A Practical Transformer Core Loss Measurement Scheme for High-Frequency Power Converter

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Abstract—**A practical method to measure transformer core loss in a high-frequency switching mode power converter is proposed. The method is also applicable when the transformer has dc bias in the magnetizing current. Practical issues to minimize the measurement error are discussed. The measurement results are very close to simulation results obtained from a time-domain finite element analysis solver. Detailed error analysis for the proposed method provides useful guidelines on transformer core loss measurement for switching mode power supplies.**

Index Terms—**Core loss, finite element analysis (FEA), pulse-width modulation (PWM), switching mode power supply (SMPS).**

I. Introduction

WITH THE TREND toward higher operation frequencies for switching mode power supplies (SMPS), the accurate characterization of transformer losses, including core loss and copper loss, plays an important role in converter design and optimization. Measurement methods are widely adopted to characterize transformer losses due to the potentially higher accuracy. The core loss measurement schemes and related issues under sinusoidal excitations have been discussed a great deal in the literature [1]–[6]. But not too much work has focused on the measurement scheme for transformer core loss in SMPS, and even less has been done on practical issues and related error analysis.

In this paper, a practical transformer core loss measurement method for SMPS is proposed. Important key issues to improve measurement accuracy are also discussed. More importantly, a detailed error analysis is conducted for the proposed measurement scheme. The results provide very useful guidelines on transformer core loss measurement for SMPS.

Section II reviews existing measurement methods and highlights their limitations. In Section III, the core loss measurement method is proposed with the detailed explanation on principles and important issues within the set-up. Section IV provides measurement results for a planar transformer core loss

using the proposed method. The results are verified by finite element analysis (FEA) simulation. Section V gives the detailed error analysis on the proposed method, while Section VI presents the conclusions.

II. Review of Existing Transformer Core Loss Measurement Methods

Direct methods are widely used to measure transformer core loss. Most of the direct schemes are based on impedance or network analyzer [1]–[6] and are only suitable for sinusoidal excitations. However, the transformer in SMPS is subject to pulsewidth modulation (PWM) excitations. They are nonsinusoidal waveforms [7]–[9] and may contain dc bias in the magnetizing current to premagnetize the core [10]–[13]. Due to the highly nonlinear nature of the B-H property of ferrite materials, linear superposition techniques such as Fourier series analysis may yield completely erroneous results [13], [14]. Therefore, the impedance or network analyzer method cannot be applied to core loss measurement for SMPS.

In [15] and [16], two measurement setups were designed to test the transformer core loss for SMPS. In [15], sine waveforms are used to verify core loss results by core material vendors and no results for SMPS are provided. In [16], a core loss measurement scheme for SMPS is proposed using a waveform generator and power amplifier. Due to the low bandwidth of the power amplifier (200 kHz), it is difficult to measure core loss under higher switching frequencies. A current probe is used to measure the magnetizing current, which is expensive and introduces significant phase-shift errors.

In the two measurement set-ups above, dc bias in the magnetizing current is not included and its effect on transformer core loss is not discussed. The error analysis of their methods is not conducted and the corresponding measurement accuracy of the experimental results is not known.

A transformer core loss measurement method for high-frequency SMPS is proposed in this paper. It is also suitable for core loss measurement under sine waveforms, and dc premagnetization is included in the scheme. Detailed error analysis for the proposed method is provided. The core loss measurement method proposed in this paper can be used to achieve optimal transformer design for high frequency dc–dc converters as discussed in [17]–[21].

The authors have published a paper [23] in a conference to discuss a new method to measure the core loss of high frequency transformers in switching power converters. This paper is an improved version of that paper.
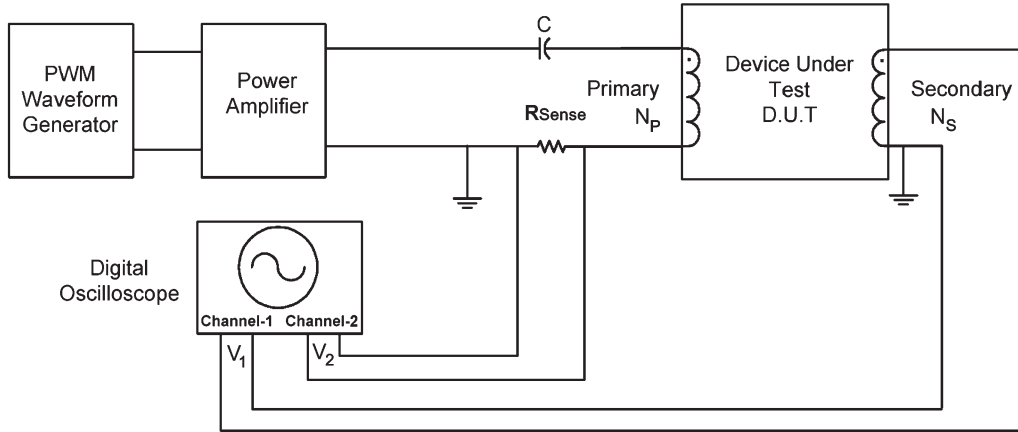
Fig. 1.   Proposed core loss measurement scheme.

