Discrete time model based multiple paths full feedforward control for three-phase inverter with output transformer

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Abstract
In a high-power three-phase voltage source inverter, a step-up transformer may be used at the output of the LC filter to achieve higher output voltage, improve electromagnetism compatibility and serve as output inductor under parallel operation. However, under standalone operation, the performance of the inverter is degraded since the feedback control does not include the output voltage of the transformer. This paper proposes a multiple paths load current full feedforward control designed in discrete time domain to improve the performance of the inverter. The proposed full feedforward controller has a concise expression due to multiple feedforward paths, and it is easy to fit in other control structures. Additionally, direct analysis in discrete time domain offers an accurate approach to investigate the relationships of the variables, and the improper fractions in the full feedforward controller can be handled with fewer trade-offs. A lower output impedance is achieved eventually. MATLAB/Simulink is used to present the handling of advance operator $z$, ringing phenomenon and transformer dynamic. Experiments of a TMS320F28335 processor based 75 kVA three-phase inverter with output transformer prove that the proposed controller can further improve the dynamic and steady-state performance of the inverter under linear and nonlinear loads.

1 | INTRODUCTION

Three-phase voltage source inverter (VSI) that aims to supply sinusoidal output voltage of high quality to critical loads is widely used in various fields, such as vessels, medical equipment, financial databases and communication systems [1]. In grid interactive ac microgrid or renewable energy conversion fields, the VSI can serve as a grid-forming inverter and provide ac voltage with fixed amplitude and frequency [2]. As the loads grow, the reliability and the power capability of the power source can be increased by replacing single VSI with multiple VSIs connected in parallel [3]. In order to avoid surge current under parallel operation, an inductor in series with the LC filter of the inverter is usually used. In some cases, a transformer is used in the VSI to protect the inverter by isolating the input and the output [4–9]. And in some special situations such as in some large vessels or remote area that uses renewable energy as its power source, the VSI needs to achieve higher output voltage to compensate the voltage drop caused by the long cables, and it also needs to have a better electromagnetism compatibility, thus a step-up transformer is usually used at the output side of the VSI. The additional advantage of this connection is that it can also make the transformer serve as an output inductor during parallel operation and avoid the switching frequency harmonic current that would increase iron loss flowing directly through the transformer.

Whether the VSIs work in parallel or not, the feedback controller should be carefully designed to make sure that the inverter has low steady-state error, fast dynamic response and low total harmonic distortions (THD). Recently for the VSI, plenty of control strategies are proposed. Based on the internal model principle, the proportional-integral (PI) based controllers are used in [10–12], the proportional-resonant (PR) based controllers are used in [13–15] and the repetitive control is applied in [1] and [16]. Besides, there are some other control methods brought up by the researchers, such as the adaptive control [17], the sliding-mode control [18, 19] and the model predictive control [20–22].
The aforementioned control strategies all aim to achieve command tracking. However, the output voltage of the VSI is also affected by the load current, and the load current is commonly regarded as a disturbance input of the VSI. One way to reduce the influence of the load current (disturbance rejection) is to increase the loop gain, but this approach is limited by the stability requirement. The other way to achieve disturbance rejection is reducing the output impedance of the VSI. Normally in a dual-loop control strategy, the capacitor current feedback is used as the inner loop to improve the disturbance rejection capability, because it can make the VSI achieve a lower output impedance when compared with the one that uses inductor current feedback as the inner loop [23]. However, the methods mentioned above can hardly improve the performance of the VSI when the step-up transformer is placed behind the LC filter, because the output voltage of the transformer is not included in the feedback control loop, the voltage drop caused by the leakage impedance cannot be compensated.

