Waveform Dependent Switching Losses in Flyback Transformer Foil Windings

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***Abstract*—The copper losses in a foil wound flyback transformer are examined using a time domain 2-D Finite Element model. The purpose of the study is to examine the losses as a function of switching waveform. It is shown that the windings carry non-negligible induced current at times when the net current in the winding is zero. As with eddy current core loss, the copper loss is shown to be functionally dependent on the time rate of change of flux density.**

I. Introduction

The transformers used in high frequency DC-DC converters must be designed to minimize copper losses in the primary and secondary windings. Classical expressions for the power loss in a foil winding [1] are widely used to estimate the copper loss in such transformers. In certain designs, one or both of the windings may be actual foils. Alternatively, the foils may be equivalent representations for a given layer of discrete wire turns. Corrections to the classical expressions have been developed to account for edge [2] and multi-layer [3,4] effects. In recent years, Finite Element (FE) tools have become available that considerable improve the capability of the designer to account for geometrical effects, at least in instances where the transformer geometry can be approximated in a 2-dimensional sense [5,6].

The transformer used in an isolated flyback converter represents a particularly interesting case in that the primary and secondary windings are excited for alternative half cycles. When the primary is conducting, the secondary is effectively open circuit, and vice versa. As a result, the transformer is designed with an air gap that, in turn, can have a significant effect on the copper loss. Special winding techniques and the use of Litz wire have been proposed to reduce such losses [7]. If sheet windings are used, loss reduction becomes particularly important in the case of the flyback transformer since it has been suggested [8], on the basis of 2-D FE analysis and certain approximations, that substantial currents can flow even when a winding is carrying zero net current; i.e. is in the off-state. If substantial off-state currents do, in fact, flow then the effective resistance of flyback transformer windings may well be much larger than predicted by the classical methods.

This paper uses time domain 2-D Finite Element modeling to examine the copper losses in typical sheet wound flyback transformer windings. The key objective is to examine the extent to which the switching waveforms that are typical of a flyback transformer influence the winding losses.

In this regard, two particular issues are of interest. First, the classical and widely used approach to loss calculations, which was described by Vandelac and Ziogas [3], uses the notion of *steady state* mmf waveforms as a basis to estimate losses in individual layers of a sheet winding. The question is whether the current in a winding that is subjected to switching can, in fact, be treated in a steady state sense.

More recently, as noted above, it has been suggested [8] that substantial local currents can flow, even when the net current in a particular winding of a flyback transformer is zero. The winding carrying the most significant current of this type (i.e. induced off-state currents) is the one that is adjacent to the center leg air gap. This issue is explored further in the present paper. In particular, time domain Finite Element models are used in an attempt to better represent the switching conditions that actually occur in flyback transformers.

II. Flyback Transformer Copper Losses

Unlike the windings of a conventional transformer, which carry steady state alternating currents that generally do not have a DC component, the primary and secondary windings of a flyback transformer carry unidirectional pulses of current, as shown in Fig.1. When the primary winding of the transformer is in a conducting mode, the net current in the secondary is zero, and vice versa. Thus, during the interval $t_0 \leq t \leq t_1$, the primary winding can be thought of as being in the on-state, whereas the secondary is in the off-state. These roles are reversed for time $t_1 \leq t \leq t_2$,

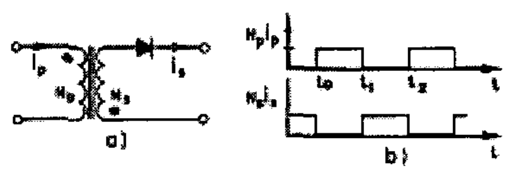

Fig. 1. (a) Ideal flyback transformer. (b) Primary and secondary ampere-turn or mmf waveforms.

