Eddy-current losses in the segmented surface-mounted magnets of a PM machine

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Abstract: A gas-turbine driven, high-speed, high-efficiency generator system intended for use in series-hybrid vehicles is developed. It consists of a permanent-magnet generator with surface-mounted magnets and a six-pulse controlled rectifier. The stator currents of the rectifier-loaded generator contain time harmonics which cause eddy-current losses in the magnets. These losses can be so high that they result in demagnetisation of the magnets. To reduce these losses, the magnets may be segmented. The aim of the paper is to model the eddy-current losses in such segmented magnets. This is done by incorporating magnet loss resistances in the equivalent circuits of the permanent-magnet machine. The model is verified by means of locked-rotor tests. To show the usefulness of the model, the losses in the magnets of a rectifier-loaded permanent-magnet machine are calculated. The eddy-current losses in the magnets can be decreased by increasing the number of magnet segments: these losses are proportional to the square of the magnet width.

1 Introduction

This paper originates from a research project, the aim of which is the development of a gas-turbine driven high-speed, high-efficiency generator system. This generator system is intended for use in series-hybrid vehicles, the drive system of which is depicted in Fig. 1. It can also be applied in aircraft, vessels, mobile ground power stations, and total energy units. A permanent-magnet (PM) generator with surface-mounted magnets and a six-pulse controlled bridge rectifier are used, because of their high efficiency, high reliability, high power density, and the possibilities for high speed.

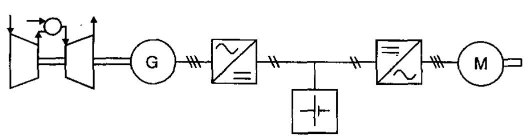

Fig. 1 *Drive system of series-hybrid vehicle, consisting of gas turbine, PM generator, rectifier, accumulator, inverter and motor*

The objective of this paper is to model the eddy-current losses in the magnets (which will be referred to as magnet losses) in a PM machine and to investigate the effect of segmenting the magnets. The magnet losses are mainly caused by the time harmonics of the stator currents of the rectifier-loaded machine. When designing high-speed PM machines, it is important to know these losses, because they heat the rotor, which can only be cooled with difficulty, and the magnets demagnetise if they become too hot. For example, NdFeB magnets may demagnetise at about 120°C.

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Usually, the magnet losses in PM machines are neglected. This is allowed for machines with plastic bonded or ferrite magnets, which have a very high resistivity. However, the resistivity of sintered rare-earth magnets is much lower (typically between 0.5 and 1.5μΩ m). There are a few indications in literature that the magnet losses in high-speed machines are important. In [1], it is stated that there are machines with large magnets that become too hot because of these losses. According to [2], the magnet losses in a flywheel machine are comparable to the iron losses in the solid rotor. In [3], it is shown that the losses in the magnet pole arcs are even larger than the iron losses in the solid rotor.

Magnet losses may be avoided by using a copper cylinder shielding the rotor [4]. The use of such a cylinder may be necessary when the rotor iron is solid. However, to compare machines with and without shielding cylinder and to investigate the effect of segmenting the magnets, a model for the losses in segmented magnets is necessary, which is derived in this paper.

In [3, 5–8], the magnet losses in PM machines are calculated. However, in [5–8], the magnets are replaced by a cylinder of magnet material. This does not give realistic results when the magnets are segmented, which is done to reduce the magnet losses [1, 8]. Therefore, in this paper, the losses in segmented magnets are modelled. In [3], the magnet losses are calculated in such a way that both segmented magnets and flux redistribution can be considered. However, the effect of segmenting the magnets is not studied: complete magnet pole arcs are considered. Besides, the method used in this paper is quite complicated.

Often, the iron losses in electrical machines are represented by 'iron loss resistances' connected in parallel to inductances in the machine equivalent circuits [9], because this simplifies the calculations. In this paper, the losses in the segmented magnets are represented by 'magnet loss resistances' in the equivalent circuits. This is also done in [10], but here this is used to study the transient performance of a PM machine; the magnet losses are not considered.

