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Active Magnetic Decoupling for Improving the Performance of Integrated *LCL*-Filters in Grid-Connected Converters

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Abstract—Magnetically integrating converter-side and grid-side inductors is an effective way to reduce the volume and weight of *LCL* filters. However, the inevitable magnetic coupling between the two coils weakens the high order harmonic attenuation ability of *LCL* filters. In this paper, an active magnetic decoupling method is proposed to tackle this issue, and the basic operation principle is to counteract the coupling inductor in the filter capacitor branch by constructing a decoupling inductor. A step-by-step design guidance of integrated *LCL* filters with the proposed magnetic decoupling method is presented in detail, and it is also found that a further volume reduction of magnetic cores can be achieved compared to the existing passive decoupling method. Experimental results finally verify the effectiveness of the proposed method.

Index Terms—Active magnetic decoupling, core volume, filter performance, grid-connected converters, *LCL* filter, magnetic integration.

I. INTRODUCTION

Power electronics converters based on insulated gate bipolar transistor (IGBT) and metal-oxide-semiconductor field-effect transistor (MOSFET) have been widely used for various energy conversion applications, e.g., photovoltaic and wind generation, energy storage systems and uninterrupted power supplies [1], [2]. However, the use of pulse width modulation (PWM) introduces significant amounts of switching harmonics, which are absolutely not permitted in grid-connected applications due to the stringent grid standards, e.g., IEEE 519-1992 [3], IEEE std. 1547-2003 [4] and IEC 1000-3 [5]. Therefore, a low-pass power filter is required to attenuate the switching harmonics to a desirable level.

Traditionally, a simple *L* filter can fulfill this function, but a large inductance usually needs to be adopted, which is costly, bulky and inefficient. Therefore, higher order power filters, such as *LCL*, *LLCL*, *LTL*, *LTCL*, etc., have been widely

investigated due to their better harmonic attenuation ability with lower total inductance [6]–[11]. However, all of these filters except *LCL* filters have one or more *LC* traps to bypass or block the switching harmonics, which are very sensitive to filter parameters variation [8]–[10]. In comparison, *LCL* filters are more preferred in practical applications because of their robustness against filter parameters variation [8], [10].

LCL filters have been popularly used in voltage-source-type grid-connected converters [6], [7], [12], [13]. However, despite the reduction in the total inductance compared to *L* filters, two individual magnetic cores are required for converter-side and grid-side inductors. In this case, filter inductors still account for most of the weight and volume of the overall power conversion system. In order to further reduce the weight and volume of *LCL* filter inductors, one way is to increase the switching frequency by using wide bandgap (WBG) devices, so that the required filter inductance can be reduced. Another more cost-effective way is to adopt magnetic integration techniques [14]–[17]. The basic principle of magnetic integration is to integrate multiple discrete inductor windings on a common magnetic core. The weight and volume of magnetic cores thus can be reduced.

Magnetic integration of filter inductors has been deeply investigated in parallel interleaved converters. In [17]–[20], two filter inductors are magnetically integrated in two-phase interleaved DC-DC power converters, which can reduce the inductor size and also achieve better steady state and transient performances. Meanwhile, in multiple-paralleled three-phase voltage-source-converters (VSCs), the integrated inductors can be applied to not only reduce the inductor size but also suppress the circulating current [21]–[25]. Even though there have been a lot of works on magnetic integration of filter inductors, only a few studies on *LCL* filter inductors can be found [26]–[28]. The existing research results show that such an integration of *LCL* filter inductors will equivalently introduce a coupling inductor in the filter capacitor branch due to the magnetic coupling, which weakens the harmonic attenuation ability of *LCL* filters [28]. In order to mitigate the coupling effect, a decoupled EIE-type core structure is employed in [27], and [28] further presented a detailed analysis of the coupling effect on *LCL* filter performance with this core structure. A comparative study shows that more switching harmonics remain in the system compared to the case of discrete *LCL* filters, and a passive decoupling method by properly selecting the cross-section area and magnetic material of the common I-type core is proposed to solve this issue. However, the reduction of

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coupling coefficient is achieved at the expense of increased volume and cost of the common I-type core.

Instead of magnetic decoupling, [29] utilized the coupling inductor in the filter capacitor branch to form an integrated *LLCL* filter with an EE-type core structure. By properly designing the coupling inductor, an *LC-trap* is tuned at the switching frequency to cancel the most dominant harmonic current [8], [9]. However, as the inverse coupling is performed, the cross-section area of the common flux path is doubled compared to that of the respective flux path, thus leading to less volume reduction compared to the case of discrete *LCL* filters. In addition, compared to *LCL* filters, the high-frequency attenuation ability of *LLCL* filters is reduced from -60 dB/dec to -20 dB/dec, and thus the current harmonics beyond the switching frequency may become significant, particularly those around the double switching frequency.

In order to eliminate the magnetic coupling between converter-side and grid-side inductor windings, an active magnetic decoupling method is proposed in this paper. With a commonly used UIU-type core structure, a decoupling winding in series with the filter capacitor is wound on the common I-type core. The proposed method can regain the -60 dB/dec harmonic attenuation ability of *LCL* filters, and meanwhile achieve a further volume reduction of magnetic cores. Following the Introduction, Section II reviews the magnetic integration of *LCL* filters with a UIU-type core structure. The effect of magnetic coupling on *LCL* filter performance is discussed briefly. An active magnetic decoupling method is then proposed in Section III. The basic principle is presented in detail. Section IV illustrates a step-by-step design guidance of integrated *LCL* filters with the proposed magnetic decoupling method, and also the volume comparison is conducted. In Section V, an integrated *LCL* filter is fabricated in the laboratory and experimental results are presented to verify the effectiveness of the proposed design method. Section VI gives a discussion of the proposed method and Section VII finally concludes the paper.

