Deadbeat Control of Induction Motor Current using State Observer with Adaptive Poles Selection

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Summary

For a high power induction motor drive, the switching frequency of inverter can not become higher than several kHz, and such switching frequency produces a large current ripple, which in turn, produces torque ripple. To minimize the current ripple, a new method based on deadbeat control theory for current regulation is proposed. The PWM pattern is determined at every sampling instant based on stator currents measurements, motor speed, current references and rotor flux vector, which is predicted by a full-order observer with adaptive poles selection, so that the stator currents are controlled to be exactly equal to the reference currents at every sampling instant.

The proposed method consists of two parts: (1) derivation of a deadbeat control; (2) construction of a full-order observer which predicts the rotor flux and the stator currents in the next sampling instant. This paper describes a theoretical analysis and computer simulations under various system conditions.

key word: Induction Motor, Deadbeat Control, Observer

1. Introduction

The vector control technique, also known as field oriented control, has made it possible for induction motor drives to be used in high performance application. However, the current controlled vector control, in general, requires the following assumptions:

(1) An exact information on the rotor flux vector is available.

(2) The actual stator currents are adjusted instantaneously and precisely to the references.

The latter condition (2) is satisfied by a high switching frequency. However, the switching frequency remains as low as a few kHz for a high power induction motor drives. As a result, large current ripples are produced, which in turn, produce torque ripples. The deadbeat technique is a type of control, by which the system exhibits a finite settling time response, as already applied to the PWM inverter in Ref. (2). In this paper the deadbeat control was extended to the induction motor drive to satisfy the above second condition. The current of induction motor is controlled without delay time by the proposed method as follows. The inverter with induction motor load is considered as a plant of a closed loop digital feedback system. Unlike the conventional bang-bang current control, the proposed control scheme determines the PWM pattern at every sampling instant based on stator currents, rotor flux, motor speed and current references. As a result the stator currents are controlled to be exactly equal to the reference currents at every sampling instant, and the torque ripples.
become less than those in conventional method.

To satisfy the former condition (1), the direct rotor flux sensing method such as by hall sensor, may be preferable in theory. From the viewpoint of practical use, this method is unreliable and expensive. Thus, the indirect flux sensing methods are more desirable, but they are dependent on machine parameters. Several methods have been proposed to minimize the effects of parameters variation in the indirect vector control method. Among the recent interesting works on analysis and design of flux observers, Ref. (13) evaluates the related works and emphasizes the importance of a predictive error feedback term in order to converge the estimated flux error to zero quickly. A minimum order flux observer using corrective prediction error was designed in Ref. (14) - (16). However these reduced-order observers are sensitive to noise in the measurements of rotor speed and stator currents. Furthermore around zero speed, these observers do not estimate the flux well. To solve this problem, the poles position of a full-order observer is shifted adaptively depending on the motor speed.

Combining the state estimator (observer) and the deadbeat controller, the stator currents are controlled at any values over a wide range of speeds.

The deadbeat controller applied for voltage source inverter fed induction motor drive is presented in Section 2 of this paper. The design of full-order observer is described in Section 3. Section 4 gives digital computer simulation results, and Section 5 concludes this paper.

2. Deadbeat control law

2.1 Induction motor state equation

The basic equation of an induction motor with the d-q coordinates fixed in the stator are expressed as,

$$\begin{bmatrix} I_s \\ \lambda_r \end{bmatrix} = \begin{bmatrix} A_{11} & A_{12} \\ A_{21} & A_{22} \end{bmatrix} \begin{bmatrix} I_s \\ \lambda_r \end{bmatrix} + \begin{bmatrix} B_1 \\ 0 \end{bmatrix} V_s$$

$$= A \begin{bmatrix} I_s \\ \lambda_r \end{bmatrix} + BV_s \quad \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots (1)$$

The output equation is

$$I_s = \begin{bmatrix} 1 & 0 \end{bmatrix} \begin{bmatrix} I_s \\ \lambda_r \end{bmatrix} = C \begin{bmatrix} I_s \\ \lambda_r \end{bmatrix} \quad \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots (2)$$

The symbol "\(^{\prime}\)" denotes the time derivative.

where,

$$A_{11} = (R_1/(\sigma L_1) + R_2(1-\sigma)/(\sigma L_2)) I$$

$$A_{12} = M/(\sigma L_1 L_2) - (R_3/L_2) I + \omega r J$$

$$A_{21} = (R_2/L_2) I$$

$$A_{22} = -(R_3/L_2) I + \omega r J$$

$$B_1 = (1/\sigma L_1) I$$

$$I = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}, \quad J = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}$$

