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Design of RF-Choke Inductors Using Core Geometry Coefficient

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***Abstract*—This paper presents a derivation of the core geometry coefficient K_g for the design of RF-choke inductors and a design example. The design example indicates that the dc winding loss is the dominant component of the total power loss of RF-choke inductors. Additionally, the advantages and disadvantages of the K_g method are discussed by using the design example results.**

***Index Terms*—Inductor design, RF-choke inductor, core geometry coefficient, dc winding loss, ac winding loss, core loss, air-gap length, fringing effect, skin effect, Dowell's equation.**

INTRODUCTION

The magnetic component design is a traditional problem, but it is not solved satisfactorily in the power electronics field. For example, the RF-power amplifier, which is one of the important circuits in many systems, has two major magnetic components: a resonant inductor and an RF-choke inductor. In particular, most of the RF-power amplifiers, which are the class A, class B, class AB, class C, class E, and class F amplifiers, have an RF-choke inductor [1]. The improvements of some kinds of switching techniques, e.g., zero-voltage switching (ZVS), zero-current switching (ZCS), and class-E ZVS and zero derivative switching (ZDS), reduce dramatically power losses in switching devices. Therefore, the importance of the design of magnetic components with low power losses increases for the design of power amplifiers and dc-dc converters. Additionally, the reduction of the magnetic-component volume is also an important aspect because the magnetic components are bottlenecked to determine the circuit size. Generally, the power loss in a magnetic component decreases when the size of the inductor increases though both the low power loss and small volume are required. This means that there is a trade-off relationship between the power loss and the magnetic component size. Therefore, the design of magnetic components, which are inductors and transformers, is one of the important challenging problems.

There are several well-known strategies for selecting a core for the design of magnetic components, for example, the area product (A_p) method and the core geometry (K_g) method [2]–[5]. These methods are primarily used in the design of inductors and transformers for switching-mode power supplies (SMPS). The concept of the A_p approach is to select a proper core satisfying both the electromagnetic conditions and the restriction of the core window area. The inductance value is adjusted by the air-gap length. The A_p method is widely used for designing the inductors and transformers for dc-dc power converters operating in CCM and DCM. On the other hand, the concept of the K_g approach is to select a proper core satisfying the electromagnetic conditions, the restriction of the core window area, and the restriction of the winding loss, simultaneously. This method is useful to design inductors and transformers with low core and low ac winding losses. By setting the dc winding loss, we can choose the core, which requires shorter length and narrower cross-sectional area of the wire compared to that obtained with the A_p method. This leads to a small core volume.

An RF-choke inductor forces a dc current, which is applicable to dc-current sources and output filters of resonant converters. Since the current through the RF-choke inductor includes a low ac component, the ac-winding loss and the core loss are very small, even if the operating frequency is high. Therefore, it is possible to focus on the dc-winding loss when the RF-choke inductor is designed. Additionally, the inductance value of the RF choke inductor is usually high to force the dc current. Therefore, the volume reduction is quite important to make an RF-choke inductor. For the reasons described above, we consider that the K_g method is suitable to design the RF-choke inductor.

This paper presents a derivation of the core geometry coefficients K_g for the design of the RF-choke inductors and a design example. The design example indicates that the dc winding loss is dominant in the power losses of the RF-choke inductor. Additionally, the advantages and disadvantages of the K_g method are discussed by using the design example results.

I. BASIC THEORY AND DERIVATION OF CORE GEOMETRY COEFFICIENT K_g

A. *Expression for Current Through RF-Choke Inductor*

The purpose of a choke is to supply a dc current. In real circuits, however, there are a little ripple on the dc current. Therefore, we assume that the current through the inductor i_L is the sum of a direct current I_{dc} and a pure sinusoidal current with the fundamental frequency equal to the operating frequency f,

$$i_L = I_{dc} + \frac{I_{pp}}{2}\sin(\omega t) = I_{dc} + (I_m - Idc)\sin(\omega t), \quad (1)$$

where I_{pp} is peak-to-peak value of the ac component of the current, I_m is a maximum value of i_L and $\omega = 2\pi f$ is the angular frequency. In this case, the peak-to peak ripple ratio to the dc current through the inductor is defined as

$$\gamma_r \equiv \frac{\Delta i_L}{I_{dc}} = \frac{I_{pp}}{I_{dc}} = \frac{2(I_m - I_{dc})}{I_{dc}}. \quad (2)$$

B. Core Geometry Coefficient K_g for RF-Choke Inductors

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

$$L = \frac{N^2}{\dfrac{l_g}{\mu_0 A_c} + \dfrac{l_c}{\mu_r \mu_0 A_c}} = \frac{\mu_0 N^2 A_c}{l_g + \dfrac{l_c}{\mu_r}}, \quad (3)$$

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, $\mu_0 = 4\pi \times 10^{-7}$ H/m is the free-space permeability, and μ_r is the relative permeability of the core material.

