Analytical Core loss Model for the Switched Reluctance Motor with Experimental Verification

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***Abstract*- An analytical model for core loss estimation in a switched reluctance motor (SRM) is presented in this paper. The flux waveforms in different parts of the magnetic circuit of the motor are obtained using a mathematical model. It is shown that excitation and switching sequences can be taken into account in the flux waveforms prediction. A mathematical equation is introduced based on Steinmetz equation which is able to calculate the core losses for non-sinusoidal flux distribution with reasonable accuracy. Measured core losses of a typical SRM are compared with the model results and a good agreement is achieved. This model can be easily applied to other SRMs.**

***Index Terms*- SRM, Core loss, flux model**

I. INTRODUCTION

It is believed that the first model for core losses estimation was introduced for a 12/10 SRM based on measured data [1]. Other models obtained flux waveforms in different parts of the magnetic circuit graphically and losses were estimated using a Fourier series which is questionable in the presence of saturation [2]. In [3], the measured or finite element (FE) computed flux in one section of the magnetic circuit has been used to determine fluxes in different sections of the SRM using a mathematical flux model. This model ignores the minor inner loops within the main hysteresis loop. Equivalent magnetic circuits can be also used for core loss estimation [4]. However, there is still no accurate and quick model, which can be easily used for development of the core loss model of the SRM. To this end, an analytical model of the static flux linkage characteristics versus phase current and rotor angular position is determined [5, 6]. The static flux-linkage characteristics and phase voltage equation are then used to determine the phase flux-linkage.When the flux distribution in different sections of the magnetic circuit of the machine are determined using the mathematical model [7], core losses are calculated in different sections, in which the successively excited phases and the phase switching sequences are taken into account. In the core loss estimation, the role of inner loops of the magnetization characteristics is considered due to the non-sinusoidal variations of the flux. In the following sections of the paper, different parts of the model are introduced and simulation results on an 8/6 SRM, rated at 1500 rpm, 4kW, are compared with the measured values.

II. PHASE FLUX-LINKAGE ESTIMATION

Static flux-linkage characteristics can be obtained using finite element techniques, magnetic equivalent circuits, analytical models or measurements. To obtain the measured flux-linkage, the motor must be available; otherwise the FE technique is the most accurate method for flux-linkage estimation, at the expense of considerable computation time. This paper uses an analytical model [5, 6], which is able to calculate the static characteristics quickly with a reasonable accuracy compared to the measured values. The calculated static flux-linkage characteristics using the analytical model as well as the measured values are shown in Fig. 1. The calculated and measured unaligned and aligned inductances are compared in Table I.

TABLE I
COMPARISON OF CALCULATED AND MEASURED INDUCTANCES IN THE PROPOSED SRM

Teeth position	Measured inductance (mH)	Calculated inductance (mH)
Unaligned	15.3	14.8
Aligned at rated current 7 A	25	27

Having the static flux-linkage characteristics, the phase flux-linkage is obtained by using the following voltage equation:

$$V = Ri + \frac{d\lambda(\theta,i)}{dt} \qquad (1)$$

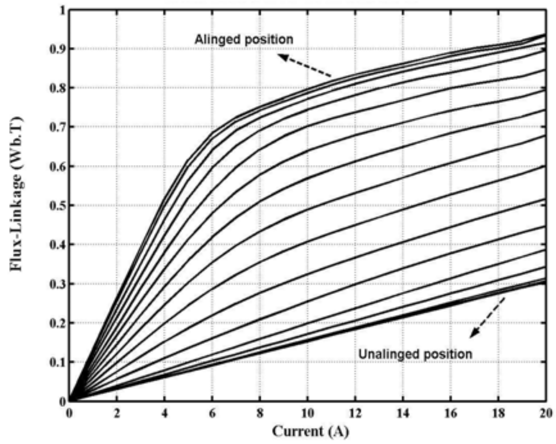
The rate of flux-linkage variation is:

$$\frac{d\lambda}{dt} = \frac{\partial\lambda}{\partial i}\frac{di}{dt} + \frac{\partial\lambda}{\partial\theta}\frac{d\theta}{dt} = L_{inc}\frac{di}{dt} + C_w\omega \qquad (2)$$