II. MAGNETIC INTEGRATION AND COUPLING OF *LCL* FILTERS

Fig. 1. An *LCL*-type single-phase grid-connected converter.

Fig. 1 shows an *LCL*-type single-phase grid-connected converter, where the *LCL* filter consists of a converter-side inductor L_1 , a grid-side inductor L_2 , and a filter capacitor C_f . i_1 and i_2 are the converter-side and grid-side current respectively. i_c is the filter capacitor current. v_{con} is the converter output voltage, and e_g is the grid voltage. u_{dc} and i_{dc} denote dc-side voltage and current respectively. As seen, there are two filter

inductors (L_1 and L_2) in an *LCL* filter, and two individual magnetic cores for these two inductors are commonly employed, thus resulting in oversize and overweight of power converters.

In order to reduce the weight and volume of filter inductors, converter-side inductor L_1 and grid-side inductor L_2 can be integrated on a single magnetic core using the magnetic integration techniques, as shown in Fig. 2 and Fig. 3. In Fig. 2 and Fig. 3, a UIU-type decoupled core structure is employed, where the I-type core is used as the common flux path to mitigate the magnetic coupling of these two inductor windings due to its very small reluctance [14]. The inductances of L_1 and L_2 can be designed by adjusting the air-gaps (g_1 and g_2) or winding turns (N_1 and N_2).

Fig. 2. Inverse coupling. (a) Core structure of integrated inductor. (b) Equivalent circuit of integrated *LCL* filter.

Fig. 3. Direct coupling. (a) Core structure of integrated inductor. (b) Equivalent circuit of integrated *LCL* filter.

In Fig. 2(a), Φ_1 and Φ_2 are the total fluxes generated by converter-side current i_1 and grid-side current i_2 respectively. Φ_{11} and Φ_{12} (Φ_{21} and Φ_{22}) are the sub-fluxes of Φ_1 (Φ_2). As seen, the fluxes generated by i_1 and i_2 in each U-type core are in the opposite direction, e.g., Φ_1 and Φ_{22} , indicating inverse coupling. However, the fluxes in the common I-type core, i.e., Φ_{11} and Φ_{21} , are in the same direction, yielding the total flux $\Phi_{11} + \Phi_{21}$. Since the air-gap reluctances are far greater than the core reluctances, $\Phi_1 \approx \Phi_{11}$ and $\Phi_2 \approx \Phi_{21}$. In this case, the cross-section area of I-type core should be almost doubled compared to that of U-type cores to avoid magnetic saturation. In contrast, Fig. 3(a) shows the direct coupling of L_1 and L_2 , where Φ_1 and Φ_{22} (Φ_2 and Φ_{12}) are in the same direction. As also seen, the flux in the common I-type core is $\Phi_{11} - \Phi_{21}$, thus the cross-section area of I-type core can be designed far less than that of U-type cores [28].

Correspondingly, Fig. 2(b) and Fig. 3(b) show the equivalent circuits of the integrated LCL filter. It can be seen that both coupling methods equivalently introduce a coupling inductor M_{12} or $-M_{12}$ in the filter capacitor branch, which is caused by the magnetic coupling between converter-side and grid-side inductor windings [28], [29].

From the above equivalent circuits, the transfer functions from converter output voltage v_{con} to grid-side current i_2 can be derived as

$$G_{vii}(s) = \frac{1 + M_{12}C_f s^2}{(L_1 L_2 C_f - M_{12}^2 C_f) s^3 + (L_1 + L_2 - 2M_{12}) s}, \quad (1)$$

$$G_{vid}(s) = \frac{1 - M_{12}C_f s^2}{(L_1 L_2 C_f - M_{12}^2 C_f) s^3 + (L_1 + L_2 + 2M_{12}) s}, \quad (2)$$

where $G_{vii}(s)$ represents the inverse coupling and $G_{vid}(s)$ represents the direct coupling. Fig. 4 shows the Bode plots of $G_{vii}(s)$ and $G_{vid}(s)$. It can be seen that the high-frequency attenuation rate of both coupling methods is reduced from -60 dB/dec to -20 dB/dec, which obviously weakens the filter performance of LCL filters despite the reduced core weight and volume.

Fig. 4. Bode plots of $G_{vii}(s)$ and $G_{vid}(s)$.

In order to improve the filter performance, [28] proposed a passive magnetic decoupling method for the case of direct coupling, which shifts the intersection frequency f_{int} shown in Fig. 4 to a higher frequency. Nevertheless, an I-type core with higher permeability is demanded and the core size is still limited to mitigate the magnetic coupling. In contrast, [29] utilized the positive coupling inductor M_{12} introduced by the inverse coupling to form an integrated $LLCL$ filter shown in Fig. 4, where the trap frequency f_i is tuned at the switching frequency. However, it results in an oversized I-type core due to the inverse coupling and high current distortion as only the harmonic at the trap frequency could be well eliminated.

III. PROPOSED ACTIVE MAGNETIC DECOUPLING METHOD FOR INTEGRATED LCL FILTERS

In order to mitigate the magnetic coupling and improve the design flexibility, this paper proposes an active magnetic decoupling method for integrated LCL filters shown in Fig. 5, where L_1 and L_2 are integrated with the direct coupling method as in Fig. 3 to greatly reduce the volume of I-type core.

As seen from Fig. 5, the filter capacitor branch is wound on the common I-type core as the active magnetic decoupling winding to counteract the coupling inductor $-M_{12}$ shown in Fig.

