

Predictive Control of a T-type Inverter in Wind Energy Conversion Systems

Guoliang YANG^{1*}, Zi YANG², Yiming FANG²

Key Lab of Power Electronics for Energy Conservation and Motor Drive of Hebei Province
Yanshan University
Qinhuangdao 066004, China

¹ E-mail: y99ygl@ysu.edu.cn; ² E-mail: 312958271@qq.com

+* Corresponding author

Abstract—A new controlling method of three level inverters is proposed in this paper. Three-level T inverters are widely used in industry for high-power, medium-voltage power conversion and drives. Topics related to power losses due to commutation and quality of the output current is relevant issues in this power range. The neutral-point balancing problem in this topology is another subject that has been studied in recent years. Among the most common control methods for this converter, the literature states are non linear techniques, like hysteresis control, and linear methods, like the use of PI controllers in conjunction with pulse width modulation. The general predictive control scheme is applied here to the inverter. The behavior of the system is predicted for each possible switching state of this kind of inverter. The switching state that minimizes a given cost function is selected to be applied during the next sampling interval following the same strategy.

Keywords—predictive control; T-type inverter; grid connected

I. INTRODUCTION

RECENTLY, the three-level T-type inverter has been proposed for high-efficiency systems in low-voltage applications such as Wind Energy Conversion Systems, power factor corrector(PFC) rectifier, and automotive inverter systems [1]. Fig. 1. shows the simplified circuit of a T-type inverter.

A. Grid Model of T-type Inverter

Taking into account the definitions of variables from the circuit shown in Fig. 1[2], leaving out the currents flow by the capacitors, the equations for load current dynamics for each phase can be written as

$$v_{kN} = L \frac{di_k}{dt} + Ri_k + e_k, \quad k = a, b, c \quad (1)$$

Considering the unitary vector $a = e^{j2\pi/3}$, which represents the 120° phase displacement between the phases, the output voltage vector can be defined as

$$v = \frac{2}{3}(v_{aN} + av_{bN} + a^2v_{cN}) \quad (2)$$

where R is the load resistance and L the load inductance.

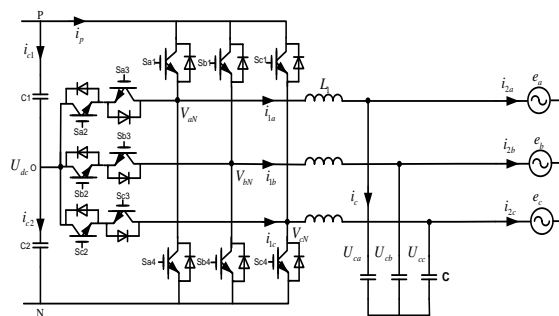


Figure 1. Grid connected model of T-type inverter

B. Discrete-time Model for Prediction

The discrete-time model will be used to predict the future value of grid current from voltages and measured currents at the t_k sampling instant [3]. So the grid current derivative di/dt is replaced by a forward Euler approximation. That is, the derivative is approximated as follows:

$$\frac{di}{dt} \approx \frac{i(k+1) - i(k)}{T_s} \quad (3)$$

which is substituted in (3) to obtain an expression that allows prediction of the future load current at time t_{k+1} , for each one of the 27 values of voltage vector $v(k)$ generated by the inverter[4]. This expression is

$$i^p(k+1) = (1 - \frac{RT_s}{L})i(k) + \frac{T_s}{L}(v(k) - e(k)) \quad (4)$$

Where $e(k)$ denotes the grid voltage, which can be measured at instant k. The superscript p denotes the predicted variables.

The reference currents can be calculated from considering the power reference $P^* = 1kw, Q^* = 0$. According to the theory of instantaneous power[5], The instantaneous active and reactive power of the system can be calculated with the following expression;

$$\begin{bmatrix} P \\ Q \end{bmatrix} = \frac{3}{2} \begin{bmatrix} e_d & e_q \\ e_q & -e_d \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix}$$

So it's obvious to get the reference currents

$$\begin{bmatrix} i_d^* \\ i_q^* \end{bmatrix} = \frac{2}{3[(e_d)^2 + (e_q)^2]} \begin{bmatrix} e_q & -e_d \\ e_d & e_q \end{bmatrix} \begin{bmatrix} P^* \\ Q^* \end{bmatrix} \quad (6)$$

