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Thermal Characterization of High-Frequency Flyback Power Transformer

Abstract

This paper presents preliminary thermographic measurements and a simple thermal analysis of power losses in high-frequency (H-F) transformers used in AC-DC power converters. The analysis is based on complex thermal impedance and time constant distribution. The main aim of this research was to consider quantitatively the contribution of different power losses in the H-F transformer, including fringing flux effect.

Keywords: IR thermography, power losses, H-F flyback transformer, thermal impedance.

1. Introduction

A change in magnetization of a core material requires energy to effect the process. A portion of this energy is consumed irreversibly due to physical mechanisms by which magnetic flux changes inside a magnetic material and is converted to heat. The structure of the material consists of a number of ordered regions with different magnetic orientation. These regions are called domains and are separated by transition regions – domains walls. The rotation of magnetic moments within a domain and walls as well as the movement of the walls themselves are responsible for the losses in the magnetic medium [1].

The existence of eddy current phenomenon is one of the factors responsible for losses in the core. Changes of magnetic flux inside magnetic material induce, according to Faraday's law, voltage in the core material, which in turn drives local currents perpendicular to the magnetic lines of force. These local currents flowing inside the core are called eddy currents [1].

The third category of losses in magnetic materials is the residual loss for which the origins have not been well defined yet. Total core loss is the sum of hysteresis, eddy current and residual losses as all occur within magnetic materials [1]. Reassuming: Core loss = Hysteresis loss + Eddy current loss + Residual loss

In addition to the core losses, transformers exhibit power dissipation through mechanisms associated with current flow in their windings. There are typical ohmic losses which are increased by the electromagnetic phenomena existing in the metal windings: the skin and proximity effects [1,6]. All of them strongly depend on frequency. In this paper, the next source of power losses and heat generation is observed and considered. It is the fringing flux effect. These losses are due to the increased leak magnetic flux near the air gap. This flux generates an extra power dissipated in the windings, next to core because of the additional eddy currents flowing there [1,6]. There are numerous papers on this effect, mostly on theoretical analysis and modelling. This paper is the one of the first where the measurements of temperature increase caused by all heat loss sources are presented. The final aim of the research is to extract the different power sources in the high frequency transformers including the fringing flux effect [7]. This paper presents the thermographic measurement and thermal impedance analysis of heat transfer near the air gap in the magnetic core and the neighbouring windings [8-10].

2. Core losses

The energy dissipated in the process of the core magnetisation is equivalent to the area of the *B-H* loop for a given magnetic material, and is known as hysteresis loss – Fig.1.

If the *N*-turn magnetic circuit is excited by voltage $v(t)$ and current $i(t)$ with waveforms of periodic character, the energy W dissipated over one cycle T is

$$W = \int_0^T v(t)i(t)\mathrm{d}t. \tag{1}$$
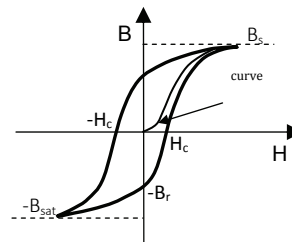

Fig. 1. *B-H* loop for typical ferromagnetic material

Substituting for $i(t)$ using Ampere's law, one gets

$$i(t) \cdot N = H(t) \cdot L_m, \tag{2}$$

where L_m – magnetic path length.
Using Faraday's law $v(t)$ takes a form:

$$v(t) = N \cdot A \cdot \frac{\mathrm{d}B(t)}{\mathrm{d}t}, \tag{3}$$

where A – cross-sectional area of a core.
Finally, the energy is expressed as

$$W = \int_0^T N \cdot A \cdot \frac{dB(t)}{\mathrm{d}t} \cdot \frac{H(t) \cdot L_m}{N}\mathrm{d}t$$

$$W = A \cdot L_m \cdot \oint H \mathrm{d}B \tag{4}$$

The integral is the area of the *B-H* loop and the term $A \cdot L_m$ is the volume of the core. One can infer from eqn. (4) that hysteresis power loss increases with the frequency of operation f as
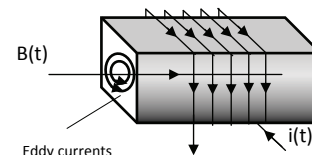
$$P_h = W \cdot f. \tag{5}$$