III. PROPOSED TRANSFORMER CORE LOSS MEASUREMENT METHOD

A. Principle

The proposed transformer core loss measurement scheme under actual PWM waveforms of SMPS is shown in Fig. 1. The measurement principle is: With the secondary side open-circuit, the averaged core loss power P_{Core} over one switching period T is defined as

$$P_{\mathrm{Core}} = \frac{1}{T} \int_0^{\mathrm{T}} v_{\mathrm{pri}}(t) \cdot i_{\mathrm{M}}(t) \, dt. \qquad (1)$$

where $v_{\mathrm{pri}}(t)$ is the primary side voltage and $i_{\mathrm{M}}(t)$ is the magnetizing current. Due to leakage inductance and resistance associated with the primary winding, the voltage $v_{\mathrm{pri}}(t)$ cannot be measured directly. A practical approach is to measure the secondary side voltage $v_{\mathrm{sec}}(t)$ and reflect it back to the primary side by the corresponding winding turns-ratio. The magnetizing current is sensed using a resistor R_{Sense}. By measuring the two voltages using a digital oscilloscope (Tektronix TDS 5054) over one switching period, we can obtain the averaged core loss directly as

$$P_{\mathrm{Core}} = \frac{N_{\mathrm{p}}}{N_{\mathrm{s}}} \frac{1}{N} \sum_{i=1}^N v_{1i} \cdot \frac{v_{2i}}{R_{\mathrm{Sense}}}. \qquad (2)$$

where N_{p} and N_{s} are the primary and secondary winding turns; v_{1i} and v_{2i} are the ith samples of the measured voltage values of the secondary winding and the current sensing resistor; and N is the number of the samples in one switching period.

A waveform generator (Agilent 33120A) and RF power amplifier are used to provide the PWM voltage source. The power amplifier should have enough bandwidth (3 MHz in this paper). The PWM waveform is distorted due to the low bandwidth of the power amplifier, mainly on the rising and falling edges. As the magnitude of the high frequency component is very small, these distortions do not introduce significant error in the measurement. A more accurate but more complicated method is to use a full bridge converter to generate the PWM waveform.

A capacitor is connected in series with the transformer primary winding to block dc current. The parasitic inductance of this capacitor should be minimized to avoid additional track inductance. The impedance of this capacitor should be very small at the test frequency, so that it will not resonate with the leakage inductor of the transformer.

Various kinds of resistors are considered as the current sensing device. In this paper, the low-inductance metal film resistor is selected in the measurement set-up due to excellent frequency characteristics, easy application and low cost. Other devices such as the current probe and coaxial shunt resistor can also be used. Care should be taken to reduce phase delay between the current and voltage. To minimize the associated inductance, several resistors (ten $10 \pm 1\%$ Ω) are connected in parallel. Its frequency response by impedance analyzer is flat up to 5 MHz. For a more accurate result, a coaxial shunt can be used to measure the current. From the analysis and experimental measurement, a metal film resistor is adequate for switching power converter application.

In some topologies such as the Flyback converter and asymmetrical half bridge converter, the core is dc premagnetized due to the dc bias in the magnetizing current. It is shown in [10]–[13] that the dc bias will impact the core loss significantly. Fig. 2 shows a circuit to introduce dc bias in the core and at the same time eliminate ac current ripple. Two identical transformers under testing are connected. Auxiliary windings with the same turns are connected in opposite polarities to cancel the ac voltage. In the ideal case, there will be no current ripple. A large inductor, L, is connected in series to further reduce the current ripple. The dc bias level can be adjusted by adjusting the dc voltage source. It is noted that the auxiliary winding can be the extra winding in the transformer and can also be a temporary winding specifically for adding dc bias. The coupling is not critical as it is used to add dc bias.

