Feasible Evaluations of Coupled Multilayered Chip Inductor for POL Converters

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Introduction

Recently, high efficiency and high power density POL (Point of load) converters have attracted great interest in networking, telecommunications, and computing applications (1–5). Usually, POL converters are directly installed around digital equipment such as MCUs (Micro Controller Unit), FPGAs (Field-Programmable Gate Array) and ASICs (Application Specific Integrated Circuit) on mother boards, in order to reduce the influence of parasitic components on the route between its output terminal and the digital equipment. In addition, the interleaved technique, magnetic integration, and application of GaN FETs are well known as good approaches to satisfy these demands in the power converters. Although coupled inductors for POL converters have been proposed in several studies, a coupled multilayered chip inductor has not been examined because of the difficulty of its construction. In this paper, a novel coupled multilayered chip inductor for interleaved POL converters is proposed. This novel coupled inductor has a pair of windings with inverse coupling in the magnetic core. Further, the magnetic material of the coupled inductor is an Fe-based metal composite powder (Fe-Si-Cr). Additionally, an Fe-based powder in the magnetic core has been processed with electrical insulation by a highly crystallized oxide nanolayer in order to reduce the eddy-current losses. Finally, the high efficiency performance of the coupled multilayered chip inductor is evaluated by prototypes of interleaved buck converters using normally off-type GaN FETs with a switching frequency of 1 MHz.

Keywords: multilayered chip inductor, coupled inductor, POL converter, interleaved converter, GaN FETs, DC superposition characteristics

Therefore, in order to achieve both, small converter size and high-efficiency, interleaved circuit topology, Gallium Nitride (GaN) devices, and magnetic integration techniques have been applied to this converter. In addition, as attractive features of the interleaved topology, this technique allows the use of small output filter capacitance because of the small voltage ripple constraints (6–7). Moreover, this technique contributes to the reduction of EMI contents and conduction loss in each passive/active device (8). On the other hand, Gallium Nitride (GaN) devices are frequently reported as low-loss and high speed switching devices in comparison to conventional silicon (Si) power devices (9–11). Therefore, using GaN FETs, it is possible to operate at a higher frequency switching and contribute to reduce the dimensions of passive components without reducing the power conversion efficiency.

Moreover, integrated magnetic components are applied to reduce the volume and the weight of magnetic components and to improve the performance of the interleaved converter (12–17). In addition, integrated magnetic components have been applied owing some attractive features. First, higher frequency operation of the inductor currents can be achieved by the effect of the mutual induction. And, second, DC fluxes, which are generated in proportion to the inductor average current, can be canceled by magnetic coupling.

Incidentally, there are two types of magnetic structures with respect to inductors in POL converters. One is wire-wound typed inductors, and the other is multilayered chip typed inductors. Although the former has a large
current-handling capacity by using wires, however, there are limitations to molding and processing a miniature drum core. Therefore, this type may be unsuitable for small POL converters. In contrast, the latter is able to realize compact and low profile inductors by the improving of multilayered technology in the recent years and achieving very high space factor for the windings. Though this type has not high inductance value as compared to the general wire-wound types, the multilayered chip inductor and high frequency driving by GaN FETs have an affinity for each other. However, “coupled multilayered chip inductor” has not examined yet because of the difficulty of its construction. Therefore, this paper proposes a novel coupled multilayered chip inductor and evaluates it by the interleaved POL converter using GaN FETs at a switching frequency of 1 MHz.

The magnetic material of the proposed coupled inductor is an Fe-based metal powder with high saturation flux density ($B_{sat}$). Generally, ferrite has been known as one of the favorite material of low loss under high switching frequency condition. However, this material is unsuitable for high power density because $B_{sat}$ is not high enough, and its inductance value rapidly decreased with the increasing current.

On the other hand, an Fe-based metal composite material (powder core) has an excellent feature of high $B_{sat}$, and its inductance value is gradual decreased even if the inductor current increases. However, this material has the eddy-current loss problem originated from the low-electrical resistivity. To solve this problem, an Fe-based magnetic material of the proposed inductor has been processed with electrical insulation by the highly crystallized oxide nanolayer without using resin (3). The purpose of this study is to discuss the performances of the novel coupled multilayered chip inductor in the interleaved POL converter as a case of study.

