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A novel approach to conductive EMI noise source modeling

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Abstract: A new approach to conductive electromagnetic interference (EMI) noise source modeling, i. e. the source internal impedance extraction, is presented. First, the impedance magnitude is achieved through an exciting probe and a detecting probe, or through calculations based on insertion loss measurement results when inserting a series high-value known impedance or a shunt low-value known impedance in the circuit. Then the impedance phase is extracted by the Hilbert transform (HT) of the logarithm of the obtained impedance magnitude. Performance studies show that the estimated phase error can increase greatly at a zero frequency in the Hilbert transform because of the existence of a singular point, and this effect can be eliminated by introducing a zero-point when the noise source does not include a series-connected capacitive component. It is also found that when the frequency is higher than 150 kHz, the estimated phase error is not sensitive to the inductive source but sensitive to the capacitive source. Finally, under the conditions of the same measurement accuracies for impedance magnitude, the accuracy of complex impedance based on the HT can be improved about 10 times when compared with the accuracy of estimated parameters based on the impedance magnitude fitting method (IMFM).

Key words: noise source modeling; impedance estimation; electromagnetic interference

In conductive EMC measurements, conventional testing for compliance with conductive electromagnetic interference (EMI) emission utilizes a line impedance stabilization network (LISN), where the noise currents excited in the power supply consist of two components. Differential-mode (DM) noise current $I_{\mathrm{DM}}(\omega)$ flows out of the live line and returns via the neutral line. Common-mode (CM) noise current $I_{\mathrm{CM}}(\omega)$ flows out of live and neutral lines and returns via the earth wire. Therefore, it can be seen that the DM currents in the live and neutral wires are equal and in the opposite direction; whereas the CM currents are equal and in the same direction. Since conductive EMI noise consists of CM noise and DM noise, an EMI filter also consists of a CM filter and a DM filter.

When the equipment under test (EUT) fails to meet EMC regulations, the EMI filter should be inserted between the EUT (i. e. noise source) and the LISN (i. e. source load) to suppress noise. In order to design an appropriate EMI filter, both the noise source impedance and the source load impedance should be known so as to calculate and determine the exact impedance of the filter. Usually, the source load impedance, which includes the CM impedance and the DM impedance coming from the LSIN, is clearly known; however, the noise source impedance is usually unknown since different types of EUT have different source impedances due to the topology and the operation mechanism. Unfortunately, at present, the determination of the noise source impedance is often made through either empirical estimation or a rough noise modeling method, and the exact source impedance is not easily achieved which results in undesirable EMI filter performance, and then results in undesirable noise suppression effects. Thus noise source modeling is necessary in EMC research[1-6]. In this paper, an approach to conductive EMI noise source modeling is proposed, where the Hilbert transform is employed for impedance phase extraction, and two-current-probe measurement or insertion loss measurement is adopted for impedance magnitude extraction. The presented approach is useful for optimizing EMI filter design applied in EMC.

1 Noise Source Modeling Approach

The purpose of EMI noise source modeling is to accurately know the impedance of a noise source in order to realize the maximum impedance mismatch between an EMI filter and a noise source. Because the source impedance involves both the magnitude and the phase, it can be simply summarized for the presented modeling approach that the magnitude is acquired by measurement and the phase is achieved by the Hilbert transform.

1.1 Impedance magnitude extraction

For source impedance magnitude extraction, in this paper, two methods are employed alternatively. One is based on the current probe measurement and the other on the insertion loss measurement.

1.1.1 Two-current-probe measurement method

In the current probe measurement, two current probes are used for signal exciting and signal detection, respectively, in the measured frequency range from 150 kHz to 30 MHz. For example, when doing CM source impedance measurements by using two current probes, the CM noise currents on live and neutral lines are joined together at the ground line by two coupling capacitors, and also both the signal exciting current probe and the detecting current probe are coupled in the main measurement circuit between the live (neutral) line

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and the ground line. Whereas the DM source impedance measurement is different from the CM impedance measurement, where live and neutral lines are connected through only one coupling capacitor and both the exciting current probe and the detecting current probe are coupled in the main measurement circuit between live and neutral lines.