In general, it can be written that

$$Ni = B\left(\frac{l_g}{\mu_0} + \frac{l_c}{\mu_0 \mu_r}\right), \quad (4)$$

where B is the flux density. Hence, the maximum flux density B_m is

$$B_m = \frac{\mu_0 N I_m}{\dfrac{l_c}{\mu_r} + l_g} = \frac{\mu_0 N I_{dc}(1 + \gamma_r/2)}{\dfrac{l_c}{\mu_r} + l_g}. \quad (5)$$

From (3) and (5), the number of turns is

$$N = \frac{L I_m}{A_c B_m} = \frac{L I_{dc}(1 + \gamma/2)}{A_c B_m}, \quad (6)$$

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

$$l_w = N l_T, \quad (7)$$

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

$$R_{wdc} = \frac{\rho_w l_w}{A_w} = \frac{\rho_w N l_T}{A_w}, \quad (8)$$

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

$$A_w = \frac{N \rho_w l_T}{R_{wdc}}. \quad (9)$$

The dc winding loss is

$$P_{wdc} = R_{wdc} I_{dc}^2. \quad (10)$$

Therefore, (9) is rewritten as

$$A_w = \frac{N \rho_w l_T I_{dc}^2}{P_{wdc}}. \quad (11)$$

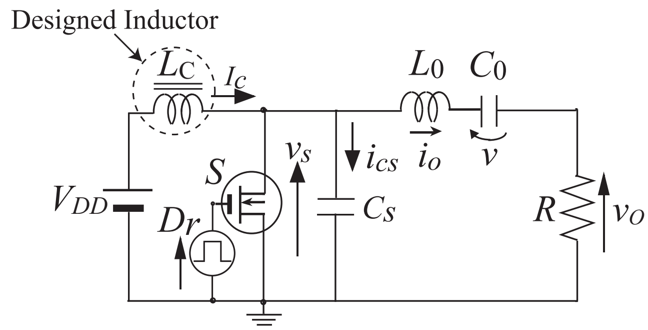

Fig. 1. Circuit topology of the class E amplifier.

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

$$K_u = \frac{A_{Cu}}{W_a} = \frac{N A_w}{W_a}. \quad (12)$$

From (11) and (12), we obtain

$$N^2 = \frac{K_u W_a P_{wdc}}{\rho_w l_T I_{dc}^2} \quad (13)$$

and

$$A_w^2 = \frac{K_u W_a \rho_w l_T I_{dc}^2}{P_{wdc}}. \quad (14)$$

These relations include both the dc winding loss condition in (11) and core-area condition in (12). The number of turns should also satisfy the electromagnetic condition in (6). Equating of the right-hand sides of (6) and (13), the *core geometry coefficient* [2] is obtained as

$$\begin{aligned} K_g &= \frac{W_a A_c^2 K_u}{l_T} \\ &= \frac{\rho_w L^2 I_m^2 I_{dc}^2}{P_{wdc} B_m^2} = \frac{\rho_w L^2 I_{dc}^4 (1 + \gamma/2)^2}{P_{wdc} B_m^2} \ (\mathrm{m}^5), \end{aligned} \quad (15)$$

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

$$K_g = \frac{\rho_w L^2 I_m^2 I_{dc}^2}{\alpha P_o B_m^2} = \frac{2\rho_w L I_{dc}^4 (1 + \gamma/2)^2}{\alpha P_o B_m^2}. \quad (16)$$

The core geometry coefficient K_g provides the core with a good combination of W_a, A_c, and l_T satisfying electromagnetic condition in (9), dc-winding-loss condition in (11), and core-area restriction in (12), simultaneously. The concrete values of K_g are presented for $K_u = 0.4$ in [3], or can be calculated from the core dimensions provided by the core manufacuturers. By using the proposed expressions for K_g in (15) and (16), we can select the core satisfying the conditions (9), (11), and (12) using only the electrical parameters.

II. Design Example

A. Design Specifications and Conditions

This section presents a design example of an RF-choke inductor for the class-E resonant power amplifier [6] whose topology is shown in Fig. 1. The specifications of the class-E amplifier are: operating frequency $f = 1$ MHz, load resistance $R_L = 10\ \Omega$, loaded quality factor $Q_L = 5$, and supply voltage $V_{DD} = 15$ V. By using the design procedure in [6], the inductance value of the RF-choke to achieve the ripple ratio $\gamma_r = 1$ % is $L_c = 1.13$ mH. In this case, the dc-current and the maximum current through the inductor is $I_{dc} = 0.807$ A and $I_m = I_{dc}(1 + \gamma_r/2) = 0.811$ A. The output power is $P_o = 11.8$ W.

The RF-choke inductor is designed to achieve the following specifications: 1) maximum current density of the wire is $J_m < 5$ A/m^2, 2) core window utilization factor is $K_u = 0.4$, 3) maximum flux density is less than the saturated flux density, $B_m < B_{sat} = 0.5$ T, and 4) dc-winding-loss coefficient is $\alpha = 0.005$.

