DESIGN ASPECTS OF BEARINGLESS SLICE MOTORS

Siegfried Silber, Wolfgang Amrhein,
LCM - Linz Center of Competence in Mechatronics
Johannes Kepler University Linz
A-4040 Linz, Austria
siegfried.silber@jku.at

Pascal Bösch, Reto Schöb, Natale Barletta
Levitronix GmbH
CH-8005 Zurich, Switzerland

Abstract

The complexity of a bearingless motor offers a multitude of freedom in constructive design because different motor setups may even have nearly the same dynamic behavior and may differ only slightly in some operation characteristic. However, the mechanical setup often has a very strong impact on the power inverter topology and the volt-ampere requirements and, thus, affects the system costs significantly.

This paper focuses on design aspects of bearingless slice motors with permanent magnet excitation and presents a methodical evaluation approach based on performance indexes.

Introduction

During the last few years several industrial products that are based on bearingless motor technology have successfully been launched on the market [5], [6]. Today development of bearingless motors has reached a state where integration of this technology into mass products becomes feasible. However, for high volume products commercial aspects become almost as important as technical ones. Therefore, cost-effective bearingless motor designs are absolutely required.

A bearingless motor design that offers advantages in regard to the costs of the mechanical setup is the bearingless slice motor [1]. This motor is characterised by a disk-shaped rotor so that three degrees of freedom are stabilised passively by means of the reluctance forces of the permanent magnet disk. Owing to the rotor design the bearingless slice motor is well suited for applications that do not demand high stiffness in axial direction.

In contrast to conventional motors the mechanical setup of a bearingless motor offers much more freedom in constructive design. In some cases the design parameters substantially influence the characteristics as well as the system costs of a bearingless drive. For this reason it is obvious that different applications require different designs. To find the most suitable motor design for a certain application, several performance indexes are introduced. These performance indexes relate to the overall system comprising both the bearingless motor and the power converter.

Analytical force and torque model

For a concept study, normally, many different constructions of bearingless motors have to be evaluated. A methodical evaluation approach should preferably be based on suitable performance indexes and an estimate of the production costs. An additional requirement for the evaluation process is that it generally has to be carried out within a short period of time. For this reason complex calculation methods like FEM analysis are not useful for deriving the performance indexes required. An alternative to numerical methods are analytical ones that offer sufficient accuracy and need only a fractional amount of time.

The force and torque model of a bearingless motor provides the basis for the performance indexes that are presented in this paper. A very general approach to an analytical force and torque model can be derived from the MAXWELL STRESS TENSOR [10]

$$\mathbf{T}_M = \mu \left[\begin{array}{ccc} H_x^2 - \frac{1}{2}H^2 & H_x H_y & H_x H_z \\ H_y H_x & H_y^2 - \frac{1}{2}H^2 & H_y H_z \\ H_z H_x & H_z H_y & H_z^2 - \frac{1}{2}H^2 \end{array} \right],$$

where $\mathbf{T}_M$ denotes the Maxwell stress tensor, $\mathbf{H}$ is the magnetic field intensity and μ is the permeability. The mechanical stress σ that acts on a surface element of the stator can then be calculated with

$$\sigma = \mathbf{T}_M \mathbf{e}_n, \tag{1}$$

where $\mathbf{e}_n$ represents the vector perpendicular to the stator surface. Taking into account that the permeability of the ferromagnetic stator is much higher than the permeability of air, the tangential component of flux density in the air gap can be neglected and (1) can be evaluated to

$$\sigma = \left[\begin{array}{c} \dfrac{B_{1n}^2}{2\mu_0} \\ B_{1n} A_s \\ 0 \end{array} \right],$$

where B_{1n} is the normal component of flux the density in the air gap and A_s is the current density distribution on the stator surface. The force acting on the rotor of the bearingless motor is determined by the surface integral

$$\mathbf{F} = \oint_f \sigma df \ ,$$

where f is referred to as the area of the stator surface. When both the current density distribution on the stator surface A_s and the normal component of the flux density in the air gap B_{1n} are expressed in terms of the stator currents $\mathbf{i}_1$ and the flux density of the permanent magnet, the simplified force and torque model results in [8]

