

Received 22 October 2022, accepted 10 November 2022, date of publication 18 November 2022, date of current version 29 November 2022.

Digital Object Identifier 10.1109/ACCESS.2022.3223436

RESEARCH ARTICLE

Experimental Investigation of the Dimensional Effect on Small-Signal Characteristics of Common-Mode Inductors

SHOTARO TAKAHASHI[®], (Member, IEEE)

Faculty of Science and Engineering, Seikei University, Kichijoji, Musashino, Tokyo 180-8633, Japan

e-mail: s.takahashi@st.seikei.ac.jp

This work was supported by the Ministry of Education, Culture, Sports, Science and Technology (MEXT) Program for Creation of Innovative Core Technology for Power Electronics under Grant JPJ009777.

ABSTRACT Due to their high permeability and high effective permittivity, manganese zinc (MnZn) ferrites exhibit magnetic resonance depending on the core dimension. Once magnetic resonance occurs, the relative permeability of MnZn ferrites decreases drastically in the high-frequency (HF) range. The decrease in the relative permeability can influence the HF noise reduction performance of common-mode inductors (CMIs). Based on the above, this study experimentally investigates the impact of the dimensional effect of MnZn ferrites on the small-signal characteristics of CMIs. First, the CM small-signal characteristics of CMIs with different core dimensions indicated that the dimensional effect of MnZn ferrites decreases the CM impedances of the CMIs in the HF range. Two types of core division (air gap insertion and core lamination) were applied to the core to mitigate the dimensional effect on the CMIs. The influence of the core divisions on the complex permeabilities and the small-signal characteristics of the CMIs were evaluated based on the experimental results. The measurement results clarified that core lamination might be the appropriate option in EMI filtering applications because it can mitigate the dimensional effect and increase the CM impedance of CMIs in the HF range without increasing the number of turns.

INDEX TERMS Common-mode inductors, EMI, MnZn ferrites, permeability, small-signal characteristic.

I. INTRODUCTION

Wide bandgap power semiconductor devices using silicon carbide and gallium nitride offer superior switching properties compared with conventional power semiconductor devices using silicon insulated-gate bipolar transistors [1]. The high switching speed of wide bandgap power semiconductor devices increases the switching frequencies and downsizes the passive components of power converters. However, the high switching speed and high switching frequency of power converters lead to an increase in electromagnetic interference (EMI) [2], [3], [4], [5], [6]. The International Electrotechnical Commission regulates the allowable EMI limit for each application. As every power converter must satisfy these EMI limits, an EMI filter is inevitably installed

The associate editor coordinating the review of this manuscript and approving it for publication was Nagesh Prabhu¹⁰.

in power converters. Therefore, researching an appropriate design procedure for EMI filters and enhancing their filtering performance are increasingly important. When designing EMI filters, power electronics engineers often use circuit simulators such as LTspice to evaluate the small-signal characteristics of the filters. Therefore, previous studies have investigated circuit simulation models of filtering components and entire power conversion systems, including EMI filters [7], [8], [9].

Fig. 1 illustrates the basic configuration of a single-stage EMI filter. A common-mode inductor (CMI), y-capacitors, differential-mode inductors (DMIs), and x-capacitors comprise the EMI filter. The CMI and y-capacitors are common-mode (CM) noise filtering components, and the DMIs and x-capacitors are differential-mode (DM) noise filtering components. For safety reasons, the values of y-capacitors are strictly limited (in most cases, several nanofarads). Thus, the



FIGURE 1. Basic configuration of single-phase EMI filter.

CMI is usually bulky, and magnetic materials with high permeability, such as nanocrystalline or manganese zinc (MnZn) ferrites, are selected for CMI magnetic cores. On the other hand, the values of the x-capacitors can be large compared to those of y-capacitors (typically several microfarads). Thus, magnetic materials with low permeability and high saturation flux density, such as iron powder, are used for DMIs. Note that the leakage inductances of CMIs can be used as DMIs to decrease the number of filter components.

Among filter components, inductors typically determine the high-frequency (HF) characteristics of EMI filters because of their complex frequency-dependent characteristics. Hence, many previous studies have presented modeling techniques to simulate the wideband small-signal characteristics of filter inductors [10], [11], [12], [13], [14], [15], [16], [17], [18], [19], [20], [21], [21]. These modeling techniques can simulate the frequency dependence of the complex permeability of magnetic materials [12], [13], [14], [15], [16], [17], [18], [19], [21] and multiple resonators due to the transmission line effect of windings [20], [21].

The frequency dependence of the complex permeability and winding stray capacitance strongly influence the smallsignal characteristics of CMIs. Furthermore, the high permeability and high effective permittivity of magnetic materials may give rise to standing electromagnetic waves that depend on the size of the magnetic core. This phenomenon is known as dimensional resonance and has been observed experimentally in MnZn ferrites [22]. Once dimensional resonance occurs, the permeability of the magnetic core decreases drastically. Therefore, the dimensional effect of MnZn ferrites may influence the small-signal characteristics of CMIs in the HF range. However, to the authors' knowledge, the influence of the dimensional effect on the small-signal characteristics of CMIs has not been investigated in detail.