(1) and (2) give:

$$\frac{di}{d\theta} = \frac{1}{\omega L_{inc}} \left(V - Ri - C_w \omega \right) \quad (3)$$

The phase current waveform i(θ) is obtained by solving differential equation (3). The phase flux-linkage λ(θ) is then obtained using the static λ(θ,i) and i(θ). Fig 2



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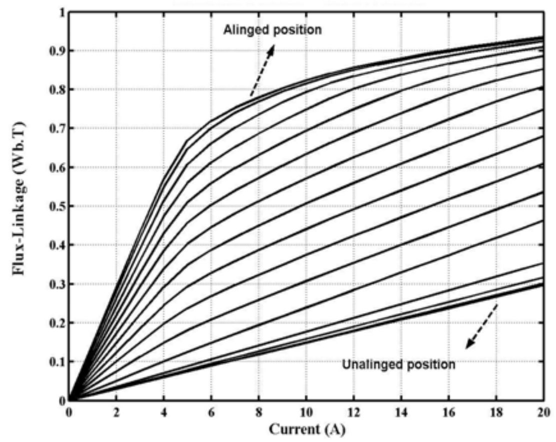


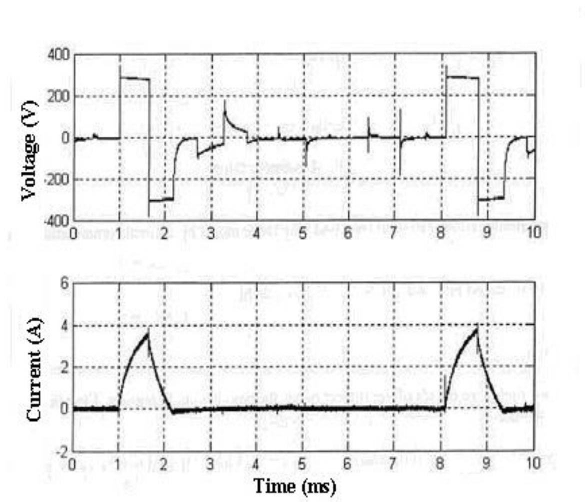
(2)

Fig. 1. Static flux-linkage characteristics: (1) measured, (2) calculated

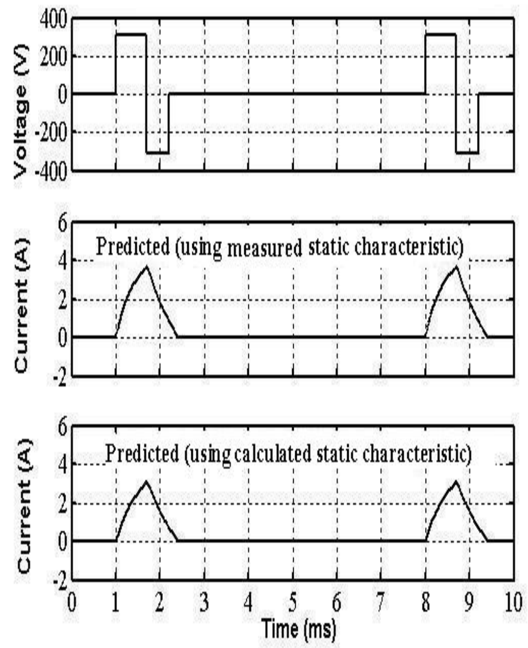
shows the predicted and measured phase voltage and current waveforms of the motor under study at the rated speed. The calculated and measured phase flux-Linkage is presented in Fig. 3. The current waveform predicted using the measured flux-linkage is in Fig. 2 which is very close to the predicted flux-linkage characteristics.