$$\begin{aligned} \mathbf{F}_r &= \left[\begin{array}{cc} \mathbf{i}_1^{\mathrm{T}} & \mathbf{0} \\ \mathbf{0} & \mathbf{i}_1^{\mathrm{T}} \end{array} \right] \left[\begin{array}{c} \mathbf{M}_{Qx}(\mathbf{x}_r,\varphi) \\ \mathbf{M}_{Qy}(\mathbf{x}_r,\varphi) \end{array} \right] \mathbf{i}_1 \qquad (2) \\ &\quad + \mathbf{M}_L(\mathbf{x}_r,\varphi)\mathbf{i}_1 + \mathbf{M}_C(\mathbf{x}_r,\varphi) \\ T_r &= \mathbf{i}_1^{\mathrm{T}} \mathbf{N}_Q(\mathbf{x}_r,\varphi)\mathbf{i}_1 + \mathbf{N}_L(\mathbf{x}_r,\varphi)\mathbf{i}_1 \ . \qquad (3) \end{aligned}$$

$\mathbf{F}_r$ and T_r are the vector of the levitation forces in the stator coordinate system and the electromagnetic torque, respectively. The stator currents of the m phase stator winding system are represented by

$$\mathbf{i}_1 = \left[\begin{array}{cccc} i_1 & i_2 & \cdots & i_m \end{array} \right]^{\mathrm{T}} .$$

The rotor position in radial direction is written as

$$\mathbf{x}_r = \left[\begin{array}{c} x_r \\ y_r \end{array} \right]$$

and the rotor angle is denoted by φ. When the investigation is restricted to bearingless motors that employ rare earth permanent magnets, the elements of the matrices $\mathbf{M}_{Qx}$, $\mathbf{M}_{Qy}$ and $\mathbf{N}_Q$ become small in comparison with the elements of the other matrices. Therefore, these terms can be neglected and the levitation force and torque model is simplified to

$$\mathbf{F}_r = \mathbf{M}_L(\mathbf{x}_r,\varphi)\,\mathbf{i}_1 + \mathbf{M}_C(\mathbf{x}_r,\varphi) \tag{4}$$

$$T_r = \mathbf{N}_L(\mathbf{x}_r,\varphi)\,\mathbf{i}_1 \ . \tag{5}$$

Normally, the matrix $\mathbf{M}_C$ in equation (4) becomes zero when the rotor is in the centred position. In this case equations (4) and (5) can be combined into one expression

$$\mathbf{Q} = \mathbf{T}_m \mathbf{i}_1, \tag{6}$$

with

$$\mathbf{Q} = \left[\begin{array}{c} F_{rx} \\ F_{ry} \\ T_r \end{array} \right], \qquad \mathbf{T}_m = \left[\begin{array}{c} \mathbf{M}_L(\mathbf{x}_r,\varphi) \\ \mathbf{N}_L(\mathbf{x}_r,\varphi) \end{array} \right] .$$

For further simplification it can be assumed that the position of the rotor is stabilised by means of an appropriate position controller. Accordingly, it can be supposed that it stays in the centred position and therefore the matrix $\mathbf{T}_m$ can be linearised about the operating point. As a result, $\mathbf{T}_m$ is only a function of the rotor angle φ

$$\mathbf{T}_m(\varphi) = \left[\begin{array}{c} \mathbf{M}_L(\varphi) \\ \mathbf{N}_L(\varphi) \end{array} \right] .$$

