Inductively coupled in-circuit impedance measurement based on two-port network characterization

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INDUCTIVELY COUPLED IN- CIRCUIT IMPEDANCE MEASUREMENT BASED ON TWO- PORT NETWORK CHARACTERIZATION

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SUMMARY

In-circuit impedance information is one of the parameters for condition monitoring of an electrical system. Besides condition monitoring, it also provides impedance information of a switching converter connected to high voltage power supply so that an optimal electromagnetic interference (EMI) filter can be designed for electromagnetic compatibility (EMC) compliance. Thus, an accurate in-circuit impedance measurement system without interrupting the normal operation of an electrical system under test is very useful for the applications mentioned earlier. Existing in-circuit impedance measurement approaches, such as transformer ratio-arm bridge, current sensing and the inductive coupling have been reviewed. The two-probe inductive coupling approach is found to be superior to other approaches, as the measurement setup does not have direct electrical contact to the electrical system being measured and it can be easily mounted on-site for any high-voltage electrical system under its normal condition without electrical safety hazard to the measurement personnel. However, the existing implementation of the two-probe inductive coupling approach has several shortcomings. To overcome these shortcomings, two-port network characterization technique is adopted so that the measurement system setup can be treated as cascaded two-port *ABCD* networks. With the accurate characterization of the inductive coupling probes, as well as the measurement setup, the setup calibration can be greatly simplified without interrupting the operation of the electrical system under test. In addition, it also enhances the measurement accuracy of the in-circuit impedance.

The theory of the two-port network characterization is described in details in the thesis and its experimental validation is also covered comprehensively. Subsequently, the value of the proposed approach is demonstrated through two practical applications. Firstly, it is applied to condition monitoring of electrical systems. Using a DC powered motor and an AC powered transformer as two examples, the deviations of in-circuit impedance of the motor and transformer are detected by varying their loading conditions. Then, with the proposed approach, an impact analysis of conducted emission measurement without and with line impedance stabilization network (LISN) is carried out. It is shown that the EMI filter designed based on conducted emission measurement with LISN may not be adequate for the actual electronic product that is connected to the AC power mains without LISN. The impact of the conducted emission measurement without LISN will allow us to know the adequacy of EMI filter, especially for high power switching converters.

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LIST OF ABBREVIATION

AWG	American Wire Gauge
CM	Common-Mode
CISPR	Comité International Spécial des Perturbations Radio électricques
DGA	Dissolved Gas-In-Oil Analysis
DM	Differential-Mode
DUT	Device Under Test
EMC	Electromagnetic Compatibility
EMI	Electromagnetic Interference
ESL	Equivalent Series Inductor
ESR	Equivalent Series Resistor
EUT	Equipment Under Test
FCC	Federal Communications Commission
FRA	Frequency Response Analysis
LISN	Line Impedance Stabilization Network
MOSFET	Metal Oxide Semiconductor Field-Effect Transistor
PCB	Printed Circuit Board
PD	Partial Discharge
SA	Spectrum Analyzer
SMA	Sub-Miniature version A
SMD	Surface-Mounted Device

SMPS	Switched-Mode Power Supply
TRA	Transformer Ratio-Arm
VNA	Vector Network Analyzer

Chapter 1

Introduction

In-circuit impedance, which is the impedance of an electrical system under its normal operating condition, is the important information for conducted electromagnetic interference (EMI) filter design and condition monitoring.

Being the inherent electrical characteristic of every electrical system, in-circuit impedance reflects the operating status of an electrical system [1]. As operating frequency and complexity increase in electronic devices, a component's actual impedance may be varied under different operating conditions. Consequently, the circuit's performance will be affected by the varying impedance. Besides operating frequency, aging [2] can cause impedance variations in the electrical system. In other words, if the electrical system's in-circuit impedance can be extracted, any incipient defect or unhealthy condition on the system can be detected through careful analysis of the extracted impedance. Thus, obtaining the in-circuit impedance profile of an electrical system over a specific range of frequencies would reveal the health condition of the electrical system.

In addition, it is well known that obtaining the switched-mode power supply's (SMPS) equivalent noise source impedance, which has to be measured under in-circuit condition, would help in analyzing the conducted emission [3-5] and designing an optimal EMI filter to suppress the conducted emission. The characteristic and

performance of the EMI filter can be estimated by obtaining the noise source impedance of the SMPS and the impedances of the filter components [6-13]. Thus, extracting the in-circuit impedance is also useful for conducted emission analysis and EMI filter design.

Acknowledging the significance of in-circuit impedance of an electrical system, the in-circuit impedance measurement and its applications are the major focuses of this thesis. Subsequent section of this chapter reviews the existing in-circuit impedance measurement approaches. Particularly, the two-probe inductive coupling approach is introduced in details. Also, advantages and drawbacks of these existing in-circuit impedance measurement approaches are compared and discussed. Finally, the motivation, objective, and scope of this thesis are covered in this chapter.

1.1 Literature Review of In-circuit Impedance Measurement Approaches

Despite of the importance of the in-circuit impedance measurement on many aspects, only a few in-circuit impedance measurement approaches are found to be demonstrated with the ability to measure the in-circuit impedances of electrical systems through a comprehensive literature review. These approaches are the transformer ratio-arm (TRA) bridge approach [14], current sensing approach [15], and the two-probe inductive coupling approach (or commonly known as two-probe approach) [16, 17]. In this section, the three approaches are reviewed and their merits and shortcomings are discussed.

1.1.1 TRA Bridge Approach

The measurement instrument of the TRA bridge approach consists of a TRA, a quadrature amplifier, and two standard capacitors. Taking Z_X as the impedance to be measured in an arbitrary circuit network shown in Fig. 1-1, the schematic circuit and measurement setup of the TRA bridge approach to measure Z_X is shown in Fig. 1-2.

Fig. 1-1 Arbitrary circuit network containing impedance to be measured Z_X [14].

Fig. 1-2 Schematic circuit and measurement setup of the TRA bridge approach [14].

As shown in Fig. 1-1, it requires points 1, 2, and 4 to be grounded when Z_X is measured. For any circuit network consisting of only passive components, grounding any points of the network will not cause any problem to the network. However, if Z_7 is replaced with a power supply, measuring Z_X with the TRA bridge approach will short the power supply and the measured Z_X is not the actual in-circuit impedance under its usual operating condition due to the shorted power supply. Thus, the TRA bridge approach can measure the in-circuit impedance of an electrical component without disconnecting it from the circuit, but it requires the power supply to be turned off during the measurement, which may not reflect the true operating condition.

Even if the power supply is at the place of Z_{10} where the power supply is not required to be shorted, the measurement of Z_X needs manual adjustment according to measurement setup shown in Fig. 1-2, in addition to grounding a few points. As shown in Fig. 1-2, C_1 and C_2 are the two adjustable standard capacitors required by the approach for the capacitive coupling. To measure Z_X , two capacitors and the TRA together need to be adjusted manually to balance the bridge circuit. The adjustment of three components may even be ineffective if the measurement frequency is higher than the resonant frequency of the capacitors. Also, the TRA bridge approach does not provide any guideline on the selection of the standard capacitors. Even if Z_X is successfully obtained through measurement, the obtained result is only the magnitude of Z_X according to the circuit configuration in Fig. 1-2 and the phase of Z_X cannot be obtained.

Hence, the measurement setup of the TRA bridge approach is cumbersome and it needs the power supply of the electrical system to be turned off, which may not be

always possible. The lack of phase information, which identifies the measured impedance's characteristic, i.e. resistive, inductive or capacitive, does not provide the full details of the impedance to be measured. For condition monitoring, the phase information of the electrical system can be associated with various faulty conditions. For EMI filter design, [13] has demonstrated the significance of knowing the phase information of the SMPS, which helps to design an effective EMI filter.

1.1.2 Current Sensing Approach

As shown in Fig. 1-3, R_X is the resistance to be measured. The measurement instrument of the current sensing approach consists of a miniature clamp-on current probe, an amplifier-detector, an AC power supply, a precision potentiometer R_R , a shunt component Z_S , and a resistor R_G with high resistance as the limiting resistor.

Fig. 1-3 Schematic circuit of the current sensing approach for measurement of resistor [15].

The two terminals of the AC power supply are connected to the two terminals of R_X and the current probe is clamped on one of the terminals. The current sensing approach

can measure the actual in-circuit impedance under the operating condition of R_X . However, it also needs manual adjustment of R_R during the measurement, and the measurement range of resistance is limited by the range of R_R . Another drawback of the current sensing approach is that the circuit in Fig. 1-3 is only applicable for the measurement of a resistor. For the measurement of capacitor, a more complex setup is shown in Fig. 1-4.

Fig. 1-4 Schematic circuit of the current sensing approach for measurement of capacitor [15].

As shown in Fig. 1-4, an inductor, a resistor, and a capacitor are added to the circuit in Fig. 1-3 when C_X is measured. Also, manual adjustment of R_R is necessary. For the measurement of inductor, the circuit is the same as Fig. 1-4, but the direction of reference current flow through the current probe has to be reversed.

In summary, the measurement setup of the current sensing approach is rather complex and the measurement ranges of the resistance, capacitance, and inductance are highly dependent on the range of R_R . The validated measurement ranges of the capacitance and inductance are 100 pF to 60 μ F and 400 μ H to 60 H, respectively. If

the impedance to be measured is a complex circuit network consisting of a combination of resistor, capacitor, and inductor, the selection of measurement circuit in Fig. 1-3 or Fig. 1-4 will be ambiguous and challenging. Since the measurement circuit and setup depends on prior information of the impedance to be measured, it may not be the case for most practical situations.

1.1.3 Basics of Current Probe and Two-Probe Approach

The measurement instruments of the two-probe approach are the vector network analyzer (VNA) and two clamp-on type current probes, one as an injecting probe and another as a receiving probe. The current probe makes in-circuit impedance measurement an attractive approach since the current probe has no direct electrical contact to the electrical system and can be easily mounted on-site for any high-voltage circuit under its usual operating condition without safety hazard concern.

In this section, the basics of the current probe are briefly introduced before the two-probe inductive coupling approach is described.

Fig. 1-5 shows various commercially available current probes.

(a)

(b)

Fig. 1-5 Current probes available in industry. (a) Large current probes for thick wire and high current rating. (b) Small current probes for thin wire and low current rating.

The current probe is designed to inject or to receive the AC current on the wire or cable being clamped [18]. The equivalent circuit of a current probe is inherently identical to a two-winding voltage transformer in that it possesses a pair of mutually coupled windings wound on a magnetic core [19]. Given the structure of the current probe, the equivalent circuit of a current probe can be modelled as shown in Fig. 1-6.

Fig. 1-6 Equivalent circuits of current probe. (a) Full equivalent circuit with parasitic components. (b) Simplified equivalent circuit.

As shown in Fig. 1-6 (a), i is the current of the wire clamped by the current probe. The ratio $1:n$ is the turn ratio between the primary and secondary windings which are equivalent to the wire and the current probe, respectively. C_p is the parasitic capacitance of the primary winding. V_s is the voltage induced across the terminals of the secondary winding due to i . L_m is the magnetizing inductance on the secondary side. R_c is the core loss equivalent parallel resistance on the secondary side. L_l is the leakage

inductance. R_s is the secondary winding resistance. C_s is the secondary stray capacitance. R is the sensing resistor and L_R is the inductance of the sensing resistor. Fig. 1-6 (b) shows the simplified equivalent circuit of the Fig. 1-6 (a) when the frequency is low and the parasitic components can be ignored. The operating bandwidth of the current probe for impedance extraction depends very much by the parasitic components. To characterize the current probe for in-circuit impedance extraction, a two-port network model will be developed to model the current probe accurately when it is used for in-circuit impedance measurement, which will be elaborated further in the subsequent section.

Besides of the equivalent circuit, another important parameter of the current probe is its transfer impedance, Z_T , which is usually provided in the probe's data sheet. Z_T is defined as the conversion ratio of the secondary voltage of the current probe to the primary current on the wire being measured by the current probe, which is $Z_T = \frac{V_s}{i}$ with unit of Ω based on the circuit in Fig. 1-6 (b). By connecting the cable of the current probe to a voltage measuring device [20], such as the oscilloscope, the current on the wire being clamped by the current probe can be calculated with the known voltage from the voltage measuring device and the known Z_T .

According to the transformer structure of Fig. 1-6 (b), i can be expressed as:

$$i = \frac{n}{R} V_s \quad (1-1)$$

Substituting (1-1) into the above expression of Z_T , there is:

$$Z_T = \frac{R}{n} \quad (1-2)$$

(1-2) shows that Z_T is a constant which is only determined by the resistance of the sense resistor and the turn ratio of a specific current probe regardless of the frequency. However, the actual Z_T , which is obtained through measurement, only remains constant at certain frequency range due to the current probe's non-linearity caused by the magnetizing current and hysteresis of the probe's magnetic core [21, 22]. As an example, Fig. 1-7 shows the frequency response of Z_T of the Tektronix CT1 current probe [23].

Fig. 1-7 Frequency response of Z_T of Tektronix CT1 current probe [23].

As shown in Fig. 1-7, Z_T remains the nominal value of 5 Ω from 1 to about 200 MHz when the DC current is 0 A, and it shows that the DC current would affect the frequency response of Z_T .

[24] provides the standard measurement setup, which is shown in Fig. 1-8, in compliance with ISO 11452-4 [25, 26] to measure Z_T .

Fig. 1-8 Measurement setup to obtain Z_T .

As shown in Fig. 1-8, the measurement setup needs the VNA to measure the S-parameters and the current probe calibration fixture, which holds the current probe and connects the 50 Ω load and VNA's port 1. The current probe connects to VNA's port 2. Then Z_T is calculated with the S_{21} measured from VNA as follows [24]:

$$Z_T = 50(S_{21}) \quad (1-3)$$

where $S_{21} = \frac{V_2^-}{V_1^+}$, V_2^- is the received voltage measured at port 2 through the current probe and V_1^+ is the signal source voltage of port 1.

After knowing the basics of the current probe, the two-probe approach is described in details.

The measurement setup of the two-probe approach to measure the in-circuit impedance of the device under test (DUT) powered by an AC or DC power source is shown in Fig. 1-9.

Fig. 1-9 Measurement setup of two-probe approach.

As shown in Fig. 1-9, the instruments for the measurement setup consist of an injecting current probe, a receiving current probe and a VNA. The DUT is powered by an AC or DC power source. A radio frequency (RF) signal is injected to the circuit loop, which consists of the DUT, the power source, and the wiring connection to the DUT and the power source, from VNA's port 1 through the injecting probe. The same signal is received at VNA's port 2 through the receiving probe. V_S is the voltage of the power source, Z_S is the impedance of the power source and Z_X is the impedance of the DUT under operating condition. Z_W is the impedance of the equivalent series resistance and series inductance of the wiring connection.

To analyze the measurement setup of Fig. 1-9, an equivalent circuit derived from the measurement setup is shown in Fig. 1-10.

Fig. 1-10 Equivalent circuit of measurement setup.

