Harmonic Currents Estimation and Compensation Method for Current Control System of IPMSM in Overmodulation Range

Smith Lerdudomsak, Mitsuhiro Kadota, Shinji Doki and Shigeru Okuma

Department of Electrical Engineering and Computer Science, Nagoya University
Chikusa, Nagoya City, Aichi Prefecture, 464-8601, Japan

Abstract— **For high performance and wide speed range control of an IPMSM, using the overmodulation range of an inverter is one of the effective methods. However, to use this overmodulation range with the closed loop currents control system, there are many problems occur and the main problem is caused by the harmonic currents generated from harmonic voltages in the inverter output voltages. Therefore, in the ususal control methods of an IPMSM in the overmodulation range, these problems are avoided by not using the closed loop currents control or by using the closed loop currents control with the filters in the feedback part. With these methods, the problems from harmonic currents can be solved, but the other new problems also occur too. The method of harmonic currents compensation is also proposed and this method seems to be more effective than the others. However, in this method, the proposed harmonic currents estimation method is quite difficult in practical use, because the real value of the inverter output voltages are used, then the carrier frequency components will cause an aliasing problem in the estimation part and this problem will become more serious when the carrier frequency is high.**

To solve the above problem, in this paper, we propose the harmonic currents estimation and compensation method based on the inverter harmonic voltages model. In our proposed method, we use only the inverter commanded voltages in the harmonic voltages estimation, then this method is more easy for the practical use. Moreover, by using the inverter harmonic voltages model, the data of both amplitude and phase angle of harmonic voltages are available, then we can estimate the harmonic currents by using only simple phasor method compared with the complex dynamic model method in the former. The experimental results are shown to confirm our proposed estimation and compensation methods.

Keywords— **IPMSM, Overmodulation range, Current control, Harmonic compensation**

I. Introduction

Because of many advantages of an interior permanent magnet synchronous motor (IPMSM) such as high efficiency, high power density and robust structure, the applications of IPMSM become much wider in many fields. Especially, when an IPMSM is used for hybrid electric vehicle or railway vehicle traction, to control an IPMSM for high response in high speed range is necessary[1]∼[6]. However, for corresponding such need, the output voltages of an inverter that supplied to an IPMSM must also be high too. But in the real situation, the output voltages of an inverter are limited by the DC link voltage, and to increase the DC link voltage will cause the other problems such as cost, size and insulation of an inverter. One way to increase the output voltages of an inverter without any increasing of the DC link voltage is by operating an inverter in the overmodulation range.

The waveform and power spectrum of inverter output phase voltage in the overmodulation range are shown in fig.1. From this figure, we can find that the fundamental component (dashed line) of output voltage in the overmodulation range can be increased considerably compared with the usual linear range (theoretically, as high as 127%)[7]. The torque-speed curve of an IPMSM in the usual linear range and the overmodulation range are also compared in Fig.2.

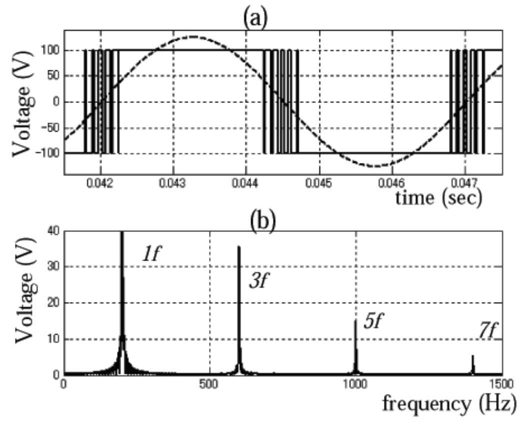

Fig. 1. (a)Wavefrom of voltage in the overmodulation range (b)Power spectrum of volatge in the overmodulation range

However, there are two main differences between the usual linear range and the overmodulation range of an inverter. First, the nonlinearity of amplitude relation between the commanded voltages and the output voltages of an inverter as shown in fig.3, here the definition of the word modulation index (m) is shown in (1) when V_m is voltage amplitude and V_{DC} represents the DC link voltage of an inverter. The modulation index ranges from 0 to 1.0 in the linear range, and 1.0 to 1.27 in the overmodulation range.

$$m = \frac{V_m}{V_{DC}/2} \tag{1}$$

Second, there are large amounts of the harmonic components generated in the output voltages as shown before