This paper does not provide a theoretical explanation for the dimensional effect on the noise reduction performance of CMIs. Instead, it explores the possible challenges encountered by wideband CMI design as switching frequency increases in wideband gap device-based power converters. For this purpose, this study experimentally investigates the dimensional effect of MnZn ferrites on the smallsignal characteristics of CMIs. First, measurements of the different numbers of turns were performed by using an impedance analyzer. The measurement results clarified that multiple factors cause resonances in inductor impedances. Next, measurements of the complex permeabilities of MnZn ferrite with several core dimensions and CM small-signal characteristics of the CMIs were carried out. The measured results clarified that the dimensional effect of MnZn ferrites decreases the CM impedances of the CMIs in the HF range. Two types of core divisions (air gap insertion and core lamination) were applied to the toroidal cores to mitigate the dimensional effect. The measurement results of the smallsignal characteristics of the fabricated CMIs verified that core lamination could mitigate the dimensional effect and increase the CM impedance of the CMIs in the HF range without increasing the number of turns. Note that every measurement in this study was performed using off-the-shelf magnetic core products.

small-signal characteristics of single-phase inductors with

II. DIMENSIONAL EFFECT OF MNZN FERRITES

A. INDUCTOR IMPEDANCE RESONANCE FACTORS

This subsection investigates the resonance factors that appear in inductor impedances. A toroidal-shaped MnZn ferrite core of N30 [23] (R50 core, part number: B64290L0082 \times 830, EPCOS) was selected as the magnetic core of a single-phase inductor under measurement. The measurements were carried out using an impedance analyzer (E4990A, Keysight) in a frequency range from 1 kHz to 50 MHz. Figs. 2(a) and 2(b) show the measured impedances and phases of the inductor when the number of turns is 3, 5, 10, 20, 30, and 50.

Fig. 2(a) indicates that the first resonance appears at frequencies lower than 1 MHz in all cases. The first resonance frequency of the inductor with 30 turns is approximately 600 kHz, and that of the inductor with 50 turns is 300 kHz. On the other hand, the first resonance frequencies in the other cases are approximately 700 kHz.

Fig. 2(b) shows a sharp phase change from 90° to -90° at approximately 300 kHz when the number of turns is 50. This acute phase change is due to self-inductance resonance and winding stray capacitance. In addition, when the number of turns is 3, 5, 10, and 20, the inductors behave as capacitive impedances at higher frequencies than the first resonance frequency. However, the phase delays are smaller than those when the number of turns is set to 30 and 50. Here, the small-signal characteristics of the inductors are typically represented by an LCR parallel circuit. Parallel capacitances obtained from the measured impedances are listed in Table 1. The calculated capacitance of the single-phase inductor with 50 turns, shown in Fig. 2(a), is 12.6 pF. On the other hand, the calculated stray capacitance of the inductor with five turns is 219.6 pF. Note that the resonance frequency of the inductor with five turns is determined as the frequency where the inductor impedance shows its maximum value. The calculated results indicate that the winding stray capacitance of inductors based on MnZn ferrites with lower turns is much



FIGURE 2. Small-signal characteristics of single-phase inductors. (a) Impedances. (b) Phases.

TABLE 1. Measured	parameters of	inductors.
-------------------	---------------	------------

Turn number	Inductance	Resonant frequency	Capacitance
3 turns	74.5 [µH]	746.3 [kHz]	610.7 [pF]
5 turns	207.1 [µH]	746.3 [kHz]	219.6 [pF]
10 turns	768.3 [µH]	838.9 [kHz]	46.8 [pF]
20 turns	3.3 [mH]	698.8 [kHz]	15.7 [pF]
30 turns	7.4 [mH]	549.0 [kHz]	11.4 [pF]
50 turns	20.3 [mH]	315.0 [kHz]	12.6 [pF]

higher than inductors with higher turns. Thus, using the LCR parallel circuit to represent the frequency characteristics of the inductors with few turns results in overestimated winding stray capacitance. In other words, the factors of the decrease in high-frequency impedance should be considered separately for the self-resonance due to the winding stray capacitance and the others. Ref. [21] described this impedance resonance as *material resonance* caused by material capacitance. With a correctly understanding of the factor of the impedance resonance, designers can determine whether to change the winding arrangement or apply the core lamination (described in subsection III-B) to improve the HF filtering performance of the filter inductors.

Furthermore, when the number of turns is set to 20, 30, and 50, multiple resonances occur periodically at frequencies

higher than 10 MHz. These multiple resonances are caused by the transmission line effect of the winding. In short, the measured results in Fig. 2 indicate that the causes of inductor impedance resonances can be classified into the following three factors.

- 1) Inherent characteristics of the magnetic material (material resonance)
- 2) Winding stray capacitance (self-resonance)
- Transmission line effect of the winding, which causes multiple resonances

The measured results when the number of turns is 3, 5, 10, and 20 suggest that material resonances occur in the first resonant frequencies of the fabricated inductors. However, the impact of the material resonance on the filter inductor impedance has not been investigated in detail [21]. The descriptions in [24], [25], [26], [27], [28], and [29] probably assumed that the natural or dimensional resonance causes this impedance resonance. The natural resonance is magnetic resonance due to the effective anisotropy field, resulting in magnetic losses [24]. Once the magnetic resonance occurs in a magnetic core, the relative permeability begins to decrease, and the resistive component becomes dominant in the inductor impedance. In general, the relative permeability of magnetic materials is inversely proportional to the frequency at which the permeability begins to decrease [24]. Snoek's limit represents the relationship between the relative permeability μ_r and the frequency f [24]. According to the material properties of N30 [23], Snoek's limit of N30 and the frequency are derived as $\mu_{\rm r} f \approx 7100 \,{\rm MHz}.$

The material properties over a wideband frequency are often unavailable from the manufacturer's datasheet. Thus, measurements of the complex permeability of N30 should be performed by the designers of the magnetic components themselves.