As shown in Fig. 1-10, V_I and the $50\ \Omega$ resistor represent the injecting voltage source and the output impedance of the VNA's port 1, respectively. The other $50\ \Omega$ resistor at the VNA's port 2 represents its input impedance. The injecting probe and the circuit loop form a mutual inductance M_1 . The receiving probe and the circuit loop form the mutual inductance M_2 . V_{P1} and V_{P2} are the voltages at ports 1 and 2, respectively. Z_{P1} and Z_{P2} are the input impedances of the injecting and receiving probes, respectively. I_1 is the current excited by V_I to the injecting probe and I_2 is the current coupled from the circuit loop by the receiving probe. I_W is the resultant current on the circuit loop. The power source is either DC or at very low power frequency and the frequency of I_1 is much higher than the frequency of V_S . At the frequency range of interest, V_S in Fig. 1-9 is a short circuit.

According to the three mutual coupled circuits in Fig. 1-10 and applying Kirchhoff's voltage law, there will be three equations as follows.

$$V_1 = (50 + Z_{P1})I_1 + (-j\omega M_1)I_W \quad (1-4)$$

$$(50 + Z_{P2})I_2 = -j\omega M_2 I_W \quad (1-5)$$

$$(Z_W + Z_S + Z_X)I_W + j\omega M_2 I_2 + (-j\omega M_1)I_1 = 0 \quad (1-6)$$

Through getting the expressions of I_1 and I_2 from (1-4) and (1-5), respectively; and substituting the expressions of I_1 and I_2 into (1-6), (1-6) can be rewritten as:

$$V_{M1} = (Z_X + Z_{M1} + Z_{M2} + Z_W + Z_S)I_W \quad (1-7)$$

where

$$Z_{M1} = \frac{(\omega M_1)^2}{50 + Z_{P1}} \quad (1-8)$$

$$Z_{M2} = \frac{(\omega M_2)^2}{50 + Z_{P2}} \quad (1-9)$$

$$V_{M1} = \frac{j\omega M_1 \cdot V_1}{50 + Z_{P1}} \quad (1-10)$$

According to (1-7), the components of the two circuits associated with VNA's port 1 and port 2 in Fig. 1-10 are reflected to the circuit loop of the DUT and a resultant equivalent circuit is shown in Fig. 1-11.

Fig. 1-11 New equivalent circuit of measurement setup.

As shown in Fig. 1-11, V_{M1} is the voltage coupled to the circuit loop of the DUT due to M_1 . Z_{M1} and Z_{M2} are the reflected impedances due to M_1 and M_2 , respectively. Since Z_X is the impedance to be measured, the sum of Z_S , Z_W , Z_{M1} , and Z_{M2} in (1-7) can be lumped together as Z_{setup} . Thus, (1-7) can be rewritten as:

$$V_{M1} = (Z_X + Z_{setup})I_W \quad (1-11)$$

From (1-11), Z_X can be determined by:

$$Z_X = \frac{V_{M1}}{I_W} - Z_{setup} \quad (1-12)$$

According to Fig. 1-10, I_W is measured by the receiving probe and thus can be expressed as:

$$I_W = \frac{V_{P2}}{Z_T} \quad (1-13)$$

where Z_T is the transfer impedance of the receiving probe. Substituting (1-10) and (1-13) into (1-12), Z_X is will be expressed by:

$$Z_X = \left(\frac{j\omega M_1 Z_T}{50 + Z_{P1}} \right) \left(\frac{V_1}{V_{P2}} \right) - Z_{setup} \quad (1-14)$$

According to the circuit associated with VNA's port 1 in Fig. 1-10, V_1 can be expressed as:

$$V_1 = \left(\frac{50 + Z_{P1}}{Z_{P1}} \right) V_{P1} \quad (1-15)$$

Substituting (1-15) into (1-14), Z_X is finally given by:

$$Z_X = K \left(\frac{V_{P1}}{V_{P2}} \right) - Z_{setup} \quad (1-16)$$

where

$$K = \frac{j\omega M_1 Z_T}{Z_{P1}} \quad (1-17)$$

From (1-16), $\left(\frac{V_{P1}}{V_{P2}}\right)$ can be measured through S-parameters measurement using the VNA. Since $V_{P1} = (S_{11} + 1)V_1$ and $V_{P2} = S_{21}V_1$, $\left(\frac{V_{P1}}{V_{P2}}\right)$ becomes:

$$\frac{V_{P1}}{V_{P2}} = \frac{S_{11}+1}{S_{21}} \quad (1-18)$$

Substituting (1-18) into (1-16), it results in:

$$Z_X = K \left(\frac{S_{11}+1}{S_{21}} \right) - Z_{setup} \quad (1-19)$$

Thus, Z_X is determined when K , Z_{setup} and $\left(\frac{S_{11}+1}{S_{21}}\right)$ are known. $\left(\frac{S_{11}+1}{S_{21}}\right)$ is directly measured by VNA when the DUT is on the circuit loop. To obtain K and Z_{setup} , the DUT is first replaced with a known precision resistor R_{std} and the S-parameters are measured. Then the DUT is made to be short-circuited and S-parameters are re-measured. With these two sets of measured S-parameters, (1-19) will result in two equations:

$$Z_X|_{Z_X=R_{std}} = K \left(\frac{S_{11}+1}{S_{21}} \right) \Big|_{Z_X=R_{std}} - Z_{setup} \quad (1-20)$$

$$Z_X|_{Z_X=0\Omega} = K \left(\frac{S_{11}+1}{S_{21}} \right) \Big|_{Z_X=0\Omega} - Z_{setup} \quad (1-21)$$

By solving (1-20) and (1-21), K and Z_{setup} are obtained. Finally, Z_X is solved by substituting the known K and Z_{setup} into (1-19).

The two-probe approach has been validated experimentally and applied for in-circuit impedance measurement on various applications, including EMI filter design [27], conducted emission analysis [28-30], signal integrity analysis [31] and characterization of passive components [32, 33].

Compared with the TRA bridge approach and the current sensing approach, the two-probe approach measures the in-circuit impedance under the operating condition of the DUT and it does not require manual adjustment on the measurement instruments. Besides, the VNA used in the two-probe approach has a much wider dynamic measurement range and is much more sensitive than the adjustable passive components used in the other two approaches, the measurement range of the two-probe approach is much wider than the other two approaches, as demonstrated in [16]. In addition, the two-probe approach can obtain both the magnitude and phase information of the impedance of the DUT through measurement without prior knowledge of the DUT, while the other two approaches can only obtain the magnitude information. Also, the inductive coupling of the two-probe approach has no direct electrical contact with a DUT that is powered by high voltage, which eliminates the safety hazard to the personnel who does the measurement.

1.2 Motivation and Objective of Thesis

As the in-circuit impedance provides significant information on various applications, existing in-circuit impedance measurement approaches are reviewed and compared in earlier sections. Although the two-probe approach outweighs the other two approaches, it still has certain drawbacks, which will be discussed in the next chapter. To address

these drawbacks, further improvement based on two-port network analysis, which also uses a VNA and two current probes as measurement instruments, will be developed through characterizing the current probes with $ABCD$ two-port network parameters. The improved approach will be validated experimentally and its applications for condition monitoring of electrical system and conducted emission analysis will also be demonstrated.

For condition monitoring, it is desirable to have an approach which can be adopted to monitor the common characteristic of most electrical systems with minimum modification to the electrical system and without interrupting the system's operation. In view of the presence of impedance as the fingerprint of every electrical system, impedance is one of the key parameters to be monitored. Also, the measurement does not require modification to the electrical system to be monitored and the RF power injected through the current probe is normally insignificant as compared with the power supply voltage to the electrical system, it is feasible to employ the proposed $ABCD$ two-port network approach to measure the in-circuit impedance of the electrical system. Subsequently, a practical example shows that the $ABCD$ two-port network approach has the ability to detect the change of in-circuit impedance of an electrical system under different operating conditions.

For conducted emission measurement, the line impedance stabilization network (LISN) provides a well-defined impedance to the SMPS. Thus, EMI filter can be designed with ease according to the standard conducted emission measurement with the presence of LISN and the known impedances of the LISN and SMPS. However, the SMPS connects the AC mains directly without LISN in reality and the impedance of

AC mains varies from location to location [34, 35], which means the conducted emission varies with the location of the AC mains. Consequently, the EMI filter which is designed according to the standard conducted emission measurement and installed on the SMPS may not effectively suppress the conducted emission in reality due to the deviation of AC mains impedance from the LISN impedance. Besides, there is no investigative study on the impact of the conducted emission measurement without LISN. Especially in certain situations, the SMPS or the device containing the SMPS can be too bulky to be tested in a laboratory and needs to be tested on site and testing with the LISN may not be possible. If the EMI filter needs to be designed for these situations, knowing the deviation of conducted emission from the actual test setup with LISN will be useful to ensure that the EMI filter designed takes care of such deviation. Thus, it is worth investigating the impact of the conducted emission measurement without LISN on conducted emission and EMI filter design [36]. Particularly, the deviation of the conducted emission without LISN from that of with LISN and the performance of the EMI filter designed according to the measurement with LISN in measurements with and without LISN are investigated and analysed from the perspective of noise impedances of the actual measurement setup including AC mains, SMPS, and LISN. To measure the noise impedances, the proposed *ABCD* two-port network approach is employed.

1.3 Scope of Thesis

This chapter introduces the significance of in-circuit impedance and reviews the existing in-circuit impedance measurement approaches. Then the motivation and objective of this thesis are described. The subsequent chapters are organized as follows.

Chapter 2 introduces the theory and measurement setup of the proposed *ABCD* two-port network as applied to two-probe inductive coupling approach. Then the *ABCD* two-port network theory for in-circuit impedance measurement is validated experimentally.

Chapter 3 introduces the background of condition monitoring and demonstrates the ability of the proposed approach on condition monitoring. Then, using an electrical motor and a transformer as two systems to be monitored, the proposed approach has demonstrated its ability to detect the change of in-circuit impedance under their usual operating conditions.

Chapter 4 describes the basics of conducted emission measurement and analyses the conducted emission deviation without LISN statistically based on the impedances measured with the proposed approach.

Chapter 5 covers the basics of EMI filter design. Then the statistical analysis in Chapter 4 is validated, and the impact of the conducted emission measurement without LISN is studied under the condition where the EMI filter designed according to the measurement with LISN is installed on the SMPS.

Chapter 6 concludes the thesis and discusses possible future work which is worth exploring further.

Chapter 2

In-Circuit Impedance Measurement Based on ABCD Two-Port Analysis

In Chapter 1, existing in-circuit impedance measurement approaches have been reviewed. Although the inductive coupling approach using two current probes is found to be the most advantageous for in-circuit impedance measurement, there are still limitations. The two-probe in-circuit measurement setup requires laborious pre-measurement calibration. It is first done with a known standard precision resistor as a DUT and then it is repeated again with the DUT replaced with a short circuit, as described in details in Chapter 1. For many practical applications, the DUT, which could be a large and complex electrical system, cannot be easily replaced with a known standard resistor due to either its size or inaccessibility. Also, the calibration under short circuit condition can only be implemented when the power supply to the DUT is turned off, which may not be always feasible if the DUT is mission-critical. In addition, the two-probe approach requires the calibration be repeated for every different circuit or system to be measured. Besides, the accuracy of the two-probe approach is found to be influenced by the value of the standard precision resistor selected for the pre-measurement calibration, which will be elaborated and explained in this chapter.

To overcome the aforementioned shortcomings of the two-probe approach, this chapter adopts the *ABCD* two-port network characterization of the measurement setup for the extraction of in-circuit impedance of any electrical system. The measurement setup is similar to Fig. 1-9, but the concept and mechanism of the *ABCD* two-port network characterization and analysis is different from the conventional two-probe approach introduced in Chapter 1.1.3.

2.1 *ABCD* Parameters of Two-Port Network

The use of the *ABCD* two-port network approach for two-probe inductive coupling approach, which uses a VNA and two current probes as measurement instruments, is to characterize the two current probes as *ABCD* two-port networks individually so that they can be cascaded with ease. For the benefit of readers, the *ABCD* parameters of a two-port network are briefly covered in this section.

The *ABCD* two-port network also known as chain parameters, allows cascading of two or more networks by simply multiplying the characterized matrices of the circuit networks. According to the two-port network in Fig. 2-1, the matrix relates input voltage and current (port 1) with output voltage and current (port 2) as follows [37]:

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} V_2 \\ I_2 \end{bmatrix} \quad (2-1)$$

where *A*, *B*, *C*, and *D* parameters are defined as:

$$A = \left. \frac{V_1}{V_2} \right|_{I_2=0} \quad (2-2)$$

$$B = \left. \frac{V_1}{I_2} \right|_{V_2=0} \quad (2-3)$$

$$C = \left. \frac{I_1}{V_2} \right|_{I_2=0} \quad (2-4)$$

$$D = \left. \frac{I_1}{I_2} \right|_{V_2=0} \quad (2-5)$$

Fig. 2-1 Two-port network represented by $ABCD$ matrix.

As shown in Fig. 2-1, I_1 and I_2 represent currents flowing into port 1 and flowing out of port 2, respectively. V_1 and V_2 represent voltages at ports 1 and 2, respectively.

According to (2-1) and Fig. 2-1, the cascade connection of multiple two-port networks can be illustrated as Fig. 2-2 [37].

Fig. 2-2 Cascade connections of multiple two-port networks.

Fig. 2-2 demonstrates that the two-port network from port 1 to port $N+1$ can be represented by the cascade connections of N two-port networks. In other words, the

$ABCD$ matrix $\begin{bmatrix} A' & B' \\ C' & D' \end{bmatrix}$ of the whole two-port network from port 1 to port 4 can be expressed as:

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} A' & B' \\ C' & D' \end{bmatrix} \begin{bmatrix} V_4 \\ I_4 \end{bmatrix} = \begin{bmatrix} A_1 & B_1 \\ C_1 & D_1 \end{bmatrix} \begin{bmatrix} A_2 & B_2 \\ C_2 & D_2 \end{bmatrix} \begin{bmatrix} A_3 & B_3 \\ C_3 & D_3 \end{bmatrix} \begin{bmatrix} V_4 \\ I_4 \end{bmatrix} \quad (2-6)$$

Hence,

$$\begin{bmatrix} A' & B' \\ C' & D' \end{bmatrix} = \begin{bmatrix} A_1 & B_1 \\ C_1 & D_1 \end{bmatrix} \begin{bmatrix} A_2 & B_2 \\ C_2 & D_2 \end{bmatrix} \begin{bmatrix} A_3 & B_3 \\ C_3 & D_3 \end{bmatrix} \quad (2-7)$$

where $\begin{bmatrix} A_n & B_n \\ C_n & D_n \end{bmatrix}$ represents the $ABCD$ matrix of the n^{th} two-port network.

