Measurement-based electrical parameters of power transformers for Frequency Response Analysis interpretation – Part I: Core analysis

Dinh Anh Khoi Pham

Abstract—**Although the standard Frequency Response Analysis (FRA) test has been approved as an efficient tool to diagnose mechanical failures in power transformers, the demand to interpret FRA traces in practical and physical way is still requested. That means physical electrical parameters of transformers should be determined reasonably based on real measurements and afterwards are applicable for the interpretation.**

For purpose of FRA interpretation in practical manner, electrical parameters of power transformers in a physical equivalent circuit should be determined. As a first step, the paper introduces a new approach in determining frequency dependent core impedances of a distribution transformer based on the combination of circuit analysis of a duality-based model, measurements of driving-point impedances and experimental formulas. From that, two important contributions can be drawn. Firstly, frequency dependent core impedances are ready as available components in the circuit for FRA interpretation in broad frequency range. Secondly, the core parameters could be useful indicators for detecting relevant failures in cases there is no more failure on transformer windings.

Index Terms—**Driving-point impedance, duality principle, electrical parameters, frequency response analysis, transformer core.**

1 INTRODUCTION

The core is an important component of power transformers, which provides magnetic links between the windings. Its magnetic property has been so far investigated in detail through

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measurements and design data based simulations [1]. However, the dual electrical equivalence of the core is still not widely approved for FRA investigations since it has been understood that the core does not play any key role in analysis of frequency responses of power transformers in medium and high frequency range. Nevertheless, it is realized that electrical impedances of core sections are necessary in determination of influence of the immeasurable winding series capacitance on total capacitance effect of the transformers, which will be mentioned in the Part II [2]. In addition, in several cases the zero-sequence impedance representing flux paths outside the windings is important at medium frequencies; as a result, a transformer circuit with physical dual magnetic-electrical core representation should be in general investigated.

In order to introduce a practical approach in analysis of the transformer core for purpose of FRA interpretation in low and medium frequency range, this paper introduces a new approach in determining frequency dependent core parameters. Two important contributions of the paper can be drawn: the applicability of the approach for any transformer, and the availableness of frequency dependent core impedances in broad frequency range for the feasibility of FRA interpretation via circuit simulation.

A test opened transformer with rating parameters 200 kVA 10.4/0.46 kV YNyn6 transformer is then investigated in this paper. Core section impedances of the transformer are analyzed based on measurements of frequency responses of open-circuit driving-point impedances at the high-voltage (HV) side at low frequencies. (Core impedances at the low-voltage (LV) side are not investigated due to their very low values). The measurements are performed by

mean of a commercial scattering-parameter vector-network analyzer (VNA) with chosen settings appropriate for FRA purpose: open-circuit voltage of 1 V_{rms}, broad frequency range from 20 Hz to 2 MHz with 801 points, receiver bandwidth of 30 Hz.

2 REVIEW OF THE DUALITY-BASED LUMPED TRANSFORMER CIRCUIT

This section recalls the duality-based transformer circuit that is originally developed for the purpose of transient and frequency response analysis [3, 4]. In this paper, only inductive components such as core, leakage and zero-sequence impedances appear in the circuit since the circuit is then analyzed at low frequencies to determine the core impedances. For the complete investigation at low and medium frequencies, the full circuit with appearance of winding capacitances will be introduced in the Part II [2].

Fig. 1 depicts the duality-based equivalent circuit of the test transformer in low frequency range with three per-phase zero-sequence impedances and one leakage inductance in context of an open-circuit driving-point impedance test performed on the HV winding of phase A by means of the VNA. In Fig. 1, $Z_1 = R_1//L_1$, $Z_y = R_y//L_y$ are non-linear core leg and yoke impedances respectively; L_3 are (per-phase) leakage inductances; $Z_4 = R_4//L_4$ are per-phase zero-sequence impedances; R_H and R_L are resistances of HV and LV windings. All of them are frequency dependent. $N_H:N_L$ is the transformer ratio.
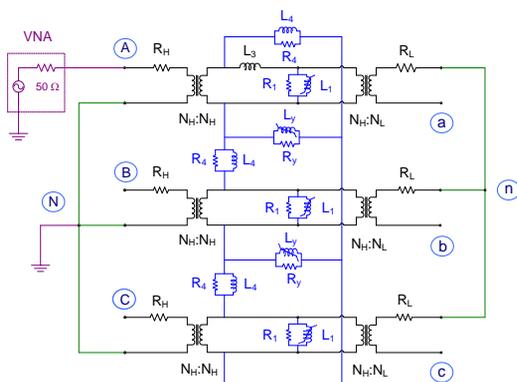
Fig. 1. Lumped circuit of the test transformer in a single-phase impedance measurement at HV side at low frequencies.