B. Core Geometry Factor and Core Selection

From (16), the core geometry coefficient is obtained as

$$\begin{aligned} K_g &= \frac{\rho_w L_c^2 I_{dc}^4 (1+\gamma_r/2)^2}{\alpha P_o B_m^2} \\ &= \frac{1.72 \times 10^{-7} \times 1.13 \times 10^{-3}}{0.005 \times 11.8 \times 0.3^2} \\ &\quad \times 0.807^4 \times (1+0.01/2)^2 \\ &= 1.768 \times 10^{-12}\ \text{m}^5. \end{aligned} \tag{17}$$

We select Magnetic PQ-42020 core with the R ferrite material whose parameters are [3]: $\mu_r = 2300$, $K_g = 1.859 \times 10^{-12}$ m^5, $A_c = 0.58$ cm^2, $W_a = 0.6$ cm^2, $l_T = 4.3$ cm, $l_c = 4.5$ cm, height of core window $g = 1.4$ cm, and core volume $V_c = 2.61$ cm^3.

C. Wire Selection

From (14), the cross-sectional area of the bare wire is given by

$$\begin{aligned} A_w &= \sqrt{\frac{K_u W_a \rho_w l_T I_{dc}^2}{\alpha P_o}} \\ &= \sqrt{\frac{0.4 \times 0.6 \times 10^{-4} \times 1.72 \times 10^{-8}}{0.005 \times 11.8}} \\ &\quad \times \sqrt{\times 4.5 \times 10^{-2} \times 0.807^2}\ (\text{m}^2) \\ &= 0.442\ \text{mm}^2. \end{aligned} \tag{18}$$

From A_w, we pick the AWG 20 copper wire with $A_w = 0.519$ mm^2, inner diameter of bare wire cross-sectional area $d_i = 0.812$ mm, and outer diameter incuding insulation $d_o = 0.879$ mm. The maximum current density of the wire is

$$\begin{aligned} J_m &= \frac{I_{dc}(1+\gamma_r/2)}{A_w} \\ &= \frac{0.807 \times (1+0.01/2)}{0.518 \times 10^{-6}} \\ &= 1.56\ \text{A/m}^2 < 5. \end{aligned} \tag{19}$$

It is confirmed that the wire satisfies the maximum-current-density condition. In the inductor design using the core geometry coefficient, the maximum-current-density condition is not guranteed. Therefore, we need to confirm this condition. From (12), the number of turns is

$$N = \frac{K_u W_a}{A_w} = \frac{0.4 \times 0.6 \times 10^{-4}}{0.518 \times 10^{-6}} = 46.2. \tag{20}$$

We pick $N = 46$.

D. Air-Gap Length Considered with Fringing Effect

For the adjustment of the inductance L_0, the air-gap length is calculated as

$$\begin{aligned} l_g &= \frac{\mu_0 A_c N^2}{L_0} - \frac{l_c}{\mu_r} \\ &= \frac{4 \times \pi \times 10^{-7} \times 0.58 \times 10^{-4} \times 46^2}{1.13 \times 10^{-3}} \\ &\quad - \frac{4.5 \times 10^{-2}}{2300}\ (\text{m}) \\ &= 0.121\ \text{mm}. \end{aligned} \tag{21}$$

We shall choose a standard value of the air-gap length $l_g = 0.1$ mm. Here, we consider the fringing effect. The fundamental theory of the fringing effect is given in [4]. In this design example, we assume the ratio of the effective width of the fringing flux cross-sectional area to the gap length as $u = 1$. In this case, the fringing area A_f is

$$\begin{aligned} A_f &= \pi u l_g (2\sqrt{A_c/\pi} + u l_g) \\ &= \pi \times 1 \times 0.1 \times 10^{-3} \\ &\quad \times (2 \times \sqrt{0.58 \times 10^{-4}/\pi} + 1 \times 0.1 \times 10^{-3}) \\ &= 0.0288\ \text{cm}^2. \end{aligned} \tag{22}$$

It is assumed that the ratio of the effective magnetic path length of the fringing flux to the gap length $k = 2$. Therefore, the fringing factor F_f is obtained as

$$\begin{aligned} F_f &= 1 + \frac{A_f}{A_c k} \\ &= 1 + \frac{0.0288 \times 10^{-6}}{0.58 \times 10^{-6} \times 2} = 1.02. \end{aligned} \tag{23}$$

The number of turns $N = 46$ should be kept for achieving the specified K_u and B_m. Therefore, The estimated value of the inductance L_c is

$$\begin{aligned} L_c &= \frac{\mu_0 A_c N^2}{l_g/F_f + l_c/\mu_r} \\ &= \frac{4\times\pi\times10^{-7}\times0.58\times10^{-4}\times46^2}{0.1\times10^{-3}/1.02+2.13\times10^{-3}/2300} \\ &= 1.33 \text{ mH}, \end{aligned} \tag{24}$$

where L_0 is 18 % larger than the specified inductance. This difference appears due to the use of the standard length value of the air gap. The purpose of RF-choke inductor is to obtain the dc-current. Therefore large value of L_c is not problem for applying to the class-E switching amplifier.