In practice, A , B , C , and D parameters can be obtained by mathematical transforms from S-parameters measured with VNA. The transforms are given by [38]:

$$A = \frac{(Z_{01}^* + S_{11}Z_{01})(1 - S_{22}) + S_{12}S_{21}Z_{01}}{2S_{21}(R_{01}R_{02})^{1/2}} \quad (2-8)$$

$$B = \frac{(Z_{01}^* + S_{11}Z_{01})(Z_{02}^* + S_{22}Z_{02}) - S_{12}S_{21}Z_{01}Z_{02}}{2S_{21}(R_{01}R_{02})^{1/2}} \quad (2-9)$$

$$C = \frac{(1 - S_{11})(1 - S_{22}) - S_{12}S_{21}}{2S_{21}(R_{01}R_{02})^{1/2}} \quad (2-10)$$

$$D = \frac{(1 - S_{11})(Z_{02}^* + S_{22}Z_{02}) + S_{12}S_{21}Z_{02}}{2S_{21}(R_{01}R_{02})^{1/2}} \quad (2-11)$$

where $*$ represents the complex conjugate and Z_{0j} is the normalizing impedance for the j^{th} port [38]. In this thesis, the two-port networks are measured by the R&S ZVB8 VNA (300 kHz to 8 GHz), Z_{01} and Z_{02} are the source and load impedances of VNA. R_{01} and R_{02} are the resistances of Z_{01} and Z_{02} . Since the VNA ports are designed to match the 50 Ω system, Z_{01} , Z_{02} , R_{01} , and R_{02} are all 50 Ω .

To obtain (2-7) over a reasonably wide frequency range, measurements on a test jig, which consists of two sections of microstrip lines and a surface-mounted device (SMD) resistor on a printed circuit board (PCB), were carried out from 300 kHz to 100 MHz, as illustrated in Fig. 2-3. The measurement setup shown in Fig. 2-3 is to obtain the S-parameters of the two-port network between port 1 and port 2. Then the *ABCD* parameters of the measured network can be calculated from (2-8) to (2-11).

Fig. 2-3 Measurement setup to validate cascade connections of two-port networks and top view of test jig.

As shown in Fig. 2-3, the two-port network consists of two sections of microstrip lines with the SMD resistor as DUT. The microstrip lines 1 and 2 are of equal length of 1 cm. So the given two-port network can be considered as a cascade connection of three two-port networks according to Fig. 2-2. The $ABCD$ parameters of each microstrip line and the SMD resistor are given in Table 2-1 [37].

Table 2-1 $ABCD$ parameters of a series impedance model and transmission line model.

Circuit Model	$ABCD$ Parameters
<p>Port 1 Port 2</p> <p>Series Impedance Model</p>	$A=1$ $B=Z$ $C=0$ $D=1$
<p>Port 1 Z_o, β Port 2</p> <p>Lossless Transmission Line Model</p>	$A=\cos\beta l$ $B=jZ_o\sin\beta l$ $C=jY_o\sin\beta l$ $D=\cos\beta l$

In Table 2-1, Z_o , β , and L represent the characteristic impedance, the phase constant, and length of the transmission line, respectively. Z represents the impedance of a series component or a system. For the proposed measurement, Z is the SMD resistor's impedance, Z_o is 50 Ω and Y_o is $1/Z_o$. L is 0.01 m, which is the length of a section of microstrip line on the test jig shown in Fig. 2-3. β is given by

$$\beta = \frac{2\pi f \sqrt{\epsilon_{eff}}}{c} \quad (2-12)$$

where ϵ_{eff} is the effective dielectric constant of the microstrip line on the test jig and C is 3.0×10^8 m/s.

From Table 2-1, the SMD resistor with known resistance and the above given parameters, the $ABCD$ matrixes of the two-port networks of microstrip line 1, SMD resistor, and microstrip line 2 can be obtained and the three cascaded $ABCD$ matrixes will give the $ABCD$ matrix of the network between port 1 and port 2 according to (2-7). On the other hand, Table 2-1 shows that B of the product of three $ABCD$ matrixes consisting of the two-port networks of two microstrip lines and the resistor will be the dominant parameter among the $ABCD$ parameters and it will reflect the impedance of the SMD resistor. Hence, a comparison between the B from direct measurement according to Fig. 2-3 and the B from the product of three matrixes is done for the validation of (2-7). Three SMD resistors with resistance of 150 Ω , 510 Ω , and 1 k Ω were soldered separately on the test jig shown in Fig. 2-3 for the validation purposes. The comparison of the measured B according to the setup in Fig. 2-3 and the calculated B from cascading the three $ABCD$ matrixes derived from Table 2-1 is shown in Fig. 2-4 from 300 kHz to 100 MHz.

Fig. 2-4 Comparison of measured and calculated B .

B_{M150} , B_{M510} , and B_{M1k} are the B from direct measurements with resistances of 150 Ω , 510 Ω , and 1 k Ω , respectively. B_{C150} , B_{C510} , and B_{C1k} are the calculated B from matrix multiplication with resistances of 150 Ω , 510 Ω , and 1 k Ω , respectively. The results show that B obtained from measurement and B calculated from matrix multiplication are in good agreement. Also, it has shown that B is the corresponding resistance of the SMD resistor under test. Hence, (2-7) has been demonstrated with measurements.

2.2 ABCD Parameters of Two-Probe Measurement Setup

As mentioned earlier, the calibration of the two-probe approach requires the power source to be disconnected so that the electrical system can be replaced with a known resistor and a short circuit. However, in reality, cutting the power off from an electrical system may not always be feasible due to the criticality of the system. Even if cutting off the power is feasible, the electrical system may not be easily accessible for the replacement of standard resistor or short circuit. To overcome these constraints, an alternative approach to extract the in-circuit impedance of the electrical system is proposed.

The current probe can be modeled and characterized with S-parameters as three-port and two-port networks [39-41]. However, using S-parameters is not practical for modeling a cascaded two-port network. Since the current probe has not been modeled as the *ABCD* two-port network before, this section first introduces how the current probes and the electrical system shown in Fig. 1-9 can be modeled as cascaded *ABCD* two-port networks. Subsequently, the basic theory and calibration of the two-probe measurement setup are introduced in details.

As illustrated in Fig. 1-6, the current probe and the wire being clamped is basically a transformer with turn ratio of $n:1$ [16]. The equivalent transformer in Fig. 1-6 (b) can be modeled as a two-port network shown in Fig. 2-5, where one port is formed by the input (or output) port of the current probe and the other port is formed by the two ends of the wire being clamped.

Fig. 2-5 Equivalent two-port network formed by the current probe and wire.

According to the above equivalent two-port network and the series impedance model in Table 2-1, the injecting probe, receiving probe and the DUT shown in Fig. 1-9 can be modeled as three cascaded two-port networks, which are shown in Fig. 2-6.

Fig. 2-6 Equivalent two-port network of the measurement setup in Fig. 1-9.

M_{inj} is the two-port inductive coupling network of the injecting probe with the wire being clamped. M_{system} is the two-port network consisting of the electrical system (DUT in Fig. 1-9), the power source, and the wiring connection. M_{rec} is the two-port inductive coupling network of the receiving probe with the wire being clamped. V_{vna} is the signal source at port 1 of the VNA. With the network representation in Fig. 2-6, the two-port network M between ports 1 and 2 is related by:

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix}_{inj} \begin{bmatrix} A & B \\ C & D \end{bmatrix}_{system} \begin{bmatrix} A & B \\ C & D \end{bmatrix}_{rec} \quad (2-13)$$

where $\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{inj}$, $\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{system}$ and $\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{rec}$ are the $ABCD$ parameters of M_{inj} , M_{system} and M_{rec} , respectively.

By solving M_{system} , one can obtain Z_{system} , which is the sum of Z_X , Z_S , and Z_W , under its in-circuit condition since the B of M_{system} equals to Z_{system} according to Table 2-1. For an operating electrical system as the DUT, Z_{system} is usually dominated by Z_X , as Z_S and Z_W are relatively small. From (2-13), M_{system} can be solved once M , M_{inj} and M_{rec} are obtained. Before the actual in-circuit impedance measurement, M_{inj} and M_{rec} must be first characterized using the test fixture as shown in Fig. 2-7.

Fig. 2-7 Test fixture to characterize injecting and receiving probes (structural illustration and actual photo).

The fabricated test fixture, which is designed to characterize two Tektronix CT1 current probes (25 kHz to 1 GHz) [23] as injecting and receiving probes for the in-circuit measurement, is a single layer double-sided PCB (1.5 mm thickness) with a full ground plane on the bottom side. The CT1 with an aperture diameter of American wire gauge (AWG) 14 is clamped onto an AWG 22 wire of 3 cm length, which is the shortest possible length to minimize the impact of the parasitic effect of the wire in the

characterized $ABCD$ parameters. One end of the wire is connected to the signal pin of the sub-miniature version A (SMA) connector and the other end connected to the ground plane through a via. The SMA connector is to facilitate the connection to the VNA. Two CT1 probes are also employed for the subsequent measurements in this thesis.

To characterize M_{inj} , ports 1 and 2 of the VNA are connected to the injecting probe and SMA connector, respectively. To characterize M_{rec} , ports 1 and 2 of the VNA are connected to SMA connector and the receiving probe, respectively. Once M_{inj} and M_{rec} are characterized, M is measured by clamping the two current probes onto the wire between the electrical system and the power source.

With M , M_{inj} and M_{rec} characterized, M_{system} can be solved easily. M_{inj} and M_{rec} are re-usable as long as the same probes are employed in the measurements and hence, only M needs to be re-measured for different electrical systems.

2.3 Experimental Validation

2.3.1 Validation

The two CT1 probes will be pre-characterized with the test fixture shown in Fig. 2-7 for the injecting and the receiving current probes connecting to ports 1 and 2 of the VNA, respectively.

Fig. 2-8 shows the PCB for experimental validation. The PCB (1.5 mm thickness) has two sections of copper traces (total length of 3.5 cm and width of 0.2 cm)

connecting two ends of two AWG 22 wires (each 3 cm in length) on the upper side of the PCB. Two vias on the PCB with a separation distance of 2 cm connect the other two ends of the wires to the ground plane on the lower side of the PCB. Passive components with known values are treated as unknown DUT placed between the two copper traces. The components are SMD resistors: 30 Ω , 51 Ω , 62 Ω , 110 Ω , 150 Ω , 510 Ω and 1 k Ω ; SMD capacitors: 4.7 pF and 6.8 pF; and axial-lead ferrite-core inductors: 4.7 μ H and 47 μ H. The circuit loop consisting of the wires, copper traces and ground plane is electrically small as compared to the wavelength of 100 MHz, the highest frequency for the measurement.

Fig. 2-8 Setup and the PCB for experimental validation (structural illustration and actual photo).

The injecting and receiving CT1 probes clamped onto the wires form M_{inj} and M_{rec} , respectively. The DUT and the wiring connection to the DUT form M_{system} . So the in-circuit impedance to be measured will be the DUT impedance and wiring impedance in series. The impedance incurred by the wiring connection is found to be far smaller than the DUT impedance throughout the frequency range up to 100 MHz. Thus, the measured impedance is dominated by the DUT impedance.

The impedance of each of the aforementioned passive components is measured separately with an impedance analyzer. The measured impedance using the two-probe approach is compared with the measured impedance using the impedance analyzer and plotted in Fig. 2-9 and Fig. 2-10 from 300 kHz to 100 MHz. The measured impedance of the wiring connection is plotted in Fig. 2-11 from 300 kHz to 100 MHz.

(a)

Fig. 2-9 Impedances of resistors measured with the proposed approach (dot lines) and with impedance analyzer (dash lines). (a) Magnitudes. (b) Phases.

Fig. 2-10 Impedances of inductors and capacitors measured with the proposed approach (dot lines) and with impedance analyzer (dash lines). (a) Magnitudes. (b) Phases.

Fig. 2-11 Impedance magnitude of the wiring connection measured with the proposed approach.

As shown in Fig. 2-9 and Fig. 2-10, the measured impedance from the proposed method (dot lines) are in close agreement with measured impedance from an impedance analyzer (dash lines) up to 100 MHz. The phase deviations above about 10

MHz in Fig. 2-9 (b) are due to the resistor's parasitic effects. For low-resistance resistors, the parasitic inductance begins to surface. For high-resistance resistors, the parasitic capacitance dominates. For the magnitude resonance of the 47 μH inductor at 10.6 MHz in Fig. 2-10 (a), it is due to the inductor's parasitic capacitance. Above 10.6 MHz, the inductor behaves more like a capacitor. Fig. 2-11 shows that the measured impedance of the wiring connection, which is inductive in nature, is very small and less than 2.6 Ω at the highest frequency of 100 MHz.

The measurement results have demonstrated that the proposed approach has the ability to measure in-circuit impedances of resistors, inductors and capacitors up to 100 MHz with good accuracy.

Actually, the maximum measurement frequency of the proposed approach and the two-probe approach is found to be highly dependent on the effective bandwidth of the current probes. Although the CT1's user manual specifies its Z_T is 5 Ω with a measurement bandwidth from 25 kHz to 1 GHz, the actual effective bandwidth with Z_T of 5 Ω is demonstrated to be from 300 kHz to 100 MHz by the subsequent measurement. It is noted that the stated Z_T is calibrated with a 50 Ω load in the circuit loop when it is configured as a receiving probe [24]. In reality, the frequency response of Z_T can vary with the load impedance in the circuit loop. Similar to the measurement setup in Fig. 1-8, another test fixture designed according to the fixture in Fig. 1-8 with the same wire for characterization, the VNA, SMD resistors (10 Ω , 51 Ω , 150 Ω , 1 k Ω), and a CT1 probe as a receiving probe are used to observe the variation of Z_T of CT1 with different load SMD resistors. The test setup is illustrated in Fig. 2-12.

Fig. 2-12 Test setup to measure Z_T of CT1 (structural illustration and actual photo).

In Fig. 2-12, the SMD resistor is soldered on the fixture where there are two short sections of copper traces connecting both ends of the SMD resistor to the SMA connector's signal pin and the ground plane of the test fixture, respectively. The fixture employs the same PCB structure for the fixture in Fig. 2-7. The SMA connector and the CT1 current probe are connected to ports 1 and 2 of the VNA, respectively. The CT1

probe clamped onto the wire receives the RF signal injected from VNA's port 1. Z_T of CT1 can be computed from:

$$Z_T = \frac{V_2^-}{i} \quad (2-14)$$

where V_2^- is the received voltage measured at port 2 through CT1, i is the current due to the injected signal from port 1, which is given by:

$$i = \frac{V_1}{Z_R} = \frac{V_1^+ (1+S_{11})}{Z_R} \quad (2-15)$$

where V_1 is the voltage across the SMD resistor and V_1^+ is the signal source voltage of port 1. Z_R is the impedance of the SMD resistor. Substituting (2-15) into (2-14) leads to:

$$Z_T = \frac{Z_R \cdot V_2^-}{V_1^+ \cdot (1+S_{11})} = \frac{Z_R \cdot S_{21}}{1+S_{11}} \quad (2-16)$$

Thus, four frequency response curves of Z_T under different resistive loads are plotted from 300 kHz to 1 GHz and shown in Fig. 2-13.

Fig. 2-13 Magnitudes of Z_T of CT1 with different SMD resistors on test fixture.

