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Wei, Baoze; Guerrero, Josep M.; Quintero, Juan Carlos Vasquez; Guo, Xiaoqiang

Published in:
IEEE Transactions on Industry Applications

Publication date:
2017

Document Version
Early version, also known as pre-print

Link to publication from Aalborg University

Citation for published version (APA):

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A Circulating-Current Suppression Method for Parallel Connected Voltage Source Inverters (VSI) with Common DC and AC Buses

Baoze Wei, Josep M. Guerrero
Juan C. Vásquez
Student Member, Fellow, Senior Member, IEEE
Aalborg University
101 Pontoppidanstræde
Aalborg, 9220, Denmark
bao@et.aau.dk, joz@et.aau.dk, juq@et.aau.dk

Xiaoqiang Guo
Senior Member, IEEE
Yanshan University
438 Hebei Street
Qinhuangdao, 066004, China
gxq@ysu.edu.cn

Abstract—This paper presents a theoretical study with experimental validation of a circulating-current suppression method for parallel operation of three-phase voltage source inverters (VSI), which may be suitable for modular parallel uninterruptible power supply systems or hybrid AC/DC microgrid applications. The basic concept of the proposed circulating-current suppression method is to modify the original current references by using the current difference among the parallel inverters. In the proposed approach, both of cross circulating-current and zero-sequence circulating-current are considered, and added into the conventional droop plus virtual impedance control. In the control architecture, the reference voltages of the inverters are generated by the primary control loop which consists of a droop control and a virtual impedance. The secondary control is used to compensate the voltage drop on the virtual impedance. Further, a circulating-current control loop is added to improve the average current-sharing performance among parallel VSIs. Experimental results are presented to show the effectiveness of the proposed control method to suppress both of the cross and zero-sequence circulating-currents.

Keywords—voltage source inverter; parallel connected; cross circulating-current; zero-sequence; common DC and AC buses

I. INTRODUCTION

The inverters are commonly used in power converter based applications, such as the uninterruptible power systems (UPS), distributed generation (DG) systems, or microgrids [1]–[5]. In case of high power demand, it is difficult to deliver large amount of power with a single inverter since that the power rating of the switching devices is often limited or constrained by technical and/or economic considerations [6]. In order to achieve high power level of a converter system, but without adversely affect the system dynamics and stability [27]. Consequently, it is necessary to develop effective methods to suppress circulating-currents, thus realizing precise average current-sharing among the parallel inverters.

Another way of suppressing the circulating-current is to use proper control methods [22], [23]. A traditional solution to achieve average current sharing is the frequency and voltage droop control method, which has the feature of wireless control among parallel inverters [24]. However, the performance of the droop method is particularly sensitive to the output impedance of the parallel units [25]. To overcome this problem, by using a virtual impedance loop, the output impedance of the inverters can be modified to acquire lower circulating-current with better current sharing performance [26]. The output impedance of the inverter system is related with the power plant. However, in a practical paralleled inverters system, it is not easy to get accurate hardware parameters, and some of them will drift under different operating conditions, such as temperature, humidity, and so on. Thus it is difficult to design properly the virtual impedance loop, which if poorly designed or implemented, it may introduce current distortion and may adversely affect the system dynamics and stability [27].

In order to solve the problems of the abovementioned circulating-current suppression strategies, this paper proposes a control method which is inserted into the conventional droop plus virtual impedance control to improve the current sharing
which will flow from one inverter to another through the common AC and DC buses [8], [17]. This problem is particularly important for applications like AC-DC hybrid microgrids and modular parallel UPS systems.

Fig. 2 depicts the circulating-current phenomena. On the one hand, Fig. 2(a) shows the possible zero-sequence circulating-current paths. On the other hand, Fig. 2(b) shows the possible cross circulating-current paths based on different switching states. The simplified equivalent circuit of the two parallel connected inverters system is shown in Fig. 3.

