Current Balancing Technique for Interleaved Voltage Source Inverters with Magnetically-Coupled Legs Connected in Parallel

Abstract—Connecting legs in parallel in a voltage source inverter is a way to increase the output current and thus, its rated power. The connection can be made using either coupled or uncoupled inductors and achieving an even contribution to the output current from all the legs is a crucial issue. Circulating currents produce additional losses and stress to the power devices of the converter. Therefore, they should be controlled and minimized. An efficient technique to attain such a balance when coupled inductors are used is presented in this paper. The proposed technique can also be used when the inductors are uncoupled, since it is a particular case where the coupling coefficients are zero. This technique does not include proportional-integral controllers and does not require any parameter tuning either. The exact control action needed to reach current balance is straightforwardly calculated and applied. Experimental results are shown in this paper to verify the efficiency of the proposed balancing method.

Index Terms—VSI, Parallel legs, Interleaving, Pulse-width modulation, Current sharing.

I. INTRODUCTION

IN order to raise the electric power handled by a voltage source inverter (VSI) either voltages, currents, or both can be increased. Multilevel converters are extensively used in high power systems because they can deal with higher voltages due to the fact that the voltage to be switched is shared among several semiconductors [1]–[3]. They also yield better output voltage spectra than two-level converters [4]. On the other hand, currents rather than voltages are increased in order to achieve higher power in some fields. This can be attained either with VSIs made up of legs connected in parallel (if all the legs are connected to a common DC bus) or by means of parallel connection of inverters (each one with its own DC bus) [5], [6]. In addition, inverters with legs connected in parallel are modular and because of that, their production and maintenance become less expensive. Moreover, they qualify for the implementation of fault-tolerant techniques thus offering improved overall reliability [7], [8].

Parallel-connected legs of a VSI require the use of inductors to obtain a single output voltage from several input legs [9]–[14]. The use of uncoupled inductors does not bring any size reduction in the amount of magnetic material needed but facilitates modularity and management in case of faulty legs [10]. The use of coupled inductors provides a high impedance path for limiting the common-mode circulating currents while providing a low impedance path for the output currents [11]. Through the use of coupled inductors, a converter is capable of responding faster to load transient depending on the coupling coefficient and control mechanism [12]. A theoretical study dealing with different options to connect multiple legs in

parallel by means of transformers is developed in [13]. It would be optimal if current sharing among the legs were balanced; however, there is no guaranty for this to happen unless a proper control is used. A great variety of techniques can be applied to reach a balanced distribution of current among the legs. Most of them are based on proportional-integral (PI) controllers and they usually include two control loops: one to control the output voltage and another one to regulate the current sharing [15]–[20]. Optimal regulators are used in [21], [22] in order to create robust feedback control systems, whereas a sliding control is proposed in [23], in spite of the serious drawback of leading to a variable switching frequency, albeit limited. By and large, all the considered methods provide good balancing performance on the whole. However, all of them need some parameter tuning and the balancing dynamics may not be optimal. Quick response of the balancing control is crucial to avoid long transitory overcurrents in any leg which might be destructive.

A space vector modulation (SVM) approach for converters with multiple legs in parallel is presented in [24]. In this paper, the SVM decouples the current sharing control and the converter output control. A study of the modulation is performed; however, in order to simplify the practical implementation, the current sharing is carried out with a hysteresis controller, in spite of its inherent variable switching frequency.

When connecting VSIs in parallel to create a local net, it is quite common that each inverter uses its own reference signal and its local controller. Owing to the fact that phase, frequency and amplitude have to be the same in order to minimize circulating currents, some kind of synchronization has to be included either providing communication lines among them [25], [26] or a self-synchronizing mechanism [27], [28].

The limitations in order to achieve a proportional load sharing in a conventional droop control scheme are revealed in [29], [30], and modified robust droop controllers are presented, notwithstanding the trade-off between voltage drop and sharing accuracy when the load voltage is not accurately measured. The multilayer control based upon a modified droop control presented in [31] allows for the parallel operation of inverters without intercommunications.

A method that avoids the use of intermodule inductors in three-phase parallel-connected inverters is presented in [32]. The proposed method compares the actual and the desired current value for each converter. The error signal is utilized by a PI compensator to provide the modulation index to a space vector modulator. The control method divides the switching cycle evenly between the equally rated converters. Accordingly, the intermodule inductors become redundant.

Fig. 1. Phase configuration of a VSI with n legs connected in parallel.

An analytical model for a parallel-connected inverter system is developed in [33]. The internal Thevenin impedance of inverters is identified as the sensitive element affecting current distribution. An active control strategy to tune the effective Thevenin impedance by using a resistor emulation approach is developed. Yet again, a parameter has to be adjusted properly to guarantee stability of the system.