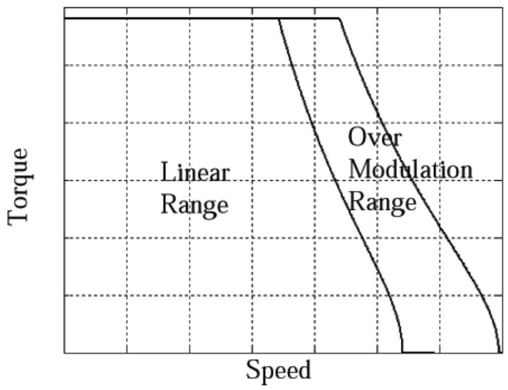

Fig. 2. Torque-Speed curve of an IPMSM when operating in linear range and overmodulation range

in fig.1(b). From the above differences, we can consider the model of an inverter in the overmodulation range to be equivalent with the nonlinear amplifier plus the harmonics disturbance generator compared with the simple linear amplifier in the case of the linear range as shown in fig.4. Moreover, the above differences will cause many problems when we use the usual linear currents control system with an inverter in the overmodulation range as will be discussed in the next section.

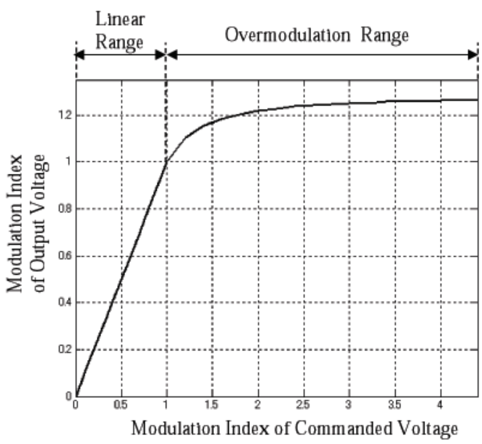

Fig. 3. The nonlinear amplitude relation between the commanded voltage and the output voltage of an inverter

II. Review of IPMSM control methods in overmodulation range

First, the nonlinearity of amplitude relation in the overmodualtion range make the output voltages of an inverter cannot be correctly controlled, then the current responses will be unstable if there is not any proper compensation. To solve this problem, the method of nonlinear amplitude compensation that uses the inverse function of the data in fig.3 is proposed[7]. This method is effective for the opened loop control system such as V/F method. However, for the closed loop current control system, using only this compensation is still not enough.

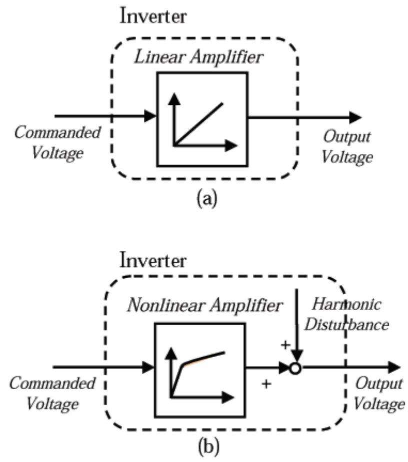

Fig. 4. Model of PWM inverter when operating in (a)linear range and (b)overmodulation range

Because in the closed loop currents control system, the harmonic voltages as shown in fig.1(b) generate the harmonic currents and these harmonic currents are fed back to the currents control loop and will cause the other problems as the belowing.

(1) The controllers in the currents control loop (such as PI controllers) try to suppress the fed back harmonic currents, however, because the output voltages of an inverter in the overmodulation range are fundamental components (same as commanded voltages) plus harmonic components, everytime that controllers try to suppress the harmonic currents, they also create the new harmonic currents too. As a result,the working of the controllers cannot suppress the harmonic currents, moreover the fundamental components of currents also deterioriate too, and this problem is more serious when the gains of controllers are high [9][10].

(2) The harmonic components will be contained in the inverter commanded voltages, and after the modulation process, because of the interaction between these harmonic components and the carrier signal, the fundamental components of inverter output voltages will be different from the commanded voltages, then the responses of currents will be much different from the desired responses [8].

For solving the above problems,to suppress the harmonic currents,by using the extra controller design method such as feedforward compensation technique in two degrees of freedom control may seem to be reasonable [11]. However, in this overmodulation range, the output voltages of an inverter are already near the limitted voltage settled by the DC link voltage, then there is not any enough surplus voltages for suppressing the harmonic currents.

Therefore, using the controllers to control only fundametal components and finding the ways that harmonic currents will not annoy the controllers is the best concept for controlling an IPMSM in the overmodulation range.

Based on the above design concept, there are many methods proposed to control an IPMSM in the overmodulation range and to avoid the problems from the harmonic currents at the same time. These proposed methods can be mainly divided into two styles, which are the methods that do not use the currents control loop, and the methods that use the currents control loop with the elimination of harmonic currents in the feedback part. The details of each proposed method are summarized as the following.

