

Current-Balancing Technique for an Interleaved Full Bridge Converter for the AC Battery Application

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(IFB) converter is studied in this paper for the Power Factor Correction (PFC) stage in the power converter system. Besides the inherent merits that make the IFB converter suitable for this domestic application, the current imbalance phenomenon is a potential drawback that leads to overheated devices and inductors, thereby reducing the stability and deteriorating the performance of the system. Firstly, based on the mathematical analysis, the cause of the current mismatch is analyzed in this paper. Then, a simple and effective solution for the current imbalance is proposed by just adding one linear controller to compensate for the deviation of the modulation index. This proposed control structure also maintains the advantages of the IFB converter and its modulation technique as well. To evaluate the unbalanced current issue and the proposed method, only the current loop is concentrated and simulated with various simulation scenarios. Finally, an experimental IFB prototype is carried out, and the good experimental results verify the effectiveness of the proposed method.

Keywords AC Battery, PFC stage, Interleaved Full Bridge converter, current imbalance.

1. Introduction

Electric Vehicles (EVs) are estimated to account for more than 60% of vehicles sold globally by 2030 and are going to replace gasoline cars to guarantee zero net emissions at the COP-26 conference [1]. Along with the proliferation of EVs, EV batteries have gained great concern because these batteries will be discarded from EVs when they are degraded to 70–80% of the initial capacity and then become insufficient for automotive use [2]. These batteries could be mainly repurposed in the battery energy storage system (BESS) and other applications, so-called Second Life Batteries (SLB) [3,5]. Recently, SLBs have been utilized in the home battery storage system [6, 10] to exchange energy with other domestic energy resources such as grid power, renewable energy, EV batteries, and household loads through the power converter system. In this study, the focus is on the AC-coupled configuration of the home battery due to its capacity to seamlessly integrate with existing home photovoltaic (PV) systems, its self-contained battery system, and its adaptability for expansion with additional modules. The energy from the PV is not only stochastic but also has its maximum power at

noon, while the peak domestic load demand occurs at around 6 a.m. and 6 p.m.

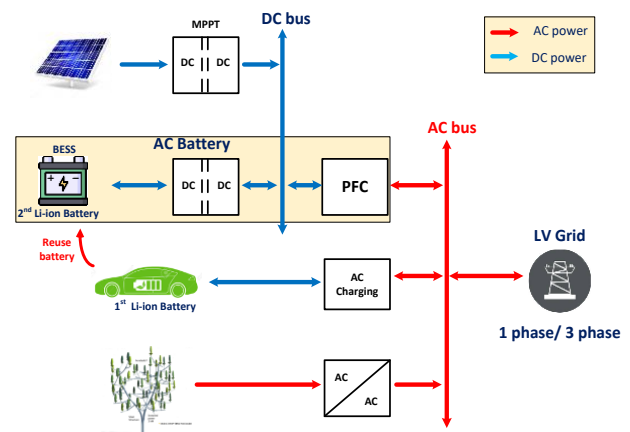


Figure 1. The overall structure of an AC Battery application

The AC Battery, which combines both the battery and power converters, functions as an energy storage to overcome the inherent disadvantage of renewable energy, particularly in this domestic application. Numerous reputable companies have recently introduced this product with an increasing power rate by incorporating multiple modules.

In this research, the power rate of the converter is chosen at 10 kW, which is similar to Energy Forverer's product. The two-stage topology in Fig. 1 is selected for flexible purposes for the AC Battery, with the maximum efficiency of the DC/DC stage and the PFC stage calculated as 98.7 % and 99 % at the nominal power rate, respectively. Hence, the maximum efficiency of the overall AC Battery that can be achieved is 97.7 %. In particular, the PFC stage is focused on in this paper. The ability to provide 3-phase grid power in different regions and the required power rate of the battery system affect the structure of the PFC converter in the domestic BESS. In the US and some countries, only the single-phase grid is supplied, while the power rate for charging EVs

is up to 10 kW. Even at a power rate of 6.6 kW, the Totem-Pole structure achieves the best performance according to the TI design [11].

However, by increasing the power up to 10kW, the current ripple and the current stress significantly increase, which leads to more losses, a bulky volume of passive components, and more electromagnetic noise. One particular solution is to use the Bridgeless interleaved boost (BLIL) [12] converter that helps reduce the current ripple significantly. However, this structure is only a unidirectional converter because it uses diodes as the upper switches, which is not suitable for bidirectional applications of the domestic AC Battery [13-15]. To address all the above problems, the Interleaved full bridge (IFB) structure is applied in this paper. By replacing the diodes of the BLIL structure with controllable MOSFETs, the converter operates as two interleaved H-bridge circuits with the ability of bidirectional power flow.

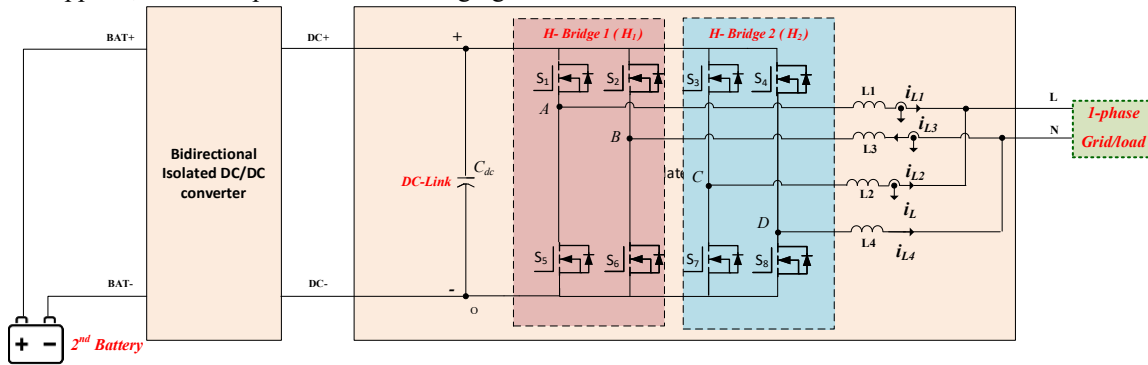


Figure 2. The IFB converter in multiple modes in an AC Battery application

Some outstanding advantages of the IFB structure can be mentioned as the reduction in the current stress and the inductor current ripple, thereby lowering the EMI circuit and the magnetic circuit size. At the same time, it also reduces high-frequency voltage ripple, hence improving efficiency for the converter. In addition, this structure is flexible in modification so that it can operate with both the 1-phase or 3-phase grid [13], which allows the AC battery to be suitable for different regions.

