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A New Simplified Sensorless Speed Control of Induction Motor using \(d\)-axis Voltage

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Abstract — This paper presents a new simplified sensorless speed control method of induction motors (IM). The output voltage of \(d\)-axis PI current controller is used to compute the flux angle and to control the speed in correspondence with its reference. The effectiveness of the proposed method has been demonstrated by simulations considered PWM and experiments.

Index Terms — \(d\)-axis voltage, root loci, speed sensorless vector control, stability

I. INTRODUCTION

Speed sensorless vector control of IM allows high performance control of torque and speed and the system is widely used. Many model reference adaptive system (MRAS) based methods are proposed [1] - [9]. However, the configurations of the systems are relatively complicated. Because MRAS based methods need state observer and many PI controllers (\(d\)-\(q\) currents, speed and speed estimation).

In this paper, we present a new simplified sensorless speed control method of IM. A linear model of the proposed system is derived in state space equation by taking a small perturbation of steady state operating point. Stability analysis is performed by showing root loci of the linear model. By the stability analysis we can determine the gains of controller.

Transient response of the linear model is compared with that by nonlinear model to check if two models of proposed method are valid. The effectiveness of the proposed method has been demonstrated by simulations considered PWM and experiments.

II. PROPOSED METHOD

To simplify control scheme, we have proposed a new sensorless vector control method shown in Fig.1. [10].

The \(d\)-\(q\) voltage equations of the induction motor in arbitrary rotating coordinate system are described as follows:

\[
e_{sd} = (R_s + \sigma L_s)p_i_{sd} - \omega \sigma L_s i_{sq} + \frac{M_p}{L_s} \psi_{rd} - \frac{aM}{L_s} \psi_{rq} \tag{1}
\]

\[
e_{sq} = \omega \sigma L_s i_{sd} + (R_s + \sigma L_s)p_i_{sq} + \frac{aM}{L_s} \psi_{rd} + \frac{M_p}{L_s} \psi_{rq} \tag{2}
\]

current model:

\[
0 = -M \frac{1}{\tau_r} i_{sq} + (\omega - \omega_0) \psi_{rd} + \left(\frac{1}{\tau_r} + p\right) \psi_{rq} \tag{3}
\]

In the steady state, the voltage model becomes:

\[
e_{sd} = R_s i_{sd} - \omega \sigma L_s i_{sq} - \frac{aM}{L_s} \psi_{rq} \tag{5}
\]

\[
e_{sq} = \omega \sigma L_s i_{sd} + R_q i_{sq} + \frac{aM}{L_r} \psi_{rd} \tag{6}
\]

In the proposed method, by using the output voltage of \(d\)-axis PI current control, we estimate the phase of rotor flux by changing it to satisfy that \(\psi_{rq}\) of (5) becomes zero.

\[
ed_d = K_r \left( i_{\text{sd}} - i_{\text{sd}}^* \right) + e_{rd} \tag{7}
\]

where, \(p e_{rd} = K_r \left( i_{\text{sd}} - i_{\text{sd}}^* \right) \)

By using the \(e_{rd}\) of Fig.1 and (5), we have following equation

\[
ed_d - R_q i_{sq}^* = -\omega \sigma M \psi_{rq} / L_r \tag{8}
\]

By controlling \(i_{\text{rd}}^*\) constant, and assuming that \(\psi_{rq} = 0\) in (3), we have

\[
\psi_{rd} = M i_{\text{rd}} \tag{9}
\]

By assuming that \(\psi_{rq} = 0\) in (4) and using (9), slip frequency \(\omega_s\) is computed as

\[
\omega_s = i_{\text{sq}} / \tau_r i_{\text{rd}} \tag{10}
\]

We estimate the flux frequency \(\omega^*\) so as to bring \(q\)-axis flux zero by using P controller shown in Fig.1

\[
\omega^* = \omega_s^* + \omega_0 - K_p \left( \omega^*_d - R_q i_{sq}^* \right) \tag{11}
\]

Fig.1. Block diagram of proposed method.
where, $K_w = \text{sign}(\omega r \dot{r})K_{aw}$. Value of $K_w$ is positive if $\omega r$ is greater than zero, and $K_w$ is negative if $\omega r$ is less than zero. Slip frequency $\omega s$ is added to speed command $\omega r$.

