Wireless-Control Strategy for Parallel Operation of Distributed-Generation Inverters

Josep M. Guerrero, Member, IEEE, José Matas, Luis García de Vicuña, Miguel Castilla, and Jaume Miret, Member, IEEE

Abstract—In this paper, a method for the parallel operation of inverters in an ac-distributed system is proposed. This paper explores the control of active and reactive power flow through the analysis of the output impedance of the inverters and its impact on the power sharing. As a result, adaptive virtual output impedance is proposed in order to achieve a proper reactive power sharing, regardless of the line-impedance unbalances. A soft-start operation is also included, avoiding the initial current peak, which results in a seamless hot-swap operation. Active power sharing is achieved by adjusting the frequency in load transient situations only, owing to which the proposed method obtains a constant steady-state frequency and amplitude. As opposed to the conventional droop method, the transient response can be modified by acting on the main control parameters. Linear and nonlinear loads can be properly shared due to the addition of a current harmonic loop in the control strategy. Experimental results are presented from a two-6-kVA parallel-connected inverter system, showing the feasibility of the proposed approach.

Index Terms—Distributed generation (DG), droop control method, microgrids, nonlinear loads.

I. INTRODUCTION

DISTRIBUTED generation (DG) is emerging as a new paradigm to produce onsite highly reliable and good-quality electrical power [1]. DG becomes a viable alternative when renewable or nonconventional energy resources are available, such as photovoltaic arrays, fuel cells, cogeneration plants, combined heat and power microturbines, or small wind turbines [2]–[4]. These resources can be connected to local low-voltage electric power networks, also called mini- or microgrids, through power conditioning ac units, i.e., inverters or ac–ac converters [5], [6], which can operate either in grid-connected mode or in island mode.

Grid-connected operation consists in delivering power to the local loads and to the utility grid. In such a case, the output-voltage reference is often taken from the grid voltage sensing, and is using a phase-locked-loop (PLL) circuit, while an inner current loop ensures that the inverter acts as a current source [7], [8]. Currently, when the grid is not present, the inverters are normally disconnected from the ac line in order to avoid islanding operation. In the coming years, inverters should be able to operate in island mode due the high penetration of DG [9]. In addition, in certain zones where a stiff grid is not accessible, e.g., some physical islands, rural or remote areas, islanding operation mode is necessary. In this situation, the output-voltage reference should be provided internally by the DG units, which operate independently without mutual intercommunication due to the long distance between them [10].

Several control techniques based on the droop method have been proposed in order to avoid using communication between DG units [11]–[16]. These control techniques consist in making tight adjustments over the output-voltage frequency and amplitude of the inverter in order to compensate for the active and reactive power unbalances [17]. Nevertheless, the standard approach only works well when linear loads are shared, since the amount of distorted power demanded by the nonlinear loads is not taken into account. In this sense, an additional control loop was proposed in order to share nonlinear loads by adjusting the output-voltage bandwidth according to the amount of delivered harmonic power [18]. In previous work, most harmonic current values were used to produce a proportional droop in the corresponding harmonic voltage term [19]. However, all these control approaches have an inherent tradeoff between voltage regulation and power sharing. Furthermore, the conventional droop method exhibits slow dynamic response, since it requires low-pass filters with reduced bandwidth to calculate the average values of the active and reactive powers [20]. In [21], a wireless controller was proposed in order to enhance the dynamic performance of the paralleled inverters by adding integral-derivative power terms to the droop-control method. Nevertheless, when an inverter is connected suddenly to the common ac bus, a current peak appears due to the initial phase error.

Another drawback of the standard droop method is that, the power sharing is degraded if the sum of the output impedance and the line impedance is unbalanced. To solve this, interface inductors can be included between the inverter and the load bus, but they are heavy and bulky [22]. As an alternative, novel control loops that fix the output impedance of the units by emulating lossless resistors or reactors have been proposed [23], [24]. However, although the output impedance of the inverter can be well established, the line impedance is unknown, which can result in an unbalance of reactive power flow [18]. This problem can be overcome by injecting high-frequency
signals through the power lines [25] or by adding external data communication signals, as in [26]. Unfortunately, such communication among DG units increases complexity and reduces reliability, since the power balance and the system stability rely on these signals.

