

Model Predictive Direct Power Control of an Inverter With Duty Cycle Optimization

Aswini.M¹, Bindu.A²

PG Scholar¹, AssistantProfessor², Department of EEE, CSI Institute of Technology, Thovalai, K.K.Dist,

Tamil Nadu, India.Email:aswinikpillai@gmail.com

Abstract—The paper proposes a model predictive direct power control (MPDPC) for a pulse width modulation (PWM) inverter. The conventional control methods achieve good steady-state and dynamic performance. However the performance relies on internal current control, proportional integral tuning and the predefined switching tables. Therefore a duty cycle optimization technique has been implemented by allocating a fraction of control period for a nonzero voltage vector and the rest time for a zero vector. The nonzero vector is selected by evaluating its effects on the cost function and its duration depends on the principle of power error minimization. Simulation and experimental results prove that, compared to the conventional methods the proposed method achieves better steady-state and dynamic performance at a reduced total harmonic distortion(THD) along with the elimination of current ripples. Furthermore nonlinearities can be included in the models at a small cost of control complexity increase.

Index Terms—Direct power control (DPC), duty cycle, model predictive control, pulse width modulation (PWM) inverter.

I.INTRODUCTION

THREE-PHASE pulse width modulation (PWM) inverters has been widely used for adjustable-speed ac drives, induction heating, stand by aircraft power supplies, uninterrupted power supplies, renewable energy systems, hvdc transmission lines and in power conversion system due to its merits of sinusoidal line current, controllable power factor, and good dc-link voltage regulation ability[1], [2]. Due to its widespread application, much effort has been done on the control of the PWM inverter. The 60 degree PWM, third harmonic PWM and space vector modulation are some of the conventional control methods of the three phase inverters.

Space vector modulation is a digital modulating technique where the objective is to generate PWM load line voltages that are in average equal to reference load line voltages. This is done in each sampling period by properly selecting the switch states of the inverter and the calculation of the appropriate time period for each state. The selection of the states and their time periods are accomplished by the space vector transformation. Although the control method generally achieves better steady-state performance and the sampling frequency can be reduced, they are usually complicated in principle.

The voltage oriented control (VOC), which decomposes the grid currents into active and reactive power components and regulate them separately [3]. Although good steady-state performance and dynamic responses are obtained using VOC, its performance relies heavily on the internal current control and fine proportional integral (PI) tuning [4]. However the most popular approach is using space vector modulation (SVM) and the high-level voltage reference value can be obtained in various ways, such as deadbeat control [14], sliding mode control [18], and PI-based control [13].VOC can also be categorized into this group due to the use of SVM.

The direct power control (DPC) is another kind of high performance control strategy for the PWM rectifier [6], which is similar to the direct torque control in motor drives [7], [8].Compared to the VOC, the DPC directly selects the desired voltage vector from a predefined switching table and eliminates the internal current loop. As a result, the dynamic response is very quick. However, the switching table in conventional DPC is obtained in a heuristic way, which cannot assure the effectiveness of the selected voltage vector [5], [9]. As a result, some authors have revised the conventional switching table to achieve performance improvement by proposing new switching tables [10], [11], using output regulation subspaces [5], [9] or fuzzy logic selection [12] to select the desired voltage vectors. However, the performance improvement is limited,

because the complete model of the PWM inverter and its future behavior are not taken into account.

The MPDPC methods have proved to be efficient over the conventional method. In the DPC method only one voltage vector is selected for the next control period, but their vector selection principles are very different. In the MPDPC, the complete model and future behavior of the PWM converter are taken into account. A cost function relating to power errors reduction is defined to evaluate the effects of each voltage vector and the one minimizing the cost function is selected. Compared to the heuristic switching table in DPC, the vector selected from the MPDPC is more accurate and effective in reducing power errors. Furthermore applying only one voltage vector during one control period fails to full exploit the potential of the technique in improving the steady state performance. Furthermore, due to the limited number of voltage vectors in the two-level converter, the performance improvement is limited and the sampling frequency should be high to ensure good performance.

