Abstract—Paralleling power converters increases the power rating and reliability of the overall system. Interleaving the carrier in parallel converters helps in reduction of output current distortion, and in reduction of EMI noise. This paper presents a common-mode design method for parallel interleaved converters which integrates the inter-phase and boost inductors together in a novel structure. It is shown that the proposed magnetic structure accommodates a wider range of desirable electromagnetic parameters. Step by step design procedure to optimize the size and the weight of the proposed integrated common-mode inductor (ICMI) is presented. The design reduces the size and the cost of overall system. The proposed ICMI is compared with the design available in literature using a design example. It is shown that the available design procedure imposes constrains on the core geometry that are not suitable for high power designs. The proposed design makes use of standard cores that improves manufacturability. Experimental results from a 7.5kW parallel single phase converter show the effectiveness of the proposed integrated common-mode inductor.

Index Terms—parallel converters, Carrier interleaving, inter-phase inductor, LCL filter, grid-connected converters.

I. INTRODUCTION

Pulse width modulated (PWM) converters are the preferred topology in applications like motor drive, grid connected converters, photovoltaic, wind power generation, etc. [1], [2]. The system power level can be increased by paralleling the smaller sized power converters [5]–[7]. Paralleling increases the power rating of the system it also increases the reliability of the system by achieving higher system redundancy and availability [8]. The grid current distortion can be reduced through carrier interleaving [9]–[12]. Phase shifting the carrier of parallel converters, shifts the harmonic components of the converters and can result in cancellation of certain harmonics. This also provides the capability to reduce the switching frequency at higher power levels for the interleaved converter designs hence, reducing the switching losses. However, carrier interleaving can cause circulation of current between the converters. The circulating current consist of low frequency as well as high frequency components. The low frequency component is due to output voltage mismatch between the converters. This causes unbalanced load sharing in the parallel connected converters.

Circulation of current increases power loss, saturates the inductors, causes instability, and over stress or even damage power devices. Several solutions have been proposed in literature to control and limit the circulating current in parallel converters. In [13] an interleaved active power filter concept with reduced size of passive components is presented. In [14] inter-phase transformers for connecting power converters in parallel are analysed to reduced the circulating current. In [15] a zero-sequence current control loop is designed to suppress the circulating current. However, it only reduces the low frequency circulating current and there is no control over the high frequency circulating current. In [16] a control method to limit the low frequency circulating current is presented, the high frequency circulating current is limited by the help of inter-phase inductors. In [17] a special inductor structure is proposed which integrates the inter-phase inductor with the boost output inductor. However, the steps involving the design procedure are not provided. The proposed inductor is not made of standard cores and needs a special core geometry.
In this paper a new design for an integrated common-mode inductor (ICMI) is presented. The proposed ICMI combines the inter-phase inductor ($L_{int}$) with the $LCL$ boost filter inductor ($L_b$). This helps reducing the component count and the size of overall system. The proposed design uses standard C cores and there is no need for special geometric dimensioning of the core. Steps involving the design of the ICMI are given along with a design example. The proposed ICMI design is compared with the design suggested in [17]. It is shown that for high power converter, where a small boost inductance is necessary, the design in [17] would need a large air-gap that is not practically feasible. The proposed ICMI has been experimentally tested in a 7.5kW, parallel single phase converter. The circuit topology of the converter is shown in Fig. 1. The results validate the effectiveness of the proposed ICMI design.

II. CIRCULATING INDUCTOR FOR PARALLEL CONVERTERS

The operation of transformer-less power converters, with common ac and dc buses, in parallel offer a number of benefits. The advantage of eliminating the transformer is the increase of the power density of the overall system. The drawback of mentioned systems is the circulating current between the parallel converters. The topology shown in Fig. 1, consists of two parallel single-phase H-bridge PWM converters with inter-phase inductor ($L_{int}$) between same phases of the converters. $LCL$ filter is used to filter out the switching frequency component in the output currents. The topology shown in Fig. 1, is considered for the analysis of this paper.

By interleaving the carrier it is possible to reduce or eliminate certain switching harmonics [12] but it causes the circulation of current between the parallel converters. These circulating current must be eliminated or limited. To achieve this, different inter-phase inductor configurations can be employed. The objectives of the inter-phase inductors are, first to offer a high impedance path for the circulating current. This makes sure that the circulating current is limited to a desired value. Second, it should provide a low impedance path for the output current. This is to prevent large voltage drop across the inter-phase inductor for normal power flow.