In this paper, we propose a novel control scheme that is able to further improve the steady state and the transient response of the parallel-connected inverters without using any communication signals. The controller uses an adaptive output impedance, which allows a good reactive power sharing with low sensitivity to the line-impedance unbalances, while keeping constant the output-voltage amplitude and the frequency in steady state. A soft-start operation of the output impedance is also proposed in order to achieve a seamless connection between the inverter and the common bus. Finally, the most significant output-current harmonics are treated separately using a bank of bandpass filters in order to share these harmonics properly without excessively increasing the output-voltage total harmonic distortion (THD). Experimental results from a two-6-kVA-inverter system are provided, showing the outstanding features of the proposed controller.

II. OUTPUT-IMPEDANCE IMPACT OVER POWER SHARING

Fig. 1 shows a general scheme of a microgrid which consists of a combination of multiple microgenerator DG units, distributed loads, and electric power interfaces that transfer energy to the local ac bus. The microgrid can be connected to the utility grid through a single point of common coupling. When the utility grid is not present, the DG units should be able to share the total power demanded by the local loads, adjusting its output-voltage references as a function of the dispatched power.

Fig. 2 shows the equivalent circuit of the DG units as inverters connected to a common ac bus through decoupling output impedances. The active and reactive power injected to the bus by every unit can be expressed as follows [27]:

\[
P = \left( \frac{EV}{Z} \cos \phi - \frac{V^2}{Z} \right) \cos \theta + \frac{EV}{Z} \sin \phi \sin \theta
\]

\[
Q = \left( \frac{EV}{Z} \cos \phi - \frac{V^2}{Z} \right) \sin \theta - \frac{EV}{Z} \sin \phi \cos \theta
\]

where \(E\) and \(V\) are the amplitudes of the inverter output voltage and the common bus voltage, \(\phi\) is the power angle, and \(Z\) and \(\theta\) are the magnitude and the phase of the output impedance, respectively. Note that both \(P\) and \(Q\) depend simultaneously on the output-voltage parameters \(\omega\) (i.e., \(\omega = d\phi/dt\)) and \(E\).

Therefore, the droop-control method can be expressed in a general form as

\[
\omega = \omega^* - m(P \sin \theta - Q \cos \theta)
\]

\[
E = E^* - n(P \cos \theta + Q \sin \theta).
\]

The output impedance angle \(\theta\) determines to a large extent the droop-control law, as shown in Table I. Traditionally, the inverter output impedance is considered to be inductive due to the high inductive component of the line impedance and the large inductor filter. However, this is not always true, since the closed-loop output impedance also depends on the control strategy [28].

By using fast-droop algorithms, different output impedances can be implemented, such as, (a) purely resistive, (b) purely inductive, (c) inductive and resistive connected in series, or (d) in parallel, as shown in Fig. 3. Resistive output impedance can be a good solution to share linear and nonlinear loads in applications such as uninterruptible-power-supply systems [23]. However, when the distance between the inverters is considerable, an inductive output impedance component appears which impoverishes active and reactive power sharing. Inductive output impedance seems to be the most natural output impedance [24]. However, it degrades the output-voltage THD too much when supplying nonlinear loads due to the large impedance value seen by the current harmonics. In order to solve these problems, two combined resistive–inductive output impedances can be considered: series and parallel. The first one [29], although it can be controlled by using (3) and (4), is still unsatisfactory,
TABLE I  
<table>
<thead>
<tr>
<th><strong>Output Impedance Dependence With Output Impedance</strong></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Active power</strong></td>
<td></td>
</tr>
<tr>
<td>$Z = jX$ (inductive: $\theta = 90^\circ$)</td>
<td>$Z = R$ (resistive: $\theta = 0^\circ$)</td>
</tr>
<tr>
<td>$P = \frac{EV}{X} \sin \phi \equiv \frac{EV}{X}$</td>
<td>$P = \frac{EV \cos \phi - V^2}{R} \equiv \frac{V}{R} (E - V)$</td>
</tr>
<tr>
<td><strong>Reactive power</strong></td>
<td></td>
</tr>
<tr>
<td>$Q = \frac{EV \cos \phi - V^2}{X} \equiv \frac{V}{X} (E - V)$</td>
<td>$Q = -\frac{EV}{R} \sin \phi \equiv -\frac{EV}{R}$</td>
</tr>
<tr>
<td><strong>Frequency droop</strong></td>
<td></td>
</tr>
<tr>
<td>$\omega = \omega^* - mP$</td>
<td>$\omega = \omega^* + mQ$</td>
</tr>
<tr>
<td><strong>Amplitude droop</strong></td>
<td></td>
</tr>
<tr>
<td>$E = E^* - nQ$</td>
<td>$E = E^* - nP$</td>
</tr>
</tbody>
</table>