Fig. 2. Eddy currents in magnetic material

The mechanism of eddy current losses is presented in Fig. 2. The magnitude of the eddy currents are limited by the resistance of the core material. Power losses due to eddy currents are essentially proportional to i^2R loss arising within the core. As the induced voltage, according to Faraday's law, increases with the rate of change of magnetic flux, the eddy current loss increases directly with frequency. The impedance of the core material is not independent of frequency and often decreases for high frequencies causing further eddy current losses.
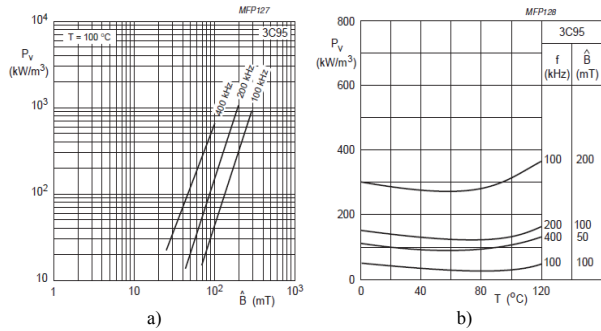
Fig. 3. Total core losses per unit volume as a function of frequency and peak flux density (a), and core temperature (b) for 3C95 core material [3]

For a given magnetic core, the manufacturers often represent the total core loss per unit volume in the form of equations such as [2]

$$P_{core} = k \cdot f^m \cdot \Delta B^n. \tag{6}$$

They also publish core data in a form of figures containing total core loss data as a function of sinusoidal excitation frequency and peak AC flux density ΔB, which can be approximated by the eqn. 6 or as a function of temperature, Fig.3.

One can note that the power density plot in Fig.3a is drawn for temperature $T = 100°C$. To account for fluctuation of losses with core temperature, eqn. (6) can be expanded into the empirical function of the form

$$P_{core} = k \cdot f^m \cdot \Delta B^n \cdot (ct - ct1 \cdot T + ct2 \cdot T^2). \tag{7}$$

The parameters are explicit for a selected magnetic material and valid only within a specific frequency range. The temperature dependant coefficients ct, $ct1$ and $ct2$ are chosen so as to the term $(ct - ct1 \cdot T + ct2 \cdot T^2)$ gave 1 at T=100°C reducing eqn. (7) to the form of eqn. (6).

3. Loss in windings

A conductive path that offers a resistance dissipates power in the form of heat when electric current flows through it. For low frequency this power loss takes the form

$$P_{DC} = I_{RMS}^2 \cdot R_{DC} \tag{8}$$

where I_{RMS} is the root-mean-square value of the current and R_{DC} is the DC resistance of windings.

The DC resistance can be expressed as:

$$R_{DC} = \frac{\rho \cdot l}{A}, \tag{9}$$

where ρ – resistivity of the winding conductor, l – the length of the winding and A is the wire cross-sectional area.

At low frequency the current is distributed homogenously throughout the wire whereas at high frequency the magnetic field resulting from current flow in windings induces voltage within a conductor, which in turn causes eddy currents to occur. These currents tend to reduce the net current density in the centre of the conductor but reinforce the main current flow at the surface – Fig.4. This specific eddy current mechanism is called the skin effect. The distance into the conductor to where the current density is 1/e of the surface one is known is the skin depth δ [4].

$$\delta = \sqrt{\frac{\rho}{\pi \cdot \mu \cdot f}}, \tag{10}$$

where μ denotes the permeability of the conductor.

