Design Considerations of A High Frequency Power Transformer

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***Abstract*—The approach of high frequency application seems to be very trendy in power transformer design. Especially when it is greatly efficient in reducing size, losses, cost etc. However, there is not much discussion that were made on how to design the high frequency power transformer using a proper approach, cost efficient and reliable to be implemented. So in this paper, some of those important parameters are briefly considered. The discussion is made throughout their geometries, core losses, winding losses and thermal model considerations accordingly.**

***Keywords*—high frequency, power transformer, ferrite core, loss.**

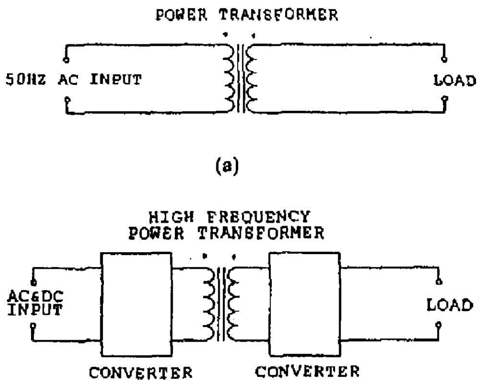
I. INTRODUCTION

A power transformer has been an indispensable component in power conversion systems. It is employed to perform several important functions such as voltage transformation, electrical isolation and noise decoupling within its primary and secondary terminal. The ideal power transformer should transfer the external electrical power source efficiently and instantaneously to its external load. However, at low operating frequencies, it is one of the heaviest and the most expensive equipment in an electrical distribution system. The concept of realizing a smaller size, lower losses, higher efficiency and smaller volume solid state transformer may be accomplish in one or a combination of this three ways: (1) by modifying the construction of open core and coil, (2) by increasing the working flux density of the core, or (3) by increasing the operating temperature of the coil [1]. However, only the second approach by means of higher frequency operation [2], utilizing higher working flux density materials are discussed in this publication.

II. THE FEATURE OF A HIGH FREQUENCY POWER TRANSFORMER

The conventional power transformer usually operated within a 50 Hz or 60 Hz generators. The design of these transformers usually consists of two or more windings wounded on a heavy and bulky laminated iron core. Even though they are simple to be built and implemented, they are always criticized as the most space consuming, heavy and costly part in the power supplies system.

For instant, we can found them to be the most apparent part in the power supplies and even some types of these power transformers can range up to several tons (used in power substations), which operate, based on these two frequencies. To overcome the problems, thus the high frequency power transformer is introduced. The comparison between conventional and the high frequency power transformer system can be visualize as in Figure 1 below.



(a)

(b)

Fig. 1 The different system arrangement of (a) Conventional power transformer and (b) High frequency power transformer.

Even though the design of the high frequency power transformer will follow the same approach as the conventional, some added converter topologies have to be assigned. At least one converter is needed to shape the source into a high frequency waveform before the high power transformer input, and one more converter just before the load.

Basically, when the source frequency applied to any transformer is increased, the inductive reactance of the transformer windings also will be increased. This situation would cause a greater ac voltage drop across the windings compared to the load. Yet, an increment of the frequency should not spoil the transformer. But, if the source frequency that applied to the transformer is decreased, then reactance of the windings is decreased followed by an increment of the flowing current through the windings. The further decrement in frequency will make the current exceeds its rated value and burnout the transformer. For this reason, a transformer is still save to be operated above its normal operating frequency, but should not be below that.