B. Key Steps in Core Loss Measurement

Some key steps are proposed to minimize errors from various sources.

1) Calibration of Winding Turns-Ratio: The proposed core loss measurement scheme is a two-part method. The primary winding voltage is obtained by measuring the open-circuit
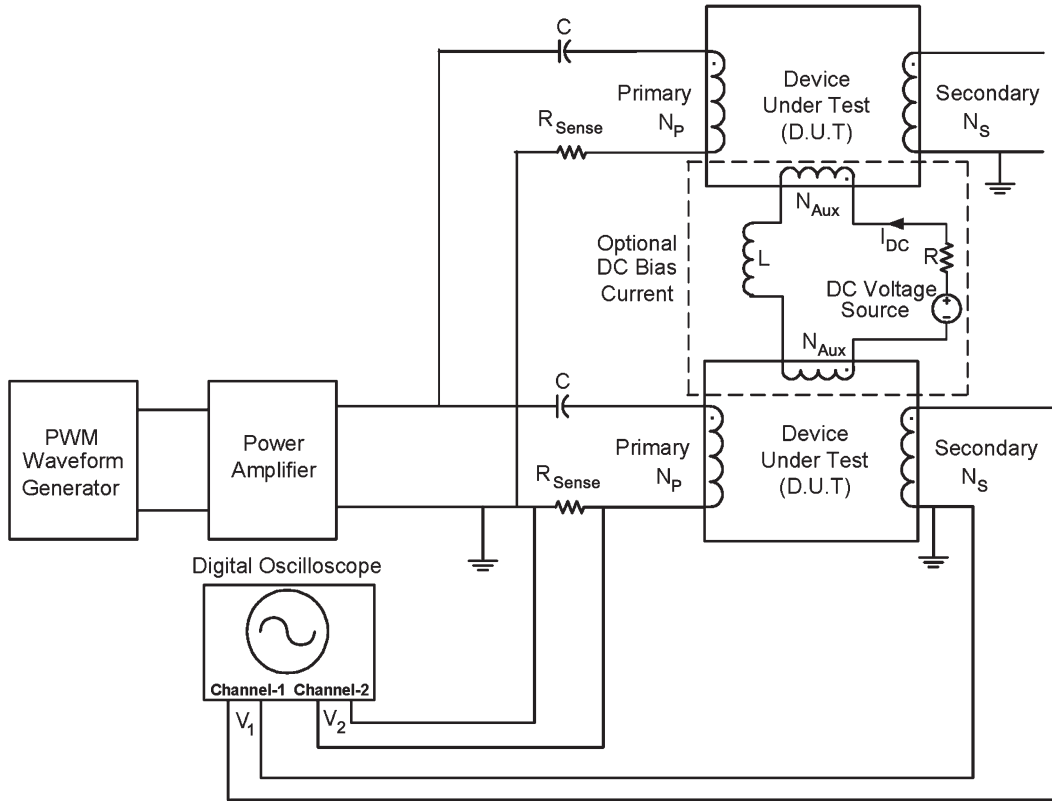
Fig. 2. Method to introduce dc bias magnetizing current core with much less current ripple.

secondary winding terminal voltage. Due to the nonperfect magnetic coupling, uncoupled magnetic flux, and winding termination, the primary and secondary winding terminal voltage ratio is different from the designed integer winding turns-ratio. Therefore, the actual turns-ratio should be calibrated and used in the loss measurement to minimize this error.

It is noted that the winding turns-ratio is a physical transformer characteristic that does not depend on excitation waveforms. Therefore, sinusoidal excitation is used for calibration, as it is simple to implement and the calibration result can be applied to PWM waveforms.