The current crowding near the surface reduces the effective cross-sectional area of the conductor, hence increased power loss at high frequency.
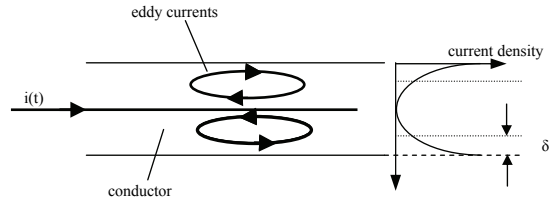

Fig. 4. High frequency current distribution

For high frequency the resulting resistance due to the skin effect can be approximated by

$$R_{AC} = \frac{\rho \cdot l}{\pi \cdot \delta(D - \delta)}, \tag{11}$$

where D is the wire diameter.

The resulting power loss in a single layer is

$$P_{AC} = I_{RMS}^2 \cdot R_{AC}. \tag{12}$$

When an AC-current carrying wire is brought together with another AC-current carrying wire, the resulting magnetic field causes the distribution of current density to be altered within both conductors. This phenomenon is known as the proximity effect. At high frequencies this effect is exacerbated and increased power loss occurs. In magnetic devices comprising many turns and layers the proximity effect is particularly evident as it generates undesirable heating.

Another mechanism behind losses in wirings is present in magnetic circuits with gapped ferromagnetic cores where fringing flux around the air gap induces excess eddy currents in the windings which cause localized heating – Fig.5 [7]. This effect can be easily observed by using IR thermography.
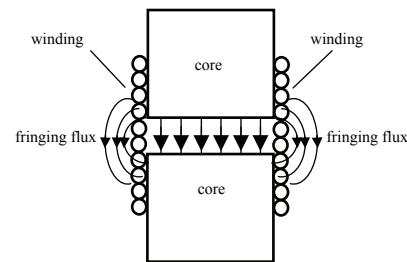

Fig. 5. Fringing flux at the gap

4. Theoretical ascertainment of losses in flyback transformer
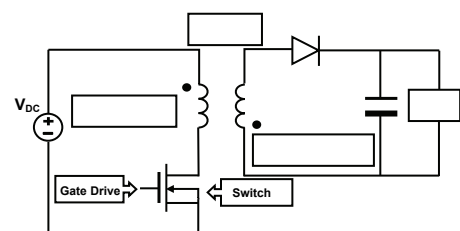
The flyback topology is shown in Fig.6.

Fig. 6. Circuit diagram of a flyback converter

Flyback transformers operate only in the first quadrant of the B-H loop of a given magnetic material, Fig. 7. To ascertain core losses the operating frequency and the maximum AC flux density ΔB are needed.
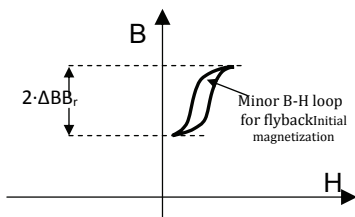

Fig. 7. Typical B-H loop for flyback converter

Core losses can be estimated by the utilization of power loss density graphs provided by the manufacturer of a given core (see Fig.3a) or the empirical function presented by eqn. (7).

The practical realization of 100 W flyback converter has been done with PQ32/30 3C95 ferrite core operating at frequency of 63.5 kHz with the maximum AC flux density ΔB swing of 135 mT and core temperature for full load $T = 62°C$. The estimate of core losses (power density) is as follows:

$$P_{core(3C95)} = 92.166434 \cdot f^{1.045} \cdot \Delta B^{2.44}$$
$$\cdot (1.332363 - 0.00794 \cdot T + 0.000046 \cdot T^2),$$

$$P_{core(3C95)} \approx 74 \ \frac{kW}{m^3}.$$

The volume of PQ32/30 core is about $12.5 \cdot 10^{-6}$ m^3 hence power dissipated in the core:

$$P_{core(diss)} \approx 0.925 \ W$$

It has to be borne in mind that the manufacturer's plot is obtained for sinusoidal excitation rather than for rectangular voltage waveforms hence the obtained core loss for the flyback transformer in Fig.6 will somewhat differ from the actual value. The precise estimation of power losses for the case described requires more advanced mathematical methods [5].

