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Analytical and field-circuit core loss prediction in induction motors

***Abstract**. The paper presents two methods for core loss prediction in induction motors that take into account harmonic flux density due to the slotting effect. The first method uses an analytical formulation to determine the flux density distribution in the motor cores. The second one is based on finite element analysis. In the both cases, an iron loss model in non sine excitation condition has to be defined. The complementarities of the both methods are displayed.*

***Streszczenie.** Artykuł prezentuje dwie metody obliczania strat w silniku indukcyjnym uwzględniające efekty związane z żłobkowaniem stojana i wirnika. Pierwsza metoda używa zależności analitycznych do określenia rozkładu indukcji w silniku. Druga wykorzystuje metodę elementów skończonych uwzględniającą ruch wirnika oraz rozwiązanie równań pola względem czasu. Dla obydwu przypadków konieczne było określenie modelu strat w żelazie przy wzbudzeniu niesinusoidalnym (wieloharmonicznym). Pokazano, że obie metody prowadzą do poprawnego wyznaczenia strat w silniku indukcyjnym. (**Analityczne i polowe prognozowanie strat w silniku indukcyjnym**)*

Keywords: Induction machine, iron losses, finite element, analytical analysis
Słowa kluczowe:silniki indukcyjne, straty w żelazie, metoda elementów skończonych, metody analityczne..

Introduction

Resulting from the recent energetic considerations, there has been a tendency to design high efficiency electrical machines. For that, the knowledge of AC machine iron losses is an important stage in the design of a machine and the designers are looking for fast, simple, accurate and reliable methods for iron loss estimation.

One of the difficulties of iron loss estimation in AC machines concern the presence of harmonics superposed to the flux density fundamental. These harmonics appear in the airgap of the machine and they mainly originate from the slotting effect. Then, these harmonic components spread within the stator and the rotor cores with a radial and a tangential component which distribution depends on the characteristic of the flux density component. These harmonics have low magnitude but high frequency so they may generates non negligible additional losses [1]. A reliable losses valuation requires a modelling of the machine able to take into account the slotting effect.

In that paper, two modelling methods are presented. The first one uses analytical formulations of the airgap flux density and its distribution in stator and rotor cores. The second one is based on time stepping finite element analysis, that considers rotor movement, associated with an electrical circuit. A numerical application is performed on a 4kW, four-pole wound rotor induction machine.

Analytical analysis

The analytical analysis does not allow an accurate estimation of the iron losses. However, it gives the contribution of each harmonic flux density and highlighted particular properties of the harmonics. The analytical model is based on three models that can be combined to each other. They concern: the airgap flux density, the stator and rotor core flux density distribution, and the iron loss model.

Air-gap flux density model

The air-gap flux density b^g is obtained by multiplying the air-gap magnetomotive force by the air-gap permeance per area unit which takes the slotting effect into account. Different hypothesis have to be performed for the b^g calculation:

- The field lines are considered radial in the airgap.
- The slots are considered rectangular with a fictitious depth equal the fifth of their opening.
- The magnetic field in the stator and rotor cores is neglected compared to the airgap one.
- Only the stator excitation at no load is taken into account. According to the Lenz law, it is supposed that the magnetic reaction of the rotor does not bring additional flux density components.
- The stator is energized by a three-phase, balanced, sine current system of I^s rms value and ω angular frequency (f frequency).

b^g will be determined in a stator referential tied to the phase one axis and in a rotor referential tied to one rotor tooth axis. In a stator referential, b^g can be expressed as follows [2]:

$$b^g = \sum_{h^s,ks,kr} \hat{B}_{h^s,ks,kr} \cos(K\omega t - Hp\alpha^s) \tag{1}$$

α^s represents the angular abscissa which locates any point of the air-gap in a stator referential, p is the machine pole pair number. The K frequency rank and the Hp pole pair number of an elementary flux density component are defined as:

$$\left.\begin{aligned} K &= 1 + krN^r(1-s) \\ H &= h^s + ksN^s + krN^r \end{aligned}\right\} \tag{2}$$

s is the slip, N^s and N^r represent the stator slots and bars number per pole pair, ks and kr are positive and negative integers, h^s is defined as h^s=6λ+1 (λ varies from -∞ to +∞). At given α^s, the flux density varies at a Kf frequency, with a magnitude $\hat{B}_K$ obtained by a vectorial summation of various components of different pole pair number Hp. It has been shown in [3] that only few component have a predominant contribution in a Kf frequency harmonic generation.