**III. PROPOSED WIRELESS POWER-SHARING CONTROLLER**

In this section, we propose a wireless control for parallel DG inverters that provides a proper output impedance, good active and reactive power-sharing capability, constant steady-state output-voltage frequency and amplitude, and fast transient response.

A. Virtual Output Impedance and Harmonic Current Sharing

A fast control loop known as virtual-output-impedance loop can be used to fix the output impedance of the inverter. Inductive output impedance around the output-voltage frequency can be implemented by drooping the output-voltage reference proportionally to the time derivative of the fundamental component of the output current $i_{o1}$. Resistive output impedance for high-order current harmonics is obtained by subtracting a voltage, which is proportional to the current harmonics from the output-voltage reference. Thus, the proposed output-voltage reference can be expressed as

$$v_{ref} = v_{ref}^* - sL_D i_{o1} - \sum_{h=3, \text{odd}}^{11} R_h i_{oh}(5)$$

where $L_D$ is the virtual output impedance, and $R_h$ is the resistive coefficient of every harmonic term $i_{oh}$. Using this loop, the output impedance presented to the fundamental and the harmonic components can be fixed independently.

Another practical issue is the desirable hot-swap capability, which consists in a seamless operation of the DG inverter when it is connected suddenly to the common ac bus. As explained in [21], the output current peak in such a situation is expressed as

$$I_p \approx \frac{E}{\omega L_D} \cdot \Delta \phi$$

where $\Delta \phi$ is the initial phase error.

In order to reduce this initial current peak, a soft-start operation of the output impedance is also proposed which achieves a seamless connection of the inverter with the common bus (hot-swap operation). The soft-start operation consists in this connection using high output impedance $L_{Do}$, which is then slowly reduced toward the nominal value $L_{Df}$.

$$L'_{D} = L_{Df} + (L_{Do} - L_{Df}) \cdot e^{-t/T_{ST}} \quad (7)$$

where $L_{Do}$ and $L_{Df}$ are the initial and the final value of the output impedance, respectively, and $T_{ST}$ is the constant time of the soft-start operation. This way, the initial current peak can be avoided regardless of the initial phase error.

B. Active and Reactive Power-Sharing Loop

In order to properly share both active and reactive powers, we propose a control strategy which avoids the use of any control wire interconnections between the modules. As we stated before, the conventional droop method presents an inherent tradeoff between the $P/Q$ sharing accuracy and the frequency/amplitude output-voltage regulation. From Table I, it can be observed that when the output impedance is inductive, the reactive power can be controlled adequately by the output-voltage amplitude, and in a conventional droop method, or by
the output impedance, as in variable active–passive reactances [31]. Therefore, taking into account that \( Q \) is reduced when \( X \) increases, we propose the following adaptive output impedance in order to regulate the \( Q \) balance [28]

\[
L_{ID} = L_{ID}^\ast + k_L Q
\]

(8)

where \( k_L \) is an adjustment constant of the inductive output-impedance gain, and \( L_{ID}^\ast \) is the reference output inductance. The proposed output impedance compensates for reactive-power differences between the modules due to output-voltage mismatches, component tolerances, or line-impedance unbalances, without deviations of the output-voltage reference amplitude. Notice that even though the active power is also affected by the output impedance, it can be controlled by adjusting the output-voltage frequency.

