시지연과 SVPWM 영향이 고려된 새로운 제어 모델에 의한 3상 전압원 PWM 컨버터의 전류 제어

민 동기 *, 안 성찬 **, 현 동석 **
* : 삼성종합기술원, ** : 한양대학교 전기공학과

Current Control of a 3φ PWM Converter Based on a New Control Model with a Delay and SVPWM effects

Dong-Ki Min*, Sung-Chan Ahn**, and Dong-Seok Hyun**
* : Samsung Advanced Institute of Technology, ** : Dept. of Electrical Eng., Hanyang University

Abstract – In design of a digital current controller for a 3φ voltage-source (VS) PWM converter, its conventional model, i.e., stationary or synchronous reference frame model, is used in obtaining its discretized version. It introduces, however, inherent errors since the following practical problems are not taken into consideration: the characteristics of the space vector-based pulsewidth modulation (SVPWM) and the time delays in the process of sampling and computation. In this paper, the new hybrid reference frame model of the 3φ VS PWM converter is proposed considering these problems. In addition, the direct digital current controller based on this model is designed without any prediction or extrapolation algorithm to compensate the time delay. So the control algorithm is made very simple. The validity of the proposed algorithm is proved by the computer simulation results.

1. INTRODUCTION

The advantages in using 3φ VS PWM converters in high performance AC motor drive system are as follows [1][2]:

- The phase of the input current with respect to the input voltage can be adjustable, which means that the input power factor is controllable.
- The DC link voltage can be regulated fast and stable against the variation in the load.
- The excess power in the load side can be regenerated into the input side.

To take these advantages fully, both the sufficient analysis of the given system and the accurate design of its controller are required. In this paper, two subjects that are usually overlooked in modeling of a 3φ VS PWM converter and designing its current controller will be discussed: characteristics of a pulsewidth modulated signal based on space vector theory and time delay between sampling the signals and manipulating the control output. In a PWM VS inverter-fed drive or converter system, the pulsewidth-modulated voltage has been approximated to only its fundamental component, which may cause some significant errors as the switching frequency decreases. In digital control system, there always is a time delay since A/D and D/A conversions and computations take time. In one of the recent papers, O. Kukrer [3] discussed the discrete-time current controller of a 3φ VS PWM inverter having a time delay between sampling and the application of the inverter voltage to the system and proposed the modified control law using a prediction algorithm to eliminate the undesirable effects of the time delay. However, the time delay must be modeled to the system and the controller must be simpler for real time applications. In this paper, a new mathematical model of the 3φ VS PWM converter is proposed based on the analysis of a pulsewidth-modulated voltage based on space vector theory and a time delay due to the process of sampling and computing. Then a direct digital current controller are designed based on this new model having no steady-state error. The validity of the proposed control algorithm is proved by the computer simulation results.

2. MODELING OF A 3φ VS PWM CONVERTER AND ITS CONVENTIONAL MODELS

In this section, a 3φ VS PWM converter system will be mathematically modeled in the time domain, then its conventional models based on the space vector theory are reviewed. The dynamic equations of a 3φ VS PWM converter can be represented in the stationary reference frame as

$$\dot{\mathbf{i}}_{\alpha\beta}(t) = \mathbf{A}_s\mathbf{i}_{\alpha\beta}(t) + \mathbf{B}\mathbf{v}_{\alpha\beta}(t) - \mathbf{B}\mathbf{e}_{\alpha\beta}(t) \qquad (1)$$

where $\mathbf{A}_s = -\frac{R}{L}\mathbf{I}_2$, $\mathbf{B} = -\frac{1}{L}\mathbf{I}_2$, and $\mathbf{I}_2$ is a 2×2 identity matrix.

The variables represented in the stationary reference frame can be transformed into the variables in the synchronous reference frame.

$$\mathbf{x}_{dq} = \mathbf{C}^{-1}(\theta)\mathbf{x}_{\alpha\beta} \qquad (2)$$

where $C(\theta) = \begin{bmatrix} \cos\theta & -\sin\theta \\ \sin\theta & \cos\theta \end{bmatrix}$ and θ is the angle between the direct axes of the two reference frames.

