Influence of Interleaving Winding Arrangement on Leakage Inductance and Winding Loss of High-frequency Transformer

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Primary and secondary interleaving is normally used to minimize the leakage inductance and winding losses in high-frequency (HF) transformers used in high-power isolated dc-dc converters. Higher switching frequencies lead to the skin and proximity effects more significant, further more, the reduction of leakage inductance and the increase of winding losses. After analyzing the nonlinear distributions of the magnetic field strength and current density along the conductor thickness at high frequency, this paper presents a novel analytical calculation method considering the strong frequency dependence of transformer parameters, which is suitable for the non-interleaving, partially-interleaving, and fully-interleaving winding arrangements. A 4.5kHz, 5kVA HF transformer test model with amorphous core is built, and finite-element method (FEM) simulation and measurement are carried out to verify the method.

***Index Terms*—HF transformer, interleaving winding, leakage inductance, winding loss, frequency-dependent parameters.**

I. INTRODUCTION

IN the low-frequency range, the current density distribution in transformer's winding is nearly uniform and the leakage inductance and the winding loss are equal to the dc parameters. However, with higher frequencies, the leakage inductance is reduced and the winding losses are increased mainly due to enhanced skin and proximity effects. In the optimization of magnetic components designing, the primary and secondary winding of HF transformer are usually sectioned and interleaved to reduce the proximity effect, further the leakage inductance and winding loss. To address the challenge of optimal HF transformer design, this work proposes an novel analytical calculation method of transformer parameters considering the strong frequency-dependent characteristics, which is suitable for non-interleaving, partially-interleaving and fully-interleaving winding arrangements. FEA simulation and measurement validate the method.

II. LEAKAGE INDUCTANCE DERIVATION

Based on the one-dimensional (1-D) Maxwell equations, the nonlinear distribution of the magnetic field strength inside the nth layer along the conductor thickness is obtained as follows

$$H_z^n(x)=\frac{H_{\mathrm{out}}^n\sinh\left[\gamma(x-x_{\mathrm{in}}^n)\right]+H_{\mathrm{in}}^n\sinh\left[\gamma(x_{\mathrm{out}}^n-x)\right]}{\sinh(\gamma d_{\mathrm{coil}})} \quad (1)$$

The leakage inductance is then calculated by using the stored energy in the stray magnetic field.

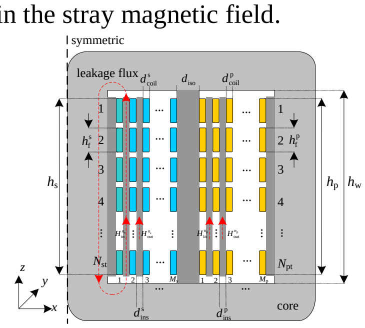


Fig. 1. Copper foils of concentric double windings HF transformer and boundary conditions.

For non-interleaved winding arrangement, the total leakage inductance referred to the primary side can be expressed by

$$\begin{aligned}L_\sigma^{\mathrm{p}}=\frac{\mu_0N_{\mathrm{pt}}^2M_{\mathrm{p}}}{h_{\mathrm{w}}}\cdot&\left\{\frac{(M_{\mathrm{p}}-1)(2M_{\mathrm{p}}-1)\mathrm{MLT}_{\mathrm{ins}}^{\mathrm{pri}}d_{\mathrm{ins}}^{\mathrm{p}}}{6}\right.\\&+\frac{M_{\mathrm{p}}(M_{\mathrm{s}}-1)(2M_{\mathrm{s}}-1)\mathrm{MLT}_{\mathrm{ins}}^{\mathrm{sec}}d_{\mathrm{ins}}^{\mathrm{s}}}{6M_{\mathrm{s}}}+M_{\mathrm{p}}\mathrm{MLT}_{\mathrm{iso}}d_{\mathrm{iso}}\\&+\frac{\mathrm{MLT}_{\mathrm{pri}}\left[k_{\mathrm{p1}}(2M_{\mathrm{p}}^2+1)+4k_{\mathrm{p2}}(M_{\mathrm{p}}^2-1)\right]}{12\gamma\sinh^2(\gamma d_{\mathrm{coil}}^{\mathrm{p}})}\\&\left.+\frac{\mathrm{MLT}_{\mathrm{sec}}M_{\mathrm{p}}\left[k_{\mathrm{s1}}(2M_{\mathrm{s}}^2+1)+4k_{\mathrm{s2}}(M_{\mathrm{s}}^2-1)\right]}{12\gamma M_{\mathrm{s}}\sinh^2(\gamma d_{\mathrm{coil}}^{\mathrm{s}})}\right\}\end{aligned} \quad (2)$$

where

$$k_{\mathrm{p1}}=\sinh(2d_{\mathrm{coil}}^{\mathrm{p}}\gamma)-2d_{\mathrm{coil}}^{\mathrm{p}}\gamma$$
$$k_{\mathrm{p2}}=d_{\mathrm{coil}}^{\mathrm{p}}\gamma\cosh(d_{\mathrm{coil}}^{\mathrm{p}}\gamma)-\sinh(d_{\mathrm{coil}}^{\mathrm{p}}\gamma)$$
$$k_{\mathrm{s1}}=\sinh(2d_{\mathrm{coil}}^{\mathrm{s}}\gamma)-2d_{\mathrm{coil}}^{\mathrm{s}}\gamma$$
$$k_{\mathrm{s2}}=d_{\mathrm{coil}}^{\mathrm{s}}\gamma\cosh(d_{\mathrm{coil}}^{\mathrm{s}}\gamma)-\sinh(d_{\mathrm{coil}}^{\mathrm{s}}\gamma).$$

