Lyapunov-Function-Based Controller for Single-Phase NPC Quasi-Z-Source Inverter with $2\omega$ Frequency Ripple Suppression

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Abstract: This paper proposes a high-performance control technique based on Lyapunov’s stability theory for a single-phase grid-connected neutral-point-clamped quasi-impedance source inverter with LCL filter. The Lyapunov function based control is employed to regulate the inverter output current, whereas the proportional resonant controller is used to produce the reference of the inverter output current that is needed in the Lyapunov function based control. Use of proportional resonant controller ensures the zero steady-state error in the grid current. An important feature of the proposed Lyapunov function based control is the achievement of resonance damping without using a dedicated damping method. Furthermore, the modified simple boost control technique is proposed to eliminate the double-line frequency ripples in the quasi-impedance source inductor currents and minimize the double-line frequency ripples in the quasi-impedance source capacitor voltages. The proposed control technique considerably reduces the inverter size, weight, and cost as well as increases overall system efficiency since the required inductances and capacitances sizes are lower. Experimental results obtained from a 2.5 kW neutral-point-clamped quasi-impedance source inverter prototype are presented to validate the performance of the Lyapunov function based control technique.

Keywords: grid-connected inverter; Lyapunov stability; nonlinear controller; proportional resonant control

1. Introduction

The power electronic converters play significant role in converting unregulated dc power into regulated ac power in photovoltaic (PV) systems. Compared to the conventional two-level configurations, the multilevel inverters (MI)s are recently promising solution for PV applications due to their significant advantages such as high performance, lower total harmonic distortion (THD), less voltage stress on the passive and active components, and so forth. The neutral-point-clamped (NPC) inverter is one of the popular MI topologies that is employed in many industrial applications [1–3].

The controller design is a challenging issue since it needs to meet several control objectives that include fast dynamic response, preferably zero steady-state error, high
stability, and low THD under various loads. To meet these objectives, many control
techniques are proposed for grid-connected inverters like deadbeat control [8], model
predictive control [9], repetitive control [10], and sliding mode control [11–13]. Among these
control techniques model predictive control and deadbeat control are the most attractive
control techniques due to their fast transient response and near-zero steady-state error.
On the other hand, these control techniques are sensitive to parameter changes in the
system model and to noise in the high sampling frequency. Therefore, these drawbacks
should be taken into account when such control techniques are designed. Another control
technique is the repetitive control that provides satisfactory steady-state performance with
excellent harmonic rejection under the non-linear loads. However, the transient response
of the repetitive control is very slow and its tracking accuracy is poor. Alternatively,
sliding mode control provides fast transient response and simplicity in implementation
for single-phase inverters. Furthermore, this control technique offers high stability against
non-linear load conditions and is not sensitive to the parameters changes in the system.
However, it suffers from the inherent drawback of chattering phenomenon because of
the variable switching frequency that results in inaccuracy in the control loop, reasonably
high switching losses, and complexity in the filter design. Although each proposed control
technique has its own advantages and disadvantages, none of these control techniques
assures the global stability of the closed-loop system against sudden perturbations which
change the operating point of the system.

In the case of grid-connected LCL-filtered inverter systems, the controller design
is even more complex due to the inherent resonance damping requirement. The res-
onance problem is usually solved by passive damping [14], or by the active damping
method [11–13,15–20]. Passive damping is realized by connecting a passive resistor in
series or in parallel with the LCL filter components. However, the passive damping method
is outdated due to the additional power losses decreasing efficiency of the system. On the
other hand, the active damping method is realized by emulating the dissipative resistor in
the control method rather than connecting it physically. Hence, it is generally referred to as
virtual-resistor (VR) based active damping. Generally, the active damping method is based
on the multi-loop control method [11–13,15–17] and the filter based control method [18,19].
Although the latter requires fewer sensors, it is sensitive to parameter variations and dis-
turbances. As mentioned in Reference [20], single loop controller designed for LCL-filtered
grid-connected inverter system cannot offer satisfactory performance. Therefore, most of
the existing control methods utilize either two [11,12,15] or three [13,16] control loops.