An inductor with N turns can be represented as a series connection of an equivalent series resistance R_s and an equivalent series inductance L_s when the winding stray capacitance is very small and negligible [25], [30]. The impedance of the inductor Z_L is described in the following equation.

$$Z_{\rm L} = R_{\rm s} + j\omega L_{\rm s} = j\omega\mu_0\mu_{\rm r}\frac{A_{\rm c}}{l_{\rm c}}N^2$$
$$= j\omega\mu_0\left(\mu_{\rm r}' - j\mu_{\rm r}''\right)\frac{A_{\rm c}}{l_{\rm c}}N^2 \qquad (1)$$

where A_c is the core cross-sectional area, l_c is the magnetic flux path length, μ_0 is the permeability of free space $(4\pi \times 10^{-7} \text{ H/m})$, and μ'_r and μ''_r are the real and imaginary parts of the complex permeability of the magnetic core. The frequency characteristics of R_s and L_s can be measured directly by using the impedance analyzer; then, μ'_r and μ''_r can be obtained from the following equations, respectively.

$$\mu_{\rm r}'(f) = \frac{l_{\rm c}}{\mu_0 A_{\rm c} N^2} \cdot L_{\rm s}\left(f\right) \tag{2}$$

$$\mu_{\rm r}^{''}(f) = \frac{l_{\rm c}}{2} \pi f \mu_0 A_{\rm c} N^2 \cdot R_{\rm s}(f)$$
(3)



FIGURE 3. Complex permeability of N30 derived by equations (2) and (3) based on inductor impedance measured by the impedance analyzer E4990A. The tested core was R50, and the turn number was six. (a) Real and imaginary parts plotted on a linear axis. (b) Real part plotted on a logarithmic axis for comparison with the Snoek's limit of N30 ($\mu_r f \approx 7100$ MHz).

The measured complex permeabilities of N30 (R50 core) are shown in Figs. 3(a) and 3(b). Fig. 3(a) shows that the real part increases slightly at approximately 400 kHz and decreases drastically. Furthermore, the real part becomes negative from 1 to 20 MHz and converges to zero. After the imaginary part rises sharply at approximately 700 kHz, it decreases sharply. The slopes of the increase and decrease are almost equal and opposite. These measured results suggest that these drastic changes in the complex permeability are caused by magnetic resonance [24].

The real part of the complex permeability is plotted on a logarithmic axis in Fig. 3(b). Fig. 3(b) also presents Snoek's limit of N30. Fig. 3(b) shows that the real part of the complex permeability begins to decrease at lower frequencies than Snoek's limit. Therefore, Fig. 3(b) indicates that the natural resonance may not cause this drastic change in the permeability. Hence, another factor, such as the dimensional resonance [22], should be investigated. In typical MnZn ferrites, crystal grains with low resistivity are surrounded by grain boundaries with high resistivity to suppress eddy-current losses. This microstructure can be represented as a lossy capacitor [27], [28], [31]. Due to the large ratio between the size of the crystal grain and the width of the grain boundary, the effective permittivity of the grain boundary ε_{eff} is significant [28]. Thus, the wavelength of electromagnetic waves inside the magnetic core is shortened because of the high permeability and high effective permittivity of MnZn ferrites [25], [28].

Assuming a lossless medium, if the relative permeability of the medium is μ_r and the relative permittivity is ε_r , then the wavelength of the electromagnetic wave propagating in the medium λ is given by

$$\lambda = \frac{c}{f\sqrt{\mu_{\rm r}\varepsilon_{\rm r}}}\tag{4}$$

where *c* is the propagation velocity of light in free space $(c \approx 3 \times 10^8 \text{ m/s})$. For example, with $\mu_r = 4 \times 10^3$ and $\varepsilon_r = \varepsilon_{eff} = 5 \times 10^4$ [28], the calculated wavelength $\lambda = 2.1$ cm at a frequency f = 1 MHz. If the smallest cross-sectional dimension of a magnetic core is half of the wavelength, then a fundamental mode standing wave will appear across the section, and magnetic resonance occurs. This phenomenon has been identified experimentally in MnZn ferrites by Brockman et al. [22].

The complex permeabilities of N30 with different core dimensions are measured and compared to investigate the dimensional dependence of the complex permeability. Fig. 4 shows the cores under the measurements, and Table 2 shows the dimensions of each core.



FIGURE 4. MnZn ferrite cores under the measurements.

Figs. 5(a) and 5(b) show the measured real and imaginary parts of the complex permeabilities of each core, respectively. Note that the measured complex permeabilities are normalized using the measured values at 10 kHz to eliminate any effect due to manufacturing errors. Figs. 5(a) and 5(b) show that the R50 core, which has the largest dimension of the five cores, indicates a drastic change in complex permeability due to the magnetic resonance. The changes in μ'_r and μ''_r in the small core are gentler than those in the large core. Moreover,

Core	Parts number	Outer diameter [mm]	Inner diameter [mm]	Height [mm]	Effective magnetic path length [mm]	Effective magnetic cross section [mm ²]
R4	B64290P0036X830	4.17	2.23	1.75	9.63	1.25
R10	B64290L0038X830	10.8	5.25	4.75	24.07	7.83
R16	B64290L0045X830	17.2	8.5	7.3	38.52	19.73
R34	B64290L0048X830	35.5	19.2	13.6	82.06	82.60
R50	B64290L0082X830	51.8	28.5	21.3	120.4	195.7

TABLE 2. Dimensions of MnZn ferrite cores.

the resonance frequencies shift to the higher frequency range as the core size decreases.

The measured real parts of the complex permeabilities are plotted on the logarithmic axis in Fig. 5(c). As the magnetic core size decreases, the change in the permeability becomes more moderate. Fig. 5(c) also indicates that the measured permeabilities become closer to Snoek's limit of N30 as the core size decreases. These results indicate that the complex permeability of MnZn ferrites strongly depends on the core size.

The wideband frequency characteristics of the complex permeabilities are shown in Figs. 5(a) and 5(b) are possibly unavailable from the datasheet in most cases. Indeed, the datasheet of N30 shows the frequency characteristic of the complex permeabilities up to around 4 MHz [24]. Furthermore, the manufacturer has measured the complex permeability of N30 on the small toroidal core, such as the R10 core, to avoid the dimensional effect [24]. Thus, the dimensional dependency of the complex permeabilities of N30, as shown in Figs. 5(a) and 5(b) are also not obtained from the datasheet.

Since this article focuses on the small-signal characteristics of CMIs, the losses of CMIs are not evaluated. However, the loss tangent follows the iron loss characteristics of inductors under large-signal excitation, as presented in the reference [32]. The loss tangent is widely used to evaluate the power loss and is derived as

$$\tan \delta_{\rm m} = \frac{\mu_{\rm r}^{''}}{\mu_{\rm r}^{\prime}} \tag{5}$$

where δ_m is the loss angle, equation (5) indicates that the magnetic loss tangent is easily obtained based on the measured complex permeabilities.

