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CALCULATION OF CORE LOSS FOR SWITCHED RELUCTANCE MOTORS

ABSTRACT *In the paper the method of core loss prediction in the switched reluctance motors is presented. The classical core loss separation technique is modified for the purpose of considering an arbitrary magnetic flux density waveform. The time-stepping finite element program is elaborated for prediction of instantaneous magnetic flux density distribution in the motor core. The method of core loss calculation is backed well by the measurements carried out on the laboratory test-rig.*

Keywords: *switched reluctance motors, FEM, core losses*

1. INTRODUCTION

A tremendous interest in core loss analysis in electrical machines relates with the need for loss reduction due to applications of switching mode power supplies and rise of operation frequency in almost all branches of industry [1-8].

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The switched reluctance motors (SRMs) are considered as prospective replacement of conventional induction machines in many applications and also as one of the best choices for the very high speed drives. Large number of switched reluctance motors is built as the machines of fractional power. That would suggest that the core loss in small SRMs can be considered negligible. Not all of them are, however built of thin size electrical sheets as significant radial forces considerably limit their usage. The estimation of core loss in these machines is thus important. Despite of that, there is relatively few works regarding an analysis of core loss in the switched reluctance motors [3, 4].

The difficulty related with core loss determination in SRMs involves highly nonsinusoidal waveforms of the magnetic flux density that have different distributions in different parts of the motor core. Additionally time-variations of magnetic flux density in certain subregions of the magnetic circuit, are affected by DC bias.

Main goal of this work is to deliver a method of core loss calculation in the switched reluctance motors. For practical design work, it is always preferable to be able to use manufacturer data instead of performing extensive measurements for particular magnetic material. In this paper the Authors modify classical core loss separation technique for the purpose of considering an arbitrary magnetic flux density waveform. The results of core loss calculations are predicted using the elaborated two dimensional time-stepping finite element program and validated by measurements carried out on the laboratory test-rig for a small two-phase, 100 Watt motor.

2. MODIFIED CORE LOSS SEPARATION TECHNIQUE

In [7] an important hypothesis about the possibility of replacing frequency in classical Steinmetz equation with time-dependent quantity was developed. The Authors limited, however their considerations to cores made of power ferrites. The general idea about generalisation of core loss equation for arbitrary excitations can also be applied to other soft magnetic materials, including laminated cores.

In general the density of core loss in electrical machines with laminated cores, can be separated in following way

$$\Delta\bar{P}_{fe} = \Delta\bar{P}_c + \Delta\bar{P}_h + \Delta\bar{P}_e, \quad (1)$$

where: $\Delta\bar{P}_c, \Delta\bar{P}_h, \Delta\bar{P}_e$ is classical eddy current, hysteresis and excess loss, respectively.

In frequency domain the distribution of core loss density, in Watts per kilogram, can be expressed as

$$\Delta \bar{P}_{fe} = c_c |B|^2 f^2 + c_h |B|^a f + c_e |B|^{1.5} f^{1.5}, \quad (2)$$

where:

f is frequency and B is magnetic flux density.

In order to obtain coefficients c_c , c_h , a and c_e it is necessary to perform a nonlinear fit of equation (2) to catalogue loss data provided by manufacturer. For a particular material used (0,5 mm, EP-20 - type polish electrical sheet) these coefficients are summarized in Table 1.

TABLE 1
Constant coefficients in core loss equation

| | |
|-------|------------|
| c_c | 0,00069963 |
| c_h | 0,00015486 |
| a | 0,93712000 |
| c_e | 0,00038978 |

By replacing frequency in (2) with $|B|^{-1} \left| \frac{dB}{dt} \right|$ the following equation can be obtained

$$\Delta \bar{P}_{fe} = c_c^* \left| \frac{dB}{dt} \right|^2 + c_h^* \left| \frac{dB}{dt} \right| |B|^{a-1} + c_e^* \left| \frac{dB}{dt} \right|^{1.5}, \quad (3)$$

Assuming that magnetic flux density varies in time sinusoidally, and using equation

$$\Delta \bar{P}_{fe} = \frac{1}{T} \int_0^T \Delta \bar{p}_{fe} dt, \quad (4)$$

it can be shown that

$$c_h^* = \frac{c_h}{4}, \quad c_c^* = \frac{c_c}{2\pi^2}, \quad c_e^* = \frac{c_e}{\sqrt{2\pi} \cdot 3,496}.$$

An important property of equation (3) is that it shows the susceptibility to DC bias.