A. The methods that do not use currents control loop

A.1 Voltage Phase Compensation Method

This control method is shown in fig.5 [5][6]. In this method,the monotone relation between the output torque of IPMSM and the phase angles of inverter output voltages is achieved by restricting the parameters and driving range of IPMSM into many constrainted conditions. With using this monotone relation, the amplitudes of inverter output voltages are constantly set to 1 pulse condition ($= (4/\pi)V_{DC}/2$), and the output torque of IPMSM is controlled by varying the phase angle Ψ of the inverter output voltages.

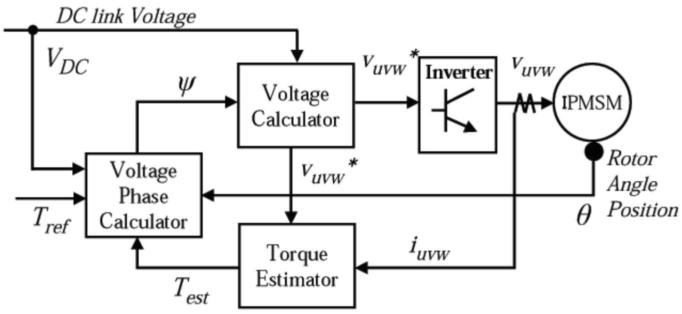

Fig. 5. Voltage Phase Compensation Method

By using this method, the problems from harmonic currents will not occur. However, there are other problems occur such as the constraints of IPMSM parameters and driving range that can use this method, the transition between this control mode and the normal mode in the linear range, and also the exceeding of currents amplitudes over the current limit at low speed.

A.2 Cascade Control Method

This control method is shown in fig.6 [1]~[4]. The concept of this method is that, the q-axis current is used to control the output torque,while the d-axis current is used to control the amplitudes of inverter commanded voltages liked in the field weakening control method.

However,in the currents control part, the d-q axis commanded voltages are calculated by using only the motor model, which neglect the derivative parts, in the feedforward fashion. Therefore, in this method, the performance of currents control part depends so much on the accuracy of the motor parameters value, and the problem of mode transition also occurs as the above method.

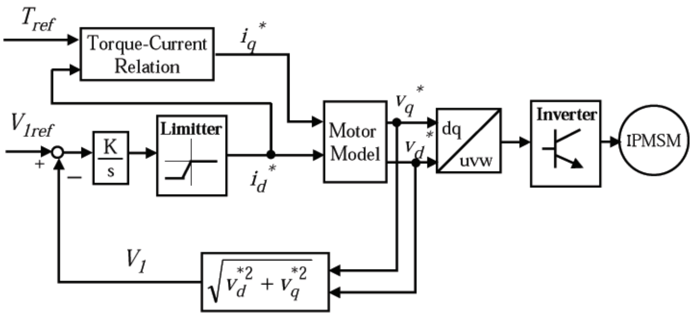

Fig. 6. Cascade Control Method

B. The methods that use currents control loop with elimination of harmonic currents in feedback part

B.1 Currents Control System with LPF Method

The currents control system of this method is shown in fig.7 [6], and we can understand easily that the low pass filters in the system are used to eliminate the harmonic currents caused by an inverter. By using this method,the problems from harmonic currents are removed and we can use the simple and usual currents control system even in the overmodulation range.

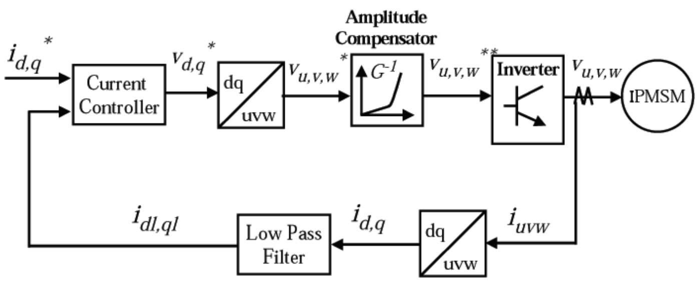

Fig. 7. Currents Control with Low Pass Filter

However, it can be cleary found that the currents transient responses in this method become worse, and how to design the low pass filters for properly removing the harmonic currents in all speed range is also the demerit of this method.

B.2 Currents Control System with Harmonic Currents Compensation Method

The current control system of this method is shown in fig.8 [9][10]. In this method, the harmonic currents are estimated and compensated from the real currents before being fed back to the currents control system. Same as the above method that use LPF, in this method, the problems from harmonic currents are removed and we can use the usual currents controllers in the overmodulation without any problem. Moreover, by using this method, the responses of currents in both steady state and transient state will be not deteriorated.