Lyapunov-function-based control has been proposed for various applications that
include hybrid active filter [21], dc/dc converters [22], single-phase inverters [23–25].
The main idea behind the Lyapunov-function-based control is to determine a control law
such that the rate of change of the Lyapunov function remains always negative. Since
the considered NPC-qZSI is interfaced to the grid via LCL filter, the energy on the LCL
filter can be used to define the Lyapunov function and implement the Lyapunov-function-
based control easily. The proposed controller requires three reference functions for the
ac-side variables (reference of the inverter current, grid current reference, and capacitor
voltage reference). The capacitor voltage reference is computed by using the defined
grid current reference via a differential equation. The reference of the inverter current is
produced using a proportional resonant (PR) controller which processes the grid current
error applied to its input. The consequence of utilizing PR controller is that zero grid
current error can be achieved. It is well known that the qZSI inverter has the ability to
boost its dc input voltage by making use of shoot-through switching state which permits
the switches to be turned on at the same inverter leg. Hence, shoot-through duty ratio
of qZS network should be controlled. To control the shoot-through state, various control
methods have been proposed that include the simple boost control, the maximum boost
control, and the maximum constant boost control [26]. Although the simple boost control
technique offers a lower modulation index and voltage gain over other control techniques,
the implementation of this technique is easier than others [26,27].
Similar to the case of conventional single-phase inverter, the double-line frequency ($2\omega$) pulsating power exist in the dc side of the single-phase NPC-qZSI that results in $2\omega$ ripple on the qZS capacitor voltages and on the qZS inductor currents [28]. These $2\omega$ ripple components should be eliminated or minimized considerably. To eliminate and minimize the $2\omega$ ripple, passive and active damping techniques have been proposed [29,30]. However, the requirement of large inductances and capacitors (for passive techniques) or extra semiconductor switches (for active techniques) cause considerably increase of the inverter volume, weight, and cost [31]. As an alternative to these techniques, the existing simple-boost technique has been modified. The modified control technique not only regulates the shoot-through duty ratio, but also mitigates the $2\omega$ power flow on the qZS network. Unlike the active damping method in Reference [29], the proposed ripple suppression method is based on passive damping which does not require additional switching devices. In Reference [30], the $2\omega$ ripple suppression method is based on modifying the shoot-through duty ratio with the small variations which are obtained by processing the inductor current ($i_{L1}$) through low pass filtering and proportional-integral (PI) controller. However, the proposed ripple suppression method requires an additional voltage sensor only without using additional PI controller and low pass filter as in Reference [30].

The proposed control strategy has two control parts: dc-side control and ac-side control. While dc-side control achieves the control of quasi impedance network variables (inductor currents and capacitor voltages), the ac-side control achieves the control of grid current. In literature, there are only several studies which investigate the control of single-phase grid-connected three-level NPC-qZSI with LCL filter [5,6]. The control method proposed in Reference [5] is for single-phase three-level NPC-qZSI. While dc-side variables are controlled by using proportional-integral-derivative (PID) control, ac-side control is achieved by PR controller. However, there is no $2\omega$ ripple elimination methodology in the dc-side control. Also, the proposed control method is verified by simulation results rather than experimental results. Furthermore, the switching frequency is selected as 100 kHz which is very high for a MI application. On the other hand, a three-phase grid-connected three-level NPC-qZSI topology is tested experimentally in Reference [6] in which the control of ac-side is not discussed at all. However, the authors proposed a new modulation technique for the shoot through. Comparing the proposed control strategy with the control strategies in References [5,6], the proposed control strategy offers important features such as $2\omega$ ripple suppression in the dc-side inductor current, fast transient response both in the dc- and ac-side variables, strengthened stability, good tracking performance and zero grid current error in the steady-state. Nevertheless, a more detailed comparison of the ac-side is also provided in Section 4.

The major contributions of this paper include—(1) The ac-side of the NPC-qZSI is controlled by Lyapunov-function-based control which possesses many features such as fast dynamic response, strengthened stability, reduced number of sensors, excellent resonance damping without using dedicated active damping technique, guaranteed zero steady-state error and low THD in grid current. (2) The existing simple-boost control method in References [26,27] is modified to eliminate the $2\omega$ power flow on the qZS inductances and minimize the $2\omega$ power flow on the qZS capacitances. The proposed control technique considerably reduces the inverter size, weight, and cost as well as increases overall system efficiency.

In order to demonstrate the effectiveness of proposed control technique in the steady-state and during transient, experimental studies have been conducted through a small-scale NPC-qZSI prototype. The remainder of the paper is organized as follows. In Section 2.1, the modeling of the NPC-qZSI is presented. In Section 2.2 the proposed control strategies are introduced and extensively analyzed. In Section 3, the experimental results obtained from a 2.5 kW prototype are presented for different operating conditions. Finally, an appraisal of the proposed methodology is given in the conclusion.
2. Materials and Methods

2.1. Mathematical Modeling of NPC-qZSI

The single-phase grid-connected NPC-qZSI with LCL filter is shown in Figure 1 [3]. The input circuit of this inverter is built by an impedance source network that provides the ability to boost the input voltage at the required level. It is well-known that impedance source inverter has two switching states namely shoot-through and non-shoot-through states and the mathematical model of the NPC-qZSI depends on these switching states as depicted in Figure 2 [32]. The capacitors and inductors of the qZS network are assumed to be identical as follows [32]

\[ L_1 = L_3, \quad L_2 = L_4, \quad C_1 = C_4, \quad C_2 = C_3. \]  

(1)

Although the values of capacitors and inductances can vary in practice, this variation is relatively small and has no influence on the performance of the proposed control method. This fact can be observed in the experimental results provided in Section 3 where theoreti-
cally identical capacitors and inductors are used which have almost no adverse effect on the performance of the proposed control method in practice. During non-shoot-through state, the NPC-qZSI is represented as a current source as shown in Figure 2a. The voltage across the inductors can be written as follows [32]

\[ V_{L2} = -V_{C1}, \quad V_{L4} = -V_{C4}, \]  

(2)

\[ V_{L13} = V_{in} - V_{C2} - V_{C3}. \]  

(3)

On the other hand, in shoot-through state, the NPC-qZSI is represented as short circuit as shown in Figure 2b. The voltage across the inductors can be written as [32]

\[ V_{L13} = V_{in} + V_{C1} + V_{C4}, \]  