3(b). In the I-type core, if the flux generated by i_c is in the same direction with that generated by i_1 , it is defined as positive coupling shown in Fig. 5(a). If the two fluxes are in the opposite direction shown in Fig. 5(b), it is defined as negative coupling. Fig. 6 shows the corresponding electrical connection diagram for the proposed design.

Fig. 5. Proposed active magnetic decoupling method for integrated LCL filters. (a) Positive coupling. (b) Negative coupling.

Fig. 6. Electrical connection diagram.

Fig. 7. Magnetic circuit model with positive coupling.

Fig. 7 shows the magnetic circuit model with positive coupling, where R_{g1} and R_{g2} denote the reluctances of air-gaps

(g_1 and g_2), R_{m1} and R_{m2} denote the reluctances of two U-type cores, R_{mc} denotes the reluctance of I-type core. Φ_{L1} and Φ_{L2} represent the fluxes through U-type cores respectively, and Φ_{Cf} represents the flux through I-type core. From Fig. 7, it can be derived as

$$\begin{cases} \Phi_{L1} = \frac{N_1 i_1}{R_1} + \frac{N_2 i_2}{R_{21}} + \frac{N_c i_c}{R_{c1}} \\ \Phi_{L2} = \frac{N_1 i_1}{R_{12}} + \frac{N_2 i_2}{R_2} - \frac{N_c i_c}{R_{c2}}, \\ \Phi_{Cf} = \frac{N_1 i_1}{R_{1c}} - \frac{N_2 i_2}{R_{2c}} + \frac{N_c i_c}{R_c} \end{cases} \quad (3)$$

where R_1 , R_2 and R_c denote the self-reluctances, and R_{12} , R_{21} , R_{1c} , R_{c1} , R_{2c} and R_{c2} denote the mutual-reluctances, expressed as

$$\begin{cases} R_1 = R_{m1} + 2R_{g1} + R_{mc} \parallel (R_{m2} + 2R_{g2}) \\ R_2 = R_{m2} + 2R_{g2} + R_{mc} \parallel (R_{m1} + 2R_{g1}) \\ R_c = R_{mc} + (R_{m1} + 2R_{g1}) \parallel (R_{m2} + 2R_{g2}) \\ R_{12} = R_{21} = R_1(R_{mc} + R_{m2} + 2R_{g2}) / R_{mc} \\ R_{1c} = R_{c1} = R_c(R_{m1} + 2R_{g1} + R_{m2} + 2R_{g2}) / (R_{m2} + 2R_{g2}) \\ R_{2c} = R_{c2} = R_c(R_{m1} + 2R_{g1} + R_{m2} + 2R_{g2}) / (R_{m1} + 2R_{g1}) \end{cases} \quad (4)$$

Since the air-gap reluctances are far greater than the core reluctances, (4) can be simplified as

$$\begin{cases} R_1 \approx 2R_{g1}, \quad R_2 \approx 2R_{g2}, \quad R_c \approx 2R_{g1}R_{g2} / (R_{g1} + R_{g2}) \\ R_{12} = R_{21} \approx 4R_{g1}R_{g2} / R_{mc} \\ R_{1c} = R_{c1} \approx 2R_{g1} \\ R_{2c} = R_{c2} \approx 2R_{g2} \end{cases} \quad (5)$$

According to (5), the reluctances in the magnetic circuit are mainly related to the reluctances of air-gaps (R_{g1} and R_{g2}) and the common I-type core (R_{mc}), which can be calculated as

$$R_{g1} = \frac{l_{g1}}{\mu_0 A_s}, \quad R_{g2} = \frac{l_{g2}}{\mu_0 A_s}, \quad R_{mc} = \frac{l_{mc}}{\mu_0 \mu_r A_{mc}}, \quad (6)$$

where l_{g1} and l_{g2} are the lengths of air-gaps for L_1 and L_2 respectively, l_{mc} is the effective path length of I-type core. A_s and A_{mc} are the cross-section areas of U-type and I-type cores respectively. μ_0 is the air permeability and equal to $4\pi \times 10^{-7}$ N/A², and μ_r is the relative permeability of I-type core.

Based on (3) and $i_1 = i_c + i_2$, the voltages across three windings can be derived as

$$\begin{cases} v_{L1} = -e_{L1} = N_1 \frac{d\Phi_{L1}}{dt} = (L_1 + M_{12}) \frac{di_1}{dt} - (M_{12} - M_{1c}) \frac{di_c}{dt} \\ v_{L2} = -e_{L2} = N_2 \frac{d\Phi_{L2}}{dt} = (L_2 + M_{12}) \frac{di_2}{dt} + (M_{12} - M_{2c}) \frac{di_c}{dt} \\ v_{Cf} = -e_{Cf} = N_c \frac{d\Phi_{Cf}}{dt} = L_c \frac{di_c}{dt} + M_{1c} \frac{di_1}{dt} - M_{2c} \frac{di_2}{dt} \end{cases} \quad (7)$$

where L_1 , L_2 and L_c represent the self-inductances, M_{12} , M_{1c} and M_{2c} represent the mutual-inductances, expressed as

$$\begin{cases} L_1 = \frac{N_1^2}{R_1} = \frac{N_1^2}{2R_{g1}}, \quad L_2 = \frac{N_2^2}{R_2} = \frac{N_2^2}{2R_{g2}} \\ L_c = \frac{N_c^2}{R_c} = \frac{N_c^2(R_{g1} + R_{g2})}{2R_{g1}R_{g2}}, \quad M_{12} = \frac{N_1 N_2}{R_{12}} = \frac{N_1 N_2 R_{mc}}{4R_{g1}R_{g2}} \\ M_{1c} = \frac{N_1 N_c}{R_{1c}} = \frac{N_1 N_c}{2R_{g1}}, \quad M_{2c} = \frac{N_2 N_c}{R_{2c}} = \frac{N_2 N_c}{2R_{g2}} \end{cases} \quad (8)$$

According to (7), if $M_{1c} = M_{2c}$, i.e., $N_1/R_{g1} = N_2/R_{g2}$ from (8), the equivalent LCL filter circuit is shown in Fig. 8.

