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Experimentally validated model of a fast switched salient pole rotor winding

R. Felicetti, C. J. D. Abrahamsson, U. Lundin

Abstract – The article proposes a model of a salient pole synchronous machine field winding based on a single transmission line model. An experimental method to derive the parameters is also presented and validated. Finally, the measured voltage distribution in the winding is compared to the model voltage distribution and the results match, demonstrating the model capabilities. The model describes the intrinsic resonance phenomena and accurately determines the voltage amplification factor.

Index Terms—Distributed parameter circuits, eddy currents, parasitic capacitance, resonance, skin effect, stray inductance.

I. INTRODUCTION

FIELD current control opens up new opportunities for controlling the electromechanical behavior of synchronous machines, as shown for example by the Unbalanced Magnetic Pull Compensation technique [1] and Synchronous Motor Start by Rotor Polarity Inversion [2]. However, faster and faster switching converters supplying the field winding pose new challenges for the insulation materials used there, especially for possible uneven turn to turn voltage distribution. This is particularly true in view of the increasing use of fast switching SiC-modules in power converters, enabling both higher DC-link voltage level and steeper commutation edges [3],[4]. This issue is made even more severe by the fact that rotor insulation material in salient pole synchronous machines is designed to cope with relatively low turn-to-turn and turn-to-ground dc voltages, usually not higher than five times the rated values [5]. Starting from the pioneer research of Blume-Boyajian [6] and Rüdenberg [7], many works have highlighted the importance of characterizing correctly the parameters of the windings in both transformers [8], [9], [10] and rotating AC machines [11], [12], [13] in order to give correct account of fast transient phenomena (e.g. atmospheric overvoltage and fast switching) and periodic or pseudo-periodic voltage surges (e.g. operational overvoltage, resonance excitation and ringing). To this end, [14] has first investigated resonance phenomena occurring in the field winding of synchronous machines which are also of interest here, but with focus on wound rotor type. However, not all the works it has been referred to agree on the kind of model that should be used, or on the simplifying assumptions that can be made. In particular:

- [8], [9], [11] make use of Multi Transmission Line Models (MTLM), whereas [12], [13], [14] prefer lumped-circuit models (LCM);
- [10] takes into account the modal dispersion of the travelling wave due to the frequency-dependent speed of propagation;
- [9], [11], [13] consider the inter-turn mutual inductance, while [8], [12], [14] do not;
- [13] deduces both inductance and capacitance matrixes from FEM 2D simulations, whereas [9], [11], [12] derive the L matrix from the calculated C matrix after having postulated Transverse Electromagnetic (TEM) propagation of the travelling waves.

Given the novelty of the internal overvoltage issue related to the excitation winding discussed in this paper, a good compromise between simplicity and reliability of the model is needed in order to rapidly obtain useful simulation results about potential breakdown or partial discharge problems in applications such as the one presented in [1] or [2]. To this end, the assumption of a simple Single Transmission Line Model (STLM) appears to be a reasonable choice at least for two reasons. The first one is illustrated in [15] where the exact calculation of constant capacitance values in a lumped circuit presupposes the knowledge of the voltage distribution, which should be an outcome of the model. The second reason is that the modal dispersion highlighted in [5] turns out to be an all but negligible phenomenon as the delays between simulated and measured data shows in [13].

Considering the difficulties of properly accounting for phenomena such as skin-effect, proximity-effect and eddy currents reaction combined, an experimental approach for determining the distributed parameters was chosen. In that sense, this paper offers an experimental method to determine the stray inductance involved in the model and proposes an innovative simple method to indirectly measure the turn-to-turn and turn-to-ground winding capacitances. The last method makes use of the voltage transfer function analysis in the frequency domain, and can be easily automated for control and diagnosis purposes.

II. THEORETICAL BACKGROUND

A. Single Transmission Line Model

An infinitesimal portion dx of the rotor winding is electrically described by the circuit of Fig. 1.