The balancing method proposed in this paper can achieve current balance very quickly since the exact required modification of the modulation signals is calculated and applied. The method is performed without distorting the output voltages or currents and thus it does not affect any external control loop. It is designed for the case of using coupled inductors; nevertheless, since it is general, it can also be applied to the case of uncoupled inductors by defining the coupling parameter to be zero.

The rest of the paper is organized as follows. In Section II, an introduction to the connection of legs in parallel by means of inductors is made. Section III introduces the proposed current balancing method for the n -leg parallel-connected case. The method utilized to attain a multiple coupling effect by means of single transformers is presented in Section IV. In Section V, some simulation and experimental results are given to prove the effectiveness of the proposed current balancing technique. The conclusions are summarized in Section VI.

II. PARALLEL-CONNECTED LEGS

A. Phase Configuration

In order to avoid shortcircuits, VSI's legs cannot be directly connected in parallel. Inductors are the optimal passive components to achieve the following benefits; (i) limiting circulating currents among the legs and (ii) averaging voltages of several legs for each output phase. Because of the averaging, the equivalent output voltage of the phase would show more than two levels [34]. In a symmetrical configuration the number of voltage levels will be $n+1$, n being the number of legs connected in parallel.

Fig. 1 shows an example of parallel-connected legs. The n legs shown in this scheme correspond to one single phase of the converter (phase a , for instance). Multiple magnetic coupling among the inductors is assumed, although the analysis could be extrapolated to different types of coupling, like the ones presented in [13], or to magnetically uncoupled inductors [35].

Fig. 2. Phase-shift disposition of the carriers for interleaving operation mode.

B. The Interleaving Technique

The interleaving technique is applied to the system shown in Fig. 1 in order to procure an apparent switching frequency n times higher than the individual switching frequency of each leg ($f_s = n f_{sw} = n / T_{sw}$).

When operating with a carrier-based modulation strategy, this is achieved by using an n number of shifted carriers v_{carr1} , v_{carr2} , ..., v_{carrn} . Fig. 2 shows the general n -case carriers' disposition although only the first three carriers have been depicted.

III. CURRENT BALANCING METHOD

The relationship between voltages and currents in each leg of the system depicted in Fig. 1 is

$$\mathbf{V}_a = \mathbf{L} \frac{d}{dt} \mathbf{I}_a + \mathbf{V}_{a0}, \quad (1)$$

where

$$\mathbf{V}_a = \begin{bmatrix} v_{a1} \\ v_{a2} \\ \vdots \\ v_{an} \end{bmatrix}, \quad \mathbf{I}_a = \begin{bmatrix} i_{a1} \\ i_{a2} \\ \vdots \\ i_{an} \end{bmatrix}, \quad \mathbf{V}_{a0} = \begin{bmatrix} v_{a0} \\ v_{a0} \\ \vdots \\ v_{a0} \end{bmatrix}, \quad (2)$$

and

$$\mathbf{L} = \begin{bmatrix} L_1 & -M_{12} & -M_{13} & \dots & -M_{1n} \\ -M_{12} & L_2 & -M_{23} & \dots & -M_{2n} \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ -M_{1n} & -M_{2n} & -M_{3n} & \dots & L_n \end{bmatrix}. \quad (3)$$

Considering a symmetrical magnetic structure for the whole set of inductors, (1) could be stated as

$$\begin{bmatrix} v_{a1} \\ v_{a2} \\ \vdots \\ v_{an} \end{bmatrix} = \begin{bmatrix} L & -M & \dots & -M \\ -M & L & \dots & -M \\ \vdots & \vdots & \ddots & \vdots \\ -M & -M & \dots & L \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{a1} \\ i_{a2} \\ \vdots \\ i_{an} \end{bmatrix} + \begin{bmatrix} v_{a0} \\ v_{a0} \\ \vdots \\ v_{a0} \end{bmatrix}. \quad (4)$$

Adding all the single equations in (4) leads to

$$\sum_{j=1}^n v_{aj} = [L - (n-1)M] \sum_{j=1}^n \frac{di_{aj}}{dt} + n v_{a0}. \quad (5)$$

Fig. 3. Averaged equivalent leg.

Provided that the output current is made up of the phase-leg currents

$$i_a = \sum_{j=1}^n i_{aj}, \quad (6)$$

and v_{acom} being the equivalent average output voltage of the n phase-legs, i.e. the voltage that would be generated from an equivalent single leg

$$v_{acom} = \frac{1}{n} \sum_{j=1}^n v_{aj}, \quad (7)$$

(5) can be written as

$$v_{acom} - v_{a0} = \left(\frac{L}{n} - \frac{n-1}{n} M \right) \frac{di_a}{dt}. \quad (8)$$

