Current-Programmed Control of Three Phase PWM AC-AC Boost Converter

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ABSTRACT

In this paper, a new scheme of current programmed control for three phase PWM AC-AC converter is presented. By considering only the magnitude components, a similar scheme in the DC-DC converter can be extended to the three phase PWM AC-AC converter. The proposed current programmed control will be well adopted into various converter topologies though three phase PWM AC-AC boost converter is treated as an example. The converter analysis is carried out by applying the vector DQ transformation to obtain physical insight into the converter operation. Finally, the experiment result shows the validity of the proposed scheme.

1. INTRODUCTION

Current programmed control has been widely used in switching converters. Current programmed DC-DC converter has been dealt with in many papers [1]-[2]. However, the current programmed control for AC-AC switching converters has rarely been investigated.

This paper proposes a practical current programmed control scheme for AC-AC switching converters. By considering only the magnitude components, a similar scheme as in the DC-DC converter can be extended to AC-AC case. An error signal is generated by comparing the sensed output voltage magnitude with a reference voltage value. Then the error signal may be processed by a Proportional-Integral compensator to generate the current command. The sensed inductor current magnitude component is compared with this current reference command to make the current error signal. Then this current error signal is processed by a Proportional-Integral-Differential controller to generate the duty control signal of the switches, so that the actual inductor current magnitude follows the current command through the inner current control loop.

The proposed current programmed control scheme has several advantages over conventional duty cycle control method. By processing the switching control signal with the current programmed controller, the small-signal control-to-output transfer function contains one less pole than that of duty cycle control, so the system order will be reduced. Such reduced system order may lead to easier controller design. Also, as the current programmed controller makes use of the sensed the inductor current information during normal converter operation, transistor failures due to excessive switch current can be prevented simply by setting the current reference value. Hence, inherent current protection is realized.

The proposed current programmed control will be well adopted into various converter topologies such as buck, boost, buck-boost, etc. The three phase PWM AC-AC boost converter is treated as an example in this paper [3]. Basic converter analysis is carried out by performing vector DQ transformation [4], thus some important characteristic relations are obtained for control purpose.

Finally, the experiment result supports the design and analysis.

2. SYSTEM DESCRIPTION

2.1 Operating principle

Fig. 1 illustrates the proposed current programmed control diagram of the three phase PWM AC-AC converter. As seen in Fig. 1, the three phase AC output voltages will be sensed so as to produce the corresponding

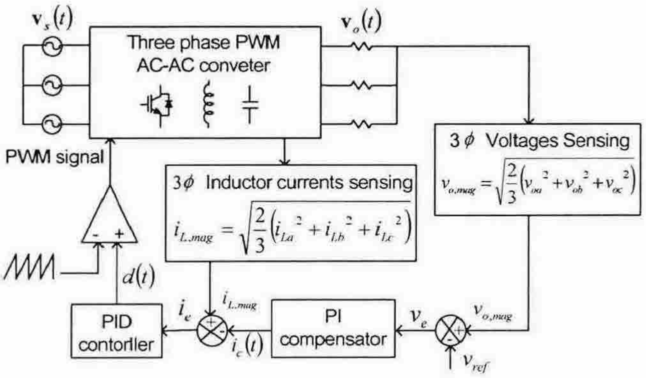

Fig .1 Proposed current programmed PWM AC-AC converter.

output voltage magnitude $v_{o,mag}$ instantly. This voltage magnitude is compared with a reference value to generate the voltage error signal v_e. The error signal is then processed by a Proportional-Integral compensator to form the inductor current magnitude command i_c.The sensed inductor current magnitude component $i_{L,mag}$ is compared with i_c to make the current error signal i_e. Then this current error signal is processed by a Proportional-Integral-Differential controller to update the duty ratio, and the PWM switching signal is made by comparing the duty with a triangular waveform with the switching frequency. Thus the actual inductor current magnitude follows the current command through the inner current control loop.

By processing the switching control signal with the current programmed controller, the inductor current is directly controlled, the behavior is somewhat like the controlled current source, which will consequently make robust output. Moreover, the current magnitude is limited by the current command, thus inherent over-current protection will be realized.

