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# Winding Position in Power Transformers to Reduce Copper Losses: Non-sinusoidal Currents

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**Abstract.** The orientation used to wind a transformer is important in power transformers carrying high-frequency currents. Analysis of copper losses due to skin and proximity effects is performed in order to choose the best configuration of the turns inside the winding area when conductors are carrying non-sinusoidal current. Initially results are a direct application of Dowell's formula for the fundamental component and the harmonics of the current flowing through the winding. Later, FEA simulations confirm the relationship between current frequency, insulation thickness, winding window width and winding position.

**Keywords.** Transformer windings, Resistance, Power transformer losses.

## I. INTRODUCTION

In high-frequency power transformers, copper losses are one of the main design elements. High number of layers and high frequencies result in high copper losses due to skin and proximity effects.

Dowell, in his famous paper from 1966 [1] solves the problem of calculating the AC resistance of conductors in a transformer. He explains the proximity effect mathematically, solving Maxwell's equations for a 1-D model of transformers. Originally, his work considers rectangular-section conductors, which occupy the entire core window. Later, adjustments are proposed to address the problem of round wires and conductors that do not fill the entire core window.

Since Dowell's 1-D analytical approach models a 3-D phenomenon, other authors have modified his approach [2-4], or developed different ones [5-8].

Several authors make use of FEA simulation to obtain more precise results [9-11]. However each simulation is specific to the design problem and it may not be applicable to other transformers. Besides that, usually FEA softwares are expensive and time-consuming, which may limit a transformer designer in an industrial environment.

A formula to calculate the optimum foil or layer thickness in multilayer transformers is presented in [12]. This formula is developed by using the RMS value of the current passing through the winding conductors and from its derivative. Also, the equations may be applied to any periodic current waveform, avoiding the need of Fourier analysis, which is computationally more complicated.

In [13], authors have provided an insight into the positioning of a certain number of turns in a given core window which reduces copper losses. Specifically, they have

determined which are the optimal number of turns in each layer and the number of layers, with regard to the current frequency.

Although [13] is useful for giving a general overview of the choice on the position of the winding, the results only apply to sinusoidal currents which is not the case for most of switch-mode power supply transformers or intercell transformers. Therefore the present article analyses the influence of non-sinusoidal current flowing through conductors in winding and how it modifies the choice of the number of turns and layers which minimize copper losses.

Discussion about the use of single-layer or multi-layer winding is also made in [14]. The authors analyse these two options by considering some non-sinusoidal current waveforms and small thickness-to-skin-depth ratio. In their analyses, authors also make use of concepts as optimum layer thickness and optimum number of layers. General conclusions from [14] verify the analysis and results presented in the present article which are simpler and more directly used in a transformer design.

## II. WINDING POSITION AND SINUSOIDAL CURRENT

In a power transformer design, sometimes the number of turns necessary on the primary and on the secondary is primarily chosen as the mutual and leakage inductances are relatively defined. Also, there are cases where the transformer designer does not have a vast option of cores available, and so the geometry of the core window is fixed. In such cases, the designer must take special attention on how many layers and how many turns in each layer he must choose so the copper losses may be minimized.

For example, if we take Fig. 1 as the basis of our analysis, the winding on the right side of the UI shape core must fit in the core window, specifically in the space reserved to it (represented by the light grey area at the right side), which has width  $W$  and height  $H$ .

As shown in Fig. 2, if the winding has, for example, 12 turns, there are 6 different ways of fitting the conductors inside the winding area, if the conductor's shape is not already defined: a) 1 layer of 12 turns, b) 2 layers of 6 turns, c) 3 layers of 4 turns, d) 4 layers of 3 turns, e) 6 layers of 2 turns, f) 12 layers of 1 turn. If conductors have round cross-section, the analyses shown here may be used but with some approximations which make part of future work.

