

A Current Control Technique for Voltage-Fed Induction Motor Drives

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Abstract- A current control scheme based on the de-coupling of the motor back emf voltage and a deadbeat feedback control to achieve fast dynamic response on the motor current is proposed in this paper. Since the de-coupling voltage is expressed as a function of motor current and control voltage, the scheme is simple for microprocessor implementation and is not sensitive to the motor parameters and the tuning of the field orientation control. The performance of the proposed control scheme was investigated both by simulation and experiment, good static and dynamic performances have been obtained in the experimental verifications.

I. INTRODUCTION

For field orientation controlled induction motor drives, de-coupling of torque and flux current components depends on the drive's ability to control the instantaneous position of the stator current vector. Error in the current controller degrades the drive's performance in the same way as de-tuning the field orientation control. Current error causes cross-coupling between the torque and flux producing currents and then introduces undesirable steady state and transient responses to the motor torque.

There has been many strategies proposed for ac motor current control. Some well known techniques are the hysteresis current control [1] and the carrier based current control such as the stator frame and the synchronous frame proportional plus integral (PI) control [2, 3]. The synchronous frame control has been acknowledged as the superior among these techniques since it allows for a fixed switching frequency and yields zero steady state error at any given command frequency. Another current control technique involves de-coupling of the motor emf voltage to improve the performance of the current regulation [4]. In that study the cross-coupling voltages between the d - q axis were de-coupled so that the induction motor becomes a linear system in the synchronous frame, therefore a simple PI or deadbeat feedback control can be utilized to regulate the current error to zero [5-6]. Nevertheless, the effectiveness of this technique is depending on the correct rotor flux field orientation and an accurate rotor flux measurement. Another study [7] attempted

to resolve the disadvantages of the de-coupling control by treating the induction motor model in the stationary frame using the discrete control technique. Although this technique has excluded the dependency on the rotor flux field orientation and the need for rotor flux measurement, however, implementation of the control is complicate since it was performed in the stationary frame; besides, no experimental verification was presented in that study.

This paper proposes a new current control scheme that overcomes the above mentioned disadvantages of the existing schemes for a voltage-fed PWM induction motor drive. The de-coupling voltage for the current to closely follow its reference has been derived using the discrete state equation of the induction motor. The model leads to a control law that does not require knowledge of rotor flux, and is independent of the tuning of the field orientation control. A deadbeat control scheme for current feedback control is incorporated into the control laws to reduce the current error as fast as possible and to stabilize the closed loop system against disturbance and motor parameter variations. Experimental verifications were performed using a voltage-fed PWM induction motor drive system controlled by a TMS320C240 DSP.

II. CONTROL STRATEGY

The equations for a three-phase squirrel cage induction motor in the synchronous d - q rotating frame can be expressed as,

$$v_{qds}^e = r_s i_{qds}^e + p \lambda_{qds}^e + j \omega_e \lambda_{qds}^e \quad (1)$$

$$0 = r_r i_{qdr}^e + p \lambda_{qdr}^e + j(\omega_e - \omega_r) \lambda_{qdr}^e \quad (2)$$

where v_{qds}^e is the motor voltage, i_{qds}^e and i_{qdr}^e are the stator and the rotor current respectively, ω_e is the stator frequency, λ_{qds}^e and λ_{qdr}^e are the stator and rotor

flux respectively, $p = \frac{d}{dt}$, and for any complex variable f , $f_{qds} = f_{qs} - j f_{ds}$. The rotor and the stator flux can also be expressed as,

$$\lambda_{qds}^e = L_s i_{qds}^e + L_m i_{qdr}^e \quad (3)$$

$$\lambda_{qdr}^e = L_m i_{qds}^e + L_r i_{qdr}^e \quad (4)$$

Substituting (3) and (4) into (1) and (2) to eliminate λ_{qds}^e and i_{qdr}^e , after rearranging the resulting equations one can obtain the following expression for the stator voltage:

$$v_{qds}^e = (r_s' + j\omega_e \sigma L_s) i_{qds}^e + \sigma L_s p i_{qds}^e - \frac{L_m}{L_r} \omega_{br} \lambda_{qdr}^e \quad (5)$$