In the rotor referential, one considers the relation: $\alpha^s=\alpha^r+\theta$, where α^r represents the angular abscissa which locates any point of the air-gap in the rotor referential, θ is the position of the rotor referential axis relatively to the stator one. Then, the airgap flux density, denoted b'^g in the rotor referential, can be expessed as follows:

$$b'^g = \sum_{h^s,ks,kr} \hat{B}_{h^s,ks,kr} \cos(K'\omega t - Hp\alpha^s) \tag{3}$$

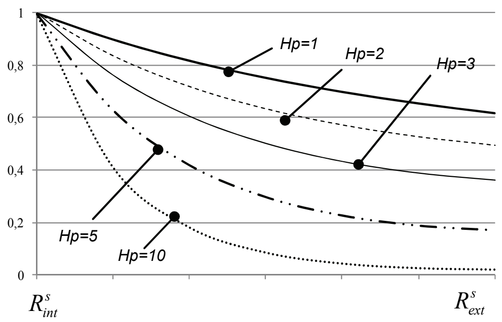
where K' is defined as:

$$K' = 1-(h^s + ksN^s)(1-s) \tag{4}$$

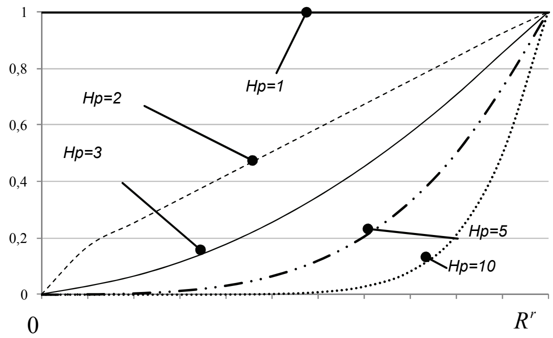
At given α^r, the flux density varies at K' frequency, with a magnitude $\hat{B}_{K'}$ still obtained by a vectorial summation of various components of different pole pair number Hp. It is obvious that the frequency related to the rotor referential are different than these related to the stator one.

Core flux density model

The analysis of the contribution of the harmonic flux density components in the iron losses requires the knowledge of their distribution within the stator and the rotor cores. Through a model based on smooth armatures, it has been shown in [4] that the penetration of a flux density component in the iron depend on its pole pair number Hp : the more H increases, the more the component is attenuated. Figure 1 gives the relative variation of the tangential component of flux density through the stator and the rotor cores.



(a)



(b)

Fig. 1 – Attenuation of the tangential component of the flux density in the stator (a) and the rotor (b) cores.

Iron loss model

Core losses in induction motors are usually represented using two components: hysteresis and eddy-current [5], [6], or following methods based on the suggestion by Bertotti [7] of having three components: classical eddy-current, hysteresis and excess losses. The specific iron losses caused by a sinusoidal magnetic flux density of $\hat{B}$ peak value and f frequency may be described by the three-component method as:

(5) $$p_T = k_e f^2 \hat{B}^2 + k_h f \hat{B}^{\alpha} + k_a f^{1.5} \hat{B}^{1.5}$$

where the coefficients k_e, k_h, k_a and α are calculated from the measured data and are normally assumed to be constant. However, to correctly specify the behaviour of specific losses for the motor under investigation, including harmonic flux density, measurements have been taken on a toroidal core excited at variable frequency. Figure 2 gives the specific core losses (W/kg) versus flux density peak for different frequencies. Each curve is then approximated by third order polynomials [8]. At 50 Hz the polynomial takes the form:

(6) $$p_{T(50)} = 0.8383\hat{B}^3 + 0.1222\hat{B}^2 + 1.3393\hat{B} - 0.0836$$

while at 250 Hz:

(7) $$p_{T(250)} = -1.4559\hat{B}^3 + 22.824\hat{B}^2 + 0.2589\hat{B} + 0.0489$$

and finally for 2000 Hz:

(8) $$p_{T(2000)} = 549.8\hat{B}^2 + 10.6\hat{B} + 0.005$$

This approach does not allow one to decompose the losses between static and dynamic losses but it provides a more accurate loss model.

Analytical results

The studied machine is a three phase, 4 pole, 230/400V, 4kW, wound rotor induction machine, with 36 stator slots and 24 rotor slots (N^s=18, N^r=12). The geometry of the machine is given in figure 3. The analysis performed in that study concerns a running at no load (s=0). The aim is not to provide numerical results of losses but it consists in presenting the properties of the iron losses associated to the various flux density harmonic components.