To eliminate any error, current through the primary winding and secondary winding should be zero to avoid the effect of its leakage inductance. Therefore, the excitation sinusoidal voltage should be applied to an auxiliary winding. The open-circuit terminal voltages of the primary and secondary windings are measured. The actual turns-ratio can be calculated using (3). Peak voltage or peak-to-peak voltage can be used. This calibration procedure should be conducted over a range of frequencies to cover the PWM harmonics frequency range. If the winding turns-ratio varies with frequencies, the averaging method can be used. In our test, the difference of measured turn-ratios at different frequencies is less than 0.5%

$$N_{\text{Turns-Ratio}} = \frac{V_{\text{pri}}}{V_{\text{sec}}}. \tag{3}$$

This auxiliary can be any winding in a multiple winding transformer other than the primary and secondary windings; or it can be an added winding wound around the same core. As the

physical structure of primary and secondary windings has been fixed and they are open-circuit in the calibration, addition of this auxiliary winding will not affect the coupling between the primary and secondary windings. It is noted that this auxiliary winding serves a different function from the auxiliary winding used to add dc bias that is discussed in Section III-A.

2) Parasitic Inductance Minimization: Due to the parasitic inductances associated with the current sensing resistor and the connection leads, error will be introduced into the core loss measurement. A good approach is to calibrate the resistor's frequency response using an impedance analyzer to ensure its frequency response is high enough. In the calibration, it is very important to make the resistor lead as short as possible. In addition, as discussed in the error analysis section, the other parasitic inductances should also be minimized.

3) Averaging Method: Averaging method is used in core loss data processing to minimize random error in the measurement. The measurement for each test condition is repeated several times and these core loss values are averaged. The offset error can be eliminated by reversing the excitation polarities to the transformer and repeating the measurements. In this paper, the test for both situations is repeated ten times, yielding a total of 20 tests for each test point.

IV. MEASUREMENT RESULTS AND FEA VERIFICATIONS

A high-frequency planar transformer is tested for core loss by the proposed scheme. As it is impossible to measure or calculate the exact core loss, the measurements are compared with simulation results by a commercial time-domain FEA solver
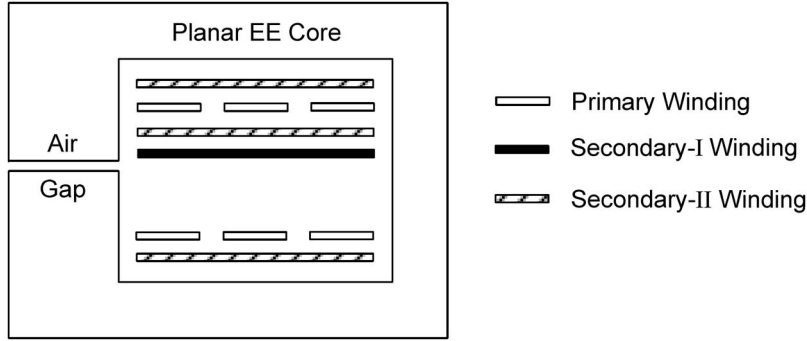
Fig. 3.   Half cross section of the designed transformer.

from ANSOFT Corporation, a software package to determine the transformer core loss.

A. Transformer Specifications

This multiwinding planar transformer is used in a 25 W 400 kHz asymmetrical half-bridge (AHB) dc/dc converter with unbalanced secondary windings for a telecommunication system power supply [22]. The half cross section of the designed transformer is shown in Fig. 3 (only the three main power windings with turns-ratio $6 : 1 : 3$ are illustrated).

For this transformer, the magnetizing current has a dc component. The circuit shown in Fig. 2 is used to measure the core loss for this AHB transformer.

B. Core Loss Measurement Results

Following the procedure proposed in Section III, the core loss under the following three operation conditions of the AHB transformer is tested using the proposed method.

1) $V_{in} = 35 \text{ V} \sim 75 \text{ V}$, no load.
2) $V_{in} = 35 \text{ V} \sim 75 \text{ V}$, load $= 5$ A.
3) $V_{in} = 48 \text{ V}$, load $= 0 \text{ A} \sim 5$ A.

Fig. 4 provides the measurement and FEA simulation results. To illustrate the effect of dc premagnetization on the core loss in SMPS, Fig. 4(a) shows the core loss measurement and simulation results at operation conditions 1) and 2) together. Fig. 4(b) provides the results for operation condition 3).

Some typical core loss measurement oscillograms are shown in Fig. 5. It is noted that due to limited bandwidth of the power source, when the duty cycle is small, as is the case in Fig. 5(b), the voltage waveform deviates from an ideal rectangular. It is suggested that a better square wave generator might be needed in real application. The time interval between the two cursors is one switching period. Compared to the FEA simulation result, the difference between the measurement and the FEA simulation is within ±5% for all the conditions. Due to the limited bandwidth of the power amplifier, there will be some distortion in the measurement waveform. This partially causes the difference between the measurement and the FEA simulation.

