Accurate Extraction of Noise Source Impedance of an SMPS Under Operating Conditions

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***Abstract*—An accurate measurement method to extract the common mode (CM) and the differential mode (DM) noise source impedances of a switched-mode power supply (SMPS) under its operating condition is developed and validated. With a proper premeasurement calibration process, the proposed method allows extraction of both the CM and the DM noise source impedances with very good accuracy. These noise source impedances come in handy to design an electromagnetic interference filter for an SMPS systematically with minimum hassle.**

***Index Terms*—Common mode (CM), differential mode (DM), electromagnetic interference (EMI), EMI filter, noise source impedance, switched-mode power supply (SMPS).**

I. Introduction

BUILT-IN power line electromagnetic interference (EMI) filters are parts of the switched-mode power supply (SMPS) designs to limit conducted EMI in the frequency range up to 30 MHz in order to comply with the international EMI regulatory requirements [1]–[4]. Unlike the filters used in communications and microwave applications, where the source and the termination impedances are well-defined as 50 Ω, the actual noise source and the termination impedances of an EMI filter in an SMPS are far from 50 Ω [5]. In the standard conducted EMI measurement setup, the noise termination of an SMPS is the line impedance stabilization network (LISN) and its impedance characteristic are well-defined [6]. Unfortunately, the noise source impedance of an SMPS varies with several parameters such as converter topology, power rating, component parasitic elements, and board layout [7]. For example, the differential mode (DM) noise source impedance is strongly influenced by the reverse recovery phenomena of the diode rectifier [8], the equivalent series resistance (ESR), and the equivalent series inductance (ESL) of the bulk capacitor [9]. As for the common mode (CM) noise source impedance, the deciding components are the parasitic capacitance between the switching device and its heat sink and the parasitic capacitance between the board and the chassis [1]–[3]. Hence, designing an EMI filter for an SMPS by assuming 50 Ω noise source and termination impedances will lead to nonoptimal EMI suppression performance from the filter. Therefore, the need of information of the noise source impedance of an SMPS is apparent [10]–[12].

Some progress has been made to measure the DM and the CM noise source impedances of an SMPS. First, the resonance method was developed to estimate the noise source impedance of an SMPS by making a simplifying assumption that the noise source is a simple Norton equivalent circuit of a current source with parallel resistive and capacitive elements [13]. By terminating at the ac power input of the SMPS with a resonating inductor, the noise source impedance can be estimated [2]. However, the process to select and tune the resonating inductor for resonance can be tedious and cumbersome. Also, when frequency increases, the parasitic effects of the nonideal reactive components become significant and the circuit topology based on which the resonance method is developed is no longer valid. This simplistic approach provides only a very rough estimate of the noise source equivalent circuit model. Later, the insertion loss method was introduced to measure the DM and the CM noise source impedances of an SMPS. This method requires some prior conditions to be fulfilled. For example, the impedances of the inserted components must be much larger or smaller than the noise source impedances [14], [15]. Hence, the accuracy deteriorates if these conditions are not met. Moreover, it only provides the magnitude information of the noise source impedance and the phase information can only be estimated with a complicated Hilbert transform process. Recently, a two-probe approach to measure the DM and the CM noise source impedances of an SMPS was developed [16]. An injecting probe, a sensing probe, and some coupling capacitors are used in the measurement setup. In order to measure the DM and the CM noise source impedances with reasonable accuracy, careful choices of the DM and the CM chokes are necessary to provide very good RF isolation between the SMPS and the LISN. Moreover, special attention is needed to ascertain that the DM and the CM chokes are not saturated for the SMPS of higher power rating. Again, this method focuses on extracting the magnitude information of the noise source impedance only.

In view of the limitations of the previously discussed methods, a direct clamping two-probe approach is proposed in this paper. Unlike the former two-probe method [16], the proposed method uses direct clamp-on type current probes, and therefore, there is no direct electrical contact to the power line wires between the LISN and the SMPS. Hence, it eliminates the needs of the coupling capacitors. Also, no isolating chokes are needed, making the measurement setup very simple to implement. With the vector network analyzer (VNA) as a measurement instrument, both

Manuscript received January 11, 2009; revised April 17, 2009. Current version published January 29, 2010. Recommended for publication by Associate Editor M. Vitelli.