It has been previously analyzed that, actually, the exciting current probe can be modeled by an equivalent voltage source V_{M1} and a voltage source impedance Z_{M1} [7]. Then in the main measurement circuit, it should involve not only the equivalent voltage source V_{M1} and the source impedance Z_{M1} coming from the exciting current probe, but also the unknown measured impedance Z_X representing the EMI noise source, mutual impedance Z_{M2} due to the detection probe coil, and the impedance Z_C due to the coupling capacitor (see Fig. 1). Let Z_{in} represent the total internal impedance consisting of Z_{M1}, Z_{M2} and Z_C seen from noise source Z_X. Then the main circuit is only organized by the voltage source V_{M1} with the total internal impedance Z_{in} and the external impedance Z_X separated by b-b′. Assuming that the current flow in the main circuit is I_w, this current component should also be the secondary coil current of the exciting current probe since the secondary coil is in series-connection in the main circuit. Therefore, the unknown noise source impedance Z_X is derived as

$$Z_X = \frac{V_{M1}}{I_w} - Z_{in} \tag{1}$$

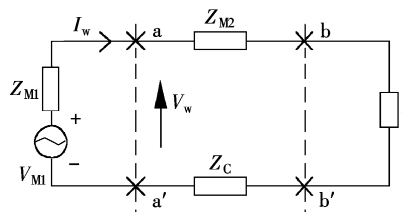

Fig. 1 Main equivalent circuit for two-current-probe measurement

In order to achieve the impedance magnitude of noise source from Eq. (1), the components V_{M1}, I_w and Z_{in} should be known separately. First, since the equivalent voltage source V_{M1} and the secondary coil current I_w of the exciting probe are both involved in the main measurement circuit and the voltage response is measured by the detecting probe, there exists an internal linear relationship between V_{M1} and I_w. Secondly, considering the linear transform between the primary coil voltage and the secondary coil voltage, there should exist a linear relationship between the signal generator output voltage V_{sig} on the primary coil of the exciting probe and the induced voltage on the secondary coil which is represented by the equivalent voltage V_{M1}. Moreover, it is regarded as a linear relationship between the measured voltage V_{p2} of the detecting probe and its mutual impedance Z_{M2}, or, in other words, there exists a linear relationship between the measured voltage V_{p2} of the detecting probe and the current flow I_w in the main circuit. Substituting the above conditions into Eq. (1), then the unknown noise source impedance can be changed into

$$Z_X = \frac{KV_{sig}}{V_{p2}} - Z_{in} \tag{2}$$

where K is a constant, and V_{p2} is the measured voltage acquired by the detecting probe and is displayed by the spectrum analyzer in the frequency domain. If the output voltage level of an RF signal generator remains unchanged, then, for a given frequency, the component KV_{sig} should also be constant. For the purpose of the source impedance Z_X estimation, two parameters, KV_{sig} and Z_{in}, should be known in the following two cases.

In case 1, by replacing the unknown measured impedance Z_X with a known large-value accurate resistance R_{std} which meets the condition $R_{std} \gg |Z_{in}|$, then the parameter KV_{sig} is obtained as $KV_{sig} \approx R_{std} V_{p2}|_{Z_x=R_{std}}$. In case 2, by replacing the unknown measured impedance Z_X with the circuit shorted, then the parameter $Z_{in} = \frac{KV_{sig}}{V_{p2}|_{Z_x=0}} = \frac{R_{std} V_{p2}|_{Z_x=R_{std}}}{V_{p2}|_{Z_x=0}}$. Finally, an actual SMPS with source impedance Z_X is placed in the measurement circuit and this unknown value is obtained by

$$Z_X = \frac{KV_{sig}}{V_{p2}|_{Z_x=\mathrm{SMPS}}} - Z_{in} = \frac{R_{std} V_{p2}|_{Z_x=R_{std}}}{V_{p2}|_{Z_x=\mathrm{SMPS}}} - Z_{in}$$

