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A Modified One-Cycle-Control Method for Modular Multilevel Converters

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Abstract: In this paper, a new One-Cycle-Control (OCC) method is designed for a modular multilevel converter (MMC) based on the principle of the equivalent resistance constant. The proposed controller has a simple structure and a small amount of calculation by cancelling the current inner loop proportional integral (PI) controller and the inverse transform in the traditional direct-quadrature (DQ) control. Compared to the traditional OCC controller, the new one separates the control method from the modulation strategy, making it possible to use not only carrier-based pulse-width modulation (PWM), but also nearest level modulation PWM to generate drive signals. Besides, the independent control of the active and the reactive power is implemented by injecting a reference current with the same phase of the supply voltage or a reference current which lags the supply voltage by $\pi/2$ into the controller, so the converter can operate in four quadrants and it can work in either a grid-connect or off-grid environment. The feasibility and the performance of the proposed OCC method have been validated by both the simulation under the MATLAB/SIMULINK (R2012a) environment and experimental results.

Keywords: modular multilevel converter; One-Cycle-Control; four-quadrant operation; carrier-based PWM; nearest level modulation PWM

1. Introduction

Modular multilevel converters (MMCs) have attracted wide attention from both industry and academia due to their advantages of full modularity, better scalability, high redundancy, and low switch voltage stress [1,2]. Consisting of a series connection of submodules (SMs), MMCs are suitable for large-voltage, high-power applications like high-voltage direct current (HVDC) transmission [3], medium-voltage high-power motor drives [4], and photovoltaic generation [5].

There are a great number of control methods which have been used in MMCs successfully since the circuit topology was proposed. Based on linear system theory, the proportional integral (PI) controller has been widely used in power electronic devices [6]. Since the PI controller cannot track the sinusoidal signal without phase error, the abc-to-dq-frame transformation is used to let the controlled variables change from alternating current (AC) to direct current (DC), and the controller output also needs coordinate transformation under this controller structure. To avoid the complicated calculations, the proportional resonant (PR) controller has also been used [7]. By replacing the integral with resonance, the PR controller can provide an infinite gain at the resonant frequency, which makes the PR controller able to track a sinusoidal signal without steady-state error. However, since the gain of the controller is rapidly reduced when the frequency is no longer equal to the resonant frequency, the resonant frequency of the PR controller must be accurately calculated. Except for the traditional control strategy, there are some new types of controllers that have been used in MMCs. Sliding

mode control (SMC) divides the state space by referencing controlled variables, and in each subspace the controlled variables are forced to slide along the boundaries by the different control structure. This control method can get good dynamic response when applied to MMCs [8]. Finite control set model predictive control (FCS-MPC) analyzes all possible operating states based on the circuit model and chooses the best one based on the cost function. The advantage of the FCS-MPC is that it can control multiple targets [9] and the disadvantage of it is the accuracy of the model plays an important role in the control effect. Repetitive control takes the previous behavior of the periodic controlled variables into consideration, obtaining a better steady-state response [10], but in order to optimize the dynamic response, it often needs to be used together with other control strategies.

One-Cycle-Control (OCC) was proposed by Keyue M. Smedley in 1991 [11]. The basic principle is to let the average controlled variable value over a switch cycle be equal to the reference. Traditional OCC lets the controlled current run in the same phase with the input voltage, therefore, it has been widely used in power factor correction (PFC) [12] and active power filters (APF) [13]. However, the traditional OCC cannot control the reactive component so that the converter can only implement the unit power factor with no ability to output leading or lagging reactive power. This problem was discussed in References [14,15]; they let the APF compensate for only the harmonic components of load currents by injecting the reactive currents. Based on the same principle, OCC working on the grid-connected inversion mode was discussed in References [16,17]. The OCC-based MMC controller has been discussed in Reference [18], but it can only work in the unity power factor rectification mode and the control method must be used in conjunction with the carrier disposition PWM, which reduces the flexibility of the system.

In this paper, a modified OCC for an MMC is proposed. As shown in Table 1, an OCC controller has less complexity and calculation demand than PI and PR controllers. Compared to other nonlinear control methods such as FCS-MPC, OCC does not rely on an accurate mathematical model and it can realize the constant switch frequency naturally. Compared with the existing OCC-based MMC controller, the modified OCC proposed in this paper can enable the bi-directional power flow, and the power factor can also be controlled, which expands the application scope of the control method. In addition, the modified OCC realizes the decoupling of the control strategy and the modulation strategy. It can work with several modulation strategies and is compatible with existing SM voltage balancing strategies.

Table 1. Comparison between several control strategies for a modular multilevel converter (MMC). PI: proportional integral; PR: proportional resonant; FCS-MPC: finite control set model predictive control; OCC: One-Cycle-Control.

Control Strategies	Phase Locked Loop (PLL)	Controller Structure	Calculation	Robustness	Modulation Strategy	Steady-State Error
PI	Yes	Common	Middle	Strong	Unlimited	No
PR	Yes	Common	Small	Less	Unlimited	No
FCS-MPC	Yes	Complex	Large	Less	Unlimited	No
OCC [18]	No	Simple	Small	Strong	Limited	Yes
Modified OCC	Yes	Simple	Small	Strong	Unlimited	No

The rest of this paper is organized as follows: the mathematical model of MMCs is briefly introduced in Section 2. The proposed OCC has been described in Section 3. In Section 4, the method that makes the OCC-controlled MMC operate in four-quadrants is described in detail. Section 5 covers the simulation and experimental results to validate the correctness and effectiveness of the proposed modified OCC, and finally, Section 6 summarizes the main points of the paper.

2. Mathematical Model of the MMC

The topology of the single-phase MMC is shown in Figure 1, in which i_{AC} , u_{AC} are the AC side current and voltage, i_p and i_q are the upper arm current and the lower arm current, respectively.

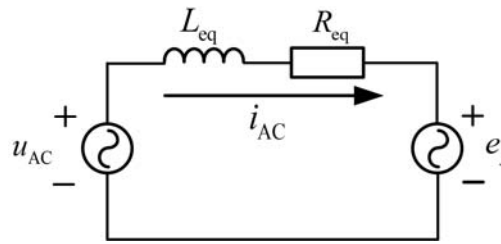


Figure 2. Alternating current (AC) side equivalent model of the MMC.

3. OCC for Multiple Modulation Strategies

The conventional OCC has been widely used in the field of unit power factor rectification [19]. The original OCC uses the input current directly compared with the carrier wave to generate the drive signal, and since there is no reference voltage generation during this process, the controller can only use the carrier phase-shift PWM modulation strategy and this limits the design of the SM voltage balancing strategy.

In this paper, the OCC is analyzed based on the equivalent resistance constant so that the AC side output voltage of the converter can be calculated and most of the existing modulation strategies can be used directly.

The basic principle of the OCC is to let the average controlled variable value over a switch cycle be equal to the reference. For the equivalent circuit shown in Figure 2, in order to make it operate in the unity power factor state, as Figure 3 shows, the MMC can be equivalent to a resistive load R_e after ignoring the equivalent parasitic resistance R_{eq} and the equivalent inductance L_{eq} .