Another solution to reduce the influence of the load current is load current feedforward control. In an early study [10], load current and load current derivative feedforward control for a single-phase VSI are presented, respectively, to achieve disturbance rejection, the feedforward gain of the load current is set to one, and the feedforward gain of the load current derivative is related to the filter inductor. The load current proportional feedforward control is commonly used to avoid the influence of the load current because of its easy implementation. In [9], a first-order load current feedforward controller based on reducing the output impedance of the three-phase VSI with output transformer is proposed; this feedforward strategy originates from the load current full feedforward control that contains the inverse model of the VSI, the improper fraction appears in the full feedforward controller, and due to its implementation issue, trade-offs are made to simplify the original full feedforward controller into a first-order feedforward controller. In [2], a load current feedforward control method based on pole-zero cancellation is used, the load current feedforward controller is used to eliminate the influence of the dominant pole; however, the influence of the nondominant poles still exist. A predictive load current feedforward is proposed in [24], the feedforward controller is also derived based on decreasing the output impedance of the VSI, and the feedforward signal is generated by the repetitive predictor; however, the improper fraction also occurs in the transfer function of the feedforward controller, but the detail to deal with this problem is not presented, besides, only the nonlinear load condition is taken into consideration. The aforementioned load current feedforward strategies are all designed in continuous time domain. In [25], the authors suggest that it is more accurate to design the controller in discrete time domain directly, because in discrete time domain, the time delay does not need to be represented in approximated rational transfer functions, in addition, direct design in discrete time domain can avoid the discrepancies introduced by different discretization methods [26]. In [27], a linear quadratic based load current feedforward control is proposed in discrete time domain to achieve dominant poles elimination, the feedforward gain is calculated by $H_{\infty}$ norm and zero dynamic method. In [28], a load current feedforward control applied in synchronous frame is presented, although this feedforward control approach is designed in discrete time domain, it only focuses on reducing the output impedance of the VSI at fundamental frequency. In addition to the load current feedforward control, the disturbance observer based approach is used to achieve disturbance rejection in [29, 30].

The research object is shown in Figure 1; it contains a three-phase VSI with a step-up transformer and different kinds of loads which are used to verify the proposed control method. In order to achieve disturbance rejection and improve the performance of the VSI that operates at standalone mode, the load current full feedforward control is more preferable, because it can theoretically eliminate the influence of the load current. However, it is difficult for the full feedforward controller designed in continuous time domain to satisfy the compensation demand for its lack of accuracy and the drawback of handling improper fraction introduced by the inverse model of the plant. Therefore, this paper proposes a discrete time model based multiple paths full feedforward control strategy according to [31]. The multiple feedforward paths make sure that the transfer functions of the feedback controllers with complicated expression are not included in the feedforward controller in order to realize full feedforward control, therefore, the proposed controller becomes more concise and it can easily fit in other control structures. Due to the direct analysis in discrete time domain, the proposed control strategy can fully consider the coupling within the discrete time model of the VSI, and the implementation of improper fraction can be handled nicely with fewer trade-offs, then the VSI with output transformer can achieve a lower output impedance at fundamental frequency as well as low-order harmonic frequencies accordingly. A lower output impedance would further improve the dynamic and steady-state performances of VSI under different loads.
2 | DISCRETE TIME MODEL OF THREE-PHASE VSI

According to Figure 1, the equivalent single-phase state-space equation of the three-phase VSI in continuous time domain is derived in Equation (1), where \( u_{op}, i_{op}, C, I_n, r, u_1, V_{DC} \) and \( i_{op} \) represent the filter capacitor voltage (primary winding voltage of the transformer), filter inductor current, filter capacitor, filter inductor, equivalent resistor in series (including the dead-time effect and so on), reference modulated wave, DC voltage and the load current that flows in primary winding of the transformer respectively.

\[
\begin{bmatrix}
\dot{u}_{op} \\
\dot{i}_{op}
\end{bmatrix} =
\begin{bmatrix}
0 & \frac{1}{C} \\
\frac{1}{L} & -\frac{r}{L}
\end{bmatrix}
\begin{bmatrix}
u_{op} \\
i_{op}
\end{bmatrix} +
\begin{bmatrix}
0 \\
\frac{V_{DC}}{L}
\end{bmatrix}
i_L +
\begin{bmatrix}
-\frac{1}{C} \\
0
\end{bmatrix}
i_{op}
\]

\[
= A\begin{bmatrix}u_{op} \\i_{op}\end{bmatrix} + B_i + W_{i_{op}}
\tag{1}
\]

The discrete time model is derived according to zero-order hold method with sample period \( T \) [32]:

\[
X(k+1) = \Phi X(k) + H_1 n_i(k) + H_2 i_{op}(k)
\tag{2}
\]

where

\[
X(k) = \begin{bmatrix} u_{op}(k) \\ i_{op}(k) \end{bmatrix}, \quad \Phi = e^{AT} = \begin{bmatrix} \phi_{11} & \phi_{12} \\ \phi_{21} & \phi_{22} \end{bmatrix}
\tag{3}
\]

\[
H_1 = \int_0^T e^{AT} dB = \begin{bmatrix} h_{11} \\ h_{21} \end{bmatrix}, \quad H_2 = \int_0^T e^{AT} dW = \begin{bmatrix} h_{12} \\ h_{22} \end{bmatrix}
\]

And Figure 2 [32] depicts the relationship of the variables in Equations (2) and (3).