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Digital Object Identifier 10.1109/TPEL.2009.2024675

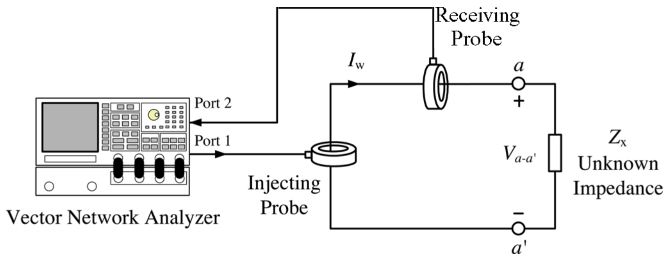

Fig. 1. Basic setup of the direct clamping two-probe measurement.

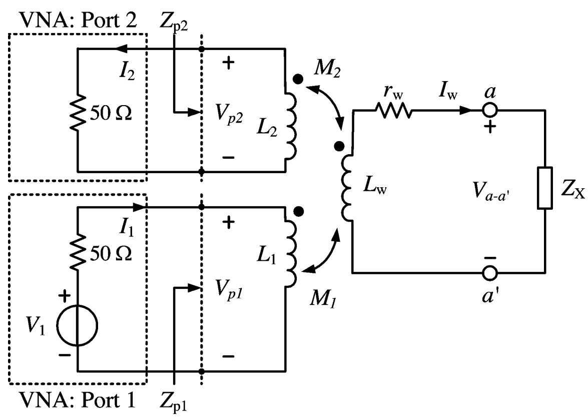

Fig. 2. Equivalent circuit of the two-probe measurement setup.

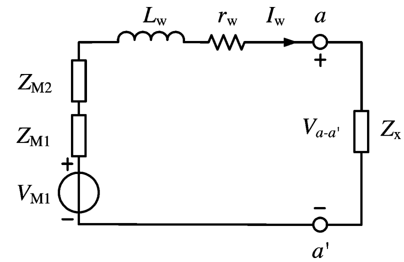

Fig. 3. Final equivalent circuit of the circuit loop connecting to the unknown impedance.

the magnitude and the phase information can be extracted directly without further processing. The proposed method is also highly accurate as it has the capability to eliminate the error introduced by the measurement setup.

This paper is organized as follows. Section II provides the necessary background theory of the direct clamping two-probe measurement technique. Experimental validation of the proposed method is given in Section III. Section IV describes the setups to measure the DM and the CM noise source impedances of the SMPS in its actual powered-up operating conditions. Finally, the conclusions are given in Section V.

II. Background Theory of the Two-Probe Measurement

The basic setup of the two-probe method to measure any unknown impedance is illustrated in Fig. 1. It consists of an injecting current probe, a receiving current probe, and a VNA. The unknown impedance to be measured (Z_x) is connected at $a - a'$. Port 1 of the VNA generates an ac signal into the closed loop through the injecting probe, and the resulting signal current in the loop is measure at port 2 of the VNA through the receiving probe.

Fig. 2 shows the complete equivalent circuit of the measurement setup shown in Fig. 1. V_1 is the signal source voltage of port 1 connected to the injecting probe and V_{p2} is the resultant signal voltage measured at port 2 with the receiving probe. The output impedance of port 1 and the input impedance of port 2 of the VNA are both 50 Ω. L_1 and L_2 are the primary inductances of the injecting and the receiving probes, respectively. L_w and r_w are the inductance and the resistance of the wiring connection that formed the circuit loop, respectively. M_1 is the mutual inductance between the injecting probe and the circuit loop and M_2 is the mutual inductance between the receiving probe and the circuit loop. Z_{p1} and Z_{p2} are the input impedances of the injecting and the receiving probes, respectively.