A. Classical Loss Method for Sheet Windings

1) Dowell Method: Depending on the application, transformer windings often consist of multiple turn, multiple layer solenoidal coils. In certain applications, there are advantages in using foils or sheets to construct the winding. Regardless of which type of conductor is used (i.e. wires or sheets), the classical approach to loss estimation, attributed to Dowell [1] but used before that in *induction heating applications* [9], is based on the use of an equivalent 1-D conducting sheet to represent a multiple turn

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solenoid. It is conventional practice [1,9] to use the coil fill factor to adjust the electrical conductivity of the equivalent sheet. Clearly, if the actual coil uses sheet windings, no adjustment is necessary.

The basic assumption used in the Dowell method, and the subsequent methods that are variations on it, was that the component of magnetic field strength tangential to the (equivalent) sheet conductor is constant along the length of the sheet. By inference, it was assumed that the waveform of the total current within the sheet was sinusoidal and could be used as a basis for estimating losses. Under this assumption, an expression was derived for the copper loss as a function of the conductor properties (geometrical and electrical), together with the sinusoidal supply frequency.

2) Vandelac and Ziogas Extension: Vandelac and Ziogas [3] provided a significant extension to the Dowell method by developing a formalism to treat multiple layer windings, including those associated with the flyback transformer application. The basic method centered on determining the waveforms of the magnetic field strength **H** on either side of each turn as a function of time, based on the switching patterns for the transformer as a whole. For example, the pulsed current waveforms shown in Fig. 1(b) lead to surface **H** waveforms that were one of: (i) square wave with DC offset, or (ii) constant amplitude. These, in turn, could be decomposed into their Fourier components. Knowing a sequence of harmonic field strength amplitudes on either side of each turn, the corresponding loss for each harmonic could be determined. By focusing on the harmonic fields on either side of each turn, it was possible to consider effects due to (i) winding layout (i.e. interleaving), and (ii) nonsinusoidal but periodic winding currents.

3) Drawbacks of the Classical Methods: The classical methods are 1-D in that the magnetic field strength at the surface of the winding has constant amplitude over that surface and is sinusoidal in time. These two constraints are important for several reasons. First, practical windings have finite height. Second, physics will dictate where current will flow, particularly in applications such as the flyback transformer where the winding currents are switched. In these cases, the total current in a given winding is constrained to be zero for a half period; this does not mean that the current everywhere in the winding is zero. The zero total current constraint cannot be decoupled from the fact that a winding is multi-dimensional (i.e. either 2-D or 3-D). Being multi-dimensional, during the off-state, current can flow subject to the constraint that zero net current is satisfied. Thus, the method of Vandelac and Ziogas may well be inappropriate for use in switching applications.

B. 2-D Time Harmonic Finite Element Approach

1) FEM Modeling of SMPS Transformers: The Finite Element Method is widely used to model static and dynamic electromagnetic devices. If the device in question can be modeled in 2-D, the FEM can be very efficiently used to model losses and extract equivalent circuit parameters. If the device must be modeled in 3-D in order to capture its electromagnetic functionality, the FEM simulations become considerably more time consuming, and thus are very costly to run. This is particularly true when the device incorporates multiple turn solenoidal coils that must be modeled on a discrete conductor basis. When used for loss estimation, particularly in 3-D, the FEM is usually limited to applications where the excitation is sinusoidally time harmonic.

The advantages offered by the FEM for the purpose of estimating losses in medium to high frequency Switch Mode Power Supply (SMPS) transformers has been increasingly recognized over the past several years [2,5-8,10-12]. Particularly when used with 2-D geometries, such models easily incorporate the finite dimensions of actual windings and the effect of gapped cores.

2) FEM Approximation for Flyback Transformer: Recently, Prieto *et al* [8] used 2-D FEM analysis to examine the copper losses in the windings of a flyback transformer. It was insightfully noted that even when a winding was in the off-state, substantial induced currents could nevertheless flow under the constraint that the net current at any instant of time was zero. It was suggested, on the basis of the FEM analysis, that the foil turn adjacent to the gapped center leg as particularly vulnerable to this effect. Although not specifically stated, it appears that the FE model used for that analysis approximated the flyback behavior by allowing a continuous sinusoidal current to flow in one winding, while simultaneously constraining the net current in the other winding to be zero.

