New High-Frequency Core Loss Measurement Method With Partial Cancellation Concept

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Abstract—As an essential part in a power converter, the magnetic cores and their design play an important role in achieving high efficiency and high power density. Accurate measurement of the core loss is important to their optimization. To improve the measurement accuracy (especially at high frequencies), previous methods proposed to cancel the reactive voltage of the testing core by using a cancellation capacitor or inductor. However, the value of the cancellation component is critical, and a small variation can induce a big measurement error, so extra effort is required to fine-tune the cancellation component value, which is a very time-consuming process and makes the standardization of the measurement instrument almost impossible. This paper presents a new measurement method with a partial cancellation concept that enables accurate core loss measurement for arbitrary wave excitation without the requirement to fine-tune the cancellation component value. The proposed method is experimentally verified up to 10 MHz.

Index Terms—Core loss measurement, high frequency, magnetic integration, magnetic loss.

I. INTRODUCTION

The desire for smaller and smarter electronics in recent years generates a significant demand to improve the power density of converters, especially in the miniaturization and integration of the traditional bulky magnetics [1]–[4]. To achieve this goal, power converter’s switching frequency has been pushed higher and higher, which makes the core loss an important factor in magnetic design and converter efficiency. Therefore, accurate core loss measurement, especially at high frequencies, plays an important role in magnetic and thermal design [5], [6], as well as the exploration of new high-frequency magnetic materials [7]–[9]. Among the existing core loss measurement methods in the prior arts [10]–[19], the two-winding method is widely used and considered a classical method [16]–[19]. Its measurement setup and equivalent circuit are shown in Fig. 1. The core loss can be extracted by integrating $v_2$ and $i_R/R_{	ext{ref}}$ over time. This method can exclude the winding loss from the measured core loss, and it is applicable to arbitrary wave excitation. However, it suffers from its sensitivity to phase discrepancy. The phase discrepancy is the phase angle difference between $i_R$ and measured $v_R$. This discrepancy usually comes from the current sensing resistor’s parasitic, a mismatch between probes, and an oscilloscope’s sampling resolution; the influence of them becomes very severe at high frequency. For example, a 0.2-Ω sensing resistor with a 1-nH ESL will produce an 8.9° phase discrepancy at 5 MHz. The measurement error $\Delta$ caused by the phase discrepancy $\Delta \phi$ for sinusoidal excitation is [20]

$$\Delta = \tan(\phi_2) \cdot \Delta \phi$$ (1)

where $\phi_2$ is the phase angle difference between $v_2$ and $i_R$, which is very close to 90° since the impedance of magnetizing inductor $L_m$ is usually much smaller than the equivalent core loss resistor $R_{\text{core}}$. This makes $\tan(\phi_2)$ very large and leads to the sensitivity to phase discrepancy. The relation between $\phi_2$ and the measurement error is shown in Fig. 2. For the rectangular excitation, a small phase discrepancy can also lead to large measurement error at high frequency [21].
Mu’s capacitive cancellation method [13] is created to overcome the drawback of two-winding method. Its idea is to decrease $\varphi_2$ from nearly 90° to nearly 0° by cancelling the reactive voltage of the testing core. The equivalent measurement circuit is shown in Fig. 3. The cancellation of the reactive voltage is achieved by adding a cancellation capacitor $C$ to resonate with the testing core, and using the voltage across the resonant tank $v_3$ to calculate core loss

$$P_{\text{core}} = \frac{f}{R_{\text{ref}}} \int_0^T v_3(t) v_R(t) \, dt$$

(2)

where $R_{\text{ref}}$ is the current sensing resistance, $f$ is the excitation frequency, and $T$ is the period of the frequency. Using this method, the measurement error $\Delta$ caused by the phase discrepancy $\Delta \varphi$ becomes

$$\Delta = \tan(\varphi_3) \cdot \Delta \varphi$$

(3)

where $\varphi_3$ is the phase angle difference between $v_3$ and $i_R$. When the cancellation capacitor exactly resonates with $L_m$ at the excitation frequency, the reactive voltage of the testing core can be perfectly cancelled out, so $\varphi_3 = 0$, and thus an accurate result can be acquired. However, despite the certain tolerable range of the phase angle $\varphi_3$ [13], the value of the cancellation capacitor is still critical to the measurement accuracy. The critical capacitance to achieve perfect cancellation (at excitation frequency $\omega / 2\pi$) is

$$C_0 = \frac{1}{\omega^2 L_m} \left[ 1 + \left( \frac{\omega L_m}{R_{\text{core}}} \right)^2 \right].$$

