PERFORMANCE IMPROVEMENT OF THE INDUCTION MOTOR DRIVE BY USING ROBUST CONTROLLER

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Abstract -: The transient response of the induction motor is obtained by using its d-q reference model. The transient response is improved by using optimal control technique because of the property of best possible control. By solving Ricatti equation, a controller gain matrix is developed such that the performance index is minimum. This gain matrix will give feedback control law. The controller will give control signal according to this law. The output is fed back and the response is analyzed. Thus the transient response is improved. This controller is robust against disturbances

Index terms – Induction motor, LQR, Stability

NOMENCLATURE

$\vec{V}_s, \vec{I}_s$	Stator voltage and current space vectors
$\vec{\psi}_s$	Stator flux space vector
$\vec{I}_r, \vec{\psi}_r$	Rotor current and flux space vectors
V_{ds}, V_{qs}	Stator voltages in d-q rotating ref. frame
I_{ds}, I_{qs}	Stator currents in d-q rotating ref. frame
ψ_{dr}, ψ_{qr}	Rotor fluxes in d-q rotating ref. frame
I_{ds}, I_{qs}	Stator currents in d-q rotor flux ref. frame
ω_r	Rotor speed (rad/s)
R_s, L_s	Stator resistance and self inductance
L_r	Rotor self inductance
τ_r	Rotor electrical Time constant
L_m	Magnetic inductance
P	Number of pole pairs
J	Total rotor inertia constant (Kgm^2)
F	Damping coefficient (Nms)
T_l	Load torque (Nm)
T_e	Electromagnetic torque (Nm)
σ	Coefficient of dispersion
ω_e	Synchronous speed (rad/s)

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INTRODUCTION

Induction machines have been the most widely used machines in fixed-speed applications for reasons of cost, size, reliability and efficiency. However, because of the involved model high nonlinearities, they require much more complex methods of control, more expensive and higher rated power converters than DC and permanent magnet machines. Nowadays, as a consequence of rapid advances in power electronics technology, vector control strategy based electrical ac drives have emerged as a powerful tool for high performance control of Induction machines. This control strategy can provide the same performance from an inverter driven Induction machine as is achieved from a separately excited DC machine.

In this thesis the author introduces a new controller called Linear Quadratic Regulator which is robust against external disturbances and provides excellent performance improvement with the improvement of stability margin. The system is simulated using Matlab and its characteristics and features are studied in this thesis.

d-q MODEL OF THE INDUCTION MOTOR

A two phase d-q model of an Induction machine rotating at the synchronous speed is introduced which will help to carry out the decoupled control concept to the induction machine. This model can be summarized by the following equations

$$\vec{V}_s = R_s \vec{I}_s + \frac{d\vec{\psi}_s}{dt} + j\omega_e \vec{\psi}_s \quad \ldots (1)$$

$$0 = R_r \vec{I}_r + \frac{d\vec{\psi}_r}{dt} + j(\omega_e - \omega_r)\vec{\psi}_r \quad \ldots (2)$$

The stator and rotor fluxes are given by the following relations:

$$\vec{\psi}_s = L_s \vec{I}_s + L_m \vec{I}_r \quad \ldots (3)$$

$$\vec{\psi}_r = L_m \vec{I}_s + L_r \vec{I}_r \qquad \ldots (4)$$

In equations 1 to 4, the voltages, currents and fluxes space vectors are function of the corresponding three-phase variables [3]. As an example, the stator current space vector is linked to the corresponding three phase currents by the following relation:

$$\vec{I}_s = \sqrt{(2/3)}\left(I_{as} + aI_{bs} + a^2 I_{cs}\right) \qquad \ldots (5)$$

Where a = e $^{j\,2\Pi/3}$. The produced electromagnetic torque is given by

$$T_e = \frac{3pL_m}{2L_r}\left(\psi \otimes \vec{I}_s\right) \qquad \ldots (6)$$

