A Two State Feedback Active Damping Strategy for the \textit{LCL} Filter Resonance in Grid-Connected Converters

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Abstract

A novel active damping strategy for the \textit{LCL} filter resonance is proposed using the grid current and the capacitor voltage. The proposed technique is deduced in the continuous time domain and a discussion for its discrete implementation is presented. According to the proposed technique, instability of the open loop system, which results in non-minimum phase behavior, can be avoided over wide range of resonant frequencies. Moreover, straightforward co-design steps for both the fundamental current regulator and the active damping loops can be used. A numerical example along with experimental results are introduced to validate the proposed strategy performance over wide range of resonant frequencies.

Key words: \textit{LCL} filter, active damping, resonance, converters, grid

I. INTRODUCTION

Due to their higher attenuation for switching harmonics with a lower size and weight, \textit{LCL} filters are widely used with grid-connected converters to limit the harmonic contents of the injected grid current to comply with the grid codes; i.e. IEEE 519-1992 [1]. However, the inherent resonance of the \textit{LCL} filters represents a challenge for control system designers. Damping techniques have to be adopted to cope with this challenge. With discrete implementation, closed loop system stability can be maintained by the inherent damping characteristic of a single grid current control loop for resonant frequencies greater than one-sixth of the control frequency [2]. However, this strategy gives rise instability due to resonant frequency variations which are likely to occur particularly in weak grids where the grid inductance changes significantly [3]. Passive damping, by using a resistor, was used to cope with this issue [4]. However, it causes power losses. Thus active damping (AD) by modifying the control algorithm is preferred [5].

Number of active damping techniques have been discussed in the literature [6]-[20]. A cascaded filter in the current control loop was used in [6] and [7]. However, this method is highly sensitive to resonant frequency variations. In addition, it causes a reduction in the system bandwidth. To overcome these issues, an inner feedback loop of one of the filter states has been employed to produce a damping effect [8]-[20]. A proportional feedback of the filter capacitor current was employed in [8]-[10]. To stabilize the closed loop system, it was proved that excitation of unstable open loop poles is mandatory for resonant frequencies greater than one-sixth of the control frequency [11]-[12]. This non-minimum phase behavior can decline the system performance especially when selective harmonic mitigation is of concern [11]. Modified feedback loops of the capacitor current have recently been proposed to avoid this behavior over wider range of resonant frequencies [11]-[15]. However, a high precision current sensor or a complicated observer loop is needed [16]. The capacitor voltage differentiation can be used to produce a damping effect. However, this technique results in noise amplification. To cope with this issue, a lead-lag network has been adopted to behave as a differentiator around the resonant frequency [17]-[18]. However, as shown in [17], this method can be used effectively over limited range of resonant frequencies between 1/3.2 and 1/3.4 of the control frequency.
A high pass filter (HPF) is employed in [19] instead of the cannot be implemented practically due to noise amplification. A high pass filter (HPF) is employed in [19] instead of the s^2 term which cannot be implemented practically due to noise amplification. This technique is further discussed in [20] where it was shown that non-minimum phase behavior can be avoided for resonant frequencies up to a theoretical limit of 0.27 of the sampling frequency. However, the co-design steps of this HPF along with the fundamental current regulator are very complicated since many iterations are needed. Moreover, avoiding non-minimum behavior at high resonant frequencies requires increasing the cutoff frequency of the HPF to its maximum practical limit of 0.5 of the sampling frequency (Nyquist frequency). However, using such high value can deteriorate the HPF performance. This in turn, makes it unsuitable for practical implementation. Due to this practical limitation, and during the performance verification presented in [20], non-minimum phase behavior has been avoided up to a maximum resonant frequency of about 0.24 of the sampling frequency.

In this paper, two feedback loops of the grid current and the capacitor voltage are proposed as a new active damping strategy. By using the proposed strategy, the non-minimum phase characteristics can be avoided over a wide range of resonant frequencies. Moreover, straightforward co-design procedures for both the fundamental current regulator and the active damping loops are proposed. For reduced number of sensors, virtual flux technique can be employed to estimate the capacitor voltage [18]. However, this is not adopted here.