Conventional DPC employ only one voltage vector during one control period. In fact, there are many control methods that use three voltage vectors during one control period to achieve reduced ripples and constant switching frequency. The three-vector-based approach has been widely applied in the power converter control [13], [14] and drives [15], [16]. It should be noted that there are also other ways to obtain the duration of each voltage vector without the help of SVM, as shown in [16] and [17]. Though all these methods assures good steady state performance, they are complicated in principle.

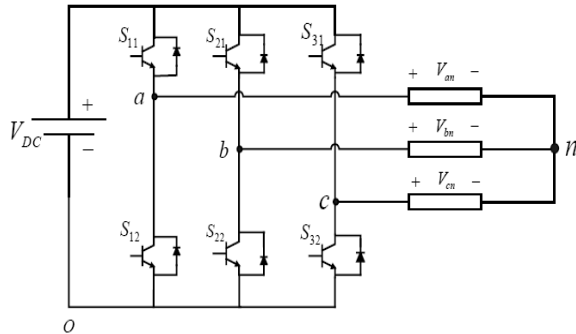


Fig. 1. Topology of a two-level PWM inverter

The paper introduces the concept of the duty cycle control in the MPDPC to achieve better performance than that of the conventional control methods. The

control period is divided into two intervals: one for a non zero vector and the other for a zero vector. The non zero vector is selected based on its effect on the cost function and duration is analytically derived based on the principle of power error minimization.

II. PREDICTIVE MODEL OF A PWM INVERTER

The circuit of a three-phase PWM inverter is shown in Fig.1, the model of a PWM converter can be expressed in stationary two-phase $\alpha\beta$ frame as

$$e = iR + L \frac{di}{dt} + v \quad (1)$$

where v , e , and i are inverter voltage vector, output voltage vector, and output current vector, respectively. The vector variables in (1) are transferred from the model in stationary abc frame [8] using the following transformation equation [20]:

$$x = \frac{2}{3} (x_a + x_b e^{j2\pi/3} + x_c e^{-j2\pi/3})$$

where x can be output voltage, output current, or inverter voltage. The complex power S at the output side can be calculated from output voltage and current vectors as

$$S = p + jq = 1.5(i^*e) \quad (2)$$

where “*” represents a conjugate operator; p and q are the active power and reactive power, respectively. Under the assumption of a balanced three-phase system, i.e., $e = |e|e^{j\omega t}$, where ω is the output frequency (rad/s), the output voltage differentiation e can be obtained as

$$\frac{de}{dt} = j\omega |e| e^{j\omega t} = j\omega e \quad (3)$$

From (1), the differentiation of grid current i can be obtained as

$$\frac{di}{dt} = \frac{1}{L} (e - v - Ri) \quad (4)$$

Substituting (3) and (4) into (2), the slope of complex power S in (2) can be obtained as

$$\frac{dS}{dt} = \frac{1}{L} [1.5(|e|^2 - v^*e) - (R - j\omega L).S] \quad (5)$$

Decomposing the real and imaginary components of (5), the differentiations of active and reactive powers are obtained as

$$\frac{dp}{dt} = \frac{3}{2}L[|e|^2 - \text{Re}(v^*e)] - \frac{R}{L}p - \omega q \quad (6)$$

$$\frac{dq}{dt} = \frac{3}{2}\text{Im}(v^*e) - \frac{R}{L}q + \omega p \quad (7)$$

From (6) and (7), the prediction of p and q at the next control period can be obtained as

$$p^{k+1} = p^k + \left(\frac{3}{2}L[|e^k|^2 - \text{Re}(\text{conj}(v^k) \cdot e^k)] - \frac{R}{L}p^k - \omega q^k\right) t_{sp} \quad (8)$$

$$q^{k+1} = q^k + \left(\frac{3}{2}L\text{Im}(\text{conj}(v^k) \cdot e^k) - \frac{R}{L}q^k + \omega p^k\right) t_{sp} \quad (9)$$

Where t_{sp} is the control period.

