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### ▶ To cite this version:

Daniel Roger, Mathieu Rossi. Over-voltage distribution between turns of the transformer coils of large SST cells. IEEE 13th International Conference on Compatibility, Power Electronics and Power Engineering (CPE-POWERENG), Apr 2019, Sonderborg, Denmark. hal-04287776

## HAL Id: hal-04287776 https://hal.science/hal-04287776

Submitted on 15 Nov 2023

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# Over-voltage distribution between turns of the transformer coils of large SST cells

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Abstract—Solid state transformers (SST) are becoming key elements for integrating an increasing part of intermittent renewable energy in electric grids [1]. They bring also compact solutions for electric transportation systems [2]. Our society needs powerful SSTs for a reasonable cost. A good technical-economic balance can be reach by designing large power SST cells with mature technologies for power electronics, magnetic cores and electrical insulation systems (EIS). The paper develop a design method able to compute the turn-to-turn voltage distribution in the transformer of large SST cells during the fast transient states imposed by the converters switchings. The proposed method makes it possible to design high voltage SST windings with a mature low voltage dry EIS based on polymers.

*Index Terms*—Solid State Transformer (SST), High-power cells, Dry polymer insulation, Mature technologies.

#### I. INTRODUCTION

Generally speaking, a solid-state transformers (SST) is a dcdc converter with a galvanic insulation. They yield a reversible power transfer between a high voltage (HV) source and a low voltage (LV) one. They are made of several elementary dc–dc cells connected in series on the HV side and in parallel on the LV one. Each dc-dc cell is made of two converters, one on each side of a transformer providing the galvanic insulation and the voltage level change.

The investigated structure consists of a single-phase transformer placed between two IGBT bridges built with available high-voltage IGBT modules [3], [4]. The transformer is designed around a rolled core made of thin grain oriented (GO) magnetic sheets. Such cores are able to operate at several kilohertz with a flux density up to 1.5T [5]. The transformer primary coil is designed with a dry insulation technology based on polymers [6]. Mature technologies for magnetic cores and coils EIS make it possible the design of high power medium frequency transformers (MFT) and elementary SST cells with good technical-economic balance.

The aim of the paper consists in studying the voltage distribution between the turns of the MFT primary coil during the fast transient that follows each fast-fronted voltage caused by IGBT bridge switchings. This analysis will be used for getting the highest possible voltage on the MFT primary coil with widespread polymer EIS developed for low-voltage applications. 2<sup>nd</sup> Mathieu Rossi Universite d'Artois - EA4025 Lab. Systemes Electrotechniques et Environnement (LSEE) Bethune, France mathieu.rossi@univ-artois.fr

Fast transients in coils were studied a long time ago for low voltage motor windings fed by IGBT inverters [7], [8]. Several research works develop high frequency (HF) equivalent circuits based on experimental measurements made with broadband impedance analyzers [9], [10]. 2D and 3D finite elements models can also be used for determining these parameters [11]–[14].

The paper proposes a geometry for the primary coil of high power SST cells based on MFTs made with a rolled GO core and a widespread low-voltage enameled rectangular copper wire [15]. A HF equivalent circuit, built at the turn level, is used for computing the maximum turn-to-turn voltage stress and estimating the maximum rated voltage that can be reached with such a mature polymer insulation technology.

#### II. MFT TOPOLOGY

The MFT of each cell must be designed for providing the capability of connecting the SST cells in series for the HV side and in parallel for the LV one. The insulation problems on the LV side remain minor but, with a series connection on the HV side, the cell placed at the top of the stack withstands the whole dc voltage of the HV source superimposed with the MFT primary ac voltage. The common mode insulation between the primary coil and the other parts of the MFT must consider this high voltage stress; distances between the primary coil, the secondary one and the magnetic core must be large. An impregnation resin must fill the gap for getting high breakdown voltage and thermal resistances as low as possible. Figure 1 shows an example of such of a MFT structure made with a standard dry insulation technology and a single-layer primary coil made of enameled rectangular copper wire that provide a good distribution of the turn-to-turn voltage stress along the primary coil.

The skin effect in the copper must be considered. For example, in a MFT working at 3kHz the skin depth in the rectangular wire is 1.48mm at  $150^{\circ}C$ . Therefore, it is possible to use a widespread enameled rectangular copper wire that thickness is 2mm. This enameled wire is isolated by two thin polymer layers. The inner one, made of polyester-imide (PEI), provides excellent adherence to the copper. The outer one, made of polyamide-imide (PAI) gives a thermal class of  $200^{\circ}C$ . The wire is able to operate at  $200^{\circ}C$  during 20000h. Lifetimes are much longer at lower temperature following the Arrhenius law [16].