1.1.2 Insertion loss measurement method

For the insertion loss measurement, the source impedance is estimated by measuring the insertion loss at the terminal of the LISN when inserting different filters. In this method, if the filter is inserted between the load of the LISN expressed by R_{load} and the source of DUT (device under test) expressed by Z_s, then the voltage drop value across the load can be reduced and this voltage reduction is defined as the insertion loss A, which is the ratio of the voltage drop value without inserting an EMI filter to the voltage drop value with inserting an EMI filter. According to the measurement results of insertion loss A and the inserted filter impedance, the source impedance can be obtained.

When $|Z_s| \gg R_{load}$, with the assumption that $|Z_{series}| \gg |Z_s|$, the measurement with the inserted impedance in a series connection is more accurate than that in a parallel connection, and the insertion loss is derived as

$$A = \frac{\dfrac{R_{load}}{R_{load}+Z_s}V_s}{\dfrac{R_{load}}{R_{load}+Z_s+Z_{series}}V_s} = 1 + \frac{Z_{series}}{R_{load}+Z_s} \approx 1 + \frac{Z_{series}}{Z_s} \tag{3}$$

Since $|A| \gg 1$, Eq. (3) can be simplified as $|Z_s| \approx |Z_{series}|/|A|$. Furthermore, because $|Z_{series}|$ is known and the insertion loss $|A|$ can be measured, the impedance magnitude of noise source can be achieved. Usually, the larger the insertion loss, the more accurate the results.

When $|Z_s| \ll R_{load}$, with the assumption that $|Z_{series}| \ll |Z_s|$, the measurement with the inserted impedance in a parallel connection is more accurate than that in a series connection, and the insertion loss is derived as

$$A = \frac{\dfrac{R_{load}}{R_{load}+Z_s}V_s}{\dfrac{R_{load}//Z_{parallel}}{R_{load}//Z_{parallel}+Z_s}V_s} = 1 + \frac{R_{load}//Z_s}{Z_{parallel}} \approx 1 + \frac{Z_s}{Z_{parallel}} \tag{4}$$

Similar to Eq. (3), Eq. (4) can be simplified as $|Z_s| \approx$

$|Z_{parallel}||A|$. Usually, the smaller the insertion loss, the more accurate the results.

1.2 Implementation of Hilbert transform for source impedance phase extraction

Though the impedance magnitude can be extracted by the current probe method or the insertion loss method[8], etc., the phase angle of the source internal impedance is not readily measured, especially in the case where the circuit system is very complex. Fortunately, since the impedance magnitude and phase are not two independent components, based on signal and system theory, it is possible to know the phase component resulting from the magnitude component theoretically[9]. Considering the source impedance as some kind of system function, we can define the current source as a system exciting source and the voltage across source as a system response. Additionally, in conducted EMI analysis the noise source can be separated into the common mode impedance and the differential mode impedance, and also these impedances can be represented by a (R, L, C) series-connected equivalent circuit. Being a system function in the Laplace domain, the noise source impedance is expressed as

$$H_1(s) = R + sL + \frac{1}{sC} \tag{5}$$

If the magnitude and the phase of the source impedance form the real and the imaginary parts of the frequency-domain causal system function, then the relationship between these two parts is obtained by using the Hilbert transform, which means that the phase estimation can be realized by the impedance magnitude[10]. This realization can be implemented by the logarithmic operation of the source impedance so as to let the logarithmic function of magnitude and phase form a real part and an imaginary part of a complex number. Now we calculate the logarithmic function of system function $H_1(s)$, i.e. $H_2(s) = \ln[H_1(s)]$.