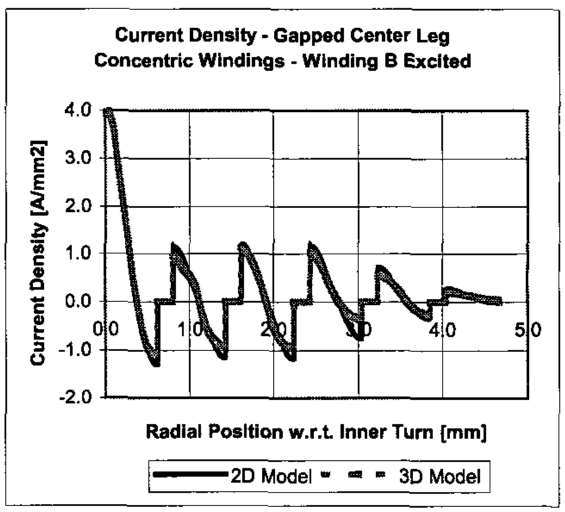

Fig. 2. Current distribution at mid-plane of 2-D transformer model [11]. Winding adjacent to gapped center leg is open circuit.

The currents induced in open circuited sheet windings were recently examined using both 2-D and 3-D FE models [11]. The large currents that were induced in the turn closest to the center leg gap, which were first reported by Prieto *et al* [8], were observed in both 2-D and 3-D FE transformer models. A typical example of the current distribution at the mid-plane of the transformer, as predicted by 2-D and 3-D FEM, is shown in Fig. 2. In this instance, the primary and secondary windings both have 3 turns, with the winding furthest from the center leg of the transformer being the one that carries a forced sinusoidal 100 kHz current. The waveforms shown in Fig. 2 are very similar to the ones that were originally reported in [8]. It should be emphasized that these results were obtained under the assumption that a flyback transformer could be approximated by forcing a sinusoidal current to flow in one winding, with the other forced to be open circuit.

C. *Geometrical and Waveform Dependency of Flyback Transformer Losses*

Prieto *et al* [8] were absolutely correct in recognizing the importance of 2-D effects in allowing significant currents to be induced in the off-state winding of a flyback transformer. In such transformers, the time varying magnetic field causes a voltage to be induced on the off-state winding. Thus, a current will be induced subject to the constraint that the total current in the winding must be zero. This implies the flow of forward and reverse compensating currents. Conventional 1-D models of a sheet winding [1,3,4], through boundary conditions, constrain the forward and reverse induced currents to flow on either side of the winding. In an actual winding, the reverse currents that flow on the upper and lower edges of a finite height winding are equally, if not more, important. It was this aspect that was captured in [8]. What was not clear in [8] was whether the model adequately represented the actual induction mechanism.

In addition to geometrical effects, the off-state winding loss in a flyback transformer is also governed, in a very important sense, by an excitation waveform dependency. As noted, the off-state currents are induced and will flow where geometry and electromagnetic constraints dictate. Just as (dB/dt), the time rate of change of the magnetic field, plays an important role in determining the magnitude of core loss [13], it will also determine the voltage that drives the off-state current. Thus, the magnitude of the off-state currents can critically depend on time dependence of the excitation waveform. Counter intuitively, a flyback winding that carries a sequence of (harmonic rich) square wave pulses may well have lower copper losses compared to when that same winding carries a fundamental frequency sinusoidal pulse.

The waveform dependency of flyback sheet winding copper losses was recently examined [14] using a Spice-derived model. In that study, a relatively simple circuit approach was used to represent spatial effects in the winding. In essence, it appears that the modeling approach used in [14] is a variation of the coupled circuit method. It is interesting to note that the latter method, as applied to eddy current and proximity effect problems, has a long and venerable history [15,16].