III. DUTY CYCLE CONTROL OF MPDPC

This paper tries to improve the performance of the control of converters by dividing the control period into two intervals for two vectors. This brings the benefit of better steady-state performance without dynamic performance degradation. There is no PI controller, so this project saves the controller tuning procedures. However, there are a number of design parameters in MPC, such as prediction horizon n , weighting matrices Q, R and sampling period.

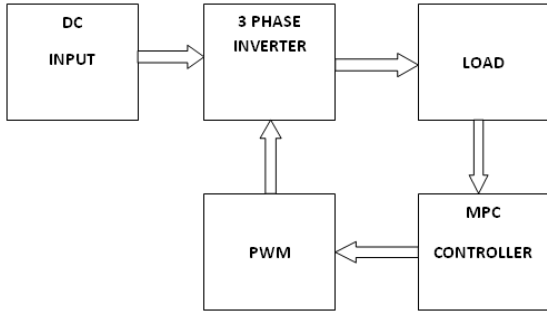


Fig.2. Control diagram of the MPDPC

The overall control diagram of the proposed MPC with duty cycle control is illustrated in Fig.2. The active power and reactive power at the k^{th} instant are calculated from the grid voltage and current using and then a two-step prediction is performed to obtain the value of p^{k+2} and q^{k+2} . The cost function defined in (15) is evaluated for each non zero voltage vector and the one minimizing is selected as the active vector. The duty cycle of the active vector is calculated by replacing p^{k+1} and q^{k+1} with p^{k+2} and q^{k+2} ,

respectively. Both the active vector and a zero vector will be applied in an appropriate sequence. In the following parts, the details regarding vector selection, vector duty cycle determination, and vector sequence are elaborated. The practical issue related to control delay compensation is also discussed.

A. Vector Selection

In the improved MPDPC with duty cycle control, there are two vectors during one control period: a nonzero (or active) vector and a zero vector. As shown in [17] and [21], generally the zero vector produces the smallest variations on active/reactive power slopes; hence, it is possible to use the zero vector along with the active vector to regulate the active power and reactive power more accurately and moderately. The selection of the active vector is based on minimizing a cost function for all possible candidate voltage vectors, which is similar to the principle of the conventional MPDPC. For the two-level converter, there are only eight discrete voltage vectors, so it is possible to evaluate the effects of each voltage vector and select the one minimizing the cost function. Generally, the cost function for power control is defined in such a way that both p and q at the end of this control period are as close as possible to the reference value, which is expressed as [22]

$$F = |S^{\text{ref}} - S^{k+1}|^2 = (p^{\text{ref}} - p^{k+1})^2 + (q^{\text{ref}} - q^{k+1})^2 \quad (10)$$

Where

$$S^{\text{ref}} = p^{\text{ref}} + jq^{\text{ref}} \text{ and } S^{k+1} = p^{k+1} + jq^{k+1}$$

are the reference value and estimated value. To achieve a unit power factor operation, the reference value of the reactive power is set to zero, so the cost function in (10) is changed to

$$F = (p^{\text{ref}} - p^{k+1})^2 + (0 - q^{k+1})^2 \quad (11)$$

According to (11), by evaluating the active and reactive powers at the next control period using (8) and (9) for each voltage vector, the best one minimizing (11) can be selected. However, there is a special case to consider for the proposed MPDPC, i.e., if the best voltage vector minimizing (11) is a zero vector, then a suboptimal vector should be selected rather than the zero vector. In other word, only the nonzero vector is necessary to be evaluated for the cost function in the proposed MPDPC. This is due to the fact that a zero vector has been selected as one of the two vectors for the proposed MPDPC with duty cycle. For this special case, the combination of

second-best vector (nonzero vector) and best vector (zero vector) satisfies the cost function (11) better than that of using zero vector only.