General relation between voltage (E), flux density (B) and frequency (f) is originate from the Faraday's law [3]:

$$E = kB_M NA_c f x\ 10^{-8} \qquad (1)$$

Where

E	is voltage applied to the primary winding (V)
k	is 4.0 (for rectangular waves)
B_M	is maximum flux density
N	is no. of primary turns
A_c	is the core cross sectional area (cm^2)
f	is the frequency of operation (Hz)

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The formula shown that, at the same input voltage level, modification or improvement can be made on parameters of B_M, N, Ac, and f. Cost and size would been the matters if improvement is based on B_M, N and Ac optimization. But when f is to be considered, the only care to be taken is on its proper selection of materials, geometry and capacity of the ready market available cores, together with some heat analysis of its losses during high frequency operation. When it is up to a certain frequency, all the losses components that involved in high frequency transformer is highly recommended to be considered individually [2], [4-8]. The transformer equivalent circuit for a high frequency analysis [2] consists of: ideal transformer (N_{prim}, N_{sec}), loss components (R_{prim}, R_{sec} ,R_{core}), magnetizing inductance (L_m) and parasitic components (L_{leak_prim}, L_{leak_sec}, C_{prim}, C_{sec}, $C_{prim\text{-}sec}$) is shown in Figure 2.

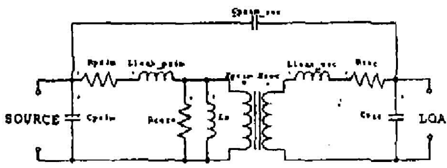

Fig. 2. 2-winding transformer equivalent circuit

Where

N_{prim} = number of primary turns
N_{sec} = number of secondary turns
R_{core} = core loss resistance
R_{prim} = primary resistance
R_{sec} = secondary resistance
L_m = magnetizing inductance
L_{leak_prim} = primary leakage inductance
L_{leak_sec} = secondary leakage inductance
C_{prim} = primary intrawinding capacitance
C_{prim_sec} = primary to secondary capacitance
C_{sec} = secondary intrawinding capacitance

III TRANSFORMER CORE SELECTION

A. Material selections

In making a selection for the transformer core, the first thing to be considered is the material of the core itself. An ideal core material should have a high flux density, low coercive force and high permeability as shown in Figure 3.

While, to reach these ideal material characteristics, certain core geometry should have a minimum air gap such as; strip wound or ferrite toroids and ring or DU laminations [9].

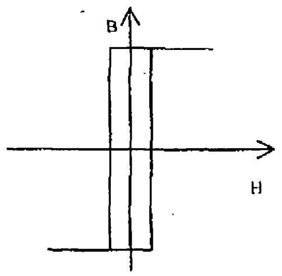

Fig. 3. Ideal B-H loops for magnetic core

While Figure 4 shows a sample of a typical B-H loop of Square Permalloy 80 material and describes the various parameters that are normally tested and defined by the core manufacturers [10].

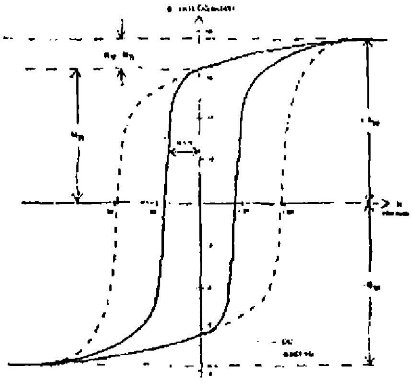

Fig. 4. Hysteresis Loops for .002" Thick Permalloy 80

The B_M represents the core saturation flux density; the higher the flux density, the smaller the size of the transformer will be in a particular design. B_M minus B_R is the difference between the maximum flux density (B_M) and the residual flux density (B_R). The smaller the number, the lower the permeability in saturation and the lower the switching losses for a given core material. While, H1/3 defines the width of the B-H loop when the core is reset 1/3 of the way from positive saturation (negative dc bias applied). H1/3 relates to the core loss; the narrower the H1/3 the lower the core loss. These two are the primarily interested parameters discussed in most inverters transformer design [11- 17].

Also shown in Figure 4 is a B-H loop for the same material at 6,000 Hz. It can be seen that, when the width expands, the core loss will increases with the increasing of frequency. The increase in core loss also depends on the strip thickness and the resistivity of the core material used. The higher the resistivity of the core material, the less core loss increases with increasing frequency for a given material thickness.