The distributed parameters, p ($F^{-1}m^{-1}$) and c (Fm^{-1}),

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represent, respectively, the specific turn-to-turn elastance and the specific turn-to-ground capacitance. The specific inductance, l (Hm⁻¹), summarizes all electromagnetic phenomena capable of inducing electromotive forces along the winding. The remaining parameters, r_w (Ωm⁻¹), r (Ωm⁻¹) and g (Sm⁻¹), represent the dissipative effects generated respectively by the conduction current, by the turn-to-turn voltage and by the turn-to-ground voltage.

Fig. 1. Electrical model with distributed parameters for an infinitesimal portion of rotor winding.

According to Heaviside's telegraph line theory, a sinusoidal voltage wave of given angular frequency, ω , propagating through the winding along the progressive x , is expressed by

$$V_x = V_0 e^{-k(\omega)x}, \quad (1)$$

where $k(\omega)$ (m⁻¹) is the complex propagation constant and V_x is the line-to-ground voltage at the generic position x .

The real and the imaginary part of $k(\omega)$, called attenuation constant, $\alpha(\omega)$ (m⁻¹), and phase constant, $\beta(\omega)$ (rad m⁻¹), can be related to the circuital parameters by

$$k(\omega) = \alpha(\omega) + i\beta(\omega) = \sqrt{z(\omega)y(\omega)}, \quad (2)$$

where:

$$z(\omega) = \left(\frac{1}{r_w + i\omega l} + \frac{1}{r} + i\frac{\omega}{p} \right)^{-1}, \quad (3)$$

and

$$y(\omega) = g + i\omega c. \quad (4)$$

B. Resonance modes

The different modes of resonance that can occur in a finite length a of a winding described by STLM have been primarily determined by [6], [7] (by solving the related telegrapher's equations) and by Abetti [16] (through the reduction of Rayleigh's principle to the winding electrical model). The deductive approach used here, for deriving them, makes use of three assumptions:

- during the resonance the losses are fed by the exciting power source;
- the superposition of travelling waves in standing waves is possible;
- the perturbation, $\delta V(x)$, of the low frequency voltage linear distribution, must be zero at the beginning and at the end of the winding (Fig. 2).

Fig. 2. Voltage perturbation of the linear low frequency voltage distribution.

From a) follows that during resonance the periodic energy exchange between inductances and capacitances can be regarded as free oscillation in a lossless circuit. In that case the propagation constant becomes purely imaginary and its module coincides with the phase constant:

$$k(\omega) = i\beta(\omega) = i \sqrt{\frac{pc}{\omega^2 l}}. \quad (5)$$

By combining assumptions b) and c) it can be stated that for the m -th mode of resonance, m half waves of $\delta V(x)$ must be generated between the extremes of the lines for a total of $m+1$ nodes (Fig. 3).

Fig. 3. First three stationary modes for the voltage perturbation $\delta V(x)$.

For the m -th mode of resonance at the angular frequency ω_m it must therefore be true that

$$\beta(\omega_m) = \sqrt{\frac{pc}{\omega_m^2 l}} = m \frac{\pi}{a}, \quad (6)$$

which leads straight to the sequence of all possible resonance frequencies compatible with constraint c)

$$\omega_m = \frac{1}{\sqrt{\left(\frac{a}{m\pi}\right)^2 lc + \frac{l}{p}}}, \quad (7)$$

where $m=1, 2, 3, \dots$.

The highest possible resonance frequency ω_∞ (limit resonance frequency) is obtained for $m \rightarrow \infty$, yielding

$$\omega_\infty = \sqrt{\frac{p}{l}}. \quad (8)$$

The voltage perturbation resonating according to (8) must have a propagation speed equal to zero due to its vanishing wavelength. From (7) and (6) it follows that the propagation speed depends on the frequency according to

$$c_m = \frac{\omega_m}{\beta(\omega_m)} = \frac{1}{\sqrt{lc + \frac{l}{p} \left(\frac{m\pi}{a}\right)^2}}. \quad (9)$$