The magnetic loss tangent of the tested MnZn ferrite cores is shown in Fig. 6. The results show that the loss tangent of the larger cores is the highest for the frequencies of interest. It can also be observed that the loss tangent of the smaller cores shows smaller values in the HF range.

C. DIMENSIONAL EFFECT ON THE SMALL-SIGNAL CHARACTERISTICS OF THE CMIS

Three single-phase CMIs with different core sizes were fabricated to investigate the impact of the dimensional effect on the small-signal characteristics of the CMIs. The cores used for the CMIs were the R16 core, R34 core, and R50 core.

Based on the AL values obtained from the datasheets, the CMIs were designed to have almost equal CM inductances in the low-frequency range. As a result, the turn numbers were set to 11, 8, and 6 in the fabricated CMIs with the R16, R34, and R50 cores, respectively.

Figs. 7(a) and 7(b) show the measured results of the CM impedances and phases of the fabricated CMIs. Fig. 7(a) indicates that the slope of the CM impedance of the CMI with the R16 core becomes moderate at approximately 500 kHz. As already shown in Figs. 5(a) and 5(b), the real part of the complex permeability of the R16 core begins to decrease at approximately 500 kHz, and the imaginary part begins to increase drastically. Hence, the resistive component is possibly dominant in the CM impedance. Furthermore, the measured CM impedance of the CMI with the R16 core begins to decrease from approximately 4 MHz. This decrease in the CM impedance may be because the real part of the complex permeability of the R16 core is negative beyond 4 MHz, and the CMI behaves as the capacitive impedance component.

On the other hand, the measured CM impedances of the CMIs begin to decrease at lower frequencies as the core dimension increases. The CMI with the R50 core shows capacitive impedance behavior beyond 1 MHz. This result is because the real part of the complex permeability of the R50 core becomes negative from approximately 1 MHz, as shown in Fig. 5(a).

The measured results are summarized as follows. The measured CM impedances of the three fabricated CMIs are almost equal in the low-frequency range up to approximately 500 kHz. However, the three measured impedance curves differ significantly in the HF range due to the dimensional dependence of the complex permeability (not to the self-resonance caused by the winding stray capacitance). In other words, the measured results indicate that the HF noise reduction performance of the CMIs based on MnZn ferrites depends not only on the winding stray capacitance but also on the core dimension.

Note that unexpected resonance and anti-resonance of loads (e.g., long power cables) connected to power converters may increase electromagnetic noise and should be appropriately suppressed. Loss resistance due to the imaginary part of the complex permeability of magnetic materials is widely used for damping the resonances. The imaginary part of the complex permeability of MnZn ferrites decreases drastically



FIGURE 5. Complex permeabilities of N30 with different core dimensions. All curves are derived by equations (2) and (3) based on the measured impedances of inductors. (a) Real parts normalized by the values at 10 kHz to eliminate manufacturing errors. (b) Imaginary parts normalized by the values at 10 kHz. (c) Real parts plotted on a logarithmic axis for comparison with the Snoek's limit of N30 ($\mu_r f \approx 7100$ MHz).

in the high-frequency range due to the dimensional effect, as already shown in Fig. 5(b). Therefore, the high-frequency loss resistance of MnZn ferrites is insignificant, and it is challenging to suppress the resonances by using MnZn ferrite cores. In practice, nanocrystalline cores are used as damping components in the frequency band of the conducted EMI (150 kHz–30 MHz) [33], and nickel-zinc ferrite cores are widely used in the frequency range of the radiated EMI (30 MHz–1 GHz) [34].



FIGURE 6. Magnetic loss tangent based on the measured complex permeabilities of N30.



FIGURE 7. Measured CM frequency characteristics of the CMIs. (a) Impedances. (b) Phases.

III. IMPACT OF CORE DIVISION ON THE FREQUENCY CHARACTERISTIC OF THE CMIS

A. AIR GAP INSERSION

1) COMPELX PERMEABILIES OF GAPPED CORES

This subsection investigates the impact of an air gap insertion on the small-signal characteristic of CMIs. First, the impact of the air gap insertion on the complex permeability of N30 is investigated. In general, air gaps are inserted into the core to avoid magnetic saturation. The slope of the B-H curve decreases due to the air gap, and magnetic saturation can be prevented. As a result, the inductor can drive higher currents. The reluctance of the air gap is greater than that of the magnetic core with high relative permeability. Thus, an impedance due to the gap reluctance is dominant in a gapped inductor. As described in the previous section, the relative permeability of MnZn ferrites decreases drastically in the HF range because of the dimensional effect. The impedance due to the core reluctance becomes dominant in the inductor impedance in the frequency range where the core reluctance is larger than the gap reluctance. Thus, the magnetic resonance frequency may shift to a higher frequency by the air gap insertion.



FIGURE 8. Toroidal core (N30, R50 core) is divided into two planes perpendicular to the magnetic path.

As shown in Fig. 8, the toroidal core (N30, R50 core) is divided into two planes perpendicular to the magnetic path. An insulating tape (its thickness is 0.05 mm) is inserted into the gap, and the thickness of the gap is adjusted by increasing or decreasing the amount of inserted insulating tape. The small-signal characteristics of the inductors with six turns are measured, and the complex permeabilities are calculated from Equations (2) and (3). Figs. 9(a) and 9(b) show the measured complex permeabilities when the air gap is not inserted (base core), and the gap length is set to 0.1 mm, 0.2 mm, and 0.5 mm. Note that the measured complex permeabilities in Figs. 9(a) and 9(b) are normalized to the measured values at 1 kHz.

Figs. 9(a) and 9(b) show that as the air gap length increases, the resonance peaks of the real and imaginary parts of the complex permeabilities are suppressed, and the frequencies at which the real part begins to decrease shift to the HF range. In other words, as the air gap length increases, the frequencies where the core reluctance becomes more significant than the air gap reluctance shift to higher frequencies. Thus, these measured results indicate that the magnetic resonance frequencies due to the core dimension shift to a higher frequency by the air gap insertion.