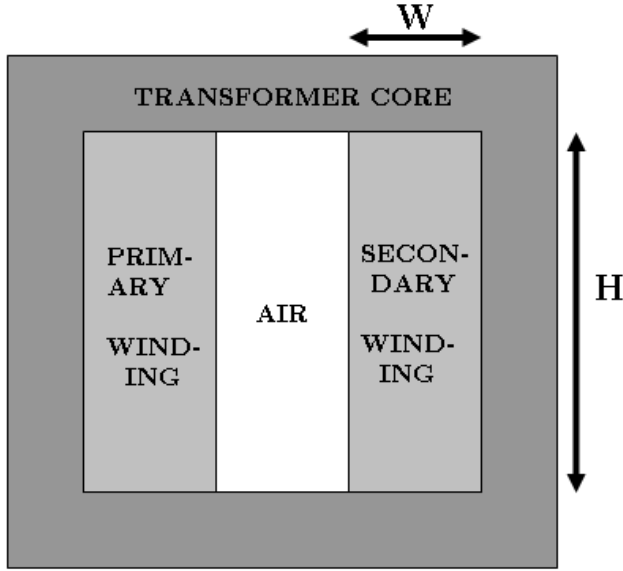


Fig. 1 – Core window and winding area

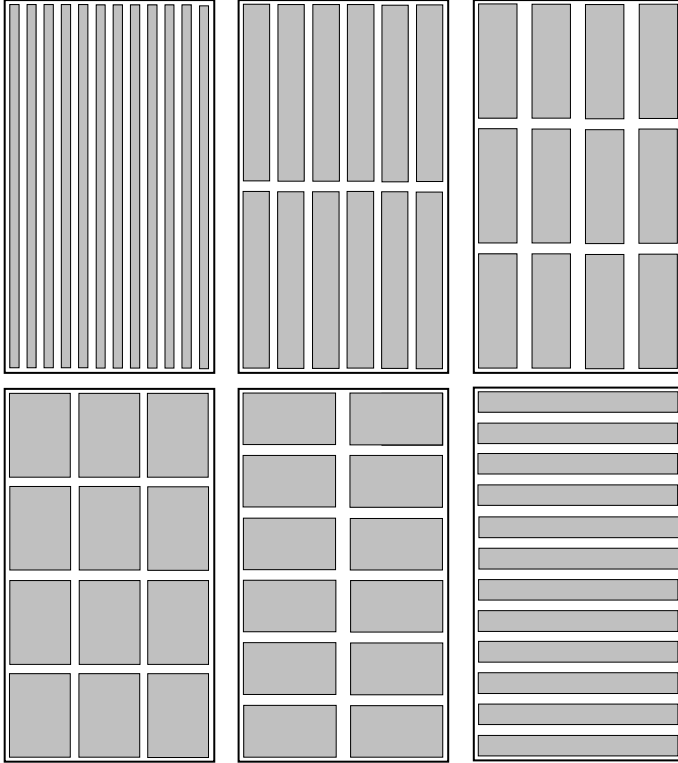


Fig. 2 – Configurations for fitting 12 conductors inside the winding area

The sections of the conductors are the same in all these configurations, but in each one of them the conductor dimensions  $Wc$  and  $Hc$  are different. For this reason, it is not direct to say that the copper losses are higher in the 12-layer configuration than in the 1-layer configuration. It depends on the frequency, as it will be seen later.

In order to calculate the copper losses or the AC resistance of each configuration, some preliminary considerations must be made:

1. Insulation between conductors are negligible (later it will be taken into account)
2. All the turns, in one specific configuration, have the same dimensions ( $Wc$  and  $Hc$ ).
3. The field inside the core window is perfectly vertical (as assumed by Dowell)

Dowell's formula, which calculates the ratio ( $Fr$ ) between the AC resistance ( $R_{ac}$ ) and the DC resistance ( $R_{dc}$ ) is presented below:

$$Fr = \frac{R_{ac}}{R_{dc}} = Q \frac{\sinh 2Q + \sin 2Q}{\cosh 2Q - \cos 2Q} + 2Q \frac{m^2 - 1}{3} \frac{\sinh Q - \sin Q}{\cosh Q + \cos Q} \quad (1)$$

where ' $m$ ' is the number of layers of the winding and  $Q$  is the ratio between the conductor's thickness ( $Wc$ ) and the skin depth ( $Dpen$ ).

In order to compare the losses associated to each configuration, (1) was used to calculate " $Fr$ ". Note that all configurations have the same  $R_{dc}$ , and consequently the one which has the lowest  $Fr$  is the one which has lower losses.

In Fig. 3,  $Fr$  was calculated for all possible configurations of a transformer with 12 turns, and for a frequency range from 10Hz to 1MHz. The winding area has dimensions  $W=7.24\text{mm}$  and  $H=13.1\text{mm}$ .

