Passivity-Based Design of Resonant Current Controllers Without Involving Partial Derivative

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Passivity-Based Design of Resonant Current Controllers Without Involving Partial Derivative

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Abstract—In terms of digital control delay, dissipation of the current controller is necessary to enhance the interactive stability between grid-following converters and the grid. The key principle is to make the real part of the converter output admittance non-negative below Nyquist frequency. Besides the fundamental current control, a phase-lead compensation should be introduced for the dissipative design of harmonic resonant controllers. However, derivation of the compensation angle requires calculating the partial derivative of real-part of converter output admittance, which is non-trivial, especially in the presence of grid voltage feedforward. This letter provides a simple calculation method for the compensation angle to achieve dissipativity, and its effectiveness is verified through experiments.

Index Terms—Grid-following converter, current control, resonant controller, compensation angle, dissipativity.

I. INTRODUCTION

As voltage source converter (VSC)-based resources continue proliferating, the power grid is undergoing a transformative shift towards a power-electronic-based structure [1], [2]. In order to tackle the converter-grid interactive instability, passivity-based controller design has received significant attention [3]. The core idea is to make the real-part of the converter output admittance non-negative below the Nyquist frequency using digital control (or called dissipation), and the VSC system can be stabilized regardless of wide-varied passive grid admittance [4].

In light of the current control of grid-following VSCs, the non-dissipative region is mainly related to the total time delay (usually 1.5 times of sampling period) in the computation and pulse width modulation (PWM) process [5], [6]. Several efforts have been made to remove the non-dissipative region such as capacitor current active damping (CCAD) [7], [8], capacitor voltage feedforward (CVFF) and CCAD [9]-[12], and only CVFF [13]. But only a few works focus on the effect of harmonic resonant (R) controllers on dissipativity [9], [14], [15]. Specifically, R controllers can introduce an extra non-dissipative region in the vicinity of harmonic frequencies even though the VSC system is dissipative without R controllers [9]. Common solution is to introduce a specific phase-lead compensation for R controllers. It is verified that neither no compensation nor compensation of computation and PWM delay can guarantee dissipativity [14], [15]. In [14], a partial derivative-based method is proposed to solve the compensation angle, where the main challenge is the complicated derivation of compensation angle through calculating the partial derivative of real-part of VSC output admittance [15]. Further, the compensation angle is strongly related to damping methods, which makes the derivation more difficult especially when using CVFF [14]. However, CVFF is often indispensable for the resonance damping and start-up of grid-following VSC [10]-[11].

To solve this problem, this letter proposes a simple compensation angle calculation method without involving partial derivatives. The rest of this paper is organized as follows. In Section II, a detailed system model is derived for a grid-following VSC with CCAD and CVFF. The proposed compensation angle calculation method is proposed in Section III. In the last, experimental results are presented in Section IV, and conclusions are drawn in Section V.

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II. SYSTEM MODELING WITH CONVERTER-SIDE CURRENT CONTROL

With switches in Fig. 1(a) and (b) switching to converter-side current, $i_{cl}(s)$, Fig. 1 presents the schematic circuit diagram and
the detailed control block diagram of the grid-connected VSC with converter-side current control. The converter is connected to the grid through an LCL filter, with $L_1$, $L_2$, and $C$ denoting the converter-side inductance, the grid-side inductance, and the filter capacitance respectively. $L_g$ and $C_g$ denote the grid inductance and capacitance respectively. The capacitor voltage ($v_c(s)$) is fed forward to a phase-locked loop (PLL) for synchronizing the current reference ($i_{ref}(s)$) with the voltage at point of common coupling (PCC), $v_{pcc}(s)$, and to current control for CVFF. The capacitor current ($i_c(s)$) is fed to the current control for CCAD. The converter-side current ($i_p(s)$) is fed back for tracking $i_{ref}(s)$. Since this letter only focuses on the dissipative region around harmonic resonant frequencies of R-controller, PLL is designed to have a low bandwidth to avoid affecting the frequency characteristics around harmonic frequencies [16]. Thus, its effect is ignored in the following analysis.