In Sect. 2, circuit analysis is carried out to show the effectiveness of the coupled inductor and its physical behavior. In Sect. 3, we will discuss the structure of the coupled multilayered chip inductor. Then, the introduction of the magnetic material and the result of its electrical insulation by the highly crystallized oxide nanolayer will be shown. In Sect. 4, DC current superposition characteristic of the coupled inductor will be investigated.

Finally, two coupled multilayered chip inductors with high and low coupling coefficient are introduced, and we evaluate these inductors to confirm the relationship between the coupling coefficient and its efficiency. The effectiveness of the novel coupled multilayered chip inductor is discussed from the theoretical and experimental point of view.

2. Interleaved POL Converter with a Coupled Multilayered Chip Inductor

2.1 Steady State Circuit Analysis of the Interleaved Converter with a Coupled Inductor

Figure 1 shows the interleaved buck converter with a coupled inductor. This converter is composed of switches $S_1$ and $S_2$, diodes $D_1$ and $D_2$ and capacitors $C_1$ and $C_o$ for smoothing or decoupling in input and output sides. The coupled inductor is mounted on the converter to reduce the volume and eliminate the dead space. In addition, the relationship between each winding is the inversely coupling in order to ensure the good characteristics described in Sect. 1. Here, $L_1$ and $L_2$ are the self-inductances in each phase and $M$ is the mutual inductance between each winding.

Firstly, to show the relationship between $M$ and the inductor ripple current, the circuit analysis is carried out. In the case of two phase interleaved converters, the switches are switched with 180 degree phase shift as shown in Fig. 2. Therefore, circuit analysis is performed separately for each mode. (It is notable that parasitic components such as capacitors and inductors on the PCB and power devices are not considered in order to show only the characteristics of the coupled inductors.) The relationships between the voltage across the inductor windings $v_{1i}$ and $v_{2i}$, and inductor currents $i_{1i}$ and $i_{2i}$ are represented as follows:

\[
\begin{align*}
\frac{dL_1}{dt} & = \frac{1}{L_1} \left( \frac{1}{2} V_i - V_o \right) + \frac{1}{L_1} \frac{1}{2} V_i \\
\frac{dL_2}{dt} & = \frac{1}{L_2} \left( \frac{1}{2} V_i - V_o \right) - \frac{1}{L_2} \frac{1}{2} V_i 
\end{align*}
\]

Generally, the inductances $L_1$ and $L_2$ are designed and manufactured with the same self-inductance value. Thus, $L_1$ and $L_2$ are approximated as $L_1 = L_2 = L$.

<Model 1: $S_1$:on-state, $S_2$:off-state> In this mode, $v_{1i} = V_i - V_o$, $v_{2i} = -V_o$. Therefore, the slopes of the inductor currents are given from (1).

\[
\begin{align*}
\frac{di_{1i,mod1}}{dt} & = \frac{1}{L_1} \left( \frac{1}{2} V_i - V_o \right) + \frac{1}{L_1} \frac{1}{2} V_i \\
\frac{di_{2i,mod1}}{dt} & = \frac{1}{L_2} \left( \frac{1}{2} V_i - V_o \right) - \frac{1}{L_2} \frac{1}{2} V_i 
\end{align*}
\]

<Model 2: $S_1$:off-state, $S_2$:on-state> This mode is a symmetrical switching condition of model1. Thus, relational equations for the inductor current are given by:

\[
\begin{align*}
\frac{di_{1i,mod2}}{dt} & = \frac{1}{L_1} \left( \frac{1}{2} V_i - V_o \right) - \frac{1}{L_1} \frac{1}{2} V_i \\
\frac{di_{2i,mod2}}{dt} & = \frac{1}{L_2} \left( \frac{1}{2} V_i - V_o \right) + \frac{1}{L_2} \frac{1}{2} V_i 
\end{align*}
\]
Mod3

