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Analysis of transformer frequency response deviations using white-box modelling

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SUMMARY

The white-box models are being used for many years in transformer design for simulation of impulse transients in the windings of power transformers and shunt reactors. In the recent years these models have been increasingly used in the studies of transformers interaction with an external network: in the network design process and for the analysis of failure accidents caused by the resonance phenomenon in transformer windings. At the present time, suggestions are increasingly being made about the appropriateness of creating tools and interfaces for the transfer of white-box models by transformer manufacturers to network design engineering and electric grid companies for the analysis of transient phenomena in the networks with transformers. Another application of the white-box models is the calculation of the frequency response characteristics of transformers (at high frequencies) for research and qualitative analysis of the effect of transformer internal faults and damages on frequency response. They can be used for the Frequency Response Analysis (FRA) results interpretation.

The report describes the white-box model used by the authors. The problems of calculating the parameters of white-box models and their adjustment for taking into account the effect of winding deformations and other transformer damages are considered. Problems related to the integration and application of models in EMTP-type tools for studying the interaction of a transformer with an external network, modelling of the FRA-measurement scheme and calculation of frequency responses are discussed.

The report presents examples of frequency response deviations for the case of shield with a floating potential. The obtained deviations of the frequency responses are explained.

KEYWORDS

Power transformers, high-frequency modelling, white-box model, winding condition assessment, frequency responses, Frequency Response Analysis, FRA interpretation.

1. INTRODUCTION

The white-box models [1] are being used for many years in transformer design for simulation of impulse transients in the windings of power transformers and shunt reactors. In the recent years these models have been increasingly used in the studies of transformers interaction with an external network: in the network design process and for the analysis of failure accidents caused by the resonance phenomenon in transformer windings.

At the present time, suggestions are increasingly being made about the appropriateness of creating tools and interfaces for the transfer of white-box models by transformer manufacturers to network design engineering and electric grid companies for the analysis of transient phenomena in the networks with transformers.

Another application of the white-box models is the calculation of the frequency response characteristics of transformers (at high frequencies) for research and qualitative analysis of the effect of transformer internal faults and damages on frequency response. They can be used for the Frequency Response Analysis (FRA) results interpretation.

Practically it is convenient to analyse white-box models and their frequency responses in EMTP-type software, where transformer model can be interconnected to take into account taps and leads connections as well as measurement scheme.

Besides advantages of white-box models for frequency-response analysis there are still a lot of issues to be solved in order to make white-box models more flexible and accurate.

Report describes specific issues of white-box models application and adjustment and presents the examples of their usage to analyse typical differences between LV and HV windings frequency responses and distortions caused by certain fault types such as missed shield earthing (or missed core earthing).

2. WHITE-BOX MODELLING SPECIFICS

The purpose of any white-box model is to reproduce the transformer electromagnetic behavior in computer simulations. These models are parametrized according to the design data of the transformer so they are usually developed and used by transformer manufacturers. For the first time, they had been developed for the simulation of impulse transients inside transformer windings when the standard lightning impulse test voltage is applied, in order to appropriately design the transformer insulation.

In the heart of any white-box model there is a system of differential (or integro-differential) equations which can be written on the basis of different physical interpretations – the transformer can be modelled as a system with distributed parameters (single- or multiconductor transmission line models [2]), or as a network with lumped self and mutual R, L and C parameters. The latter approach became widely used among the transformer designers [1]. This has historical and practical reasons – it is easier to manually tune the lumped parameters model by including additional or by correcting existing capacitances or other elements in the model. In this report the lumped parameters white-box model is discussed.

To calculate R-L-C parameters of white-box models the software dedicated to impulse transient simulations in transformer windings can be used. There are several approaches existing to calculate the model parameters and today there is still no clear understanding of which one is the best. It is within the scope of currently active CIGRE JWG A2-C4.52 “High-Frequency Transformer and Reactor Models for Network Studies” to identify the best practices for parameters calculation. In this report we briefly summarize the approaches that we use in the software of our own development.