The output current is filtered by the help of $LCL$ filter. The $LCL$ filter consists of grid side inductor $L_f$, filter capacitor $C_f$, and the boost inductor $L_b$. It is suggested in [17] to integrate the boost inductor with the inter-phase inductor as one magnetic structure. In that configuration the main inductance is used as inter-phase inductor, while the leakage inductance is serving as a boost inductor. The structure of inductor designed in [17] is depicted in Fig. 2(a). Although this design reduces the cost and the weight of the overall system, but yet it is not an optimized design in terms of factors such as core geometry and limits of the inductances values that are feasible. The windings are on the sides limbs, while the center limb which has double cross section does not have any winding. Hence, for the structure shown in Fig. 2(a), lesser inductance can be obtain for a same core and the same number of turns with respect to the case when the windings are placed on the center limb. Second, the core requires different air-gap for the center and side limbs. Such a core structure is not a standard geometry. So, manufacturing this magnetic structure would be difficult, especially for amorphous cores. Third, in high power application with higher dc bus voltage, high main circulating path inductance and small and accurate leakage inductance are needed. This is especially so for $LCL$ filters where the optimum inductance requirement can be small [18]. In the shown configuration in Fig. 2(a), the minimum leakage inductance occurs when the center limb is completely removed. Lesser leakage inductance would not be geometrically feasible. These can be large tolerance in the value of the inductance with large air-gap due to uncertainty in the fringing flux effects. The proposed ICMI is able to overcome these design issues of the CM inductor used for interleaved operation of parallel connected inverters.

III. THE PROPOSED ICMI

Fig. 2(b) and Fig. 2(c) show the structure of the proposed ICMI. Fig. 2(b) shows the flux paths for circulating current using dotted lines, and the output current using dashed line. The winding connections are shown in Fig. 2(c). In the
proposed configuration the windings are split into two identical side windings and a center winding. The ratio of splitting the windings is dictated by the ratio of the boost inductance to inter-phase inductance. The center winding is split into two identical windings, the windings are connected in series with the center tap connection “M” taken out as shown in Fig. 2(c). The windings on center limb are wound in a bifilar manner. Hence, the leakage inductance of the center winding is very small and is neglected.

As said before the leakage inductance of the proposed CM inductor is used as boost inductor. The leakage inductance of the shown topology is equal to the inductance of a side winding plus leakage inductance of the center winding. As the leakage inductance of the center winding can be neglected, the leakage inductance of the proposed CM inductor would be equal to the inductance of a side winding. Hence, for a given core and air-gap, the boost inductor value can be controlled by changing the number of turns in the side windings. The main inductance, which is the inductance between point A1 to A2, is used as inter-phase inductor. The main inductance value is controlled by changing the number of turns in center winding. So, for a given air-gap and core size, the number of turns in side windings to have the desired boost inductance is first chosen. Then, the number of turns in the center winding is chosen to get the required value of inter-phase inductance. The steps for optimum design of ICMI using standard C type amorphous cores are given in the next section.

IV. PROPOSED ICMI DESIGN

The flowchart for designing the CM inductor is depicted in Fig. 3. The steps involving the design procedure are explained here.

**Step 1**- calculate the desired inductance values for boost inductor and inter-phase inductor based on the power converter requirements [18].

**Step 2**- calculate the wire gauge required based on the rating of the individual H-bridge converter.

**Step 3**- choose the smallest core size from the available cores under evaluation. The available cores are chosen based on the area product of the cores. A sufficient number of core sizes with area product higher than the calculated value in (1) are considered for evaluation.

\[
W_a A_c = \frac{L_{eq} I_p^2}{B_{max} J K_w}
\]

(1)

Here, \(W_a\) is window area of the core \((cm^2)\), \(A_c\) is cross section of the core \((cm^2)\), \(L_{eq}\) is the inductance due to half of the total turns which would be \(L_{total} / 2\), \(I_p\) is the peak current \((A)\), \(B_{max}\) is the maximum peak flux density (Tesla), \(J\) is the current density \((A/cm^2)\), and \(K_w\) is the winding factor.

**Step 4**- calculate the minimum air-gap required for the chosen cores such that the variation in the core reluctance will effect the total inductance by only 10%-20%. Set the air-gap required equal to the minimum air-gap. Fig. 4 shows the variation in core reluctance by total reluctance versus air-gap for different core size. This plot can be used to define the minimum air-gap required.

**Step 5**- using the traditional inductor design procedure [19], calculate the number of turns, \(N_1\) and \(N_2\), required for the side windings to obtain the inductance value needed as boost inductance \(L_b\). Using (6), given in appendix A, calculate the total number of turns needed for the center winding to obtain the required value of main inductance \(L_{main}\).