(4)

And the phase angle $\varphi_3$ is

$$\varphi_3 = \arctan \left[ \left( 1 - \frac{C_0}{C} \right) \cdot \tan(\varphi_2) \right].$$

(5)

Based on (5), the relation between $\varphi_3$ and the cancellation capacitance $C$ can be plotted in Fig. 4. It can be seen that a small difference between $C$ and $C_0$ can lead to a significant jump in $\varphi_3$, thus resulting in a big measurement error [according to (3)], especially for low loss magnetic material (whose $\varphi_2$ is very close to 90°). Therefore, it is necessary to design and fine-tune the cancellation capacitor for each testing sample at each testing condition (frequency, flux density, temperature, etc.). This is a very time-consuming process and makes the standardization of the measurement instrument almost impossible.

Mu’s inductive cancellation method [14] is based on the same principle of Mu’s capacitive cancellation method, whereas it uses an inductor instead of a capacitor to cancel the reactive voltage (see Fig. 5), which extends its application range from sinusoidal excitation to arbitrary excitation. Similar to the capacitive cancellation method, the accuracy of this method is also sensitive to the cancellation inductance value $L$. For sinusoidal excitation, the critical inductance needed to achieve the perfect cancellation is

$$L_0 = L_m \cdot \frac{R_{\text{core}}^2}{R_{\text{core}}^2 + (\omega L_m)^2}.$$  

(6)

And the phase angle $\varphi_3$ is

$$\varphi_3 = \arctan \left[ \left( 1 - \frac{L}{L_0} \right) \cdot \tan(\varphi_2) \right].$$

(7)

Similar to the capacitive cancellation method, a small difference between $L$ and $L_0$ can lead to a significant jump in $\varphi_3$ (as shown in Fig. 6), and thus results in a big measurement error.

II. PROPOSED METHOD WITH PARTIAL CANCELLATION CONCEPT

A new method is proposed to overcome the accuracy sensitivity to the cancellation component. The equivalent circuits for both the capacitive and inductive cancellation versions are shown in Fig. 7. The capacitive cancellation version can only be used for sinusoidal excitation, and the inductive cancellation version is applicable for arbitrary excitation. By comparing Figs. 7 to 3 and 5, it can be seen that the proposed method is a better utilization of the similar circuit configuration of Mu’s cancellation method, whereas the voltage across the cancellation component ($v_L$ or $v_C$) is measured instead of $v_3$ to implement
the partial cancellation concept. The working principle of both the inductive and capacitive cancellation version will be demonstrated and analyzed in the following discussion.

Let us consider the inductive cancellation version under sinusoidal excitation first. In the proposed method, three parameters are measured: \(i_R\) (by measuring \(v_R\)), \(v_2\) and \(v_L\). So three types of phase discrepancy may be involved in the measurement: \(\Delta \phi_1\) (phase discrepancy between measured \(v_R\) and real \(i_R\)), \(\Delta \phi_2\) (phase discrepancy between measured \(v_2\) and real \(v_2\)), and \(\Delta \phi_L\) (phase discrepancy between measured \(v_L\) and real \(v_L\)). We will analyze them in two steps.

Step 1: consider \(\Delta \phi_1\) only (assuming \(\Delta \phi_2 = \Delta \phi_L = 0\)). At an excitation frequency of \(\omega/2\pi\), we have

\[
\begin{align*}
i_R &= I_R \cdot \sin (\omega t) \\
v_2 &= V_2 \cdot \sin (\omega t + \phi_2) \\
v_L &= V_L \cdot \sin (\omega t + 90^\circ)
\end{align*}
\]

(8) (9) (10)

According to [20], the measured loss of the core under test is

\[
f \frac{f}{R_{ref}} \int_0^T v_2 v_R dt = P_{core} + \Delta P_{core}
\]

\[
= P_{core} + V_2 \cdot \sin \phi_2 \cdot I_R \cdot \Delta \phi_1
\]

(11)

where \(P_{core}\) is the real core loss and \(\Delta P_{core}\) is the error caused by \(\Delta \phi_1\). Similarly, the measured power loss of the reference air core is

\[
f \frac{f}{R_{ref}} \int_0^T v_L v_R dt = V_L \cdot I_R \cdot \cos (90^\circ)
\]

\[
+ V_L \cdot I_R \cdot \sin (90^\circ) \cdot \Delta \phi_1 = V_L \cdot I_R \cdot \Delta \phi_1
\]

(12)

The result of (12) is proportional to the error in (11) \((\Delta P_{core})\), with the proportion of

\[
k = \frac{V_L}{V_2 \cdot \sin \phi_2}
\]

(13)

This \(k\) factor is named cancellation factor, which represents the percentage of cancelled reactive voltage to the total reactive voltage.

