

# Time-domain distributed-parameter modeling of transformer windings for fast front transients

P. Gómez, J. C. Escamilla, P. Moreno

**Abstract**— This work presents the application of a recently developed time domain multiconductor line model for the representation of a transformer winding for high frequency transients, including the frequency dependence of the winding's electrical parameters. The model is based on the method of characteristics but, in contrast to its conventional implementation, it does not require of any space discretization. This is particularly useful to reduce the computer burden related to the simulation of the transient response of a transformer winding consisting of a large number of turns. Besides, since the final representation of the model consists of a nodal form, the inclusion of the model in a transient simulation program is straightforward. The winding model obtained in this work is applied for the computation of transient overvoltages and dielectric stresses due to the injection of a fast front impulse. The results are compared to those from a frequency domain method. The parameters required by the model are computed by means of FEM simulations and analytical expressions.

**Keywords:** Fast front transients, time domain analysis, transformer winding, method of characteristics.

## I. INTRODUCTION

ONE of the most important standard tests for the design and assessment of power transformers is the impulse test, which defines for most cases the highest dielectric stress that the transformer will suffer once in operation. Therefore, the insulation system is designed in general to withstand this test. However, to date commercial programs for the simulation of transient response of electrical components and networks (such as EMTP, Simulink, PSpice, etc.) do not include models able to predict the response of a transformer winding to an impulse test.

Several winding models have been proposed in the literature to analyze the electromagnetic transients produced by the propagation of fast front impulses. Lumped and

distributed parameter representations have been applied (see for instance [1]-[4]), but it has been shown that lumped parameter representations are only valid for transients containing frequencies up to around 1 MHz, while distributed parameter representations can reproduce transients well above such frequencies [4]. A winding representation based on a multiconductor transmission line model has demonstrated accurate results when compared to experimental measurements, given its ability to consider the inductive and capacitive coupling between turns, as well as the wave propagation along the winding. However, this model is usually described in the frequency domain, which precludes its direct implementation in time domain simulation programs. The representation of this model in time domain by means of the method of characteristics has also been reported in the past [5]. Still, this method requires the discretization of each winding turn in a number of segments and, since a detailed winding model can consist of hundreds or thousands of turns, the resulting system can be extremely large and, in consequence, the computer times can be excessive.

This work presents the application of a recently developed time domain multiconductor line model [6] for the representation of a transformer winding for high frequency transients, including the frequency dependence of the winding's electrical parameters. The model is based on the method of characteristics [7] but, in contrast to its conventional implementation, it does not require of any space discretization. This is particularly useful to reduce the computer burden related to simulate the transient response of a transformer winding consisting of a large number of turns. Besides, since the final representation of the model consists of a nodal form, the inclusion of the model in a transient simulation program is straightforward.

The winding model obtained in this work is applied for the computation of transient overvoltages and dielectric stresses due to the injection of a fast front impulse. The results are compared to those from a frequency domain method (the numerical Laplace transform [8]). The parameters required by the model are computed by means of finite element method (FEM) simulations and analytical expressions.

## II. MTL MODEL WITH FREQUENCY DEPENDENT ELECTRICAL PARAMETERS

The telegrapher equations in time domain for multiconductor transmission lines with frequency dependent electrical parameters are defined as [9]:

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P. Gómez is with SEPI-ESIME Zacatenco, Instituto Politécnico Nacional, México (e-mail: pgomezz@ipn.mx).

P. Moreno is with CINVESTAV, Unidad Guadalajara, México (e-mail: pmoreno@gdl.cinvestav.mx).

J. C. Escamilla is with CENALTE, Comisión Federal de Electricidad, México (e-mail: escamilla\_14@hotmail.com).

$$\frac{\partial \mathbf{v}}{\partial z} + \mathbf{L}_0 \frac{\partial \mathbf{i}}{\partial t} + \frac{\partial}{\partial t} \int_0^t \mathbf{r}'(t-\tau) \mathbf{i}(\tau) d\tau = \mathbf{0} \quad (1a)$$

$$\frac{\partial \mathbf{i}}{\partial z} + \mathbf{C} \frac{\partial \mathbf{v}}{\partial t} + \frac{\partial}{\partial t} \int_0^t \mathbf{g}'(t-\tau) \mathbf{v}(\tau) d\tau = \mathbf{0} \quad (1b)$$

