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## *Core Geometry Coefficient For Resonant Inductors**

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A resonant inductor is required to have a small size, a low power loss, and good heat dissipation. In particular, it is difficult to design an optimal inductor for high-frequency and high-power applications. The design methods for resonant inductors presented until now are based on the trial-and-error approach. There is no criterion to pick the candidates for the core for resonant inductors from different core-product companies.

The core geometry coefficient ($K_g$) is one of the useful criteria to select the core [1]-[5]. By using the $K_g$ method, it is possible to select the core satisfying the acceptable wire loss, the electromagnetic conditions, and a core area restriction. However, there are no considerations and examples of the design procedure of resonant inductors using the $K_g$ method.

This paper presents expressions of the core geometry coefficient for the resonant inductor design. Additionally, a design example of a resonant inductor for a class-E power amplifier is given. By using the proposed expressions, the core geometry coefficient is determined from the electrical specifications of the loaded quality factor of a series-resonant circuit, the output power, the operating frequency, the maximum flux density, and the maximum wire loss. It is a good criterion to select the core from different manufacturers.

### *Basic Theory And Derivation Of Core Geometry Coefficient $K_g$*

The inductance of an inductor with an air gap is given by

$$L = \frac{N^2}{\frac{l_g}{\mu_{0A_c}} + \frac{l_c}{\mu_r \mu_{cA_c}}} = \frac{\mu_0 N^2 A_c}{l_g + \frac{l_c}{\mu_r}}, \qquad (1)$$

where $N$ is the number of wire turns, $l_g$ is the are-gap length, $l_c$ is the core length, $A_c$ is the cross-sectional area of the magnetic core, $\mu_0$ is the free-space permeability, and $\mu_r$ is the relative permeability of a core material.

In general, it can be written that

$$Ni = B\left(\frac{l_g}{\mu_0} + \frac{l_c}{\mu_0 \mu_r}\right), \qquad (2)$$

where $B$ is the magnetic flux density. Hence,

$$B_m = \frac{\mu_0 N I_m}{\frac{l_c}{\mu_r} + l_g}, \qquad (3)$$

where $B_m$ and $I_m$ are the maximum flux density and the maximum inductor current. From (1) and (3), the number of turns is

$$N = \frac{LI_m}{A_c B_m}, \quad (4)$$

which is a requirement from electromagnetic point of view. The length of the winding wire is

$$l_w = Nl_T, \quad (5)$$

where $l_T$ is the mean length of a single turn (MLT). The dc winding resistance is

$$R_{wdc} = \frac{\rho_w l_w}{A_w} = \frac{\rho_w N l_T}{A_w}, \quad (6)$$

where $A_w$ and $\rho_w$ are the cross-sectional area of the winding bare wire and the resistivity of the copper, respectively. Therefore,

$$A_w = \frac{N \rho_w l_T}{R_{wdc}}. \quad (7)$$

The dc and low-frequency wire loss is

$$P_{wdc} = R_{wdc} I_{rms}^2, \quad (8)$$

where $I_{rms}$ is the root-mean-square value of the inductor current. Therefore, (7) is rewritten as

$$A_w = \frac{N \rho_w l_T I_{rms}^2}{P_{wdc}}. \quad (9)$$

The core window utilization factor is defined as the ratio of the cross-sectional area of the winding bare wire $A_{Cu}$ to the window cross-sectional area of a core $W_a$

$$K_u = \frac{A_{Cu}}{W_a} = \frac{NA_w}{W_a}. \quad (10)$$

From (9) and (10), we obtain

$$N^2 = \frac{K_u W_a P_{wdc}}{\rho_w l_T I_{rms}^2} \quad (11)$$

and

$$A_w^2 = \frac{K_u W_a \rho_w l_T I_{rm}^2}{P_{wdc}}. \tag{12}$$

These relations include both the wire-loss condition in (9) and the core-area condition in (10). The number of turns should also satisfy the electromagnetic condition in (4). Equating the right-hand sides of (4) and (11), the *core geometry coefficient* [1] is obtained as

