Novel Integrated Common-Mode Inductor Design for Parallel Interleaved Converters

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The basics of this paper [1] has been presented at IEEE International Conference on Power Electronics, Drives and Energy System (IEEE PEDES 2014) on 16-19 December 2014, IIT Bombay, Mumbai, India. However, this paper is upgrade by more than 30% to 40% with respect to the conference paper.

**Abstract**

Paralleling power converters can increase 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 inductor design 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 of the proposed integrated common-mode inductor (ICMI) is presented. The proposed ICMI design results in fewer components, reduces the size and the cost of overall system. The proposed ICMI is compared with a design available in literature using an example.
Existing design procedure imposes constraints on the core geometry that are not suitable for high power designs. The proposed design makes use of standard C cores that improves manufacturability. Experimental results from a 7.5kVA parallel single-phase power converter show the effectiveness of the proposed ICMI.

Index Terms

Parallel converters, carrier interleaving, inter-phase inductor, LCL filter, grid-connected converters, finite element methods.

I. INTRODUCTION

Pulse width modulated (PWM) voltage source converters are the preferred topology in applications like motor drives, grid connected converters, photovoltaic and, wind power generation, [2]–[5]. The system power level can be increased by paralleling smaller rated power converters [6]–[8]. Paralleling increases the power rating of the system. It can also increase the reliability of the system by achieving higher system redundancy and availability [9], [10]. The grid current distortion and current ripple can be reduced through carrier interleaving [11]–[15]. Phase shifting the carrier of parallel converters, shifts the harmonic components of the converters and can result in cancellation of harmonics. This also provides the ability to reduce the switching frequency at higher power levels for the interleaved converters, thus, 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 close to the fundamental frequency and the high frequency component is around the switching frequency. The high frequency circulating current is due to turning on the top switch of one converter and bottom switch of the second converter, which belong to the same phase group, at the same time. This circulating current is limited by the help of inter-phase transformer or coupled inductor [16]. 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 overstress or even damage power semiconductor devices. Several solutions have been proposed in literature to control and limit the circulating current in parallel converters. In [18] an interleaved
active power filter concept with reduced size of passive components is presented. In [19] inter-phase transformers for connecting power converters in parallel are analysed to reduced the circulating current. In [20] 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 [21] a control method to limit the low frequency circulating current is presented, the high frequency circulating current is limited with 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 proposed inductor is not made of standard cores and needs a special core geometry with unequal size of the cores.
legs which makes it difficult to manufacture. In [1] an ICMI is proposed. The proposed ICMI combines boost inductor of LCL filter with inter-phase inductor into a single magnetic structure to reduce the size and the cost of the overall system. However, the comparison between the proposed ICMI and other structures, thermal evaluation, and finite element analysis are not considered. These issues are addressed in this paper.

In this paper a novel design for an ICMI is proposed which combines the inter-phase inductor ($L_{int}$) and the LCL boost filter inductor ($L_b$). This helps in reducing the component count and the size of overall system. The proposed design make use of 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 and accurate boost inductance is necessary, the existing design would need a large air-gap that may not be practically feasible. The proposed ICMI has been experimentally tested in a 7.5kVA, parallel single-phase converter. The circuit topology of the converter is shown in Fig. 1(a). The circuit configuration shown in Fig. 1(a) is a general grid connected converter configuration. This power converter can be used in applications such as ac-dc, dc-ac, and Static VAR Compensator (STATCOM). The experimental results from the laboratory prototype controlled as a single phase STATCOM validate the effectiveness of the proposed ICMI design.

II. CIRCULATING CURRENT 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 primary drawback of such systems is the circulating current between the parallel converters. The topology shown in Fig. 1(a), consists of two parallel single-phase H-bridge PWM converters with inter-phase inductor, $L_{int}$, between similar phases of the converters. An LCL filter is used to filter the switching frequency component in the converter output currents. The topology shown in Fig. 1(a), is considered for analysis in this paper.