Pulse width modulation techniques for the IFB converter have been studied [14]. Unipolar double-frequency (PWM) modulation has a high-frequency circulating current, and its influence on split-filter elements can be minimized. Unipolar Frequency-PWM modulation has a high-frequency circulating current, but its effect on split inductors cannot be minimized. In particular, unipolar-PWM modulation has been proven to have no high-frequency circulating currents in any operating mode of the converters [14]. At the same time, based on the idea of the BLIL converter, the phase of the carrier for the two H-bridges of the IFB converter will be shifted by 180 degrees to reduce the ripple of the output current. This method of pulse width modulation is defined as the interleaved unipolar PWM (IU-PWM), which also helps to achieve low output voltage THD

compared to the Unipolar PWM method, and reduces leakage current as well.

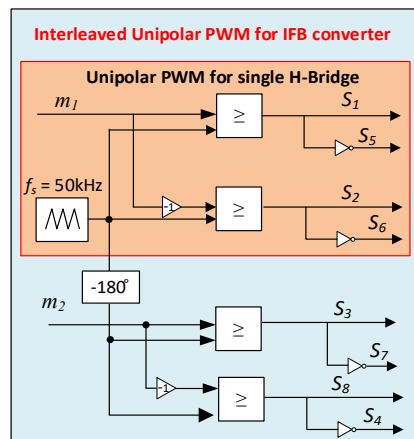
However, paralleled power modules often have current imbalance issues due to parameter mismatches [15]. With regard to the IFB converter, utilizing the IB-PWM method makes the advantages of the converter with the chosen modulation technique not be maintained. This issue limits the application of this PFC structure. The unbalanced current causes high current stress, additional power loss, and overheating on the split inductors and the power switches [16], which in turn affects the lifespan of the device and the quality of the converter operation. The current imbalance has been mentioned and solved in [17] for the Unipolar Frequency- PWM modulation; however, the only mentioned cause in this modulation technique is the differences between reference currents or the measurement amplification, which does not take the effect of the inductor imbalance into account. Meanwhile, the difference between internal resistance and inductance greatly affects the balance of current through each branch of the converter. Therefore, no study has provided a detailed explanation for the cause of inductor current imbalance in the IFB converter with IU-PWM modulation, as well as the solution to balance the current without changing the structure or other modulation rules.

This paper will handle all these problems by proposing an improved control structure that will allow the IFB to operate well with the IU-PWM modulation, which only requires one additional current sensor and retains the advantages of both the hardware structure and the modulation method. In detail, the problem of current imbalance among phases of the Interleaved Unipolar- IFB converter is first defined, and the reason is analyzed with equivalent circuits and control loop structure. Then, a new control structure was proposed to eliminate the current imbalance with various simulation scenarios. Furthermore, an experimental prototype for controlling the current loop of the IFB converter is carried out to verify the proposed control algorithm.

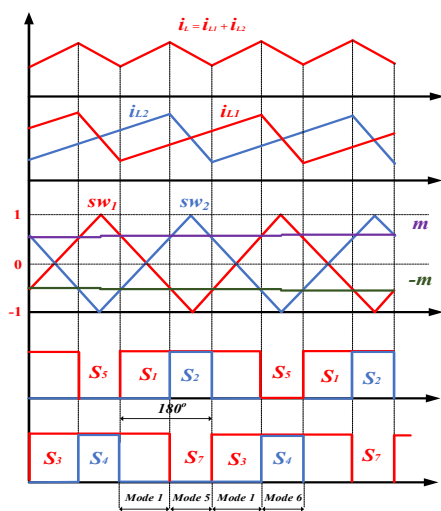
2. Preliminary mathematics

2.1 Operational principle of the IFB converter with the IB-PWM method

In this section, the interleaved unipolar PWM modulation is defined and applied to the IFB converter with its structure shown in Fig. 2.



a)



b)

Figure 3. IU-PWM method for the IFB converter

a) Pulse generation; b) Typical pulse pattern
Switches S_1 , S_2 , S_5 , and S_6 are in the first H-Bridge (H_1), and two branches connected to 2 split inductors L_1 and L_2 are symbolized as branch 1 and branch 2, respectively. Similarly, switches S_3 , S_4 , S_7 , and S_8 belong to the second H- Bridge (H_2), with branch 3 and branch 4 connecting to two split inductors L_3 and L_4 , respectively.

The Unipolar PWM technique for an H-Bridge converter uses a modulation signal to compare with a single carrier. With regard to the interleaved Unipolar PWM technique for the IFB converter, the pulse diagram is distributed as shown in Fig. 3a. In the ideal case, the two carriers of m_1 and m_2 are the same, The phase of these two carriers is guaranteed to shift exactly 180 degrees, so-called the Interleaved unipolar PW method. The control signals are applied to the switches S_1 to S_8 in Fig. 3a, respectively, with the corresponding pulse pattern shown in Fig. 3b.

2.2. Current imbalance phenomenon

Fig. 4 shows the overall control structure for the IFB converter, in which this paper exclusively focuses on the tested current loop to verify the current responses in the balanced and unbalanced cases. In detail, a DC source is applied at the DC side of the converter. On the other hand, the output of the inductors is connected, and the inductor current of each H-Bridge (current through inductors L_1 and L_2) is measured and controlled independently by two independent controllers. This ensures steady-state quality, dynamic response, and the ability to protect the switches of each converter in the ideal case.

The control structure in Fig. 4 is verified by numerical simulation; hence, the unbalanced current problem is first detected and shown in this section. There are a total of four simulation cases with the corresponding parameters given in Table 1.

