The Application of Variable Integral Gain Double Closed-Loop Control for Vehicle Inverters

Sheng Lu¹, Mingjie Liu², Hong Cui³, Long Chen⁴ and Changhao Piao⁵

^{1,2,5}*Institution of Pattern Recognition and Application, Chongqing University of Posts and Telecommunications, Chongqing 400065, China*

³*State Grid Sichuan Electric Power Research Institute, Chengdu 610072, China*

⁴*Chongqing kk-Qianwei Windpower Equipment Co., Ltd, Chongqing 400065, China*

{¹lusheng, ⁵piaoch}@cqupt.edu.cn, {²lkj_lmj, ³CuiH}@163.com,
⁴408748792@qq.com

Abstract

The variable integral gain voltage and current double closed-loop control technique based on a PWM inverter is studied in this paper. To improve the response speed and steady-state accuracy of the system, the voltage and current double closed-loop control model is established at first, and then the fixed integral of double closed-loop control strategy is analyzed based on this model. Because variable integral gain control method can improve the control precision based on the double closed loop, the voltage and current double closed-loop control strategy with variable integral gain is proposed. Finally, a series of simulation and experiment results are accomplished to verify the validity of this method. It is shown that the dynamic response time is 2ms and the maximum voltage change rate is only at 0.5%, and the steady-state accuracy can reach 99.55%.

Keywords: Variable integral gain; Full-bridge; Double closed-loop control; Electric vehicle

1. Introduction

Thanks to numerous advantages in energy conservation and environmental protection, electric vehicles (EVs) have been actively studied and developed and widely viewed as an important transition towards sustainable transportation [1,2]. In EVs, a DC-AC inverter supplies the high power electric vehicle motors torques of the propulsion system, utility loads and automotive electronics [3]. However, the cost, power density, efficiency and reliability still present major barriers to the implementation of inverter power supply equipments on a large scale [4]. Besides, the working conditions of EVs have strict requirements for inverters such as scale and voltage ratio.

It is an appropriate topology and control methods that are crucial to inverters. From the perspective of control strategy, as a branch of development of EVs power electronics system control, it has attracted increasing interests in this field [2]. Paper [5-10] present some control methods for inverter, such as deadbeat control and repetitive control, but most of them are single closed-loop control which cannot achieve good performances. To solve the problem, a study of a control scheme with voltage and current double loops control method becomes popular in the field [7,11,12]. However, these conventional methods with fixed integral coefficient always cause disturbances in external characteristics of the inverter, which is very difficult to eliminate. As a result, overshoot and static error are produced.

In this paper, a push-pull+full-bridge inverter is chosen, as shown in Figure 1. For

enhancing the system's performances, the variable integral gain voltage and current double closed-loop control method is proposed. The paper is laid out as follows. In section 2, the models of the inverter and its control scheme are established first. Then, some analyses and mathematical derivations are implemented. Also, the controller for the inverter is designed based on the proposed method. In section 3 and 4, some simulation and experiment results are accomplished to verify the validity of the proposed method respectively. At last, section 5 concludes the paper.

Figure 1. The Push-Pull + Full Bridge Inverter

2. Models of the Inverter and Control System

First, in order to design a good inverter, its model is established. Then, its characteristics will be analyzed as well. Finally, based on these, the controller will be designed using variable integral voltage and current double closed-loop control method.

2.1. Model of the Inverter

In order to facilitate the analyses, the full-bridge topology in Figure 1 is briefed, as shown in Figure 2. Clearly, the inverter is a double-input single-output system. We select output voltage U_o and capacitor current i_c as the state variables to establish the state space description and its expression is presented as (1).

$$\begin{bmatrix} \dot{i}_c \\ \dot{u}_o \end{bmatrix} = \begin{bmatrix} -\frac{r}{L} & -\frac{1}{L} \\ -\frac{1}{C} & 0 \end{bmatrix} \begin{bmatrix} i_c \\ u_o \end{bmatrix} + \begin{bmatrix} \frac{1}{L} & -\frac{r}{L} & -1 \\ 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} u_i \\ i_0 \\ i_o \end{bmatrix} \quad (1)$$

Figure 2. Single-Phase Full-Bridge Inverter

In Figure 2, U_{dc} , U_i and U_o represent the DC input voltage, output voltage and the output voltage after filtering, respectively. The parameters i_L , i_c and i_o are the currents of inductor L , capacitor C and load Z respectively. And r is an equivalent series resistance (ESR) of the filter inductor and circuit, and i_o is the output current.