where

$$\sigma = 1 - \frac{L_m^2}{L_s L_r}, \quad r_s' = r_s + r_r \frac{L_m^2}{L_r^2} \quad \text{and} \quad \omega_{br} = \frac{r_r}{L_r} - j\omega_r. \quad \text{The}$$

first two terms in the right hand side of (5) are the voltage drop due to the winding resistance and inductance, and the last term is the voltage caused by the rotor rotation. Also, it can be seen that all the terms with ω_e or ω_r have cross-coupling between the d and the q-axis. The proposed control strategy is consisted of a de-coupling and a feedback control. The de-coupling control cancels the emf voltage in the motor in order for fast tracking of the current reference. The feedback control provides fast dynamic and steady state responses for current error regulation.

A. Decoupling Control

Since the current control is typically performed at a few thousand times per second, the execution rate is much faster than the motor mechanical and rotor flux dynamic response. Therefore Eq. (5) can be simplified by assuming the motor emf voltage, i.e. the sum of terms associate with the rotor flux, is constant within a current control period. Then the motor voltage equation becomes,

$$v_{qds}^e = (r_s' + j\omega_e \sigma L_s) i_{qds}^e + \sigma L_s p i_{qds}^e + E_o \quad (6)$$

where E_o is a constant relate to the motor speed and rotor flux at a current control instant. A discretized state equation can be derived from Eq. (6) since the motor voltage is constant between the k_{th} and $(k+1)_{th}$ sampling instants for a voltage-feed PWM drive:

$$\begin{bmatrix} i_{qs}^e(k+1) \\ i_{ds}^e(k+1) \end{bmatrix} = \begin{bmatrix} 1 - \frac{r_s' T}{\sigma L_s} & -\omega_e T \\ \omega_e T & 1 - \frac{r_s' T}{\sigma L_s} \end{bmatrix} \begin{bmatrix} i_{qs}^e(k) \\ i_{ds}^e(k) \end{bmatrix} + \frac{T}{\sigma L_s} \begin{bmatrix} v_{qs}^e(k) \\ v_{ds}^e(k) \end{bmatrix} + \begin{bmatrix} E_{qd}(k) \\ E_{dq}(k) \end{bmatrix} \quad (7)$$

where T is the sampling period. Letting

$$\bar{I}_s = \begin{bmatrix} i_{qs}^e \\ i_{ds}^e \end{bmatrix}, \quad \bar{V}_s = \begin{bmatrix} v_{qs}^e \\ v_{ds}^e \end{bmatrix}, \quad \bar{A} = \begin{bmatrix} 1 - \frac{r_s' T}{\sigma L_s} & -\omega_e T \\ \omega_e T & 1 - \frac{r_s' T}{\sigma L_s} \end{bmatrix}, \quad \bar{B} = \frac{T}{\sigma L_s}$$

then Eq. (7) can be rewritten as follows for convenience,

$$\bar{I}_s(k+1) = \bar{A} \bar{I}_s(k) + \bar{B} \bar{V}_s(k) + \bar{E}_o(k) \quad (8)$$

Since the state equation at the $(k-1)_{th}$ sampling instant is

$$\bar{I}_s(k) = \bar{A} \bar{I}_s(k-1) + \bar{B} \bar{V}_s(k-1) + \bar{E}_o(k-1) \quad (9)$$

Because E_o is approximately constant between the two sampling instants, therefore combining (8) and (9) to eliminate $E_o(k)$ and $E_o(k+1)$ yields an expression consisting only the discretized motor voltages and currents.

$$\bar{V}_s(k) = \bar{V}_s(k-1) + \frac{1}{\bar{B}} \left[\left(\bar{I}_s(k+1) - \bar{I}_s(k) \right) - \bar{A} \left(\bar{I}_s(k) - \bar{I}_s(k-1) \right) \right] \quad (10)$$