$$K_g = \frac{W_a A_c^2 K_u}{l_T} = \frac{\rho_w L^2 I_m^2 I_{rms}^2}{P_{wdc} B_m^2} = \frac{2\rho_w L W_m I_{rms}^2}{P_{wdc} B_m^2}\ (\mathrm{m}^5), \tag{13}$$

where the maximum energy stored in the inductor is

$$W_m = \frac{L I_m^2}{2}. \tag{14}$$

It is convenient to express the dc wire loss as the ratio of the output power $P_{wdc} = \alpha P_o$, resulting in

$$K_g = \frac{\rho_w L^2 I_m^2 I_{rms}^2}{\alpha P_{o} B_m^2} = \frac{2\rho_w L W_m I_{rms}^2}{\alpha P_o B_m^2}. \tag{15}$$

The core geometry coefficient $K_g$ provides the core with a good combination of $W_a$, $A_c$, and $l_T$ satisfying the electromagnetic condition in (7), the dc-wire-loss condition in (9), and the core-area restriction in (10) simultaneously. In [2], $K_g$ of many cores at $K_u = 0.4$ is presented or can be calculated from the core dimensions.

## *Core Geometry Coefficient For Resonant Inductors*

We consider the design of resonant inductors conducting a sinusoidal current $i_L = I_m \sin \omega t$, where $\omega = 2\pi f$ is the angular frequency and $I_m = \sqrt{2} I_{rms}$ is the amplitude of the sinusoidal current. From (15), we obtain

$$K_g = \frac{\rho_w L^2 I_m^4}{2\alpha P_o B_m^2} = \frac{2\rho_w W_m^2}{\alpha P_o B_m^2}. \tag{16}$$

Generally, the loaded quality factor of a series-resonant $L - C - R$ circuit is defined as

$$Q_L = \frac{\omega L}{R} = \frac{\omega L I_m^2}{2 P_o} = \frac{\omega W_m}{P_o}. \tag{17}$$

Thus, the core geometry coefficient for resonant inductors is given by

$$K_g = \frac{2\rho_w Q_L^2 P_o}{\alpha \omega^2 B_m^2}. \tag{18}$$

By using the proposed expressions for $K_g$ in (16) and (18), we can select the core satisfying the conditions (7), (9), and (10), using only the electrical parameters.

### *Design Example*

This paper presents a design example of a resonant inductor for the class E resonant power amplifier [7], whose topology is shown in Fig. 1. We design the resonant inductor $L$ to meet the following specifications: $f = 100$ kHz, $P_o = 80$ W, $R_L = 70\,\Omega$, and $Q_L = 5$. From these specifications, we obtain $I_m = \sqrt{2P_o/R_L} = 1.51$ A, and $L = QR/\omega = 557\mu$H. The inductor specifications are: $J_m < 5\ \mathrm{A/mm^2}$, $K_u = 0.4$, and $B_m = 0.2$ T, where $J_m$ is the maximum current density of the wire. Design examples of the dc-feed inductor $L_c$ are given in [5].