From equation (1), it is easy to get the relationship between inputs and output as well as its equivalent continuous-time model. Both of them are shown in equation (2) and Figure 3, respectively.

$$U_o(s) = \frac{1}{LCs^2 + rCs + 1} U_i(s) + \frac{-(Ls + r)}{LCs^2 + rCs + 1} I_o(s) \quad (2)$$

$$= G_i(s) \cdot U_i(s) - G_{io}(s) \cdot I_o(s)$$

Figure 3. Equivalent Continuous-Time Model of the Inverter

2.2. Analyses of Inverter's Characteristics

From equation (2), it is known that the inverter is a second-order system. Its resonant frequency and damping ratio are expressed as below respectively.

$$\omega = 1/\sqrt{LC} \quad (3)$$

$$\xi = \frac{r}{2} \sqrt{\frac{C}{L}} \ll 1 \quad (4)$$

When the system is under no-load conditions, its dynamic performances will be bad. From equation (2), it is known that the output includes two parts. One is the response of the output voltage with applying input voltage only, *i.e.*, the no-load response of the inverter. Another response is the one with applying output current independently. In the second part, the coefficient $G_{io}(s)$ is treated as the system's open-loop output impedance. To describe its characteristics in detail, we set the values of inductance, capacitance and the ESR in equation (2) are 2mH, 10uF and 0.1Ω respectively. In the open-loop mode, the Bode plot of output impedance is shown in Figure 4. From this picture, it is known that the output impedance increases with frequency in low band. When there exists load current, current i_o will form a voltage drop. The bigger load is the more drastic voltage drop will be. This is the reason why the output voltage always be reduced under an open-loop circumstance, also it explains that why the output voltage changes when the load has step variations.

Due to dead zone of switching tubes and over modulation, the output voltage waveform will be distorted in the open-loop mode. And when the inverter connects with big nonlinear loads, the output voltage waveform will be distorted too. Generally, nonlinear loads always have rectifying and filtering installations, when the instantaneous value of output voltage is bigger than that of filter capacitor, the generated current will go across rectifier diode. As a result, the inverter will switch its states between load and no-load conditions. Shown on the waveform, the output voltage will be reduced.

Figure 4. The Bode Plot of Output Impedence in Open-Loop Mode

2.3. Voltage and Current Double Closed Loop

As mentioned above, due to output impedance, over modulation and nonlinear load, the output voltage waveforms will be distorted and have bad dynamic performances. So it is necessary to design a suitable controller to improve them. At present, there are several control methods such as voltage instantaneous value feedback, deadbeat control and repetitive control [5-7]. However, majority of them are single closed-loop controls which can improve output performances in some degree, but they have a weak anti-load-disturbance ability. In some researches, it is said that double closed-loop and multi closed-loop controls can enhance the system's performances [6,7]. As a double closed-loop control, the external voltage loop and instant voltage internal loop control [13] based on output voltage cannot solve the problems which caused by load disturbances. By contrast, the inverter uses external voltage loop and inner current loop control [14] achieves better performances, because the inner current loop improves the bandwidth to make the dynamic responses faster.

As to voltage and current double closed-loop control, the inner current loop has two types, that is, inductor current loop and capacitor current loop [15]. Compared with capacitor current, inductor current is more stable. Nevertheless, the output current i_o treated as a disturbance which locates at the outside of inductor loop, as shown in Figure 5, cannot restrain load disturbances especially the nonlinear load disturbances [12]. To solve the problem, a feedforward is introduced, as shown in Figure 6.

Figure 5. The Block Diagram of Double Closed-Loop Control System

Figure 6. The Block Diagram of Double Closed-Loop Control System With A Feedforward

2.4. Design of the Controller and Analyses

Form Figure 6, it is easy to get equation (5) to (8).

$$U_v(s) = (U_r(s) - U_o(s)) \square G_v(s) \quad (5)$$

$$U_i(s) = (U_v(s) + \alpha \square I_o(s)) \square G_i(s) \quad (6)$$

$$U_i(s) = U_o(s) + (sL + r)I_L(s) \quad (7)$$

$$I_L(s) = I_o(s) + sCU_o(s) \quad (8)$$

Where coefficients $G_v(s)$ and $G_i(s)$ represent the voltage external loop and current inner loop regulators respectively.