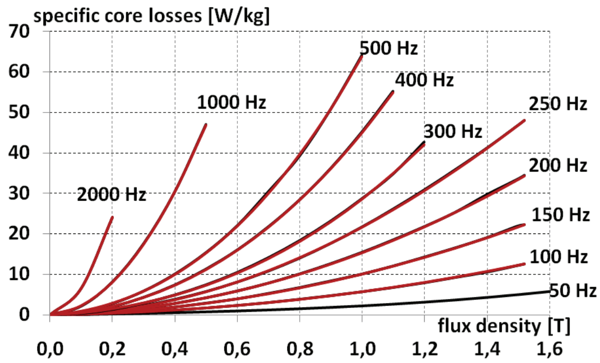


Fig. 2 – Specific core losses

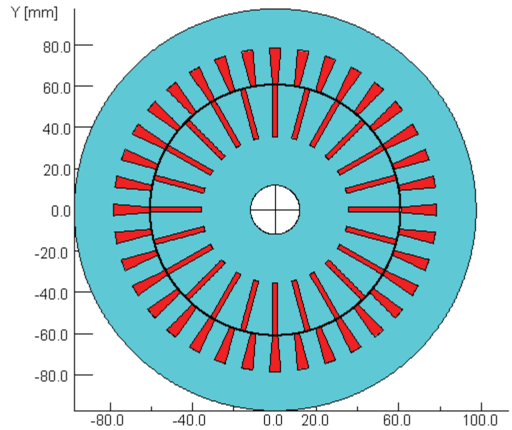


Fig. 3 – Machine geometry

In the following, various harmonic at Kf (or $K'f$) frequencies are analysed. For each, one will give the predominant elementary component at given h^s, ks, kr.

Losses in the stator armature:

- h^s=1, ks=0, kr=0 leads to K=1, H=1, Kf=50Hz : this component correspond to the fundamental of the flux density, it is not very attenuated, consequently, the associated losses occurs in all the stator core.

- h^s=1, ks=0, kr=±1 leads to K=-11, +13, H=-11, +13, $|Kf|$=550Hz, 650Hz. These components of high frequency have also a high polarity, so they are strongly attenuated and the corresponding losses are located at the top of the stator teeth.
- $h^s=1$, ks=0, kr=±2 leads to K=-23, +25, H=-23, +25, $|Kf|$=2350Hz, 2450Hz. These components have the same property than the previous ones but with lower effect due to lower magnitude and higher polarity (much more attenuated).

Losses in rotor armature:
- h^s=-5, +7 , ks=0, kr=0, leads K'=±6, H=-5, +7, $|K'f|$=300Hz. These components correspond to stator space harmonics which generate time variation flux density in the rotor. The polarity is not so high than the slotting harmonics ($ks \neq 0$ or $kr \neq 0$) and the components are not so attenuated.
- h^s=1, ks=±1, kr=0, leads to K'=±18, H=-17, 19, $|K'f|$=900Hz. These components of high frequency and high polarity are consequently located at the top of the rotor teeth.

Losses in stator and rotor armature: Slotting resonance harmonics
- h^s=1, ks=±2, kr=±3, leading to K=-35, +37, K'=±36, H=1, $|Kf|$=1750Hz, 1850Hz, $|K'f|$=1800Hz. Contrary to the previous ones these particular components act in the stator and in the rotor, with different frequencies. Another peculiarity concerns the low polarity of those components, which is equal to the polarity p of the machine but also this of the fundamental. Consequently, the attenuation in the stator and rotor cores is the same than the fundamental and the corresponding losses are located in the same area than these generated by the fundamental. A study presented in [9] based on dynamic losses analysis, gives an approximation of the ratio between the dynamic losses P_{dK} generated these harmonics and P_{d1}^{s} generated by the fundamental in the stator:

(9) $P_{dK} / P_{d1}^{s} \cong 2(1+\lambda_d)(\Delta\hat{B}_K)^2\left[1+(1-s)^2 K^2\right]$

λ_d is the ratio between the rotor and the stator volume (λ_d=0.5 for the considered machine). It is supposed that stator and rotor iron have the same losses characteristic. $\Delta\hat{B}_K$ is defined by: $\hat{B}_K / \hat{B}_1$, $\hat{B}_1$ is the magnitude of the fundamental flux density. According to the machine geometry, the numerical application leads to $\Delta\hat{B}_K$ =0.0068, leading to $P_{dK} / P_{d1}^{s} = 0.18$. It can be noticed that a very low magnitude harmonic has a strong contribution in the dynamic losses.