Today it is clear that self and mutual inductances of disks and turns should to be calculated with the magnetic core taken into account. In our calculations we use the analytical method developed in [3], in which the Poisson equation for vector potential is solved assuming the magnetic core is infinitely high with a circular cross-section. Yokes are omitted so the problem can be solved in 2D with axial symmetry. This approach shows a more stable behavior in mutual inductances calculations than the well-known Rabins’ method [4] and its accuracy was confirmed many times by comparison of calculated and measured transformer responses to impulse excitation.

The calculation of the capacitances is more straightforward but tricky in details. We use an approach based on the electrostatic power conservation principle with an assumption of linear voltage distribution along a disk, which is explained in details in [5]. A very good introduction to the calculation of winding disk capacitances with wound-in-shields is presented in [6].

The accurate calculation of resistances is important to reproduce damping of transients and transformer behavior in resonance conditions. In the FRA simulations resistances mostly affect peaks in frequency responses. The resistances in the transformer equivalent circuit are strongly frequency dependent, but calculation of this dependency constitutes one of the most complicated problems in transformer numerical modelling. The simplified approach consists in the calculation of the resistances at the power frequency with further multiplication of the obtained values by the k-factor [7] accounting for the increase of losses at high-frequencies. In this report we will show that this approach allows getting the estimations suitable for the analysis of transformer frequency responses.

After the R-L-C parameters have been calculated the frequency response calculations can be performed. The software program developed by authors of this report has an option to export the model parameters to other simulation packages. For the network studies we developed another program which features a detailed transformer model, initialized by the R-L-C parameters of transformer equivalent circuit accompanied by information on their connections.

Typically these parameters are calculated for disc-by-disc winding representations, but in the case of complicated interconnections of turns within disks (non-standard interleaving schemes etc.) the turn-by-turn winding representation is required. This representation results in significant increase of the model size and respectively of the number of nodes in the equivalent circuit. Not all of these nodes are necessary to be shown in the simulator, so the procedure of the data import into the program implicates the labeling of the nodes to be shown to the user. These labeled nodes are accessible for the connections with external circuit and with virtual measuring tools.

So the white-box model is completely defined by five matrices namely R, L and C, the matrix of interconnections T, and the matrix of labeled nodes N.

It needs to be mentioned that the formulae generally used for the calculation of R-L-C parameters are based on the assumptions such as winding axial symmetry and vertical alignment of winding conductors that do not allow simulating local deformations of windings such as buckling and tilting. One of possible approaches to take into account the impact of winding local deformations is to use FEM software to evaluate correction factors which can be used to fix white-box model parameters. For instance, the rotation of winding wires caused by tilting yields to changing of the capacitances between adjacent turns and adjacent discs and corresponding correction factors can be obtained from FEM simulation of the capacitance matrix elements for tilted and not tilted discs. The correction factors for the elements of the inductance matrix can be obtained similarly.

3. MODELLING OF WINDINGS FREQUENCY RESPONSES

Transformer windings frequency responses can be evaluated from white-box model using approach described in [8, 9]. Summarizing the above mentioned white-box modelling specifics it is based on a combined application of two types of software programs:

1) Software for generation of high-frequency models of windings based on the equivalent circuits with lumped parameters which further can be imported via R-L-C matrices.

2) EMTP-type tools to connect the transformer windings with measuring circuit and to calculate the frequency responses and transfer functions using a frequency analyzer simulation.

It should be noted that white-box models based on equivalent circuits with lumped parameters which were originally intended for the simulation of impulse overvoltages in transformer windings are based on a number of assumptions, as a result of which the frequency response characteristics determined by white-box models differ a priori from the measured responses. Although this is not a significant impediment to using white-box models for qualitative analysis of the influence of various factors on frequency responses there is a need for further development of the white-box models for more accurate calculation of the frequency characteristics and transfer functions of the transformer windings.