The same approximation of the derivative can be used for the capacitor voltages for a sampling time T_s

$$\frac{dv_{cx}}{dt} \approx \frac{v_{cx}(k+1) - v_{cx}(k)}{T_s} \quad (7)$$

giving the following discrete-time equations:

$$v_{c1}^p(k+1) = v_{c1}(k) + \frac{1}{C} i_{c1}(k) T_s \quad (8)$$

$$v_{c2}^p(k+1) = v_{c2}(k) + \frac{1}{C} i_{c2}(k) T_s \quad (9)$$

where currents $i_{c1}(k)$ and $i_{c2}(k)$ depend on the switching state of the inverter and the value of the output currents, and can be calculated using the following expressions:

$$i_{c1}(k) = i_{dc}(k) - H_{1a} i_a(k) - H_{1b} i_b(k) - H_{1c} i_c(k) \quad (10)$$

$$i_{c2}(k) = i_{dc}(k) + H_{2a} i_a(k) + H_{2b} i_b(k) + H_{2c} i_c(k) \quad (11)$$

where i_{dc} is the current supplied by the voltage source V_{dc} . Variables H_{1x} and H_{2x} depend on the switching states and are defined as

$$H_{1x} = \begin{cases} 1, & \text{if } S_x = "+" \\ 0, & \text{otherwise} \end{cases} \quad (12)$$

$$H_{2x} = \begin{cases} 1, & \text{if } S_x = "-" \\ 0, & \text{otherwise} \end{cases} \quad (13)$$

with $x = a, b, c$.

Hence, (7)-(13) allow us to predict the effect of selecting a given switching state on the variation of the capacitor voltages [6].

II. PREDICTIVE CURRENT CONTROL METHOD

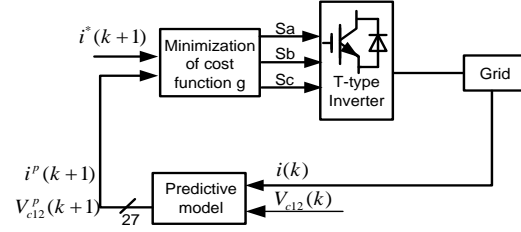


Figure 2. Predictive current control method for the T-type inverter

The control requirements for the T-type inverter are Grid current reference tracking, DC link capacitor voltages balance and reduction of the switching frequency.

These requirements can be formulated in the form of a cost function to be minimized. The cost function for the T-type inverter has the following composition:

$$g = |i_\alpha^* - i_\alpha^p| + |i_\beta^* - i_\beta^p| + \lambda_{dc} |v_{c1}^p - v_{c2}^p| + \lambda_n n_c \quad (14)$$

The first two terms are the load current errors in orthogonal coordinates, where i_α^p and i_β^p are the real and imaginary components of the predicted current vector i^p , respectively, and i_α^* and i_β^* are the real and imaginary components of the reference current vector i^* .

A. Minimization of the Switching Frequency

As in power converters, one of the major measures of control effort is the switching frequency. It is important in many applications to be able to control or limit the number of commutations of the power switches [7].

To directly consider the reduction in the number of commutations in the cost function, a simple approach is to include a term in it that covers the number of switches that change when the switching state $S(k)$ is applied, with respect to the previously applied switching state $S(k-1)$.

The resulting cost function is expressed as $g = (i_\alpha^* - i_\alpha^p) + (i_\beta^* - i_\beta^p) + \lambda_n \cdot n$. Considering the three-phase inverter as an example, the switching state vector $S = (S_a, S_b, S_c)$, defines the switching state of each inverter leg [8]. Then the number of switches changing from time $k-1$ to time k is

$$n = |S_a(k) - S_a(k-1)| + |S_b(k) - S_b(k-1)| + |S_c(k) - S_c(k-1)| \quad (15)$$

B. Capacitor Voltage Balance

One of the most interesting aspects of the predictive control method is the simplicity for implementing voltage balance in the DC link [12]. This feature was tested by disconnecting the middle point of the DC link from the source and applying the predictive control method with the following cost function:

$$g = |i_{\alpha}^* - i_{\alpha}^p| + |i_{\beta}^* - i_{\beta}^p| + \lambda_{dc} |v_{c1}^p - v_{c2}^p| \quad (16)$$

Nevertheless, to take advantage of the possibilities offered by this control method, it is necessary to adjust parameters λ_{dc} and λ_n . First, the designer should consider the different units and magnitudes of the variables involved in the cost function g . This will give some idea about the order of magnitude of the weighting factors for equal importance of all terms. If the designer wants to maintain voltage balance in the DC link only by selecting the appropriate switching state from the redundant states that generate a given voltage vector, then a small value of λ_{dc} should be used. The smallest value allowed by the implementation platform will work for that purpose. The same criteria can be applied to λ_n . When increasing λ_n , the method could choose switching states that are not within the optimal voltage vector

in terms of reference tracking, but imply fewer commutations.

Summarizing, the predictive current control method was implemented, confirming the observations made in simulations. The strategy succeeded in maintaining voltage balance in the DC link and reducing the switching frequency. Working at the same switching frequency, the presented method achieved better reference tracking than the carrier-based method.

