High performance current control of single-phase grid-connected converter with harmonic mitigation, power extraction and frequency adaptation capabilities

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Abstract
Grid-connected converters in distributed generation systems are required to operate under highly distorted voltage with a wide frequency range, and to provide active/reactive power information for ancillary services and power control. This paper presents multiple unbalanced synchronous reference frame control for regulating and mitigating the grid current distortion of a voltage source converter caused by grid voltage distortion and converter deadtime. The proposed control technique has intrinsic power extraction and frequency adaptation capabilities. The multiple unbalanced synchronous reference frame control is compared with the proportional multi-resonant control. A new controller which combines the unbalanced synchronous reference frame control for regulation of the fundamental component current and the multi-resonant control for harmonic compensation is presented. The proposed control techniques were verified with a 1.5-kVA LCL-filtered grid connected inverter. The multiple unbalanced synchronous reference frame control with odd harmonic orders 3rd to 13th exhibited excellent harmonic rejection with a total harmonic distortion (THD) less than 0.60% under a highly distorted voltage. The proposed combined control technique provided slightly inferior performance with the grid current THD less than 0.65%. The combined control scheme retains functionalities as the multiple unbalanced synchronous reference frame control with a significant reduction of computational effort.

1 INTRODUCTION

Power electronic converters play a vital role in modern energy systems by the integration of renewable energy sources and energy storage devices with the smart grid. Single-phase voltage source converters (VSC) are widely used in small scale systems for solar [1–3], wind [4] and hydro energy [5] as well as in electric vehicle battery chargers [6]. Grid standards have been imposed to ensure that the VSC current injected to the grid contains low harmonics under a certain frequency range [7, 8]. The injected harmonic current is caused by the dead time in each leg of the VSC [9] and the voltage harmonics at the point of common coupling (PCC) [10]. Moreover, the control scheme of the VSC with accurate power extraction capability are required for ancillary services [6, 11, 12], synchronous generator emulation [13], and for the estimation of the DC bus ripple voltage [14]. Thus, the current control of the VSC, the most inner control loop, should have the following capabilities to support higher hierarchy control systems:

1. Rejection of harmonics due to the converter deadtime and the distorted PCC voltage.
2. Maintain satisfactory performance under grid frequency variation.
3. Inherent power extraction.

Current control of VSCs can be implemented in a stationary reference frame and in a synchronous referent frame. A proportional-integral (PI) controller in the stationary frame and with a feedforward of the PCC voltage achieves satisfactory...
performance [2, 15]. However, there is an inherent current error due to a finite regulator gain at the grid frequency [15]. Among the stationary frame controllers, the proportional-resonant (PR) regulator is widely adopted in the grid current control loop of the VSCs [6, 16–18]. The PR regulator has a relatively large gain at the grid frequency which forces the current error to zero. Multiple resonant regulators at low frequency harmonics are plugged into the fundamental component controller for selective harmonic compensation [10, 19]. Frequency adaptation can be achieved by implementing the resonant controller from two integrators. Great care should be taken on discretisation methods of the resonant controller especially for higher harmonic orders since it can cause deterioration of the performance or even the stability of the control system [17].

A repetitive controller based on the internal model principle has attracted the attention of the research community for harmonic mitigation [9, 20–22]. It is equivalent to a bank of multiple resonant controllers up to the Nyquist frequency. Considering the robustness of the control loop, the repetitive controller is normally incorporated with a low pass filter and a phase lead filter. The transient performance of the repetitive controller is sluggish because it stores an integer number of data cells for a whole period. Thus, the repetitive controller is normally plugged in parallel with the fundamental component controller, i.e. a PR controller [21] or a dead beat controller [22] to compensate for the slow dynamic. On the other hand, the conventional repetitive controller has an identical gain over the frequency range, making it impossible to optimally select individual harmonics to be compensated. Furthermore, the performance of the repetitive controller deteriorates when the grid frequency deviates from the nominal value. This causes the number of data points in a cycle to be fractional. Online sampling frequency adjustment or compensation of the fractional delay using a Lagrange polynomial interpolation are the intuitive solutions to maintain the harmonic mitigation performance [20]. Other control approaches applied to the single-phase converter in the stationary reference frame include hysteresis [3], model predictive control [1] and state feedback control [11].

The aforementioned control techniques in the stationary reference frame lack power extraction capability. An additional power estimator is needed when a stationary reference frame current controller is adopted [6, 11, 13].

Current control in the synchronous reference frame has inherent power extraction from the decoupled current components. The grid current is assigned as the \( \alpha \)-axis component of the stationary frame whereas an orthogonal signal generator (OSG) creates the quadrature (\( \beta \)-axis component) of the grid current. A Park transformation is then used to convert from the AC components in the \( \alpha \beta \) axes to the DC components in the synchronous reference frame (dq axes). From there, standard PI controllers force the current errors to zero. The reference voltage in the \( \alpha \)-axis from the inverse Park transformation is sent to the pulse-width modulator (PWM) to produce the converter terminal voltage. Common OSG techniques include phase shifting methods and signal reconstruction methods. The phase shifting methods include a quarter period delay [23], an all-pass filter [24, 25], and differentiation [26]. The OSGs introduce a slow dynamic into the control loop and even cause unstable operation if the PI regulators are not properly tuned [24]. The signal reconstruction methods include a second order generalised integrator (SOGI) [5], an enhanced adaptive filter (EAF) [27], and a fictive-axis emulator (FAE) [28]. The SOGI and EAF methods reconstruct the fundamental component of the input signal with a \( -\pi/2 \) phase shift, which works well with the sinusoidal grid current. When the grid current is distorted, however, all the harmonic components go through the \( \alpha \) component of the Park transformation only. However, there is no reported evidence that this always occurs under these circumstances. The FAE technique emulate the VSC signals orthogonal to the physical axis inside a digital signal processor (DSP), then the emulated orthogonal current is used as the \( \beta \)-axis component for the Park transformation [28]. This method relies on accurate parameters of the converter and oscillation occurs when there is a parameter mismatch. Moreover, the FAE technique was tested only with an \( L \)-filtered VSC.