One of the key reasons for the difference between calculated and measured frequency responses is that the frequency dependence of the losses in the windings is not taken into account. This is due to the fact that in the calculation of impulse overvoltages it is not necessary to reproduce the transient attenuation with high accuracy. In the majority of white-box models the values of the winding element resistances (the R-matrix) are determined at some fixed frequency. Thus, the frequency dependence of winding losses is not taken into account despite the substantial increase in losses with increasing frequency. When using constant R-matrices, attenuation at high frequencies is not sufficient to reproduce peaks in frequency responses correctly. But increasing the values of R-matrix elements for more accurate behavior at high frequencies will lead to unnecessary losses and lower accuracy at low frequencies.

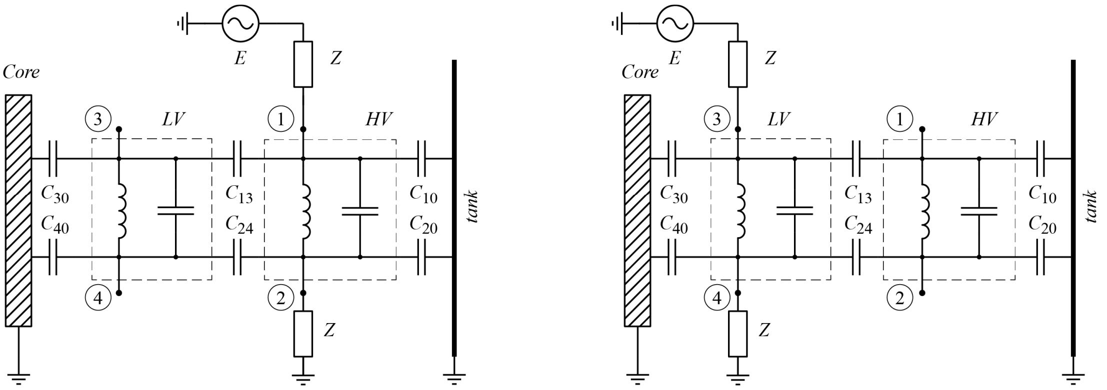
4. EXAMPLES OF WHITE-BOX APPLICATION

4.1. The difference between frequency responses of inner and outer windings

Frequency responses of LV and HV transformer windings have some basic differences:

- – for LV winding the characteristic typically goes higher, especially at low and medium frequencies, which is caused by a smaller number of turns and inductance and by a smaller electric length of LV winding compared to HV winding;
- – LV windings are typically located closer to the magnetic core and this proximity determines the fundamental differences in the frequency response characteristics of internal and external windings: the outer windings often have a pronounced V-shaped frequency response.

Let's consider the reasons of differences in the frequency responses of inner and outer windings. The specific distinctions can be explained using a simplified equivalent circuit (Fig. 4.1.1), in which the capacitance of LV winding to the magnetic circuit (magnetic core shield), the capacitance between LV and HV windings and the capacitance of HV winding to the tank are divided into parts and connected to the terminals of these windings: $C_{35} = C_{45}$; $C_{13} = C_{24}$; $C_{10} = C_{20}$.



a) *b*)

Figure 4.1.1 – Simplified equivalent circuit of two winding transformer when measuring the frequency response of HV (*a*) and LV (*b*) windings

Outer HV windings, as a rule, have large insulation distances to earthed parts and smaller capacitances to ground compared to LV windings. The electrostatic coupling between HV windings and the ground turns out to be weaker, and there is a strong electromagnetic coupling with the neighboring LV winding. When measuring the frequency response of the HV winding (Figure 4.1.1a), the voltage is applied to node 1, and impedance 50 Ω of the measuring device is connected to node 2, which is practically equivalent to earthing of this node. The voltage on the LV winding, which is coupled with the HV winding, is induced. The terminals of the LV winding are not earthed during the measurement, they remain under floating potential. In the general case, due to low inductance and low series capacitance of LV winding the following voltage ratios hold: $U_1 > U_3$ and $U_2 < U_4$. The last relation determines the current direction in the branch 2–4: the current is directed towards the node 2 and flows further to ground through a relatively small impedance of the measuring unit. Thus, in the current through the measuring impedance in addition to the component associated with the HV winding current, there is also an additional capacitive component caused by the current flowing to ground from the LV winding. This capacitive component determines the rising of the frequency response characteristic at medium and high frequencies.