where  $\mathbf{v}$  and  $\mathbf{i}$  are the vectors of voltages and currents propagating along the line ( $z$  is the propagation axis);  $\mathbf{L}_0$ ,  $\mathbf{C}$ ,  $\mathbf{r}'$  and  $\mathbf{g}'$  are the matrices of geometrical inductances, capacitances, series transient resistances and shunt transient conductances of the line, respectively. In (1a) the convolution term is solved using a recursive scheme and the transient resistance is fitted with a sum of rational functions of the form:

$$\mathbf{R}'(s) = \sum_{i=1}^N \frac{1}{s + p_i} \mathbf{K}_i + \frac{\mathbf{K}_0}{s} + \mathbf{K}_\infty \quad (2)$$

where  $\mathbf{R}'(s)$  is the frequency domain image of  $\mathbf{r}'(t)$ ,  $N$  is the order of the approximation,  $\mathbf{K}_i$  is the  $i$ -th residues matrix and  $p_i$  the corresponding pole,  $\mathbf{K}_0$  is the residues matrix at  $s = 0$  and  $\mathbf{K}_\infty$  is the residues matrix at  $s = \infty$ . Considering that the shunt conductances matrix  $\mathbf{G}$  does not depend on frequency the telegrapher's equations become of the following form:

$$\frac{\partial \mathbf{v}}{\partial z} + \mathbf{D} \frac{\partial \mathbf{i}}{\partial t} + \mathbf{R}_x \mathbf{i} + \mathbf{\Psi} = \mathbf{0} \quad (3a)$$

$$\frac{\partial \mathbf{i}}{\partial z} + \mathbf{C} \frac{\partial \mathbf{v}}{\partial t} + \mathbf{G} \mathbf{v} = \mathbf{0} \quad (3b)$$

where  $\mathbf{D}$ ,  $\mathbf{R}_x$  and  $\mathbf{\Psi}$  are defined as:

$$\mathbf{D} = \mathbf{K}_\infty + \mathbf{L}_0 \quad (4a)$$

$$\mathbf{R}_x = \mathbf{R}_{dc} + \sum_{i=1}^N \mathbf{K}_i \quad (4b)$$

$$\mathbf{\Psi}(z, t) = -\sum_{i=1}^N p_i \mathbf{K}_i \left[ e^{-p_i t} * \mathbf{i}(z, t) \right] \quad (4c)$$

Multiplying (3a) and (3b) by  $\mathbf{D}^{-1}$  and  $\mathbf{C}^{-1}$ , respectively, gives the following system:

$$\frac{\partial}{\partial z} \mathbf{U} + \mathbf{A} \frac{\partial}{\partial t} \mathbf{U} + \mathbf{B} \mathbf{U} + \mathbf{W} = \mathbf{0} \quad (5)$$

where

$$\mathbf{A} = \begin{bmatrix} \mathbf{0} & \mathbf{C}^{-1} \\ \mathbf{D}^{-1} & \mathbf{0} \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} \mathbf{C}^{-1} \mathbf{G} & \mathbf{0} \\ \mathbf{0} & \mathbf{D}^{-1} \mathbf{R}_x \end{bmatrix} \quad (6a), (6b)$$

$$\mathbf{U} = \begin{bmatrix} \mathbf{V} \\ \mathbf{I} \end{bmatrix}, \quad \mathbf{W} = \begin{bmatrix} \mathbf{0} \\ \mathbf{D}^{-1} \mathbf{\Psi} \end{bmatrix} \quad (6c), (6d)$$

Diagonalization of products  $\mathbf{DC}$  and  $\mathbf{CD}$  are defined by:

$$\mathbf{T}_v^{-1} \mathbf{DC} \mathbf{T}_v = \mathbf{\Lambda} \quad (7a)$$

$$\mathbf{T}_i^{-1} \mathbf{CD} \mathbf{T}_i = \mathbf{\Lambda} \quad (7b)$$

where  $\mathbf{T}_v$  and  $\mathbf{T}_i$  are the modal matrices of voltages and currents respectively. Besides, modal electrical parameters can be expressed as follows:

$$\tilde{\mathbf{D}} = \mathbf{T}_v^{-1} \mathbf{D} \mathbf{T}_v \quad (8a)$$

$$\tilde{\mathbf{C}} = \mathbf{T}_i^{-1} \mathbf{C} \mathbf{T}_i \quad (8b)$$

and

$$\tilde{\mathbf{DC}} = \mathbf{\Lambda} \quad (8c)$$

The characteristic impedance and admittance matrices are given by:

$$\mathbf{Z}_w = \sqrt{\tilde{\mathbf{D}} \tilde{\mathbf{C}}^{-1}} \quad (9a)$$

$$\mathbf{Y}_w = \sqrt{\tilde{\mathbf{C}} \tilde{\mathbf{D}}^{-1}} \quad (9b)$$

Moreover, from the diagonalization of  $\mathbf{A}$  the propagation velocities are obtained:

$$\mathbf{M}_L \mathbf{A} \mathbf{M}_R = \begin{bmatrix} \mathbf{\Gamma} & \mathbf{0} \\ \mathbf{0} & -\mathbf{\Gamma} \end{bmatrix} \quad (10)$$

where

$$\mathbf{\Gamma} = \sqrt{\tilde{\mathbf{C}}^{-1} \tilde{\mathbf{D}}^{-1}} = \sqrt{\mathbf{\Lambda}^{-1}} \quad (11)$$

The left ( $\mathbf{M}_L$ ) and right ( $\mathbf{M}_R$ ) eigenvectors of  $\mathbf{A}$  are defined as

$$\mathbf{M}_L = \frac{1}{\sqrt{2}} \begin{bmatrix} \mathbf{T}_v^{-1} & \mathbf{Z}_w \mathbf{T}_i^{-1} \\ \mathbf{T}_v^{-1} & -\mathbf{Z}_w \mathbf{T}_i^{-1} \end{bmatrix} \quad (12a)$$

$$\mathbf{M}_R = \frac{1}{\sqrt{2}} \begin{bmatrix} \mathbf{T}_v & \mathbf{Z}_w \mathbf{T}_v \\ \mathbf{T}_i \mathbf{Y}_w & -\mathbf{T}_i \mathbf{Y}_w \end{bmatrix} \quad (12b)$$

Eq. (5) is left-multiplied by  $\mathbf{M}_L$  and, knowing that along the characteristics

$$\left( \frac{\partial}{\partial t} + \gamma_j \frac{\partial}{\partial z} \right) = + \frac{d}{dt} \quad (13a)$$

$$\left( \frac{\partial}{\partial t} - \gamma_j \frac{\partial}{\partial z} \right) = - \frac{d}{dt} \quad (13b)$$

the following system in modal domain is obtained:

$$\begin{aligned} \frac{d}{dt}V_{mj} + Z_{wj} \frac{d}{dt}I_{mj} + \frac{dz_j}{dt} \sum_{k=1}^n \tilde{R}_{Xjk} I_{mk} \\ + \frac{dz_j}{dt} Z_{wj} \tilde{G}_j V_{mj} + \frac{dz_j}{dt} \psi_{mj} = 0 \end{aligned} \quad (14a)$$

$$\begin{aligned} \frac{d}{dt}V_{mj} - Z_{wj} \frac{d}{dt}I_{mj} + \frac{dz_j}{dt} \sum_{k=1}^n \tilde{R}_{Xjk} I_{mk} \\ - \frac{dz_j}{dt} Z_{wj} \tilde{G}_j V_{mj} + \frac{dz_j}{dt} \psi_{mj} = 0 \end{aligned} \quad (14b)$$

where

$$\gamma_j = \pm \frac{dz}{dt} \quad (14c)$$

Equations (14) are an Ordinary Differential Equations system that represents the Partial Differential Equations system given by (1a) and (1b). The central finite differences method is used to integrate (14). The values at the sending and receiving ends are approximated using the values at a travel time back. This approach, conversely to the conventional Method of Characteristics, *does not require of any spatial discretization*. After further algebraic manipulation a dual Norton model for the transmission line in phase domain is obtained. This model can be included into any simulation program based on the Nodal or the Modified Nodal Method. Further explanation of this modeling approach can be found in [7].

