A practical technique to measure transformer losses in high frequency SMPS

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Abstract

All electronic gadgets today employ high frequency Switch Mode Power Supplies. As the number of such gadgets is increasing rapidly, the focus is on improving the efficiency of power conversion and better utilization of energy. It is always a challenge for the practicing power supply engineers to exactly apportion and evaluate the total losses in the power converters. While it is relatively easy to compute and measure the power losses in semiconductor devices, it is practically difficult to measure the power losses in magnetic components. The data sheets provided by the manufacturers, for core loss and thermal resistance, do act as a starting point in the design stage but, always the designers felt the need to know the exact power loss in the ferrite high frequency transformers and their thermal resistance values. The data sheets provide thermal resistance values only for the cores and it is necessary to quantify value for a fully wound transformers. Copper losses in high frequency transformers are very significant and too involved to compute. In this paper, we propose a simple and practical technique, which precisely depicts the actual power loss in a high frequency ferrite core transformer. With this technique various ferrite cores can be characterized for their hysteresis losses at varying flux densities and frequencies. We can also precisely compute the thermal resistance of the Ferrite transformers.

Keywords High frequency transformer · Ferrite core loss · Measurement of losses · SMPS · Practical power supply design

1 Introduction

Electronics has become an indispensable part of all facets of human life. Entertainment, communication, lighting, health care, information, transport etc. are all invariably linked to utilizing electrical energy. However, the electric supply from the utility grid is in the form of AC and the electrical gadgets need a stable DC supply for their operation. This calls for power supplies which take AC and convert to a stable DC power to the target equipment [1].

In view of the product design requirements, compactness, in terms of size and volume also has become a major specification in the product. This has necessitated the adoption of switch mode power converters which operate at high frequencies [2].

A general purpose AC to DC switch mode converter is detailed below.

AC–DC converter has in its front end a typical full bridge rectifier, which converts the incoming AC mains voltage into a unidirectional time varying voltage at twice the mains frequency. This is followed up by a large high voltage bulk reservoir capacitor which filters out the time varying voltage to an unregulated steady DC voltage. Typically, for an input AC voltage of 230 V AC, the rectified DC voltage at the reservoir capacitor would be about 325 V DC. However, this voltage has some ripple content, based on the reservoir capacitor value and the load current drawn from this source. In addition, the DC voltage would be unregulated and varies with mains voltage. For an input voltage range of 180 V AC to 264 V AC, The DC voltage can be in the range of 225 V DC to 390 V DC. The power

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converter is supposed to handle this large input DC voltage variation and deliver a rock steady DC output voltage in the range of say 5 V DC, or 12 V or 24 V or 48 V as needed by the end equipment and also should provide galvanic isolation from the input mains supply for safety and EMI/EMC related issues [3].

As a next step, the unregulated DC voltage is connected to a transformer with a ferrite core and a semiconductor switch, either a power transistor or a power MOSFET. The transformer would be capable of operating at high frequencies (around 100's of kHz) and the semiconductor switch is driven ON and OFF at high frequency. Thus, the circuit acts as a chopper at high frequency and the transformer sees a high frequency AC voltage. Secondary of the transformer, which is electrically isolated from the primary, is rectified with fast recovery rectifier diodes and then connected to a low pass filter comprising of a ferrite core inductor and electrolytic capacitors. Output of the low pass filter is a DC voltage which can be regulated to a precise value, by controlling the ON or OFF time of the semiconductor power switch [3].

A stable reference, high gain error amplifier, sampled output voltage together generate an error voltage which controls the ON or OFF time of the power switch and is transferred to power switch with a galvanic isolated drive. Thus, the AC DC converter delivers a stable lower DC voltage to the end appliance with total isolation from the high voltage utility mains [3].

2 Problem statement

As discussed earlier, automation has created a sudden demand for the power converters and also for the electrical energy. However, the generation of electricity which is predominantly by fossil fuels has impact on the environment [1]. Hence, the focus is on improving the energy efficiency of electronic gadgets so that the available resources are best utilized. The importance of discussion on the power losses and especially in transformers can be well appreciated by referring to [4, 5], as it discusses the power loss in an electronic power transformer as against a conventional power transformer.

Having said about the efficiency improvement, the first element in the chain of a product is the power converter sitting right at the front end and it would be critically looked at for efficiency improvement. Front end converter handles the complete power requirement of the electrical gadget and therefore needs to be highly efficient.