The situation is different with the LV winding when measuring its frequency response, the voltage is applied to node 3, and the impedance of 50 Ω is connected to node 4 (Figure 4.1.1b). Typically, for two-winding transformers the capacitances C_{13}, C_{24}, C_{10} and C_{20} have comparable values. The HV winding at high frequencies has big impedance. So, the typical voltages relation is: $U_3 > U_1$, $U_4 > U_2$ or $U_4 \approx U_2$. The current in branch 2–4 is directed towards the node 2 or is practically absent, so the significant capacitive current from the winding HV is not flowing to the ground through the measuring impedance. Thus, the tendency of the LV winding frequency response characteristic to rise in the high-frequency region is not noticed or not explicitly expressed.

This effect can be illustrated by the results of measurements on a physical model with two identical coil windings (Figure 4.1.2) placed on separate magnetic cores. Each winding contains 54 coils with 7 turns each, the mean diameters of the windings are about 430 mm and the heights are about 900 mm. The frequency response of a single winding, measured according to the standard FRA measurement scheme, is shown in Figure 4.1.3. Its behavior is representative for the inner winding: a decrease in the low-frequency region, the presence of resonance peaks in the typical for the natural frequencies of the winding oscillations frequency range (in this case, 160 kHz and above) and the absence of an explicitly pronounced trend to increase with frequency.

If the neutral end of the winding is earthed, and the ground of the measuring circuit is decoupled from the ground of the measurement object as proposed in [10] for measuring the diagonal elements of the admittance matrix, the frequency response will take the form of a V-shaped curve which is typical for external high-voltage windings. Figure 4.1.3 shows that in the red curve (Y_{AA}) the individual resonant peaks corresponding to the natural oscillation frequencies of the winding become smoother, and the high-frequency part of the curve has a pronounced capacitive behavior, which manifests itself in admittance increasing with frequency.

Figure 4.1.2 – Exterior view of the model

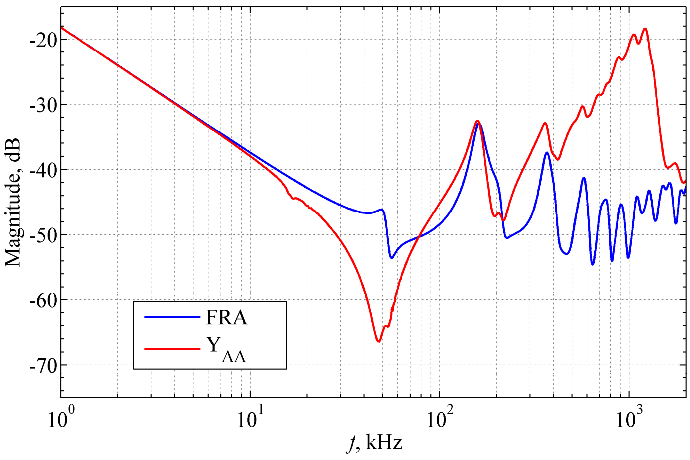


Figure 4.1.3 – The measured frequency response of the winding

It is known that measurements on physical models have their limitations on the possible combinations of the parameters of the models under investigation, and the use of numerical modelling allows removing these limitations. Results similar to the above can be obtained with a white-box model inside of EMTP-type program. Figure 4.1.4 shows the computer model of the admittance Y_{AA} measurement circuit. In this model the voltage is applied to the measurement object through an ideal transformer, which allows the ground of the measurement object to be decoupled from the voltage source ground and to provide a return of the total current through the measuring impedance R3. Figure 4.1.5 shows the calculated frequency response characteristics for the standard FRA circuit and the scheme for measuring the diagonal elements of the admittance matrix.