The frequency-domain system function of the causal system $H_2(s)$ is written as

$$H_2(\omega) = H_{2R}(\omega) + jH_{2I}(\omega) = \ln[|H_1(\omega)|] + j\arg[H_1(\omega)] \tag{6}$$

where $H_1(\omega)$ is the Fourier transform of $h_1(t)$ and $\arg[\cdot]$ represents the phase of a complex number. Then

$$H_{2R}(\omega) = H_{2I}(\omega) * \frac{1}{\pi\omega} = \frac{1}{\pi}\int_{-\infty}^{\infty}\frac{H_{2I}(\xi)}{\omega-\xi}\mathrm{d}\xi = \text{Hilbert}[H_{2I}(\omega)] \tag{7}$$

$$H_{2I}(\omega) = H_{2R}(\omega) * \frac{-1}{\pi\omega} = \frac{-1}{\pi}\int_{-\infty}^{\infty}\frac{H_{2R}(\xi)}{\omega-\xi}\mathrm{d}\xi = -\text{Hilbert}[H_{2R}(\omega)] \tag{8}$$

where $*$ means the convolution operation and Hilbert$[\cdot]$ represents the Hilbert transform. Thus

$$\arg[H_1(\omega)] = \frac{-1}{\pi}\int_{-\infty}^{\infty}\frac{\ln[|H_1(\xi)|]}{\omega-\xi}\mathrm{d}\xi = -\text{Hilbert}[\ln[|H_1(\omega)|]] \tag{9}$$

The impedance is estimated by using the Hilbert transform based on the measurement results of the impedance magnitude performance.

2 Performance Study on Hilbert Transform of Impedance Phase Estimation

2.1 Influence of singular points

It is found in Eq. (8) that if $H_{2R}(\xi)$ is not equal to zero at $\xi=\omega$, then $\xi=\omega$ should be a singular point of the integral function and numerical overflow can be generated in an integral operation. In order to eliminate its influence, Eq. (8) is simplified as

$$\begin{aligned} H_{2I}(\omega) &= \frac{1}{\pi}\int_{-\infty}^{\infty}\frac{H_{2R}(\xi)}{\xi-\omega}\mathrm{d}\xi = \\ &\frac{1}{\pi}\int_{-\infty}^{\infty}\frac{H_{2R}(\xi)-H_{2R}(\omega)}{\xi-\omega}\mathrm{d}\xi = \\ &\frac{1}{\pi}\int_{-\infty}^{\infty}\frac{[H_{2R}(\xi)-H_{2R}(\omega)](\xi+\omega)}{\xi^2-\omega^2}\mathrm{d}\xi = \\ &\frac{2\omega}{\pi}\int_{0}^{\infty}\frac{H_{2R}(\xi)-H_{2R}(\omega)}{\xi^2-\omega^2}\mathrm{d}\xi \end{aligned} \tag{10}$$

where $H_{2R}(\omega) = \ln[|H_1(\omega)|]$ is the even function of ω, and $1/\omega$ is an odd function of ω. If so, only when function $H_{2R}(\xi)$ is continuous at $\xi=\omega$ can the zero points of the integrated function at $\xi=\omega$ cancel the pole points at $\xi=\omega$. Thus the singular point problem no longer appears.

First, we consider a (R, L) series-connected circuit where $R = 2.29\ \Omega$ and $L = 3.2\ \mu\text{H}$. As shown in Figs. 2(a) and (b), compared with the phase estimation at $\omega = 0$ using the