<Mode3: S1:off-state, S2:off-state> In this mode, the switches in each phase are in off-state together. Accordingly, the following equation is obtained:

\[
\frac{dh_{1, \text{mod3}}}{dt} = \frac{dh_{2, \text{mod3}}}{dt} = \frac{1}{(L - M)} \cdot (-V_o) \cdots (4)
\]

<Mode4: S1:on-state, S2:on-state> In the same way, two switches are in on-state. Therefore, the slopes of the inductor currents are as follows:

\[
\frac{dh_{1, \text{mod4}}}{dt} = \frac{dh_{2, \text{mod4}}}{dt} = \frac{1}{(L - M)} \cdot (V_i - V_o) \cdots (5)
\]

If these current behaviors are summarized mathematically from (2)–(5), each current will have the following relationships:

\[
\begin{align*}
\frac{dh_1}{dt} &= \frac{dl_{\text{com}}}{dt} + \frac{dl_{\text{wh}}}{dt} \\
\frac{dh_2}{dt} &= \frac{dl_{\text{com}}}{dt} - \frac{dl_{\text{wh}}}{dt} \\
\frac{dl_{\text{com}}}{dt} &= \frac{1}{(L - M)} \left( (s_1 + s_2) \cdot \frac{1}{2} V_i - V_o \right) \\
\frac{dl_{\text{wh}}}{dt} &= \frac{1}{(L + M)} \cdot (s_1 - s_2) \cdot \frac{1}{2} V_i
\end{align*}
\]

Here, \(s_1\) and \(s_2\) are logic functions which show the switching conditions, and are given by the following expression:

\[
\begin{align*}
\{ s_1 = 1 \ (S_1:\text{ON}), \\
\{ s_1 = 0 \ (S_1:\text{OFF}) \\
\{ s_2 = 1 \ (S_2:\text{ON}), \\
\{ s_2 = 0 \ (S_2:\text{OFF}) \cdots (7)
\end{align*}
\]

As seen in (6), the inductor current can be separated into the common current \(i_{\text{com}}\) in each phase, and the wheeling current \(i_{\text{wh}}\). The common current is one of the current components which is related to the leakage inductance \(L_k\) (\(L_k = L - M\)), and can be observed in all the switching modes. On the other hand, the wheeling current is a current component which presents the same absolute value and reverse direction in each phase. Mainly, \(i_{\text{wh}}\) relates to a transformer in the coupled inductor by the inductance value of \(L + M\), and it is generated in Modes 1 and 2.

Considering the above electrical behavior, the routes of each inductor current component in Mode 1 and 2 can be illustrated as shown in Fig. 3. As seen in this Fig. 3, the wheeling current \(i_{\text{wh}}\) flows between the windings. For these reasons, \(i_{\text{wh}}\) has no relation to the power supply in the output side. Therefore, only \(i_{\text{com}}\) has possibility to contribute in the power transfer to the output side. In other words, \(i_{\text{com}}\) contains DC current components, \(i_{\text{wh}}\) has AC current components.

2.2 Inductor Ripple Current of the Coupled Inductor

Based on the above results of the circuit analysis, the relationship between the inductor ripple current and each inductance are investigated theoretically in this subsection.

By using (6), the electrical behaviors of the inductor current including its components such as \(i_{\text{com}}\) or \(i_{\text{wh}}\) are illustrated in Fig. 4. As seen in this figure, \(i_{\text{com}}\) operates at a higher frequency in comparison to the single-phase converters. In particular, change of the slope of \(i_{\text{com}}\) synchronizes with the transition of the switching modes. On the other hand, the change of the slope of \(i_{\text{wh}}\) occurs when the switching state is different in each phase.

Here, inductor ripple current \(I_{\text{rpp}}\) is the sum of the current ripples \(i_{\text{com}}\) and \(i_{\text{wh}}\). Therefore, \(I_{\text{rpp}}\) is given by;

\[
I_{\text{rpp}} = |I_{\text{compp}}| + |I_{\text{whpp}}| \cdots \cdots \cdots \cdots \cdots \cdots (8)
\]