From Figure 6, it is clearly see that the voltage error is obtained by comparing the given voltage with the output voltage firstly. And then, the given inner current u_i is generated by $G_v(s)$. Finally, along with the flow path, u_i is compared with feedback capacitor current i_c to generate the current error, which will be used to generate the

control signal u_1 by $G_i(s)$ to control the inverter. In the double closed-loop control method, as to inner current loop, the output voltage is a disturbance. The coefficient α is smaller than 1, i.e., $\alpha < 1$.

In the double closed-loop control method, we design the external voltage loop as a PI controller by which the output voltage can follow the given voltage, and the inner current loop is designed as a PI controller too, which can increase the damping ratio to make the system more stable. It is assumed that $G_v(s)$ and $G_i(s)$ are expressed as below.

$$G_v(s) = K_{1p} + \frac{K_{1i}}{s} \quad (9)$$

$$G_i(s) = K_{2p} + \frac{K_{2i}}{s} \quad (10)$$

Where K_{1p} and K_{1i} represent the coefficients of proportional and integral parts of $G_v(s)$, respectively. And K_{2p} and K_{2i} are those of $G_i(s)$ respectively.

Substitution of equation (9) and (10) into equation (5) to (8), we get the system's transfer function (11) and its characteristic equation (12),

$$U_o(s) = \frac{\frac{k_{1p}k_{2p}}{LC}s^2 + \frac{k_{1p}k_{2i} + k_{1i}k_{2p}}{LC}s + \frac{k_{1i}k_{2i}}{LC}}{D(s)} U_r(s) + \frac{[\frac{1}{C}s^2 + \frac{r + k_{2p}(1-\alpha)}{LC}s + \frac{k_{2i}(1-\alpha)}{LC}]}{D(s)} I_o(s) \quad (11)$$

$$= G_{ur}(s)U_r(s) + G_{io}(s)I_o(s) \quad (12)$$

$$D(s) = s^4 + \frac{(r + k_{2p})}{L}s^3 + \frac{k_{1p}k_{2p} + Ck_{2i} + 1}{LC}s^2 + \frac{k_{1p}k_{2i} + k_{2p}k_{1i}}{LC}s + \frac{k_{1i}k_{2i}}{LC}$$

Where the coefficient $G_{ur}(s)$ presents the traceability of the output voltage aiming to the given signal, and its poles can show the dynamic characteristics of the traceability. The coefficient $G_{io}(s)$ can be treated as the output impedance.

From equation (11) and (12), it is clearly known that the denominators of $G_{ur}(s)U_r(s)$ and $G_{io}(s)I_o(s)$ are the same. The feedforward coefficient α has an influence on the zeroes of $G_{io}(s)$ except the poles in closed-loop condition. Therefore, in voltage and current double closed-loop control, the characteristics equation of the double closed-loop control with load current feedforward is the same as that of the double closed-loop control with capacitor current. From equation (12), it is known that the system is forth-order. We know that the performances of the system are determined by predominant poles. In this paper, we set the expected predominant poles and the others as equation (13) and (14) respectively. Based on these, the corresponding characteristics equation (15) is obtained.

$$s_{1,2} = -\xi\omega_n \pm j\omega_n\sqrt{1-\xi^2} \quad (13)$$

$$s_3 = -m\xi\omega_n, s_4 = -n\xi\omega_n \quad (14)$$

$$D(s) = (s^2 + 2\xi\omega_n s + \omega_n^2) \cdot (s + n\xi\omega_n) \cdot (s + m\xi\omega_n) \quad (15)$$

Rearranging equation (12) and (15), we get equation (16). In equation (17), coefficient K_{2i} has three solutions which include two complex roots and a real one. Substituting the real one into (17), we get equation (18).