E. Power Loss Estimation

From the height of the core, we can obtain the maximum number of turns per one winding layer N'

$$\begin{aligned} N\text{'} &= \frac{g}{d_o} \\ &= \frac{1.4\times10^{-2}}{0.879\times10^{-3}} = 15.9 \text{ turns/layer}. \end{aligned} \tag{25}$$

Therefore, 15 turns are allowed per one winding layer. The number of the winding layer N_l is

$$\begin{aligned} N_l &= \frac{N}{N\text{'}} \\ &= \frac{46}{15} = 3.06. \end{aligned} \tag{26}$$

Strictly speaking, we need 4-layer windings to realize the inductor. But 4th layer is only one turn. Therefore, we use $N_l = 3$ to estimate the winding loss. The length of the winding wire is

$$\begin{aligned} l_w &= N l_T \\ &= 46\times4.3\times10^{-2} = 1.98 \text{ m}. \end{aligned} \tag{27}$$

The dc winding resistance is

$$\begin{aligned} R_{wdc} &= \frac{\rho_w l_w}{A_w} \\ &= \frac{1.72\times10^{-8}\times1.02}{0.518\times10^{-6}}(\Omega) = 62.5 \text{ m}\Omega, \end{aligned} \tag{28}$$

where $\rho_w = 1.72\times10^{-8}$ Ωm is the resistivity of the copper at T =20 °C. The dc winding power loss is

$$\begin{aligned} P_{wdc} &= R_{wdc} I_{dc}^2 \\ &= 62.5\times10^{-3}\times0.807^2 \text{ (W)} = 50.2 \text{ mW}. \end{aligned} \tag{29}$$

The skin depth of copper at $f = 1$ MHz is,

$$\begin{aligned} \delta_w &= \frac{\rho_w}{\pi\mu_0 f} \\ &= \sqrt{\frac{1.724\times10^{-8}}{\pi\times4\times\pi\times10^{-7}\times10^6}} \text{ (m)} \\ &= 66.0\ \mu\text{m}. \end{aligned} \tag{30}$$

The porosity factor η_p is

$$\begin{aligned} \eta_p &= \frac{d_o N}{g N_l} \\ &= \frac{0.879\times10^{-3}\times46}{1.4\times10^{-2}\times3} = 0.94. \end{aligned} \tag{31}$$

We estimate the ac winding loss using Dowell's equation [4]. The factor of Dowell's equation for round wire is

$$\begin{aligned} A &= \left(\frac{\pi}{4}\right)^{\frac{3}{4}} \frac{d}{\delta_w}\sqrt{\eta_p} \\ &= \left(\frac{\pi}{4}\right)^{\frac{3}{4}} \times \frac{0.879\times10^{-3}}{66\times10^{-6}} \times \sqrt{0.94} = 10.8. \end{aligned} \tag{32}$$

By using the factor A, we obtain the winding ac-to-dc resistance ratio is

$$\begin{aligned} F_R &= A\left[\frac{\sinh(2A)+\sin(2A)}{\cosh(2A)-\cos(2A)}\right. \\ &\qquad \left.+\frac{2(N_l^2-1)}{3}\frac{\sinh(A)-\sin(A)}{\cosh(A)+\cos(A)}\right] \\ &= 10.8\times\left[\frac{\sinh(2\times10.8)+\sin(2\times10.8)}{\cosh(2\times10.8)-\cos(2\times10.8)}\right. \\ &\qquad \left.+\frac{2(3^2-1)}{3}\frac{\sinh(10.8)-\sin(10.8)}{\cosh(10.8)+\cos(10.8)}\right] \\ &= 64.6. \end{aligned} \tag{33}$$

The ac power loss is

$$\begin{aligned} P_{wac} &= F_R R_{wdc}(\gamma_r I_{dc}/2)^2 \\ &= 64.6\times33.9\times10^{-3} \\ &\quad \times(0.01\times0.807/2) \text{ (W)} \\ &= 0.04 \text{ (mW)}, \end{aligned} \tag{34}$$

which is neglegible compared with the dc winding loss because of a little ac component in the current through the inductor L_c.

The peak-to-peak value of the ac component of the flux density is

$$\begin{aligned} B_{pp} &= \frac{L_c\gamma_r I_{dc}}{N A_c} \\ &= \frac{1.33\times10^{-3}\times0.01\times0.807}{46\times0.58\times10^{-4}} \text{ (T)} = 4.02 \text{ mT}. \end{aligned} \tag{35}$$

The core power loss per unit volume is obtained from

$$P_v = a f^c (10 B_{pp}/2)^d, \tag{36}$$

where a, c, and d are coefficients to calculate the core power loss provided by the core manufacturers. Additionally, for the Magnetics R ferrite material, these coefficients for $f = 1$ MHz are $a = 0.00806$, $c = 1.66$, and $d = 2.68$. Therefore, the core power loss P_c is obtained

$$\begin{aligned} P_c &= P_v V_c = a f^c (10 B_{pp}/2)^d V_c \\ &= 0.00806\times(10^6)^{1.66} \\ &\quad \times(10\times4.02\times10^{-3}/2)^{2.68}\times2.61\times10^{-3}\text{(W)} \\ &= 0.203 \text{ mW}. \end{aligned} \tag{37}$$