III. FEM WAVEFORM DEPENDENT ANALYSIS

A. *3-D vs. 2-D FE Models*

The lower portion of a transformer having foil wound primary and secondary windings is shown below in Fig. 3. Spiral 3-turn windings have been used for both the primary and secondary, and the design has been based on a Magnetics Inc. ETD-43434 ferrite core. Note that the tab shown on each winding was included in the model to allow for the presence of current source excitation.

This model was used to examine several accuracy issues related to the 2-D and 3-D modeling of transformers [10,11], among them: (i) The extent to which the actual 3-D core and winding geometries affect predicted circuit parameters of the device, and (ii) The order of magnitude error that is made when an equivalent 2-D model is used to represent the actual 3-D device. Hoke and Sullivan [17] also examined similar issues in a recent publication.

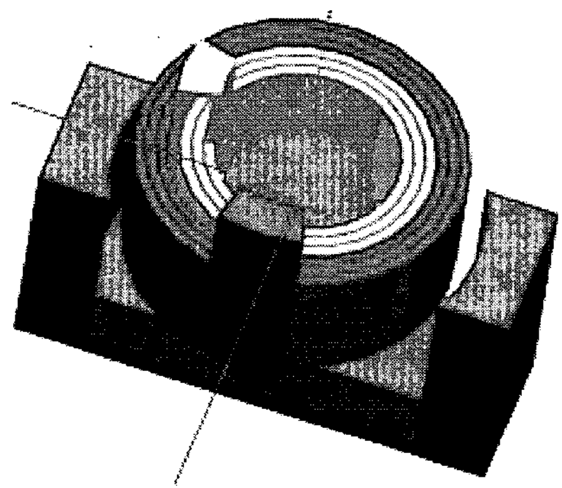

Fig. 3. Partial solid model of ETD core transformer with spiral primary and secondary windings

The 2-D vs. 3-D FE comparison suggested that: (i) The error made in replacing 3-D spiral foil windings by concentric, discrete turns of foil was in the order 5% [10]; (ii) A simple 2-D approximate model consistently over estimated copper losses by 20-30% [11,17]; (iii) In agreement with [8], 2-D and 3-D models both showed very large currents induced in the turn adjacent to a gapped center leg [10,11] of a flyback transformer. In the latter instance, the excitation consisted of a continuous sinusoidal signal applied to one winding, with the other constrained to be open circuit. Clearly, this only approximates the actual behavior of the flyback transformer.

As noted above, 2-D approximate models, in which an equivalent pot core [6] is used, tend to over estimate coil losses by as much as 30% when the core is highly three-dimensional, as is the case with E-cores. On the other hand, a 2-D FE model provides an extremely efficient means of exploring design issues such as, for example, the dependency of the copper loss on geometrical factors. A parametric study of this issue using full 3-D FEM would simply not be possible.

B. 2-D Time Dependent FE Model

A time domain 2-D FEM provides a convenient and very efficient means to explore issues related to the waveform dependency of copper losses high frequency transformer foil windings. Based on previous studies [11,17], it is known that the losses will be overestimated, within reasonable bounds, relative to values predicted by accurate 3-D models. Nevertheless, the identified error bounds are judged to be tolerable.

The present study considers a transformer based on an E-core, operating at 100 kHz, with 3-turn primary and secondary foil windings. The equivalent 2-D pot core approximation was defined as per [6]. The 2-D model is shown in Fig. 4 and critical dimensions are summarized in Table 1.