It can be observed from the measurement results that the transformer core loss under a PWM waveform depends on
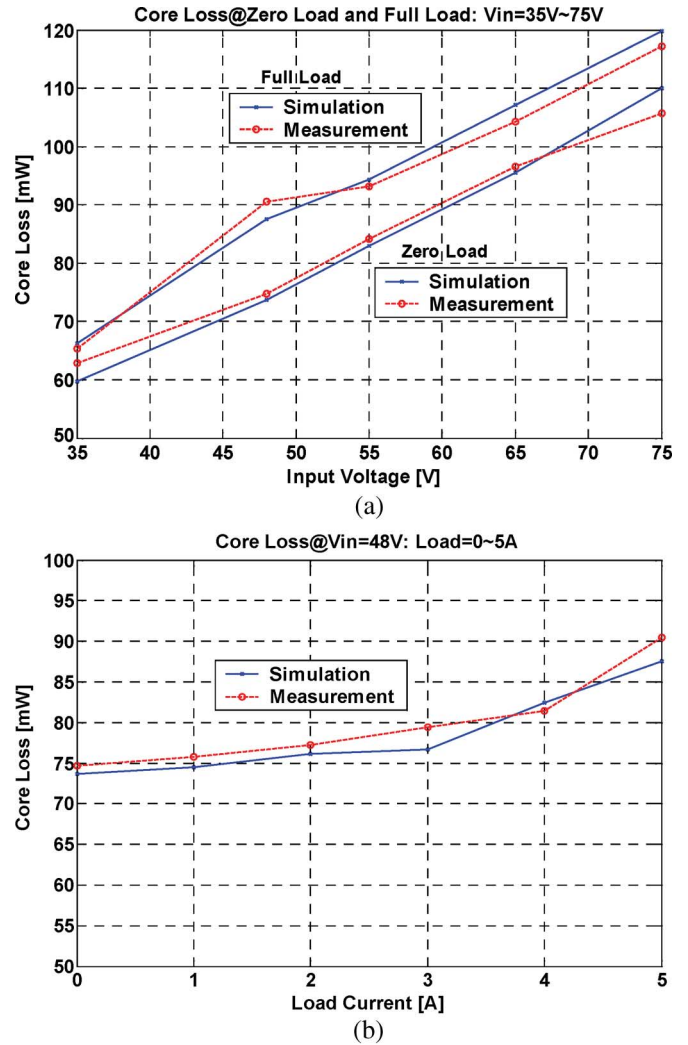

(a)

(b)

Fig. 4.   AHB transformer core loss measurement and simulation results at different conditions. (a) Operating conditions 1) and 2). (b) Operating condition 3).

the duty ratio. When this ratio of the switching converter decreases from 50%, the core loss increases. The transformer core loss will increase when the dc magnetizing current in the transformer is increased even though the voltage-second product remains the same. It is a function of both the dc and ac flux densities as well as the switching frequency.
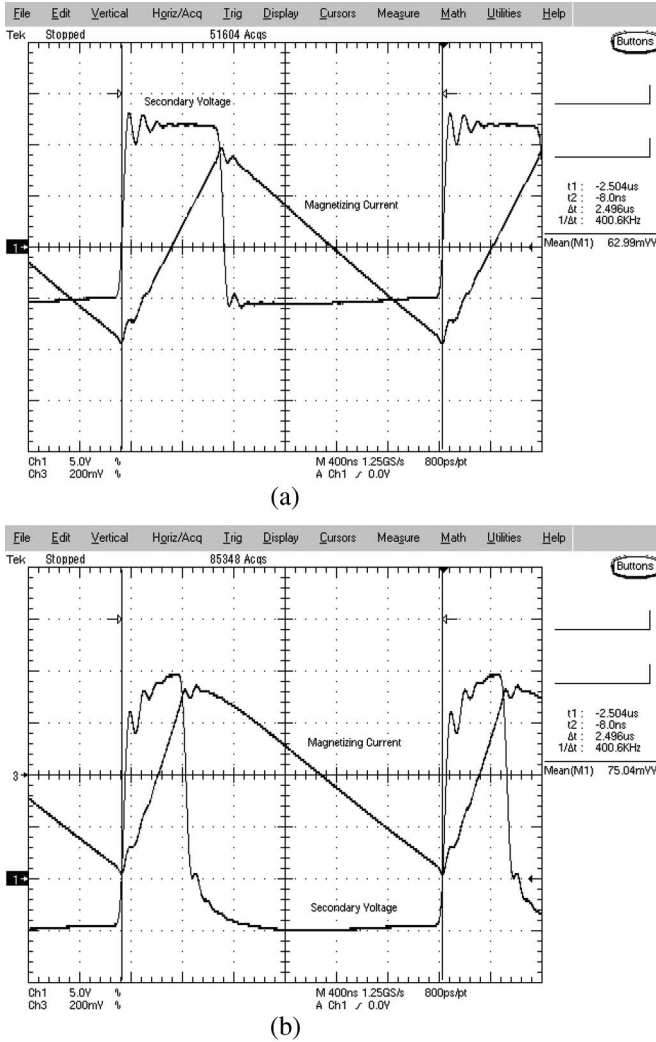
Fig. 5. Oscillograms of secondary voltage and magnetizing current: (a) $V_{\text{in}} = 35$ V; (b) $V_{\text{in}} = 48$ V.