3 MEASUREMENT-BASED DETERMINATION PROCEDURE OF TRANSFORMER CORE IMPEDANCES AT LOW FREQUENCIES

In this section, a new procedure in determining core impedances (Z_1, Z_y) based on measurements of open-circuit driving-point impedances on star winding in low frequency range is proposed. The low frequency range that is different with different transformers is defined within 20 Hz to a frequency where the measured impedances are still purest inductive.

3.1 Necessary measurement configurations

Since core influence is dominant in open-circuit impedances at low frequencies, it is logical to analyze only core impedances that are referred into HV side in the equivalent circuit in Fig. 1 (R_H, L_3, and Z_4 are short-circuited, R_X has no influence, the LV phase windings are left open).

In order to determine core impedances, different configurations of open-circuit measurements associated with different core impedance combinations shown in Table 1 are necessary. In Table 1, equivalent impedances are input impedances seen from the excited terminals. Afterwards, a data fitting procedure is applied to determine the core impedances with a selected number of measurement configurations.

TABLE 1
CONFIGURATIONS OF INPUT IMPEDANCE MEASUREMENTS ON STAR WINDING

No	Configuration	Calculated $Z_{equivalent}$
1	A-N excited (in Fig. 1)	$(Z_1//Z_y + Z_1)//(Z_1//Z_y)$
2	A-N excited, B-N shorted	$(Z_1//Z_y)//(Z_1//Z_y)$
3	A-N excited, C-N shorted	$Z_1//(Z_1//Z_y)$
4	B-N excited	$(Z_1//Z_y + Z_1//Z_y)//Z_1$
5	B-N excited, C-N shorted	$(Z_1//Z_y)//Z_1$
6	B-N excited, A-N shorted	$(Z_1//Z_y)//Z_1$
7	C-N excited	$(Z_1//Z_y + Z_1)//(Z_1//Z_y)$
8	C-N excited, B-N shorted	$(Z_1//Z_y)//(Z_1//Z_y)$
9	C-N excited, A-N shorted	$Z_1//(Z_1//Z_y)$

3.2 Data fitting procedure

The procedure is designed to solve a mathematic problem with two non-linear variables (Z_1 and Z_y) based on measurement configurations in Table 1. The problem can be solved efficiently through a non-linear least-square solver whose objective function is established as norm of real and imaginary impedance errors.

In Table 1, there are nine different measurement configurations and four different calculated equivalent impedances. That means, to determine reasonable core impedances, at least

two configurations that have different calculated equivalent impedances should be taken into account. In fact, the best solution is using all configurations for data fitting.

To prove this argument, the so-called maximum magnitude error ΔZ_{max} between measured and calculated impedances introduced in Table 1 is determined for two cases: a) three first configurations of measurements on phase A and b) all nine configurations are analyzed. The ΔZ_{max} is of the 4th configuration in Table 1 (B-N excited) with values of 10.29 % (case a) and 1.45 % (case b). Obviously, analysis of all measurement configurations provides better results, which cannot be achieved by using measurements on a single phase as proposed in the approach in [4].