In order to avoid the steady-state frequency deviation of the conventional droop method, the following control scheme is proposed:

\[
\omega = \omega^\ast - m_p \bar{P} - m_d \frac{d\bar{P}}{dt}
\]

(9)

where \( m_p \) and \( m_d \) are, respectively, the proportional and derivative coefficients of the active power with no dc component, \( \bar{P} \), which can be obtained by using a high-pass filter such as

\[
\bar{P} = \frac{s}{s + \tau^{-1}} P
\]

(10)

and \( \tau \) is the time constant of the transient droop action. The derivative term of \( \bar{P} \) is added to the conventional scheme in order to improve the transient response of the system, as will be shown in the following Section. Note that the \( \hat{P} - \omega \) function has no frequency deviation in steady state. The transient droop function ensures a constant frequency regulation under steady-state conditions and, at the same time, achieves active power balance by adjusting the inverter frequency during load transients. The proposed control scheme allows us to modify the transient response by acting on the control coefficients and, at the same time, keeping constant the steady-state frequency and amplitude of the output-voltage reference [30].

IV. SMALL-SIGNAL MODELING AND CONTROL DESIGN RULES

A small-signal analysis is proposed in order to investigate the stability and the transient response of the system. Thus, the closed-loop system dynamics is derived, taking into account the well-known stiff load-bus approximation [20], [21]. The small-signal dynamics of the active and reactive power (\( \dot{p} \) and \( \dot{q} \)) are obtained by linearizing (1) and (2), with \( Z \angle \theta = Z \angle 90^\circ \), and modeling the low-pass filters with a first-order description

\[
\dot{p} = \frac{\omega_c}{s + \omega_c} \frac{1}{X} [EV \cos \Phi \cdot \dot{\phi} - P \cdot \dot{x}]
\]

(11)

\[
\dot{q} = \frac{\omega_c}{s + \omega_c} \frac{1}{X} [-EV \sin \Phi \cdot \dot{\phi} - Q \cdot \dot{x}]
\]

(12)

where \(^^\wedge\) denotes perturbed values, capital letters mean equilibrium point values, \( X \) is the output impedance at the fundamental frequency, and \( \omega_c \) is the cutoff angular frequency of the low-pass filters, which is fixed over one decade below the line frequency. For the sake of simplicity, the high-frequency impedance values are not considered in this analysis since they have little effect over the system dynamics.

Subsequently, by linearizing (8) and (10), and using (11) and (12), we obtain

\[
\dot{x} = k_L \omega_c EV \sin \Phi \cdot \dot{\phi} - \omega_c EV \cos \Phi \cdot P \cdot \dot{x}
\]

(13)

Finally, substituting (14) into (13), we can find the small-signal dynamics of the closed-loop system

\[
s^3 \ddot{\phi} + As^2 \dot{\phi} + Bs \dot{\phi} + C \phi = 0
\]

(15)

where

\[
A = \frac{1}{X} [\tau^{-1} + \omega_c + m_d EV \cos \Phi + X \omega_c + k_L \omega_c Q] \]

\[
B = \frac{1}{X^2} \left[ Pm_d k_L \omega_c^2 EV \sin \Phi + Xm_p \omega_c EV \cos \Phi \right. \]

\[
+ \left. (X \tau^{-1} + X \omega_c + m_d EV \cos \Phi)(X \omega_c + k_L \omega_c Q) \right]
\]

\[
C = \frac{1}{X^2} \left[ (X \omega_c \tau^{-1} + m_p \omega_c EV \cos \Phi)(X \omega_c + k_L \omega_c Q) \right. \]

\[
+ \left. Pm_p k_L \omega_c^2 EV \sin \Phi \right]
\]

By using (15), the stability of the closed-loop system can be evaluated, and a desired transient response can be selected following a linear third-order dynamics. After studying the eigenvalues of the system \((\lambda_1, \lambda_2, \lambda_3)\) through a series of root locus diagrams, as in [20] and [21], and using the values listed in Table II, we can conclude that both the output impedance \( X \) and the coefficient \( k_L \) have little effect on the location of the roots in comparison with \( m_p \) and \( m_d \) coefficients. Fig. 4(a) depicts the family of root locus diagrams, as in [20] and [21], and using the values listed in Table II, we can conclude that both the output impedance \( X \) and the coefficient \( k_L \) have little effect on the location of the roots in comparison with \( m_p \) and \( m_d \) coefficients.
TABLE II