By analyzing these curves and several other curves created for windings with " $p$ " turns, some conclusions may be made:

1. For lower frequencies, the configuration with  $p$  layers is the one which has lower losses.
2. For higher frequencies, the configuration with 1 layer is the one which has lower losses.
3. With regard to copper losses, the other configurations are never better than the 2 extreme ones ( $p$  layers of 1 turn or 1 layer of  $p$  turns).

By the use of MATLAB 7.1, from MATHWORKS, innumerable graphs were generated and it was observed that the 3 conclusions stated above are valid independently of the dimensions of the winding area. Those graphs have the same shape as the one in Fig. 3 but the frequency in which Configuration 1 has less losses than Configuration 6 changes with the winding area width  $W$ . From now on, this frequency will be called the "limit frequency" ( $Flim$ ).

Note that  $Fr$  and  $Flim$  can only vary with the winding area width but not with its height since Dowell's formula is a 1-D approach.

The limit frequency is the frequency where one configuration becomes more advantageous than the other. So, in order to analytically find the limit frequency, we must equal Dowell's equation for the two configurations. By doing that, we find that there is a unique  $Q$  for each winding of  $p$  turns. Consequently the product of the Limit Frequency and the winding area width have the following formula

$$fW^2 = \frac{(pQ)^2}{\sigma\pi\mu} \quad (2)$$

From equation (2) we can verify that the product of the limit frequency and the square of the winding area width is a constant for each winding with  $p$  turns, if we assume that permeability and conductivity of the conductor are constant. These constants are shown in Table I, for windings from 2 to 30 turns. This table was generated for conductors made of copper at 20°C, which have permeability  $\mu=4\pi \cdot 10^{-7} \text{N/A}^2$  and conductivity  $\sigma=58.10^6 \text{S/m}$ .

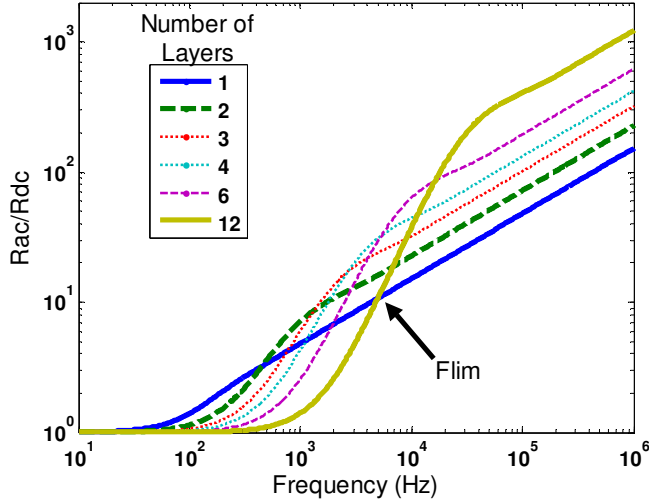


Fig. 3 – Variation of the AC resistance with the frequency for each of the configurations of Fig. 2.

TABLE I  
CONSTANTS USED TO DECIDE WHICH CONFIGURATION RESULTS IN LESS LOSSES

$p$	$Flim * W^2$	$p$	$Flim * W^2$	$p$	$Flim * W^2$
2	0.0451	11	0.4395	21	1.0563
3	0.0759	12	0.4945	22	1.1246
4	0.1115	13	0.5513	23	1.1940
5	0.1507	14	0.6095	24	1.2653
6	0.1929	15	0.6695	25	1.3373
7	0.2378	16	0.7308	26	1.4093
8	0.2851	17	0.7932	27	1.4839
9	0.3346	18	0.8570	28	1.5582
10	0.3860	19	0.9226	29	1.6346
		20	0.9886	30	1.7101

This table is an easy tool used to choose which configuration is the one resulting in less losses. The designer should only multiply the frequency of the current flowing through the transformer and the square of the winding area width. If the resulting number is greater than the constant in Table I corresponding to the number of turns of his winding, then the configuration with 1 layer of  $p$  turns results in less losses, otherwise it is the configuration with  $p$  layers of 1 turn which is preferable. Obviously, final choice of the winding orientation must take into account not only copper losses but also technological and fabrication constraints.

### III. THE INFLUENCE OF NON-SINUSOIDAL CURRENTS

In switch mode power supplies, the waveform of the current flowing through transformers and inductors may assume different shapes when compared to a sinusoidal one. For example, in forward converters the current waveform is typically pulsed, while in a full bridge converters it is squared. In inductors and coupled inductors current has typically a triangular waveform.