According to [31], the cross circulating-current \( I_{\text{cir}} \) can be defined as

\[
I_{\text{cir}} = \frac{(I_a - I_b)}{2}
\]  

being \( I_a \) and \( I_b \) are the output currents of the parallel inverters. Considering the output impedances of the parallel inverters \( Z_1 \) and \( Z_2 \), the \( I_{\text{cir}} \) can be calculated as

\[
I_{\text{cir}} = \frac{1}{2} \left( \frac{E_1}{Z_1} - \frac{E_2}{Z_2} \right)
\]  

being \( E_1 \) and \( E_2 \) are the output voltages of the two inverters. If assuming that the output impedances are equal to each other, \( Z_1 = Z_2 = Z \), then the cross circulating-current can be calculated as

\[
I_{\text{cir}} = \frac{(E_1 - E_2)}{2Z}
\]  

But as discussed in Section I, in a practical system, it is difficult to guarantee that the output impedances of the inverters are equal to each other because of different parameters of the power plants or working conditions. By using a virtual impedance loop, the cross circulating-current can be calculated as

\[
I_{\text{cir}} = \frac{1}{2} \left( \frac{E_1}{Z_1 + Z_{\text{vir1}}} - \frac{E_2}{Z_2 + Z_{\text{vir2}}} \right)
\]

being \( Z_{\text{vir1}} \) and \( Z_{\text{vir2}} \) are the virtual impedances of the two inverters respectively. The zero-sequence current \( I_z \) can be calculated as

\[
I_z = \frac{(I_a + I_b + I_c)}{3}
\]

where \( I_a \), \( I_b \) and \( I_c \) are the three phase output currents of the inverter [8], [20], then the zero-sequence circulating-current among two inverters can be obtained as following

\[
I_{\text{seq}} = \frac{(I_{\text{a1}} - I_{\text{a2}})}{2}
\]
B. Proposed Control Method

Fig. 4 shows the control scheme for one of the parallel units of the two parallel connected inverters system. The other one inverter will use the same control principle. The zero-sequence current of the inverter can be easily calculated by using the abc/dqz transformation of the inductor currents. Equation (7) is used to complete the transformation from stationary coordinates to synchronous rotating coordinates. Then with (6), the zero-sequence circulating-current among the inverters can be calculated.

The control scheme contains the conventional droop plus virtual impedance, secondary control, the proposed circulating-current control loops, and the basic voltage/current control loops. In which, the droop control is responsible to generate the reference voltages for the voltage control loop, the virtual impedance is applied to obtain a similar output impedances of the parallel inverters, and the secondary control is added into the control scheme to recover the voltage drop on the virtual impedances. In Fig. 4, \( R_{V} \) and \( L_{V} \) are the virtual resistor and virtual inductor respectively. The basic concept of the proposed circulating-current control is to appropriately revise the reference currents which are generated by the voltage control loop. The original current references are added with the current errors among the two parallel inverters, including d-axis, q-axis and zero-axis currents. Then, new reference currents can be generated with this compensation. Finally, the switching signals are generated by using the conventional Sinusoidal Pulse Width Modulation (SPWM) method which is commonly used for the three-phase two level inverter based applications [32], [33].

\[
\begin{bmatrix}
  i_{d} \\
  i_{q} \\
  i_{z}
\end{bmatrix} =
\begin{bmatrix}
  \cos \theta & \cos(\theta - \frac{2\pi}{3}) & \cos(\theta + \frac{2\pi}{3}) \\
  -\sin \theta & -\sin(\theta - \frac{2\pi}{3}) & -\sin(\theta + \frac{2\pi}{3}) \\
  \frac{1}{2} & \frac{1}{2} & \frac{1}{2}
\end{bmatrix}
\begin{bmatrix}
  i_{d} \\
  i_{q} \\
  i_{z}
\end{bmatrix}
\]

The detailed process of generating new current references with the proposed circulating current control loops is shown as follows (8) – (19), including the d-axis, q-axis and zero-axis.