Following this introduction, section II presents the continuous time domain derivation of the proposed active damping strategy. In section III, discrete implementation of the proposed strategy is presented along with the co-design steps of the fundamental current regulator and the active damping loops. In section IV, a numerical example along with experimental work are introduced to verify the proposed strategy performance at different resonant frequencies. Finally, section V presents some conclusions.

II. PROPOSED ACTIVE DAMPING STRATEGY

A. System Description

Fig. 1 shows a single phase inverter connected to the grid through an LCL filter. The block diagram of the capacitor-current-based active damping system is shown in Fig. 2, where a proportional feedback (H_d) of the capacitor current is used to actively damp the filter resonance. The un-damped filter transfer function is denoted as G_c(s) and is expressed in (1), where \( \omega_{res} \) - expressed in (2) - is the LCL filter resonant frequency. A proportional resonant (PR) controller with a transfer function of \( G_i(s) \) – expressed in (3) – is employed for fundamental current regulation.

\[
G_i(s) = \frac{s}{L_i g_i s (s^2 + \omega_{res}^2)}
\]

(1)

\[
\omega_{res} = \sqrt{\left( L_i + g_i \right) / \left( C L_i g_i \right)}
\]

(2)

\[
G_c(s) = K_p + \frac{2 \omega_c s}{s^2 + 2 \omega_c s + \omega_c^2}
\]

(3)

where \( \omega_c \) and \( \omega_i \) are the fundamental frequency and the bandwidth of the resonant part of the PR regulator, respectively.

According to Fig. 2, the transfer function of the actively damped filter is expressed in (4).

\[
F_{ad}(s) = \frac{1}{C L_i g_i s \left( s^2 + \frac{H_d}{L_i} s + \omega_{res}^2 \right)}
\]

(4)

In Figs. 3(a) till 3(d), the capacitor-current-based active damping system is manipulated using signal flow graph manipulation. In Fig. 3(a), the capacitor current is replaced by the difference between the inverter output current and the grid injected current. With further manipulation, it is shown in Fig. 3(c) that the capacitor current feedback is equivalent to using three feedback loops of the grid current (i_g), the capacitor voltage (v_c), and the modulated inverter voltage (v_i). By further manipulation, the modulated voltage feedback is augmented in the main loop as a HPF, which is denoted as \( G_h(s) \) and expressed in (5), with a cut off frequency of \( \omega_h = H_d / L_i \); this system is shown in Fig. 3(d). The typical range for \( \omega_h \) can be calculated by expressing the transfer function of the actively damped filter \( (F_{ad}) \) in terms of \( \omega_h \) and writing it in a standard form as in (6). The damping ratio \( \zeta \) is typically around 0.7 [8], [21]. Therefore, the typical range for \( \omega_h \) can be determined as \( \omega_{res} < \omega_h < 2 \omega_{res} \).

\[
G_h(s) = \frac{1}{\omega_h} \cdot \frac{s}{1 + s / \omega_h}
\]

(5)

\[
F_{ad}(s) = \frac{1}{C L_i g_i s \left( s^2 + \omega_h s + \omega_h^2 \right)} = \frac{1}{C L_i g_i s \left( s^2 + 2 \omega_{res} s + \omega_{res}^2 \right)}
\]

(6)

Note that the capacitor voltage feedback is an integrator, denoted as \( G_h(s) \) in Fig. 3(d), with a time constant of \( L_i \).
The presence of the HPF ($G_i$) in cascade with the main control loop can deteriorate the system disturbance rejection capability. This deterioration can be determined by comparing the system transfer function to the grid voltage ($v_g$). The HPF, expressed in terms of a new variable ($K_d$) as in (9) and (10), respectively.

$$\begin{align*}
G_{ad}(s) &= \frac{sL_i}{1+s/\omega_h} \\
F_{new}(s) &= \frac{G_{ig}(s)}{1-G_{ad}(s)G_{ig}(s)(1+sL_iG_i(s))} \\
&= \frac{(1+s/\omega_h)}{s^2(\frac{cL_iL_g}{\omega_h}^2+\frac{cL_iL_g}{\omega_h}+\frac{\omega_i}{\omega_h})}
\end{align*}$$  \\
\tag{8}

The $s^2$ term in the denominator of $F_{new}(s)$ results in a constant phase of -180° in the open loop bode plot. This in turn, can dramatically deteriorate the phase margin. As a result, more modifications are necessary.