2) ANALYSIS OF THE IMPACT OF THE GAP INSERTION ON CM IMPEDANCES OF GAPPED CMIS

A CM single-phase equivalent circuit model of a gapped CMI (shown in Fig. 10) is used to analyze the impact of the insertion of an air gap on CM impedance. The analytical model shown in Fig. 10 consists of CM inductance L_{CM} , core loss resistance R_{CM} , air gap inductance L_g , and stray capacitance C_{CM} . L_{CM} and R_{CM} are frequency-dependent



FIGURE 9. Impact of the air gap length I_g on the complex permeability of MnZn ferrite N30. The tested core was the toroidal R50 core, and the turn number was six. All curves are derived by equations (2) and (3) based on the measured impedances of inductors. (a) Real parts normalized by the values at 1 kHz to eliminate manufacturing errors. (b) Imaginary parts normalized by the values at 1 kHz.



FIGURE 10. CM single-phase equivalent circuit model of the gapped CMI.

components due to the complex permeability of the magnetic material. L_{CM} , R_{CM} , and L_{g} can be expressed in the following equations (6)–(8), respectively.

$$L_{\rm CM}(f) = \mu_0 \mu_{\rm r}'(f) \frac{A_{\rm c}}{l_{\rm c}} N^2$$
(6)

$$R_{\rm CM}(f) = 2\pi f \mu_0 \mu_{\rm r}^{''}(f) \frac{A_{\rm c}}{l_{\rm c}} N^2$$
(7)

$$L_{\rm g} = \mu_0 \frac{A_{\rm c}}{l_{\rm g}} N^2 \tag{8}$$

In equation (8), l_g is the total length of the air gap. The real and imaginary parts of the complex permeability are taken to be the measured values mentioned in the previous subsection. Note that stray capacitance is neglected for simplicity.



FIGURE 11. Calculated results of CM impedances.

Fig. 11 shows the calculated impedance of the magnetic core with 11 turns ($Z_{CM-core} = R_{CM} + j\omega L_{CM}$), the impedance of the air gap of length 0.2 mm ($Z_{CM-gap} = j\omega L_g$), and CM impedance without an air gap.

Fig. 11 shows that the CM impedance of the gapped CMI is significantly increased in the frequency range beyond 1 MHz, compared to the case in which the air gap was not inserted $(l_g = 0 \text{ mm})$. It can also be confirmed that the impedance of the air gap $(Z_{\text{CM}-\text{gap}})$ is dominant in the frequency range below 3 MHz and the impedance of the magnetic core $(Z_{\text{CM}-\text{core}})$ becomes dominant at frequencies higher than 3 MHz.



FIGURE 12. Calculated CM impedance of gapped CMIs with different gap lengths.

CM impedances for different gap lengths of 0, 0.1, 0.2, and 0.5 mm were calculated. Fig. 12 shows the calculated CM impedances. Fig. 12 confirms that increasing the gap length widens the frequency range in which the gapped CMI behaves as an inductive impedance, and that the CM impedance increases significantly compared to the case when an air gap is not inserted, at frequencies higher than 1 MHz. It should be noted that the required number of turns for the desired inductance increases as the gap length increases. In this study, a single-layered winding was selected to minimize stray

capacitance. Therefore, the window of the selected toroidal core and the diameter of the windings limit the maximum gap length that can be inserted into the core.

3) MEASUREMENT OF SMALL-SIGNAL CHARACTERIS-TICS OF GAPPED CMIS

To evaluate the impact of the gap insertion on the CM smallsignal characteristics of the CMIs, single-phase CMIs based on the gapped ferrite cores were fabricated. The CM and DM impedances of the gapped CMIs were measured using an impedance analyzer. A photograph of the gapped singlephase CMI is shown in Fig. 13. The frequency characteristics of the fabricated CMI were measured (with respect to CM and DM) according to the connections of windings shown in Fig. 14. The measurements were performed for total gap length values of 0.1, 0.2, and 0.5 mm. The measured CM and DM frequency characteristics are shown in Figs. 15 and 16, respectively. Note that Figs. 15 and 16 also show the results for when an air gap is not inserted ($l_g = 0$ mm).



FIGURE 13. Photograph of the fabricated gapped single-phase CMI.



FIGURE 14. Connections of CMI's windings for CM and DM impedance measurements.

First, Fig. 15(a) confirms that CM impedance increases at frequencies higher than 1 MHz, when an air gap is inserted. In the case where the total gap length was set to 0.1 mm, the CM impedance was more than that of the ungapped CMI in the frequency range of 1 MHz to 50 MHz. This result is in agreement with the analytical result shown in Fig. 12. The results for gap lengths of 0.2 mm and 0.5 mm do not match with the analytical results in the HF range. In particular, for the gap length of 0.5 mm, there is no increase in CM impedance compared to the case with gap length set to 0.2 mm, at frequencies higher than 7 MHz. It can be confirmed from the phase measurement results shown



FIGURE 15. Measurement results of CM frequency characteristics of the gapped CMIs. (a) Impedance. (b) Phase.

in Fig. 14(b) that the phase approaches -90° in the high-frequency range above 4 MHz with increasing gap length. This indicates that the influence of stray capacitance becomes dominant in the HF range with the increase in gap length.

Next, it can be confirmed from the measured DM frequency characteristics (shown in Fig. 16) that the DM impedance increases significantly in the frequency range of 10 kHz to 10 MHz, with the increase in total gap length. This is because the required number of turns increases with the increase in the total gap length. From Fig. 16, we can determine that the self-resonance frequencies of the DM impedance shift to lower values with the increase in the length of the air gap, due to increased DM inductances.