For a round enameled wire, the high-performance flexible insulating layer makes it possible to operate sharp bendings, which radius are similar to the wire one. This property is also valid for enameled rectangular wires. With a good winding machine imposing a high tensile force and adequate tools, it is possible to bend a rectangular enameled wire as along its width as shown in figure 2. The minimum internal bending radius Ri must be of the order of the wire width W. The sharp edges of the magnetic core corners must be softened for avoiding electric field concentration that may cause corona effects and partial discharges (PD) [17]. With a  $14 \times 2mm$ enameled rectangular wire for example, the current can be up to  $120A_{RMS}$  depending on thermal conditions. Higher currents are possible by winding simultaneously two or more similar enameled rectangular wire connected in parallel.



Fig. 1. MFT structure providing high breakdown voltage for the common mode insulation of the primary coil.



Fig. 2. Cross-section of the MFT primary coil made with an enameled rectangular wire. The large distance between the core and the makes it possible to withstand the high common-mode voltage stress.

A SST cell prototype of reduced power (800V - 50kW) has been built for getting an experimental experience of such SST cells. The connections between the IGBT bridges and the

MFT are as short as possible (20cm). Voltage measurements made on this prototype shows short oscillating over-voltages after each voltage fast fronted voltages imposed by the IGBT bridge switchings. The natural frequency of the transient states (500kHz) measured on the MFT prototype gives a starting point building a HF equivalent circuit that will help the design of MFT with higher rated voltages and powers.

#### III. PRIMARY COIL HIGH FREQUENCY MODEL

#### A. Preliminary experimental assessments

Experimental assessments have been made with single layer coil for evaluating the influence of the core at high frequencies. The cross section of this coil is the same as the primary of the MFT prototype. The axial distances between turns were increased for lowering the turn-to-turn capacitances. The turn number is lower. With this coil it is possible to investigate the inductive couplings between turns with a very low influence of the turn-to-turn capacitances. Figure 3 shows the impedance spectrum modulus of this coil without any magnetic core and with the rolled GO core. It can be seen that the natural frequency of the coil is at about 20MHz in air and a little less with the magnetic core. The lower maximum value of the modulus shows the influence of core losses. This figure shows that, at the natural frequency measured on the MFT prototype, the influence of the core cannot be neglected.



Fig. 3. Impedance modulus of a low turn-to-turn capacitance coil.

#### B. Rolled GO core equivalent HF permeability

The core is made of many thin GO sheets rolled together for getting the correct cross-section; then, the core is cut for facilitating the MFT assembly. With this technology each magnetic sheet works everywhere in its rolling direction. At high frequencies, the skin effect has a major influence. A classical analytic approach is made for sine operations, supposing linear conditions. The complex formulation is used; it defines the skin depth (1) form angular frequency  $\omega = 2\pi f$ , the electric conductivity  $\sigma$  and the magnetic permeability  $\mu = \mu_0 \ \mu_R$ . The magnetic field, at a given point x inside the sheet results of two damped waves moving in opposite directions, from one side of the sheet toward the other one [18]. In the resulting expression (3), the sheet thickness is a; the origin x = 0 is situated at the centre or the considered sheet;  $H_0$  is the excitation field imposed at the two borders  $x = \pm a/2$  and  $\gamma$  the wave propagation constant (2).

$$\delta = \sqrt{\frac{2}{\omega\mu\sigma}} \tag{1}$$

$$\gamma = \frac{1+j}{\delta} \tag{2}$$

$$\underline{H}(x) = H_0 \frac{\cosh(\gamma \ x)}{\cosh(\gamma \ a/2)} \tag{3}$$

The average flux density  $\langle B \rangle$  is computed from (3) by integration for a given excitation field  $H_0$  on both side of the considered sheet. Then the equivalent magnetic permeability is computed from the ratio (4) that yields expression (5).

$$\underline{\mu_{eq}} = \frac{1}{\mu_0} \frac{\langle B \rangle}{H_0} \tag{4}$$

$$\underline{\mu_{eq}} = \frac{2 \ \mu_R}{\gamma \ a} \ tanh(\gamma \frac{a}{2}) \tag{5}$$

In these expressions the local parameters  $\mu$  and  $\sigma$  characterize the GO steel itself in the rolling direction. The conductivity can be supposed constant but not the permeability, which strongly depends on the flux density in the core. Consequently, the exact determination of the parameters of a HF equivalent circuit able to describe the coil transient behaviour after an IGBT bridge switching is not possible because they depends on the flux density in the core just before the switching. The parameters are not constant, they depend on the converter command. However, it is possible to consider a wide range of material permeabilities for computing the complex equivalent permeability. Fig. 4 plots results for two material relative permeabilities ( $\mu_R = 5000$  and  $\mu_R = 10000$ ) and a sheet thickness  $a = 165\mu m$ . The solid lines represent the real parts and the dotted lines the imaginary ones.