However,the difficulty of this method is that how to estimate the harmonic currents correctly. The harmonic cur-

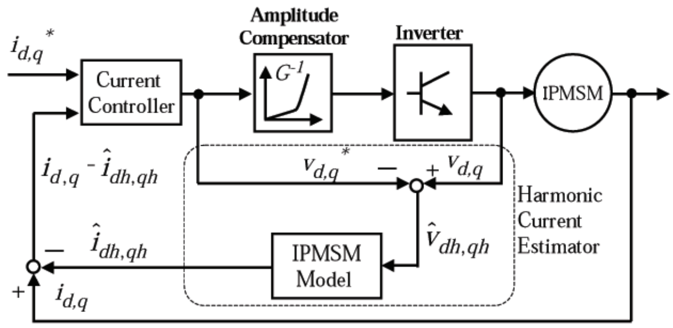

Fig. 8. Current control system with harmonic currents compensation

rents estimation and compensation methods are discribed as the following.

From fig.8, first the harmonic voltages $\hat{v}_{dh,qh}$ are estimated by the difference between the inverter commanded voltages $v^*_{d,q}$ and the output voltages $v_{d,q}$. Second,using the IPMSM model with these estimated harmonic voltages $\hat{v}_{dh,qh}$, the harmonic currents $\hat{i}_{dh,qh}$ can be estimated. Next, delete the estimated harmonic currents $\hat{i}_{dh,qh}$ from the real IPMSM currents $i_{d,q}$, only the fundametal currents are fed back to the currents control loop.

As the above discribing , this method seem to be most effective compared with the others. However, the main disadvantage of this method is the use of real inverter output voltages $v_{d,q}$ in the estimation of harmonic voltages $\hat{v}_{dh,qh}$. Because, in the inverter output voltages $v_{d,q}$, they also contain the carrier frequency components of PWM inverter. Then, for preventing the aliasing problem, the harmonic currents estimation period must be shorter than the carrier period of an inverter (at least 10 times for moderate accuracy). This restriction makes it is difficult to use this method in the real situation, especially when the carrier frequency is high.

By considering all the above proposed method, we found that the last method that uses the harmonic currents compensation seems to be most effective which can be summarized again as the belowing reasons.

(1)The mode transition method between linear and overmodulation range is not necessary (seamless control), and only one controller design technique can be used in both ranges.

(2)Because the output torque of motor directly depends on the currents, using the currents control loop is the effective way of torque control.

(3)When there are currents control loops, the overcurrent problem will not occur.

(4)Compared with the method that use LPF, in this method both the steady state and transient state performances are good.

However, as mentioned above, the method of harmonic currents estimation is need to be improved. The improved method proposed by this paper will be detailly discribed in the next section

III. Proposed Harmonic Currents Compensation Method and Harmonic Voltages Model

From fig.1(b), in the overmodulation range, the main harmonic components in the three phases output voltages of an inverter are the $3^{rd}, 5^{th}$and7^{th} harmonic components. However, the 3^{rd} harmonic component does not appear in the line voltages, then the main harmonic components in the overmodulation range are 5^{th}and7^{th} components. After transformation from the three phases system to the d-q axis system, these harmonic components become the 6^{th} harmonic components as shown in (2). When v_{d6}, v_{q6} are the 6^{th} harmonic voltages in the d-q axis system, $v_{u5}, v_{v5}, v_{w5}, v_{u7}, v_{v7}, v_{w7}$ are the 5^{th}and7^{th} harmonic voltages of three phases system respectively and θ_{re} is the rotor angle (rad) of IPMSM in electrical degree.

$$\begin{bmatrix} v_{d6} \\ v_{q6} \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} cos(\theta_{re}) & cos(\theta_{re} - \frac{2\pi}{3}) & cos(\theta_{re} + \frac{2\pi}{3}) \\ -sin(\theta_{re}) & -sin(\theta_{re} - \frac{2\pi}{3}) & -sin(\theta_{re} + \frac{2\pi}{3}) \end{bmatrix} \begin{bmatrix} v_{u5} + v_{u7} \\ v_{v5} + v_{v7} \\ v_{w5} + v_{w7} \end{bmatrix} \quad (2)$$

Then, in this paper, we will pay attention to this 6^{th} harmonic components. The proposed current control system with harmonic currents compensation of this paper is shown in fig.9. In this proposed method, for calculating harmonic voltages $\hat{v}_{d6,q6}$, only the inverter commanded voltages $v^*_{d,q}$ are used, then the aliasing problem will not occur and the period of harmonic currents estimation can be as long as the currents control period that make this method become easy for pratical use.