As shown in Fig. 15(a), increasing the turn number is inevitable to obtain the required CM inductance. The winding pitch of adjacent windings decreases as the number of turns increases. As a result, the stray capacitance between adjacent windings becomes considerable. In other words, the self-resonance of the gapped CMI limits the CM filtering performance of CMIs in the HF range. Thus, it is necessary to carefully adjust the gap length with considering the selfresonance frequency of the gapped CMI. In addition, the leakage flux due to the air gap insertion may increase and cause unexpected magnetic coupling between filtering components.



FIGURE 16. Measurement results of DM frequency characteristics of the gapped CMIs. (a) Impedance. (b) Phase.

Previous studies have pointed out that such parasitic coupling degrades EMI filter noise reduction performance [35], [36], [37]. Furthermore, the increase in the winding loss caused by DM currents due to the increase in windings is also a drawback of the air gap insertion. The increase in the winding loss causes a decrease in the efficiency of power converters. Based on the above, air gap insertion is not considered an appropriate option for improving the HF characteristics of inductors for EMI filtering applications.

B. CORE LAMINATION

1) FREQUENCY CHARACTERISTICS OF LAMINATED MNZN FERRITE CORES

As described in [28], magnetic resonance due to the dimensional effect occurs when the smallest cross-sectional dimension of the core equals a half-wavelength of an electromagnetic wave within the core. Suppose the toroidal core is divided in the radial direction and laminated. In that case, the cross-sectional area of the entire core does not change, but the smallest cross-sectional dimension of the core becomes small. Therefore, the magnetic resonance frequency may be shifted to the HF range, producing the same effect as the smaller magnetic core. Hence, core lamination is

possibly able to mitigate the drastic change in the complex permeability due to the dimensional effect.



FIGURE 17. Picture of the laminated cores.

This subsection investigates the effect of core lamination on the CM impedance of CMIs. Fig. 17 shows a photograph of the laminated cores (N30, R50 core). Here, four laminated cores are fabricated with cores divided into one (base core), two, three, and four.



FIGURE 18. Measured complex permeabilities of the laminated cores (N30, R50 core). (a) Real parts. (b) Imaginary parts.

First, the complex permeabilities of the laminated MnZn ferrite cores were measured. Fig. 18 shows the measured results of the complex permeabilities of each laminated core calculated from Equations (2) and (3). The measured real and imaginary parts are normalized using the measured values at 1 kHz. In the base core, the real part decreases sharply at approximately 700 kHz, and a sharp resonance peak of the imaginary part appears. This result is similar to that seen in the measurements taken of the R50 core shown in Figs. 5(a) and 5(b). As the number of laminations increases, the change in the real part becomes gentle, the resonance peak

of the imaginary part is also suppressed, and the resonance frequency of 700 kHz shifts to the HF range. This tendency is similar to when the core size decreases, which is verified in subsection II-B.



FIGURE 19. Magnetic loss tangent of laminated MnZn ferrite cores.

The magnetic loss tangent of the laminated MnZn ferrite cores derived by equation (5) is shown in Fig. 19. Fig. 19 shows that the magnetic loss tangent of the MnZn ferrite core with the higher number of laminations is lower in the HF range. The loss tangent for the smaller size cores presents a smaller value in the HF range as already shown in Fig. 6. Figs. 6 and 19 confirm that increasing the number of laminations also shows the same tendency for the magnetic loss tangent as decreasing the core size.

Fig. 20 shows the measurement results of the CM frequency characteristics of the single-phase CMIs with the base core and the four-laminated core. Fig. 20 indicates that the measured CM impedances increase at frequencies higher than 1 MHz by the core lamination. This result may be because the change in permeability becomes gentle by core lamination, as confirmed in Fig. 18. The above results indicate that the core lamination can mitigate the drastic change in the complex permeability due to the dimensional effect and increase the CM impedance of the CMIs in the HF range.

2) CM NOISE FILTERING PERFORMANCE OF LAMINAT-ED CORE CMI

In order to verify the filtering performance improvement by the core lamination, CM currents were measured in a pulsewidth-modulated (PWM) inverter-fed motor drive system. A configuration of an experimental system is shown in Fig. 21. A DC power source and the three-phase PWM inverter (HEK-INV-A, Headspring) were connected via line impedance stabilization networks (LISNs, model no. LI-325C, COM-POWER). The DC power source output was 200 V, output frequency of the three-phase PWM inverter was 50 Hz, and carrier frequency was 100 kHz. A 0.75 kW induction motor with no load was connected to the inverter using a 1-m three-core cable (cross-section of conductors: 2.0 mm²). The base core CMI and/or the laminated core CMI was connected to the input-side of the inverter. The inverter,



FIGURE 20. Measured CM frequency characteristics of the laminated CMIs. (a) Impedances. (b) Phases.



FIGURE 21. Configuration of the experimental system for measuring CM current.

motor, power cables, and fabricated CMI were placed on styrofoam blocks (thickness: 5 cm). An inverter heat sink and the motor frame were connected to an aluminum plate that imitated the ground plane via 0.1 m ground wires (conductor cross-section: 2.0 mm^2). A current monitor probe (F-33-1, FCC) and spectrum analyzer (FPL1003, Rohde & Schwarz) were used for the measurement of CM currents. The measurement point of CM currents *I*_{CM} is the input-side of the inverter, and CM currents were measured by clamping the DC input wires of the inverter as a whole. A picture of the experimental setup is shown in Fig. 22.

Fig. 23 shows the measured CM currents when no filtering component, base core CMI, and laminated core CMI were connected. The measurements were performed in the frequency range of 100 kHz–50 MHz. Fig. 23 shows that the base core CMI reduces the CM current from 400 kHz. When the laminated core CMI was connected, the measured CM current had lower values than the results obtained



FIGURE 22. Picture of the experimental setup.



FIGURE 23. Comparison of the measured CM currents.

when the base core CMI was connected above 1 MHz. The improvement of the attenuation reaches 8.5 dB in 2.3 MHz. This result is due to the core lamination increasing the CM impedance of the base core CMI at frequencies higher than 1 MHz, as shown in Fig. 20(a).

