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Magnetic Integration of *LTL* Filter with Two *LC*-Traps for Grid-Connected Power Converters

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Abstract—High-order inductor-traps-inductor (*LTL*) filters recently attracted a lot of research attentions in grid-connected converter applications due to their excellent harmonic attenuation ability, low cost, volume and weight. However, more magnetic cores for trap-inductors are required compared to conventional *LCL* filters. In order to further reduce the volume and weight of output power filter, this paper proposes a magnetic integrated *LTL* filter with two *LC*-traps. On the basis of conventional integrated *LCL* filters, the proposed method introduces a coupling winding wound on the central-limb of an EE-type magnetic core, and two equivalent trap-inductors are introduced by the magnetic coupling between different windings. A step-by-step design guideline of the proposed magnetic integration method and the parameters design of *LTL* filters are also presented. The feasibility and effectiveness of the proposed method are finally verified by experimental results.

Index Terms—Grid-connected converter, *LC*-trap inductor, *LTL* filter, magnetic coupling, magnetic integration, switching harmonic.

I. INTRODUCTION

The increasing energy demands give rise to the rapid developments of power electronics based renewable generation and smart grid technologies. Grid-connected voltage-source converters (VSCs), as common AC/DC or DC/AC energy conversion units, have been widely used in modern power systems for easy integration of renewable sources and energy storages [1]-[5]. However, the use of pulse width modulation (PWM) in VSCs introduces large amounts of switching harmonics. These harmonics may violate the regulations of the IEEE standard 1547-2003 [6], in which the grid-current harmonics higher than 35th should be stringently limited below 0.3% of the fundamental component. Therefore, the converters are commonly interfaced to the unity grid through low-pass power filters, so that the high-frequency PWM harmonics can be attenuated to a satisfied level.

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Traditionally, a single *L* filter can be used to attenuate high-frequency PWM harmonics [7]. Nonetheless, such a simple filter requires large inductance to obtain the desirable harmonic performance, which definitely results in oversize, overweight and high cost of power converters. The increased inductance would also limit the dynamics of grid current. To address this concern, the *LCL* filter, a third-order low-pass filter, provides an alternative topology and excellent harmonic attenuation with much reduced inductance [8]-[10]. Due to the ever-increasing price of copper, inductors become progressively expensive. In order to further reduce the grid-side inductance and the filter cost and volume, many state-of-the-art high-order filters are recently developed on the basis of *LCL* filter, such as *LTL* [11]-[15], *LTCL* [16]-[20], *SPRLCL* [21] and *L(LCL)₂* [22], where the notation “*T*” represents one or multiple *LC*-traps. All of these modified filters have at least one *LC*-trap to bypass or block selected harmonics. Generally, the switching harmonics are centered on the switching frequency and its multiples. If the resonant frequencies of *LC*-traps are tuned at these frequencies, the output harmonics caused by the PWM can be remarkably decreased. In addition, as the resonant frequencies of *LC*-traps are normally equal to or greater than the switching frequency, the inductances of trap-inductors could be very small, thus leading to further reduction of the total inductance. Actually, multilevel converters can also reduce switching harmonic currents greatly with less filter inductance [2]-[4]. However, in low-voltage applications, using high-order output filters is a more preferred option.

Among the trap-based high-order filters, the *LTL* filters have been widely explored. For this kind of power filters, having more *LC*-traps in parallel allows better harmonic attenuation, but it may inevitably result in excessive trap-inductors. Actually, the dominant harmonics are mainly located around the switching frequency and double switching frequency, thus two *LC*-traps are preferred in consideration of cost, volume and weight in practical applications. Nevertheless, two additional inductors are still required compared to *LCL* filters.

Most of previous studies focus on the reduction of the total inductance, so that the cost, volume and weight of power filters can be reduced. Practically, magnetic cores account for most of the weight and volume of the overall power filters. Even though the total inductance is greatly reduced, the trap-based high-order filters introduce additional magnetic cores, which may even increase system cost, weight and volume. The common solution to this issue is the magnetic integration

technique whose primary principle is to integrate multiple discrete inductor windings on a common magnetic core [23]-[30], so that magnetic core can be fully utilized. In [31] and [32], an EIE-type magnetic core is adopted for magnetic integration of single- or three-phase *LCL* filters, aiming to reduce the weight and volume of magnetic core and further improve the power density of converters. In [33], a delta-yoke composite core is proposed for magnetic integration of three-phase *LCL* filters, which can achieve almost 10% weight reduction compared to that with an EIE-type core. However, magnetic integration inescapably results in magnetic coupling even with decoupled core structures, such as EIE-type and delta-yoke-type. The magnetic coupling will equivalently introduce a coupling inductor in the filter capacitor branch, which compromises the harmonic attenuation ability of *LCL* filters. In order to suppress the magnetic coupling effects, [32] employed an I-type core of high permeability in the EIE-type magnetic core. In contrast, [34] utilized the coupling inductor generated by the magnetic coupling to form an *LLCL* filter. However, in order to obtain the required trap inductance, a specified air gap needs to be inserted in the common flux path of EE-type core, which lacks design flexibility and makes parameter adjustment difficult after being fabricated. Moreover, only one *LC*-trap is achieved in [34], thus leading to weak harmonic attenuation ability. To tackle aforesaid concerns, this paper proposes a new magnetic integration method for *LTL* filters, where a coupling winding is wound on the central-limb of an EE-type core instead of inserting an air gap, and two equivalent trap-inductors can be formulated by the magnetic coupling between different windings.

