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Frequency-Dependent Behavior of the Lumped Parameter Model of the Windings of Electrical Machines in Transient Simulations

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This paper presents a method to take into account the frequency-dependent behavior of the elements of lumped parameter models of electrical machine windings. First, frequency-dependent inductances, resistances and capacitances are determined thanks to time harmonic finite element (FE) computations in the frequency range of interest. Then, equivalent electrical networks are deduced using the vector-fitting (VF) algorithm, which allows for their inclusion in transient circuit solvers. The resulting time-domain voltage distributions can therefore be extracted and the whereabouts of partial discharges (PD) predicted using Paschen's law. This will enable the usage of this method as a diagnostic tool before/during the design phase of the winding of electrical machines.

Index Terms—Finite elements method (FEM), frequency-dependent parameters, transient simulations, vector-fitting.

I. INTRODUCTION

The high $\frac{dv}{dt}$ imposed by the new generation of semiconductor devices generates intense inter-turn electric fields in the winding of electric machines. These high electric fields may exceed the partial discharge inception voltage resulting in partial discharges. Occurrence of partial discharges may result in an inter-turn local fault, increasing the thermal stress electrical machines have to withstand. Correspondingly, an expansion of the fault can take place followed by an eventual premature breakdown of the machines. Hence the need of a well parameterized model of the winding of machines serving as a predictive diagnostic tool. Previous research works [1]–[4] have presented a phenomenological lumped parameter circuit (LPC) of the winding of electrical machines. However, the values of the elements of the LPC were computed at a single frequency regardless of their frequency-dependent behavior. This paper presents a method to take into account the frequency-dependency of these elements in transient simulations.

II. LUMPED PARAMETER MODEL

The phenomenological approach proposed in [1,2] enables to model each of the occurring physical phenomena within the winding of the machine using an electrically passive component. Iron losses, Joule losses (including both skin and proximity effects) and dielectric losses (both turn-to-turn and turn-to-ground) are modeled using resistances R_p , R_s , R_t , and R_m , respectively. Across each turn, the appearing electromotive force is modeled using an inductance L_p . Finally, the capacitive coupling (turn-to-turn and turn-to-ground) is modeled using capacitances C_t and C_m , respectively. The elementary cell modelling each turn is represented in Fig. 1.

III. FREQUENCY-DEPENDENT PARAMETERS

Since the input voltage is rich in harmonics, it is important to account for the frequency-dependent behavior of the elements of the elementary cell. Accordingly, the values of the resistances R_s and the inductances L_p are computed over the

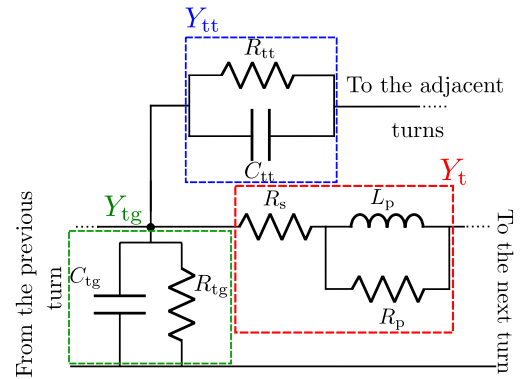


Fig. 1. The elementary cell of one turn of the winding of the machine [2]. The three blocks Y_t , Y_{tt} and Y_{tg} correspond to the self admittance of the turn, the coupling admittance of the turn to the remaining turns, and the coupling admittance of the turn to ground, respectively.

frequency range of interest in a magnetodynamic (“magneto-quasistatic”, where displacement currents are neglected) finite element approximation to consider skin and proximity effects. In the magnetodynamic approximation, the magnetic vector potential \mathbf{a} and the scalar electric potential v satisfy:

$$\mathbf{curl}(\mu^{-1}\mathbf{curl}\mathbf{a}) + \sigma\left(\mathbf{grad}v + \frac{\partial\mathbf{a}}{\partial t}\right) = 0, \quad (1)$$

where μ and σ are the magnetic permeability and conductivity, respectively.

The frequency-dependent behavior of the capacitive coupling and the dielectric losses, i.e. C_t , C_m , R_t and R_m , is taken into account using an electrostatic approximation for each frequency of interest f :

$$\mathbf{div}(-\varepsilon(f)\mathbf{grad}(v)) = 0, \quad (2)$$

where v is the scalar potential, $\varepsilon(f)$ is the dielectric permittivity of the used insulation, characterized experimentally in terms of the frequency. It is worth mentioning that the frequency-dependent behavior of the capacitive coupling is

only considered for adjacent turns. For non-adjacent turns, only the capacitance computed at $f = 0$ Hz is used.