With V_1 as the exciting signal source, it induces a signal current I_w in the circuit loop through the injecting probe. From Fig. 2, three circuit equations are resulted as follows:

$$\begin{bmatrix} V_1 \\ 0 \\ -V_{a-a'} \end{bmatrix} = \begin{bmatrix} 50\,\Omega + Z_{p1} & 0 & -j\omega M_1 \\ 0 & 50\,\Omega + Z_{p2} & +j\omega M_2 \\ -j\omega M_1 & +j\omega M_2 & r_w + j\omega L_w \end{bmatrix} \times \begin{bmatrix} I_1 \\ I_2 \\ I_w \end{bmatrix}. \tag{1}$$

Eliminating I_1 and I_2 from (1) gives

$$V_{M1} = V_{a-a'} + (Z_{M1} + Z_{M2} + r_w + j\omega L_w) I_w \tag{2}$$

where $Z_{M1} = (\omega M_1)^2/(50\,\Omega + Z_{p1})$, $Z_{M2} = (\omega M_2)^2/(50\,\Omega + Z_{p2})$, and $V_{M1} = V_1\,(j\omega M_1/(50\,\Omega + Z_{p1}))$.

According to the expression in (2), the injecting probe can be reflected in the closed circuit loop as an equivalent voltage source V_{M1} in series with a reflected impedance Z_{M1}, and the receiving probe can be reflected in the same loop as another impedance Z_{M2}, as shown in Fig. 3. For frequencies below 30 MHz, the dimension of the coupling circuit loop is electrically small as compared to the wavelengths concerned. Therefore, the current distribution in the coupling circuit is uniform throughout the loop, and V_{M1} can be rewritten as

$$\begin{aligned} V_{M1} &= (Z_{M1} + Z_{M2} + r_w + j\omega L_w + Z_x) I_w \\ &= (Z_{\text{setup}} + Z_x) I_w. \end{aligned} \tag{3}$$

The equivalent circuit seen at $a - a'$ by the unknown impedance Z_x can be substituted by a voltage source V_{M1} in series with an impedance due to the measurement setup Z_{setup}. From (3), Z_x can be determined by

$$Z_x = \frac{V_{M1}}{I_w} - Z_{\text{setup}}. \tag{4}$$

The current I_w measured by the receiving probe is

$$I_w = \frac{V_{p2}}{Z_{T2}} \tag{5}$$

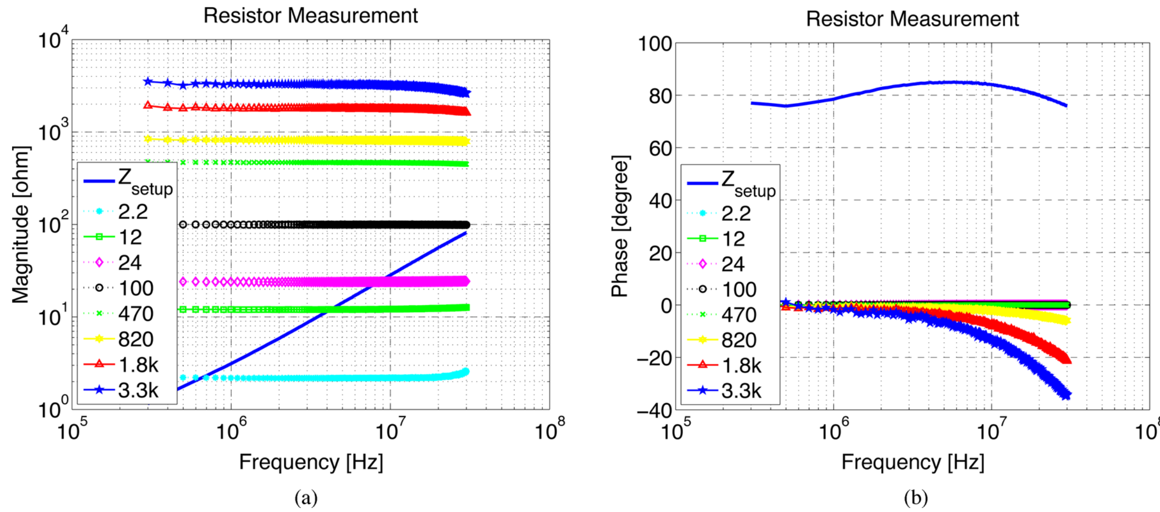

Fig. 4. Measured results of selected resistors using the proposed two-probe approach. (a) Magnitude. (b) Phase.