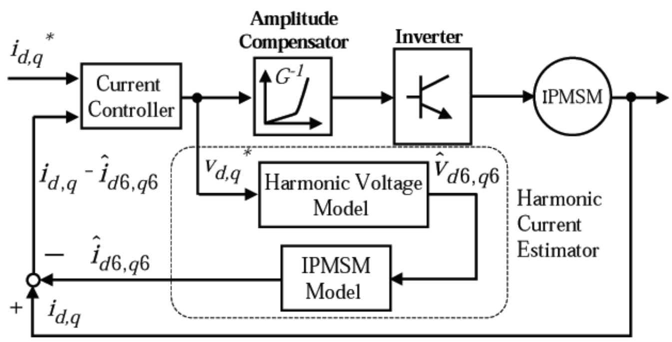

Fig. 9. Proposed currents control system with harmonic currents compensation

However, in this method, knowledge about the relation between the inverter commanded voltages $v^*_{d,q}$ and the harmonic voltages $v_{d6,q6}$, which we call "harmonic voltages model" as shown in fig.10 is necessary. Although this harmonic voltages model depends on the modulation technique of an inverter, the method of finding this model in each modulation technique is same. In this paper we will consider the usual sine PWM (SPWM) technique, however the method that shown below also be appliable in other modulation technique too. For finding the harmonic voltages model, first we define the three phases fundametal voltages and the $5^{th}$and 7^{th} harmonic voltages as in (3)~(5). An

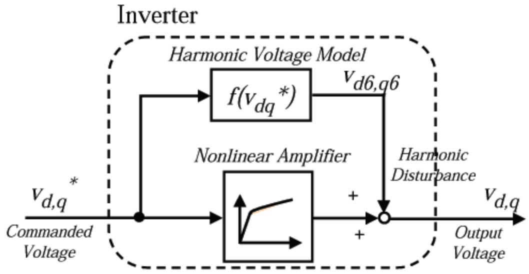

Fig. 10. Inverter model with harmonic voltages model

amplitude V_1 and phase angle ϕ_1 in (3) can be easily calculated from the d-q axis commanded voltages $v^*_{d,q}$ as shown in (6) and (7).

$$\begin{bmatrix} v_{u1} \\ v_{v1} \\ v_{w1} \end{bmatrix} = \begin{bmatrix} V_1 sin(\theta_{re} + \phi_1) \\ V_1 sin(\theta_{re} - \frac{2\pi}{3} + \phi_1) \\ V_1 sin(\theta_{re} + \frac{2\pi}{3} + \phi_1) \end{bmatrix} \tag{3}$$

$$\begin{bmatrix} v_{u5} \\ v_{v5} \\ v_{w5} \end{bmatrix} = \begin{bmatrix} V_5 sin(5\theta_{re} + \phi_5) \\ V_5 sin(5\theta_{re} + \frac{2\pi}{3} + \phi_5) \\ V_5 sin(5\theta_{re} - \frac{2\pi}{3} + \phi_5) \end{bmatrix} \tag{4}$$

$$\begin{bmatrix} v_{u7} \\ v_{v7} \\ v_{w7} \end{bmatrix} = \begin{bmatrix} V_7 sin(7\theta_{re} + \phi_7) \\ V_7 sin(7\theta_{re} - \frac{2\pi}{3} + \phi_7) \\ V_7 sin(7\theta_{re} + \frac{2\pi}{3} + \phi_7) \end{bmatrix} \tag{5}$$

$$V_1 = \sqrt{\frac{2}{3}}\sqrt{v_d^{*2} + v_q^{*2}} \tag{6}$$

$$\phi_1 = tan^{-1}(\frac{v_q^*}{v_d^*}) \tag{7}$$

The amplitudes of harmonic voltages V_5, V_7 shown in equation (4) and (5) depend only on the modulation index of an inverter and by using FFT analysis with the inverter output voltages, the relation between these ampltudes V_5, V_7 and inverter modulation index is shown in fig.11.

For the phase angles of harmonic voltages ϕ_5, ϕ_7, they are also depend on the modulation index of an inverter, however the relation can be mainly divided into 2 cases as shown in (8) and (9). By test results, the relation between the phase angles of harmonic voltages and the modulation index of an inverter can be shown as in fig.12.

$$\phi_5 = 5\phi_1, \phi_7 = 7\phi_1 \tag{8}$$

$$\phi_5 = 5\phi_1 + \pi, \phi_7 = 7\phi_1 + \pi \tag{9}$$

And after substitution of harmonic voltages equations (4) and (5) into (2), the relation between the d-q axis 6^{th} harmonic voltages $\hat{v}_{d6,q6}$ and the amplitudes V_5, V_7 and phase angles ϕ_5, ϕ_7 of the 5^{th} and 7^{th} harmonic voltages can be written in compact form as (10) and (11). By using equation (10) and (11) with the relation shown in fig.11 and fig.12, we can model the relation between the inverter