### III. ZIG-ZAG CONNECTION FOR WINDING MODEL

Wave propagation phenomena along the winding can only be reproduced accurately with a distributed parameter model. However, the need to consider the turn-to-turn inductances can be a serious shortcoming. A model based on the multi-conductor transmission line theory, initially proposed by Rabins [10] and developed for electrical machines by Guardado et al. [11], has been used successfully to account for the mutual inductance between turns in a distributed parameter model for transformer windings (see for instance [1], [2]). The model is based on a zig-zag connection of the different conductors, each of them representing a winding section (disc or turn) of the complete winding, as shown in Fig. 1. To preserve continuity, the end of each conductor is topologically connected to the beginning of the next conductor, resulting in a zig-zag connection. The winding model is obtained from the telegrapher equations of a multiconductor transmission line, so that it can be solved using the approach described in section II.

In Fig. 1, the equivalent impedance  $Z_{eq}$  connected at the end of the  $N$ -th element can be used to represent the remaining part of the winding, when only a section of the winding needs to be modeled in detail.  $Z_{eq}$  can also represent the neutral impedance.

### IV. DETERMINATION OF WINDING PARAMETERS

Computation of parameters for high-frequency transformer models is based on similar approaches from those applied for low- and mid-frequency models. They can be classified in three: (1) the application of reduced scope expressions obtained from simplified geometries or empirical equations; (2) direct experimental determination of the parameter through laboratory tests; (3) electromagnetic field simulations. [12]

Regardless of the model employed for the simulation of the transformer response, inductive, capacitive and loss components of the model are required to accurately describe the behaviour of the winding at high frequencies.

The flux penetration into the core is usually neglected for very fast transients assuming that the core acts as a flux barrier. [13]

#### A. Capacitance

The most common approach to compute the winding capacitances is based on the well-known formula for parallel plates. However, this method neglects fringe effects. A more general and accurate calculation can be obtained from electrostatic field simulations. In this work, the finite element method (FEM) is used for this purpose, considering a 2D axial-symmetric arrangement, as shown in Fig. 2. Self and mutual capacitances are obtained using the electrostatic energy method. Further explanation of this computation approach can be found in [3].

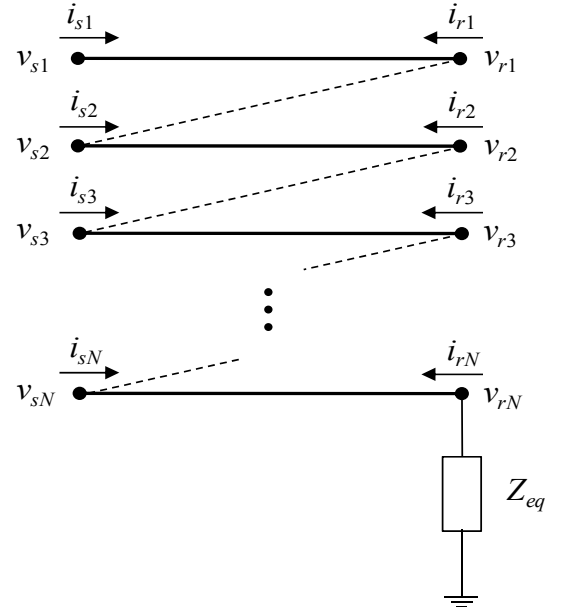


Fig. 1 Multiconductor transmission line model for the transformer winding [12]

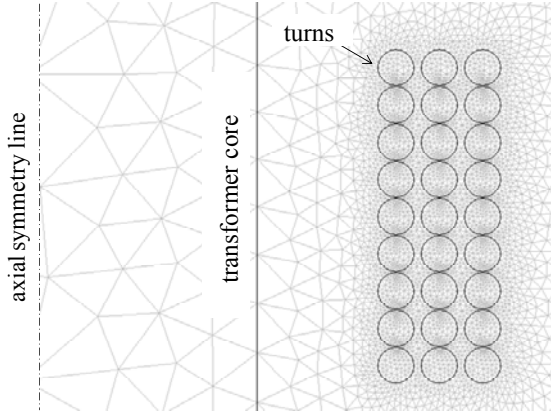


Fig.2. 2D axial-symmetric FEM model of the transformer winding for electrostatic simulations.