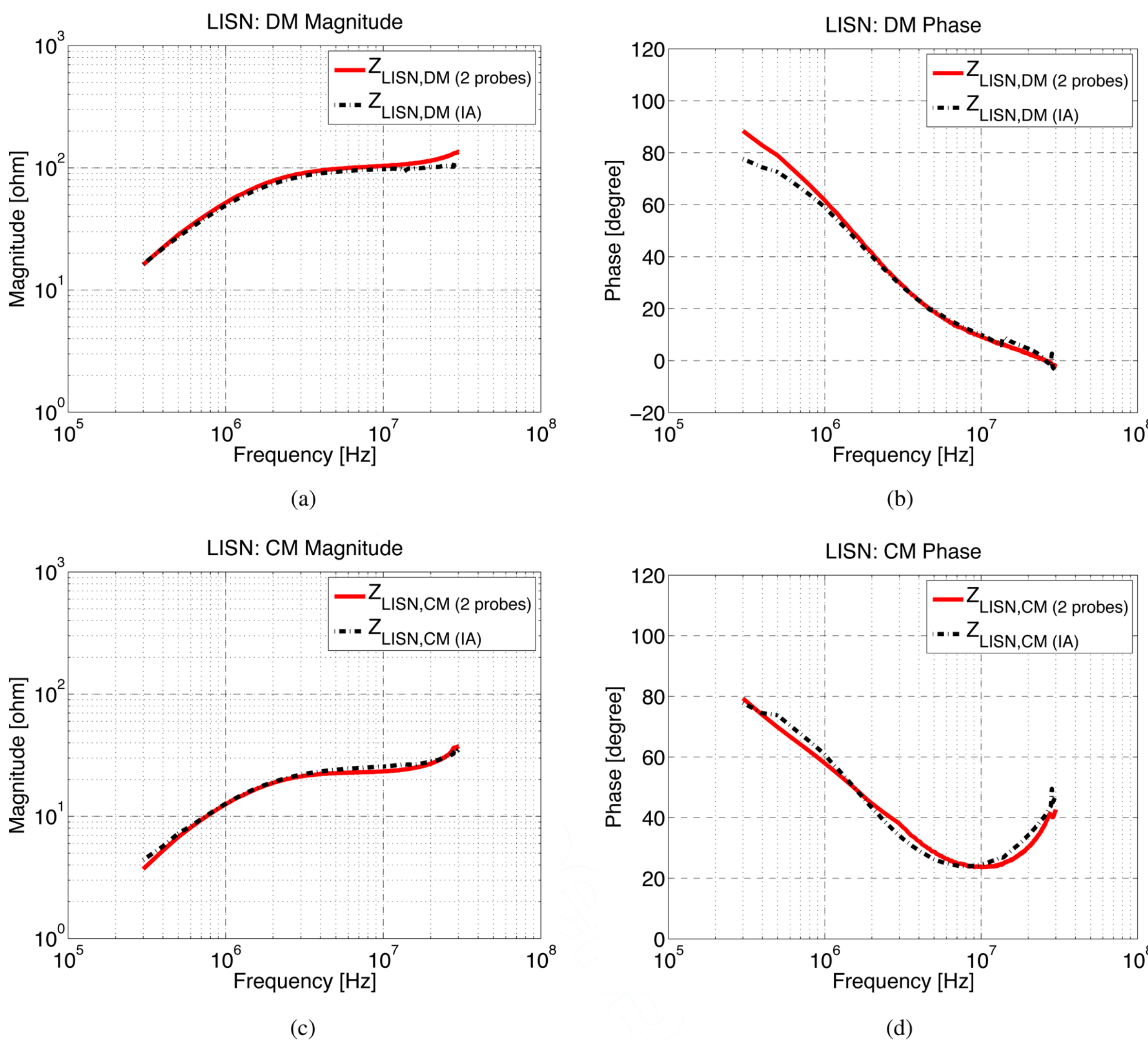

Fig. 5. Comparison of measured results for LISN. (a) DM magnitude. (b) DM phase. (c) CM magnitude. (d) CM phase.