$H_i$ denotes the CCAD coefficient with converter-side current control. The transfer function of CVFF is $H_i(s)$. The current controller, $G_i(s)$, is a proportional resonant (PR) controller expressed as:

$$G_i(s) = K_p + \sum_{n} K_n \frac{s \cos \phi_n - h \omega_n \sin \phi_n}{s^2 + h \omega_n^2}$$

(1)

where $\omega_i$ is the fundamental angle frequency, $h$ is the order of the harmonics of concern, $K_p$ is the proportional gain, $K_n$ is the resonant gain of the R controller tuned at $h \omega_n$ ($R_i(s)$), and $\phi_n$ is the compensation angle at the resonant frequency, $\omega_i$. The delay from the voltage reference to the output voltage is modeled as

$$G_y(s) = e^{-s\tau_d}$$

(2)

where $\tau_d = 1.5 T_i$ is the total delay caused by computation and pulse-width modulation (PWM) with $T_i$ being the sampling period. With $C$ and $L_2$ regarded as a part of grid impedance, $Z_g$, the admittance model of the VSC with $L_1$ filter is derived as:

$$i_p(s) = G_i(s)i_{ref}(s) - Y_o(s)v_c(s)$$

(3)

where the closed-loop transfer function, $G_o(s)$, and the output admittance, $Y_o(s)$, are respectively given by:

$$Y_o(s) = \frac{1 - sH_i Y_i(s)}{sL_i + G_i(s)G_y(s)}$$

(5)

According to the passivity theory, two conditions guarantee the system’s stability, 1) $G_o(s)$ is stable; 2) $Z_o(s)$ and $Y_o(s)$ are both dissipative below the Nyquist frequency.

The first condition can be fulfilled by a properly designed $K_p$. It is recommended that the bandwidth of the open loop transfer function, $\alpha_c$, should be smaller than a tenth of sampling angle frequency, $\omega_s/10$, and herein $\alpha_c = \omega_s/10$ is adopted. Then, $K_p$ is given by $K_p = \alpha L_i$ [9], which ensures the first condition.

Provided that $Z_o(s)$ is composed of passive components, $Z_o(s)$ is passive. The dissipativity of $Y_o(s)$ relies on proper control parameters, whose design will be detailed below.

Since R controllers only affect the admittance around the resonant frequencies, they can be neglected when designing other parameters. Consequently, the real part of $Y_o(s)$ is calculated as:

$$\text{Re}[Y_o(j \omega)] = A_1(\omega) + A_2(\omega)$$

(6)

where $A_1(\omega)$ is the real part of $Y_o(s)$ without CVFF, and $A_2(\omega)$ is the real part of $Y_o(s)$ introduced by CVFF, which are respectively expressed as:

$$A_1(\omega) = \frac{(K_p - H_i L_c \omega^2) \cos(\omega T_d)}{(K_p \cos(\omega T_d))^2 + (\alpha L_1 - H_i \sin(\omega T_d))^2}$$

(7)

$$A_2(\omega) = \frac{-0.5K_p}{(K_p \cos(\omega T_d))^2 + (\alpha L_1 - H_i \sin(\omega T_d))^2}$$

(8)

$$+ \frac{0.5K_p}{(K_p \cos(\omega T_d))^2 + (\alpha L_1 - H_i \sin(\omega T_d))^2} \cdot H_i$$

(9)

$H_i$ is designed such that $A_1(\omega) \geq 0$ for $0 < \omega < \omega_s / 2$. Thus, $(K_p - H_i L_c \omega^2)$ should change its sign at the critical frequency, $\omega_s / 6$, where $\cos(\omega T_d)$ changes its sign, which gives

$$H_i = \frac{4K_p T_d^2}{\pi^2 L C}$$

(10)

According to [12], CVFF using moving average filter is in favor of dissipativity of $Y_o(s)$, and is herein adopted. Thus, $H_o(s)$ is given by

$$H_o(s) = K_{ff} [0.5 + 0.5e^{-sT_d}]$$

(10)

where $K_{ff}$ is the CVFF coefficient, which should be designed smaller than one to ensure the low-frequency dissipativity, and is set to 0.9 in this letter.

For now, $Y_o(s)$ has been designed to be dissipative without R controllers. In the next section, the dissipative design of $Y_o(s)$ with R controller will be discussed.