TABLE I
DATA FOR EQUIVALENT 2-D POT CORE TRANSFORMER

Core Radius (inner leg)	5.4	mm
Window Height	10.6	mm
Inner Radius (1st turn)	6.745	mm
Thickness (per turn)	0.62	mm
Height (per turn)	9.62	mm
Spacing (between turns)	0.19	mm

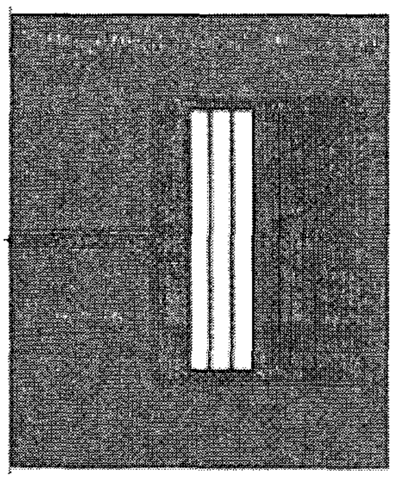

Fig. 4. 2-D equivalent pot core model of E-core transformer with foil wound primary and secondary.

For the purpose of this study, the following core configurations were considered: (i) No gap (considered for completeness), (ii) A single 0.50 mm gap in the center leg, and (iii) Split 0.25 mm gaps – center leg and outer shell. In each instance, concentric windings, as shown in Fig. 4, as well as fully interleaved windings, were considered.

One purpose of the study was to determine the extent to which the flyback transformer losses could be determined using the approximate excitation model; i.e. where one winding had a continuous sinusoidal excitation of 1 A peak, 100 kHz and the other was treated as being open circuit. Recall that this was the model that predicted large induced losses for the turn immediately adjacent to the center leg air gap.

Full transient analyses were performed for each of the various models using one of two 100 kHz, 1 A peak waveforms: (i) Half period (5μs) pulses of sinusoidal current, applied sequentially to each winding, and (ii) Half period (5μs) pulses of trapezoidal current, again, applied sequentially. In the latter case, each pulse had 0.5μs rise and fall times. Each simulation was run over a sufficient number of periods to ensure convergence. It was determined during numerical testing that a time step of 0.1 μs (50 steps per half period) was sufficient to ensure good convergence and numerical stability. The predicted losses did not changed when the time step was decreased to 0.01μs.

IV. RESULTS AND DISCUSSION

A. Comparison of Continuous and Pulsed Sinewave Excitation

The time domain model was initially used to compare the losses predicted for alternating half period sine pulses in each winding to the losses when one winding carries a continuous sine wave current, the other winding being open [6,11]. The center leg of the core has a 0.50 mm gap in this instance. The results are summarized in TABLE II, where the turn closest to the center leg is A01. TABLE II shows the actual turn-by-turn loss as well as the continuous to pulsed ratio. These results clearly show a 5 to 8-fold discrepancy between the time domain losses and the losses under time harmonic excitation.

TABLE II
TURN-BY-TURN LOSSES
CONTINUOUS VS. PULSED SINE WAVE EXCITATION

Turn	Continuous Sine Wave	Pulsed Sine Wave	Ratio Cont./Pulsed
	mW	mW	
A01	4.43	0.59	7.48
A02	2.67	0.32	8.31
A03	2.90	0.43	6.80
B01	2.26	0.39	5.80
B02	0.90	0.16	5.53
B03	0.26	0.05	5.01
Coil A	10.00	1.34	7.46
Coil B	3.42	0.60	5.66
Total	13.42	1.94	6.90

Numerous simulations were run, all with results similar to those shown in TABLE II; namely, time harmonic models cannot be used to approximate the behavior of devices that inherently operate in a discontinuous fashion. Note further that since one of the windings in the time harmonic model was open circuit, it was not possible to extract a meaningful equivalent resistance for that winding.

B. Pulsed Sine: Effect of Gap and Winding Strategy

The AC to DC resistance ratio is shown in Fig. 5 and Fig. 6 for the case where the windings are concentric and interleaved, respectively. In both instances, the current excitation consists of alternating half period pulses of sine wave current (1 A peak; 100 kHz). Interleaving the turns and distributing the total gap between the center and outer legs minimizes the total loss. Relative to the case where concentric windings are used with a single gap, the total loss reduction is in the ratio 3.2:1.