III. SIMULATIONS

When the predictive control is implemented experimentally, the same code is rewritten in C language with alpha and beta currents calculated separately. Results using the control algorithm implemented in MATLAB/Simulink are shown next, considering (6) for load current prediction and (9) for back-emf estimation. The system parameters $V_{dc} = 800V$, $L = 10mH$, $R = 0.02\Omega$, and $e = 31V_{peak}$ have been considered for simulations. Current and voltage in one phase of the load are shown in Fig. 4 for a sampling time $T_s = 10\mu F$. There is no steady state error in the current but there is a noticeable ripple. This ripple is reduced considerably when a smaller sampling time is used at sampling time $T_s = 3\mu F$. However, by reducing the sampling time, the switching frequency is increased.

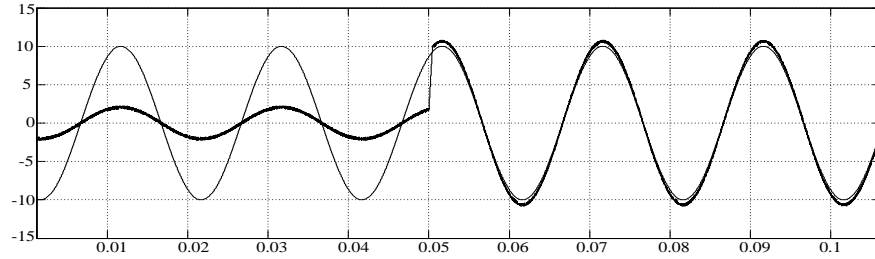


Figure 3. Phase voltage and phase current

However, the proposed method requires a greater sampling frequency or data acquisition frequency. It is important to mention that the sampling instant is always located at a fixed position within the sampling period, making easy the acquisition of measurement data, and avoiding problems with switching the power devices. The dSPACE system used to obtain the results had no problem running the algorithm at the sampling time selected

$T_s = 100\mu s$. In fact, it took only $52\mu s$ to execute the entire algorithm, including voltage balance and reduction of the switching frequency. The algorithm was also implemented on a DSP from Texas Instruments, TMS320F2812, using the same sampling frequency and achieving similar results in terms of processing times.

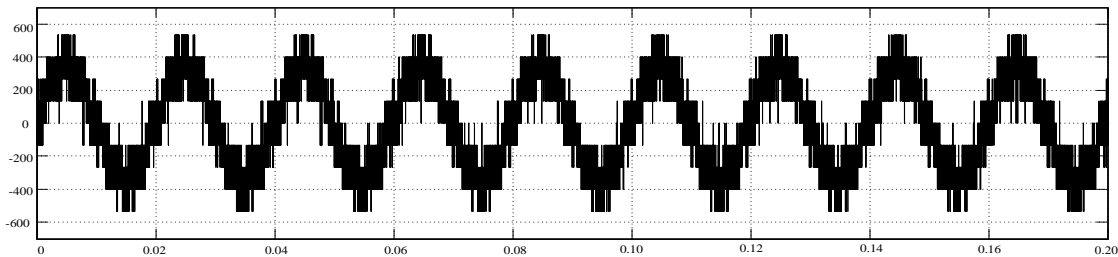


Figure 4. Output voltage of phase A

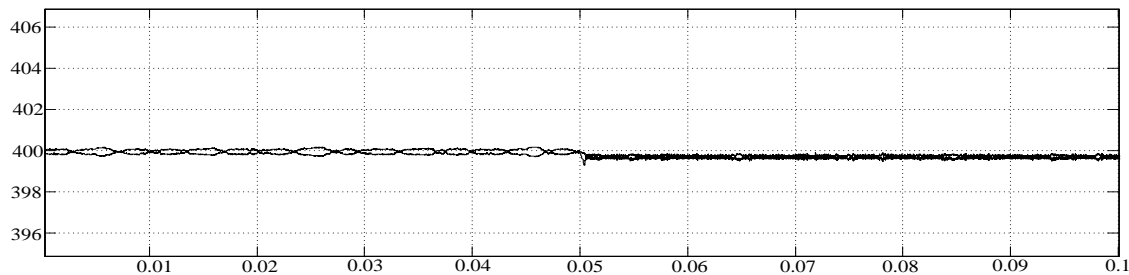


Figure 5. Capacitor voltage of C1,C2

IV. CONCLUSIONS

The predictive current control method presented does not require any kind of linear controller or modulation technique. It effectively controls the load current and compares well to established control methods like PWM, achieves a comparable dynamic response and reference tracking, and works at lower switching frequencies. One of the remarkable aspects of the method is the use of costs assigned to each objective to achieve reference tracking, balance in the DC link, and a reduction in the switching frequency. The simplicity of the theory makes it easy to understand and implement. The strategy allows the designer to adjust the λ parameters to fit the requirements in terms of switching frequency, voltage balance, and reference tracking. A systematic way to determine the weighting factors is a challenge for future work.

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