This paper starts with the calculation of the magnet losses in a PM machine. Next, these losses are represented by magnet loss resistances in the machine model. Subsequently, the derived model is verified by means of locked-rotor tests. To illustrate the usefulness of the model, it is used to calculate the magnet losses in a PM machine with rectifier load. Finally, conclusions are drawn.

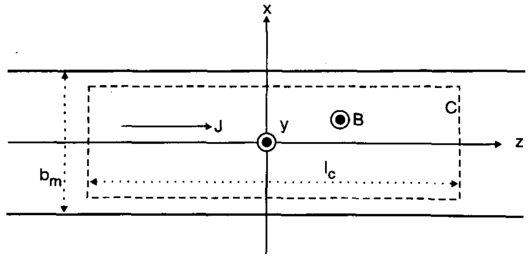

Fig. 2 *Cross-section of magnet segment in rectangular co-ordinate system*

2 Eddy-current losses in the magnets of a PM machine

2.1 Eddy-current losses per unit of magnet volume

Fig. 2 depicts a cross-section of a magnet segment in a rectangular co-ordinate system. The current density in the magnets is calculated with the second of Maxwell's equations:

$$\oint_C \boldsymbol{E} \cdot d\boldsymbol{s} = -\frac{d}{dt} \iint_S \boldsymbol{B} \cdot d\boldsymbol{a} \tag{1}$$

This equation is applied to the dashed closed path C in Fig. 2, which has length l_C in the (axial) z-direction. In this equation, the following assumptions are used.

(i) The magnetic flux density is perpendicular to the plane of the drawing.

(ii) The effect of eddy currents in the magnets on the magnetic field is negligible.

(iii) End effects are negligible, so that the current density only has a component in the z-direction, and the two sides of the closed path parallel to the x-axis do not contribute to the line integral. This assumption is reasonable if the magnet length in the z-direction is much larger than the magnet width b_m. If this is not the case, the losses are overrated, up to about a factor two when the magnet length and the magnet width are comparable.

(iv) The magnet segments are so small that the magnetic flux density can be considered constant over the magnet width. Consequentially, the current density is an odd function of x: $J_z(-x) = -J_z(x)$. When large or cylindrical magnets are used, other methods [3, 5–8] may give better results.

Further, the electric field strength is replaced by the product of the current density and the resistivity of the magnet: $\boldsymbol{E} = \rho_m \boldsymbol{J}$. The resulting expression for the current density is:

$$J_z(x) = \frac{x}{\rho_m} \frac{dB}{dt} \tag{2}$$

With this, the eddy-current losses per unit of magnet volume are calculated as:

$$k_m = \frac{1}{b_m} \int_{-b_m/2}^{b_m/2} \rho_m J_z^2(x) dx = \frac{b_m^2}{12\rho_m} \left(\frac{dB}{dt} \right)^2 \tag{3}$$

Comparable expressions are derived in, for example [11, 12] for the eddy-current losses in laminated iron.

Because of the assumptions used, the losses are not calculated very accurately. However, it is expected that for machines with segmented magnets, the results are much better than the results calculated assuming a cylinder of magnet material.

2.2 Eddy-current losses in a PM machine

It is assumed that the magnet losses caused by the space harmonics of the stator windings and the stator slotting are negligible. Calculations in [7, 13] indicate that these losses are much smaller than the losses caused by the time harmonics of the stator currents. Therefore, the magnetic flux density causing the losses can be written as:

$$B(\alpha_r) = \hat{B} \cos(p(\alpha_r - \beta)) \tag{4}$$

where p is the number of pole pairs of the machine. Both β and $\hat{B}$ may be functions of time, because in a rectifier-loaded machine in steady state, β is a function of time, while during the locked-rotor tests (Section 4) $\hat{B}$ is a function of time.