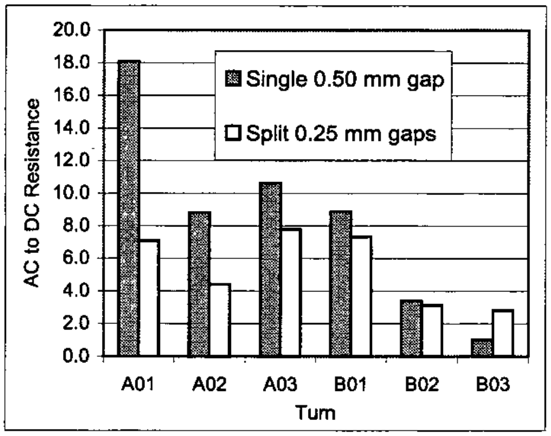

Fig. 5. AC to DC ratio for concentric foil windings and split gap core. Half period sinusoidal current.

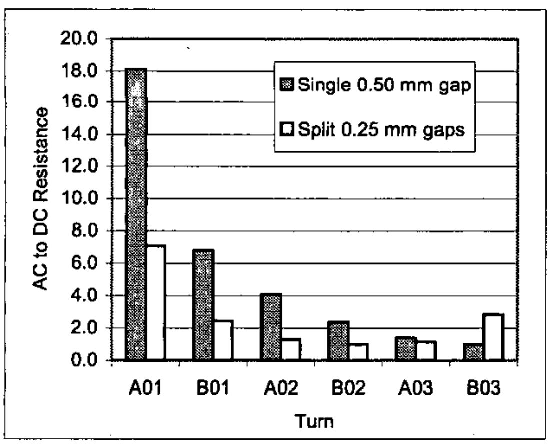

Fig. 6. AC to DC ratio for interleaved foil windings and split gap core. Half period sinusoidal current.

C. Pulsed Sine vs. Pulsed Trapezoid

The effect of current wave shape on the turn-by-turn resistance is illustrated in Fig. 7. Here, losses for half period pulses of 100 kHz 1 A peak sine and trapezoidal waveforms wave are compared. These results are for the case where the gap is split between the center and outer legs.

This result is particularly interesting since it not only illustrates the effect that current wave shape can have, but also brings out the point that losses in this transformer cannot be estimated on the basis of the steady state waveforms and their harmonic components. In Fig. 7, the resistance of turns A03 through B02 is essentially independent of wave shape. However, wave shape does have a significant impact on the resistance of turns adjacent to an air gap. Moreover, counter intuitively; the resistance is higher when the wave shape is sinusoidal. In other words, for the innermost turn of this particular winding arrangement, the copper loss is almost 6 times lower when the current is a trapezoidal pulse of magnitude I_0, as compared to the case where the winding carries only the fundamental component of that pulse.

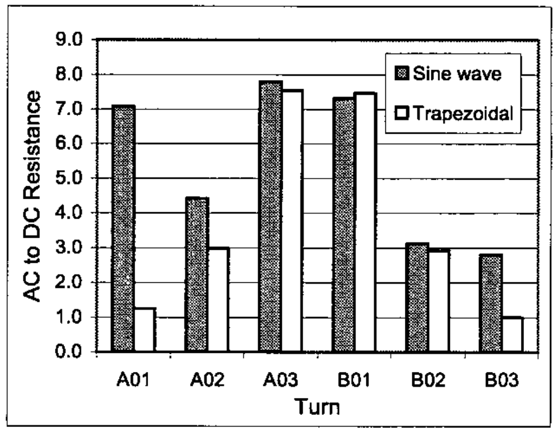

Fig. 7. AC to DC resistance ratio for concentric windings and split gap core. Sine vs. trapezoidal waveform