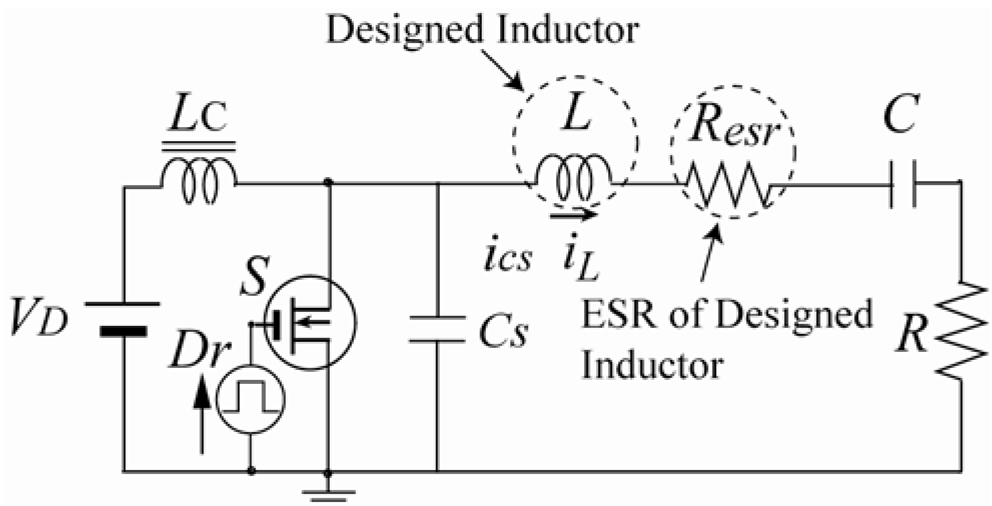


*Fig. 1. Class-E power amplifier.*

### Wire Loss, Core Loss, And Current Density

In the first design example, we allow 0.5% of the output power to be the wire loss in the inductor, that is, $\alpha = 0.005$. From (18), the core geometry coefficient increases with the decrease in the wire loss, which usually implies a large core volume. Therefore, the core loss increases with the decrease of the wire loss. The core loss is large in resonant inductors because of a high frequency and a large value of $I_m$. Therefore, the selection of the value of $\alpha$ is crucial in the design of the resonant inductors.

From (18), the core geometry coefficient is obtained as

$$K_g = \frac{2\rho_w Q_L^2 P_o}{\alpha \omega^2 B_m^2} = \frac{2 \times 1.72 \times 10^{-7} \times 5^2 \times 80}{0.005 \times 2 \times \pi \times 100 \times 10^3 \times 0.2^2} = 8.73 \times 10^{-13} (\mathrm{m^5}) = 0.00873\ \mathrm{cm^5}. \tag{19}$$

We select NEC/TOKIN FEE-30W core with the following parameters [2], [8]: $K_g = 0.0100\ \mathrm{cm^5}$, $A_c = 0.412\ \mathrm{cm^2}$, $W_a = 1.36\ \mathrm{cm^2}$, $l_T = 5.1$ cm, $l_c = 4.7$ cm, $V_c = 1934\ \mathrm{mm^3}$. From (12), the bare wire cross-sectional area is given by

$$A_W = \sqrt{\frac{K_u W_a \rho_w l_T I_m^2}{2\alpha P_o}}$$
$$= \sqrt{0.4 \times 1.36 \times 10^{-4} \times 1.72 \times 10^{-8}}$$
$$\times \sqrt{5.1 \times 10^{-2} \times 1.51^2}$$
$$/\sqrt{2 \times 0.005 \times 80} \quad (\mathrm{m}^2)$$
$$= 0.275\ \mathrm{mm}^2\ . \qquad (20)$$

The maximum current density of the wire is

$$J_m = \frac{I_m}{A_w} = \frac{1.51}{0.275 \times 10^{-6}} \quad (\mathrm{A/m^2})$$
$$= 5.49\ \mathrm{A/mm^2} > 5\ \mathrm{A/mm^2}. \qquad (21)$$

Since the current density is higher than that given by the regulation, the NEC/TOKIN FEI-25 core cannot be used for this inductor. The current-density restriction is rarely not satisfied and can be neglected for the design of inductors used in dc-dc converters in CCM and DCM [1]. This result indicates that the current density is one of the bottle-neck for the design of resonant inductors and we should carefully consider the current density restriction. For a low current density, a thick wire is required as shown in (20). A thicker wire yields a lower wire loss and smaller $\alpha$, which provides a higher $K_g$ value as shown in (18).

### Inductor Design

We reset $\alpha = 0.0025$ and obtain

$$K_g = \frac{2\rho_w Q_L^2 P_o}{\alpha \omega^2 B_m^2}$$
$$= \frac{2 \times 1.72 \times 10^{-7} \times 5^2 \times 80}{0.0025 \times 2 \times \pi \times 100 \times 10^3 \times 0.2^2}$$
$$= 1.75 \times 10^{-12} (\mathrm{m}^5) = 0.0175\ \mathrm{cm}^5. \qquad (22)$$