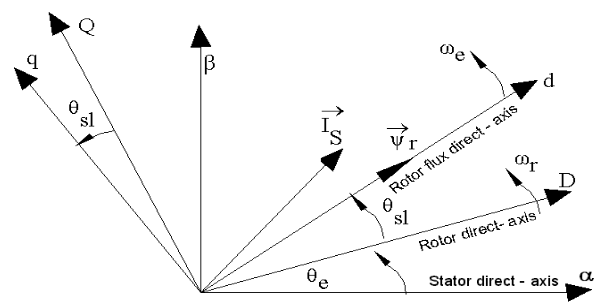

Figure 1. Reference frames and space vector representation

Using the d-q coordinate system, as illustrated in Figure 1, and separating the machine variables state vectors into their real and imaginary parts, the well-known Induction motor model expressed in terms of the state variables is obtained from equations 1 to 6, and is given by:

$$\frac{d}{dt}\begin{bmatrix} I_{ds} \\ I_{qs} \\ \psi_{dr} \\ \psi_{qr} \\ \omega_r \end{bmatrix} = \begin{bmatrix} -\left(\frac{R_s}{\sigma L_s} + \frac{1-\sigma}{\sigma\tau_r}\right)I_{ds} + \omega_e I_{qs} + \frac{L_m}{\sigma L_s L_r \tau_r}\psi_{dr} + \frac{L_m \omega_r}{\sigma L_s L_r}\psi_{qr} \\ -\omega_e I_{ds} - \left(\frac{R_s}{\sigma L_s} + \frac{1-\sigma}{\sigma\tau_r}\right)I_{qs} - \frac{L_m \omega_r}{\sigma L_s L_r}\psi_{dr} + \frac{L_m}{\sigma L_s L_r \tau_r}\psi_{qr} \\ \frac{L_m}{\tau_r}I_{ds} - \frac{1}{\tau_r}\psi_{dr} + (\omega_e - \omega_r)\psi_{qr} \\ \frac{L_m}{\tau_r}I_{qs} - (\omega_e - \omega_r)\psi_{dr} - \frac{1}{\tau_r}\psi_{qr} \\ \frac{p^2 L_m}{J L_r}\left(I_{qs}\psi_{dr} - I_{ds}\psi_{qr}\right) - \frac{F}{J}\omega_r - \frac{P}{J}T_l \end{bmatrix} + \begin{bmatrix} \frac{1}{\sigma L_s}V_{ds} \\ \frac{1}{\sigma L_s}V_{qs} \\ 0 \\ 0 \\ 0 \end{bmatrix} \qquad (7)$$

In (7), the coefficient of dispersion σ is given by:

$$\sigma = 1 - \frac{L_m^2}{L_s L_r}$$

As shown in Figure 1, the d-axis is aligned with the rotor flux space vector. Under this condition we have; $\psi_{qr} = 0$ and $\psi_{dr} = \psi_r$. Consequently, the induction motor model established in the rotor flux field coordinate is then given by the equations 9 to 12.

$$\frac{d}{dt}\begin{bmatrix} I_{ds} \\ I_{qs} \\ \psi_r \\ \omega_r \end{bmatrix} = \begin{bmatrix} -\left(\frac{R_s}{\sigma L_s} + \frac{1-\sigma}{\sigma\tau_r}\right)I_{ds} + \omega_e I_{qs} + \frac{L_m}{\sigma L_s L_r \tau_r}\psi_r \\ -\omega_e I_{ds} - \left(\frac{R_s}{\sigma L_s} + \frac{1-\sigma}{\sigma\tau_r}\right)I_{qs} - \frac{L_m \omega_r}{\sigma L_s L_r \tau_r}\psi_r \\ \frac{L_m}{\tau_r}I_{ds} - \frac{1}{\tau_r}\psi_r \\ \frac{p^2 L_m}{J L_r}\left(I_{qs}\psi_r\right) - \frac{F}{J}\omega_r - \frac{P}{J}T_l \end{bmatrix} + \begin{bmatrix} \frac{1}{\sigma L_s}V_{ds} \\ \frac{1}{\sigma L_s}V_{qs} \\ 0 \\ 0 \end{bmatrix}$$