PARAMETERS OF THE POWER-SHARING CONTROLLER

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Value</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal frequency</td>
<td>$\omega^*$</td>
<td>2$\pi$.50</td>
<td>rad/s</td>
</tr>
<tr>
<td>Nominal amplitude</td>
<td>$E^*$</td>
<td>311</td>
<td>V</td>
</tr>
<tr>
<td>Nominal virtual output impedance</td>
<td>$L_{o^*}$</td>
<td>800</td>
<td>$\mu$H</td>
</tr>
<tr>
<td>L-Q proportional boost coefficient</td>
<td>$k_L$</td>
<td>$10^{-7}$</td>
<td>H/VAr</td>
</tr>
<tr>
<td>$\tilde{P}$-$\omega$ proportional droop coefficient</td>
<td>$m_p$</td>
<td>$3 \cdot 10^{-4}$</td>
<td>rad/W</td>
</tr>
<tr>
<td>$\tilde{P}$-$\omega$ derivative droop coefficient</td>
<td>$m_d$</td>
<td>$6 \cdot 10^{-5}$</td>
<td>rad/s/W</td>
</tr>
<tr>
<td>$\tilde{P}$-$\omega$ time constant</td>
<td>$\tau$</td>
<td>0.5</td>
<td>s</td>
</tr>
<tr>
<td>Power filters cut-off frequency</td>
<td>$\omega_0$</td>
<td>10</td>
<td>rad/s</td>
</tr>
</tbody>
</table>

Fig. 4. Family of root locus diagram: (a) $m_d = 0, 6 \cdot 10^{-5},$ and $12 \cdot 10^{-5}$ for $0 \leq m_p \leq 4 \cdot 10^{-4}$, (b) $0 \leq \tau^{-1} \leq 10$.

since the influence over the dominant pole location is small, as shown in Fig. 4(b).

The nominal output inductance of the inverter $X$ should be selected by taking into account the following tradeoff. Increasing the output impedance reduces the circulating current produced by the differences between the power lines or between the unit parameters, such as the output filter values or the frequency and amplitude set points. Nevertheless, as Fig. 5 depicts, increasing $X$ also reduces the maximum active power that can be delivered to the common bus

$$P_{\text{max}} = \frac{EV}{X^*}. \quad (16)$$

On the contrary, decreasing the output impedance forces the power angle to become smaller

$$\phi = \sin^{-1}\left(\frac{PX}{EV}\right). \quad (17)$$

In that case, the approximations $\sin \theta \approx \theta$ and $\cos \theta \approx 1$ are good enough, and consequently, the active and reactive power expressions become

$$P \approx \frac{EV}{X^*} \cdot \phi \quad (18)$$
$$Q \approx \frac{V}{X^*} \cdot (E - V) \quad (19)$$

which leads to a good decoupling between $P/Q$ and the frequency/amplitude output voltage.

Thus, there is a tradeoff in the output-impedance design. In our case, we must make sure that the output impedance has a proper value in a range of full load conditions ($Q_{\text{max}}$) from capacitive to inductive. This value also affects the selection of the $k_L$ coefficient, since

$$X^* - k_L Q_{\text{max}} < X < X^* + k_L Q_{\text{max}}. \quad (20)$$

V. DSP-CONTROLLER IMPLEMENTATION

Fig. 6 depicts the block diagram of the proposed power-sharing controller, which includes the transient $P - \omega$ droop and the adaptive output impedance loop. The average active power $P$, without the dc component, can be obtained by
multiplying the output voltage $v_o$ by the output current $i_o$ and filtering this product using a bandpass filter [30]

$$\hat{P} = \frac{\omega_c s}{s + \tau^{-1}} \frac{P_i}{s + \omega_c}$$  \hspace{1cm} (21)

where $\tau^{-1}$ must range between 0 and $\omega_c$. This expression is derived by combining the high-pass filter (10), which avoids the dc component, and the low-pass filter (11), which averages the instantaneous active power. In order to adjust the output-voltage frequency, (9) is implemented, which corresponds to a PD controller applied to the transient active power signal $\hat{P}$.