Then, the transfer function of the VSI in discrete time domain is expressed as follows:

\[
u_{op}(k) = n_i(k) INV_a + i_{op}(k) INV'_a
\tag{4}
\]

where

\[
INV_a = \frac{h_{11} \xi - h_{12} \phi_{22} + h_{23} \phi_{12}}{\xi^2 - (\phi_{11} + \phi_{22}) \xi + \phi_{11} \phi_{22} - \phi_{12} \phi_{21}}
\tag{5}
\]

3 | DISCRETE TIME MODEL BASED MULTIPLE PATHS FULL FEEDFORWARD CONTROL

3.1 | Outline of the whole control structure

First of all, there is a brief introduction of the whole control structure before the derivation of the proposed feedforward controller. In Figure 3, the feedback control loop is marked with grey. A dual-loop PR control scheme and a proportional feedback of the reference modulated wave is used in \( \alpha \beta \) frame to guarantee the stability and the performance of the three-phase VSI. The VSI block in Figure 3 stands for the model of the inverter in Figure 2. The PR controllers \( PR_a \) and \( PR_r \) for fundamental frequency are used in inner loop and outer loop, respectively, and the parameter \( K \) is a proportional gain. In order to suppress harmonic distortion of the output voltage, several harmonic resonant controllers are added in the voltage control loop (they are not depicted in Figure 3). The controllers \( PR_s \) and \( PR_t \) for fundamental frequency can be expressed as

\[
INV'_a = \frac{b_{12} \xi - b_{11} \phi_{22} + b_{22} \phi_{12}}{\xi^2 - (\phi_{11} + \phi_{22}) \xi + \phi_{11} \phi_{22} - \phi_{12} \phi_{21}}
\tag{6}
\]

Notice that the transfer functions in Equations (5) and (6) possess the same denominator polynomial.

Compared with the continuous time model of the VSI, the discrete time model is more complicated, because it contains more coupling terms. In order to precisely investigate the influence of the load current, it is necessary to analyze directly in discrete time domain, since the discrete time model is more accurate and it can reveal the coupling relationships of different variables. Moreover, when it comes to the digital control implementation, the advance operator \( \xi \) introduced by the improper fraction can be handled with fewer trade-offs in discrete time domain. Therefore, the feedforward controller that is designed directly in discrete time domain can make the VSI with output transformer have a lower output impedance and achieve a better performance. The detailed analysis would be shown in the following sections.
follows:

\[
PR_u(s) = \frac{K_{pu1} s + K_{pu1} \omega_u}{s^2 + 2\omega_u + \omega_u^2} \\
PR_i(s) = \frac{K_{pi1} s + K_{pi1} \omega_i}{s^2 + 2\omega_i + \omega_i^2}
\]

(7)

The harmonic resonant controller with a lead compensator is

\[
K_{rh} \left( \cos \phi_u - \sin \phi_u \right) s^2 + 2\omega_h s + \omega_h^2
\]

(8)

where \( h \) refers to the harmonic order. During the design, the controllers are mapped to discrete time domain by means of bilinear transformation with frequency pre-warping.

In Figure 3, \( u_{u0} \), \( 1/z \), \( Z_T \), and \( u_o \) represent the output voltage reference, one sample period delay of digital control, leakage impedance of the transformer and equivalent output voltage of secondary winding of the transformer (converted from \( u_{o0} \) in Figure 1), respectively.

Because of the existence of the output transformer, the parameters in the secondary winding and the leakage impedance of the transformer are converted into the primary for the convenience of analysis. And the ultimate output of the three-phase VSI is the secondary winding of the transformer, and the equivalent output voltage is given by:

\[
u_o(k) = u_{op}(k) - Z_T i_{op}(k)\]

(9)

According to (9), the equivalent output voltage is influenced by the output voltage reference and the voltage drop caused by the load current. In order to reduce the influence of the load current, as well as further improve the performance of the VSI, the discrete time model based multiple paths full feedforward control is proposed. In Figure 3, the proposed full feedforward controller is encircled by the dashed line.

### 3.2 Multiple paths full feedforward control

The proposed full feedforward control is based on reducing the output impedance of the three-phase VSI with output transformer. The reason for setting multiple feedforward paths is that the transfer function of the feedforward controller does not include the expressions of the PR controllers; a further explanation would be presented later. For the physical meaning of these three feedforward paths, they can be considered as an additional output reference, inductor current reference and reference modulating wave, respectively.