Table 1. Balanced and unbalanced parameters

Ideal case (Case 1)	
DC voltage	100 VDC
Reference current	$i_L^* = 10\sqrt{2} \sin(314t)$
Split-inductor	Inductance $L = 500\mu H$ Internal resistor $r_L = 0.8\Omega$
Unbalanced case	
Case 2	$L_1 = 70\%L$; $L_2 = L_3 = L_4 = 100\%L$ $r_{L1} = 70\%r_L$; $r_{L2} = r_{L3} = r_{L4} = 100\%r_L$

Case 3	$L_1 = 70\%L; L_2 = 130\%L; L_3 = L_4 = 100\%L$ $r_{L1} = 70\%r_L; r_{L2} = 130\%r_L; r_{L3} = r_{L4} = 100\%r_L$
Case 4	$L_1 = 70\%L; L_2 = 70\%L; L_3 = L_4 = 100\%L$ $r_{L1} = 70\%r_L; r_{L2} = 70\%r_L; r_{L3} = r_{L4} = 100\%r_L$

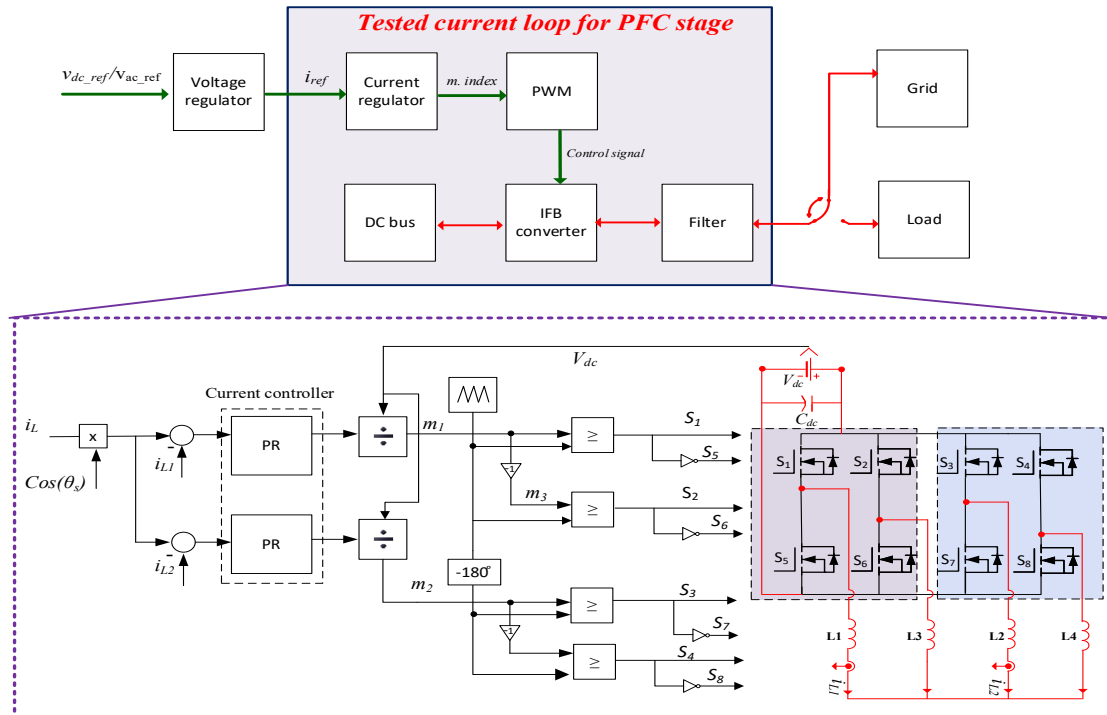


Figure 4. Tested current loop for the IFB converter

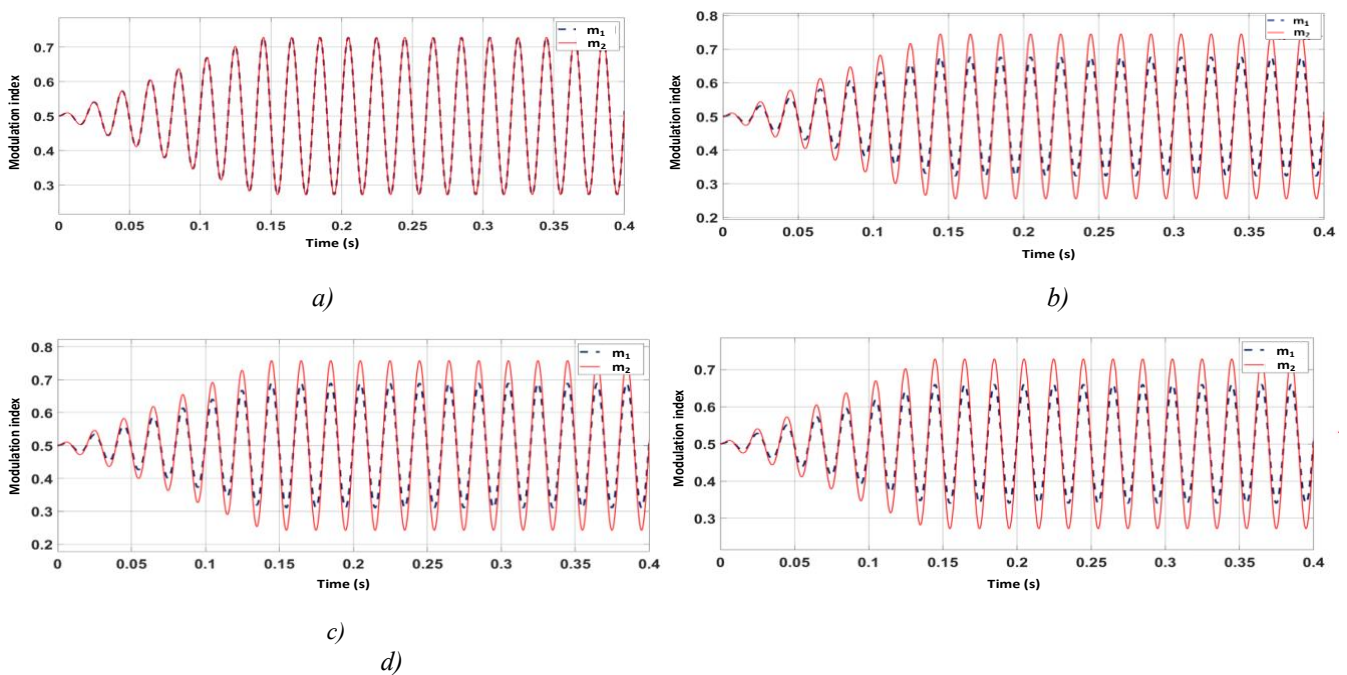


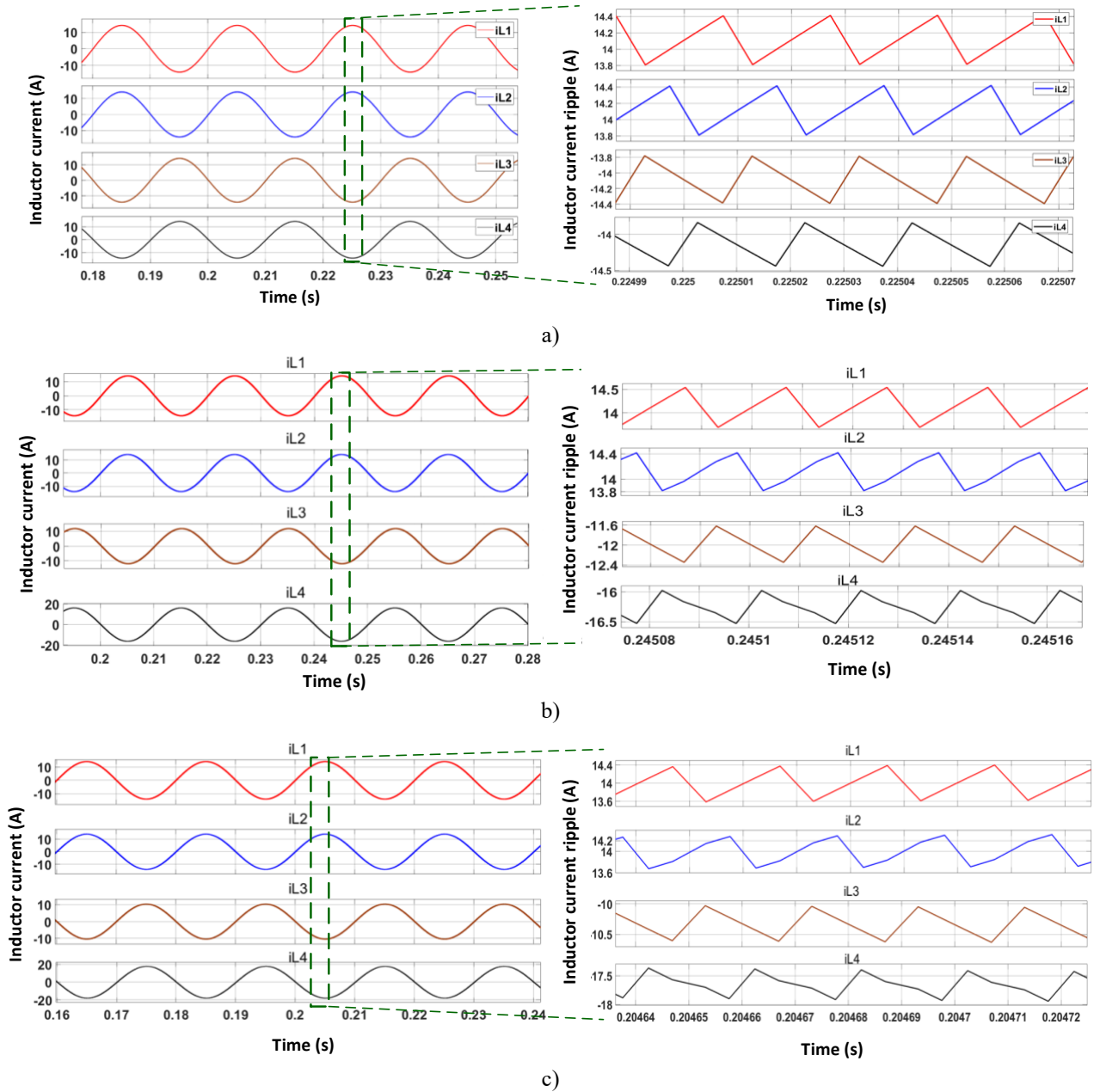
Figure 5. Response of modulation indices in different cases:

a) Case 1; b) Case 2; c) Case 3; d) Case 4

In case 1, the parameters of the converter are ideal, and the four split inductors are equal, so a balanced response can be achieved among the inductor currents through four phases, which are shown in Fig. 6a with the same Root Mean square (RMS) values and peak values. At this time, the modulation indices for the two H-Bridges are equal, as shown in Fig. 5a. This is the basis for the comparison with the next unbalanced cases.

In practice, the internal resistance and inductance values of inductors are not ideally equal. Some cases of unbalanced inductors are given for analysis and comparison. Case 2 illustrates that only one inductor is reduced, taking L_1 for instance. Case 3 applies the change for two inductors in the same H-Bridge; for example, inductor L_1 in H_1 decreases, and inductor L_3 increases by the same amount of 30% of the