Total power loss of the inductor is

$$\begin{aligned} P_t &= P_{wdc} + P_{wac} + P_c \\ &= 50.2 + 0.04 + 0.203 = 50.4 \text{ mW} \end{aligned} \tag{38}$$

The equivalent series resistance (ESR) for the inductor is expressed as

$$\begin{aligned} R_{esr} &= \frac{Pt}{I_{dc}^2} \\ &= \frac{50.4 \times 10^{-3}}{0.807^2}(\Omega) = 62.5 \text{ m}\Omega. \end{aligned} \tag{39}$$

F. Confirmation of The Specified Conditions

From (6), the maximum flux density is calculated as

$$\begin{aligned} B_m &= \frac{LI_{dc}(1+\gamma_r/2)}{NA_c} \\ &= \frac{1.33 \times 10^{-3} \times 0.807 \times (1+0.01/2)}{46 \times 0.58 \times 10^{-4}} \\ &= 0.404 \text{ T}, \end{aligned} \tag{40}$$

which is less than the saturated flux density $B_{sat} = 0.5$ T, but 33 % larger than the specified value of $B_m = 0.3$ T. This difference occurs due to the normalization of the air-gap length and the difference in the K_g value of the selected core. From this result, it can be stated that we should give the value of B_m with a sufficient margin from B_{sat}.

The ratio of the dc winding loss to output power is

$$\begin{aligned} \alpha &= \frac{P_t}{P_o} \\ &= \frac{62.5 \times 10^{-3}}{11.9} = 0.423 \text{ \%}, \end{aligned} \tag{41}$$

which is 15 % lower than the required value because of the difference of the cross-sectional area of the wire and the K_g value.

The core window utilization factor is

$$\begin{aligned} K_u &= \frac{NA_w}{W_a} \\ &= \frac{46 \times 0.518 \times 10^{-6}}{0.6 \times 10^{-4}} = 0.397, \end{aligned} \tag{42}$$

which is almost the same as the required value. This is because we fix the number of turn N for adjusting the fringng effect.

Fundamentally, the design method using core geometry coefficient gurantees inductance value, maximum flux density, and ratio of the dc winding loss to output power as shown in Section II. However some differences apperar due to air-gap length normalization, difference of selected wire size, and difference in the K_g value of the selected core.

III. Discussion of Results

The inductance value of the RF-choke inductor L_c of the class E switching circuit becomes small when the ripple ratio of the current through it becomes large. Small L_c value provides the small core volume V_c and a short length of the wire l_w. However, the increase of the ac component of the current make the large I_m and B_{pp}. Therefore, it is difficult to predict what happens in the inductor when the ripple ratio of the current varies. In this section, we investigate this problem. The all designed inductor in this section consists of Magnetics ferrite core from References [3] and [7] and AWG standardized wire.