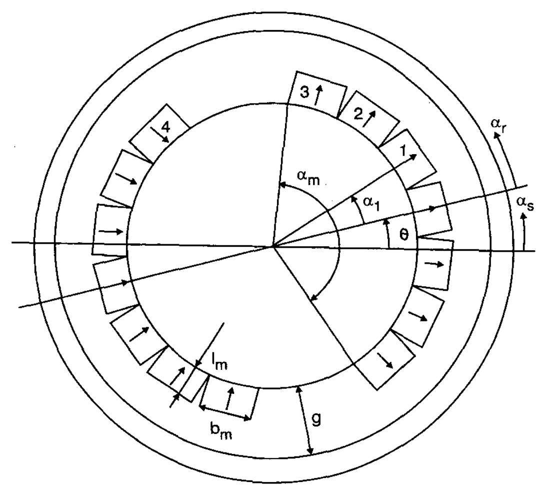

Fig. 3 *Cross-section of two-pole PM machine*

The magnets are numbered 1 to N_m, and the axis of kth magnet lays at rotor co-ordinate α_k (Fig. 3). The magnetic flux density was assumed to be constant over the magnet width b_m. With this, the magnetic flux density in the kth magnet can be written as:

$$B_k = B(\alpha_k) = \hat{B} \cos(p(\alpha_k - \beta)) \tag{5}$$

When this is used in eqn. 3, the eddy-current losses per unit of magnet volume in the kth magnet are calculated as:

$$k_{m,k} = \frac{b_m^2}{12\rho_m} \left(\frac{d}{dt} \left\{ \hat{B} \cos(p(\alpha_k - \beta)) \right\} \right)^2 \tag{6}$$

Multiplication of this expression by the volume of the magnet gives the losses in the kth magnet. Summation over all magnets results in the total magnet losses:

$$P_m = l_s l_m b_m \sum_{k=1}^{N_m} \frac{b_m^2}{12\rho_m} \left(\frac{d}{dt} \left\{ \hat{B} \cos(p(\alpha_k - \beta)) \right\} \right)^2$$
$$\approx 2pr_s l_s l_m \frac{b_m^2}{12\rho_m} \times \int_{-\alpha_m/2}^{\alpha_m/2} \left(\frac{d}{dt} \left\{ \hat{B} \cos(p(\alpha_r - \beta)) \right\} \right)^2 d\alpha_r$$

$$= \frac{r_s l_s l_m b_m^2}{12\rho_m} \times \left\{ (p\alpha_m + \sin(p\alpha_m)) \left(\frac{d}{dt} \left\{ \hat{B} \cos(p\beta) \right\} \right)^2 + (p\alpha_m - \sin(p\alpha_m)) \left(\frac{d}{dt} \left\{ \hat{B} \sin(p\beta) \right\} \right)^2 \right\} \quad (7)$$

where l_m is the thickness of the magnets, α_m is the magnet pole arc (see Fig. 3), r_s is the air-gap radius, and l_s is the stack length of the machine. The approximation in this equation is based on the assumption that the magnet width b_m is small.

For each time harmonic of the stator currents, the magnetic flux density can be written as in eqn. 4, and the magnet losses may be calculated with eqn. 7. However, a different way is followed here: the losses are incorporated into the machine model by means of a magnet loss resistance for the following reason. In the calculations, it has been assumed that the effect of eddy currents on the magnetic field is negligible. However, this assumption is not valid at very high frequencies, as appears from the locked-rotor tests (Section 4). When the magnet losses are incorporated into the machine model by means of a magnet loss resistance, this effect is also visible from the model: the magnet loss resistance becomes smaller than the reactance it is connected in parallel to. Therefore, incorporating a magnet loss resistance in the equivalent circuit is a means to check if the mentioned assumption is valid. In [13], it is even shown that this rough way of incorporating the effect of eddy currents on the magnetic field makes sense.

