Modeling and Analysis of Linear Synchronous Motors in High-Speed Maglev Vehicles

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Accurate knowledge of magnetic field distribution and thrust and normal force calculation in linear synchronous motors are essential for assessing performances and design of the motors. In this paper, two-dimensional field distribution produced by dc excitation of secondary and ac currents of primary of a high-speed single-sided wound secondary linear synchronous motor is obtained by an analytical method solving Maxwell equations in the motor layers. Further, the determination of electromagnetic forces of this type of motor is presented. The results are compared with those obtained from an adopted base model and the finite-element method (FEM). Anisotropy, field harmonics and primary slot effects are investigated. Good correlations between the results obtained by the proposed method and the finite-element method confirm the superiority of the former method over the base method.

Index Terms—Electromagnetic analysis, electromagnetic fields, electromagnetic forces, finite-element analysis, linear synchronous motors, modeling.

I. INTRODUCTION

EXTRA high-speed rail transportation has progressed by different Maglev technologies for decades [1]. Two major competing technologies are German Transrapid and Japanese superconducting systems [2]. The former uses electromagnetic levitation while the latter enjoys a totally different electrodynamic levitation. However, both systems are propelled by wound secondary linear synchronous motors due to salient features of this type of motor [3], [4].

A linear synchronous motor (LSM) enjoys high efficiency due to a lack of slip losses and high magnetizing current. In Maglev applications the motor does not require contactless high power transmission as it is essential for induction motors. Also, the machine power factor can be controlled to higher values than a fixed power factor which is obtained by a comparable induction motor at the same output power and speed. Higher efficiency and power factor lead to a significant reduction of inverter rating, resulting in a substantial cost saving.

Many aspects of LSMs have been studied in the literature, including their modeling, analysis, design and control [5]–[10]. Among these studies the machine modeling plays a fundamental role since it is required for all other studies. Early attempts on LSM modeling are just marginal modifications of rotary synchronous machine modeling in which a rotational speed is replaced by a translational velocity [11]. This type of modeling ignores the essential topology of LSMs, i.e., the machine flatness and its consequences like the phenomenon of end effect. This phenomenon in particular, has significant effect on machine performance especially at high speeds. Therefore, the traditional LSM modeling is not appropriate to high-speed applications.

Recent efforts overcome the shortcoming of early modeling methods by finite element method (FEM), taking into account the linear topology of the machines [5]–[7]. However, a FEM model although accurate in studying machine performance, lacks analytical studies. Therefore, they face limitations in tasks like design optimization of machines. A FEM analysis can evaluate a final design but it is not time efficient during design procedures, since it needs many long iterations to reach a desirable design. The thrust and levitation force characteristics of an electrodynamically levitated linear synchronous motor are calculated analytically [8]. However, the field distribution of the motor is assumed to be known for example from a FEM model. Some researchers have presented approximated d-q models of LSMs [9], [10]. These rather simple models are useful for machine analysis and control. However, they are not accurate enough for design and optimization purposes.

More recently an analysis and design of a short primary linear synchronous motor for high-speed applications has been presented, ignoring the effect of permeability of the iron core of electromagnets on the field distribution [12]. Many recent studies have focused heavily on permanent-magnet LSMs [13]–[26]. Different analytical and numerical methods for such machines have been presented [27]–[29]. However, permanent-magnet LSMs are mainly used in low-power automation applications. They still are not considered heavily in high-power high-speed applications like Maglev.

This paper aims at presenting a full analytical model of wound secondary linear synchronous motors (WSLSMs) considering some features of the motors missing in the existing literature, e.g., the field harmonics and anisotropy. A primary attention is paid to the field and force calculations by taking into account the flatness of machines. A layer model approach is followed and rather accurate expressions are obtained for the machine characteristics. The main contribution of this paper is modeling of a wound secondary motor instead of a machine with a coreless or a permanent-magnet secondary as reported in the literature [30], [31]. Therefore, the iron core of the secondary is taken into account. A second model is also developed in this paper for WSLSMs as a basis of comparison by improving a model already available for permanent-magnet LSMs [32]. Both models plus a FEM based model are applied...
to a WSLSM and their results are compared. It is shown that the motor characteristics obtained by the proposed model are closer to FEM results.

II. MACHINE STRUCTURE

Fig. 1 shows a schematic view of a single-sided WSLSM with a long primary. The primary includes a three-phase iron-core winding which is extended along the motion path. The windings are made of copper and laid in the open slots of the primary iron core. The slot-opening of the windings is three slots. The three phases are supplied from several inverter stations which are placed along the path.