initial value. Case 4 is similar to case 3, but both inductors L_1 and L_3 decrease. The modulation indices in cases 2, 3, and 4 with the differences are shown in Fig. 5b, 5c, and 5d, respectively. For the case in which there is only one variable inductor, as shown in Fig. 6b, the inductor currents i_{L1} and i_{L2} keep tracking the reference current as in case 1, but the inductor currents i_{L3} and i_{L4} are unbalanced: the current i_{L3} is lower than the reference value, and the modulation index for the H_1 also decreases compared to the ideal value. At the same time, the current i_{L4} increases along with a growth in the modulation index for H_2 compared to case 1. The RMS value of i_{L1} , i_{L2} , i_{L3} , and i_{L4} are 9.92A, 9.95A, 8.28A, and 11.2A, respectively.



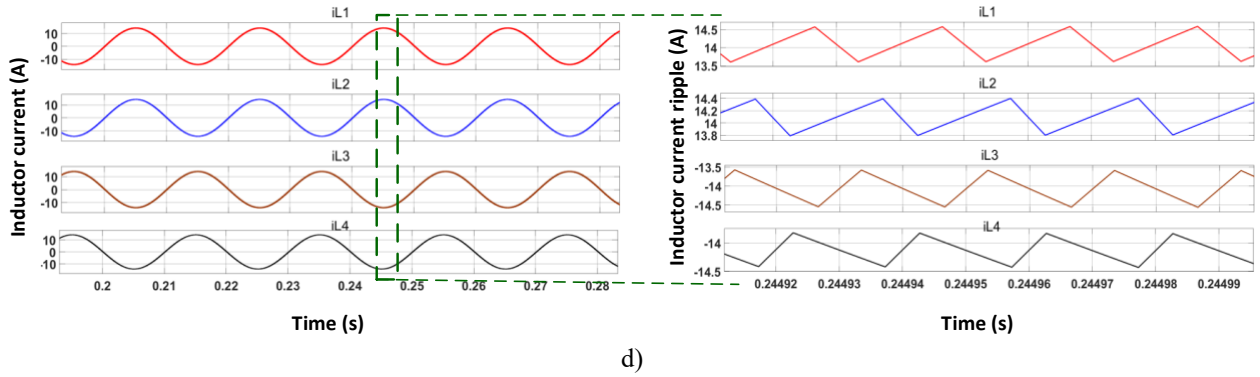


Figure 6. Inductor current responses in various cases:

a) Case 1; b) Case 2; c) Case 3; d) Case 4

For the case experiencing the deviation of two inductors, firstly in case 3 as shown in Fig. 6c, in which the currents i_{L1} and i_{L2} continue to balance and be equal to the reference value, the current imbalance issue occurring on the currents i_{L3} and i_{L4} becomes more serious than that in case 2. Correspondingly, the 2 modulation indices also deviated much further from the original value than in case 2. Conversely, when the parameters of the inductor L_3 decrease by an equivalent amount to those of inductor L_1 in case 4, the simulation result in Fig. 6d illustrates that the inductor currents i_{L3} and i_{L4} tend to rebalance with a negligible difference from the reference value.