Using (2), the dynamic equation of (1) can be rewritten as

$$\dot{\mathbf{i}}_{dq}(t) = \mathbf{A}_e\mathbf{i}_{dq}(t) + \mathbf{B}\mathbf{v}_{dq}(t) - \mathbf{B}\mathbf{e}_{dq}(t) \qquad (3)$$

where $\mathbf{A}_e = \begin{bmatrix} -R/L & \omega \\ -\omega & -R/L \end{bmatrix}$ and ω is the angular speed of the source voltage vector.

3. HYBRID REFERENCE FRAME MODEL

In the design of a direct digital controller [4], the continuous-time model is discretized and simplified with the assumption of a constant control input during a control period T. In the 3φ VS PWM converter model in the stationary or synchronous reference frame, however, this assumption is not satisfied. The sinusoidal source voltage vector in the stationary reference frame in (1), $\mathbf{e}_{\alpha\beta}$, is not constant during the current control period T but the source voltage vector in the synchronous reference frame, $\mathbf{e}_{dq}$, in (3) is constant. The converter voltage is

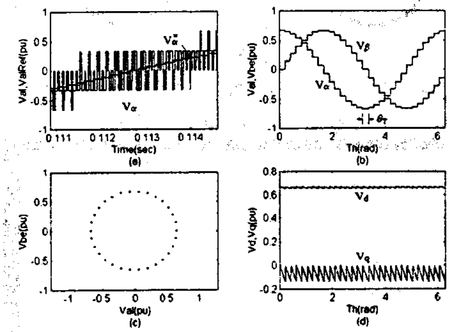

Fig. 1. Converter voltage modulated by SVPWM (normalized by DC link voltage) (a) real and reference voltage waveforms (T=200μs) (b) in the stationary reference frame (θ_T=2π/30) (c) in the voltage space vector plane (d) in the synchronous reference frame

synthesized by its two adjacent switching vectors, $\mathbf{V}_a$ and $\mathbf{V}_b$ to be equal to a given voltage reference vector $\mathbf{V}^*_{\alpha\beta}$ in average of a switching period T [6]: Therefore, the resultant voltage could be equal to the reference voltage vector that is piecewise constant during T as shown in Fig. 1(a). Then, direct and quadrature components of the converter voltage normalized by the DC link voltage are piecewise constant in the stationary reference frame as shown in Fig. 1(b). Its trajectory in the voltage space vector plane is shown in Fig. 1(c). In the synchronous reference frame, direct and quadrature components of the converter voltage vector are not piecewise constant as shown in Fig. 1(d). Also the delay in the process of sampling and computation, which cannot be be avoidable in digital control system, is not considered in these models.

Now the new model considering the characteristics of the signals and a time delay will be derived. The stationary reference frame model with one unit of the time delay in applying the converter voltage can be expressed as

$$\dot{\mathbf{i}}_{\alpha\beta}(t)=\mathbf{A}_s\mathbf{i}_{\alpha\beta}(t)+\mathbf{B}\mathbf{v}_{\alpha\beta}(t-T)-\mathbf{B}\mathbf{e}_{\alpha\beta}(t) \quad (4)$$

Then using (2), (4) can be transformed into

$$\dot{\mathbf{i}}_{dq}(t)=\mathbf{A}_c\mathbf{i}_{dq}(t)+\mathbf{B}\mathbf{C}^{-1}(\theta)\mathbf{v}_{\alpha\beta}(t-T)-\mathbf{B}\mathbf{e}_{dq}(t) \quad (5)$$

We defined this new continuous-time model as a hybrid reference frame model because both frames are referred. Also, note that all signals are piecewise constant during T so that this model can be simply discretized without any errors.