B. Vector Duty Cycle Control

After selecting the active vector and zero vector from Section III-A, it is vital to decide the durations of each vector, or the duty cycle of the active vector. In this paper, a least-square optimization method is used to obtain the duration of the active vector by minimizing the equation in (11). Suppose the slopes of active power are s_1 and s_2 for the active vector and the zero vector, and the slopes of reactive power are s_{11} and s_{22} , which can be easily obtained from (6) and (7). The active and reactive powers at the end of control period can be expressed as

$$p^{k+1} = p^k + s_1 t_v + s_2 (t_{sp} - t_v) \quad (12a)$$

$$q^{k+1} = q^k + s_{11} t_v + s_{22} (t_{sp} - t_v) \quad (12b)$$

where t_v is the duration of the active vector. The optimal duration of t_v that minimizes F during a control period satisfies the following condition:

$$\frac{\partial F}{\partial t_v} = 0 \quad (13)$$

Solving (13), the duration of the active vector can be obtained as

$$t_v = \frac{(p^{ref} - p^{k+1})(s_1 - s_2) + (0 - q^{k+1})(s_{11} - s_{22})}{(s_1 - s_2)^2 + (s_{11} - s_{22})^2} + \frac{t_{sp}(s_2^2 + s_{22}^2 - s_1 s_2 - s_{11} s_{22})}{(s_1 - s_2)^2 + (s_{11} - s_{22})^2} \quad (14)$$

It should be noted that the value of t_v is saturated to 0 if $t_v < 0$ or t_{sp} if $t_v > t_{sp}$ for the protection of t_v .

C. Vector Sequence

Generally, the active vector will be applied first, followed by an appropriate zero vector with minimal switching jumps. For example, if the voltage vector “110” is selected as the active voltage vector, the appropriate voltage vector will be “111” rather than “000.” However, the zero vector may be applied first if the prior vector sequence contains the same zero vector. For example, if the vectors during the last period are “100” and “000” with “000” at the end, and the vectors to be applied in the next period are “001” and “000,” in that case, “000” instead of “001” will be applied first to decrease the switching frequency. The vector durations should be changed accordingly if there is a vector sequence change.

D. Control Delay Compensation

It is well known that there is a one-step delay in digital implementation. In other word, the voltage vector decided at the k^{th} instant will not be applied until the $(k+1)^{\text{th}}$ instant. To eliminate this delay, the value at the $(k+2)^{\text{th}}$ instant should be used rather than the $(k+1)^{\text{th}}$ instant, which requires a two-step prediction. As a result, the cost function should be changed to

$$F = (p^{ref} - p^{k+2})^2 + (0 - q^{k+2})^2 \quad (15)$$

where p^{k+2} and q^{k+2} are predicted from p^{k+1} and q^{k+1} .

The values of p^{k+1} and q^{k+1} have been predicted from using the voltage v^k decided at the $(k-1)^{\text{th}}$ instant. Consequently, the values of p^{k+2} and q^{k+2} can be calculated for each non zero rectifier voltage vector v^{k+1} with initial states of p^{k+1} and q^{k+1} . The voltage vector minimizing the cost function is selected as the best voltage vector. The equations for predicting p^{k+2} and q^{k+2} are

$$p^{k+2} = p^{k+1} + \left(\frac{3}{2} L |e^{k+1}|^2 - \text{Re}(\text{conj}(v^{k+1}) \cdot e^{k+1})\right) - \frac{R}{L} p^{k+1} - \omega q^{k+1} t_{sp} \quad (16)$$

$$q^{k+2} = q^{k+1} + \left(\frac{-3}{2} L \text{Im}(\text{conj}(v^{k+1}) \cdot e^{k+1})\right) - \frac{R}{L} q^{k+1} - \omega p^{k+1} t_{sp} \quad (17)$$

The voltage vector at the $(k+1)^{\text{th}}$ instant is predicted

$$e^{k+1} = e^{j\omega t_{sp}} e^k \approx (1 + j\omega t_{sp}) e^k \quad (18)$$

IV. SIMULATION AND EXPERIMENTAL RESULTS

A MATLAB/Simulink routine was constructed to determine the behavior of the proposed MPDPC of 3 phase inverter. The active power and reactive power at the k^{th} instant are calculated from the voltage and current and then a two-step prediction is performed to obtain the value of p^{k+2} and q^{k+2} . The cost function is evaluated for each nonzero voltage vector and the one minimizing the cost function is selected as the active vector. The duty cycle of the active vector is calculated by replacing p^{k+1} and q^{k+1} with p^{k+2} and q^{k+2} , respectively. Both the active vector and a zero vector will be applied in an appropriate sequence.