Fig. 8. Equivalent LCL filter circuit with positive coupling.

From Section II, the equivalent coupling inductor in the filter capacitor branch weakens the filter performance of LCL filters. In order to counteract the coupling inductor, it can be derived from Fig. 8 as

$$L_c + 2M_{1c} = M_{12}. \quad (9)$$

Substituting (8) into (9) yields

$$2N_c^2(R_{g1} + R_{g2}) + 4N_1 N_c R_{g2} = N_1 N_2 R_{mc}. \quad (10)$$

Similarly, if the negative coupling shown in Fig. 5(b) is performed, the equivalent LCL filter circuit is shown in Fig. 9, and the constraints for magnetic decoupling are derived as

$$\begin{cases} N_1 / R_{g1} = N_2 / R_{g2} \\ 2N_c^2(R_{g1} + R_{g2}) - 4N_1 N_c R_{g2} = N_1 N_2 R_{mc} \end{cases} \quad (11)$$

Fig. 9. Equivalent LCL filter circuit with negative coupling.

According to the above derivations, $N_1/R_{g1} = N_2/R_{g2}$ should be met for both coupling methods, then the flux through the common I-type core can be obtained from (3) and (5) as

$$\begin{cases} \Phi_{Cf} = \frac{[N_c(R_{g1} + R_{g2}) + N_1 R_{g2}] i_c}{2R_{g1}R_{g2}} \quad (\text{Positive coupling}) \\ \Phi_{Cf} = \frac{[N_c(R_{g1} + R_{g2}) - N_1 R_{g2}] i_c}{2R_{g1}R_{g2}} \quad (\text{Negative coupling}) \end{cases} \quad (12)$$

For the negative coupling, it can be derived as $N_c \approx 2N_1 R_{g2} / (R_{g1} + R_{g2})$ from (11) due to $R_{mc} \ll R_{g1}$ and R_{g2} , then $\Phi_{Cf} \approx N_1 i_c / 2R_{g1}$. Since i_c is generally limited to be less than 5% of grid current i_2 [6]–[9], the flux Φ_{Cf} is thus very small, and a very

small size of I-type core can be employed. However, the number of winding turns N_c in this case is comparable with N_1 and N_2 , thus leading to higher copper loss and cost. In contrast, N_c can be far less than N_1 and N_2 for the positive coupling from (10), then Φ_{cf} can be also approximated as $N_1 i_c / 2R_{g1}$. Therefore, the positive coupling is performed in this paper.

Comparing Fig. 8 with Fig. 3(b), it can be seen that the proposed method actually constructs a decoupling inductor to counteract $-M_{12}$ caused by the magnetic coupling between converter-side and grid-side inductor windings. The decoupling inductor is $(L_c + 2M_{1c})$ from (9). Ideally, if $L_c + 2M_{1c} = M_{12}$, converter-side and grid-side inductor windings can be fully decoupled. However, in practice, the tolerances of air-gap length and core size, etc., may lead to the mismatch between the decoupling inductor $(L_c + 2M_{1c})$ and M_{12} . Considering the tolerance of 10%, i.e., $L_c + 2M_{1c} = 0.9 M_{12}$ or $1.1 M_{12}$, the equivalent coupling inductor in the filter capacitor branch is finally $\pm 0.1 M_{12}$. By contrast, the equivalent coupling inductor is always $-M_{12}$ for the conventional integration method without magnetic decoupling. Therefore, the proposed method always exhibits the superior performance compared to the conventional integration method.

IV. DESIGN GUIDANCE AND VOLUME COMPARISON

A. Design Guidance of Integrated LCL Filters

Table I lists the parameters of the system investigated in experiments, where grid-side inductor L_2 is chosen to be identical to L_1 for maximizing the utilization of inductance.

TABLE I
PARAMETERS OF SYSTEM

Description	Symbol	Value
Converter-side inductance	L_1/mH	0.45
Grid-side inductance	L_2/mH	0.45
Filter capacitance	$C_f/\mu\text{F}$	10
Resonant frequency	f_r/kHz	3.36
Grid voltage	e_g/V	110
Dc-voltage	u_{dc}/V	210
Output power	P_o/kW	0.65
Fundamental frequency	f_0/Hz	50
Switching frequency	f_s/kHz	10
Sampling frequency	f_s/kHz	10

1) Selection of U-Type Magnetic Cores

Due to the low cost and high permeability, a PC40-type MnZn ferrite core with minimum initial permeability $\mu_r = 1725$ N/A² is chosen in this paper. Its saturated flux density B_{sa} is 0.51 T at 25 °C, and the maximum flux density $B_{\max}$ is designed as 0.35 T with 30% margin. According to the well-known area-product method [30], the core geometrical constant K_g should meet the following constraint.

$$K_g = A_s W_A > \frac{L I_{\max} A_l}{K_u B_{\max}}, \quad (13)$$

where W_A denotes the window area of U-type core, K_u denotes

the utilization coefficient of W_A , specified as 0.5 here, A_l denotes the bare area of wires for inductor windings, and $I_{\max}$ denotes the maximum current.