Following the introduction, Section II briefly reviews the topology and mathematic model of *LTL* filters. Section III presents the magnetic model of a single inductor for high-order power filters. In Section IV, the magnetic integration of conventional *LCL* filters is first illustrated. On this basis, a magnetic integration method for *LTL* filters with two *LC*-traps is proposed, where two equivalent trap-inductors are introduced by the magnetic coupling between different windings. A step-by-step design guideline of the proposed method and the parameter design of *LTL* filters are presented in Section V. In Section VI, an integrated *LTL* filter was fabricated in the laboratory, and experimental results are also presented to verify the effectiveness of the proposed design. Finally, Section VII concludes this paper.

II. CIRCUIT CONFIGURATION AND MATHEMATICAL MODEL OF *LTL* FILTERS

$$v_{con}(t) = \frac{4U_{dc}}{\pi} \sum_{n=1}^{\infty} \frac{1}{q_1} J_n(q_1 \frac{\pi}{2} M) \sin([q_1 + n] \frac{\pi}{2}) \left| \sin n \frac{\pi}{2} \right| \cos(n\omega_1 t) \\ + \frac{4U_{dc}}{\pi} \sum_{m=1}^{\infty} \sum_{\substack{n=-\infty \\ (n \neq 0)}}^{\infty} \frac{1}{q_2} J_n(q_2 \frac{\pi}{2} M) \sin([q_2 + n] \frac{\pi}{2}) \left| \sin n \frac{\pi}{2} \right| \cos(m\omega_{sw} t + n\omega_1 t) \quad (1)$$

Fig. 1(a) illustrates a single-phase full-bridge *LTL*-filtered grid-connected converter. i_c and i_g are the converter-side and grid-side current respectively. v_{con} is the converter output voltage and e_g is the grid voltage. U_{dc} is the dc-side voltage. Fig. 1(b) shows the circuit configuration of *LTL* filters, which consists of a converter-side inductor L_c , a grid-side inductor L_g , and multiple *LC*-traps (L_{t1} - C_{t1} and L_{t2} - C_{t2} , etc.) tuned at the switching frequency and other dominant switching frequencies. i_{t1} and i_{t2} represent the *LC*-trap currents.

Fig. 1. (a) Single-phase full-bridge *LTL*-filtered grid-connected converter. (b) Circuit configuration of *LTL* filters.

With the unipolar, sine-triangle and symmetrical regular sampled PWM, the output voltage $v_{con}(t)$ of grid-connected converters in the time domain can be derived as (1) at the bottom of the page [35], where M is the modulation index, ω_{sw} is the switching frequency, ω_1 is the fundamental frequency, m represents the multiple of switching frequency ω_{sw} , n represents the multiple of fundamental frequency ω_1 , $q_1 = n\omega_1/\omega_{sw}$, $q_2 = m + n\omega_1/\omega_{sw}$, and $J_n(x)$ is referred as the integrals of the Bessel function, expressed as

$$J_n(x) = \frac{1}{2\pi} \int_0^{2\pi} \cos(nt - x \sin t) dt \quad (2)$$

Then the harmonic magnitude of converter output voltage

v_{con} around the switching frequency or higher can be obtained as

$$V_{con}(m, n) = \left| \frac{4U_{dc}}{\pi} \frac{1}{q_2} J_n(q_2 \frac{\pi}{2} M) \sin([q_2 + n] \frac{\pi}{2}) \sin n \frac{\pi}{2} \right|. \quad (3)$$

According to (3), Fig. 2 shows the switching harmonic spectra of converter output voltage v_{con} under the condition of $U_{dc}=200V$, $M=0.7$, $f_{sw}=10$ kHz and $f_1=50$ kHz. As can be seen, all the switching harmonics are located at the even times switching frequencies, because the equivalent switching frequency is $2f_s$ due to the unipolar sinusoidal pulse width modulation (SPWM). Moreover, it should be noted that the harmonics around the double and fourfold switching frequencies are significant compared to those of higher switching frequencies. Considering the cost and volume, two LC-traps are preferred in practical applications and also explored in this paper.

Fig. 2. Switching harmonic spectra of converter output voltage v_{con} .

From Fig. 1(b), the transfer function from output voltage v_{con} to grid-side current i_g can be derived as

$$G_{v1}(s) = \frac{(L_{l1}C_{l1}s^2 + 1)(L_{l2}C_{l2}s^2 + 1)}{s(a_4s^4 + a_2s^2 + a_0)}, \quad (4)$$

where,

$$\begin{cases} a_4 = C_{l1}C_{l2}[L_cL_gL_{l1} + L_cL_gL_{l2} + L_cL_{l1}L_{l2} + L_gL_{l1}L_{l2}] \\ a_2 = L_cL_gC_{l2} + L_cL_gC_{l1} + L_cL_{l1}C_{l1} + L_gL_{l1}C_{l1} \\ \quad + L_cL_{l2}C_{l2} + L_gL_{l2}C_{l2} \\ a_0 = L_c + L_g \end{cases} \quad (5)$$

From (4), it can be found that $G_{v1}(s)$ has two pairs of conjugate zeros at $\omega_{l1} = \sqrt{1/(L_{l1}C_{l1})}$ and $\omega_{l2} = \sqrt{1/(L_{l2}C_{l2})}$ respectively, which indicates that there exist two magnitude traps at these two frequencies, as shown in Fig. 3. If these two frequencies are set at the dominant switching frequencies, the corresponding switching harmonics can be attenuated effectively. Moreover, it can also be seen from Fig. 3 that there exist two resonance peaks in the magnitude-frequency characteristic of $G_{v1}(s)$. The resonance peaks may cause system instability. In order to ensure system stability, many damping methods have been developed, such as passive

damping and active damping [8], [14], [21], [36]-[38]. In addition, the single-current-loop control with inevitable digital control delay can also stabilize the system by properly designing the filter resonant frequency [9], [10], [13], [18]. Considering extra power losses of passive damping methods and high sensor costs of active damping methods, the single-current-loop control is preferred to be employed. Moreover, parasitic resistances of the filter can also provide effective damping to promote system stability. As this research work mainly focuses on the magnetic integration of *LTL* filters, the detailed stability analysis is not presented here, which has already been elaborated in existing studies [9], [10], [13], [18].