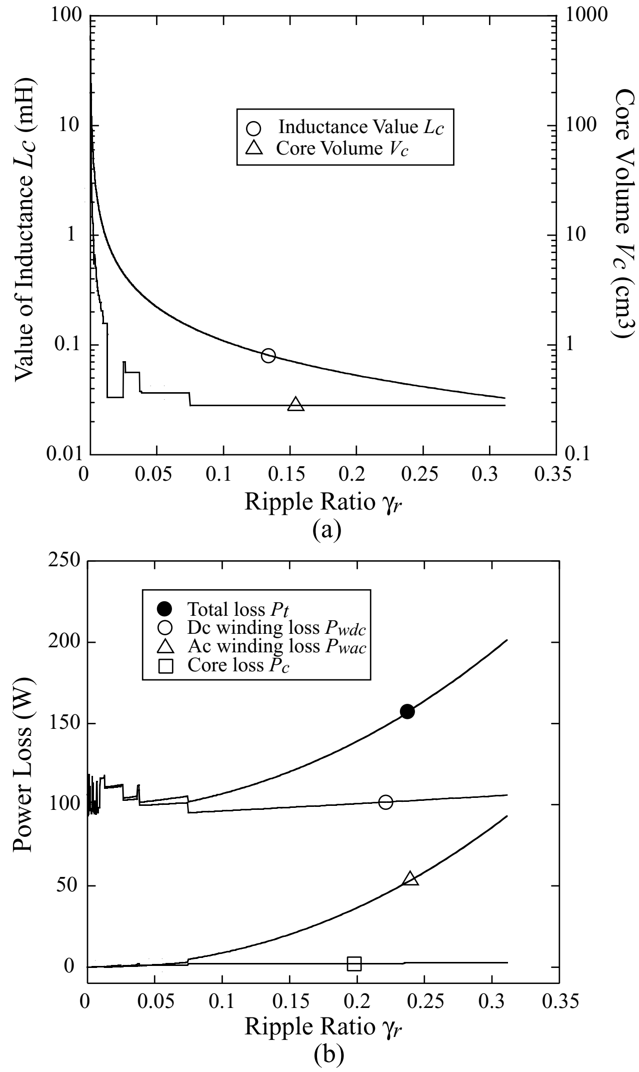

Fig. 2. Characteristics of the designed RF-choke inductor L_c as a function of the ripple ratio of the current γ_r for α = 0.01, I_{dc} = 0.807 A, and f = 1 MHz. (a) Inductance value L_c and core volume V_c of the designed inductor. (b) Dc winding losses P_{wdc}, ac winding loss P_{wac}, core loss P_c, and total power loss P_t of the designed inductor.

Figure 2 shows the characteristics of the RF-choke inductor of the class E amplifier designed by K_g method as a function γ_r. For the design of the class E amplifier, the specifications, which are $f = 1$ MHz, $V_{DD} = 15$ V, and $R = 10\ \Omega$, are given. Additionally, α = 0.01, B_m= 0.3 T, and $J_{max} = 5$ A/mm^2 are also given as specifications of the inductor design. Figure 2(a) shows the characteristics of the inductance value L_c and the core volume V_c. As ripple ratio γ_r decreases, the inductance value increases. Because the high-inductance value prevent ac current from flowing through the inductor, high inductance value is necessary to generate the current with low ripple ratio. Therefore, the core volume V_c becomes large with the decrease in the ripple ratio γ_r. The core volume is constant when $\gamma_r > 0.075$. This is because this core volume is the smallest for all prepared cores. It is seen from Fig. 2(b) that the dc winding loss and core loss are almost constant regardless of γ_r. The concept of K_g method is to achieve the required dc winding loss. The output power is constant against the ripple ratio variation. Therefore, the dc winding loss is also constant for all the ripple ratio. The designer can achieve the required maximum flux density B_m by using K_g method. Therefore, the core loss is also constant regardless of γ_r, which is small enough to neglect. When the ripple ratio increases, the ac component of the current through the inductor also increases. Therefore, ac winding loss increases with the increase in the ripple ratio. For large γ_r, the ac winding loss is almost same as the dc winding loss. This means that it is important to consider the ac winding loss when the ripple ratio γ_r is high.

For $\gamma_r > 0.32$, we cannot design the inductor by using K_g method. In this range, the maximum current density is higher than J_m. This means that the cross-sectional area of the wire is too narrow. For avoiding this problem, we need to set a smaller value of α because a small power loss requires a wide cross-sectional area of the wire. This process, however, is a nonsense. In this case, we should not use K_g method. This is because it is guaranteed that the dc winding loss is lower than the required power loss, and we do not need to consider dc winding loss. Namely, the A_p method should be used in this situation.

Figure 3 shows the characteristics of the RF-choke inductor of the class E amplifier designed by K_g method as a function of α. In this case, $L_c = 1.33$ mH, $I_{dc} = 0.807$ A and $\gamma_r = 0.01$ are fixed for any α. When the cross sectional area of the wire is narrow, the dc winding loss becomes high. Therefore, the increase of α tends to the narrow cross-sectional area of the wire. Therefore, the core volume V_c becomes small as the α increases as shown in Fig. 3(a). Since the large α allows the large dc-winding loss, the dc winding loss is proportional to the α as shown in Fig. 3(b). Conversely, the variations of the ac winding and the core losses are very small compared with the variation of the dc winding loss because the I_m and B_m are fixed when the α is changed. Strictly, the core loss becomes small with the increase in the α since the core volume V_c becomes small. It is seen from Fig. 3(b), both the ac winding loss and the core loss are negligible for all α. For $\alpha > 0.078$, we cannot the design the inductor by using K_g method. Because the cross-sectional area of the wire is too narrow, the current density problem occurs in this range.