At a duty ratio of \(d < 0.5\), the peak-to-peak amplitude of \(I_{\text{rpp}}\) can be calculated during Mode 1. Hence, the values of ripple current of \(I_{\text{compp}}\) and \(I_{\text{whpp}}\) can be obtained from (6):

\[
\begin{align*}
I_{\text{compp}, d < 0.5} &= \frac{1}{L_k} \cdot \left( \frac{1}{2} - d \right) \cdot V_i \cdot d \cdot T_s \\
I_{\text{whpp}, d < 0.5} &= \frac{1}{(L_k + 2M)} \cdot \frac{1}{2} \cdot V_i \cdot d \cdot T_s \quad \text{(d < 0.5)}
\end{align*}
\]

Identically, \(I_{\text{rpp}}\) is shown in Mode 2 in the case of \(d \geq 0.5\). Therefore, the following relationships can be obtained:

\[
\begin{align*}
I_{\text{compp}, d \geq 0.5} &= \frac{1}{L_k} \cdot \left( d - \frac{1}{2} \right) \cdot V_i \cdot (1-d) \cdot T_s \\
I_{\text{whpp}, d \geq 0.5} &= \frac{1}{(L_k + 2M)} \cdot \frac{1}{2} \cdot V_i \cdot (1-d) \cdot T_s \quad \text{(d \geq 0.5)}
\end{align*}
\]

By using (8)–(10), inductor ripple current can be calculated theoretically. Therefore, to confirm the validity of (8)–(10), a circuit simulation is carried out. Simulated circuit parameters are shown in Table 1. The output voltage is fixed at 1 V considering a power supply to digital ICs, and the input voltage is greatly varied between 1 V–10 V in order to evaluate the relationship between the duty ratio and the
inductor ripple current. Simulated and theoretical results in all the ranges of duty ratio are shown in Fig. 5. In this figure, the solid line shows the theoretical value using (8)–(10), and the dots mean the simulated results. As seen in this figure, the theoretical values agree with the simulated values closely. Therefore, the validity of (8) is confirmed from the simulated point of view. On the other hand, focusing on the relationship between these currents and the duty ratio, $I_{\text{compp}}$ is the smallest at $d = 0.5$, and effectively reduced around $d = 0.5$. This is because the voltages across the leakage inductances in each phase are zero at $d = 0.5$. On the other hand, $I_{\text{whpp}}$ has a constant value at $d < 0.5$, and it is reduced at $d \geq 0.5$. $I_{\text{lep}}$ is determined by the sum of $I_{\text{compp}}$ and $I_{\text{whpp}}$, also it has been confirmed from simulation results.

Then, to indicate the effectiveness of the coupled inductor, the inductor ripple current is compared to the non-coupled inductor in interleaved converters. The inductor ripple current $I_{\text{Lnon}}$ of the non-coupled inductor can be represented by the following equation:

$$I_{\text{Lnon}} = \frac{1}{L_{\text{non}}} \cdot V_i \cdot (1 - d) \cdot d \cdot T_s,$$

where $L_{\text{non}}$ is the self-inductance of the non-coupled inductor. As a comparison condition of the ripple current, $I_{\text{lep}}$ of the coupled inductor has equal values than $L_{\text{non}}$. In addition, the coupling coefficients of the coupled inductor are changed from 0.2 to 0.9 in order to investigate the coupling effect.

Here, the coupling coefficient $k$ is given by:

$$k = \frac{M}{L} = \frac{M}{L_d + M} \tag{12}$$

Therefore, if (12) is replaced into (9) and (10) and then (8)/(11) is calculated, the ratio of the inductor ripple current can be calculated. The result is shown in Fig. 6. As seen in this figure, the high coupling coefficient is very effective for reducing the ripple current as compared to the non-coupled inductor. The reduction effect of the inductor ripple current is effective with a focus on $d = 0.5$, especially.

Therefore, coupled inductors have good characteristics from the electrical points of view.

### 3. Structure of the Coupled Multilayered Chip Inductor and its Magnetic Material

#### 3.1 Magnetic Structure

Structures of the coupled multilayered chip inductor are discussed in this section. Magnetic core structure is one of the important parts of the magnetic design. Generally, the windings of multilayered chip inductors are covered by magnetic cores. The advantage of the multilayered chip inductors is the ability of getting high winding factor. In addition, control circuits for POL converters are immune to the effect of the external leakage flux of the windings because this core structure covers magnetic core with higher relative permeability ($\mu_r$) in comparison to the permeability of the free space ($\mu_0$).