2. Both the gain of $G_{ad}(s)$ and the time constant of $G(s)$ are expressed in terms of a new variable ($K_d$) as in (9) and (10), respectively.

$$\begin{align*}
G_{ad}(s) &= \frac{(K_d-L_g)s}{1+s/\omega_h} \\
G_i(s) &= \frac{1}{(K_d-L_g)s}
\end{align*}$$  \\
\tag{9}
\tag{10}

Substituting (9) and (10) into (8), $F_{new}(s)$ is re-written as:

$$\begin{align*}
F_{new}(s) &= \frac{(1+s/\omega_h)}{s^2(\frac{cL_iL_g}{\omega_h}^2+\frac{cL_iL_g}{\omega_h}+\frac{\omega_i}{\omega_h})}
\end{align*}$$  \\
\tag{11}

Using Routh’s criteria, $K_d$ has to follow the constraint in (12) to guarantee the open loop system stability and hence minimum phase behavior.

$$0 < K_d < (L_i + L_g)$$  \\
\tag{12}

To generalize the following analysis, $K_d$ is expressed in terms of the above maximum limit ($L_i + L_g$) as in (13), where $0 < \beta_d < 1$ for a stable open loop system.

$$K_d = \beta_d (L_i + L_g)$$  \\
\tag{13}

Substituting (13) into (11), the actively damped filter of the proposed system is finally expressed in (14).

$$F_{new}(s) = \frac{(1+s/\omega_h)}{s^2(\frac{cL_iL_g}{\omega_h}^2+\frac{cL_iL_g}{\omega_h}+\frac{\omega_i}{\omega_h})-s\beta_d(L_i+L_g)}$$  \\
\tag{14}

III. DISCRETE IMPLEMENTATION

A. System Discretization

The discrete system representation of the proposed active damping strategy is shown in Fig. 5 where the DSP delay is represented by one sample delay. Using Tustin approximation with pre-warping at the fundamental frequency, the discrete PR regulator is determined in (15) where $T_i$ is the sampling period.

$$G_c(z) = K_p + K_r \frac{\sin(\omega T_i)}{2\omega} \frac{z^{2\pi-1}}{(z^2-2\cos(\omega T_i)z+1)}$$  \\
\tag{15}

In addition to $G_d(s)$, expressed in (1), two other transfer functions should be defined for system discretization:
and (21), respectively. According to this representation, both the actively damped filter transfer function \( F_{\text{new-d}}(z) \) and the loop transfer function \( T_{\text{loop-d}}(z) \) are expressed in (24) and (25), respectively.

\[
F_{\text{new-d}}(z) = \frac{(1+s/\omega_d)G_d(s)}{CL_d\beta z^2+\omega_{\text{res}}^2(1+s/\omega_d)-s\beta_d (L+L_g)G_d(s)}
\]

\[
T_{\text{loop-d}}(z) = G_c(z)F_{\text{new-d}}(z)
\]

It was shown in [12] that the resonant frequency changes with discrete implementation. The new resonant frequency will be denoted as \( \omega_{\text{res-ad}} \). At this resonant frequency the gain of \( F_{\text{new-d}} \) can be approximately expressed in (26).

\[
|F_{\text{new-d}}(j\omega_{\text{res-ad}})| \approx \frac{|1+j\omega_{\text{res-ad}}/\omega_h|}{1-j\omega_{\text{res-ad}}\beta_d (L+L_g)} \quad (26)
\]

According to (26), higher values of \( \omega_h \) should be used to acquire better damping effect. Theoretically, for discrete implementation, \( \omega_h \) can be extended up to 0.5\( \omega_0 \) (Nyquist sampling theory, where \( \omega_0 \) is the control frequency in rad/sec). However, such high value can deteriorate the discretization process. A value of \( \omega_h = 0.4\omega_0 \) is adopted here.

