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Inductively coupled on-line impedance measurement for condition monitoring of electrical equipment

Sooriya Bandara Rathnayaka¹ ✉, Kye Yak See¹, Kangrong Li¹

¹School of Electrical and Electronic Engineering, Nanyang Technological University, Singapore, Singapore

✉ E-mail: sooriyab001@e.ntu.edu.sg

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Abstract: Off-line periodic maintenance has been used to keep critical electrical equipment in good condition. However, there is an increasing demand for on-line condition monitoring of the electrical equipment for preventive maintenance purposes. An on-line impedance extraction technique based on an inductive coupling approach is explored for the detection of early signs of defects of electrical equipment. The proposed method is demonstrated experimentally using a custom-made transformer with emulated defects as equipment-under-test. To enhance the defect detection capability for impedance deviation, various statistical diagnostic indicators are also investigated. By combining the proposed on-line impedance extraction method and a suitable diagnostic indicator, early detection of the defect in electrical equipment becomes feasible.

1 Introduction

Condition monitoring (CM) techniques can be deployed to identify the early signs of defects or deterioration in critical electrical equipment to avoid catastrophic damage and to minimise the economic impact due to system downtime [1–5]. A typical real-time CM system consists of sensing, data acquisition and fault diagnosis subsystems, which can be deployed round the clock [6, 7]. Some commonly adopted techniques to identify electrical equipment's defects are partial discharge (PD), frequency response analysis (FRA), thermal analysis and vibration analysis [8–13]. In this paper, the focus will be on defect detection based on FRA.

The off-line FRA has become a well-established diagnostic technique for fault detection of an electrical equipment [14]. For example, any faults in a transformer's winding change its inductance and capacitance and such deviation will be reflected in the measured frequency response of the transformer's impedance [15, 16]. However, the off-line FRA can only be carried out when the electrical equipment is shut down for periodic maintenance. Any serious defects that surface between the window periods of the scheduled maintenance can cause unexpected damage to the equipment. This paper aims to develop an on-line impedance extraction technique without the need of shutting down the equipment-under-test (EUT). To avoid the direct electrical contact to the EUT that operates at high voltage, an inductive-coupled method can be adopted [17, 18]. With the EUT connected with high-voltage power supply under its normal operating condition, an inductive coupling probe induces a signal into the EUT and another inductive coupling probe receives the same signal across the frequency range of interest. With a pre-measurement characterisation of both probes, the on-line impedance's frequency response can be extracted through a two-port network analysis. As the proposed method has no direct electrical contact with the EUT, its setup requires no modification on the EUT and can be easily implemented on site for CM purpose.

This paper will be organised as follows. Firstly, the basic theory behind the proposed inductively coupled method for in-circuit impedance extraction is described and validated in Section 2. Section 3 demonstrates that the proposed method is defect detection using a custom-made transformer as EUT with emulated defects. To enhance the capability of defect detection through impedance variation, a few statistical analysis techniques are evaluated so that a suitable technique is chosen. Finally, Section 4 concludes the paper.

2 Theoretical background and validation

Fig. 1a illustrates the basic measurement setup for the extraction of impedance information of an electrical equipment connected to high-voltage power supply. The impedance of the electrical equipment is denoted as Z_E and the high-voltage power supply is represented by its equivalent open-circuit voltage V_S and equivalent source impedance Z_S . The impedance of the wiring connections between $a-b$ and $a'-b'$ is denoted as Z_W . Ports 1 and 2 of the vector network analyser (VNA) inject and receive a sinusoidal signal through the two clamp-on-type inductive probes. Fig. 1b shows the equivalent circuit model of Fig. 1a, in terms of three cascaded two-port networks of injecting inductive probe (N_{IP}), impedance to be measured (N_X) and receiving inductive probe (N_{RP}). Z_X is the total impedance of Z_E , Z_W and Z_S in series, which is the impedance to be measured.

In most practical situations, Z_X is usually dominated by Z_E , as Z_S and Z_W are relatively small. By measuring the S -parameters of overall cascaded two-port network N_T , its $ABCD$ parameters can be obtained mathematically and the $ABCD$ parameters of N_T can be expressed as (1).