III. Determination Of Flux In Different Sections Of The Magnetic Circuit

It is necessary to develop an accurate mathematical model for the flux waveforms in the SRM. In [2], the flux waveforms are modeled based on a graphical technique which is not convenient for design purpose. [3] and [8] have introduced mathematical models of the flux-linkage where switching and excitation sequences have not been taken into account in an SRM with low number of phases such as a 4-phase, 8/6



(1)



(2)

Fig. 2. Voltage and current waveforms: (1) measured, (2) calculated

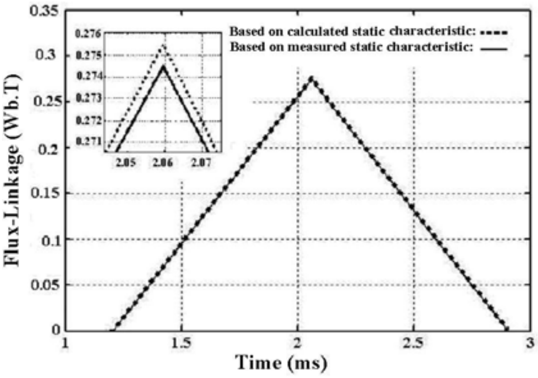


Fig. 3. Phase flux-linkage waveform at rated speed

machine, phases are always excited in sequence. For a motor with a higher number of phases, the excitation and switching sequences must be taken into account otherwise large errors in the predicted flux characteristic can result. In the proposed 8/6 SRM, the phases are

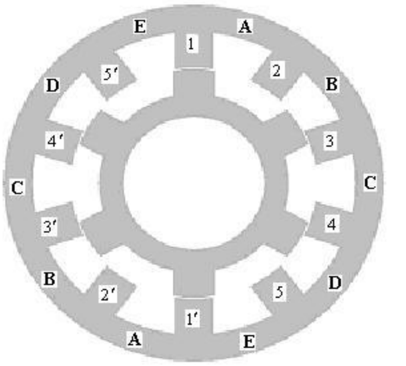


Fig. 4. A cross-section of 10/6 SRM

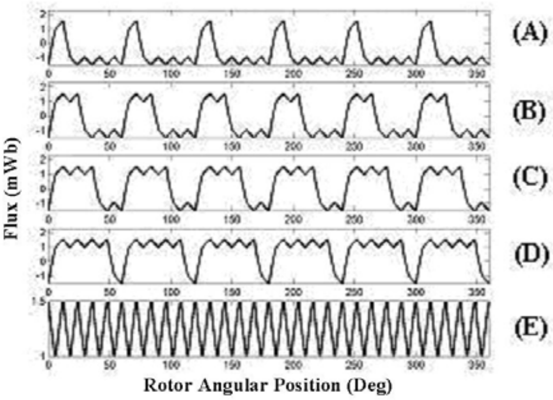


Fig. 5. Flux in different sections of stator yoke when phases excited successively

excited successively and complicated switching and excitation sequences are not proposed. In order to study the importance of the phase excitation sequence in the flux estimation, a 10/6 SRM shown in Fig. 4 is proposed. For successive excitation of the phases (1-2-3-4-5) the flux waveforms are as shown in Fig. 5. But in order to have the maximum torque, the excitation sequence of (1-4-2-5-3) is required. In this case, fluxes in different sections of the stator yoke are as shown in Fig. 6. As seen, the flux linkage profile in sections B and C are different in Figs. 5 and 6. This leads to a considerable error in the core losses estimation.

Therefore, in the core loss model a mathematical flux model [7] is used where proper excitation and switching sequences are determined based on the design parameters. The flux waveforms in all magnetic circuit sections of the machine are then estimated using the phase flux-linkage waveform. The simulation results for the proposed SRM are shown in Figs. 7-10. In the proposed SRM, phases are excited successively and the flux profiles in the stator poles are obtained.

As shown in Fig. 8, the stator yoke has been divided into four sections where the core losses are equal in section A and C. In an SRM with q phases and N_R teeth