Several studies have been discussed and proposed regarding magnetic coupling core structures for wire-wound type inductors. E-E or E-I cores with typical wire-wound type are proposed in order to simply the manufacturing process (13). CCTT core split-winding structure is effective for reducing the parasitic capacitance between the windings or the external leakage flux of the windings (14). E-E or E-I core structure with laminated or bifilar windings (15) (16) and E-I-E core structure (17) for obtaining high coupling coefficients are proposed as well.

Some magnetic core structures for the coupled multilayered chip inductors can be considered based on the above references, and they are shown in Figs. 7(a)–(c).

In the case of (a), although this structure is relatively easy to manufacture, high coupling coefficient cannot be obtained because the magnetic flux has a straightness characteristic. In addition, installing an air-gap for obtaining high coupling coefficient has the difficulty to keep a constant quality or reliability of the product in case of the multilayered chip inductors. (b) is effective for reducing the parasitic capacitances between the windings in the case of a relatively large core for high power applications. However, the internal structure of the multilayered chip inductors is likely to be complicated.

Last is the structure (c). The advantage of the structure (c) is the very high coupling coefficient that is easy to obtain. The effectiveness of this structure has been reported at the close-coupled inductor method which is separated into an energy storage inductor and a close-coupled transformer (18).
and Fe-based powder materials have been discussed\(^{(18)}\)–\(^{(20)}\).

Cation, several magnetic materials such as amorphous, ferrite high frequency converters such as POL and high power application reduce iron losses and to downsize magnetic components. For this purpose, Fe-Si-Cr metal composite powder is processed with electrical insulation by the highly crystallized oxide nanolayer. Figure 10 shows the Transmission Electron Microscope (TEM) images of the metal composite material for the proposed coupled multilayered chip inductor. This metal composite material has thin oxide layers with the thickness of a few hundred nanometers on the metal powder surface. By applying this insulating process by the highly crystallized oxide nanolayer, three advantages are obtained in comparison to a layer of resin\(^{(18)}\).

First, high $\mu_r$ can be obtained by a thin insulating layer, and the result contributes to get high inductance or to reduce the winding turns. Figure 11 shows the measurement results of magnetic flux density ($B_{sat}$) is not high enough, and the inductance value is rapidly decreased with the increasing inductance current. Further, they cannot tolerate high temperature because the curie temperature is around 200°C.

On the other hand, an Fe-based composite material is an attractive material for the inductor, because of the higher $B_{sat}$ than ferrite materials. Therefore, this material is effective for downsizing magnetic components on power converter. Based on these facts, reference\(^{(15)}\) has achieved downsizing magnetic components by applying powder cores.

In this paper, Fe-Si-Cr chemical components shown in Fig. 9 are applied as magnetic material for powder cores. This image is taken by the scanning electron microscope. Concerning about the material, Hysteresis loss characteristic of Fe-Si-Cr composite materials are the same as the general silicon steel.

The material of winding paste is silver (Ag). Silver has a characteristic of lower electrical resistivity in comparison to other metals such as Cu and Al. Therefore, the reduction of copper losses will be achieved. For these reasons, the proposed coupled multilayered chip inductor uses a silver paste winding as the internal electrode material.

### 3.3 Electrical Insulation of Iron Powder

Although Fe-based composite materials have some good characteristics, they may not be suitable for high frequency operation because the electrical resistivity of the material is lower in comparison to ferrite materials, etc. Hence, Fe-based powder cores have the possibility of increasing eddy current losses at high frequency operation.

Usually, Fe-based powder cores are conducted electric insulating treatment between powders by a resin layer. However, if insulated layers of thick resin are produced, the relative permeability ($\mu_r$) of the magnetic cores will be much smaller, and the result means an increasing number of winding turns to get the required inductances. From the above, it is important to reduce the thickness of the insulating layer between the powders.