As illustrated in Fig. 2-13, Z_{T10} , Z_{T51} , Z_{T150} , and Z_{T1k} denote four different Z_T due to the SMD resistors of 10 Ω , 51 Ω , 150 Ω , and 1 k Ω , respectively. All four Z_T remain a constant of 5 Ω below 100 MHz but begin to deviate above 100 MHz. It shows that the effective bandwidth of the probe is from 300 kHz to 100 MHz where Z_T is insensitive to the large variation of the impedance to be measured. Thus, it is demonstrated that the CT1 probes provide good accuracy up to 100 MHz for a large variation of impedance to be measured.

2.3.2 Comparison of Accuracy

As mentioned at the beginning of Chapter 2, the measurement accuracy of the two-probe approach in [6, 16, 27] is sensitive to the value of the precision resistor chosen for the setup calibration. To demonstrate this issue, a 110 Ω SMD resistor is used as a DUT. It is first measured with the two-probe approach when two different precision resistors ($R_{std} = 30 \Omega$ and 1 k Ω) are chosen for the setup calibration. Then it is re-measured again with the *ABCD* network method. The measurement results are plotted in Fig. 2-14 from 300 kHz to 100 MHz for comparison with the measured impedance using an impedance analyzer as a reference. For ease of comparison, the deviations of the measured impedances from the two approaches with reference to the impedance analyzer are tabulated in Table 2-2.

Fig. 2-14 Measured impedance of the 110 Ω SMD resistor with conventional calibration (solid lines), proposed *ABCD* network calibration (dot line) and impedance analyzer (dash line).

Table 2-2 Accuracy comparison between two approaches.

Frequency (MHz)	Deviation of Conventional Two-probe Approach (%)		Deviation of the Improved Approach (%)
	Setup Calibration with $R_{std}=30\ \Omega$	Setup Calibration with $R_{std}=1\ \text{k}\Omega$	
1	+ 1.4	+ 2.2	+ 0.3
10	+ 1.9	+ 2.2	+ 0.1
30	+ 1.5	+ 3.6	+ 0.6
50	- 0.5	+ 5.8	+ 0.7
70	- 2.7	+ 9.8	+ 1.1
100	- 8.5	+ 15.5	+ 3.6

Fig. 2-14 shows that measured impedance with the improved approach (dot line) is in close agreement with that measured with impedance analyzer (dash line) from 300 kHz to 100 MHz. However, the measured impedances using the conventional two-probe approach (solid lines) with different precision resistors ($R_{std}=30\ \Omega$ and $R_{std}=1\ \text{k}\Omega$) for calibration start to deviate at when the frequency exceeds 30 MHz and deviation becomes significant with increasing frequency. Table 2-2 shows that the deviation of the two-probe approach at 100 MHz can be as high as -8.5% if 30 Ω is used for the

setup calibration or +15.5% if 1 k Ω is chosen for the setup calibration. The worst deviation of the proposed approach is only +3.6% at 100 MHz. The results reveal that the conventional two-probe approach can only achieve good measurement accuracy at higher frequencies if the calibration resistor is as close to the impedance of the DUT as possible. Unfortunately, this prior knowledge may not be readily available in reality. On the other hand, the proposed approach overcomes this shortcoming and provides highly accurate in-circuit impedance extraction for many potential applications.

Based on the *ABCD* two-port network theory, this chapter proposes a characterization method to extract in-circuit impedance of any electrical system accurately under its operating conditions. Using a precision 110 Ω SMD resistor as a DUT, the proposed approach has demonstrated its measurement accuracy from 300 kHz to 100 MHz with a maximum deviation of 3.6%. As opposed to the conventional two-probe calibration method, the proposed method eliminates the need of a standard precision resistor for its setup calibration and therefore its measurement accuracy is independent of the setup resistor value.

Chapter 3

Application of the Proposed Method for Condition Monitoring of Electrical System

After introducing and validating the two-probe measurement setup based on *ABCD* two-port network characterization, this chapter demonstrates the practical applications of the proposed method for condition monitoring of electrical systems.

3.1 Condition Monitoring of Electrical Systems

Condition monitoring is a technique or a process of monitoring the characteristics or behaviors of a system in such a way that changes and trends of the monitored parameter can be used to predict the health or abnormality of the system before serious deterioration or breakdown occurs [42]. Before the emergence and development of condition monitoring techniques, electrical systems are inspected or maintained at a routine schedule and the systems have to be shut off. In other words, any faults emerged during the system's operation time cannot be detected timely for immediate remedy. Consequently, the undetected incipient faults may deteriorate gradually and lead to major system breakdown before next scheduled maintenance. On the other hand,

the offline scheduled maintenance would lead to unnecessary off-service hours of the system, which may have economic implications. In view of the aforementioned limitations, condition monitoring techniques, which can detect the operating system in real time, will be very useful.

Condition monitoring techniques of electrical systems, such as transformer, generator and motor in power plants, are widely adopted in industry because they have the potential to reduce the operating cost, enhance the reliability of the operation, and improve power supply and service to customers [43-52]. The beauty of the condition monitoring techniques is that they do not require the system to be turned off, so the operating electrical system can be monitored continuously. With the help of the techniques, faults in the operating system can be detected timely and the system breakdowns can be minimized.

Condition monitoring techniques are vital for monitoring the critical electrical infrastructures such as the power grid and power plant due to their significance to the society. Particularly, some condition monitoring techniques are developed to monitor the high-power cables and high-power transformers, which are the key assets of the power grid and power plant. For techniques to monitor the high-power transformer [52-66], there are dissolved gas-in-oil analysis (DGA), on-line frequency response analysis (FRA), vibration monitoring, temperature monitoring, etc. For techniques to monitor the high-power cable, there are on-line partial discharge (PD) methods [67-73] using capacitive couplers and various signal processing techniques, which are still under development. The aforementioned techniques are specifically developed for either the transformer or cable by monitoring the chemical, electrical, thermal, or mechanical

characteristics [55] of the system. Besides, these techniques need mechanical or electrical modification on the existing system and the electrical modification of the on-line FRA have to be carefully designed and implemented to avoid potential safety hazards.

3.2 In-circuit Impedance Measurement for Condition Monitoring

In view of the increasing number and complexity of critical electrical systems, there is an increasing demand for a condition monitoring technique which can be adopted in most of the electrical systems with minimum modification to the electrical systems and without interrupting the system's operation. Among various characteristics of the electrical system, impedance, which exists in every electrical system and can reflect the electrical status and health condition of the electrical system, is one of the most important parameters to be monitored in all electrical systems. Abnormalities, such as overloading, aging, partial short circuit, etc., can be detected by comparing and analyzing the impedance changes over a certain frequency range and over a period of time.

If the electrical system's in-circuit impedance can be extracted without interrupting its operation or changing its circuitry, it will be useful to detect the system's abnormality by monitoring the in-circuit impedance of the electrical system. By comparing and analyzing the extracted in-circuit impedance, the health condition of the electrical system can be detected early.

Based on the *ABCD* two-port network characterization method described in Chapter 2, this chapter shows how the in-circuit impedance extraction can be applied to condition monitoring of a DC motor and a transformer under their usual operating conditions.

A FA-130 DC motor powered by a Caltron PR-3504D DC power supply is used as a first example. The source voltage is kept at 3 V and the developed method is employed to measure the in-circuit impedance under two different operating conditions. For the first operating condition, there is nothing attached to the motor's axle (denoted as unloaded condition). For the second operating condition, an object weighed 6 g is attached to the motor axle (denoted as loaded condition). As expected, for the unloaded condition, the motor spins at higher speed and for the loaded condition, the motor spins at lower speed. The measurement setup is shown in Fig. 3-1.

Fig. 3-1 In-circuit impedance measurement setup on the operating DC motor.

As shown in Fig. 3-1, two AWG 22 wires, each 4 cm in length, connect the motor to the power supply. An RF signal is induced to the electrical system that consists of the

power supply, the connecting wires and the motor, through an injecting CT1 probe. The RF signal is picked up by the receiving CT1 probe.

As mentioned in Chapter 2.2, since the *ABCD* parameters of the two CT1 probes have been pre-characterized, only two more measurements are needed to extract the in-circuit impedances of the powered up motor system under unloaded and loaded conditions. The in-circuit impedances are measured and extracted and shown in Fig. 3-2 from 300 kHz to 100 MHz.

Fig. 3-2 In-circuit impedances of the operating motor system under unloaded and loaded conditions.

Fig. 3-2 shows that the impedance is inductive in nature, as it is dominated by the motor's coil winding. The parasitic capacitance of the coil causes a resonant frequency at around 16 MHz, and above 16 MHz, the impedance is dominated by the parasitic capacitance. The impedance under unloaded condition is noticeably higher than that of the loaded condition. Based on the impedance response below the resonant frequency,

the inductances of the motor coil under unloaded and loaded conditions are about 11 μH and 8 μH , respectively.

The drop in coil inductance is expected due to the saturation of the soft iron core where the coil windings are wound. The DC supply current to the motor increases from 0.65 A to 1.80 A when the 6 g load is applied to the motor. The increase in DC supply current to the motor has resulted in core saturation. Through this example, it has demonstrated that the proposed method has the ability to monitor the impedance change of an electrical system under different operating conditions. The ability to monitor the change is useful so that any unusual behavior of the system can be detected early and rectified.

The in-circuit measurement of a DC motor under its usual operation has also demonstrated that the *ABCD* two-port network approach can detect the change of the in-circuit impedance for real-time condition monitoring purpose. The proposed approach does not require the power supply to the system to be disconnected for setup calibration. The injecting and receiving probes used need only one-time characterization to obtain the *ABCD* parameters and the characterized *ABCD* parameters are reusable for subsequent measurements using the same probes. Although the example shown is a DC motor operated at low-voltage DC, the proposed approach can be extended to high-voltage electrical system, of course with a larger current probe and bigger wire size for pre-characterization. The proposed approach opens the doors for an accurate real-time condition monitoring of high-voltage electrical system for detection of early signs of potential faults before the system breaks down.

The second practical example for condition monitoring is a 1.5 kW transformer. The transformer is powered by the 230 V AC mains at the primary side. The method developed in Chapter 2 is employed to measure the in-circuit impedance at the primary side of the transformer under two different operating conditions. To emulate the two different operating conditions, two different secondary load resistors, 220 Ω and 330 Ω are chosen. Through measuring the in-circuit impedance at the primary side, the impedance change at the output load variation can be detected at the primary side. The measurement setup is shown in Fig. 3-3.

Fig. 3-3 In-circuit impedance measurement setup on the 2:1 step-down transformer.

As shown in Fig. 3-3, two AWG 10 wires, each 60 cm in length, connect the primary of the transformer to the AC mains with a mains voltage V_{AC} and an impedance Z_{AC} . An RF signal is induced to the electrical system that consists of the AC mains, the connecting wires and the reflected primary impedance of the transformer with secondary loaded, through an injecting probe. The RF signal is picked up by the

receiving probe. Since the measurement is done on a high-power transformer, the Solar 9144-1N (4 kHz to 100 MHz) and Solar 9134-1 (20 Hz to 100 MHz) current probes, which have much higher power handling capability than the CT1 current probes, are chosen as the injecting and receiving probes for the measurement, respectively. According to the specs of two probes, which show that the transfer impedance of the receiving probe and the insertion loss of the injecting probe remain constant from 100 kHz to 10 MHz, the effective measurement frequency range is from 100 kHz to 10 MHz. Considering the transformer's magnetizing inductance, which is a dominant factor of the transformer impedance on the primary side, is very large and the apparent impedance change on the primary side of the transformer normally happens at the parallel resonant frequency due to parasitic capacitance of the primary winding, usually around several hundreds kHz. Thus, the frequency range for monitoring is kept within 100 kHz and 1 MHz. To cater to the frequency range for measurement, the Bode 100 VNA with measurement frequency range from 1 Hz to 40 MHz is employed.

Before measuring the in-circuit impedance, the injecting and receiving probes must be characterized in terms of $ABCD$ parameters from 100 kHz to 1 MHz. Similar to the CT1 current probes' characterization setup described in Chapter 2.2, the characterization setups of the injecting and receiving probes are done with the Solar 9125-1 characterization fixture and shown in Fig. 3-4.

(a)

(b)

Fig. 3-4 Characterization setups of (a) Solar 9144-1N injecting probe (b) Solar 9134-1 receiving probe.

In Fig. 3-4 (a), the injecting probe connecting the VNA's port 1 is clamped on the inner conductor of the fixture. One end of the fixture is connected to the VNA's port 2 and the other end is terminated with a coaxial short. In Fig. 3-4 (b), the receiving probe connecting the VNA's port 2 is clamped on the inner conductor of the fixture. Similarly,

one end of the fixture is connected to the VNA's port 1 and the other end is terminated with a coaxial short. Through measuring the S-parameters from 100 kHz to 1 MHz according to the procedure described in Chapter 2, the $ABCD$ parameters of the injecting and receiving probes are characterized.

Once both probes are properly characterized, the in-circuit impedances under two different loading conditions ($Z_{T,220\Omega}$ and $Z_{T,330\Omega}$) are measured using the setup in Fig. 3-3 with the 220 Ω and 330 Ω resistors connected on the secondary side, respectively. The measured in-circuit impedances are shown in Fig. 3-5 from 100 kHz to 1 MHz.

Fig. 3-5 In-circuit impedances of the operating transformer system with two different resistors on the secondary side.

As expected, Fig. 3-5 shows that the impedance is inductive in nature, as it is dominated by the transformer's magnetizing inductance. The parasitic capacitance of the transformer's primary winding causes a resonant frequency at around 360 kHz, and above 360 kHz, the impedance is dominated by the parasitic capacitance of about 0.13

nF. The impedance measured with 220 Ω resistor is noticeably higher than that of with 330 Ω resistor from 200 to 400 kHz. The reason of lower loading resistance results in the earlier mentioned behavior can be explained through analyzing the transformer's equivalent circuits [74-76] shown in Fig. 3-6.

Fig. 3-6 Transformer's equivalent circuits (a) with primary and secondary sides (b) with secondary side reflected to primary side.

Fig. 3-6 (a) shows the equivalent circuit of a transformer. L_m is the magnetizing inductance on the primary side. C_p and C_s are the parasitic capacitances in the primary

and secondary windings, respectively. C_w is the parasitic capacitance between the primary and secondary windings. R_{mp} and R_{ms} are the core resistances on the primary and secondary sides, respectively. L_{lkp} and L_{lks} are the leakage inductances of the primary and secondary windings, respectively. R_{wp} and R_{ws} are the winding resistances of the primary and secondary windings, respectively. Z_{wire} is the resultant impedance of two 60 cm wires. R_s is the resistance of the resistor connected on the secondary side. The impedance to be measured is Z_T , which is the resultant impedance on the right side of the dash line a.