Equation (9) proves that the adopted model takes into account the modal dispersion. It shows also that the TEM propagation mode, at speed

$$c_0 = 1/\sqrt{lc}, \quad (10)$$

is applicable without error only at very slowly varying perturbations. A refractive index of the winding n_m (always greater than 1) can be here defined as

$$n_m = \frac{c_0}{c_m} = \sqrt{1 + \frac{1}{pc} \left(\frac{m\pi}{a}\right)^2}. \quad (11)$$

C. Model inductance

In [6], the l of Fig. 1 is shown to be the specific stray inductance of the winding since the principal flux is assumed to not vary along the x -axis. Even though [11], [17] give a correct account of the turn-to-turn mutual coupling [7], [11] show that its contribution to the specific inductance of the winding is negligible due to the eddy currents formed in the main flux path. It is possible to show that also the stray inductance is affected by the eddy currents induced in the iron

lamination, in the damper bars (if present) and in the stator winding. Fig. 4 shows the geometry of a 12 poles, 110 kW, 400 V synchronous generator rotor pole and winding.

Fig. 4. MMF distribution, Θ , between two adjacent poles under DC excitation.

The principal dimensions of the previous figure are shown in Tab. I.

TABLE I ROTOR POLE PRINCIPAL DIMENSIONS		
Symbol	Description	Measure
r_i	coil initial radius	212 mm
r_f	coil final radius	346 mm
h	pole tip height	10 mm
b	coil width	28 mm
w	pole-core width	74 mm
w_{tip}	pole-shoe-width	130 mm
l_a	axial rotor length	303 mm
γ	angular pole-pitch	30°
N	turns per pole	162
$a/12$	pole winding length	136.42 m
A	wire cross-section	11.2 mm ²

By following the approach of [18], which assumes the MMF, Θ , increasing linearly in radial direction as long as the integration path in Ampere's law links a portion of the excitation windings, the stray inductance can be calculated as

$$L = 4\mu_0 l_a \frac{1}{N^2} \int_{r_i}^r \frac{N^2 (r-r_i)}{s(r-r_i)} d(r-r_i) = 4\mu_0 l_a N^2 \lambda, \quad (12)$$

where the specific permeance $\lambda (-)$ is equal to

$$\lambda = \frac{1}{N^2} \int_{r_i}^r \frac{N^2 (r-r_i)}{s(r-r_i)} d(r-r_i). \quad (13)$$

By considering the specific permeances for intrapolar vane λ_{core} and for pole tip λ_{tip} separately, it can be obtained

$$\lambda_{core} = \frac{1}{\gamma} \left[\frac{1}{2} - \xi_{core} + \log \left(1 + \frac{1}{\xi_{core}} \right) \xi_{core}^2 \right] \quad (14)$$

and

$$\lambda_{tip} = \frac{1}{\gamma} \log \left(1 + \frac{1}{\xi_{tip}} \right), \quad (15)$$

where

$$\xi_{core} = \frac{r_i}{(r_f-r_i)} - \frac{w}{(r_f-r_i)\gamma} \quad (16)$$

and

$$\xi_{tip} = \frac{r_f}{h} - \frac{w_{tip}}{h\gamma}. \quad (17)$$

The pole stray inductance for unsaturated iron and for infinitely slowly varying current can be then estimated by (12)

$$L = 4\mu_0 l_a N^2 (\lambda_{core} + \lambda_{tip}) = 27.9 \text{ mH}. \quad (18)$$

As soon as the excitation current varies rapidly, eddy currents are induced in the laminated iron of the rotor and the

stator as well as in the copper circuits. It can be demonstrated that already at 10 kHz the stator iron (M235-35) is not fully penetrated due to the skin effect. Considering the specific resistivity of the iron being $\rho_{iron} = 480 \text{ n}\Omega\text{m}$ and assuming a reasonable relative permeability $\mu_r = 100$ for the unsaturated core lamination, the penetration depth δ becomes:

$$\delta_{iron}(10\text{kHz}) = \sqrt{\frac{\rho_{iron}}{\pi f \mu_0 \mu_r}} \cong 0.35 \text{ mm}. \quad (19)$$