For this problem, Fe-Si-Cr metal composite powder is processed with electrical insulation by the highly crystallized oxide nanolayer. Figure 10 shows the Transmission Electron Microscope (TEM) images of the metal composite material for the proposed coupled multilayered chip inductor. This metal composite material has thin oxide layers with the thickness of a few hundred nanometers on the metal powder surface. By applying this insulating process by the highly crystallized oxide nanolayer, three advantages are obtained in comparison to a layer of resin\(^{(18)}\).

First, high $\mu_r$ can be obtained by a thin insulating layer, and the result contributes to get high inductance or to reduce the winding turns. Figure 11 shows the measurement results of
Coupled Multilayered Chip Inductor for POL Converter (Jun Imaoka et al.)

Fig. 10. The magnetic core and its electrical insulation

Fig. 11. Comparative results of the permeability due to the difference of insulating layers.

Fig. 12. DC superposition characteristics of the high coupled, low coupled and non-coupled inductors.

Fig. 13. Magnetic field simulation results.

the permeability due to the difference of the insulating layers. In this case, the magnetic materials of the cores have the same Fe-Si-Cr composite materials. In addition, the magnetic core structure is toroidal for measuring magnetic characteristics. From this figure, the permeability of the magnetic core with conventional resin layer is around 19.5 and the case of the highly crystallized oxide nanolayer has a permeability of 28 at a range of 1 kHz~20 MHz. Therefore, this insulating treatment is effective for increasing the permeability, when there is a limit of the thickness of the resin layer. Although, the permeability of the core starts to increase from 20 MHz, this phenomenon does not mean that the permeability increase. Consequently, it is a resonant phenomenon due to the inductance and the parasitic capacitance, presented between the windings.

Second is to get higher mechanical strength. While the powder with resin shows almost 8 \times 10^4 Pa of mechanical strength, the powders with the highly crystallized oxide nanolayer show 1.5 \times 10^5 Pa.

Third, it can obtain high breakdown voltage characteristic. The breakdown voltage of the highly crystallized oxide nanolayer is 3.6 \times 10^4 V/m, and it is higher than the breakdown voltage of the resin layer (2.5 \times 10^4 V/m). The results suggest that highly crystallized oxide nanolayer has large electrical insulation. Therefore, considering the miniaturized chip inductor for POL converters, these attractive features are beneficial to improve the size and the product reliability from the magnetic point of view.

4. DC Superposition Characteristics of Coupled Inductor

Proposed coupled inductor consists of powder metal composite. Therefore, there is a need to investigate the DC superposition characteristics for the coupled inductor with the proposed structure. In order to perform this analysis, we conducted a simulation by JMAG (JSOL Corporation). In addition, we compared the coupled inductor with high or low coupling coefficients and the non-coupled inductor to confirm effectiveness of the coupled inductor. Concerning to each inductor size, coupled inductors with high coupling and low coupling have almost the same size, and a non-coupled inductor per phase has half size of the coupled inductors. In addition, inductor with high coupling has many winding turns as compared to low coupling inductors to get high inductance, and it has shortened distance x between the windings to obtain high coupling coefficient. Further, when the DC current condition at 0A, the coupled inductor with high coupling has \( L_{lk} = 1.2 \mu H, M = 0.63 \mu H \), and low coupling has \( L_{lk} = 1.17 \mu H, M = 0.17 \mu H \). On the other hand, the self-inductance \( L_{non} \) of the non-coupled inductor is 1.12 \\mu H. As seen in this condition, \( L_{lk} \) and \( L_{non} \) have almost the same value because these inductances are related to the DC flux which is generated according to the DC current.

Figure 12 shows the DC superposition characteristics. In case of the coupled inductor with high coupling, the decreasing rate of the \( L_{lk} \) is relatively early as compared to other inductors. To confirm the reason, Fig. 13 shows the magnetic field simulation results of each inductor under the same DC current condition. From this figure, it is understood that the leakage flux path of the coupled inductor with high coupling is high. Therefore, \( L_{lk} \) decreases earlier, because the relative permeability \( \mu_r \) of the magnetic material in this area decreases in an early stage. However, as understood from the analysis result in Section 2, not only \( L_{lk} \) value but also \( M \) value contributes to the inductor ripple current in case of the coupled inductor. Therefore, DC superposition
characteristics of \( L_{\text{lk}} \) and \( M \) of the coupled inductor with high coupling are shown in Fig. 14. From this figure, it is understood that \( M \) value does not decrease easily because each winding are inversely coupling and DC magnetic fluxes are effectively canceled. In addition, the leakage flux path with high magnetic reluctance contributes to keep the high \( M \) at high current condition. As a result, coupling coefficient \( k \) increases at high DC current condition.