Copper loss within the windings can be determined to a certain degree of accuracy with the knowledge of the dimensions of transformer windings and their configuration.

Calculation of the DC copper loss requires the length and cross-sectional area of the wire as well as the *RMS* values of transformer currents to be known. Subsequent substitution into eqn. (8) and eqn. (9) yields the DC power loss in the windings.

The *RMS* values of the primary and secondary current waveforms for the constructed converter are about 0.76 A and 9.01 A, respectively (Fig. 8).
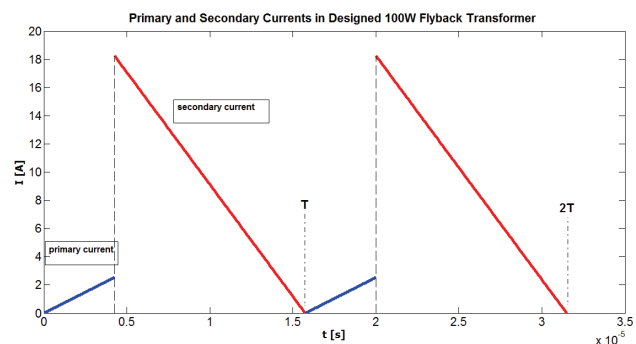

Fig. 8. Primary (blue) and secondary (red) current waveforms for the designed flyback transformer

The average length of turn for PQ32/30 coil former is $l = 66.7$ mm and the turn ratio of the transformer $\frac{Np}{Ns}$ equals $\frac{30}{4}$. The selected wire cross-sectional area is 0.6362 mm^2 for the primary side, and 4.3641 mm^2 for the secondary one which was built using a multi-strand wire. The resistivity of copper for a winding working at a temperature $T_w = 67°C$ is about $1.989 \cdot 10^{-8}$ Ω·m. Multiplication and subsequent substitution of the above data into eqn. (9) results in $R_{DCp} = 6.256 \cdot 10^{-2}$ Ω for primary winding and $R_{DCs} = 1.216 \cdot 10^{-3}$ Ω for secondary one. The corresponding power losses are as follows:

$$P_{DCp} = 0.76^2 \cdot 6.256 \cdot 10^{-2} \approx 0.0362 \ W,$$

and

$$P_{DCs} = 9.01^2 \cdot 1.216 \cdot 10^{-3} \approx 0.0987 \ W.$$

The skin depth for 63.5 kHz increases copper losses to:

$$P_{skin\_p} \approx 0.0411 \ W,$$

$$P_{skin\_s} \approx 0.0992 \ W.$$

The secondary winding comprises a number of 0.63 mm diameter wires to reduce high frequency loss due to the skin effect. For multiple layer windings the proximity effect is a major contributor to power loss. This is particularly evident in high frequency designs as the power loss increases rapidly with frequency. The complexity of winding geometry as well as difficulty of modelling interactions between wires in windings makes the estimation of the proximity loss in windings very problematic. There are numerous methods that can be utilized to achieve this goal. One of them is the Dowel method where theoretical formulas are derived for the factors by which the DC resistance is multiplied to get corresponding AC values [6]. The results are presented in a graphical form.

Another approach to compute the proximity loss is the use of numerical methods, the finite element method etc., through which the losses can be found for any given winding configuration with a required degree of accuracy. However, these methods are usually very time-consuming.

5. Thermal characteristics

The thermal characterisation of the flyback transformer was performed using dynamic thermographic measurements. The experiment started at room temperature about 22°C when the transformer was off. In thermal analysis, as long as the process is linear, one always considers the temperature surplus over the ambient. The AC-DC converter was switched on. The heating pulse was assumed to be the power step. The entire thermal process lasted more than 1 h. An exemplary thermal image extracted from the sequence recorded is presented in Fig. 9.
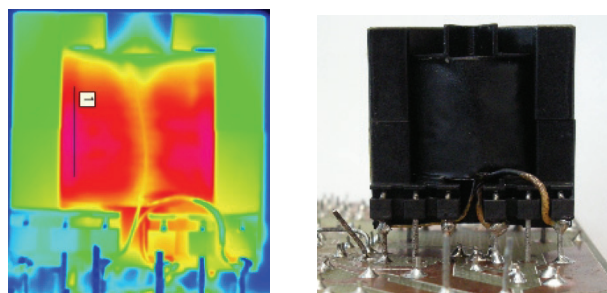