Assuming that the winding has  $N$  layers and  $n$  turns per layer, the following capacitive values need to be computed:

- $C_{s,o}$  Self capacitance of any turn at the outer layer ( $N$ )
- $C_{s,i}$  Self capacitance of any turn at the inner layer ( $i$ )
- $C_{s,m}$  Self capacitance of any turn at any interior layer (2, ...  $N-1$ )
- $C_{it,o}$  Mutual capacitance between any two adjacent turns at the outer layer ( $N$ )
- $C_{it,i}$  Mutual capacitance between any two adjacent turns at the inner layer ( $i$ )
- $C_{it,m}$  Mutual capacitance between any two adjacent turns at any interior layer (2, ...  $N-1$ )
- $C_{il,o}$  Mutual capacitance between the  $i$ -th turn at the outer layer and the  $i$ -th turn at the following interior layer
- $C_{il,m}$  Mutual capacitance between the  $i$ -th turns of any 2 interior layers

### B. Inductance

An expression used extensively to compute mutual inductances is the exact formula derived by Maxwell for two thin wire coaxial loops [14]. This has been extended by Lyle to compute the mutual inductances between winding turns or group of turns [15], [16]. Analytical expressions to compute the self-inductance of a circular coil have been defined by Grover [17] and Gray [18]. However, all these formulations neglect the magnetic shielding effect of the core at high frequencies.

Similarly to the capacitance matrix, the inductance matrix can be computed directly from FEM analysis using the energy method to obtain more accurate results for realistic arrangements [19]. Alternatively, the geometric inductance matrix can be obtained directly from the inverse of the capacitive matrix computed from FEM simulations:

$$\mathbf{L}_0 = \mu \varepsilon \mathbf{C}^{-1} \quad (15)$$

where  $\mu$  and  $\varepsilon$  are the permeability and permittivity of the surrounding medium.

### C. Losses

Both series and shunt elements of the windings at high

frequencies are frequency dependent.

Expressions for the conductor and core losses at high frequencies are based on the concept of the complex flux penetration depth in the windings and in the core. In this work, only conductor losses due to skin and proximity effects are considered, which is acceptable for very fast transients. These are computed from the following expression [2]:

$$\mathbf{R} = \frac{1}{d} \sqrt{\frac{2\omega}{\sigma_c \mu_c}} \mathbf{L}_0 \quad (16)$$

where  $d$  is the distance between layers,  $\omega$  is the angular frequency,  $\sigma_c$  is the conductivity of the winding conductor and  $\mu_c$  is its permeability. According to the Laplace transform of (1a), the transient resistances matrix is related to the conductor losses matrix by  $\mathbf{R}' = \mathbf{R} / s$ .

At very high-frequency, the conductance representing the capacitive loss in the winding's dielectric also depends on frequency. The capacitive loss in the insulation material can be computed directly from the capacitance matrix making use of the loss factor of the winding insulation,  $\tan \delta$ , which is obtained from the concept of complex permittivity. Capacitive losses are a linear function of frequency; additionally, it has been observed that the loss factor in oil treated cellulose papers is also frequency dependent [4]. These losses are included in this work in the form of a shunt conductance matrix given by

$$\mathbf{G} = (\omega \tan \delta) \mathbf{C} \quad (17)$$

## V. TEST CASE

A small prototype (200 turns: 4 layers, 50 turns each) of a transformer winding with dielectric distances designed for a voltage level of 13.2 kV (BIL = 95 kV) is considered. Its main geometrical data is listed in Table I.

The transient response of the winding is analyzed by means of the injection of a linearly rising and decaying waveform (60 ns of front time and 25 μs for decay to half value) at the high-voltage terminal of the winding (located at the outermost layer of the winding). For the simulation of the response of the winding to this waveform, the distributed parameter model described in Section II is applied, with its parameters computed as explained in Section III.

TABLE I  
MAIN GEOMETRICAL DATA OF THE WINDING UNDER STUDY

Core diameter [mm]	250
Distance between windings and between winding and core [mm]	10
Winding conductor diameter [mm]	3
Distance between layers [mm]	0.6
Distance between turns [mm]	0.1
Number of layers	4
Number of turns per layer	50

Fig. 3 shows a comparison of the transient voltages at different turns of the winding with the proposed model and by means of a model based also on a multiconductor line representation, but defined in the frequency domain and solved in time domain by means of the inverse numerical Laplace transform [8]. Waveforms are very similar for all the turns shown in the figure.