The designers, while at the drawing board stage would allocate efficiency figure for the power converter as a part of the product specification. Based on the output power requirement, the efficiency apportionment puts up a

figure for the total power loss that can be allowed in the power converter. The main elements in the power converter where the power loss occurs are

- (a) The input rectifier bridge,
- (b) The high frequency power transformer,
- (c) Power MOSFET,
- (d) Output filter inductor,
- (e) Output rectifier diodes.

Miscellaneous losses such as in the bias circuits, house-keeping circuits, filter capacitors, PCB track losses etc. can be neglected as they form a small percentage of loss and their improvement may not add much to the target. Among them the losses in the semiconductor devices are primarily computable, with the available math equations [6]. They can be easily verified also, by mounting them on heat sinks which are pre calibrated for their thermal resistance R_{th} . Thermal resistance is defined as the temperature rise above ambient for every watt of power dissipated and has the units $^{\circ}\text{C}/\text{watt}$ [7]. By mounting the semiconductors on the heat sinks with known R_{th} and measuring the temperature rise of the heat sink directly measures the power loss in the device.

The power loss measurement of the ferrite transformers/inductors is not that simple [8–10]. Thermal resistance of the cores as depicted from the data sheet cannot be adopted as it is. Data sheets provide thermal resistance for only the core and not for the complete transformer [11, 12]. A wound transformer has a different property altogether.

Also, it is a practical problem that when the power converter is ready and evaluated for its losses, it does not match 100% with the apportioned power loss on all components and the exact loss in the magnetic components becomes a debating point. Losses in ferrite transformers occur mainly due to

- (a) Core loss or hysteresis loss and
- (b) Copper losses in the winding.

Both losses are very complicated, non-linear in nature and too involved to theoretically compute [13]. Core loss depends on the frequency of operation, flux swing, geometry of the core and also the grade of ferrite. Likewise copper losses depend on Skin effect, proximity effect, type of topology of converter and the construction of the transformer layers [14].

Therefore, it is matters of importance to at least quantify the exact amount of core loss in ferrite transformer and also characterize the thermal resistance of the transformer. With this, half of the total transformer loss can be precisely estimated. If the accurate thermal resistance value can be

obtained for the transformer, then we can also indirectly compute the copper loss.

We propose in this paper a technique to precisely measure the core loss in the designed transformer and also estimate the thermal resistance of the transformer. With these two, exact losses in the transformer can be measured.

Further, this technique can also be used to create a data base for the core losses in various cores at different frequencies and a variety of flux density swings.

3 Methodology

The proposed methodology is very simple and easily implemented in the real life applications.

Hysteresis losses of a ferrite core are due to the work done during every switching cycle in magnetizing and demagnetizing the core. The same is well discussed widely under the heading called B-H curve [15]. Core losses depend on frequency of operation, flux density swing, core volume and core material grade. Lot of literature is available from the manufacturers' data sheets.

When any transformer primary is excited with alternating current (AC) voltage and the secondary is left unconnected, then the current drawn from the primary source voltage indicates the core loss. This is very simple in case of low frequency (50 Hz) transformer with sinusoidal voltage input. Whereas, in high frequency applications with non sinusoidal input voltage and varying ON and OFF times, the measurement of core loss is not that simple. Therefore a simple technique to measure the core loss in a high frequency transformer with non sinusoidal excitation and varying ON or OFF times is essential in practical applications. The following methodology suggests a solution for the above problem. Most important of this technique is that the actual core loss is obtained in the in situ conditions for the exact field conditions.

Let us consider a power converter set up of Full bridge configuration as shown in Fig. 1 wherein the transformer is fed with bidirectional voltage in every cycle and does not contain any DC component due to presence of a coupling capacitor in series with the transformer primary [13].

Switches, S1 to S4 are driven by a PWM controller and the output is tightly regulated with a feedback control loop. A working model of say 200 or 300 W power converter is fabricated. Input power source, a variable DC voltage is connected to the Bridge converter via a power meter, which accurately measures the input power drawn. Yokogawa make WT210 power meter was used in the present context which can precisely detect, even a minute change in the input power. By varying the input voltage or the output load, we can vary the duty cycle (ON and OFF

Fig. 1 Setup for measurement of core loss

times of the switches S1 to S4) which would impinge input voltage across the primary of the transformer for varying time duration. The aim of the experiment is to evaluate the core loss in the power transformer for a given input/output condition. Therefore, One more transformer identical to the one used in the full bridge converter is fabricated (DUT) and the primary of DUT is connected in parallel to the existing transformer with a switch. Initially the switch is left open and so also all the secondaries of DUT are also unconnected. The following steps detail the core loss measurement procedure:

- The designed power converter unit is powered ON at a nominal load at the output and with a stable DC input voltage.
- Input DC current and DC voltage is recorded. The unit is kept switched ON for about 30 min to stabilize all the parameters. Input power is also recorded and so also the output power.
- DUT, exactly similar to the one used in the power converter is wound and kept ready.
- Now, the power converter is switched OFF and the primary of the DUT transformer is connected in parallel to the working unit, with all secondaries of DUT left unconnected.
- The unit is powered ON again with exactly same input and output conditions as in step (a).
- After a warm up of about 30 min, again the input voltage and input current are recorded. There would be increase in the input current, and hence the input power now, for the same output power.
- The increment in the input power is entirely due to the core loss of the DUT transformer. This is noted as the core loss for the given switching frequency and the flux density swing determined by the ON time of the switches S1 to S4.

Fig. 2 Waveform across DUT

Fig. 3 Core loss measurement setup

- (h) The flux swing in DUT is easily computed by monitoring the voltage impinged across the primary of the DUT transformer (which is same for the working unit also) and recording the time duration.
- (i) We know the core cross section area and hence we can easily arrive at the flux density swing, because we know, voltage across the transformer, its primary turns and also the core area and the time duration for which the voltage is applied. The typical waveform across the DUT is shown in Fig. 2 and the test set up in Fig. 3
- (j) Thus, for a given transformer, the core losses are estimated in situ.
- (k) The DUT is kept ON in the above condition, till the thermal equilibrium is reached and then the temperature of the DUT is recorded.

- (l) Since we know the power dissipated in DUT as per step (g), and also the rise in temperature of DUT above ambient, the accurate R_{th} of the DUT is also arrived at.

Thus at one go, we can compute the core loss and the thermal resistance of the DUT. If the full bridge converter is designed to be of a variable frequency switcher, where in its switching frequency varies with line/load, then by varying the load and line and repeating the above steps, we can easily characterize the core losses for DUT for different frequencies and different flux density swings.

4 Discussion

Calculation of the power loss in a high frequency transformer is very involved and a lot of study has been done and reported [8, 16]. For a normal practicing power supply manufacturers and engineers, the theoretical part appears to be too complicated to follow. However, it is very important to know the actual losses in all the power components of a converter, in order to improve the product performance, reliability and product design. Therefore a practical technique to assess the losses in magnetic components is well appreciated and need of the hour.

Transformer losses include hysteresis loss, eddy current loss, and copper loss. All these are dealt in detail from a very long time in theory [17]. The idea of choosing a power converter with half bridge or full bridge configuration in this proposal is due to the fact that the power transformer is auto reset and the flux swing is truly bi directional. This is so because the transformer is coupled through a DC blocking capacitor. Under such condition, additional transformer, if connected across the working transformer has no effect not on the normal performance of the power converter. Had a forward converter or a fly back configuration is taken for power converter, then addition of an extra transformer would affect the performance of the converter and the results obtained are erroneous. Further, if the chosen power converter is powered ON and delivering the load, the input current drawn from the DC input would have accounted for all the other losses such as MOSFET switching losses, snubber losses and all other parasitic losses. Under such conditions, if only a transformer primary is paralleled to the already working unit, then the increase in input current exactly depicts the added transformer core loss at the operating frequency and magnetic flux density. Flux density swing is computed by the equation

$$\text{Flux density swing} = V_{in} * T_{on} / N_p * A_e \quad (1)$$

Table 1 Losses for EE25 double stack transformer at various flux densities and frequencies

EE 25 double stack N87		
ΔB (mT)	Frequency (kHz)	Loss (mW)
100	100	360
200	100	1350
400	100	4830
100	200	600
200	200	1800
400	200	8100

Table 2 Losses for EE42 single stack transformer at various flux densities and frequencies

EE 42 single stack N87		
ΔB (mT)	Frequency (kHz)	Loss (mW)
100	100	772
200	100	3280
320	100	6800
100	200	900
200	200	4500
450	200	17,000

where V_{in} is the applied voltage to the transformer, T_{on} is the time for which the input voltage is impressed, N_p is the primary number of turns and A_e is the cross sectional area of the core. The equation above is the standard transformer equation depicting the relation between applied voltage and change in magnetic flux linkages.