3.3 Influence of zero-sequence impedances

It is realized that per-phase zero-sequence impedances have small but certain influence on core analysis at frequency from 20 Hz to 100 Hz for the test transformer due to their high values since the transformer has no tank. The calculation of equivalent impedances is still based on analysis of the circuit in Fig. 1 but not presented due to its complexity. Fig. 2 plots fundamental parameters of core sections (R_1, R_y, L_1, L_y) in frequency range from 20 Hz to 400 Hz calculated from the core impedances (Z_1, Z_y) of the test transformer when two zero-sequence impedances of outer phases are considered. In general, when transformers have tank, influence of other components than core impedances can be neglected at low frequencies.
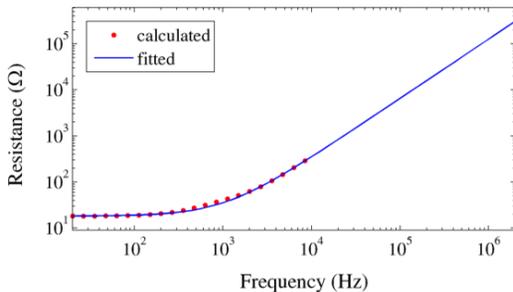
Fig. 2. Core section components calculated from 20 Hz to 400 Hz.

4 FORMULA-BASED DETERMINATION PROCEDURE OF TRANSFORMER CORE IMPEDANCES AT HIGH FREQUENCIES

In this section, only the core impedances of the test transformer are calculated based on formulas

since as frequency is higher than 400 Hz, core impedances of the test transformer are no longer dominant in measured open-circuit impedances. Since the main goal of the research is to interpret FRA measurements, all inductive impedances in the equivalent circuit should be developed versus frequency in broad frequency range. This section will discuss how the core impedances are determined at frequencies higher than 400 Hz for application.

4.1 Phenomena in transformer core at low field applied in the FRA measurements

1) Eddy current (skin effect): It is a main effect caused by low alternating magnetic field within transformer core laminations. When such magnetic field generated by frequency-varying source penetrates a ferromagnetic material, induced circulating current loops appear and are called as eddy currents whose density has a non-linear distribution depending on frequency of the magnetic field. As a result, a non-linear distribution of magnetic flux (towards the surface) is associated. This phenomenon results in the so-called eddy current loss which brings a rise of effective resistance and a reduction of inductance when frequency increases [1, 5]. Eddy current loss is proportional to the square of lamination thickness, square of frequency and square of effective value of flux density [6].

2) Hysteresis phenomenon: It is a characteristic of a magnetic material to retain magnetization or to oppose a change in magnetization, which depends on applied magnetic field magnitude. When a low field is applied, there is reversible change in domain magnetic structures which causes the magnetization returns to its initial value on removal of the external magnetic field. As field strength is high enough for the so-called magnetic domain wall motion, both reversible and irreversible processes take place so that the magnetization does not return to its original value on applied filed removal [1]. In case applied field is cyclic and its magnitude is high enough, there exists a hysteresis (B-H) loop whose loss is proportional to the area of the B-H loop [6]. However, under a low applied field for FRA measurements (~1 V), the corresponding hysteresis loop is a small elliptical loop whose area is very small and therefore, hysteresis loss is considered insignificant compared with eddy current loss at high frequencies [1].

3) Anomalous losses: There exists another

complicated phenomenon in the core which is called as diffusion effect in the crystal grid (domain structures) of the core material causing anomalous losses [7].

Conclusion: Under low field and no-load conditions, the (classical) skin effect has most influence whereas other phenomena can be neglected [1]. From that, core impedances are determined based on relevant experimental formulas.

4.2 Calculation of core impedances based on skin effect

1) Magnetic reluctance: The reluctance of the core (leg) section in the magnetic circuit representing a relation between the magnetomotive force F and corresponding flux Φ can be determined based on magnetic and design parameters of the core as follows [8, 9]:

$$\Re = \frac{F}{\Phi} = \frac{\ell}{\hat{\mu}_{eff}\,\mu_0 A} \qquad (1/H) \qquad (1)$$

where ℓ (m) is the magnetic flux path length of the core section; $\hat{\mu}_{eff}$ is the complex effective relative permeability of core lamination in rolling direction, i.e., along the core section; μ_0 (H/m) is the permeability constant of free space and A (m^2) is the cross-sectional area of the core section.