where V_{p2} is the signal voltage measured at port 2 of the VNA and Z_{T2} is the calibrated transfer impedance of the receiving probe provided by the probe manufacturer. Substituting V_{M1} and (5) into (4) yields

$$Z_x = \left(\frac{j\omega M_1 Z_{T2}}{50\,\Omega + Z_{p1}}\right)\left(\frac{V_1}{V_{p2}}\right) - Z_{\text{setup}}. \tag{6}$$

The excitation source V_1 of port 1 of the VNA and the resultant voltage at the injecting probe V_{p1} is related by

$$V_1 = \left(\frac{50\,\Omega + Z_{p1}}{Z_{p1}}\right) V_{p1}. \tag{7}$$

Substituting (7) into (6), the unknown impedance can finally be expressed as

$$Z_x = K\left(\frac{V_{p1}}{V_{p2}}\right) - Z_{\text{setup}} \tag{8}$$

where $K = (j\omega M_1 Z_{T2}/Z_{p1})$, which is a frequency-dependent coefficient. The ratio V_{p1}/V_{p2} can be obtained through the S-parameters measurement using the VNA. Since $V_{p1} = (S_{11}+1)V_1$ and $V_{p2} = S_{21}V_1$, the ratio of the two probe voltages is given by

$$\frac{V_{p1}}{V_{p2}} = \frac{S_{11}+1}{S_{21}}. \tag{9}$$

The coefficient K and the setup impedance Z_{setup} can be obtained by the following steps. First, measure V_{p1}/V_{p2} by replacing impedance Z_x with a known precision standard resistor R_{std}. Then, measure V_{p1}/V_{p2} again by short-circuiting $a - a'$. With these two measurements and (8), two equations (10) and (11) with two unknowns K and Z_{setup} are obtained. Hence, K and Z_{setup} can be obtained by solving (10) and (11). Once K and Z_{setup} are found, the two-probe setup is ready to measure any unknown impedance using (8)

$$Z_x|_{Z_x=R_{\text{std}}} = K\left(\frac{V_{p1}}{V_{p2}}\right)|_{Z_x=R_{\text{std}}} - Z_{\text{setup}} \tag{10}$$

$$Z_x|_{Z_x=\text{short}} = K\left(\frac{V_{p1}}{V_{p2}}\right)|_{Z_x=\text{short}} - Z_{\text{setup}}. \tag{11}$$