Considering the copper damping bars instead with $\rho_{copper} = 17 \text{ n}\Omega\text{m}$ and relative permeability $\mu_r \cong 1$:

$$\delta_{copper}(10\text{kHz}) = \sqrt{\frac{\rho_{copper}}{\pi f \mu_0}} \cong 0.69 \text{ mm} \quad (20)$$

Under these conditions, the eddy currents can be regarded as infinite distributed secondary circuits of a transformer closed on resistive loads, which turn out to be all in parallel. Then, the excitation winding experiences a current which is essentially the primary short-circuit current, the MMF of which must be completely balanced by all secondary currents in order to keep the main flux constant. Fig. 5 assumes the reaction of all secondary currents (marked in green) as uniformly distributed along the main flux path.

Fig. 5. MMF distribution, Θ , between two adjacent poles under AC excitation.

In the same figure it can be observed that the MMF, Θ , responsible for the intrapolar flux is now different from the one shown in Fig. 4., in spite of the unchanged maximal intensity of the excitation current I . Θ starts negative (blue shaded area), experiences an inversion point and then it grows positive (red shaded area). Since the inductance is defined as the ratio between the linked flux Ψ and the current I , where

$$\Psi \sim N\Theta, \quad (21)$$

the stray inductance due to the resultant MMF in presence of reaction currents, $\Theta_{r.c.}$, is then

$$L_{r.c.} = \frac{\Psi_{r.c.}}{I} = \frac{\Psi}{I} \frac{N\Theta_{r.c.}}{N\Theta} = L \frac{\Theta_{r.c.}}{\Theta}. \quad (22)$$

The principal dimensions of the main flux path represented in Fig. 5 are shown in Tab. II.

TABLE II PATHS LENGTHS				
Symbol	s_1	Δr_2	Δr_3	s_4
Measure	72.3 mm	134.0 mm	72.0 mm	180.2 mm

The result of using the simplistic assumption of reaction currents MMF uniformly distributed along the main path is

$$\Theta_{r.c.} = \Theta - \Theta \frac{s_1 + 2\Delta r_2}{s_1 + 2\Delta r_2 + 2\Delta r_3 + s_4} = \Theta \frac{2\Delta r_3 + s_4}{s_1 + 2\Delta r_2 + 2\Delta r_3 + s_4}. \quad (23)$$

By substituting (23) in (22) and considering Tab. II, the following estimate of the stray inductance is obtained:

$$L_{r.c.} = L \frac{2\Delta r_3 + s_4}{s_1 + 2\Delta r_2 + 2\Delta r_3 + s_4} \cong 13.6 \text{ mH} \quad (24)$$

The performed estimation (24) shows that the reaction currents cause a reduction of the stray inductance (18).

The theory of operational impedances on the other hand [19], [20] assumes a constant asymptotic value for the excitation winding inductance when $f \rightarrow \infty$ as seen in Fig. 6.

Fig. 6. Contributions to the winding inductance at high frequency.

In this case, the subtransient inductance seen from the field winding terminals depends on armature stray inductance $L_{\sigma,a}$, excitation winding stray inductance $L_{\sigma,f}$, damping circuit stray inductance $L_{\sigma,D}$, magnetization inductance L_m and mutual inductance between damping bars and field winding $M_{D,f}$. In [21] is shown that in salient pole synchronous machines with damping bars and massive rotor poles $M_{D,f} < L_m$, which explains from another point of view why $L_{r.c.}$ can be smaller than $L_{\sigma,f}$ when reaction currents arise.

D. Model AC conduction resistance

The analytical estimation of the AC field winding resistance, made by considering the conductor cross section determined only by the current penetration depth, is shown in Fig. 7.

Fig. 7. Frequency dependence of the winding AC-resistance.