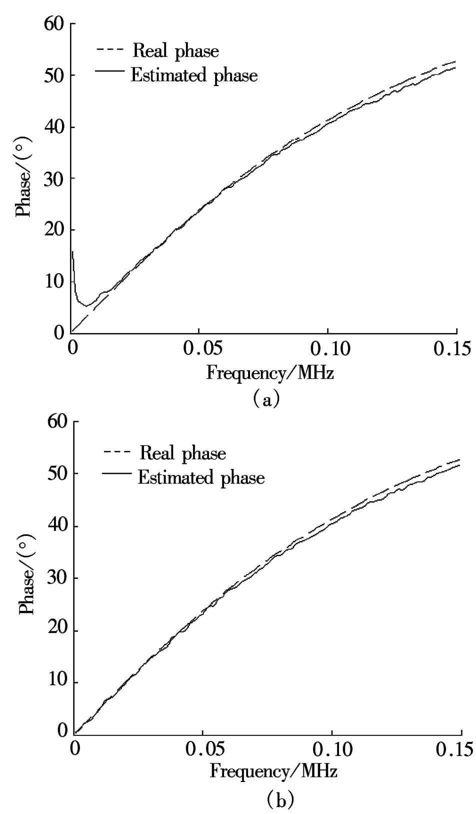

Fig. 2 Comparison between real and estimated impedance phases of the (R, L) circuit near singular point. (a) Singular point processing is not done; (b) Singular point processing is done

original formula (8), the phase estimation error using the modified formula (10) is greatly decreased because of the convergence of the impedance system at $\omega = 0$.

Secondly, we consider a series-connected capacitor involved in the circuit, where $R = 2.29\ \Omega$, $L = 3.2\ \mu\text{H}$, $C = 22\ \mu\text{F}$. As shown in Figs. 3 (a) and 3(b), no matter whether using the original formula (8) or the modified formula (10), the phase estimation error is always relatively great near $\omega = 0$ because of non-convergence of the impedance system at this point.

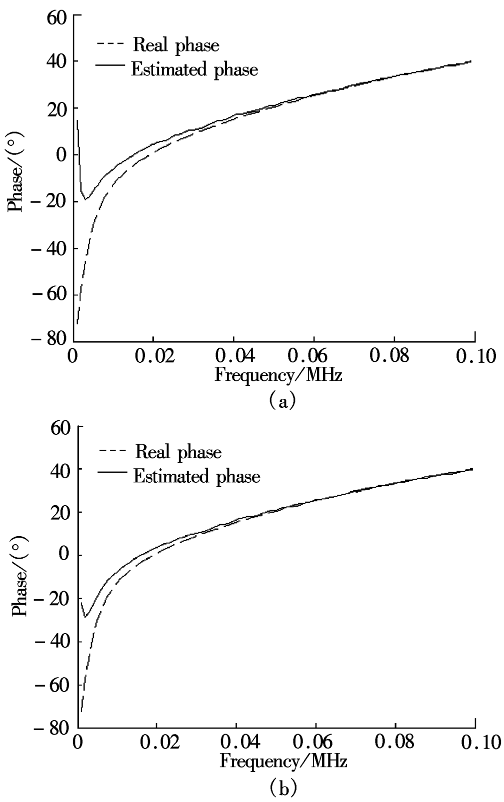

Fig. 3 Comparison between real and estimated impedance phases of the (R, L, C) circuit near singular point. (a) Singular point processing is not done; (b) Singular point processing is done

2.2 Influence of the inductive component in DM impedance modeling

In order to have a better understanding of the influence of the inductive component on phase estimation using the Hilbert transform, the performance of a (R, L) series-connected circuit representing the typical DM source impedance is researched with simulations. Assuming that the interval of measurement frequencies is 1 kHz and the maximum measurement frequency is 60 MHz, then the error of the impedance magnitude measurement can be ignored. Let the resistance be a fixed value $R = 2.29\ \Omega$. Then the relationship between the phase estimation error and the measured frequency is achieved by changing the parameter value of R/L which is shown in Fig. 4. It is found that in DM source impedance phase estimation, the accuracy is not sensitive to the influence of the inductive component, especially at high frequencies. Thus, there is no need to consider the influence on the accuracy from the ratio of R/L in noise sources.