Another control technique in the synchronous reference frame is to generate the \( \beta \)-axis current independent of the grid current [29–34] which can be zero [30, 31] or can be the reference quadrature signal [29, 32–35]. This technique has been called the unbalanced synchronous reference frame control, which is equivalent to an ideal PR regulator. Somkun and Chunkag [35] proposed a unified structure of the unbalanced synchronous reference frame control in which an arbitrary signal is assigned as the \( \beta \)-axis component that is simultaneously fed into the Park transformations on the reference side and the feedback side. This control technique benefits from the power extraction capability of the \( dq \)-axes currents. It can be also reconfigured as a stationary frame controller with a fraction of Park and inverse Park transformations for simplification of the control system. Control stability is guaranteed due to the absence of an OSG function. Frequency adaptation is intrinsic through a phase-locked loop (PLL). The unbalanced synchronous reference frame control schemes presented in [29, 32–35] were successfully validated with the sinusoidal grid voltage. Multi-frequency parallel connections of this controller have been used to mitigate the output harmonics of a single-phase stand-alone inverter caused by a non-linear load [36]. However, an application of the unbalanced synchronous reference frame controller for regulation of the grid current under a distorted grid voltage and/or a large VSC dead time has not been reported.

This paper reports on our investigation of a multiple unbalanced synchronous reference frame control to regulate the grid current of a single-phase LCL-filtered VSC, as shown in Figure 1, and to mitigate harmonics from the grid voltage and the VSC dead time. Frequency adaptation and power extraction abilities were also evaluated. The unbalanced synchronous reference frame control scheme was compared with the multi resonant regulator. A mixed control scheme where the unbalanced synchronous reference frame control regulates the fundamental component with the multi resonant regulator for the harmonic mitigation is presented. The proposed schemes were validated with a 1.5 kVA grid-connected inverter.
2 | SYSTEM DESCRIPTION

2.1 | Decoupling power control

The single-phase LCL-filtered grid-connected inverter in Figure 1 was selected for this study. A damping resistor \( R_f \) was added to the LCL filter to ensure control stability with presence of a large grid impedance \( L_g \) and \( R_g \). The current control loop with the grid current feedback receives the reference grid current from the power control loop with the active and reactive power references \( p_{\text{ref}} \) and \( q_{\text{ref}} \). The PCC voltage \( v_{\text{PCC}}(t) \) contains harmonics given by:

\[
v_{\text{PCC}}(t) = \hat{V}_1 \cos \omega t + \sum_{n=2}^{N} \hat{V}_n \cos (n\omega t + \phi_n) \tag{1}
\]

The grid current \( i_{\text{g}}(t) \) is controlled to be:

\[
i_{\text{g}}(t) = \hat{I}_1 \cos (\omega t + \varphi). \tag{2}
\]

For the synchronous reference frame control, the grid current is decoupled into the \( dq \) axes using the Park transformation given by:

\[
\begin{bmatrix} i_d(t) \\ i_q(t) \end{bmatrix} = \begin{bmatrix} \cos \omega t & \sin \omega t \\ -\sin \omega t & \cos \omega t \end{bmatrix} \begin{bmatrix} i_{\text{g}}(t) \\ i_{\text{g}}(t) \end{bmatrix} = \begin{bmatrix} \hat{I}_1 \cos \varphi \\ \hat{I}_1 \sin \varphi \end{bmatrix}, \tag{3}
\]

where the orthogonal current \( i_{\beta}(t) = \hat{I}_1 \sin (\omega t + \varphi) \) generated by an OSG. Thus, the grid current can be written as:

\[
i_{\text{g}}(t) = i_d(t) \cos \omega t - i_q(t) \sin \omega t. \tag{4}
\]

Considering the fundamental component of the PCC voltage, the instantaneous power and reactive power are given as follows:

\[
p(t) = \frac{\hat{V}_1}{2} \hat{I}_1 \cos \varphi = \frac{\hat{V}_1}{2} i_d(t) \tag{5}
\]

\[
q(t) = -\frac{\hat{V}_1}{2} \hat{I}_1 \sin \varphi = -\frac{\hat{V}_1}{2} i_q(t). \tag{6}
\]

It is noted that the negative sign in (6) is applied when the converter is defined as the inverter, as depicted in Figure 1. Although the synchronous reference frame control enables power extraction capability, control complexity and slow dynamic performance hinder it from being widely applicable in single-phase VSCs.

2.2 | Causes of grid current distortion

The grid current control block diagram is depicted in Figure 2(a) where the winding resistance of the inverter-side and grid-side inductors \( L_1 \) and \( L_2 \) are neglected. The transfer function of the pulse-width modulated VSC is modelled as:

\[
G_{\text{PWM}}(s) = V_{\text{D}} e^{-sT_d} \tag{7}
\]

where \( T_d = 1.5T_s \) is the sampling and transportation delay, and \( T_s \) is the sampling period of the control loop synchronised with the PWM carrier signal. The grid current control block diagram is rearranged as illustrated in Figure 2(b) where the PCC voltage \( v_{\text{PCC}}(t) \) and the harmonic voltage due to the converter deadtime voltage \( v_{\text{DT}}(t) \) are considered as the disturbances. The admittance transfer function with \( v_{\text{DT}}(t) \) is given by

\[
Y_{\text{DT}}(s) = \frac{i_{\text{g}}(s)}{v_{\text{DT}}(s)} = \frac{G_{\text{LCL}}(s)}{1 + G_{\text{LCL}}(s) + G_{\text{PWM}}(s) G_{\text{LCL}}(s)} \tag{8}
\]

where \( G_{\text{LCL}}(s) \) is the current controller in the stationary reference frame and \( G_{\text{LCL}}(s) = \frac{i_{\text{g}}(s)}{v_{\text{LCL}}(s)} \) is the transfer function of the LCL filter given by

\[
G_{\text{LCL}}(s) = \frac{sC_lR_f + 1}{C_lL_1L_2s^3 + C_l(L_1 + L_2)R_fs^2 + C_l(L_1 + L_2)s}. \tag{9}
\]