The meanings of parameters are as follows,

$$\sigma = 1 - M^2/L_1 L_2 : \text{leakage coefficient},$$

$$R_1, R_2 : \text{stator and rotor resistance},$$

$$L_1, L_2 : \text{stator and rotor self inductance},$$

$$M : \text{stator/rotor mutual inductance},$$

$$\omega r : \text{electrical rotor angular velocity},$$

$$V_{as}, V_{qs} : \text{d-axis and q-axis stator volt-ages},$$

$$I_{ds}, I_{qs} : \text{d-axis and q-axis stator currents}$$

$$\lambda_{ar}, \lambda_{qr} : \text{d-axis and q-axis rotor fluxes}$$

The torque equation is then

$$T_e = (3/2) M L_q (I_{qs} \lambda_{ar} - I_{ds} \lambda_{qr}) \quad \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots (3)$$

and the state equation of the mechanical system is as follows,

$$\dot{\omega}_r = -D/\omega_r + 1/J \pi(T_e - T_L) \quad \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots (4)$$

where,

$$T_e : \text{electromagnetic torque},$$

$$T_L : \text{load torque} J_{\pi} : \text{moment of inertia},$$

$$D : \text{damping factor}$$

2.2 Deadbeat control theory

Fig. 1 is the proposed digital control system of the induction motor drive. Assuming that the inverter is for high power application, 1.8 kHz sampling frequency is chosen. To derive a sampled-data model of Eqs. (1) and (2), inverter output voltages V_{ab}, V_{sc} and V_{ca} are chosen as one or two voltage pulses of magnitude +E or -E, satisfying the relation of V_{ab} + V_{sc} + V_{ca} = 0 at any moment as shown in Fig. 2. \(\Delta T_{ab}(k), \Delta T_{sc}(k), \Delta T_{ca}(k)\) are the pulse widths of V_{ab}, V_{sc} and V_{ca} respectively in the k-th sampling interval. The two phase pulse widths \(\Delta T_d(k)\) and \(\Delta T_q(k)\) are given by Eqs. (5) and (6).
As long as these pulses are symmetrical at the center of the sampling interval and so are \( V_{ds} \) and \( V_{q} \), the sampled-data model of Eqs. (1) and (2) becomes,

\[
\begin{bmatrix}
\Delta T_{ab}(k) \\
\Delta T_{bc}(k) \\
\Delta T_{ca}(k)
\end{bmatrix} =
\begin{bmatrix}
1 & -1/\sqrt{3} \\
0 & 2/\sqrt{3} \\
-1 & -1/\sqrt{3}
\end{bmatrix}\begin{bmatrix}
\Delta T_s(k) \\
\Delta T_s(k) \\
\Delta T_s(k)
\end{bmatrix}
\]

\[
\begin{bmatrix}
\Delta T_a(k) \\
\Delta T_b(k) \\
\Delta T_c(k)
\end{bmatrix} =
\begin{bmatrix}
1/2 & 0 & -1/2 \\
0 & \sqrt{3}/2 & 0
\end{bmatrix}\begin{bmatrix}
\Delta T_{ab}(k) \\
\Delta T_{bc}(k) \\
\Delta T_{ca}(k)
\end{bmatrix}
\]

Thus, using Eq. (7) and replacing \( I_s(k+1) \) with the reference \( I_{sref}(k+1) \), the deadbeat control law is given as follows:

\[
\Delta T(k)=[\Delta T_s(k), \Delta T_s(k)]^T. F, H, I_s(k), \lambda_s(k) \text{ and } \lambda_s(k) \text{ are the values at the sampling instant of } kT.
\]

This control law forces the output currents to be exactly equal to the reference signal at the \( (k+1) \)th sampling instant. For computation of Eq. (8), the values of \( F_{in}, F_{in} \) and \( H_1 \), which are derived from the actual motor speed and plant parameters, have to be calculated or searched from look-up table, and \( L_u(k) \) and \( \lambda_u(k) \) must be provided at every sampling instant. The references are one sampling ahead preview values of stator currents, which are provided by the vector control method. If \( I_s(k) \) is detected at \( kT \) from which \( \bar{I}_s(k) \) is estimated, followed by computation of Eq. (8), then the pulse width \( \Delta T(k) \) in Eq. (8) cannot be as large as the sampling interval \( T \) due to the finite calculation time. To get rid of this, \( I_s(k) \) and \( \lambda_s(k) \) are predicted at the \( (k-1) \)th sampling instant, and Eq. (8) is computed before \( k \)th sampling instant, which enables the maximum pulse width to be equal to \( T \), as shown in Fig. 3.