4. DIRECT DIGITAL CURRENT CONTROLLER DESIGN

From the complete solution of (5), the current vector at $k+1$ step can be obtained as

$$\begin{aligned}\mathbf{i}_{dq}(kT+T)&=e^{\mathbf{A}_cT}\mathbf{i}_{dq}(kT)\\&+\left(\int_0^T e^{\mathbf{A}_c\tau}\mathbf{C}^{-1}(\omega(kT+T-\tau))\mathbf{B}d\tau\right)\mathbf{v}_{\alpha\beta}(kT-T)\\&-\left(\int_0^T e^{\mathbf{A}_c\tau}\mathbf{B}d\tau\right)\mathbf{e}_{dq}(kT)\\&=\mathbf{A}_{cd}\mathbf{i}_{dq}(kT)+\hat{\mathbf{B}}_d\mathbf{v}_{dq}(kT-T)-\mathbf{B}_d\mathbf{e}_{dq}(kT)\end{aligned} \quad (6)$$

where $\mathbf{A}_{cd}=e^{\mathbf{A}_cT}$, $\hat{\mathbf{B}}_d=\left(\int_0^T e^{\mathbf{A}_c\tau}\mathbf{C}(\omega\tau)\mathbf{B}d\tau\right)\mathbf{C}^{-1}(2\omega T)$

$\mathbf{B}_d=\int_0^T e^{\mathbf{A}_c\tau}\mathbf{B}d\tau$.

The matrix $\hat{\mathbf{B}}_d$ can be approximated as

$$\hat{\mathbf{B}}_d\approx\left(\int_0^T e^{\mathbf{A}_c\tau}\mathbf{B}d\tau\right)\mathbf{C}\left(\frac{\omega T}{2}\right)\mathbf{C}^{-1}(2\omega T)=\mathbf{B}_d\mathbf{C}^{-1}\left(\frac{3\omega T}{2}\right) \quad (7)$$

The structure of the digital current controller is given as

$$\begin{aligned}\mathbf{v}^*_{dq}(kT)&=\mathbf{L}_1\mathbf{i}_{dq}(kT)+\mathbf{L}_2\mathbf{i}_{dq}(kT-T)\\&+\mathbf{M}_1\mathbf{i}^*_{dq}(kT)+\mathbf{N}_1\mathbf{e}_{dq}(kT)\end{aligned} \quad (8)$$

where $\mathbf{L}_1$, $\mathbf{L}_2$, $\mathbf{M}_1$, and $\mathbf{N}_1$ are 2x2 gain matrices to be determined.

Substitution of (8) in (6) and defining the matrix $\hat{\mathbf{X}}=\hat{\mathbf{B}}_d\mathbf{X}$, $\mathbf{X}\in\{\mathbf{I},\mathbf{V},\mathbf{E}\}$ result in

$$\begin{aligned}\mathbf{i}_{dq}(kT+T)&=\mathbf{A}_{cd}\mathbf{i}_{dq}(kT)+\hat{\mathbf{L}}_1\mathbf{i}_{dq}(kT-T)\\&+\hat{\mathbf{L}}_2\mathbf{i}_{dq}(kT-2T)+\hat{\mathbf{M}}_1\mathbf{i}^*_{dq}(kT-T)\\&+\hat{\mathbf{N}}_1\mathbf{e}_{dq}(kT-T)-\mathbf{B}_d\mathbf{e}_{dq}(kT)\end{aligned} \quad (9)$$

If the current vector, $\mathbf{i}_{dq}$, is convergent to its reference vector, $\mathbf{I}^*_{dq}$, then the following steady state equation can be obtained as

$$\left(\mathbf{I}_2-\mathbf{A}_{cd}-\hat{\mathbf{L}}_1-\hat{\mathbf{L}}_2-\hat{\mathbf{M}}_1\right)\mathbf{I}^*_{dq}=\left(\hat{\mathbf{N}}_1-\mathbf{B}_d\right)\mathbf{e}_{dq} \quad (10)$$

Two conditions for no steady-state errors in the input current control are

$$\hat{\mathbf{M}}_1=\mathbf{I}_2-\mathbf{A}_{cd}-\hat{\mathbf{L}}_1-\hat{\mathbf{L}}_2, \qquad \hat{\mathbf{N}}_1=\mathbf{B}_d \quad (11)$$

Since the source voltage is not a function of time, (11) ensures that the source voltage can not affect current control.