Combining (11)–(13), we can get the real core loss as

\[
P_{core} = f \frac{f}{R_{ref}} \int_0^T v_2 v_R dt - \frac{1}{k} f \frac{f}{R_{ref}} \int_0^T v_L v_R dt
\]

(14)

Then, the remaining problem is how to find the cancellation factor \(k\). The difficulty in finding \(k\) is that \(V_2 \cdot \sin \phi_2\) cannot be measured out directly. Here, we propose a method to find \(k\) by adding a phase perturbation \(\Delta \phi_1\) into \(v_R\) with the de-skew function of the oscilloscope. If we call the measured voltage of the current sensing resistor after the phase perturbation as \(v'_R\), we have

\[
P_{core} = f \frac{f}{R_{ref}} \int_0^T v_2 v'_R dt - \frac{1}{k} f \frac{f}{R_{ref}} \int_0^T v_L v'_R dt
\]

(15)

Combining (14) and (15), we have

\[
k = \frac{\int_0^T v_L v'_R dt - \int_0^T v_L v_R dt}{\int_0^T v_2 v'_R dt - \int_0^T v_2 v_R dt}
\]

(16)

After \(k\) is known, (14) can be used to extract the real core loss of the testing core.

To make a proper choice of \(\Delta \phi_1\), two factors should be considered: 1) it should be a small perturbation to validate the linear approximation and 2) a too small value should also be avoided because the difference before and after the perturbation is used to calculate \(k\) in (16). In our measurement, we found that \(\Delta \phi_1 \approx 1^\circ\) can be a proper choice.

Step 2: Consider all three phase discrepancy \((\Delta \phi_1, \Delta \phi_2, \Delta \phi_L)\) together (now the measured voltages are \(v'_2, v'_L\), and \(v_R\)).

Following the same procedure in step 1, we can derive

\[
f \frac{f}{R_{ref}} \int_0^T v'_2 v_R dt - \frac{1}{k} f \frac{f}{R_{ref}} \int_0^T v'_L v_R dt
\]

\[
= P_{core} + V_2 \cdot \sin \phi_2 \cdot I_R \cdot (\Delta \phi_2 - \Delta \phi_L)
\]

(17)
The results show that there exists a measurement error of \( [V_2 \cdot \sin \varphi_2 \cdot I_R \cdot (\Delta \varphi_2 - \Delta \varphi_L)] \). The error comes from the remaining phase discrepancy \( (\Delta \varphi_2 - \Delta \varphi_L) \), which is caused by the phase mismatch between the two probes for \( v_2 \) and \( v_L \). To diminish this error, we can use the two probes to measure the same voltage \( v_2 \) simultaneously (so that two \( v_2 \) are generated by the two probes), and then observe the integration \( \frac{f}{R_{ref}} \int_0^T v_2 v_R dt \) calculated by each of them. The integration result from the two probes are \( [P_{core} + V_2 \cdot \sin \varphi_2 \cdot I_R \cdot (\Delta \varphi_1 + \Delta \varphi_2)] \) and \( [P_{core} + V_2 \cdot \sin \varphi_2 \cdot I_R \cdot (\Delta \varphi_1 + \Delta \varphi_2)] \), so the difference between them is \( [V_2 \cdot \sin \varphi_2 \cdot I_R \cdot (\Delta \varphi_2 - \Delta \varphi_L)] \), which is the same as the error involved in the measurement result. Then, by adjusting \( \Delta \varphi_2 \) or \( \Delta \varphi_L \) using the de-skew function, the difference between the two integration results from the two probes can be diminished, and thus the measurement error is diminished too.