From the aforementioned simulation results, if there is a mismatch among the four split-inductors of the IFB converter applying the IU-PWM technique, the current in the phase at which the current sensor is set to measure and control will balance to the reference current, while the current of the remaining phases will be unbalanced with different trends. With the same analysis, the current imbalance issue caused by the misalignment in inductors of the H_2 or both H-Bridges is similarly assessed.

2.2. Analysis of current imbalance cause

Generally, m_1 and m_3 are the modulation indices for 2 legs with inductors L_1 and L_3 of the first H-Bridge, respectively. Applying Kirchhoff 2 law to the circuit diagram of H_1 in Fig. 4:

$$(m_1 - m_3) \frac{V_{dc}}{2} = r_{L1} i_{L1} + L_1 \frac{di_{L1}}{dt} - r_{L3} i_{L3} - L_3 \frac{di_{L3}}{dt} \quad (1)$$

In which:

$$r_{L1} = r_L + \Delta r_{L1}; r_{L3} = r_L + \Delta r_{L3}; L_1 = L + \Delta L_1; L_3 = L + \Delta L_3$$

In the ideal case, the inductance L and internal resistance r_L are balanced among inductors:

$$\Delta r_{L1} = \Delta r_{L3} = \Delta L_1 = \Delta L_3 = 0 \quad (2)$$

With conditions (2), (1) becomes:

$$(m_1 - m_3) \frac{V_{dc}}{2} = r_L (i_{L1} - i_{L3}) + \frac{L d(i_{L1} - i_{L3})}{dt} \quad (3)$$

By applying the PR controller [18], the current i_{L1} will track the reference current i_L under any case. The corresponding modulation index in the ideal case is defined as m . The rule of traditional unipolar PWM determines: $m_1 = -m_3 = m$. Hence, (3) is rewritten as:

$$m V_{dc} = r_L (i_L - i_{L3}) + L \frac{d(i_L - i_{L3})}{dt} \quad (4)$$

A similar analysis for the second H-Bridge:

$$m V_{dc} = r_L (i_L - i_{L4}) + L \frac{d(i_L - i_{L4})}{dt} \quad (5)$$

Applying Kirchhoff 1 law to the circuit diagram in Fig. 4:

$$i_{L1} + i_{L2} + i_{L3} + i_{L4} = 0 \quad (6)$$

$$\Rightarrow r_L (2i_L + i_{L3} + i_{L4}) + L \frac{d(2i_L + i_{L3} + i_{L4})}{dt} = 0 \quad (7)$$

By adding (4), (5), and (7):

$$2m V_{dc} = 4r_L i_L + \frac{4L di_L}{dt} \Rightarrow m \frac{V_{dc}}{2} = r_L i_L + L \frac{di_L}{dt} \quad (8)$$

Substituting (8) into (4), (5):

$$\begin{cases} m \frac{V_{dc}}{2} = -r_L i_{L3} + L \frac{d(-i_{L3})}{dt} \\ m \frac{V_{dc}}{2} = -r_L i_{L4} + L \frac{d(-i_{L4})}{dt} \end{cases} \quad (9)$$

Equation (9) proves that if all the split-inductors are the same, even though not being measured and controlled, the current through inductors L_3 and L_4 is still balanced with the current through L_1 and L_2 . Furthermore, equation (8) presents an important relationship between modulation index m and the initial resistance r_L and inductance L of the inductor in the ideal case of the IFB converter operation.

In the non-ideal case, the inductor mismatch problem is taken, for instance, with inductors L_1 and L_3 of the H_1 , where $\Delta r_{L1}, \Delta r_{L3}, \Delta L_1, \Delta L_3 \neq 0$. These conditions make the transfer function from current to modulation index different from that

$$(2m + \Delta m_1 + \Delta m_1) \frac{V_{dc}}{2} = (r_L + \Delta r_{L1}) i_L + \frac{(L + \Delta L_1) di_L}{dt} - (r_L + \Delta r_{L3}) (-i_L - \Delta i_{L3}) - \frac{(L + \Delta L_3) d(-i_L - \Delta i_{L3})}{dt} \quad (13)$$

$$\Rightarrow (m + \Delta m_1) V_{dc} = 2(r_L i_L + L \frac{di_L}{dt}) + \Delta r_{L1} i_L + \Delta L_1 \frac{di_L}{dt} + \Delta r_{L3} i_L + r_{L3} \Delta i_{L3} + \Delta L_3 \frac{di_L}{dt} + L_3 \frac{d\Delta i_{L3}}{dt} \quad (14)$$

Substituting (8) into (13):

$$r_{L3} \Delta i_{L3} + L_3 \frac{d\Delta i_{L3}}{dt} = V_{dc} \Delta m_1 - (\Delta r_{L1} + \Delta r_{L3}) i_L - (\Delta L_1 + \Delta L_3) \frac{di_L}{dt} \quad (15)$$

With equation (15), it is clear that the current imbalance on inductor L_3 is created by 2 main components: $V_{dc} \Delta m_1$ that depends on the modulation index difference and $(\Delta r_{L1} + \Delta r_{L3}) i_L + \frac{(\Delta L_1 + \Delta L_3) di_L}{dt}$ that depends on the inductor imbalance. The latter component also explains the phenomenon that occurs in simulation cases 3 and 4. Assuming that the inductance and internal resistance of the inductor L_1 experience a decrease of 30%. If the internal resistance and inductance of L_3 also decrease by the same amount, the imbalance current issue becomes more serious. However, in the circumstance that the internal resistance and inductance of L_3 decrease by the same amount, this discrepancy among inductor currents gets much smaller, considered to be negligible.