Fig. 3. Bode diagram of the transfer function $G_{v1}(s)$.

III. MAGNETIC MODEL OF A SINGLE FILTER INDUCTOR

Fig. 4(a) shows the schematic diagram of a single filter inductor, where an EI-type magnetic core is employed [32]. Generally, the key parameters to fabricate a filter inductor consist of material and dimensions of magnetic core, winding turns N and length l_g of air gaps. The inserting of air gaps can avoid magnetic saturation, but decreases the effective magnetic permeability, thus resulting in more winding turns N to achieve a desired inductance. Fig. 4(b) shows the magnetic circuit model, where R_{e1} , R_{e2} , R_{e3} , R_{e4} , R_{e5} , R_{l1} and R_{l2} represent the reluctances of limbs and yokes of magnetic core respectively, and R_g represents the reluctance of air gaps, calculated as

$$R_x = \frac{l_x}{\mu_0 \mu_r A_{sx}}, \quad R_g = \frac{l_g}{\mu_0 A_{sx}}, \quad (6)$$

where l_x is the corresponding length of limbs or yokes, l_g is the length of air gaps, A_{sx} is the cross-section area of central-limb or side-limbs, μ_0 is the air permeability and equal to $4\pi \times 10^{-7}$ N/A², and μ_r is the relative permeability of magnetic core.

Fig. 4. A single filter inductor. (a) Schematic diagram. (b) Magnetic circuit model.

Since air gap reluctances are far greater than core reluctances, the flux Φ through the central-limb can be derived from Fig. 4(b) as

$$\Phi = \frac{Ni}{R_g + R_{e5} + (R_{e1} + R_{e2} + R_{l1} + R_g) \parallel (R_{e3} + R_{e4} + R_{l2} + R_g)} \approx \frac{2Ni}{3R_g}. \quad (7)$$

Further, the voltage across the winding can be obtained as

$$v = -e = N \frac{d\Phi}{dt} = L \frac{di}{dt}, \quad (8)$$

where L is the inductance with $L=2N^2/3R_g$.

IV. PROPOSED MAGNETIC INTEGRATION METHOD FOR *LTL* FILTERS

Section III briefly presents the design principle of a single filter inductor. For conventional *LCL* filters, two such inductors should be fabricated for grid-side and converter-side inductors respectively. While for *LTL* filters, as illustrated by Fig. 1(b), if two *LC*-traps are employed, two additional inductors need to be fabricated besides grid-side and converter-side inductors. Therefore, even with the reduction of the total inductance compared to conventional *LCL* filters, *LTL* filters still suffer from excessive weight, volume and cost due to two magnetic cores required for trap-inductors. In order to further improve the power density and save the cost, magnetic integration is an effective solution, which has already been applied in *LCL* filters.

In order to facilitate further comparison and analysis, Fig. 5(a) shows the schematic diagram of magnetic integration for conventional *LCL* filters [34], where two E-type magnetic cores are employed to form a closed-magnetic circuit, and two air gaps are inserted in the side-limbs to avoid magnetic

saturation. As seen, two windings for converter-side and grid-side inductors denoted by *Winding c* and *Winding g* respectively, are wound on two side-limbs of the resultant magnetic core. Since these two windings are inversely coupled, the fluxes in the central-limb generated by the respective winding are in the same direction. In order to avoid magnetic saturation, the cross-section area of the central-limb is doubled compared to that of the side-limbs. Actually, the common E-type cores in industrial applications have been manufactured with such dimensions. Fig. 5(b) shows the electrical connection diagram of the integrated *LCL* filter. As seen, *Winding c* is for converter-side inductor L_c , while *Winding g* is for grid-side inductor L_g .

Fig. 5. Magnetic integration for *LCL* filters. (a) Schematic diagram. (b) Electrical connection diagram.

Based on the magnetic integration of *LCL* filters, Fig. 6(a) illustrates the schematic diagram of the proposed magnetic integration for *LTL* filters, and Fig. 6(b) shows the corresponding electrical connection diagram. As seen, there are three windings (*Winding c*, *Winding m* and *Winding g*) wound on three different limbs of magnetic core with their respective winding turns N_c , N_m and N_g . According to the flux directions in each limb of the magnetic core, *Winding c* and *Winding g*, *Winding c* and *Winding m*, *Winding g* and *Winding m* are all inversely coupled. From Fig. 6(b), these windings are connected in series, and two shunt capacitors C_{l1} and C_{l2} are placed at the junctions of two windings respectively. The basic idea of the proposed magnetic integration is to construct the trap-inductors by utilizing the

magnetic coupling between different windings, which are named as active trap-inductors. In the proposed design, *Winding m* introduces two coupling inductors for the two LC-traps and is called the coupling winding.

Fig. 6. Proposed magnetic integration for LTL filters. (a) Schematic diagram. (b) Electrical connection diagram.