### 2.3 Pulse pattern selection

The three phase pulse widths \( \Delta T_{ab}(k), \Delta T_{bc}(k), \text{ and } \Delta T_{ca}(k) \) are calculated in Eq. (5) from \( \Delta T_s(k) \)

\[
\Delta T_s(k)=[\Delta T_s(k), \Delta T_s(k)]^T. F, H, I_s(k), \lambda_s(k) \text{ and } \lambda_s(k) \text{ are the values at the sampling instant of } kT.
\]
and $\Delta T_s(k)$ obtained in Eq. (8). The method used for selection of actual pulse is the same as in Ref. (1). An example of pulse pattern selection is shown in Fig. 2.

3. **Full-order observer**

3.1 **Design of observer with corrective prediction error feedback**

Recently, there are several approaches published in which a full-order observer theory was applied to rotor flux estimation \(^{(17)-(19)}\). However, none of these papers deal with low speed problem even in the case of no change in motor parameters.

The full-order observer state equation with corrective prediction error feedback becomes as follows \(^{(21)}\),

$$\begin{bmatrix}
\hat{I}_s(k+1) \\
\hat{\lambda}_r(k+1)
\end{bmatrix} = \begin{bmatrix}
F_{11} & F_{12} \\
F_{21} & F_{22}
\end{bmatrix} \begin{bmatrix}
\hat{I}_s(k) \\
\hat{\lambda}_r(k)
\end{bmatrix} + \begin{bmatrix}
H_1 \\
H_2
\end{bmatrix} \Delta T(k)$$

$$+ \begin{bmatrix}
G_1 \\
G_2
\end{bmatrix} \left[I_s(k) - \hat{I}_s(k)\right]$$

(9)

The symbol "−" denotes the predicted quantities.

where, $G = \begin{bmatrix}
G_1 \\
G_2
\end{bmatrix}$ is the gain matrix.

The error dynamics of the observer is,

$$e(k+1) = (F - GC) e(k)$$

(10)

where,

$$e(k) = \begin{bmatrix}
I_s(k) - \hat{I}_s(k) \\
\lambda_r(k) - \hat{\lambda}_r(k)
\end{bmatrix}$$

(11)

The gain matrix $G$ is designed such that the poles of Eq. (10) take the desired values. Thus the pole placement technique is used, in which, Eq. (7) is first transformed into a controllable canonical form, then 4 poles of the observer are assigned. With choice of the observer gain matrix $G$ as shown in Eq. (12), it is possible to place the poles at any specified conjugate complex pairs in the $s$ domain \(^{(12)-(17)}\).

$$G_1 = g_1I + g_2J = \begin{bmatrix}
g_1 & -g_2 \\
g_2 & g_1
\end{bmatrix}$$

$$G_2 = g_3I + g_4J = \begin{bmatrix}
g_3 & -g_4 \\
g_4 & g_3
\end{bmatrix}$$

(12)

4. **Simulations**

4.1 **Deadbeat control simulations**

In order to verify the proposed deadbeat control law Eq. (8), digital simulations were carried out, assuming that the flux vector could be measured. The motor parameters used for simulation are listed in the appendix. Fig. 4 shows the waveforms of stator current $I_{ds}$, electromagnetic torque $T_e$ and stator voltage $V_{ds}$. Fig. 5 shows the comparison of torque and currents ripple between the proposed and the conventional hysteresis comparator method in which the phase currents are forced to track the reference generated by the vector controller. The switching frequency of the proposed pulse pattern in Fig. 2 is about 2/3 of the sampling frequency, thus the average of the switching frequency of the hysteresis comparator method is adjusted to be equal to it. From Fig. 5, the torque ripple of the proposed control method is about 40% less than that of hysteresis method.