Calling L_{eq} the equivalent output inductance of that fictitious leg

$$L_{eq} = \frac{L}{n} - \frac{n-1}{n} M, \quad (9)$$

(8) becomes

$$v_{acom} - v_{a0} = L_{eq} \frac{di_a}{dt}. \quad (10)$$

The locally-averaged operator is defined as follows

$$\bar{x}(t) = \frac{1}{T_w} \int_{t-T_w}^t x(\tau) d\tau. \quad (11)$$

If the window-width (T_w) used in this operator is defined to be the same as the converter switching period, switching frequency ripples in the voltages and currents will be completely filtered and canceled; as a consequence, those variables will become continuous. Applying the locally-averaged operator to (10), and considering that the averaged variable $\bar{v}_{acom}$ becomes the global reference voltage of the phase ($\bar{v}_{acom} = v_{aref}$), (10) becomes

$$v_{aref} - \bar{v}_{a0} = L_{eq} \frac{d\bar{i}_a}{dt} \quad (12)$$

that corresponds to the averaged equivalent leg of the whole phase that can be seen in Fig. 3.

If there were no current balancing control in the system, the voltage reference provided to each leg and the global reference of the phase would be the same, i.e. $\bar{v}_{aj} = v_{aref}$ for $j = \{1, 2, \dots, n\}$ or

$$\bar{\mathbf{V}}_a = \begin{bmatrix} \bar{v}_{a1} \\ \bar{v}_{a2} \\ \vdots \\ \bar{v}_{an} \end{bmatrix} = \begin{bmatrix} v_{aref} \\ v_{aref} \\ \vdots \\ v_{aref} \end{bmatrix}. \quad (13)$$

Fig. 4. Time sequence of the sampling process for current balancing.

In order to provide a control law for each leg current, each individual voltage is modified as follows:

$$\bar{v}_{aj} = v_{aref} + \Delta \bar{v}_{aj} \quad \text{for } j = \{1, 2, \dots, n\}, \quad (14)$$

where $\Delta \bar{v}_{aj}$ are the control signals.

Applying the locally-averaged operator to (1) allows it to be written as

$$\bar{\mathbf{V}}_a = \mathbf{L} \frac{d}{dt} \bar{\mathbf{I}}_a + \bar{\mathbf{V}}_{a0}. \quad (15)$$

When including the effect of the control variables ($\Delta \bar{v}_{aj}$) into (15), this becomes

$$\bar{\mathbf{V}}_a + \Delta \bar{\mathbf{V}}_a = \mathbf{L} \frac{d}{dt} (\bar{\mathbf{I}}_a + \Delta \bar{\mathbf{I}}_a) + \bar{\mathbf{V}}_{a0} + \Delta \bar{\mathbf{V}}_{a0}. \quad (16)$$

The Δ terms that derive from such control variables can be isolated by subtracting (15) from (16), that leads to

$$\Delta \bar{\mathbf{V}}_a = \mathbf{L} \frac{d}{dt} \Delta \bar{\mathbf{I}}_a + \Delta \bar{\mathbf{V}}_{a0}, \quad (17)$$

where

$$\Delta \bar{\mathbf{V}}_a = \begin{bmatrix} \Delta \bar{v}_{a1} \\ \Delta \bar{v}_{a2} \\ \vdots \\ \Delta \bar{v}_{an} \end{bmatrix}, \quad (18)$$

$$\Delta \bar{\mathbf{I}}_a = \begin{bmatrix} \bar{i}_{a1} - \bar{i}_a/n \\ \bar{i}_{a2} - \bar{i}_a/n \\ \vdots \\ \bar{i}_{an} - \bar{i}_a/n \end{bmatrix}, \quad (19)$$

and

$$\Delta \bar{\mathbf{V}}_{a0} = \begin{bmatrix} \Delta \bar{v}_{a0} \\ \Delta \bar{v}_{a0} \\ \vdots \\ \Delta \bar{v}_{a0} \end{bmatrix}. \quad (20)$$

Considering the control variables introduced in (14), (7) can be stated as

$$\begin{aligned} \bar{v}_{acom} &= \frac{1}{n} \sum_{j=1}^n (v_{aref} + \Delta \bar{v}_{aj}) \\ &= v_{aref} + \frac{1}{n} \sum_{j=1}^n \Delta \bar{v}_{aj}. \end{aligned} \quad (21)$$

Bearing in mind that the control variables should not affect the output voltage generated by the leg ($\bar{v}_{acom} = v_{aref}$), one can conclude that the control voltages have to meet the following condition:

$$\sum_{j=1}^n \Delta \bar{v}_{aj} = 0. \quad (22)$$

Fig. 5. Current balancing control diagram.