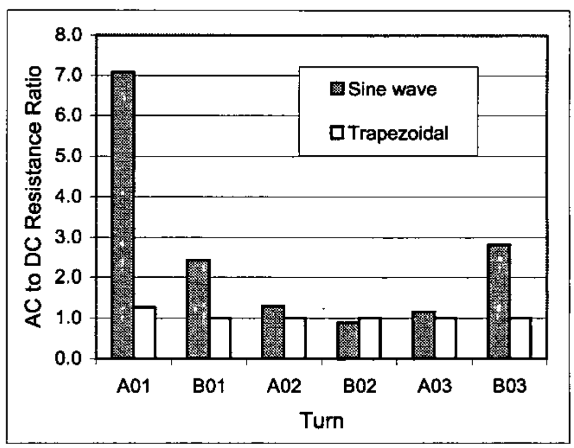

Fig. 8. AC to DC resistance ratio for interleaved windings and split gap core. Sine vs. trapezoidal waveform.

The corresponding results for an interleaved winding arrangement are shown in Fig.8. Interleaving will have no impact on the resistance of the inner and outer most windings. However, the resistance of the other windings is reduced. Moreover, when the waveform is trapezoidal, the resistance is close to the DC value.

C. Copper Loss as Function of Time

The loss in windings A01 through A03 is shown in Fig. 9 as a function of time in the steady state. The winding arrangement is concentric and there is a single center leg gap. The excitation is trapezoidal (1 A peak, 100 kHz). The wave shape effect is clearly evident. The losses rise sharply on the leading and trailing edge of the current. However, during the period when the current waveform is flat, the losses decay exponentially due to the decaying transient component of current. The greatest loss is in turn A03, which is adjacent to Winding B, and not in Turn A01, which is adjacent to the center leg air gap.

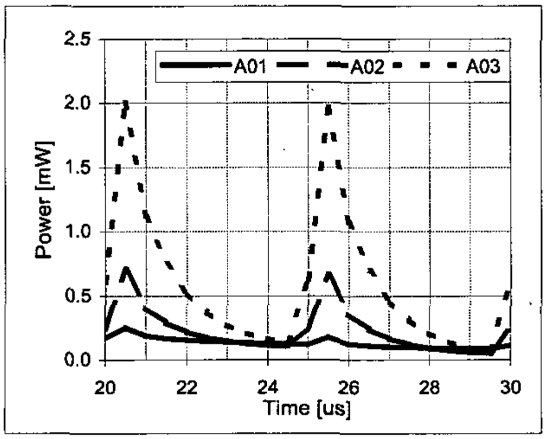

Fig. 9. Steady state time variation of loss. Concentric winding; Single gap. Trapezoidal current.

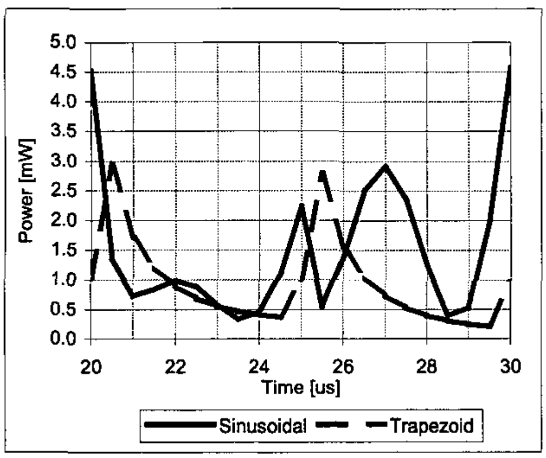

Fig. 9. Total Winding A loss - trapezoidal vs. sinusoidal half pulse. Single gap. Concentric winding..

A comparison of total loss in Winding A is given in Fig. 10 for the case of sinusoidal and trapezoidal half period pulses, both 1 A peak, 100 kHz. The winding arrangement is concentric and the center leg is gapped. The time effects are clearly evident in the sine pulse loss, particularly for the period that Winding A is in the off-state ($25 \leq t \leq 30$ μs). It is this period of operation that contributes to the higher losses.