2.2 Converter circuit

Fig. 2 shows the three-phase PWM Boost AC-AC converter. As seen in Fig. 2, the system requires six IGBTs. In Fig. 2, *d* means the duty ratio of the switches where they turn on or off in the way of simultaneous switching. Therefore, it should be noted that the system under analysis has only one control variable, *d*.

The source voltages with angular speed, ω are assumed ideal and balanced and are given as follows

$$\mathbf{v}_{sabc} = \begin{bmatrix} v_{sa} \\ v_{sb} \\ v_{sc} \end{bmatrix} = \sqrt{\frac{2}{3}} \cdot V_s \begin{bmatrix} \sin(\omega t) \\ \sin(\omega t - 2\pi/3) \\ \sin(\omega t + 2\pi/3) \end{bmatrix} \tag{1}$$

where V_s is the *rms* line-to-line AC source voltage.

3. CONVERTER ANALYSIS

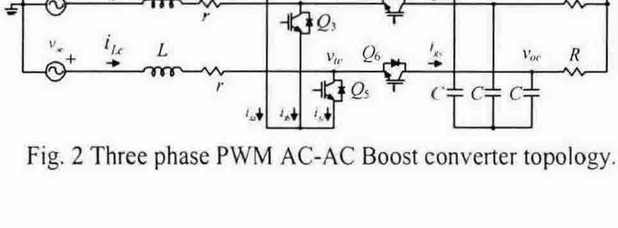

Fig. 2 Three phase PWM AC-AC Boost converter topology.

3.1 Vector DQ transformation

A vector DQ transformation of three phase quantities $\mathbf{x}_{yabc}$ into a complex vector $\mathbf{x}_y$ is defined as

$$\mathbf{x}_y = \mathbf{K}\mathbf{x}_{yabc} = \begin{bmatrix} x_{yq} & x_{yd} & x_{yo} \end{bmatrix}^T \tag{2}$$

$$\mathbf{K} = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos(\omega t) & \cos\left(\omega t - \frac{2\pi}{3}\right) & \cos\left(\omega t + \frac{2\pi}{3}\right) \\ \sin(\omega t) & \sin\left(\omega t - \frac{2\pi}{3}\right) & \sin\left(\omega t + \frac{2\pi}{3}\right) \\ 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \end{bmatrix} \tag{3}$$

$$\mathbf{x}_y = x_{yd} + jx_{yq} \tag{4}$$

where $\mathbf{x}_{yabc} = \begin{bmatrix} x_{ya} & x_{yb} & x_{yc} \end{bmatrix}^T$.

After obtaining the averaged circuit equations and applying the vector DQ transform, one can obtain the converter equations as follows

$$\mathbf{v}_s - \mathbf{v}_t = r\mathbf{i}_L + L \cdot \frac{d}{dt}\mathbf{i}_L + j\omega \mathbf{i}_L \tag{5}$$

$$C \cdot \frac{d}{dt}\mathbf{v}_o + j\omega C\mathbf{v}_o + \frac{1}{R}\mathbf{v}_o = \mathbf{i}_g \tag{6}$$

$$\mathbf{i}_g = (1-d)\mathbf{i}_L \tag{7}$$

$$\mathbf{v}_t = (1-d)\mathbf{v}_o \tag{8}$$

where $\mathbf{v}_s = v_{sd} + jv_{sq} = V_s$.

In (5) and (6), the term $j\omega L(j\omega C)$ does not mean conventional impedance(admittance) but an element that represents the relationship of *d*- and *q*-components. Note that in (5) to (8) every voltage or current value can be regarded as a complex variable which is a function of time.

3.2 Small signal model and transfer function

The small signal model is established by introducing some perturbation to the control variable *d* whereas the input voltage is not perturbed. Then, the circuit variables consist of DC and AC components. The perturbed component is indicated by the diacritical mark '^' of the corresponding variable to distinguish it from the quiescent value denoted by capital letter as follows

$$\mathbf{i}_L = \mathbf{I}_L + \hat{\mathbf{i}}_L\ ,\quad \mathbf{v}_o = \mathbf{V}_o + \hat{\mathbf{v}}_o\ ,\quad d' = D' - \hat{d} \tag{9}$$

where $d' = 1 - d$ and $D' = 1 - D$.