The following integral controller computes the angular frequency $\omega i$ that control torque and speed.

$$p\omega i = \frac{K_w}{T_w} \left( e^r_i - R_i \dot{e}^p_i \right)$$  \hspace{1cm} (12)

The $q$-axis voltage $e^q_i$ is computed by neglecting the voltage drop of stator resistance, but it is compensated by $\omega i$.

$$e^q_i = \omega^q_i L_i i^q_{sd}$$  \hspace{1cm} (13)

where, $\omega^q_i = \omega^r_i + \omega^s_i$.

In steady-state, $\psi_{eq}$ must be zero by the action of integral controller, and then the following relation is valid.

$$\omega^r_i + \omega^s_i = \omega^r_i$$  \hspace{1cm} (14)

When (14) is satisfied, the rotor speed is equal to its command. In proposed system of Fig.1, any observer is not used and $d$-axis PI controller and $P$ and $I$ frequency controllers are only used. Therefore, the proposed system is much simpler than conventional systems.

### III. Stability Analysis

Taking a small perturbation at steady-state operating point, a linear model of proposed system is derived. System stability is discussed by using root loci. A linear model of the proposed system is derived in form

$$p\Delta x = A\Delta x + B\Delta \omega^i + B_t\Delta T_e$$  \hspace{1cm} (15)

$$\Delta x = \begin{bmatrix} \Delta e_{rd} \\ \Delta \psi_{eq} \\ \Delta \psi_{rd} \\ \Delta \omega_r \\ \Delta e_{sd} \\ \Delta \omega_s \end{bmatrix}^T$$

$$a_i = \frac{R_i}{\sigma L_i} + \frac{M^2}{\sigma L_i \tau_r}, \ a_s = \frac{M p^2}{4 L_i}$$

$$A = \begin{bmatrix} -a_i - \frac{K_p}{\sigma L_i} & 0 & \frac{M}{\sigma L_i \tau_r} \\ -\omega^r_i - \frac{I_a K_w}{\sigma L_i} & -a_i - \frac{1}{\tau_r} - \frac{1}{\tau_r} & \frac{M \omega_s}{\sigma L_i} \\ \frac{M}{\tau_r} + \frac{\psi_{eq}}{\tau_{eq}^p} & \frac{\psi_{eq}}{\tau_{eq}^p} & -1 \\ \psi_{rd} K_p & \psi_{rd} K_p & \omega^r_i - \omega^s_i \\ -a_i \psi_{rd} & a_i \psi_{rd} & a_i \dot{i}_{sd} \\ -K_i & 0 & 0 \\ \frac{K_w}{T_w} & 0 & 0 \end{bmatrix}$$

$$B = \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix}$$

System is stable when all eigenvalues of matrix $A$ are in the left half of the $s$-plane.

Fig.2 shows root loci when the slip speed $N_{sl}$ and $T_e$ of I controller are changed. In Fig.2 (a), the speed command $N_r^r$ is 700 rpm, $K_w = 3.0$, $T_e$ varies from 0.01 to 0.05, and $N_{sl}$ varies from -80 rpm to 80 rpm. System becomes unstable by the root near the point $j60$ especially at motoring operation when $T_e = 0.01$. In Fig. (b), the increase in $T_e$ causes instability by the root near the point $j5$ especially at regenerating operation.

Fig.3 shows root loci when the slip speed $N_{sl}$ and the gain $K_w$ of P and I frequency controllers are changed. The speed command $N_r^r$ is at 700 rpm, $T_e = 0.05$, $K_w$ varies from 0.01 to 10.0, and $N_{sl}$ varies from -80 rpm to 80 rpm. The value of $T_e$ is chosen from root loci of Fig.2 (a) as 0.05. When $K_w = 0.1$ and slip speed $N_{sl}$ is in regenerating operating points, all poles are in unstable region on positive real axis. From Fig.3, stable operations are obtained in all load condition when $K_w = 3.0$ or $K_w = 10.0$, then we choose that $K_w = 3.0$ in experimental system.

Fig.4 shows the root loci when $N_r^r$ is 700 rpm and $N_{sl}$ is 35 rpm. In this figure, $K_w$ varies from 0.01 to 100.0 and $T_e$ from 0.01 to 0.10. It is observed that the small values of $K_w$ and $T_e$ result in unstable or oscillating operation.

To examine the stability at low speed operation and at higher speed one, root loci at 100 rpm and 1500 rpm are computed for selected parameter.

In Fig.5, the root loci are shown when $T_e = 0.05$, $K_w = 3.0$, and $N_r^r = 100$ rpm for the change of slip speed $N_{sl}$. The system can work in stable region in motoring and regenerating operation.
Fig. 2. Root loci with parameters $T_\omega$ and $N_{sl}$.