Eq. (10) gives a relationship for the motor voltages at the consecutive sampling instants. By using this equation the applied motor voltage at any control instance can be synthesized with the voltage applied at the previous control instance and the motor currents. Based (10), the de-coupling voltage required for the current to follow its reference without any error at the k_{th} sampling instance is

$$\bar{V}_{ff}^*(k) = \bar{V}_s(k-1) + \frac{1}{\bar{B}} \left[\left(\bar{I}_s^*(k+1) - \bar{I}_s(k) \right) - \hat{A} \left(\bar{I}_s^*(k) - \bar{I}_s(k-1) \right) \right] \quad (11)$$

where $\bar{I}_s^*(k+1)$ and $\bar{I}_s^*(k)$ are current commands, \hat{B} and \hat{A} are the estimated \bar{B} and \bar{A} respectively. Note that because $\bar{I}_s^*(k+1)$ is not available for the calculation of $\bar{V}_{ff}^*(k)$, Eq. (11) is delayed one sampling period so that all the currents are available at the time of the calculation. The final form of the de-coupling voltage is found by expanding (11) into the d-q axis, and the results are

$$\begin{aligned} v_{qff}^* = & v_{qs}(k-2) + \frac{\sigma L_s}{T} [i_{qs}^*(k) - i_{qs}^e(k-1)] + (r_s' - \frac{\sigma L_s}{T}) \\ & [i_{qs}^*(k-1) - i_{qs}^e(k-2)] + \omega_e \sigma L_s [i_{ds}^*(k-1) - i_{ds}^e(k-2)] \end{aligned} \quad (12)$$

$$\begin{aligned} v_{dff}^* = & v_{ds}(k-2) + \frac{\sigma L_s}{T} [i_{ds}^*(k) - i_{ds}^e(k-1)] + (r_s' - \frac{\sigma L_s}{T}) \\ & [i_{ds}^*(k-1) - i_{ds}^e(k-2)] - \omega_e \sigma L_s [i_{qs}^*(k-1) - i_{qs}^e(k-2)] \end{aligned} \quad (13)$$

Eqs. (12) and (13) require only the motor control voltage and current feedback for the calculation, there is no need to calculate rotor or stator flux. Also \bar{V}_{ff}^* calculation is not dependent upon the tuning of the field orientation control.

B. Feedback Control

Since the total voltage applied to the motor is:

$$\bar{V}_s(k) = \bar{V}_{ff}^*(k) + \bar{V}_{fb}^*(k), \quad \text{where } \bar{V}_{fb}^* \text{ is the voltage}$$

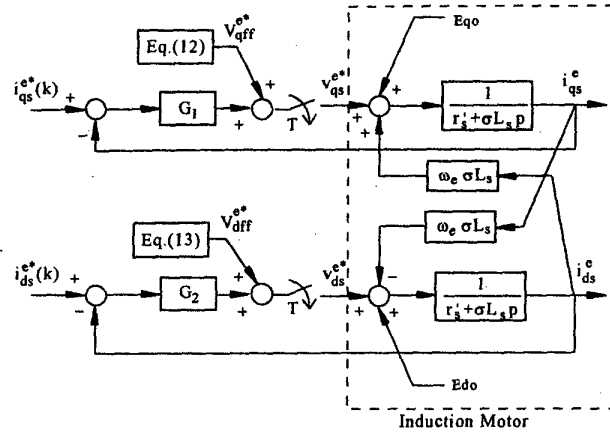


Fig. 1 Block diagram of the current controller

from the feedback controller. Combining (10) and (11) one can obtain an expression for $\bar{V}_{fb}^*(k)$,

$$\bar{V}_{fb}^*(k) = \left(\frac{1}{B} \bar{I}_s(k+1) - \frac{1}{B} \bar{I}_s^*(k+1) \right) - \left(\frac{\bar{A}}{B} \bar{I}_s(k) - \frac{\hat{A}}{B} \bar{I}_s^*(k) \right) - \left(\frac{1}{B} - \frac{1}{\bar{B}} \right) \bar{I}_s(k) + (\bar{A} - \hat{A}) \bar{I}_s(k-1) \quad (14)$$