The moving part of the motor is a short secondary which consists of a back iron and dc excitation electromagnetic poles. The secondary is located above the primary with an air gap, \( g_c \), and may experience a contactless motion along the path. However, for being compatible with a more common layer model, depicted in the next section, Fig. 1 is drawn upside down as shown. The electromagnetic poles of secondary consist of dc excited copper windings and iron cores which are connected to the back iron. The parameters and dimensions of the motor are as follow: \( \tau_s \) is the slot pitch, \( b_s \) is the slot width, \( h_m \) is the electromagnetic pole height, \( \tau_p \) is a pole pitch, and \( \tau_m \) is the electromagnet width which is width of iron core of a pole plus almost half of a dc coil width, \( h_a \) is the slot height, and \( L \) is the motor width along the z direction.

III. THE PROPOSED LAYER MODEL

In this section, a method for developing a layer model for the motor presented in the previous section is present. All harmonics of the magnetic field produced by dc-excited electromagnets are considered. The method is based on solving Maxwell equations in several layers of the motor.

A. The Physical Model

Fig. 2 shows a layer model of the machine assuming that the primary is not excited. The layer representing dc excitation consists of iron core electromagnetic poles and the gap between them. As the iron permeability is much greater than the air permeability, the permeability throughout this layer is not monotonous. Therefore, this layer is modeled by an anisotropic layer with different permeability along \( x \) and \( y \) directions. The magnetic vector potentials in

\[
\begin{align*}
\mu_{xmn} &= \mu_0 \left[ \frac{\mu_r}{1 + [(\tau_p - \tau_m)/\tau_p] \times (\mu_r - 1)} \right] \\
\mu_{ymn} &= \mu_0 \left[ \frac{\tau_p - \tau_m)}{\tau_p} + \mu_r (1 - (\tau_p - \tau_m)/\tau_p) \right]
\end{align*}
\] (1)

where \( \mu_r \) is the relative permeability of iron. In addition to this assumption, the following major assumptions are used in deriving the mathematical model:

1) all regions are extended to infinity in \( \pm X \) direction;
2) permeability of back iron and primary yoke are equal to infinity;
3) analysis model behaves linearly;
4) primary slots are ignored and instead the air gap is increased by a Carter coefficient \( (K_c) \) [33]:

\[
K_c = \frac{\tau_s}{(\tau_s - \gamma g')}
\] (2)

where \( \gamma \) and \( g' \) are given respectively by

\[
\gamma = \frac{4}{\pi} \left\{ \frac{b_y}{2g'} \tan^{-1} \left( \frac{b_y}{2g'} \right) - \ln \left[ 1 + \left( \frac{b_y}{2g'} \right)^2 \right] \right\}
\] (3)

\[
g' = g + \frac{h_m}{\mu_r}
\] (4)

where \( \mu_r \) is the relative permeability of iron. Therefore, the equivalent air gap, \( g_e \), is calculated as

\[
g_e = g + (K_c - 1)g'
\] (5)

5) each electromagnetic pole is modeled by a MMF source \( M_p = N_{lk}I_{dc} \). At, representing an excitation with a square wave function along \( x \) direction, provided by a dc excitation coil of \( N_{dc} \) turns. \( I_{dc} \) is the dc current of each coil.

B. Field Calculation

The Maxwell equations lead to Laplace and Poisson equations as follows [34]:

\[
\begin{align*}
\frac{\partial^2 A_y(x, y)}{\partial x^2} + \frac{\partial^2 A_y(x, y)}{\partial y^2} &= 0 \\
\frac{\partial}{\partial x} \left( \frac{\partial A_{1x}}{\mu_{1x}} \right) + \frac{\partial}{\partial y} \left( \frac{\partial A_{1y}}{\mu_{1y}} \right) &= -J_m(x)
\end{align*}
\] (6)

where \( \mu_{xmn} \) and \( \mu_{ymn} \) are the permeability of pole region in \( x \) and \( y \) directions, \( A_I \) and \( A_{II} \) are the magnetic vector potentials in
layers I and II, and \( J_m \) is the equivalent current density of the electromagnetic pole given by

\[
J_m(x) = \sum_{n=1,3,...} J(n) \sin \left( \frac{n\pi x}{\tau_p} \right) \tag{7}
\]

where \( J(n) \) is the maximum value of this current density for different harmonics given by

\[
J(n) = \frac{4}{\tau_p} \frac{N_k}{h_m} \sin \left( \frac{\eta n\pi}{2} \right) \tag{8}
\]