In order to get the expression of the proposed feedforward controller, the transfer function of the output impedance that relates to \( u_o(k) \) and \( i_o(k) \) should be figured out. In order to simplify the design procedure, some equivalent transformations of the block diagrams are necessary. In the first place, the relationship between \( u_{op}(k) \), \( i_{op}(k) \) and \( i_{op}(k) \) should be derived. According to Figure 2, \( i_{op}(k) \) can be derived as

\[
i_{op}(k) = \frac{FF_1 PR_{PR, INV_u} u_o(k) + FF_2 PR_{PR, INV_u} u_o(k) + FF_3 INV_u}{\frac{K}{z-K} + \frac{K}{z-K} + \frac{K}{z-K}}
\]

(12)
In an ideal situation, the full feedforward control can totally eliminate the influence of the load current; in other words, the output impedance can be set to zero due to the application of the full feedforward controller. Therefore, the proposed full feedforward controller can be derived as follows by letting Equation (13) equal to zero:

\[ FF_1 = Z_T \]  

\[ FF_2 = -\frac{b_{11}}{\phi_{12}} + \frac{b_{11}}{INV'_u} \left( \frac{b_{21} - b_{22} + b_{21}^2}{b_{11} - b_{12}^2} \right) z - K \]  

\[ FF_3 = \left( -\frac{INV'_u}{INV_u} + \frac{Z_T}{INV_u} \right) (z - K) \]  

Now, suppose that there is only one feedforward path as in the conventional feedforward control method. In order to make the output impedance equal to zero, the full feedforward controller would become complicated for it contains the transfer functions of the PR controllers. That is the reason why the multiple paths full feedforward control strategy is proposed.

4 | DESIGN CONSIDERATIONS

Before implementation, some details should be discussed and some modifications of the proposed feedforward controller would be made.

4.1 | Implementation issue of advance operator \( \xi \)

Notice that in Equation (16), the improper fraction terms \( 1/INV'_u \) [inverse of Equation (5)] and \( (z-K) \) are in \( FF_3 \). It means that the order of numerator polynomial is higher than the order of denominator polynomial in the transfer function. Unfortunately, the inverse of one sample period delay cannot be realized physically \([33]\). An approximation should be made to deal with these two terms to make them realizable and keep their frequency responses close to their original expressions.

One possible approach is to use Taylor expansion to deal with the advance operator \( \xi \). Notice that there is a relationship between the continuous time domain and discrete time domain:

\[ \xi = e^{\xi T} \]  

and then, the advance operator \( \xi \) can be expanded as follows:

\[ \xi = e^{\xi T} = 1 + \xi T + \frac{(\xi T)^2}{2!} + \frac{(\xi T)^3}{3!} + \ldots \]  

Discretizing the first and the second terms of the Taylor expansion by means of backward difference method, the advance operator \( \xi \) can be derived:

\[ \xi \approx 2 - \frac{1}{\xi} \]  

And then, the transfer functions \( 1/INV'_u \) and \( (z-K) \) can be replaced by:

\[ \frac{1}{INV'_u} \approx \frac{2z - 1 - (\phi_{11} + \phi_{22})}{b_{11} - b_{12}^2} z + \phi_{11} \phi_{22} - \phi_{12}^2 \]  

\[ z - K \approx 2 - K - \frac{1}{\xi} \]  

The other solution is to increase the order of the denominator polynomial in both \( 1/INV'_u \) and \( (z-K) \) by multiplying one sample period delay \( 1/z \). And they can be expressed as follows:

\[ \frac{1}{INV'_u} \approx \frac{z^2 - (\phi_{11} + \phi_{22}) z + \phi_{11} \phi_{22} - \phi_{12} \phi_{21}}{z(b_{11} - b_{12}^2 + b_{21} \phi_{12})} \]  

\[ z - K \approx \frac{z}{\xi} - K \approx 1 - K \]  

The bode diagrams of approximation results are plotted for comparison. In Figure 5, the approximation result Equation (22) retains the resonance peak of the original transfer function,
thus the corresponding approximation result matches $1/INV_{u}$ perfectly in magnitude, but an obvious phase shift appears when the frequency is above 1 kHz. While the Taylor expansion method shows less accuracy in both magnitude and phase, because the approximate transfer function Equation (20) is an inverse of a one-order transfer function that does not reflect the resonance peak of a second-order transfer function. In Figure 6, the approximation result in Equation (23) has a lower magnitude amplification at high frequency, while the Taylor expansion method shows a large magnitude amplification at high frequency due to the derivative term. Considering that there is a tiny phase lead at fundamental frequency of the original $(z-K)$, the approximation result in Equation (23) is acceptable. In conclusion, the second approximation by multiplying one sample period delay $1/z$ is more preferable than the one using Taylor expansion.

### 4.2 Ringing elimination

In particular, the ringing (inter-sample ripple) may occur when a digital controller is designed by means of the inverse of the discrete time model of a plant, such as the Dahlin algorithm. If the poles of a digital controller are close to -1, the output of the controller would oscillate at half of the sample frequency when there is a step input. The ringing of the controller would have a negative influence on dynamic response and it must be eliminated.