Even though the proximity effect is not taken into account there, it is possible to observe that, already at 100 Hz, the winding reactance, ωl_a , is at least one order of magnitude bigger than the conduction resistance, $r_w(\omega)a$, independently from the shape of the wire cross-section (rectangular or circular). The specific resistance, r_w , in the model of Fig. 1, is therefore negligible, as soon as f is greater than 100 Hz.

Fig. 9. Supplying circuit a), its equivalent representation b) and dynamic circuit without DC-bias.

III. METHOD

A. Determination of the specific stray inductance

In order to find an experimental estimate of the field winding stray inductance for rapid variation of the supply voltage, the current response of the entire excitation circuit (12 poles) to a steep edge step voltage has been measured. In Fig. 8 the applied voltage and the resulting transient field current are depicted together with a polynomial fit for the current plot.

Fig. 8. Excitation current response to a fast edge step voltage.

The 15-th order polynomial fit has the following expression:

$$i_{fit}(t) = \sum_{i=0}^{15} a_i t^i. \quad (25)$$

Imagining that in a very narrow right neighborhood of $t=0$ the current can be approximated by

$$i_{app}(t) = \frac{V_{DC}}{R^*} \left(1 - e^{-\frac{R^*}{L^*}t} \right), \quad (26)$$

it must hold true that

$$\left. \frac{di}{dt} \right|_{t=0} \cong a_1 = \frac{V_{DC}}{L^*} \quad (27)$$

and, at the same time, that

$$\left. \frac{d^2i}{dt^2} \right|_{t=0} \cong 2a_2 = -\frac{U_{DC}}{R^*} \left(\frac{R^*}{L^*} \right)^2. \quad (28)$$

By substituting $a_1 = 95.3754 \text{ As}^{-1}$ and $a_2 = -6.6422 \cdot 10^4 \text{ As}^{-2}$ from the polynomial fit and $V_{DC}=12 \text{ V}$ in (26) and (27), it results:

$$R^* \cong 175.25 \Omega \text{ and } L^* \cong 125.8 \text{ mH}. \quad (29)$$

The estimated subtransient asymptotic field winding inductance for a pole is therefore

$$L_{r.c.}(\infty) = \frac{L^*}{12} \cong 10.5 \text{ mH} \quad (30)$$

and the specific one for unit wire length

$$l = \frac{L_{r.c.}(\infty)}{a/12} \cong 77.0 \mu\text{Hm}^{-1}. \quad (31)$$

B. Determination of the winding specific capacitances

Consider the square-wave inverter of Fig. 9a. When a series of $2n$ poles is supplied by it, the middle point O becomes a node of fixed potential $V_{DC}/2$ as shown in Fig. 9b.

The voltage distribution of Fig. 2 is produced on each half of the n poles circuit. The two distributions are antisymmetric with respect to O due to the signs of the generators given in Fig. 9c. Fig. 10 shows the first mode of resonance on a set of 6 poles, achieved by supplying the entire series of 12 poles according to the concept of figure 9a.

Fig. 10. Antisymmetric fundamental resonance over half rotor wheel.

Fig. 11 shows instead the first mode of resonance for a series of 3 poles achieved by supplying a chain of only 6 poles according to the same concept.

Fig. 11. Antisymmetric fundamental resonance over $1/4$ of the rotor circuit. By making use of (7) at the resonance shown in Fig. 10 it must be true that:

$$\omega_{1,1/2} = \frac{1}{\sqrt{\left(\frac{a/2}{\pi}\right)^2 lc + \frac{l}{p}}} \quad (32)$$

In the same way, for the resonance shown in Fig. 11 the following identity must be fulfilled:

$$\omega_{1,1/4} = \frac{1}{\sqrt{\left(\frac{a/4}{\pi}\right)^2 lc + \frac{l}{p}}} \quad (33)$$

By considering the fundamental harmonics of both V_{in} and V_{out} in the measurement settings of Fig. 10 and Fig. 11, the following transfer function can be defined:

$$A(\omega) = \frac{V_{out,1}(\omega)}{V_{in,1}(\omega)} \quad (34)$$

Fig. 12 represents the magnitude of (34) obtained by measuring V_{in} and V_{out} in both named settings at different switching frequencies.