III. PASSIVITY-BASED DESIGN OF RESONANT CONTROLLER WITH CONVERTER-SIDE CURRENT CONTROL

A. Conventional Compensation Angle Calculation Methods

The compensation angle, $\phi_n$, of R controller significantly influences the dissipativity characteristics of the output

| TABLE I |
| MAIN PARAMETERS OF A THREE-PHASE GRID-CONNECTED VSC |

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$S_n$</td>
<td>Rated power 7 kVA</td>
</tr>
<tr>
<td>$V_e$</td>
<td>Grid phase voltage (RMS) 220 V</td>
</tr>
<tr>
<td>$V_d$</td>
<td>DC-link voltage 700 V</td>
</tr>
<tr>
<td>$L_1$</td>
<td>Converter-side inductance 4 mH</td>
</tr>
<tr>
<td>$L_2$</td>
<td>Grid-side inductance 2 mH</td>
</tr>
<tr>
<td>$C$</td>
<td>Filter capacitance 10 mF</td>
</tr>
<tr>
<td>$C_1$</td>
<td>Grid capacitance 0.5 mF</td>
</tr>
<tr>
<td>$C_2$</td>
<td>Grid capacitance 30 mF</td>
</tr>
<tr>
<td>$f_{sw}$</td>
<td>Switching frequency 4 kHz</td>
</tr>
<tr>
<td></td>
<td><strong>Double-Sampling Mode Parameters</strong></td>
</tr>
<tr>
<td>$f_i$</td>
<td>Sampling frequency 8 kHz</td>
</tr>
<tr>
<td>$K_p$</td>
<td>Proportional current controller gain 20 Ω</td>
</tr>
<tr>
<td>$K_{n,h}$</td>
<td>Resonant current controller gain 4000 Ω/s</td>
</tr>
<tr>
<td></td>
<td><strong>Single-Sampling Mode Parameters</strong></td>
</tr>
<tr>
<td>$f_i$</td>
<td>Sampling frequency 4 kHz</td>
</tr>
<tr>
<td>$K_p$</td>
<td>Proportional current controller gain 10 Ω</td>
</tr>
<tr>
<td>$K_{n,h}$</td>
<td>Resonant current controller gain 1000 Ω/s</td>
</tr>
</tbody>
</table>
admittance around resonant frequency of the R controller \[14\]. A direct way to calculate the compensation angle, \( \phi_h \), is to compensate for the phase delay of \( G_d(s) \) \[9\], i.e.,

\[
\phi_h = h\omega T_d. \tag{11}
\]

However, such a compensation angle cannot ensure dissipativity when CCAD or CVFF exists. To demonstrate this, the Nyquist plot of the corresponding impedence, \( Z_d(s) = 1/Y_d(s) \), is depicted in Fig. 2 with \( \phi_h \) selected as (11), \( h \in \{1, 5\} \), and other parameters listed in Table I. Note that the dissipativity of \( Y_d(s) \) is equivalent to the dissipativity of \( Z_d(s) \). The asymptotic lines of the Nyquist curve of \( Z_d(s) \) are straight lines (dashed line in Fig. 2) as \( s \) tends to \( \pm \infty \). As long as the asymptotic line is not vertical to the real axis, the Nyquist curve is bound to pass through the left half plane, and the dissipativity is lost. This is the case of \( Z_d(s) \) with (11) adopted (see the circled part in Fig. 2), proving (11) fails to ensure dissipativity.

Then a partial derivative-based method was proposed to solve this problem \[14\], which selects \( \phi_h \) such that

\[
\left. \frac{\partial \text{Re}\{Y_d(j\omega)\}}{\partial \omega} \right|_{\omega=\omega_0} = 0. \tag{12}
\]

Meanwhile, the second derivative should be positive:

\[
\left. \frac{\partial^2 \text{Re}\{Y_d(j\omega)\}}{\partial \omega^2} \right|_{\omega=\omega_0} > 0. \tag{13}
\]

The principle is that since \( Y_d(j\omega_0) = 0 \), i.e., \( \text{Re}\{Y_d(j\omega_0)\} = 0 \), \( j\omega_0 \) should be a local minimum point of \( \text{Re}\{Y_d(j\omega)\} \) otherwise \( \text{Re}\{Y_d(j\omega)\} \) will be negative in the neighborhood of \( \omega_0 \). Thus, (12) and (13) should be satisfied.

Even though the idea is clear, the solving process is very complex \[9\], \[14\], \[15\]. For that reason, another calculation method for \( \phi_h \) will be proposed below, which proves to be equivalent to the partial derivative-based method but it is much simpler.