Based on Fig. 6(a), Fig. 7 shows the magnetic circuit model of the proposed magnetic integration method, where R_{l1} , R_{l2} , R_{l3} , R_{y1} , R_{y2} , R_{y3} and R_{y4} represent the reluctances of limbs and yokes of magnetic core respectively, R_{g1} and R_{g2} represent the reluctances of air gaps, R_{wc} , R_{wm} and R_{wg} represent the leakage reluctances of three windings, Φ_c , Φ_m and Φ_g represent the fluxes through three limbs respectively. Considering the symmetry of the magnetic core shown in Fig. 6(a), $R_{g1}=R_{g2}=R_g$, $R_{y1}=R_{y2}=R_{y3}=R_{y4}$ and $R_{l1}=R_{l2}$. Reluctances of magnetic core and air gaps can be calculated by (6), and leakage reluctances of three windings can be obtained according to the method presented in [39]. From Fig. 7, it can be derived as

$$\begin{cases} \Phi_c = \frac{N_c i_c}{R_{cc}} - \frac{N_g i_g}{R_{gc}} - \frac{N_m i_m}{R_{mc}} \\ \Phi_g = \frac{N_g i_g}{R_{gg}} - \frac{N_c i_c}{R_{cg}} - \frac{N_m i_m}{R_{mg}} \\ \Phi_m = \frac{N_m i_m}{R_{mm}} - \frac{N_c i_c}{R_{cm}} - \frac{N_g i_g}{R_{gm}} \end{cases} \quad (9)$$

where R_{cc} , R_{mm} and R_{gg} are the self-reluctances, and R_{gc} , R_{cg} , R_{mc} , R_{cm} , R_{g2} and R_{cg} are the mutual-reluctances. Since

reluctances of the air gaps are far greater than those of the magnetic core, it then can be derived as

$$\begin{cases} R_{cc} \approx R_{g1} = R_g \\ R_{gg} \approx R_{g2} = R_g \\ R_{mm} \approx R_{g1} || R_{g2} || R_{wc} || R_{wm} || R_{wg} \\ R_{cm} = R_{mc} \approx R_{g1} = R_g \\ R_{gm} = R_{mg} \approx R_{g2} = R_g \\ R_{cg} = R_{gc} \approx \frac{R_{g1} R_{g2}}{R_{g2} || R_{wc} || R_{wg} || R_{l3}} + R_{g2} \end{cases} \quad (10)$$

Fig. 7. Magnetic circuit model of the proposed magnetic integration method.

As cross-section area of the central-limb is equal to two times that of the side-limbs and permeability of the magnetic core is far greater than that of the air, the core reluctance R_{l3} is thus very small compared to R_g , R_{wc} and R_{wg} . It then can be derived from (10) that R_{cg} or R_{gc} is very large and far greater than R_{cc} , R_{gg} , R_{mg} and R_{mc} . Therefore, the magnetic coupling between *Winding c* and *Winding g* can be neglected, and (9) can be approximated as

$$\begin{cases} \Phi_c \approx \frac{N_c i_c}{R_{cc}} - \frac{N_m i_m}{R_{mc}} \\ \Phi_g \approx \frac{N_g i_g}{R_{gg}} - \frac{N_m i_m}{R_{mg}} \\ \Phi_m = \frac{N_m i_m}{R_{mm}} - \frac{N_c i_c}{R_{cm}} - \frac{N_g i_g}{R_{gm}} \end{cases} \quad (11)$$

With $i_c = i_m + i_{l1}$ and $i_m = i_g + i_{l2}$, the voltages across three windings can be derived from (11) as

$$\begin{cases}
v_c = -e_c = N_c \frac{d\Phi_c}{dt} = \frac{N_c^2}{R_{cc}} \frac{di_c}{dt} - \frac{N_c N_m}{R_{mc}} \frac{di_m}{dt} \\
= L_c \frac{di_c}{dt} + L_{t1} \frac{di_{t1}}{dt} \\
v_g = -e_g = N_g \frac{d\Phi_g}{dt} = \frac{N_g^2}{R_{gg}} \frac{di_g}{dt} - \frac{N_g N_m}{R_{mg}} \frac{di_m}{dt} \\
= L_g \frac{di_g}{dt} - L_{t2} \frac{di_{t2}}{dt} \\
v_m = -e_m = N_m \frac{d\Phi_m}{dt} = \frac{N_m^2}{R_{mm}} \frac{di_m}{dt} - \frac{N_c N_m}{R_{mc}} \frac{di_c}{dt} - \frac{N_g N_m}{R_{mg}} \frac{di_g}{dt} \\
= L_m \frac{di_m}{dt} - L_{t1} \frac{di_{t1}}{dt} + L_{t2} \frac{di_{t2}}{dt}
\end{cases} \quad (12)$$

where,

$$\begin{cases}
L_c = \frac{N_c^2}{R_{cc}} - \frac{N_c N_m}{R_{mc}} = \frac{N_c (N_c - N_m)}{R_g} \\
L_g = \frac{N_g^2}{R_{gg}} - \frac{N_g N_m}{R_{mg}} = \frac{N_g (N_g - N_m)}{R_g} \\
L_{t1} = \frac{N_c N_m}{R_{mc}} = \frac{N_c N_m}{R_g} \\
L_{t2} = \frac{N_g N_m}{R_{mg}} = \frac{N_g N_m}{R_g} \\
L_m = \frac{N_m^2}{R_{mm}} - \frac{N_c N_m}{R_{mc}} - \frac{N_g N_m}{R_{mg}} = \frac{N_m^2}{R_{mm}} - \frac{N_m (N_c + N_g)}{R_g}
\end{cases} \quad (13)$$

Fig. 8. Equivalent circuit of the proposed magnetic integration method.

In (13), L_c and L_g are the converter-side and grid-side inductances respectively. L_{t1} and L_{t2} are the trap-inductances introduced by the magnetic coupling of *Winding c* and *Winding m*, and *Winding g* and *Winding m* respectively. Based on (12) and (13), the equivalent circuit of the proposed magnetic integration method is shown in Fig. 8. It can be seen that the proposed method can achieve two equivalent trap-inductors by utilizing the magnetic coupling between different windings. As a result, two magnetic cores can be saved, and the cost, volume and weight of *LTL* filters can be further reduced. In addition, it can be noted that there is an additional inductor L_m between the two *LC*-traps in

comparison to Fig. 1(b), and the similar filter topology can also be found in [22], named *L(LCL)*₂.