Fig. 12. Voltage transfer function modules related to the measurement settings of Fig. 10 and Fig. 11.

In Fig. 12 it is easy to detect the resonance frequencies in both cases, given the good selectivity of the obtained bells. Their values are reported in Tab. III

TABLE III PATHS LENGTHS		
	12 - poles	6-poles
Resonance frequency	43.4 kHz	73.0 kHz

Once the resonance frequencies are known it must be stressed that c and p are the only unknowns to be found in (32) and (33). Their values can therefore be calculated, and were found to be

$$c = 2.2 \text{ pFm}^{-1} \text{ and } p = 41.5 \text{ }\mu\text{F}^{-1}\text{m}^{-1}. \quad (35)$$

IV. RESULTS

A. Determination of the specific stray inductance

Two measurements on the series of six rotor poles led by a Q-meter and shown in Tab. IV confirm that the stray inductance is reduced by the effect of reaction currents.

TABLE IV PATHS LENGTHS				
f (Hz)	$L_{6 \text{ poles}}$ (mH)	R_6 (Ω)	Q (-)	$L_{1 \text{ pole}}$ (mH)
100	158.0	62.7	1.59	26.3
1000	73.6	263	1.75	12.3

As shown by Tab. IV at low frequency (< 100 Hz) the stray inductance converges to (18) but for higher frequency (> 1000 Hz) it tends to (30).

B. Comparison with analytically calculated capacitances

The structure of the excitation winding consists of $q=5$ layers of 2.8 mm x 4.0 mm copper conductors spaced by insulation material as shown in Fig. 13.

Fig. 13. Winding cross section showing the insulation coordination.

Assuming the insulation material having everywhere an effective permittivity $\epsilon_r=2$ and with $d_l=2$ mm, $d_g=5$ mm and $d_i=0.18$ mm, by [22] it is possible to find:

$$c \cong \frac{\epsilon_0 \epsilon_r 2(l_a + w)(r_f - r_i)}{d_g a_{12}} = 2.6 \text{ pFm}^{-1} \quad (36)$$

and

$$p \cong \frac{3q^2}{4(q-1)} \frac{d_l}{\epsilon_0 \epsilon_r 2(l_a + w + 2b)(r_f - r_i) a_{12}} = 33.4 \text{ }\mu\text{F}^{-1}\text{m}^{-1} \quad (37)$$

(36) and (37), with all limitations of simple analytic estimations, confirm the order of magnitude of the specific capacitance values, found by the proposed experimental method in (35).

C. Voltage distribution profiles

In steady state, and by considering the line representing six poles in series short-circuited in O, (1) becomes

$$v(x) = \frac{V_x}{V_0} = \frac{\sin[(\alpha + j\beta)(\frac{a}{2} - x)]}{\sin[(\alpha + j)\frac{a}{2}]} \quad (38)$$

From (38) it is possible to determine the values of α and β , once the voltage profiles at the frequencies (32) and (8) are known:

$$f_{1,a/2} = \frac{1}{2\pi \sqrt{(\frac{a/2}{\pi})^2 lc + \frac{l}{p}}} = 43.4 \text{ kHz} \quad (39)$$

and

$$f_{\infty} = \frac{1}{2\pi} \sqrt{\frac{p}{l}} = 116.8 \text{ kHz}. \quad (40)$$

Table V offers the values of real and imaginary parts for the propagation constant at the frequencies (39) and (40), found by fitting (38) with the measured voltage profiles of Fig. 14 in two points, except the extremal ones. Once the propagation constant for a given frequency is known, by using (2), (3) and (4) the parameters r and g for the model can be algebraically derived.