Since the resonant gain of the PR regulator is mainly effective at the fundamental frequency, the PR controller can be approximated as (27).

\[
G_c(\omega) = \begin{cases} K_p & \text{for } \omega > \omega_0 \\ K_r & \text{for } \omega = \omega_0 \\ 0 & \text{for } \omega < \omega_0 \end{cases}
\]

At the crossover frequency (\( \omega_c \)), which should be sufficiently higher than \( \omega_0 \) and below both \( \omega_{\text{res}} \) and the adopted \( \omega_h (0.4\omega_0) \), the loop gain can be approximated as (28).

\[
T_{\text{loop-d}}(j\omega_c) = \frac{K_p}{\omega_c (L+L_g)} \frac{e^{-j1.5Ts\omega_c}}{|1-j\omega_c \beta_d e^{-j1.5Ts\omega_c}|} = 1 
\]

Using Trigonometry, this gain is reduced to (29).

\[
T_{\text{loop-d}}(j\omega_c) = \frac{K_p}{\omega_c (L+L_g)} \frac{1}{|A_c e^{j\theta_c}|} = \frac{K_p}{\omega_c (L+L_g)A_c} = 1 
\]

where

\[
A_c = \sqrt{1 + \beta_d^2 - 2\beta_d \cos(1.5Ts\omega_c)}
\]

\[
\theta_c = \sin^{-1} \frac{-\beta_d \sin(1.5Ts\omega_c)}{A_c}
\]

Hence, for certain value of \( \beta_d, K_p \) should be calculated as in (31) to obtain certain crossover frequency.

\[
K_p = \omega_c (L+L_g)A_c
\]

Substituting (31) into (23), the loop transfer function is expressed in (32)

\[
T_{\text{loop-d}}(z) = A_c \omega_c (L+L_g) F_{\text{new}}(z)
\]
At the fundamental frequency, the loop gain can be approximated as in (33).

\[ |T_{lo}op-d(j\omega_o)| = \frac{K_r}{\omega_o(L_i+L_g)A_o} \quad (33) \]

where \( A_o = \sqrt{1+\beta_d^2 - 2\beta_d \cos(1.5T_o\omega_o)} \)

This is expressed in dB in (34) from which \( K_r \) can be determined from (35) for certain fundamental loop gain (\( T_{fo} \)).

\[ T_{fo} = 20 \log_{10}\frac{K_r}{\omega_o(L_i+L_g)A_o} \quad (34) \]

\[ K_r = \omega_o(L_i + L_g)A_o \cdot 10^\frac{T_{fo}}{20} \quad (35) \]

Using the above-derived expressions, the following steps are proposed to co-design the control system parameters.

1. Plot the pole map of \( F_{new}(z) \), expressed in (22), by sweeping \( \beta_d \). Select \( \beta_d \) so that it corresponds to the farthest resonant poles inside the unit circle to achieve the best damping.
2. For a certain value of the fundamental loop gain (\( T_{fo} \)) along with the selected value for \( \beta_d \), use (35) to determine \( K_r \).
3. For a certain value of the crossover frequency (\( \omega_c \)) along with the selected value of \( \beta_d \), use (31) to determine \( K_p \).
4. Plot a bode diagram for the loop transfer function expressed in (23). Check the resonant peak. If the resonant peak is more than 0 dB, then decrease the pre-specified crossover frequency (\( \omega_c \)) and repeat steps 3 and 4.

IV. VERIFICATION

A. Numerical Example

Table I lists the parameter values of the grid-connected inverter shown in Fig. 1. Four capacitance values, corresponding to resonant frequencies of 0.143\( \omega_o \), 0.179\( \omega_o \), 0.209\( \omega_o \) and 0.241\( \omega_o \), are used to verify the performance of the proposed system over a wide range of resonant frequencies with respect to the control frequency. These resonant frequencies are denoted as \( \omega_{res1} \), \( \omega_{res2} \), \( \omega_{res3} \) and \( \omega_{res4} \), respectively. The HPF cut off frequency (\( \omega_{hi} \)) value is taken as 0.4\( \omega_o \) to mitigate the resonant peak as much as possible. In addition, a value of 60 dB is adopted for the fundamental loop gain (\( T_{fo} \)). Finally, an initial value for the crossover frequency of 0.3 of each corresponding resonant frequency is adopted.