where \( \eta \) stands for an electromagnet width to pole pitch ratio. The corresponding general solutions of (6) are obtained as

\[
A_I(x,y) = \sum_{n=1,3,...} \left( C_1 e^{\frac{n\pi y}{\tau_p}} + C_2 e^{\frac{-n\pi y}{\tau_p}} \right) \sin \left( \frac{n\pi x}{\tau_p} \right)
\]

\[
A_{II}(x,y) = \sum_{n=1,3,...} \left( C_3 e^{\sqrt{\frac{\mu_{ym}}{\mu_{ym}}} \frac{n\pi y}{\tau_p}} + C_4 e^{-\sqrt{\frac{\mu_{ym}}{\mu_{ym}}} \frac{n\pi y}{\tau_p}} \right) + k(n) \sin \left( \frac{n\pi x}{\tau_p} \right) \tag{9}
\]

where

\[
k(n) = \frac{\mu_{ym} I(n)}{n^2 \pi^2 \tau_p^2} \tag{10}
\]

Assuming that the permeability of back iron is infinity, the boundary conditions are fulfilled as [24]

\[
\begin{align*}
  \{ y = g_e & \rightarrow H_{xI} = 0 \\
  y = 0 & \rightarrow H_{xI} = H_{xII} \& B_{xI} = B_{xII} \\
  y = -h_m & \rightarrow H_{xII} = 0
\end{align*} \tag{11}
\]

Therefore, the constants \( C_1 - C_4 \) will be found as

\[
C_1 = \frac{k(n)}{1 + e^{\frac{2\pi n \tau_p}{g_e}}} - \sqrt{\frac{\mu_{re} \mu_{ym}}{\mu_0}} \left( 1 - e^{-\frac{2\pi n \tau_p}{g_e}} \right)
\]

\[
C_2 = C_1 e^{\frac{2\pi n \tau_p}{g_e}}
\]

\[
C_3 = C_1 \sqrt{\frac{\mu_{re} \mu_{ym}}{\mu_0}} \left( 1 - e^{\frac{2\pi n \tau_p}{g_e}} \right) \left( 1 - e^{-\sqrt{\frac{\mu_{ym}}{\mu_{ym}}} \frac{2\pi n \tau_p}{g_e} h_m} \right)
\]

\[
C_4 = C_3 e^{-\sqrt{\frac{\mu_{ym}}{\mu_{ym}}} \frac{2\pi n \tau_p}{g_e} h_m} \tag{12}
\]

It is known that the flux density in each layer can be achieved by the curl of magnetic vector potential of the layer, which has been given by (9). Because of the symmetry, the direction of magnetic vector potential is perpendicular to the \( x-y \) plane. Therefore, the flux density distribution is obtained as

\[
\mathbf{B} = \nabla \times \mathbf{A} = \frac{\partial A_y}{\partial x} - \frac{\partial A_x}{\partial y} \tag{13}
\]

As a result, the normal component of flux density at \( y = y_0 \) is given by

\[
B_y(x) = -\frac{\partial A}{\partial x} \tag{14}
\]
current (rms), $M_p$ is the dc-excitation MMF source which is equal to $N_{dc}I_{dc}$. $\alpha_0$ is the primary current angle, and $K_{uv1}$ is a winding factor. In this paper, the current sheets will be considered in the middle of the electromagnet height $h_m$ and slot height $h_s$, respectively, i.e., $y_m = h_m/2$ and $y_s = h_s/2$. From Maxwell equations, $H_x$ and $H_y$ in each region have already been obtained [32]. However, the ten constants of those equations, i.e., $A, B, C, D, E, H_{x3'}, H_{x3''}, H_{y2}, H_{y2'}, H_{x3''}$, have not been presented in the literature. Here, these constant are obtained as

$$H_{x1'} = \frac{1}{\sinh K_s(h_m + g - h_t)} \times [H_{x3'} \left( \frac{\sqrt{\mu_0 \mu_m g y_m}}{\mu_0} \sinh \frac{\pi}{\tau_p} g \cosh K_m h_m + \cosh \frac{\pi}{\tau_p} g \sinh K_m h_m \right)$$

$$+ A_{1p} \cosh K_s(h_m + g - (y_t - y_s)) - A_{1p} \left( \frac{\sqrt{\mu_0 \mu_m g y_m}}{\mu_0} \sinh \frac{\pi}{\tau_p} g \cosh K_m(h_m - y_m) + \cosh \frac{\pi}{\tau_p} g \sinh K_m(h_m - y_m) \right)]$$