Considering the switching current ripples in converter-side current i_1 and 1.2 times inrush current during startup or step change of loads, $I_{1\max} = 14$ A and $I_{2\max}$ is set the same as $I_{1\max}$. Litz wires with $A_l = 0.5\pi \text{ mm}^2$ are employed for inductor windings, then K_g should be greater than $0.57 \times 10^{-7} \text{ m}^4$ according to (13). The selected U-type cores for L_1 and L_2 have the same dimensions due to $L_1 = L_2 = 0.45 \text{ mH}$, shown in Fig. 10, where $A_s = e \cdot f = 7.84 \times 10^{-4} \text{ m}^4$, and the core geometrical constant K_g is $6.7 \times 10^{-7} \text{ m}^4$, thus (13) is valid.

Fig. 10. Dimensions of U-type magnetic cores.

2) Calculation of Winding Turns and lengths of air-gaps for L_1 and L_2

The design of winding turns for L_1 and L_2 should avoid magnetic saturation of the U-type cores, i.e.,

$$N > \frac{L I_{\max}}{A_s B_{\max}}. \quad (14)$$

Additionally, the constraint $N_1/R_{g1} = N_2/R_{g2}$ should also be satisfied from Section II. From (8), it can be derived as

$$\frac{L_1}{L_2} = \frac{N_1^2}{N_2^2} \cdot \frac{R_{g2}}{R_{g1}}. \quad (15)$$

Then, the constraint $N_1/R_{g1} = N_2/R_{g2}$ can be transformed into $L_1/N_1 = L_2/N_2$. With this expression, the same constraint can also be found in [28]. In this paper, $L_1 = L_2$, thus N_1 is equal to N_2 , and set as 27 turns to ensure that (14) is valid.

According to (6) and (8), the lengths of air-gaps are calculated as

$$l_{g1} = l_{g2} \approx 0.8 \text{ mm}. \quad (16)$$

3) Selection of I-Type Magnetic Core and calculation of Winding Turns N_c

The same magnetic material is chosen for the I-type core, and the width and length are the same as those of the U-type cores shown in Fig. 11. With these dimensions, the effective path length l_{mc} of I-type core is 59 mm calculated by $(a-f) \cdot f$ is shown in Fig. 10.

Fig. 11. Dimensions of I-type magnetic core.

In Fig. 11, h denotes the height of I-type core, which is limited by the maximum flux through I-type core and subject to the following constraint,

$$B_{\max} A_{mc} > \Phi_{Cf \max}, \quad A_{mc} = e * h. \quad (17)$$

From (12), the flux Φ_{Cf} with positive coupling is approximated as $N_1 i_c / 2R_{g1}$, while the fluxes Φ_{L1} and Φ_{L2} through U-type cores are approximated as $N_1 i_1 / 2R_{g1}$ and $N_2 i_2 / 2R_{g2}$ from (3) and (5). Thus, it can be yielded as

$$\frac{A_{mc}}{A_s} > \frac{\Phi_{Cf \max}}{\Phi_{L1 \max}} = \frac{I_{Cf \max}}{I_{1 \max}}. \quad (18)$$

Considering the fundamental reactive power and switching current ripples, $I_{Cf \max} = 3.5$ A can be obtained from simulation results, thus $I_{Cf \max} / I_{1 \max} \approx 0.25$. The height h of I-type core should be greater than 3.5 mm by (18) to avoid magnetic saturation.

In order to counteract the coupling inductor in the filter capacitor branch, (10) serving as the second constraint should be met. According to (10) and (6), it can be derived as

$$h = \frac{N_1 N_2 I_{mc}}{\mu_0 \mu_r e [2N_c^2 (R_{g1} + R_{g2}) + 4N_1 N_c R_{g2}]}. \quad (19)$$

Fig. 12 shows the curve of height h varying with winding turns N_c with $\mu_r = 1725$ N/A². To ensure $h > 3.5$ mm, N_c should be less than N_{cm} shown in Fig. 12. As N_c is an integer, $N_c = 1$ is chosen, and h is selected as 5 mm. An I-type core with this dimension is available in the laboratory, and its height is close to 4 mm.

Fig. 12. Curve of h varying with winding turns N_c .

In the above design example, the ferrite core is selected for the proposed magnetic integration method. However, in practical applications, the iron powder core and amorphous core are also commonly used in *LCL* filter inductors. According to the above analysis, the magnetic coupling is actually caused by the non-zero reluctance of the common I-type core. When using the iron powder core or amorphous core, the magnetic coupling

of the windings also exists, and particularly, for the iron powder core, the coupling coefficient even is greater than that with the ferrite core due to its very low permeability. The basic operation principle of the proposed method is to construct a decoupling inductor by introducing an additional winding, so that the coupling inductor in the filter capacitor branch can be counteracted. It is a general method and actually can be applied in many types of magnetic material. Therefore, for the powder iron core and amorphous core, the proposed method still can be effective through proper design.

B. Volume Comparison

According to the above design parameters, the volume of magnetic core with the proposed integrated design can be calculated as

$$V_1 = 2V_U + V_I = 2(ace - bde) + aeh = 1.94 \times 10^5 \text{ mm}^3. \quad (20)$$

where V_U denotes the volume of U-type core and V_I denotes the volume of I-type core.

