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Fast Models to Predict the Stray Magnetic Field Created by Faulty Electrical Machines

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Abstract—The analysis of the magnetic stray field created by electrical machine is now known as a powerful method for fault monitoring. This paper proposes two approaches to compute magnetic stray field created by faulty electrical machine. The first one presents a homogenized FEM method. The second one is based on a combination of an analytical expression of the magnetic field in the air gap with an integral method. Both models lead to accurate results while saving significantly CPU time and memory and highly simplify the machine description. Then, it is possible to use them to model electrical machines in faulty operation modes and to compute the radiated magnetic field.

I. INTRODUCTION

The computation of magnetic stray field created by electrical machines concerns lots of potential applications (magnetic silencing for military applications [1], fault monitoring with magnetic sensors [2], EMC,...). However, this task is not without complexity, the main difficulties being the need of a 3D model with a ratio of 10^6 between the internal and the external magnetic fields. An accurate formulation has been proposed in [3]. It is based on a FEM computation with a magnetic moments method in post-processing. A first magnetostatic scalar FEM resolution is achieved. Once the scalar potential is obtained in the whole machine, the magnetization \mathbf{M} in ferromagnetic parts V_{mag} is computed. Then it is possible to compute the stray field thanks to a volume integral equation which links the magnetization \mathbf{M} to the outside magnetic induction \mathbf{B} .

$$\mathbf{B} = \frac{\mu_0}{4\pi} \iiint_{V_{\text{mag}}} \left(3 \frac{\mathbf{M} \cdot \mathbf{r}}{r^5} \mathbf{r} - \frac{\mathbf{M}}{r^3} \right) dV + \mathbf{B}_0 \quad (1)$$

where \mathbf{r} is the distance between the point where the magnetic induction is computed and the integration point, \mathbf{B}_0 is the induction created by winding. This integral post-processing approach is called moment method has been successfully validated by measurements in [3].

However, this kind of 3D FEM modeling is very memory consuming. In the case of faulty machines, geometry or physical symmetries are lost. Therefore, the number of unknowns becomes too much significant to allow the modeling of the whole machine. The aim of this paper is to propose simplified models to evaluate the influence of faulty operation modes on the stray magnetic field of a synchronous machine, while keeping an acceptable number of unknowns.

In this paper, two models are proposed. The first part presents a homogenized FEM approach. This method is already known [4] but is originally adapted to the stray field computation. The second one is an original use of a classical integral method. Both models are accurate to compute stray field created by healthy electrical machine and can be used to model faulty operation mode too.

II. HOMOGENIZED FEM MODEL

The main assumption of the first proposed approach lies in an idealized distribution of induction flux inside the machine. The fields lines are tangential (according to a circular way) in the yoke and become radial in a zone located in the vicinity of the air gap and delimited, on both sides, by the end of the slots of the stator and the rotor. In the absence of a significant induced reaction (during no-load operation mode), this assumption is realistic because it does not modify the total reluctance associated inductor flux.

To schematize the behavior of the induction flux in teeth, let us consider the figure 1, in which B_t and B_s are average induction values respectively in the teeth and the slots.

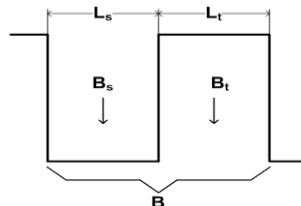


Fig. 1. Simplified tooth and slot and associated induction in the electrical machine

Average induction \mathbf{B} on this tooth and slot area is given by the expression:

$$B(H) = \frac{B_t(H) \times L_t + B_s(H) \times L_s}{L_t + L_s} \quad (2)$$

where L_t and L_s present the widths of tooth and slot. Let us suppose that the average values of the excitation field in the iron and in the vacuum are both equals to the given value \mathbf{H} .

Then, if the characteristic $B_t(H)$ of the magnetic materials is known, it is possible for us to replace the tooth and slot area by an equivalent homogenized continuous media (see fig. 2) whose characteristic is given by the following formula:

$$B(H) = \frac{B_t(H) \times L_t + \mu_0 \times H \times L_s}{L_t + L_s} \quad (3)$$

When materials law is linear, equation (2) leads to a homogenized equivalent permeability for each area (tooth and slot). Moreover, as the magnetic flux density is mainly forced to remain radial, the permeability is then anisotropic. The radial μ_1 and vertical μ_3 components of this permeability are thus:

$$\mu_1 = \mu_3 = \frac{\mu_r \times L_t + L_s}{L_t + L_s} \quad (4)$$

where μ_r is the permeability of the ferromagnetic material.