For ease of analyzing Z_T , the components on the secondary side of the transformer are reflected to the primary side and the new equivalent circuit is shown in Fig. 3-6 (b). After reflecting the components to the primary side, C_s , R_{ms} , L_{lks} , R_{ws} and R_s become $1/4C_s$, $4R_{ms}$, $4L_{lks}$ (L_e), $4R_{ws}$ and $4R_s$, respectively. In Fig. 3-6 (b), C_w is neglected since it is usually very small as compared with other parasitic capacitances. C_{ps} is the capacitance of C_p and $1/4C_s$ in parallel. R_m is the resistance of R_{mp} and $4R_{ms}$ in parallel. R_e is the resistance of $4R_{ws}$ and $4R_s$ in series. Thus, Z_T is given as:

$$Z_T = j2\pi f L_{lkp} + R_{wp} + Z_{AC} + Z_{wire} + Z_{parallel} \quad (3-1)$$

where $Z_{parallel}$ is the impedance of R_m , C_{ps} , L_m and $(L_e + R_e)$ in parallel. From (3-1), it is $Z_{parallel}$ that causes the difference between $Z_{T,220\Omega}$ and $Z_{T,330\Omega}$. And since the impedance of R_m does not vary with frequency, the difference between $Z_{T,220\Omega}$ and $Z_{T,330\Omega}$ is caused by the parallel circuit of C_{ps} , L_m and $(L_e + R_e)$. Normally, L_m is at tens of mH level and L_e is only a few μ H. Thus, the impedance of L_m is much larger than that of $(L_e + R_e)$ from 100 kHz to 1 MHz when R_s is 220 Ω or 330 Ω . Under such a condition, the parallel

circuit of C_{ps} , L_m and (L_e+R_e) can be represented by the parallel circuit of C_{ps} and (L_e+R_e) . The impedance (Z_{LCR}) of C_{ps} and (L_e+R_e) in parallel is expressed as:

$$Z_{LCR} = \frac{(j2\pi f L_e + R_e) \frac{1}{j2\pi f C_{ps}}}{j2\pi f L_e + R_e + \frac{1}{j2\pi f C_{ps}}} \quad (3-2)$$

When $f = \frac{1}{2\pi\sqrt{L_e C_{ps}}}$, the magnitude of Z_{LCR} reaches the maximum, which is

$$|Z_{LCR}|_{max} = \sqrt{\left(\frac{L_e}{C_{ps} R_e}\right)^2 + \frac{L_e}{C_{ps}}}. \text{ Since } L_e \text{ and } C_{ps} \text{ are fixed values and } R_e \text{ varies with } R_s,$$

smaller R_s will result in larger $|Z_{LCR}|_{max}$, which results in larger peak magnitude of Z_T at resonance. Hence, the measurement result is consistent with the analysis.

The second application example demonstrates that the proposed method has the ability to monitor the in-circuit impedance change of an electrical system connecting to high voltage, if proper current probes are chosen. As the transformer is a critical equipment in a power delivery system, the demonstration of this practical example indicates the potential of applying the proposed method to monitor the impedance change of other critical electrical infrastructure.

Chapter 4

Using In-Circuit Impedance to Study the Impact on Conducted Emission Measurement without LISN

Besides condition monitoring, the method developed in this thesis can also be employed for the analysis of conducted emission of a SMPS connected to the power mains. This chapter investigates the deviation of the conducted emission measurement without and with LISN from the perspective of the in-circuit impedance. Before investigating the conducted emission measurement without LISN, the basics of the standard conducted emission measurement with LISN will be briefly described.

4.1 Conducted Emission Measurement with LISN

Conducted emission is the unwanted noise current which is caused by a SMPS and flows to the AC mains through cables [77]. Hence, from EMI analysis perspective, the SMPS and AC mains can be treated as the noise source and noise load, respectively. Contaminated by the noise current, the AC mains becomes noisy and may affect the operation of the other equipment connected to the same AC mains.

In general, conducted emission I_{noise} can be defined as follows:

$$I_{noise} = \frac{V_{noise}}{Z_{noise,circuit}} \quad (4-1)$$

where V_{noise} and $Z_{noise,circuit}$ represent the equivalent noise source voltage and the circuit impedance of the noise current path, respectively. V_{noise} is produced by the SMPS and $Z_{noise,circuit}$ consists of the noise source impedance of the SMPS, noise termination impedance of the AC mains and impedance of the cable connecting the SMPS and AC mains. Hence, the conducted emission is dependent on the circuit impedance of the noise current path.

4.1.1 Source of Conducted Emission

As compared with the linear power supply, which dissipates power constantly due to the linear operating region of the regulating transistor, the SMPS has much higher efficiency since the metal oxide semiconductor field-effect transistor (MOSFET) in the SMPS operates as a switching element which dissipates very little power [78]. Besides, the SMPS is much lighter in weight than the linear power supply since the SMPS needs a much smaller magnetic core for the transformer with operating frequency of tens or hundreds of kHz, which is much higher than that of the linear power supply. Due to the superiorities of the SMPS over the linear power supply on efficiency and weight, it has gained popularity and widely adopted in electronic products. However, the high frequency switching of the MOSFET generates harmonics with frequencies extending into the MHz range.

To understand noise generation of the SMPS, the operating principle of the SMPS is explained with a typical flyback converter [34], which is shown in Fig. 4-1.

Fig. 4-1 Schematic circuit of a flyback converter.

As shown in Fig. 4-1, the SMPS connects to the AC mains on the left. On the primary side, the 50/60 Hz AC waveform is first rectified by the full-wave bridge rectifier and then smoothed by the bulk capacitor C_{B1} . The voltage across C_{B1} is DC with a constant voltage, which is close to the peak voltage of the AC waveform. A pulsed waveform V_G , which is generated from the pulse width modulator, fed to the gate of the MOSFET Q_1 to switch it on and off repeatedly. A resistor R_G connecting the gate of Q_1 is to vary the rise/fall times of V_G . Increasing and decreasing the resistance of R_G will lead to longer and shorter rise/fall times of V_G , respectively. Since Q_1 dissipates power and produces heat during its switching, a grounded heat sink is attached to the body of Q_1 to help dissipate the heat effectively. Thus, a parasitic capacitance C_{para} is formed between the heat sink and Q_1 . With the switching of Q_1 and the DC voltage supply across C_{B1} , the voltage across the primary winding of the transformer T becomes a pulsed waveform with alternating polarity. Through coupling, the voltage across the secondary winding of T also has the pulsed waveform and its

magnitude is determined by the turn ratio of the primary and secondary windings. Through the rectification of the full-wave rectifier, smoothing of the bulk capacitor C_{B2} , and filtering of the low-pass filter on the secondary, the pulse waveform on the secondary winding finally becomes DC with specific constant value at the SMPS's output, where it supplies DC power to electronic circuits. To maintain a stable DC output voltage in response to the variation of load, the output voltage is sampled by a detection circuit and then fed to the pulse width modulator, which varies the duty cycle of V_G according to the output voltage changes. Besides converting the input AC voltage to the desired output DC voltage, the SMPS generates unwanted RF noise currents which flow to the AC mains due to the switching of Q_I . There are two conduction modes for the RF noise currents.

The first conduction mode of RF noise current is called the common-mode (CM) current. The pulse voltage on Q_I due to the switching causes the charging and discharging of C_{para} between Q_I and the heat sink. Thus, the CM current is generated along with the charging and discharging of C_{para} , which becomes the CM noise source. Fig. 4-2 shows the path of the CM current in the SMPS.

Fig. 4-2 Illustration of the CM current path in SMPS.

As illustrated in Fig. 4-2, the CM current (I_{CM}) is flowing along the dash line entering into the AC mains through the live and neutral lines and leaving the AC mains through the ground line. Hence, the live and neutral lines can be treated as parallel as one conductor between the SMPS and the AC mains. The CM current passes Q_1 , C_{B1} , two turn-on diodes of the rectifier, live and neutral lines, AC mains, and finally returns back to C_{para} through the ground line and the grounded heat sink.

The second conduction mode of RF noise current is called the differential-mode (DM) current. A pulsed current is drawn from C_{B1} to the primary winding side when Q_1 is switching. Since C_{B1} is not ideal and has equivalent series inductor (ESL) and resistor (ESR) which both dominate the impedance of C_{B1} at the switching frequency, the current flowing through C_{B1} generates a noise ripple voltage across C_{B1} , which becomes a DM noise voltage source causing DM noise current flowing to the AC mains. Fig. 4-3 shows the path of the DM current in the SMPS.

Fig. 4-3 Illustration of the DM current path in SMPS.

As illustrated in Fig. 4-3, the DM current (I_{DM}) is flowing along the dash line entering into the AC mains through the live line and leaving the AC mains through the neutral line. I_{DM} passes C_{BI} , one of the turn-on diodes of the rectifier, live line, AC mains, and finally returns back to C_{BI} through the neutral line and another turn-on diode of the rectifier.

With I_{CM} and I_{DM} , the live (I_L) and neutral (I_N) line noise currents can be expressed phasor sum and phasor difference as follows:

$$I_L = I_{CM} + I_{DM} \quad (4-2)$$

$$I_N = I_{CM} - I_{DM} \quad (4-3)$$

4.1.2 CM and DM Impedances

Besides of the CM and DM sources, the other key factor determining the conducted emission is the CM and DM impedances. According to the above illustrations of the

CM and DM current paths, there are equivalent CM and DM impedances in the SMPS, cables, and AC mains.

The CM impedance of the SMPS is dominated by C_{para} between Q_1 and heat sink since the impedance of C_{para} is much larger than the impedance of C_{BI} at the frequency of interest. Thus, the CM impedance of the SMPS is capacitive in nature. The DM impedance of the SMPS is dominated by the ESR and ESL of C_{BI} at the switching frequency and thus the DM impedance is an inductive element and resistive element in series.

The CM impedance of the cables is the series impedance of the ground line and the parallel of the live and neutral lines. The DM impedance of the cables is the series impedance of the live and neutral lines. Both the CM and DM impedances of the cables are resistive and inductive elements in series.

The CM and DM impedances of the AC mains are undetermined since the AC mains' impedance between live and ground lines and impedance between neutral and ground lines vary from location to location [34, 35] and usually not readily available. Thus, the CM and DM currents of a specific SMPS with specific cables can vary with different AC mains.

4.1.3 LISN and Conducted Emission Measurement

In order to design an EMI filter to suppress conducted emission, the magnitude of the conducted emission must be first obtained through measurement. Since different AC mains connecting to the same SMPS yields different conducted emission, taking the

AC mains as the noise terminating load to quantify and evaluate the conducted emission can be ambiguous. Hence, international standards require the AC mains supplying power through LISN, which provides a well-defined mains impedance to the SMPS. Besides stabilizing the line impedance, LISN prevents external conducted EMI on the AC mains from contaminating the measurement without disturbing the 60 or 50 Hz AC power to the SMPS [79-81]. The schematic circuit of a typical LISN and the measurement setup of I_L are shown in Fig. 4-4.

Fig. 4-4 Schematic circuit of LISN and measurement setup of I_L .

As shown in Fig. 4-4, the LISN in the dash box is inserted between the AC mains and the SMPS. The SMPS with a DC load is the equipment under test (EUT) whose I_L

is measured with a spectrum analyzer (SA). The LISN consists of two 1 μF capacitors, two 0.1 μF capacitors, two 50 μH inductors, and two 1 $\text{k}\Omega$ resistors. The 1 μF capacitors between live and ground lines and between neutral and ground lines are to bypass any external noise from the AC mains to ground so that the measured conducted emission produced by the SMPS would not be contaminated by other noise. Similarly, the 50 μH inductors on the live and neutral lines are to block the high frequency noise from the AC mains without affecting the 50 Hz/60 Hz power supply. The 0.1 μF capacitors between live and ground lines and between neutral and ground lines are to provide low impedance paths for the I_L and I_N test ports, which are designed to connect the SA for the measurement of I_L and I_N . The 1 $\text{k}\Omega$ resistors between live and ground lines and between neutral and ground lines are to provide static charge paths to discharge the 0.1 μF capacitors when two test ports are not connected to SA. The actual photo of the LISN is showing the LISN's front side, whose live, neutral, and ground ports are designed to connect the input terminals of the SMPS, and the back side of the LISN connects the AC mains. I_L is measured by connecting the SA 50 Ω (input impedance) to the I_L test port with a coaxial cable and terminating the I_N test port with a 50 Ω dummy load. Consequently, connecting the SA to the I_L test port is equivalent to parallel the SA's 50 Ω input termination to the 1 $\text{k}\Omega$ resistor which is between live and ground lines, and the 50 Ω dummy load on the I_N test port is equivalent to parallel the 50 Ω to the 1 $\text{k}\Omega$ resistor which is between neutral and ground lines. Similarly, I_N is measured by connecting the SA to the I_N test port and terminating the I_L test port with a 50 Ω dummy load. Both I_L and I_N are measured from 150 kHz to 30 MHz according to

international standards, such as federal communications commission (FCC) part 15 and Comité International Spécial des Perturbations Radioélectriques (CISPR) 22.

With two equivalent $50\ \Omega$ resistors paralleling two $1\ \text{k}\Omega$ resistors respectively when I_L or I_N is measured, the effect of two $1\ \text{k}\Omega$ resistors can be neglected, as $1\ \text{k}\Omega \gg 50\ \Omega$. Hence the circuit of the LISN between the live and ground lines or between the neutral and ground lines can be simplified to Fig. 4-5.

Fig. 4-5 Simplified circuit of LISN between the live and ground lines or between the neutral and ground lines.

According to Fig. 4-5, the LISN line impedance (Z_{LISN}) seen from the SMPS side is given by:

$$Z_{LISN} = \frac{(R_1 + \frac{1}{j\omega C_1})(j\omega L_1 + \frac{1}{j\omega C_2})}{R_1 + \frac{1}{j\omega C_1} + j\omega L_1 + \frac{1}{j\omega C_2}} \quad (4-4)$$

Based on (4-4), Z_{LISN} can be determined from 150 kHz to 30 MHz and plotted in Fig. 4-6.

Fig. 4-6 Magnitude of Z_{LISN} .

As shown in Fig. 4-6, Z_{LISN} approaches 50 Ω when frequency increases. This implies that the equivalent 50 Ω resistor dominates Z_{LISN} at high frequency. Thus, Z_{LISN} between the live and ground lines and between the neutral and ground lines can be approximated to be 50 Ω from 150 kHz to 30 MHz. From the perspective of conducted emission, the LISN's capacitors can be assumed to be short circuits and the LISN's inductors can be assumed to be open circuits. With the aforementioned approximation and assumption, I_L and I_N would only flow through the equivalent paralleled 50 Ω resistors and return to SMPS through the ground line when I_L and I_N are measured. The simplified equivalent circuit of the LISN is shown in Fig. 4-7 and I_L and I_N are expressed in terms of the voltages (V_L and V_N , respectively) measured with the SA as:

$$V_L = 50I_L \quad (4-5)$$

$$V_N = 50I_N \quad (4-6)$$

Substituting (4-2) and (4-3) into (4-5) and (4-6), respectively, leading to:

$$V_L = (I_{CM} + I_{DM})50 = V_{CM} + V_{DM} \quad (4-7)$$

$$V_N = (I_{CM} - I_{DM})50 = V_{CM} - V_{DM} \quad (4-8)$$

where $V_{CM} = 50I_{CM}$ and $V_{DM} = 50I_{DM}$ represent the CM and DM voltages measured by the SA, respectively. Normally, the voltage magnitudes of V_L and V_N are expressed in unit of dB μ V (i.e. dB with respect to μ V) against the limits given in FCC part 15 and CISPR 22 standards.