Since $\bar{v}_{acom}$ becomes unaltered if restriction (22) is applied, from (12) and its equivalent circuit in Fig. 3, it can be inferred that $\bar{i}_a$ and $\bar{v}_{a0}$ will not be affected by the control variables either. As a consequence, $\Delta \bar{V}_{a0} = 0$ and therefore (17) becomes

$$\Delta \bar{V}_a = L \frac{d}{dt} \Delta \bar{I}_a. \quad (23)$$

When the phase current is equally shared among the phase legs, i.e. $\bar{i}_{aj} = \bar{i}_a/n$ for $j = \{1, 2, \dots, n\}$, then $\Delta \bar{I}_a = 0$. If there were a current imbalance ($\Delta \bar{I}_a \neq 0$) at the instant kT_s , the discrete representation of (23)

$$\Delta \bar{V}_a(k) = \frac{1}{T_s} L [\Delta \bar{I}_a(k+1) - \Delta \bar{I}_a(k)] \quad (24)$$

could be used to compute the control voltages necessary to achieve a fairly shared current in the next sampling period by imposing the condition $\Delta \bar{I}_a(k+1) = 0$, that would render

$$\Delta \bar{V}_a(k) = -\frac{1}{T_s} L \Delta \bar{I}_a(k). \quad (25)$$

A timing diagram for this on-line process can be seen in Fig. 4. Fig. 5 shows the control diagram of the current balancing technique for phase a . The local averager block in Fig. 5 is based on a moving average filter (MAF) with a window width equal to the carrier period, i.e. the switching period (T_{sw}). As a consequence, the current ripple of the circulating currents is completely removed. On the other hand, the sampling period of the converter controller, which includes the current balancing control, is at the apparent switching period, i.e. T_s , as shown in Fig. 4. This period is smaller than T_{sw} ($T_s = T_{sw}/n$) and therefore, the dynamic of the converter becomes faster. Ideally, all the reference signals should be updated at any sampling period (T_s). However, in some practical implementations, updating the reference signal may not be feasible until the corresponding carrier signal of the specific leg completes a PWM cycle. In those cases, only one reference signal is updated at a time, just before its carrier signal cycle starts.

It can be seen in Fig. 5 that n current sensors are used to provide the information needed. The current of each leg (i_{aj})

is sensed and then, locally averaged ($\bar{i}_{aj}$) with a window width of T_{sw} . Afterwards, those averaged currents are synchronously sampled every apparent switching period T_s (see Fig. 2). The values of the Δv_{aj} variables are calculated according to (25) and, after checking that they will not cause over-modulation, applied to the specific modulation signal for each leg. In case that the Δv_{aj} values required to achieve the current balance were such that some of the them might cause over-modulation in their legs, i.e. $|v_{aref} + \Delta \bar{v}_{aj}| > 1$, the biggest one would have to be trimmed down to its maximum possible value and the rest of them rescaled accordingly in order to make sure that condition (22) is always met. That is to avoid distortion in the overall output voltage phase, even if some of the Δv_{aj} values had to be limited because of the aforementioned restriction. This task is performed by the "Overmodulation Preventer" block that can be seen in Fig. 5.

IV. IMPLEMENTATION OF MULTI-COUPLED INDUCTORS

The implementation of a multi-coupled set of inductors by means of a common magnetic core with several windings implies some assembly difficulties if a total symmetry is to be achieved. In other words, it would be difficult to attain that all the M_{ij} terms (for $i, j = \{1, 2, \dots, n\}$ and $i \neq j$) included in (3) be of the same value. An equivalent multi-coupling effect can be achieved by using only one-to-one coupled inductors (transformers) as described below.

Assuming that all the magnetic cores of the transformers have the same reluctance ($\mathcal{R}$) and calling i_p and i_s the currents in their primary and secondary windings respectively, the magnetic flux in each core is

$$\Phi = \frac{N_1 i_p - N_2 i_s}{\mathcal{R}} \quad (26)$$

where N_1 and N_2 are the number of turns of each winding. The inductance parameters in each transformer are

$$L_1 = N_1 \frac{d\Phi}{di_p} = \frac{N_1^2}{\mathcal{R}}, \quad (27)$$

$$L_2 = -N_2 \frac{d\Phi}{di_s} = \frac{N_2^2}{\mathcal{R}}, \quad (28)$$

Fig. 6. Cyclic cascade connection.

Fig. 7. Transformers between two generic legs in a combinatorial cascade connection: (a) original configuration and (b) equivalent configuration.

$$M = N_1 \frac{d\Phi}{di_s} = -N_2 \frac{d\Phi}{di_p} = \frac{N_1 N_2}{\mathfrak{R}}. \quad (29)$$

In a cyclic cascade connection [13], two transformers are used in every leg, as shown in Fig. 6. Every transformer links the current in one leg to the current in the next one, except for the last one that is linked to the first one. In this configuration, (3) becomes

$$\mathbf{L} = \begin{bmatrix} L_1 + L_2 & -M_{12} & 0 & \dots & -M_{12} \\ -M_{12} & L_1 + L_2 & -M_{12} & \dots & 0 \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ -M_{12} & 0 & \dots & -M_{12} & L_1 + L_2 \end{bmatrix}. \quad (30)$$