When a coil is placed on this core, its inductance depends on the coil geometry and on the core equivalent permeability. Noting K the constant factor that depends on geometry and the turn number, the coil inductance is a complex number defined by (6). The impedance of such a coil, at a given angular frequency  $\omega$ , is the complex number (7) corresponding to a series equivalent circuit. The real part of the inductance yields the classical magnetic influence of the core. The imaginary part denotes the core losses due to eddy currents.

$$\underline{L}(\omega) = K \ \mu_R(\omega) \tag{6}$$

$$\underline{Z_S} = j\omega \ \underline{L}(\omega) = +jX_{Seq}(\omega) + R_{Seq}(\omega)$$
(7)

Fig. 4 shows that, for low frequencies, the real part of the complex permeability is the material one; the imaginary part

tends toward zero. Figure 5 is drawn for high frequencies. It shows that the magnetic core keeps an influence for the frequency in the range of the natural frequency of the transient measured on the MFT prototype.



Fig. 4. Equivalent complex relative permeability of the rolled core versus frequency.



Fig. 5. Equivalent complex relative permeability for high frequencies.

#### C. Inductances and core losses

The self inductance of each turn and the magnetic couplings between turns are determined experimentally at 500kHz with the experimental setup presented in figure 6. Two 10-turn rectangular coils of the same size as the primary winding are placed around the magnetic core. The coils are made of a thin copper wire with a thick insulation for minimizing turnto-turn capacitances and getting a natural frequency much higher than 500kHz. The two coils are identical. The first coil is fed by a sine generator while a broadband current probe measures the current. The self inductance is computed from the voltage measured on the first coil. The mutual inductances are determined from he voltage is measured on the second coil at several distances d between coils.



Fig. 6. Experimental setup for determining inductances.





Fig. 7. Self and mutual inductances.

In the air, the phase lag between the measured voltage and current is very close to  $90^{\circ}$  the inductances are real numbers. With the rolled core at the centre of the coils, the phase lag is much lower because of eddy currents. The inductances becomes complex numbers.

The decrease of the inductances look like exponential curves. Therefore the polynomial fitting has been made on the natural logarithm of the ratio  $L_{ij}/L_{max}$ . Figure 8 presents results for a second degree fitting.

The inductance model is given by (8); d is the distance between the coils expressed in meters; the fitting coefficients are in table I.  $L_{max}$  is the real part of the self inductance without any core and with the rolled core.



Fig. 8. Fitting curves.

$$L_{ij} = L_{max} \exp\left(a_2 d^2 + a_1 d + a_0\right)$$
(8)

TABLE I POLYNOMIAL FITTING COEFFICIENTS

	L <sub>max</sub>	$a_2$	$a_1$	$a_0$
Air	$12.73 \mu H$	$74.87 \ m^{-2}$	$-36.85 \ m^{-1}$	-0.0975
Re(Core)	$26.23 \mu H$	$38.33 \ m^{-2}$	$-25.33 m^{-1}$	-0.097
Im(Core)	$26.23 \mu H$	$-15.46 \ m^{-2}$	$-2.51 \ m^{-1}$	-1.192

With this inductance model, it is possible to compute the symmetrical  $N \times N$  matrix inductance of the N-turn primary coil from the distance between the turn *i* and the turn *j* using (8) after dividing  $L_{max}$  by 100 because the model was made with two flat ten-turn coils. The distance  $d = T_w |j-i|$  between the considered turns that depends on the overall thickness  $T_w$  of the en enameled rectangular wire.

The inductance model was used for predicting global impedance of the long coil used in section *III.A* (fig. 3) at 500kHz. For this coil and at this frequency, the influence of the turn-to-turn capacitances can be neglected; the global inductance of the coil is the simple sum of all the terms of the inductance matrix. Table II gives the results where  $R_{eq} = \Im(L_G)\omega$ . Measurements made without any core show that the wire resistance is negligible. The accuracy of the model is more or less 10%.

TABLE II Model validation at 500kHz

	Model		Measurements	
	$\Re(L)$	$R_{eq}$	$L_s$	$R_s$
Air	$5.85 \mu H$	_	$6.08 \mu H$	$0.17\Omega$
Core	$14.3 \mu H$	$26.9\Omega$	$13\mu H$	$24.4\Omega$

#### D. Capacitance estimation

Turn-to-turn capacitances are computed with the classical plane capacitance formulae, considering the the average turn length, the enameled rectangular wire width, the total insulation thickness and its permittivity. The result is  $C_T = 35pF$ , with a large uncertainty because of the deformations of the enameled rectangular wires during winding operations. The common-mode capacitances ( $C_{CM}$ ) of each turn is much lower because of the small thickness of each turn (2mm) and the large insulation distances. The turn-to-ground capacitance is estimated to  $C_{CM} = 0.2pF$  for the internal turns and to  $C_{CM0} = 5pF$  for the first turn and the last one. With standard polymers, the dielectric losses can be neglected.