For the inductive cancellation version under rectangular excitation, according to the analysis in [21], the previous (11)–(13) for sinusoidal excitation become the following equations:

\[
\frac{f}{R_{ref}} \int_0^T v_2 v_R dt = P_{core} + \frac{V_{pp}^2 I_{pp} \Delta t}{T} \tag{18}
\]

\[
\frac{f}{R_{ref}} \int_0^T v_L v_R dt = \frac{V_{Lpp}^2 I_{pp} \Delta t}{T} \tag{19}
\]

\[
k = \frac{V_{Lpp}}{V_{pp}} \tag{20}
\]

where \( \Delta t \) is a small time delay caused by phase discrepancy, \( I_{pp} \) is the peak-to-peak value of magnetizing current, \( V_{2pp} \) and \( V_{Lpp} \) are the peak-to-peak values of \( V_2 \) and \( V_L \) shown in Fig. 7(b). Then, the same procedure discussed above for sinusoidal excitation can be applied for rectangular excitation.

For the capacitive cancellation version, it works in the same principle as the inductive cancellation version under sinusoidal excitation, whereas the previous (10), and (12)–(16) become the following equations:

\[
v_C = V_C \cdot \sin(\omega t - 90^\circ) \tag{21}
\]

\[
\frac{f}{R_{ref}} \int_0^T v_C v_R dt = V_C \cdot I_R \cdot \cos(90^\circ) + V_C \cdot I_R \cdot \sin(-90^\circ) \cdot \Delta \varphi_1 = -V_C \cdot I_R \cdot \Delta \varphi_1 \tag{22}
\]

\[
k = \frac{V_C}{V_2 \cdot \sin \varphi_2} \tag{23}
\]

\[
P_{core} = \frac{f}{R_{ref}} \int_0^T v_2 v_R dt + \frac{1}{k} \frac{f}{R_{ref}} \int_0^T v_C v_R dt \tag{24}
\]

\[
P_{core} = \frac{f}{R_{ref}} \int_0^T v_2 v_R' dt + \frac{1}{k} \frac{f}{R_{ref}} \int_0^T v_C v_R' dt \tag{25}
\]

\[
k = \frac{\int_0^T v_C v_R dt - \int_0^T v_C v_R' dt}{\int_0^T v_2 v_R' dt - \int_0^T v_2 v_R dt} \tag{26}
\]

And the discussion in step 2 about diminishing the error caused by phase discrepancy between \( \Delta \varphi_2 \) and \( \Delta \varphi_L \) holds the same for the phase discrepancy between \( \Delta \varphi_2 \) and \( \Delta \varphi_C \) in the capacitive cancellation version.

In another point of view, the effect of partial cancellation mechanism can be regarded as to create a virtual voltage measurement of \( v_3 = v_2 - \frac{1}{k} v_C \) (for inductive cancellation) and \( v_3 = v_2 + \frac{1}{k} v_C \) (for capacitive cancellation) to replace the initial \( v_3 = v_2 - v_L \) (for inductive cancellation) and \( v_3 = v_2 + v_L \) (for capacitive cancellation) in Mu’s cancellation method. As a result, the effective \( \varphi_3 \) can keep almost zero even when the cancellation component value \( L \) or \( C \) has a mismatch with its critical value \( L_0 \) or \( C_0 \) (in contrast to the significant jump of \( \varphi_3 \) shown in Figs. 4 and 6), and therefore the measurement error can be minimized.

It is worth mentioning that the proposed method is applicable not only to the core loss measurement, but also to the total inductor loss measurement. By simply replacing the core under test with an inductor, and using the \( v_2 \) probe to measure the voltage across the inductor (with everything else kept the same), the measurement result will provide the total inductor loss including core loss and winding loss.

### III. Experimental Verification

To verify the proposed method, the core loss of a 3F4 ferrite toroid core from Ferroxcube (TN10/6/4) is measured under sinusoidal excitation with a 2-MHz frequency, a 20-mT peak flux density, and 100°C. The excitation voltage is supplied by a power amplifier (Amplifier Research® 25A250A) driven by a function generator (Tektronix AFG3102). The Tektronix TDP1000 differential probe is used to sense the voltage signal. The core loss is first measured using Mu’s inductive cancellation method with a different cancellation inductance value \( L \). The waveform and measurement result are shown in Fig. 8. It shows that the measurement result is sensitive to the cancellation inductance value \( L \). The perfect cancellation \( (v_3 \) and \( v_R \) in phase) is achieved at \( L = 3.7 \mu \text{H} \), and the accurate core loss density is 351 kW/m³. Then, the core loss is measured using the proposed method. The waveform working and corresponding result is shown in Fig. 9. It shows that the measurement result with a different \( L \) can all match the accurate result from Mu’s inductive cancellation method at the perfect cancellation condition (with an error \( \Delta \leq 3\% \)). Another comparison between Mu’s method and the new proposed method is performed at 10 MHz with the core loss measurement of Ferroxcube’s 4F1 ferrite material, and the measurement results are summarized in Fig. 10.
Fig. 9. Working waveform and measurement result using proposed method for sinusoidal wave excitation. $v_2$ and $v_L$ in 5 V/div, $v_R$ in 50 mV/div.