Then, with the DC superposition characteristics, it is possible to investigate the inclusion of the effect of \( M \). In this way, to realize the same inductor ripple current between the coupled inductor and the non-coupled inductor, the effective inductance \( L_{\text{eff,coupled}} \) of the coupled inductor is defined by the following equation, from (8)–(11).

\[
L_{\text{non}} = L_{\text{eff,coupled}} = \begin{cases} 
\frac{2(1-d) \cdot L_{\text{lk}} \cdot (L_{\text{lk}} + 2M)}{(1-2d) \cdot (L_{\text{lk}} + 2M) + L_{\text{lk}}} & (d < 0.5) \\
\frac{2d \cdot L_{\text{lk}} \cdot (L_{\text{lk}} + 2M)}{(2d-1) \cdot (L_{\text{lk}} + 2M) + L_{\text{lk}}} & (d \geq 0.5)
\end{cases}
\]

\( L_{\text{eff,coupled}} \) depends on the duty ratio \( d \), \( M \), and \( L_{\text{lk}} \) values. By using (13), Fig. 15 shows an effective inductance comparison between the coupled inductor with high coupling and the non-coupled inductor. As seen in this figure, \( L_{\text{eff,coupled}} \) increases in comparison to the self-inductance \( L_{\text{non}} \) of the non-coupled inductor at all duty ranges. In addition, \( L_{\text{eff,coupled}} \) increases effectively at around duty 0.5, because the common ripple current \( I_{\text{comp}} \) does not occur in this duty range. Therefore, coupled inductor is effective to use in this duty range. The result strongly suggests that \( L_{\text{lk}} \) is possible to be reduced by high \( M \).

This property of the coupled inductor carries another advantage. Usually, DC magnetic flux \( \Phi_{\text{dc}} \) shown in Fig. 8, can be obtained by the following equation. (The deriving process of this equation is shown in Appendix 1)

\[
\Phi_{\text{dc}} = \frac{L_{\text{lk}} \cdot I_{\text{dc}}}{N} \quad \text{.......................... (14)}
\]

From (14), DC flux is related to only leakage inductance values because the windings between each phase are inversely coupled, and DC fluxes are effectively canceled on the path where magnetizing flux flows into. This feature carries an advantage. For example, if coupled inductor is designed at the same inductor ripple current with non-coupled inductor, \( L_{\text{lk}} \) of the coupled inductor can be reduced by high \( M \) value, and it can reduce the flux density under the same sectional area with the non-coupled inductor and high DC current conditions. In other words, coupled inductor can reduce the sectional area considering DC superposition characteristics with the effect of the mutual inductance. As a result, it contributes to the downsizing of magnetic components, or high reduction effect of the inductor ripple current can be obtained.

5. Experimental Results

To confirm the effectiveness of the coupled multilayered chip inductor, experimental evaluation is conducted in this section. The circuit parameters for this evaluation are shown in Table 2. First of all, two coupled multilayered chip inductors which have a different coupling coefficient are prepared in order to investigate the relationship between the power conversion efficiency and the coupling coefficient. Magnetic parameters of the two inductors are shown in Table 3.

The prototypes of the interleaved POL converter and the two coupled inductors which have a different coupling coefficient are shown in Fig. 16. In addition, \( C_1 \) and \( C_0 \) are multilayered ceramic capacitors (MLCC) for smoothing or decoupling considering low ESR and ESL characteristics.

Experimental waveforms using high and low coupling
In Sect. 2, inductor ripple current analysis is conducted and we use 1 MHz bandwidth limiting function in the oscilloscope. In this paper, the coupled multilayered chip inductor for POL converter applications. This work was partially supported by Grant-in-Aid for JSPS Fellows (No. 26110031).