$$T_e = \frac{pL_m \psi_r}{L_r} I_{qs} \qquad \ldots (10)$$

$$\psi_r = \frac{L_m}{(1 + s\tau_r)} I_{ds} \qquad \ldots (11)$$

$$\omega_e - \omega_r = \frac{L_m I_{qs}}{\tau_r \omega_r} \qquad \ldots (12)$$

In ordinary use, only stator voltages, currents and rotor speed are available for measurement. In this case, the d-q stator voltages and currents are obtained from the corresponding α−β stationary reference frame variables through an appropriate transformation involving rotor flux space vector angle θ_e**,** as shown in Figure 1. This transformation is given by:

$$\begin{bmatrix} x_d \\ x_q \end{bmatrix} = \begin{bmatrix} \cos(\theta_e) & \sin(\theta_e) \\ -\sin(\theta_e) & \cos(\theta_e) \end{bmatrix} \begin{bmatrix} x_\alpha \\ x_\beta \end{bmatrix} \qquad \ldots (13)$$

In equation 13, "x" is a voltage, a current or a flux. As mentioned before, θ_e is the rotor flux space vector angle. In direct vector control, the rotor flux is available for measurement or is estimated from measured stator voltages and currents. The rotor flux angle is then given by:

$$\theta_e = a\tan\frac{\psi_{\alpha r}}{\psi_{\beta r}}$$

The rotor flux amplitude is obtained by solving equation 11, and its spatial position is given by:

$$\theta_e = \int (\omega_r + \frac{L_m I_{qs}}{\tau_r \psi_r})dt$$

The Indirect vector control strategy can now satisfactorily be achieved since both amplitude of rotor flux vector and its spatial position are known. As in DC machines, the torque and the flux are controlled independently: The electromagnetic torque T_e is controlled by I_{qs} (torque producing current), and the flux is controlled by I_{ds} (flux producing current).

LQR QUADRATIC REGULATOR

A Linear Quadratic Regulator (LQR) is used to determine the SVFB K such that the Performance Index J is minimized. It comes under Optimal Control. It is called so because in every control step the performance index is reduced to a minimum. Furthermore, it has a comparable high robustness against parameter changes.

A system can be expressed in state variable form as

$$\dot{x} = Ax + Bu \tag{14}$$

With $x(t) \in R^n, u(t) \in R^m$. The initial condition is x (0). We assume here that all the states are measurable and seek to find a state-variable feedback (SVFB) control.

$$u = -Kx \tag{15}$$

That gives desirable closed-loop properties. The closed – loop system using this control becomes

$$\dot{x} = (A - BK)x + Bu = A_c x + Bu \tag{16}$$

With Ac the closed-loop plant matrix and u(t) the new command input. The output matrices C and D are not used in SVFB design. If there is only one input so that m=1, then Ackermann's formula gives a SVFB K that places the poles of the closed-loop system as desired. However, it is very inconvenient to specify all the closed-loop poles, and a technique is needed that works for any number of inputs. The optimal controllers require least control energy for control the system.

Since many naturally occurring systems are optimal, it makes sense to design man-made controllers to be optimal as well. To design a AVFB that is optimal, a term performance index (PI) is to be considered.

$$J = \frac{1}{2}\int_0^\infty x^T Qx + v^T Rvdt \tag{17}$$

Substituting the SVFB control into this yields

$$J = \frac{1}{2}\int_0^\infty x^T (Q + k^T Rk)xdt \tag{18}$$

The objective in optimal design is to select the SVFB K that minimizes the performance index J. The performance index J can be interpreted as an energy function, so that making it small keeps small the total energy of the closed-loop system. If both the state x (t) and the control input u (t) are weighted in J, so that if J is small, then neither x (t) nor u (t) can be too large. If J is minimized, then it is certainly finite, and since it is an infinite integral of x (t), this implies that x (t) goes to zero as t goes to infinity. This in turn guarantees that the closed-loop system will be stable.

The two matrices Q (an n x n matrix) and R (an m x m matrix) are selected by the design engineer. Depending on how these design parameters are selected, the closed-loop system will exhibit a different response. Generally speaking, selecting Q large means that, to keep J small, the state x (t) must be smaller. On the other hand selecting R larger means that the control input u (t) must be smaller to keep J small. This means that larger values of Q generally result in the poles of the closed-loop system matrix Ac = (A –BK) being further left in the s-plane so that the state decays faster to zero. On the other hand, the larger R means that less control effort is used, so that the poles are generally slower, resulting in larger values of the state x(t).