In cases where the current is not sinusoidal Dowell's formula can also be used [15-17], but firstly the amplitude of each harmonic of the current must be found, by means of Fourier series, as the one shown below:

$$i(t) = I_{dc} + \sum_{n=1}^{\infty} a_n \cos(n\omega t + \theta_n) \quad (3)$$

where  $I_{dc}$  is the continuous value of the current,  $a_n$  is the amplitude of the Nth harmonic with corresponding phase  $\theta_n$ .

Copper losses in the transformer is given by the sum of the losses of each harmonic, which is shown below:

$$P = R_{dc} \left( I_{dc}^2 + \sum_{n=1}^{\infty} Fr_n \frac{a_n^2}{4} \right) \quad (4)$$

where  $Fr_n$  is the ratio between the AC resistance of the Nth harmonic and the DC resistance, and is given by the equation below:

$$Fr_n = \sqrt{n}Q \frac{\sinh 2\sqrt{n}Q + \sin 2\sqrt{n}Q}{\cosh 2\sqrt{n}Q - \cos 2\sqrt{n}Q} + 2\sqrt{n}Q \frac{p^2 - 1}{3} \frac{\sinh \sqrt{n}Q - \sin \sqrt{n}Q}{\cosh \sqrt{n}Q + \cos \sqrt{n}Q} \quad (5)$$

where  $Q$  is the ratio between the conductor's thickness and the skin depth related to the fundamental frequency

$$Q = \frac{W_c}{Dpen_1} = W_c \sqrt{\sigma \pi f_1} \quad (6)$$

In order to verify the influence of non-sinusoidal currents on the "Limit Frequency", initially Dowell's formula for each harmonic will be used. Later simulations using a FEA software will be carried out with the purpose of obtaining more precise results and evaluating the use of Dowell's formula.

#### A. Analytical Explanation about the influence of non-sinusoidal waveforms on the Constants

As observed in section II, the two configurations which result in fewer copper losses are those with 1 layer of  $p$  turns and  $p$  layers of 1 turn. And the limit frequency is the frequency where one configuration becomes more advantageous than the other. In this case, in order to analytically find the limit frequency, we must equal the equation of losses (4) for the two configurations.  $P_1$  is the equation for the configuration of 1 layer of  $p$  turns, and  $P_2$  is the equation for the configuration of  $p$  layers of 1 turn.

$$P_1 = R_{dc} \left( I_{dc}^2 + \sum_{n=1}^{\infty} \frac{a_n^2}{4} \sqrt{n} Q_1 \frac{\sinh 2\sqrt{n} Q_1 + \sin 2\sqrt{n} Q_1}{\cosh 2\sqrt{n} Q_1 - \cos 2\sqrt{n} Q_1} \right)$$

$$P_2 = R_{dc} \left[ I_{dc}^2 + \sum_{n=1}^{\infty} \frac{a_n^2}{4} \left( \sqrt{n} Q_2 \frac{\sinh 2\sqrt{n} Q_2 + \sin 2\sqrt{n} Q_2}{\cosh 2\sqrt{n} Q_2 - \cos 2\sqrt{n} Q_2} + 2\sqrt{n} Q_2 \frac{p^2 - 1}{3} \frac{\sinh \sqrt{n} Q_2 - \sin \sqrt{n} Q_2}{\cosh \sqrt{n} Q_2 + \cos \sqrt{n} Q_2} \right) \right] \quad (7)$$

Since

$$Q_1 = W / D p n_1 = W \sqrt{\sigma \pi \mu f} \quad (8)$$

$$Q_2 = W / p / D p n_p = \frac{W}{p} \sqrt{\sigma \pi \mu f} = \frac{Q_1}{p} \quad (9)$$

when we equal  $P_1$  and  $P_2$ , we find the following expression:

$$\sum_{n=1}^{\infty} a_n^2 \left( p \frac{\sinh 2\sqrt{n} p Q_2 + \sin 2\sqrt{n} p Q_2}{\cosh 2\sqrt{n} p Q_2 - \cos 2\sqrt{n} p Q_2} - \frac{\sinh 2\sqrt{n} Q_2 + \sin 2\sqrt{n} Q_2}{\cosh 2\sqrt{n} Q_2 - \cos 2\sqrt{n} Q_2} - 2 \frac{p^2 - 1}{3} \frac{\sinh \sqrt{n} Q_2 - \sin \sqrt{n} Q_2}{\cosh \sqrt{n} Q_2 + \cos \sqrt{n} Q_2} \right) = 0 \quad (10)$$