(4)

\[ V_{L24} = V_{C2} + V_{C3}. \]  

(5)

From (2) to (5), inductor voltages are written as [32]

\[ V_{L1} = T_0 \left( \frac{V_{in} + V_{C2} + V_{C4}}{2} \right) + T_1 \left( \frac{V_{in} - V_{C2} - V_{C4}}{2} \right) = 0, \]  

(6a)

\[ V_{L2} = T_0 \left( \frac{V_{C2} + V_{C4}}{2} \right) + T_1 (-V_{C1}) = 0, \]  

(6b)

\[ V_{L3} = T_0 \left( \frac{V_{in} + V_{C2} + V_{C4}}{2} \right) + T_1 \left( \frac{V_{in} - V_{C2} - V_{C4}}{2} \right) = 0 \]  

(6c)

\[ V_{L4} = T_0 \left( \frac{V_{C2} + V_{C4}}{2} \right) + T_1 (-V_{C4}) = 0, \]  

(6d)

where \( T_0 \) is the period of shoot-through whereas \( T_1 \) is the period of non-shoot-through states. By making use of (6), the capacitor voltages can be written as [32]

\[ V_{C1} = V_{C4} = \frac{d_{ST} V_{in}}{2 - 4d_{ST}}, \quad V_{C2} = V_{C3} = \frac{(1 - d_{ST}) V_{in}}{2 - 4d_{ST}}, \]  

(7)

where \( d_{ST} \) is duty-cycle of the shoot-through state. Applying Kirchhoff’s voltage law to the second loop involving capacitor voltages in Figure 2a, the expression of dc-link voltage \( V_{PN} \) can be obtained as follows [32]

\[ V_{PN} = V_{C1} + V_{C2} + V_{C3} + V_{C4}. \]  

(8)

The boost factor (\( B \)) is written as

\[ B = \frac{V_{PN}}{V_{in}} = \frac{V_{C1} + V_{C2} + V_{C3} + V_{C4}}{V_{in}} = \frac{1}{1 - 2d_{ST}}. \]  

(9)

It is worth noting that the correctness of the Equations (1)–(9) is already verified in the literature [3,6,7]. This clearly shows the validity of the models in Figure 2. The inverter output voltage (\( v_{inv} \)) and current (\( i_1 \)) are

\[ v_{inv} = \sqrt{2} V_{inv} \sin \omega t, \quad i_1 = \sqrt{2} I_1 \sin(\omega t - \theta), \]  

(10)

where \( \omega \) is the fundamental angular frequency, \( \theta \) is the phase angle, \( V_{inv} \) and \( I_1 \) are the rms output voltage and current, respectively. The inverter output power is written as

\[ P_o = V_{inv} I_1 \cos \theta - V_{inv} I_1 \cos(2\omega t - \theta). \]  

(11)

Equation (11) shows that output power has two components which flow through
dc and ac sides of the inverter. The first component of the equation represents active power and the second component is the $2\omega$ power that causes fluctuations (ripples) on voltages and currents in the dc-side of the qZSI. The frequency of this ripple is $2\omega$. These ripples, unfortunately, cannot be eliminated completely because of the inherent nature of the single-phase inverter. However, the impact of the $2\omega$ ripples can be minimized through some design and/or control approaches.

2.2. The Proposed Control Technique

The block diagram of the control structure is given in Figure 3. The proposed control technique handles the grid current ($i_2$) control and the dc-link voltage ($V_{PN}$) regulation with $2\omega$ power flow mitigation. These control objectives are described in detail as follows.

![Figure 3. The block diagram of the developed control technique; (a) Lyapunov-function based control for ac-side variables; (b) dc-side control with $2\omega$ ripple suppression; (c) Gate signal generation.](image)

2.2.1. Control of AC-Side Variables

To control the grid current ($i_2$), the Lyapunov-function-based control technique is utilized. It is worth noting that the derivative of Lyapunov function should be guaranteed in this control technique. The determination of the Lyapunov function is usually based on the energy dissipation of the system under consideration. More detailed information about Lyapunov function is available in [23–25,33].

Now, let the Lyapunov function be defined as follows

$$V(x) = 0.5L_1x_1^2 + 0.5L_2x_2^2 + 0.5C_f x_3^2,$$  (12)

where $x_1$, $x_2$, and $x_3$ are the state variables defined as

$$x_1 = i_1 - i_1^* \quad x_2 = i_2 - i_2^* \quad x_3 = v_C - v_C^*,$$  (13)

where $i_1$, $i_2$, and $v_C$ represent the inverter output current, the grid current, and the filter capacitor voltage, respectively. The reference values of these variables are marked with *.
superscript symbol. It can be seen from (12) that the Lyapunov function $V(x)$ contains terms which are the energy stored in the LCL filter components ($L_i$, $L_o$, and $C_f$). The derivative of $V(x)$ is given by

\[
V(x) = x_1 L_i \dot{x}_1 + x_2 L_o \dot{x}_2 + x_3 C_f \dot{x}_3.
\] (14)