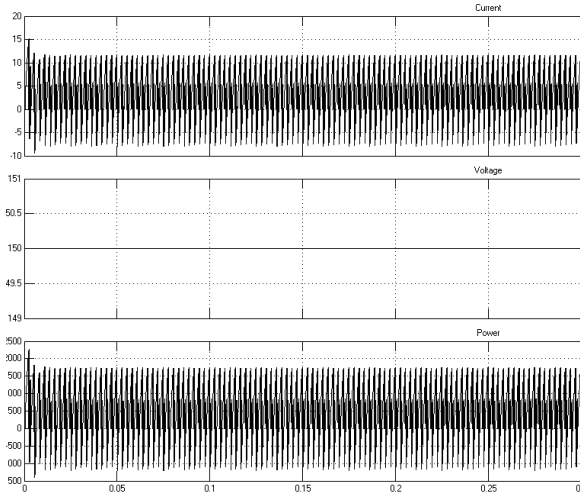


Fig.3.Input current,voltage and power waveforms.

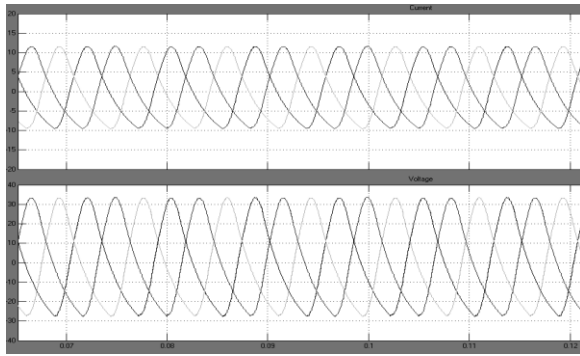


Fig.4. Output current and voltage waveforms.

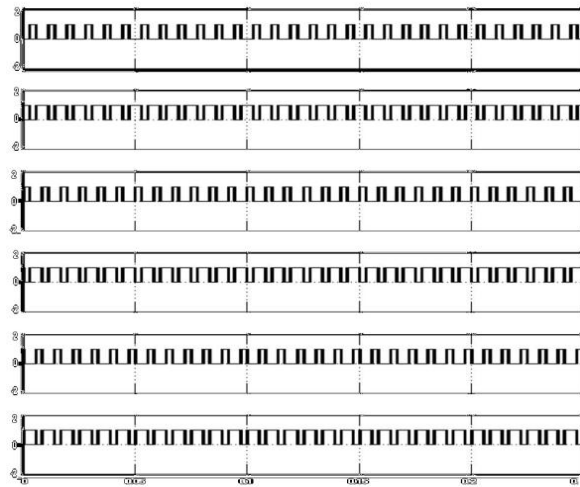


Fig.5. PWM output.

It is observed that the proposed MPDPC with duty cycle control presents less power ripples and lower current harmonics. This is further confirmed by the

harmonic spectrum of currents. By introducing the duty cycle control in the MPDPC, the total harmonic distortion (THD) of current is reduced effectively exhibiting excellent performance in improving the quality of current. For the conventional MPDPC, the harmonics are mainly distributed in the range of below 20 kHz, which may not be easy to be filtered. For the proposed MPDPC with duty cycle control, the low-order harmonic contents are lower and there are some high-order harmonics which may facilitate the filter design. There is only small drop in the dc voltage and it returns to its reference value quickly, showing strong robustness against load disturbance.