3. MODEL FOR CALCULATION OF INSTANTANEOUS MAGNETIC FIELD DISTRIBUTION

To calculate the magnetic flux density distribution in the motor the time-stepping voltage-driven finite element model is elaborated. The equation governing the magnetic field in the motor can be written in form

$$\nabla \times (\nu \nabla \times \mathbf{A}) = \mathbf{J}_s - \sigma \left(\frac{d\mathbf{A}}{dt} - \nabla V \right), \quad (5)$$

where, \mathbf{A} is magnetic vector potential, ν is magnetic reluctivity, \mathbf{J}_s is vector of source current density, σ is an electric conductivity, V is an electric scalar potential. The formulation also includes equations describing connection of electric circuit

$$e_u = Ri_u + L_e \frac{di_u}{dt} + \frac{d}{dt} \oint_{C_u} \mathbf{A} \cdot d\mathbf{l}, \quad (6)$$

where u stands for phase name ($u = a, b, c$, etc.), R is resistance, L_e is and inductance due to coils end-region and C_u denotes integration path along coils. If the magnetic field distribution is restricted to two dimensions, equation (5) can be simplified to

$$-\nabla \cdot (\nu \nabla A_z) = \sum_u \sum_{j=1}^{Q_w} \frac{n_t \eta_j \eta_u}{S_j} i_{pu} - \sigma \frac{dA_z}{dt}, \quad (7)$$

where A_z is z-th component of magnetic vector potential, S_j is total cross section area of j -th coil in u -th winding, Q_w is a number of coils per phase belt, n_t is a number of turns per coil, $\eta_j = \pm 1$ indicates current sense in coil, η_u is equal 1 if point belongs to coil in u -th phase belt and is equal zero, otherwise.

The weighted residual approach and the Galerkin method [9] applied to equation (7) result in the system of equations of form

$$\nu \int_{\Omega} (\nabla N_i \cdot \nabla N_j) \varphi_i dS = \sum_u \sum_{j=1}^{Q_w} \frac{n_t \eta_j \eta_u}{S_j} i_u \int_{\Omega} N_i dS - \sigma \int_{\Omega} (N_i N_j) \frac{d\varphi_i}{dt} dS, \quad (8)$$

where: φ_i is an i -th nodal value of A_z , N is an element shape function, Ω is region of analysis. Equation (8) can be written as

$$e_u = Ri_u + L_e \frac{di_u}{dt} + \ell \sum_{j=1}^{Q_w} \frac{n_t \eta_j}{S_j} \int_{\Omega} N_i \frac{d\varphi_i}{dt} dS, \quad (9)$$

where: ℓ is a machine length, and action of η_j is same with that in equation (7). The equations (8) and (9) are further discretised in time domain using backward Euler formula [9] to give time-stepping system of algebraic equations. The model also enables for rotor movement by implementing classical moving band technique where the air-gap elements are cyclically distorted and the interface nodes are renumbered to provide continuity of solution as the rotor rotates.

4. NUMERICAL CALCULATIONS AND PHYSICAL VALIDATION OF CORE LOSS MODEL

The calculations have been done for a small two-phase motor with technical specifications summarized in Table 2.

TABLE 2
Technical specifications of laboratory motor

| | |
|-------------------------------|------------------|
| Number of stator poles | 4 |
| Number of rotor poles | 2 |
| Outer/inner diameter/length | 61,2/31,75/60 mm |
| Radial air-gap length min-max | 0,25-0,35 mm |
| Nominal voltage | 24 V |
| Nominal power | 100 W |
| Nominal speed | 1400 RPM |

Figure 2 depicts the structure of the motor as well as the photograph of the laboratory test-rig.