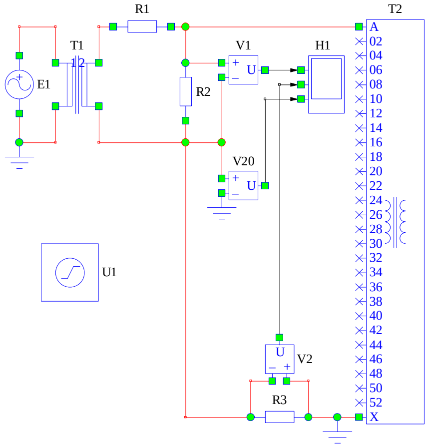


Figure 4.1.4 – The computational model, assembled according to the Y_{AA} measurement scheme with unearthed power supply side

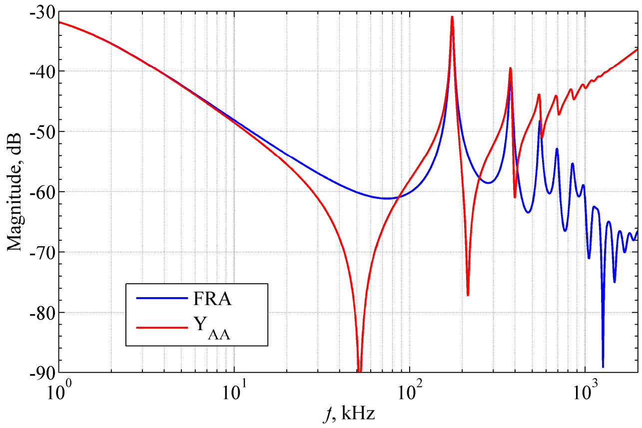


Figure 4.1.5 – Calculated frequency response characteristics

4.2. Frequency response characteristics in the presence of unearthed shields

The above effect with an additional capacitive component in the measured frequency response can also be caused by the appearance of shields with a floating potential or the unearthing of the magnetic circuit. In service the cases of internal damages to transformers which lead to the appearance of a floating potential on electrostatic shields are not uncommon. This can be caused by burnout and breakage of the shield earthing conductor, for example, if the current of the internal short-circuit fault flows to the shield and further to the grounded magnetic circuit. Also, the appearance of a shield floating potential can be caused by a poor contact of the grounding conductor, sparking and subsequent conductor burnout. In addition, there are cases when FRA measurements are taken after other diagnostic measurements performed with the magnetic core or shields being ungrounded (for example, measuring of the magnetic circuit or shields insulation resistance). By mistake the magnetic circuit or core shields may be left unearthed, which affects the obtained frequency response characteristics. When shields with a floating potential appear there is a significant change in the frequency response characterized, as a rule, by its shift up or down in a wide frequency range and by the displacement of individual resonance frequencies in the mid and high frequency region. When applying approaches based on correlation analysis, for example, DL/T 911 [11], such changes in frequency response characteristics can be interpreted as serious winding deformation (severe or obvious deformations in terms of DL/T 911), although, in fact, the measured winding may not have any deformation.

Below are the results of white-box modelling of a two-winding transformer having a shield around a magnetic core that was grounded or left under floating potential. Figure 4.2.1 shows the calculated frequency response characteristics for a two-winding generator step-up transformer 500 kV, obtained with the help of the white-box model and the approach described above.

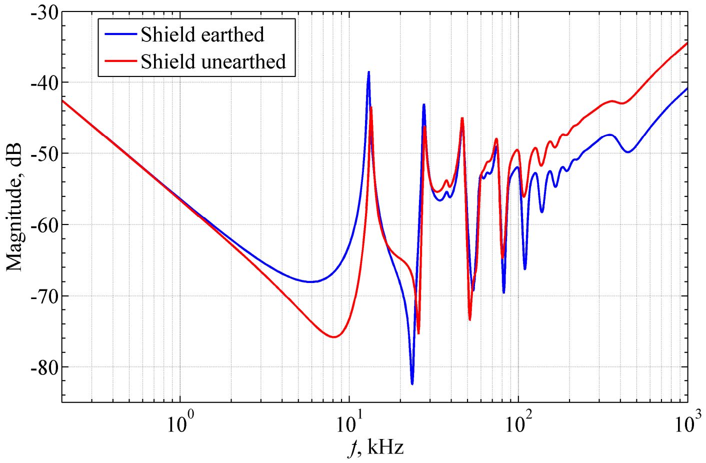


Figure 4.2.1 – Calculated frequency response of HV winding

It can be seen from Figure 4.2.1 that when the shield was unearthed, the natural frequencies of the HV winding remained practically unchanged, while the frequency response at frequencies of the order of hundreds of kHz shifted up by 4–6 dB.