For the operation of a bearingless motor in a closed control loop it is necessary to supply the motor phases with such currents that desired radial forces and torque are generated. In a mathematical sense this requires that equation (6) has to be solved for the phase currents $\mathbf{i}_1$. However, a unique solution of the form

$$\mathbf{i}_1 = \mathbf{T}_m^{-1} \mathbf{Q}$$

can only be found if the bearingless motor has three phases. Unfortunately, the bearingless motors that are under consideration in this paper have more than three phases. This means that any desired radial force and torque might then be realised by many different sets of phase currents. Since many solutions might be possible it is the task to find a set of currents that generates the required radial forces and torque in the best way. One feasible solution of finding the best set of currents is to minimise the resistive power losses. This leads to the following optimisation problem:

$$\min_{\mathbf{i}_1} \mathbf{i}_1^{\mathrm{T}} \mathbf{R}_1 \mathbf{i}_1$$

subject to

$$\mathbf{T}_m \mathbf{i}_1 - \mathbf{Q} = \mathbf{0} \ ,$$

where $\mathbf{R}_1$ is the matrix of the phase resistances. The solution to this optimisation problem yields [9]

$$\mathbf{i}_1 = \mathbf{K}_m(\varphi)\mathbf{Q} \ , \qquad (7)$$

with the decoupling matrix

$$\mathbf{K}_m(\varphi) = \mathbf{T}_m^{\mathrm{T}} \left(\mathbf{T}_m \mathbf{T}_m^{T}\right)^{-1} .$$

For some of the motor conceptions that are considered in this paper star connection of the motor phases leads to a simplified inverter topology. For these types of bearingless motors the sum of the phase currents is restricted to zero for any mode of operation. To meet this additional requirement it is possible to impose an extra constraint on the optimisation problem of the form

$$\begin{bmatrix} 1 & 1 & \cdots & 1 \end{bmatrix} \mathbf{i}_1 = \mathbf{1}^{\mathrm{T}} \mathbf{i}_1 = 0$$

to force the sum of the phase currents to zero. Solving the optimisation problem leads to the following decoupling matrix

$$\mathbf{K}_m(\varphi) = \begin{array}{l} \left(\mathbf{T}_m^{\mathrm{T}} - \dfrac{1}{m}\mathbf{1}\mathbf{1}^{\mathrm{T}}\mathbf{T}_m^{\mathrm{T}}\right) \cdot \\ \cdot \left(\mathbf{T}_m \left(\mathbf{T}_m^{\mathrm{T}} - \dfrac{1}{m}\mathbf{1}\mathbf{1}^{\mathrm{T}}\mathbf{T}_m^{\mathrm{T}}\right)\right)^{-1} . \end{array}$$

Performance indexes

The freedom in constructive design has brought about many new conceptions of bearingless PM motors recently (e.g. [2]). In many cases, the required torque and bearing forces can be generated by motor designs that differ in the winding configurations, the number of poles, and even in the number of phases. Furthermore, the motor design also has an impact on the inverter topology. The evaluation of the different motor conceptions is hardly feasible without

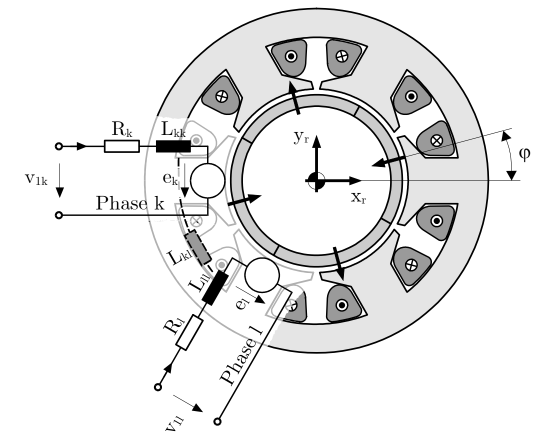

FIGURE 1: Equivalent circuit of a polyphase bearingless motor

the availability of appropriate performance figures or performance indexes. For this reason some numerical values are introduced in this section that enable the design engineer to find the most suitable motor construction for a certain application.