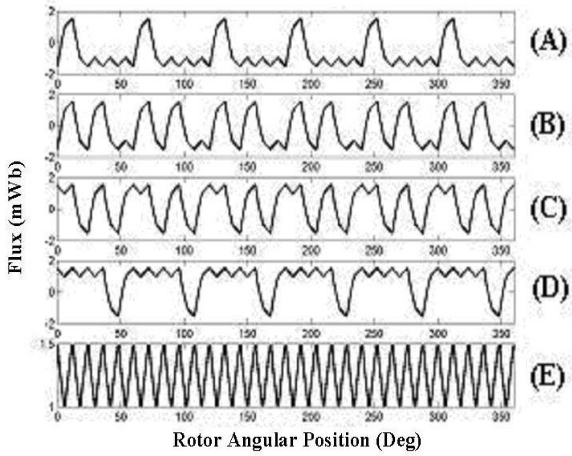


Fig. 6. Flux in different sections of stator yoke when proper phases are excited

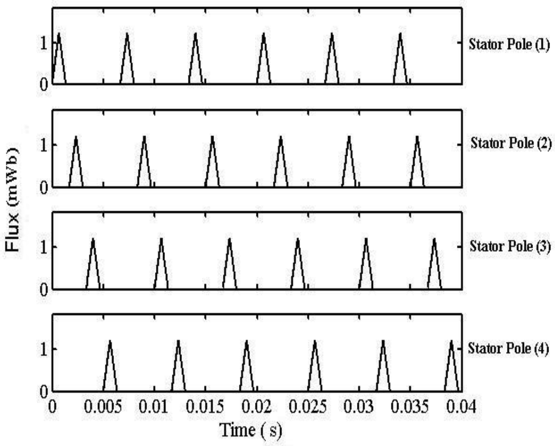


Fig. 7. Flux waveform in stator pole

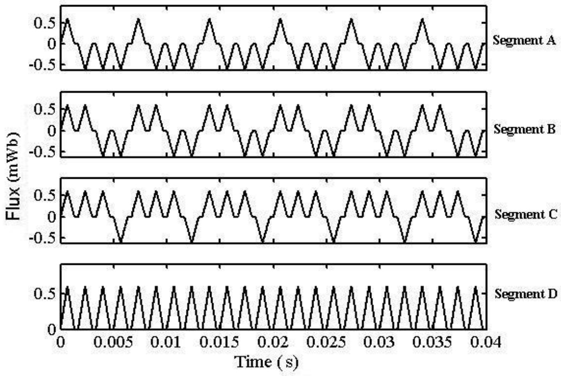


Fig. 8. Flux waveform in stator yoke

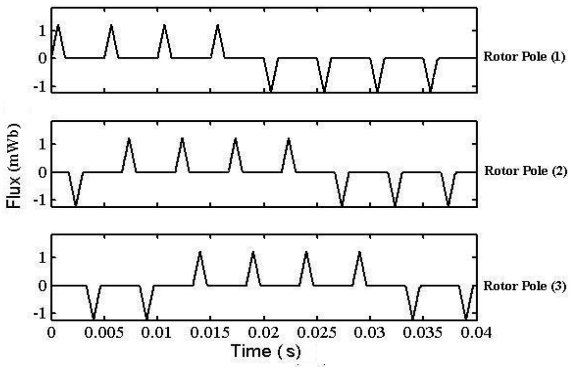


Fig. 9. Flux waveform in rotor pole

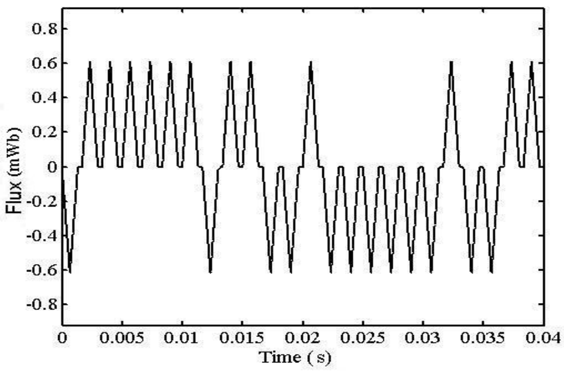


Fig. 10. Flux waveform in rotor core

on the rotor, the switching period of the phases are as follows:

$$TR1 = \frac{2\pi}{\omega_R q N_R} \tag{4}$$

where ω_R is the rotor angular speed in rad/s. Flux distribution in the rotor poles are similar and only they display by TR2:

$$TR2 = \frac{2\pi}{\omega_R N_R} \tag{5}$$

Fig. 9 shows the flux distribution of the rotor poles. The core of 8/6 SRM is divided into six sections where all sections carry the same flux, as shown in Fig. 10.