\[
\begin{align*}
\left[0 - (I_{d1} - I_{d2})/2\right]G_{cd1} &= I_{de1} \\
\left[0 - (I_{q1} - I_{q2})/2\right]G_{cq1} &= I_{qe1} \\
\left[0 - (I_{d1} - I_{d2})/2\right]G_{cd2} &= I_{de2} \\
\left[0 - (I_{q1} - I_{q2})/2\right]G_{cq2} &= I_{qe2} \\
\left[0 - (I_{d1} - I_{d2})/2\right]G_{cz1} &= I_{ze1} \\
\left[0 - (I_{q1} - I_{q2})/2\right]G_{cq2} &= I_{ze2} \\
I_{de1} + I_{de2} &= I_{de}^* \\
I_{qe1} + I_{qe2} &= I_{qe}^* \\
I_{de1} + I_{de2} &= I_{de}^* \\
I_{qe1} + I_{qe2} &= I_{qe}^* \\
I_{ze1} + I_{ze2} &= I_{ze}^* \\
I_{ze1} + I_{ze2} &= I_{ze}^*
\end{align*}
\]

Taking the generation of d-axis reference current \( I_{de}^* \) for example, then \( I_{de1}^* \) and \( I_{de2}^* \) will use the same principle (x=1, 2). After the abc/dqz transformation of the currents \( I_{d1} \) and \( I_{d2} \) for VSI\(_1\) and VSI\(_2\), one can obtain the d-axis currents \( I_{dn1} \) and \( I_{dn2} \). With (8), the d-axis compensation current \( I_{de1} \) for VSI\(_1\) is obtained. Note that this calculation is completed in the controller of VSI\(_1\), so that \( I_{dn} \) is in the position of minuend in (8). In (8), the difference between \( I_{dn1} \) and \( I_{dn2} \) will be compared with 0. The purpose of doing this is to decide the compensation direction, increasing or decreasing the reference currents. Another consideration is that “0” is like a reference value because the compensation currents should be 0 if there is no circulating-current. The gain “1/2” is used in this particular case to calculate the cross circulating-current from (1). \( G_{cq} \) is the controller in the d-axis circulating-current control loop. Then, by substituting (8) into (14), one can obtain the new reference current \( I_{de1}^* \) for the VSI\(_1\). Being \( I_{de1} \) is the original d-axis reference current.

To explain the functions more clearly, supposing that the currents \( I_{dn1} > I_{dn2} \), the correct direction of the compensation should be decreasing reference \( I_{de1} \) for VSI\(_1\) and increasing \( I_{de2} \) for VSI\(_2\). The calculation results from (8) and (10) will be \( I_{dn1} > 0, I_{dn2} < 0 \). Substituting (8) into (14), and (10) into (16), the new d-axis reference current \( I_{de1}^* \) will be subtracted from \( I_{dn1} \), and \( I_{de2}^* \) will be increased from \( I_{dn2} \), which indicates the correct compensation direction. So that, with proper design of the circulating-current controllers, the cross and the zero-sequence circulating-currents can be effectively suppressed.
III. INFLUENCE ANALYSIS OF THE PROPOSED CIRCULATING-CURRENT CONTROL LOOP

In order to analyze the influence and parameters sensitivity of the additional circulating-current control loop, a continuous model by using the transfer function of the control method has been developed in this paper. With the pole map of the characteristic equation, one can analyze the stability of the system and obtain the stability range of the control parameters.

The purpose of this paper is to present a circulating current suppression method to get a better current sharing performance for the existing control strategies. From Fig. 4, one can notice that, the output values of the circulating-current control loops are added to the original current loops. The basic control structure of the droop plus virtual impedance loop, and the secondary control will not be changed. Some previous works discussed the stability of control framework based on droop control plus the virtual impedance loop [34]-[41], while this paper puts more emphasis on analyzing the influence of the additional circulating-current control loop to the output voltage control. The reason of this is that since except the output impedance, the voltage difference among the inverters is another important issue that may cause circulating-currents which can also be seen from the circulating-current calculation equation (2) in Section II.A. Meanwhile, the conventional voltage and current controls will be compared with the proposed control method to illustrate more clearly the difference between them.