For partially-interleaved winding arrangement, the total leakage inductance referred to the primary side is

$$\begin{aligned}L_\sigma^{\mathrm{p}}=\frac{\mu_0N_{\mathrm{pt}}^2M_{\mathrm{p}}}{h_{\mathrm{w}}}&\left\{\frac{\mathrm{MLT}_{\mathrm{ins}}^{\mathrm{pri}}d_{\mathrm{ins}}^{\mathrm{p}}}{2}+\frac{\mathrm{MLT}_{\mathrm{ins}}^{\mathrm{sec}}d_{\mathrm{ins}}^{\mathrm{s}}}{2}+2\mathrm{MLT}_{\mathrm{iso}}d_{\mathrm{iso}}\right.\\&\left.+\frac{\mathrm{MLT}_{\mathrm{pri}}\left[5k_{\mathrm{p1}}+12k_{\mathrm{p2}}\right]}{12\gamma\sinh^2(\gamma d_{\mathrm{coil}}^{\mathrm{p}})}+\frac{\mathrm{MLT}_{\mathrm{sec}}\left[5k_{\mathrm{s1}}+12k_{\mathrm{s2}}\right]}{12\gamma\sinh^2(\gamma d_{\mathrm{coil}}^{\mathrm{s}})}\right\}.\end{aligned} \quad (3)$$

For fully-interleaved winding arrangement, the total leakage inductance referred to the primary side is

$$L_\sigma^{\mathrm{p}}=\frac{\mu_0N_{\mathrm{pt}}^2M_{\mathrm{p}}}{4\gamma h_{\mathrm{w}}}\left\{4\gamma\mathrm{MLT}_{\mathrm{iso}}d_{\mathrm{iso}}+\frac{k_{\mathrm{p1}}\mathrm{MLT}_{\mathrm{pri}}}{\sinh^2(\gamma d_{\mathrm{coil}}^{\mathrm{p}})}+\frac{k_{\mathrm{s1}}\mathrm{MLT}_{\mathrm{sec}}}{\sinh^2(\gamma d_{\mathrm{coil}}^{\mathrm{s}})}\right\}. \quad (4)$$

III. WINDING LOSS DERIVATION

The nonlinear distribution of the current density inside the nth layer along the conductor thickness is given by

$$J_y^n(x)=\gamma\cdot\frac{H_{\mathrm{in}}^n\cosh\left[\gamma(x_{\mathrm{out}}^n-x)\right]-H_{\mathrm{out}}^n\cosh\left[\gamma(x-x_{\mathrm{in}}^n)\right]}{\sinh(\gamma d_{\mathrm{coil}})} \quad (5)$$

The time-average power loss in the nth layer is

$$P=\frac{h_w}{2\sigma}\int_{x_{in}^n}^{x_{out}^n}\left[J_y^n(x)\right]^2 dx=\frac{h_w}{4\sigma\delta}\Bigg[\left|H_{out}^n-H_{in}^n\right|^2\cdot \frac{\sinh\Delta+\sin\Delta}{\cosh\Delta-\cos\Delta}+\left|H_{in}^n+H_{out}^n\right|^2\cdot\frac{\sinh\Delta-\sin\Delta}{\cosh\Delta+\cos\Delta}\Bigg] \tag{6}$$

Using the boundary conditions for H at the inner and outer surface of the nth winding layer, the winding loss for non-interleaved winding can be expressed by

$$P=\frac{Mh_w}{4\sigma\delta}\left[\left(\frac{I_m}{h_w}\right)^2 A_1(\Delta)+\frac{(4M^2-1)}{3}\left(\frac{I_m}{h_w}\right)^2 A_2(\Delta)\right] \tag{7}$$

where

$$A_1(\Delta)=\frac{\sinh\Delta+\sin\Delta}{\cosh\Delta-\cos\Delta},\quad A_2(\Delta)=\frac{\sinh\Delta-\sin\Delta}{\cosh\Delta+\cos\Delta}.$$

For partially-interleaved winding arrangement, the winding loss can be expressed by

$$P=\frac{Mh_w}{4\sigma\delta}\left[\left(\frac{I_m}{h_w}\right)^2 A_1(\Delta)+5\left(\frac{I_m}{h_w}\right)^2 A_2(\Delta)\right]. \tag{8}$$