$$H_{x3'} = H_{x3}/\cosh K_s h_s$$

$$H_{x2} = \frac{\sqrt{\mu_0 \mu_m g y_m}}{\mu_0} \times [H_{x3'} \cosh K_m h_m - A_{1p} \sinh K_m(h_m + y_m)]$$

$$H_{x3} = \frac{\varepsilon}{\tau_0} A_{1p} + \frac{\varepsilon}{\tau_0} A_{1s}$$

$$H_{x3''} = H_{x3'} \cosh K_m y_m$$

$$A = E = 0$$

$$B = [H_{x3'} \sinh K_m y_m - A_{1p}]/H_{x3''}$$

$$C = [H_{x3'} \sinh K_m y_m - A_{1p} \cosh K_m y_m]/H_{x2}$$

$$D = [H_{x3'} \sinh K_m y_m - A_{1s}]/H_{x3''}$$

where

$$v_{000} = \frac{\sqrt{\mu_0 \mu_m g y_m}}{\mu_0} \left[ - \cosh K_s(h_m + g - (y_t - y_s)) \times \sinh K_s(h_m + g - h_t) \right]$$

$$v_{00} = \frac{\sqrt{\mu_0 \mu_m g y_m}}{\mu_0} \left[ \frac{\sqrt{\mu_0 \mu_m g y_m}}{\mu_0} \sinh \frac{\pi}{\tau_p} g \cosh K_m(h_m - y_m) + \cosh \frac{\pi}{\tau_p} g \sinh K_m(h_m - y_m) \right]$$

$$- \tanh K_s(h_m + g - h_t) \times \left[ \frac{\sqrt{\mu_0 \mu_m g y_m}}{\mu_0} \cosh \frac{\pi}{\tau_p} g \sinh K_m(h_m - y_m) + \sinh \frac{\pi}{\tau_p} g \cosh K_m(h_m - y_m) \right]$$

$$v_0 = \frac{\sqrt{\mu_0 \mu_m g y_m}}{\mu_0} \left[ \frac{\sqrt{\mu_0 \mu_m g y_m}}{\mu_0} \sinh \frac{\pi}{\tau_p} g \cosh K_m h_m + \cosh \frac{\pi}{\tau_p} g \sinh K_m h_m \right]$$

Therefore, the normal component of flux density in the air gap is given by

$$B_{y2}(y) = \frac{\mu_0 H_{y2}}{\tau_p} \left[ \cosh \frac{\pi}{\tau_p} (y - h_m) + C \sinh \frac{\pi}{\tau_p} (y - h_m) \right]. \quad (20)$$

V. THRUST AND NORMAL FORCE CALCULATION

Since the motor is a single-sided WSLSM, in addition to the thrust $F_x$, a normal force $F_n$ of attraction type is also developed between the primary and secondary cores. These forces can be calculated from the electromagnetic field distribution in the air gap. The thrust for the fundamental harmonic of electromagnetic field produced by armature currents can be found on the basis of Lorentz equation. The force increment acting on an electromagnetic pole with the coordinate $x = \tau_m/2$ and a current density of dc-excitation equal to $J_m = N_{dc}I_{dc}/h_m$ is obtained as $dF_x = B_y(x = \tau_m/2, y) \times J_m L dy$. Therefore, the thrust is given by [2]

$$F_x = 2 \times 2 \times L \times J_m \int_{0}^{h_m} B_y(x = \tau_m/2, y) dy \quad (21)$$

where $B_y(x = \tau_m/2, y)$ is obtained from (20). Another way of calculating the thrust is based on a conventional d-q electrical model of the machine in a synchronously rotating reference frame. In this model, the flux distribution in air-gap—which is only produced by electromagnetic poles—is assumed to be sinusoidal and the magnetic saturation is not considered. The motor thrust is then obtained as [35]

$$F_{x(av)} = \frac{3}{2} \frac{\pi}{\tau_p} [\lambda_{dc-excitation} + (I_d - L_q) \times i_d] \times i_q \quad (22)$$

where $\lambda_{dc-excitation}$ is the flux linkage produced by the dc excitation in a pole pitch, $L_q$ and $L_q$ are the d- and q-axis mutual inducances, respectively, and $i_d$ and $i_q$ are the d- and q-axis stator current components respectively. In the motor considered in this paper $L_d$ is not equal to $L_q$, but for $i_d = 0$ the second term of thrust vanishes, so (22) is simplified to