A. Analysis of Conventional Voltage and Current Control

The linear control model of conventional voltage and current control is shown in Fig. 5. Typically, it contains the model of control part and the model of power plant part. In the control part, \( G_1(s) \) and \( G_2(s) \) represent the voltage controller and current controller respectively. And in the power plant part, \( K_{PWM} \) represents the model of the inverter and it can be expressed as (20) in which \( T_s \) is the sampling time [38], \( rL \) represents the parasitic resistance of the output line, and it is determined by the equivalent series resistance (ESR) of the filter inductor and other parasitic elements. So the \( rL \) is not easy to measure or estimate [41].

\[
K_{PWM} = \frac{1}{1 + 1.5 T_s} \tag{20}
\]

Defining that:
and based on Fig. 5, one can deduce the close loop transfer function of the conventional voltage and current control as:

\[
V_c(s) = \frac{-Ls^2 + (rL + K_p L_{PWM})s + (K_p L_{PWM} + 1)s + K_p L_{PWM}}{D_1(s)}
\]

(22)

where \(D_1(s)\) is the characteristic equation expressed as

\[
D_1(s) = L C s^3 + (rL + K_p L_{PWM})Cs^2 + (K_p L_{PWM} + 1)s + K_p L_{PWM}.
\]

So that, if the parameters of the voltage/current control loops and the power plant are obtained, based on the pole map of \(D_1(s)\), one can analyze the stability of the system.

Table I shows the parameters of the filter and controllers used in the experiments of this paper, in which \(rL=0.2\Omega\) is considered as the ESR of the filter inductor, and \(K_{PWM}\) is considered as \(K_{CS}=1\) since the \(T_S\) is very small under high switching frequency. By substituting the parameters into (23), one can get the following characteristic equation:

\[
D_1(s) = 4.86e^{-8}s^3 + (5.4e^{-6} + 2.7e^{-5}K_p)s^2 + (1 + 1.5K_p)s + 10K_p.
\]

(24)

By using the pole map analysis tool in Matlab, with the increase of \(K_p\) from 1 to 20, and the increment step of 1, one can obtain the pole map as shown in Fig. 6. It is obviously that all the poles are located in the left filed of imaginary axis, which indicates that the system is stable with these controller parameters.

### B. Model of Proposed Control Method

In order to analyze the transfer function of the proposed circulating current control loop in a simple way, the current control loop with circulating-current control will be introduced first, then it will be combined in the analysis of the voltage control loop.

Fig. 7 shows the linear control model of the current control loop with the proposed circulating current control loop. Compared with Fig. 5, there is one more variable \(I_4(s)\) added to the original current reference \(I_{ref}(s)\) to get a new current reference \(I_{ref}(s)\). And \(I_4(s)\) comes from the circulating-current control loop, which is the compensation value to suppress the circulating-current. From Fig. 7, one can obtain

\[
I_{ref}(s) = I_{ref}(s) + I_4(s)
\]

(25)

\[
I_4(s) = \frac{1}{2} [0 - (I_{L1}(s) - I_{L2}(s))] G_c(s)
\]

(26)

where \(G_c(s)\) represents the controller of the circulating-current control loop. Substituting (26) into (25), and based on Fig. 7, one can deduce the transfer function as

\[
I_{L1}(s)(Ls + rL) = \left[ I_{ref}(s) + \frac{1}{2}(I_{L2}(s) - I_{L1}(s)) G_c(s) - I_{L1}(s) \right]
\]

\[
\times G_c(s) K_{PWM} = V_c(s)
\]

(27)

Furthermore, one can obtain (28) from (27), it depicts the influence of the reference current \(I_{ref}(s)\), the capacitor voltage \(V_c(s)\) and the inductor current \(I_{L1}(s)\) from the neighbor inverter to the local inductor current \(I_{L1}(s)\).

\[
I_{L1}(s) = F_1(s)I_{ref}(s) + F_2(s)I_{L2}(s) - F_3(s) V_c(s)
\]

(28)

In (28):
can be written as
parallel-connected inverter to the output voltage:
capacitor voltage
controller as
(Circulating-current controller)
inverter is considered in the proposed control method.
Compared with the traditional voltage and current control, the
Describing the influence of the reference voltage
Substituting (21), (29), (30) and (31) into (32), one can obtain:
\[
D_2(s) = a_0s^4 + a_1s^3 + a_2s^2 + a_3s + a_4.
\] (35)

The parameters of the coefficients \(a_0, a_1, a_2, a_3\) and \(a_4\) are listed in Table II.