Using the voltage and current laws of Kirchhoff, the following expressions can be obtained with the circuit model

\[
L_i \frac{di_1}{dt} + R_i i_1 = v_{inv} - v_C,
\] (15)

\[
L_o \frac{di_2}{dt} + R_o i_2 = v_C - v_g,
\] (16)

\[
C_f \frac{dv_C}{dt} = i_1 - i_2,
\] (17)

where $v_g$ denotes the grid voltage, $v_C$ denotes the filter capacitor voltage, and $v_{inv}$ is the inverter output voltage. Furthermore, $R_i$ and $R_o$ are the the resistance of the filter inductances $L_i$ and $L_o$. The output voltage of the inverter can be defined in terms of dc input voltage as follows

\[
v_{inv} = d V_{PN},
\] (18)

where $d$ represents the switching function defined as a combination of its steady-state and perturbed terms as

\[
d = d_{ss} + d_p.
\] (19)

The expression of switching function in the steady-state can be obtained from (15) with the assumption that $i_1 = i_1^*$ and $v_C = v_C^*$

\[
d_{ss} = \frac{1}{V_{PN}} \left( L_i \frac{di_1^*}{dt} + R_i i_1^* + v_C^* \right).
\] (20)

Now, substituting (13) into (15)–(17) results in

\[
L_i \dot{x}_1 = d_p V_{PN} - R_i x_1 - x_3,
\] (21)

\[
L_o \dot{x}_2 = x_3 - R_o x_2,
\] (22)

\[
C_f \dot{x}_3 = x_1 - x_2,
\] (23)

where $\dot{x}$ represents the time derivative of $x$. Now, substituting (21)–(23) into (14) yields

\[
\dot{V}(x) = d_p V_{PN} x_1 - R_i x_1^2 - R_o x_2^2.
\] (24)

It is evident from (24) that $\dot{V}(x)$ is always negative if $d_p$ is selected as

\[
d_p = K_c V_{PN} x_1,
\] (25)

where $K_c < 0$. The overall switching function can be written by combining (20) and (25) as

\[
d = \frac{1}{V_{PN}} \left( L_i \frac{di_1^*}{dt} + R_i i_1^* + v_C^* \right) + K_c V_{PN} x_1.
\] (26)
The capacitor voltage reference in (26) is defined as

\[ v_C^* = L_o \frac{di_1^*}{dt} + R_o i_2^* + v_s. \]  

The reference for the inverter current \((i_1^*)\) can be computed using \(i_1^* = i_1^* + i_2^*\) where \(i_2^* = C_f dv_C^*/dt\). However, \(i_2^*\) would be computed inaccurately when \(L_o, R_o,\) and \(C_f\) deviate from their actual values. In this case, \(i_1^*\) will be inaccurate as well which leads to steady-state error in the grid current \(i_2\). Hence, in order to avoid these problems, \(i_1^*\) is produced through a PR controller which uses the following transfer function

\[ G_{PR}(s) = K_p + \frac{2K_r \omega_c s}{s^2 + 2\omega_c s + \omega_r^2}, \]  

where \(K_p\) and \(K_r\) denote the proportional and resonant gains, \(\omega_c\) denotes the cut-off frequency and \(\omega\) denotes the resonant frequency. The transfer function \(G_{PR}(s)\) produces \(i_1^*\) by processing the grid current error \((i_2^* - i_2)\) applied to its input. As a consequence of using PR controller, not only \(i_1^*\) is generated accurately, but also \(i_2\) tracks its reference in the steady-state with zero error. However, when (26) is used to control the ac-side variables \((i_1, i_2, \text{and } v_C)\), the closed-loop system may not exhibit satisfactory response (or even may become unstable) since \(d\) does not contain capacitor voltage feedback. Therefore, the perturbed duty cycle in (25) is modified by adding the capacitor voltage loop as follows

\[ d_p = K_v v_{PN} x_1 - K_v x_3, \]  

where \(K_v\) is a constant. Now, the total switching function can be obtained as

\[ d = \frac{1}{V_{PN}} \left( L_i \frac{di_1^*}{dt} + R_i i_1^* + v_C^* \right) + K_v V_{PN} x_1 - K_v x_3. \]  

The closed-loop transfer function which relates the reference grid current to actual grid current is an important tool in determining the behavior of the closed-loop system in the frequency domain. In order to investigate the performance of the closed-loop system at the worst case, the closed-loop transfer function is derived by neglecting \(R_i\) and \(R_o\) (since \(R_i\) and \(R_o\) provide passive damping). Now, substituting (30) into (15) yields

\[ L_i \frac{di_1}{dt} = L_i \frac{di_1^*}{dt} + v_C^* + K_v v_{PN} (i_1 - i_1^*) - K_v v_{PN} (v_C - v_C^*) - v_C. \]  

The expressions for \(i_1\) and \(v_C\) in terms of \(i_2\) when \(v_C = 0\) can be written as

\[ i_1 = L_o C_f \frac{d^2 i_2}{dt^2} + i_2, \]  

\[ v_C = L_o \frac{di_2}{dt}. \]  

Substituting (27), (32), and (33) into (31) results in

\[ L_i C_f L_o \frac{d^3 i_2}{dt^3} - K_v v_{PN}^2 C_f L_o \frac{d^2 i_2}{dt^2} + L_i \frac{di_2}{dt} - K_v v_{PN} i_2 = L_i \frac{di_1^*}{dt} - K_v v_{PN} i_1^*. \]  