Notice that in the transfer functions of the proposed feedforward controller in Equations (15) and (16), they all share the same fraction $1/(h_{11}z - h_{11}phi_{22} + h_{22}phi_{12})$ whose pole is very close to -1. The existence of this negative pole in the controller is caused by the poor damping effect of the equivalent series resistors $r$. The negative pole would lead to ringing that degrades the dynamic response of the three-phase VSI. The ringing also can be seen from the bode diagram in Figure 7. The negative pole leads to a huge magnitude amplification at high frequency, and it is not good for the dynamic response of the VSI when there is a step input of the load current.

According to [34], Vogel–Edgar controller is proposed to deal with the ringing. Based on its idea, the fraction above is modified as follows to cancel the negative pole:

$$\frac{1}{h_{11}z - h_{11}phi_{22} + h_{22}phi_{12}} = \frac{1}{(h_{11} - h_{11}phi_{22} + h_{22}phi_{12})z}$$

(24)

In Figure 7, the modified transfer function in Equation (24) eliminates the magnitude amplification, and the phase shift at low frequency is relatively small. For example, at 1 kHz, the magnitude error is less than 0.25 dB, the phase of the original fraction is -11.6°, and the phase of the modified one is -23.1°. The magnitudes and the phases of these two fractions are almost overlapped at low frequency.

After the approximation and modification, according to Equations (5) and (6), the proposed feedforward controller is now updated as follows:

$$FF_1 = Z_T$$

(25)

$$FF_2 = -\frac{h_{12}}{\phi_{12}} + \frac{h_{11}(h_{12}z - h_{12}phi_{22} + h_{22}phi_{12})}{\phi_{12}(h_{11} - h_{11}phi_{22} + h_{22}phi_{12})z} + \frac{Z_T(h_{21}z - h_{21}phi_{11} + h_{11}phi_{21})}{(h_{11} - h_{11}phi_{22} + h_{22}phi_{12})z}$$

$$= FF_{2-1} + FF_{2-2} + FF_{2-3}$$

(26)
According to Equations (4) and (9), the output impedance of the VSI with output transformer contains $\text{INV}_i$ and $Z_T$. And $\text{INV}_i$ is far less than $Z_T$. For the convenience of the analysis, the term $\text{INV}_i$ can be temporarily neglected and the capacitor voltage can be considered as an ideal voltage source in series with the transformer. The equivalent circuit is shown in Figure 9, where $L_T$, $R_T$ and $R_{\text{Load}}$ represent the leakage inductance, leakage resistance and resistive load, respectively. Suppose that the switch is on when $t = 0$ s, the load current is derived as follows:

$$i(t) = -\frac{U_m}{\sqrt{(R_T + R_{\text{Load}})^2 + (\omega L_T)^2}} \cos (\Theta) e^{-\frac{R_T + R_{\text{Load}}}{L_T}t} + \frac{U_m}{\sqrt{(R_T + R_{\text{Load}})^2 + (\omega L_T)^2}} \cos (\omega t + \Theta)$$

(28)

where

$$\Theta = -\tan^{-1}\frac{\omega L_T}{R_T + R_{\text{Load}}} \approx 0$$

(29)

Then, the voltage of leakage inductance $L_T$ is

$$L_T \frac{di}{dt} = \frac{(R_T + R_{\text{Load}})U_m}{\sqrt{(R_T + R_{\text{Load}})^2 + (\omega L_T)^2}} \cos (\Theta) e^{-\frac{R_T + R_{\text{Load}}}{L_T}t} - \frac{R_T + R_{\text{Load}}}{L_T} U_m \omega L_T \sin (\omega t + \Theta)$$

(30)
TABLE 1 Parameters of the 75 kVA three-phase VSI

| Parameter                        | Value         |
|---------------------------------|---------------|
| Rated power                     | 75 kVA        |
| Line-to-line voltage (RMS)      | 390 V         |
| Frequency                       | 50 Hz         |
| Sample frequency (1/T)          | 15.6 kHz      |
| DC voltage (V<sub>dc</sub>)     | 500 V         |
| L                               | 300 μH        |
| C                               | 440 μF        |
| r                               | 0.07 Ω        |
| Rated capacity of transformer   | 75 kVA        |
| Rated frequency of transformer  | 50 Hz         |
| Z<sub>T</sub>                   | (0.02625+0.04172) Ω |

FIGURE 10 Simulation result of the three-phase VSI with the proposed feedforward control from no load to 30 kW. (a) Overall dynamic response. (b) Details of transformer dynamic