**TABLE II**

<table>
<thead>
<tr>
<th>Coefficients of the Characteristic Equation</th>
<th>P controller of (G_s(s))</th>
<th>PI controller of (G_s(s))</th>
</tr>
</thead>
<tbody>
<tr>
<td>(a_0)</td>
<td>2LC</td>
<td>2LC</td>
</tr>
<tr>
<td>(a_1)</td>
<td>(2rL + (2 + \frac{K_{PC}}{K_{PWM}})K_{PWM}K_r)C</td>
<td>(2rL + (2 + \frac{K_{PC}}{K_{PWM}})K_{PWM}K_r)C</td>
</tr>
<tr>
<td>(a_2)</td>
<td>(2(1 + K_{PC}K_r)K_{PWM})</td>
<td>(2(1 + K_{PC}K_r)K_{PWM})</td>
</tr>
<tr>
<td>(a_3)</td>
<td>(2K_{PC}K_{PWM}K_r)</td>
<td>(2K_{PC}K_{PWM}K_r)</td>
</tr>
</tbody>
</table>

Fig. 9 shows the pole map using the controller parameters shown in Table III. In Fig. 10, \(\lambda_i (i=1, 2, 3)\) represents the eigenvalues of the characteristic equation. With the increasing of the proportional parameter of the circulating-current controller \(K_{PC}\) from 1 to 20 with increment step of 1, the eigenvalues change from \(\lambda_i (i=1, 2)\) to \(\lambda_{i'} (i=1, 2)\). It can be seen that there will be always an eigenvalue of \(\lambda_3 = -6.17\). So, all the poles of the close loop transfer function are located in the left field of imaginary axis, so that the system will remain stable.

Fig. 10 shows the pole map when \(K_{PC}\) is fixed with parameters shown in Table IV. The eigenvalues are always located in the left side of imaginary axis. The location of the poles does not change much with the increasing of the integral parameter of the circulating-current controller. It means that the integral parameter barely has influence to the system stability. It also can be seen clearly from the expression of the eigenvalues for the P and PI circulating-current controller, only the third eigenvalues \(a_2\) are slightly different. So that, for the circulating-current control loop, a proportional controller is chosen for the experiments.

**TABLE III**

| Parameters of the Controllers with Variable Proportional Part of Circulating-current Controller |
|-----------------------------------------------|-------|-------|-------|-------|-------|-------|
| Coefficient | \(K_{PC}\) | \(K_r\) | \(K_{ PWM}\) | \(K_{PWM}\) | \(K_r\) | \(K_{PC}\) |
| Value       |       |       |       |       |       |       |
| 1.5         |       |       |       |       |       |       |
| 10          |       |       |       |       |       |       |
| 8           |       |       |       |       |       |       |
| 5           |       |       |       |       |       |       |
| 1-20        |       |       |       |       |       |       |
| 0.2         |       |       |       |       |       |       |

**Fig. 9** Pole map when changing the proportional term value of the circulating-current controller.
By analyzing the pole map of the transfer function of the proposed circulating-control method with proper control parameters, one can guarantee the system stability.

C. The Selection of Virtual Impedance

The output impedance of the closed-loop inverter determines the droop control strategy [37]. The conventional droop scheme $P - \omega$ and $Q - V$ is often adopted. With the droop control, the frequency and the amplitude of the inverter output-voltage reference can be expressed as [37]:

$$\omega = \omega' - m_p P$$

$$E = E' - m_Q Q$$

(36)

(37)

where $\omega'$ and $E'$ are the output voltage reference frequency and amplitude, $m_p$ and $m_Q$ are the droop coefficients. More details of droop control can be found in the literatures [40]–[43] and for the conventional droop control scheme, a highly inductive output impedance at fundamental frequency is required to decouple the influence of $P$ and $Q$ to the frequency and voltage amplitude [42], [44].