From Fig. 8, the transfer function from output voltage v_{con} to grid-side current i_g can be derived as

$$G_{vi2}(s) = \frac{(L_{t1}C_{t1}s^2 + 1)(L_{t2}C_{t2}s^2 + 1)}{s(b_4s^4 + b_2s^2 + b_0)}, \quad (14)$$

where,

$$\begin{cases}
b_4 = C_{t1}C_{t2}[L_cL_gL_{t1} + L_cL_mL_g + L_cL_mL_{t2} \\
+ L_cL_gL_{t2} + L_mL_gL_{t1} + (L_c + L_m + L_g)L_{t1}L_{t2}] \\
b_2 = L_cL_gC_{t2} + L_cL_mC_{t1} + L_cL_gC_{t1} + L_mL_gC_{t2} \\
+ (L_c + L_m + L_g)L_{t1}C_{t1} + (L_c + L_m + L_g)L_{t2}C_{t2} \\
b_0 = L_c + L_m + L_g
\end{cases} \quad (15)$$

Comparing (14) with (4), it is easy to find that transfer function $G_{vi2}(s)$ has the same two pairs of conjugate zeros with $G_{vi1}(s)$, which indicates that the design of trap frequencies is the same as the *LTL* filters with discrete inductors.

Actually, utilizing the coupling inductor to construct the *LC*-trap is first proposed in [34]. However, in order to obtain the required trap inductance, a specified air gap needs to be inserted in the central-limb of *EE*-type core, which lacks design flexibility and makes parameter adjustment difficult after being fabricated. Moreover, only one *LC*-trap is achieved in [34], thus resulting in weak harmonic attenuation ability. While in the proposed design, a coupling winding is wound on the central-limb of *EE*-type core instead of inserting an air gap, which thereby is more flexible. Meanwhile, two trap-inductors can be formulated by the magnetic coupling between different windings, leading to two equivalent *LC*-traps and therefore better harmonic attenuation ability.

V. DESIGN GUIDELINE OF PROPOSED MAGNETIC INTEGRATED *LTL* FILTER

A. Design Procedure

Based on the constraints derived above, a step-by-step design procedure of the proposed magnetic integration method and the parameter design of *LTL* filters will be presented as follows.

1) Converter-side Inductance L_c

Converter-side inductance L_c should be designed to satisfy the maximum ripple current requirement. For unipolar modulated single-phase full bridge converters, the design constraint of L_c is given as [9], [17], [21]

$$\frac{\Delta I_{\max}}{I_{\text{ref}}} = \frac{U_{dc}}{8f_{sw}L_cI_{\text{ref}}} \leq 60\% \quad (16)$$

where $\Delta I_{\max}$ represents the maximum peak-to-peak ripple current of converter-side inductor, and I_{ref} represents the peak value of the rated fundamental current.

2) Inductances L_t and Capacitances C_t of *LC*-traps

For the unipolar SPWM, the equivalent switching

frequency is $2f_{sw}$. Therefore, resonant frequencies of the two LC -trap filter branches should be tuned at the double and fourfold switching frequencies respectively, which can be derived as

$$\frac{1}{2\pi} \sqrt{\frac{1}{L_{t1}C_{t1}}} = 2f_{sw}, \quad \frac{1}{2\pi} \sqrt{\frac{1}{L_{t2}C_{t2}}} = 4f_{sw}. \quad (17)$$

According to (17), the inductances (L_{t1} and L_{t2}) and capacitances (C_{t1} and C_{t2}) of LC -traps can be obtained. However, in order to satisfy the power factor requirement, the total capacitance C_t is limited by the absorbed reactive power at rated conditions, i.e.,

$$C_t = C_{t1} + C_{t2} < \frac{5\%P_{rated}}{E_g^2\omega_1}, \quad (18)$$

where P_{rated} is the rated output power, ω_1 is the fundamental frequency, and E_g is the root-mean-square (RMS) value of the fundamental grid voltage.

3) Grid-side Inductance L_g

According to (13), grid-side inductance L_g can be derived as

$$L_g = \frac{L_{t2}^2(L_c + L_{t1})}{L_{t1}^2} - L_{t2}. \quad (19)$$

Here, two constraints shown below need to be checked.

① The total inductance L_{total} ($=L_c+L_m+L_g$) should be less than 0.1 pu to limit the ac voltage drop on the inductors. Considering that N_m is very small compared to N_c and N_g , L_m can be approximated as $-(L_{t1}+L_{t2})$ from (13). Thus, the total inductance L_{total} can be approximately calculated as

$$L_{total} \approx L_c + L_g - L_{t1} - L_{t2}. \quad (20)$$

② For LTL filters with two LC -traps, the uppermost harmonics will appear around the sixfold switching frequency. The harmonic currents around this frequency should be attenuated less than 0.3% [6].

If any of the above constraints is not satisfied, the design procedure returns to step 1) or step 2) to redesign converter-side inductance L_c or inductances L_t and capacitances C_t of LC -traps.

4) Material and Dimensions of Magnetic Core

With the inductances designed above, the following steps will present the design guideline of the proposed magnetic integration method. The prior task is the selection of EE-type magnetic core, including material and dimensions.

Generally, the selection of magnetic material depends on the cost, relative permeability μ_r , saturation flux density B_{sat} and resistivity, etc. High saturation flux density or relative permeability can reduce the turns of inductor windings and thus the weight and volume of inductor, while high resistivity can reduce the eddy-current losses. To obtain a cost-effective inductor, the magnetic material should be selected to make a trade-off among the cost, weight, volume and efficiency.

The dimensions of magnetic core can be selected according to the well-known area-product method [34], [40], [41]. Following this method, a geometrical constant K_g is used to evaluate the core dimensions, given as

$$K_g = A_s W_A > K_m = \frac{LI_{max}A_l}{K_u B_{max}}, \quad (21)$$

where A_s is the cross-section area of side-limbs, W_A is the window area of magnetic core, K_u is the utilization coefficient of W_A , A_l is the bare area of wires for inductor windings, B_{max} is the maximum flux density of magnetic core and set as λB_{sat} with $0 < \lambda < 1$, and I_{max} is the maximum current through windings.