The last two terms of Eq. (14) exist only when the motor parameters are not correctly estimated, whereas the first two terms are due to the inaccurate de-coupling voltage and the disturbance to the current control. When the modeling errors are treated as disturbance to the current control, then (14) can be simplified and expressed in state space form with the current error as state variables,

$$\Delta \bar{I}_s(k+1) = \bar{A} \Delta \bar{I}_s(k) + \bar{B} \bar{V}_{fb}^*(k) \quad (15)$$

where $\Delta \bar{I}_s(k) = \bar{I}_s^*(k) - \bar{I}_s(k)$ is the current error. Let the feedback control gain matrix be \bar{G}_c , and

$$\bar{G}_c = \begin{bmatrix} G_1 & 0 \\ 0 & G_2 \end{bmatrix} \quad (16)$$

Since $\bar{V}_{fb}^*(k) = -\bar{G}_c \Delta \bar{I}_s(k)$, then (15) becomes

$$\Delta \bar{I}_s(k+1) = (\bar{A} - \bar{B} \bar{G}_c) \Delta \bar{I}_s(k) \quad (17)$$

In order to regulate the current error $\Delta \bar{I}_s(k+1)$ to zero as fast as possible and to minimize the current ripple, deadbeat control scheme is used to design \bar{G}_c . That is to place all the poles of $\bar{A} - \bar{B} \bar{G}_c$ at the origin of the z-plane [8]. The resulting control gains are:

$$G_1 = -\frac{\hat{\sigma} \hat{L}_s}{T} \left(1 - \frac{\hat{r}_s'}{\hat{\sigma} \hat{L}_s} T - \omega_e T \right) \quad (18)$$

and

$$G_2 = -\frac{\hat{\sigma} \hat{L}_s}{T} \left(1 - \frac{\hat{r}_s'}{\hat{\sigma} \hat{L}_s} T + \omega_e T \right) \quad (19)$$

Note the feedback control gains are functions of motor resistance and leakage inductance, and are varying with ω_e . The variable feedback control gains allows for the same current transient response as the motor speed varies.

A block diagram depicting the current control scheme is shown in Fig. 1. The de-coupling voltages V_{qff}^* and V_{dff}^* are calculated using Eq. (12) and (13) respectively. Since the de-coupling voltages are derived using the digital control theory and its calculation requires only the motor current and control voltage, therefore the technique is simple for microprocessor implementation and is independent of the tuning of the field orientation control. The feedback controller is designed based on the deadbeat control theory to achieve fast dynamic response so that the motor current can track its reference precisely.

III. PARAMETER SENSITIVITY ANALYSIS

The influence of motor parameters to the de-coupling voltage is analyzed in this section. A one horsepower induction motor is used for the analysis, the motor parameters are shown in Appendix A. The motor current was set to its rated values and the stator frequency was varied from 0-120 Hz in the calculation. Variations of V_{qff}^* to the stator frequency when r_s , r_r , L_m and σL_s varied from 50% to 200% of their nominal values are calculated and shown in Fig. 2. Similarly, variation of V_{dff}^* vs. the stator frequency when r_s , r_r , L_m and σL_s varied from 50% to 200% of their nominal values are shown in Fig. 3. It can be seen from these plots that the d-axis voltage is significantly smaller than the q-axis voltage. This is

because the de-coupling voltage is dominated by the motor emf, and it can be seen from Eq. (5) that λ_{qr} is nearly zero while λ_{dr} is approximately constant. Therefore V_{dff}^{e*} is much smaller than V_{qff}^{e*} .

Also from Figs. 2-3, both V_{qff}^{e*} and V_{dff}^{e*} are not sensitive to L_m . Again this can be explained from Eq. (5) that the mutual inductance is not a dominate

parameter in the de-coupling voltage calculation. On the other hand, despite r_s and r_r caused offset errors to the de-coupling voltages, but the errors are small comparing to the total voltages. The errors caused by the variation of σL_s are proportional to the stator frequency, but again they are small in comparison to the total de-coupling voltages except when the stator frequency is very high.