Equation (10) depends basically of 3 parameters: the number of turns in the winding ( $p$ ), the number of harmonics in the current ( $n$ ) and the amplitude of each harmonic ( $a_n$ ). Having these 3 values, we can find  $Q_2 = Q_0$  which solves this equation and since  $Q_2$  is given by (9), we can note that

$$Q_0 = \frac{W}{p} \sqrt{\sigma \pi \mu F_{lim}} \quad (11)$$

or

$$F_{lim} W^2 = \frac{(p Q_0)^2}{\sigma \pi \mu} \quad (12)$$

The influence of each harmonic ( $n$  and  $a_n$ ) in the value of  $Q_0$  is difficult to be analytically analysed since (10) is given by a relatively complicated expression. However, Fig. 4 gives us an insight of such influence, for a winding of 12 turns ( $p=12$ ). In this figure, the x-axis is the ratio between the amplitude of the harmonic and the amplitude of the fundamental. We have plotted the influence of the 3<sup>rd</sup> and the 5<sup>th</sup> harmonics.

Note that in both graphs adding harmonics decreases the value of  $Q_0$ . The higher the amplitude of the harmonic, the smaller the value of  $Q_0$ , and consequently the smaller the Limit Frequency. Also, the higher the frequency of the harmonic, the smaller is the value of  $Q_0$ .

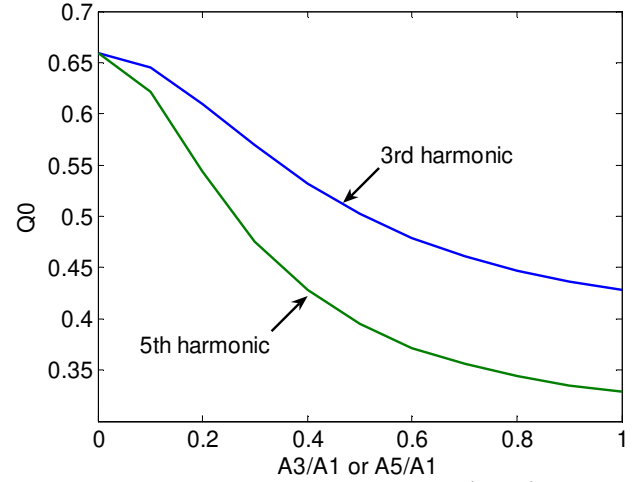


Fig. 4 – Variation of  $Q_0$  with the amplitude of the 3<sup>rd</sup> and 5<sup>th</sup> harmonics

For a given type of waveform, the harmonics content changes depending on the duty cycle ( $D$ ) of the static converter. As a result, the value of  $Q_0$  also changes when the duty cycle is modified. As an example, Fig. 5 shows the variation of  $Q_0$  in regards of the duty cycle, for a triangular current. Since the harmonic content of the triangular current is symmetric about the value  $D=0.5$ , we only plot the values of  $Q_0$  for duty cycles from 0.05 to 0.5, for a number of layers  $p$  from 2 to 15. Note that the closer the duty cycle from 0.5, the higher is the value of  $Q_0$  since the harmonic content is lower for this duty cycle.

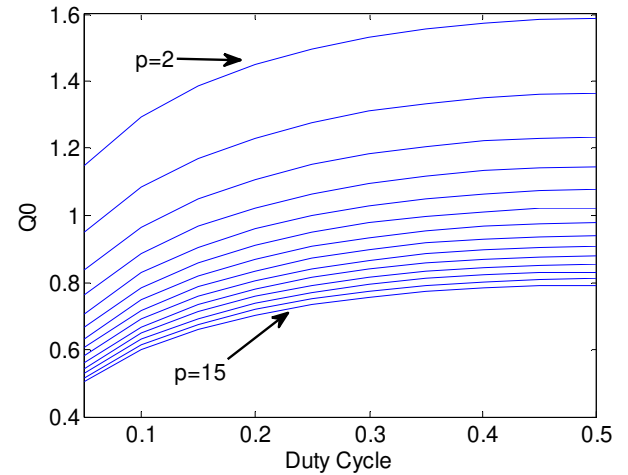
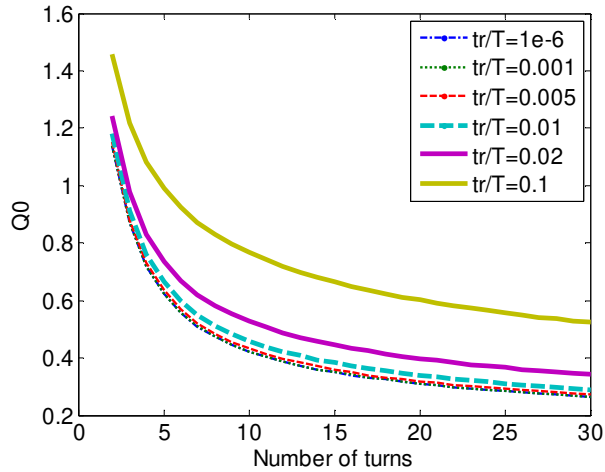