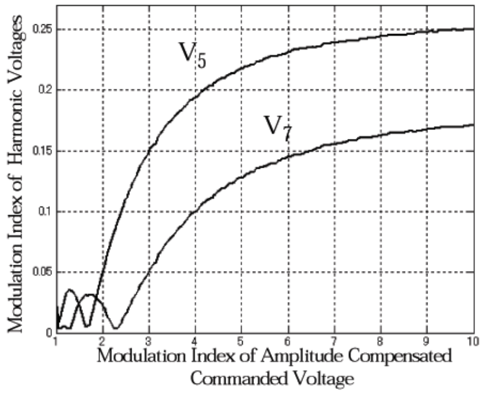

Fig. 11. The relation between the harmonic voltages amplitude and modulation index

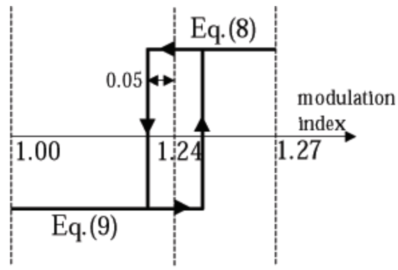

Fig. 12. The relation between the harmonic voltages phase angles and modulation index

commanded voltages and the respective harmonic voltages. Based on this harmonic voltages model, the harmonic currents can be estimated as be discribed in the next section.

$$v_{d6} = \sqrt{\frac{3}{2}}(V_5 cos(6\theta_{re} + \phi_5) + V_7 cos(6\theta_{re} + \phi_7)) \tag{10}$$

$$v_{q6} = \sqrt{\frac{3}{2}}(-V_5 sin(6\theta_{re} + \phi_5) + V_7 sin(6\theta_{re} + \phi_7)) \tag{11}$$

IV. Estimation Method of Harmonic Currents

The calculation method of harmonic currents based on the value of harmonic voltages calculated from the previos section and the harmonic model of IPMSM shown in (12). When R is resistance, L_d, L_q are d-q axis inductances, p is derivative operator and ω_{re} is rotation speed in electrical degree of IPMSM respectively.

$$\begin{bmatrix} v_{d6} \\ v_{q6} \end{bmatrix} = \begin{bmatrix} R + pL_d & -\omega_{re} L_q \\ \omega_{re} L_d & R + pL_q \end{bmatrix} \begin{bmatrix} i_{d6} \\ i_{q6} \end{bmatrix} \tag{12}$$

For adapting equation (12) to digital computation, in [10], the method for calculating harmonic currents is proposed by using the bilinear transformation of equation (12), and the results are shown again here as (13) and (14), when T_s is the sampling time of harmonic currents estimation loop.

$$i_{d6}(k) = \frac{T_s}{2L_d + RT_s}(v_{d6}(k) - L_q \omega_{re} i_{q6}(k) + v_{d6}(k-1)$$

TABLE I

PARAMETERS OF IPMSM USED IN EXPERIMENT

pole pairs P	2
EMF constant K_e	0.104V/(rad/s)
windings resistance R	0.45Ω
d-axis inductance L_d	4.15mH
q-axis inductance L_q	16.74mH

$$-L_q\omega_{re}i_{q6}(k-1)) + \frac{2L_d - RT_s}{2L_d + RT_s}i_{d6}(k-1) \quad (13)$$

$$i_{q6}(k) = \frac{T_s}{2L_q + RT_s}(v_{q6}(k) - L_d\omega_{re}i_{d6}(k) + v_{q6}(k-1)$$

$$-L_d\omega_{re}i_{d6}(k-1)) + \frac{2L_q - RT_s}{2L_q + RT_s}i_{q6}(k-1) \quad (14)$$

However, when using the above estimation method in the real situation, because of the derivative operator in the equation, we found that only the small amount of noise signal can make the estimation wrong easily.

Moreover,as shown in fig.9, the proposed method of this paper use only the inverter commanded voltages in the estimation process, and the noise amount of the commanded voltages $v^*_{d,q}$ is larger than $(v_{d,q} - v^*_{d,q})$ in the case of fig.8. Then, the harmonic currents estimation method shown in (13) and (14) are not proper for the proposed method.

By using the proposed harmonic voltages model discribed in the previous section, not only the instantaneous values of harmonic voltages can be calculated, but also the instantaneous values of amplitudes and phase angles of harmonic voltages can be known too. Therefore, with these amplitudes and phase angles data, we can estimate the harmonic currents easily by using only the simple phasor method as shown in (15) and (16), when the overarrow represent the phasor quantity and $j = \sqrt{-1}$.

$$\overrightarrow{i_{d6}}(k) = \frac{\overrightarrow{v_{d6}}(k) + L_q\omega_{re}\overrightarrow{i_{q6}}(k-1)}{R + j6\omega_{re}L_d} \quad (15)$$

$$\overrightarrow{i_{q6}}(k) = \frac{\overrightarrow{v_{q6}}(k) - L_d\omega_{re}\overrightarrow{i_{d6}}(k-1)}{R + j6\omega_{re}L_q} \quad (16)$$

With the proposed estimation method of harmonic currents discribed above, the experimental results of the estimation and compensation for harmonic currents are shown in the next section.