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_T = \begin{bmatrix} A & B \\ C & D \end{bmatrix}_{IP} \begin{bmatrix} A & B \\ C & D \end{bmatrix}_X \begin{bmatrix} A & B \\ C & D \end{bmatrix}_{RP} \quad (1)$$

where $\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{IP}$, $\begin{bmatrix} A & B \\ C & D \end{bmatrix}_X$ and $\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{RP}$ are the two-port $ABCD$ matrices of N_{IP} , N_X and N_{RP} , respectively. By defining voltage and the current relation of the two-port network of the impedance to be measured, $ABCD$ parameters of N_X can be determined by (2).

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_X = \begin{bmatrix} \left. \frac{V_{x1}}{V_{x2}} \right|_{I_{x2}=0} & \left. \frac{V_{x1}}{I_{x2}} \right|_{V_{x2}=0} \\ \left. \frac{I_{x1}}{V_{x2}} \right|_{I_{x2}=0} & \left. \frac{I_{x1}}{I_{x2}} \right|_{V_{x2}=0} \end{bmatrix} \quad (2)$$

where V_{x1} , I_{x1} , V_{x2} and I_{x2} are input voltage, input current, output voltage and output current of N_X , respectively; as illustrated in Fig. 1b. A , B , C and D are open-circuit voltage transfer ratio, short-circuit transfer impedance, open-circuit transfer admittance and short-circuit current transfer ratio, respectively. Prior to the

Fig. 1 Basic measurement setup for the extraction of impedance information of an electrical equipment connected to high-voltage power supply

(a) Basic measurement setup, (b) Two-port equivalent circuit

Fig. 2 Characterisation of inductive probes

(a) Receiving probe, (b) Injecting probe

Fig. 3 Extracted 33 Ω and wire loop impedance

measurement, $ABCD$ parameters of N_{IP} and N_{RP} must first be characterised as explained later. Once the $ABCD$ parameters of the injecting and receiving inductive probes are obtained, $ABCD$ parameters of N_X can be determined through matrix manipulation from (3). By solving parameter B in N_X , Z_X can be extracted [19] through (4).

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{N_X} = \begin{bmatrix} A & B \\ C & D \end{bmatrix}_{N_{IP}}^{-1} \begin{bmatrix} A & B \\ C & D \end{bmatrix}_{N_{RP}} \begin{bmatrix} A & B \\ C & D \end{bmatrix}_{N_T}^{-1} \quad (3)$$

$$Z_X = B_X = \frac{A_{RP}(D_{IP}B_T - B_{IP}D_T) - B_{RP}(A_TD_{IP} - C_TB_{IP})}{(A_{IP}D_{IP} - C_{IP}B_{IP})(A_{RP}D_{RP} - C_{RP}B_{RP})} \quad (4)$$

To validate the measurement setup described in this paper, Solar 9144-1N (bandwidth 10 kHz–100 MHz, maximum current 80 A) and Solar 9134-1 (bandwidth 20 Hz–100 MHz, maximum current 500 A) current probes are chosen as injecting and receiving probes, respectively [20]. In reality, all inductive coupling probes are non-ideal due to leakage inductance, core loss and inter-winding capacitance. Hence, both inductive coupling probes have to be properly characterised using Solar 9125-1, a circular cross-section transmission line calibration fixture. To characterise the receiving probe, it is clamped onto the inner conductor of the fixture and its output is connected to Omicron node 100 VNA's port 2, as shown in Fig. 2a. With one port of the fixture terminated with a short, the other port is connected to VNA's port 1. The inner conductor of the test jig is used to emulate the clamped wire. With one end of the calibration fixture short-circuited so that the overall impedance for port 1 is 50 Ω instead of 100 Ω. Then the two-port S -parameters of the receiving probe are measured with the VNA. Similarly, to characterise the injecting probe, ports 1 and 2 are swapped, as illustrated in Fig. 2b. With the measured S -parameters [19] of both receiving and injecting probes, they can be converted to their respective $ABCD$ parameters using (5)–(8). The equivalent circuit models of both receiving and injecting probes are shown in Figs. 2a and b, respectively; where Z_P is the impedance of the inner conductor of the calibration fixture; Z_M and Z_S are equivalent core loss impedance, and winding impedance of the current probe, respectively.