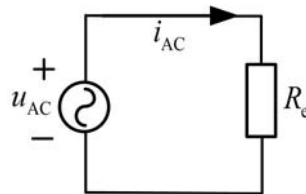


Figure 3. Equivalent circuit in unit power factor rectification mode.

In order to let the behavior of the MMC in a switch cycle T be equivalent to an emulated resistance R_e , the control equation of the conventional OCC can be expressed as follows:

$$\frac{1}{T} \int_0^T \frac{e_j}{i_{AC}} dt = R_e. \quad (7)$$

Due to the high switching frequency of the MMC, the change of the current i_{AC} in a switch cycle can be ignored, so the control law of the conventional OCC can be expressed as:

$$e_j = R_e i_{AC}. \quad (8)$$

Once the e_j is determined, the arm voltages, u_p and u_n can be determined based on Equation (6). The inserted SM numbers of the upper and lower arms are also determined. However, which SM should be inserted has not been determined. In addition to making the AC side current change as expected, another major control target is to keep the SM capacitor voltages in balance. That can be implemented by several modulation strategies [20–23], which balances the SM capacitor voltage by controlling the insertion and bypass time of each SM.

One of the most commonly used modulation strategies is nearest level modulation (NLM) [20,21]. As Figure 4 shows, the basic principle of NLM is to determine the number of inserted SMs according to the required bridge arm voltage and the SM voltage to minimize the error between the actual bridge arm voltage and the reference voltage. The algorithm sorts the SMs based on the voltage of the SM

and insert the SMs from large to small, or from small to large, according to the direction of current to balance the SM capacitor voltages.

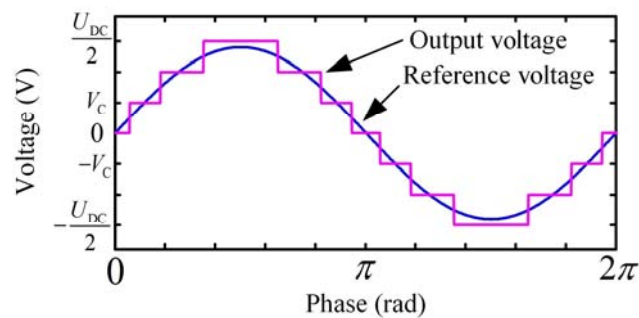


Figure 4. Principle of the nearest level modulation (NLM) strategy.

Another popular modulation strategy for the MMC is carrier phase-shift PWM (CPS-PWM) [22]. The basic principle of CPS-PWM is shown in Figure 5. The switching signals are generated by comparing the reference voltage with several carrier waves. For an $N+1$ level converter, N carrier waves are needed and each one has a phase shift of $\frac{2\pi}{N}$. The outputs are the control signals for the upper arm SMs, and the control signals for the lower arm SMs can be generated by inverting the upper ones.

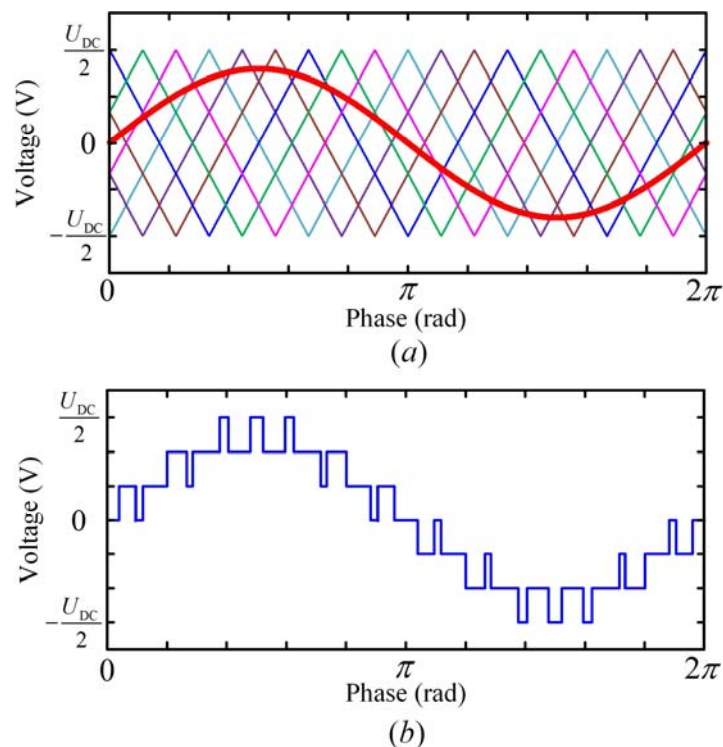


Figure 5. Principle of carrier phase-shift (CPS)-PWM. (a) Reference voltage and carrier waves; (b) output voltage.

Both the modulation strategies have their own advantages and disadvantages [23]. Compared to NLM, CPS-PWM can generate more accurate voltage, and because the switching devices under the CPS-PWM strategy have nearly the same turn-on time, the loss of the electronic components are basically identical. However, because the traditional CPS-PWM cannot balance the SM capacitor voltages, additional controllers are often required for each SM to balance the SM capacitor voltages in actual applications. The additional controllers increase the calculations and this will put higher

requirements for the performance of the microprocessor. Therefore, CPS-PWM is suitable for medium- and low-voltage applications needing fewer SMs such as a variable-frequency drive, while for the high-voltage applications such as HVDC, NLM is more suitable because in these occasions the system often requires a large number of SMs. For the OCC described in this paper, since the controller calculates the AC side output voltage e_j , compared to traditional OCC, the modulation strategy is no longer limited, and both NLM and CPS-PWM can be used.

The basic form of the proposed OCC controller is shown in Figure 6. An outer DC-side voltage control loop implemented by a PI controller is designed to maintain the DC bus voltage u_{DC} tracking the reference value u_{DC}^* . In order to make the controlled variable positively related to the controller output, the PI controller output is set as the admittance of equivalent resistance R_e and it determines the input active power. Then, dividing the AC side input current i_{AC} by the admittance, which is equivalent to multiplying R_e and i_{AC} , the AC side output voltage of the MMC e_j can be derived. Finally, the switching signals are generated by the modulation strategy.

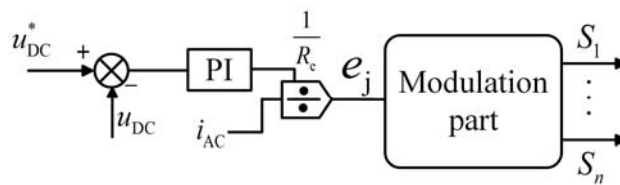


Figure 6. OCC algorithm for the unit power factor rectifier.

Compared to other control strategies, the OCC has a simpler control structure and less calculation. This method can be implemented without a PLL or even an AC side voltage sensor. However, this control strategy can only implement the unity power factor rectifier and the power factor cannot be set. In addition, because the influence of the inductance is neglected in the theoretical analysis, there is an error in the power factor, so it is only suitable for some limited applications, such as the active front-end (AFE) rectifier, which allows for a certain power factor error.