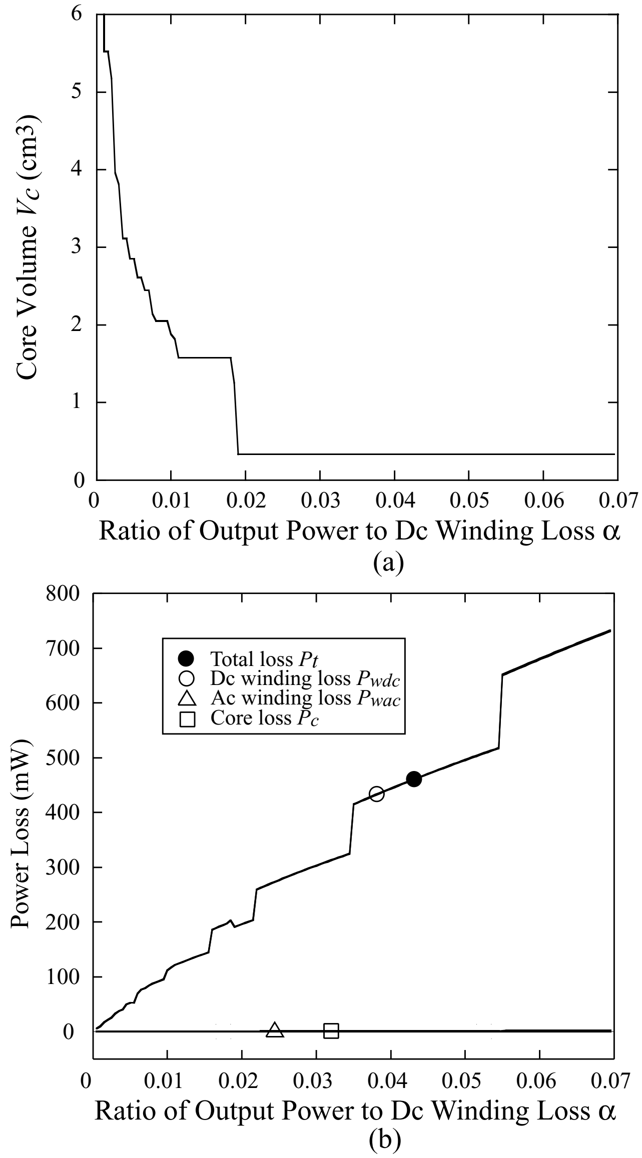

Fig. 3. Characteristics of the designed RF-choke inductor L_c as a function of the ratio of the output power to the dc winding loss α for $\gamma_r = 0.01$, I_{dc} = 0.807 A, and f = 1 MHz. (a) Core volume V_c of the designed inductor. (b) Dc winding losses P_{wdc}, ac winding loss P_{wac}, core loss P_c, and total power loss P_t of the designed inductor.

Figure 4 shows the characteristics of the RF-choke inductor of the class E amplifier designed by K_g method as a function of the I_{dc}. When we vary the dc-supply voltage V_{DD} of the class E amplifier, the output power P_o becomes high. The element values, however, do not vary when the dc-supply voltage V_{DD} varies. Therefore, $L_c = 1.33$ mH, $\gamma_r = 0.01$, and $\alpha = 0.01$ are fixed in Fig. 4. The maximum amplitude of the current I_m is high when the dc-current is high. For avoiding the maximum flux density B_m is larger than the saturated magnetic density B_{sat}, the core volume should be large with the increase in I_{dc} as shown in Fig. 4(a). The output power increases with the increase in I_{dc}, namely, V_{DD}. Because of

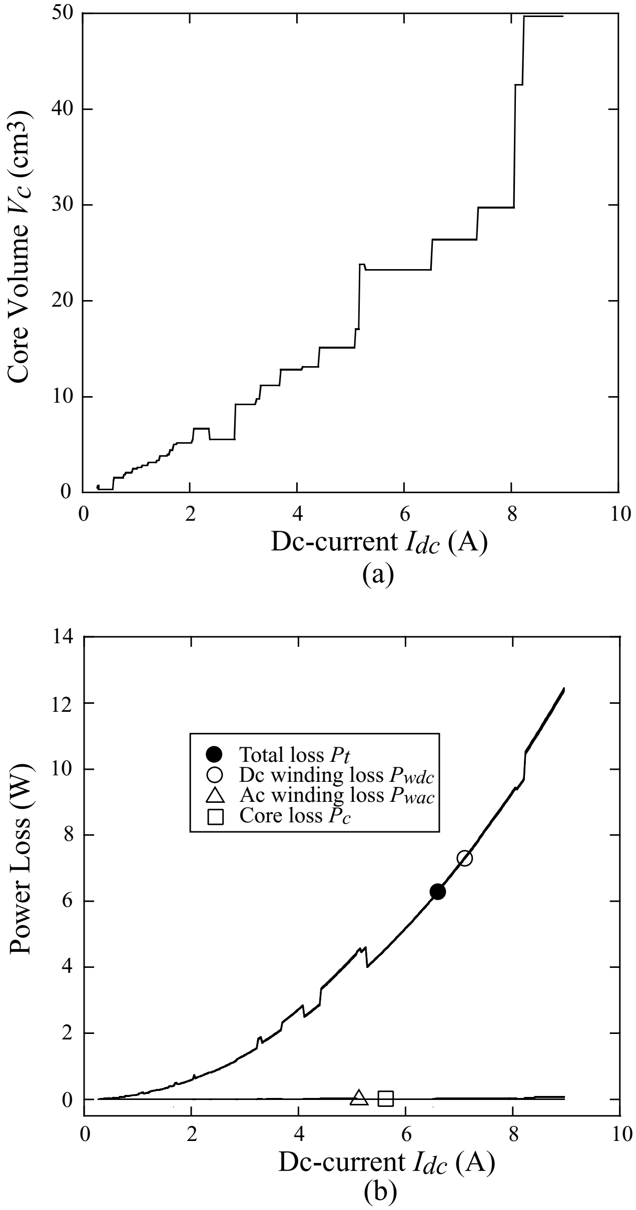

Fig. 4. Designed inductor characteristics of RF-choke inductor L_c as a function of the magnitude of dc current I_{dc} for $\gamma_r = 0.01$, α = 0.01, and f = 1 MHz. (a) Core volume V_c of the designed inductor. (b) Dc winding losses P_{wdc}, ac winding loss P_{wac}, core loss P_c, and total power loss P_t of the designed inductor.