A measure that is closely associated with the efficiency of the motor is the resistive loss in the windings. The instantaneous value of the resistive power is obtained as

$$p_L = \mathbf{i}_1^{\mathrm{T}} \mathbf{R}_1 \mathbf{i}_1 .$$

Substituting Eq. (7) into the above equation gives:

$$p_L = \mathbf{Q}^{\mathrm{T}} \mathbf{K}_m^{\mathrm{T}}(\varphi) \mathbf{R}_1 \mathbf{K}_m(\varphi) \mathbf{Q} \ . \qquad (8)$$

Assuming that the demanded bearing forces and torque are constant - this is the case at steady-state operation of the bearingless motor - the instantaneous resistive power is only a function of the rotor angle. For evaluation purposes, however, the average resistive power at steady-state operation is a more meaningful value than the instantaneous power. This figure can be calculated from (8) as

$$P_L = \frac{1}{2\pi} \int_0^{2\pi} \mathbf{Q}^{\mathrm{T}} \mathbf{K}_m^{\mathrm{T}}(\varphi) \mathbf{R}_1 \mathbf{K}_m(\varphi) \mathbf{Q} \, d\varphi \ . \qquad (9)$$

Another very useful figure for the comparison of different motor conceptions is the peak phase voltage that is required for a specified load condition. According to the equivalent circuit of the bearingless motor, shown in Fig. 1, the phase voltage is given as:

$$\mathbf{v}_1 = \mathbf{R}_1 \mathbf{i}_1 + \mathbf{T}_m^{\mathrm{T}}(\varphi) \begin{bmatrix} \dot{\mathbf{x}}_r \\ \omega \end{bmatrix} + \mathbf{L}_1 \dot{\mathbf{i}}_1 \ , \qquad (10)$$

where $\dot{\mathbf{x}}_r$ denotes the velocity of the rotor in radial direction, ω is the rotational speed, and $\mathbf{L}_1$ is the inductance matrix. In this equation the three terms on the right-hand side represent the resistive voltage drop, induced emf and inductive voltage drop, respectively. With (7) the time derivative of the phase currents results in:

$$\begin{aligned} \dot{\mathbf{i}}_1 &= \frac{d\left(\mathbf{K}_m(\varphi)\mathbf{Q}\right)}{dt} \\ \dot{\mathbf{i}}_1 &= \frac{\partial \mathbf{K}_m(\varphi)}{\partial \varphi}\omega \mathbf{Q} + \mathbf{K}_m(\varphi)\frac{d\mathbf{Q}}{dt} \ . \end{aligned} \tag{11}$$

A simplified expression for the phase voltages can be found when the resistive voltage drop $\mathbf{R}_1\mathbf{i}_1$ is neglected and when it is assumed that the velocity of the rotor in radial direction is $\dot{\mathbf{x}}_r$ small in magnitude. Moreover, when the phase voltages are calculated at a steady-state operating point the generalised force vector $\mathbf{Q}$ is constant. Thus, the time derivative term of $\mathbf{Q}$ in Eq. (11) vanishes and the vector of the phase voltages, then, is

$$\mathbf{v}_1 = \mathbf{T}_m^{\mathrm{T}}(\varphi)\begin{bmatrix} \mathbf{0} \\ \omega \end{bmatrix} + \mathbf{L}_1\frac{\partial \mathbf{K}_m}{\partial \varphi}\omega \mathbf{Q} \ . \tag{12}$$

The peak value of the phase voltage is a very helpful figure for the required DC-link voltage of the inverter. However, it has to be taken into consideration that for dynamic operation a higher DC-link voltage is required. This additional voltage is mainly affected by the required dynamics of the bearing part of the bearingless motor. A detailed derivation of the voltage requirement for dynamic operation of magnetic bearings can be found in [7].

For the selection of the power semiconductors the volt-ampere requirement of the inverter is an important figure. Analyses of the VA rating for different types of motors were made by several authors (e.g. [4]) in the past. Different definitions for the VA ratings were established, such as the inverter peak VA rating and the VA rating in terms of inverter peak voltage and rms current. In the context of bearingless motors which are mainly used at higher rotational speed, the VA rating in terms of inverter peak voltage and rms current is more meaningful. Because power electronic silicon die sizing is based on thermal considerations and device losses rather than on peak currents. Mathematically, the VA rating of the inverter can be written as:

$$P_{VA} = \mathbf{V}_{1peak}^{\mathrm{T}}\mathbf{I}_{1rms} \ , \tag{13}$$

where $\mathbf{V}_{1peak}$ is the vector of the peak phase voltages that can simply be evaluated from Eq. (12) and $\mathbf{I}_{1rms}$ is the vector of the rms currents resulting from Eq. (7).