4. Four-Quadrant Operation for OCC Controlled MMC

4.1. Reactive Power Control

For the rectifier with the reactive power output, it can also be equivalent using the method described in the previous section. Similar to Figure 3, the MMC rectifier with reactive power output can be equivalent to a resistor in parallel with a capacitor or an inductor, which is determined by the power factor angle. Figure 7 shows the equivalent circuit of a rectifier with a lagging power factor, where R_e is the equivalent resistance and L_e is the equivalent inductance. $i_{AC,d}$ and $i_{AC,q}$ are the currents flowing through the resistance and the inductor, respectively, which can also be considered as the active or reactive component of the input current i_{AC} , respectively. The relationship between $i_{AC,d}$, $i_{AC,q}$, and i_{AC} can be written as

$$i_{AC} = i_{AC,d} + i_{AC,q}. \quad (9)$$

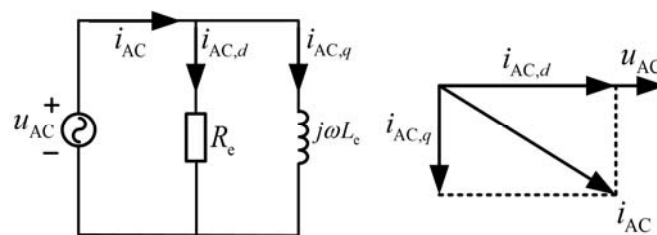


Figure 7. Equivalent circuit for reactive power output.

In addition, the output voltage of the MMC can be written as

$$e_j = R_e i_{AC,d} = R_e (i_{AC} - i_{AC,q}). \quad (10)$$

The reactive component of the input current $i_{AC,q}$ can be expressed as Equation (11). It shows that the $i_{AC,q}$ lags u_{AC} by $\frac{\pi}{2}$. Thus, a sine wave in phase with $i_{AC,q}$ can be obtained by a PLL or by delaying the u_{AC} by a quarter of a cycle. The amplitude of the $i_{AC,q}$ determines the value of reactive power.

$$i_{AC,q} = \frac{u_{AC}}{j\omega L_e} \quad (11)$$

The block diagram of the OCC with a reactive power controller is shown in Figure 8. The active power controller is the same as the controller described in Section 3. The control of reactive power is implemented by injecting a reactive current $i_{AC,q}$ into i_{AC} . The phase of the virtual current is obtained by delaying the AC voltage u_{AC} by a quarter of a cycle and the amplitude can be obtained by a PI controller, which is used to maintain the reactive power Q tracking the reference voltage Q^* . Positive amplitude means output current lags system voltage while negative amplitude means output current leads system voltage.

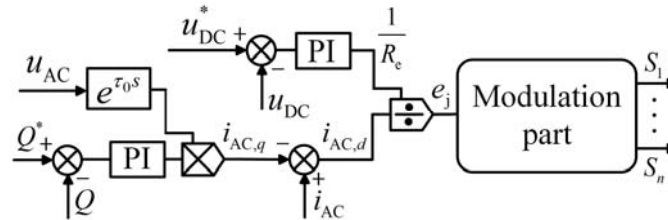


Figure 8. OCC algorithm with reactive power controller.

Compared to the traditional control strategies based on coordinate transformation, although the proposed controller still needs a PLL and DQ transform, the inverse transformation is no longer needed. In addition, since the DQ transform or the PLL is only used in the outer power loop and the outer loop has a lower frequency than the inner current loop, the proposed one has less calculation and because it requires less hardware, it also has cost advantages. However, it also has some limitations; because the controller can only output the positive active power, the MMC can only work in the rectifier mode, therefore it is still not suitable for applications requiring four-quadrant operations such as HVDC.

4.2. Inversion

When R_e is negative, the system will have a pole at the right half of the s-plane and it will cause a stability problem [17]. Thus, R_e must be positive and it is impossible to make the circuit work in the inverter mode by setting the R_e as a negative value directly. Therefore, for the OCC control algorithm in the inverter mode, the injection method described in the previous section is still used.

The equivalent circuit for the inverter mode under both the grid-connected and off-grid condition is shown in Figure 9. The output current i_{AC} is the sum of i_s and the fictitious current i_f . If i_f is negative and $|i_f| > |i_s|$, the converter will work under the inverter mode.

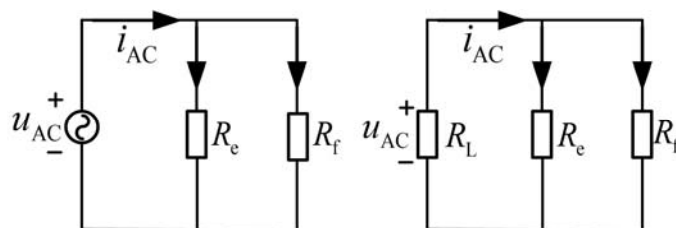


Figure 9. Equivalent circuit for inverter mode.

Similar to Equation (10), the output voltage of the MMC can be written as

$$e_j = R_e i_s = R_e (i_{AC} - i_f). \quad (12)$$

As for the grid-connected condition, the i_f is a sine wave with the same frequency and phase as the input voltage u_{AC} , so it can be easily derived by sampling the u_{AC} . Denote the amplitude of the i_f and i_s as I_f , I_s , respectively. If $I_f > -I_s$, the converter will work under the rectifier mode and when the $I_f < -I_s$, the converter will work under the inverter mode. For the off-grid condition, the frequency of the i_f determines the frequency of the output voltage.

The block diagram of the overall control strategy for the four-quadrant OCC-controlled MMC is shown in Figure 10. Both the active power and the reactive power have been controlled by the injecting strategy. The emulated resistance R_e can be given as any positive value, but in order to control the i_f conveniently, the R_e can be given as the maximum input voltage divided by the converter's maximum allowable current I_{max} , the amplitude of i_f can be limited as $[0, 2I_{max}]$ and the phase is always opposite to the phase of the u_{AC} . Thus, when the amplitude of i_f belongs to $[0, I_{max}]$, the system works under the rectifier mode and when it belongs to $[I_{max}, 2I_{max}]$, the system works under the inverter mode.

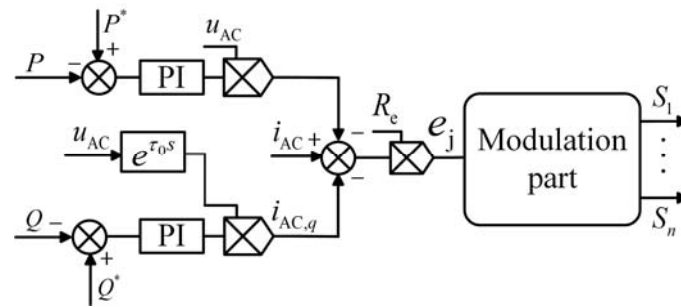


Figure 10. The block diagram of the overall control strategy.

This control module can implement four-quadrant operation for the OCC-controlled MMC with less computation, but compared to other control modules introduced in this paper, its structure is slightly complicated. This chapter introduces three different OCC control modules and they meet the requirements for the MMC in most applications, so the most suitable control modules can be selected to balance the features and costs.

5. Simulation and Experimental Results

In order to verify the correctness of the proposed control scheme, some simulations and experiments based on single-phase MMC have been carried out. To ensure the accuracy of the simulations and the experiments, both of them use the same parameters which have been listed in Table 2.