V. ERROR ANALYSIS

To ensure that measurement of the core loss for the AHB transformer using the proposed scheme is accurate and valid, it is necessary to analyze various error sources in the measurement and obtain measurement accuracy.

From (2), the error sources in the core loss measurement are identified as: 1) the voltage measurement error; 2) the error from the current sensing resistor tolerance; 3) the winding turns-ratio error; 4) and the error introduced to the time delay between the voltage and magnetizing current in the measurement.

A. Voltage Measurement Error

The digital oscilloscope, Tektronix TDS5054, has 8-bit ADC. Because the averaging acquisition mode is used in the measurement, the effective sampling resolution can be as high as 11-bit. Both the digitizing and linearity errors in the measurement are $\pm 1/2$LSB for full scale. Combining them together, we can assume the voltage measurement error is ± 1LSB of 11-bit under full scale.
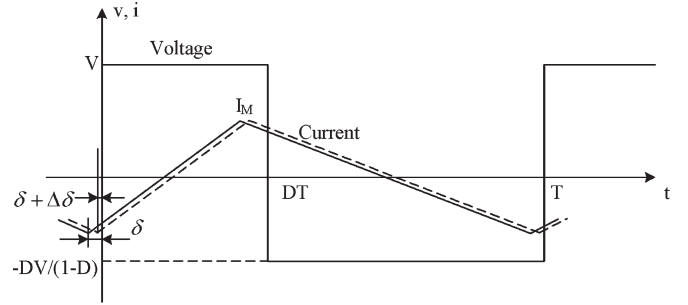

Fig. 6. Ideal and typical transformer voltage and magnetizing current waveforms in SMPS.

In the measurement, location of the peak value of the voltage waveform in the digital oscilloscope scale will decide voltage measurement accuracy. For the worst case scenario, the peak value is kept above 10% (due to asymmetry of the PWM waveform) of the full scale. Therefore, for the vertical scale, the voltage measurement error at worst is less than 0.489%.

The four measurement channels in the TDS5054 digital oscilloscope share one ADC. The digitizing error can be considered the same for each channel. From (2), the relative error on the core loss power due to the voltage measurement error in the worst case can be calculated by (4), which is less than 1%

$$\frac{\Delta P_{\text{Core}}}{P_{\text{Core}}} = \left| \frac{\frac{N_{\text{p}}}{N_{\text{s}}} \frac{1}{N} \sum_{i=1}^{N} v_{1i}(1 \pm 0.489\%) \cdot \frac{v_{2i}(1 \pm 0.489\%)}{R_{\text{Sense}}}}{\frac{N_{\text{p}}}{N_{\text{s}}} \frac{1}{N} \sum_{i=1}^{N} v_{1i} \cdot \frac{v_{2i}}{R_{\text{Sense}}}} - 1 \right|$$
$$= 0.98\%. \tag{4}$$

B. Winding Turns-Ratio Error

As explained in the key steps of the proposed measurement, calibration on the winding turns-ratio will help eliminate the resulting error. In the turns-ratio calibration, the peak voltage is kept close to the full scale of the oscilloscope. The digitizing error in the measurement for the turns-ratio calibration can be considered as ± 1LSB of the ADC, which is $1/2^{11} = 2^{-11}$. From (3), the relative error for the turns-ratio calibration is less than 0.1% at worst. Therefore, the error due to winding turns-ratio is neglected.