5) Winding Turns N_c and Length l_g of Air Gaps

Winding turns N_c of *Winding c* for converter-side inductor should be designed to avoid the magnetic saturation of the left-side limb shown in Fig. 6(a). According to (10), (11) and (13), it can be derived as

$$\Phi_c \approx \frac{N_c i_c}{R_{cc}} - \frac{N_m i_m}{R_{mc}} = \frac{N_c i_c - N_m i_m}{R_g}, \quad L_c = \frac{N_c(N_c - N_m)}{R_g}. \quad (22)$$

Considering that LC -trap current i_{t1} is very small, i_m can be approximated as i_c , then the design constraint of winding turns N_c is obtained as

$$\begin{aligned} \Phi_{cmax} < \Phi_{max} &\Rightarrow \frac{(N_c - N_m)I_{cmax}}{R_g} < B_{max} A_s \\ &\Rightarrow \frac{N_c(N_c - N_m)I_{cmax}}{R_g} < N_c B_{max} A_s \\ &\Rightarrow N_c > \frac{L_c I_{cmax}}{B_{max} A_s} \end{aligned} \quad (23)$$

where Φ_{cmax} represents the maximum flux through the left-side limb, Φ_{max} represents the maximum allowed flux, I_{cmax} represents the maximum current of converter-side inductor.

The reluctance R_g of air gaps can be derived from (13) as (24), and then the length l_g of air gaps can be calculated based on (6).

$$\begin{cases} L_c = \frac{N_c(N_c - N_m)}{R_g} \\ L_{t1} = \frac{N_c N_m}{R_g} \end{cases} \Rightarrow R_g = \frac{N_c^2}{L_c + L_{t1}}. \quad (24)$$

6) Winding Turns N_m

According to (13) and (24), winding turns N_m of *Winding m* can be calculated with $N_m = N_c L_{t1} / (L_c + L_{t1})$.

7) Winding Turns N_g

Similarly, according to (13) and (24), winding turns N_g of *Winding g* for grid-side inductor can be calculated with $N_g = N_c L_{t2} / L_{t1}$. Since converter-side inductance L_c is normally designed to be greater than grid-side inductance L_g , winding turns N_c is thus greater than N_g from (13). For the proposed design, the right-side limb wound by *winding g* will not be saturated, because the maximum flux Φ_{gmax} through the right-side limb is always less than Φ_{cmax} derived as (25), and hence is also less than Φ_{max} .

$$(11) \xrightarrow{i_c \approx i_m \approx i_g} \begin{cases} \Phi_{c \max} \approx \frac{(N_c - N_m) I_{c \max}}{R_g} \\ \Phi_{g \max} \approx \frac{(N_g - N_m) I_{c \max}}{R_g} \end{cases} \quad (25)$$

$$\xrightarrow{N_c > N_g} \Phi_{c \max} > \Phi_{g \max}$$

To clearly present the step-by-step design procedure of the proposed integrated *LTL* filter, Fig. 9 shows the design flowchart.

Fig. 9. Flowchart of the step-by-step design procedure of the proposed magnetic integrated *LTL* filter.

B. Design Example

Following the above design procedure, a design example will be presented for experimental verification under the condition of $U_{dc}=200V$, $E_g=110V$, $P_{rated}=0.8 kW$, $f_{sw}=f_s=10 kHz$ and $f_1=50 Hz$, where f_s is the sampling frequency.

1) According to (16), converter-side inductance L_c is selected to be $600 \mu H$ with about 40.5% current ripple under the rated conditions.

2) For the paralleled two *LC*-trap filter branches, $L_{r1}=42 \mu H$ and $C_{r1}=1.5 \mu F$ for the double switching frequency, while $L_{r2}=21 \mu H$ and $C_{r2}=0.75 \mu F$ for the fourfold switching frequency. The total capacitance is $2.25 \mu F$, and the reactive

power is limited to 1.1%.

3) According to step 3), grid-side inductance L_g is calculated as $140 \mu H$, then the total inductance L_{total} is approximated as $677 \mu H$ by (20), which is $0.014 pu$ ($< 0.1 pu$). The harmonic attenuation of grid-side current i_g around the sixfold switching frequency can be verified by (26).

$$|V_{con}(6, n)G_{vi2}(j\omega)|_{\omega=6\omega_{sw}+n\omega_1 (n=0, \pm 1, \pm 2, \dots)} \leq 0.3\% I_{ref}, \quad (26)$$

where $V_{con}(m, n)$ and $G_{vi2}(j\omega)$ have been presented in (3) and (14) respectively. By calculation, (26) is valid.

4) Due to the low cost, high permeability and resistivity, N87-type *MnZn* ferrite material from TDK with typical initial permeability $\mu_r=2200 N/A^2$ is chosen in this work. Its saturated flux density B_{sa} is $0.49 T$ at $25^\circ C$. In order to ensure sufficient margin, the maximum flux density B_{max} is designed as $0.35 T$. The utilization coefficient K_u of window area W_A is designed as 0.5 , and Litz wires with $A_f=0.5\pi mm^2$ are used for inductor windings [34], [40]. Considering the switching current ripple and inrush current during startup, the maximum converter-side current I_{cmax} is designed as $14 A$. According to (21), it can be calculated as $K_m=7.41 \times 10^{-8} m^4$. Referring to the product catalog of TDK, two pairs of E 70/33/32 are used for the proposed integrated *LTL* filter, shown in Fig. 10. Since three limbs of E-type magnetic core have the same length, a small piece of rectangular magnetic core is inserted in the central-limb to achieve two same air gaps in the side-limbs. From Fig. 10, the cross-section area A_s of side-limbs and the window area W_A are calculated as $7.17 \times 10^{-4} m^2$ and $5.69 \times 10^{-4} m^2$ respectively, then the geometrical constant K_g is equal to $4.08 \times 10^{-7} m^4$, and thereby (21) is valid for converter-side *Winding c*. Since $I_{cmax} > I_{gmax}$ and $L_c > L_g$, (21) is also valid for grid-side *Winding g*.