Table 2. Parameters used in the simulations and experiments.

Symbol	Description	Value
u_{AC}	AC system voltage	110 V(rms)
f	Power frequency	50 Hz
u_{DC}	DC-side voltage	330 V
N	Number of SMs in each arm	4
C_{SM}	SM capacitance	780 μ F
L_0	Arm inductance	14.7 mH
f_c	Control frequency	10 kHz
f_e	Equivalent control frequency	40 kHz

5.1. Simulation Results

In order to verify both the steady-state performance and dynamic characteristics of the proposed control strategies, several simulations have been carried out based on the MATLAB/ SIMULINK platform and all the three control modules mentioned in this paper have been tested.

5.1.1. Basic OCC Controller

Figure 11 shows the steady-state performance of the basic unit power factor OCC controller which is shown in Figure 6. Figure 11a,b shows the situation when $I_{AC} = 6$ A and $I_{AC} = 12$ A, respectively. Both of them have sinusoidal current waveforms with few ripples, but as described in Section 4.1, the phases of the voltage and current are not exactly the same. The current always lags behind the voltage and the situation becomes more severe as the current increases.

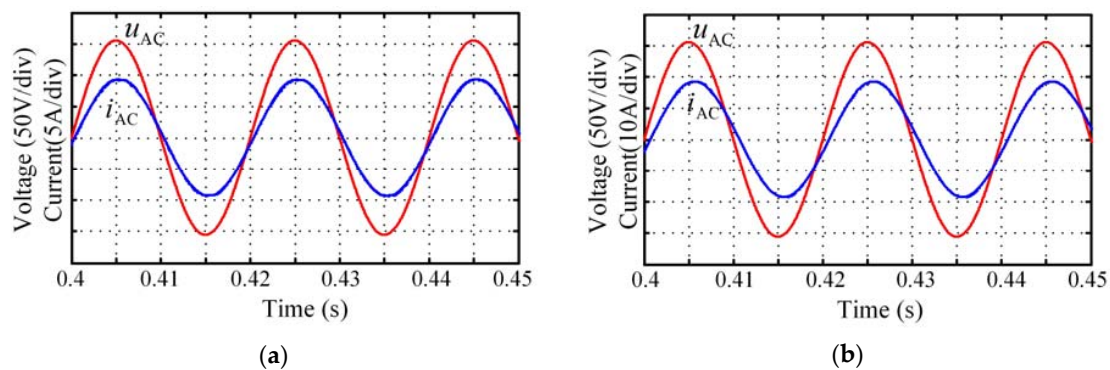


Figure 11. The steady-state performance of the unit power factor OCC controller: (a) $I_{AC} = 6$ A; (b) $I_{AC} = 12$ A.

The dynamic characteristics of this control strategy is shown in Figure 12. When $t = 0.4$ s, the DC voltage reference u_{DC}^* increased from 330 to 390 V. Then, because of the effect of the PI controller, the current increases rapidly and the DC voltage gradually rises. By the same time, the current begins to decrease gradually and after a short adjustment, when $t = 0.5$ s, the entire system regains a new steady state, and compared to the original state, the new one has a larger DC side ripple due to the increased DC side power.

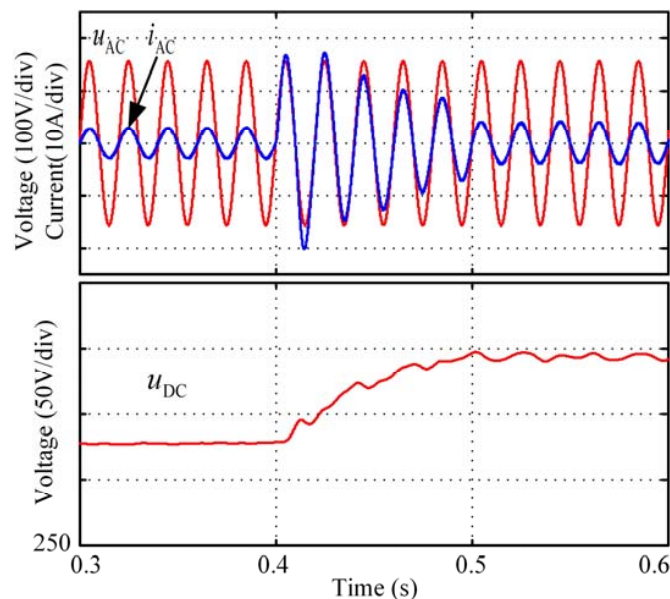


Figure 12. The situation when the direct current (DC) voltage reference changed.

5.1.2. Reactive-Power-Controlled OCC Controller

For the reactive-power-controlled OCC controller, which is shown in Figure 8, the steady-state performances are shown in Figure 13. Figure 13a,b shows the input current and voltage waveforms when the power factor angles φ_S are $\frac{\pi}{4}$, $-\frac{\pi}{4}$, respectively, and all of them have the same active current. From Figure 13a,b it can be seen that the controller has good performance in both the capacitive and the inductive reactive power output situations. Figure 13c shows the current and voltage under the unit power factor situation, and comparing Figure 13c with Figure 11b shows that both of them have almost the same current amplitude but Figure 13c is more likely to implement the unit power factor. Figure 13d shows the equivalent AC side output voltage of the MMC e_j calculated by Equation (6); the five-level voltage output can be clearly seen from the figure and at the same time the voltage fluctuation caused by the charge and the discharge of the SM capacitors can also be seen from the figure.