By interleaving the carrier it is possible to reduce or eliminate certain switching harmonics [14], 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 configu-
The objectives of the inter-phase inductors are, first to offer a high impedance path for the circulating current. This ensures that the circulating current is limited to a small 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 an LCL filter. The LCL filter consists of grid side inductor \( L_g \), filter capacitor \( C_f \), and boost inductor \( L_b \). It is suggested in [17] to integrate the boost inductor and the inter-phase inductor into a single magnetic structure. In the suggested configuration the main inductance is used as inter-phase inductor, while the leakage inductance serves as the boost inductor. The structure of inductor designed in [17] is depicted in Fig. 1(b) and represents an ICMI with unequal air-gaps. Although this design reduces the cost and weight of the overall system, there are constraints in terms of factors such as the need for special core geometry and limits on the value of inductances that can be obtained. The windings are placed 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. 1(b), for a condition of equal air-gap and given number of turns, lower inductance is obtained when compared to the case when the windings are placed on the center limb. Second, the core requires different air-gap lengths for the center and side limbs. Standard core geometries of C cores and E cores lead to uniform air-gaps. Third, in high power applications with higher operating current, 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 [22]. In the shown configuration in Fig. 1(b), the minimum leakage inductance occurs when the center limb is completely removed. Lesser leakage inductance would not be geometrically feasible. There can be large tolerance in the value of the obtained inductance, in case of large air-gap, due to uncertainty in the fringing flux effects. If an air-gap is introduced in the magnetic core, flux paths are established in regions even beyond the core cross section area. This is referred to as fringing flux, and the amount of fringing increases with the air-gap dimension. The proposed ICMI is able to overcome these limitation in the design of the CM inductor.

III. THE PROPOSED ICMI

Fig. 1(c) and Fig. 1(d) show the structure of the proposed ICMI. Fig. 1(c) shows the flux paths for circulating current using dotted lines, and the output current using dashed line. The winding
connections are shown in Fig. 1(d). 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. 1(d). 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. Use of standard non-bifilar windings is also feasible and will result in a higher value of boost inductance, $L_b$, compared to the analytical expression for $L_b$ derived in this paper.

The leakage inductance of the proposed ICMI 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 points $A_1$ and $A_2$, is used as inter-phase inductor. The inter-phase 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 obtain 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 design of ICMI using standard C type amorphous cores are given in the next section.

IV. PROPOSED ICMI DESIGN

The flowchart for designing the ICMI is depicted in Fig. 2. The steps involving the design procedure are explained here.

**Step 1**- calculate the desired inductance values for boost inductor, $L_b$, [22] and inter-phase inductor, $L_{int}$, based on the power converter requirements.

**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 [23]. A sufficient number of core sizes
with area product higher than the calculated value in (1) are considered for evaluation.

\[ W_a A_c = \frac{1.25 L_b I_p^2 + 2.5 L_{int} I_{circ} I_p}{B_{de} J K_w (10^{-4})} \]  

(1)

Where, \( W_a = 2CD \) is window area of the core (cm\(^2\)) as shown in Fig. 8(c), \( A_c = AB \) is cross section of the core (cm\(^2\)), \( I_p \) is the peak current of individual converter (A), \( I_{circ} \) is the peak circulating current between converters (A), \( B_{de} \) is the desired peak flux density in the core (Tesla), \( J \) is the current density (A/cm\(^2\)), and \( K_w \) is the winding factor. The winding factor is a
factor that shows the feasibility of occupying the window area with the wire and it is calculated by dividing the total wire conductor area by the core window area. Different winding approaches and conductor shapes have different winding factor. It is assumed, in (1), that the flux density due to fundamental and switching current is 80% of $B_{de}$ and the flux density due to circulating current is 20% of $B_{de}$ in the core.

**Step 4** - calculate the minimum air-gap required for the chosen cores such that the variation in the core reluctance will not affect the total inductance by more than 10%. Set the air-gap required to be equal to the minimum air-gap. Fig. 3(a) 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 [23], calculate the number of turns, $N_1$ and $N_2$, required for the side windings to obtain the desired boost inductance value $L_b$. Using (10), given in appendix A, calculate the total number of turns needed in the center winding to obtain the required value of inter-phase inductance $L_{int}$.

**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 evaluate starting from step 4. $K_w = 0.4$ is normally chosen for windings with round conductors.

**Step 7** - calculate the maximum flux density in the core based on the peak current of the power converters. If the maximum flux density is higher than the saturation flux density or any desired flux density level $B_{de}$, then increase the air-gap, by 10%, and re-evaluate starting from step 5. For amorphous cores $B_{sat} = 1.5$ T is provided by the manufacturer. The $B_{max}$ used for actual design is lower than this value.