We reselect NEC/Tokin FEE-25W core with the following parameters [2], [8]: $K_g = 0.0204\ \mathrm{cm}^5$, $A_c = 0.417\ \mathrm{cm}^2$, $W_a = 1.49\ \mathrm{cm}^2$, $l_T = 5.1\ \mathrm{cm}$, $l_c = 7.2\ \mathrm{cm}$, $V_c = 3010\ \mathrm{mm}^3$. The core dimensions $C, D, E$ and $F$, defined in Fig. 2, are $C = 7\ \mathrm{mm}, D = 12.2\ \mathrm{mm}, E = 25.3\ \mathrm{mm}$, and $F = 6.8\ \mathrm{mm}$

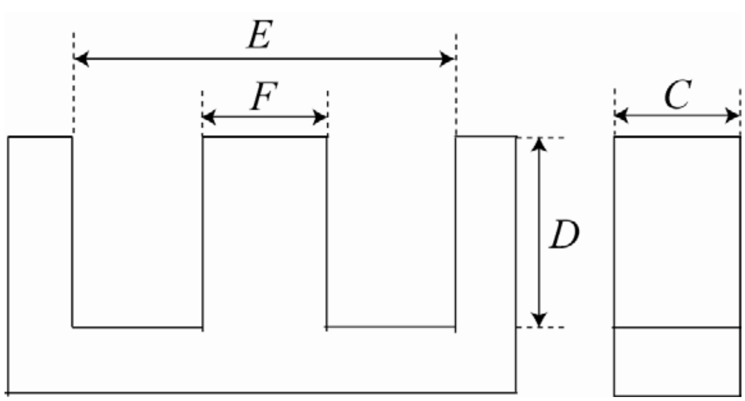


*Fig.2 Mechanical parameters of E core.*

Additionally, BH1 core material is chosen with $\mu_r = 2300 \pm 20\%$ and $P_v = 400\,\mathrm{kW/m^3}$ for $f = 100\,\mathrm{kHz}$ and $B_m = 0.2\,\mathrm{T}$, where $P_v$ is the core power loss per unit volume. Following the same procedure as that in (20) and (21), we can obtain

$$
\begin{aligned}
A_W &= \sqrt{\frac{K_u W_a \rho_w l_T I_m^2}{2\alpha P_o}} \\
&= \sqrt{0.4 \times 1.49 \times 10^{-4} \times 1.72 \times 10^{-8}} \\
&\quad \times \sqrt{7.2 \times 10^{-2} \times 1.51^2} \\
&\quad /\sqrt{2 \times 0.0025 \times 80} \quad (\mathrm{m^2}) \\
&= 0.542\ \mathrm{mm^2}
\end{aligned}
\qquad (23)
$$

and

$$
\begin{aligned}
J_m &= \frac{I_m}{A_w} = \frac{1.51}{0.542 \times 10^{-6}} \quad (\mathrm{A/m^2}) \\
&= 2.77\ \mathrm{A/mm^2} < 5\ \mathrm{A/mm^2}.
\end{aligned}
\qquad (24)
$$

This core satisfies the current-density restriction.

**Single-Strand Winding**. From the calculated cross-sectional area $A_w = 0.542\ \mathrm{mm^2}$ in (20), an AWG 19 copper wire with $A_w = 0.653\ \mathrm{mm^2}$ and a bare wire diameter $d = 0.912\ \mathrm{mm}$ is selected. Because the nominal outer diameter of the AWG 19 wire is $d_o = 0.98\ \mathrm{mm}$, the number of turns is obtained from

$$
\begin{aligned}
N &= \frac{K_u W_a}{A_w} \\
&= \frac{0.4 \times 1.49 \times 10^{-4}}{0.653 \times 10^{-6}} = 91.2\ \text{turns}.
\end{aligned}
\qquad (25)
$$

We pick $N = 91$ turns. For the adjustment of the inductance $L$, the air-gap length is calculated as

$$
\begin{aligned}
l_g &= \frac{\mu_0 A_c N^2}{L} - \frac{l_c}{\mu_r} \\
&= \frac{4 \times \pi \times 10^{-7} \times 0.417 \times 10^{-4} \times 91^2}{557 \times 10^{-6}} \\
&\quad - \frac{7.2 \times 10^{-2}}{2300} (\mathrm{m}) = 0.748\ \mathrm{mm},
\end{aligned}
\qquad (26)
$$

where $\mu_0 = 4\pi \times 10^{-7}\,\mathrm{H/m}$ is the free-space permeability.