$$F_{x(av)} = \frac{3}{2} \frac{\pi}{\tau_p} \lambda_{dc-excitation} i_q = 3I_s \times (B_{1g} L_{Np} h_{w1}) \quad (23)$$

where $B_{1g}$ is the maximum of fundamental harmonic of field in the air gap produced by the electromagnetic poles which is obtained from (14), and $I_s$ is the maximum of primary current.
The normal force of the motor obtained from virtual work principle is given by [36]

\[ F_n = \frac{1}{2\mu_0} B^2 \rho L \tau_p \]  
(24)

where

\[ B^2 = B_{1a}^2 + B_{1p}^2 + 2B_{1a} B_{1p} \cos \beta_0 \]  
(25)

and \( B_{1p} \) and \( B_{1a} \) are the maximum value of the first harmonic of the normal fields produced by the primary and secondary currents obtained from (20) and (14), respectively, and is the angle between these two fields.

VI. COMPARISON AND EVALUATION

In this section, the results of field and force calculations of the three models—the proposed model, the base model, and a 2-D FEM—are presented and compared to evaluate the proposed method. To achieve this, an 18-pole WSLSM of the type depicted in Fig. 1 is selected. The geometric parameters of the motor are listed in Table I. Considering the frequency of the power supply at 215 Hz, the motor speed will be 111 m/s or 400 km/h. Field current is 57 A and the number of turns in dc-excitation coil is 134. It means that the dc-excitation MMF source is equal to \( N_{ph} = 7635 \) A.

Fig. 4 shows the normal component of the electromagnetic field in a pole pitch of the motor due to the dc excitation obtained by 2-D FEM and the proposed layer model. The flux density under the electromagnetic pole is almost smooth. Three deeps in a pole pitch are due to the three slots of the primary in those places. The big magnitude of field in the middle of the pole causes a bigger slot effect there. A good agreement between the FEM results and the analytical results is seen. This is mainly due to the fact that different field harmonics are taken into account by the proposed method. As a result, the approximations used in the analysis are justified and the effectiveness of the proposed method is confirmed. Fig. 5 shows the distribution of electromagnetic field under a pole pair of the slotless motor. In this case, the agreement between the analytical and the FEM results is even closer due to the lack of slot effects.

Figs. 6, 7 show two-dimensional flux lines in the middle and at the end of the slotted motor obtained by FEM. Figs. 8, 9 show flux lines in the middle and at the end of the slotless motor respectively produced by dc excitation. The substantial differences seen between the flux lines of Figs. 6, 7 and Figs. 8, 9 are because of the primary teeth. In fact, the flux lines are more inclined to path through the teeth and therefore the flux density of teeth in the slotted primary is much more than the one of the slotless primary. For both cases, the magnitude of field is the same under every pole in the middle of the motor. However, this magnitude substantially varies in the two cases at the end of the motor. This is due to the fact that a leakage flux is generated at the end-part because the magnetic circuit of the machine is open. It can be corrected by a compensating winding which is not included in the motor.

The current of phase A of the primary is considered as \( I_a = 1200 \times \cos(\omega t) \). Figs. 10–15 show field distributions and flux lines produced by the primary currents at two different instances: \( \omega t = \pi /2 \) and \( \omega t = 0 \), respectively. The initial instance of \( \omega t = 0 \) corresponds to \( \beta_0 = 9 \) or \( i_a = 0 \) and \( \omega t = \pi /2 \) corresponds to \( \beta_0 = \), as it is clear in Figs. 10, 11.
For both instances, flux lines are repetitive and the same in the middle of the motor; but they differ considerably at the end of the motor due to the linear topology of the machine and its long primary. As the primary currents exist outside the secondary region, the flux lines produced by these currents also exist outside the secondary region where they have to close their path through the air.

For \( \omega t = \pi/2 \), the primary phase currents are \( I_a = 0 \) and \( I_b = -I_c = 1030 \) A. Because of zero current of phase A, any flux lines does not circle this phase. For \( \omega t = 0 \), currents are \( I_a = 1200 \) A and \( I_b = I_c = -600 \) A. The same currents of phases B and C cause the special flux lines shown in Figs. 14, 15. For both cases, normal and horizontal fields are affected by the primary slots. Under every slot the horizontal fields have a deep [Author: please finish this sentence].