One should select Q to be positive semi-definite and R to be positive definite. This means that the scalar quantity $x^T Qx$ is always positive or zero at each time t for all functions x (t), and the scalar quantity $u^T Ru$ is always positive at each time t for all values of u (t). This guarantees that J is well-defined. In terms of eigenvalues, the eigenvalues of Q should be non-negative, while those of R should be positive. If both matrices are selected diagonal, this means that all the entries of R must be positive while those of Q should be positive, with possibly some zeros on its diagonal. Note that then R is invertible.

The use of Linear Quadratic Regulator (LQR) is to determine the SVFB K such that it minimizes the

Performance Index J. The word 'regulator' refers to tracker problems, where the objective is to make the output follow a prescribed (usually nonzero) reference command.

To find the optimal feedback K it is proceeded as follows. Suppose there exists a constant matrix P such that

$$\frac{d(x^T Px)}{dt} = -x^T (Q + k^T Rk)x \tag{19}$$

Then, substituting into equation (17) yields,

$$J = -\frac{1}{2}\int_0^\infty \frac{d(x^T Px)dt}{dt} = \frac{1}{2}x^T(0)Px(0) \tag{20}$$

Where it is assumed that the closed-loop system is stable so that x (t) goes to zero as time t goes to infinity. Equation (20) now implies that J is now independent of K. It is a constant that depends only on the auxiliary matrix P and the initial conditions.
Differentiating (17) and then substituting from the closed-loop state equation (14) it is seen that (17) is equivalent to

$$x^T (A_c^T P + PA_c + Q + k^T Rk)x = 0 \tag{21}$$

It has been assumed that the external control v(t) is equal to zero. Now note that the last equation has to hold for every x(t). Therefore, the term in brackets must be identically equal to zero. Thus, proceeding it is seen that

$$A^T P + PA + Q + k^T Rk - k^T B^T P - PBk = 0 \tag{22}$$

This is a matrix quadratic equation. Exactly as for the scalar case, one may complete the squares. Though this procedure is a bit complicated for matrices, suppose if

$$k = R^{-1}B^T P \tag{23}$$

Then, it results in

$$A^T P + PA + Q - PBR^{-1}B^T P = 0 \tag{24}$$

This result is of extreme importance in modern control theory. Equation (24) is known as the algebraic Riccati equation (ARE). It is a matrix quadratic equation that can be solved for the auxiliary matrix P given (A, B, Q, R). Then, the optimal SVFB gain is given by (23). The minimal value of the PI using this gain is given by (22), which only depend on the initial condition. This mean that the cost of using the SVFB (24) can be computed form the initial conditions before the control is ever applied to the system.
The design procedure for finding LQR feedback K is:
Select design parameter matrices Q and R
Solve the algebraic Riccati equation for P
Find the SVFB using

The matrixes Q and R can be found out by trial and error method or using GA technique. There are very good numerical procedures for solving the ARE. The MATLAB routine that performs this is named lqr (A, B, Q, R).The LQR design procedure is guaranteed to produce a feedback that stabilizes the system as long as some basic properties hold.

LQR THEOREM:

Let the system (A, B) be reachable. Let R be positive definite and Q be positive semi-definite. Then the closed loop system (A-BK) is asymptotically stable. Note that this holds regardless of the stability of the open-loop system. Recall that reachability can be verified by checking that the reachability matrix has full rank n.

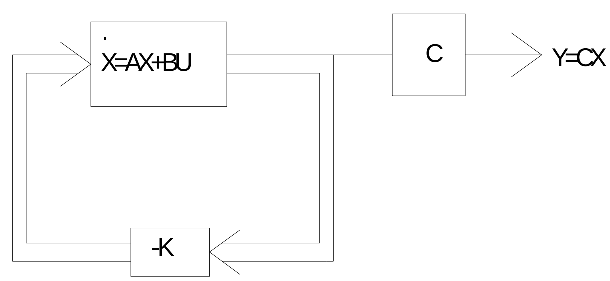

Figure2. System Block Diagram

SIMULATION RESULTS

Simulations, using Matlab-Simulink software package, have been carried out to verify the effectiveness of the proposed control method. The results are shown in figures 3. 4., & 5. Figure 3 shows the unit step response before and after applying controller. Figure 4 shows the current curves ids and iqs of the motor drive. Figure 5 shows the locations of poles before and after applying controller. It also shows how the stability is enhanced by modifying the pole locations.