V. Experimental Results

In this section,the experimental results when using the proposed estimation and compensation methods in controlling of an IPMSM in the overmodulation range are shown. The control block diagram used in this experiment is shown in Fig.13 and the parameters of IPMSM and setting of experiment conditions are shown in TableI and II respectively.

First, we will show the effectiveness of the proposed harmonic currents estimation method, by comparing the waveform of the d-axis real current and the estimated current. When speed of IPMSM is 1,500rpm at no load condition,

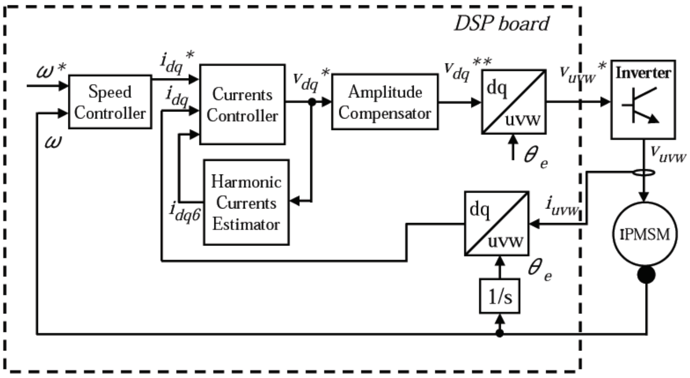

Fig. 13. Experimental Setup

TABLE II

SETTING CONDITIONS OF EXPERIMENT

DC Link Voltage	50 V
Inverter Carrier Frequency	10kHz
Speed Control Period	$500\mu s$
Current Control Period	$100\mu s$
Current Controller Gain	500rad/s

which modulation index of an inverter is about 1.27, the waveform of real and estimated currents are shown together in Fig.14 and the estimation error also shown in Fig.15. From Fig.15, we found that by using the proposed estimation method, the maximum estimation erros is just about 15% of the amplitude of harmonic current.

Next,we show the results of current waveforms when an IPMSM is operated in the overmodulation range with and without harmonic currents compensation. When speed of IPMSM is 1,480 rpm at no load condition, and the modulation index of an inverter is about 1.26. Without the harmonic currents compensation , the waveforms of d and q axis currents are shown in Fig.16. From this figure, we found that, because of the effect of the harmonic currents that fed back to the currents control loop, the waveforms of the currents, especially d-axis current, have become deterioriated.

On the other hand, when there is the compensation for the harmonic currents in the feedback part, the waveforms of d and q axis currents become as Fig.17. The waveforms of currents in this case are better, because the currents controllers can control the fundamental components of the currents effectively without any effect from the harmonic currents. However, the 6^{th} harmonic components that appear in the currents can not be avoided when using the overmodulation range.

VI. Conclusion

The operating range of an IPMSM become much wider when using the overmodulation range of an inverter. However, there are many problems occur and the main problem for closed loop currents control system is the large amount of harmonic components in the output voltages of an in-

