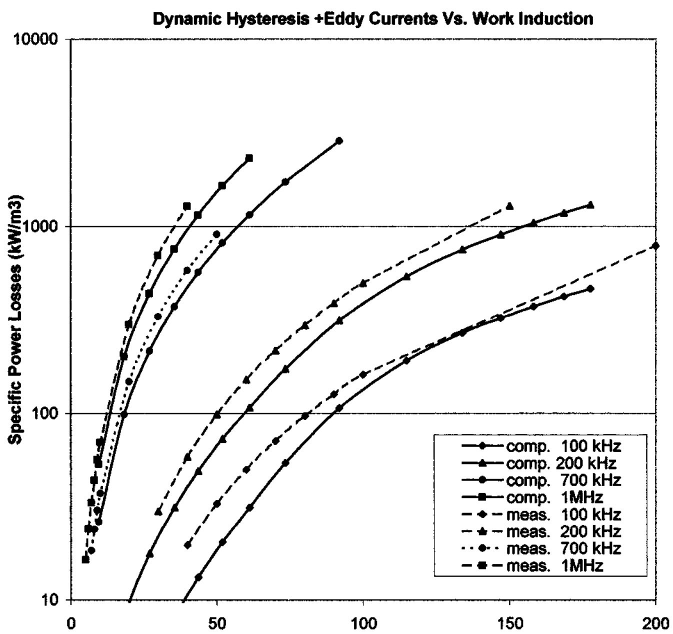

Fig. 4. Dynamic losses vs. the rate of magnetic induction. Same case as Fig. 3.

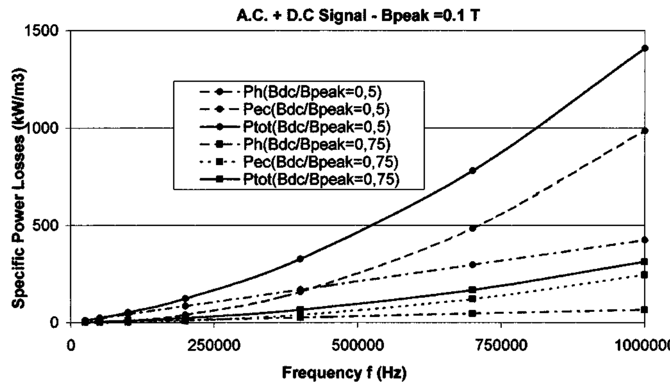

Fig. 5. Computed dynamic magnetic core losses vs. frequency with distorted wave-forms (linear scale), for different ratio of Bdc on Bpeak. Bpeak=Bdc+Bac. Bac sinusoidal. Same core and temperature as in Fig. 1.

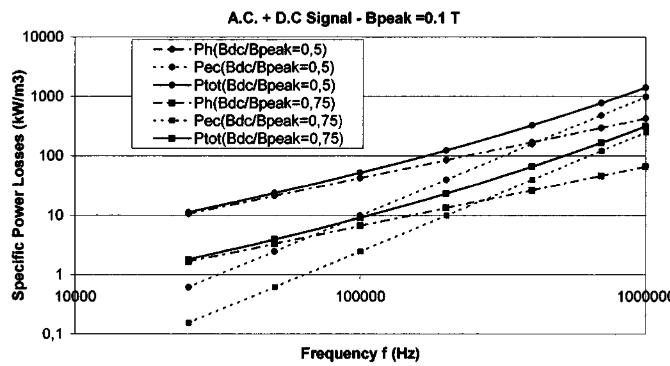

Fig. 6. Computed dynamic losses vs. frequency with distorted wave-forms (logarithmic scale). Same case as Fig. 5.

the figures are also shown separately the hysteresis and the eddy current components of the power losses. We have investigated several cases and we have found substantial agreement with the experimental data up to 1 MHz, when most likely due to the presence of consistent dielectric losses the measured and predicted curves diverge. In any case the influence of the conductive, and of the displacement currents in the core on the actual value and orientation of the magnetic and electric field deserves further attention.

V. Conclusions

We have presented in this paper a magnetic modeling technique in order to predict the dynamic losses in Mn–Zn ferrite cores. The procedure consumes an acceptably low amount of computer time and needs only for general data of the material, usually available on the data sheets given by the main producers. The principal features of the modeling technique described above include the separate treatment of hysteresis and eddy currents and the possibility of working in time-domain, allowing the analysis of dynamic losses of non-sinusoidal wave-forms. Experimental verification has revealed that up to the level of dielectric losses, the modeling technique is sufficiently accurate for usual design requirements.

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