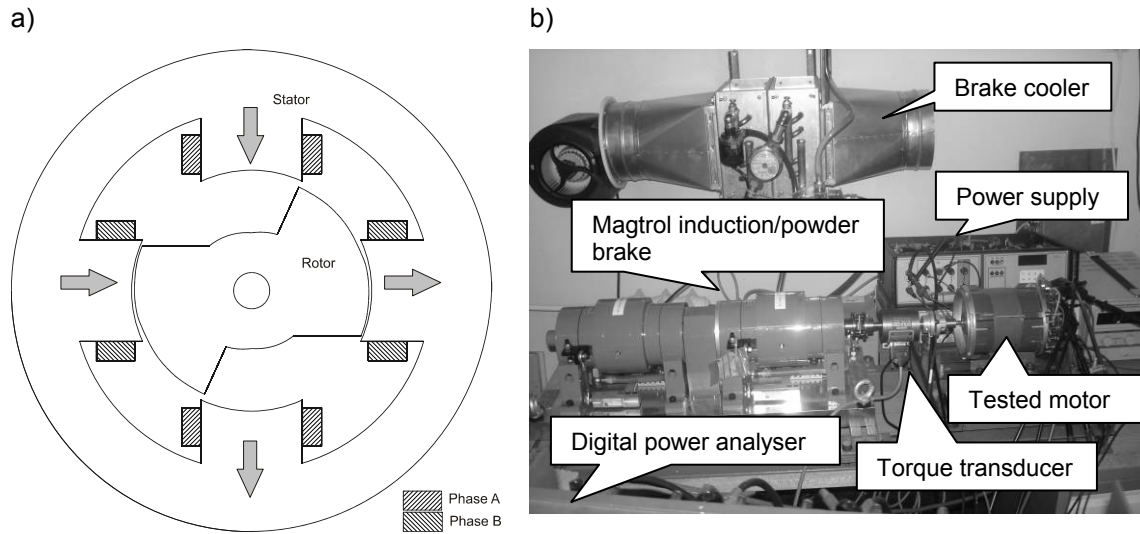


Fig. 2. Motor structure - a) (arrows indicate sense of linkage fluxes produced by coils), laboratory test-rig for core loss measurement - b)

The motor core loss was measured by taking difference between input electrical power measured at winding terminals P_i , resistive winding losses P_Ω , mechanical power P_{mo} and internal mechanical power loss P_{mi}

$$\Delta P_{fe} = P_i - P_\Omega - P_{mo} - P_{mi} . \quad (10)$$

The input power P_i was recorded by digital power analyser. The weak point of the measurement method is lack of direct measurement of instantaneous temperature of the motor winding. In calculations the temperature rise of the winding could only be estimated assuming adiabatic heating during given time interval. An inaccuracy is therefore expected in core loss measurements due to uncertainty in determination of real winding resistance. The mechanical power P_{mo} was measured directly using torque transducer. The mechanical power loss P_{mi} was estimated by performing run-down test from 6000 RPM to 0 RPM from which the braking torque vs. rotational speed function was determined.

The finite element model described in previous section was used for calculations of core loss for the motor considered. The machine was driven from 24 V fixed voltage battery source. It was operated by forcing desired rotational speed at time $t = 0$. When the steady-state was reached, the quantities of interest were calculated.

Figures 3a-b compare the finite element predictions of basic characteristics of the motor with the measurements. It is clear from the comparison that the validity of the elaborated finite element model is high. In the next step the power loss in the motor core is predicted. The results of the calculations compared with measurements are shown plotted in Fig. 4.

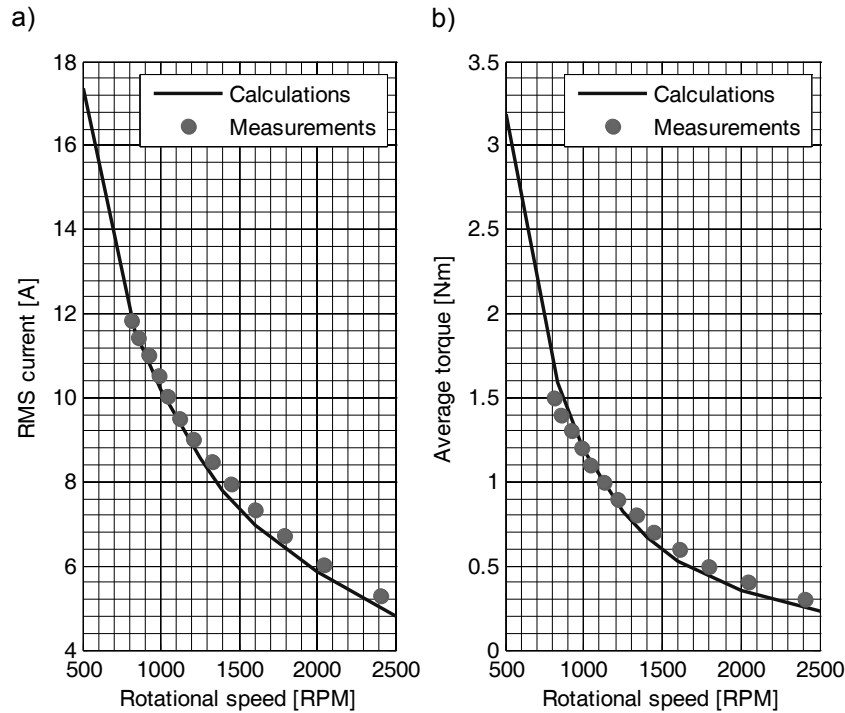


Fig. 3. Finite element predictions of RMS phase current a) and torque b) vs. rotor speed compared with measurements

It can be observed that the agreement between calculated and measured values is relatively good. The inaccuracy is mainly due to estimated resistance of coils. In spite of that complication, based on the results presented at this stage of the study, it can be said that the computational algorithm presented here can be used for loss prediction in the switched reluctance motors that are enough accurate for design purposes.