Assessment of bearingless motor designs

In this section, different embodiments of bearingless slice motors should be assessed with regard to the performance indexes presented in this paper. The target application which the bearingless motor should be employed in is specified by the maximum rotational speed, the rated torque, and the maximum bearing force. For the design of the different types of bearingless motors the following assumptions are taken:

- Rotor diameters and the stator bores are the same for all motor designs.
- The volume of the magnetic material is the same.
- Current density in the windings is equal at rated torque.
- Flux density in the stator core caused by the permanent magnets is the same.

The bearingless motors that are considered are shown in Fig. 2 and should be referred to as follows:

(a) Bearingless motor with four concentrated coils and four-pole PM rotor.

(b) Bearingless motor with five concentrated coils and two-pole PM rotor.

(c) Bearingless motor with six concentrated coils and two-pole PM rotor.

(d) Same as (c), but with separate bearing and motor windings.

It is the characteristic of the motor embodiments (a) to (c) that only concentrated coils are used which contribute to force and torque generation at the same time. In contrast, motor (d) features a separate winding system for force and torque generation. The corresponding inverter topologies for the different motor embodiments are shown in Fig. 4.

The scaled resistive power loss of the different motor embodiments is depicted in Fig. 3 for rated output power and as a function of the bearing forces required. The scaling factor that is used in this figure is the resistive power loss of motor (a) at rated power and when no bearing forces are demanded.