The admittance transfer function with the PCC voltage is written as follows:

\[
Y_{\text{PCC}}(s) = \frac{i_{\text{g}}(s)}{v_{\text{PCC}}(s)} = -G_{\text{FW}}(s) Y_{\text{DT}}(s) \tag{10}
\]

\[
Y_{\text{PCC}}(s) = \frac{i_{\text{g}}(s)}{v_{\text{PCC}}(s)} = -\left( \frac{L_1C_l^2 + 1}{C_lR_f + 1} \right) G_{\text{LCL}}(s). \tag{11}
\]

Equations (8) and (11) indicate that the transfer function \( G_{\text{LCL}}(s) \) should be as small as possible to minimise the
effects of the dead time and the PCC voltage distortion. However, this causes a large current ripple in the inductor $L_1$. On the other hand, the current controller $G_{ci}(s)$ should have a large gain to attenuate the grid current harmonics. However, a PR regulator normally used in the grid current control loop has an infinite gain at the grid frequency which is not adequate to suppress such harmonics. Feedforward of the PCC voltage (the dashed line in Figure 2(b)) can reduce the effect of the PCC voltage distortion to some extent. If the delay in the $G_{PWM}(s)$ is neglected, the $Y_{pcc}(s)$ is
approximated as:
\[
Y_{\text{pcc}} (s) = \frac{i_g (s)}{r_{\text{pcc}} (s)} \approx \frac{G_{\text{LCL}} (s)}{1 + G_{\text{c}} (s) G_{\text{PWM}} (s) G_{\text{LCL}} (s)}. \tag{12}
\]

For the case of the L filter \( G_l = 0 \), the PCC voltage harmonics can be nearly eliminated. For the LCL filter, however, the values of \( L_1 \) and \( C_l \) should be carefully selected to lower the influence of the PCC voltage distortion to the grid current.

Grid current distortion due to the dead time is normally mitigated by using compensation schemes which requires another current sensor to detect the direction of the inverter current \( i_1 (t) \). Distortion in the grid current, due to the dead time and the PCC voltage, can be simultaneously mitigated by using a harmonic compensator \( G_{\text{c}} (s) \) plugged in parallel with the fundamental current controller \( G_{\text{c}} (s) \). The harmonic compensator provides a large gain at the target frequencies, which behaves as a set of resonant filters to bypass the harmonic current from the dead time and the PCC voltage.

3 | UNBALANCED SYNCHRONOUS REFERENCE FRAME CONTROL

The unified structure of the unbalanced synchronous reference frame control is depicted in Figure 3(a) where the reference current \( i_{\text{ref}} \) and the grid current \( i_g \) are defined as the \( \alpha \)-axis component of the Park transformations on both the reference side and the feedback side. However, the arbitrary signal \( i_1 \) is assigned as the \( \beta \)-axis component on both the feedback side and the reference side. The symbol \( \omega \) represents the variables in the unbalanced synchronous reference frame. The angle \( \theta \) for the axis transformations is the estimated value of the phase angle \( \omega t \) with respect to the fundamental component of the grid voltage from the PLL. Then, the error signals \( e_{\alpha} \) and \( e_{\beta} \) in the \( d \) and \( q \) axes are regulated by the PI current controllers \( G_{\text{c}} (s) \). Only the \( \alpha \)-axis voltage reference from the inverse Park transformation is used as the modulation signal for the PWM. The unified structure in Figure 3(a) can be reconfigured as a stationary reference frame controller, as shown in Figure 3. The grid current error \( e_{\alpha} \) is only the input signal whereas the error \( e_{\beta} \) in \( \beta \)-axis is zero. Thus, the error signals \( e_{\alpha} \) and \( e_{\beta} \) in the unbalanced synchronous reference frame are:
\[
\begin{bmatrix}
    e_{\alpha}' \\
    e_{\beta}'
\end{bmatrix} =
\begin{bmatrix}
    \cos \theta \\
    -\sin \theta
\end{bmatrix} \cdot e_{\alpha}.
\tag{13}
\]

According to the inverse Park transformation, the grid current error \( e_{\alpha} \) assigned to the \( \alpha \) axis is given by,
\[
e_{\alpha} = e_{\alpha}' \cos \theta - e_{\beta}' \sin \theta \tag{14}
\]
where \( e_{\alpha}' \) and \( e_{\beta}' \) are the error signals in the synchronous reference frame. Substituting Equation (14) into (13) yields,
\[
\begin{bmatrix}
e_{\alpha}' \\
e_{\beta}'
\end{bmatrix} = \frac{1}{2} \begin{bmatrix}
1 + \cos 2\beta \sin 2\beta \\
-\sin 2\beta (1 - \cos 2\beta)
\end{bmatrix} \cdot \begin{bmatrix}
e_{\alpha} \\
e_{\beta}
\end{bmatrix}. \tag{15}
\]

The current errors \( e_{\alpha}' \) and \( e_{\beta}' \) in the unbalanced synchronous reference frame, coupled with the current errors \( e_{\alpha} \) and \( e_{\beta} \) in the synchronous reference frame, contain DC and \( 2\beta \) components. For the balanced system such as the three-phase grid connected inverter, the error \( e_{\beta} \) that is orthogonal to the error \( e_{\alpha} \) exists in the Park transformation, which leads to the absence of \( \sin 2\beta \) in Equation (15). The current errors \( e_{\alpha}' \) and \( e_{\beta}' \) are forced to zero by the PI controllers \( G_{\text{c}} (s) \) making the current errors \( e_{\alpha} \) and \( e_{\beta} \) equal to zero. Thus, the grid current error \( e_{\alpha} \) in the physical axis is zero too. The waveforms of \( e_{\alpha}' \) and \( e_{\beta}' \) change with the \( \beta \)-axis current \( i_{\beta}' \) with similar \( e_{\alpha}' \) and \( e_{\beta}' \). The stationary equivalence of the synchronous reference frame controller is given by [24]
\[
G_{\text{c}} (s) = K_p + \frac{K_i}{s^2 + \omega^2} - \frac{\omega K_{i1}}{s^2 + \omega^2} \cdot \text{OSG} (s). \tag{16}
\]
where \( K_p \) and \( K_{i1} \) are the proportional and integral gains of the PI controller, and the OSG(\( s \)) is the OSG function. It is possible for the synchronous reference frame, \( G_{\text{c}} (s) \) to have a negative DC gain due to the OSG(\( s \)) resulting in an instability. For the unbalanced synchronous reference frame, no OSG(\( s \)) is associated. Hence, its stationary frame equivalence becomes
\[
G_{\text{c}} (s) = K_p + \frac{K_{i1}}{s^2 + \omega^2}. \tag{17}
\]