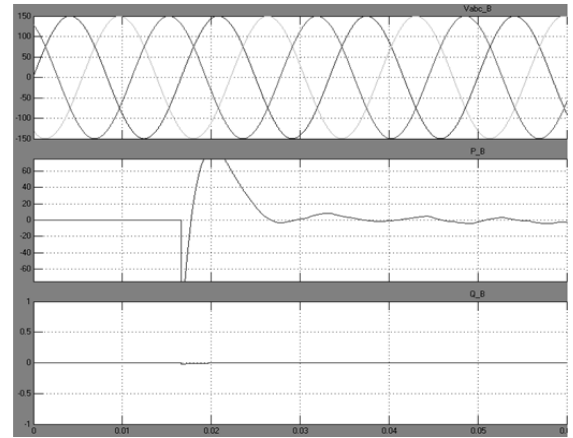


Fig.5.Active and reactive power output

There is very insignificant difference in the dynamic process for both methods, but the power ripples in the proposed MPDPC are smaller, validating its superiority in steady-state performance improvement. Furthermore it is observed that if the inductance value used in the control is 50% of the real value, the ripple in the active and reactive powers will increase and become more irregular. Furthermore, some dc offset in the reactive power will appear, which affects the unit power factor operation of the system. On the other hand, if the inductance value used in the control is increased up to 100%, the power ripples will increase significantly. However, the unit power factor operation is maintained. The results indicate that the variation of the inductance has influence on the steady-state performance of the proposed MPDPC, but does not influence the system stability at least in the range of 50% to 100% inductance variations. To maintain the control performance, it is better to use a online inductance identification technique.

V.CONCLUSION

An improved MPDPC for a PWM inverter is proposed in this paper by using a duty cycle control.

Different from the conventional control methods, the proposed MPDPC which applies one non zero vector and one zero vector during one control period achieves improved steady state performance. The duration of the non zero vector is obtained based on the principle of minimizing the errors of both active power and reactive power at the end of control period. The issues of vector selection, vector duty cycle, vector sequence, and control delay compensation are discussed in detail. Furthermore, extensive simulation results are discussed. It is observed that the proposed MPDPC can achieve reduced power ripples and lower current THD while maintaining high dynamic response. Considering the hardware burden for lower sampling frequency in the case of conventional control methods it is concluded that the proposed MPDPC is more favorable and practical to achieve performance improvement.