Fig. 9. Flyback transformer operating in AC-DC converter – thermal and visual images

1-dimensional heat transfer from the heat source to the ambient was presumed. The point on the transformer's surface located over the air gap, which itself is positioned in the central limb of the core, was chosen as a measurement point. It is because the maximum temperature was observed in the vicinity of the air gap (Fig. 10) although temperature distribution is almost certainly governed by a number of factors. In consequence, one can conclude that it was an extra power loss there due to the fringing flux effect. This paper is one of the first showing thermography measurements of transformer's temperature near the air gap where the leak magnetic flux increases power dissipated.
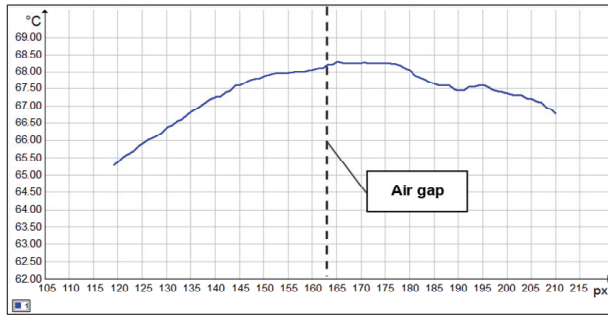

Fig. 10.   Temperature distribution along line 1 shown in Fig. 9

Temperature rise is shown in Fig. 11. The power losses in the transformer were estimated at about 2.5 W by using the electrical measurements of voltages and currents on primary and secondary sides. It corresponded to about 97.5% efficiency which is a satisfactory result.
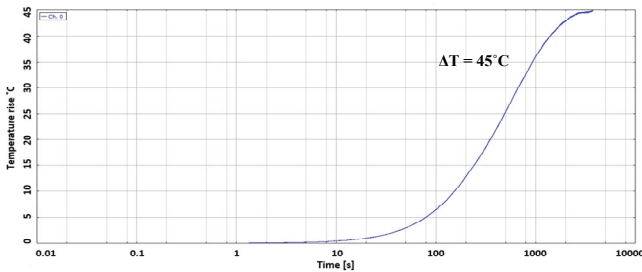

Fig. 11.   Heating curve after switching on AC-DC converter

Using the concept of thermal impedance, one can use Nyquist plot which represents the thermal impedance in frequency domain – Fig. 12 [9,10]. The important result derived from Nyquist plot is that so-called transfer thermal impedance was measured. It means that the plot exceeds −90° phase angle between temperature and the power dissipated in the transformer. It is for $\omega \approx 0.012$ 1/s and denotes that the selected measurement point was located at a certain distance from the heat source. The heat source is distributed over the entire structure of the transformer (core and windings) and its major component is definitely somewhere deeper.
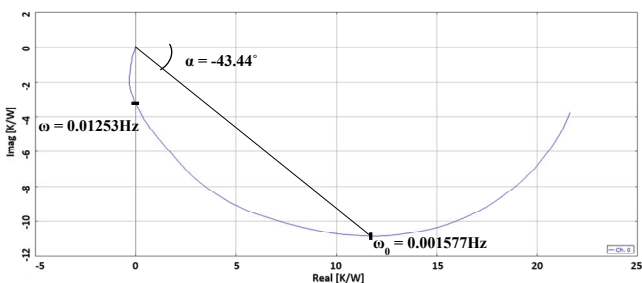