4 | PROPOSED CURRENT CONTROL SCHEMES

4.1 | Multiple unbalanced synchronous reference frame control

Figure 4 shows the proposed grid current control derived from the unbalanced synchronous reference frame concept. The proposed control scheme consists of the fundamental component current controller with power extraction capability, referred to as the DQ controller, and the harmonic controller \( H (t) \). These together mitigate the grid current distortion due to the PCC voltage harmonics and the switching dead time. The control scheme is referred as the DQH controller hereafter. The fundamental component controller is adopted from Figure 3(a) by setting \( i_g = i_{\text{ref}} \). The reference currents \( i_{\text{direct}} \) and \( i_{\text{rect}} \) are directly used as the references of the PI controllers in the \( dq \) axes. The reference orthogonal current \( i_{\text{orth}} \) can be written as:
\[
i_{\text{orth}} (t) = i_g (t) + i_{\beta} (t) \tag{18}
\]
where \( i_g (t) = i_g (t) e^{-i\theta/2} \) is the imaginary current that is orthogonal to the physical grid current \( i_g (t) \), and \( i_{\beta} (t) \) is the
FIGURE 4  Multiple unbalanced synchronous reference frame current control with harmonic controller

orthogonal current error of the physical grid current error. Then, the transformed currents $i'_d(t)$ and $i'_q(t)$ can be written as follows

$$
\begin{bmatrix}
i'_d(t) \\
i'_q(t)
\end{bmatrix} = \begin{bmatrix}
\cos \theta & \sin \theta \\
\sin \theta & \cos \theta
\end{bmatrix} \cdot \begin{bmatrix}
i_d(t) \\
i_q(t)
\end{bmatrix} + e_i(t) \cdot \begin{bmatrix}
\sin \theta \\
\cos \theta
\end{bmatrix}.
$$

(19)

$$
\begin{bmatrix}
i'_d(t) \\
i'_q(t)
\end{bmatrix} = \begin{bmatrix}
i_d(t) \\
i_q(t)
\end{bmatrix} + e_i(t) \cdot \begin{bmatrix}
\sin \theta \\
\cos \theta
\end{bmatrix}.
$$

(20)

TABLE 1  Converter parameters

| Variables | Values |
|-----------|--------|
| DC voltage $V_D$ | 400 V |
| PCC nominal voltage $V_{pcc}$ | 230 V 50 Hz |
| Grid frequency $f$ | 10 kHz 20 kHz |
| PWM carrier frequency $f_{pwm}$ | |
| Sampling frequency $f_s$ | |
| Inverter-side inductor $L_1$ | 1 mH |
| $L_1$ winding resistance $R_1$ | 0.07 Ω |
| Grid-side inductor $L_2$ | 1 mH |
| $L_2$ winding resistance $R_2$ | 0.07 Ω |
| Filter capacitor $C_f$ | 2 μF |
| Damping resistance $R_d$ | 2.2 Ω |

Substituting $e_i(t) = e_i(t) \sin \theta + e_i(t) \cos \theta$ from the inverse Park transformation and with trigonometric identities, Equation (20) can be expressed as:

$$
\begin{bmatrix}
i'_d(t) \\
i'_q(t)
\end{bmatrix} = \begin{bmatrix}
i_d(t) \\
i_q(t)
\end{bmatrix} + \frac{1}{2} \begin{bmatrix}
-(1 - \cos 2\theta) & \sin 2\theta \\
\sin 2\theta & (1 + \cos 2\theta)
\end{bmatrix} \cdot \begin{bmatrix}
e_i(t) \\
e_i(t)
\end{bmatrix}.
$$

Equation (21) indicates that the currents $i'_d(t)$ and $i'_q(t)$ in the unbalanced synchronous reference frame approach the
FIGURE 7  Frequency response of the open loop transfer function of the grid current control loop

FIGURE 8  Frequency response of the output admittance with the PCC voltage

FIGURE 9  Frequency response of the output admittance with the deadtime voltage
The current errors $i'_{dq}(t)$ and $i''_{dq}(t)$ in Equation (24) are identical to those in Equation (15) which proves that the fundamental component controller in Figure 4 is equivalent to the unbalanced synchronous reference frame controller in Figure 3. The outputs of the PI regulators provide the modulation signals $m'_{d1ref}(t)$ and $m'_{q1ref}(t)$. Then the modulation signal $m_{1ref}(t)$ of the fundamental component is determined from the $\alpha$ component of the inverse Park transformation given by,

$$m_{1ref}(t) = m'_{d1ref}(t) \cos \theta - m'_{q1ref}(t) \sin \theta. \quad (25)$$

The harmonic controller $H(j)$ is also implemented in the unbalanced synchronous reference frame. The main role of the harmonic controller is to provide a relatively large gain at the compensated frequency $\omega$ where $n$ is the harmonic order number. So, the integral gain $K_{in}$ is only used to correct the current error at $\omega_n$. Power extraction at each harmonic frequency is not necessary so the stationary reference frame structure in Figure 3(b) is selected due to its simplicity. Since $e_{dq} = 0$, the Park transformation of the current errors $e_{eq}(t)$ at the $n$th harmonic can be reduced to:

$$[e'_{d}(t)] = e_{eq}(t) \cdot \begin{bmatrix} \cos n\theta \\ -\sin n\theta \end{bmatrix}. \quad (26)$$