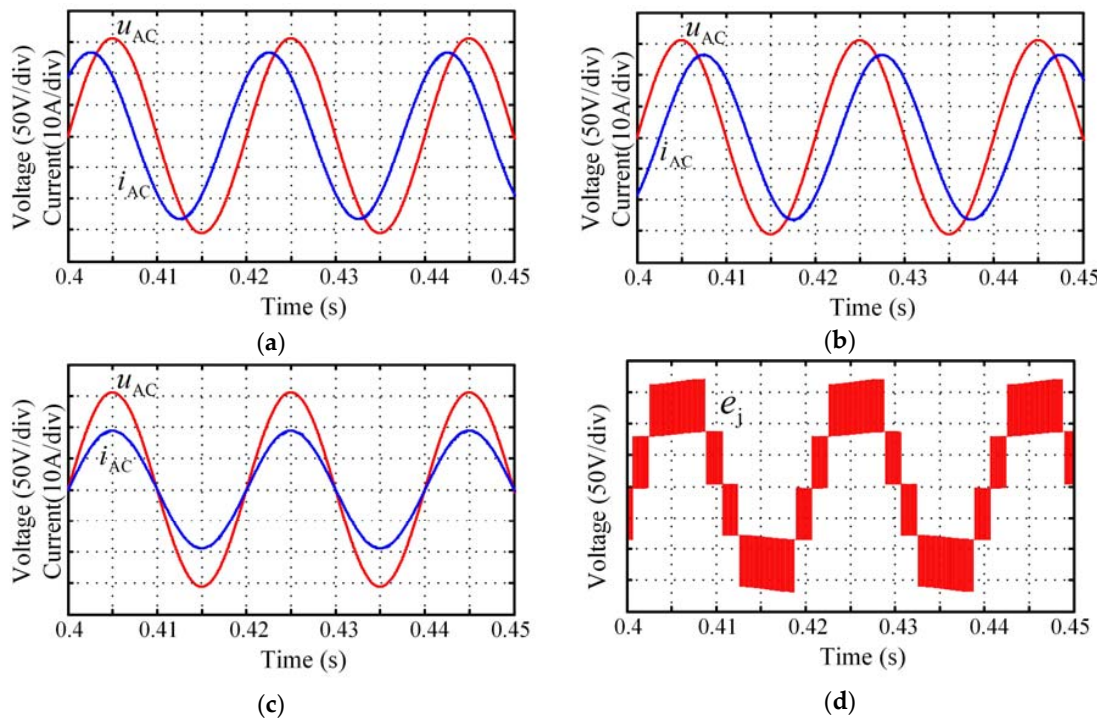


Figure 13. The steady-state performance of the reactive-power-controlled OCC controller: (a) $\varphi_S = \frac{\pi}{4}$; (b) $\varphi_S = -\frac{\pi}{4}$; (c) $\varphi_S = 0$; (d) the equivalent AC side output voltage of the MMC.

The dynamic characteristic is shown in Figure 14. When $t = 0.4$ s, let the reactive power reference become the opposite of the original value. The current changes quickly after the reference changes and is basically stable after half a grid voltage cycle.

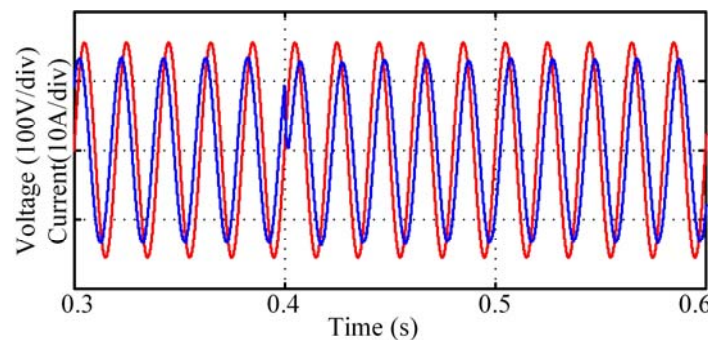


Figure 14. The dynamic state performance of the reactive-power-controlled OCC controller.

5.1.3. Four-Quadrant OCC Controller

The steady-state performance of the four-quadrant OCC controller which has been described in Section 4.2 is shown in Figure 15. In these cases, the DC side is connected to a 400 V DC power supply. Figure 15a,b shows the situation in the inversion mode under the grid-connect or off-grid situation, respectively. For the off-grid MMC, the AC power is replaced by a $10\ \Omega$ resistor and the reactive power controller is no longer needed. Since the AC power supply no longer needs to output active power to maintain the DC side voltage, it is possible to only output the reactive power to the AC side. Figure 15c,d shows the case when the current lags or leads the AC source voltage $\frac{\pi}{2}$, respectively.

The dynamic characteristic of the four-quadrant OCC controller is shown in Figure 16. When $t = 0.4\text{ s}$, the active power reference P^* changes from positive to negative, which means the system switches from rectification mode to inversion mode. The current changes quickly after the mode changes and is basically stable after half a grid voltage cycle.

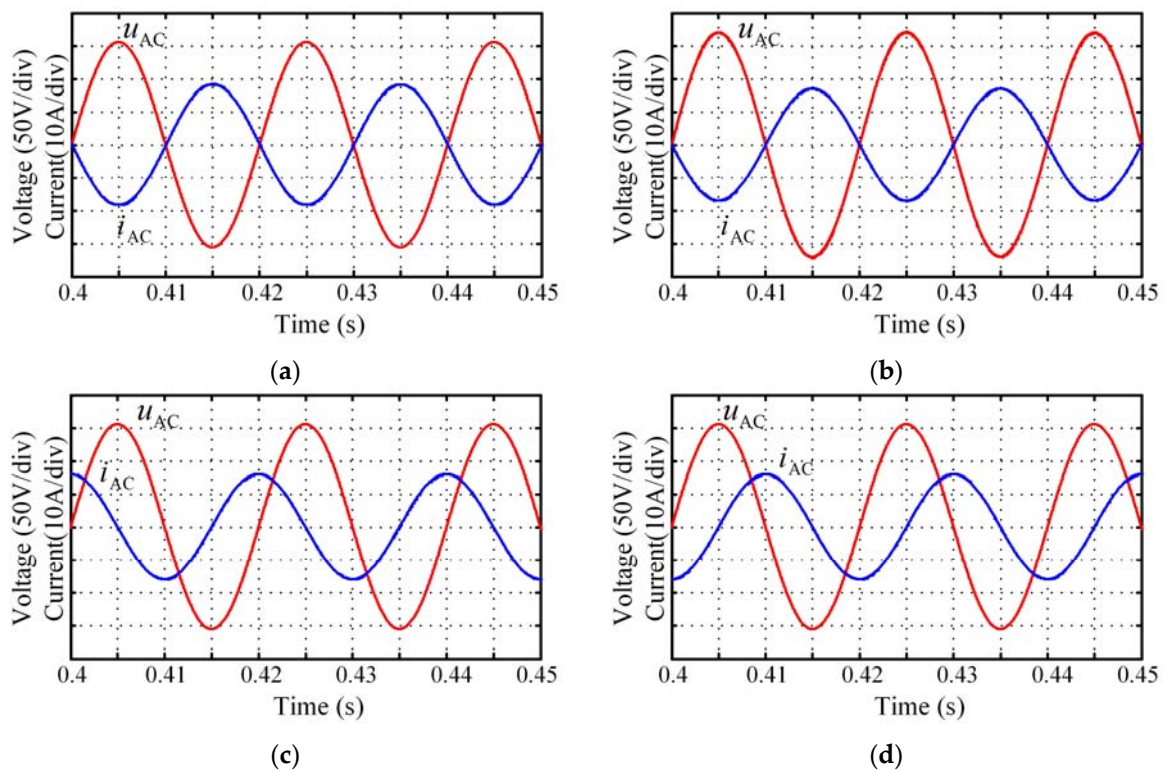


Figure 15. The steady-state performance of the four-quadrant OCC controller: (a) Grid-connect inversion; (b) off-grid inversion; (c) capacitive reactive power output; (d) inductive reactive power output.

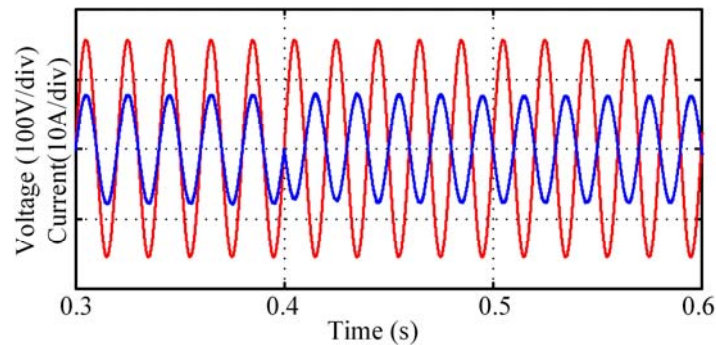


Figure 16. The dynamic state performance of the four-quadrant OCC controller.