2.3. Handling the current imbalance issue

The difference between the ideal modulation index and the modulation index for branch 1 in the H_1 , combined with the imbalance of internal resistance and inductance, generates the difference in current i_{L3} from the reference one. The main idea of the proposed method is to focus on the

in the ideal case. However, because the current i_{L1} is still measured and controlled to track the reference i_L ($\Delta i_{L1} = 0$), the modulation index for branch 1 in H_1 has to change an amount as defined:

$$m_1 = m + \Delta m_1 \quad (10)$$

Based on the rule of unipolar PWM:

$$m_3 = -(m + \Delta m_1) \quad (11)$$

As the simulation results in the previous section, there is a difference between i_{L3} and $-i_L$:

$$i_{L3} = -i_L - \Delta i_{L3} \quad (12)$$

Substituting (10), (11), and (12) into (2):

modulation index, where 2 different modulation indices are applied for two branches of H_1 instead of only one as in the traditional Unipolar PWM rule. The output of the controller for i_{L1} is the modulation index for branch 1, while the compensated modulation index is obtained based on the proposed method for the unbalanced branch. In detail, instead of using the conditions $\Delta m_3 = -\Delta m_1$ in (11), the modulation index of branch 3 will be compensated by an amount of $2\Delta m$:

$$\Delta m_1 - \Delta m_3 = 2(\Delta m_1 + \Delta m) \quad (16)$$

Substituting (16) into (15):

$$\begin{aligned} r_{L3} \Delta i_{L3} + L_3 \frac{d\Delta i_{L3}}{dt} &= (\Delta m_1 + \Delta m) V_{dc} - (\Delta r_{L1} + \Delta r_{L3}) i_L \\ &\quad + (\Delta L_1 + \Delta L_3) \frac{di_L}{dt} \end{aligned} \quad (17)$$

The current imbalance is solved when $\Delta i_{L3} = 0$, resulting in:

$$r_{L3}\Delta i_{L3} + L_3 \frac{d\Delta i_{L3}}{dt} = 0 \quad (18)$$

Therefore, the compensated modulation is determined from the root of the unbalanced current problem:

$$V_{dc}\Delta m = -\left[V_{dc}\Delta m_1 - (\Delta r_{L1} + \Delta r_{L3})i_L + (\Delta L_1 + \Delta L_3)\frac{di_L}{dt}\right] \quad (19)$$

Ideally, if the deviation amounts of the inductors are known in advance, a direct feed-forward compensation of the modulation coefficient Δm may be implemented. However, if this inductor mismatch cannot be determined in a black-box system or varies during system operation. Therefore, the

proposed compensation in this paper is determined as the output of a current difference Δi_{L3} controller which will be updated every sampling period. The transfer function from the current error to the modulation index is derived by substituting (17) into (19):

$$V_{dc}\Delta m = -(r_{L3}\Delta i_{L3} + L_3 \frac{d\Delta i_{L3}}{dt}) \quad (20)$$

Therefore:

$$\frac{\Delta i_{L3}}{V_{dc}\Delta m} = -\frac{1}{L_3 s + r_{L3}} \quad (21)$$

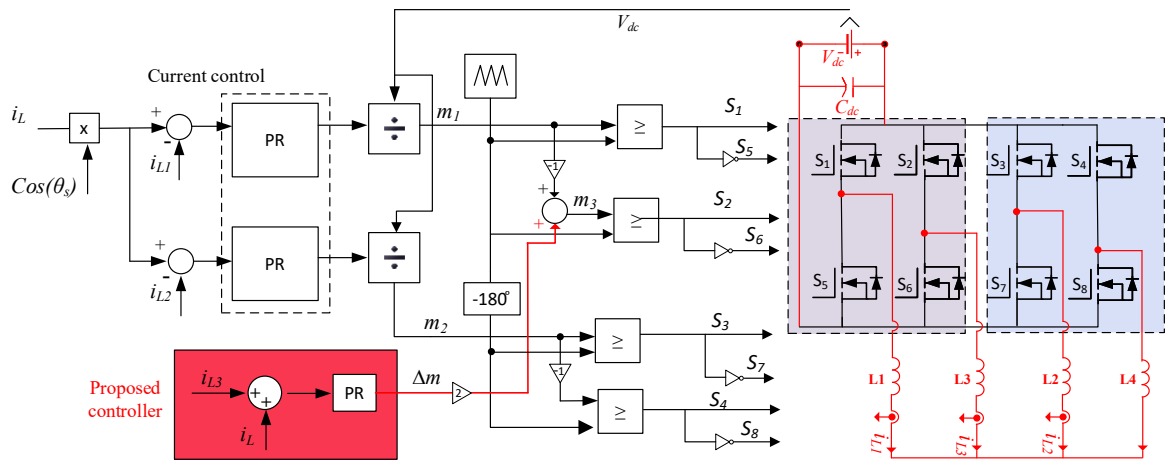
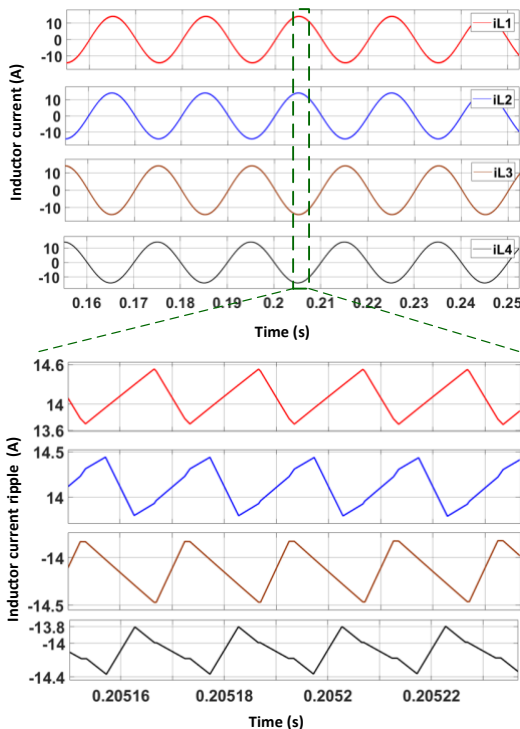
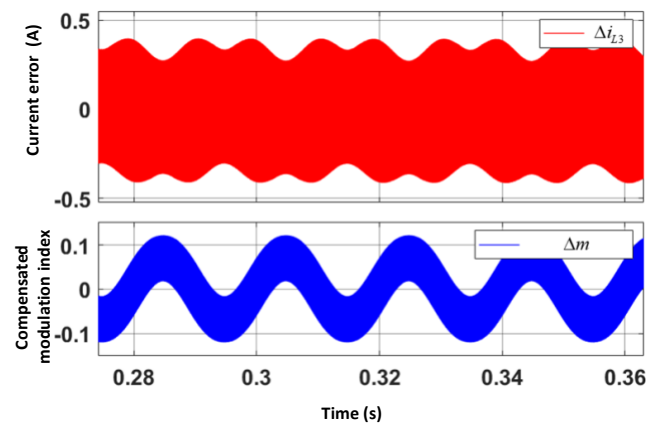