Fig. 3. Root loci with parameters $K_\omega$ and $N_{sl}$.

Fig. 4. Root loci with parameters $T_\omega$ and $K_\omega$.

Fig. 5. Root loci at $N'_{sl} = 100$ rpm.

Fig. 6. Root loci at $N'_{sl} = 1500$ rpm.
In Fig.6, the root loci are shown when $T_w = 0.05$, $K_w = 3.0$, and $N_r' = 1500$ rpm. The system is stable in all operating points.

Comparison between Fig.5 and Fig.6 shows that the system works more stable at higher speeds both motoring and regenerating operation because with increasing speed, the roots have smaller real part.

Fig.7 shows the transient responses for the step change of speed command from 700 rpm to 710 rpm and back to 700 rpm. The load torque $T_L$ is 4 Nm, $T_w = 0.05$ and $K_w = 0.5$.

The result of linear model (15) is shown in Fig. (a), and nonlinear model result is shown in Fig. (b). The nonlinear model uses the equations which are obtained when the linear model is derived. The results of the linear model are close to those of the nonlinear model. It is confirmed that the derivation of the linear model (15) is correct.

![Fig.7. Transient responses for the step change of speed command.](image)

IV. EXPERIMENTAL AND SIMULATION RESULTS

The proposed control system is implemented by a DSP (TMS3200C32)-based PWM inverter as shown in Fig.8. Sampling period used in this system is 200 $\mu$s. Because the dead time and the non-ideal features of IGBT influence the output voltage of the inverter, a compensating algorithm is developed for the experimental system [8]. Parameters of IM are: number of pole $p=4$, stator resistant $R_s=1.54 \Omega$, rotor resistant $R_r=0.787 \Omega$, stator and rotor inductance $L_s=L_r=0.0115 H$, mutual inductance $M=0.11 H$, and moment of inertia $J=0.0126 \text{kgm}^2$.

In Figs.9 – 11, the speed command is changed from 700 rpm to 800 rpm and then 700 rpm in the monitoring operation at load torque of 4 N-m for three different values of $K_w$.

When $K_w = 3.0$, the transient response is shown in Fig.9. The nonlinear simulation result of (a) is obtained by considering PWM switching in detail. Because of the PWM switching, high frequency ripples are observed in the waveforms of the synchronous speed $N_r'$, $d$-axis voltage $e_d'$ and $q$-axis current $i_q$. Experimental results of (b) are similar to those of nonlinear simulation. On the other hand, the results of linear model (15) do not contain high frequency ripples as shown in Fig.(c). However, the low frequency responses are similar to those of Figs.(a) and (b).

Fig.10 gives the results when $K_w = 0.5$. By comparing the results of Fig.10 with those of Fig.9, it is observed that the oscillation is increased and the settling time of actual speed to speed command becomes longer, and the overshoot of speed is larger. Good correlation between the nonlinear simulation result and experimental one is also obtained. However, the result of linear model has slightly moderate response because of nonlinear characteristic.

Fig.11 gives the results when $K_w = 5.0$. By comparing Fig.9 and Fig.11, the ripple of $N_r'$ in Fig.11 becomes larger in the nonlinear simulation and experimental result. In experimental result, the effect of noise is expanded by large gain of $K_w$ in Fig.11 (b). However, the actual speed smoothly converges to the speed command and the results of actual speed in all cases are almost same.

![Fig.8. Experimental system.](image)
Fig. 9. Responses for the step change of speed command.

(a) Nonlinear simulation result.

(b) Experimental result.

(c) Linear model result.

(700 $\rightarrow$ 800 $\rightarrow$ 700 rpm, $\tau_x = 4$ N-m, $K_\omega = 3.0$, $T_{\omega x} = 0.05$).

Fig. 10. Responses for the step change of speed command.

(a) Nonlinear simulation result.

(b) Experimental result.

(c) Linear model result.

(700 $\rightarrow$ 800 $\rightarrow$ 700 rpm, $\tau_x = 4$ N-m, $K_\omega = 0.5$, $T_{\omega x} = 0.05$).
V. CONCLUSIONS

The conclusions drawn from this study are summarized as follows:

1. We have proposed a new simplified sensorless vector control method of IM.
2. P control for flux angle computation and I control for torque and speed are realized by using the output voltage of $d$-axis PI controller.
3. A linear model is derived and the root loci are used to design the control parameters.
4. It is demonstrated that the proposed system can realize stable operation in both motoring and regenerating modes by the result root loci.
5. The experimental results agree with those of nonlinear simulation and the usefulness of proposed method is verified.

REFERENCES