Fig. 5 – Variation of  $Q_0$  with the duty cycle

Also, for some types of waveforms, the rise and fall times ( $tr$ ) play an important role in the harmonics and, as a consequence, they modify the value of  $Q_0$ . If we take a square waveform as an example, the value of  $Q_0$  changes with regards to the the rise and fall times as shown in Figure 6.

Fig. 6 – Variation of  $Q_0$  with the rise and fall times

High values of  $tr$  result on less square-shape currents. This contributes to diminish the harmonic content of the current and consequently to increase the value of  $Q_0$ .

With the values of  $Q_0$  found for each winding of  $p$  turns and for each type of current waveform, we can set up a table for each type of current waveform containing the values of  $F_{lim} * W^2$  used to decide which configuration leads to lower losses.

Instead of showing one table for each type of waveform, we decided to approximate their values by a polynome of 4<sup>th</sup> degree and show all results in a compact form in Table II. Note that this table contains the coefficients to be multiplied by the powers of the number of turns, as shown in the equation below, for each of the main types of current waveforms found in switch mode power supplies, shown in Fig. 7. In this table, we used  $D=0.5$  and  $tr/T=0.01$ .

$$F_{lim} * W^2 = C_4 p^4 + C_3 p^3 + C_2 p^2 + C_1 p + C_0 \quad (13)$$

In order to illustrate the difference in the product  $F_{lim} * W^2$  between different waveforms, we have calculated, using the coefficients in Table II, this product for windings containing 2, 10 and 20 turns. We included these values in the final columns of Table II.

Comparing the exact values found by (10) and those approximated by the 4<sup>th</sup> order polynomes, it was found that the error is below 3.4% for  $p=2$  and far below 1.2% for  $p>2$ .

Similar to Table I, Table II is an easy tool used to choose which configuration is the one resulting in fewer losses. The designer should only multiply the frequency of the current flowing through the transformer and the square of the winding area width. If the resulting number is greater than the result of the polynome described by the coefficients in Table II corresponding to current waveform, then the configuration with 1 layer of  $p$  turns results in fewer losses, otherwise it is the configuration with  $p$  layers of 1 turn which is preferable.