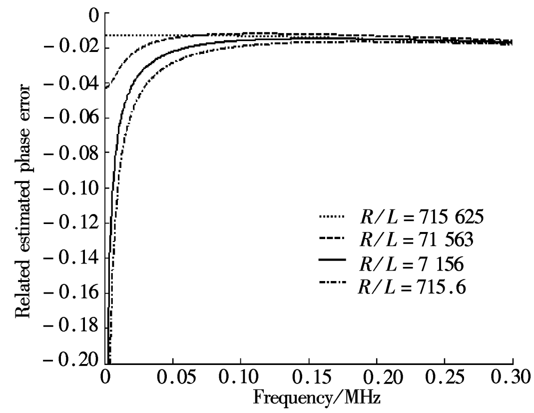

Fig. 4 Performance of estimated error of DM source impedance circuit with changes of parameter R/L

2.3 Influence of capacitive component in CM impedance modeling

Similar to the inductive component influence research, the capacitive component influence is also researched by simulations for an equivalent (R, C) series-connected circuit representing typical CM source modeling. Let the resistance be a fixed value; i. e., $R = 1.05\ \Omega$. Then the relationship between the phase estimation error and the measured frequency is achieved by changing the parameter value of $1/(RC)$ as shown in Fig. 5. It can be seen that, in CM source impedance phase estimation, the accuracy is sensitive to the influence of the capacitive component. Thus, different from DM source modeling, we should pay more attention to the phase estimation process in CM source modeling when the capacitive component is relatively great in the use of the Hilbert transform.

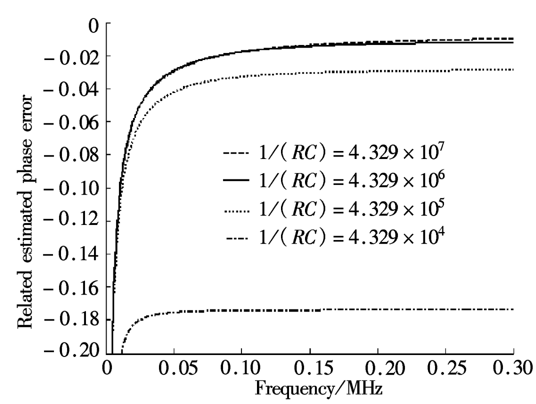

Fig. 5 Performance of estimated error of (R, C) series-connected circuit with changes of parameter $1/(RC)$

3 Application

In order to check the accuracy differences between the conventional impedance magnitude fitting method (IMFM) given in Ref. [7] and the Hilbert transform method (HTM) presented in this paper, a comparison is made between these two methods where they are both utilized for impedance phase extraction after obtaining the impedance magnitude-frequency information. In this example, the series-connected circuit parameters involving a coupling circuit are $R = 1.12\ \Omega$, $L = 0.24\ \mu\text{H}$, $C = 2.2\ \mu\text{F}$. The impedance magnitude curve is shown in Fig. 6 with a frequency interval of 1 kHz and a maximum measurement frequency of 30 MHz. Also assume that there exists an average distributed measurement error of 1% for the magnitude curve. Then two methods, the IMFM and the HTM, are respectively employed for source impedance modeling and the results are compared (see Tab. 1). The formulae used for the IMFM and the HTM are given below.

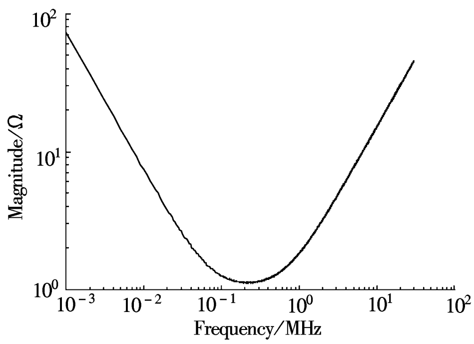

Fig. 6 Measured impedance magnitude with assumption of 1% measurement error

Tab. 1 Results comparison of circuit parameter estimation between the conventional IMFM and the HTM %

Methods	Estimation error		
	Resistance R/Ω	Inductance L/μH	Capacitance C/μF
IMFM	2	184	−30. 55
HTM	0. 21	5. 71	−2. 28