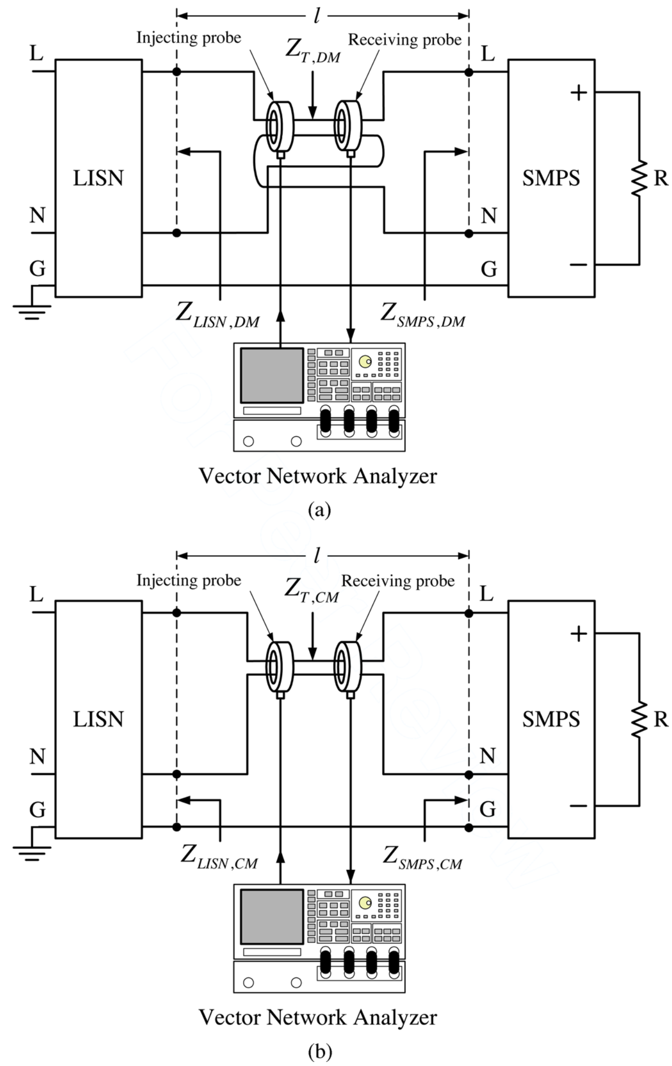

Fig. 6. Noise source impedance measurement setup of the SMPS. (a) DM. (b) CM.

III. Experimental Validation

In the experimental validation that follows, the Solar 9144-1N current probe (10 kHz–100 MHz) and the Schaffner CPS-8455 current probe (10 kHz–1000 MHz) are chosen as the injecting and the receiving current probes, respectively. The R&S ZVB8 VNA (300 kHz–8 GHz) is selected for the S-parameter measurement.

Practically, the DM noise source impedance of SMPS ranges from several ohms to several tens of ohms, and CM noise source impedance is capacitive in nature and is in the range of several kilohms [1], [2], [12]. In the validation, a precision resistor R_{std} (620 $\Omega \pm 1\%$) is chosen, as it is somewhere in the middle of the range of unknown impedance to be measured (tens of ohms to a few kilohms). Based on the procedure described in Section II, K and Z_{setup} are determined accordingly. Once K and Z_{setup} are found, a few resistors of known values (2.2 Ω, 12 Ω, 24 Ω, 100 Ω, 470 Ω, 820 Ω, 1.8 kΩ, and 3.3 kΩ) are treated as unknown impedances and measured by the direct clamping two-probe setup. The wire loop of the resistor under measurement is made as small as possible to avoid any loop resonance below 30 MHz. Even by making the loop very small, there is still a finite impedance due to the measurement setup (Z_{setup}). Z_{setup} comprises of the effects of the injecting and the receiving probes, the wire connection to the resistor, and the coaxial cable between the current probes and the VNA. The ability to measure Z_{setup} and to subtract it from the two-probe measurement eliminates the error due to the setup and provides highly accurate measurement results. The measurement frequency range is from 300 kHz to 30 MHz. As shown in Fig. 4(a) and (b), the magnitude and the phase of the so-called unknown resistors are measured by the proposed method. The measured results are in close agreement with the stated resistance values of the resistors.