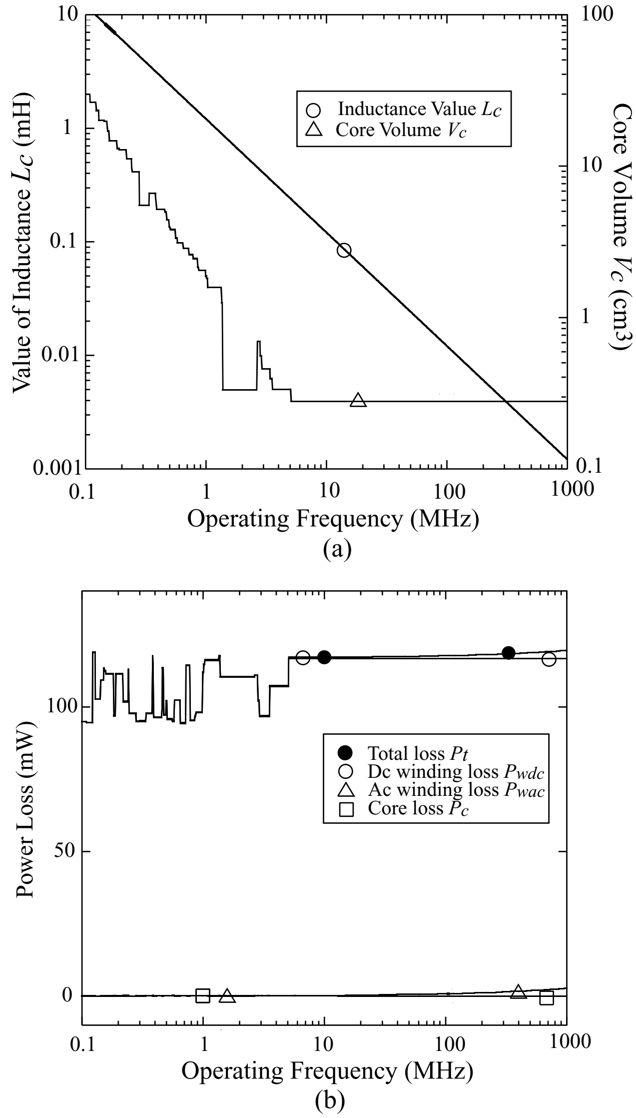

Fig. 5. Designed inductor characteristics of RF-choke inductor L_c as a function of the operating frequency f for $\gamma_r = 0.01$, α = 0.01, and $I[dc]$ = 0.807 A. (a) Inductance value L_c and core volume V_c of the designed inductor. (b) Dc winding losses P_{wdc}, ac winding loss P_{wac}, core loss P_c, and total power loss P_t of the designed inductor.

the fixed $\alpha = 0.01$, the dc winding loss, which is $P_{wdc} = \alpha P_o$, is also increases with the increase in I_{dc}. The dc winding loss is the dominant factor of the total power loss P_t as shown in Fig. 4(b), and the ac winding loss and the core loss are negligible. This figure shows the effectiveness to use the core geometry coefficient for the design of RF-choke inductors.

Figure 4 shows the characteristics of the RF-choke inductor of the class E amplifier designed by K_g method as a function of the operating frequency f. Dc-current through the inductor I_{dc}, $\gamma_r = 0.01$, and $\alpha = 0.01$ are fixed for all the designs of the inductors in this figure. A higher frequency makes the inductors be small. Therefore, the L_c decreases as increase of f as shown in Fig. 4(a). The core volume V_c also decreases. The designers should concern the core loss at high-frequency operations. The core loss for RF-choke inductor, however, generates low core loss as shown in Fig. 4(b). The ac-winding loss is also problem for high-frequency current because of the skin and proximity effects. It is seen from Fig. 4(c) that the ac-winding loss is also generated in RF-choke inductor. This is because the current has few ac component because of small γ_r. As a result, the dc-winding loss P_{wdc} is the dominant factor of the total loss P_t regardless of f. This result shows we can use ferrite core in spite of the high-frequency operations when the RF-choke with low ripple-ratio current is designed.

IV. Conclusion

This paper has presented the fundamental theory to derive the K_g factor for the design of RF-choke inductor and design example. The design example indicates that the dc winding loss is dominant in the power losses of the RF-choke inductor. Fundamentally, it can be stated that the K_g method is useful for the design of RF-choke inductors because the required dc winding loss can be achieved. In some cases, however, the current density is a bottleneck for the design of the inductor. In the cases, the A_p method is better than the K_g method. This is because the coefficient of the A_p method takes into account the maximum current density instead of the dc winding loss.

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