The current error signal $e_{eq}$ consists of the fundamental component $e_{eq1}(t)$ and harmonic components $e_{eqj}$ as follows:

$$e_{eq}(t) = e_{eq1}(t) + \sum_{j=3}^{k} e_{eqj}(t)$$

$$e_{eq1}(t) = \hat{e}_1 \cos (\theta + \phi_1) + \sum_{j=3}^{k} \hat{e}_j \cos (j\theta + \phi_j)$$

where $\hat{e}_1$ and $\hat{e}_j$ are the amplitude of the current error at the first and the $j$th harmonics. Substituting Equation (28) into (26) creates the DC components in the harmonic current errors $e'_{d}(t)$ and $e'_{q}(t)$ the $dq$ axes at the rotating angle $n\theta$ and multiple frequency AC components as illustrated as follows:

$$[e'_{d}(t)] = \frac{1}{2} \begin{bmatrix} \cos \phi_d + \cos (2n\theta + \phi_d) \\ \sin \phi_d - \sin (2n\theta + \phi_d) \end{bmatrix} + \hat{e}_1 \cos (\theta + \phi_1) + \sum_{j=3}^{k} \hat{e}_j \cos (j\theta + \phi_j) \cdot \begin{bmatrix} \cos n\theta \\ -\sin n\theta \end{bmatrix}.$$
The integral regulators $K_{in}/s$ in the $dq$ axes at the $n$th order harmonic force the DC and the $2n\theta$ components simultaneously to zero, and do not respond to the other AC components. Then, the modulation signal $m_{\text{rect}}(t)$ at the $n$th harmonic is determined from the modulation signal in the $dq$ axes $m_{dq\text{ rect}}(t)$ and $m'_{dq\text{ rect}}(t)$ given by,

$$m_{\text{rect}}(t) = m'_{dq\text{ rect}}(t) \cos n\theta - m'_{dq\text{ rect}}(t) \sin n\theta.$$  \hfill (31)

Therefore, the total modulation signal for the PWM in Figure 4 is written as:

$$m_{\text{ref}}(t) = v_{\text{pcc}}(t)/V_D + m_{1\text{ ref}}(t) + \sum_{n=3}^{k} m_{n\text{ ref}} \cdot \text{Feedforward}.$$ \hfill (32)

Note that the feedforward of the PCC voltage in Equation (32) is optional and it is not necessary when using the harmonic controllers as discussed in Section 5.2. The equivalent transfer function of the proposed controller in the stationary reference frame becomes

$$G_{ci}(s) = K_p + \frac{K_{i1}s}{s^2 + \omega^2} + \sum_{n=3}^{k} \frac{K_{in}s}{s^2 + (n\omega)^2}.$$ \hfill (33)

### 4.2 PR controller with multi-resonant harmonic controller

The PR controller mainly used in the single-phase VSC was chosen for comparison with the DQH controller. Figure 5 illustrates the regulator structure which comprises a proportional gain $K_p$ and multiple resonant controllers. This controller is then referred to as the PMR control scheme. Each resonant regulator is implemented using two integrators with a gain of $K_{in}$. This form allows frequency adjustment with the estimated grid.
FIGURE 12 Experimental results of the PR control scheme under the sinusoidal PCC voltage

frequency \( \tilde{\omega} \) obtained from the PLL. The transfer function of this controller is similar to that of the DQH controller given in Equation (25). The PMR control structure is simpler compared with the DQH control scheme. Calculation of sine and cosine at individual harmonics is not needed, which reduces the computation burden of the microcontroller. However, direct power extraction is not available as in the DQH control scheme.

4.3 | Unbalanced synchronous reference frame control with multi-resonant harmonic controller

The unbalanced synchronous reference frame control has inherent power extraction and frequency adaption but requires large computational effort for harmonic compensation. The resonant regulator is much simpler and imposes only a small computation burden whereas frequency adaption is still maintained with the double integrator form. Thus, the benefits of the two controllers can be combined. Figure 6 shows the unbalanced synchronous reference frame current regulator with multi-resonant harmonic controller which is referred as the DQMR controller. This proposed control scheme overcomes the lack of power extraction capability of the PMR regulator. The multi-resonant harmonic controller also reduces computation effort due to the absence of sine and cosine calculations at harmonic frequencies. The transfer function of this DQMR controller in the stationary reference frame is also identical to the DQH and PMR controllers given in Equation (33).

5 | CONTROLLER DESIGN AND PERFORMANCE ANALYSIS

5.1 | Controller design

The stationary reference frame equivalence of the DQH, PMR and DQMR are identical. Thus, the regulator parameters are
FIGURE 13  Experimental results of the DQ control scheme with a deadtime of 1 μs under the distorted PCC voltage

designed in the stationary reference frame. The parameters \( K_p \) and \( K_i \) for the fundamental component are determined first. Then, the parameters \( K_i \) are calculated as ratios of \( K_i \) with a satisfied stability margin. The crossover frequency \( \omega_c \) is normally chosen to be located between the grid frequency and the resonant frequency of the LCL filter. In such a frequency range the LCL filter characteristic is similar to the L filter. The open loop transfer function can be simplified as:

\[
G_o(s) = K_p \left( 1 + \frac{K_i}{K_p} \times \frac{s}{\omega_c^2 + \omega_c^2} \right) \frac{1}{sL_\alpha + R_i} 
\]

where \( L_\alpha = L_1 + L_2 \) and \( R_i = R_1 + R_2 \). The phase angle of the forward path can be written as:

\[
\angle G_o(j\omega_c) \approx \tan^{-1}\left( \frac{K_p}{K_i} \right) - \frac{\pi}{2} - \omega_c T_d - \tan^{-1}\left( \frac{\omega_c L_i}{R_i} \right) 
\]

\[ (34) \]