### B. Proposed Compensation Angle Calculation Method

As mentioned above, the dissipativity of \( Y_d(s) \) is equivalent to the dissipativity of \( Z_d(s) \). The basic principle to make \( Z_d(s) \) dissipative is to design a \( \phi_h \) such that the asymptotic line at the resonant frequency is vertical to the real axis. The inclination angle of the asymptotic line can be obtained by limit calculation with \( \omega \) tending to \( h\omega_0 \). First, the left limit is:

\[
\lim \angle Z_d(j\omega) = \lim_{\omega \to +h\omega_0} \angle \left( \frac{j\omega L_s + G_s(j\omega)G_d(j\omega)}{1 - j\omega H_jCG_d(j\omega) - H_s(j\omega)G_d(j\omega)} \right) = \lim_{\omega \to +h\omega_0} \angle \left( \frac{G_s(j\omega)}{1 - j\omega H_jCG_d(j\omega) - H_s(j\omega)G_d(j\omega)} \right) \tag{14}\]

Since the magnitude of \( R(s) \) is infinite at \( h\omega_0 \), any limited value added to it can be neglected. \( \Sigma_k R(s) \) does not include a resonant frequency of \( h\omega_0 \), and has a finite magnitude at \( h\omega_0 \). Thus, the first part of (14) can be neglected. Then,

\[
\lim \angle Z_d(j\omega) = \lim_{\omega \to +h\omega_0} \angle \left( \frac{R_s(j\omega)G_s(j\omega)}{1 - j\omega H_jCG_d(j\omega) - H_s(j\omega)G_d(j\omega)} \right) = \lim_{\omega \to +h\omega_0} \angle \left( \frac{G_s(j\omega)}{1 - j\omega H_jCG_d(j\omega) - H_s(j\omega)G_d(j\omega)} \right) \tag{15}\]

where \( \phi_p \) denotes the second part of (15). For the first part of (15), the limit calculation result is:

\[
\lim_{\omega \to +h\omega_0} \angle Z_d(j\omega) = \frac{\pi}{2} + \phi_h. \tag{16}\]

Substituting (16) into (15) gives

\[
\lim_{\omega \to +h\omega_0} \angle Z_d(j\omega) = \frac{\pi}{2} + \phi_h + \phi_p. \tag{17}\]

This is the inclination angle of the asymptotic line as \( \omega \) tends to \( h\omega_0 \). In the same way, the right limit of \( \angle Z_d(j\omega) \) can be obtained:

\[
\lim_{\omega \to -h\omega_0} \angle Z_d(j\omega) = -\frac{\pi}{2} + \phi_h + \phi_p. \tag{18}\]

Thus, the phase range of \( \angle Z_d(j\omega) \) at \( h\omega_0 \) is \([-\pi/2 + \phi_h + \phi_p, \pi/2 + \phi_h + \phi_p]\). To align the asymptotic line vertically with the real axis, featuring an inclination angle of \( \pi/2 \) and a phase range of \([-\pi/2, \pi/2]\) due to its positioning on the right-half plane, \( \phi_h \) should be set to

\[
\phi_h = -\phi_p = -\angle \left( \frac{G_s(j\omega)}{1 - j\omega H_jCG_d(j\omega) - H_s(j\omega)G_d(j\omega)} \right) \tag{19}\]

which can be solved easily with MATLAB function \textit{bode}. The result does not involve \( L_2 \) and \( L_p \), because they do not affect \( Y_d(s) \) or do not appear in (5). To validate the correctness of (19), the Nyquist plot of \( Z_d(s) \) is depicted in Fig. 2 with \( \phi_h \) selected as (19), \( h \in \{1, 5\} \), and other parameters listed in Table I. The asymptotic line as \( \omega \) tends to \( h\omega_0 \) (dashed line) is vertical to the real axis, indicating a phase range of \([-\pi/2, \pi/2]\), and therefore having no negative real part.

The proposed method is equivalent to the partial derivative-based method. This can be verified by two special cases. First, by omitting the CVFF path, or setting \( H_s(s) \) to zero, the control scheme in Fig. 1 becomes the same as that in \[9\]. (19) is then reduced to,

\[
\phi_h = -\angle \left( \frac{G_s(j\omega)}{1 - j\omega H_jCG_d(j\omega)} \right) = \angle \left( e^{\phi_h T_d} - j\omega H_jC \right) \tag{20}\]
which is exactly the same as the result in [9] derived by the partial derivative-based method. Second, by omitting the CCAD path, or simply setting \( H_t \) to zero, (20) reduces to \( \phi_t = h_0 T_d \), which is also exactly the same as the result in [14].