IV. Procedure For Core Losses Estimation

Generally when a material is exposed to a sinusoidal flux density waveform with variable amplitude and frequency, core losses are computed from the following traditional Steimetz equation:

$$P_c = P_h + P_e = k_h \omega_s B_{\max}{}^n + k_e \omega_s{}^2 B_{\max}{}^2 \tag{6}$$

This equation cannot be directly applied to an SRM where a non-sinusoidal flux density waveform exists. Therefore, this equation must be modified for each flux density waveform. Since eddy currents are induced due to the magnetic flux variations, it seems that it is more convenient to replace the magnetic flux variations in the place of frequency in Eqn. 6. For a sinusoidal flux density:

$$B = \hat{B}\sin(\omega_s t) \;\Rightarrow\; \frac{dB}{dt} = \omega_s \hat{B}\cos(\omega_s t)$$

$$\Rightarrow \left(\frac{dB}{dt}\right)^2_{avg.} = \frac{\left(\omega_s \hat{B}\right)^2}{2} \tag{7}$$

By comparing (6) and (7), the magnitude of the eddy current losses can be modeled in the term of the rate of flux density as follows:

$$P_e = 2k_e \left(\frac{dB}{dt}\right)^2_{avg.} \tag{8}$$

Two coefficients, k_{h1} and k_{h2} are used to modify (6) and calculate the hysteresis loss.

$$P_h = k_{h1} k_{h2} k_h\, \omega_s B_{\max}{}^n \tag{9}$$

When the flux density changes between zero and peak maximum value of $B_{\max}, k_{h1} = 0.5$, but this coefficient will be 1 when the flux density changes between $-B_{\max}$ and $B_{\max}$. In order to take into account the minor loops the following coefficient is used [9]:

$$k_{h2} = 1 + \frac{0.65}{\mathrm{B}_{\max}} \sum \Delta B_i \tag{10}$$

where ΔB_i is the amplitude of the returned i[th] flux. Thus, (6) for a non-sinusoidal waveform is as follows:

$$P_c = P_h + P_e = k_{h1}k_{h2}k_h\omega_s B_{\max}{}^n + 2k_e\left(\frac{dB}{dt}\right)^2_{avg.} \quad (11)$$

Constants n, k_h and k_e of the Steimetz equation are obtained from the core loss curves of the lamination (P/B/f) and the least square sum of errors for at least two ranges of low and high frequencies. For better realization the modification factors k_{h1} and k_{h2}, are introduced to calculate the magnitude of the hysteresis losses in various sections of the magnetic circuit of an 8/6 SRM. Referring to Fig. 7, the flux in the stator pole varies between 0 and B_{max} with frequency $N_R\omega_R$. Since there is no inner loop, k_{h2}=1 and magnitude of the hysteresis losses are as follows:

$$P_h = 0.5k_h N_R\omega_s B_{\max}{}^n \quad (12)$$

Referring to Fig. 8, the magnetic flux varies between $-B_{max}$ and $+B_{max}$ with frequency $N_R\omega_R$ in three sections of the stator. For every inner loop, it may be assumed that the density of the hysteresis losses is as follows:

$$P_h = k_h k_{h2} N_R\omega_s B_{\max}{}^n \quad (13)$$

where

$$k_{h2} = 1 + \frac{0.65}{\mathrm{B}_{\max}}2\Delta B \quad (14)$$

but the magnetic flux in one section of the stator varies between 0 and B_{max} with frequency $4N_R\omega_R$. In this case there is no inner loop and:

$$P_h = 0.5k_h 4N_R\omega_R B_{\max}{}^n \quad (15)$$

Referring to Fig. 9, the magnitude of the hysteresis losses must be calculated using six inner loops with maximum amplitude B_{max} and frequency ω_R:

$$P_h = k_h k_{h2}\omega_R B_{\max}{}^n \quad (16)$$

where

$$k_{h2} = 1 + \frac{0.65}{\mathrm{B}_{\max}}6B_{\max} = 4.9 \quad (17)$$

Finally in the rotor core (Fig. 10), five complete loops and 14 inner loops can be assumed and the magnitude of the hysteresis losses is as follows:

$$P_h = 5k_h k_{h2}\omega_R B_{\max}{}^n \quad (18)$$

where

$$k_{h2} = 1 + \frac{0.65}{\mathrm{B}_{\max}}3\Delta \quad (19)$$

The predicted core losses in different sections of the magnetic circuit of the proposed motor at the rated speed are summarized in Table II. Calculated core losses using the measured static flux linkage are also shown in this figure which indicates the merit of the proposed analytical model for the estimation of the core losses.