According to the parameters in Table 1, at the step moment of 30 kW, the voltage of leakage inductance climbs to nearly 0.999 V<sub>um</sub>, and after 1/7800s (two sample periods), the voltage of leakage inductance drops to 0.011 V<sub>um</sub>, therefore, the leakage inductance has a voltage peak during a short time. This phenomenon occurs when the load current has a step change, and it leads to a sudden drop and recovery on the output voltage. The MATLAB/Simulink simulation result is shown in Figure 10 to indicate the dynamic process of the transformer, where V<sub>AB</sub>, V<sub>BC</sub> and V<sub>CA</sub> are the line-to-line voltage of the secondary winding of the transformer, and i<sub>OA</sub>, i<sub>OB</sub> and i<sub>OC</sub> are the load currents that flow in the secondary winding of the transformer. Notice that the output voltage almost drops to 0 when a step change occurs in load current. Because of the tiny time constant of the circuit and the proposed feedforward controller, the voltage recovers to a higher level quickly.

Obviously, the sudden change of the voltage is caused by the leakage inductance of the transformer, and it cannot be compensated by the feedforward controller. However, in some parts of the feedforward controller (FF<sub>1</sub>, FF<sub>2,3</sub> and FF<sub>3,2</sub>), there is a Z<sub>T</sub> term that would lead to a large derivative output when step load current occurs. The unnecessary input signal of the feedforward controller would degrade the dynamic performance, thus the output values of the transfer functions that contain Z<sub>T</sub> should be limited. The upper and lower limit values are chosen based on the steady-state output values of the transfer functions which include Z<sub>T</sub> under 30% overload condition. The limit is set to avoid the large derivative output, and it also makes sure that the three-phase VSI has the ability to operate under overload condition in a short time. The specific output limiting values of FF<sub>1</sub>, FF<sub>2,3</sub>, FF<sub>3,2</sub> are ±10.58, ±1.46, ±0.0265, respectively.

4.4 Robustness analysis

In practice, the parameter drifting may appear in the filter inductor and filter capacitor; therefore, the investigation on the parameter sensitivity of the proposed feedforward controller is necessary. Suppose that the actual filter inductor L<sub>Real</sub> varies from 0.9L to 1.1L, and the actual filter capacitor C<sub>Real</sub> varies from 0.9C to 1.1C. The comparisons of different situations are presented in Figure 11.

The results show that the output impedance of the VSI almost stays the same with the drifting parameter. In order to verify the robustness of the proposed full feedforward control by experiments, in each phase, an extra capacitor of 120 μF is connected in parallel with the original filter capacitor to make the actual capacitor reach 560 μF (1.272 C). The corresponding output impedance is shown in Figure 11, and it has the highest output impedance at low-order harmonic frequencies among (a) to (g) in Figure 11.

5 EXPERIMENTAL RESULTS

Several groups of experiments are carried out via a 75 kVA three-phase VSI with output transformer, its parameters are shown in Table 1. The details of the experimental setup are given in Figure 1. The control algorithm is implemented in the digital signal processor TMS320F28335. According to Table 1 and Equations (25)–(27), the specific expression of the proposed full feedforward controller is shown in Equations (31)–
The nonlinear load is used to verify the command tracking performance as well as the ability of harmonic distortion suppression; the results are shown in Figure 13. Notice that harmonic resonant controllers are used in feedback control loop, in order to prove that the proposed feedforward control can also reduce the voltage distortion to a certain extent, an extra group of experiments in which the harmonic resonant controllers are taken away is presented in Figure 14. The experimental data of output line-to-line voltages (RMS value), the maximum steady-state error (Errormax) and the maximum THD (THDmax) of the one with the proposed full feedforward control further reduces the voltage drop and the settling time after the step moment. In addition, the results of parameter variation are also better than those of the other three methods.

5.2 Steady-state response

The dynamic performance of the three-phase VSI is verified through a step change of 30 kW. In Figure 12, the load is suddenly added when one of the output line-to-line voltages reaches its peak. The experimental data is recorded in Table 3; the percentage of the instantaneous voltage drop is calculated using the data that excludes the dynamic process of transformer, because the first voltage drop is inescapable. The voltage drop and settling time with the proposed feedforward control is about 47% lower and 87% shorter than the one without feedforward control, respectively, about 46% lower and 87% shorter than the one with the feedforward control in [2], respectively, and about 24% lower and 63% shorter than the one with conventional proportional feedforward control, respectively. The experimental results indicate that the proposed full feedforward control further reduces the voltage drop and the settling time after the step moment. In addition, the results of parameter variation are also better than those of the other three methods.