REFERENCES

- [1] B. Singh, B. N. Singh, A. Chandra, K. Al-Haddad, A. Pandey, and D. P. Kothari, "A review of three-phase improved power quality ac-dc converters," *IEEE Trans. Ind. Electron.*, vol. 51, no. 3, pp. 641–660, Jun. 2004.
- [2] J. Rodriguez, J. Dixon, J. Espinoza, J. Pontt, and P. Lezana, "PWM regenerative rectifiers: State of the art," *IEEE Trans. Ind. Electron.*, vol. 52, no. 1, pp. 5–22, Feb. 2005.
- [3] V. Blasko and V. Kaura, "A new mathematical model and control of a three-phase ac-dc voltage source converter," *IEEE Trans. Power Electron.*, vol. 12, no. 1, pp. 116–123, Jan. 1997.
- [4] P. Antoniewicz and M. Kazmierkowski, "Virtual-flux-based predictive direct power control of ac/dc converters with online inductance estimation," *IEEE Trans. Ind. Electron.*, vol. 55, no. 12, pp. 4381–4390, Dec. 2008.
- [5] J. Alonso-Martinez, J. E. Carrasco, and S. Arnaltes, "Table-based direct power control: A critical review for micro grid applications," *IEEE Trans. Power Electron.*, vol. 25, no. 12, pp. 2949–2961, Dec. 2010.
- [6] T. Noguchi, H. Tomiki, S. Kondo, and I. Takahashi, "Direct power control of PWM converter without power-source voltage sensors," *IEEE Trans. Ind. Appl.*, vol. 34, no. 3, pp. 473–479, May–Jun. 1998.
- [7] I. Takahashi and T. Noguchi, "A new quick-response and high-efficiency control strategy of an induction motor," *IEEE Trans. Ind. Appl.*, vol. IA-22, no. 5, pp. 820–827, Sep. 1986.
- [8] Y. Zhang and J. Zhu, "Direct torque control of permanent magnet synchronous motor with reduced torque ripple and commutation frequency," *IEEE Trans. Power Electron.*, vol. 26, no. 1, pp. 235–248, Jan. 2011.
- [9] G. Escobar, A. Stankovic, J. Carrasco, E. Galvan, and R. Ortega, "Analysis and design of direct power control (DPC) for a three phase synchronous rectifier via output regulation subspaces," *IEEE Trans. Power Electron.*, vol. 18, no. 3, pp. 823–830, May 2003.
- [10] A. Baktash, A. Vahedi, and M. A. S. Masoum, "Improved switching table for direct power control of three-phase PWM rectifier," in *Proc. Australasian Univ. Power Eng. Conf.*, 2007, pp. 1–5.
- [11] A. Bouafia, J.-P. Gaubert, and F. Krim, "Analysis and design of new switching table for direct power control of three phase PWM rectifier," in *Proc. 13th Power Electron. Motion Control Conf.*, 2008, pp. 703–709.
- [12] A. Bouafia, F. Krim, and J.-P. Gaubert, "Fuzzy-logic-based switching state selection for direct power control of three-phase PWM rectifier," *IEEE Trans. Ind. Electron.*, vol. 56, no. 6, pp. 1984–1992, Jun. 2009.
- [13] M. Malinowski, M. Jasinski, and M. Kazmierkowski, "Simple direct power control of three-phase PWM rectifier using space-vector modulation (DPC-SVM)," *IEEE Trans. Ind. Electron.*, vol. 51, no. 2, pp. 447–454, Apr. 2004.
- [14] A. Bouafia, J.-P. Gaubert, and F. Krim, "Predictive direct power control of three-phase pulse width modulation (PWM) rectifier using space-vector modulation (SVM)," *IEEE Trans. Power Electron.*, vol. 25, no. 1, pp. 228–236, Jan. 2010.
- [15] Y. Zhang, J. Zhu, W. Xu, and Y. Guo, "A simple method to reduce torque ripple in direct torque-controlled permanent-magnet synchronous motor by using vectors with variable amplitude and angle," *IEEE Trans. Ind. Electron.*, vol. 58, no. 7, pp. 2848–2859, Jul. 2011.

[16] Y. Zhang and J. Zhu, "A novel duty cycle control strategy to reduce both torque and flux ripples for DTC of permanent magnet synchronous motor drives with switching frequency reduction," *IEEE Trans. Power Electron.*, vol. 26, no. 10, pp. 3055–3067, Oct. 2011.

[17] S. A. Larrinaga, M. A. R. Vidal, E. Oyarbide, and J. R. T. Apraiz, "Predictive control strategy for dc/ac converters based on direct power control," *IEEE Trans. Ind. Electron.*, vol. 54, no. 3, pp. 1261–1271, Jun. 2007.

[18] J. Hu, L. Shang, Y. He, and Z. Q. Zhu, "Direct active and reactive power regulation of grid-connected dc/ac converters using sliding mode control approach," *IEEE Trans. Power Electron.*, vol. 26, no. 1, pp. 210–222, Jan. 2011.

[19] N. Flourentzou, V. G. Agelidis, and G. D. Demetriades, "VSC-based HVDC power transmission systems: An overview," *IEEE Trans. Power Electron.*, vol. 24, no. 3, pp. 592–602, Mar. 2009.

[20] P. Cortes, J. Rodriguez, P. Antoniewicz, and M. Kazmierkowski, "Direct power control of an AFE using predictive control," *IEEE Trans. Power Electron.*, vol. 23, no. 5, pp. 2516–2523, Sep. 2008.

[21] Y. Zhang, Z. Li, Y. Zhang, W. Xie, Z. Piao, and C. Hu, "Performance improvement of direct power control of PWM rectifier with simple calculation," *IEEE Trans. Power Electron.*, vol. 28, no. 7, pp. 3428–3437, Jul. 2013.

[22] S. Kouro, P. Cortes, R. Vargas, U. Ammann, and J. Rodriguez, "Model predictive control—a simple and powerful method to control power converters," *IEEE Trans. Ind. Electron.*, vol. 56, no. 6, pp. 1826–1838, Jun. 2009.