**Step 8** - the design outputs are the core size, the air-gap, and the number of turns selected for the ICMI.

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 wire considered is 9.4 US$ per kilograms where as the amorphous material is 14 to 18.75 US$ 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 density below the saturation flux density, $B_{sat}$, or below a desired flux density level, $B_{de}$. The steps mentioned above ensure that the core is chosen properly and the cost and the size are kept low. Additionally, the air-gap length is selected such that the flux density in the core is neither too small and the core is underutilized, nor too large that the core operates near its saturation limit.

V. DESIGN EXAMPLE

In this section an ICMI design example is given for the topology shown in Fig. 1(a). The total power rating of the considered system is 7.5kVA with grid voltage of 240V rms. The current rating of each converter is about 16A rms, while operating at UPF full load. The switching frequency is 10kHz. The design steps 1 to 8 from Section IV are explained below in a design example.

Step 1- The designed values for inter-phase and boost inductances are: $L_{int} = 20\text{mH}$, and $L_{b} = 900\mu\text{H}$ respectively. 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 of 20A; the current density chosen as, $J = 3 \text{A.mm}^{-2}$.

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. 3(a). 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 one tenth. This is to make the ICMI design less sensitive to core parameter variations.

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:

$$N_p = \frac{W_a K_w}{A_w}$$

(2)

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 cross section of a single wire. The possible number of turn $N_p$ along with the required number of turns $N_1 + N_3$ have
TABLE I
ICMI CORE AND WINDING PARAMETERS EVALUATION

<table>
<thead>
<tr>
<th>Core</th>
<th>PART No.</th>
<th>Side Windings $N_1$</th>
<th>Center Windings $N_3$</th>
<th>$B_{max}$ (T)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 AMCC0200</td>
<td>46</td>
<td>107</td>
<td>——</td>
<td>——</td>
</tr>
<tr>
<td>2 AMCC0250</td>
<td>42</td>
<td>99</td>
<td>——</td>
<td>——</td>
</tr>
<tr>
<td>3 AMCC0320</td>
<td>43</td>
<td>101</td>
<td>0.72</td>
<td></td>
</tr>
<tr>
<td>4 AMCC0400</td>
<td>39</td>
<td>90</td>
<td>0.63</td>
<td></td>
</tr>
<tr>
<td>5 AMCC0500</td>
<td>40</td>
<td>92</td>
<td>0.64</td>
<td></td>
</tr>
</tbody>
</table>

been calculated and depicted in Fig. 3(b). From the figure it can be seen that the first two cores can not hold the required number of turns. Hence, they are omitted and is not used.

Step 7- The contribution of circulating current and output current should be factored in the calculation of the maximum flux density, $B_{max}$. The summation of these two flux density terms should be lesser than the saturation flux density and the desired peak flux density. $B_{max}$ can be calculated using:

$$B_{max} = \left\{ \frac{L_{int} i_{cir}}{2(N_1 + N_3)} + \frac{L_{b} i_i}{N_1} \right\} \frac{1}{A_{side}}$$  \hspace{1cm} (3)

Where, $i_{cir}$ is the peak high frequency circulating current $i_{cir} = i_2 - i_3$ shown in Fig. 1(a) and from Fig. 8(c), $A_{side} = AB$ is the cross sections area of the side limb. $i_i$ is the maximum output peak current of individual power converters. The calculated $B_{max}$ is tabulated in TABLE I. It can be observed that none of the core flux densities is 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 ICMI.

VI. Finite Element Analysis of the Proposed ICMI

To analyse the proposed ICMI core flux density levels and flux linkage paths, the ICMI has been modelled using finite element method (FEM). Numerical analyses are carried out in 2D using FEMM package [24]. The grid used in the numerical simulation consists of 19690 nodes,
and the boundary is kept far enough away from the magnetic structure. Simulations have been performed with three current profiles. First, with only output current and no circulating current. This is used to evaluate the boost inductance $L_b$. Second, with the circulating current only. This provides the inter-phase inductance $L_{int}$ value. Third, with combined output and circulating current shown in Fig. 4(c). This corresponds to the case of maximum flux density, $B_{max}$, in the core. This simulation corresponds to the situation where the circulating current is at its peak value, and having a polarity that is the same as that of the current in the left side windings ($N_1$). This occurs when the circulating current polarity is the same as that of the left side winding, $N_1$, and the currents are at the peak. This results in maximum flux density in the left limb of the ICMI. It can be observed that the maximum flux density shown in Fig. 4(c) is 0.74T which is in agreement with the calculated maximum flux density given in Section V. The error between the calculated and simulated maximum flux density is less than 3%.
The simulated values of $L_b$ and $L_{int}$ are $985\mu$H and $19.54$mH respectively. The error between calculated values and simulated values of $L_b$ and $L_{int}$ are $9.4\%$ and $2.3\%$ respectively.