Substituting (9) into (5) to (8), and applying Laplace transform, one can obtain

$$\frac{\hat{\mathbf{v}}_o}{\hat{\mathbf{i}}_L}=\frac{a-jb}{(sD'CR+2D)+j2D'\omega CR} \tag{10}$$

where $a=\omega^2 LCR+D'^2 R-rC+sL$

$b=\omega L+rC^2\omega R+sL\omega CR$.

In (10), an assumption is made that the inductor current magnitude perturbation $\hat{\mathbf{i}}_L$ is identical to the programmed current command perturbation $\hat{\mathbf{i}}_c$ That is $\hat{\mathbf{i}}_L(s)\approx\hat{\mathbf{i}}_c(s)$.This is valid to the extent that the controller is stable, and that the magnitude of the inductor current ripple is sufficiently small.

From (10), it can be found that the small-signal control-to-output transfer function $\hat{\mathbf{v}}_o(s)/\hat{\mathbf{i}}_c(s)$ contains only one pole and thus the system order is reduced.

The magnitude relationship of $\hat{\mathbf{v}}_o(s)/\hat{\mathbf{i}}_c(s)$ in decibels,

$$20\log_{10}\left|\frac{\hat{\mathbf{v}}_o}{\hat{\mathbf{i}}_L}\right|=10\log_{10}\frac{gs^2+hs+k}{ms^2+ns+p} \tag{11}$$

where

$g=L^2\omega^2C^2R^2+L^2 \qquad h=2LrC^3\omega^2R^2+2LCr-2LD'^2 R$

$k=r^2C^4\omega^2R^2+\omega^4L^2C^2R^2+r^2C^2+2\omega^2D'^2 R^2CL-2rCD'^2 R+D'^4 R^2+\omega^2L^2$

$m=D'^2 R^2C^2 \qquad n=4D'^2 RC$

$p=4D'^2+4D'^2 \omega^2C^2R^2$.

It is worth noting that in (11), generally, the numeric value of k and p is 10^7 times greater than that of g and m respectively, which implies that in low frequency operation, it mainly shows proportional gain effect.

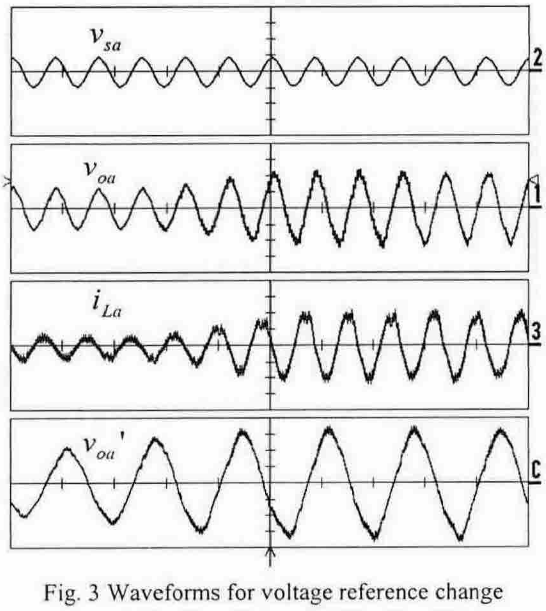

Fig. 3 Waveforms for voltage reference change
v_{sa}(25V/div, 20ms/div); i_{La}(5A/div, 20ms/div);
v_{oa}(25V/div, 20ms/div); v_{oa}'(17V/div, 10ms/div).

4. EXPERIMENT

To confirm the validity of the design and analysis of the proposed scheme, an experimental setup was made and TMS320F2812 DSP was applied. The circuit parameters are

v_s=30 V, f=60 Hz, L=1.2 mH,
r=0.01 Ω, C=20 μF, R=15 Ω.

Also, the switching frequency is 6kHz.

Fig. 3 shows the experiment waveforms of the input voltage, output voltage and inductor current when voltage reference is abruptly changed from 40V to 60V. It can be seen from Fig. 3 that the output well follows the reference voltage with the transient time about one cycle. In Fig. 3, v_{oa}' is the extended waveform of v_{oa}.

5. CONCLUSION

This paper deals with the current programmed control of three phase PWM AC-AC boost converter. A similar scheme of DC-DC converter in current programmed mode is extended to AC-AC case by considering the magnitude components. Circuit analysis is carried out by using vector DQ transform method. Small signal model and transfer function are investigated. The experiment results support the validity of the design and analysis.

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