The tangential component is:

$$\mu_2 = \frac{L_t + L_s}{L_s} \quad (5)$$

As the consequence, with this approach, we only need a coarse mesh in each area of the tooth and slot, in comparison with standard FEM modeling where to take into account the real slot geometry leads to a very fine mesh. Thus, the number of unknowns is largely decreased, which allows to model the complete motor geometry without symmetry. In particular, several default operation modes can be modeled rotor short-circuit, for example. However, this kind of 3D modeling is still difficult and heavy to manage (geometry description and mesh, need of CPU memory, problem of cancellation error). Moreover, some defaults operation modes, like eccentricity, are still difficult to manage with a 3D FEM model. This is why, we propose in this paper a simpler model.

III. 3D “AIR GAP” SURFACE MODEL

Let us consider an electric machine made of windings, an air gap, a stator and rotor. In a first assumption, we consider that the material law is linear. Maybe, this assumption seems to be a little bit strong, but it must be pointed out that the presented approach is original and very simple compared to the complexity of the problem. This model is a first step and it could be generalized thereafter to take into account non-linear material law.

Under the assumption material linearity, the stray magnetic field is created by [5]:

- Inductor coils in which currents flow. The field can easily computed by Biot and Savart law.
- Charge surface density $\mathbf{M} \cdot \mathbf{n}$ located on surfaces which delimits the interface between ferromagnetic material and air regions (\mathbf{M} is the magnetization of the material and \mathbf{n} the external normal going from the material to the air).

In an electrical machine, three main surfaces must be considered. The first one is the external boundary (air / stator), and the two others deals with the air gap (air gap / stator and air gap / rotor) which are closed to each other (fig. 2) [6].

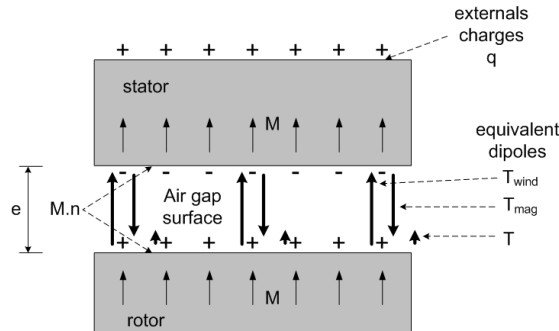


Fig. 2. Air gap surface with the dipoles and externals charges localization.

These two last surfaces carry nearly same charge distribution with opposite signs (the induction in the air gap

is mainly radial and continuous so associated magnetizations are the same and both normal have opposite directions). If we neglect the slot geometry, both surfaces can be replaced by an average one of the air gap carrying a normal distribution of dipoles \mathbf{T}_{mag} .

In most part of electrical machine, this dipole distribution is easy to determine, the field \mathbf{B} in the air gap being well-known and directly linked to \mathbf{M} . Many analytical harmonic expressions already exist which depend, for example, on the pole number and on the maximum induction in the air gap. So, for a 2p-pole electrical machine with a sinusoidal air gap induction, this distribution created by the ferromagnetic material can be represented:

$$\mathbf{T}_{\text{mag}}(\theta, t) = e \times \frac{\mu_r - 1}{\mu_0 \mu_r} \times \mathbf{B}_{\text{airgap_max}} \times \cos(p\theta - \omega_s t) \quad (6)$$

where e is the thickness of the air gap, μ_r is the relative permeability of the material, θ is the mechanical position of the rotor and ω_s is the electrical pulsation and t is the time.

Let us now consider the rotor windings. The field created by all the windings can be represented, in a first approximation, by a sinusoidal distribution of normal dipoles located on the average air gap surface [6].

$$\mathbf{T}_{\text{wind}}(\theta, t) = -A_{t_max} \times \cos(p\theta - \omega_s t) \quad (7)$$

where A_{t_max} is the maximum number of Ampere-turn for one pole. Let us notice that with such a distribution, end-winding effect is taken into account in the model.

Thus a distribution of normal dipoles \mathbf{T} is obtained on the cylindrical surface in the air gap which is the sum of \mathbf{T}_{mag} or \mathbf{T}_{wind} and takes into account both ferromagnetic and windings effects.

It remains now to determine the charge distribution on the external surface (stator / air region interface). Now, we consider the machine as a whole ferromagnetic cylinder with the internal field source normal dipoles \mathbf{T} described previously (its influence can be expressed with a simple numerical integration on the air gap surface).