To derive the decoupling condition, the transfer function matrix, $\mathbf{T}(z)$, is necessary and it can be obtained from the z-transformation of (9) as

$$\mathbf{I}_{dq}(z)=\mathbf{T}(z)\mathbf{I}^*_{dq}(z)=\mathbf{P}^{-1}(z)\hat{\mathbf{M}}_1\mathbf{I}^*_{dq}(z) \quad (12)$$

where $\mathbf{P}(z)=z^2\mathbf{I}_2-z\mathbf{A}_{cd}-\hat{\mathbf{L}}_1-z^{-1}\hat{\mathbf{L}}_2$.

The transfer function matrix is given as

$$\begin{aligned}\mathbf{T}(z)&=\mathbf{P}^{-1}(z)\hat{\mathbf{M}}_1\\&=\frac{1}{\det\mathbf{P}(z)}\begin{bmatrix}p_{22}(z)&-p_{12}(z)\\-p_{21}(z)&p_{11}(z)\end{bmatrix}\begin{bmatrix}\hat{m}_{11}&\hat{m}_{12}\\\hat{m}_{21}&\hat{m}_{22}\end{bmatrix}\end{aligned} \quad (13)$$

where p_{ij} and $\hat{m}_{ij}$ are entries of $\mathbf{P}(z)$ and $\hat{\mathbf{M}}_1$ respectively.

The decoupling control, or non-interactive control can be achieved if the off-diagonal terms of the transfer function matrix, $\mathbf{T}(z)$, is null or becomes zero as

$$\hat{m}_{12}p_{22}(z)-\hat{m}_{22}p_{12}(z)=0\ ,\ -\hat{m}_{11}p_{21}(z)+\hat{m}_{21}p_{11}(z)=0 \quad (14)$$

Although there are many ways to determine the parameters in (14) as it is convenient that the off-diagonal terms of $\hat{\mathbf{M}}_1$ set to zero and the off-diagonal terms of $\mathbf{P}(z)$ shrink to zero by setting the off-diagonal terms of $\hat{\mathbf{L}}_1$ and $\hat{\mathbf{L}}_2$.

$$\hat{m}_{12}=\hat{m}_{21}=0 \quad (15)$$

$$\hat{l}_{1,12}=-\hat{l}_{1,21}=-2a_{12}, \quad \hat{l}_{2,12}=-\hat{l}_{2,12}=a_{12} \quad (16)$$

where a_{ij} is an entry of $\mathbf{A}_{cd}$.

5. SIMULATION RESULTS

To show the validity of the proposed direct digital current controller for the 3ϕ VS PWM converter with a time delay, the simulation has been done. The system parameters are given in Table 1.First, the simulated waveforms in the steady state for the predictive current controller designed in the synchronous reference are shown in Fig. 2. The space current vector is shown in Fig. 2(a) and the input voltage and the input current are shown in Fig. 2(b). The input current is not sinousoidally controlled with a large amount of lower-order harmonics. The direct and quadrature components of the input current in the synchronous reference frame are shown in Fig. 2(c). This unstability results mainly from the time delay.

On the other hand, in case of the direct digital current controller, the input current is sinusoidally controlled without any steady-state errors. Fig. 3(a) shows that the trajectory of the space current vector is circle. Fig. 3(b) shows that the input current is sinusoidally controlled with unit power factor. Fig. 3(c) shows that no steady state errors is

TABLE 1
SYSTEM PARAMETERS

Parameters	Value
Input Inductance	1.2mH
DC Link Capacitance	2300μF
Input Voltage(Line-to-Line, rms)	220V
Load Resistance	18Ω
Current Control Period	200μsec
DC Link Voltage Control Period	2msec
DC Link Voltage	400V

produced in the input current in the synchronous reference frame.