Comparing (30) to (3) and (4), it can be observed that:

$$L = L_1 + L_2 = \frac{N_1^2 + N_2^2}{\mathfrak{R}}, \quad (31)$$

and the term M is either zero or

$$M = M_{12} = \frac{N_1 N_2}{\mathfrak{R}}. \quad (32)$$

The M/L ratio can be adjusted by means of the number of turns in each winding (N_1 and N_2) according to the expression

$$\frac{M}{L} = \frac{N_1 N_2}{N_1^2 + N_2^2} = \frac{1}{\frac{N_1}{N_2} + \frac{N_2}{N_1}}, \quad (33)$$

and its maximum value is

$$\frac{M}{L} = \frac{1}{2}. \quad (34)$$

In the cycling cascade connection, some mutually coupling coefficients of the matrix are zero. The combinatorial cascade connection [13] can achieve that all the mutually coupling

Fig. 8. Generic leg in a combinatorial cascade connection.

Fig. 9. (a) Three multi-coupled inductors (b) cyclic/combinatorial cascade connection.

coefficients be equal and different from zero. In such a connection, the current in each leg should be magnetically coupled to the current in each of the other $n - 1$ legs. In order to ensure a complete magnetic symmetry that would require the use of $2(n - 1)$ transformers; $n - 1$ to be connected through their primary winding to the other legs, and $n - 1$ to be connected through their secondary winding. Fig. 7(a) shows the transformers between two legs in the combinatorial cascade connection, where N_1 and N_2 are the turns in the primary and secondary windings of the $2(n - 1)$ transformers, L_1 and L_2 their respective inductances, and M_{12} the mutually coupling coefficient. This theoretical set of $2(n - 1)$ transformers, is equivalent to a set of $(n - 1)$ transformers with a number of turns N in each of their windings and their magnetic parameters would be $L_{eq} = L_1 + L_2$ and $M_{eq} = 2M_{12}$. An equivalent transformer between two generic legs is shown in Fig. 7(b). Fig. 8 presents the magnetic links among the inductors connected in series in a generic leg and the coupled ones connected in the rest of the legs.

In such a combinatorial cascade configuration (3) becomes

$$\mathbf{L} = \begin{bmatrix} (n - 1)L_{eq} & -M_{eq} & \dots & -M_{eq} \\ -M_{eq} & (n - 1)L_{eq} & \dots & -M_{eq} \\ \vdots & \vdots & \ddots & \vdots \\ -M_{eq} & -M_{eq} & \dots & (n - 1)L_{eq} \end{bmatrix}. \quad (35)$$

Comparing (35) to (3) and (4), it can be observed that $L = (n - 1)L_{eq}$ and $M = M_{eq}$. Therefore, the ratio between these parameters is:

$$\frac{M}{L} = \frac{M_{eq}}{(n - 1)L_{eq}} = \frac{2}{(n - 1) \left(\frac{N_1}{N_2} + \frac{N_2}{N_1} \right)}. \quad (36)$$

The disposition of the transformers in a four-leg cyclic cascade coupling is shown in Fig. 10(a), whereas the combinatorial one is shown in Fig. 10(b). For the three-coupled-leg case (Fig. 9), if $N_1 = N_2$ both the cyclic cascade and the combinatorial cascade structures are equivalent.

V. SIMULATION AND EXPERIMENTAL RESULTS

The proposed balancing technique has firstly been simulated on a Matlab-Simulink model. Then, it has been implemented

Fig. 10. Coupling four voltage sources with transformers: (a) Cyclic cascade coupling, and (b) combinatorial cascade coupling.

TABLE I
SPECIFICATIONS OF THE LABORATORY PROTOTYPE

Symbol	Parameter	Value
V_{dc}	DC bus voltage	60 V
f	Reference signal frequency	50 Hz
m_a	Modulation index	0.8
n	Number of legs	2/3
f_s	Carrier/switching frequency	2 kHz
f_{sw}	Apparent switching frequency	4/6 kHz
R_L	Load resistance	3.3 Ω
L	Self inductance	8.8 mH
M	Coupling inductance	2.8 mH

on a dSpace DS1103 PPC Controller board and verified on a single-phase laboratory prototype whose schemes are depicted in Fig. 11. The single-phase converter has two/three legs connected in parallel and drives a resistive load connected between the output ("a") and the dc neutral point ("0"). The main parameters are shown in Table I.

In order to achieve a multi-coupling effect among the inductors (Fig. 9(a)) by using only one-to-one coupled inductors, the cyclic cascade connection shown in Fig. 9(b) has been used.