Since the core losses are non-linearly related to flux swing and frequency of operation, it is in the best interest that the power converter chosen is of variable frequency type. This greatly simplifies in evaluating the core loss at different frequencies and different Flux density levels.

We have conducted the characterization of two ferrite cores of geometry EE42/21/15 core and also a double stacked EE25/13/7 cores. The results are tabulated. The following tables show the core loss characterization of two types of cores, EE25/13/7 double stack and EE42/21/15

obtained through the proposed scheme. Core losses for various flux density swings at 100 kHz and 200 kHz are recorded and presented in Tables 1 and 2.

5 Experimental verification and validation

In order to demonstrate the efficacy and usefulness of the scheme proposed, a practical fly back converter of 150 W was designed, fabricated and evaluated to accurately assess the power loss in the power transformer.

The chosen topology for the practical converter was a soft switching fly back configuration due to the fact that it has only a single transformer and fewer components. Input voltage chosen was 190 V DC. Output voltage was selected as 150 V DC so that the output current is only 1.0 A and hence does not add up to unnecessary PCB track losses and diode losses. The configuration of the fly back was a soft switching Boundary Control Mode BCM control topology as cited in [18]. BCM allows the diode losses to be at minimal. BCM control IC FAN7527 was used as the controller. This IC senses the complete energy transfer point by measuring the zero current in the secondary and again turning ON the primary Mosfet switch. BCM control automatically has the switching frequency as a variable for various Line and load conditions. The Block diagram of the flyback converter and the photograph of the fabricated unit are shown in Figs. 4 and 5.

Specifically, the soft switching converter with BCM controller is chosen, so as to minimize all the losses which makes it easy to concentrate on the transformer loss alone.

In such configuration, the losses are only due to

- Input bridge rectifier,
- MOSFET switch,
- Output rectifier, and
- The main power transformer.

Bias and drive losses are catered to by a separate bias supply. Such a unit was designed and fabricated and

Fig. 4 150 W flyback converter block diagram

Fig. 5 150 W flyback converter

tested. The test data such as input power, output power is recorded for a stable DC input voltage. With this data, total losses were arrived at.

The converter designed has the following specifications

1. INPUT voltage: 190 V DC
2. OUTPUT voltage: 150 V DC
3. Output current: 1.0 A
4. Switching Frequency anticipated to be around 150 kHz (Since the BCM flyback is a variable frequency converter)
5. Configuration: soft switching BCM

The designed transformer has the following characteristics.

- (a) Core: ferrite EE25/13/7 double stack
- (b) Material: N87
- (c) Primary inductance $L_p = 100$ micro henries
- (d) Primary turns $N_p = 20$
- (e) Primary wire size = 34SWG, 6wires twisted and bundled
- (f) Secondary inductance $L_s = 59$ micro henries
- (g) Secondary turns $N_s = 15$
- (h) Secondary wire size: 34SWG, 4 wires twisted and bundled
- (i) Leakage inductance $L_l = 1.9$ micro henries
- (j) Primary DC resistance $R_p = 80$ mOhms
- (k) Secondary DC resistance $R_s = 90$ mOhms

Construction was split primary and secondary sandwiched between two half primaries.

Flux swing anticipated

$$= 150V * 2.5 * 10^{-6} / (15 * 100 * 10^{-6}) = 250 \text{ mT} \quad (2)$$

In order to evaluate the exact core losses in the above transformer, a 250 W bridge converter is built to operate at 175 kHz. Input voltage and output load were adjusted to obtain 150 V DC bipolar and for 2.5 ms to be impinged across the transformer. Next, an Identical transformer as the one used in the flyback converter is fabricated and

the secondary consisting of 15 turns, is connected in parallel to the main transformer of the half bridge converter. The primary of the flyback transformer with 20 turns was left open. With this condition the half bridge converter was run for about 4 h to attain thermal equilibrium.

Under this condition, the observed increment in the input power of the half bridge converter is noted to be 1.75 W. This is the core loss in the fly back transformer at 175 kHz and 250 mT flux swing.

Simultaneously, the rise in the flyback transformer temperature was recorded as 32 °C above ambient.

Therefore, the computed Thermal Resistance of the flyback transformer is

$$R_{th} = 32/1.75 = 18.3 \text{ }^{\circ}\text{C/W} \quad (3)$$

The Flyback converter with designed transformer is then run at 150 W load continuously till thermal equilibrium and temperature rise of transformer is recorded. The obtained switching frequency was 175 kHz.