For IMFM processing, the parameter estimation can be accomplished by extraction from three non-zero frequency points of impedance magnitude, i. e. $|H_1(\omega_1)|$, $|H_1(\omega_2)|$ and $|H_1(\omega_3)|$. Then

$$\hat{C}=\sqrt{\frac{\dfrac{1}{\omega_1^2\omega_2^2}-\dfrac{1}{\omega_1^2\omega_3^2}}{\dfrac{|H_1(\omega_3)|^2-|H_1(\omega_1)|^2}{\omega_3^2-\omega_1^2}-\dfrac{|H_1(\omega_2)|^2-|H_1(\omega_1)|^2}{\omega_2^2-\omega_1^2}}}$$

$$\hat{L}=\sqrt{\frac{|H_1(\omega_2)|^2-|H_1(\omega_1)|^2}{\omega_2^2-\omega_1^2}+\left(\frac{1}{\hat{C}^2}\right)\frac{1}{\omega_2^2\omega_1^2}}$$

$$\hat{R}=\sqrt{|H_1(\omega_1)|^2+\frac{2\hat{L}}{\hat{C}}-\hat{L}^2\omega_1^2-\frac{1}{\hat{C}^2}\frac{1}{\omega_1^2}}$$

For HTM processing, the parameter estimation can be accomplished by extraction from two non-zero frequency points of complex impedance value, i. e. $H_1(\omega_1)$ and $H_1(\omega_2)$. Then

$$\hat{R}=\mathrm{Re}[H_1(\omega_1)]$$

$$\hat{L}=\frac{\omega_1\mathrm{Im}[H_1(\omega_1)]-\omega_2\mathrm{Im}[H_1(\omega_2)]}{\omega_1^2-\omega_2^2}$$

$$\hat{C}=\frac{\dfrac{1}{\omega_2^2}-\dfrac{1}{\omega_1^2}}{\dfrac{\mathrm{Im}[H_1(\omega_1)]}{\omega_1}-\dfrac{\mathrm{Im}[H_1(\omega_2)]}{\omega_2}}$$

It can be concluded from Tab. 1 that the IMFM is sensitive to the measurement error, whereas the HTM is not sensitive; the HTM is more robust and accurate than the IMFM.

4 Conclusion

Currently the most efficient technique used to suppress the conducted EMI noise on power lines is to insert a power-line EMI filter. The EMI filter performance greatly depends on the impedance-match condition between the filter and the noise source. Thus, the modeling of the noise source is important for source impedance extraction. However, the conventional modeling methods sometimes are too empirical or very rough and cannot provide an acceptably accurate source impedance for EMI filter design. Therefore, in this paper a new approach, based on the Hilbert transform in cooperation with current probe measurement, is presented for conductive EMI noise source modeling, and it can provide good source impedance estimation in frequency ranges up to 30 MHz.

Though the approach has been widely analyzed in literature, some issues are still necessary to be discussed further. For example, in the numerical solution for impedance phase estimation based on the Hilbert transform, it is necessary to implement a convolution operation for the impedance magnitude logarithm and the frequency reciprocal.

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一种传导电磁干扰噪声源建模新方法

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摘要:提出一种传导噪声源建模即内阻抗提取新方法. 该方法首先通过激励和检测探头测得阻抗模,或通过串联已知高阻抗、并联已知低阻抗的插损测量,计算得到阻抗模. 然后利用得到的阻抗模的对数进行 Hilbert 变换(HT)提取阻抗相位. 对该方法的性能研究表明,Hilbert 变换在零频率处存在的奇异点会引起该处相位估计误差骤增,对不含串联电容的情况可引入零点以消除该影响;当频率高于 150 kHz 时,估计误差对感性噪声源的变化不敏感,而对容性噪声源的变化敏感;同样阻抗模测量精度下,基于 HT 获取复阻抗比基于阻抗模拟合的参数估计精度提高近 10 倍.

关键词:噪声源建模;阻抗估计;电磁干扰

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