IV. TIME-DOMAIN SIMULATION

At this stage, the elements of the LPC seen in Fig. 1 are vectors whose coordinates correspond to their values at different frequencies within the range of interest. To account for this frequency-dependent behavior in transient circuit solvers, the three admittances Y_t , Y_{tt} and Y_{tg} presented in Fig. 1 are expressed as

$$\begin{cases} Y_t(f) = \frac{1}{R_s(f) + \frac{1}{\frac{1}{jL_p(f)\omega} + R_p(f)}}, \\ Y_{tt}(f) = \frac{1}{R_{tt}(f)} + jC_{tt}(f)\omega, \\ Y_{tg}(f) = \frac{1}{R_{tg}(f)} + jC_{tg}(f)\omega, \end{cases} \quad (3)$$

with ω the angular frequency, and are fitted using vector-fitting (VF) [5]. This allows for them to be described analytically in the frequency domain and included in time domain simulations using the equivalent passive network depicted in Fig. 2. The elements of the latter are of constant values. Taking $Y_{tg}(f)$ as an example, its VF expression is

$$Y_{tg\text{fit}}(s) = d + se + \sum_{i=1}^{N_r} \left(\frac{c_{r_i}}{s - a_{r_i}} \right) + \sum_{j=1}^{N_c} \left(\frac{c_{c_j}}{s - a_{c_j}} \right), \quad (4)$$

where $s = j\omega$, d and e are real numbers, and c_{r_i} , c_{c_j} , a_{r_i} and a_{c_j} are the real and complex residues with their corresponding real and complex poles, respectively. The order of approximation is $N = N_r + N_c$, with N_r being the number of the used real poles and N_c the number of the used complex poles for the fitting.

The fitting $Y_{tg\text{fit}}$ could also be represented in the form of the electrical network depicted in Fig. 2, so its expression also equals the equivalent admittance of the network:

$$Y_{tg\text{fit}}(s) = \frac{1}{R_0} + sC_0 + \sum_{i=1}^{N_r} \left(\frac{1}{sL_{1_i} + R_{1_i}} \right) + \sum_{j=1}^{N_c} \left(\frac{1}{sL_j + R_j + \frac{1}{G_j + sC_j}} \right). \quad (5)$$

The constant values of the parameters R_0 , L_0 , R_{1_i} , L_{1_i} , R_j , L_j , C_j and G_j according to the computed residues and poles of the fitting can be deduced from (4) and (5) by identification as described in [5].

Finally, to guarantee the stability of the transient simulation, the passivity of the equivalent network is enforced using the positive-definite criterion. The performed correction is based on constrained minimization by quadratic programming as proposed in [6].

V. PRELIMINARY RESULTS

The proposed method was tested on a case study where the considered winding is of 34 turns with no magnetic core. The result of the frequency analysis is seen in Fig. 3, where the

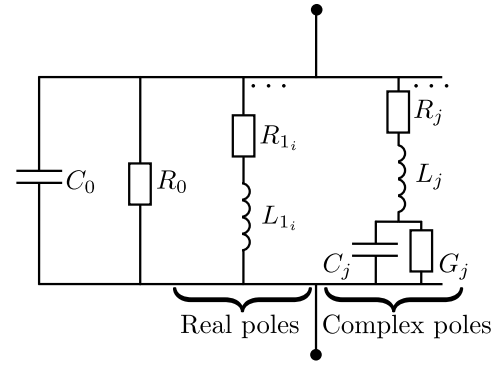


Fig. 2. The equivalent network used to take into account the frequency-dependent behavior of either Y_t , Y_{tt} or Y_{tg} [5].

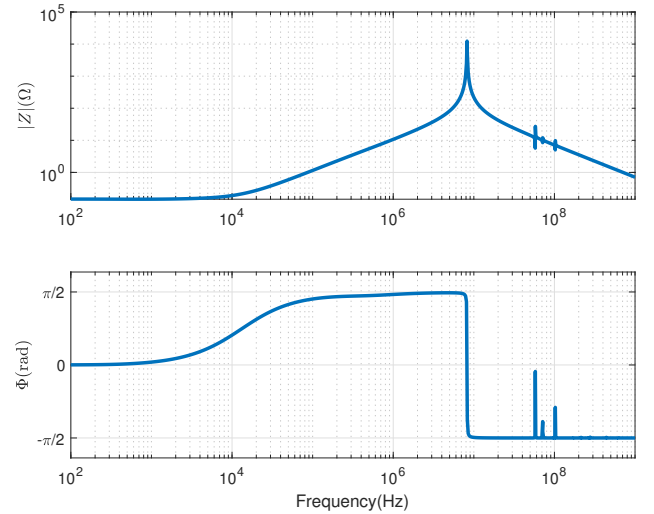


Fig. 3. The simulated frequency response, amplitude and phase, of the studied 34 turns winding.

high frequency resonances are noticed. This result and the time domain response of the model will be compared with the results of previous research work [2] in the extended version of the paper.

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