Figure 7. Proposed control structure for current imbalance issue in IFB converter



a) Responses of 4 inductor currents



b) Current discrepancy and the compensated modulation index for branch 3

Figure 8. Simulation results in the nonideal case with the proposed control method

To implement the proposed method, simply an additional current measurement sensor for inductor L_3 is

required, and the PR controller is designed with a similar principle to that for i_{L1} and i_{L2} . With the wide bandwidth, the controller could operate properly regardless of the inductor parameter changes within the transfer function in (21). Thanks to the controller, the compensated modulation index for branch 3 is directly obtained from the difference of current i_{L3} every sampling period. Note that the output of the controller is always added to the modulation index of the balanced branch to get the regulated modulation index for the unbalanced branch. If Δi_{L3} is smaller than zero, then a positive amount is added to the modulation index for branch 1, and else if is larger than zero, then a negative amount is added.

From all the above analysis, the proposed control structure for the current imbalance elimination is presented in Fig. 7. To verify the correctness and effectiveness of the proposed method, only the current loop is implemented in which one side of all the split-inductors is matched. The main advantages of this technique include: a simple control strategy with only one additional linear controller instead of the complex nonlinear control structures, simple hardware reconfiguration with one additional current measurement sensor instead of changing the structure of the converter, and maintaining the merits of the Interleaved unipolar PWM method for the IFB converter. The proposed control structure is first verified by numerical simulation, with the parameter of the unbalanced case 2 in Table 1. An important note is that the higher the power rate of the operating converter is, the more serious the current imbalance issue occurs.

By compensating the modulation index with a PR controller as in Fig. 7, the current imbalance problem among the four branches is effectively solved. Instead of being 9.92A, 9.95A, 8.28A, and 11.2A as in the previous simulation results, the RMS value of 4 inductor currents returns balanced at 10.02A, 10.02A, 10.02A, and 10.03A, respectively. The waveforms and ripples of all inductor currents before and after applying the proposed method are presented in Fig. 6a, Fig. 8a, and Fig. 8b, respectively. In addition, the compensated modulation index determined from the controller is shown in Fig. 8c. The input of the controller representing Δi_{L3} is kept around zero. The good effect of the regulated modulation index theoretically proves the effectiveness of the proposed method.

3. Experimental verification

The experimental system for verifying the proposed control method is carried out with the detailed parameters shown in Table 2. The algorithm is embedded in a TI DSP TMS320F379D kit, ensuring a phase shift of 180 degrees of the IB-PWM method. The applied IC for the current

measurement is ACS724LMATR-65AB-T with high bandwidth and accuracy to ensure the measured results. The current label labeled MD9272 is used to measure and display the RMS current through each split-inductor. Furthermore, the SiC MOSFET technology is selected for the power switches in this AC Battery application to improve the power density and efficiency of the system.

Table 2. Experimental parameters

Reference current	$i_L^* = 10\sqrt{2} \sin(314t)$
Split-inductor	$L_1 = L_2 = L_4 = 500\mu H;$ $L_3 = L_{test} = 350\mu H$
Switching frequency	50kHz
DC voltage	100VDC
SiC Mosfet	C2M0080120D

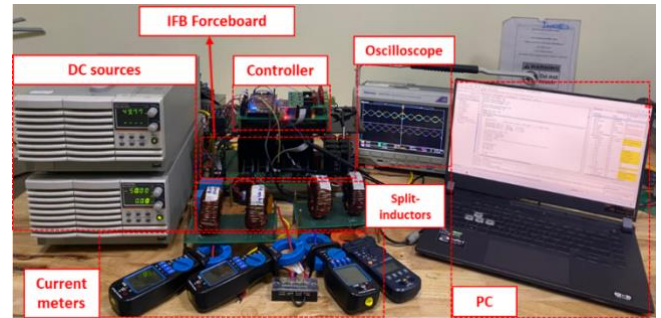
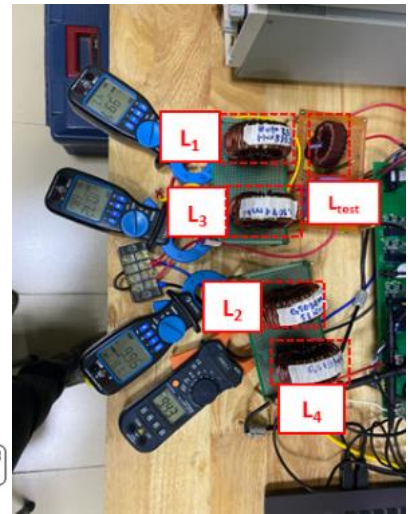
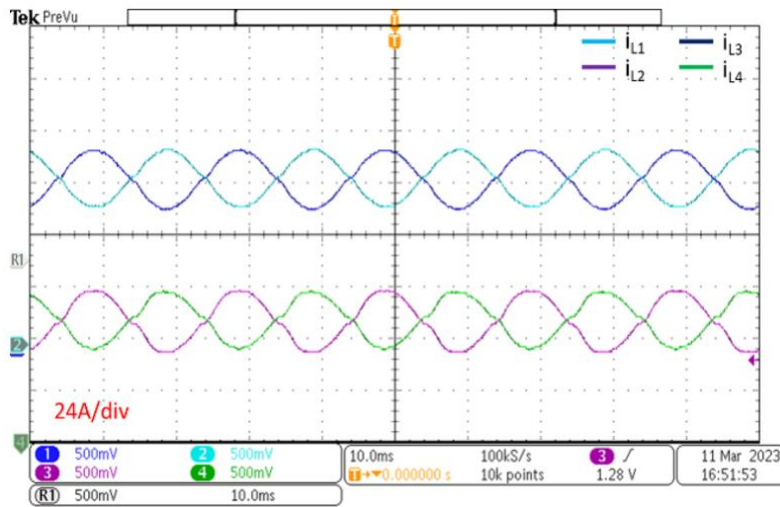