At \( \omega_c \), the phase contributions from terms \( \omega_c K_p / K_i \) and \( \omega_c L_i / R_i \) approach \( \pi / 2 \). Therefore, the maximum crossover frequency \( \omega_{c,\text{max}} \) for a given phase margin \( \phi_m \) can be written as [15]:

\[
\omega_{c,\text{max}} = \frac{\pi/2 - \phi_m}{T_d} 
\]

\[ (35) \]

The possible maximum \( K_p \) is approximated as:

\[
K_p \approx \frac{\omega_{c,\text{max}} L_\alpha}{V_D} = \frac{\pi/2 - \phi_m}{T_d} \times \frac{L_\alpha}{V_D} 
\]

\[ (36) \]
The value of $K_{i1}$ is then determined from

$$K_{i1} = \frac{\omega_{c,\text{max}}}{10K_p}$$

at which $\tan^{-1}(\omega_c K_p / K_{i1}) = 85^\circ$.

Table 1 lists the converter parameters used in this study. The LCL filter has a resonant frequency $f_r$ of 5.03 kHz, which lies in the stable frequency range between $f_s/6$ to $f_s/2$ for the grid current feedback [37]. The phase margin $\phi_m$ is normally selected around 45° to achieve a higher crossover frequency ensuring a fast dynamic performance. However, adding the harmonic controllers reduces the phase margin $\phi_m$. Thus, the phase margin is conservatively selected at $\phi_m = 60^\circ$. The maximum crossover frequency is found at $\omega_{c,\text{max}} = 2222\pi$ rad/s with $T_d = 75$ μs, at which the proportional gain $K_p$ and the integral gain $K_{i1}$ are calculated from Equations (29) and (30).

Figure 7 shows the comparison between the frequency response of the open loop transfer function with and without the harmonic controllers, orders 3rd to 13th. It is found that setting $K_m = K_{i1}/3$ makes all the selected harmonic gains greater than 100 dB, whereas the phase margin reduces to 40° still high enough to ensure a stable operation. The crossover frequency slightly shifts to $\omega_{c1} = 2468\pi$ rad/s. There are two other crossover frequencies $\omega_{c2}$ and $\omega_{c3}$ around $f_r$ of the LCL filter, of which phase margins $\phi_{m2}$ and $\phi_{m3}$ satisfy the stability criterion [37].

5.2 Analysis of harmonic rejection performance

The controller parameters obtained from the previous section are used to analyse the harmonic rejection performance.
TABLE 3 Experimental results of the prototype inverter under the sinusoidal PCC voltage

| Control methods | tDT (μs) | Iref (A) | THD (%) | Ppcc (W) | λ |
|----------------|---------|---------|---------|----------|---|
| DQ            | 1       | 6.617   | 1.51    | 1523     | -0.9996 |
| DQ            | 4       | 6.636   | 1.67    | 1524     | -0.9979 |
| PR            | 1       | 6.625   | 1.64    | 1523     | -0.9996 |
| PR            | 4       | 6.632   | 1.66    | 1524     | -0.9978 |
| DQH           | 1       | 6.621   | 0.69    | 1522     | -0.9997 |
| DQH           | 4       | 6.622   | 2.09    | 1524     | -0.9995 |
| DQMR          | 1       | 6.622   | 0.72    | 1523     | -0.9997 |
| DQMR          | 4       | 6.626   | 2.20    | 1524     | -0.9995 |
| PMR           | 1       | 6.614   | 0.75    | 1523     | -0.9997 |
| PMR           | 4       | 6.632   | 2.29    | 1524     | -0.9995 |

Figure 8 compares the frequency responses of the output admittance with the PCC voltage \( Y_{pcc}(s) = \frac{i_g(s)}{v_{pcc}(s)} \) plotted from Equations (11) and (12). If the fundamental component controller without \( r_{pcc}(t) \) feedforward (dotted line) is used, the current control loop is susceptible to the harmonics of \( Y_{pcc}(s) \) with a damped response at the frequency of \( (L_1C_1)^{-1}/2 \). When applying the feedforward of \( r_{pcc}(t) \) (dashed line), the current control system becomes less susceptible to the voltage harmonics in the frequency range below the crossover frequency \( \omega_{c1} \). However, the control system amplifies the voltage harmonics above \( \omega_{c1} \). It should be noted that the delay in the feedforward path is excluded for simplification as explained in Section 2.2. Thus, the experimental performance can be slightly more deteriorated than the analysis. When applying the harmonic controllers, the current control system attenuates \( r_{pcc}(t) \) at the selected harmonic orders, which is far more effective than using the voltage feedforward method. If the harmonic controller is used together with the voltage feedforward, it is possible that the high order selective harmonic compensator close to \( \omega_{c1} \) could trigger a resonance.

Figure 9 shows the frequency response of the output admittance with the deadtime voltage \( Y_{DT}(s) = \frac{i_g(s)}{v_{DT}} \). It indicates that the grid current is also susceptible to the converter deadtime. Applying the selective harmonic controller, the grid current is effectively attenuated from such deadtime harmonics.

6 PERFORMANCE VALIDATION AND DISCUSSION

6.1 Experimental setup

The proposed control methods were validated with a 1.5-kVA grid-connected inverter using the parameters listed in Table 1, and the experimental setup is displayed in Figure 10. The control schemes were implemented on a Texas Instruments 32-bit TMS320F28069 microcontroller. The integral regulators in the unbalanced synchronous reference frame control were discretised using the backward difference approximation. The forward integrator of the resonant regulator was discretised using the forward difference approximation whereas the feedback integrator was approximated by the backward difference approximation [17]. Internal signals inside the microcontroller were sent to 14-bit digital- to-analogue converters for monitoring on an oscilloscope. An inverse Park transformation PLL with a bandwidth of 10 Hz was used for grid synchronisation and frequency adaptation. A distorted PCC voltage with THD of 5.74% was emulated by a Chroma 61860 60-kVA grid simulator with the voltage harmonics, as shown in Table 2. The deadtime was adjusted on the microcontroller. The current control loop was set to inject a power of 1.5 kW with a unity power factor (\( \lambda = 1.0 \)) into the grid. A Yokogawa WT-3000E power analyser was used to measure voltage and current and power components of the prototype inverter.