Here, we consider the fringing effect. The fundamental theory of the fringing effect is given in [5]. It is assumed that the ratio of the effective width of the fringing flux cross-sectional area $w_f$ to the gap length is $u = w_f/l_g = 1$, and the ratio of the effective magnetic path length of the fringing flux to the gap length $k = 2$. The fringing flux factor $F_f$ is

$$F_f = 1 + \frac{2ul_g\left(C + F + 2ul_g\right)}{kCF} = 1 + 2 \times 1 \times 0.748 \times 10^{-3} \times \frac{(7 + 6.8 + 2 \times 1 \times 0.748) \times 10^{-3}}{2 \times 7 \times 6.8 \times 10^{-6}} = 1.24. \quad (27)$$

The number of turns including the fringing effect is

$$N' = \sqrt{\frac{L\left(l_g/F_f + l_c/\mu_r\right)}{\mu_0 A_c}} = \sqrt{557 \times 10^{-6}} \times \sqrt{\frac{0.748 \times 10^{-3}/1.24 + 7.2 \times 10^{-2}/2300}{4 \times \pi \times 10^{-7} \times 0.417 \times 10^{-4}}} = 82.1 \text{ turns}. \quad (28)$$

We pick an 82-turn winding for realizing the inductor. Therefore, the number of winding layers is

$$N_l = \frac{d_o N'}{2D} = \frac{0.98 \times 10^{-3} \times 82}{2 \times 12.2 \times 10^{-3}} = 3.29. \quad (29)$$

We need about a 3-layer winding to realize the inductor. The length of the winding wire is

$$l_w = N' l_T = 82 \times 5.1 \times 10^{-2} = 4.18 \text{ m}. \quad (30)$$

The low-frequency winding resistance, which is almost equal to the dc winding resistance, is given by

$$R_{wdc} = \frac{\rho_w l_w}{A_w} = \frac{1.72 \times 10^{-8} \times 4.64}{0.653 \times 10^{-6}} = 110\ \text{m}\Omega, \tag{31}$$

where $\rho_w = 1.72 \times 10^{-8}\ \Omega\text{m}$ is the resistivity of the copper at $T = 20\ °\text{C}$. The dc winding power loss without the skin and proximity effect is

$$P_{wdc} = \frac{R_{wdc} I_m^2}{2} = \frac{0.11 \times 1.51^2}{2} = 0.125\ \text{W}. \tag{32}$$

The skin depth of copper at $f = 100\ \text{kHz}$ is

$$\delta_w = \sqrt{\frac{\rho_w}{\pi \mu_0 f}} = \sqrt{\frac{1.72 \times 10^{-8}}{\pi \times 4 \times \pi \times 10^{-7} \times 100 \times 10^3}}\ (\text{m}) = 0.209\ \text{mm}. \tag{33}$$

We estimate the ac winding loss using Dowell's equation. The factor of Dowell's equation for a round wire, which is obtained in [4] and [6], is expressed as

$$A = \left(\frac{\pi}{4}\right)^{\frac{3}{4}} \frac{d}{\delta_w} \sqrt{\frac{d}{d_o}} = \left(\frac{\pi}{4}\right)^{\frac{3}{4}} \frac{0.912 \times 10^{-3}}{0.209 \times 10^{-3}} \sqrt{\frac{0.912 \times 10^{-3}}{0.98 \times 10^{-3}}} = 3.51. \tag{34}$$

By using $A = 3.51$, we obtain the winding ac-to-dc resistance ratio as

$$\begin{aligned} F_R &= A\left[\frac{\sinh(2A)+\sin(2A)}{\cosh(2A)-\cos(2A)}\right] \\ &\quad + \frac{2A(N_l^2-1)}{3}\left[\frac{\sinh(2A)-\sin(2A)}{\cosh(2A)+\cos(2A)}\right] \\ &= 3.51\left[\frac{\sinh(2\times 3.51)+\sin(2\times 3.51)}{\cosh(2\times 3.51)-\cos(2\times 3.51)}\right] \\ &\quad + \frac{2\times 3.51\times(3^2-1)}{3} \\ &\quad \times\left[\frac{\sinh(2\times 3.51)-\sin(2\times 3.51)}{\cosh(2\times 3.51)+\cos(2\times 3.51)}\right] \\ &= 22.2 \end{aligned} \quad (35)$$