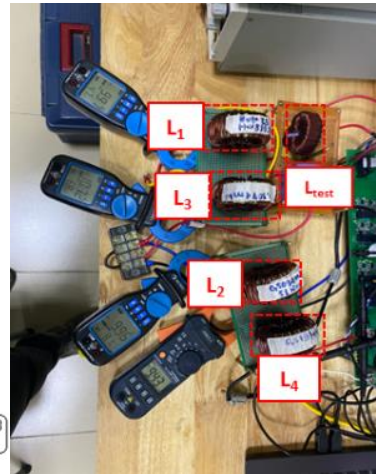
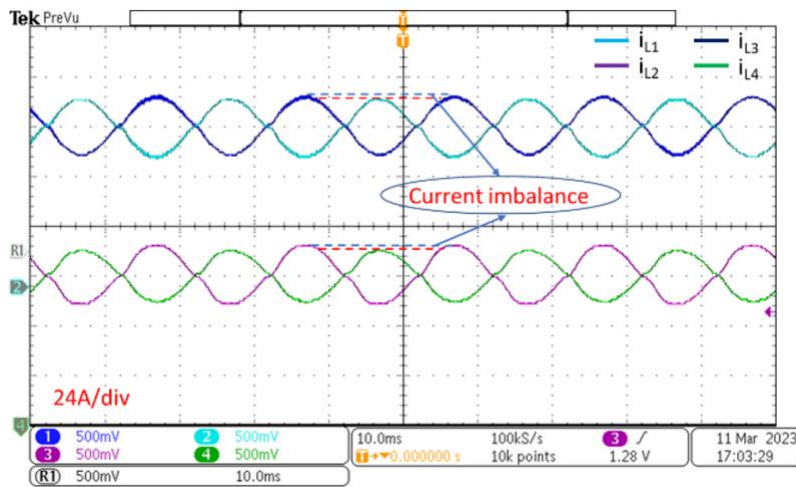
Figure 9. Experimental prototype for the IFB converter

Fig. 9 illustrates the current responses through four split inductors when the system operates with the interleaved unipolar PWM and the balanced inductors. The RMS values of inductor currents displayed on the current meter in Fig. 10a are 10.00A, 10.18A, 9.94A, and 10.00A for inductors L_1 , L_3 , L_2 , and L_4 , respectively. All four currents are equal to the reference current of 10A with the corresponding waveform in Fig. 10a.

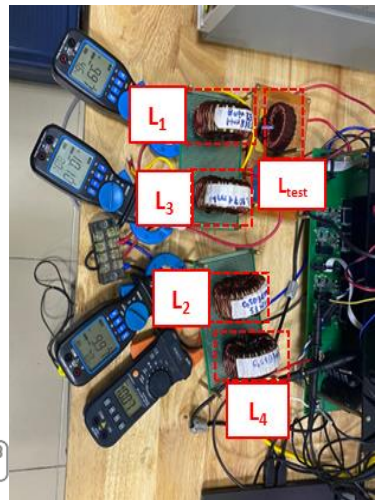
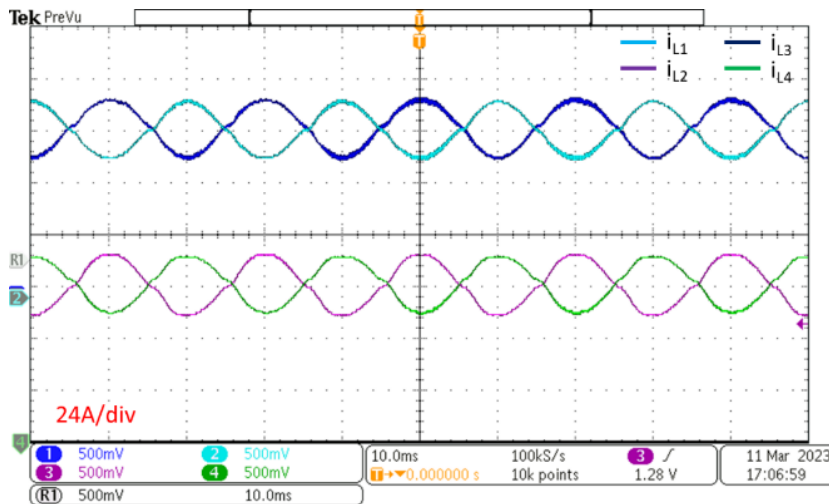
In the unbalanced case, the initial inductor L_3 is replaced by a test inductor; hence, the parameters of L_3 decrease by 30%, and the imbalance current problem occurs. The RMS values of inductor currents are recorded as 9.97A, 10.72A, 9.96A, and 9.43A for the currents through the inductors L_1 , L_3 , L_2 , and L_4 , respectively, in Fig. 10b. In this case, the difference of RMS current i_{L3} and i_{L4} is 7.2% and -5.7%, where the current waveforms in Fig.10b also show the difference between inductor currents. This error is proportional to the reference current and the mismatch value among inductors, which has to be eliminated.



a) Current responses in the balanced case



b) Current responses in unbalanced cases without the proposed method



c) Current responses in unbalanced cases with the proposed method

Figure 10. Experimental results

The linear controller is applied to handle the imbalance current problem with a specific amount of compensated modulation index for branch 3. It can be seen in Fig. 10c that the RMS value of inductor currents has returned to balance with 9.97A, 10.10A, 9.97A, and 10.07A for inductors L_1 , L_3 , L_2 , and L_4 , respectively. The rebalanced current waveforms in Fig. 10c have proven the effectiveness of the proposed method, which shows good agreement with the theoretical analysis in the previous section.

4. Conclusion

The paper proposes a control structure that solves the current imbalance issue for the IFB converter in an AC Battery application. This problem is caused by the imbalance in inductance and internal resistance of split-inductors, which deteriorates the performance of the system. Firstly, this paper theoretically analyzes the reason for the unbalanced current problem with the IB-PWM method. Secondly, a simple and effective control structure is proposed to balance the inductor current while maintaining the advantages of the converter structure along with the modulation technique. Then, various simulation scenarios are conducted to verify the problem and its proposed solution as well. Finally, the experimental prototype is carried out, and the results show good agreement with the previous theoretical analysis, which proves the effectiveness of the current imbalance elimination method for the IFB converter in particular and the parallel-operated converter in general.

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Author Contributions

Tuan Anh Do and Phuong Vu were responsible for the conceptualization, validation, resources, data curation, software development, and project administration. Quang Dich Nguyen jointly contributed to the methodology, formal analysis, investigation, original draft preparation, review and editing, visualization, supervision, and funding acquisition. All authors have read and agreed to the published version of the manuscript.

Conflict of Interest

The author(s) declared no potential conflicts of interest with respect to the research, authorship, and/or publication of this article.

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