The results obtained can be explained using a simplified equivalent circuit (Figure 4.2.2) where C_{50} is the capacitance of the shield to the magnetic circuit (which is earthed); $C_{50} >> C_{13}$; the remaining symbols are the same as in Figure 4.1.1.

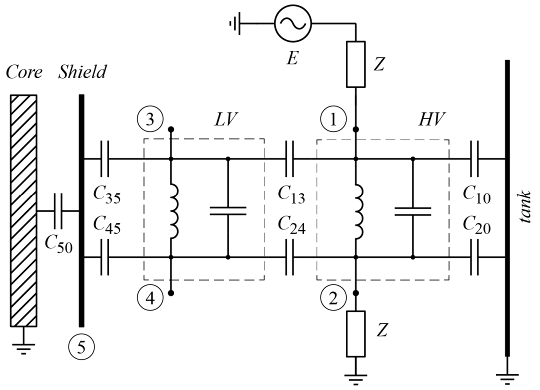


Figure 4.2.2 – Simplified equivalent circuit of two winding transformer with unearhed shield

Consider the case when C_{35} and C_{45} are comparable in value to C_{13} and C_{24}. When measuring the frequency response of the HV winding according to the standard FRA scheme the voltage is applied to node 1 and impedance Z = 50 Ω of the measuring device is connected to node 2 which is practically equivalent to grounding of this node. The voltage is induced in the LV winding which has an electromagnetic coupling with the HV winding. The terminals of the LV winding during the measurement are not grounded, they remain under floating potential. In the general case the following voltage relations are fulfilled: $U_1 > U_3$ and $U_2 < U_4$. The last relation determines the current direction in the branch 2–4: the current is directed towards the node 2 and flows to ground through a relatively small measuring impedance. Thus, in the current through the measuring impedance in addition to the component associated with the HV winding current, there is also an additional capacitive component caused by the current flowing to ground from the LV winding.

The unearthing of the shield yields to a decrease of coupling of LV winding to the ground, an increase in the induced potentials at nodes 3 and 4 of this winding, in the potential difference between nodes 2 and 4 and in the capacitive current flowing through the measuring impedance.

Thus, in the case where C_{35} and C_{45} are comparable to C_{13} and C_{24} the unearthing of shield should lead to an upward shift of the HV winding frequency response. In this case, some changes in the resonant frequencies corresponding to the inter-winding interactions are possible but the natural frequencies of the HV winding will remain practically unchanged.

The upward shift of the frequency response characteristic of the HV winding increases with increasing voltage induced in node 4. This voltage is determined by the relation of the capacitances C_{50}, C_{35} (C_{45}) and C_{13} (C_{24}). If the capacitance of the LV winding to the shield is much less than the capacitance between the windings that is $C_{35} << C_{13} << C_{50}$ then the voltage U_4 turns out to be comparatively large and the unearthing of the shield (magnetic circuit) practically does not lead to a significant shift in the frequency characteristic of the HV winding.

As the results of the preliminary calculations show, with a certain ratio of the capacitances even a decrease in the voltage U_2 (downward shift of the frequency response) is possible but this decrease is usually negligible. Similar conclusions are valid for the LV winding: if the value of C_{35} is close to the value of C_{50} the frequency response of the LV winding after unearthing of the shield shifts upward; if C_{35} is much less than C_{50} the change in the frequency response is negligible.

5. CONCLUSIONS

1. White-box models are suitable for qualitative evaluation of frequency response characteristics of transformer windings and can be useful for the interpretation of FRA results.
2. The reason for the differences in the frequency responses of the LV and HV (inner and outer) windings is shown. The frequency response of the HV winding typically is V-shaped, which is caused by the additional capacitive current flowing through the measuring impedance from the unearthed LV winding.
3. The effect of the shield with floating potential on the frequency response is revealed and explained. The unearthing of the shield yields to the redistribution of capacitive currents and the HV (outer) winding frequency response generally shifts up.

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