IV. CONCLUSION

This study experimentally investigated the dimensional effect of MnZn ferrites on the small-signal characteristics of CMIs. The CM small-signal characteristics of the CMIs with different core dimensions indicated that the dimensional effect of MnZn ferrites decreases the CM impedances of the CMIs in the HF range. Two types of core division (air gap insertion and core lamination) were applied to the core to mitigate the dimensional effect on the CM impedances of the CMIs. The influence of the core divisions on the complex permeabilities and the small-signal characteristics of the CMIs were evaluated based on the experimental results. The measurement results clarified that air gap insertion can improve the HF performance of CMIs. However, many turns are required to obtain the same CM impedance of the ungapped CMI. This result suggests that the stray capacitance may be increased by inserting the air gap. On the other hand, core lamination might be the appropriate option in EMI filtering applications because it can mitigate the dimensional effect and increase the CM impedance of the CMIs in the HF range without increasing the number of turns.

REFERENCES

- J. Biela, M. Schweizer, S. Waffler, and J. W. Kolar, "SiC versus Si—Evaluation of potentials for performance improvement of inverter and DC–DC converter systems by SiC power semiconductors," *IEEE Trans. Ind. Electron.*, vol. 58, no. 7, pp. 2872–2882, Jul. 2011.
- [2] D. Han, C. T. Morris, W. Lee, and B. Sarlioglu, "Comparison between output CM chokes for SiC drive operating at 20- and 200-kHz switching frequencies," *IEEE Trans. Ind. Appl.*, vol. 53, no. 3, pp. 2178–2188, May 2017.
- [3] D. Han, S. Li, Y. Wu, W. Choi, and B. Sarlioglu, "Comparative analysis on conducted CM EMI emission of motor drives: WBG versus Si devices," *IEEE Trans. Ind. Electron.*, vol. 64, no. 10, pp. 8353–8363, Oct. 2017.
- [4] G. Engelmann, A. Sewergin, M. Neubert, and R. W. De Doncker, "Design challenges of SiC devices for low- and medium-voltage DC–DC converters," *IEEJ J. Ind. Appl.*, vol. 8, no. 3, pp. 505–511, May 2019.
- [5] B. Zhang and S. Wang, "A survey of EMI research in power electronics systems with wide-bandgap semiconductor devices," *IEEE J. Emerg. Sel. Topics Power Electron.*, vol. 8, no. 1, pp. 626–643, Mar. 2020.
- [6] S. Takahashi, K. Wada, H. Ayano, S. Ogasawara, and T. Shimizu, "Review of modeling and suppression techniques for electromagnetic interference in power conversion systems," *IEEJ J. Ind. Appl.*, vol. 11, no. 1, pp. 7–19, Sep. 2022.
- [7] Y. Koyama, M. Tanaka, and H. Akagi, "Modeling and analysis for simulation of common-mode noises produced by an inverter-driven air conditioner," *IEEE Trans. Ind. Appl.*, vol. 47, no. 5, pp. 2166–2174, Sep. 2011.
- [8] B. Touré, J.-L. Schanen, L. Gerbaud, T. Meynard, J. Roudet, and R. Ruelland, "EMC modeling of drives for aircraft applications: Modeling process, EMI filter optimization, and technological choice," *IEEE Trans. Power Electron.*, vol. 28, no. 3, pp. 1145–1156, Mar. 2013.
- [9] E. Rondon-Pinilla, F. Morel, C. Vollaire, and J.-L. Schanen, "Modeling of a buck converter with a SiC JFET to predict EMC conducted emissions," *IEEE Trans. Power Electron.*, vol. 29, no. 5, pp. 2246–2260, May 2014.
- [10] Q. Yu, T. W. Holmes, and K. Naishadham, "RF equivalent circuit modeling of ferrite-core inductors and characterization of core materials," *IEEE Trans. Electromagn. Compat.*, vol. 44, no. 1, pp. 258–262, Feb. 2002.
- [11] A. Eroglu, "Complete modeling of toroidal inductors for high power RF applications," *IEEE Trans. Magn.*, vol. 48, no. 11, pp. 4526–4529, Nov. 2012.
- [12] C. R. Sullivan and A. Muetze, "Simulation model of common-mode chokes for high-power applications," *IEEE Trans. Ind. Appl.*, vol. 46, no. 2, pp. 884–891, Mar./Apr. 2010.
- [13] M. L. Heldwein, L. Dalessandro, and J. W. Kolar, "The three-phase common-mode inductor: Modeling and design issues," *IEEE Trans. Ind. Electron.*, vol. 58, no. 8, pp. 3264–3274, Aug. 2011.
- [14] M. Kovacic, Z. Hanic, S. Stipetic, S. Krishnamurthy, and D. Zarko, "Analytical wideband model of a common-mode choke," *IEEE Trans. Power Electron.*, vol. 27, no. 7, pp. 3173–3185, Jul. 2012.
- [15] K. Nomura, T. Kojima, and Y. Hattori, "Straightforward modeling of complex permeability for common mode chokes," *IEEJ J. Ind. Appl.*, vol. 7, no. 6, pp. 462–472, Sep. 2018.
- [16] S. Takahashi, S. Ogasawara, M. Takemoto, K. Orikawa, and M. Tamate, "A modeling technique for designing high-frequency three-phase common-mode inductors," in *Proc. IEEE Energy Convers. Cong. Expo.*, Nov. 2018, pp. 6600–6606.
- [17] B. Wunsch, S. Skibin, V. Forsstrom, and T. Christen, "Broadband modeling of magnetic components with saturation and hysteresis for circuit simulations of power converters," *IEEE Trans. Magn.*, vol. 54, no. 11, pp. 1–5, Nov. 2018.
- [18] S. Takahashi and S. Ogasawara, "A novel simulation model for commonmode inductors based on permeance-capacitance analogy," in *Proc. IEEE Energy Convers. Cong. Expo.*, Dec. 2020, pp. 5862–5869.
- [19] S. Takahashi and S. Maekawa, "Wideband small-signal model of commonmode inductors based on stray capacitance estimation method," *IEEJ J. Ind. Appl.*, vol. 11, no. 3, pp. 514–521, 2022.
- [20] I. Stevanovic, S. Skibin, M. Masti, and M. Laitinen, "Behavioral modeling of chokes for EMI simulations in power electronics," *IEEE Trans. Power Electron.*, vol. 28, no. 2, pp. 695–705, Feb. 2013.
- [21] C. Cuellar, N. Idir, and A. Benabou, "High-frequency behavioral ring core inductor model," *IEEE Trans. Power Electron.*, vol. 31, no. 5, pp. 3763–3772, May 2016.
- [22] F. G. Brockman, P. H. Dowling, and W. G. Steneck, "Dimensional effects resulting from a high dielectric constant found in a ferromagnetic ferrite," *Phys. Rev.*, vol. 77, pp. 85–93, Jan. 1950.