$$A = \frac{(1 + S_{11})(1 - S_{22}) + S_{12}S_{21}}{2S_{21}} \quad (5)$$

$$B = Z_0 \frac{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}}{2S_{21}} \quad (6)$$

$$C = \frac{1}{Z_0} \frac{(1 - S_{11})(1 - S_{22}) - S_{12}S_{21}}{2S_{21}} \quad (7)$$

$$D = \frac{(1 - S_{11})(1 + S_{22}) + S_{12}S_{21}}{2S_{21}} \quad (8)$$

The validation setup is similar to Fig. 1a, except that $a-a'$ are not connected to the AC power source but shorted instead to form a closed-circuit loop. Several known passive components of different values are chosen as Z_E . To reduce the influence of the inductance due to the wiring connection of $a-b$ and $a'-b'$ to the Z_E , it is shortened to its minimum possible length. Both probes are clamped onto the circuit loop and connected to VNA. The two-port $ABCD$ parameters of N_T are measured with the setup and N_X is obtained using (3) with the characterised $ABCD$ parameters of injecting and receiving probes. The impedance Z_X , which is the resultant impedance of the wire loop and Z_E , can be extracted from parameter B of N_X as described in (4). Similarly, the wire loop impedance can also be extracted using the proposed method by replacing the Z_E with a short. The wire loop impedance will be de-embedded from Z_X to extract Z_E . For example, a 33 Ω resistor is chosen as Z_E . Fig. 3 shows extracted Z_X and Z_E . The wire loop impedance is also plotted. Based on the measured impedance, the loop has an inductance of 386 nH. For Z_E with low impedance, the loop inductance has a significant impact on the accuracy of the extracted impedance at high frequency, if the impedance of the loop is not de-embedded. Fig. 3 shows that by de-embedding the inductive reactance from Z_X , Z_E can be extracted with good accuracy. To illustrate the ability of the proposed method in extracting impedance of various passive components with a wide range of impedances, Figs. 4–6 shows the extracted impedances from 100 kHz to 10 MHz using the proposed method for three resistors (120 Ω, 1.2 kΩ and 12 kΩ), three capacitors (33 nF, 330 nF and 3.3 μF) and three inductors (33 μH, 330 μH and 3.3 mH), respectively. The starting frequency is set at 100 kHz, as the

Fig. 4 Extracted impedance of resistors using the proposed method (IC) and IA
(a) Magnitude, (b) Phase

Fig. 5 Extracted impedance of capacitors using the proposed method (IC) and IA
(a) Magnitude, (b) Phase

Fig. 6 Extracted impedance of inductors using the proposed method (IC) and IA
(a) Magnitude, (b) Phase

transfer characteristics of the two inductive probes below 100 kHz are poor and can compromise the accuracy of the measurement. The extracted impedance's magnitude and phase are in close agreement with the direct measurement results using Agilent 4294A impedance analyser.

3 On-line impedance for condition monitoring

3.1 On-line impedance monitoring

To demonstrate the ability of the proposed method for extraction of impedance frequency response of electrical equipment as a mean of defect detection, a customised 1.5 kW step-down transformer as shown in Fig. 7a is designed and fabricated as the EUT. It has extra 1, 2 and 3 turns in both primary and secondary windings. They are made accessible externally and can be short-circuited at either the

Fig. 7 Fabricated single-phase transformer winding configuration
(a) External view, (b) Winding configuration

Fig. 8 Setup of on-line impedance FRA measurement
(a) Actual view, (b) Equivalent circuit

primary or the secondary to emulate winding defect within the transformer as shown in Fig. 7b.