For the discrete *LCL* filter, two individual filter inductors are required, thus leading to two U-type cores and two I-type cores. The cross-section area of I-type cores should be the same as that of U-type cores to avoid magnetic saturation. Therefore, the height h of each I-type core is 14 mm. Then the total volume of magnetic cores for the two inductors can be obtained as

$$V_2 = 2V_U + 2V_I = 2(ace - bde) + 2aeh = 2.89 \times 10^5 \text{ mm}^3. \quad (21)$$

Furthermore, for volume comparison, the magnetic integration method in [28] is also investigated. In [28], in order to ensure -60 dB/dec attenuation ability at the frequencies of dominate switching harmonics, the intersection frequency f_{int} shown in Fig. 4 should be set above the double switching frequency due to unipolar SPWM [31]. Here, f_{int} is set as 25 kHz to ensure enough margin, then h can be calculated as 32.2 mm by combining (22), (8) and (6), where (22) has been given in [28].

$$f_{int} = \frac{1}{2\pi} \sqrt{\frac{L_1 L_2 + 2M_{12}(L_1 + L_2) + 3M_{12}^2}{(L_1 L_2 - M_{12}^2)M_{12} C_f}}. \quad (22)$$

The core volume can be calculated as

$$V_3 = 2V_U + V_I = 2(ace - bde) + aeh = 3.06 \times 10^5 \text{ mm}^3. \quad (23)$$

Comparing (21) with (23), it can be seen that this integration design method cannot reduce the volume of magnetic cores. Therefore, [28] further proposed to employ an I-type core with higher permeability, e.g., silicon steel with $\mu_r = 5660$ N/A², then h is decreased to 9.8 mm. The core volume can be obtained as

$$V_4 = 2V_U + V_I = 2(ace - bde) + aeh = 2.14 \times 10^5 \text{ mm}^3. \quad (24)$$

From (20), (21), (23) and (24), the integration method in [28] with *MnZn* ferrite core yields a 5.9% ($= (V_3 - V_2) / V_2$) volume increase compared to the discrete inductors while a 26% ($= (V_2 - V_4) / V_2$) volume decrease with silicon steel core. Even though the core volume can be smaller with a higher permeability of I-type core, it simultaneously leads to a higher core cost. For the proposed method in this paper, it can obtain a 32.9% ($= (V_2 - V_1) / V_2$) volume reduction even with *MnZn* ferrite core, indicating a higher power density.

V. EXPERIMENTAL RESULTS

A single-phase grid-connected experimental platform has

been constructed to verify the performance of the proposed active decoupling method for integrated *LCL* filters. The circuit topology is shown in Fig. 1, and the system parameters are listed in Table I.

An integrated inductor and two discrete inductors are fabricated respectively in the laboratory for experimental verification, shown in Fig. 13. As seen from Fig. 13(a), Winding *c* is wound on the common I-type core for active magnetic decoupling.

(a)

(b)

Fig. 13. Photographs of inductors. (a) Integrated inductor. (b) Two discrete inductors.

A. Harmonic Attenuation

Fig. 14 shows the experimental results of the discrete *LCL* filter, where the harmonic spectra of grid-side current i_2 are also presented by using the FFT function of a Lecroy Wavesurfer 3024 digital oscilloscope. As seen, the harmonic current at $2f_s$ is obvious as the unipolar SPWM is adopted, and the harmonic currents at higher frequencies are attenuated effectively.

Fig. 14. Experimental results of the discrete *LCL* filter.

Fig. 15 shows the experimental results of the integrated *LCL* filter without magnetic decoupling, where the integrated inductor shown in Fig. 13 is employed but the decoupling winding wound on the common I-type core is removed. As seen, all switching harmonic currents are increased compared to those in Fig. 14, indicating that the direct magnetic integration weakens the filter performance of *LCL* filter.

Fig. 15. Experimental results of the integrated *LCL* filter without magnetic decoupling.

Fig. 16 shows the experimental results of the integrated *LCL* filter with the proposed active magnetic decoupling method. Since the equivalent coupling inductor in the filter capacitor branch can be counteracted by the decoupling winding on the common I-type core, this integrated *LCL* filter can regain a -60 dB/dec roll-off rate at switching frequencies for better harmonic attenuation. Therefore, the current harmonic at $2f_s$ is decreased significantly compared to that in Fig. 15, and other switching harmonic currents are also attenuated effectively.

Fig. 16. Experimental results of the integrated LCL filter with the proposed active magnetic decoupling.

Fig. 17 shows the comparison of switching harmonic currents at different frequencies for the above three cases. As seen, with the proposed active magnetic decoupling method, the current harmonics above $2f_s$ can be effectively attenuated to the same level as those of the discrete LCL filter. Moreover, the current harmonic at $2f_s$ is noted to be lower than that of the discrete LCL filter, because the equivalent total inductance of this magnetically integrated LCL filter shown in Fig. 8 is $L_1+L_2+2M_{12}$, which is greater than L_1+L_2 of the discrete one. For the integrated LCL filter without magnetic decoupling, all switching current harmonics become significant, especially the one at $2f_s$ which even exceeds 0.3% of the rated fundamental current and fails to comply with the IEEE std. 1547-2003 [4].

Fig. 17. Comparison of switching harmonic currents at different frequencies.

B. Transients and Saturation Current Analysis

In order to avoid current saturation, sufficient margin has been reserved in each design step of integrated inductors, seen from Section IV Part A. For the design example in this paper, the maximum current is set as 14 A, i.e., $I_{1\max}=I_{2\max}=14$ A. To investigate whether the current could be saturated for the designed maximum value, the system transients have been tested, shown in Fig. 18, where a step change of loads is carried out. As seen, the system still can work well during the transients even when converter-side current i_1 reaches the maximum value of 17 A.

Fig. 18. Transient results with the proposed integrated LCL filter.