To further verify the performance of the proposed controller, the AC currents under different operating conditions have been analyzed in the frequency domain. The total harmonic distortions (THDs) in the inversion and rectification conditions are 0.54% and 0.45%, respectively, and during the capacitive and inductive reactive power output conditions, the THDs are 0.4% and 0.43% separately. The spectrum analysis results under the four conditions are shown in Figure 17. As shown in the figure, although there are slight differences in the THDs, the AC currents have almost no harmonic contents under all the operating conditions.

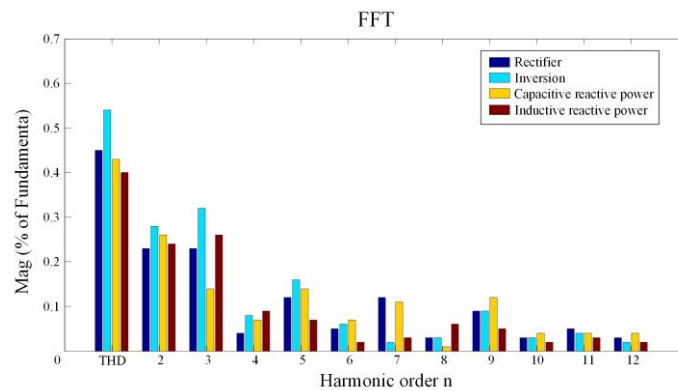


Figure 17. Harmonic comparison in different modes.

5.2. Experiment Results

To test the working condition of the controller in the real environment, a 200 W MMC prototype was built in the laboratory environment. The half bridge SM was made up by the IGBT IKCM15F60GA (600 V/15 A, Infineon). The digital control was implemented on a floating-point digital signal process (DSP) TMS320F28335 and a field-programmable gate array (FPGA) Xilinx-XC6SLX25. As Figure 18 shows, the DSP was used to calculate the output voltage and the FPGA generated the switching signals. The signal sample was implemented by the analog-to-digital converter (ADC) chip AD7606. Other parameters are shown in Table 1 and the experiment prototype is shown in Figure 19.

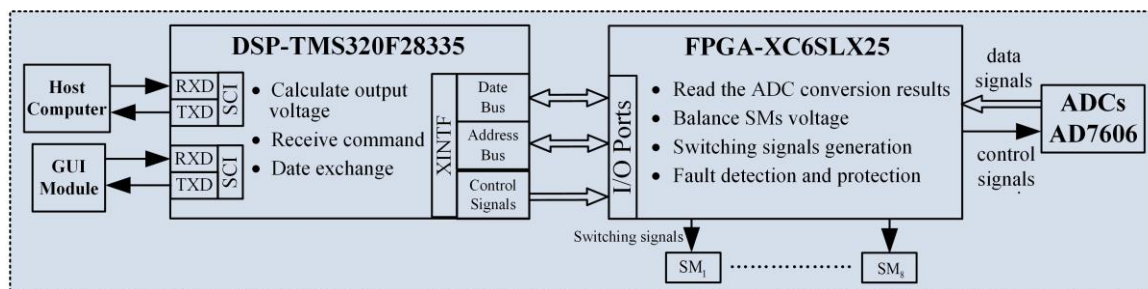


Figure 18. The hardware architecture and functional introduction of the control system. DSP: digital signal process.

The current and voltage waveforms under the unit power factor situation controlled by the conventional power factor OCC controller and the reactive-power-controlled OCC controller are shown in Figure 20a,b, respectively, and both of them have sinusoidal current waveforms with fewer ripples. Compared with the simulation results, the current in the experiment is smaller due to the limitation of the hardware conditions, but the conclusion that the current waveform shown in Figure 20b is closer to the unit power factor relative to Figure 20a can be obtained by observing the zero-crossing point of the current waveform. Figure 21 shows the equivalent AC side output voltage of the MMC and the five-level voltage output can be clearly seen from that figure.

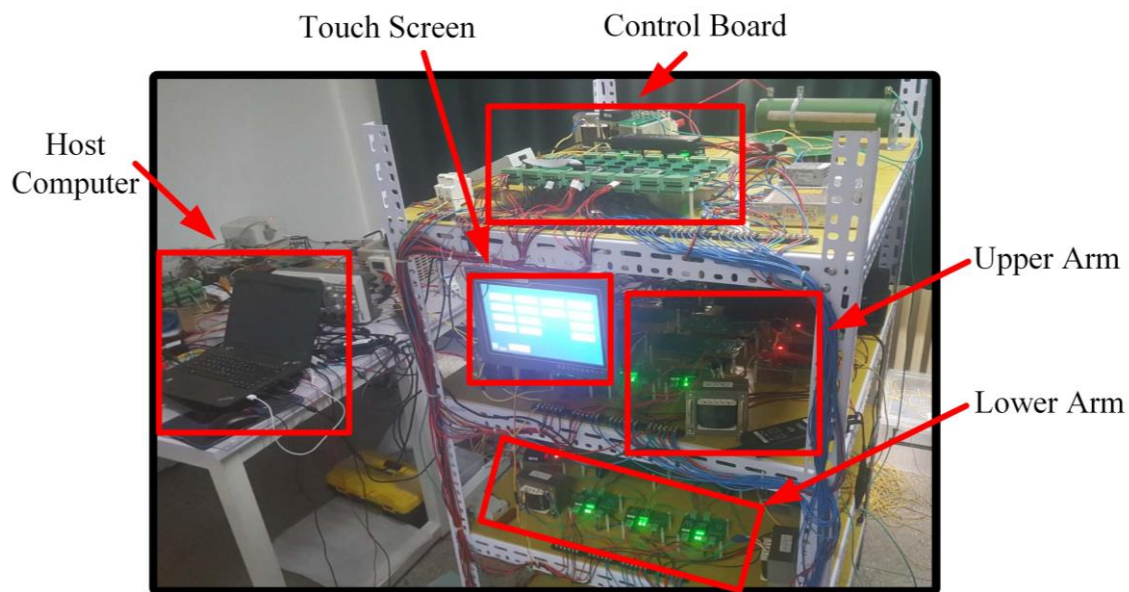


Figure 19. Experiment prototype.