Then, we can use a classical global integral equation to obtain external charges q [5].

$$\frac{\mu_r + 1}{\mu_r - 1} q(P) - \frac{1}{2\pi} \iint_S q(M) \frac{\mathbf{r} \cdot \mathbf{n}}{r^3} dS = 2(n \cdot \mathbf{B}_0(P)) \quad (8)$$

where \mathbf{r} is the vector between a point P of the surface and M the integration point and \mathbf{B}_0 is the induction created by windings and the internal normal dipoles distribution. By meshing the surface and using a point matching approach, we get a full linear system resolved with LU decomposition. Then, q distribution is obtained and the magnetic stray field can be computed.

Once the values of normal dipoles \mathbf{T} and external charges q at each time step is obtained, the magnetic stray field created on a sensor located somewhere in the air region can be calculated thanks to a relation:

$$\mathbf{B} = \frac{\mu_0}{4\pi} \iint_{S_{\text{airgap}}} \left(3 \frac{\mathbf{T} \cdot \mathbf{r}}{r^5} \mathbf{r} - \frac{\mathbf{T}}{r^3} \right) dS + \frac{\mu_0}{4\pi} \iint_{S_{\text{external}}} q \frac{\mathbf{r}}{r^3} dS \quad (9)$$

As a consequence, this model is thus easy to implement by considering two cylindrical surfaces (air gap and external surfaces) meshed into rectangular elements. The air gap one is carried out with uniform normal dipoles distribution. Note that in this model, the mesh does not have any motion. The computation of the integral (8) and (9) is provided by Gauss numerical integration method.

One of the main advantages of this approach is the simplicity of the machine description. In particular, lots of references provide analytical expression of the field in the machine air gap in faulty operation modes. For instance, [7] provides analytical harmonic expressions of the induction in air gap of electric machines for short-circuit rotor windings and rotor dynamic eccentricity operation modes.

$$\mathbf{T}(\theta, t) = k \times \cos(p\theta - \omega_s t) + \lambda_{cc} \times \cos(\theta - \omega_s t/p) \quad (10)$$

$$\mathbf{T}(\theta, t) = k \times \cos(p\theta - \omega_s t) \times (1 + \lambda_{de} \times \sin(\theta - \omega_s t/p)) \quad (11)$$

where λ_{cc} and λ_{de} represent the constants linked to the fault amplitude for the short-circuit and the eccentricity respectively.

IV. NUMERICAL RESULTS

In this section, the models are applied to a no-load operation mode modeling of a complex 4-poles synchronous machine ($\mu_r = 1000$, $I_{\text{rotor}} = 18A$, $2916At$ max). In a first step, magnetic stray field of the healthy machine is computed thanks to our both approaches and the results are compared to those obtained by our reference. This reference is a standard FEM modeling (fig. 3.a.) in Flux3D software. A post-processing magnetic

moments method [3] which is now a well-known method to compute stray magnetic field is used. Once both models are validated, different rotor faults (short-circuit windings, dynamic eccentricity) are introduced and their influence on the magnetic stray field are computed.

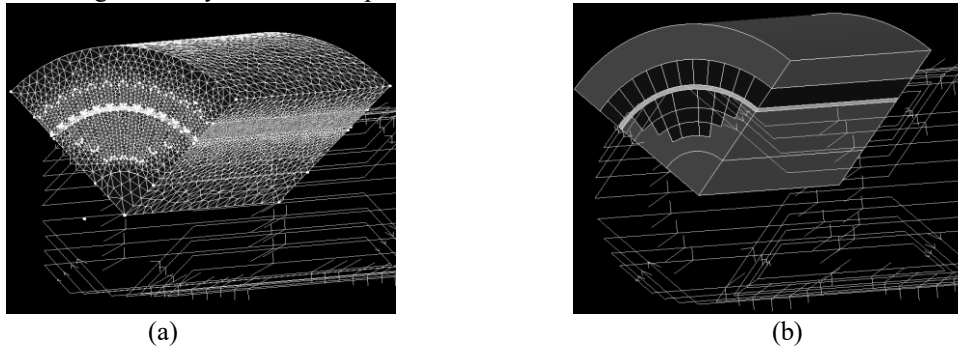


Fig. 3. a. 3D FEM model of the synchronous machine. The whole geometry (rotor and stator slots) are meshed, with second order nodes: 596945. The magnetic stray field is computed by a moment method post-processing.

Fig. 3. b. Homogenized model of the machine. Tooth and slots are presented by a homogenized equivalent material. Number of second order nodes: 180422. Different regions correspond to different permeability values and directions. The number of unknowns is divided by 3.3, thus it is now possible to model the whole machine.

The first model based on homogenized approach has been carried out (fig. 3.b.). To avoid the field singularity around conductors crossing the material, the rotor windings are set in the air gap. Because of the FE numerical noise in the air region, stray magnetic fields are also calculated by a post-processing magnetic moments method. Let us notice that the number of elements is highly reduced.