The data recorded in the practical 150 W flyback converter are as under.

1. Frequency of operation: 175 kHz
2. MOSFET ON time T_{on} obtained: 2.0 ms
3. D_s Secondary diode ON time D_{son} : 2.5 ms
4. Delay time T_d : 1.2 ms
5. Time period $T = T_{on} + T_d + D_{son} = 5.7$ ms
6. Duty cycle $D = T_{on}/T = 0.351$
7. D_s ON duty ratio = 0.439
8. Input voltage = 190 V DC
9. Input current $I_{indc} = 0.846$ A
10. Output voltage = 151.2 V DC
11. Output current = 1.0 A
12. Total losses recorded = 9.54 W (input power–output power)
13. The computed flux density swing in the practical converter is 250 mT

Temperature rise of the flyback transformer after thermal stabilization = 68 °C above ambient.

From the above data, it is estimated that in the fly back converter at 150 W, the power loss in transformer is

$$\text{Power loss} = 68/18.3 = 3.72 \text{ W} \quad (4)$$

Semiconductor components of the fly back converter such as MOSFET and diode, were mounted on calibrated heat sinks and the power loss in them is recorded to be as

$$\text{MOSFET loss} = 1.5 \text{ W}$$

$$\text{Diode loss} = 2.0 \text{ W.}$$

The input bridge rectifier loss is measured to be 1.3 W.

The current sense resistor and the output sense network together consumed about 1.3 W.

With this the total losses measured in the converter of 150 W are 6.1 W, not accounting for transformer. The total converter losses are noted to be 9.54 W. Hence, from this angle; the estimated power loss in transformer is $9.54 - 6.1 = 3.44 \text{ W}$. However, the power loss computed from the proposed scheme by measuring the temperature rise and R_{th} is 3.72 W which is in close acceptance range. Thus, the proposed method of core loss measurement is validated in a practical set up

Out of the total power loss in transformer, the core loss is recorded as 1.75 W from the proposed scheme. This concludes that the copper losses in the power transformer at 150 W are

$$\text{Copper losses} = 3.72 - 1.75 = 1.97, \quad \text{say } 2.0 \text{ W} \quad (5)$$

It is interesting to compute the increased copper losses due to proximity, skin effect losses, which can be termed as AC copper losses. It is done as follows

$$\text{Input DC current } I_{inavg} = 0.846 \text{ A}$$

$$\text{Duty cycle } D = 0.351$$

Therefore, the primary peak current

$$I_p = I_{inavg} * 2/D = 4.82 \text{ A} \quad (6)$$

Primary RMS current

$$I_{prms} = I_p * (\sqrt{D/3}) = 1.65 \text{ A} \quad (7)$$

Similarly, the secondary RMS current

$$I_{srms} = 2.46 \text{ A}$$

DC primary and secondary copper losses are

$$\text{DC copper loss} = 1.65^2 * 0.080 + 2.46^2 * 0.090 = 0.763 \text{ W} \quad (8)$$

Thus, a DC copper loss of 0.763 W is manifested to be 2.0 W of total copper loss

$$\text{Increased copper losses due to skin, proximity effect} = 2.0 - 0.763 = 1.237 \text{ W} \quad (9)$$

$$\text{Increased AC copper loss to DC copper loss ratio} = 1.237/.763 = 1.62$$

6 Conclusions

Measuring the power loss in a high frequency ferrite transformer is very essential for all power supply engineers. Improving the efficiency of the converter can be done, only if the losses occurring at various levels are known. Therefore, any effort in contributing to the loss estimation is a welcome phenomenon and this paper proposes a simple, implementable technique to practically estimate the losses in high frequency transformers.

Core loss data for two types of cores are measured and presented. Thermal resistance derived practically is also depicted. A practical 150 W flyback converter is built and evaluated for its losses and verified with the proposed technique. Quantum of increased copper loss due to skin and proximity effects is clearly arrived at.

The proposed scheme accurately arrives at the losses in a high frequency power transformer. This enables designers to try out various design options and arrive at an optimum configuration for maximizing the efficiency. The designed 150 W fly back converter achieved an efficiency of 94.06%.

Compliance with ethical standards

Conflict of interest The authors declare that they have no conflict of interest.

Human and animal rights There is no involving human participants and/or animals in this research work.

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