The average reactive power is obtained by delaying the output voltage by 90° through a circular buffer, multiplying by $i_o$, and using a low-pass filter. These filters were discretized through the bilinear transformation, obtaining infinite-impulse response solutions. The virtual output impedance is implemented by using the adaptive output impedance, which is regulated by the reactive power, as (8) shows.

The fundamental and harmonic components of the output current can be extracted by using a bank of bandpass filters. Each filter extracts the selected component, following the procedure described in [32]

$$H_i(s) = \frac{2k_i s}{s^2 + 2k_i s + \omega_i^2}$$  \hspace{1cm} (22)

where $k_i = \zeta_i \omega_i$ is the coefficient of the filter, $\zeta_i$ is the damping factor, and $\omega_i$ is the frequency of the $i$-harmonic ($i = 3, 5, \ldots, 11$). Note that each filter implementation includes the resistive gain coefficient $R_i$. Using (5) and (22), the following virtual impedance expression can be found:

$$Z_V(s) = \frac{2k_1 s^2}{s^2 + 2k_1 s + \omega_1^2} + \sum_{i=3}^{n} \frac{2k_i s}{s^2 + 2k_i s + \omega_i^2}.$$  \hspace{1cm} (23)

Fig. 7 shows a Bode-diagram representation of this impedance.

By properly adjusting $R_i$ coefficients, we can obtain a good tradeoff between harmonic current sharing and output-voltage THD, in contrast with the conventional techniques. In addition, this sum of filters can be easily implemented in a digital signal processor (DSP) by using finite-impulse response digital filters.

VI. SIMULATION AND EXPERIMENTAL RESULTS

The proposed controller was simulated for a two-parallel-inverter system sharing a load in order to show the feasibility of the proposed controller. Fig. 8(a) and (b) depicts the transient response of a two-parallel-inverter system at the startup, and for load step changes. It is shown that when the system stays synchronized, the load changes do not affect power-sharing accuracy. Fig. 8(c) and (d) shows the reduced circulating current between the inverters, and the capability to restore the frequency in front of the conventional droop method. Notice that when the startup transient ends, the power sharing is unaffected by the load step changes.

Two 6-kVA single-phase inverter units were built and tested, confirming experimentally the validity of the proposed approach. Each inverter consisted of a single-phase IGBT full
bridge with a switching frequency of 20 kHz and an LC output filter, with the following parameters: \( L = 500 \mu H \), \( C = 100 \mu F \), \( V_{in} = 400 \text{ V} \), \( \nu_o = 220 \text{ V}_{\text{rms}}/50 \text{ Hz} \). The impedances of the lines connected between the inverters and the load were intentionally unbalanced, \( Z_{L1} = 0.12 + j0.028 \Omega \), and \( Z_{L2} = 0.24 + j0.046 \Omega \). The controllers of these inverters were based on three loops: an inner current loop, an outer PI controller that ensures voltage regulation [21], and the load-sharing controller, based on (8)–(10), using the parameters shown in Table II. The first two loops were implemented into a TMS320LF2407A, fixed-point 40-MHz DSP from Texas Instruments, and the wireless load-sharing controller was performed by using a TMS320C6711, floating-point 200-MHz DSP. The connection between the two-DSPs was made through the host-port-interface characteristic of the floating-point processor. The voltage sampling frequency was rated at 10 kHz, and the current sampling frequency was at 20 kHz. The DSP controller also includes a PLL block in order to synchronize the output voltage of the inverter in frequency and phase with the common bus. When the output-voltage synchronization is completed, the static bypass switch is turned ON, and the soft-start operation and the droop-based control are initiated.