3 Magnet loss resistance

To represent the magnet losses by magnet loss resistances in equivalent circuits, we need the voltage equations of the machine. Here, a condensed derivation of these voltage equations is given; for a more thorough derivation is referred to [9].

3.1 Magnetic flux density of the stator currents

The losses caused by the space harmonics of the stator windings are assumed to be negligible. Therefore, only the fundamental of the stator winding distribution is considered. The conductor density (the number of conductors per radian) of phase a is then given by:

$$n_{sa}(\alpha_s) = \frac{1}{2} N_s \sin(p\alpha_s) \quad (8)$$

where N_s is the number of turns of the fundamental of the stator winding distribution, which is related to the actual number of turns N by:

$$N_s = \frac{4}{\pi} k_w N \quad (9)$$

where k_w is the winding factor of the actual winding.

The magnetic flux density in the air gap caused by the stator currents is

$$B_s(\alpha_s) = \frac{\mu_0 N_s}{2gp} \left\{ i_{sa} \cos(p\alpha_s) + i_{sb} \cos\left(p\alpha_s - \frac{2}{3}\pi\right) + i_{sc} \cos\left(p\alpha_s - \frac{4}{3}\pi\right) \right\} \quad (10)$$

where g is the effective air gap (see Fig. 3), which includes the magnets, because the relative magnetic permeability of the magnets is assumed to be one.

3.2 Flux linkages of the stator windings

At an arbitrary moment, the magnetic flux density in the air gap can be written as a Fourier series:

$$B(\alpha_s) = \sum_{k=1,3,5,\ldots}^{\infty} \hat{B}_k \cos(kp(\alpha_s - \beta_k)) \quad (11)$$

To obtain an expression for the flux linkage of a stator winding, first the flux linkage of a full-pitch turn $\psi_t(\alpha')$ at stator co-ordinate α' is calculated:

$$\psi_t(\alpha') = \iint_S \boldsymbol{B} \cdot d\boldsymbol{a} = \int_{\alpha'-\pi/p}^{\alpha'} B(\alpha_s) l_s r_s d\alpha_s \quad (12)$$

Using eqns. 8, 11 and 12, the flux linkage of stator phase a ψ_{sa} is obtained by integration:

$$\psi_{sa} = p \int_0^{\pi/p} n_{sa}(\alpha')\psi_t(\alpha')d\alpha' = \frac{\pi r_s l_s N_s}{2p} \hat{B}_1 \cos(p\beta_1) \quad (13)$$

In the same way, the flux linkages of the other stator phases can be calculated:

$$\begin{bmatrix} \psi_{sa} \\ \psi_{sb} \\ \psi_{sc} \end{bmatrix} = \frac{\pi r_s l_s N_s}{2p} \hat{B}_1 \begin{bmatrix} \cos(p\beta_1) \\ \cos\left(p\beta_1 - \frac{2}{3}\pi\right) \\ \cos\left(p\beta_1 - \frac{4}{3}\pi\right) \end{bmatrix} \quad (14)$$

3.3 Voltage equations

When R_s is the resistance of a stator phase, the stator voltages can be written as

$$\begin{bmatrix} u_{sa} \\ u_{sb} \\ u_{sc} \end{bmatrix} = R_s \begin{bmatrix} i_{sa} \\ i_{sb} \\ i_{sc} \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \psi_{sa} \\ \psi_{sb} \\ \psi_{sc} \end{bmatrix} \quad (15)$$

The flux linkages in this equation are separated into different contributions, namely:

(i) the flux linkage due to leakage fields $\psi_{s\sigma}$, and

(ii) the flux linkage due to the air-gap field, which again is separated into:

(*a*) the flux linkage due to the field of the magnets ψ_{sm}, and

(*b*) the flux linkage due to the field of the stator currents ψ_{ss}.