Table 1 listed the summary of various materials for saturable transformer that are available in the market. Also listed is their recommended frequency of operation [18]. Whereas Figure 5 shows the B-H loops of those saturable core materials:

TABLE 1
SATURABLE CORE MATERIAL CHARACTERISTIC

METAL OR ALLOY	COMPOSITION IN % (Balance Iron)	FLUX DENSITY (Kilogauss)	CURIE TEMP. OC	Recommended Maximum Frequency for Power Transformer
Silicon-iron	3.3% Si (Grain Oriented)	18	750	100-5000Hz
Cobalt-Iron	49% Co 2% V (Magnetic Anneal)	22	940	800-3000Hz
Nickel-Iron	45-50% Ni (Grain Oriented)	15	500	100-30 000Hz
Nickel-Iron	79% Ni 4% Mo	7	460	400-150 000Hz

Amorphous (iron base)	3.5% Si 13.5% B 2% C	16	370	25 000Hz
Amorphous (Cobalt-base)	66% Co 15% Si 14% B	5	205	500 000Hz
Ferrites	MnZn Permeability equals 5000	4	170	33 000Hz
Ferrites	MnZn Permeability equals 2300	4	170	50 000Hz

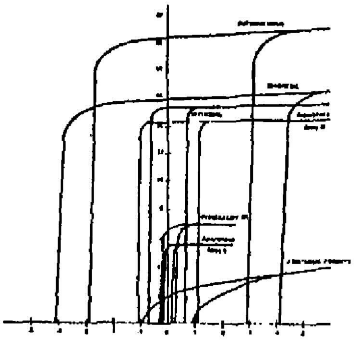

Fig. 5. B-H Loops of Saturable Materials

As can be seen from the Table 1 and Figure 5, when size is the most important consideration in the design, Supermendur (cobalt-iron) material, hold the highest flux density compare to others, therefore it would give the smallest core size. While the ferrite material, having the lowest flux density, would be the largest. Whilst, Amorphous Alloy E has the lowest coercive force, therefore it was the lowest core loss of any other core materials available. However, this figure of dc B-H loops does not necessarily give the true picture of what happens to core loss at higher frequencies [9].

Ideally the transformers are designed for smallest size, highest efficiency, lowest cost, and widest environmental conditions. Unfortunately, the material producing the smallest size may have the poorest efficiency and highest cost. Materials giving the highest efficiency may have high cost and large size. Therefore, there would be some trade-off between the characteristics of the material selection.

B. Core Size Selection

After selecting the transformer core material and material thickness, the next step that should be considered is the selection of core size within the operating frequency and output power. The core size also should be selected so that the wire size would completely fill the window opening to reduce the excessive core losses [19,13-15] due to the unnecessary magnetic path length of the core. Therefore, it is better in a certain cases to select a core with a smaller diameter, but with the same cross-sectional area, to insure that the windings will completely fill the core window. The power handling capability of a core can easily verify by this core area product ($W_A A_C$). A rough indication of the required area product is given by [3]

$$W_A A_C = [P_O/(K \Delta B f_{sw})]^{4/3} \qquad (2)$$

Where

W_A is the core window area (circular mils)

A_C is the core effective cross sectional area (cm²)

P_O output power (W)

ΔB is the flux density swing (Tesla)

f_{sw} is the switching frequency (Hz)

K is the winding factor

The curves in Figure 6 shows the required $W_A A_C$ product for some of the common core materials plotted against transformer output power for a given frequency [10].

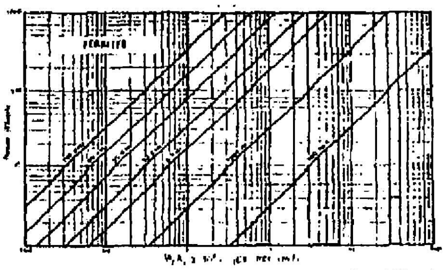

Fig. 6. Relationship of Core $W_A A_C$ to Output Power Capability for ferrite materials

C. Core Geometries

For a high frequency application, the most efficient and economical cores are those made up from ferrite materials [2,6,19-21]. There are many types of ferrite core geometries and shapes that has been applied and analyzed for their characteristic and advantages. Most of their dimensions follow IEC standards so that there is interchangeability between manufacturers. From the analysis and report made by most of the designers, it can assume that each of the geometries come with their own advantages. But the selection depends totally to the designer those matched it with their aims and application:

The Pot Core types, when assembled, are found to surround nearly the entire wound bobbin. This type will aids in shielding the coil from absorption of an outside sources EMI. However, they are more expensive compare to other shapes of the same size. But pot cores for high power applications (>1kW) are still not available.