- [24] EPCOS AG. (May 2017). Ferrites and Accessories SIFERRIT Material N30. [Online]. Available: https://www.tdk-electronics.tdk.com/dow nload/540182/05a0bcaec69256cbbdbec9fecb243311/pdf-n30.pdf
- [25] G. R. Skutt and F. C. Lee, "Characterization of dimensional effects in ferrite-core magnetic devices," in *Proc. PESC Record*, 27th Annu. *IEEE Power Electron. Specialists Conf.*, vol. 2. Baveno, Italy, Jan. 1996, pp. 1435–1440.
- [26] M. Kacki, M. S. Rylko, J. G. Hayes, and C. R. Sullivan, "Magnetic material selection for EMI filters," in *Proc. IEEE Energy Convers. Congr. Expo.* (ECCE), Oct. 2017, pp. 2350–2356.
- [27] A. Furuya, Y. Uehara, K. Shimizu, J. Fujisaki, T. Ataka, T. Tanaka, and H. Oshima, "Magnetic field analysis for dimensional resonance in Mn–Zn ferrite toroidal core and comparison with permeability measurement," *IEEE Trans. Magn.*, vol. 53, no. 11, pp. 1–4, Nov. 2017.
- [28] B. Wunsch, T. Christen, S. Skibin, and V. Forsstrom, "Broadband circuit model of a ferrite core, including dimensional resonance, saturation, and hysteresis," *IEEE Trans. Magn.*, vol. 55, no. 7, pp. 1–5, Jul. 2019.
- [29] S. Takahashi, S. Ogasawara, M. Takemoto, K. Orikawa, and M. Tamate, "Experimental evaluation of the relationship between filter inductor impedances and dimensional resonances of MnZn ferrites," in *Proc. IEEE* 4th Int. Future Energy Electron. Conf. (IFEEC), Nov. 2019, pp. 1–8.
- [30] L. Dalessandro, W. G. Odendaal, and J. W. Kolar, "HF frequency characterization and nonlinear modeling of a gapped toroidal magnetic structure," *IEEE Trans. Power Electron.*, vol. 21, no. 5, pp. 1165–1175, Sep. 2006.
- [31] F. Fiorillo, C. Beatrice, O. Bottauscio, and E. Carmi, "Eddy-current losses in Mn–Zn ferrites," *IEEE Trans. Magn.*, vol. 50, no. 1, pp. 1–9, Jan. 2014.
- [32] M. Kacki, M. S. Rylko, J. G. Hayes, and C. R. Sullivan, "Measurement methods for high-frequency characterizations of permeability, permittivity, and core loss of Mn–Zn ferrite cores," *IEEE Trans. Power Electron.*, vol. 37, no. 12, pp. 15152–15162, Dec. 2022.
- [33] A. Roc'h and F. Leferink, "Nanocrystalline core material for highperformance common mode inductors," *IEEE Trans. Electromagn. Compat.*, vol. 54, no. 4, pp. 785–791, Aug. 2012.
- [34] J. Yao, Y. Li, H. Zhao, and S. Wang, "Design of CM inductor based on core loss for radiated EMI reduction in power converters," in *Proc. IEEE Appl. Power Electron. Conf. Expo.*, Mar. 2019, pp. 2673–2680.
- [35] S. Wang, F. C. Lee, D. Y. Chen, and W. G. Odendaal, "Effects of parasitic parameters on EMI filter performance," *IEEE Trans. Power Electron.*, vol. 19, no. 3, pp. 869–877, May 2004.
- [36] S. Wang, F. C. Lee, W. G. Odendaal, and J. D. van Wyk, "Improvement of EMI filter performance with parasitic coupling cancellation," *IEEE Trans. Power Electron.*, vol. 20, no. 5, pp. 1221–1228, Sep. 2005.
- [37] C. Cuellar and N. Idir, "Reduction of the parasitic couplings in the EMI filters to improve the high frequency insertion loss," in *Proc. 44th Annu. Conf. IEEE Ind. Electron. Soc. (IECON)*, Oct. 2018, pp. 5766–5771.



SHOTARO TAKAHASHI (Member, IEEE) was born in Hokkaido, Japan. He received the B.S., M.S., and Ph.D. degrees in electrical engineering from the Graduate School of Information Science and Technology, Hokkaido University, Japan, in 2015, 2017, and 2020, respectively.

From April 2020 to March 2021, he was a Project Assistant Professor at Tokyo Metropolitan University, Hachioji, Tokyo, Japan. He has been an Assistant Professor with Seikei University,

Kichijoji, Tokyo, since April 2021. His current research interests include the electromagnetic compatibility of high-frequency switching power converters, active electromagnetic interference filters, and modeling of magnetic components. He is a member of the Institute of Electrical Engineers of Japan (IEEJ) and the Institute of Electronics, Information and Communication Engineers (IEICE). He was a recipient of the 20th International Conference on Electrical Machines and Systems (ICEMS2017) Best Paper Award (Third Prize), in 2017, and the IEEJ Industry Applications Society Distinguished Transaction Paper Award, in 2018.