The self-inductances of the leakage flux of the stator phases are called $L_\sigma + M_{s\sigma ab}$, the mutual inductances of the leakage flux between the different stator phases are called $M_{s\sigma ab}$. Hence, the flux linkage due to leakage fields can be written as

$$\begin{bmatrix} \psi_{s\sigma a} \\ \psi_{s\sigma b} \\ \psi_{s\sigma c} \end{bmatrix} = \begin{bmatrix} L_\sigma + M_{s\sigma ab} & M_{s\sigma ab} & M_{s\sigma ab} \\ M_{s\sigma ab} & L_\sigma + M_{s\sigma ab} & M_{s\sigma ab} \\ M_{s\sigma ab} & M_{s\sigma ab} & L_\sigma + M_{s\sigma ab} \end{bmatrix} \begin{bmatrix} i_{sa} \\ i_{sb} \\ i_{sc} \end{bmatrix} \quad (16)$$

The magnetisation of the magnets is constant. Therefore, the flux linkage due to the field of the magnets ψ_{sm} only depends on the rotor position angle θ (see Fig. 3). The time derivative of this flux linkage is the no-load voltage:

$$\begin{bmatrix} e_{pa} \\ e_{pb} \\ e_{pc} \end{bmatrix} = \frac{d}{dt}\begin{bmatrix} \psi_{sma} \\ \psi_{smb} \\ \psi_{smc} \end{bmatrix} \tag{17}$$

The stator flux linkages due to the field of the three-phase stator (eqn. 10) are calculated in the same way as in eqn. 14 as

$$\begin{bmatrix} \psi_{ssa} \\ \psi_{ssb} \\ \psi_{ssc} \end{bmatrix} = \frac{1}{3}L_m \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix}$$

$$L_m = \frac{3\mu_0 \pi r_s l_s N_s^2}{8gp^2} \tag{18}$$

Incorporating this into the voltage equations (eqn. 15) and transforming them to the dq-system by means of the Park transformation [14], given by:

$$\begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} = \boldsymbol{P}\begin{bmatrix} i_{sa} \\ i_{sb} \\ i_{sc} \end{bmatrix} \qquad \begin{bmatrix} u_{sd} \\ u_{sq} \end{bmatrix} = \boldsymbol{P}\begin{bmatrix} u_{sa} \\ u_{sb} \\ u_{sc} \end{bmatrix}$$

$$\begin{bmatrix} \psi_{sd} \\ \psi_{sq} \end{bmatrix} = \boldsymbol{P}\begin{bmatrix} \psi_{sa} \\ \psi_{sb} \\ \psi_{sc} \end{bmatrix} \qquad \begin{bmatrix} 0 \\ e_{pq} \end{bmatrix} = \boldsymbol{P}\begin{bmatrix} e_{pa} \\ e_{pb} \\ e_{pc} \end{bmatrix}$$

$$\boldsymbol{P} = \sqrt{\frac{2}{3}}\begin{bmatrix} \cos(p\theta) & \cos\left(p\theta - \frac{2}{3}\pi\right) & \cos\left(p\theta - \frac{4}{3}\pi\right) \\ -\sin(p\theta) & -\sin\left(p\theta - \frac{2}{3}\pi\right) & -\sin\left(p\theta - \frac{4}{3}\pi\right) \end{bmatrix} \tag{19}$$

results in

$$\begin{bmatrix} u_{sd} \\ u_{sq} \end{bmatrix} = \begin{bmatrix} 0 \\ e_{pq} \end{bmatrix} + R_s \begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} + (L_\sigma + L_m)\frac{d}{dt}\begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} + p\Omega(L_\sigma + L_m)\begin{bmatrix} -i_{sq} \\ i_{sd} \end{bmatrix} \tag{20}$$

where the zero-components were omitted, because they are zero, and where Ω is the mechanical angular speed of the rotor.