Fig. 10. Measurement results at different cancellation inductance value for 4F1 ferrite material at 10 MHz, 6-mT peak flux density. Triangular points: Mu’s cancellation method; circular points: proposed method.

For the rectangular wave excitation, the core loss of an iron powder material -52 from Micrometals, Inc. is measured at 0.5 MHz, with a 10-mT peak flux density, and room temperature. The working waveform and measurement results using Mu’s inductive cancellation method and proposed method are shown in Figs. 11 and 12. It can be seen that the proposed method can effectively overcome the accuracy sensitivity to the $L$ value shown in Mu’s inductive cancellation method. Furthermore, the core loss of 3F35, 3F4, and 3F45 ferrites from Ferroxcube are also measured using the proposed method under the rectangular excitation, and the core loss ratio under rectangular and sinusoidal excitations are summarized in Fig. 13.

IV. ERROR ANALYSIS

To ensure the accuracy of the proposed measurement method, it is necessary to identify and analyze various error sources in the system, including amplitude discrepancy, phase discrepancy, parasitic capacitance, component tolerance, etc.

A. Amplitude Discrepancy

The oscilloscope used in the measurement (Tektronix MSO5104) has an 8-bit resolution for an analog-to-digital (ADC) conversion. Considering the ADC digitizing error of ±1/2 least significant bit, the measurement error caused by the amplitude discrepancy is ±0.2%. Moreover, the amplitude sensing error of the voltage probe is minimized by the Tektronix calibration fixture: 067-1686-00, and the random error in the measurement is further reduced by taking the average value of a large sampling number.

The amplitude measurement discrepancy can also cause an error in the measurement of the cancellation factor $k$ in (16), especially when two different datasets are acquired before and after the de-skew process in the oscilloscope. But this error can be reduced by an alternative (and also more convenient) way of performing the de-skew process: instead of using the oscilloscope’s de-skew function, one can save all the testing data of $v_2$, $v_L$, and $v_R$ in a computer, and then use a simple computer program to perform the effective de-skew (i.e., a small phase shift) of $v_R$ to get $v'_R$. In this way, the $v_R$ and $v'_R$ data used in the $k$ calculation comes from a single data acquisition, and thus
their amplitude discrepancy will cancel out in the numerator and denominator in (16). Therefore, it is safe to ignore the impact from \( v_{R} \) and in the worst case the error can be calculated by (27), which is 0.4% 

\[
\frac{\Delta k}{k} = \left| \frac{\int_0^T v_L (1 \pm 0.2\%) v_R' dt - \int_0^T v_L (1 \pm 0.2\%) v_R dt}{\int_0^T v_L v_R' dt - \int_0^T v_L v_R dt} - 1 \right|
\]

\[
= 0.4\%.
\]

(27)

B. Phase Discrepancy

The source of the phase discrepancy in the measurement system includes the ESL of the current sensing resistor, the phase mismatch between the voltage probes, and the scope sampling resolution limit (a detailed procedure of evaluating phase discrepancy is described in [22]). With the partial cancellation mechanism demonstrated aforesaid, it is safe to have \(< 1\%\) measurement error due to the phase discrepancy, which is not a significant contributor to the overall error.

C. Error Caused by Parasitic Capacitance

The source of the parasitic capacitance in the system includes a transformer intrawinding capacitance, a transformer interwinding capacitance, and a voltage probe input capacitance. 