Flux linkage outside the magnetic component (external flux) should be kept as small as possible. The external flux heats up any nearby metallic objects and causes increased losses. Extra care should be taken while designing a magnetic structure to keep the external flux low and provide paths inside the cores. The flux lines are shown in Fig. 4. It can be seen that the most of the flux lines are inside the core and the external flux is negligible.

Fig. 4. Flux density and flux linkage plot of the proposed ICMI due to, (a) output current only, (b) circulating current only, (c) combined output and circulating current.
VII. COMPARISON BETWEEN PROPOSED ICMI AND OTHER DESIGNS

A. Proposed structure and existing design

Analysis based on reluctance calculation is used to carry out the comparison between the proposed structure and existing structure [17]. Magnetic equivalent circuit of CC core structure is given in Fig. 8(b). The reluctance offered to the circulating current in the available structure is calculated as,

$$ R_{ava} = \frac{1}{F}(2R_{gs} + 4R_t + 4R_s). \quad (4) $$

The reluctance offered to circulating current in the proposed structure is derived in Appendix and given as:

$$ R = \frac{1}{F}(R_{gs} + R_t + 2R_s) \quad (5) $$

It can be observed that the reluctance of the available design, $R_{ava}$, is larger than the proposed structure reluctance, $R$, by a factor of more than 2. Hence, for the same number of turns, air-gap, and core size the proposed structure obtains an inductance which is more than twice of that obtained by the available structure. Hence, for the same inductance value the proposed method needs smaller core size compared to available design.

In the available design, to get the desired inductance value for boost inductor, the center limb air-gap should be adjusted. It is found that for small values of boost inductance even by removing the center limb completely the desired value can not be obtained. However, in the proposed structure, the boost inductance value is controlled by choosing side windings number of turns. Hence, side windings turns can be chosen to obtain small values of boost inductance.

B. Separated Inter-phase and Boost Inductor Design

Alternative approach of a separate inter-phase and boost inductor base design is compared with the proposed ICMI. Three cases are considered,

Case (a) separate inter-phase inductor and two boost inductors, one for each legs, as shown in Fig. 5(a).

Case (b) separate inter-phase inductor and a boost inductor between the inter-phase inductor and the output filter, as shown in Fig. 5(b).
Case (c) same as case (a) but with air-gap-less inter-phase inductor.

Case (d) same as case (b) but with air-gap-less inter-phase inductor.

Case (e) proposed ICMI, shown in Fig. 1(d).

For the case of design with air-gap, the minimum air-gap is chosen based on the ratio of core reluctance to total reluctance. This ratio is chosen to be 10%, as shown in Fig. 3(a). $B_{de}=0.7T$ is selected as maximum flux in the core for all cases. The winding factor of 0.4 is selected for the design.

It is to be noted that the inter-phase inductors in case (a) and case (b) are not identical. Additionally, the boost inductance value in case (b) is half of the boost inductance value in case (a). The current rating of the boost inductor in case (b) is double with respect to case (a).

1) Inter-phase Inductor Design: The inter-phase inductor is used to reduce the circulating current between the power converters. The inter-phase inductor is a normal inductor with a center tap. The circulating current is quite small due to high inter-phase inductance value. However, half of the output current, $i_1$, is passing through each windings of the inter-phase inductor. Hence, the conductors current rating should be equal to rms of $i_1 + i_{cir}$. In case (a), it can be seen that, the boost inductor also limits the circulating current. Hence, to get the same effective inter-phase inductance, as in case (b), the inter-phase inductor value in case (a) should be $L_{int_a} = L_{int} - 2L_b$.

Whereas, in case (b), the inter-phase inductor value is equal to $L_{int_b} = L_{int}$. Among the available cores given in TABLE I, the smallest core which can hold the needed number of turns for inter-phase inductor to meet the design conditions with and without the air-gaps are AMCC0320 and
TABLE II

DEIGNED PARAMETERS OF THE SEPARATED INTER-PHASE AND BOOST INDUCTORS.