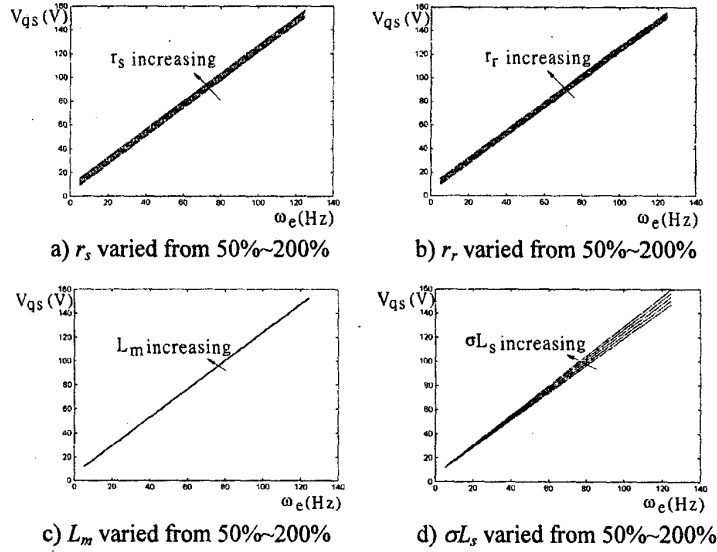


Fig.2 Influence of the motor parameter variations on the q -axis de-coupling voltage

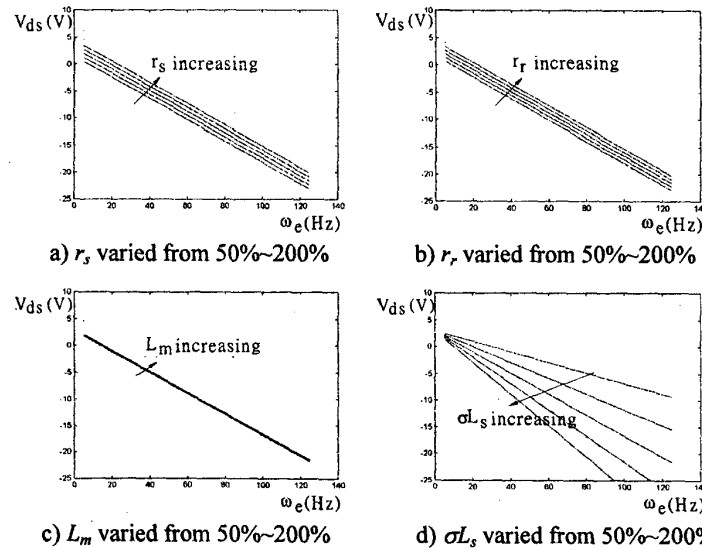
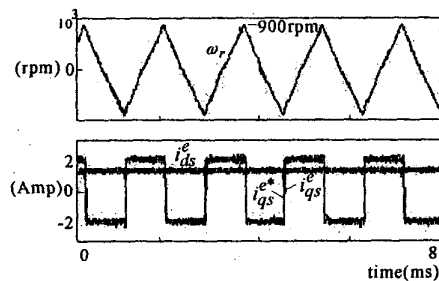
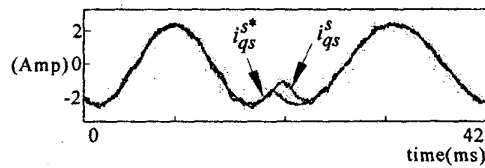
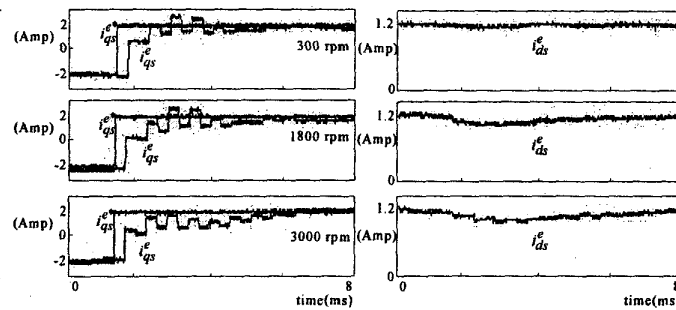
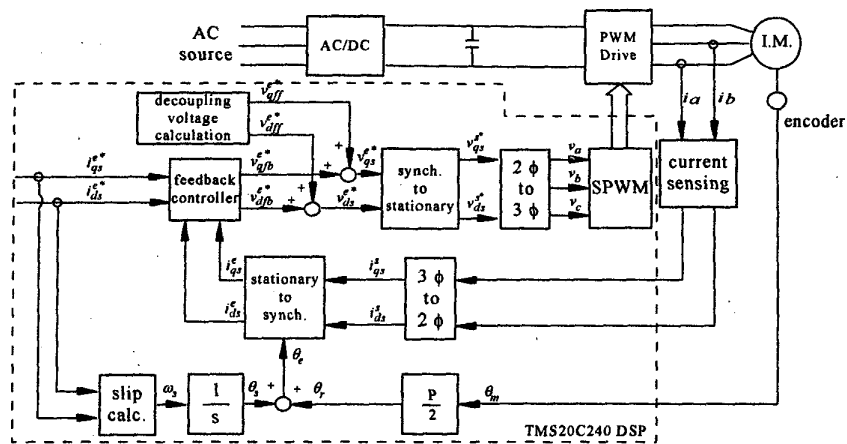