**Step 6**- cross check whether calculated total number of turns \((i.e. N_1 + N_3)\) for the selected core can be held in the window area of the core, considering the winding factor \(K_w\). If the selected core fails to hold the total turns calculated, then choose next larger core size from available cores and perform step 4. \(K_w = 0.4\) is chosen for windings with round conductors.

**Step 7**- calculate the maximum flux in the core based on the peak current of the power converters. If the maximum flux is higher than the saturation flux or any desired flux level \(B_{dc}\), then increase the air-gap by around 10\% and perform step 5. For amorphous cores \(B_{sat}\) of 1.5T is used for the design.

**Step 8**- the outputs are the core size, the air-gap, and the number of turns designed for the CM inductor.

The inductance value is proportional to the cross section of core, it is also proportional to square of the number of turns. With respect to cost, the copper wire is cheaper than Amorphous cores. Cost of copper is about $10 per kilograms where as the amorphous material is about $15 to $20 per
kilograms. To reduce the cost and the size of the inductor it is preferred to use less core material when compared to the copper. Extra care is needed in this case to keep the maximum flux below the saturation flux or below a desired level. Steps given here make sure that the core is chosen properly and the cost and the size is optimized. Additionally, the air-gap length is selected such that the flux in the core is neither small and the core is underutilized nor too large that the core operates near the saturation limit.

V. DESIGN EXAMPLE AND COMPARISON

In this section a CM inductor design example is given for the topology shown in Fig. 1. The total power rating of the considered system is 7.5kW with grid voltage of 240V rms. The current rating of each converter is about 16A rms and the dc bus voltage is 700V with switching frequency of 10kHZ. The design steps 1 to 8 from Section IV are explained below in a design example.

Step 1- The designed values for main and boost inductances are: \( L_{\text{main}} = 20mH \), and \( L_b = 900\mu H \). This makes sure that the circulating current is less than 2A peak or 1.2A rms which is around 4% of the total current rating.

Step 2- The wire gauge chosen is SWG 11. This wire can carry rms current about 27A.

Step 3- Metglas AMCC series have been chosen, based on the area product calculation (1), and listed in TABLE I.

Step 4- To calculate the air-gap, the core reluctance by total reluctance have been calculated for different air-gaps. The results are plotted in Fig. 4. From the figure it can be seen that by choosing the air-gap of 1.5mm for each limb, the effect of variation in the total inductance due to change in the core reluctance would be around 10%.

Step 5- The number of turns for the side windings and center winding is tabulated in TABLE I.

Step 6- The number of turns that can be held in a window of a core is calculated as:

\[
N_p = \frac{W_a K_w}{A_w}
\]

Where, \( N_p \) is the possible number of turns that can be occupied in the window area of core, \( W_a \) is the window area, \( K_w \) is the winding factor, and \( A_w \) is the area of a single wire. The possible number of turn \( N_p \) along with the required number of turns \( N_1 + N_3 \) have been calculated and depicted in Fig.5. From the shown figure it can be seen that the first two cores can not hold the required number of turns hence, they are omitted and can not be used.

Step 7- To calculate the maximum flux \( B_{\text{max}} \) in the core we have to calculate the maximum flux due to circulating current and the maximum flux due to the output current, the summation of these two fluxes should be less than the saturation flux. \( B_{\text{max}} \) can be calculated using:

\[
B_{\text{max}} = \left( \frac{L_{\text{main}} i_{\text{cir}}}{2(N_1 + N_3)} + \frac{L_b i_1}{N_1} \right) \frac{1}{A_{\text{side}}}
\]

Where, \( i_{\text{cir}} \) is the peak high frequency circulating current \( i_2 - i_3 \) shown in Fig. 1, \( A_{\text{side}} \) is the cross sections area of the side limb, and \( i_1 \) is the maximum output peak current of individual power converters. The calculated \( B_{\text{max}} \) is tabulated in TABLE I. It is seen that none of the core flux densities would be higher than the saturation flux density. Hence, none of them needs to be omitted.
Step 8- The smallest core among the remaining cores is AMCC0320 which is selected for making the CM inductor.

To compare proposed CM inductor design given in this paper with the inductor design proposed in [17], an inductor has been designed based on the proposed method in [17]. For the sake of comparison, the side leg air-gap and the core size were kept same for both methods. It was found that the total number of turns needed in their design were 208 turns to obtain the desired value of $L_{main}$. To get the desired inductance value for boost inductor, the center limb air-gap should be adjusted. It is found that even by removing the center limb completely the desired value of $900\mu H$ can not be obtained.