C. System Efficiency and Inductor Thermal Characteristics

Even though the integrated LCL filter has the advantage of less weight and volume compared to the discrete one, their inductor power losses should also be investigated for a fair comparison. The power loss of inductors normally consists of winding copper loss and core loss. From Fig. 13, it can be seen that the inductors of these two filters have the same winding structure and U-type cores, which indicates the same winding copper loss and U-type core loss due to the same winding currents and flux densities of U-type cores. Note that winding copper loss of Winding c can be neglected due to small turns N_c and current i_c . The only difference of these two filters is the I-type core. Since converter-side output voltage v_{con} and current i_1 contain large amounts of switching harmonics, it is difficult to directly measure or calculate the power losses of inductors. Therefore, the comparison of inductor power losses between the integrated LCL filter and discrete one is conducted by calculating system efficiency under the same conditions. The system efficiency is calculated with P_o/P_{in} , where P_{in} is the input active power calculated with $P_{in}=E_g I_{2\text{rms}}$ under unit power factor operation and P_o is the dc-side output power calculated with $P_o=U_{dc} I_{dc}$. E_g and $I_{2\text{rms}}$ are the root-mean-square (RMS) values of grid voltage e_g and grid-side current i_2 respectively, and U_{dc} and I_{dc} are the average values of u_{dc} and i_{dc} respectively. These four variables can be directly measured by the multimeter. Fig. 19 shows the curves of system efficiency under different output power P_o . As seen, system efficiency of the integrated LCL filter is always slightly greater than that of the discrete one, which thus indicates that the integrated LCL filter cannot introduce extra power loss. The slight improvement of system efficiency is because the only difference of these two filters is the I-type core, and the I-type core volume of the integrated LCL filter is much less than that of the discrete one, which leads to less I-type core loss.

Fig. 19. Curves of system efficiency under different output power P_o .

Fig. 20 further shows the thermal images of inductors under the rated output power, which are captured by a Fluke thermal imager Ti55. The two filters are tested with the same running time. As seen, both filter inductors have almost the same thermal distribution, because they have the same winding structure and U-type cores seen from Fig. 13. Through experimental measurement, the highest temperature of integrated inductor is 40.9 °C while that of discrete inductors is 41.6 °C. Note that the converter-side inductor winding has higher temperature compared to the grid-side inductor winding due to larger amounts of switching harmonics in converter-side current i_1 .

Fig. 20. Thermal images of inductors. (a) Integrated inductor. (b) Two discrete Inductors.

VI. DISCUSSION

Normally, the grid voltage varies in specified range ($\pm 10\%$) and therefore the current drawn by the paralleled filter capacitor branch also varies. Actually, no matter how the current varies, the relationship of $i_1 = i_c + i_2$ is always valid, and then (7) is

always valid. Consequently, the equivalent circuit in Fig. 8 derived from (7) is unchanged regardless of grid voltage variation, thus the switching harmonic attenuation performance of the filter could not be affected. However, it should be noted that the increased capacitor current due to grid voltage swells may cause the I-type core saturation. Therefore, the grid voltage variation should also be considered in the design process of the filter to avoid the core saturation.

VII. CONCLUSION

This paper proposes an active magnetic decoupling method for integrated LCL filters. With a commonly used UIU-type core structure, a magnetic decoupling winding in series with the filter capacitor is wound on the common I-type core. According to a detailed design guidance presented in this paper, the equivalent coupling inductor in the filter capacitor branch can be counteracted or greatly reduced. Meanwhile, comparison of core volume shows that the proposed method can obtain a 32.9% volume reduction, which is greater than 26% of the existing passive decoupling method with a higher permeability silicon steel core. Finally, an integrated LCL filter with the proposed decoupling design is fabricated in the laboratory, and experimental results verify that the proposed method can regain the high harmonic attenuation ability of LCL filter.

REFERENCES

- [1] F. Blaabjerg, R. Teodorescu, M. Liserre, and A. V. Timbus, "Overview of control and grid synchronization for distributed power generation systems," *IEEE Trans. Ind. Electron.*, vol. 53, no. 5, pp. 1398-1409, Oct. 2006.
- [2] R. Teodorescu, M. Liserre, and R. Rodríguez, *Grid converters for photovoltaic and wind power systems*. Hoboken, NJ: Wiley, 2011.
- [3] *IEEE Recommended Practices and Requirements for Harmonic Control in Electric Power Systems*, IEEE Standard 519-1992, 1992.
- [4] *IEEE Standard for Interconnection Distributed Resources with Electric Power System*, IEEE Std. 1547-2003, July 28, 2003.
- [5] *Electromagnetic Compatibility (EMC)-Part 3: Limits-Section 6: Assessment of Emission Limits for Distorting Loads in MV and HV Power Systems*, IEC 1000-3-6, 1996.
- [6] M. Liserre, F. Blaabjerg, and S. Hansen, "Design and control of an LCL-filter-based three-phase active rectifier," *IEEE Trans. Ind. Appl.*, vol. 41, no. 5, pp. 1281-1291, Sep./Oct. 2005.
- [7] A. Reznik, M. G. Simões, A. Al-Durra, and S. M. Mueen, "LCL filter design and performance analysis for grid-interconnected systems," *IEEE Trans. Ind. Appl.*, vol. 50, no. 2, pp. 1225-1232, Mar./Apr. 2014.
- [8] W. Wu, Y. He, and F. Blaabjerg, "An LLCL power filter for single-phase grid-tied inverter," *IEEE Trans. Power Electron.*, vol. 27, no. 2, pp. 782-789, Feb. 2012.
- [9] J. Xu, J. Yang, J. Ye, Z. Zhang, and A. Shen, "An LTCL filter for three-phase grid-connected converters," *IEEE Trans. Power Electron.*, vol. 29, no. 8, pp. 4322-4338, Aug. 2014.
- [10] W. Wu, Y. Sun, Z. Lin, T. Tang, F. Blaabjerg, and H. S.-H. Chung, "A new LCL-filter with in-series parallel resonant circuit for single-phase grid-tied inverter," *IEEE Trans. Ind. Electron.*, vol. 61, no. 9, pp. 4640-4644, Sep. 2014.
- [11] M. Sanatkar-Chayjani, and M. Monfared, "Stability analysis and robust design of LCL with multituned traps filter for grid-connected converters," *IEEE Trans. Ind. Electron.*, vol. 63, no. 11, pp. 6823-6834, Nov. 2016.
- [12] Y. Tang, W. Yao, P. C. Loh, and F. Blaabjerg, "Design of LCL-filters with LCL resonance frequencies beyond the Nyquist frequency for grid-connected inverters," *IEEE J. Emerg. Sel. Top. Power Electron.*, vol. 4, no. 1, pp. 3-14, Mar. 2016.
- [13] S. Jayalath, and M. Hanif, "Generalized LCL filter design algorithm for grid-connected voltage source inverter," *IEEE Trans. Ind. Electron.*, vol. 64, no. 3, pp. 1905-1915, Mar. 2017.