4.2 **Simulation of state observer with fixed poles placement**

The dynamic of the observer is determined by the selection of poles position. Simulation results are shown in Fig. 6, using the gain matrix given by Eq. (12) and setting the poles of the observer in $s$ domain, for example, as follows,

$$a_1 \pm jb_1 = -2,000 \pm j2,000$$

$$a_2 \pm jb_2 = -1,000 \pm j1,000$$

Fig. 6 shows that the stator currents and the rotor flux are estimated without any error, except at very low speed where the predicted flux does not coincide with the real one. The observer poles should be
chosen in such a way that the observer can predict all state variables at any speed and reduce the prediction error as fast as possible. Different kinds of fixed poles were simulated for the best poles allocations, but fixed poles did not provide reasonably small prediction errors in the low speed range. To overcome this problem, a new method, called an adaptive poles placement, is presented in the following section.

4.3 Simulation of the observer with adaptive poles placement

The induction motor poles as a function of rotor speed are computed as shown in Fig. 7. Poles in this figure are eigenvalues of the controllable canonical form of the matrix $A$ in Eq. (1). In other words, they are open loop system poles. Thus if the
observer poles are fixed by adjusting the observer gains, these gains become very high at low speed, which deteriorates the prediction. In the proposed adaptive poles placement technique, the poles of observers are shifted when motor poles move as a function of rotor speed. Then, the error of the estimated state variables was averaged through the reverse speed operation, and the observer poles $\alpha \pm j\beta_1$ and $\alpha_2 \pm j\beta_2$, which provided the minimum average error, were selected, as follows,

$$\alpha \pm j\beta_1 = 4.5(\gamma_1 \pm j\gamma_1)$$
$$\alpha_2 \pm j\beta_2 = 3(\gamma_2 \pm j\gamma_2)$$

(13)

where, $\gamma_1$ and $\gamma_2$ are the real parts of the motor poles in Eq. (1).

This choice provides the best results through simulations. The flux is well predicted even at low speed. Now that all state variable are precisely predicted, deadbeat controller with observer under vector control (DBOV) is implemented in the system, and Fig. 8 shows the reverse speed operation using the proposed controller.

Since the whole sampling interval (0.555 ms) is available for necessary calculation, the actual implementation can be made by a digital signal processor, which can execute the required computation using look-up tables for coefficients of deadbeat control and adaptive observer poles. The look-up table for poles should be prepared using Eq. (13). However, the speed step has to be chosen carefully taking into account the desired performance as well as the hardware constraints.

4.4 Parameters sensitivity

The coefficients in Eqs. (8) and (9) are all dependent on motor parameters. Since these
parameters may vary during on-line operation due to temperature or saturation effects, it is important to investigate the sensitivity of the complete system (DBOV) to parameters changes. One of the most significant changes in motor parameters is the rotor resistance $R_2$. It was shown in Ref. (1) that the deadbeat control is not affected by the practical range of $R_2$ change. Note that the sensitivity of the complete system to $R_2$ change increased as the motor speed decreased. A variation of $\pm 5\%$ of $R_2$ does not affect the complete drive system (DBOV). Fig. 9 shows the system under the proposed method at $\omega_r=1$ rpm with $+5\%$ variation of $R_2$. Note that the slip frequency $(\omega_s)$ is 100 rpm.

At very low speed, whenever $R_2$ variation exceeds 10\%, the predicted and the actual rotor flux do not coincide. As rotor resistance variation is increased, the difference between predicted and actual rotor flux is increased. For example, $+10\%$ variation of $R_2$ causes about $+6\%$ change in rotor flux magnitude and $+4$ degrees deviation in phase, while $+20\%$ variation of $R_2$ causes respectively $+9\%$ and $+8$ degrees. An additional compensating method has to be considered for these cases, but it is beyond this paper.

5. Conclusions

A new technique based on deadbeat control theory, using a state observer with adaptive poles selection, was proposed to control induction motor stator currents under vector control. From the theoretical analysis and digital simulations, the following conclusions were made.

The deadbeat controller enabled low current ripple with lower switching frequency, which, in turn, resulted in low torque ripple.

The state observer with adaptive poles selection showed the deadbeat current control to be effective around zero rotor speed, and therefore contributed to performance improvement over a wide range of rotor speeds.

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References

Appendix

Induction motor parameters used for simulations
60 hp, 3 phases, 60 Hz, 2 poles, 440 V, induction motor.

Moment of inertia \( J_m = 0.1 \text{ N} \cdot \text{m}^2 \)
Damping factor \( D = 0.01 \text{ N} \cdot \text{m} / \text{s} \)
\( R_1 = 0.0793 \Omega, R_2 = 0.0785 \Omega \)
\( L_1 = 31.25 \text{ mH}, L_2 = 31.19 \text{ mH} \)
\( M = 30.50 \text{ mH} \)

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