Now the first harmonic of the normal and horizontal fields of the primary and secondary obtained by 2-D FEM are compared with the results of the analytical models of the machine. Figs. 16–18 show respectively, normal and horizontal components of electromagnetic fields produced by the primary currents and the dc excitation obtained by 2-D FEM and the analytical
models. It is seen that the main field in the air gap is produced by the dc excitation of the electromagnetic poles.

Comparing the first harmonic of dc excitation field obtained by both analytical models with the one obtained by FEM, as presented in Fig. 18, the proposed model is confirmed to be more accurate than the base model. Another merit of the proposed model in comparison with the base model is the calculation of all harmonics of field produced by electromagnetic poles. Therefore, the proposed model can be used to analyze thrust ripples. However, the base model only calculates the main harmonic of field produced by the primary currents besides the main harmonic of field produced by poles in \( x \) and \( y \) directions. Therefore, the proposed model can be regarded as a basis for the analysis, design and optimization purposes of a WSLSM.

Also, the normal force and thrust of the motor in the airgap calculated by the proposed model are compared with the FEM results. An average thrust obtained by the 2-D FEM is \( F(x) = 17.9 \text{ kN} \). In comparison, calculating \( B_{1y} \) from the proposed model and using (23), the average thrust is obtained as \( F(x) = 18.2 \text{ kN} \). Using (21) where \( B_y \) is produced solely by the primary currents, the thrust is obtained by the base model as \( F(x) = 24.3 \text{ kN} \). It is seen that the thrust calculated from the proposed model is closer to the FEM result. Fig. 19 shows the thrust and its average in a pole pitch by analytical and FEM methods and Fig. 20 shows these data for the normal force. It is seen that both thrust and the normal force are affected by the primary slots. Each curve has a deep under every slot. Figs. 19 and 20 are obtained by considering displacement due to the motor movement as a time function in a certain time interval. Accordingly, the stator currents change in the time interval in a way that \( i_d = 0 \). Fig. 21 shows the normal force obtained by the proposed method using (24) and by the 2-D FEM versus angle between the first harmonic of normal fields of the primary and secondary. This angle is a function of distance between these two fields and pole pitch. Depending on the secondary position, the place of the first harmonic of the secondary normal field is determined. Also, depending on the primary current, the place of the first harmonic of the primary normal field is obtained. So, the distance between peaks of these two fields can be calculated and shown in terms of an angle. If this distance is equal to one pole pitch, the angle will be \( \pi \) radians. As it is obtained from these three figures, the error in the normal force is a little bigger than the thrust error. In general the results presented above, confirm that the proposed layer model is appropriate for the analyses and design optimization of WSLSMs.

VII. CONCLUSION

The main contribution of the paper is the analytical—rather than the numerical—modeling of a wound secondary linear synchronous motor instead of a machine with a coreless or a permanent-magnet secondary reported in the literature. Therefore, the iron core of the secondary is taken into account. A layer model for the motor was presented. Magnetic field, thrust and normal force calculations were carried out using the model. A series of rather accurate expressions were obtained for the machine char-
In this paper, a 2-D FEM model of the motor with the assumption of an infinite motor width is used. Such a model may result in errors due to the neglect of the edge effect caused by a large air gap. However, since the motor width in high-speed applications is much larger than the air gap, the assumption of an infinite width seems acceptable and the edge effect is rather small. However, for more accurate modeling of the motor, a 3-D FEM based modeling and experimental evaluation may be regarded useful.

Appendix A: Details of 3-D FEM Model

Characteristics. A second model was also developed in this paper for WSLSMs as a basis of comparison by improving an existing model. Both models plus a FEM based model were applied to a slotted and a slotless primary WSLSM and their results were compared. It is shown that the motor characteristics obtained by the proposed model are closer to the FEM results. This is mainly due to the fact that different field harmonics are taken into account by the proposed method. The proposed model also provides a more accurate distribution for the first harmonic of dc excitation.

It is shown that the normal and horizontal fields are affected by the primary slots causing inaccuracy in the modeling. As a result, the proposed method can model slotless machines more accurately. Also, it is confirmed that the flux density of a primary tooth in the slotted primary is much more than the one at the primary of the slotless machine.

The proposed method can be regarded as a time saving and flexible modeling basis for the analysis, design, and optimization of different types of iron core linear synchronous motors. However, for more accurate analysis and design of the motor, it is necessary to consider the iron saturation effect. Also, a fine motor design to decide on the motor details, e.g., slot shape, needs accurate modeling methods like finite-element methods which are time consuming. In any case, the proposed method provides a suitable first step modeling to determine many motor parameters/characteristics with acceptable accuracy before focusing on less important details.

References


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