Using the tuning steps presented in the last section, a pole-map of \( F_{new}(z) \) is plotted with variation of \( \beta_d \). These pole maps are plotted in Figs. 7(a), 8(a), 9(a) and 10(a) for the resonant frequencies \( \omega_{res1} \), \( \omega_{res2} \), \( \omega_{res3} \) and \( \omega_{res4} \), respectively. To achieve the best damping effect, the values of \( \beta_d \) corresponding to the farthest resonant poles inside the unit circle are selected. These values are determined as 0.55, 0.45, 0.3 and 0.15 for \( \omega_{res1} \), \( \omega_{res2} \), \( \omega_{res3} \) and \( \omega_{res4} \), respectively. Using the selected values of \( \beta_d \) along with the pre-specified values of \( \omega_o \) and \( T_{fo} \), the corresponding values of \( K_p \) and \( K_r \) are determined from (31) and (35), respectively.
For $\omega_{\text{res}1}$ and $\omega_{\text{res}2}$, Figs. 7(b) and 8(b) show bode plots of the loop transfer function, expressed in (23), respectively. It is shown that the resonance peak is less than 0 dB. For $\omega_{\text{res}3}$ and $\omega_{\text{res}4}$, it is found that the frequency response exhibits a resonant peak of more than 0 dB. To overcome this issue, a reduction in the crossover frequency has to be adopted. For $\omega_{\text{res}3}$, it is found that a reduction of the crossover frequency of 0.120$\omega_{\text{res}3}$ can reduce the resonant peak to less than 0 dB. However, for $\omega_{\text{res}4}$, a large crossover frequency reduction is required to obtain a resonant peak of less than 0 dB. Such a reduction can deteriorate the system dynamic performance.

Moreover, the phase lag introduced by the PR controller at low frequencies dramatically reduces the phase margin. Therefore, only a reduction of the crossover frequency to 0.1$\omega_{\text{res}4}$ is adopted. Figs. 9(b) and 10(b) show the frequency responses for $\omega_{\text{res}3}$ and $\omega_{\text{res}4}$ respectively.

Table I summarizes the designed control parameters and the achieved performance of the phase margin (PM), $\omega_c$ and $T_{\text{fo}}$. These results indicate the well damped performance of the proposed method over a wide range of resonant frequencies while meeting the pre-specified values of $\omega_c$ and $T_{\text{fo}}$.

### Table I

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Quantity</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P$</td>
<td>Rated power</td>
<td>400 W</td>
</tr>
<tr>
<td>$V_s$</td>
<td>Grid voltage</td>
<td>100 V</td>
</tr>
<tr>
<td>$F_s$</td>
<td>Grid Frequency</td>
<td>50 Hz</td>
</tr>
<tr>
<td>$V_{\text{DC}}$</td>
<td>DC Voltage</td>
<td>200 V</td>
</tr>
<tr>
<td>$L_i$</td>
<td>Inverter side inductance</td>
<td>1.85 mH</td>
</tr>
<tr>
<td>$L_g$</td>
<td>Grid side inductance</td>
<td>1.3 mH</td>
</tr>
<tr>
<td>$C$</td>
<td>Capacitance</td>
<td>16.3 µF, 10.4 µF, 7.6 µF, 5.7 µF</td>
</tr>
<tr>
<td>$F_{\text{sw}}$</td>
<td>Switching Frequency</td>
<td>10 kHz</td>
</tr>
<tr>
<td>$F_s$</td>
<td>Sampling Frequency</td>
<td>10 kHz</td>
</tr>
</tbody>
</table>

In real operation, the grid side inductance ($L_g$) may vary significantly. To investigate system robustness against such variations, the pole maps of the closed loop system $T_{\text{closed}}$, expressed in (36), are plotted in Fig. 11 while sweeping $L_g$ between 100-300% of its original value.

$$T_{\text{closed}}(z) = \frac{T_{\text{loop}}(z)}{1 + T_{\text{loop}}(z)}$$

(36)

For the considered resonant frequencies, it is shown that the closed loop poles move inside the unit circle with an increasing $L_g$. These plots reflect the system robustness against grid inductance variations.
to the existing capacitor voltage/current based AD methods, the limitations of these methods are clarified under the same parameters used in the aforementioned numerical example.