5.1 Dynamic response

The parameters of the feedback controllers are presented in Table 2. The waveforms of experimental results are the line-to-line voltage and the load current of secondary winding of the transformer. The proposed feedforward controller 8.71% 0.46 ms

| Parameter | Value | Parameter | Value | Parameter | Value |
|-----------|-------|-----------|-------|-----------|-------|
| $K_{pv1}$ | 0.275 | $K_{v1}$  | 8     | $\varphi_1$ | 1.38 rad/s |
| $K_{v1}$  | 471   | $K_{v1l}$ | 8     | $\varphi_{1l}$ | 1.87 rad/s |
| $K_{pv}$  | 0.00330 | 6.67 | $\varphi_3$ | 2.11 rad/s |
| $K_{v1}$  | 0.533 | $K_{v1l}$ | 6.67 | $\varphi_{1l}$ | 2.53 rad/s |
| $K$      | -0.119 | $K_{v23}$ | 5.71 | $\varphi_{19}$ | 2.71 rad/s |
| $K_{v5}$  | 10    | $K_{v25}$ | 5.71 | $\varphi_{23}$ | 3.04 rad/s |
| $K_{v7}$  | 10    | $\varphi_5$ | 1.17 rad/s | $\varphi_{25}$ | 3.19 rad/s |

The nonlinear load is used to verify the command tracking performance as well as the ability of harmonic distortion suppression; the results are shown in Figure 13. Notice that harmonic resonant controllers are used in feedback control loop, in order to prove that the proposed feedforward control can also reduce the voltage distortion to a certain extent, an extra group of experiments in which the harmonic resonant controllers are taken away is presented in Figure 14. The experimental data of output line-to-line voltages (RMS value), the maximum steady-state error (Errormax) and the maximum THD (THDmax) of the one with the proposed feedforward control further reduce the voltage drop and the settling time after the step moment. In addition, the results of parameter variation are also better than those of the other three methods.

5.2 Steady-state response

The dynamic performance of the three-phase VSI is verified through a step change of 30 kW. In Figure 12, the load is suddenly added when one of the output line-to-line voltages reaches its peak. The experimental data is recorded in Table 3; the percentage of the instantaneous voltage drop is calculated using the data that excludes the dynamic process of transformer, because the first voltage drop is inescapable. The voltage drop and settling time with the proposed feedforward control is about 47% lower and 87% shorter than the one without feedforward control, respectively, about 46% lower and 87% shorter than the one with the feedforward control in [2], respectively, and about 24% lower and 63% shorter than the one with conventional proportional feedforward control, respectively. The experimental results indicate that the proposed full feedforward control further reduces the voltage drop and the settling time after the step moment. In addition, the results of parameter variation are also better than those of the other three methods.

5.1 Dynamic response

The parameters of the feedback controllers are presented in Table 2. The waveforms of experimental results are the line-to-line voltage and the load current of secondary winding of the transformer.
FIGURE 12  Dynamic experiments from no load to 30 kW. (a) With the proposed feedforward controller. (b) Without feedforward controller. (c) With the feedforward control in [2]. (d) With conventional proportional feedforward controller. (e) With the proposed feedforward controller (drifting parameter $C_{\text{Real}} = 1.272C$).

FIGURE 13  Steady-state experiments under nonlinear load (the feedback controller includes the harmonic resonant controllers). (a) With the proposed feedforward controller. (b) Without feedforward controller. (c) With the feedforward control in [2]. (d) With conventional proportional feedforward controller. (e) With the proposed feedforward controller (drifting parameter $C_{\text{Real}} = 1.272C$).
Figure 14  Steady-state experiments under nonlinear load (the feedback controller does not include the harmonic resonant controllers). (a) With the proposed feedforward controller. (b) Without feedforward controller. (c) With the feedforward control in [2]. (d) With conventional proportional feedforward controller. (e) With the proposed feedforward controller (drifting parameter $C_{\text{Real}} = 1.272C$).
TABLE 4  Steady-state experimental results under nonlinear load

| Working condition | Feedback controller includes harmonic resonant controllers | Feedback controller excludes harmonic resonant controllers |
|-------------------|-------------------------------------------------------------|-----------------------------------------------------------|
|                   | $U_{AB}$, $U_{BC}$, $U_{CA}$ | $U_{AB}$, $U_{BC}$, $U_{CA}$ | $U_{AB}$, $U_{BC}$, $U_{CA}$ |
| The proposed feedforward controller | 391.2 V, 0.31% 1.52% | 390.1 V, 0.23% 3.50% |
| No feedforward controller | 388.6 V, 0.87% 1.72% | 385.4 V, 1.69% 5.29% |
| Feedback control in [2] | 387.6 V, 0.62% 1.87% | 385.2 V, 1.67% 5.85% |
| Proportional feedforward controller | 387.8 V, 0.77% 1.74% | 386.8 V, 1.23% 4.54% |
| The proposed feedforward controller with drifting parameter ($C_{Real} = 1.272C$) | 390.5 V, 0.21% 1.66% | 389.9 V, 0.15% 3.68% |

(RMS value) and the maximum steady-state error of the output voltage are recorded in Table 5.