The complex effective relative permeability $\hat{\mu}_{eff}$ in the rolling direction of the magnetic flux is expressed as [1, 10]:

$$\hat{\mu}_{eff} = \mu'_{eff} - j\mu''_{eff} = k_{fe}\mu_r \frac{\tanh\{(1+j)b/\delta\}}{(1+j)b/\delta} \qquad (2)$$

where k_{fe} is stacking factor representing fraction of core steel in the total cross section, μ_r is local relative permeability in the rolling direction, b (m) is half of a lamination thickness and δ (m) is skin depth or depth of penetration of eddy currents.

The skin depth δ is frequency dependent and calculated as [5]:

$$\delta = \sqrt{2/(\omega\sigma\mu_0\mu_r)} \qquad (3)$$

where ω (radian/s) is angular frequency and σ (S/m) is the electrical conductivity of the core lamination.

2) Dual electrical impedance: The complex reluctance $\Re$ accounting for skin effect has a dual electrical impedance Z whose equivalent circuit consists of a resistance (R_s) in series with an inductance (L_s) [8]:

$$Z = j\omega N^2 \frac{1}{\Re} \equiv R_s + j\omega L_s \qquad (4)$$

where N is the number of turns of the excited winding on the phase of the core section.

From that, the resistance R_s (Ω) and inductance L_s (H) can be determined as:

$$R_s = -\omega N^2 \operatorname{Im}\{1/\Re\} \qquad (5)$$

$$L_s = N^2 \operatorname{Re}\{1/\Re\} \qquad (6)$$

Due to the fact that the equivalent electrical circuit of a core section in Fig. 1 includes only resistance and inductance in parallel, R_s and L_s should be converted into R_p and L_p respectively.

3) Calculation of core impedance: Thanks to the magnetic parameters which can be in general applied for a transformer core [1, 10] and approximate design parameters of the core leg of the test transformer in Table 2, components of the core leg in both series and parallel models are derived and shown in Fig. 3. A comparison between the measurement- and formula-based components of the core leg at the so-called joint frequency, i.e. 400 Hz, in Fig. 2 and 3 respectively shows that there is small error between L_1 and L_p, which indicates that the calculation of a core section based on skin effect is acceptable and therefore can be used at high frequencies. Error between R_1 and R_p at 400 Hz is not important since it contributes only to a little change of damping of frequency responses at first resonance frequencies.

TABLE 2
MAGNETIC AND DESIGN PARAMETERS FOR CORE IMPEDANCE CALCULATION

Para-meter	Quantity	Value
μ_0	permeability of free space	$4\pi \times 10^{-7}$ H/m
μ_r	relative permeability	500
σ	conductivity	5×10^6 S/m
$2b$	lamination thickness	0.35×10^{-3} m
k_{fe}	stacking factor	0.92
ℓ	core section length	0.36 m
A	core cross-sectional area	25.6×10^{-3} m^2
N	number of turns	736 turns
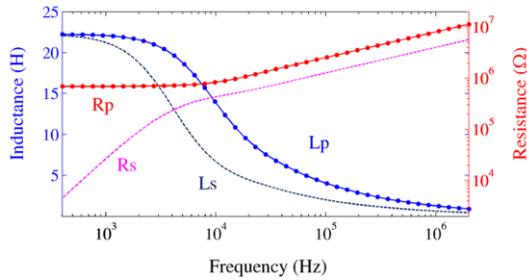
Fig. 3. Skin-effect-based core leg components from 400 Hz to 2 MHz.