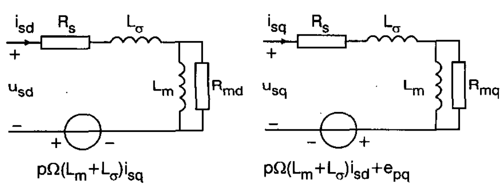

Fig. 4 *Direct-axis and quadrature-axis equivalent circuits of PM machine*

3.4 Magnet loss resistance

The magnet losses can be represented by magnet loss resistances in the equivalent circuits of the PM machine, as illustrated in Fig. 4. To determine the resistances, the losses in these resistances caused by the magnetic flux density of eqn. 4 are calculated and compared to the losses calculated in eqn. 7. The stator flux linkages due to the magnetic flux density of eqn. 4 are calculated in the same way as in eqn. 14. They are transformed to the dq-system with the Park transformation (eqn. 19):

$$\begin{bmatrix} \psi_{sd} \\ \psi_{sq} \end{bmatrix} = \sqrt{\frac{3}{2}}\frac{\pi r_s l_s N_s}{2p}\hat{B}\begin{bmatrix} \cos(p\beta) \\ \sin(p\beta) \end{bmatrix} \tag{21}$$

The induced voltages are the time derivatives of these fluxes. Therefore, the losses in the magnet loss resistances R_{md} and R_{mq} are given by:

$$P_m = \frac{1}{R_{md}}\left(\frac{d\psi_{sd}}{dt}\right)^2 + \frac{1}{R_{mq}}\left(\frac{d\psi_{sq}}{dt}\right)^2 \tag{22}$$

Comparison of this equation with eqn. 7 shows that the magnet losses can be represented by magnet loss resistances if the values of these resistances are given by

$$R_{md} = \frac{9\rho_m \pi^2 r_s l_s N_s^2}{2 l_m p^2 b_m^2 \left(p\alpha_m + \sin(p\alpha_m)\right)}$$
$$R_{mq} = \frac{9\rho_m \pi^2 r_s l_s N_s^2}{2 l_m p^2 b_m^2 \left(p\alpha_m - \sin(p\alpha_m)\right)} \tag{23}$$

4 Locked-rotor tests

The derived model is verified by comparing the calculated and the measured impedances of a PM machine with locked rotor. The rotor of the 6-pole PM machine is nearly covered with small magnets (b_m = 5mm), so there is hardly any difference between the direct-axis and the quadrature-axis impedances. Therefore, only the direct-axis locked-rotor tests are reported.

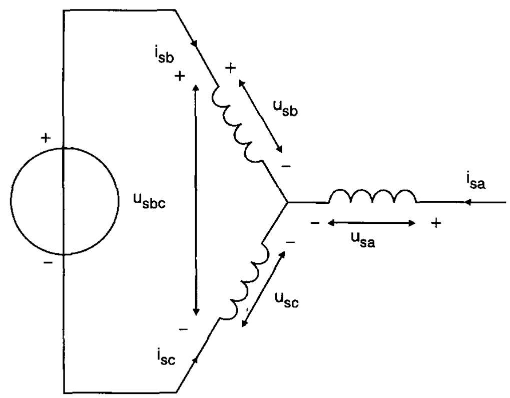

Fig. 5 *Measurement circuit*

During these tests, a sinusoidal voltage is supplied to phases b and c, which are connected in series, as depicted in Fig. 5. The stator field axis coincides with the direct axis ($p\theta = \pi/2$). With the Park transformation (eqn. 19), the dq-components of the currents and voltages are calculated. The direct-axis impedance is:

$$\underline{Z} = \frac{\hat{\underline{u}}_{sbc}}{\hat{\underline{i}}_{sb}} = 2\frac{\hat{\underline{u}}_{sd}}{\hat{\underline{i}}_{sd}} = 2\left\{R_s + j\omega L_\sigma + \frac{j\omega L_m R_{md}}{R_{md} + j\omega L_m}\right\}$$
$$R = \mathrm{Re}(\underline{Z}) \qquad L = \frac{1}{\omega}\mathrm{Im}(\underline{Z}) \tag{24}$$