Fig. 14(a) shows the simplified equivalent circuit with \( C_p \) (the total parasitic capacitance of the core under test, including transformer’s intrawinding capacitance, and \( v_2 \) probe’s input capacitance). The existence of \( C_p \) causes an extra current flowing through it, and thus introduces a voltage drop on the sensing winding’s parasitic \( L_2 \) and \( R_2 \). To quantify the error caused by the parasitic capacitance, we assume

\[
R_{\text{core}} \ll \omega L_m \quad (32)
\]

\[
\omega L_m \ll \frac{1}{\omega C_p} \quad (33)
\]

\[
R_1, R_2, \omega L_1, \omega L_2 \ll \frac{1}{\omega C_p}. \quad (34)
\]

The error caused by \( C_w \) can be simplified as

\[
\Delta C_w = \frac{P_{\text{measured}} - P_{\text{actual}}}{P_{\text{actual}}} = \frac{\int v_2 i_R dt - \frac{1}{2} \int v_L i_R dt - \int v_m i_{\text{core}} dt}{\int v_m i_{\text{core}} dt}
\]

\[
\approx \frac{-R_{\text{core}} C_w}{L_m} \left( R_1 L_2 + R_2 L_1 \right). \quad (35)
\]

In the testing example in Section III, \( C_p \approx 8 \mu F, L_m \approx 3.7 \mu H, L_2 \approx 100 \mu H, R_2 \approx 0.1 \Omega, \omega = 2\pi \cdot 2 MHz, \) and \( R_{\text{core}} \approx 4.6 \Omega, \) so \( \Delta C_p \approx 0.11\% \).

Besides \( C_p \), the transformer interwinding capacitance \( C_w \) can also cause an extra current and introduce a measurement error, as shown in Fig. 14 (b). Assume that

\[
R_{\text{core}} \ll \omega L_m \quad (32)
\]

\[
\omega L_m \ll \frac{1}{\omega C_p} \quad (33)
\]

\[
R_1, R_2, \omega L_1, \omega L_2 \ll \frac{1}{\omega C_p}. \quad (34)
\]
For the measurement in the example, $C_w \approx 15 \, pF, L_m \approx 3.7 \, \mu H, R_1 \approx R_2 \approx 0.1 \, \Omega, L_3 \approx L_4 \approx 100 \, nH, and R_{core} \approx 4.9 \, k\Omega$, so $\Delta C_{pc} \approx -0.011\%$.

Besides the parasitic capacitance of the core under test, the parasitic capacitance associated with the reference core $C_{pc}$ and $C_{wc}$ are also considered as shown in Fig. 15.

Assume that

$$\omega L \ll \frac{1}{\omega C_{pc}}$$

(36)

$$\omega L \ll \frac{1}{\omega C_{wc}}$$

(37)

$$R_3, R_4, \omega L_3, \omega L_4 \ll \frac{1}{\omega C_{pc}}$$

(38)

$$R_3, R_4, \omega L_3, \omega L_4 \ll \frac{1}{\omega C_{wc}}$$

(39)

The error caused by $C_{pc}$ and $C_{wc}$ can be simplified as

$$\Delta C_{pc} \approx -\frac{C_{pc}}{k} \left( \frac{R_3 R_{core}}{L_m} + \omega^2 L_3 \right)$$

(40)

$$\Delta C_{wc} \approx \frac{R_{core} C_{wc}}{k L_m^2} \left( R_4 L_3 + R_3 L_4 \right).$$

(41)

For the worst case in the measurement example, $k=0.09$, $C_{pc} \approx 5 \, pF, C_{wc} \approx 20 \, pF, L_m \approx 3.7 \, \mu H, R_3 \approx R_4 \approx 0.2 \Omega, L_3 \approx L_4 \approx 200 \, nH, \omega = 2\pi \Delta 2 \, MHz, and R_{core} \approx 4.9 \, k\Omega$, so $\Delta C_{pc} \approx -1.5\%, \Delta C_{wc} \approx 0.64\%$.

D. Current Sensing Resistor Variation

The variation of current sensing resistor is directly transferred to the final measurement error. The film resistor used in the measurement has a tolerance, thus a 1% relative error can be introduced in the measurement result. Besides, the resistance value also changes with frequency, as shown in Fig. 16. The resistance value at the frequency of measurement should be used to diminish the error caused by the frequency variation of the sensing resistor.

V. Conclusion

A new core loss measurement method for arbitrary excitation with a partial cancellation concept is proposed and experimentally verified with both sinusoidal and rectangular wave excitation. It overcomes the accuracy sensitivity to the cancellation component in the existing methods and enables an accurate measurement without fine-tuning of the cancellation component value. The proposed method is experimentally verified up to 10-MHz frequency. It can also be easily extended to measure the total inductor loss.

REFERENCES


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