- Capacitor-voltage-based AD method: Fig. 12 shows a discrete representation of this method, where a lead-lag network of \( G_{adv} (z) \) is used for AD. The s-domain counterpart of this network is expressed as \( G_{adv}(s) \) in (37). Using Fig. 12, the discrete loop transfer function can be expressed as (38). It was demonstrated in [17] that this method can behave effectively over the limited range of resonant frequencies between 1/3.2 and 1/3.4 of the sampling frequency \( (\omega_s) \). To emphasize the difficulty of using this method outside specified limits, the AD loop design procedures presented in [17] are used for the resonant frequencies \( \omega_{res} \) and \( \omega_{res2} \) \( (<\omega_s/3.2) \) as follows. The value of \( K_f \) is determined using (39) to achieve a maximum network angle \( (\varphi_{max}) \) of 75 degree at a frequency of \( \omega_{max} = \omega_c \). Then, the minimum value of \( K_f \) is determined as \( (K_{dmin} = L_g/3T_g) \). Following this, the root locus of the closed loop system, expressed in (40), is plotted while sweeping \( K_f \) starting the from \( K_{dmin} \) as shown in Figs. 13(a) and 13(b) for the resonant frequencies \( \omega_{res1} \) and \( \omega_{res2} \), respectively.

\[
G_{ad-v}(s) = K_d C \omega_{res} \frac{s+K_f \omega_{res}}{K_f s+\omega_{res}} \quad (37)
\]

\[
T_{open-v}(z) = \frac{z^{-1}G_c(z)G_{ad}(z)G_{tg}(z)}{1+z^{-1}G_{ad-v}(z)G_{tg}(z)} \quad (38)
\]

C. Comparative Study

To show the superiority of the proposed method compared to the existing capacitor voltage/current based AD methods, the limitations of these methods are clarified under the same parameters used in the aforementioned numerical example.

- Capacitor-voltage-based AD method: Fig. 12 shows a discrete representation of this method, where a lead-lag network of \( G_{adv} (z) \) is used for AD. The s-domain counterpart of this network is expressed as \( G_{adv}(s) \) in (37). Using Fig. 12, the discrete loop transfer function can be expressed as (38). It was demonstrated in [17] that this method can behave effectively over the limited range of resonant frequencies between 1/3.2 and 1/3.4 of the sampling frequency \( (\omega_s) \). To emphasize the difficulty of using this method outside specified limits, the AD loop design procedures presented in [17] are used for the resonant frequencies \( \omega_{res} \) and \( \omega_{res2} \) \( (<\omega_s/3.2) \) as follows. The value of \( K_f \) is determined using (39) to achieve a maximum network angle \( (\varphi_{max}) \) of 75 degree at a frequency of \( \omega_{max} = \omega_c \). Then, the minimum value of \( K_f \) is determined as \( (K_{dmin} = L_g/3T_g) \). Following this, the root locus of the closed loop system, expressed in (40), is plotted while sweeping \( K_f \) starting the from \( K_{dmin} \) as shown in Figs. 13(a) and 13(b) for the resonant frequencies \( \omega_{res1} \) and \( \omega_{res2} \), respectively.

\[
G_{ad-v}(s) = K_d C \omega_{res} \frac{s+K_f \omega_{res}}{K_f s+\omega_{res}} \quad (37)
\]

\[
T_{open-v}(z) = \frac{z^{-1}G_c(z)G_{ad}(z)G_{tg}(z)}{1+z^{-1}G_{ad-v}(z)G_{tg}(z)} \quad (38)
\]
Fig. 13. Closed loop pole map using capacitor-voltage-based AD method with sweeping $K_d$. a) for $\omega_{res1}=0.143\omega_s (K_p=8.47)$, b) for $\omega_{res2}=0.179\omega_s (K_p=10.6)$.

Fig. 14. Closed loop pole map of capacitor-current-based AD method with sweeping $L_g$. a) for $\omega_{res2} (K_p=10.6, H_f=5)$, b) for $\omega_{res3} (K_p=4.96, H_f=1)$.