The outstanding features of the parallel system are experimentally evaluated in the case of the two-unit system sharing a linear load, as depicted in Fig. 9. Fig. 10 shows the output currents of each unit and the circulating current \( (i_1−i_2) \). These results show an excellent dynamic response of the proposed controller for load step changes. As it can be seen, the circulating current remains very small, even for no-load conditions. Fig. 11 depicts a detail of the steady state and the dynamic performances of the system, revealing the equal current sharing.
Another experimental test consists in connecting one inverter while the other is supplying the load all the time. As can be seen in Fig. 12, transient output currents are not higher than inverter nominal current, and the transient circulating current stays low.

Finally, the behavior of the two-parallel-inverter system was evaluated when a nonlinear load with a crest factor of three is supplied. Fig. 13 shows the load voltage and current, and the output current of the two units. The measured output-voltage THD was 2.1%. As it can be seen, the load sharing capability is also very good when supplying nonlinear loads.

VII. CONCLUSION

In this paper, a novel power-sharing controller for parallel inverters has been proposed. Based on the droop method, the controller avoids the use of communication signals among the units. In a clear-cut contrast with the conventional droop method, the presented controller is able to modify the dynamic response of the paralleled system by tuning the control gain parameters. The controller consists of a transient frequency droop loop and an adaptive output-impedance loop which allow sharing active and reactive power, respectively, without sacrificing frequency or amplitude regulation in steady state.

In addition, the proposed adaptive virtual output-impedance further improves hot-swap operation and provides nonlinear load sharing in parallel-connected DG inverters. Furthermore,
it is interesting to note that the proposed control can be extended to other output impedances if necessary.

Experimental results have been reported to validate the proposed control approach, showing good power sharing when supplying linear and nonlinear loads, even when the line impedances are different. The excellent features of this wireless controller show its applicability to parallel-connected inverters in distributed power systems, such as DG systems or microgrids.

REFERENCES


Josep M. Guerrero (S’01–M’03) received the B.S. degree in telecommunications engineering, the M.S. degree in electronics engineering, and the Ph.D. degree in power electronics from the Universitat Politècnica de Catalunya, Barcelona, Spain, in 1997, 2000, and 2003, respectively.

From 1998 to 2004, he was an Assistant Professor in the Department of Automatic Control Systems and Computer Engineering, Universitat Politècnica de Catalunya. In 2004, he became a Senior Lecturer at the same university, where he teaches digital signal processing, control theory, and microprocessors. Since 2004, he has been the Responsible of the Sustainable Distributed Generation and Renewable Energy research group at the Escola Universitària d’Enginyeria Tècnica Industrial de Barcelona (EUETIB). His research interests include DSP-/FPGA-based control, uninterruptible power systems, inverters for photovoltaic applications, and wind energy conversion in microgrids.

Dr. Guerrero is an Associate Editor of the IEEE TRANSACTIONS ON INDUSTRIAL ELECTRONICS. He is a Guest Editor of the Special Issue of the IEEE TRANSACTIONS ON POWER ELECTRONICS “Power Electronics for Wind Energy Conversion” and the Special Section of the IEEE TRANSACTIONS ON INDUSTRIAL ELECTRONICS “Uninterruptible Power Supply (UPS) Systems.” He has organized and chaired sessions at several IEEE IECON, APEC, and PESC Conferences. He is listed in Marquis’ Who’s Who in the World and Marquis’ Who’s Who in Science and Engineering.

José Matas received the B.S., M.S., and Ph.D. degrees in telecommunications engineering from the Universitat Politècnica de Catalunya, Barcelona, Spain, in 1988, 1996, and 2003, respectively.

Since 1997, he has been an Associate Professor with the Department of Electronic Engineering, Universitat Politècnica de Catalunya. His research interests include power electronics modeling, simulation, and control, active power filtering, and high-power-factor ac–dc conversion.

Luis García de Vicuña received the M.S. and Ph.D. degrees in telecommunications engineering from the Universitat Politècnica de Catalunya, Barcelona, Spain, in 1980 and 1990, respectively, and the Dr. Sci. degree from the Université Paul Sabatier, Toulouse, in France, 1992.

From 1980 to 1982, he was an Engineer with Control Applications. He is currently an Associate Professor with the Department of Electronic Engineering, Universitat Politècnica de Catalunya, where he teaches power electronics. His research interests include power electronics modeling, simulation, and control, active power filtering, and digital control.