Considering the transfer function (22) of the conventional voltage and current control, it can be written as (38) in which $G(s)$ can be seen as voltage gain and $Z(s)$ is the output impedance of the inverter [42], [44]. And the frequency-domain behavior of the output impedance can be analyzed through the Bode diagram [42].

$$V_r(s) = G(s)V_{ref}(s) - Z(s)I(s)$$

(38)

Considering virtual impedance in the voltage loop, $V_{ref}(s)$ in (38) should be replaced by $V_{ref}^*(s)$, in which $R_v(s)$ represents the virtual impedance.

$$V_{ref}^*(s) = V_{ref}(s) - R_v(s)I(s)$$

(39)

Supposing that the virtual impedance is chosen as (40), where $R_v$ is the virtual resistance, and $L_v$ is the virtual inductance.

$$R_v = R_v + sL_v$$

(40)

By deducing the transfer function of the output voltage with virtual impedance, one can obtain:

$$V_r(s) = \frac{K_p K_i K_{PWM} s + K_p K_{PWM} V_{ref}(s)}{D_v(s)} - \frac{F(s)}{D_v(s)}I(s)$$

(41)

In which:

$$F(s) = (L + K_p K_{PWM} L_q) s^2 + [rL + K_p K_{PWM} (1 + K_p R_v + K_{PWM})] s + K_p K_{PWM} R_v$$

(42)

With virtual impedance, the output impedance can be expressed as (43), in which $D_v(s)$ is as same as (22).

$$Z_v(s) = F_v(s)/D_v(s)$$

(43)

So that if the parameters of the power plant and the controller are obtained, one can analyze the frequency-domain behavior of the output impedance by using Bode diagram. Through the frequency-domain behavior, one can choose a proper virtual impedance to have a highly inductive output impedance for the conventional droop control.

The Bode diagram of the output impedance with virtual impedance is shown in Fig. 11. Four different parameters of virtual inductance $L_v$ are analyzed, they are 0.2 mH, 0.5 mH, 0.9 mH and 1.8 mH respectively, and the virtual resistance is fixed as $R_v=0.1$. The $X_L/X_R$ ratio should be kept in mind to make sure a highly inductive virtual impedance is applied [43], [44]. From Fig. 11, one can notice that with the increasing of $L_v$, the impedance angle of the output impedance is closer to $90^\circ$ at fundamental frequency, which behaves as a highly inductive output impedance. So based on the frequency-domain behavior of the output impedance and experimental results, one can select the virtual impedance. More details of the virtual impedance design and droop control can be found in literatures [40]–[44].
IV. EXPERIMENTAL RESULTS

In order to verify the effectiveness of the proposed circulating-current control strategy, experiments were performed with a dSPACE 1006 system and two 2.2kVA three-phase two-level inverters from Danfoss. The parameters of the controllers and the power stage are shown in Table V and Table VI respectively.

### TABLE V

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Voltage controller $K_{vp}$</td>
<td>1.5, $K_{iv}$ = 10</td>
</tr>
<tr>
<td>Current controller $K_{pc}$</td>
<td>8</td>
</tr>
<tr>
<td>Circulating-current controller $K_c$</td>
<td>15</td>
</tr>
<tr>
<td>$R_{v1}$, $L_{v1}$</td>
<td>0.1Ω, 0.9mH</td>
</tr>
<tr>
<td>$R_{v2}$, $L_{v2}$</td>
<td>0.15Ω, 0.9mH</td>
</tr>
<tr>
<td>Droop coefficients $m_p$, $m_q$</td>
<td>0.0001, 0.0001</td>
</tr>
</tbody>
</table>

### TABLE VI

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Linear load current</td>
<td>5.4A</td>
</tr>
<tr>
<td>Linear load power</td>
<td>3.7kVA</td>
</tr>
<tr>
<td>Nonlinear load peak current</td>
<td>9.4A</td>
</tr>
<tr>
<td>Nonlinear load configuration</td>
<td>3φ rectifier connected with a resistor (60Ω) and a capacitor (215μF)</td>
</tr>
<tr>
<td>Switching frequency</td>
<td>10kHz</td>
</tr>
<tr>
<td>Load voltage (RMS)</td>
<td>230V/50Hz</td>
</tr>
<tr>
<td>Filter inductances</td>
<td>$L_{f1}=L_{f2}=1.8$mH</td>
</tr>
<tr>
<td>Filter capacitances</td>
<td>$C_{f1}=C_{f2}=27$μF</td>
</tr>
</tbody>
</table>

Experimental results are shown below. Figs. 12 to 18 are the experimental results when the linear load was connected. Figs. 19 to 26 are the experimental results when nonlinear load was sharing.