The Double Slab cores are in advantage of having the large openings of the core, to allow large size of wires to be accommodated and also assist in removing heat from the assembly. While RM cores are also similar to pot cores, but with their design, they are found to minimize board space at least 40% savings in mounting area [18]. Their solid center post also helped in generates less core loss, therefore minimizing their heat buildup.

However, when compare in price, E cores are less expensive than pot cores. They are also reported to have the advantages of simple bobbin winding plus easy assembly. Unfortunately, E cores do not offer self-shielding. These E cores can also be pressed into a different thickness, providing a selection of cross-sectional areas. Bobbins for these different cross sectional areas are often available commercially. E cores become a popular shapes due to their lower cost, ease of assembly and winding.

EC, ETD and EER cores come with shapes look like across between E cores and pot cores. Like E cores, they provide a wide opening on each side. This gives adequate space for the large size wires required for low output voltage switched mode power supplies. It also allows for an airflow, which keeps the assembly cooler. The center post is round, same like the pot core. One of the obvious advantages of the round center post is that the winding has a shorter path length around it (11% shorter) than the wire around a square center post with an equal area. This reduces the losses of the windings by 11% [10] and enables the core to handle a higher output power. The round center post also eliminates the sharp bend in the wire that occurs with winding on a square center post.

PQ cores have been the chosen one in the design of switched mode power supplies. Providing an optimized ratio of volume to winding area and surface area is their principal advantage. As a result, it would provide a maximum power output with a minimum assembled transformer weight and volume and taking up a minimum amount of area on the printed circuit board.

EP Cores with a round center-post cubical shape, are found to enclose the coil completely except for the printed circuit board terminals. The particular shape minimizes the effect of air gaps formed at the mating surfaces of the magnetic path and provides a larger volume ratio to total space used. Shielding is most excellent in this EP core.

Toroids are reported to be the most economical in manufacturing process, hence they are least costly of all comparable core shapes. Since no bobbin is required, accessory and assembly costs are nil. Their shielding is found to be relatively good.

The summary of the parameters is shown in Table 2.

TABLE II
SUMMARIES OF FERRITE CORE COMPARATIVE GEOMETRY CONSIDERATIONS

	Pot Core	Double slab, RM Cores	E Core	EC, ETD, EER Cores	PQ Core	EP Core	Toroid
Core Cost	H	H	L	M	H	M	V
Bobin Cost	L	L	L	M	H	H	N
Winding Cost	L	L	L	L	L	L	H
Winding Flexibility	G	G	E	E	G	G	F
Assembly	S	S	S	M	S	S	N
Mounting Flexibility	G	G	G	F	F	G	P
Heat dissipation	P	G	E	G	G	P	G
Shielding	E	G	P	P	F	E	G

**V- Very Low, L- Low, M- Medium, H- High, S- Simple, N- None, F- Fair, P- Poor, E- Excellent, G- Good

IV. LOSSES AND RATING CAPACITY

Performances of the power transformers are also limited by their losses and temperature rise limit. In a consumer or industrial applications, a transformer temperature rise of 40-50°C may be acceptable, resulting in a maximum internal temperature of 100°C [9]. Losses of the transformer are quite difficult to predict with accuracy especially when they operate within a higher frequency range. The data from core manufacturers are always undependable, partly because measurements are made only under sinusoidal drive conditions. For instance, a low frequency winding losses are easy to calculate, but high frequency eddy current losses are difficult to determine accurately, because of the high frequency harmonic content of the switched rectangular current wave shape. Therefore, general losses consideration made by most of the designers are reviewed here:

A. Core Losses

Transformer design generally is a quiet simple task, but involving a lot of parameters to be considered on. However, some of the designers had simplified the calculation in their design [4,6,12,17,22]. The total losses at flux densities below saturation are considered as the sum of three-loss mechanism: hyterisis, eddy current and residual current. In his paper, Petkov [17] had used the combination between the accuracy and simplicity of his curve fitting method, where this equation of total core loss is applied:

$$P_c = K_1 * f^{K2} * B^{K3} \tag{3}$$

Where

P_c	is the core power loss density (kW/m^3)
f	is the frequency (Hz)
B	is the flux density (T)
K_1, K_2, K_3	is the curve fitting formula constants[17].

Thus in the case of 3C80 grade material, the values are given as $K_1 = 16.7$, $K_2 = 1.3$, $K_3 = 2.5$ at the frequency range between 10 to 100 kHz. And K2 is found to increase up to 1.65 when the frequency raised to 200 kHz [23].

While others, calculate the losses separately with some added coefficients. Sippola [2] for instant, quoted a *hysteresis loss* modeled early by Mulder [11] using Steinmetz equation of:

$$P_{hyst} = K_1 * f^{K2} * B^{K3} (ct_2 T^2 - ct_1 T + ct) \tag{5}$$

Where

P_{hyst}	Hysteresis power loss density (kW/m^3)
Ct, ct1, ct2	temperature coefficients
T	temperature (^{0}C)

Since the loss constants K are only applicable for sinusoidal magnetizing excitation, only, Albach *et al.* [14] extended them for a non-sinusoidal magnetizing currents by the concept of equivalent frequency

$$P_{hyst} = (1/\tau)\, K_1 * f_{eq}^{K2-1} * B^{K3} (ct_2 T^2 - ct_1 T + ct) \tag{6}$$

Where

τ = switching period (s),

f_{eq} = equivalent frequency (Hz).

The equivalent frequency can be calculated from the instantaneous flux density changes during the total switching period [2].

To obtain more accurate total core loss, the *eddy current loss* that are induced to the core by the changing magnetic flux is also should be included [23]. Sippola [2] also has expended the equation disscuss by Mulder [11] to match it with the rectangular high frequency switching pattern:

$$P_{eddy} = (U_{in} D / 2N_{prim})^2 l_e \pi / 4\rho \tag{7}$$

where

P_{eddy}	is eddy current loss density (kW/m^3)
U_{in}	is the input voltage
D	is duty cycle
N_{prim}	is the no. of primary turns
l_e	is the magnetic flux path length (m)
ρ	is the resistivity of the core (Ωm)

B. Winding Losses

Winding losses of two winding transformer carrying an AC current can be considered as:

$$P_w = I_p^2 R_p + I_s^2 R_s \qquad (8)$$

where

P_w total winding losses (W)
I_p, I_s primary and secondary rms current (A)
R_p, R_s primary and secondary dc-resistance (Ω)

However for a simplified discussion or analysis, ones can consider the followed equation, which has been demonstrate by Petkov [17]:

$$P_w = K_r * \rho * I^2_{rms} / A_c \qquad (9)$$

Where

P_w is conductor copper or winding loss (W)
Kr is an AC- resistance coefficient (Rac/Rdc)
ρ is the core permeability
A_c is the core cross sectional area

Actually, there are many papers have been published regarding this subject. Started basically by Dowell [22] and then the extended analysis that had followed for a better and accurate interpretation. However, is has been reported that by using Dowel's interpretation above, is more accurate especially for the case of high power transformer analysis[17].