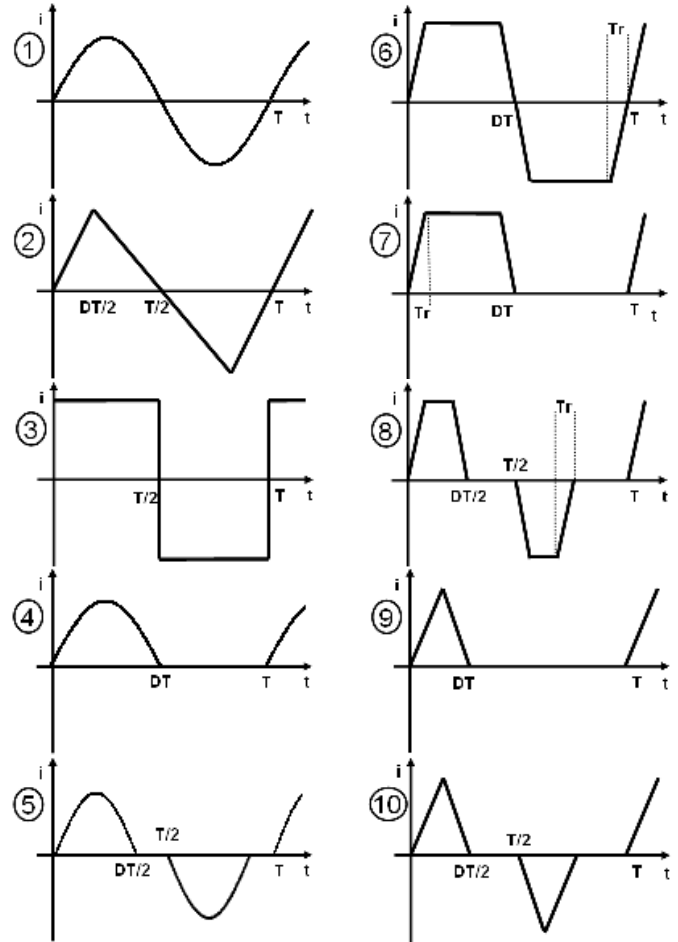


Fig. 7 – Typical current waveforms found in switch mode power supplies

TABLE II  
4<sup>TH</sup> ORDER POLYNOME COEFFICIENTS RELATED TO (13) AND RESULTS

waveform	C4	C3	C2	C1	C0
1	$4.37 \times 10^{-7}$	$-4.09 \times 10^{-5}$	0.001954	0.023832	-0.011582
2	$3.81 \times 10^{-7}$	$-3.54 \times 10^{-5}$	0.001688	0.023247	-0.010411
3	$-3.22 \times 10^{-8}$	$9.62 \times 10^{-7}$	$1.15 \times 10^{-4}$	0.005359	0.012158
4	$3.17 \times 10^{-7}$	$-2.96 \times 10^{-5}$	0.001414	0.019137	-0.007055
5	$1.93 \times 10^{-7}$	$-1.80 \times 10^{-5}$	$8.63 \times 10^{-4}$	0.011759	-0.002811
6	$-1.30 \times 10^{-8}$	$-4.00 \times 10^{-7}$	$1.59 \times 10^{-4}$	0.006581	0.011075
7	$-1.36 \times 10^{-8}$	$-3.91 \times 10^{-7}$	$1.61 \times 10^{-4}$	0.006591	0.011053
8	$-1.87 \times 10^{-8}$	$5.55 \times 10^{-8}$	$1.45 \times 10^{-4}$	0.006564	0.009554
9	$2.97 \times 10^{-7}$	$-2.76 \times 10^{-5}$	0.001307	0.017984	-0.006425
10	$1.70 \times 10^{-7}$	$-1.59 \times 10^{-5}$	$7.66 \times 10^{-4}$	0.010906	-0.003275

$F_{lim} * W^2$ ( $p=2$ )	$F_{lim} * W^2$ ( $p=10$ )	$F_{lim} * W^2$ ( $p=20$ )
0.0436	0.3856	0.9893
0.0426	0.3592	0.9072
0.0233	0.0779	0.1681
0.0366	0.2993	0.7554
0.0240	0.1850	0.4644
0.0249	0.0923	0.2012
0.0249	0.0926	0.2020
0.0233	0.0896	0.1964
0.0346	0.2795	0.7031
0.0215	0.1682	0.4214

#### IV. SIMULATION IN A FEA SOFTWARE

In order to evaluate the accuracy of Dowell's formula and Table II, simulations using a FEA software were performed. Fig. 8 shows the result of a simulation performed in FEMM4.2, for 12-turn windings. To illustrate the problem, Fig. 8 shows a core window with width equal to 10mm, 12mm for each winding. In Fig. 8a the winding has 12 layers of 1 turn and on Fig. 8b the winding has 1 layer of 12 turn.

Since the software FEMM only simulates sinusoidal currents we were obligated to simulate for each frequency separately and then sum up all losses to post calculate an equivalent AC resistance in order to compare with analytical calculation using Dowell's formula. The equation of the equivalent AC resistance is shown below, based on (4):

$$Fr_{eq} = \frac{R_{aceq}}{R_{dc}} = \sum_{n=1}^{\infty} Fr_n a_n^2 \quad (14)$$

where  $a_1$  is always equal to 1 for the purpose of facilitating the calculation. Note that  $R_{aceq}$  multiplied by  $\frac{1}{4}$  of the square of the amplitude of the fundamental current passing through the winding gives us the AC copper losses of this winding.

