Fast iron loss prediction method in the pre-design stage of SRMs

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Abstract: To effectively find a valid solution in the development process of electrical machines, it is essential to predict machine iron loss in the pre-design stage. In guidance on the choice of the most suitable configuration of the switched reluctance machine (SRM) is given, but iron loss prediction is neglected. Based on the study of Burkhart et al., this study proposes a simplified model to efficiently calculate iron loss in the design procedure of SRMs. For this purpose, a maximum co-energy loop control method is implemented to facilitate the estimation of flux waveform and machine torque. Three different iron loss calculation methods are compared in terms of accuracy and time. The accuracy of the simplified model is verified by comparison with finite element simulation results. Furthermore, the impact of different parameters on the iron loss is discussed, and the iron loss of various machine geometries is comparatively analysed.

1 Introduction

Due to the simple working principle and robust mechanical construction, switched reluctance machines (SRMs) have received increasing attention by drive designers for various kinds of industrial applications, such as electrical grids, aircraft starter/generator, and electric vehicles [1–3]. Most publications about the subject of SRMs design or the optimal design are based on iterative design approaches, which contain several empirical parameters [4, 5]. A trial-and-error procedure is inevitable to determine these parameters, which is time-consuming and cannot provide designer scientific physical understanding.

The concept of the generalised SRM model was first introduced into the procedure of the SRM design in [6]. This model could uncover physical relationships between machine torque and design parameters without the aid of empirical values. Based on [6], a comprehensive design procedure was proposed in [7], which contributed to reduce the initial design period. Burkhart et al. [8] proposed a solution space-based pre-design approach, which can help a designer to select the most suitable configuration in an intuitive way for a specific torque requirement. In this approach, the torque production ability is pre-calculated by finite element analysis (FEA), stored in a database, and applied for the power density optimisation. The scheme of solution space-based design approach is shown in Fig. 1. The core of this method is the construction of the solution database, which is built with a given design range of CTSR, $R_{1}$, and machine configurations. Here, $R_{1}$ is the outer radius of the stator and CTSR is the coil-to-slot ratio.

However, Burkhart et al. [8] do not include an iron loss prediction method, which is important for more accurate stack length calculation, machine efficiency optimisation, and thermal behaviour evaluation.

In a pre-design stage, a large number of machines need to be compared. In addition to the accuracy of the prediction results, the algorithm speed is also very vital to the machine designer. This paper presents a fast analytical iron loss prediction method for the pre-design of SRMs. The proposed method allows a quick and accurate view on the losses, to compare the dynamic performance of machines with different configurations of stator and rotor poles.

2 Iron loss model

The iron loss of SRMs is not only related to the geometry dimensions, but also determined by the control parameters. Generally speaking, it is very hard to find optimal control parameters during the pre-design stage since in this stage, as described in [8], magnetic characteristics only at aligned and unaligned positions are simulated by FEA to save space and time. Due to the limited information, it is very difficult to compare the dynamic performance at the optimal control point. In this case, comparing the iron loss with a designated control method seems to be a good trade-off. The maximum co-energy loop control method (MCLC), which was proposed by Fuengwarodsakul et al. [6] to estimate machine average torque, is applied for the flux model construction in this paper.

2.1 Maximum co-energy loop control method

The MCLC method is applied for utilising the entire co-energy loop at a certain peak value $\Theta_{\text{peak}}$ of the magnetomotive force (MMF) $\Theta$. Fig. 2 presents the magnetic characteristics of one stator pole at aligned and unaligned positions with the actual and assumed co-energy loop for a three-phase 12/8 SRM. In order to achieve the entire co-energy loop at $\Theta_{\text{peak}}$, the MMF $\Theta$ should be a square waveform, which is obviously impossible for SRM due to the existence of phase inductance. Although the idealised square waveform cannot be realised for SRM, the entire co-energy loop can be approximately achieved if the following requirements are satisfied:

(i) $\Theta$ rises to $\Theta_{\text{peak}}$ at the position $\Theta_{1}$ where a rotor pole and a stator pole begin to overlap.

(ii) $\Theta$ begins to decrease at the position $\Theta_{2}$ where the approaching edge of a rotor pole and the edge corner of a stator pole are aligned.