Fig. 4 shows, by means of a contour map, the behavior of the winding along all its turns and for all the time range under consideration. This type of plot allows identifying the region of maximum voltage stress and the time period in which this occurs. Finally, as an example of the possible application of the proposed model for insulation design, Fig. 5 shows the inter-turn dielectric stresses along the first layer, computed as described in [3]. The region of maximum stress is observed in the last turns, with values around 70 MV/m, indicating that dielectric distances and insulation material between turns of the winding prototype require improvement to avoid breakdown.

A complete study of dielectric stresses should consider all the layers, the stress between HV and LV windings and in particular the inter-layer stresses, which can be even greater than the inter-turn stresses due to the large potential difference between turns from different layers. However, the scope of this paper is only to establish the accuracy of the proposed model, with aims of its future application in commercial EMTP-type programs.

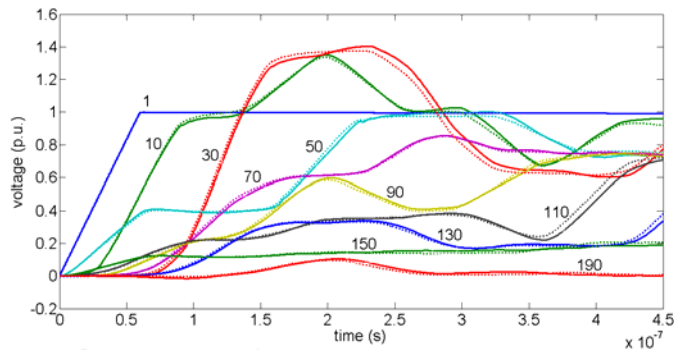


Fig.3. Transient voltages at different turns of the winding. Results from the proposed model are shown in continuous line. Results from the numerical Laplace transform are shown in dotted line.

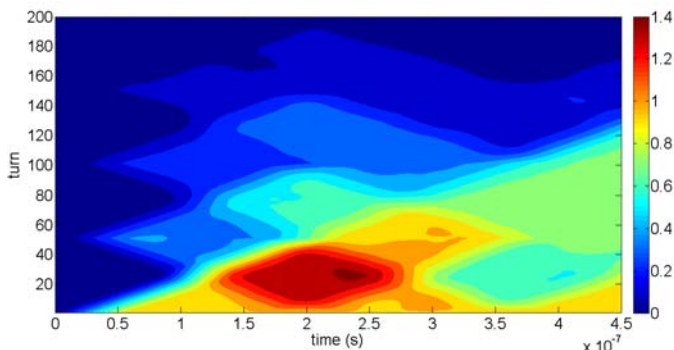


Fig.4. Contour map of the transient voltages (absolute value, in p.u.) along the winding

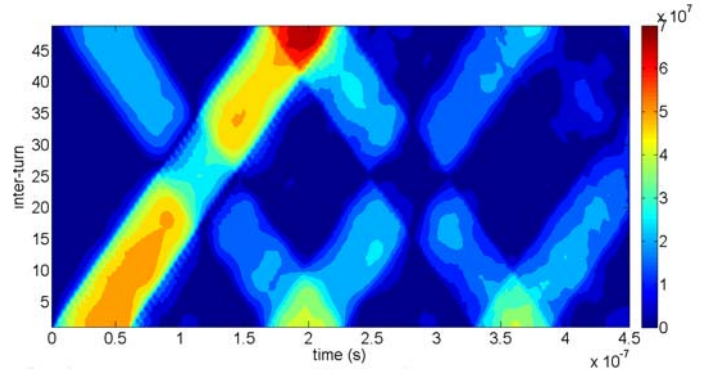


Fig.5. Contour map of inter-turn dielectric stresses (in V/m) along the first 50 turns of the winding.

## VI. CONCLUSIONS

A new time-domain modeling approach for the study of fast transients in transformer windings has been presented in this paper. This approach is based on the method of characteristics and the multiconductor line model but, conversely to the conventional method of characteristics, it does not require of any spatial discretization, making it more suitable for its implementation in EMTP-type programs, which do not include this type of models to date.

The parameters required by the winding model are computed by means of FEM simulations and analytical expressions. The model is able to include frequency dependence of series and shunt losses.

Comparisons with the results from the numerical Laplace transform for a 200 turns winding have shown that the proposed modeling approach has very good precision. Further results using contour maps of transient voltages and dielectric stresses were included to show possible application of the model for insulation design.

## VII. ACKNOWLEDGMENT

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