Field circuit analysis

The simulation, performed with the Opera-2D/RM solver yield the time variation of the flux density at each node of the mesh (figure 4). Then a Discrete Fourier Transform is applied to the flux density at each node, what enables the use of the iron loss model within an elementary volume. The iron losses associated to one frequency are obtained by summation on the whole elementary volumes constituting the machine.

Figures 5 and 6 show the distribution of the magnitude of the magnetic flux density within the stator and the rotor cores for four given harmonics. The simulation confirms the analytical results concerning the location of the losses, depending on the polarity of the component.

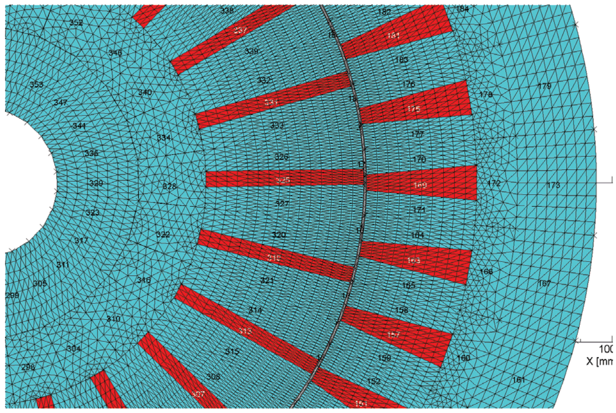
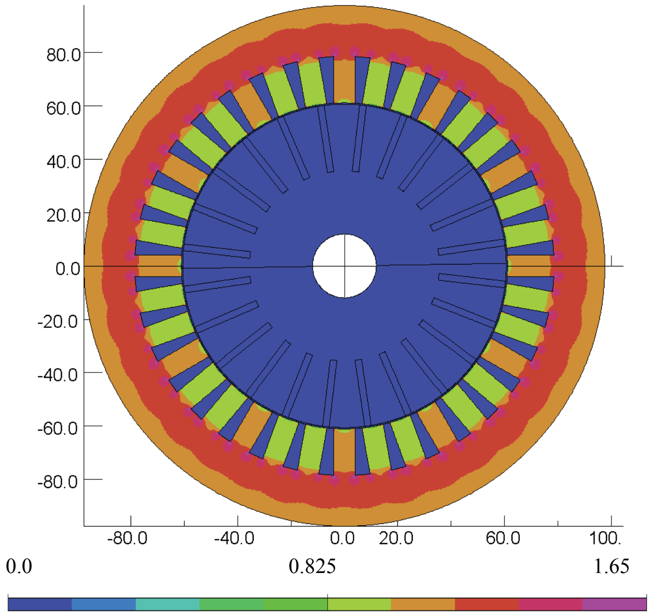
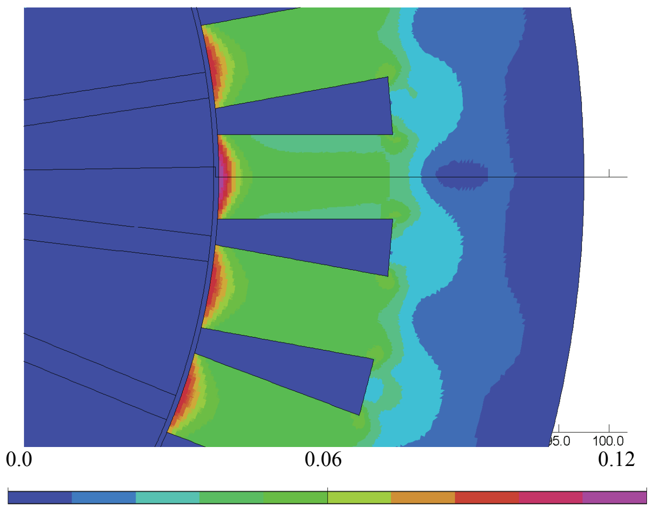