![Fig. 1 Solution space approach proposed by Burkhart et al. [8]](image-url)
Phase resistance and fringing effect are neglected.

MCLC method is implemented.

The phase voltage in Stage II is equivalent to the flux is assumed to rise linearly from position

The entire co-energy loop is assumed to be achieved when the energy loop. So, the control method which can produce the above-
to the changes of flux, the waveforms of one electric period are calculated. Fig. 3 shows the dynamic waveforms

Based on the above simplifications and assumptions, the flux waveform can be calculated. Fig. 3 shows the dynamic waveforms of flux $\phi$ and $\Theta$ in one stator pole and the curves of the idealised inductance and phase voltage with the MCLC method. According to the changes of flux, the waveforms of one electric period are divided to three stages, which are illustrated in Fig. 3. In this paper, all the windings for different poles are connected in parallel.

In Stage I, the flux $\phi$ rises linearly to the angle $\theta_1$ when $\Theta$ rises to the reference value $\Theta_{peak}$. The flux $\phi(\theta)$ in Stage I is determined by

$$\phi(\theta) = \frac{ku}{w}(\theta - \theta_{on})$$  \hspace{1cm} (1)

where $\omega$ is the machine speed and $u_w$ is the voltage in one winding. $u_w$ can be calculated from the phase voltage $u$:

$$u_w = \frac{u}{N_w N_{serials}}$$  \hspace{1cm} (2)

where $N_w$ is the number of turns per pole and $N_{serials}$ is the number of serial paths. In this paper, all the poles are assumed to be connected in parallel.

The turn-on angle $\theta_{on}$ can be determined by substituting $(\theta_1, \phi_a)$ into (1):

$$\theta_{on} = \theta_1 - \frac{\omega u_w}{ku} \phi_a$$  \hspace{1cm} (3)

In Stage II, $\phi$ rises linearly from $\phi_a$ to $\phi_2$. The flux $\phi(\theta)$ in Stage II is calculated by

$$\phi(\theta) = \frac{ku}{w}(\theta - \theta_{on}) + \phi_a$$  \hspace{1cm} (4)

where $k$ is the duty cycle and its value is derived by

$$k = \frac{\omega \phi_a - \phi_2}{u_w \theta_2 - \theta_1}$$  \hspace{1cm} (5)

The duty cycle should be no bigger than 1, which is given by

$$k = \frac{\omega \phi_a - \phi_2}{u_w \theta_2 - \theta_1} \leq 1$$  \hspace{1cm} (6)

From (6), the limitation of $N_w$ can be deduced by

$$N \leq \frac{u}{N_{serials} \omega \theta_2 - \theta_1}$$  \hspace{1cm} (7)

In Stage III, the MMF decreases from $\phi_{peak}$ to 0 at the angle $\theta_3$. The flux $\phi(\theta)$ in Stage III is written as

$$\phi(\theta) = \frac{-u_w}{\omega}(\theta - \theta_{on})$$  \hspace{1cm} (8)

where $\theta_3$ is the angle where $\Theta$ extinguishes and it is given by

$$\theta_3 = \theta_1 + \frac{\omega}{u_w} \phi_a$$  \hspace{1cm} (9)

To avoid continuous conduction, the difference between $\theta_3$ and $\theta_{on}$ should be no more than one electric period:

$$\theta_3 - \theta_{on} = \theta_1 - \theta_3 + \frac{\omega}{u_w}(\phi_1 + \phi_a) \leq 360^\circ$$  \hspace{1cm} (10)

From (8), another constraint on $N_w$ can be derived, which is given by

$$N_w \leq \frac{u}{N_{serials} \omega \phi_a}$$  \hspace{1cm} (11)

Combining (1), (4), and (8), the flux $\phi(\theta)$ at any angle in an electrical period is given by

$$\phi(\theta) = \begin{cases} \frac{u_w}{\omega}(\theta - \theta_{on}) (\theta_{on} < \theta \leq \theta_1) \\ \frac{ku}{w}(\theta - \theta_{on}) + \phi_a (\theta_1 < \theta \leq \theta_2) \\ -\frac{u_w}{\omega}(\theta - \theta_3) (\theta_2 < \theta \leq \theta_3) \\ 0 (0^\circ < \theta \leq \theta_{on}, \theta_3 < \theta \leq 360^\circ) \end{cases}$$  \hspace{1cm} (12)

2.2 Flux density model

A flux density model is proposed in this section based on waveform decomposition and synthesis. To analytically determine the flux density, some simplifications have to be made firstly:
(i) The machine is divided into sections, and flux density in each section is assumed to be homogeneous.
(ii) Leakage and mutual coupling are neglected.