Under unbalanced load condition, the percentage of voltage unbalance is calculated according to [35]. The percent unbalance and the Error$_\text{max}$ of the one with the proposed full feedforward control is about 73% and 73% lower than the one without feedforward control, respectively, about 70% and 69% lower than the one with the feedforward control in [2], respectively, and about 72% and 71% lower than the one with proportional feedforward control, respectively.

Under full load condition, the Error$_\text{max}$ of the output voltage that with the proposed feedforward control is about 90% lower than the one without feedforward control, about 90% lower than the one with the feedforward control in [2] and about 89% lower than the one with proportional feedforward control.

Under linear load condition, the steady-state performance is further improved by the proposed full feedforward control. And the parameter variation does not bring in too much influence on the steady-state performance.

The experimental results show that the proposed full feedforward control further improves the performance of the three-phase VSI with output transformer when compared with the results of those without feedforward control, with the feedforward control in [2] and with conventional proportional feedforward control. The proposed method reduces the output impedance of the VSI at fundamental frequency as well as low-order harmonic frequencies. Therefore, the instantaneous error and the settling time of the output voltage are reduced after the step change of load. Under steady-state condition, the voltage drop is compensated and the harmonic distortion caused by nonlinear load is suppressed. Besides, the results also indicate that the proposed full feedforward controller possesses good robustness.

Although the proposed full feedforward controller contains more expressions, it is worth to be applied because it can distinctly improve both dynamic and steady-state response of the VSI with output transformer. And additionally, as it is shown in the bode diagram and the experimental results, the output impedance of the VSI with the feedforward control in [2] and with proportional feedforward controller is not low enough to minimize the influence of the load current.
TABLE 5  Steady-state experimental results under linear load

| Working condition | Unbalance load | Full load |
|-------------------|---------------|-----------|
|                   | $U_{AB}$, $U_{BC}$, $U_{CA}$ | Percent voltage unbalance, $\text{Error}_{\text{max}}$ | $U_{AB}$, $U_{BC}$, $U_{CA}$ | $\text{Error}_{\text{max}}$ |
| No feedforward controller | 390.5 V, 384.5 V, 388.5 V | 0.84% 1.41% | 382.2 V, 381.7 V, 381.2 V | 2.51% |
| Feedforward control in [2] | 391.0 V, 385.3 V, 388.5 V | 0.76% 1.22% | 380.5 V, 381.2 V, 381.2 V | 2.50% |
| Proportional feedforward controller | 390.9 V, 384.8 V, 388.2 V | 0.82% 1.33% | 382.5 V, 382.3 V, 380.8 V | 2.36% |
| The proposed feedforward controller with drifting parameter ($C_{\text{diff}} = 1.272C$) | 391.5 V, 391.2 V, 390.6 V | 0.13% 0.38% | 391.2 V, 391.2 V, 390.7 V | 0.31% |

6 | CONCLUSION

This paper proposes a multiple paths full feedforward control strategy for a three-phase VSI with output transformer that works in standalone mode. The multiple feedforward paths make the full feedforward controller almost independent of the feedback controller, and the proposed controller can easily fit in other control structures with a concise expression. Analysis in discrete time domain is more accurate and it can fully consider the coupling within the discrete time model, and the implementation issue of the transfer functions that include improper fraction can be handled more nicely to retain the original frequency response as much as possible. In addition, the details of handling the advance operator $\zeta$ introduced by improper fraction and ringing are provided in this paper. The proposed feedforward control contributes to a lower output impedance and a better compensation performance when compared with the existing feedforward control strategies. Experimental results prove that the proposed full feedforward control can distinctly reduce the voltage drop, settling time, steady-state error and the THD of the output voltage, and it also owns good robustness. The dynamic and steady-state performance of the three-phase VSI with output transformer under different loads is improved. Therefore, the proposed method can improve the power quality of the VSI that serves as the power supply of critical loads. Besides, the proposed feedforward control can be simply disabled without bringing in any influence during parallel operation.

ACKNOWLEDGEMENTS

The authors are grateful for the support provided by the National Natural Science Foundation of China, grant No. 51577078.

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