Fig. 4-7 Simplified equivalent circuit of LISN and current paths of I_L and I_N in LISN.

According to (4-2) and (4-3), the current paths of I_{CM} and I_{DM} in the simplified equivalent circuit of the LISN are shown in Fig. 4-8.

Fig. 4-8 Current paths of I_{CM} and I_{DM} in LISN.

As shown in Fig. 4-8, I_{CM} flows along the dash arrows and I_{DM} flows along the dot arrows. According to the CM current path, the CM impedance ($Z_{LISN,CM}$) of the LISN is equivalent to two $50\ \Omega$ resistors in parallel connection, which is $25\ \Omega$. According to the DM current path, the DM impedance ($Z_{LISN,DM}$) of the LISN is equivalent to two $50\ \Omega$ resistors in series connection, which is $100\ \Omega$.

To measure V_{CM} and V_{DM} , a CM-DM noise separation network, is needed, which is shown in Fig. 4-9. It will be inserted between the LISN and the SA.

Fig. 4-9 Schematic circuit of the CM-DM noise separation network.

As shown in Fig. 4-9, the schematic circuit of the EMI noise separation network is in the dash box, and the $50\ \Omega$ resistor is the input termination of the SA. Two ports on one side of the separation network connect to the LISN's I_L and I_N test ports, respectively, and the other port of the separation network connects to the SA. The switch at the V_N side is to change the polarity of V_N so that it is either V_N or $-V_N$ on the secondary side of the transformer. Through two 1:1 wideband RF transformers (model: LF428, VARI-L Co.) which have an essentially flat frequency response from 10 kHz to 50 MHz and center-tapped secondaries and are DC isolated [34, 82], the separation network's output port connecting the SA has the voltage of $V_L + V_N$ or $V_L - V_N$, which is $2V_{CM}$ or $2V_{DM}$, respectively. Thus, the CM and DM conducted emissions measured with the separation network and SA are $2V_{CM}$ and $2V_{DM}$, respectively.

Besides of the aforementioned measurement setups in Fig. 4-4 and Fig. 4-9, the conducted emission can be measured with the current probe and SA together, which is introduced later in this chapter.

Knowing the magnitude of the conducted emission from measurement, the measured conducted emission is then compared with the conducted EMI limits of FCC part 15 or CISPR 22 accordingly so that the excessive conducted emission which should be suppressed can be known through comparison. The conducted EMI limits of CISPR 22 are shown in Fig. 4-10.

Fig. 4-10 Conducted EMI limits of CISPR 22.

With the known excessive conducted emission after comparing with the conducted EMI standard, an EMI filter will be designed to bring the emission below level lower than the conducted EMI limit. The filter design procedure is introduced in next chapter.

4.2 Analysis of Conducted Emission Measurement without LISN

As mentioned in last section, the standard conducted emission measurement requires the presence of LISN and the corresponding measurement setups with LISN have been

illustrated. The LISN is to ensure the repeatability of conducted emission measurements [83] so that the measurement can be done at any location with a well-defined AC mains impedance. It is also well known that the performance of an EMI filter depends on the impedances of noise source and noise termination. Several methods have been developed to estimate or measure these impedances for EMI filter design purposes [4, 5, 9, 16, 84]. For the AC mains impedance, there has been some work reported on the representativeness of the LISN impedance to be treated as the representative of the AC mains impedance for conducted emission measurement [3, 7, 10]. Although [85] provide statistical AC mains impedance data and conclude that the LISN impedance is statistically acceptable to represent the commercial AC mains impedance, no further investigative study has been conducted on how the deviation of mains impedance from the LISN impedance affects conducted emission measurement [86]. [87] demonstrates that conducted emission measurements without and with LISN can be very different. Hence, the actual performance of the EMI filter designed according to the measurement with LISN would be different from its expected performance when the SMPS with the designed filter is connected directly to the AC mains. In certain situations, the SMPS for very high power application can be too bulky to be tested in a laboratory and needs to be tested on site and testing with the LISN may not be possible. If the EMI filter needs to be designed for these situations, the deviation of conducted emission from the actual test setup with LISN will be helpful to ensure that the EMI filter designed takes care of such deviation.

In this section, the deviation of the conducted emission measurement without LISN from that of with LISN will be investigated from the perspectives of both CM and DM

impedances of the actual measurement setup including AC mains, SMPS, and LISN. To cater for the AC mains impedance variation from location to location, a total of 10 AC mains from various locations are selected for the investigative study to ensure the conclusion is representative for most situations. To extract the CM and DM impedances under the actual operating condition, the proposed approach based on *ABCD* two-port network is employed [88]. Through analyzing the extracted CM and DM impedances of the AC mains, LISN, SMPS, and wiring connection statistically, the deviations of the CM and DM conducted emissions without LISN can be evaluated. With the evaluated CM and DM deviations due to variation of AC mains impedances, a guideline for EMI filter design is recommended.

4.2.1 CM and DM Impedances and Deviations

Since the SMPS employs an active switching device, which has the on and off states, it will have the on-state and off-state impedances. However, only the on-state CM and DM impedances are the major concern for the analysis on conducted emission because the conducted emission is highest when the rectifier bridge and the MOSFET are turned on for the conduction of the CM and DM currents from the SMPS to the AC mains or LISN. Hence, the measured impedance from the proposed measurement method measures the on-state CM and DM impedances. During the CM/DM impedance measurement with the proposed method, the SMPS switches periodically and the VNA also receives the on-state and off-state RF signals periodically. Given the measurement condition, the VNA usually detects and records the stronger received RF signal, which is usually the on-state condition of the SMPS. According to the

illustrations of Fig. 4-2 and Fig. 4-3, the SMPS's on-state CM/DM impedance is much smaller than that of the off-state, so the received on-state RF signal at the receiving probe will be much higher than the received off-state signal. Although the SMPS cannot be fixed at the on or off state to compare the on-state and off-state impedances to verify the above explanation, the on-state and off-state impedances of a diode circuit shown in Fig. 4-11 can be measured to verify the explanation. According to Fig. 4-11, the circuit loop consists of a PCB with a $30\ \Omega$ SMD resistor R_D and a SMD schottky diode D in series connection and is powered by a DC power supply. The in-circuit impedance of the circuit when the diode is turned on and off can be measured separately.

Fig. 4-11 Measurement of on-state and off-state impedances of a circuit loop with a diode.

The diode and resistor are connected to the DC power supply with two AWG 22 wires (each 4 cm in length). The diode is turned on and off with 3 V and -3 V by changing the polarities of the DC power supply, respectively. Then the on-state (Z_{on})

and off-state (Z_{off}) impedances of the circuit loop are measured with the proposed method and plotted in Fig. 4-12 from 300 kHz to 100 MHz.

Fig. 4-12 On-state and off-state impedances of the circuit loop.

As shown in Fig. 4-12, Z_{on} is much smaller than Z_{off} from 300 kHz to 100 MHz. The high impedance of Z_{off} is mainly due to the junction resistance and capacitance of D when D is turned off with -3 V. Thus, the on-state RF signal will be dominant in the VNA's received signals when the CM and DM impedances of the SMPS are measured with the proposed method, and the measured CM and DM impedances of the SMPS in the subsequent sections are on-state impedances, which actually causes the highest conducted emission and worst case.

Using the proposed method based on $ABCD$ two-port network, the CM and DM impedances of AC mains at 10 different locations, an unfiltered SMPS (model: VTM22WB, 15 W, +12 V_{DC}/0.75 A, -12 V_{DC}/0.5 A) which is treated as an EUT and connects the resistive load R at its output terminal, LISN (model: Electro-Metrics MIL

5-25/2), and the wiring connection are measured. The CT1 current probes are chosen for measurement, as they have been pre-characterized as injecting and receiving probes from 300 kHz to 100 MHz. Also, the small size of CT1 probe allows the wiring connection of the conducted emission measurement to be as short as possible so that the CM and DM impedances of the wiring connection have little effects on the measurement. The wires used in subsequent measurements consist of the AWG 20 multi-strand wire connecting to the three-pin power plug.

To investigate the deviation of conducted emission without LISN, the study and analysis will be carried out with the LISN first and then followed by without the LISN.

A. CM Impedance Extraction and CM Deviation Analysis

Referring to the measurement setup in Fig. 1-9, the measurement setups to extract the CM impedances of the LISN, SMPS and AC mains are shown in Fig. 4-13 (a), (b), and (c), respectively.

(b)

(c)

Fig. 4-13 Measurement of CM impedances of (a) LISN (b) SMPS (c) AC mains.

As shown in Fig. 4-13, the live and neutral wires clamped by two CT1 probes are treated as one wire for the CM impedance measurement. According to the illustration of Fig. 2-6 and the above measurement setup, the aforementioned CM impedances can be measured.

Fig. 4-13 (a) measures the CM impedance of the LISN ($Z_{LISN,CM}$) and the setup connection ($Z_{Setup,CM}$) consisting of wires and the two 3.3 nF capacitors in series. The use of the 3.3 nF capacitors in Fig. 4-13 (a) and (c) is to provide a low-impedance CM path. The capacitance value is chosen such that it does not introduce excessive earth leakage current that causes tripping of the circuit breaker. $Z_{LISN,CM}$ can be obtained by subtracting $Z_{Setup,CM}$ from the measured CM impedance. $Z_{Setup,CM}$ can be obtained by switching off the mains and shorting the three wire terminals on the LISN side.

Fig. 4-13 (b) measures the CM impedance of the LISN, the wiring connection ($Z_{W,CM}$) and the SMPS ($Z_{SMPS,CM}$) in series under actual powered up condition. $Z_{W,CM}$ can be obtained by shorting both terminals of the three wires. $Z_{SMPS,CM}$ can be determined by subtracting the known $Z_{LISN,CM}$ and $Z_{W,CM}$.

Finally, Fig. 4-13 (c) measures the CM impedance of the AC mains ($Z_{Mains,CM}$) and $Z_{Setup,CM}$ in series. $Z_{Mains,CM}$ is extracted by subtracting the known $Z_{Setup,CM}$ from the measured CM impedance.

According to the measurement setup shown in Fig. 4-13 (b), both the CM current and the injected RF signal from the VNA are flowing on the live and neutral wires and will be picked up by the CT1 probes, so the CM current may affect the measurement accuracy of $Z_{SMPS,CM}$ if the CM current is higher than the current of the injected RF signal. To ensure the measurement of $Z_{SMPS,CM}$ is not affected by the CM noise current, the power levels of the CM current and the injected RF signal are measured according to the measurement setup in Fig. 4-14 and compared in Fig. 4-15.

Fig. 4-14 Measurement of power levels of CM current and injected RF signal from VNA.

As shown in Fig. 4-14, two CT1 probes clamping on the live and neutral wires are connected to the port 1 of the VNA and the input of the SA, respectively. The model of the SA is R&S FSL6, which is also employed for the subsequent measurements in this thesis. The power source at port 1 of the VNA injects RF signal with power level of 0 dBm from 300 kHz to 30 MHz to the live and neutral wires through the CT1 probe, and the SA is set to capture the maximum signal on the live and neutral wires throughout the spectrum through another CT1 probe. As long as the captured power level ($P_{inj,CM}$) is much higher than the power level (P_{CM}) of the CM current, the impedance measurement according to Fig. 4-13 (b) will not be affected. And according to Fig. 4-14, P_{CM} can be captured with the same SA setting for capturing $P_{inj,CM}$ but disconnecting the VNA from the CT1 probe. $P_{inj,CM}$ and P_{CM} are compared and plotted in Fig. 4-15 from 300 kHz to 30 MHz.

(a)

(b)

Fig. 4-15 (a) $P_{inj,CM}$ (b) P_{CM} .

As shown in Fig. 4-15, $P_{inj,CM}$ is generally much higher than P_{CM} above 2.5 MHz except for some frequencies. Hence, the measurement of $Z_{SMPS,CM}$ above 2.5 MHz will not be affected with the CM noise current. Below 2.5 MHz, $P_{inj,CM}$ and P_{CM} have little difference, as $Z_{SMPS,CM}$ is expected to be very high at low frequency due to its capacitive nature according to the illustration in Fig. 4-2 and it causes the CM current and the current of the injected RF signal to be lower than the noise floor of the SA. Hence, both $P_{inj,CM}$ and P_{CM} below 2.5 MHz in Fig. 4-15 are not showing the actual

power levels. But that does not mean that the $Z_{SMPS,CM}$ below 2.5 MHz cannot be measured. Since the VNA has a dynamic measurement range of 120 dB from 300 kHz to 30 MHz [89], which is 40 dB higher than that of the SA [90], the VNA has much higher sensitivity than the SA and can distinguish the difference between $P_{inj,CM}$ and P_{CM} . The capacitive profile of the measured $Z_{SMPS,CM}$ in Fig. 4-16 validates the capability of the VNA for impedance measurement with the presence of conducted emission. If P_{CM} is higher than or comparable to $P_{inj,CM}$, the power level of the VNA output source can be increased accordingly.

With the aforementioned impedance measurement procedure, $Z_{Mains,CM}$ can be measured and extracted at 10 different locations ($Z_{Mains,CM(n)}$, $n = 1$ to 10, where n denotes the measured CM mains impedances at the n^{th} location). The mean value of the CM mains impedances, $\bar{Z}_{Mains,CM}$, will be used for a quick analysis. $\bar{Z}_{Mains,CM}$ is plotted in Fig. 4-16 together with $Z_{LISN,CM}$, $Z_{W,CM}$, and $Z_{SMPS,CM}$.

Fig. 4-16 Magnitudes of $Z_{LISN,CM}$, $Z_{W,CM}$, $Z_{SMPS,CM}$, and $\bar{Z}_{Mains,CM}$.

$Z_{SMPS,CM}$ is capacitive in nature with a capacitance of 40 pF. As expected, $Z_{LISN,CM}$ is inductive initially and converges to 25 Ω with increasing frequency. $Z_{W,CM}$ is inductive with an inductance of about 0.16 μ H. Base on the trend of $\bar{Z}_{Mains,CM}$ with frequency, the CM mains impedance is generally following an inductive trend and it increases from a few Ω to about 100 Ω .

The CM conducted emission with LISN ($I_{LISN,CM}$) and without LISN ($I_{w/oLISN,CM(n)}$, $n = 1$ to 10, where n denotes the CM current measured at the n^{th} location) can be expressed as:

$$I_{LISN,CM} = \frac{V_{SMPS,CM}}{Z_{LISN,CM} + Z_{SMPS,CM} + Z_{W,CM}} \quad (4-9)$$

$$I_{w/oLISN,CM(n)} = \frac{V_{SMPS,CM}}{Z_{Mains,CM(n)} + Z_{SMPS,CM} + Z_{W,CM}} \quad (4-10)$$

where $V_{SMPS,CM}$ is the equivalent CM noise source voltage of the SMPS, which will be the same for situation without or with LISN. When the AC mains is connected to the LISN, the CM current will not flow into the AC mains since the 50 μ H inductors of the LISN block the CM current and the 0.1 μ F capacitors of the LISN bypass the CM current back to the SMPS at the frequency range of interest. Hence, $Z_{Mains,CM(n)}$ will not affect $I_{LISN,CM}$ and therefore will not be taken into consideration in (4-9). Similarly, the DM impedance of the AC mains will not affect the DM current with LISN and will also not be taken into consideration in the expression of the DM current with LISN.