The stator poles and yokes are divided into $N_s$ sections, and $N_r$ sections for the rotor poles and yokes, as can be seen in Fig. 4. The flux density fundamental periods for different sections are not equal, but one mechanical period ($N_r \times 360^\circ$) is the common period for all sections. In one mechanical period, each stator pole sees $N_s$ flux pulses flowing through the nearest rotor pole. If the initial position is assumed to be at unaligned position, the flux flows through the first stator pole and the first rotor pole is given by

$$
\phi_{sp}(0) = \sum_{j=1}^{N_r} K \phi_{sp,j}(\theta) = \sum_{j=1}^{N_r} K \phi_{sp,j}(\theta + (j - 1)\frac{N_s}{N_r} - 1)$$

The flux flowing through the first stator pole and the rotor pole $j$ can be deduced by the rotor angular shift, which is shown as

$$
\phi_s,j(\theta) = \phi_{sp,j}(\theta + (j - 1)\frac{N_s}{N_r} - 1)$$

The flux flowing through the stator pole $i$ and the rotor pole $j$ is further represented by the stator angular shift:

$$
\phi_{sp,i}(\theta) = K \phi_{sp,i}(\theta + (j - 1)\frac{N_s}{N_r} - 1)$$

where $K$ is the sign vector, which represents the conductor arrangement direction and its element is either 1 or -1.

The flux waveform in the stator pole $i$ can be obtained by the superposition of all the flux waveforms flowing through the stator pole $i$:

$$
\phi_{sp,i}(\theta) = K \phi_{sp,i}(\theta + (j - 1)\frac{N_s}{N_r} - 1)$$

Similarly, the flux waveform in the rotor pole $j$ is

$$
\phi_{sp,j}(\theta) = \sum_{i=1}^{N_s} K \phi_{sp,i,j}(\theta) = \sum_{i=1}^{N_s} K \phi_{sp,i,j}(\theta + (j - 1)\frac{N_r}{N_s} - 1)$$

For modelling the flux in the stator and rotor yokes, the method in [9] is adopted in this paper. Once the flux $\phi$ in all sections is found, the flux density in each section can be obtained by dividing the flux by its cross-sectional area through which it flows:

$$
B = \frac{\phi}{A}
$$

### 2.3 Iron loss calculation

Klein-Hessling et al. [10] discussed three kinds of methods for the iron loss calculation of SRM: Steinmetz method, Bertotti method, and IEM5 method. The parameters of these formulas are determined by curve-fitting the formulas to the measured values provided by manufacturers. However, these fits may be inaccurate due to the lack of iron loss data for flux density $>1.5$ T, which is also referred to in [10]. In this paper, the predictive capability of three methods is tested and compared based on the existing measured data. For this purpose, the measured data is divided into two groups: all data points $>1$ T are removed and the formulas are fitted to the remaining data. The deviation from the existing points between 1 and 1.5 T (test data) can then be evaluated as the extrapolation error. Fig. 5 shows the iron loss density prediction results of three methods. From this figure, we can derive the following conclusions.

(i) The prediction error between the analysis formula calculation and test data increases with flux density and frequency.
(ii) The prediction errors for the three methods only slightly differ.

Though the IEM5 formula offers best fitting results for variable frequencies, the errors caused by the missing data points are much larger than the improvement achieved. Table 1 presents the comparisons of computing time for one calculation for the exemplary 12/8 SRM, as shown in Fig. 4. The iron loss calculation was performed on a desktop PC with an Intel Core i5 3.2 GHz, 16 GB RAM, and a 64-bit operating system. The Bertotti method requires about three times as long as the Steinmetz method, while the IEM5 method takes nearly four times as long. Since thousands of cases have to be evaluated during the pre-design stage, this adds up to an immense increase in calculation time. Considering the computation time and accuracy, the Steinmetz method is adopted in this paper for the calculation of the iron loss.