The resistance and the inductance of the PM machine are determined from the measured voltage U_{sbc}, current I_{sb}, and dissipation P in the machine as

$$R = \frac{P}{I_{sb}^2} \qquad L = \frac{1}{\omega}\sqrt{\left(\frac{U_{sbc}}{I_{sb}}\right)^2 - R^2} \tag{25}$$

Fig. 6 depicts the measured and the calculated resistance and inductance and gives rise to three remarks.

(i) Differences between the measured and the calculated resistance are probably caused by the neglect of the iron losses in the laminated stator and rotor iron. However, the measurements make sense, because at high frequencies, the iron losses are comparable to, or smaller than, the magnet losses, as can be seen in the following way. At high frequencies, the eddy-current losses in iron dominate the iron losses. The eddy-current losses in the iron and the magnets can be estimated as (compare eqn. 3):

$$P_{Fe} \simeq \frac{V_{Fe} b_{Fe}^2 \hat{B}_{Fe}^2 \omega^2}{12\rho_{Fe}} \qquad P_m \simeq \frac{V_m b_m^2 \hat{B}_m^2 \omega^2}{12\rho_m} \quad (26)$$

• The lamination thickness b_{Fe} is an order of magnitude smaller than the magnet width b_m.

• The iron volume V_{Fe} is an order of magnitude larger than the magnet volume V_m.

• The frequency ω, the resistivity ρ, and the magnetic flux density $\hat{B}$ in iron and magnets have the same order of magnitude.

Hence, the iron losses are comparable to or smaller than the magnet losses.

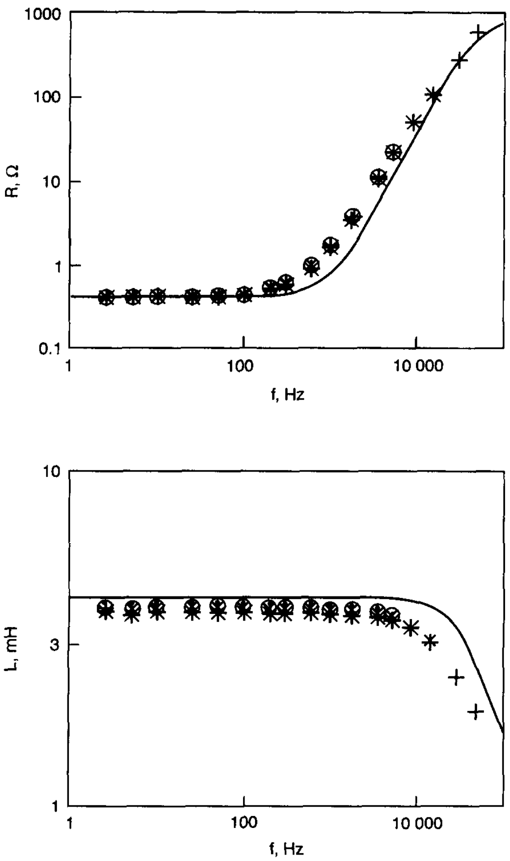

Fig. 6 *Measured and calculated resistance and inductance in the direct axis*
+ I_{sb} = 0.2A, meas. ○ I_{sb} = 2A, meas.
× I_{sb} = 0.5A, meas. ● I_{sb} = 5A, meas.
✳ I_{sb} = 1A, meas. ——— calculated

(ii) The model was derived on the assumption that the effect of eddy currents on the magnetic field was negligible. Above 10kHz, this assumption is not valid, as appears from the decrease of the inductance. However, the agreement between measurements and calculations remains reasonable.