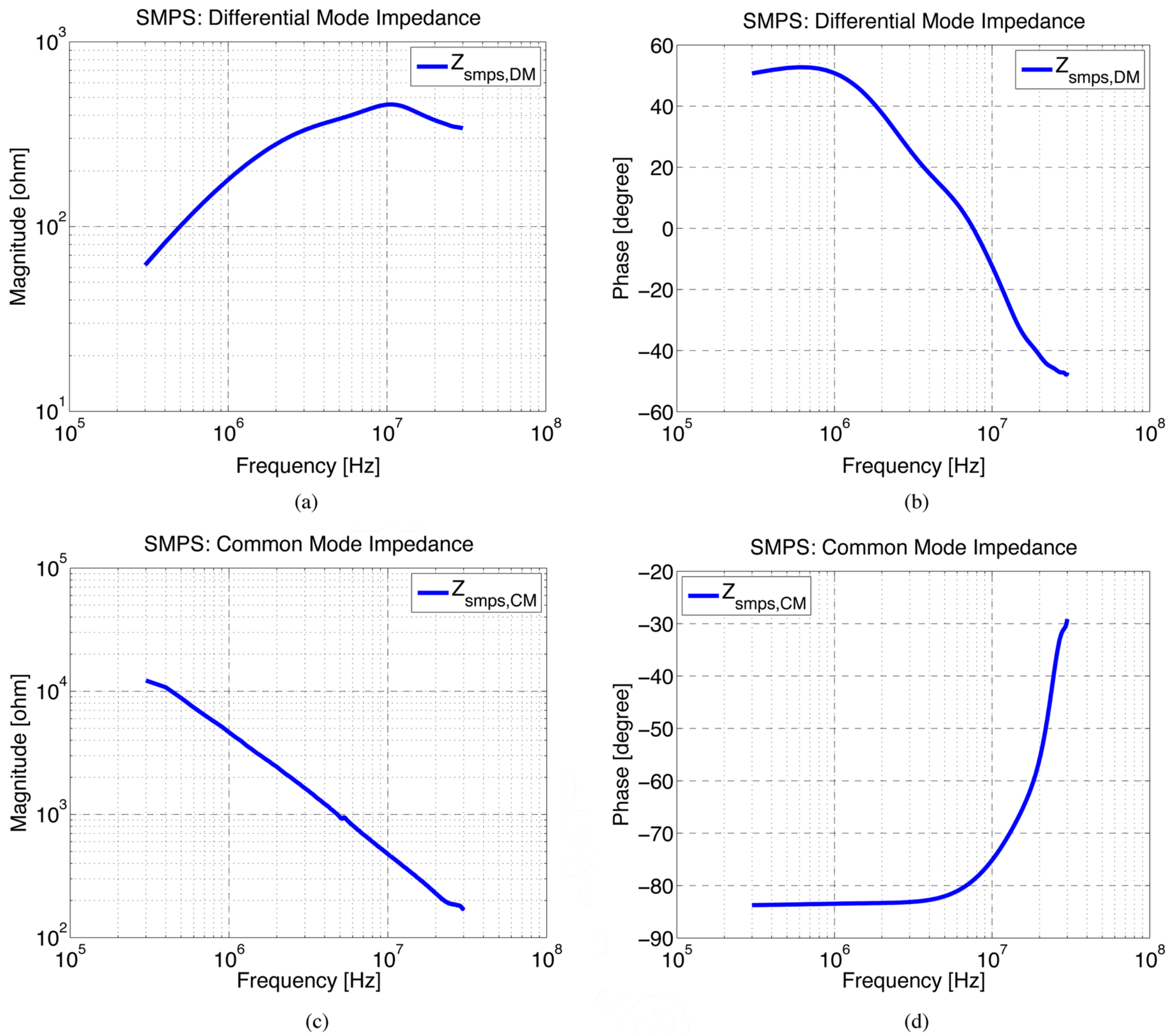

Fig. 7. Noise source impedance measurement. (a) DM magnitude. (b) DM phase. (c) CM magnitude. (d) CM phase.

For large-resistance resistors such as 1.8 and 3.3 kΩ, the roll-off at higher frequency is expected due to the parasitic capacitance that is inherent to large-resistance resistors, but the parasitic effect is negligible for small-resistance resistors. In Fig. 4, Z_{setup} is also plotted to show its relative magnitude and phase with respect to the measured resistances. It shows that Z_{setup} is predominantly inductive and can be as high as 100 Ω at 30 MHz. Hence, for small-resistance resistors, where their values are comparable to Z_{setup}, the error contributed from Z_{setup} can be very large, if it is not subtracted from the measurement.

For further validation purposes, the DM as well as the CM output impedances of an LISN (Electro-Metrics MIL 5-25/2) are measured using the proposed method and the HP4396B impedance analyzer (100 kHz–1.8 GHz). The measured DM impedance $(Z_{\mathrm{LISN,DM}})$ and the measured CM impedance $(Z_{\mathrm{LISN,CM}})$ of the LISN using both methods are compared. By using the two-probe method, the LISN can be measured with the ac power applied. However, the measurement using the impedance analyzer can only be made with no ac power applied to the LISN to prevent damage to the measuring equipment. For the two-probe method, ac power is applied to the input of the LISN and one or two 1-μF "X class" capacitors are connected at the output of the LISN to implement an ac short circuit. A 1-μF capacitor is connected between line and neutral for DM measurement. For CM measurement, two 1-μF capacitors are needed: one connected between line and ground and another connected between neutral and ground. For the DM output impedance measurement, the line wire is treated as one single outgoing conductor and the neutral wire is treated as the returning conductor. In the case of CM measurement, the line and the neutral wires are treated as one single outgoing conductor, and the safety ground wire is treated as the returning conductor. The length of the connecting wire between the LISN and capacitor is chosen to be as short as possible to eliminate the parasitic inductance of the connecting wires. The comparisons of the measured results of the output impedance of LISN using the direct clamping two-probe approach $(Z_{\mathrm{LISN,DM(2probes)}}$ and $Z_{\mathrm{LISN,CM(2probes)}})$ and using the impedance analyzer $(Z_{\mathrm{LISN,DM(IA)}}$ and $Z_{\mathrm{LISN,CM(IA)}})$ are given in Fig. 5. Again, close agreement between the two measurement methods is demonstrated.