### 3 Torque and stack length calibration

In the process of electromagnetic design, the torque production capability is calculated by dividing the co-energy by the stator pole arc [7, 8]. However, neglecting iron loss may cause inaccurate in the calculation of stack length, especially for a high-speed iron-loss-dominated machine.

Considering the iron loss, the machine torque equation in [7] can be calibrated by

$$
T_{ca} = \frac{P_a}{P_a - P_{iron}} T
$$
where $T$ is the machine-designed torque, $T_{ca}$ is the machine-calibrated torque, $P_n$ is the machine-designed power, and $P_{iron}$ is the iron loss.

Due to the existence of iron loss, $T_{ca}$ is smaller than $T$. To ensure the machine torque achieves the designed value, the stack length should also be calibrated:

$$L_{stk, ca} = \frac{T}{T_{ca}}L_{stk}$$

(20)

where $L_{stk}$ is originally calculated stack length $L_{stk}$ and $L_{stk, ca}$ is the calibrated stack length.

With the calibrated stack length $L_{stk, ca}$ the machine iron and copper loss can be updated by

$$P_{iron, ca} = \frac{T}{T_{ca}}P_{iron}$$

(21)

$$P_{cu, ca} = P_{cu, end} + \frac{T}{T_{ca}}(P_{cu} - P_{cu, end})$$

(22)

where $P_{iron, ca}$ and $P_{cu, ca}$ are calibrated iron and copper loss, respectively. $P_{cu, end}$ is the end winding copper loss.

### 4 Simulation results

#### 4.1 Evaluation of the simplified model

In order to verify the rationality of the simplifications and assumptions made in the previous sections, a comparison between the simplified model and the detailed FEM model is given for a three-phase 12/8 SRM. The specifications of this machine are given in Table 2.

Fig. 6 presents the iron and dc copper loss comparisons between the FEM and simplified model for different numbers of turns. In Fig. 6, the average machine output torque for the FEM and simplified model is kept constant, both at $\sim 127$ Nm for a different number of turns. In the FEM model, the $\Theta_{peak}$ is slightly adjusted to get the same expected torque output. It is readily observed that the FEM and simplified model results have good agreements, which indicates that the simplified model is good enough to be applied for the fast evaluation of the machine performance during pre-design.

It can also be found that there is a trade-off between iron loss and copper loss for the selection of $N_w$. The iron loss decreases with the increase of $N_w$. The iron loss is proportional with the rate of change of $B$ and an increased $N_w$ will reduce it. In contrast, the copper loss increases with $N_w$. This is mainly because the inductance increases. It takes a longer time for $\Theta$ to reach $\Theta_{peak}$, which eventually causes an increased $\Theta_{rms}$. With a constant total copper area, the copper loss increases with the increase of $\Theta_{rms}$ based on Joule's first law. For the example high-speed machine, the iron loss dominates. So the optimal efficiency is obtained at the maximum number of turns.

#### 4.2 Iron loss comparisons

Fig. 7 depicts the iron and dc copper loss results for different $\Theta_{peak}$ and machine configurations for 200 kW and 15,000 rpm at $R_2 = 150$ mm and CTSR = 0.6. The corresponding stack length and $N_w$ results are given in Figs. 8 and 9, respectively. At all $\Theta$ levels, more rotor poles are advantageous for stack length. Before flux saturation, both iron and copper losses have a slight upward tendency with the increase of $\Theta_{peak}$ for all configurations. The increase of copper loss is caused by increasing the end-winding copper loss. The fluctuation of iron loss is mainly because $N_w$ is round down to the nearest whole number. Obviously, low $\Theta_{peak}$ will lead to poor power density. However, for copper-loss-dominated machines, the flux saturation point is a good trade-off between machine loss and power density. After flux saturation,
5 Conclusion

In this paper, a fast iron loss modelling method is implemented to quickly evaluate and compare a large number of machine configurations in the pre-design stage. Besides, this method can also be applied for obtaining a more accurate stack length prediction. The results show that the efficiency of iron loss-dominated machines can be improved by adopting a higher number of turns, higher peak $\Theta$, and fewer pole pair configurations. The results also lay ground for further efficiency optimisation of SRM in the pre-design stage.

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7 References

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