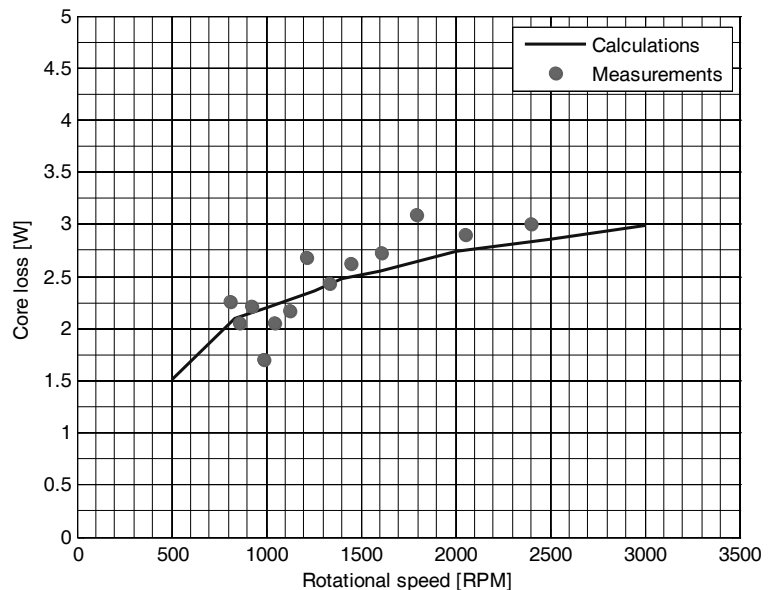


Fig. 4. Variation of core loss vs. rotor speed

An unexpected result is that the total core loss for the motor considered, operating at nominal point of operation, is nearly 3% of nominal power what is relatively high value for that machine type. This suggests that better quality electrical sheet should be considered for improved motor design. Figure 5 shows variations of instantaneous core loss at 1400 RPM. It is evident that main part of core loss is concentrated in the stator, but the rotor loss is also significant.