$$k_{2p} = \frac{a_3}{C} - r$$

$$k_{1p}k_{2p} + Ck_{2i} = a_2 - 1 \quad (16)$$

$$k_{1p}k_{2i} + k_{2p}k_{1i} = a_1$$

$$k_{1i}k_{2i} = a_0$$

$$Ck_{2i}^3 + (1-a_2)k_{2i}^2 + a_1k_{2p}k_{2i} - k_{2p}^2a_0 = 0 \quad (17)$$

$$k_{1p} = \frac{a_2 - Ck_{2i} - 1}{k_{2p}} \quad (18)$$

$$k_{1i} = \frac{a_0}{k_{2i}}$$

Where

$$\begin{aligned} a_0 &= LCmn\xi^2\omega_n^4 \\ a_1 &= LC(m+n+2mn\xi^2)\xi\omega_n^3 \\ a_2 &= LC[1+(2m+2n+mn)\xi^2]\omega_n^2 \\ a_3 &= LC(2+m+n)\xi\omega_n \end{aligned} \quad (19)$$

In this paper, we selected the coefficients $\xi=0.8$, $\omega_n=2500$, $m=n=10$, $L=0.43\text{mH}$ and $C=140\mu\text{F}$. Substitution of these into equation (16) to (18), we got the parameters of the controller, as expressed in equation (20). From Figure 4, it is easy to get the open-loop and closed-loop transfer functions of the inner current loop, as shown in equation (21) and (22) respectively. And their Bode plots are shown in Figure 7 and 8 respectively.

$$\begin{cases} k_{1p} = 18.82 \\ k_{1i} = 156170 \\ k_{2p} = 0.597 \\ k_{2i} = 963.70 \end{cases} \quad (20)$$

$$G_{iop}(s) = \frac{C(k_{2p}s + k_{2i})}{LCs^2 + rCs + 1} \quad (21)$$

$$G_{icl}(s) = \frac{C(k_{2p}s + k_{2i})}{LCs^2 + (r+k_{2p})Cs + Ck_{2i} + 1} \quad (22)$$

Figure 7. The Open-Loop Bode Plot of Current Loop

Figure 8. The Closed-Loop Bode Plot of Current Loop

From Figure 7 and 8, it is known that the inner current controller makes the inner-loop phase margin big enough. It is about 75° which means the open inner loop has a better stability. Compared with the open loop, the cross-over frequency of closed loop is bigger, which increases the bandwidth and make the system response faster. Consequently, the disturbance resisting ability has been enhanced.

Similarly, the transfer functions of open loop and closed loop of the external voltage loop are obtained and are expressed as equation (23) and (24) respectively. And their Bode plots are shown in Figure 9 and 10, respectively. From Figure 9 and 10, it is known that the external voltage loop has enough phase and gain margins, it means that the loop has a good stability.

$$G_{uop}(s) = \frac{(k_{1p}s + k_{1i})(k_{2p}s + k_{2i})}{s^2[LCs^2 + (r + k_{2p})Cs + Ck_{2i} + 1]} \quad (23)$$

$$G_{uc}(s) = \frac{(k_{1p}s + k_{1i})(k_{2p}s + k_{2i})}{LCS^4 + (r + k_{2p})Cs^3 + (k_{1p}k_{2p} + Ck_{2i} + 1)s^2 + (k_{1p}k_{2i} + k_{2p}k_{1i})s + k_{1i}k_{2i}} \quad (24)$$

Figure 9. The Open-Loop Bode Plot of Voltage Loop

Figure 10. The Closed-Loop Bode Plot of Voltage Loop

The output impedance of the closed loop is expressed as equation (25). The Bode plot is shown in Figure 11. And it shows that the output impedance is always smaller than 1, which means that the system has a good disturbance resisting ability.

$$G_{io}(s) = \frac{(Ls^2 + (r + k_{2p}(1-\alpha))s + k_{2i}(1-\alpha))s}{LCS^4 + (r + k_{2p})Cs^3 + (k_{1p}k_{2p} + Ck_{2i} + 1)s^2 + (k_{1p}k_{2i} + k_{2p}k_{1i})s + k_{1i}k_{2i}} \quad (25)$$

Figure 11. The Bode Plot of Output Impedence

As to fundamental current disturbance, the output impedance is expressed as equation (26). Where $f=50\text{Hz}$ and $\omega=2\pi f$. By rearranging (26), we get equation (27). When there exists nonlinear load disturbance, the load current will contain a large number of harmonic components. When the current harmonic is a fifth harmonic, the output impedences is expressed as equation (28) and (29).

$$20\lg|G_{io}(j\omega)|_{\omega=314\text{rad/s}} = -43 \quad (26)$$

$$|Z_o| = |G_{io}(j\omega)| = 0.007\Omega \quad (27)$$

$$20\lg|G_{io}(j\omega)|_{\omega=1570\text{rad/s}} = -18.4 \quad (28)$$

$$|Z_o| = |G_{io}(j\omega)| = 0.12\Omega \quad (29)$$

From what described above, the output impedance of the closed loop is just 0.007Ω under the circumstance of fundamental frequency, which means that the closed loop has a good disturbance resisting ability.