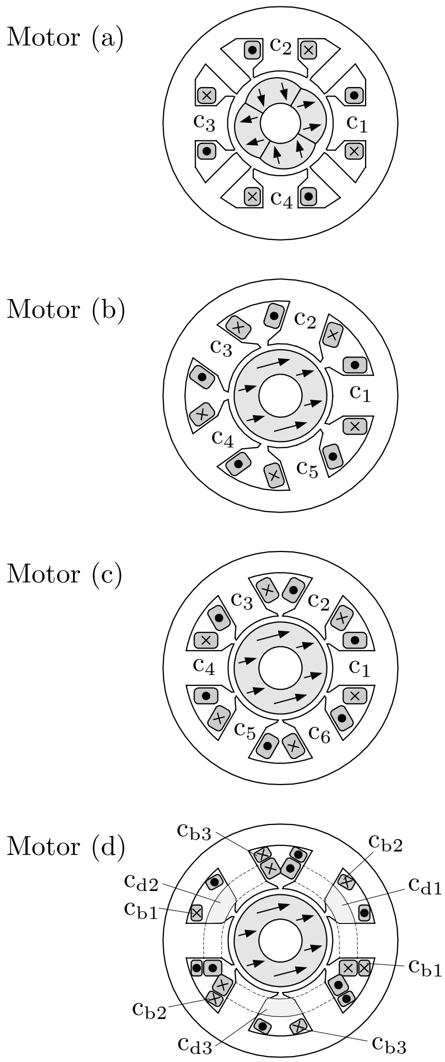

FIGURE 2: Motor designs

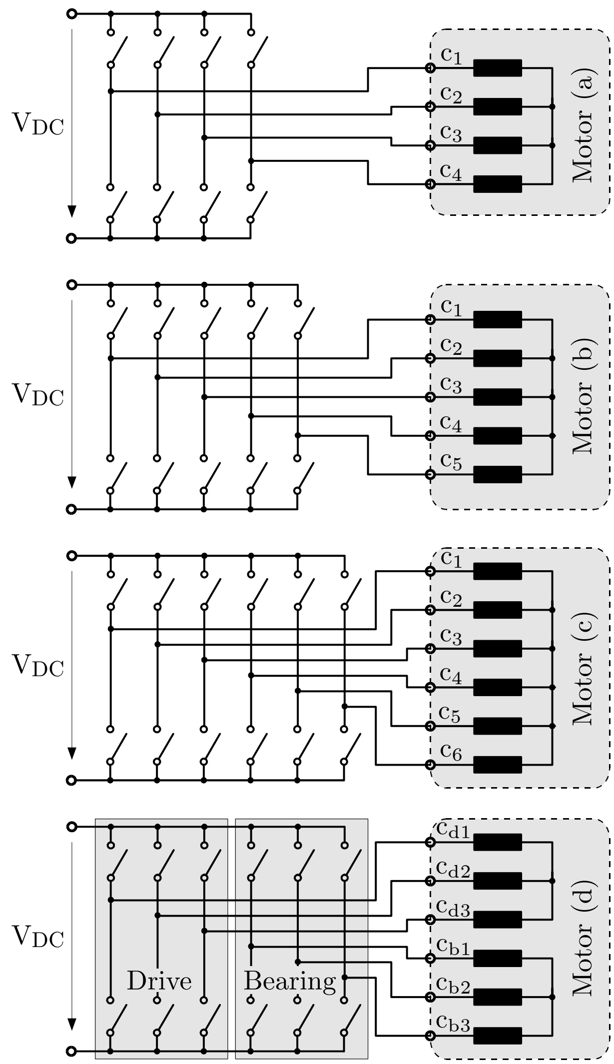

FIGURE 4: Inverter topologies

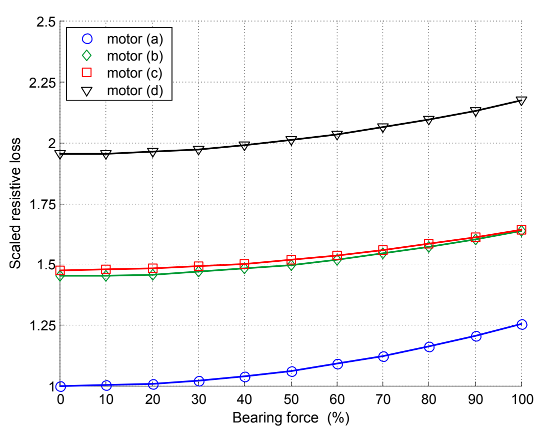

FIGURE 3: Scaled losses

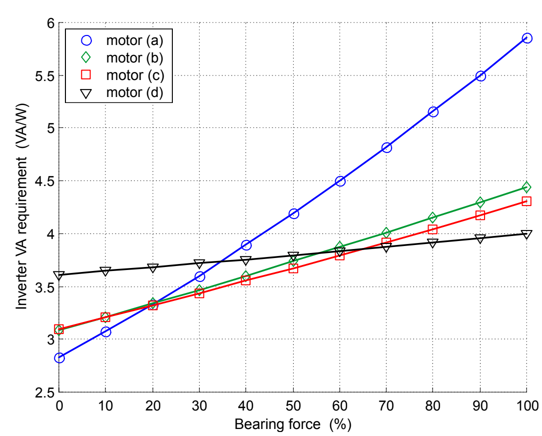

FIGURE 5: Inverter VA requirement

	Motor (a)	**Motor (b)**	**Motor (c)**	**Motor (d)**
Copper mass	233 g	414 g	421 g	456 g
Number of half bridges	4	5	6	6

TABLE 1: Figures that affect the system cost

The VA requirement in terms of peak phase voltage and rms current scaled to the rated mechanical power of the motor is shown in Fig. 5 for the different motor designs. In this figure the efficiency of the motor is not taken into account, and additional phase voltages for dynamic operation are neglected. Since the peak phase voltage is contained in the VA requirements this performance index is not considered separately.

When the bearingless motor should be used for mass products commercial aspects become almost as important as technical ones. For this reason two additional figures that have a strong impact on the ststem cost are given in Table 1. The first is the mass of copper needed for a specific motor setup. The second concerns the inverter, namely the number of half bridges of the inverter.

Conclusion

By means of the performance indexes which are defined in this paper, it becomes feasible to find a design of a bearingless motor that best meets the requirements of a certain application. Additionally, it is even possible to find the most cost-effective motor design that just suffices for the application. However, application of the performance indexes is only possible when the required forces and the rated torque are exactly known. This requires for in-depth knowledge of the application which the bearingless motor should be employed in.

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