Fig. 10. Magnetic core for the proposed integrated *LTL* filter. (a) 3-D diagram. (b) Front view.

5) According to step 5), winding turns N_c of *Winding c* for converter-side inductor should be greater than 33 turns and designed as 42 turns. Then, the reluctance R_g and length l_g of air gaps are calculated as $2.75 \times 10^6 \text{ H}^{-1}$ and 2.5 mm respectively.

6) According to step 6), winding turns N_m of *Winding m* is calculated as 2.7 turns.

7) According to step 7), winding turns N_g of *Winding g* for grid-side inductor is calculated as 21 turns.

Due to the fringing effect of air gaps and non-precise craftsmanship in the fabrication process, the above design parameters are slightly adjusted, and the final version is shown in Table I.

TABLE I
PARAMETERS OF THE INTEGRATED *LTL* FILTER

Parameters	Value	Parameters	Value
L_c	580 μH	L_g	170 μH
L_{l1}	41 μH	N_c	40 turns
C_{l1}	1.5 μF	N_g	21 turns
L_{l2}	20.5 μH	N_m	2 turns
C_{l2}	0.75 μF	l_g	3.5 mm

C. Volume Comparison

In order to show the advantage of the proposed integrated *LTL* filter, volume comparison with the discrete one is performed. From Fig. 10, the core volume of the proposed integrated *LTL* filter can be calculated as

$$V_1 = 4V_E = 4 \times (32 \times 70.5 \times 33.2 - 2 \times 32 \times 21.9 \times 13) = 2.267 \times 10^5 \text{ mm}^3 \quad (27)$$

While for the discrete *LTL* filter, four individual inductors, L_c , L_g , L_{l1} and L_{l2} , need to be fabricated respectively with the method presented in Section III. To make the core volume comparable, the selection of magnetic cores for the converter-side and grid-side inductors should have the same geometrical constant K_g with the integrated inductor shown in Fig. 10. From Fig. 4, the winding of single inductor is wound on the central-limb of EI-type magnetic core, thus A_s in (21) denotes the cross-section area of the central-limb,

which is equal to two times that of the side-limbs. Moreover, by comparing Fig. 4 with Fig. 6, the window area W_A of EI-type magnetic core is one half of that of EE-type magnetic core. According to (21), in order to keep K_g unchanged, two pairs of E 70/33/32 are also chosen for the converter-side and grid-side inductors respectively. The structure of magnetic cores is shown in Fig. 11(a), and the detailed dimensions are listed in Table II. Considering the fundamental reactive power and switching current ripple, $I_{l1\text{max}}$ and $I_{l2\text{max}}$ are designed as 3 A and 2 A respectively. According to the area-product method, one E 36/18/11 is chosen for L_{l1} with $K_g = 9.99 \times 10^{-9} \text{ m}^4$ and $K_m = 1.1 \times 10^{-9} \text{ m}^4$, and one E 25.4/10/7 is chosen for L_{l2} with $K_g = 1.67 \times 10^{-9} \text{ m}^4$ and $K_m = 3.68 \times 10^{-10} \text{ m}^4$. The structure of magnetic cores is shown in Fig. 11(b), and the dimensions are given in Table II.

Fig. 11. Magnetic cores for the discrete *LTL* filter. (a) Converter-side and grid-side inductors. (b) Trap-inductors.

TABLE II
DIMENSIONS OF MAGNETIC CORES FOR THE DISCRETE *LTL* FILTER

Parameters	Value			
	L_c	L_g	L_{l1}	L_{l2}
A/mm	32	32	11.5	6.5
B/mm	70.5	70.5	36	25.4
C/mm	33.2	33.2	18	9.78
D/mm	21.9	21.9	12	6.48
E/mm	13	13	7.1	6.1
F/mm	22	22	10.2	6.5
G/mm	11.2	11.2	6	3.3

With the dimensions listed in Table II, the total volume of magnetic cores can be obtained as

$$V_2 = V_{Lc} + V_{Lg} + V_{Ll1} + V_{Ll2} = 3.374 \times 10^5 \text{ mm}^3. \quad (28)$$

Comparing (27) with (28), the proposed magnetic integration method can obtain a 32.8% volume reduction.

VI. EXPERIMENTAL RESULTS

In this section, the proposed magnetic integrated *LTL* filter is experimentally verified by comparing with the discrete *LTL* filter and integrated *LCL* filter. According to the design parameters listed in Table I, a magnetic integrated *LTL* filter was fabricated in the laboratory. Fig. 12 shows the integrated inductor for the *LTL* filter. As seen, three windings (*Winding c*, *Winding m* and *Winding g*) are wound on three limbs of the EE-type magnetic core respectively. Through experimental measurement, the equivalent series resistances of these three windings are 0.1074 Ω , 0.0113 Ω and 0.0536 Ω respectively, while the equivalent series resistances of two paralleled capacitors C_{l1} and C_{l2} are 0.0588 Ω and 0.0628 Ω respectively.

Fig. 12. Photograph of a magnetic integrated inductor for the *LTL* filter.

A single-phase grid-connected experimental prototype has also been constructed, shown in Fig. 13. In the experiments, a Chroma 61502 programmable ac power supply and a dc resistive load were connected to the ac-side and dc-side of the power converter respectively. The overall control algorithm was executed on a DSP+FPGA digital microcontroller, where a TMS320F28335 DSP from Texas Instruments (TI) was used as the main algorithm controller to implement the dc-side voltage control and ac-side current control, while a XC3S400 FPGA from Xilinx was used as the assistant controller to implement PWM signals output and fault protection, etc. A Lecory Wavesurfer 3024 digital oscilloscope was used for capturing all the experimental results. The system parameters are given in Section V Part B.