TABLE V
PATHS LENGTHS

$f(\text{kHz})$	$\alpha(\text{m}^{-1})$	$\beta(\text{rad m}^{-1})$	$r(\Omega \text{ m}^{-1})$	$g(\text{S m}^{-1})$
43.4	$9.27 \cdot 10^{-4}$	$36.36 \cdot 10^{-4}$	61.77	$7.83 \cdot 10^{-8}$
116.8	$95.93 \cdot 10^{-4}$	$93.84 \cdot 10^{-4}$	110.52	$3.60 \cdot 10^{-8}$

The obtained values for the dissipative parameters shown in Tab. V offer a first rough estimation of their dependency on the frequency:

$$r(f) \cong r(f_{1,1/2}) \sqrt{\frac{f}{f_{1,1/2}}} \propto \sqrt{f} \quad (41)$$

and

$$g(f) \cong g(f_{1,1/2}) \frac{f_{1,1/2}}{f} \propto \frac{1}{f}. \quad (42)$$

Figure 14 shows the good fit between the simulated data provided by the STLM-model and the voltage distributions obtained by considering half of the voltages measured between the antisymmetric poles of Fig. 10. The measurements, which relate to a machine available at the Electricity Division, already characterized in figure 4 and in table I, have been performed at the at low frequency (9 kHz), at the first resonance mode (43.4 kHz) and at the limit resonance frequency (116.8 kHz).

Fig. 14. Voltage distribution profile at canonical frequencies.

Figure 14 also shows that the highest surge in the winding ($\sim 1.3 V_{in}$) happens at the first resonance mode close to the middle point of the six poles series (actually internally to the third pole). But the most important achievement is the possibility to observe the strongest turn to turn voltage gradients which happen at the limit resonance frequency (40)

at the beginning of the line ($e_{i,max,0}$) and for the first mode resonance frequency (39) at the end of the line ($e_{i,max,a/2}$), as it is shown in Fig. 15.

Fig. 15. Voltage gradients distributions at canonical frequencies.

Called l_t the length of a single turn, it results:

$$e_{i,max,0} = \frac{1}{V_{in}} \frac{\Delta V_0}{d_t} = -\frac{l_t}{d_t} \sqrt{pc} = -44.7 \text{ m}^{-1}. \quad (43)$$

and

$$e_{i,max,a/2} = \frac{1}{V_{in}} \frac{\Delta V_{a/2}}{d_t} = -\frac{l_t}{d_t} \frac{\sqrt{\alpha^2 + \beta^2}}{\sqrt{\sinh^2(\alpha a/2) + \sin^2(\beta a/2)}} = -21.0 \text{ m}^{-1}. \quad (44)$$

Equation (43) refers to the classical lightning-stroke voltage gradient [6] here achieved by an investigation in the frequency domain. Equation (44) represents a voltage gradient due to a winding natural-frequency (eigenfrequency).

V. CONCLUSIONS

The investigation described in the paper highlights that the Single Transmission Line Model is able to accurately reproduce voltage distributions and voltage gradients in the excitation winding of a salient pole synchronous generator once its distributed parameters are correctly determined. Two experimental methods, easy to implement and automatize, have been suggested in order to determine the value of the specific stray inductance and the specific series and parallel parasitic capacitances of the winding. The winding stray inductance, obtained by the first method, has proven to be consistent with its value obtained through direct measurement. The distributed parasitic capacitance values found out by the second method have shown to be coherent with their analytical estimations. The STLM winding model based on them correctly takes into account the modal dispersion in the propagation phenomena so that the simulated results obtained by it are in good agreement with the experimental results. The simulation of the winding voltage profile, at its first natural resonance frequency and at its limit resonance frequency, have shown the highest voltage gradients respectively at the end and at the beginning of the rotor circuit. The theory of STLM applied in here has proven to be able to predict both of them. Furthermore, the turn-to-iron core relative voltage surge, detected at the first resonance frequency for a salient pole rotor, has been shown to have a lower amplitude than the one reported by [14] for a wound rotor type. Finally, this investigation has offered some insight into the frequency dependency of the dissipative parameters used in the model, but some further and deeper understanding of that must be gained by future work.

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(*) *Calculation of Electrical Machines*

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