Fig. 3 Influence of the motor parameter variations on the d -axis de-coupling voltage



a) the slip gain was tuned correctly

Fig. 7 Synchronous frame d - q axis currents and speed response as i_{qs}^* reverse its direction at +900 rpm

IV. EXPERIMENTAL RESULTS

The digital current control scheme shown in Fig. 1 was implemented with a TMS320C240 DSP based development system to control the induction motor used for the analysis in the previous section. A rotor flux field orientation control was also implemented in the system to assist the verification of the current control scheme. Sinusoidal comparison PWM was used for the voltage modulation, the sampling rate was set to about 3.3 KHz. A schematic diagram of the experimental setup is shown in Fig. 4.

Fig. 5 shows the transient responses of i_{qs}^e and i_{ds}^e when the q-axis current command i_{qs}^{e*} reversed its direction from -2 to 2 Amps as the motor was running at 300, 1800 and 3000 rpm respectively. i_{ds}^{e*} was set to about half of its rated value on purpose to allow the motor to run up to the twice of its rated speed without having to weaken the field. It can be seen that in all three motor speeds, i_{qs}^e reached the reference value within four sampling periods, and then oscillated slightly before settling into their final values. Also at the instance i_{qs}^{e*} reversed its direction i_{ds}^e was disturbed slightly but returned to their original values in several sampling periods. The disturbance to i_{ds}^e was more noticeable at high speed. This is because the effective voltage available for current regulation is less when the motor is running at high speed. Fig. 6 shows the response of the stationary frame q-axis current and its reference input as i_{qs}^{e*} reversed its direction from -2 to 2 Amps and the motor was running at 1800 rpm. The i_{qs}^s response is consistent with the results shown in Fig. 5 since it has catch up with its reference in approximately four sampling periods.

Figs. 7 show the current and speed response of entire speed reversal cycles. In a), the slip gain was tuned to the correct value while in Fig. 6 the slip gain was tuned to 4 times the correct value. It can be seen from these plots that i_{qs}^e essentially overlapped with i_{qs}^{e*} in both figures. This has verified that the current control scheme is independent of the tuning of the field orientation.

V. CONCLUSIONS

A current control scheme combines a de-coupling voltage control and a deadbeat feedback control for voltage-fed PWM induction motor drive is presented in this paper. The de-coupling voltage is derived using the discrete state space equation and its

calculation requires only the motor current and control voltage. The feedback control is designed using the deadbeat control theory to minimize the response time such that the stator current can be adjusted instantaneously and precisely to its reference. From the parameter sensitivity analysis, modeling error of the motor leakage inductance will cause calculation error of the de-coupling voltage, but its influence is small. It was also demonstrated from the experimental verifications that the control scheme has good transient and static characteristics, and is not sensitive to the tuning of field orientation control.

APPENDIX A

The motor used in this paper was a 1 hp, 4 poles, 220VAC three phase induction motor, parameters are:

stator resistance r_s	3.0 Ω
rotor resistance r_r	2.7 Ω
leakage inductances L_{ls}, L_{lr}	0.008 H
mutual inductance L_m	0.18 H

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