TABLE II

CORE LOSSES IN DIFFERENT SECTIONS OF THE MAGNETIC CIRCUIT OF THE PROPOSED SRM

Section	Hysteresis losses (W)	Eddy current losses (W)
Stator pole	1.76	1.43
Stator yoke	26.95	18.48
Rotor pole	1.65	0.91
Rotor core	6.61	3.36
Stator	29.73	19.7
Rotor	8.26	4.28
Rotor+stator	36.39	24.15
Core losses	62.16	

Fig. 11. Measured mechanical losses of 8/6 SRM

V. CORE LOSSES MEASUREMENTS

In order to measure the mechanical losses, the no-load 8/6 SRM is run by a dc motor at different speeds, in each case the output power of the dc machine is the mechanical losses (P_{mec}) of the SRM obtained as follows:

$$P_{mec} = T\omega \quad (20)$$

where T is the average torque, ω is the angular speed of the rotor. Fig. 11 presents the mechanical losses of the SRM at different speeds.

The measurement setup is shown in Fig. 12. Input power and stator winding copper losses are calculated as follows:

$$P_{in} = \frac{1}{T}\int_{\tau}^{\tau+T} u.i.dt \quad (21)$$

$$P_{cu} = \frac{1}{T}\int_{\tau}^{\tau+T} R.i^2.dt \quad (22)$$

where T is the phase excitation period, R is the stator winding resistance. Finally, by calculating the input power (P_{in}), stator copper losses (P_{cu}) and mechanical losses, core losses are estimated as follows:

$$P_{iron} = P_{in} - P_{cu} - P_{mec} \quad (23)$$

The estimated and measured losses are shown in Fig. 13. The estimated and measured core losses of the proposed SRM at the rated speed are 62.16 W and 67.6 W respectively, which shows an 8% error. The reasons for this error include:

1. Changes in the core losses characteristics due to stress and forming process.

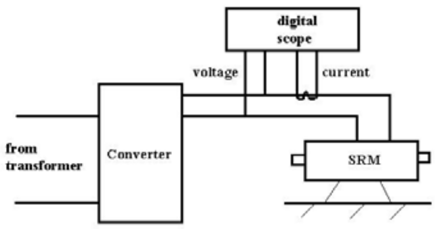


Fig. 12. Core loses test setup for SRM

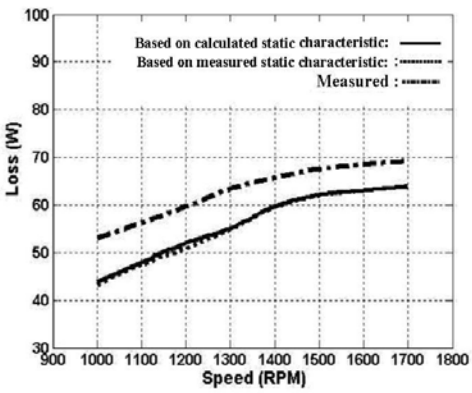


Fig.
Fig. 13. Calculated and measured core losses of 8/6 SRM

2. Increase in the eddy current losses due to lamination
contact at the edges caused by manufacturing.
3. Errors in the estimated flux at the corners of the stator
and rotor poles.

VI. Conclusions

In the developed core loss model attempts were made to include the following:

a) Saturation in the magnetic circuit.
b) End winding effects upon the phase inductance.
c) The sequence of phase excitation for flux estimation in different parts of the magnetic circuit of the SRM.
d) Core losses characteristics of the laminations.
e) The effect of the minor inner loops in the hysteresis losses estimation.

The developed core losses model is quick and accurate and can be used during the optimization design process for different types of SRMs.

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