The Laplace transform of (34) can be written as

\[ I_2(s) \left( L_i C_f L_o s^3 - K_v v_{PN}^2 C_f L_o s^2 + L_i s - K_v v_{PN}^2 \right) = I_1^*(s) \left( L_i s - K_v v_{PN}^2 \right). \]
The Laplace transform of $i_1^s$ can be written as

$$I_1^s(s) = (I_2^s(s) - I_2(s)) \frac{K_ps^2 + 2\omega_c(K_p + K_r)s + K_p\omega^2}{s^2 + 2\omega_c s + \omega^2}.$$ (36)

Now, substituting (36) into (35) and collecting similar terms, one can obtain the transfer function from reference grid current to actual grid current as follows

$$H(s) = \frac{I_2(s)}{I_2^s(s)} = \frac{a_3s^3 + a_2s^2 + a_1s + a_0}{b_5s^5 + b_4s^4 + b_3s^3 + b_2s^2 + b_1s + b_0},$$ (37)

where

$$a_3 = L_iK_p + L_o(1 + K_vV_{PN})$$
$$a_2 = 2\omega_cL_i(K_p + K_r) - K_pK_vV_{PN}^2 + 2\omega_cL_o(1 + K_vV_{PN})$$
$$a_1 = \omega^2L_iK_p - 2\omega_cK_vV_{PN}^2(K_p + K_r) + \omega^2L_o(1 + K_vV_{PN})$$
$$a_0 = -\omega^2K_pK_vV_{PN}^2$$
$$b_5 = L_oC_fC_o$$
$$b_4 = 2\omega_cL_oC_fL_o - C_fL_oK_vV_{PN}^2$$
$$b_3 = L_i + L_o + \omega^2L_oC_fL_o - 2\omega_cC_fL_oK_vV_{PN}^2 + L_iK_p + L_oK_vV_{PN}$$
$$b_2 = 2\omega_c(L_iK_p + L_iK_r + L_i + L_o + L_oK_vV_{PN}) - K_vV_{PN}^2(1 + K_p + \omega^2C_fL_o)$$
$$b_1 = \omega^2(L_i + L_o + L_iK_p + L_oK_vV_{PN}) - 2\omega_cK_vV_{PN}^2(K_p + K_r + 1)$$
$$b_0 = -\omega^2K_vV_{PN}^2(1 + K_p).$$

The denominator of $H(s)$ constitutes the characteristic equation of the closed-loop system. All coefficients in the characteristic equation are positive constants, since $K_c < 0$, $K_v > 0$, $K_p > 0$, and $K_r > 0$. It is obvious from the characteristic equation that there are five poles. Locations of these poles determine the dynamic response of the closed-loop system. The root locus of the poles, obtained with the system parameters shown in Table 1 by varying $K_c$ in the interval $[-0.0001, -0.003]$ while $K_v$ is maintained at 0.875, 0.5, and 0.3, are shown in Figure 4. It can be seen that while three poles ($p_1$, $p_2$, $p_3$) are strongly affected from $K_c$ and $K_v$, the other poles ($p_4$, $p_5$) are slightly affected from $K_c$ and $K_v$. From Figure 4, it can be easily observed that when $K_v$ is increased, while break-in point becomes larger, the break-away point decreases. Hence, as a consequence of varying $K_v$, all poles (including $p_4$ and $p_5$) move away from the imaginary axis until break-in point occurs. This means that $K_v$ has the ability to move all poles away from the imaginary axis to some extent (until break-in point). Moving poles away from the imaginary axis, not only strengthens the stability, but also makes the dynamic response faster. However, when $|K_c|$ is increased further, while $p_1$ continues to move away from the imaginary axis, $p_2$ and $p_3$ change direction and move toward the imaginary axis as shown in Figure 4a,c,e. The other poles ($p_4$, $p_5$) also move away from the imaginary axis when $|K_c|$ is increased as shown in Figure 4b,d,f. But their effect on the dynamic response is limited. Although the system in (15)–(17) is linear, its control using a linear controller is a challenging issue due to the inherent resonance introduced by the LCL filter. That’s why nonlinear controllers are widely used to achieve the desired performance for such inverter systems [4,8–10,12–15].
2.2.2. DC-Side Control with 2ω Ripple Suppression

In order to achieve the control of voltages \(V_{C1}, V_{C2}, V_{C3},\) and \(V_{C4}\) and currents \(I_{L1}, I_{L2}, I_{L3},\) and \(I_{L4}\) in the dc-side, the shoot-through duty cycle \(d_{ST}\) should be controlled. Once, the control of \(V_{C2}, V_{C3},\) and \(I_{L1}\) are accomplished successfully, the control of other dc variables are done automatically. In this study, the control of dc-side variables is achieved by using proportional-integral (PI) controllers. The inductor current reference \(I_{L1}^*\) is generated as follows