An occasional instantaneous disturbance Δv_a is applied to the modulation of the reference signals in the captures in Fig. 12 to Fig. 15. It can be observed that when the balance controller is activated such distorting source is not significantly affecting the output currents. Comparing the current waveforms without balancing control (Figs. 12 and 14) to the ones with balancing control (Figs. 13 and 15) it is noteworthy that the currents are fairly shared when the controller is on.

Permanent current imbalances can happen on account of different voltage drops across the power devices when they are in the on state. Such drifts might produce different dc offsets among the output voltages of the legs. As a result, provided that inductors cannot stop dc circulating currents, the resistances among the legs are the ultimate elements capable of limiting this type of circulating currents. Since the internal resistor values of the inductances are very small, even small differences among the voltage drops across the transistors can cause large current imbalances. It should be remarked that these kinds of permanent imbalances are much

Fig. 11. Laboratory prototype schemes with (a) two and, (b) three legs in parallel.

Fig. 12. Two legs connected in parallel. Disturbance effect without balancing control: (a) simulation, and (b) experimental results.

Fig. 13. Two legs connected in parallel. Disturbance arresting with balancing control: (a) simulation, and (b) experimental results.

Fig. 14. Three legs connected in parallel. Disturbance effect without balancing control: (a) simulation, and (b) experimental results.

more dangerous for the legs than those produced by transitory processes.

Fig. 16(a) shows a VSI with two legs connected in parallel where the lower switch $\bar{s}_{a1}$ has an additional voltage drop (ΔV_{dc}) compared to the other switches. This will produce a dc voltage component in the leg-to-leg voltage. The equivalent circuit for the dc components is shown in Fig. 16(b). In the steady state, the dc circulating current would be:

$$I_{dc} = \frac{\Delta V_{dc}}{2R}. \quad (37)$$

In order to emulate the case of different voltage drops in the transistors of the legs, a 1-V battery has been connected in series with the lower transistor of one leg. This creates a permanent current imbalance. Fig. 17 shows current waveforms from the system starting with such current imbalance. When the control is activated, the currents are balanced almost instantaneously as the proposed controller can compensate for the different dc output voltages produced in the legs. In real high power systems, where resistors associated to huge power inductors are very small, the dc component of the circulating current can be significant.

One can observe how the output current has practically no distortion due to compensation. This is because the controller

only produces differential voltage for current compensation, but it does not change the global output voltage generated by the phase.

VI. CONCLUSION

Achieving evenly shared currents among the parallel-connected legs of a power inverter is a remarkable challenge. The control technique presented in this paper is capable of achieving a fair current distribution with very fast dynamics. It is based on calculating the exact control actuation needed for current balance; therefore, no PI controller is required. Because of its generic formulation, the proposed balancing technique can be applied to converters with any number of legs connected in parallel. Provided that the use of uncoupled inductors is only a particular case—where all the M coefficients in (4) are zero—the balancing method can be applied regardless of whether the inductors are magnetically coupled or not.

Experimental results are provided to validate the behavior of the proposed compensator. The results show that currents are quickly balanced no matter what the reason for the imbalance is. Permanent current imbalances are the most dangerous and can be provoked by slight differences in the voltage drops across the power semiconductors.

(a)

(b)

Fig. 15. Three legs connected in parallel. Disturbance arresting with balancing control: (a) simulation, and (b) experimental results.

Fig. 16. Example of two legs connected in parallel where one switch $\bar{s}_{a1}$ has a larger voltage drop: (a) circuit diagram and (b) equivalent leg-to-leg circuit for the dc components.

Motor drives and grid-connected converters are some application fields in which VSIs with legs connected in parallel are applied, especially for high power systems such as those used in wind generation [36], [37]. The main benefit is that the

(a)

(b)

Fig. 17. Current imbalance caused by a small dc voltage difference for the 2-leg case: (a) simulation, and (b) experimental results.

low-voltage regulations have to be met instead of the highly demanding medium-voltage regulations. Furthermore, for very high power systems, multilevel converters with parallel legs can be considered.