Fig 4 : mesh of the geometry



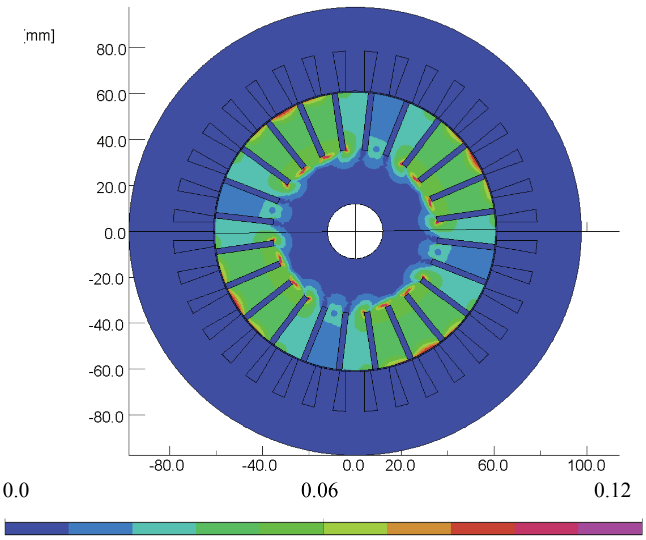
(a) - 50Hz (K=1)



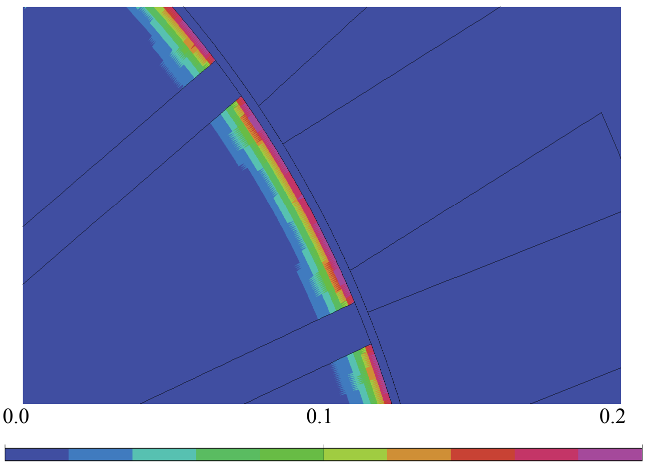
(b) – 550 Hz (K=-11)

Fig. 5 – Magnitude of the magnetic flux density for the 50Hz (a) and the 550Hz (b) components

contribution in the iron losses. The analytical and the finite element approaches have to be combined to define procedure for iron loss reduction, by acting on the design or on the supply.



(a) - 300Hz (K'=6)



(b) - 1800Hz (K'=36)

Fig. 6 – Magnitude of the magnetic flux density in rotor distribution in rotor for the 300Hz (a) and the 1800Hz (b) components

Figure 7 gives the losses associated to each harmonic flux density. The fundamental, not given in the figure, is equal to 44W. The data merge the stator and the rotor losses. The losses associated to the rank K=35,37, K'=36 correspond to 10% of the fundamental wheras the analytical approach that use equation (9) lead to 18%. The difference can be justifiy by the used loss models. Actually, the analytical approach [9] considers only the dynamic losses whereas the numerical application uses a model which does not differenciate static, dynamic and excess losses. It is well known that the static losses affect essentially the fundamental, taken as the reference in the given percentages.

Conclusion

Two methods for induction machine iron loss analysis are presented. The analytical approach provides qualitative information related to the origin of the flux density components, their frequency and the location of the associated losses in the machine. A simple expression allows estimating the losses generated by the slotting resonance harmonics. The numerical approach provide much quantitative information, thank's to an accurate iron loss model. The both methods display that the harmonic flux density may have a very low magnitude but a high

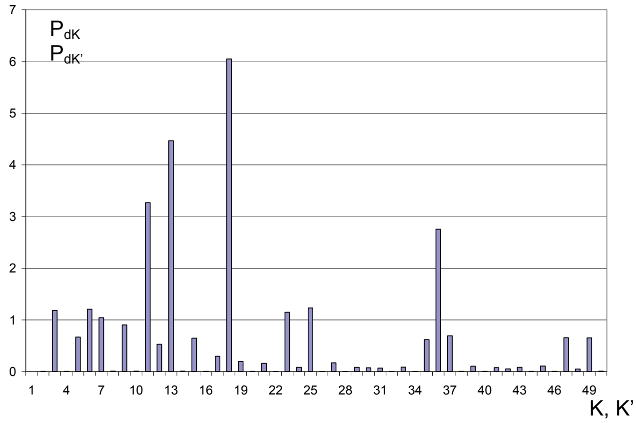


Fig. 7 – Iron loss harmonic distribution

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