The high-frequency ac resistance is

$$\begin{aligned} R_{wac} &= F_R R_{wdc} \\ &= 22.2\times 0.11 = 2.44\ \Omega. \end{aligned} \quad (36)$$

Therefore, the high-frequency ac winding power loss is

$$\begin{aligned} P_{wac} &= F_R P_{wdc} \\ &= 22.2\times 0.125 = 2.78\ \mathrm{W} \end{aligned} \quad (37)$$

The core power loss per unit volume at $f = 100\ \mathrm{kHz}$ is obtained from the catalog [8] as $P_v = 400\times 10^3\ \mathrm{W/m^3}$. Therefore, the total core loss is

$$\begin{aligned} P_c &= P_v V_c = 400\times 10^3\times 3010\times 10^{-9} \\ &= 1.20\ \mathrm{W}. \end{aligned} \quad (38)$$

The equivalent series resistance (ESR) representing the core loss is expressed as

$$\begin{aligned} R_c &= \frac{2P_c}{I_m^2} \\ &= \frac{2\times 1.2}{1.51^2} = 1.05\ \Omega. \end{aligned} \quad (39)$$

Therefore, the total power loss in the inductor is

$$P_{cw} = P_c + P_{wac} = 1.2 + 2.78 = 3.98 \text{ W}. \quad (40)$$

The equivalent series resistance of the inductor is

$$R_{esr} = R_{wac} + R_c = 2.44 + 1.05 = 3.49 \ \Omega. \quad (41)$$

The quality factor of the inductor is

$$Q = \frac{\omega L}{R_{esr}} = \frac{2 \times \pi \times 100 \times 10^3 \times 557 \times 10^{-6}}{3.49} = 100. \quad (41a)$$

**Multi-Strand Winding**. In this section, the design of the resonant inductor with multiple-strand winding is carried out to avoid the skin and proximity effects. For this purpose, the diameter of a single wire strand should be

$$d_s < 2\delta_w = \sqrt{\frac{\rho_w}{\pi \mu_0 f}} = \sqrt{\frac{1.724 \times 10^{-8}}{\pi \times 4 \times \pi \times 10^{-7} \times 100 \times 10^3}} \ \text{(m)} = 0.418 \text{ mm}. \quad (42)$$

The thickest wire less than $2\delta_w$ is AWG 26 with the bare wire diameter $d_s = 0.405 \text{ m}$, the insulated wire diameter $\delta_{so} = 0.452 \text{ mm}$, and the bare wire cross-sectional area $A_{ws} = 0.128 \text{ mm}^2$. The number of strands is

$$N_{strand} = \frac{A_w}{A_{ws}} = \frac{0.547}{0.128} = 4.2 \ . \quad (43)$$

The number of strands $N_{strand} = 4$ is selected. From (10), the number of turns is

$$N_s = \frac{K_u W_a}{A_{wo}} = \frac{K_u W_a}{\frac{N_{strand} \pi d_{so}^2}{4}} = \frac{0.4 \times 1.49 \times 10^{-4}}{\frac{4 \times \pi \times (0.452 \times 10^{-3})^2}{4}} = 92.7 \ . \quad (44)$$

We pick $N_s = 93$. The inductance value is adjusted by tuning the gap length, which is

$$l_{gs} = \frac{\mu_0 N^2 A_c}{L} - \frac{l_c}{\mu_r} = \frac{4 \times \pi \times 10^{-7} \times 93^2 \times 0.417 \times 10^{-4}}{557 \times 10^{-6}} - \frac{7.2 \times 10^{-2}}{2300} \ (\mathrm{m}) = \ 0.782 \ \mathrm{mm}. \quad (45)$$