5 FREQUENCY DEPENDENT CORE PARAMETERS

The frequency dependent functions of core parameters (resistances, inductances) in broad frequency range can be obtained from measured values (at low frequencies) and calculated values (at high frequencies), simply by combining the fitting functions [11] of parameter values in whole frequency range. The fitting functions yield smooth representation of measurement based values (see Fig. 2) and can be used to adapt the formula based values (see Fig. 3) so that there is no difference between them at the joint frequency. Fig. 4 and 5 plot frequency dependent functions of inductances and resistances of core legs and yokes in broad frequency range.
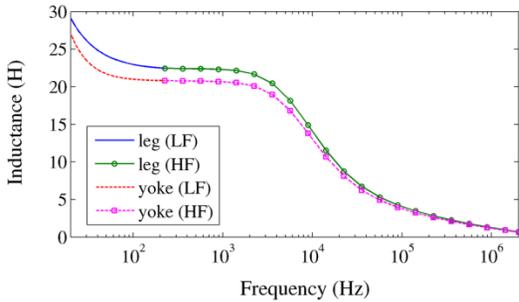
Fig. 4. Core section inductances from fitted frequency dependent functions.
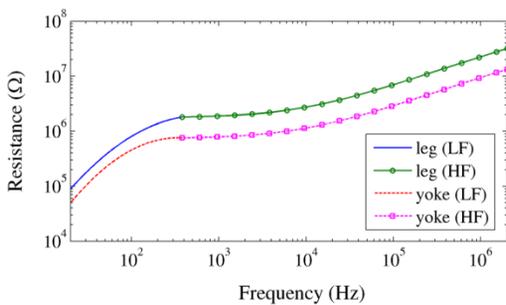
Fig. 5. Core section resistances from fitted frequency dependent functions.

It is realized in Fig. 4 that core inductances vary significantly at very low frequencies, e.g.,

from 20 Hz to 100 Hz. It is not clear whether the changes at such frequencies come from the transformer core itself (saturation condition) or from the measurement system. In the measurement system, the problem may be caused by the difference of characteristic impedances of the measuring cable (not 50 Ω from 20 Hz to several tens kHz) and the instrument source (50 Ω) since the measurement is performed by means of the VNA. However, a good measurement with a known resistance in broad frequency range after calibration shows that the calibration removes the effect very well. It is concluded that, for the purpose of FRA application, the parameter functions plotted in Fig. 4 and 5 can be used in whole frequency range.

6 CONCLUSION

This paper presents a complete approach in determining core section impedances of a distribution transformer in a duality-based circuit based on measurements of frequency responses of driving-point impedances at low frequencies and skin-effect-based formulas at high frequencies. Several important contributions of the paper can be drawn as follows.

Physical electrical parameters of two-winding three-legged transformers are introduced in the duality-based circuit for the purpose of FRA interpretation at low and medium frequencies. As a first step, frequency dependent functions of core parameters in whole range of frequencies are determined and ready as available components for simulation.

Formula-based calculation of core parameters of the test transformer at high frequencies is proposed since the parameters cannot be determined directly from measurements at these frequencies. More importantly, the formula-based solution provides "reference curves" from which core impedances of any transformer can be obtained reasonably in broad range of frequencies based on only the measurements at low frequencies. This contributes to large application for FRA purpose.

The calculated constant values of core parameters at a "joint frequency" can be used as physical indicator to diagnose core problem. It is suggested from the research that core parameters should be calculated based on the proposed approach as "fingerprint" data when transformers

are in healthy condition for any comparison later.

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Thông số điện máy biến áp lực cho phân tích đáp ứng tần số - Phần I: Lõi thép

Phạm Đình Anh Khôi

Tóm tắt— Mặc dù các phép đo đáp ứng tần số đã được áp dụng hiệu quả trong việc chẩn đoán các sự cố cơ trong máy biến áp lực (MBA), cách thức phân tích kết quả đo trong thực tế vẫn chưa được đề xuất cụ thể, vốn cần phải dựa trên việc xác định các thông số điện của MBA từ các phép đo.

Để phân tích đáp ứng tần số ứng dụng thực tế, các thông số điện trong một sơ đồ mạch tương đương của MBA cần phải được xác định. Bài báo giới thiệu một giải pháp mới xác định tổng trở lõi thép theo tần số của một MBA phân phối dựa trên sự kết hợp giữa phép phân tích mạch tương đương MBA xây dựng trên nguyên lý tương hỗ, các phép đo tổng trở đầu cực và các công thức thực nghiệm. Kết quả của nghiên cứu là giá trị các tổng trở của lõi thép MBA nhận được trong vùng tần số rộng áp dụng cho phân tích đáp ứng tần số. Giá trị tổng trở này có thể ứng dụng để chẩn đoán sự cố của riêng lõi thép, trong trường hợp MBA không có các sự cố khác trên cuộn dây.

Từ khóa— Tổng trở đầu cực, nguyên lý tương hỗ, thông số điện, phân tích đáp ứng tần số, lõi thép máy biến áp.