<table>
<thead>
<tr>
<th>Core Size</th>
<th>Inductance Value / p.u.</th>
<th>Air-gap</th>
<th>$B_{max}$</th>
<th>Turns</th>
</tr>
</thead>
<tbody>
<tr>
<td>Inter-phase Inductor case (a) AMCC0320</td>
<td>18.2mH / 74.5%</td>
<td>1.5mm</td>
<td>0.16T</td>
<td>138</td>
</tr>
<tr>
<td>Inter-phase Inductor case (b) AMCC0320</td>
<td>20mH / 81.8%</td>
<td>1.5mm</td>
<td>0.15T</td>
<td>145</td>
</tr>
<tr>
<td>Inter-phase Inductor case (c) AMCC0200</td>
<td>18.2mH / 74.5%</td>
<td>0</td>
<td>0.71T</td>
<td>30</td>
</tr>
<tr>
<td>Inter-phase Inductor case (d) AMCC0200</td>
<td>20mH / 81.8%</td>
<td>0</td>
<td>0.68T</td>
<td>32</td>
</tr>
<tr>
<td>Boost Inductor cases (a), (c) AMCC0063</td>
<td>900µH / 3.68%</td>
<td>2.5mm</td>
<td>0.72T</td>
<td>71</td>
</tr>
<tr>
<td>Boost Inductor cases (b), (d) AMCC0125</td>
<td>450µH / 1.84%</td>
<td>4mm</td>
<td>0.72T</td>
<td>51</td>
</tr>
</tbody>
</table>

AMCC0200 respectively. The design parameters are tabulated in TABLE II. It can be seen that the maximum flux densities in the gapped inter-phase inductor cores are 0.16T and 0.15T for the case (a) and (b) respectively, and the cores are under utilized. This is due to the need for large window area capable of holding the windings. So, big cores with large window areas are required. It is possible to use material like ferrite core which has the saturation flux density 0.2T to 0.5T but, to have a fair comparison amorphous core is chosen for the inter-phase inductor.

2) Boost Inductor Design: The boost inductors for the cases shown in Fig. 5 are designed and their parameters are given in TABLE II. In case (a) two boost inductors for each legs are needed. The AMCC0063 core size is used with 71 turns for each inductors. In case (b) one boost inductor is used. The core size is AMCC0125 with 51 turns. The conductor area for boost inductor in case (b) should be double with respect to case (a). SWG8 is chosen for the boost inductor in case (b).

C. Comparison between ICMI and Separated Boost and Inter-phase Inductor

The comparison between the aforementioned cases are tabulated in TABLE III. The proposed ICMI needs only two sets of AMCC0320. Whereas, case (a) needs two sets of AMCC0320 along with two sets of AMCC0063, case (b) needs two sets of AMCC0320 along with one set of AMCC0125, case (c) needs two sets of AMCC0200 along with two set of AMCC0063, and case (d) needs two sets of AMCC0200 along with one set of AMCC0125. The total cores weight
TABLE III

COMPARISON BETWEEN DIFFERENT INDUCTOR DESIGN CASES AND ICMI.

<table>
<thead>
<tr>
<th>Case</th>
<th>Number of CC Cores</th>
<th>Core Weight</th>
<th>Copper Weight</th>
<th>Length of Copper Wire</th>
</tr>
</thead>
<tbody>
<tr>
<td>(a)</td>
<td>4 sets</td>
<td>5.74kg</td>
<td>4.27kg</td>
<td>70.1m (SWG11)</td>
</tr>
<tr>
<td>(b)</td>
<td>3 sets</td>
<td>5.5kg</td>
<td>4.26kg</td>
<td>44.6m+11.6m (SWG11) (SWG8)</td>
</tr>
<tr>
<td>(c)</td>
<td>4 sets</td>
<td>4.73kg</td>
<td>1.97kg</td>
<td>32.2m (SWG11)</td>
</tr>
<tr>
<td>(d)</td>
<td>3 sets</td>
<td>4.5kg</td>
<td>1.91kg</td>
<td>7.14m+11.58m (SWG11) (SWG8)</td>
</tr>
<tr>
<td>(e)</td>
<td>2 sets</td>
<td>4.33kg</td>
<td>2.7kg</td>
<td>44.3m (SWG11)</td>
</tr>
</tbody>
</table>

of proposed ICMI is 4.334 kg. This is 24% and 21% smaller than the case (a) and case (b) respectively. Comparing cases (c) and (d) with proposed ICMI, case (e), the ICMI total core weight is reduced by 8.5% and 3.7% respectively. This shows that the proposed ICMI effectively reduces the total core weight needed.