From (4-9) and (4-10), the deviation of the CM conducted emission without LISN from that of with LISN, in dB, can be expressed as:

$$20 \log |I_{w/oLISN,CM(n)}| - 20 \log |I_{LISN,CM}| =$$

$$20 \log \left| \frac{Z_{LISN,CM} + Z_{SMPS,CM} + Z_{W,CM}}{Z_{Mains,CM(n)} + Z_{SMPS,CM} + Z_{W,CM}} \right| \quad (4-11)$$

Hence, the average deviation of the CM conducted emission without LISN from that of with LISN ($\overline{Dev}_{CM}$) is given by:

$$\overline{Dev}_{CM} = \frac{1}{10} \cdot \sum_{n=1}^{10} 20 \log \left| \frac{Z_{LISN,CM} + Z_{SMPS,CM} + Z_{W,CM}}{Z_{Mains,CM(n)} + Z_{SMPS,CM} + Z_{W,CM}} \right|. \quad (4-12)$$

The CM emissions without and with LISN can be measured directly according to the measurement setup in Fig. 4-17 with a CT1 probe and the SA. The average deviations based on CM emission measurement and based on (4-12) are plotted in Fig. 4-18 from 300 kHz to 30 MHz.

Fig. 4-17 Measurement setup to measure CM emissions without and with LISN.

Fig. 4-18 Comparison of average CM deviations based on CM emission measurement and CM impedance calculation.

In general, Fig. 4-18 indicates that the trend of the average CM deviation with frequency based on (4-12) is consistent with the average CM deviation based on CM emission measured without and with LISN. The average CM deviation based on measurement has limitation to obtain the true deviation at some cases where the CM conducted emission is too low to be detected due to the noise floor of the SA. Hence, the deviation based on (4-12) provides more representative results for in-depth analysis. It also reveals the actual cause of the deviation. According to Fig. 4-16, below 15 MHz, the overall CM impedance without LISN is dominated by the CM impedance of the SMPS and therefore the expected deviation is negligible. From 15 to 25 MHz, the inductive CM mains impedance becomes comparable to the capacitive CM SMPS impedance. Consequently, a series resonance happens around this frequency range and the overall CM impedance without LISN drops significantly and results in much higher CM emission without LISN. Above 25 MHz, the overall CM impedance without LISN

is expected to be dominated by the CM mains impedance and results in lower CM emission without LISN.

As the analysis based on the average CM mains impedance may not provide a complete picture, further statistical analysis is necessary. With the 10 sets of measured CM impedances, the standard deviation of the 10 sets of deviations based on the measured CM impedances (σ_{CM}) can be computed. For a 99.7% confidence level, the expected deviation falling within $\overline{Dev}_{CM} \pm 3\sigma_{CM}$ is plotted in Fig. 4-19 from 300 kHz to 30 MHz.

Fig. 4-19 CM deviation without LISN (n =10, confidence level = 99.7%).

Fig. 4-19 shows that below 6 MHz, the deviation without LISN is insignificant. Above 6 MHz, the deviation starts to surface and becomes significant. At 23 MHz, the deviation can be as high as ± 16 dB. From the EMI filter design perspective, such a significant deviation must be accounted for to ensure that the filter is adequately designed for situation without LISN.

B. DM Impedance Extraction and DM Deviation Analysis

The measurement setups to extract the DM impedances of the LISN, AC mains, and SMPS are shown in Fig. 4-20 (a), (b), and (c), respectively.

(a)

(b)

(c)

Fig. 4-20 Measurement of DM impedances of (a) LISN (b) AC mains (c) SMPS.

Fig. 4-20 (a) and (b) measure the DM impedance of the LISN ($Z_{LISN,DM}$) and the wiring ($Z_{W,DM}$) in series, and the DM impedance of the AC mains ($Z_{Mains,DM}$) and $Z_{W,DM}$ in series, respectively. The 1 μ F capacitor provides a low-impedance DM path for the injected RF signal from VNA's port 1. $Z_{LISN,DM}$ and $Z_{Mains,DM}$ can be obtained by subtracting $Z_{W,DM}$ from the two measured DM impedances, respectively. $Z_{W,DM}$ can be obtained by shorting the two wire terminals on the LISN side with the power switched off.

Fig. 4-20 (c) measures the DM impedance of $Z_{W,DM}$ and the SMPS ($Z_{SMPS,DM}$) in series under actual powered up condition. $Z_{SMPS,DM}$ can be obtained by subtracting $Z_{W,DM}$ from the measured DM impedance.

According to the measurement setup shown in Fig. 4-20 (c), both the live-to-neutral current and the injected RF signal from the VNA are flowing on the live and neutral

wires and will be picked up by the CT1 probes. In order to ensure that the live-to-neutral conducted emission does not affect the measurement of $Z_{SMPS,DM}$, the injected RF signal must be higher than the emission level. The power levels of the live-to-neutral conducted emission and the injected RF signal are measured according to the measurement setup in Fig. 4-21 and compared in Fig. 4-22 from 300 kHz to 30 MHz.

Fig. 4-21 Measurement of power levels of live-to-ground current and injected RF signal from VNA.

As shown in Fig. 4-21, the two CT1 probes clamping on the live wire are connected to the port 1 of the VNA and the SA, respectively. The power source at port 1 of the VNA injects RF signal with power level of 0 dBm from 300 kHz to 30 MHz to the live wire through the CT1 probe, and the SA is set to capture the maximum signal on the live wire throughout the spectrum through the CT1 probe. As long as the captured injected power level ($P_{inj,LN}$) is much higher than the power level (P_{LN}) of the live-to-neutral conducted emission, the validity of the impedance measurement according to

Fig. 4-20 (c) will be ensured. $P_{inj,LN}$ and P_{LN} are compared and plotted in Fig. 4-22 from 300 kHz to 30 MHz.

(a)

(b)

Fig. 4-22 (a) $P_{inj,LN}$ (b) P_{LN} .

As shown in Fig. 4-22, $P_{inj,LN}$ is generally much higher than P_{LN} from 300 kHz to 30 MHz. Hence, the measurement of $Z_{SMPS,DM}$ is validated.

With the aforementioned impedance measurement procedure, $Z_{Mains,DM}$ can be extracted at the 10 different locations ($Z_{Mains,DM(n)}$, $n = 1$ to 10, where n denotes the

measured DM mains impedances at the n^{th} location). The mean value of the DM mains impedances, $\bar{Z}_{Mains,DM}$, is plotted in Fig. 4-23 together with $Z_{LISN,DM}$, $Z_{W,DM}$, and $Z_{SMPS,DM}$ from 300 kHz to 30 MHz.

Fig. 4-23 Magnitudes of $Z_{LISN,DM}$, $Z_{W,DM}$, $Z_{SMPS,DM}$, and $\bar{Z}_{Mains,DM}$.

$Z_{SMPS,DM}$ can be represented as a resistance of 80 Ω and an inductance of 3 μH in series. As expected, the $Z_{LISN,DM}$ converges to about 100 Ω resistance with increasing frequency. $Z_{W,DM}$ is inductive in nature with an inductance of about 0.2 μH . Base on the trend of $\bar{Z}_{Mains,DM}$ with frequency, the DM mains impedance is inductive in general, with impedance increases from a few Ω to about 150 Ω .

The average DM deviation of the conducted emission measurement without LISN from that of with LISN ($\overline{Dev}_{DM}$) is given by:

$$\overline{Dev}_{DM} = \frac{1}{10} \cdot \sum_{n=1}^{10} 20 \log \left| \frac{Z_{LISN,DM} + Z_{SMPS,DM} + Z_{W,DM}}{Z_{Mains,DM(n)} + Z_{SMPS,DM} + Z_{W,DM}} \right| \quad (4-13)$$

The DM emissions without and with LISN can be measured directly according to the measurement setup in Fig. 4-24 with a CT1 probe and the SA. The average deviations based on DM emission measurement and based on (4-13) are plotted in Fig. 4-25 from 300 kHz to 30 MHz.

Fig. 4-24 Measurement setup to measure DM emissions without and with LISN.

Fig. 4-25 Comparison of average DM deviations based on DM emission measurement and DM impedance calculation.

In general, Fig. 4-25 shows that the trend of average DM deviation with frequency based on (4-13) is consistent with that based on measurement without and with LISN. From Fig. 4-23, below 7 MHz, the overall DM impedance without LISN is expected to be smaller than the overall DM impedance with LISN, which results in higher DM emission without LISN. From 7 to 12 MHz, the overall DM impedance without LISN is dominated by the DM mains impedance, which is higher than the DM impedance of the LISN, and results in lower DM emission without LISN. Above 12 MHz, the overall DM impedance without LISN is dominated by the DM impedance of the SMPS and therefore the deviation is insignificant.

Similarly, further statistical analysis on the DM deviation is also studied. With the 10 sets of measured DM impedances, the standard deviation of the 10 sets of deviations based on the measured DM impedances (σ_{DM}) can be computed. The expected deviation for 99.7% confidence level ($\overline{Dev}_{DM} \pm 3\sigma_{DM}$) is plotted in Fig. 4-26.

Fig. 4-26 DM deviation without LISN (n =10, confidence level = 99.7%).

Fig. 4-26 shows that below 1 MHz, the deviation is rather small. Above 1 MHz, the deviation becomes significant and it can be about ± 11 dB under the worst case scenario. Again, the DM EMI filter must take this deviation into consideration to ensure that the filter is adequately designed for situation without LISN.

By extracting the CM and DM impedances with the *ABCD* two-port network approach, the impacts of CM and DM conducted emissions without LISN are investigated systematically. The causes of the deviations without LISN from that of with LISN can be explained through the known extracted impedances. Based on a sample size of 10 measured impedances and 99.7% confidence level, the worst deviation can exceed 10 dB.

Chapter 5

EMI Filter Performance without LISN

In this chapter, the CM and DM deviations evaluated based on the statistical analyses of the extracted CM and DM impedances are validated by comparing the conducted emissions measured with and without LISN. Besides, the impact of the conducted emission measurement without LISN is further demonstrated under the condition where an EMI filter designed according to the measurement with LISN is installed on the SMPS. With the installed filter, the conducted emissions measured with and without LISN are compared again.

Since an EMI filter is designed according to the measurement with LISN for demonstration, the EMI filter components and EMI filter design procedure for the measurement with LISN are briefly introduced in this chapter.

5.1 EMI Filter Components

The passive components commonly used for EMI filter design are CM and DM chokes and X and Y capacitors. The characteristics, configurations, and operating mechanisms of these components are covered in this section.

5.1.1 CM and DM Chokes

A CM choke can suppress the CM noise currents effectively. It consists of a ferrite core with high relative permeability, and two sets of windings wound on the same core. The two windings are wound such that the magnetic fluxes due to CM currents add within the core and the magnetic fluxes due to DM currents cancel within the core [91, 92]. Hence, it introduces high inductance for CM current path but negligible inductance for DM current. Equivalent circuits of a CM choke under CM and DM conditions are illustrated in Fig. 5-1 (a) and Fig. 5-1 (b), respectively.

Fig. 5-1 Configurations and simplified equivalent circuits of CM choke as EMI filter under (a) CM current (b) DM current.

In Fig. 5-1 (a), both CM currents on the live and neutral wires produce two magnetic fluxes which are indicated with dash clockwise arrows and added together in the core. Thus, two produced fluxes are equivalent to two parallel inductors which suppress two CM currents on the live and neutral wires, respectively. The resultant CM inductance of two equivalent parallel inductors is L_{CM} and the live and neutral wires are equivalent to one wire under the CM condition.

In Fig. 5-1 (b), the high frequency DM and 50 Hz/60 Hz AC (I_{AC}) currents produce magnetic fluxes. The fluxes which are indicated with dash counter clockwise and clockwise arrows and canceled in the core. Due to the 50 Hz/60 Hz AC current whose magnitude is much larger than the high frequency DM noise current. Because of the flux cancellation, the AC current does not saturate the core. The high frequency DM currents generate leakage fluxes, which are indicated with dash straight arrows. The leakage fluxes cause incomplete cancelation of magnetic fluxes, which lead to finite DM inductance that suppresses the high frequency DM currents in the live and neutral wires. For the DM condition, the two DM inductors are in series and the resultant leakage inductance (L_{leak}) of two series inductors is $L_{leak,L}+L_{leak,N}$, which is usually much smaller than L_{CM} since the magnitude of the leakage fluxes is much smaller than that of the fluxes due to the CM currents.

Since the CM choke has four terminals and two different inductances, the measurements of L_{CM} and L_{leak} with impedance analyzer are different from the measurement of the inductance of an ordinary two-terminal inductor and are illustrated in Fig. 5-2.

(a)

(b)

Fig. 5-2 Measurement setups to obtain (a) L_{CM} (b) L_{leak} .

In Fig. 5-2, T1, T2, T3 and T4 represent the four terminals of the CM choke. Fig. 5-2 (a) shows the measurement of L_{CM} with the impedance analyzer. T1 and T3 are shorted and connect to impedance analyzer's port 1, and T2 and T4 are shorted and connect to impedance analyzer's port 2 [93].

Fig. 5-2 (b) shows that L_{leak} can be measured when T1 and T2 connect together and T3 and T4 connect to impedance analyzer's ports 1 and 2, respectively [93].

For a DM choke, its working principle is the same as the ordinary two-terminal inductor. It consists of an iron powder magnetic core and one set of wire winding wound around the core. Unlike the CM choke, the magnetic fluxes canceled within the

core, the DM choke carries very high fluxes produced by the AC current and thus the core of the DM choke is required to handle the strong magnetic fluxes without saturation [34]. The equivalent circuit of the DM choke is illustrated in Fig. 5-3.

Fig. 5-3 Configuration and simplified equivalent circuit of DM choke as EMI filter.

As shown in Fig. 5-3, two identical DM chokes with each inductance of L_{DM} are needed to be placed on the live and neutral wires respectively. When CM currents pass two DM chokes, two DM chokes are equivalent to two inductors in parallel and thus the resultant inductance of two parallel chokes for CM current suppression is $L_{DM}/2$. When DM currents pass two DM chokes, two DM chokes are equivalent to two inductors in series and thus the resultant inductance of two series chokes for DM current suppression is $2L_{DM}$. The DM choke does not present high impedance to the AC current since the frequency of the AC current is too low. In addition, the DM chokes are not always needed for EMI filter if the leakage inductance of the CM choke is large enough to double up as DM choke.