V. Conclusions

This paper has used a time domain Finite Element model to examine the effect of current waveform on the losses in a foil wound flyback transformer. It has been shown that widely used steady state loss models do not apply to this application. The copper loss includes induced current effects that depend on the time rate of change of the excitation. Counter intuitively, a harmonic rich waveform may result in lower winding resistance that than in obtained for sinusoidal excitation at the same frequency.

References

1. P.L. Dowell, "Effects of eddy currents in transformer windings", *Proc. IEE*, vol. 113, no. 8, pp. 1387-1394, 1966.
2. F. Robert, P. Mathys, J.-P. Schauwers, "A closed-form formula for 2D ohmic loss calculation in SMPS transformer foils", *Proc. APEC.*, pp. 199-205, 1999.
3. J.P. Vandelac, P. Ziogas, "A novel approach for minimizing high frequency transformer copper losses", *IEEE Trans. Power Electronics*, vol. 3, no. 3, pp. 266-277, 1988.
4. N.H. Kutkut, D.M. Divan, "Optimal air-gap design in high frequency foil windings", *Trans. Power Electronics*, Vol. 13, No.5, pp. 942-949, 1998.
5. R. Asensi, J.A. Cobos, O. Garcia, R. Prieto, J. Uceda, "A full procedure to model high frequency transformer windings", *Proc. PESC 1994*, pp. 856-863, 1994.
6. R. Prieto, L. Ostergaard, J.A. Cobos, J. Uceda, "Axisymmetric modeling of 3D magnetic components", *Proc. APEC*, pp. 213-219, 1999.
7. C.R. Sullivan, T. Abdallah, T. Fujiwara, "Optimization of a flyback transformer winding considering two-dimensional field effects, cost and loss", *Proc. APEC*, pp. 116-122, 2001.
8. R. Prieto, J.A. Cobos, O. Garcia, P. Alou, J. Ueda, "High frequency resistance in flyback type transformers", *Proc. APEC*, pp. 714-719, 2000.
9. K. Reichert, "The calculation of coreless induction furnaces" *ETZ-A* (in German), *Archiv für Elektrotechnik*, vol. 49, pp. 376-397, 1965.
10. J.D. Lavers, "On- and off-state losses in foil windings", *Proc. PCC-Osaka*, pp. 1096-1101, Osaka, JP, April 2-5, 2002
11. J.D. Lavers, E.D. Lavers, "An accuracy assessment of 2D and 3D finite element models for ferrite core, sheet wound transformers", *Proc. APEC*, pp. 158-164, Dallas, TX, March 10-14, 2002.
12. F. Robert, P. Mathys, J.-P. Schauwers, "Ohmic loss calculation in SMPS transformers: numerical study of Dowell's approach accuracy", *IEEE Trans. Magnetics*, vol. 34, no. 4, pp. 1255-1257, 1998.
13. J. Reinhart, A. Brockmeyer, R.W. De Donker, "Calculation of losses in ferro- and ferromagnetic materials based on the modified Steinmetz equation", *Proc. IAS Conf.*, pp. 2087-2092, 1999.
14. L. Heinemann, J. Helfrich, "Modeling and accurate determination of winding losses of high frequency transformers in various power electronic applications", *Proc. APEC*, pp. 647-653, 2000.
15. R. Dudley, P.E.Burke, "The prediction of current distributions in induction heating installations", *Trans. Ind. Appl.*, vol. IA-8, pp. 565-571, 1972.
16. F. Hegewaldt, "The calculation of skin effect by a differential method", (in German), Paper 632, *Vth Int. Congress on Electroheat*, Wiesbaden, West Germany, 1963.
17. A.F. Hoke, C.R. Sullivan, "An improved two-dimensional modeling method for E-core transformers", *Proc. APEC*, pp. 151-157, Dallas, TX, March 10-14, 2002.