6. Conclusion

In this paper, the coupled multilayered chip inductor for POL converter was proposed and evaluated. In the Sect. 2, inductor ripple current analysis is conducted to confirm the effect of magnetic coupling. As a result, the separation of the inductor ripple currents into common and wheeling ripple currents were succeeded. Especially, high mutual inductance is effective for reducing wheeling ripple current. In addition, it is understood that the common ripple current does not occur when the duty ratio is equal to 0.5 or around. The validity of this analysis result is confirmed by circuit simulation. In Sect. 3, magnetic core structure and magnetic material were discussed. The proposed magnetic core structure has simple structure, and it contributes to an easy manufacturing and design. In addition, the proposed coupled inductor has Fe-based metal composite, and this metal composite has processed by the highly crystallized oxide nanolayer in order to reduce the eddy-current losses. In Sect. 4, DC superposition characteristics of the proposed coupled inductors are investigated. Although the leakage inductance is easily decreased in the case of the coupled inductor with high coupling, the mutual inductance does not easily decrease because each winding is inversely coupled and DC magnetic fluxes are effectively canceled. As a result, the effective inductance of the coupled inductor, which has the same ripple current as the non-coupled inductor, is increased by the high mutual inductance. And also, its effect increases when the duty ratio is equal to 0.5. Finally, two coupled multilayered chip inductors, which have different coupling coefficients, are prepared and evaluated at a switching frequency of 1 MHz using GaN FETs. The coupled inductor with high coupling coefficient achieves a power conversion efficiency of 90%. This result is higher than the coupled inductor with low coupling coefficient. From the above, coupled multilayered chip inductor with high coupling is effective for POL converter applications.

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References

The relationship can be obtained:

\[ R_{lk} = R_{mt} + R_{nw} / R_{mt} \]  \hspace{1cm} (A1)

\[ R_{mt} \text{ and } R_{lw} \text{ are the magnetic reluctances of each leg. However, this}
\text{magnetic circuit model is complex. Therefore, (b) shows a simplification magnetic}
\text{circuit model. In order to change the magnetic circuit model to the simple model,}
\text{the following equation is used:}

\[ N_l1 = \frac{\phi_1}{R_{mt}} + 2\phi_{lk} / R_{lk} \]  \hspace{1cm} (A2)

\[ N_l2 = \frac{\phi_2}{R_{mt}} + 2\phi_{lk} / R_{lk} \]  \hspace{1cm} (A3)

Where, \( \phi_1 \) and \( \phi_2 \) are the magnetic fluxes in the inner sectional area. \( \phi_{lk} \) is the leakage flux which flows into the space between the windings. Here, \( i_1, i_2, \phi_1, \phi_2 \) are changed to \( I_{dc1}, I_{dc2}, \Phi_{dc1}, \) and \( \Phi_{dc2} \) for analyzing DC components. Furthermore, \( I_{dc1} \) and \( I_{dc2} \) are assumed as the same value (\( I_{dc1} = I_{dc2} = I_{dc} \)). From this equation, DC magnetic flux can be calculated as follows:

\[ \Phi_{dc1} = \Phi_{dc2} = \frac{N \cdot I_{dc}}{R_{mt} + 2 \cdot R_{lk}} \]  \hspace{1cm} (A4)

On the other hand, from this magnetic circuit model, each inductance can be represented by the following equation:

\[ L_{dc} = \frac{N^2}{M} \cdot \frac{R_{lk}}{R_{mt}^2 + 2 \cdot R_{mt} \cdot R_{lk}} \]  \hspace{1cm} (A5)

Therefore, by using Eqs. (A3) and (A4), the following equation can be obtained:

\[ \Phi_{dc1} = \Phi_{dc2} = \frac{L_{dc} \cdot I_{dc}}{N} \]  \hspace{1cm} (A6)

Conclusively, coupled inductors are effective for downsizing because the inductor ripple current can be reduced by high mutual inductance. In addition, the leakage inductance \( L_{dc} \) which is directly proportional to the DC flux can be reduced.

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Coupled Multilayered Chip Inductor for POL Converter  (Jun Imaoka et al.)

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