The fringing flux factor $F_f$ is

$$F_{fs} = 1 + \frac{2ul_{gs}(C + F + 2ul_{gs})}{kCF} = 1 + 2 \times 1 \times 0.782 \times 10^{-3} \times \frac{(7 + 6.8 + 2 \times 1 \times \ 0.782) \times 10^{-3}}{2 \times 7 \times 6.8 \times 10^{-6}} = 1.25. \quad (46)$$

The number of turns including the fringing effect is

$$N_s{}' = \sqrt{\frac{L(l_{gs}/F_{fs} + l_c/\mu_r)}{\mu_0 A_c}} = \sqrt{557 \times 10^{-6}} \times \sqrt{\frac{0.782 \times 10^{-3}/1.25 + 7.2 \times 10^{-2}/2300}{4 \times \pi \times 10^{-7} \times 0.417 \times 10^{-4}}} = 83.6 \ \text{turns}. \quad (47)$$

We pick $N_s' = 84$. The multi-strand wire length is

$$\begin{aligned} l_{ws} &= N_s' l_T \\ &= 84 \times 5.1 \times 10^{-2} = 4.28 \text{ m.} \end{aligned} \tag{48}$$

Therefore, the dc winding resistance is

$$\begin{aligned} R_{wdcs} &= \frac{\rho_w l_{ws}}{A_{wo}} = \frac{\rho_w l_{ws}}{\frac{N_{strand} \pi d_{so}^2}{4}} \\ &= \frac{1.72 \times 10^{-8} \times 4.28}{\frac{4 \times \pi \times (0.452 \times 10^{-3})^2}{4}} = 115 \text{ m}\Omega. \end{aligned} \tag{49}$$

Because the skin and proximity effects can be neglected, the winding loss is

$$\begin{aligned} P_{ws} &= \frac{R_{wdcs} I_m^2}{2} \\ &= \frac{0.115 \times 1.51^2}{2} = 0.131 \text{ W.} \end{aligned} \tag{50}$$

Therefore, the total power loss in the inductor is

$$\begin{aligned} P_{cw} &= P_c + P_{ws} \\ &= 1.2 + 0.131 = 1.33 \text{ W.} \end{aligned} \tag{51}$$

The equivalent series resistance of the inductor is

$$\begin{aligned} R_{esrs} &= R_{wdcs} + R_c \\ &= 0.115 + 1.05 = 1.16 \ \Omega. \end{aligned} \tag{52}$$

The quality factor of the inductor with multiple-strand winding is

$$\begin{aligned} Q &= \frac{\omega L}{R_{esrs}} \\ &= \frac{2 \times \pi \times 100 \times 10^3 \times 557 \times 10^{-6}}{1.16} \\ &= 302. \end{aligned} \tag{53}$$

Table 1 gives the design results for both $\alpha = 0.005$ and $\alpha = 0.0025$. Since the example resonant inductor is designed for the high-frequency and high-power application, the current density restriction is a bottle-neck of the design. Additionally, it is shown that the core loss is much larger than the wire loss.

Table 1: Results for Resonant Inductor Designs for Multiple-Strand Winding.

| Required $P_w$ | Core | Material | $V_c$ (mm$^3$) | Wire | $N_{strand}$ | $N$ | $P_w$ (W) | $P_c$ (W) | $J_m$ (A/mm$^2$) |
|---|---|---|---|---|---|---|---|---|---|
| 0.4 | FEI25 | BH1 | 1934 | AWG26 | 2 | 83 | 0.402 | 0.77 | **5.49** |
| 0.2 | FEE25W | BH1 | 3010 | AWG26 | 4 | 84 | 0.115 | 1.20 | 2.77 |

## *Conclusion*

This paper has presented a novel core-selection criterion, taking into account the core loss. The proposed criterion indicates the core candidates, which satisfy the core-window-area restriction and the electromagnetic conditions with the guarantee of the permissible core-loss. It is confirmed from the design example that the designed inductor satisfies multiple restrictions simultaneously. The geometry coefficient $K_g$ method is recommended for designing inductors for resonant power converters [9] and PWM dc-dc power converters [10], [11].

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### About The Authors

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