\[
I_{L1}^* = K_{P1}(V_{C2}^* - V_{C2} + K_{ω}\overline{V}_{L1}) + \frac{1}{K_{I1}}\int (V_{C2}^* - V_{C2} + K_{ω}\overline{V}_{L1})dt + K_{P1}(V_{C3}^* - V_{C3} + K_{ω}\overline{V}_{L1}) + \frac{1}{K_{I1}}\int (V_{C3}^* - V_{C3} + K_{ω}\overline{V}_{L1})dt, \tag{38}
\]

where \(K_{ω}\) is a constant for 2ω ripple suppression and \(\overline{V}_{L1}\) denotes the average of \(V_{L1}\). The voltage across \(L_1\) is measured by a voltage sensor and its average can be computed by using the built-in block available in the real time controller platform such as Dspace and OPAL-RT. In the conventional dc-side control, \(I_{L1}^*\) involves 2ω ripple which occurs on \(I_{L1}\) as a result of the applied control [24,28]. In order to suppress 2ω ripple on \(I_{L1}\), it is required to produce \(I_{L1}^*\) without 2ω ripple. If \(I_{L1}^*\) does not involve 2ω ripple, then \(I_{L1}\) will not contain 2ω ripple provided that \(I_{L1}\) tracks \(I_{L1}^*\) successfully. Here, if \(K_{ω}\overline{V}_{L1}\) is fed back into \((V_{C2}^* - V_{C2})\) and \((V_{C3}^* - V_{C3})\), then \(I_{L1}^*\) can be produced without 2ω ripple. Figure 5 shows the simulation results of \(I_{L1}\) and \(I_{L1}^*\) without and with 2ω ripple suppression method.
Having generated $I_{L1}^*$, it can be used in the input of third PI controller to generate the shoot-through duty cycle as

$$d_{ST} = K_{I2}(I_{L1}^* - I_{L1}) + K_{I2} \int (I_{L1}^* - I_{L1}) dt. \quad (39)$$

Now, $d_{ST}$ can be combined with $d$ so that the desired pulse width modulation (PWM) signals for the switches of NPC-qZSI can be generated.

![Figure 5](image_url)

**Figure 5.** Reference and actual inductor currents when (a) $2\omega$ suppression is not activated, (b) $2\omega$ suppression is activated.

### 2.2.3. Gate Signal Generation

The presented block diagram in Figure 3c illustrates the gate signal generation scheme. To generate gate signals for switches of NPC-qZSI, the level shifted pulse width modulation approach is utilized.

The output of Lyapunov-function-based control block constitutes the source for the PWM signal generation process. The switching functions (modulation signals) should have $180^\circ$ phase difference as seen in Figure 3c. To generate the PWM signals, the modulation signals should be compared with the level-shifted triangular signals. To do that, the carrier signals (Car1 and Car2), are generated which have the same phase shift, but are shifted vertically from each other.

Although the single-phase NPC-qZSI, which is illustrated in Figure 1, consists of eight switching devices, four switching signals are required to generate three different voltage levels. The generated switching signals ensure that the NPC-qZSI operates in buck mode. To allow the inverter to operate in boost mode, the shoot-through state should be considered in the gate signal generation scheme. It can be seen that the $d_{ST}$ is obtained from dc-side control loop. To combine non-shoot through duty ratio and $d_{ST}$, primitive logic gates are used. The switching devices on the first leg are turned-on when either Car1 is greater than $d_{ST}$ or Car2 is greater than $d_{ST} - 1$. The shoot-through state generation can be expressed as

$$I_{f} \left( \begin{array}{c} \text{Car1} > d_{ST} \\ \text{OR} \\ \text{Car2} > d_{ST} - 1 \end{array} \right) \text{then turn on } S_{1a} - S_{4a}. \quad (40)$$

### 3. Results and Discussion

The proposed control idea and theoretical considerations are verified experimentally with a 2.5 kW grid-connected 3L-NPC-qZSI prototype. The system and control parameters are shown in Table 1.
Table 1. System and Control Parameters.

| Parameter                           | Value   |
|-------------------------------------|---------|
| Input dc voltage \((V_{in})\)       | 200 V   |
| qZS network inductances \((L_1 - L_4)\) | 0.5 mH  |
| qZS network capacitances \((C_1 - C_4)\) | 470 µF  |
| Filter inductance \((L_{i1}, L_{i2})\) | 1.5, 0.5 mH |
| Filter resistance \((R_i, R_o)\)     | 0.1, 0.05 Ω |
| Filter capacitance \((C_f)\)         | 22 µF   |
| Grid voltage \((v_g)\)               | 220 Vrms|

AC-Side Controller

| Proportional gain \((K_p)\)            | 5       |
| Resonant gain \((K_r)\)               | 1000    |
| Control gains \((K_c, K_v)\)          | −0.0008, 0.875 |

DC-Side Controller

| Proportional gains \((K_{P1}, K_{P2})\) | 1.72, 1.2 |
| Integrator gains \((K_{I1}, K_{I2})\)  | 3.03, 2.1 |
| Control gain \((K_ω)\)                 | 20       |