IV. Measurement of Noise Source Impedance of the SMPS

The measurement setups to extract the DM noise source impedance $(Z_{\mathrm{SMPS,DM}})$ and the CM noise source impedance $(Z_{\mathrm{SMPS,CM}})$ of an SMPS are shown in Fig. 6(a) and (b), respectively. The model and technical specifications of the SMPS are VTM22WB, 15 W, $+12\ \mathrm{V_{dc}}$/0.75 A, $-12\ \mathrm{V_{dc}}$/0.5 A. The SMPS is powered through the MIL 5-25/2 LISN to ensure stable and repeatable ac mains impedance. A resistive load is connected at the output of the SMPS for loading purposes. The DM impedance $(Z_{\mathrm{LISN,DM}})$ and the CM impedance $(Z_{\mathrm{LISN,CM}})$ of the LISN have been measured earlier in Section III and presented in Fig. 5.

To extract the DM and the CM noise source impedances of the SMPS, first, $Z_{\mathrm{setup,DM}}$, $Z_{\mathrm{setup,CM}}$, and the frequency-dependent coefficient K of the measurement setup are determined. The transmission-line effect of the wire connection can be ignored as the length of connecting wires (l) from the LISN to the SMPS is 70 cm, which is much shorter than the wavelength of the highest frequency of interest (30 MHz). The impedance measured by the direct clamping two-probe method is the total impedance (Z_T) of the circuit loop connecting the SMPS and LISN under actual ac powered-up operating condition. This total impedance is given by

$$Z_T = Z_{\mathrm{LISN}} + Z_{\mathrm{SMPS}} + Z_{\mathrm{setup}} \tag{12}$$

where Z_{LISN} is the impedance of the LISN, Z_{SMPS} is the noise source impedance of the SMPS, and Z_{setup} is the impedance due to the measurement setup.

With known Z_{LISN} and Z_{setup}, once Z_T is measured, the noise source impedance of the SMPS can be evaluated easily using (12).

Fig. 7(a) and (b) shows the magnitude and the phase of the extracted DM noise source impedance $(Z_{\mathrm{SMPS,DM}})$ in the frequency range from 300 kHz to 30 MHz. In general, the DM noise source impedance is dominated by the series resistive and inductive components at low frequencies and above 10 MHz, the effect of the diode junction capacitance of the full-wave rectifier begins to kick in. Fig. 7(c) and (d) shows the magnitude and the phase of the extracted CM noise source impedance $(Z_{\mathrm{SMPS,CM}})$. The CM noise source impedance is dominated by the effect of the heat-sink-to-ground parasitic capacitance.

V. Conclusion

Based on a direct clamping two-probe measurement approach, the DM and the CM noise source impedances of any SMPS under its operating condition can be extracted with good accuracy. As compared to previously reported methods, the measurement setup is simple and it also allows both the magnitude and the phase of the DM and the CM noise source impedances of an SMPS to be extracted with ease. The major feature of the proposed method is its ability to eliminate the error due to the impedance of the measurement setup. With a careful premeasurement process to determine the setup impedance, practically, the measurement result is almost error-free. Further work will be carried out to characterize the complete EMI filter under in-circuit operating condition.

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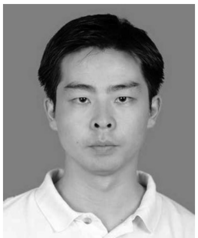

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