The length of copper wire needed for ICMI is the least among cases corresponding to gapped cores, in (a), (b), and (e). In fact the ICMI needs lesser copper wire with respect to the inter-phase inductors in cases (a) and (b). This is due to higher diameter of the outer layers of the inter-phase inductor. In the inter-phase inductor all the turns are around the center limb with higher diameter than the side limb diameter. For cases where there is no air-gap in (c) and (d), the copper wire required is lesser than the ICMI. The reason is that by removing the air-gap from inter-phase inductor lesser number of turns are needed. However, cases (c) and (d) have some disadvantage over proposed ICMI. They are more sensitive to core magnetic parameter changes and also more number of core with smaller size are needed. Hence, cost associated with cores is lesser for ICMI compared with other designs. Optimization of the proposed ICMI design is out of scope of the paper and would be covered in future work.
VIII. EXPERIMENTAL RESULTS

The proposed ICMI has been built in the laboratory, the designed parameters are given in TABLE IV and the photograph of the ICMI is shown in Fig. 6(a). The fabricated inductors have been tested with a 7.5kVA two parallel single-phase grid connected converter with interleaved carrier. The power converter configuration is as shown in Fig. 1(a). Unipolar PWM method with switching frequency of 10kHz is employed as the modulation technique. The setup has been tested for continuous operation at full load as a STATCOM, for about two hours. A thermal image of the proposed ICMI is obtained using a FLUKE Ti20 thermal imager camera which is shown in Fig. 6(b). It can be seen, from Fig. 6(b), that the maximum temperature of the proposed ICMI is 88°C. In this case, the ICMI is cooled by natural convection. The operating temperature of the Amorphous cores are specified in the data sheet to be from -20°C to +155°C. Class F insulation grade is used along with these cores. It can be seen that the maximum temperature obtained is within the range of operating temperature of the core and the insulation. A 30W fan is attached to the ICMI to evaluate its temperature profile with forced cooling. Under ambient temperature of 25°C, it is observed that the maximum temperature settles down at 70°C, which is 18°C lower than the case of natural cooling.

Converter currents and their harmonic spectra are shown in Fig. 7(a) and Fig. 7(b) respectively. It can be seen that the fundamental currents supplied by both converters are almost identical and the circulating current between the converters is small. The circulating current is measured and shown in Fig. 7(c) 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 1.5A. Grid voltage and current and their spectra are shown in Fig. 7(d). 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. The THD of the grid current is calculated to be 1.93%.

The power loss in the ICMI is measured at full load condition using a WT1600 YOKOGAWA digital power meter. In Fig. 1(d), measurement of voltage between points A1 and M along with the current i2, are used to calculate $P_1$. Similarly, the voltage between points A2 and M along with the current i3 are measured to calculate $P_2$. The total power loss in the ICMI is the summation of $P_1$ and $P_2$. The ICMI power loss at full load condition was measured to be 50.2W, while
Fig. 6. (a) The fabricated ICMI for a 7.5kVA, 240V, single-phase power converter. (b) The thermal image of the ICMI after running the setup for two hours at full load condition under natural convection cooling.

operating at a power level of 7.5kVA.

### TABLE IV
**Designed Parameters of the ICMI.**

<table>
<thead>
<tr>
<th>SL.NO.</th>
<th>PARAMETER</th>
<th>Value / p.u.</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Inter-phase Inductance $L_{int}$</td>
<td>20mH / 81.8%</td>
</tr>
<tr>
<td>2</td>
<td>Boost Inductance $L_b$</td>
<td>900 $\mu$H / 3.68%</td>
</tr>
<tr>
<td>3</td>
<td>Side Windings $N_1$, $N_2$</td>
<td>40 turns</td>
</tr>
<tr>
<td>4</td>
<td>Center windings $N_3$</td>
<td>100 turns</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>AMCC0320</td>
</tr>
<tr>
<td>8</td>
<td>Measured $L_{int}$</td>
<td>19.8mH / 80.98%</td>
</tr>
<tr>
<td>9</td>
<td>Measured $L_b$</td>
<td>940 $\mu$H / 3.84%</td>
</tr>
</tbody>
</table>