It has been reported that the grid current with the unipolar modulation scheme is more distorted than that with the bipolar modulation scheme. Therefore, the unipolar modulation was selected in this investigation so as to validate the proposed harmonic compensation schemes. The parameters of the LCL filter also play an important role in attenuation of the grid current harmonics. The ratio \( L_1/(L_1 + L_2) \) has been reported to be optimised at 0.5 with a negligible deadtime. With a larger deadtime, the optimised ratio \( L_1/(L_1 + L_2) = 0.5 \), the harmonic controllers will play the main role in attenuation of the grid current harmonics with a large deadtime. The inductor \( L_1 \) is normally selected to create the ripple current to be less than 10–20% of the rated current. A higher current ripple will cause an excessive loss in the inductor winding and in the inductor core. From that, the inductor \( L_2 \) and the capacitor \( C_1 \) are chosen to have a resonant frequency in the stable range between \( f_s/6 \) to \( f_s/2 \) for the grid current feedback.
6.2 | Harmonic mitigation

Figure 11 shows the grid current $i_g(t)$ and the grid current error $e_{iA}(t)$ under the sinusoidal PCC voltage using the unbalanced synchronous reference frame control without the voltage feedforward. With the deadtime $t_{DT} = 1 \mu s$, the grid current distortion $THD_i = 1.64\%$ is within the IEEE1547 limit. Increasing the deadtime to $t_{DT} = 4 \mu s$ gives rise to the low order dominant harmonics, especially the third harmonics, exceeding the IEEE1547 limit. Similar results were found for the PR controller as shown in Figure 12 with slightly higher current distortions. The DQ and PR controllers are susceptible to the PCC voltage harmonics with the deadtime $t_{DT} = 1 \mu s$ as displayed in Figures 13 and 14. Tables 3 and 4 summarise the experimental results under the sinusoidal PCC voltage and the distorted PCC voltage respectively. The current distortions under the two control methods are far greater than the IEEE1547 limit. Applying the feedforward of the PCC voltage on the DQ and PR controllers (then referred to as the DQFW and PRFW controllers respectively) mitigates the grid current to be in compliance with the IEEE1547 standard.

However, the voltage feedforward method cannot effectively attenuate the higher order harmonics as analysed in Figure 8. This agrees with the amplitudes of the current harmonic orders 9th, 11th and 13th shown in both Figures 13 and 14 that are greater than the 7th order even though the voltage harmonic orders 7th, 9th, 11th and 13th have an identical amplitude of 1%.

The grid current is far more distorted under the distorted PCC voltage and the deadtime $t_{DT} = 4 \mu s$ as depicted in Figure 15 for the DQ control scheme and Figure 16 for PR control scheme. The strong presence of the PCC voltage harmonics and the large deadtime causes a large current error $e_{iA}$ and gives rise to the grid current distortion $THD_i = 38.86\%$ for the DQ control and $THD_i = 39.25\%$ for the PR control as listed in Table 4. The application of the PCC voltage feedforward for the DQFW and PRFW control methods partially suppresses the grid current harmonics caused by the PCC voltage harmonics. As analysed in Figure 8, the PCC voltage feedforward cannot attenuate the harmonics due to the deadtime. Thus, the dominant grid current harmonics, orders third and fifth of both the DQFW and PRFW schemes are still greater than the IEEE1547.
The grid frequency was adjusted between 48 and 52 Hz. Figure 20 shows the transient responses of $i_g(t)$ and $e_{ia}(t)$ when the grid frequency was stepped from 48 to 52 Hz under the distorted PCC voltage with the deadtime $\delta_{DT} = 1 \, \mu s$ at the nominal power of 1.5 kW. The current error of the DQH control method takes about 120 ms to recover back to zero, whereas the estimated frequency reaches the steady state after 220 ms. The responses of the $i_g(t)$ and $e_{ia}(t)$ under the DQMR and PMR methods have the same characteristics as the DQH control scheme. Grid frequency changes between 48 to 52 Hz do not affect the grid current distortion at the steady state, as illustrated in Figure 21. The three control schemes accommodate tracking of the grid frequency. The tracking response is mainly governed by the PLL. In fact, frequency change in an electrical grid is not as extreme as that illustrated in Figure 20. The inverse Park transformation PLL used in this study is sufficient for providing the estimated frequency for the current control system.
Transient performance and power extraction capability

This section simultaneously validates the transient performance and the power extraction capability of the proposed control schemes. The prototype inverter was initially set to inject 1 kW of active power into the grid with the deadtime \( t_{DT} = 1 \mu s \), then a step of 1-kVar (inductive) reactive power was applied at \( \omega t \approx -\pi/2 \). Figure 22 shows the transient performance of the DQ and PR control schemes under the sinusoidal PCC voltage. The grid current error signals \( e_{ia}(t) \) of the two control methods smoothly recover to the steady state within approximately 1 ms after the step of the reactive power reference. However, a steady state error in the grid current is present for both the DQ and PR controllers as discussed in Section 6.2. The transient results of the DQH, DQMR and PMR schemes under the sinusoidal and distorted PCC voltage are shown in Figures 23 and 24 respectively. The grid current error signals \( e_{ia}(t) \) of the three control methods are very close which proves that they are identical in the stationary reference frame. Figures 23 and 24 indicate that the distorted PCC voltage has no effect on the transient response of the grid current \( i_g(t) \) and the grid current error \( e_{ia}(t) \) waveforms. The harmonic controllers introduce oscillations in the grid current during the transient period, and the grid current error signals \( e_{ia}(t) \) reach zero within a grid voltage period. Thus, design of the grid current with harmonic mitigation for optimal dynamic and steady state performances will be a topic of interest.