Figure 6 shows the steady-state experimental results of dc- and ac-side variables. The reference capacitor voltage \((V_{C2}^* \text{ and } V_{C3}^*)\) is 175 V and the reference grid current amplitude \((i_2^*)\) is 10 A (max) during this test. It is obvious that the grid current \((i_2)\) is in phase with the grid voltage \((v_g)\). The measured THD of the grid current is 2.2% which is reasonably small and within the international standards for grid-connected inverters (see Figure 6b). It can be seen that the capacitor voltages are \(V_{C2} = V_{C3} = 175\) V and \(V_{C1} = V_{C4} = 75\) V. The dc-link voltage \((V_{PN})\), which is sum of the capacitor voltages, is 500 V. This result verifies that experimental results are are in good agreement with the theoretical analysis, which are presented in Section 2.1. Furthermore, the inductor current \((I_{L1})\) is continuous and \(2\omega\) ripple is eliminated by the proposed approach, which reduces the stress in the qZS network. It is important to note that \(2\omega\) ripple is eliminated only in the inductor currents not in the capacitors and dc-link voltages. However, the magnitude of these ripples are very small and depend on the qZS network design.

Figure 6. Experimental results of the steady-state analysis. (a) ac and dc side variables; (b) Harmonic spectrum of grid current. \((i_1 = 20\) A/div, \(i_2 = 5\) A/div, \(v_X = 100\) V/div, \(v_{inv} = 150\) V/div, \(I_{L1} = 10\) A/div, \(V_{C1-4} \), \(V_{PN} = 100\) V/div).

Figure 7 presents the steady-state results of without and with the \(2\omega\) ripple suppression technique. Please note that all system parameters are kept exactly the same for both tests. Figure 7a shows the experimental results when the dotted and highlighted part in
Figure 3b is disabled. It can be seen that the qZS inductor has around 4 A peak-to-peak $2\omega$ ripple current. The relevant fast Fourier transform (FFT) analysis in Figure 7c shows that the inductor current contains the DC component as well as 100 Hz ($2\omega$) ripple current component. On the other hand, Figure 7c shows the experimental results when the dotted and highlighted part in Figure 3b is enabled. Comparing Figure 7a with Figure 7b, one can see that the $2\omega$ ripple on the inductor currents is eliminated by the proposed dc-side controller. In addition, the relevant FFT analysis in Figure 7d shows that the inductor current does not contain 100 Hz ($2\omega$) ripple current component. Although there are high frequency components on the inductor current, their amplitudes are very low and can be neglected.

Figure 7. (a) Experimental steady-state results without double-line frequency reduction, (b) Experimental steady-state results with double-line frequency reduction, (c) Simulated spectrum of $I_{L1}$ in part (a), and (d) Simulated spectrum of $I_{L1}$ in part (b). ($i_1 = 20$ A/div, $i_2 = 5$ A/div, $v_g = 100$ V/div, $I_{L1} = 10$ A/div, $V_{C1-C4}, V_{PN} = 100$ V/div).

Figure 8a presents the transient responses of ac-side variables. Please note that during this test the $V_{C2}^*$ and $V_{C3}^*$ are constant and their values are 175 V. The initial grid current reference amplitude ($i_{2}^*$) is 5 A. Then, the $i_{2}^*$ is stepped up from 5 to 10 A. It can be seen that the proposed control strategy shows a fast transient response to this abrupt reference variation without endangering the stability and damping. Although there are oscillations on the dc-side variables, they were damped in a short time (less than 60 ms) by the controller.

To test the transient performance of the dc-side controller, the capacitor voltage reference is stepped up from 150 to 175 V as shown in Figure 8b. It can be seen that initial capacitor voltages are $V_{C2} = V_{C3} = 150$ V and $V_{C1} = V_{C4} = 50$ V, which are consistent with the theoretical calculations. Results of this test show that the capacitor voltages follow their references accurately and reasonably small rising and settling times. Please note that the dynamic response on the dc-side does not affect the ac side operation.
Figure 8. Experimental result of the dynamic response. (a) step on the grid current; (b) step on the capacitor voltage ($i_1 = 20$ A/div, $i_2 = 5$ A/div, $v_g = 100$ V/div, $I_{L1} = 10$ A/div, $V_{C1-C4}, V_{PN} = 100$ V/div).

Figure 9 shows the steady-state results obtained with $K_v = 0.15$. It should be noted that the other control parameters are exactly the same as in previous results. These results reveal that the performance of the closed-loop system in the ac-side is adversely affected when $K_v$ is decreased. The main reason of this degradation comes from the fact that some poles are placed close to the imaginary axis.

To verify the proposed method under different power factor, the simulation studies were performed. The performance of the system under lagging, unity, and leading power factors with and without $2\omega$ ripple reduction technique are shown in Figure 10a–c, respectively. In these results, while the active power is denoted by $P_g$, the reactive power is denoted by $Q_g$. Furthermore, the relevant FFT analysis of these operation modes are presented in Figure 11. The results show that the inductor current does not contain 100 Hz ($2\omega$) ripple current component when the $2\omega$ ripple reduction technique is enabled. Although there are high frequency components on the inductor current when the $2\omega$ ripple reduction technique is disabled, their amplitudes are very low and can be neglected. To sum up, the results show that the performance of the $2\omega$ ripple reduction technique is satisfactory for all operating conditions.