2.5. Variable Integral Double Closed-Loop Control

In this paper, the double closed-loop control is achieved based on digital control. So the continuous double closed-loop control has to be discretized. We use accumulation and first-order difference equation to replace the integral and differential equation respectively, and the proportional equation remains unchanged. Then we get equation (30) to (32). The control parameters are defined as well, as expressed in equation (33) and (34).

$$K_p \frac{1}{T_i} \int_0^t e(t)dt \cong K_p \frac{T}{T_i} \sum_{j=0}^k e(j) \quad (30)$$

$$K_p T_D \frac{de(t)}{dt} \cong K_p \frac{T_D}{T} [e(k) - e(k-1)] \quad (31)$$

$$K_p e(t) = K_p e(k) \quad (32)$$

$$K_i = K_p \frac{T}{T_i} \quad (33)$$

$$K_d = K_p \frac{T_D}{T} \quad (34)$$

Where $t=kT$, and coefficient k is the discrete-time series, $k=1, 2, 3, \dots$. T_i and T_D represent the sampling periods of the integral and differential equations respectively.

Based on what was mentioned above, the PID control law in digital system is obtained and is shown in equation (35).

$$u(k) = K_p e(k) + K_i \sum_{j=0}^k e(j) + K_D [e(k) - e(k-1)] \quad (35)$$

Xu [16] and Chen [17] indicated that the integral coefficients of the voltage external loop are the main factor which has an influence on the system's characteristics. In conventional double closed-loop control method, the integral coefficient K_i of external loop is always a constant so that the increase of integral remains unchanged during the whole control process. When the coefficient is too big, it will cause not only overshoot but also integral saturation. And when it is too small, it will take long time to eliminate the static error. In practice, we hope that when the error is big the function of integral can be slender or negligible, and when the error is small the function can be enhanced. Thus it can be seen, how to set a suitable coefficient to change integral gain based on the errors is vital to the enhance the system's performances.

The variable integral gain control method is a good way to solve the problem. The main idea of this method is to make the integral gain change with the error, that is, the bigger error is the slower integrating speed will be, otherwise, it speeds up [18]. In the paper, we set a factor $f[e(k)]$ which is a function of system error $e(k)$. When $e(k)$ is increasing $f[e(k)]$ will decrease, and when it is decreasing $f[e(k)]$ will increase. The equations of inner current loop and the external loop control are established. They are shown as (36) and (37) respectively.

$$u_i(k) = k_{ip} e(k) + k_{ii} \left\{ \sum_{i=0}^{k-1} e(i) + f[e(k)] e(k) \right\} T \quad (36)$$

$$u_v(k) = k_{vp} e(k) + k_{vi} \left\{ \sum_{i=0}^{k-1} e(i) + f[e(k)] e(k) \right\} T \quad (37)$$

where k_{ip} and k_{ii} represent the coefficients of proportional and integral terms of inner current loop control, respectively. Parameters k_{vp} and k_{vi} represent the coefficients of proportional and integral terms of external voltage loop control respectively. And k is the discrete-time series and $k=0, 1, 2, \dots$ T is the sampling time. The relation between $f[e(k)]$ and the present error $|e(k)|$ is linear or nonlinear, and we set $f[e(k)]$ as,

$$f[e(k)] = \begin{cases} 1 & |e(k)| < B \\ \frac{A - |e(k)| + B}{A} & B \leq |e(k)| \leq A + B \\ 0 & |e(k)| > A + B \end{cases} \quad (38)$$

Where A and B are the boundaries of $|e(k)|$. Fortunately, A and B are not required strictly, so it is easy to get them. When the error of external voltage loop is bigger than $A+B$, it means that the system has entered the saturation region. Under this circumstance, the integral will not work so that the system will be controlled by proportional control, and the response speed of the inverter will be faster. As to the inverter, it seems that the system is under a load step-change condition. When the error is smaller than $A+B$ but is bigger than B , the integral will be enhanced with the decreased error and the response will be smoother. This result is just like that under the condition of load disturbance changed from strong to weak. And when the error is smaller than B , there has a maximum integral and the static error will be eliminated quickly. Under this condition, it means that the system is in steady state.