- [14] P. Zumel, O. Garcia, J. A. Cobos, and J. Uceda, "Magnetic integration for interleaved converters," in *Proc. IEEE APEC*, 2003, pp. 1143-1149.
- [15] Y. Yang, D. Yan, F. C. Lee, "A new coupled inductors design in 2-phase interleaving VRM," *IEEE 6th Int. Power Electron. and Motion Control Conf.*, pp. 344-350, May 2009.
- [16] Y. Yang, J. Ma, C. N.-M. Ho, and Y. Zou, "A new coupled-inductor structure for interleaving bidirectional DC-DC converters," *IEEE J. Emerg. Sel. Top. Power Electron.*, vol. 3, no. 3, pp. 841-849, Sep. 2015.
- [17] P.-L. Wong, P. Xu, B. Yang, and F. C. Lee, "Performance improvements of interleaving VRMs with coupling inductors," *IEEE Trans. Power Electron.*, vol. 16, no. 4, pp. 499-507, Jul. 2001.
- [18] F. Yang, X. Ruan, Y. Yang, and Z. Ye, "Interleaved critical current mode boost PFC converter with coupled inductor," *IEEE Trans. Power Electron.*, vol. 26, no. 9, pp. 2404-2413, Sep. 2011.
- [19] Z. Dang, and J. A. A. Qahouq, "Permanent-magnet coupled power inductor for multiphase DC-DC power converters," *IEEE Trans. Ind. Electron.*, vol. 64, no. 3, pp. 1971-1981, Mar. 2017.
- [20] H. Kosai, S. Mcneal, B. Jordan, J. Scofield, B. Ray, and Z. Turgut, "Coupled inductor characterization for a high performance interleaved boost converter," *IEEE Trans. Magn.*, vol. 45, no. 10, pp. 4812-4815, Oct. 2009.
- [21] D. Shin, J.-P. Lee, D.-W. Yoo, and H.-J. Kim, "Stability improvement of interleaved voltage source inverters employing coupled inductors for grid-connected applications," *IEEE Trans. Ind. Electron.*, vol. 62, no. 10, pp. 6014-6023, Oct. 2015.
- [22] G. Gohil, L. Bede, R. Teodorescu, T. Kerekes, and F. Blaabjerg, "An integrated inductor for parallel interleaved VSCs and PWM schemes for flux minimization," *IEEE Trans. Ind. Electron.*, vol. 62, no. 12, pp. 7534-7546, Dec. 2015.
- [23] G. Gohil, L. Bede, R. Teodorescu, T. Kerekes, and F. Blaabjerg, "Optimized integrated harmonic filter inductor for dual-converter fed open-end transformer topology," *IEEE Trans. Power Electron.*, vol. 32, no. 3, pp. 1818-1831, Mar. 2017.
- [24] J. Ewanchuk, and J. Salmon, "Three-limb coupled inductor operation for paralleled multi-level three-phase voltage sourced inverters," *IEEE Trans. Ind. Electron.*, vol. 60, no. 5, pp. 1979-1988, May 2013.
- [25] G. Gohil, L. Bede, R. Teodorescu, T. Kerekes, and F. Blaabjerg, "Magnetic integration for parallel interleaved VSCs connected in a whiffletree configuration," *IEEE Trans. Power Electron.*, vol. 31, no. 11, pp. 7797-7808, Nov. 2016.
- [26] Y. Liu, K. Y. See, K. J. Tseng, R. Simanjorang, and J.-S. Lai, "Magnetic integration of three-phase LCL filter with delta-yoke composite core," *IEEE Trans. Power Electron.*, 2016, Publish online.
- [27] K.-J. Lee, N.-J. Park, R.-Y. Kim, D.-H. Ha, and D.-S. Hyun, "Design of an LCL filter employing a symmetric geometry and its control in grid-connected inverter applications," in *Proc. IEEE PESC*, 2008, pp. 963-966.
- [28] D. Pan, X. Ruan, C. Bao, W. Li, and X. Wang, "Magnetic integration of the LCL filter in grid-connected inverters," *IEEE Trans. Power Electron.*, vol. 29, no. 4, pp. 1573-1578, Apr. 2014.
- [29] J. Fang, H. Li, and Y. Tang, "A magnetic integrated LLCL filter for grid-connected voltage-source converters," *IEEE Trans. Power Electron.*, vol. 32, no. 3, pp. 1725-1730, Mar. 2017.
- [30] R. W. Erickson, and D. Maksimovic, *Fundamentals of Power Electronics*, Springer, 2001.
- [31] D. G. Holmes, and T. A. Lipo, *Pulse Width Modulation for Power Converters: Principles and Practice*, New York: Wiley, 2003.

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