However, when each of the harmonic components are to be included, it seems like Sepponen [2] had mentioned the best best winding losses equation to be reffered:

$$P_w = \sum_{h=0}^{\infty} \sigma \frac{l_w N^2_p}{A_w \alpha} \left(\frac{1}{n(1-n)} \right) I^2_{ph} \left(\frac{R_{ac}}{R_{dc}} \right)_h \qquad (10)$$

Where

σ is the resistivity of copper (Ωm)
lw is the average length of winding turn (m)
N_p is the number of primary turn
A_w is the total winding are (m2)
α is the fill factor
n is the ratio of primary winding to total winding area
I_{ph} is a primary rms current at h^{th} harmonic (A)
$(R_{ac}/R_{dc})_h$ is the ratio of winding AC and DC resistance at the frequency of h^{th} harmonic component
h is the index for harmonic components (DC=0).

C. Thermal Modeling

Temperature rise on the transformer depends not only upon its losses, but also on the thermal resistances, type of geometries and mounting condition itself. Thermal resistances has been the key parameter discussed in many publications, and was a very difficult to define with a reasonable degree of accuracy. Generally there are three components of heat transfer mechanisms [2] (conduction, convection and radiation) as illustrated in Figure 7. However, the linear thermal resistance analysis is valid only for the conductive heat transfer analysis. While the convective and radiative heat transfer are nonlinear and depend on the instantaneous temperature differences between the object and its ambient.

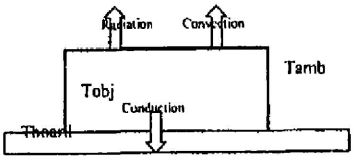

Fig. 7. Heat transfer mechanism for an abject mounted on a board.

The heat conducted, P_{cond} from the object to mounting board, PCB for instant is:

$$P_{cond} = R_{obj\_bd} (T_{obj} - T_{bd}) \qquad (11)$$

where

$R_{obj\text{-}bd}$ is the conductive thermal resistance from object to board (K/W)
T_{obj} is the object temperature (K)
T_{bd} is the printed circuit board (PCB) temperature (K)

In *convection* the heat is transferred by a moving fluid or air (*natural convection*) or for example by a fan (*forced convection*). Convective heat transfer capacity includes non-linear heat transfer coefficient *β*:

$$P_{conv} = \beta A (T_{obj} - T_{amb}) \qquad (12)$$

where

β is the convective heat transfer coefficient
A is the surface area (m^2)
T_{amb} is the ambient temperature (K)

Radiative heat exchange between object and ambient can be shown as:

$$P_{rad} = \sigma B \epsilon A (T^4_{obj} - T^4_{amb}) \qquad (13)$$

where

σB is the Boltzmann coefficient
ϵ is the emissivity

V. DISCUSSION

An increasing in frequency operation of the transformer has called for a better design consideration. The materials, geometries, and the losses analysis have to be considered properly. Any analysis of high frequency transformer cannot be started from the overall design usually quoted only, since they involved empirical assumptions that would not remain constant as the frequency is increased [23]. Instead, it is more preferable to treat all the analysis of the components separately.

For this high frequency transformer operation(>50Hz, typ. 1kHz - 1MHz) power transformer, quite numbers of papers have been presented. High frequency design (i.e. analysis of losses and temperature rise) has attained considerable attention in the literature as well. However it is acceptable to neglect the DC bias component and the losses caused by the magnetizing current for the means of simplicity. Typically, core losses can be modeled using Steinmetz equation, additional harmonic analysis and form factor correction [2]. While winding losses can be modeled using DC-resistance [17], analytical expressions for winding AC resistance [2-4,17] or Finite Element Method (FEM) derived winding AC resistance factor [14]. Heat transfer can been modeled as a constant thermal resistance [13,16,17] or as a non-linear heat transfer coefficient [7].

VI. CONCLUSION

The design of transformer in a higher frequency operation is quiet complex. Since there are so many parameters to be look at and most of them involved with the non-sinusoidal load current that need a proper analysis to be performed. The attention should be given properly on their losses distribution and the hot-spot temperature rise. However, the selection of frequency range also should be considered consequently for it to be worth enough for the whole design. Also, the result of the theoretical design should be verified with the designed transformer that are examined under the worst-case conditions which the power supply is expected to operate over a long periods of time, rather than in a transient conditions for its efficiency proof.

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VIII. BIOGRAPHIES

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