In the paper, we first set the initial values of A and B as $0.3/e(k)_{\max}$ and $0.2/e(k)_{\max}$, respectively. Of course, they will be amended according to system's responses. For reducing the too frequent control to the inverter and eliminating the output voltage oscillation and decreasing THD, a wider dead zone is introduced and the control expression is showed as (39),

$$e(k) = \begin{cases} 0 & |e(k)| \leq |e_0| \\ e(k) & |e(k)| > |e_0| \end{cases} \quad (39)$$

where $e(k)$ is the error and e_0 is an adjustable parameter whose value will be selected based on the real situation. In equation (39), if e_0 is too small, it will cause too frequent

control and unstable output voltage and even a large number of harmonic waves; and if e_0 is too big, the output voltage will have some lags. So, the selection of e_0 should be careful.

3. Simulation Results

According to the system's parameters, based on the fixed integral gain double closed-loop control and variable integral gain double closed-loop control, the simulations are implemented. Because the update of waveform is lag to the sampling time, we call it time delay computation. When it comes to 50Hz output condition, the delay is negligible.

From Figure 12, it is clearly known that the output voltage variation of the inverter is more than 30% when the load occurs step changes. By contrast, the variation of output voltage in Figure 13 is less than 20% under the same condition, and the output voltage returns to normal value quickly.

Figure 12. The Output Voltage and Current Waveforms with Fixed Integral Gain Double Closed-Loop Control for Adding A 2kw Impedance Load

Figure 13. The Output Voltage and Current Waveforms with Variable Integral Gain Double Closed-Loop Control for Adding A 2kw Impedance Load

Figure 14 and 15 show the changes of the root-mean-square (RMS) value of output voltage with a 2kW impedance load. From Figure 14, it is easy to understand that the maximum fluctuation of the output voltage is 6V and the change is at 2.73% while those in Figure 15 are 3.7V and 1.68% respectively.

Figure 14. The Output Voltage RMS Value with Fixed Integral Gain Double Closed-Loop Control

Figure 15. The Output Voltage RMS Value with Variable Integral Gain Double Closed-Loop Control

Figure 16 and 17 show the harmonic contents of the output voltage with fixed integral gain double closed-loop control and variable integral gain double closed-loop control. It is clearly see that the THDs are at 1.04% and 0.83%, respectively.

Figure 16. The Output Voltage THD with Fixed Integral Gain Double Closed-Loop Control Under a 2kw Resistive Load Condition

Figure 17. The Output Voltage THD with Variable Integral Gain Double Closed-Loop Control Under a 2kw Resistive Load

In Figure 18 and 19, it is clearly known that the voltage and current waveforms with these two control methods are different. The output waveforms of the system controlled by variable integral gain double closed-loop method, which have few spikes and steady state errors, are superior to those with another method.

Figure 18. The Output Voltage and Current Waveform with Fixed Integral Gain Double Closed-Loop Control Under a 2kw Rectifier Load Condition

Figure 19. The Output Voltage and Current Waveform with Variable Integral Gain Double Closed-Loop Control Under a 2kw Rectifier Load Condition

Figure 20 and 21 show the harmonic contents of the system controlled by these two methods, and the THDs are at 1.35% and 1.17%, respectively.

Figure 20. The Output Voltage THD with Fixed Integral Gain Double Closed-Loop Control Under a 2kw Rectifier Load Condition

Figure 21. The Output Voltage THD with Variable Integral Gain Double Closed-Loop Control Under a 2kw Rectifier Load Condition

4. Experiment Results

In this section, some experiments are performed on a prototype. And the specifications of the inverter are showed in Table 1.