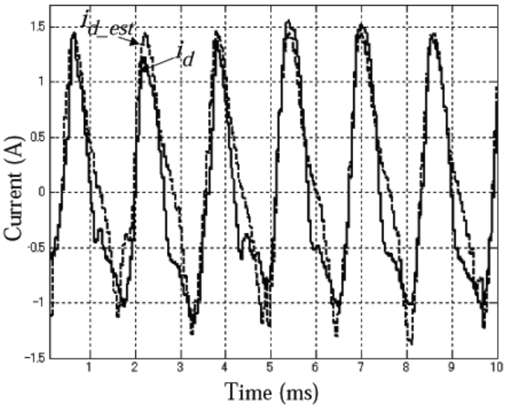

Fig. 14. Real an estimated d-axis currents

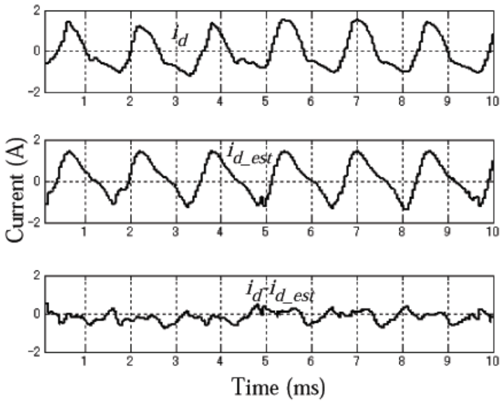

Fig. 15. Estimation error of d-axis current

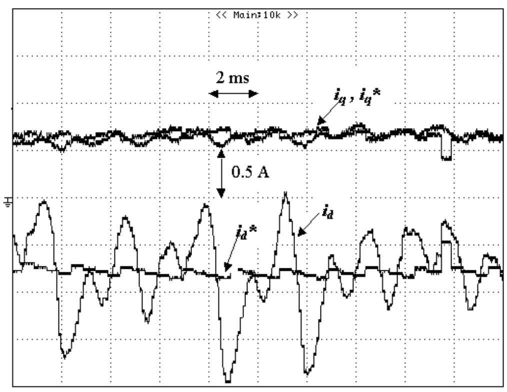

Fig. 16. d,q-axis currents without harmonic currents compensation

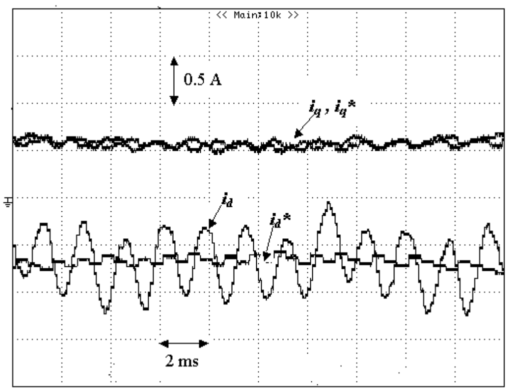

Fig. 17. d,q-axis currents with harmonic currents compensation

verter. With the feedback of the harmonic currents, generated from the above harmonic voltages, to the currents control loops, the responses of the currents become worse.

Many methods that have been proposed for controlling an IPMSM in the overmodulation range are summarized in this paper. And we found that the method that use the usual currents control system with the compesation of harmonic currents is most interesting. However,the harmonic currents estimation method is quite difficult to do in the pratical use.

Therefore, in this paper, we proposed the harmonic voltages estimation method that based on the harmonic voltages model and harmonic currents estimation method that based on the simple phasor method. The proposed estimation and compensation methods is easy for practical use and give the good results as shown in the experimental results section.

References

[1] K.Kondo, K.Matsuoka and Y.Nakazawa "A Designing Method in Current Control System of Permanent Magnet Synchronous Motor for Railway Vehicle Traction," *IEEJ Trans. Industrial Application*, vol. 118,no.7/8,Jul. 1998 pp. 900–907 (in Japanese)

[2] Y.Nakazawa, S.Toda and I.Yasuoka "A New Vector Control for Induction Motor Drives in Full Block Mode of Inverters," *IEEJ Trans. Industrial Application*, vol. 118,no.9,Nov. 1998 pp. 1071–1080 (in Japanese)

[3] K.Kondo, K.Matsuoka, Y.Nazakawa and H.Shimizu "Torque Feed-Back Control for Salient Pole Permanent Magnet Synchronous Motor at Weakening Flux Control Range," *IEEJ Trans. Industrial Application*, vol. 119,no.10,Sep. 1999 pp. 1155–1164 (in Japanese)

[4] Y.Ito, K.Tobari and T.Endo "Large Torque Control of PMSM in the Region of Voltage Saturation," *IEEJ Industrial Application Society Annual Meeting*,2004, pp.175–178,(in Japanese).

[5] H.Nakai, H.Ohtani, E.Satoh and Y.Inaguma "Development ans Testing of the Torque Control for the Permanent-Magnet Synchronous Motor," *IEEE Trans. Industrial Electronics*, vol. 52,no.3,Jun. 2005 pp. 800–806.

[6] H.Nakai, H.Ohtani and Y.Inaguma "Novel Torque Control Technique for High Efficiency/High Power Interior Permanent Magnet Synchronous Motors," *R&D Review of Toyota CRDL*, vol. 40,no.2,Mar. 2005 pp. 44–49.

[7] A.H.Hava,R.J.Kerkman and T.A.Lipo "Carrier-Based PWM-VSI Overmodulation Strategies:Analysis, Comparison, and Design," *IEEE Trans. Power Electronics*, vol. 13,no.4,Jul. 1998 pp. 674–689.

[8] S.Lerdudomsak,M.Kadota,S.Doki and S.Okuma "Inverter Modeling and Current Control System Design for Improving Current Response of PM Synchronous Motor in Overmodulation Range," *IEEE-IECON 2006*, pp. 1275-1280

[9] A.M.Khambadkone,J.Holtz "Compensated Synchronous PI Current Controller in Overmodulation Range and Six-Step Operation of Space-Vector-Modulation-Based Vector-Controlled Drives" *IEEE Trans. Industrial Electronics*, vol. 49,no.3,June. 2002 pp. 574–580.

[10] H.W.Kim,N.V Nho and M.J.Youn "Current Control of PM Synchronous Motor in Overmodulation Range," *IEEE-IECON 2004*, pp. 896-901

[11] K.Ogata, *Modern Control Engineering,* Prentice-Hall Inc., U.S.A., 1997.