#### E. Equivalent circuit automatic building

The real parts of the inductance matrix correspond to tuns self-inductances and magnetic couplings between turns. The imaginary parts define the core losses at a given frequency. A simplifying hypotheses consists in adding a resistance per turn that corresponds to a part of the global equivalent resistance. For a N-turn coil, this resistance is given by (9).

$$R_T = \frac{1}{N} \sum_{i=1}^{N} \sum_{j=1}^{N} \Im(L_{ij}) \omega$$
(9)

The HF equivalent circuit of the primary coil is presented in figure 9 with 2 nodes per turn  $(In_i \text{ and } R_i \text{ for the turn } i)$ . Common mode capacitances are connected between each turn input  $In_i$  and the ground (node 0). The wires between the source and the coil are added  $(0.1\Omega - 50nH)$ . An OCTAVE script creates the netlist defining the HF equivalent circuit.



Fig. 9. Equivalent HF circuit of the primary coil.

The SPICE solver computes the transient voltage distribution with a time step of 4ns during transients following each voltage edge for a square excitation. The rise and fall times are set to 40ns.

#### IV. RESULTS AND DISCUSSION

The spice simulation was made for a 40-turn coil, 40 selfinductances and 780 mutual ones. Figure 10 presents the turnto-ground voltages for a full-wave command of the inverter and a dc bus at 1000V. A perfect voltage inverter produces the source voltage ( $V_S$  - black curve). Damped oscillating transients can be seen for the inner turns. The figure shows common mode voltages on tuns 1, 10, 20, 30 and 40 situated at the input of the coil, 1/4, 1/2, 3/4 of the coil length and at the end of the coil. For this coil, the transient state duration is more or less  $1\mu s$ . Another faster transient of small magnitude can be observed on  $V_1$ , It is due to the inductances of the connection wires.

More details on the turn-to-turn voltages are given in figure 11 for the adjacent turns situated at the same positions in the coil. The maximum voltage stress is obtained between turns 10 - 11 and 30 - 31 which are situates at 1/4 and 3/4 of the coil length. There is nearly no voltage spike for turns situated at the centre of the coil length (20 - 21).



Fig. 10. Turn-to-ground voltages for a DC bus of 1000V.

These simulations are made for a dc bus of 1000V, in linear conditions, for a 40-turn coil and for a given core size and insulation distances. The main practical result is a maximum turn-to-turn voltage stress of 46V during transients. This value can be compared to the same stress stress endured in steady state, after the transient, which is only  $\frac{1000}{40} = 25V$ , which correspond to a maximum overshoot of 84%

For designing a high power MFT with the primary coil topology proposed in section II, the partial discharge inception voltage (PDIV) of the turn-to-turn insulation is a key parameter. The maximum stress must remain under this value, with a safety margin, for guaranteeing the absence of PD and the long lifetime of the Arrhenius law [16]. The PDIV measured on a set of standard enameled rectangular wires, at sea-level pressure and room temperature is about  $1200V_{Peak}$  for a grade 2 low-cost insulation [19]. With this key parameter, the proposed equivalent circuit can be used for helping the design of high power MFT for high input voltage SST cells.

The weak point of the proposed analysis is the lack of accuracy for determining several elements of the HF equivalent circuit. Indeed, they are determined with a sine excitation at a given frequency, while the model is used in time domain. Therefore, the robustness of the model was evaluated with respect to three main parameters: the turn capacitance  $C_T$ , the common mode one  $C_{CM}$  and the maximum value of the



Fig. 11. Turn-to-turn voltages after a positive and a negative edge.

inductance matrix  $L_{max}$ . One of the above parameter was doubled while the others remain constant; results are given in table III. It can be seen that the most sensitive parameter is the common mode capacitance.

TABLE III Model robustness

	Overshoot	Difference
Reference	84%	-
Doubling $C_T$	68%	-16%
Doubling $C_{CM}$	133%	+49%
Doubling $L_{max}$	108%	+24%

#### V. CONCLUSION

A HF model of the primary coil of the MFT of a high power SST cell has been proposed for determining the voltage maximum stress of the turn-to-turn insulation during fast transients. The model consists on a HF equivalent circuit built at the turn level and an experimental method for determining its parameters. This model is applied to a winding structure based on a widespread rectangular enameled wire commercialized for low-voltage applications. The transient analysis shows that this low-cost solution can be used for designing high power SST cells able to work in medium frequencies up to 10kV.

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