REFERENCES

- [1] J. Rodriguez, L. Franquelo, S. Kouro, J. Leon, R. Portillo, M. Prats, and M. Perez, "Multilevel converters: an enabling technology for high-power applications," *IEEE Proc.*, vol. 97, no. 11, pp. 1786–1817, Nov. 2009.
- [2] S. Kouro, M. Malinowski, K. Gopakumar, J. Pou, L. G. Franquelo, B. Wu, J. Rodriguez, M. A. Perez, and J. I. Leon, "Recent advances and industrial applications of multilevel converters," *IEEE Trans. Ind. Electron.*, vol. 57, no. 8, pp. 2553–2580, Aug. 2010.
- [3] D. Floricau, E. Floricau, and G. Gateau, "New multilevel converters with coupled inductors: properties and control," *IEEE Trans. Ind. Electron.*, vol. 58, no. 12, pp. 5344–5351, Dec. 2011.
- [4] D. G. Holmes and B. P. McGrath, "Opportunities for harmonic cancellation with carrier-based PWM for a two-level and multilevel cascaded inverters," *IEEE Trans. Ind. Appl.*, vol. 37, no. 2, pp. 574–582, Mar./Apr. 2001.
- [5] T. Itkonen, J. Luukko, and R. Pollanen, "Analysis of current characteristics of parallel three-phase voltage source inverters," in *Proc. Eur. Conf. Power Electron. Appl. (EPE 2009)*, Barcelona, 8–10 Sep., pp. 1–10.
- [6] J. S. Siva Prasad and G. Narayanan, "Minimization of grid current distortion in parallel-connected converters through carrier interleaving," *IEEE Trans. Ind. Electron.*, vol. 61, no. 1, pp. 76–91, Jan. 2014.
- [7] Y. J. Ko, K. B. Lee, D. C. Lee, and J. M. Kim, "Fault diagnosis of three-parallel voltage-source converter for a high-power wind turbine," *IET Power Electron.*, vol. 5, no. 7, pp. 1058–1067, Aug. 2012.