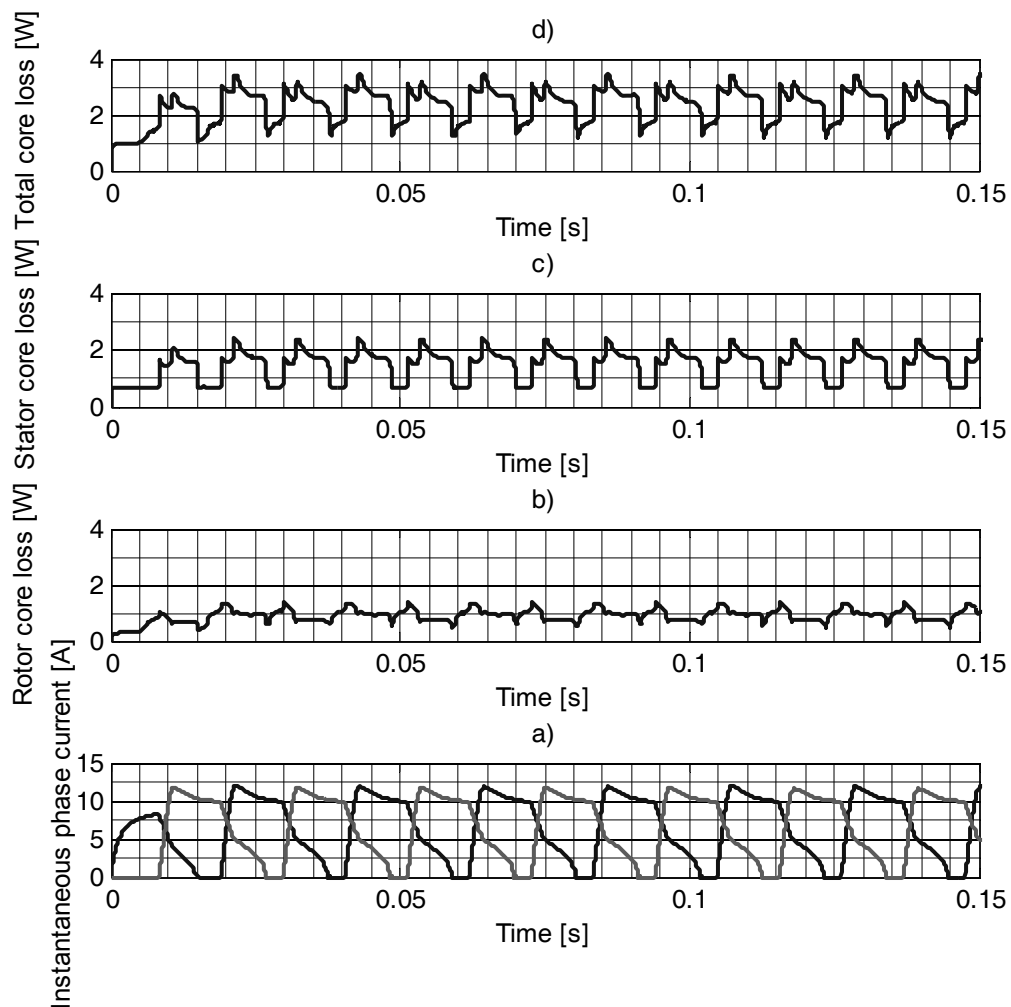


Fig. 5. Computed instantaneous waveforms at rotational speed of 1400 RPM:
a) phase currents, b) rotor core loss, c) stator core loss, d) total core loss.

5. CONCLUSIONS

In the work the model for core loss predictions in the switched reluctance motors was presented. The voltage driven finite element model was elaborated for calculation of instantaneous waveforms of magnetic flux density in the motor core. These waveforms were used as an input data for modified core loss equation. Experimental validation of the proposed model shows its applicability for accurate determination of core loss for the machines fed with single pulse current waveform. An advantage of the presented method is that it practically does not require additional measurements of unit power loss for particular electrical sheet over those provided by manufacturer.

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OBLICZANIE STRAT W ŻELAZIE W PRZEŁĄCZALNYM SILNIKU RELUKTANCYJNYM

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STRESZCZENIE *W pracy przedstawiono model do obliczeń strat w żelazie w przełączalnym silniku reluktancyjnym. Podstawowa trudność związana z wyznaczeniem strat w żelazie w tego rodzaju maszynach jest związana z silnie odkształconymi przebiegami lokalnych wartości indukcji magnetycznej w rdzeniu. Ponadto, w większości punktów obwodu magnetycznego przebiegi te zawierają niezerową składową stałą. W takim przypadku nie można użyć bezpośrednio podejścia, stosowanego w maszynach prądu przemiennego, polegającego na podziale strat i wykorzystaniu typowych krzywych stratności jednostkowej materiału magnetycznego podawanych przez producentów.*

W celu rozwiązania tego problemu w niniejszej pracy zastosowano metodę polegającą na przekształceniu równania opisującego stratność jednostkową materiału ferromagnetycznego, zgodnie z hipotezą przedstawioną w pracy [7]. W wyniku analitycznego przekształcenia równania (2), otrzymano równanie (3), które jest określone w dziedzinie czasu, a nie w dziedzinie częstotliwości. Równanie to zawiera tylko stałe, co równanie oryginalne. Stałe te można względnie łatwo wyznaczyć metodą optymalizacji nieliniowej. Wyniki obliczeń tych stałych zawiera tabela 1.

W celu wyznaczenia przebiegów lokalnej wartości indukcji w obwodzie magnetycznym, opracowano model polowo-obwodowy silnika o dwóch pasmach fazowych oraz mocy znamionowej 100 W, w oparciu o metodę elementów skończonych. Wykorzystując obliczone przebiegi indukcji magnetycznej w całym obwodzie magnetycznym oraz równanie (3) opisujące stratność jednostkową, wyznaczono całkowite straty w rdzeniu w funkcji prędkości obrotowej. Przeprowadzono weryfikację eksperymentalną opracowanego modelu w oparciu o skonstruowane stanowisko pomiarowe (rys. 2b) uzyskując dużą zgodność wyników obliczeń z danymi doświadczalnymi (rys.4). Pozytywne wyniki weryfikacji pozwalają uznać przedstawione podejście za wystarczająco dokładną oraz ogólną metodę obliczania strat w żelazie w silnikach reluktancyjnych zasilanych jednopulsowo.