(iii) The increase of the measured resistance due to skin effect in the stator conductors is negligible [13]. The reasonable agreement between measurements and calculations shows that the proposed model is useful.

5 Rectifier-loaded PM machine

To illustrate the usefulness of the proposed model, the PM machine is loaded with a controlled bridge rectifier and the losses are calculated.

The steady-state performance at full load and 6000rpm is calculated with the calculation method described in [15]. Fig. 7 depicts measured and calculated voltage and current waveforms. Because the effect of eddy currents in the magnets on the magnetic field is negligible, they do not affect the voltage and current waveforms. The current waveform is only used to calculate the magnet losses.

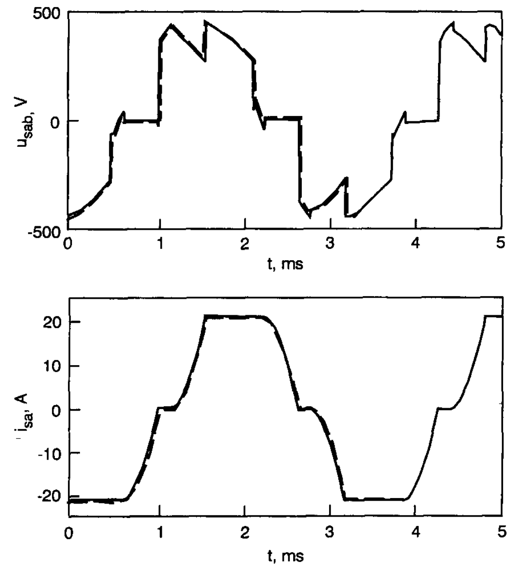

Fig. 7 *Line voltage and phase current*
——— measured
- - - - calculated

Table 1: Losses in a rectifier-loaded PM machine at different speeds

n (rpm)	P_m (W)	P_{Cu} (W)	P_{gen} (kW)
6000	11	163	8.0
30000	246	163	40

In Table 1, the calculated magnet losses P_m are given. For comparison, the stator copper losses P_{Cu} and the generated power P_{gen} are also given. Also the calculated powers at 30000 rpm with the same firing angle and the same current in the direct-current circuit are given, assuming that the stator resistance does not depend on the frequency. This shows that, at high speeds, the magnet losses may be a serious problem, because they are roughly proportional to the square of the frequency, as also follows from eqn. 3.

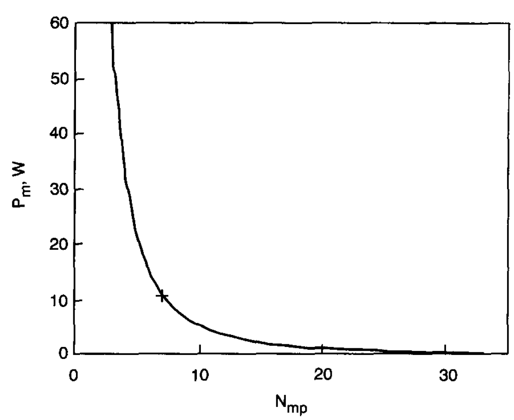

Fig. 8 *Calculated magnet losses P_m as a function of the number of magnet segments per pole N_{mp}*
+ indicates the case of the actual PM machine

Fig. 8 depicts the magnet losses in the rectifier-loaded machine at 6000rpm as a function of the number of magnet segments per pole. This shows that increasing the number of magnet segments is a very effective way to reduce the magnet losses: they are proportional to the square of the magnet width, as could be expected from eqn. 3.

6 Conclusion

When designing high-speed PM machines, it is important to know the magnet losses, because they may be large. Therefore, this paper introduces a model of a PM machine including the magnet losses due to the time harmonics of the stator currents. These losses are represented by magnet loss resistances in the equivalent circuits. The model is verified by means of locked-rotor tests. The magnet losses can be decreased by increasing the number of magnet segments: these losses are proportional to the square of the magnet width.

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