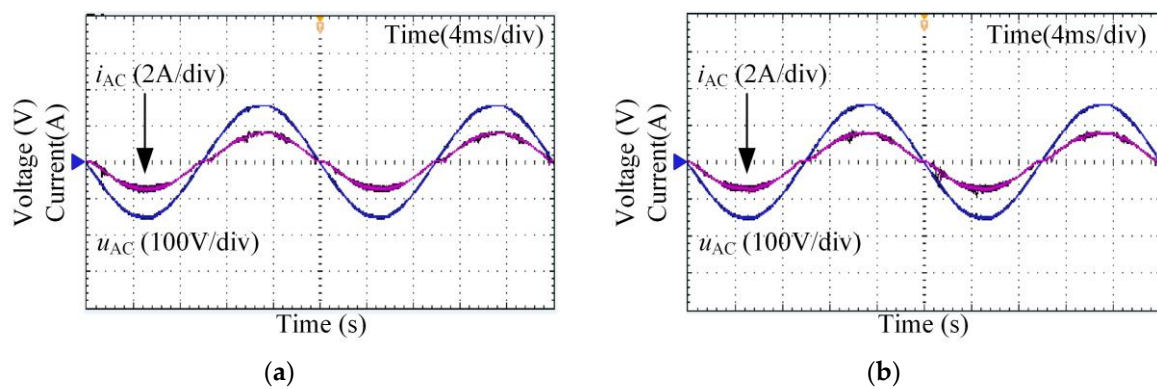


Figure 20. The current and voltage waveforms under the unit power factor situation controlled by: (a) traditional unit power factor OCC controller; (b) reactive-power-controlled OCC controller.

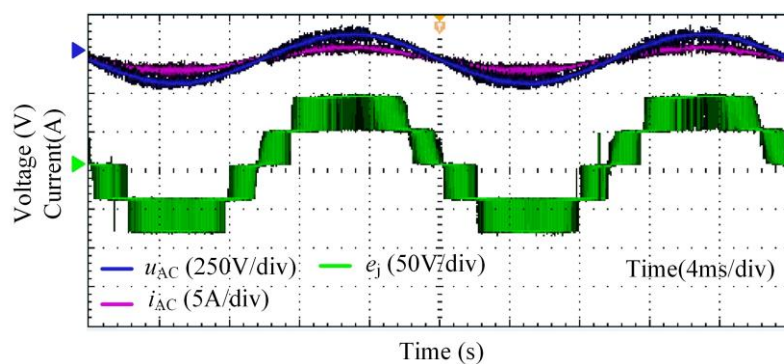


Figure 21. The equivalent AC side output voltage of the MMC.

The dynamic characteristics of the conventional unit power factor OCC controller are shown in Figure 22. After DC voltage reference u_{DC}^* changed, because of the effect of the PI controller, the current increases rapidly and the DC voltage gradually rises. The variation of three SM capacitor voltages are also shown, where V_{SM1} and V_{SM2} are the upper arm SM voltages and V_{SM3} is a lower arm SM voltage. It can be seen that the voltages of the three SMs are basically the same throughout the entire process and the voltage oscillation increases as the current increases.

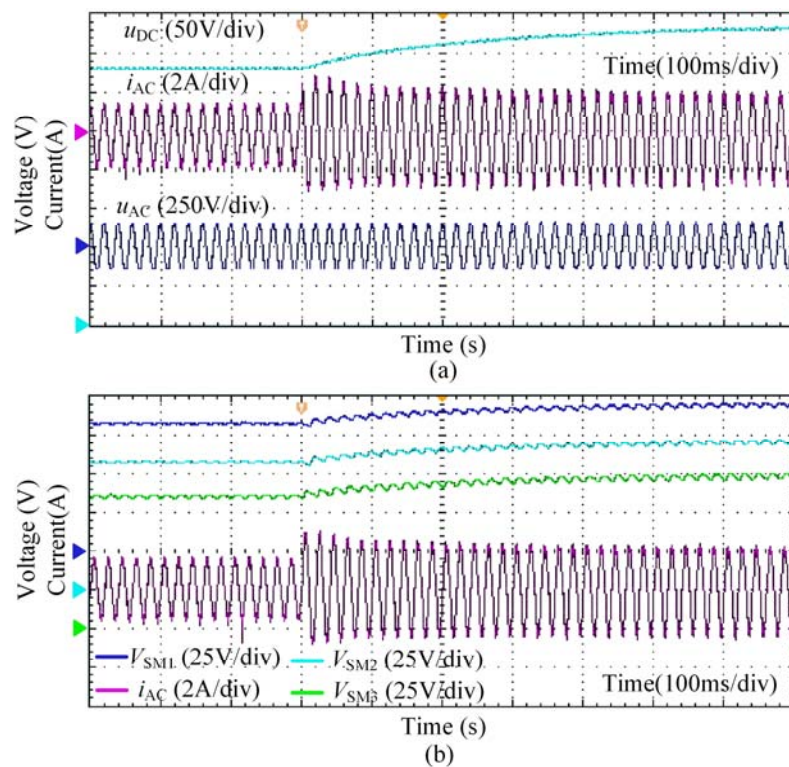


Figure 22. Experiment results for the dynamic response when the DC voltage reference changed: (a) DC bus voltage; (b) SM voltages.

For the reactive-power-controlled OCC controller, the situations are shown in Figure 23, the power factor angles φ_S are $\frac{\pi}{4}$, $-\frac{\pi}{4}$, respectively, and all of them have the same active current. The dynamic response when the reactive power reference becomes the opposite of the original value is shown in Figure 24. As can be seen from the figure, the adjustment of the reactive power is completed in two cycles and although the DC side voltage has some slight fluctuation due to the influence of the disturbance, the steady-state value of the DC voltage has not changed significantly, which means the active power has not been affected by the change of reactive power.

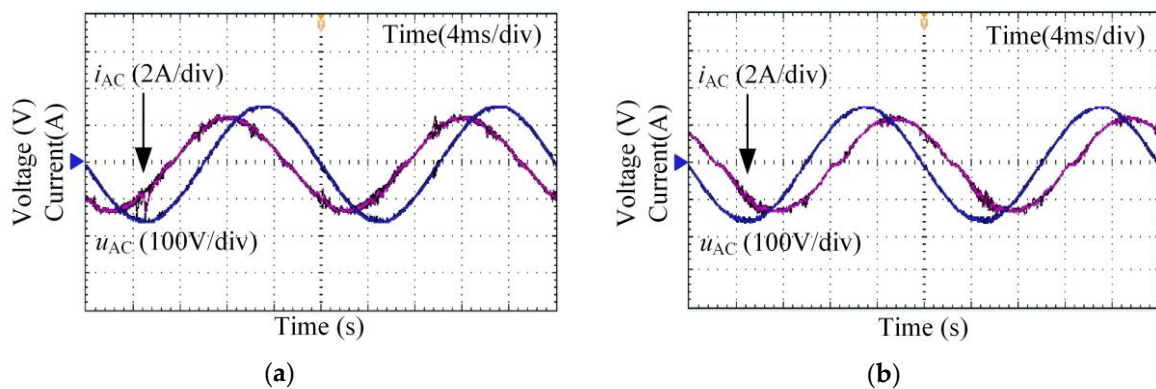


Figure 23. Experiment results for the steady-state performance of the reactive-power-controlled OCC controller: (a) $\varphi_S = \frac{\pi}{4}$; (b) $\varphi_S = -\frac{\pi}{4}$.