C. Tolerance of Current Sensing Resistor

The metal film resistor used in the core loss measurement has tolerance of $\pm 1\%$. Theoretically, 1% relative error is introduced into the measurement based on (2) due to this tolerance. The error introduced by the resistor parasitic inductance is considered in the next section.

D. Time Delay Error

For transformers in SMPS, waveforms for the voltage across the magnetizing inductor and current through the inductor are typically in the wave shapes shown in Fig. 6 (as shown by the solid lines) due to the core loss. The voltage across the magnetizing inductor is measured directly and the error is considered

in Section V-A. The magnetizing current is measured by the sense resistor added to the system. If this resistor were an ideal resistor without parasitic inductance, the voltage across the component will have the same shape as the magnetizing current. Unfortunately, since the resistor has parasitic inductance, the measured current will have phase-shift. This is the main reason for time delay that causes measurement error. Other sources of time delay may include the difference between oscilloscope probes of different channels in the measurement. However, they are very minor compared with the current sensing resistor. In calculating the time delay error, first we need to obtain δ, which is the time delay between the magnetizing voltage (measured at winding terminals) and the magnetizing current (measured through voltage across current sensing resistor) using (6) for different measurement conditions. We can use the P_{Core} value from either the measurement value or FEA simulation value as they are very close to each other. $\Delta\delta$ value is related to the current sensing resistor and can be calculated using (10) in this paper. With these two parameters known, we can calculate the time delay error using (9).

Defining the time delay between the voltage and the magnetizing current to be δ as illustrated, from (2) we can derive the following equation for core loss power over one switching period as: (details of the derivations are in Appendix)

$$P_{\text{Core}} = V \cdot I_{\text{M}} \cdot \frac{2D\delta T - 2D^2\delta T - \delta^2}{(1-D)^2 DT^2} \qquad (5)$$

where V is the amplitude of the positive part of the voltage waveform; I_{M} is the amplitude of the magnetizing current; D is the duty ratio of the PWM waveform; T is the switching period of the circuit; and δ is the time delay between the voltage and magnetizing current.

As the magnetizing current is sensed by a resistor, (5) can be written as

$$P_{\text{Core}} = V \cdot \frac{V_{\text{R}}}{R_{\text{Sense}}} \cdot \frac{2D\delta T - 2D^2\delta T - \delta^2}{(1-D)^2 DT^2}. \qquad (6)$$

If some error $(\Delta\delta)$ is introduced into δ, the magnetizing current waveform will be shifted away from the actual waveform in the time axis, as illustrated by the dashed line in Fig. 6. The increment of the core loss power due to this error can then be calculated as

$$\Delta P_{\text{Core}} = \frac{\partial P_{\text{Core}}}{\partial \delta} \Delta\delta. \qquad (7)$$

From (6) and (7), the relative error of core loss power due to $\Delta\delta$ can be derived as

$$\left| \frac{\Delta P_{\text{Core}}}{P_{\text{Core}}} \right| = \left| \frac{2DT - 2D^2T - 2\delta}{2D\delta T - 2D^2\delta T - \delta^2} \right| |\Delta\delta| \qquad (8)$$

δ is usually much smaller than DT and D^2T. And (8) can be approximated as

$$\left| \frac{\Delta P_{\text{Core}}}{P_{\text{Core}}} \right| \approx \left| \frac{\Delta\delta}{\delta} \right|. \qquad (9)$$

It is clear that relative error is very sensitive to small δ values. If a large time delay error is introduced in measurement, the relative error will become big. Therefore, care should be taken to reduce the time delay error caused by various sources.

The sources for this error $(\Delta\delta)$ could be: poor frequency response of the current sensing device; parasitic inductance picked up by the measurement circuit; and trigger jitter and delays for different channels of the oscilloscope. For the digital oscilloscope used, the trigger jitter is typical at 8-ps and can be ignored. The deskewed method can be used to further reduce the oscilloscope measurement error. The inductances associated with the current sensing resistor and picked up by the measurement circuit loop are the two important sources for the phase delay error. Since parasitic inductance due to the flux picked up by the measurement circuit is proportional to the area enclosed by the circuit, one useful method is to reduce this area by shortening the connection wire length. It is a good practice to make lead connection of the current sensing resistor as short as possible. Another important point for accurate measurement is that in measuring the magnetizing current, the oscilloscope probe should be placed as close to the resistor body as possible.