In reality, the load of the transformer secondary is unknown. However, for validation purpose, a $100\ \Omega$ resistor (Z_L) is connected to the secondary of the transformer as a nominal load, as shown in Fig. 8. The primary terminals of the transformer ($d-d'$) are connected to 230 V/50 Hz power supply ($c-c'$) through power cables. For the given load, the primary current is about 0.6 A, which is well below the maximum current ratings of the inductive probes and hence, the no issue on the magnetic core saturation the probes. The injecting and receiving inductive coupling probes are clamped onto the power cable connecting between the power supply and primary of the transformer to extract the impedance response looking into $d-d'$. The frequency response of the

primary's impedance will be extracted to monitor any deviation of the impedance of the transformer from its norm.

Before connecting the transformer to the AC mains, the source impedance can be extracted by adding a suitable bypass capacitor ($2.2\ \mu\text{F}$) across $c-c'$, which serves as an AC short and creates an electrical loop in series with source impedance. Firstly, the bypass capacitor is measured off-line with an impedance analyser. Then, the bypass capacitor is connected across $c-c'$ and the total impedance of the source and bypass capacitor is measured with the proposed method. The impedance of the bypass capacitor can be de-embedded from the measured total impedance to recover the source impedance. The extracted source impedance is resistive at low frequency but dominated by the inductive reactance at high frequency, as shown in Fig. 9. From the source impedance response, the estimated inductance is about $1.8\ \mu\text{H}$. Once the source impedance is known, the bypass capacitor will be removed.

For all the measurements that follow, the output of the VNA is set to 13 dBm (20 mW) so that the injected signal to be measured by the receiving probe is at least 10 dB above the background noise. Figs. 10a and b show the measured impedance responses looking into $d-d'$ with the emulated inter-turn shorts of primary and secondary windings, respectively. Although the measurement is validated up to 10 MHz, the measurement here focuses on the frequency range around the resonant frequency the transformer primary winding, which is between 100 kHz and 2 MHz. Both the measured impedance responses show that the deviation of impedance due to inter-turn short can be detected. The impedance deviation is due to the changes of the transformer's winding inductance and capacitance from its normal condition. As both the primary and secondary windings are wound on the same core and according to Lenz's law, shorted turns in the primary or secondary winding do influence the magnetising inductance, where it decreases with the higher number of shorted turns. Inter-winding capacitance remains constant for both cases. The shorted turns in the primary and secondary windings reduce the overall parasitic primary winding capacitance. Fig. 10a shows that shorted turns in the primary winding are more noticeable. As expected, the shorted turns in the secondary winding are less visible, as shown in Fig. 10b. Also, a parallel resonance about 250 kHz due to the winding inductance and capacitance is clearly observed. Below 250 kHz, the impedance is dominated by the transformer primary inductance, as well as the reflected inductance and load impedance from secondary to primary. With more turns shorted in the primary winding, the primary inductance reduces, and the reflected secondary inductance is also reduced due to lower effective turn ratio. This inductance reduction is clearly observed below 250 kHz, as the impedance is dominated by the inductive reactance. On the other hand, below 250 kHz, the inductive reactance increases with more shorted turns in the secondary winding, as observed in Fig. 10b. The parasitic capacitance of the transformer winding is more dominant in the primary impedance above 250 kHz. With the emulated inter-turn short in the primary and secondary windings, the frequency response from the extracted primary impedance of the transformer has demonstrated its ability to detect the deviation of the transformer's internal parameters such as inductance and capacitance due to emulated defects. Since the inductance of the AC mains impedance is about $1.8\ \mu\text{H}$ and the transformer winding's inductance is in the order of mH, the impact of AC mains impedance variation to the measurement results is minimal.

3.2 Enhancement of defect detection capability

The proposed on-line impedance extraction method has shown its ability to detect the emulated defects of the transformer when it is connected to the power supply and operated under its normal operating condition. Although the defects in the secondary winding result in less visible impedance changes, we can deploy some statistical indicators to enhance the detection of small impedance deviation. Statistical methods such as root-mean-square error (RMSE), cross-correlation coefficient (CCC), absolute logarithmic error (ASLE) are used to provide a clear indication of the variation of measured parameter relative to its norm. Let us consider measured impedance of the transformer without and with an

Fig. 9 Impedance magnitude of high-voltage source impedance (Z_s)