Fig. 12.   Nyquist plot of the transfer thermal impedance

The thermal structure of the transformer can be modelled as a single time constant equivalent circuit. It is because the Nyquist plot is almost a semi-circle, and the minimum is for almost −45° phase angle [10].
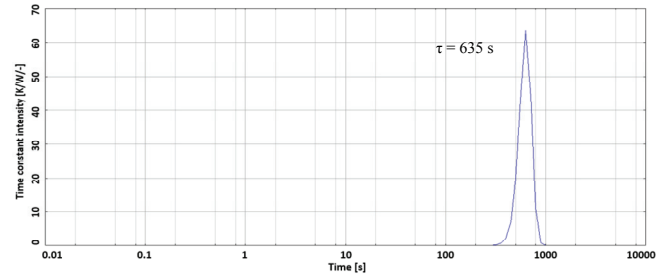

Fig. 13.   Time constant spectrum

From the complex thermal impedance, using deconvolution one can derive the time constant distribution (Fig. 13) [8-10]. Again one can see, that the transformer can be considered as a single time constant thermal structure. This thermal constant corresponds to power dissipation to the ambient by natural convection.

Let's assume that the transformer is a cylinder of 26 mm diameter consisting of different layers as it is shown in Fig. 14.
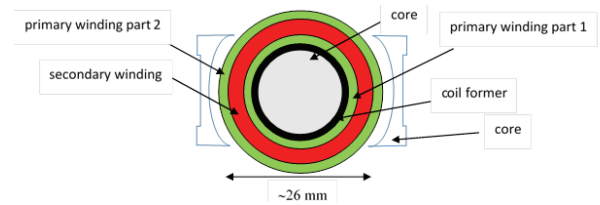

Fig. 14.   Cross-section of the high frequency transformer

Thermal time constant τ is

$$\tau = R_{th}C_{th}. \tag{13}$$

Thermal resistance R_{th} and capacitance C_{th} are approximated by the following equations:

$$R_{th} = \frac{1}{hS} = \frac{1}{h(2\pi rH + 2\pi r^2)}, \tag{14}$$

$$C_{th} = \rho Vc = \rho\pi r^2 Hc, \tag{15}$$

where $H = 22$ mm is the height of the windings, $r = 13$ mm is the outer radius of windings (Fig. 14). Symbols ρ and c denote the effective, average density and specific heat of the transformer structure, h is the heat transfer coefficient by convection and radiation.

From (14) and (15) one can derive the effective, average volumetric thermal capacity of the main core and windings of the transformer:

$$C_{vth} = \rho c = \tau_{th} h \left(\frac{2}{r} + \frac{2}{H}\right) \tag{16}$$

Assuming that $h = 20$ W/m²K and taking from Fig. 12 or 13 time constant value $\tau = 635$ s, one can get $C_{vth} = 3.1 \cdot 10^6$ J/Km³. For copper one can get $C_{vthCu} = 3.4 \cdot 10^6$ J/Km³ ($\rho = 8900$ kg/m³ and $c = 396$ J/kgK) while for typical ferrites $C_{vthferrite} = 3.7 \cdot 10^6$ J/Km³ ($\rho = 5000$ kg/m³ and $c = 750$ J/kgK). Taking into account that there are many air free spaces in the transformer, the obtained value of C_{vth} is quite realistic – eqn. (16).

6. Conclusions

This paper presents preliminary results of the research on power losses in H-F transformers. The special attention was paid to the impact of the fringing effect on areas adjacent to the air gap in the device. The analysis was performed using complex thermal impedance approach well known and widely used in power electronics [10]. An exemplary 100 W flyback transformer was identified as a single time constant thermal structure. The experiment confirmed the main heat source is somewhere inside the core or/and the windings, probably due to the secondary winding AC power losses and the fringing flux. Time constant identified by the experiment corresponded to heat removal by convection. All the measurements were performed using dynamic, time-dependent thermographic measurements. The cooled photon camera was used in the research operating at low frame rate 25 Hz only. It was the reason that the heat flow inside the core and windings was not observed. In the next step of the research an attempt to estimate the contribution of fringing flux to the total power losses in H-F transformer is to be undertaken.

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