Then, we have modeled the machine with our integral approach (see fig. 4). The internal cylinder is the air gap surface with normal dipoles. The external surface represents the charge distribution obtained after the resolution of the linear system. The expression of the (6) is obtained thanks to a simple 2D modeling of the electric machine. Integral in (8) and (9) are computed with a simple gauss quadrature.

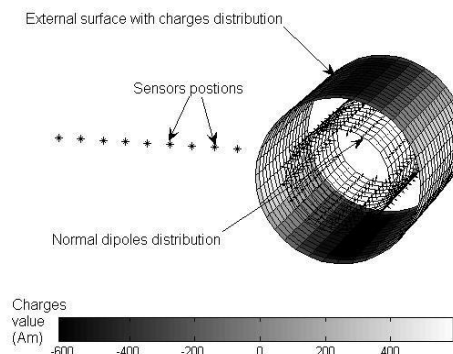


Fig. 4. Boundary integral modeling of the machine. The inner cylinder carries the normal dipoles distribution. The external charge distribution appears on the external cylinder.

The 3D homogenized model, the “air gap” surface model and the 3D real geometry model (with slots) have been compared. The stray fields obtained on a line outside the device (see fig. 5) are presented and show a very good adequacy (see fig. 5) for all modelings, which enables us to validate our approaches.

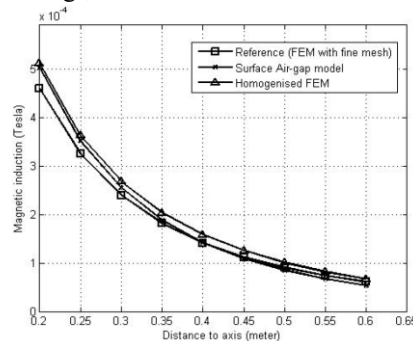


Fig. 5. Comparison between stray magnetic field obtained with our models and standard FEM.

Thanks to our these new models, faulty machines can be studied with an acceptable number of elements. The first fault considered is a one-ninth short circuit of one pole of the rotor windings. The model used is the homogenized one. By analyzing the radial component of the stray magnetic field on a magnetic sensor outside the machine, we obtain the following spectrum:

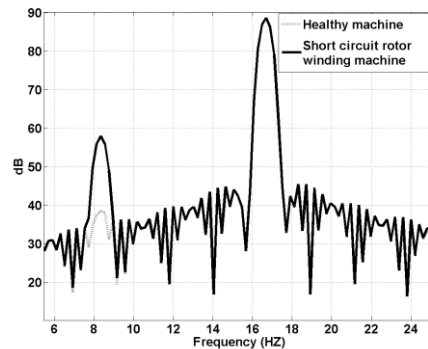


Fig. 6. Frequency spectrum of the stray magnetic field obtained by the homogenized model - Comparison between healthy machines (dashed line) and short-circuit rotor windings machine (simple line).

The stray magnetic field can be approximated by the rotation of a quadrupole (4-poles electrical machine) at the frequency 16.67 Hz. If we introduce a fault (rotor winding short circuit), the machine loses its symmetry and a new spectral component appears at the rotation speed (8.33 Hz).

The "air gap" surface model has also used to predict the magnetic stray field created by a faulty machine with an eccentricity. The spectrum is obtained below:

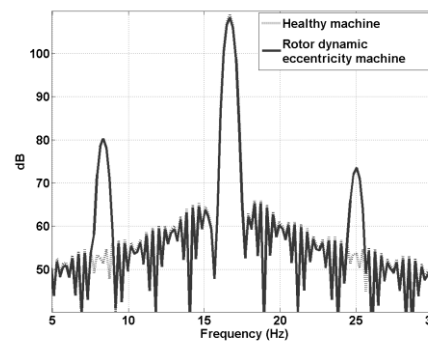


Fig. 7. Frequency spectrum of the stray magnetic field by the "air gap" surface model - Comparison between healthy machines (dashed line) and a machine with an eccentricity (simple line)..

If we introduce an eccentricity fault, the machine loses its symmetry and two new spectral components appear at 8.33 Hz and 24.99 Hz.

V. CONCLUSION

In this paper, two new approaches are presented for computing the stray magnetic field created by electrical machines. The first one presents a well-known homogenized technique, but originally used to compute 3D stray magnetic field. The second is a new, very light and easy to implement one. These two models permit us to precisely predict the frequency contents of stray magnetic field created by faulty machines and the associated amplitudes. These models can easily be used to develop new tools methods for fault monitoring.

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