Table 1. Specifications of the Inverter

Description	Parameter	Nominal Value
Output voltage	U_{rms}	220V
Voltage frequency	F	50Hz
Output power	P_o	1kW
Output inductor	L	0.33mH
Output capacitor	C	110uF
ESR	r	0.1 Ω

The experimental waveforms are showed in Figure 22 to 24. The output voltage RSM value of the inverter is 220V under a no-load condition. Figure 22 shows us the dynamic responses of the voltage and current of the inverter when the impedance load surges by 400W. It is obvious to see that the RMS value of output voltage is 219.3V and the steady state error is at 0.32%. Also, when there has a step variation, we can clearly understand that the voltage drop is less than 1.0V and the setting time is just 2ms. Above all, it is self-evident that this control method has some merits such as short setting time, low voltage change rate, high steady accuracy and good waveforms and low deformation, even if the inverter is under a condition of impedance load change.

Figure 22. The Output Voltage and Current For Adding a 400W Impedance Load

The waveforms of the voltage and current of the inverter with a 500VA rectifier load are showed in Figure 23. Obviously, the output voltage RMS value is 218.9V and the steady accuracy is at 99.5%. At the same time, the output current RMS value is 2.777A and the current peak is 12A and the crest factor is 4.321. Under this condition, the voltage deformation rate is at 1.33%. From what has been described, it is too can say that this control method has good performances even though the inverter is under a nonlinear load variation. The advantages of high steady accuracy, fast dynamic response and low voltage deformation are outstanding.

Figure 23. The Output Voltage and Current Waveforms of the System with 500VA Rectifier Load

Figure 24 shows us some details of the output voltage and current under a full-load condition. It is easy to see that the waveforms approach the perfect sine wave and the voltage RMS value is 219.0V and the steady accuracy is at 99.55%. It is no exaggeration to say that the inverter has good performances, including high steady accuracy and strong load ability even if it is under a full-load condition

Figure 24. The Output Voltage and Current Waveforms with A Full Load

Comparing with the simulation and experiment results, it is known that the results showed in Figure 12 to 17 are accorded with those in Figure 22. Also, the simulation results in Figure 18 to 21 agree with those in Figure 23 approximately when the inverter is under a condition of rectifier load variation. However, the crest factor in experiments is 4.321, which is bigger than that in simulations. This circumstance has relations with the output capacitance and the accuracy of the sampling time of capacitive current. Above all, although there exists some minor flaws, the inverter controlled by the proposed method still has some good performances under both resistive and rectifier load conditions. The setting time, maximum voltage change rate and steady accuracy are at 2ms, 0.5% and 99.55% respectively. No doubt, the proposed control method is feasible.

5. Conclusions

In this paper, the models of the inverter and its control scheme are established at first. Some analyses and mathematical derivations are implemented as well. And then, the parameters of the voltage external loop and current inner loop of the inverter are designed based on pole assignment method. To solve the problems caused by fixed integral gain voltage and current double closed-loop control, the variable integral control method for voltage external loop is proposed. At last, a series of simulations and experiments are accomplished by which the feasibility of the proposed method is verified under conditions of resistive and rectifier loads. In a word, the inverter using the variable integral gain voltage and current double closed-loop control method has both good dynamic and steady performances and the proposed control method is validated and feasible.

Acknowledgments

This work is supported by the Natural Science Foundation of CQ CSTC of P. R. China (CSTC2013yykfC60005, CSTC2014jcyjA60004, CSTC2013jcsf-jcssX0022).