[9] and listed in below the corresponding plots). It is shown that the closed loop poles are very close to the unit circle. This in turn, demonstrates the ineffective damping performance of this method for resonant frequencies of more than one-sixth of the sampling frequency. Moreover, as shown in the zoomed part, the system stability violates around a certain value of the grid inductance corresponding to a resonant frequency of one-sixth of the sampling frequency. On the other hand, it has been shown that avoiding such non-minimum behavior and high robustness against grid inductance variations can be achieved over a wide range of resonant frequencies using the proposed AD method.

D. Experimental Work

Using the system parameters listed in Table I, a single phase inverter prototype has been built and connected through an LCL filter to an AC power supply to emulate a grid. The control algorithm has been implemented using the PE-Expert3 platform, which consists of a C6713-A DSP development board along with a high-speed PEV board for analog-to-digital conversion and PWM signal generation. To verify the dynamic response, the reference current is stepped up from 2 A ($0.5I_{p,ref}$) to 4 A ($I_{p,ref}$). Using the designed parameters listed in Table II, some tests are carried out with and without the proposed active damping method.

For $\omega_{res1}$, which is lower than one-sixth of the sampling frequency, the system cannot be stabilized without active damping (AD). Thus, removing the active damping loop for this case causes a high oscillatory current as shown in Fig. 15(a). On the other hand, Fig. 15(b) shows the waveforms when using active damping loops.

For $\omega_{res2}$, $\omega_{res3}$ and $\omega_{res4}$, the system can be stabilized without active damping as shown in Figs. 16(a), 17(a) and 18(a). However, it can recognize the dynamic oscillations which are caused by weak damping (there is some damping introduced by the small resistance of the coils). Figs. 16(b), 17(b) and 18(b) show the waveforms when using the proposed active damping loops. It can recognize the mitigation effect of the dynamic oscillations when using the proposed active damping method. This mitigation effect can be further clarified in Figs. 19 and 20. These figures show the spectrum of the grid current for each resonant frequency with and without the proposed active damping method.
Fig. 15. Experimental waveforms of grid current ($i_g$) and grid voltage ($v_g$) for $\omega_{res1} = 0.143\omega_s$ (a) without AD, (b) with AD.

Fig. 16. Experimental waveforms of grid current ($i_g$) and grid voltage ($v_g$) for $\omega_{res2} = 0.179\omega_s$ (a) without AD, (b) with AD.

Fig. 17. Experimental waveforms of grid current ($i_g$) and grid voltage ($v_g$) for $\omega_{res3} = 0.209\omega_s$ (a) without AD, (b) with AD.

Fig. 18. Experimental waveforms of grid current ($i_g$) and grid voltage ($v_g$) for $\omega_{res4} = 0.241\omega_s$ (a) without AD, (b) with AD.
For experimental verification of its ineffective damping for resonant frequencies of more than one-sixth of the sampling frequency, the capacitor current based AD method has been used for the resonant frequency $\omega_{\text{res}2} = 0.179 \omega_s$, and Fig. 21 shows the corresponding experimental waveforms. It can be seen that the resonant current oscillations are still present in this case. On the other hand, the damping of the proposed AD method at the same resonant frequency has been clarified in Fig. 16(b).

These results, along with the frequency response analysis introduced in the above numerical example, reflect satisfactory steady state and transient performances along with resonance damping over a wide range of resonant frequencies using the proposed active damping method and the control parameters tuning steps.

V. CONCLUSION

A novel active damping strategy using two feedback loops of the grid current and filter capacitor voltage is proposed in this paper. Compared to the previous active damping methods, the proposed one can offer the following merits.

- Compared to the capacitor-current-based method, the cost can be reduced by omitting the high cost current sensor. Moreover, the non-minimum phase behavior can be avoided over a wide range of resonant frequencies.
- Compared to the capacitor-voltage-based method, the proposed strategy can behave effectively over a wide range of the resonant frequencies without stability violations.
- Compared to the grid current based method, a straightforward co-design method for the fundamental current regulator and the active damping loops are proposed.

A numerical example has been introduced to verify the performance of the proposed method over a wide range of resonant frequencies. To show the superiority of the proposed method, the drawbacks of the capacitor voltage/current based methods have been clarified. This example along and experimental results reflect the satisfactory performance of the proposed method.

REFERENCES

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