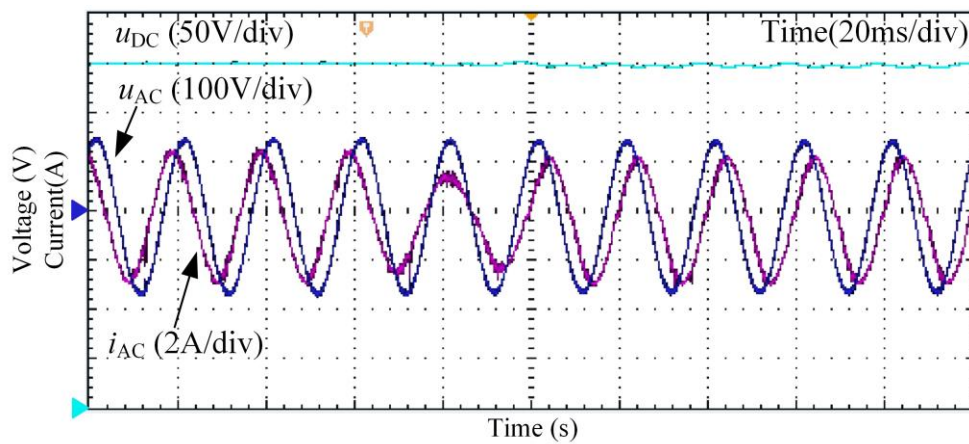


Figure 24. Experiment results for the dynamic response when the reactive power reference changed.

The steady-state performance of the four-quadrant OCC controller is shown in Figure 25. During these tests, the DC side is connected to a 400 V DC power supply. Figure 25a,b shows the situation in the inversion mode under the grid-connect or off-grid situation, respectively. For the off-grid MMC, the AC power is replaced by a 100 Ω resistor. Figure 25c,d shows the case when the current lags the AC source voltage $\frac{\pi}{2}$, respectively. The dynamic characteristic is shown in Figure 26; the working state of the system finishes the change from rectification mode to inversion mode in two cycles, which shows the superior dynamic characteristic of the controller.

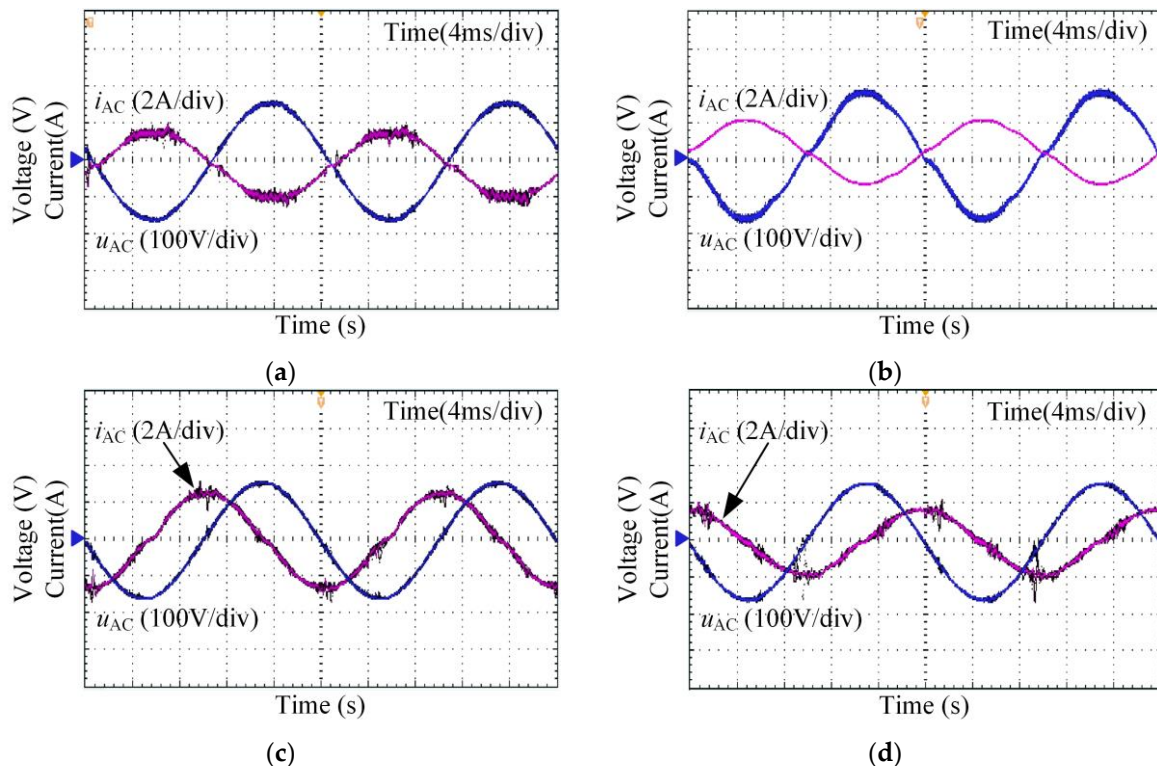


Figure 25. Experiment results for the steady-state performance of the four-quadrant OCC controller: (a) grid-connect inversion; (b) off-grid inversion; (c) capacitive reactive power output; (d) inductive reactive power output.

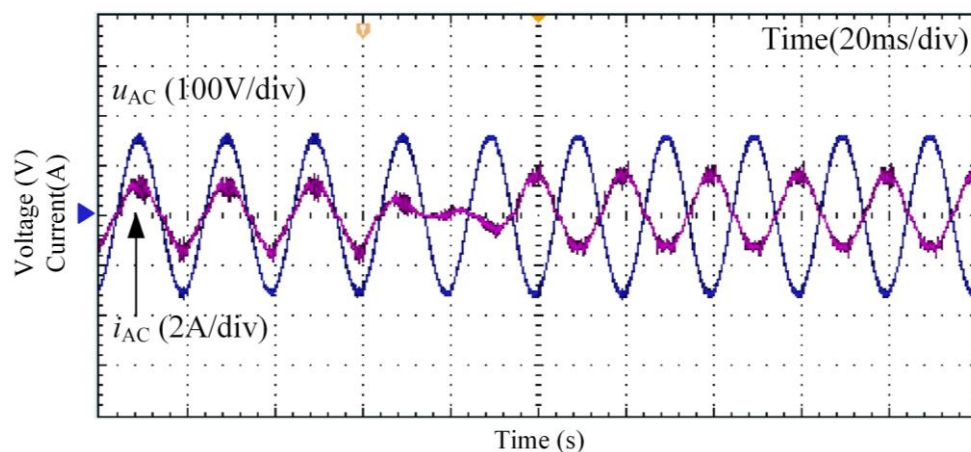


Figure 26. The performance when changed from rectification mode to inversion mode.

6. Conclusions

This paper proposes a modified OCC for MMCs. It uses the equivalent resistance constant principle so that the modulation strategy can be separated from the control strategy. By adding the equivalent impedance into the controller, the proposed OCC-based MMC controller can control reactive power output and makes MMC operate in the inverter mode. Compared with the traditional OCC used in MMCs, the proposed one has the following advantages: (1) unlike the traditional controller which can only use CPS-PWM to generate the switch signals, the modified one can be combined with both NLM and CPS-PWM; (2) compared to the traditional one which can only operate at the unity power factor, the proposed one expands the range of applications for the OCC-based MMC controller. Finally, the proposed controllers have been evaluated and validated by both simulation under the MATLAB/SIMULINK platform and an experiment on a five-level single-phase MMC prototype.

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