Figure 9. Experimental results when $K_v = 0.15$ ($i_1 = 20$ A/div, $i_2 = 5$ A/div, $v_g = 100$ V/div, $I_{L1} = 10$ A/div, $V_{C1-C4}, V_{PN} = 100$ V/div).
Figure 10. Simulation results of different power factors with and without the $2\omega$ reduction technique; (a) $P_g = 1.5$ kW, $Q_g = -0.5$ kVAR; (b) $P_g = 1.5$ kW, $Q_g = 0$ kVAR; (c) $P_g = 1.5$ kW, $Q_g = +0.5$ kVAR.
The advantages and disadvantages of the existing ac-side control methods and the proposed control method are summarized in Table 2. According to the comparison in Table 2, the proposed control method offers advantages in terms of fast dynamic response, strengthened stability, reduced number sensors, excellent resonance damping, zero steady-state error and low THD in grid current.
Table 2. Existing AC-side control methods compared with the proposed control method.

| Comparison Category | [14] | [11–13] | [15] | [16] | [18,19] | Proposed Control |
|---------------------|------|---------|------|------|---------|-----------------|
| Type of damping     | Passive damping | VR-based active damping | VR-based active damping | VR-based active damping | Filter-based active damping | VR-based active damping |
| Advantages          | [14]: Easy implementation | [11–13]: Easy implementation, fast dynamic response, strong robustness and disturbance rejection | [15]: Fast dynamic response, extensive stability analysis, requires two sensors | [16]: Easy implementation, fast dynamic response, low THD in $i_2$ | [18,19]: Requires one sensor | Proposed control: Fast dynamic response, strengthened stability, requires two sensors, zero steady-state error and low THD in $i_2$, and very good damping feature |
| Disadvantages       | [14]: Additional power losses, reduced efficiency, and not reliable | [11–13]: Chattering and variable-switching frequency, [13]: Requires three sensors | [15]: Sensitive to parameter variations, complicated implementation, high THD and steady-state error in $i_2$ | [16]: Fixed switching frequency with additional computation, no stability analysis, requires three sensors | [18,19]: Sensitive to parameter variations and disturbances | Proposed control: Increased computations |

4. Conclusions

In this study, a high-performance control technique based on Lyapunov’s stability theory for a single-phase neutral-point-clamped quasi-impedance source inverter was proposed. The Lyapunov function based control is employed to regulate the inverter output current, whereas the proportional resonant controller is used to produce the reference of the inverter output current that is needed in the Lyapunov function based control. Use of proportional resonant controller ensures the zero steady-state error in the grid current.

It is shown that the modified simple boost control technique eliminates the double-line frequency ripple in the quasi-impedance source inductor currents and minimizes the double-line frequency ripples in the quasi-impedance source capacitor voltages. The proposed control technique considerably reduces the inverter size, weight, and cost as well as increases overall system efficiency since the required inductances and capacitances sizes are lower.

Experimental results obtained from a 2.5 kW neutral-point-clamped quasi-impedance source inverter prototype are presented to validate the performance of the controller under various operating conditions, including steady-state and transient responses. The results show that the proposed Lyapunov function based control has a fast response and nearly zero steady-state error. Furthermore, the proposed modified simple boost control technique eliminates the ripples on the inductances and minimizes the ripples on the capacitors.

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Nomenclature

The following abbreviations and nomenclature are used in this manuscript:

Abbreviations

- AC: Alternating current
- DC: Direct current
- MI: Multilevel inverters
- NPC: Neutral point clamped
- NPC-qZSI: Neutral point clamped quasi-impedance source inverter
- PI: Proportional-integral
- PID: Proportional-integral-derivative
- PR: Proportional resonant
- PV: Photovoltaic
- qZSI: quasi-impedance source inverter
- THD: Total harmonic distortion
- VR: Virtual resistor

Nomenclature

- $B$: Boost factor
- $C_f$: The filter capacitance (µF)
- $C_1 - C_4$: qZS network capacitances (µF)
- $d_{ST}$: Shoot-through duty-cycle (%)
- $d$: The switching function
- $d_{ss}$: The steady state term of the switching function
- $d_p$: The perturbed term of the switching function
- $G_{PR}$: The transfer function of proportional resonant controller
- $i_L$: Inductor current (A)
- $I_1$: The output current (A)
- $K_p$: AC-side proportional gain
- $K_r$: AC-side resonant gain
- $K_c, K_v$: AC-side control gains
- $K_{p1}, K_{p2}$: DC-side proportional gains
- $K_{i1}, K_{i2}$: DC-side integral gains
- $K_\omega$: DC-side control gain
- $L_1 - L_4$: qZS network inductances (mH)
- $L_f, L_o$: The filter inductances (mH)
- $P_o$: The inverter output power (W)
- $R_i, R_o$: The resistances of filter inductances (ohm)
- $T_0$: Shoot-through period (sec)
- $T_1$: Non-shoot-through period (sec)
- $V_{PN}$: DC-link voltage (V)
- $v_{inv}$: The output voltage (V)
- $V_{in}$: Input dc voltage (V)
- $V_{L1-L4}$: The voltage across the inductors (V)
- $V_{C1-C4}$: The voltage across the capacitors (V)
- $v_C$: The filter capacitor voltage (V)
- $\omega$: Fundamental angular frequency (rad/s)
- $\theta$: Phase angle (rad)
- $2\omega$: Double-line frequency (rad/s)

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