The signals \( i_d'(t) \) of the DQH and DQMR control methods are available for calculation of the instantaneous reactive power \( q(t) \) using Equation (6), whereas the instantaneous active power \( p(t) \) is determined from the signal \( i_d'(t) \) which is not shown in Figures 23 and 24. The current signals \( i_d'(t) \) and \( i_q'(t) \) are not available for the PMR controller. The instantaneous active and
reactive powers can be controlled through the power references $p_{ref}(t)$ and $q_{ref}(t)$ which become correct once the grid current error to reach zero. In other words, the PMR control has an indirect power control capability.

7 | DISCUSSION

The harmonic mitigating performance of the DQH scheme is slightly better than both the DQMR and PMR control techniques in every condition. The discrete transfer function of the resonant controller is given by:

$$G_{rn}(z) = \frac{K_{in} T_s (z^{-1} - z^{-2})}{1 + (n^2 \omega^2 T_s^2 - 2) Z^{-1} + z^{-2}}. \quad (39)$$

Figure 25 compares the frequency responses of the 13th order resonant controller in the continuous time domain and in the discrete time domain using Equation (39). The resonant peak of the discrete controller is approximately shifted from the continuous controller by +1.1 Hz, which is caused by the computational delay [17]. This shifted resonant peak reduces the controller gain at the target frequency which deteriorates the harmonic mitigating performance. This effect becomes prominent as the increasing resonant frequency. The computational delay effect can be optimally compensated by factors $\cos(2n\omega T_s)$ and $\sin(2n\omega T_s)$. However, this will result in the need for a more complicated resonant controller. Direct implementation of the resonant controller using the Tustin approximation with frequency pre-wrapping provides a better performance. However, frequency adaptation is not applicable [17].

The harmonic mitigating results of the three control techniques are comparable to the results presented in the literature [9]. With the grid voltage THD of 5.08% and $\tau_{JT} = 3.25$ μs [9], the current distortion of the 2-kVA inverter using a PMR

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**FIGURE 18** Experimental results of the DQMR control scheme under the distorted PCC voltage
controller with orders 3rd to 11th was found to be approximately 4%, and 1.2% for a PR controller plus a repetitive controller. However, the PR applied together with the repetitive controller in that study, lacks the power extraction and frequency adaptation functionalities which are already included in the DQH and DQMR control techniques in this study. The unbalanced synchronous reference frame control can be implemented together with a repetitive controller to improve the higher order harmonic mitigating performance.

Although the DQH control technique has the best performance and functionality, it requires a larger computational effort. Table 5 compares the number of mathematical operations for the DQH, PMR and DQMR methods. Sine/cosine calculation of the fundamental component is not included as it is embedded in the PLL. The DQH control scheme uses a large number of mathematical operations which have an execution time of 9.08 μs for the TMS320F28069 microcontroller running at 90 MHz. The calculation of sine and cosine at the harmonic orders 7th, 9th, 11th and 13th takes about 4.44 μs to complete. The PMR control technique has the smallest computation time with a sacrifice of the power extraction capability. The PMR control method is then suitable for simple applications such as

| Mathematic operations          | Control methodology |
|--------------------------------|---------------------|
| Multiplications               | DQH  | PMR  | DQMR |
| Additions/subtractions        | 51   | 23   | 34   |
| Sin/cosine calculations       | 37   | 31   | 37   |
| Saturation limits             | 10   | –    | –    |
| Execution cycles              | 2    | –    | 2    |
| Execution time                | 817  | 231  | 315  |
| THD                            | 2.31%| 0.76%| –    |
| IEEE1547 limit                | –    | –    | –    |

**TABLE 5** Computational effort of the control schemes

![FIGURE 19](image_url) Experimental results of the PMR control scheme under the distorted PCC voltage
Multiple unbalanced synchronous reference frame control is presented for current control of grid-connected converters with selective harmonic mitigation, frequency adaptation and power extraction capabilities. The proposed control technique is compared with the proportional multi-resonant controller. A new controller combining the unbalanced synchronous reference frame control for regulating the fundamental component current together and a bank of multi-resonant controllers for harmonic compensation is also presented. The three control schemes have identical characteristics in the physical stationary reference frame.

Experimental results of a 1.5-kVA grid-connected LCL-filtered converter under a highly distorted voltage at the point of common coupling indicate that the three control techniques effectively attenuate the grid current harmonics at the selected orders. Even with a large converter deadtime of 4 μs, the grid current is still maintained within the IEEE1547 limit. The DQMR control scheme is a good alternative to the DQH control technique, requiring only a slightly greater computation time than the PMR controller, while retaining the power extraction capability as the DQH technique.

CONCLUSION

FIGURE 20  Transient responses of the DQH, PMR and DQMR control schemes under the distorted PCC voltage and the deadtime of 1 μs when the grid frequency changing from 48 to 52 Hz.

FIGURE 21  Grid current THD of the DQH, PMR and DQMR control schemes under the distorted PCC voltage with the deadtime of 1 and 4 μs and the grid frequency from 48 to 52 Hz.

FIGURE 22  1-kVar step reactive power responses of the DQ and PR control schemes under the sinusoidal PCC voltage and the deadtime of 1 μs.
multiple unbalanced synchronous reference frame control has the best harmonic mitigating performance but with the largest computational effort. The proportional multi-resonant control requires the shortest execution time but with slight inferior performance and lack of the power decoupling capability. The proposed combined control technique requires a slightly larger computational burden than the proportional multi-resonant control while retaining the same functionalities as the multiple unbalanced synchronous reference frame control.

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FIGURE 23 1-kVar step reactive power responses of the DQH, PMR and DQMR control schemes under the sinusoidal PCC voltage and the deadtime of 1 μs

FIGURE 24 1-kVar step reactive power responses of the DQH, PMR and DQMR control schemes under the distorted PCC voltage and the deadtime of 1 μs

FIGURE 25 Magnitude of the resonant controller order 13th plotted from the continuous and discrete time domains
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