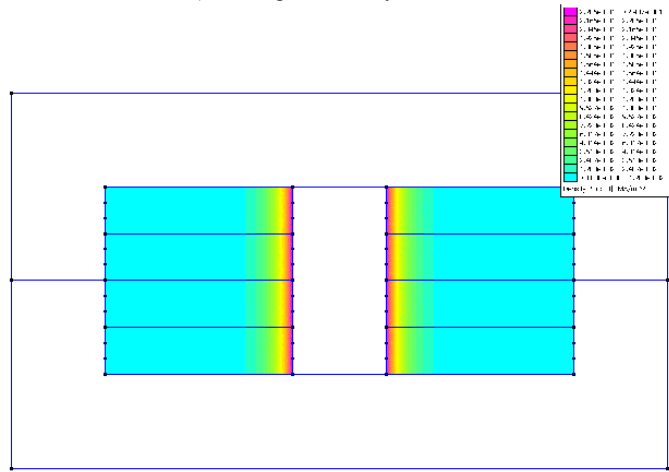
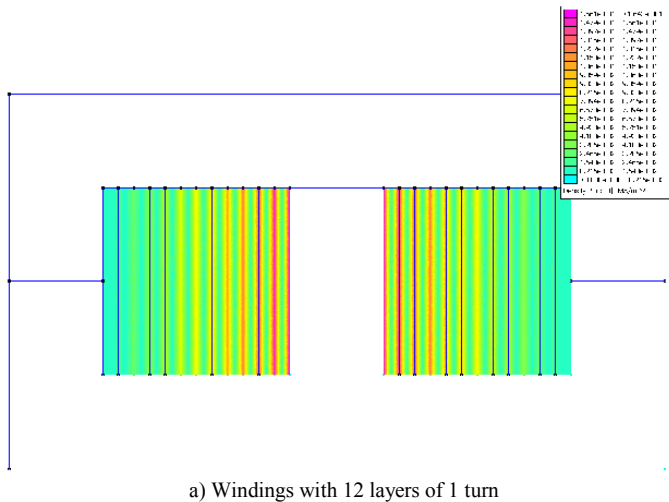


Fig. 8 – Transformers simulated

Simulations of a triangular current for several different fundamental frequencies were performed and the ratio  $R_{aceq}/R_{dc}$  was obtained for each fundamental frequency. Each triangular current was simulated up to the 11<sup>th</sup> harmonic. This is plotted in Fig. 9. Also in this figure we plot the curves calculated by Dowell's equation for this specific configuration. Note that Dowell's formula and FEA simulation give us very close results.

Also using FEMM, a set of simulations was performed varying the number  $p$  of turns in the windings, the winding area width  $W$  and the type of current waveform, in order to find the limit frequencies. After finding all those limit frequencies, we could calculate the constants ( $F_{lim} * W^2$ ) for each winding of  $p$  turns, to compare with the values found using (13) and Table II. For all types of current waveform shown in this article, there is no error greater than 4.0% for  $p=2$  and 1.74% for  $p>2$ , which validates Table II.

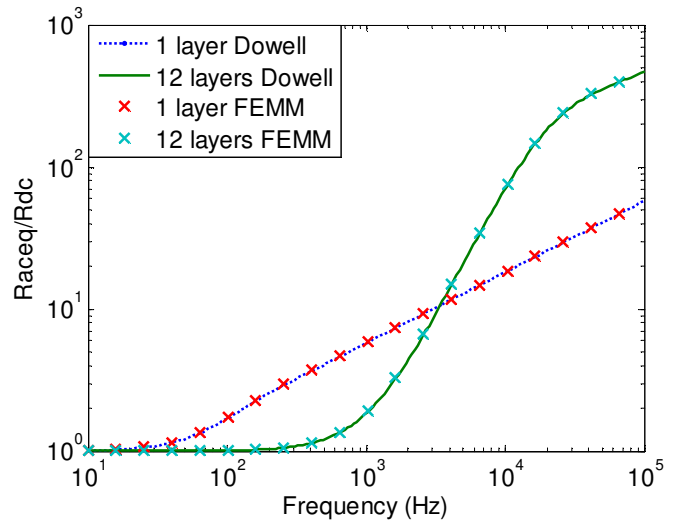


Fig. 9 – AC resistance variation by simulation in FEA software and by analytical calculation

#### CONCLUSIONS

Basing on Dowell's formula, constants which relate the number of turns of a winding, the fundamental frequency of the current passing through the winding and the winding area width were found for different type of current waveforms. These constants, shown in Table II, are an easy tool for transformer designers. They indicate the orientation of the conductors inside the winding area which results in less AC copper losses.

Since Dowell's formula is a 1-D approximation of a 3-D phenomenon, intensive 2-D FEA simulation was carried-out to verify the validity of the constants found. Insulation between conductors was not taken into account, as it is part of future work.

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