For fully-interleaved winding arrangement, the winding loss can be expressed by

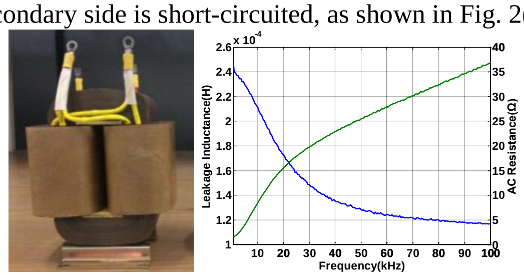
$$P=\frac{Mh_w}{4\sigma\delta}\left[\left(\frac{I_m}{h_w}\right)^2 A_1(\Delta)+\left(\frac{I_m}{h_w}\right)^2 A_2(\Delta)\right]. \tag{9}$$

It should be noted that the above analytical expression is suitable for round conductors when Δ is substituted by

$$\Delta' =\left(\frac{\pi}{4}\right)^{\frac{3}{4}}\frac{d_r}{\delta}\sqrt{\frac{N_t d_r}{h_w}}. \tag{10}$$

IV. Test Model and Calculation Method Verification

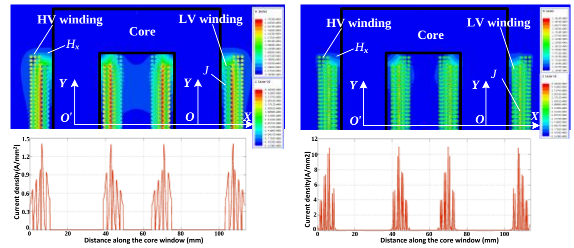
A 5kVA/4.5kHz HF transformer test model has been designed and manufactured. Photo and 2-D schematic diagram of the test model are shown in Fig. 2(a). The impendence of the test model has been measured from the primary side while the secondary side is short-circuited, as shown in Fig. 2(b).



(a) (b)

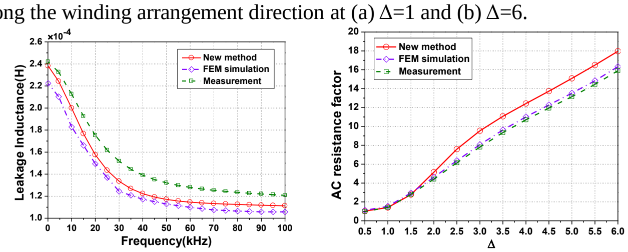
Fig. 2. (a) Test model of High-frequency transformer. (b) Measurement values of leakage inductance and ac resistance.

The leakage inductance refered to the primary side and ac resistance of the primary winding in a wide frequency band is analyzed and calculated by using the 2-D simulation model and the finite-element method. The distribution of the current density J inside all the conductors and the leakage magnetic field strength H between the layers of windings at Δ=1 and 6 is illustrated in the top part of Fig. 3(a) and (b), respectively. Fig. 4(a) and (b) shows a comparison for leakage inductance and ac resistance factor obtained by new methods, FEA simulation and measurement in the wide frequency band, respectively.



(a) Δ=1 (b) Δ=6

Fig. 3. Leakage magnetic field distribution, current density and its strength along the winding arrangement direction at (a) Δ=1 and (b) Δ=6.

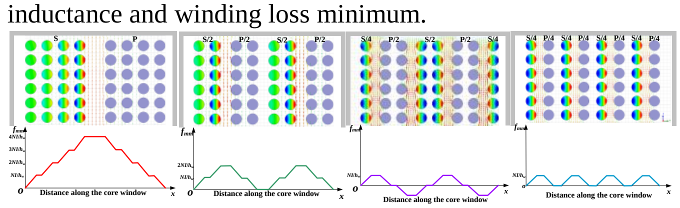


(a) (b)

Fig. 4. Leakage inductance refered to the primary winding and ac resistance factor of the primary winding obtained by the new method, FEA simulation, and experimental method.

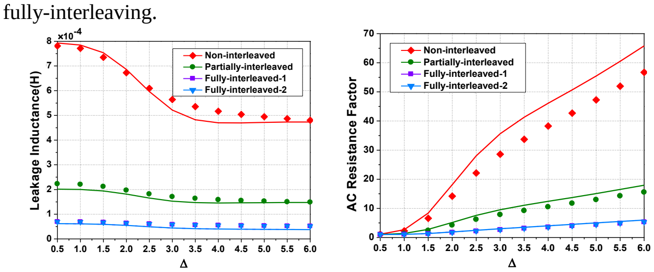
V. Validity Investigation for Winding Interleaving Configuration

The proximity effect is partially weakened when the partially-interleaving winding in Fig. 5(b) is utilized, and the maximum f_{mm} inside the insulation layers has been halved. Moreover, the leakage inductance and winding loss are partially weakened. For the fully-interleaving winding shown in Fig. 5(c) and (d), the proximity effect is virtually eliminated, which results in four times less f_{mm} compared to the first arrangement mentioned earlier, which makes the leakage inductance and winding loss minimum.



(a) (b) (c) (d)

Fig. 5. Current density and leakage field distribution of typical winding configurations (a) non-interleaving, (b) partially-interleaving, and (c)~(d) fully-interleaving.



(a) (b)

Fig. 6. Comparison of (a) leakage inductance and (b) ac resistance factor under four winding configurations (Solid lines are analytical results, and points are FEA simulation results).

References

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