VI. EXPERIMENTAL RESULTS

The proposed CM inductor has been built in the laboratory, the designed parameters are given in TABLE II and the photograph of the inductor is depicted in Fig. 6. The built inductors have been tested with a 7.5kW two parallel single phase grid connected converter with interleaved carrier. Unipolar PWM method with switching frequency of 10kHz is employed as the modulation technique. Converter currents and their harmonic spectra are shown in Fig. 7(a). It can be seen that the fundamental currents supplied by each converter are almost the same and the circulating current between the converters is small. The circulating current is measured and shown in Fig. 7(b) along with its harmonic spectrum. From this, it can be seen that the fundamental circulating current between the converters is almost zero and the high frequency peak circulating current (i.e. $10kHz$ component) is around $1.2A$. Grid voltage and current and their spectra are shown in Fig. 8. It can be seen that the switching frequency ripple component is effectively filtered out in the grid current by the $LCL$ filter using the proposed ICMI as the boost side inductor.

<table>
<thead>
<tr>
<th>SL.NO.</th>
<th>PARAMETER</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Main Inductance $L_{main}$</td>
<td>20mH</td>
</tr>
<tr>
<td>2</td>
<td>Leakage Inductance $L_b$</td>
<td>$900\mu H$</td>
</tr>
<tr>
<td>3</td>
<td>$N_1$, $N_2$</td>
<td>40turns</td>
</tr>
<tr>
<td>4</td>
<td>$N_3$</td>
<td>100turns</td>
</tr>
<tr>
<td>5</td>
<td>air-gap $g$</td>
<td>1.5mm</td>
</tr>
<tr>
<td>6</td>
<td>Conductor Gauge</td>
<td>SWG 11</td>
</tr>
<tr>
<td>7</td>
<td>Core Size</td>
<td>AMCC-320</td>
</tr>
</tbody>
</table>

VII. CONCLUSION

An ICMI magnetic structure and its design is proposed, which integrate the boost inductor and the inter-phase inductor for use in parallel interleaved converters. It makes use of standard core geometries that improves manufacturability. A step by step design procedure to optimize the size and the weight of the CM inductor is presented. The proposed design reduces the size and the cost of overall system. Experimental
results on a 7.5 kW parallel single phase converters show that the fundamental circulating current is about zero and the high frequency peak circulating current between the converters is less than 1.2 A. Although a parallel single phase power converter is presented here, the proposed method can be extended for parallel three phase power converter.

![Diagram](image_url)

Fig. 9. Proposed ICMI with individual windings voltage drops and fluxes due to circulating current.

VIII. APPENDIX

A. Main Inductance Calculation

The proposed ICMI with individual windings voltage drops and fluxes due to circulating current is shown in Fig. 9. The main inductance, inductance between A1 to A2, limits the circulating current. It is calculated as given below.

\[ V_{\text{main}} = V_1 + V_3 + V_2 \]  

(2)

Each voltage drop can be written in terms of number of turns in the winding and the flux in the corresponding limb as,

\[ V_{\text{main}} = N_1 \frac{d\varphi_1}{dt} + N_3 \frac{d\varphi_3}{dt} + N_2 \frac{d\varphi_2}{dt} \]  

(3)

The number of turns in windings 1 and 2 are equal, \( N_1 = N_2 \). The circulating current is same in these windings. Hence, from the symmetrical structure of the core and windings it can be concluded that,

\[ \varphi_1 = \varphi_2 = \frac{\varphi_3}{2} = \frac{\varphi}{2} \]  

(4)

The (3) is simplified using (4) as,

\[ V_{\text{main}} = 2N_1 \frac{d\varphi_1}{dt} + N_3 \frac{d\varphi_3}{dt} = (N_1 + N_3) \frac{d\varphi}{dt} \]  

(5)

Equation (5) shows that the structure shown in Fig. 9 is equivalent to an inductor with a same core structure but only one center winding with number of turns equal to \( N = N_1 + N_3 \). So, the total number of turns can be obtained as given in (6).

\[ N = \sqrt{L_\text{int}} R \]  

(6)

Where, \( R = R_\text{core} + R_\text{gap} \) is the reluctance of the structure shown in Fig. 9, \( R_\text{core} \) is the reluctance of the core and the \( R_\text{gap} \) is the reluctance of the total air gap.

B. Boost Inductance Calculation

As mentioned before the center winding does not have a significant effects on the boost inductance. The number of turn for the side windings can be design using (7).

\[ N_1 = \sqrt{L_0 \cdot 2R} \]  

(7)

IX. ACKNOWLEDGEMENT

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REFERENCES