Fig. 10 On-line impedance frequency response at d-d' with shorted turns in the (a) Primary winding, (b) Secondary windings

emulated defect as two functions $A(f)$ and $B(f)$, respectively. RMSE of $A(f)$ and $B(f)$ can be calculated using the following mathematical expression [21]:

$$\text{RMSE}_{(A,B)} = \sqrt{\frac{\sum_{i=1}^N [A(i) - B(i)]^2}{N-1}} \quad (9)$$

where $A(i)$ and $B(i)$ are the i th values of measured impedance for the transformer without and with an emulated defect at a specific frequency, respectively; where N is the total number of sampled impedance values. Since the impedance near the resonant frequency can vary significantly and affect the repeatability of the indicators, RMSE is computed below and above the resonant

Fig. 11 Calculated indicators for a turn to turn short at the primary and secondary windings (a) RMSE, (b) CCC, (c) ASLE

frequency for more reliable indication of changes in impedance. For the study in this paper, the sample size $N = 801$, sampled values $N = 1-17$ are below resonance and sampled values $N = 18-801$ are above resonance. Using the measured impedance frequency responses in Figs. 10a and b, the computed RMSE is plotted in Fig. 11a. Based on the previous literature reviews, a deviation $>1\%$ from its base value translate to $\text{RMSE} > 1$, which is an indication 1% deviation of impedance's frequency response of the transformer without inter-turn short, as shown in Fig. 11a. RMSE has shown its ability to provide an alert of either 1, 2 or 3 turns short in both primary and secondary windings but tends to be oversensitive for the inter-turn short in the primary below the resonant frequency. The small frequency shift of the resonance in impedance frequency responses also tends to dominate the RMSE indicator.

CCC is another technique, which can be used to compare the similarity of two data sets and it is computed using (10).

$$CCC_{(A,B)} = \frac{\sum_{i=1}^N A(i)B(i)}{\sqrt{\sum_{i=1}^N [A(i)]^2 \sum_{i=1}^N [B(i)]^2}} \quad (10)$$

Its critical region is defined with $CCC < 0.9998$. Fig. 11b shows that the calculated CCC only alerted 2 or 3 turns short in the primary winding above the resonant frequency but failed to alert for other cases, which clearly indicates its inadequacy to provide a good indicator for defects that result in small impedance change. CCC approaches 0 if the impedance's frequency responses are very different and approaches 1 if the impedance's frequency responses are of a similar trend. For example, considering a particular condition of $A(i) = cB(i)$ at some frequencies where c is a constant coefficient, CCC approaches 1 if c is large. Hence, it may mislead the final decision. Therefore, it is also not a good choice to differentiate impedance's frequency responses with the similar trend but with different magnitude.

ASLE provides more equitable indication compared with RMSE and CCC, as it is a fully logarithmic comparison as given by:

$$ASLE_{(A,B)} = \frac{\sum_{i=1}^N |20\log_{10}(A(i)/B(i))|}{N} \quad (11)$$

Its critical region is defined with $ASLE > 0.3$, as shown in Fig. 11c. ASLE has shown its ability to provide an alert of either 1, 2 or 3 turns short in both primary and secondary windings. In ASLE, equidistant logarithmic vertical and horizontal axis of the impedance frequency responses overcomes the issues faced in CCC and RMSE. Therefore, it is chosen as for the detection of impedance variation for the transformer.

4 Conclusion

Based on an inductive coupling approach, an on-line impedance extraction method has been described and validated. It has the ability to detect impedance variation in critical electrical equipment, which can be served as one of the parameters for an early sign of defects within the electrical equipment. As the proposed method requires no direct electrical contact with high-voltage power supply, it eliminates the risk of electrical hazards and eases the implementation on site. Using a transformer powered up by the high-voltage power supply as an example, the deviation of the extracted impedance response from its norm serves as good indicative signs of defects in the transformer. To enhance the detection of a very small variation of impedance, proper choice of statistical indicators can be chosen to allow the less visible deviation of impedance response to be detected. Further work will be carried out to characterise the impedance behaviours of other

critical electrical infrastructures, such as motors and switch gears, as well as non-linear loads.

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