In the transient state, the direct digital current controller gives a good dynamic response as shown in Fig. 4. The load is changed abruptly from no load to 8.9kW.

6. CONCLUSION

In this paper, the current control for 3ϕ VS PWM converter with a time delay has been dealt. After investigating the pulsewidth-modulated signals, the new control model, hybrid reference frame model, is proposed considering a time delay. From this continuous-time model, the discrete-time model can be easily obtained since the converter voltage and the source voltage are represented in the stationary and synchronous reference frames respectively. The gain matrices in the direct digital current controller are determined from the no steady-state error condition and decoupling control condition. This current controller has a simple structure and its validity is proved by the computer simulation results.

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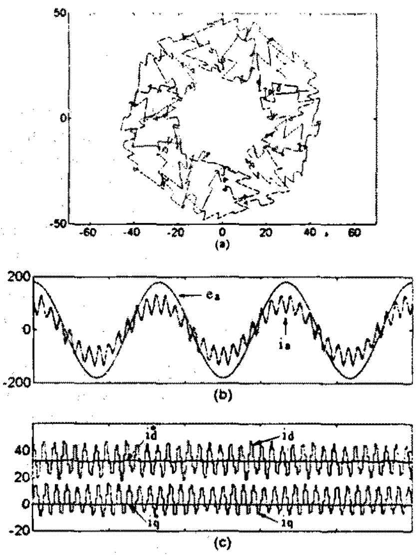

Fig. 2. Simulation waveforms in steady state for the predictive current controller in the synchronous reference frame without consideration of SVPWM effects and a time delay.
(a) the space current vector (b) the input current(×3) and input voltage(10msec/div.) (c) the direct and quadrature components of the input current(10msec/div.)

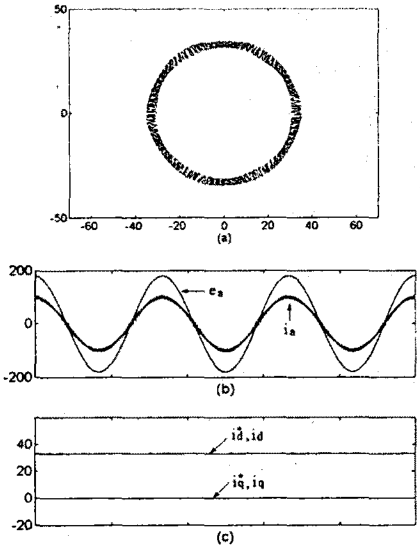

Fig. 3. Simulation waveforms in steady state for the direct digital current controller.
(a) The space current vector (b) the input current (×3) and the input voltage (10msec/div.) (c) The direct and quadrature components of the input current (10msec/div.)

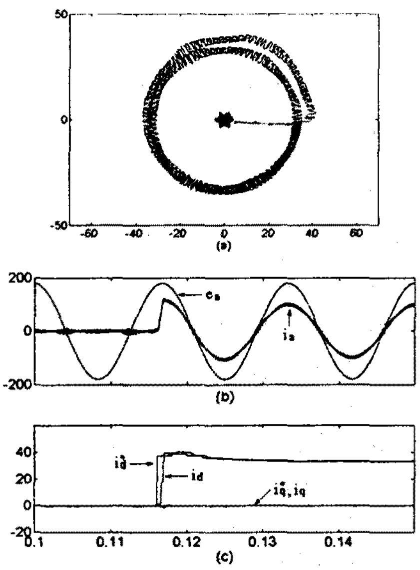

Fig. 4. Simulation waveforms in transient state from no load to 8.9kW for the direct digital current controller.
(a) The space current vector (b) the input current (×3) and the Input voltage (10msec/div.) (c) the direct and quadrature components of the input current(10msec/div.)