As for another parameter of the R controller, the resonant gain, \( K_{rh} \), should be set to strike an equilibrium between fast dynamic response and good harmonic selectivity [4]. Furthermore, it is important to note that an elevated \( K_{rh} \) can potentially compromise the dissipativity when the CVFF strategy is applied [4]. Considering above factors, \( K_{rh} \) is set to 4000 for all \( h \) under a sampling frequency of 8000 Hz.

Indeed, this letter focuses on the analysis of converter-side current-controlled VSC to exemplify the proposed method. However, it is important to emphasize that this specific illustration does not preclude the applicability of the method to a broader range of systems. As demonstrated further, the method will also be applied to grid-side current control in the upcoming section.

IV. EXTENSION TO GRID-SIDE CURRENT CONTROL

With the switches in Fig. 1(a) and (b) switching to grid-side current, \( i_{gd}(s) \), Fig. 1(a) and (b) present the schematic circuit diagram and the detailed control block diagram of VSC with grid-side current control, respectively. The admittance model of the VSC with LCL filter can be expressed as:

\[
i_{gd}(s) = G_{cl,g}(s)Y_{o,g}(s) - Y_{o,g}(s)v_{pcc}(s) \tag{21}
\]

where \( G_{cl,g}(s) \) represents the closed-loop transfer function, and \( Y_{o,g}(s) \) is the output admittance whose expression (22) is shown at the bottom of this page. With a similar manner as converter-side current control, the CCAD coefficient \( H_{tg} \) for grid-side current control is derived as [11]:

\[
H_{tg} = \frac{4K_p T_d^2}{\pi^2 L_s C} - K_p. \tag{23}
\]

With the proposed method elaborated in Section III B, the compensation angle, \( \phi_t \), of \( R_o(s) \) with grid-side current control can be derived as:

\[
\phi_t = -\frac{G_c(\omega)}{1 - \omega^2 L_s C - j\omega H_c G_{cl}(\omega) - H_c(\omega)G_d(\omega)} \Big|_{\omega=\omega_{01}} \tag{24}
\]

which ensures the dissipativity of \( Y_{o,g}(s) \).

The fulfillment of the second condition of the passivity theory is now accomplished. To satisfy the first condition (internal stability), the stability analysis is undertaken through the transformation to converter-side current control, as expounded in [11]. According to [11], the equivalence between grid-side current control and converter-side current control is maintained, preserving the integrity of all controllers depicted in Fig. 1(b), with the sole alteration being the replacement of the CCAD coefficient with:

\[
H_{t,g} = H_{t,g} + G_c(s). \tag{25}
\]

As illustrated in Fig. 3, the admittance model of VSC with grid-side current control (see Fig. 3(a)) can be equivalently transformed to Fig. 3(b) comprising an admittance model of VSC with converter-side current control. Thus, by selecting the same \( K_p \) as converter-side current control, the independent current source \( G_{cl}(s)Y_{o,g} \) is stabilized. Moreover, \( Y_{o,g}(s) \) is dissipative due to the dissipativity of \( Y_{o,g}(s) \). This is evident from the fact that the disparity between these two models involves only two passive components, \( C \) and \( L_2 \), both characterized by zero resistance.

Due to the equivalence, the two independent current sources in Fig. 3(a) and (b) have the following relationship:

\[
G_{cl}(s)\frac{Y_{o,g}(s)}{1 + sL_2 Y_{o,g}(s)} = \frac{1}{1 + sL_2 Y_{o,g}(s)} \tag{26}
\]

where \( Y_{cl,g} = sC + 1/(sL_2) \). To interpret (26) from the framework of passivity theory, the dissipativity of \( Y_{o,g}(s) \) and the stability of \( G_{cl}(s) \) together ensure the stability of \( G_{cl}(s) \), i.e., the internal stability of grid-side current control.

\[
\begin{align*}
Y_{o,g}(s) &= \frac{1 + s^2 L_2 C - sCH_{cl,g} G_d(s) - H_c(s)G_d(s)}{s^2 L_2 C - s^2 L_2 CH_{cl,g} G_d(s) + s(L_1 + L_2) - sL_2 H_c(s)G_d(s) + G_c(s)G_d(s)} \tag{22}
\end{align*}
\]