From the measurement result by impedance analyzer, the phase shift of the current sensing resistor is less than $0.01°$ at 400 kHz, less than $0.02°$ at 800 kHz and $0.05°$ at 2 MHz. If we use the phase shift at 400 kHz and transfer it to delay time for 400 kHz, the time error can be calculated as

$$\Delta\delta = \frac{0.01°}{360°} \times 2.5 \ \mu s = 69.5 \ \text{ps}. \qquad (10)$$

As the relative error depends on both the time delay (δ) and the error $(\Delta\delta)$ introduced into δ, its value will change for different operation conditions. For example, for the AHB transformer core loss at $V_{\text{in}} = 48$ V, δ is calculated as 5.9 ns from (6). With $\Delta\delta = 69.5$ ps, we can obtain relative core loss error less than 1.2% using (9). This calculation procedure can be applied to other operation conditions and yield relative error of less than 2.5%.

Using the above result, the relative error for three measurement conditions is calculated and listed in Table I, which shows that the measurement accuracy using the proposed scheme is within 5%.

VI. CONCLUSION

A new and practical scheme to measure transformer core loss under actual operating conditions in high frequency SMPS is proposed. It is simple and easy to implement. It can also consider the impact of dc magnetization current. Measurement results are presented for a planar transformer operating in a high-frequency dc/dc power converter. Detailed error analysis

TABLE I
RELATIVE ERROR ANALYSIS RESULTS IN MEASURING AHB
TRANSFORMER CORE LOSS FOR THREE OPERATING CONDITIONS

Condition	Voltage Measurement Error	Resistor Tolerance	Time Delay Error	Total Error
V_{in}=35V	1%	1%	<2%	<4%
V_{in}=48V	1%	1%	<2%	<4%
V_{in}=75V	1%	1%	<2.5%	<5%

is provided for the proposed core loss measurement scheme. It shows that the proposed scheme can provide accurate transformer core loss measurement for SMPS. With the proposed method, the core loss can be measured quickly and accurately. This will help to minimize core loss in practical transformer and power converter design.

APPENDIX

TRANSFORMER CORE LOSS CALCULATION FOR SMPS

From (2) and Fig. 6, we can divide the integration into four subperiods over one switching period as: $[0, DT - \delta]$, $[DT - \delta, DT]$, $[DT, T - \delta]$, $[T - \delta, T]$. The transformer core loss is calculated as

$$
\begin{aligned}
P_{\text{Core}} \\
= \frac{1}{T} \Bigg\{ &\int_0^{DT-\delta} V \cdot I_{\text{M}} \cdot \left[\frac{2}{DT}(t+\delta) - 1 \right] dt \\
&+ \int_{DT-\delta}^{DT} V \cdot I_{\text{M}} \\
&\times \left[1 - \frac{2}{(1-D)T}(t - DT + \delta) \right] dt \\
&+ \int_{DT}^{T-\delta} \frac{VD}{1-D} \cdot I_{\text{M}} \\
&\times \left[1 - \frac{2}{(1-D)T}(t - DT + \delta) \right] dt \\
&+ \int_{T-\delta}^{T} \frac{VD}{1-D} \cdot I_{\text{M}} \cdot \left[\frac{2}{DT}(t - T + \delta) - 1 \right] dt \Bigg\}.
\end{aligned}
$$
(A-1)

By integration, the following result can be obtained for the transformer core loss in SMPS:

$$
P_{\text{Core}} = V \cdot I_{\text{M}} \cdot \frac{2D\delta T - 2D^2\delta T - \delta^2}{(1-D)^2 DT^2}.
\tag{A-2}
$$

As the magnetizing current is measured by a sensing resistor, (A-2) can be rewritten as

$$
P_{\text{Core}} = V \cdot \frac{V_{\text{R}}}{R_{\text{Sense}}} \cdot \frac{2D\delta T - 2D^2\delta T - \delta^2}{(1-D)^2 DT^2}.
\tag{A-3}
$$

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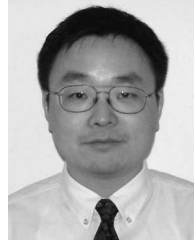
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