Fig. 13. Photograph of the experimental prototype.

Fig. 14 shows the experimental results of the proposed integrated *LTL* filter. It can be seen from the harmonic spectra of converter-side current i_c , large amounts of switching harmonic currents are visible around even-order switching frequencies as the unipolar SPWM is adopted, and the dominant harmonic currents are located around the double and fourfold switching frequencies, i.e., 20 kHz and 40 kHz. Fig. 14(b) shows the harmonic spectra of grid-side current i_g . As can be observed, the switching harmonic currents have been dramatically reduced, especially those around 20 kHz and 40 kHz where two *LC*-traps are set. Table III presents the detailed content of grid-side switching harmonic currents. As seen, all the harmonic currents are less than 0.3% of its rated current, and thus meet the requirements of the IEEE standard 519-2014.

Fig. 14. Experimental results of the integrated *LTL* filter. (a) Experimental waveforms and harmonic spectra of converter-side current i_c . (b) Harmonic spectra of grid-side current i_g .

TABLE III
MAXIMUM HARMONICS (% OF THE FUNDAMENTAL) OF GRID-SIDE CURRENT

Filter Type	Max harmonics around $2f_{sw}$	Max harmonics around $4f_{sw}$	Max harmonics around $6f_{sw}$
Integrated <i>LTL</i> filter	0.07%	0.05%	0.01%
Discrete <i>LTL</i> filter	0.09%	0.02%	0.01%
Integrated <i>LCL</i> filter	1.55%	0.03%	0.01%

For comparison, the experiment with the discrete *LTL* filter was also performed. Since there exist some tolerances in the process of inductor fabrication, the converter-side harmonic currents of the discrete *LTL* filter are slightly different from those of the integrated one, as shown in Fig. 14(a) and Fig. 15(a). Fig. 15(b) shows the harmonic spectra of grid-side current i_g , and Table III lists the corresponding switching harmonic content. As seen, the switching harmonic currents of these two methods at the *LC*-trap frequencies are almost the same, which verifies the effectiveness of the proposed method. However, compared to the discrete *LTL* filter, the proposed method can save two magnetic cores for trap inductors, thus leading to lower cost, weight and volume.

(a)

Fig. 15. Experimental results of the discrete *LTL* filter. (a) Experimental waveforms and harmonic spectra of converter-side current i_c . (b) Harmonic spectra of grid-side current i_g .

In order to further show the superiority of the proposed method, Fig. 16 shows the experimental results of the integrated *LCL* filter. In this experiment, *Winding m* of the integrated *LTL* filter is removed from the central-limb of magnetic core to form the integrated *LCL* filter, as shown in Fig. 5, and the total capacitance is equal to $C_{l1}+C_{l2}$. Since the integrated *LCL* filter and integrated *LTL* filter have the same magnetic core structure as well as converter-side and grid-side inductor windings, their converter-side harmonic currents are almost the same, as shown in Fig. 14(a) and Fig. 16(a). From Fig. 16(b), even though the grid-side harmonic currents above the fourfold switching frequency are attenuated effectively, the harmonic current around the double switching frequency is remarkable, which is far beyond the 0.3% limit of the rated current, as shown in Table III. In order to meet the requirements of grid standards, a larger grid-side inductance or/and a larger filter capacitance have to be employed. By contrast, to better attenuate grid-side switching harmonic currents, the proposed method only requires a coupling winding with very small number of turns wound on the central-limb of magnetic core.

(a)

Fig. 16. Experimental results of the integrated *LCL* filter. (a) Experimental waveforms and harmonic spectra of converter-side current i_c . (b) Harmonic spectra of grid-side current i_g .

Fig. 17 shows the transient experimental result with the proposed integrated *LTL* filter, where a 30 V step-up change of the dc reference voltage (190V→220V) was carried out. As seen, the proposed method still exhibits excellent switching harmonic attenuation ability during the transient period.

Fig. 17. Transient experimental results with the proposed integrated *LTL* filter.

Finally, the system efficiency with the proposed integrated *LTL* filter was measured under different output power P_o , shown in Fig. 18. As seen, the efficiency is always greater than 96%, indicating low system power losses. Furthermore, the system efficiency with the integrated *LCL* filter was also presented for a fair comparison, because they have the same magnetic core structure as well as grid-side and converter-side inductor windings. As can be observed, the efficiency with the integrated *LTL* filter is always slightly less than that with the integrated *LCL* filter, because an additional coupling winding with very small turns is introduced for the integrated *LTL* filter.

Fig. 18. System efficiency under different output power P_o .

VII. CONCLUSION

In this paper, a magnetic integrated *LTL* filter is proposed for grid-connected power converters to reduce the weight and volume of filter inductor. On the basis of conventional integrated *LCL* filters, an additional coupling winding is wound on the central-limb of an EE-type magnetic core in the proposed method. By utilizing the magnetic coupling of the windings, two coupling inductors can be introduced in the two paralleled filter capacitor branches respectively, thus leading to two equivalent *LC*-traps. With a step-by-step design procedure, these two *LC*-traps can be tuned at specified harmonic frequencies. Therefore, the proposed method exhibits the same harmonic attenuation performance with the discrete *LTL* filter, but can save two magnetic cores for trap-inductors. Moreover, compared to the integrated *LCL* filter, the proposed method has the same magnetic core structure but features better harmonic attenuation performance. Finally, an integrated *LTL* filter was fabricated in the laboratory, and three groups of experimental results are presented to verify the effectiveness of the proposed method.

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