5.1.2 Line-to-Line (X) and Line-to-Ground (Y) Capacitors

Line-to-line capacitor, or commonly known as X capacitor, which is placed across the live and neutral wires, is for DM current suppression. Line-to-ground capacitor, also known as Y capacitor, is for CM current suppression, where two Y capacitors are needed and placed across the live and ground wires and across the neutral and ground wires, respectively. The capacitance (C_X) of the X capacitor is usually large and in μF range. On the contrary, the capacitance (C_Y) of the Y capacitor has to be small to prevent excessive earth leakage for electrical safety consideration, and C_Y is normally limited to several thousand pF. The equivalent circuit of X and Y capacitors as the EMI filter is illustrated in Fig. 5-4.

Fig. 5-4 Equivalent circuit of X and Y capacitors as EMI filter.

In Fig. 5-4, only the two Y capacitors are effective for CM current suppression and all three capacitors are effective for DM current suppression. For CM current suppression, two Y capacitors are treated as parallel and thus the resultant capacitance of two parallel Y capacitors is $2C_Y$. For DM current suppression, the X capacitor is in parallel with two series Y capacitors and thus the resultant capacitance of three capacitors is $C_X + C_Y/2$.

5.2 EMI Filter Design

To further demonstrate the impact of the conducted emission measurement without LISN under the condition where an EMI filter is installed on the SMPS, an EMI filter designed according to the measurement with LISN is described in this section.

There are two approaches in EMI filter design [12, 13]. The first approach [12] is relatively simple but ignores the consideration of filtering configurations and impedances of the SMPS and LISN. By ignoring the impedances of the SMPS and LISN, the filter designed is often over-designed, which was reported in [13]. On the contrary, the second approach [13] provides a systematic EMI filter design procedure, which takes the impedances and filter configurations into account to avoid over-designing. Thus, the later approach is employed to design the EMI filter according to the measurement with LISN for further demonstrating the impact on the EMI filter performance without LISN.

The very first step of designing an EMI filter is to determine the required CM and DM attenuations, which can be determined by subtracting the measured CM and DM emissions from one of the conducted EMI limits described in Fig. 4-10. However, the conducted EMI limits, which are defined with the conducted emission measured with LISN and in unit of $\text{dB}\mu\text{V}$, do not apply to the subsequent measured emissions for demonstration, which is measured in terms of current in unit of $\text{dB}\mu\text{A}$, as the measurement of conducted emission includes emissions measured without LISN. In order to determine the required attenuations for this case, the CISPR 22 class A quasi-peak limit is converted to the current limit in unit of $\text{dB}\mu\text{A}$ by using the limit, which is

in terms of $\text{dB}\mu\text{V}$, to subtract $34 \text{ dB}\Omega$, which is the logarithmic expression of 50Ω . The newly defined limit and the CM and DM emissions measured with LISN are shown in Fig. 5-5.

Fig. 5-5 CM and DM emissions measured with LISN and limit line.

In Fig. 5-5, $I_{CM,LISN}$ and $I_{DM,LISN}$ denote CM and DM conducted emissions (in $\text{dB}\mu\text{A}$) measured with LISN, respectively. $I_{CM,LISN}$ is generally below the limit line from 300 kHz to 30 MHz and it can be further attenuated a few dB below 1 MHz. $I_{DM,LISN}$ shows that it needs attenuation below 1.5 MHz and it needs the maximum attenuation of 15 dB at 300 kHz. Thus, the required CM and DM attenuations are determined.

After determining the required CM and DM attenuations, the filter configuration has to be determined. Since the SMPS impedance is larger than the LISN impedance at CM and DM over the frequency range of measurement, the filter configurations for CM and DM are determined as the CL-filter [82, 94-96], where the shunt capacitor and series inductor of the filter are placed in parallel at the high impedance side and in series with the low impedance side, respectively. Thus, an X capacitor, two Y capacitors and a CM

choke are chosen to form the CL-filter. The DM choke is only needed when the required DM attenuation cannot be achieved. The X and Y capacitors are placed in parallel with the SMPS, and the CM choke is placed in series with the LISN. The equivalent circuits of the determined CL-filter for CM and DM are shown in Fig. 5-6.

Fig. 5-6 Equivalent circuits of determined CL-filter for (a) CM (b) DM.

As shown in Fig. 5-6 (a) and (b), L_{CM} and two parallel C_Y form the CM filter, whereas $L_{leak,L}$, $L_{leak,N}$, C_X and two series C_Y form the DM filter.

Based on the above CM and DM configurations in Fig. 5-6 (a) and Fig. 5-6 (b), respectively, the estimated CM ($A_{CM,est}$) and DM ($A_{DM,est}$) attenuations of the filter can be calculated by [97-99]:

$$A_{CM,est} = 20 \log \left| s^2 \left(\frac{L_{CM} 2C_Y Z_{SMPS,CM}}{Z_{LISN,CM} + Z_{SMPS,CM}} \right) + s \left(\frac{L_{CM} + 2C_Y Z_{SMPS,CM} Z_{LISN,CM}}{Z_{LISN,CM} + Z_{SMPS,CM}} \right) + 1 \right| \quad (5-1)$$

$$A_{DM,est} = 20 \log \left| s^2 \left(\frac{L_{leak}(C_X + \frac{C_Y}{2})Z_{SMPS,DM}}{Z_{LISN,DM} + Z_{SMPS,DM}} \right) + s \left(\frac{L_{leak} + (C_X + \frac{C_Y}{2})Z_{SMPS,DM}Z_{LISN,DM}}{Z_{LISN,DM} + Z_{SMPS,DM}} \right) + 1 \right| \quad (5-2)$$

where $s=j\omega$. The filter components in (5-1) and (5-2) are assumed to be ideal, so L_{leak} , L_{CM} , C_X , and C_Y are treated as frequency invariant from 300 kHz to 30 MHz. From (5-1) and (5-2), L_{leak} , L_{CM} , C_X , and C_Y can be estimated.

Since L_{leak} depends on the chosen CM choke, L_{CM} and C_Y are determined first before L_{leak} and C_X . For CM attenuation according to (5-1), L_{CM} and C_Y have to be chosen such that $A_{CM,est}$ meets the required CM attenuation. Since the required CM attenuation is rather small, $A_{CM,est}$ is should be kept sufficient enough to avoid over-designing. By choosing C_Y to be 1000 pF, L_{CM} is determined by substituting C_Y , $Z_{SMPS,CM}$, and $Z_{LISN,CM}$ into (5-1) once $A_{CM,est}$ is known. Thus, L_{CM} is determined to be 0.8 mH. By choosing a CM choke with L_{CM} of 0.8 mH, L_{leak} is measured and found to be 3.7 μ H using the setup shown in Fig. 5-2 (b). For DM attenuation, L_{leak} , C_Y , and C_X have to be selected such that $A_{DM,est}$ meets the required DM attenuation, which is 16 dB at 300 kHz. Since L_{leak} and C_Y are already determined, C_X is determined by substituting L_{leak} , C_Y , $Z_{SMPS,DM}$, and $Z_{LISN,DM}$ into (5-2) it is found that C_X is 0.6 μ F. The inductors and capacitors of the above equations determining $A_{CM,est}$ and $A_{DM,est}$ are assumed to be ideal inductors and capacitors that meet the required CM and DM attenuations, respectively; and $A_{CM,est}$ and $A_{DM,est}$ are plotted in Fig. 5-7 from 300 kHz to 30 MHz. The actual impedances of L_{leak} (Z_{Lleak}), L_{CM} (Z_{LCM}), C_X (Z_{CX}), and C_Y (Z_{CY}) are measured from 300 kHz to 30 MHz and shown in Fig. 5-8. With the determined filter components and configuration, the EMI filter designed according to the measurement with LISN is formed.

Fig. 5-7 $A_{CM,est}$ and $A_{DM,est}$ determined by chosen ideal inductors and capacitors.

Fig. 5-8 Actual impedances of L_{leak} (Z_{leak}), L_{CM} (Z_{CM}), C_X (Z_X), and C_Y (Z_Y).

5.3 Deviation Validation and Analysis of Conducted Emissions

In this section, the CM and DM conducted emissions are measured using the setups shown in Fig. 4-17 and Fig. 4-24, respectively; firstly with the SMPS powered through LISN and then with the SMPS powered directly from the mains (one of the 10

locations mentioned in Chapter 4 is selected for the study). The measured CM and DM emissions with the previously designed filter are compared with those measured without filter and plotted in Fig. 5-9 and Fig. 5-10 from 300 kHz to 30 MHz, respectively.

Fig. 5-9 Measured CM conducted emissions under four operating conditions.

In Fig. 5-9, $I_{CM,Mains1}$, $I_{CM,LISN,Filter}$ and $I_{CM,Mains1,Filter}$ denote measured CM conducted emissions (in dBμA) under three operating conditions: mains without filter, LISN with filter and mains with filter, respectively. $I_{CM,LISN}$ is attenuated as expected. Both $I_{CM,LISN}$ and $I_{CM,Mains1}$ are rather similar, which is consistent with the statistical analysis of the CM deviation given in Fig. 4-19. After the filter is installed on the SMPS, $I_{CM,LISN,Filter}$ and $I_{CM,Mains1,Filter}$ are rather similar because $Z_{SMPS,CM}$ is much higher than the total resultant impedance of $Z_{Mains,CM(1)}$, $Z_{LISN,CM}$ and $Z_{W,CM}$ below 15 MHz. Therefore, with the filter added, the change in CM conducted emission is expected to be insignificant. The CM conducted emission with the SMPS powered directly from mains does not

deviate much from that of powered through the LISN. Hence, the impact of CM filter attenuation without the LISN is rather small.

Fig. 5-10 Measured DM conducted emissions under four operating conditions.

In Fig. 5-10, $I_{DM,Mains1}$, $I_{DM,LISN,Filter}$ and $I_{DM,Mains1,Filter}$ denote measured DM conducted emissions (in dBμA) under the three similar conditions as mentioned earlier. $I_{DM,LISN}$ is attenuated as expected. Fig. 5-10 shows that $I_{DM,Mains1}$ can be up to 7 dB higher than $I_{DM,LISN}$ below 6 MHz, which is also consistent with the statistical analysis of the DM deviation shown in Fig. 4-26. $I_{DM,LISN,Filter}$ and $I_{DM,Mains1,Filter}$ show that with LISN, DM conducted emission is well suppressed by the filter below 10 MHz with the maximum attenuation of 15 dB at 300 kHz. However, without LISN, DM conducted emission is only suppressed by the filter below 3 MHz with negligible attenuation at 300 kHz, which reveals that the DM attenuation of the filter designed according to the measurement with LISN may not provide adequate attenuation when the SPMS is powered by the mains directly. The reason of such a large variation between

Fig. 5-12 Measured neutral-to-ground conducted emissions under four operating conditions.

Fig. 5-13 Measured live-to-ground conducted emissions under four operating conditions.

In Fig. 5-12, $I_{N,LISN}$, $I_{N,Mains1}$, $I_{N,LISN,Filter}$ and $I_{N,Mains1,Filter}$ denote the measured neutral-to-ground conducted emissions (in dBμA) under the four conditions mentioned earlier.

In Fig. 5-13, $I_{L,LISN}$, $I_{L,Mains1}$, $I_{L,LISN,Filter}$ and $I_{L,Mains1,Filter}$ denote the measured live-to-ground conducted emissions (in dB μ A) under the four conditions mentioned earlier.

Both Fig. 5-12 and Fig. 5-13 demonstrate that the actual conducted emissions without LISN are much higher emissions due to much higher DM emission without LISN. In addition, both the neutral and live emissions with LISN and filter are lower than the limit due to the effective attenuation on CM and DM emissions with LISN. Hence, the filter designed according to the measurement with LISN may not reflect the true performance of the filter when the SMPS is powered directly from the mains.

The above study of conducted emissions has demonstrated that the actual DM conducted emissions without LISN tend to be higher than that of with LISN and therefore the designed filter based on measurement with LISN may not provide sufficient attenuation under the actual operating condition without LISN. The DM conducted emission deviation between measurements with and without LISN can result in an EMI filter designed based on measurement with LISN to be ineffective in reality. Hence, for a high power converter system with a specific installation location, knowing both the CM and DM impedances of the AC mains and the system allows an EMI filter to be tailored according to the mains impedances of the specific location to ensure EMC.

Chapter 6

Conclusion and Future Work

6.1 Conclusion

The significance and the practical application aspect of the in-circuit impedance information have been discussed. A comprehensive review of the existing in-circuit impedance measurement approaches has also been covered. This thesis has proposed the concept of using the $ABCD$ two-port network characterization technique and analysis to improve the shortcomings of a two-probe inductive coupling approach for in-circuit impedance measurement.

The theory behind the proposed $ABCD$ two-port network method is first described and then validated experimentally. With the injecting and receiving current probes carefully characterized, the proposed method enhances the accuracy of the two-probe approach with a maximum measurement deviation of +3.6% up to 100 MHz. As compared with the conventional characterization and calibration method of the two-probe approach, the proposed method has demonstrated the following merits:

- There is no need to cut off the power supply to the DUT for pre-measurement calibration; which may not be possible for critical electrical system.

- There is no need to replace the DUT with a known standard resistor or short circuit for pre-measurement calibration, which may not be practical for on-site high power system.
- Once the current probes are characterized and calibrated, they are re-usable for different DUT for in-circuit impedance measurement.
- Better measurement accuracy.

The proposed method is found to be useful for condition monitoring of electrical system without the need of shutting down the power when installing the setup. Using an operating DC-powered motor and an operating AC-powered transformer as two examples, the proposed method has successfully detected the impedance variation under two operating conditions.

Also, the proposed method is applied to extract the noise impedances for analyzing the impact of the conducted emission measurement without LISN. Through extracting the CM and DM impedances of the SMPS, LISN, wiring connection, and AC mains at 10 locations with the proposed approach, the CM and DM deviations of the conducted emission measurement without LISN from that of with LISN are analyzed based on the extracted impedances and evaluated statistically with a sample size of 10 mains and confidence level of 99.7%. Subsequently, the worst deviation due to the 10 mains is derived and an insight for designing the EMI filter is given. Finally, the statistically evaluated CM and DM deviations are validated with measurements and the impact of the conducted emission measurement without LISN is further demonstrated by comparing the attenuation performances of the designed EMI filter in conducted emission measurements with and without LISN.

6.2 Future Work

The proposed approach can be further extended to applications for condition monitoring switchgear and cable under their usual operating condition.

Detecting the potential fault of the high-voltage cable is not enough for most practical cases. Knowing the fault location of a long high-voltage cable without disconnecting the cable from the power network is highly desirable so that the fault can be localized and rectified. The high-voltage cable employed in infrastructure for power delivery is mission-critical and not easily accessible for frequent inspection. Thus, with the same measurement instruments as the proposed approach, the algorithm of the proposed approach can be modified to include the transmission line theory for fault localization of the cable.

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K. Li, K. Y. See, and X. M. Li, "Inductive Coupled In-Circuit Impedance Monitoring of Electrical System Using Two-Port ABCD Network Approach," *IEEE Transactions on Instrumentation and Measurement*, vol. 64, no. 9, pp. 2489-2495, 2015.

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