V. EXPERIMENTAL VALIDATION

To further verify the theoretical analysis, experiments are carried out on a three-phase grid-connected VSC with an LCL filter, as shown in Fig. 4. The grid is emulated with a linear amplifier APS 15000. The VSC and the control platform are a PEB-SiC-8024 module and a B-BOX RCP control platform from Imperix, respectively. To verify the effectiveness of the proposed method below and beyond the current control bandwidth, \( \alpha_c, h \) is selected to be 1, 5, 7 (below the bandwidth), 17, and 19 (beyond the bandwidth).

A. Converter-Side Current Control

In this subsection, grid impedance is assumed to contain only an inductor \( L_g \), and double-sampling mode is adopted. Relevant parameters are presented in Table I. Fig. 5 shows the frequency responses of \( Y_{o,g}(s) \), with different compensation angle selections, and the equivalent grid admittance \( Y_{o,g} = 1/(sL_2 + sL_3) + sC \). As
when the DC-link voltage is established, the q-axis reference is shifted to 15A. As it can be seen, the system is stable with proposed $\phi_h$ (see Fig. 6(a)) since dissipativity is achieved. However, the system with $\phi_h = ho_n T_d$ is unstable (see Fig. 6(b)). This instability leads to a gradual escalation of harmonics due to the loss of dissipativity. These findings align consistently with those presented in Fig. 5.

B. Grid-Side Current Control

In this subsection, grid-side current control performance is assessed. Additionally, in Fig. 1(a), $C_g$ is linked to the grid for testing under varied conditions corresponding a grid admittance: $Y_g(s) = 1/(s L_g) + s C_g$. Both double-sampling mode ($f_s=8kHz$) and single-sampling mode ($f_s=4kHz$) are implemented. Accordingly, $K_p$ of single-sampling mode is halved to align with the principle that bandwidth relates proportionally to sampling frequency. Other parameters can be found in Table I.

Fig. 7 shows the frequency responses of the grid admittance, $Y_g(s)$, and output admittances of VSC, $Y_o(s)$, as implemented in both double-sampling and single-sampling modes. As can be seen from Fig. 7, both output admittances are dissipative beneath their respective Nyquist frequencies (4kHz for double-sampling, 2kHz for single-sampling). Additionally, their positive phase margins signify the stability of the system in both sampling modes. Notably, it is important to exclude consideration of the output admittance beyond the Nyquist frequency range, given that the employed s-domain analysis method holds precision solely within the confines of the Nyquist frequency. Importantly, for the single-sampling mode, the 19th harmonic frequency (950Hz) surpasses the bandwidth of 400Hz by a substantial margin. This stark contrast conclusively demonstrates the efficacy of the proposed method in stabilizing the system, independent of the relationship between the resonant frequencies of the R controllers and the bandwidth.

Fig. 8 presents experimental waveforms illustrating the startup of the grid-side current-controlled VSC, comparing both double-sampling mode (Fig. 8(a)) and single-sampling mode (Fig. 8(b)). The grid voltage integrates 10% 5th harmonic and 10% 7th harmonic components. Notably, the single-sampling
mode’s startup current is substantially higher, around 14A, in contrast to the double-sampling mode’s startup current of approximately 5A. This discrepancy can be attributed to the extended delay intrinsic to the single-sampling mode, adversely affecting overall system performance.

Fig. 9 captures the experimental waveforms of the double-sampling grid-side current-controlled VSC, following a step change in current reference from 7.5A to 15A. The grid voltage, $v_g$, comprises 10% 5th harmonic and 10% 7th harmonic components. The displayed dynamics highlight the rapid response of the system using double-sampling, attaining settling within approximately 0.02s.

V. CONCLUSION

This letter introduces an innovative and straightforward calculation approach for determining the compensation angle in the design of passivity-based R controllers, eliminating the need for intricate partial derivative calculations. The proposed method ensures the dissipative nature of the output admittance of the VSC below the Nyquist frequency. The step-by-step procedure of this approach is meticulously elucidated, initially demonstrated within the framework of converter-side current control, and subsequently extended to encompass grid-side current control. The validity of the proposed method is substantiated through experiments.

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