### IX. Conclusion

A novel ICMI 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
Fig. 7. Current waveform in the parallel converters. (a) Converter currents, (b) converter currents harmonic spectra, (c) circulating current with its harmonic spectrum, and (d) Grid voltage and grid current and their harmonic spectra.
core geometries that improves manufacturability. A step by step design procedure of the ICMI is presented. The proposed design reduces the size and the cost of overall system and is insensitive to core parameter tolerances. Experimental results on a 7.5kVA parallel single-phase converter show that the fundamental circulating current is about zero and the high frequency peak circulating current between the converters is less than 1.5A. The power loss in the ICMI is measured to be 50.2W, which represents about 0.67% of system rating. Although a parallel single-phase power converter is presented here, the proposed method can also be extended to parallel three-phase power converters.

X. APPENDIX

A. Inter-phase Inductance Calculation

The proposed ICMI with individual windings voltage drops and fluxes due to circulating current is shown in Fig. 8(a). The inter-phase inductance, inductance between terminals $A_1$ and $A_2$, limits the circulating current. It is calculated as given below.

$$V_{int} = V_1 + V_3 + V_2$$ (6)

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

$$V_{int} = N_1 \frac{d\varphi_1}{dt} + N_3 \frac{d\varphi_3}{dt} + N_2 \frac{d\varphi_2}{dt}$$ (7)
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,

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

(8)

The expression in (7) is simplified using (8) as,

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

(9)

Equation (9) shows that the structure shown in Fig. 8(a) is equivalent to an inductor with a same core structure but only 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 (10).

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

(10)

Where, \( \mathcal{R} \) is the reluctance of the structure shown in Fig. 8(b) and defined in Appendix X-C.

B. Boost Inductance Calculation

As explained in Section III, 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 (11).

\[
N_1 = \sqrt{L_b \ 2\mathcal{R}}
\]

(11)

C. Reluctance Calculation

Fig. 8(b) shows the magnetic equivalent circuit of the proposed ICMI for circulating current. The equivalent circuit is drawn with \( N \) effective turns on the center limb of the ICMI. Fig. 8(c) shows the dimensions of the core used in the ICMI. The reluctances \( \mathcal{R}_t, \mathcal{R}_c, \) and \( \mathcal{R}_s \) are the corresponding reluctances of the core. \( \mathcal{R}_{gs}, \) and \( \mathcal{R}_{gc} \) are the reluctance of the side and center air-gap respectively. These are calculated as

\[
\mathcal{R}_t = \frac{L_{c1}}{\mu_0 \mu_r A_c}, \quad \mathcal{R}_s = \frac{L_{c2}}{\mu_0 \mu_r A_c}, \quad \mathcal{R}_c = \frac{L_{c2}}{\mu_0 \mu_r 2A_c}
\]

\[
\mathcal{R}_{gs} = \frac{g}{\mu_0 A_c}, \quad \mathcal{R}_{gc} = \frac{g}{\mu_0 2A_c}
\]

(12)
Where, the $A_c$ is the cross section of the core shown in Fig. 8(c) and is calculated as $A_c = A \times B$. $L_{c1}$ and $L_{c2}$ are the magnet length shown in Fig. 8(c). $\mu_o$ and $\mu_r$ are the magnetic permeabilities of air and the core material respectively. The core and air-gap reluctances are calculated as

$$\mathcal{R}_{\text{core}} = \mathcal{R}_t + \mathcal{R}_s + 2\mathcal{R}_c \quad (13)$$

$$\mathcal{R}_{\text{gap}} = \mathcal{R}_{gc} + \frac{1}{2} \mathcal{R}_{gs} \quad (14)$$

Simplifying (13) and (14), using $\mathcal{R}_{gc} = 0.5 \mathcal{R}_{gs}$ and $\mathcal{R}_c = 0.5 \mathcal{R}_s$, the total reluctance is calculated as

$$\mathcal{R} = \frac{1}{F} (\mathcal{R}_{\text{core}} + \mathcal{R}_{\text{gap}})$$

$$= \frac{1}{F} (\mathcal{R}_{gs} + \mathcal{R}_t + 2\mathcal{R}_s) \quad (15)$$

Where, $F$ is the fringing factor [25] and is equal to:

$$F = 1 + \frac{2g}{\sqrt{A_c}} \ln\left(\frac{2C}{g}\right). \quad (16)$$

REFERENCES