Fig. 12. Experimental results with the proposed control strategy when linear load is connected. (a) Load voltage. (b) Phase A currents of two inverters and the load.

Fig. 13. The phase A output currents and cross circulating-current with the proposed circulating-current control loop when sharing linear load.

Fig. 14. The phase A output currents and cross circulating-current with the conventional droop plus virtual impedance control when sharing linear load.

Fig. 15. The zoomed-in cross circulating-current with the proposed circulating-current control loop when sharing linear load.

Fig. 16. The zoomed-in cross circulating-current with the conventional droop plus virtual impedance control when sharing linear load.
Fig. 17. The zoomed-in zero-sequence circulating-current with the proposed circulating-current control loop when sharing linear load.

Fig. 18. The zoomed-in zero-sequence circulating-current with the conventional droop plus virtual impedance control when sharing linear load.

Fig. 19. The phase A output current and load current with the proposed circulating-current control loop when sharing nonlinear load.

Fig. 20. The phase A output current and load current with the conventional droop plus virtual impedance control when sharing nonlinear load.

Fig. 21. The phase A output current and the cross circulating-current with the proposed circulating-current control loop when sharing nonlinear load.

Fig. 22. The phase A current and the cross circulating-current with conventional droop plus virtual impedance control when sharing nonlinear load.

Fig. 23. The zoomed-in cross circulating-current with the proposed circulating-current control loop when sharing nonlinear load.

Fig. 24. The zoomed-in cross circulating-current with conventional droop plus virtual impedance control when sharing linear load.
When load was sharing, from the experimental waveforms, one can notice that the peak value of the cross circulating-current and zero-sequence circulating-current are about 50mA and 75mA respectively with the proposed circulating-current control strategy. However, the values are about 200mA and 300mA when using the conventional droop plus virtual impedance control method.

More details about the obtained experimental results are shown in Table VII, which shows the approximate peak value of the circulating-currents. It can be seen that, independently from sharing linear or nonlinear loads, compared with the droop plus virtual impedance control, both of the cross circulating-current and the zero-sequence circulating-current can be effectively suppressed with the proposed control method. Thus a better performance of average current-sharing is realized.

| TABLE VII | EXPERIMENTAL RESULTS COMPARISON (APPROXIMATE PEAK VALUE) |
| | Using the proposed control method | Using the conventional droop plus virtual impedance control |
| | Linear load | Nonlinear load | Linear load | Nonlinear load |
| | \( I_{xc} \) | \( I_{zc} \) | \( I_{xc} \) | \( I_{zc} \) |
| Linear load | 50mA | 75mA | 200mA | 200mA |
| Nonlinear load | 75mA | 100mA | 300mA | 400mA |

V. CONCLUSION

Parallel-connected inverters systems are widely applied for high-power/high-reliable electrical supply requirements. In these systems, an accurate average current-sharing performance is necessary. In this paper, a circulating-current suppression method is proposed for paralleled VSIs with common AC and DC buses, which can be found in applications such as AC-DC hybrid microgrids and modular parallel UPS systems.

In the proposed approach, both of the cross and zero-sequence circulating-currents are considered and successfully suppressed. In comparison with the conventional droop plus virtual impedance control framework, it presents superior performances. The proposed concept to suppress the circulating-currents is based on the distributed control strategy. The system modeling is presented to study the influence of the additional circulating-current control loop to the output voltage. Then, the performance of the proposed control strategy is analyzed by using experimental results obtained from a parallel VSI system when sharing linear and nonlinear loads. The results demonstrate that the average current-sharing performance of the proposed control strategy is more accurate than that using the conventional droop plus virtual impedance control, pointing out that both of zero-sequence and cross circulating-currents among the parallel inverters can be effectively suppressed.

REFERENCES