- [8] X. Yu and A. M. Khambadkone, "Reliability analysis and cost optimization of parallel-inverter system," *IEEE Trans. Ind. Electron.*, vol. 59, no. 10, pp. 3881–3889, Oct. 2012.
- [9] E. Laboure, A. Cuniere, T. A. Meynard, F. Forest, and E. Sarraute, "A theoretical approach to intercell transformers, application to interleaved converters," *IEEE Trans. Power Electron.*, vol. 23, no. 1, pp. 464–474, Jan. 2008.
- [10] F. Forest, E. Laboure, T. A. Meynard, and V. Smet, "Design and comparison of inductors and intercell transformers for filtering of PWM inverter output," *IEEE Trans. Power Electron.*, vol. 24, no. 3, pp. 812–821, Mar. 2009.
- [11] J. Ewanchuk and J. Salmon, "Three-limb coupled inductor operation for paralleled multi-level three-phase voltage sourced inverters," *IEEE Trans. Ind. Electron.*, vol. 60, no. 5, pp. 1979–1988, 2013.
- [12] G. Zhu, B. McDonald, and K. Wang, "Modeling and analysis of coupled inductors in power converters," *IEEE Trans. Power Electron.*, vol. 26, no. 5, pp. 1355–1363, May 2011.
- [13] I. G. Park and S. I. Kim, "Modeling and analysis of multi-interphase transformers for connecting power converters in parallel," in *Proc. IEEE Power Electron. Spec. Conf. (PESC 1997)*, vol. 2, St. Louis, MO, 22–27 Jun., pp. 1164–1170.
- [14] B. Ge, X. Lu, X. Yu, M. Zhang, and F. Z. Peng, "Multiphase-leg coupling current balancer for parallel operation of multiple mw power modules," *IEEE Trans. Ind. Electron.*, vol. 61, no. 3, pp. 1147–1157, Mar. 2014.
- [15] M. Xie, Y. Li, K. Cai, P. Wang, and X. Sheng, "A novel controller for parallel operation of inverters based on decomposing of output current," in *Proc. IEEE IAS Annu. Meeting Ind. Appl. Conf. (IAS 2005)*, vol. 3, 2–6 Oct., pp. 1671–1676.
- [16] C. Zhang, F. Zu, Q. Zhao, W. Wu, and G. Chen, "An interactive following current-sharing control strategy for paralleled inverters in full digital," in *Proc. IEEE IAS Annu. Meeting Ind. Appl. Conf. (IAS 2005)*, vol. 2, pp. 1080–1083.
- [17] H. M. Hsieh, T. F. Wu, Y. E. Wu, H. S. Nien, and Y. K. Chen, "A compensation strategy for parallel inverters to achieve precise weighting current distribution," in *Proc. IEEE IAS Annu. Meeting Ind. Appl. Conf. (IAS 2005)*, vol. 2, 2–6 Oct., pp. 954–960.
- [18] M. Yu, Y. Kang, Y. Zhang, M. Yin, S. Duan, H. Shan, and G. Chen, "A novel decoupled current-sharing scheme based on circulating-impedance in parallel multi-inverter system," in *Proc. IEEE Ind. Electron. Conf. (IECON 2007)*, Taipei, 5–8 Nov., pp. 1668–1672.
- [19] I. W. Jaskulski, H. Pinheiro, and L. Mariotto, "Multi-leg voltage source converter for grid connected wind turbines," in *Proc. Int. Conf. Clean Electr. Power (ICCEP '07)*, Capri, 21–23 May, pp. 229–235.
- [20] C. Wen, J. Li, X. Zhu, and H. Xu, "Research on circulation of parallel three-phase converters in mw wind power system," in *Proc. Work. Power Electron. Intell. Transport. Syst. (PEITS '08)*, Guangzhou, 2–3 Aug., pp. 349–354.
- [21] Y. K. Chen, T. F. Wu, Y. E. Wu, and C. P. Ku, "A current-sharing control strategy for paralleled multi-inverter systems using microprocessor-based robust control," in *Proc. IEEE Reg. 10 Int. Conf. Electr. Electron. Technol. (TENCON 2001)*, vol. 2, 19–22 Aug., pp. 647–653.
- [22] N. Talebi, M. A. Sadrnia, and S. M. R. Rafiei, "Current and voltage control of paralleled multi-module inverter systems," in *Proc. Mediterr. Conf. Control Autom. (MED'09)*, Thessaloniki, 24–26 Jun., pp. 1498–1503.
- [23] R. Ramos, D. Biel, E. Fossas, and F. Guinjoan, "Sliding-mode control design applied to parallel-connected modular inverters through fpga-based implementation," *IET Control Theory Appl.*, vol. 3, no. 12, pp. 1611–1624, Dec. 2009.
- [24] I. W. Jaskulski, I. J. Gabe, J. P. da Costa, M. Stefanello, and H. Pinheiro, "Space vector modulation extended to voltage source converters with multiple legs in parallel," in *Proc. Eur. Conf. Power Electron. Appl. (EPE 2007)*, Aalborg, 2–5 Sep., pp. 1–10.
- [25] Y. Zhang and H. Ma, "Theoretical and experimental investigation of networked control for parallel operation of inverters," *IEEE Trans. Ind. Electron.*, vol. 59, no. 4, pp. 1961–1970, Apr. 2012.
- [26] —, "Analysis of networked control schemes and data-processing method for parallel inverters," *IEEE Trans. Ind. Electron.*, vol. 61, no. 4, pp. 1834–1844, Apr. 2014.
- [27] T. Lazzarin, G. Bauer, and I. Barbi, "A control strategy for parallel operation of single-phase voltage source inverters: analysis, design and experimental results," *IEEE Trans. Ind. Electron.*, vol. 60, no. 6, pp. 2194–2204, Jun. 2013.
- [28] K. Zhi-zhong, Z. Chun-jiang, Z. Bo, and W. Xiao-huan, "Analysis of the conflict between close-loop control and the current-sharing of ac parallel inverters and parallel control strategy," in *Proc. IEEE Int. Power Electron. Motion Control Conf. (IPEMC'09)*, Wuhan, 17–20 May, pp. 368–372.
- [29] W. Yao, M. Chen, J. Matas, J. M. Guerrero, and Z.-M. Qian, "Design and analysis of the droop control method for parallel inverters considering the impact of the complex impedance on the power sharing," *IEEE Trans. Ind. Electron.*, vol. 58, no. 2, pp. 576–588, Feb. 2011.
- [30] Q. C. Zhong, "Robust droop controller for accurate proportional load sharing among inverters operated in parallel," *IEEE Trans. Ind. Electron.*, vol. 60, no. 4, pp. 1281–1290, Apr. 2013.
- [31] M. Hua, H. Hu, Y. Xing, and J. M. Guerrero, "Multilayer control for inverters in parallel operation without intercommunications," *IEEE Trans. Power Electron.*, vol. 27, no. 8, pp. 3651–3663, Aug. 2012.
- [32] B. M. H. Jassim, B. Zahawi, and D. Atkinson, "Simple control method for parallel connected three-phase pwm converters," in *Proc. IET Int. Conf. Power Electron. Mach. Drives (PEMD 2012)*, Bristol, 27–29 Mar., pp. 1–5.
- [33] B. Shi and G. Venkataramanan, "Parallel operation of voltage source inverters with minimal intermodule reactors," in *Proc. IEEE IAS Annu. Meeting (IAS 2004)*, vol. 1, 3–7 Oct., pp. 156–162.
- [34] B. Cougo, G. Gateau, T. Meynard, M. Bobrowska-Rafal, and M. Cousineau, "Pd modulation scheme for three-phase parallel multi-level inverters," *IEEE Trans. Ind. Electron.*, vol. 59, no. 2, pp. 690–700, Feb. 2012.
- [35] J. Pou, J. Zaragoza, G. Capella, I. Gabiola, S. Ceballos, and E. Robles, "Current balancing strategy for interleaved voltage source inverters," *EPE Journal*, vol. 21, no. 1, pp. 29–34, Jun. 2011.
- [36] M. Hua, H. Hu, Y. Xing, and Z. He, "Distributed control for ac motor drive inverters in parallel operation," *IEEE Trans. Ind. Electron.*, vol. 58, no. 12, pp. 5361–5370, Dec. 2011.
- [37] R. Li and D. Xu, "Parallel operation of full power converters in permanent-magnet direct-drive wind power generation system," *IEEE Trans. Ind. Electron.*, vol. 60, no. 4, pp. 1619–1629, Apr. 2013.