References

- [1] X. Hu, S. Li H, Peng and F. Sun, "Changing time and loss optimization for LiNMC and LiFePO₄ batteries based on equivalent Circuit models", *Journal of Power Sources*, vol. 239, pp. 449–457, (2013).
- [2] C. Piao, C. Jiang, H. Qiao, C. Cho and S. Lu, "Study on the Control Algorithm of Two-Stage DC-DC Converter for Electric Vehicles", *Advances in Mechanical Engineering*, (2014).
- [3] B. Allaoua and A. Laoufi, "Application of a robust fuzzy sliding mode controller synthesis on a buck-boost DC-DC converter power supply for an electric vehicle propulsion system", *Journal of Electrical Engineering and Technology*, vol. 6, no. 1, (2011), pp. 67–75.
- [4] Q. Zhang, D. Zhang, H. Hu, J. Shen and I. Batarseh, "Controlled-Type ZVS Technique without Auxiliary Components for Micro-inverters", *Journal of Power Electronics*, vol. 12, no. 6, (2013), pp. 919-927.
- [5] W. Guo, S. Duan and X. Kong, "A Modified Deadbeat Control for Single-phase Voltage-source PWM Inverters based on Asymmetric Regular Sample", *IEEE-PESC*, vol. 19, no. 2, (2001), pp. 962-967.
- [6] C. Piao, C. Jiang, H. Qiao, C. Cho and S. Lu, "Modeling and Implementation of Fixed Switching Frequency Sliding Mode Control to Two-Stage DC-DC Converter", *International Journal of Control Automation and Systems*, vol. 12, no. 6, (2014), pp. 1225-1233.
- [7] P. C. Loh, M. J. Newman and D. N. Zmood, "A Comparative Analysis of Multi-loop Voltage Regulation Strategies for Single and Three-phase UPS System", *IEEE Trans. On Power Electronics*, vol. 18, no. 9, (2003), pp. 1176-1185.
- [8] X. Jin, F. Ichihashi, A. Mano, N. Yamamoto, M. Suzuki, T. Yasue, T. Koshikawan and Y. Takeda, "High Quantum Efficiency and High Spin-polarization Transmission-type Photocathode", *International Journal of Future Engineering*, vol. 2015, ID.3315108, (2015), pp. 1-5.
- [9] Z. Salam and A. Jusoh, "Bidirectional high-frequency link inverter with regenerative snubber and deadbeat control", *International Journal of Electronics*, vol. 98, no. 3, (2011), pp. 339-355.
- [10] T. Hrník and Q. C. Zhong, "A Current-Control Strategy for Voltage-Source Inverters in Microgrids Based on H-infinity and Repetitive Control", *IEEE Trans. on Power Electronics*, vol. 26, no. 3, (2011), pp. 943-952.
- [11] S. Buso, S. Fasolo and P. Mattavelli, "Uninterruptible Power Supply Multiloop Control Employing Digital Predictive Voltage and Current Regulator", *IEEE Trans. on Power Electronics*, vol. 37, no. 11,

- (2001), pp. 1846-1854.
- [12] Q. Lei, F. Z. Peng and S. Yang, "Multiloop control method for high-performance microgrid inverter through load voltage and current decoupling with only output voltage feedback", IEEE Trans. on Power Electronics, vol. 26, no. 3, (2001), pp. 953-960.
- [13] H. Wu, D. Lin, D. Zhang, K. Yao and J. Zhang, "A Current-Mode Control Technique with Instantaneous Inductor-Current Feedback for UPS Inverters", IEEE-APEC'9, (1999)9, pp. 951-957.
- [14] K.H. Jung, Y.H. Kim, S. Lim, B. Ahn and G.A. Lee, "Effect of friction on critical damage value in compressive deformation", International Journal of Future Engineering, vol.2015, ID.3315112,(2015), pp.1-5.
- [15] S. Xu, "Design of Photovoltaic High Frequency Link Inverter Based on Push-pull Forward Converter. Communications, Circuits and Systems (ICCCAS)", 2010 International Conference. Chengdu, (2010), pp. 593 ~596.
- [16] A. Xu and S. Xie, "External Characteristics of Digital Control Inverter with Instantaneous Dual-Loop Feedback", Journal of Nanjing University of Aeronautics & Astronautics, vol. 38, no. 4, (2006), pp. 513-518.
- [17] L.Chen, L. Xiao and G. Yan, "Research on Output Characteristics of Double Close-loop Controlled Voltage Source Inverter", Power Electronics, vol. 38, no. 4, (2004), pp. 6-7.
- [18] J. Wu, H. Wu, J. An, Y. Bai, H. Li and M. Xuan, "Implementation of Variable Structure and Speed-changed Integration Control for Flywheels", Optics and Precision Engineering, vol. 18, no. 1, (2010), pp. 149-155.

Authors

Sheng Lu, He received his B.S. from Xi'an Jiaotong University in 2004, China. And in 2009, he received his M.S. and D.S. from Inha University, Republic of Korea. Now he is an associate professor of Chongqing University of Posts and Telecommunications, China. His interests include automobile electronics, power electronics, micro-nano mechanics, and intelligent materials.