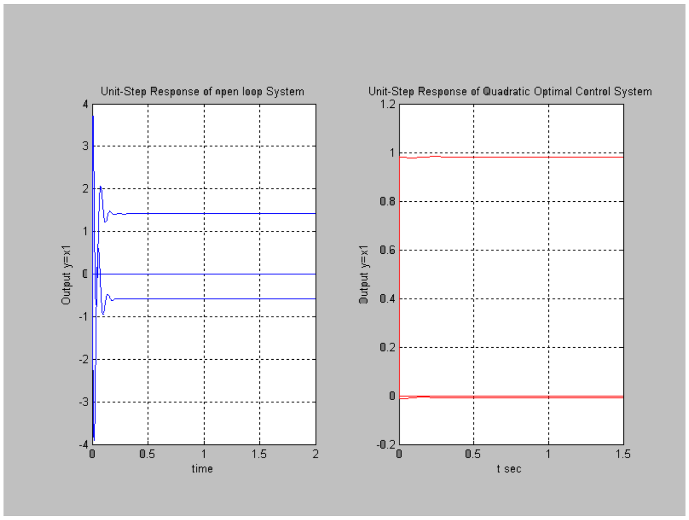

Figure 3. Response of system before and after applying LQR

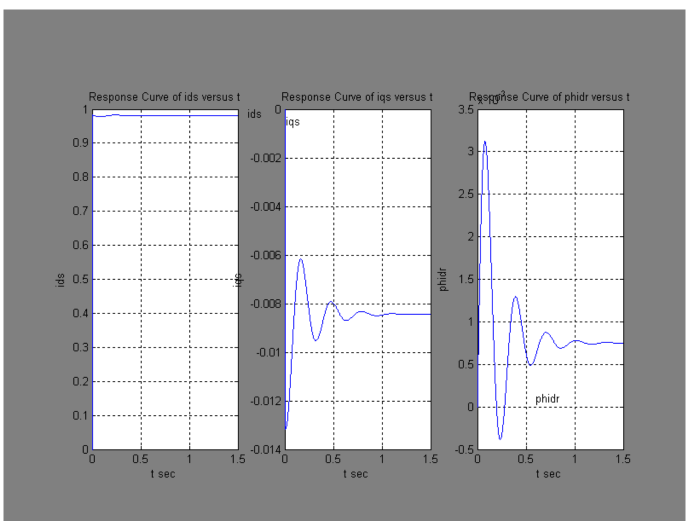

Figure 4. Response curves of ids, iqs and phidr

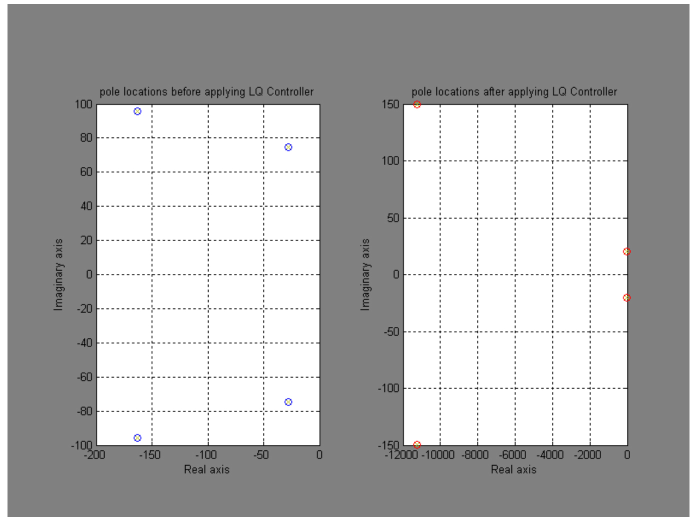

Figure 5. Pole locations before and after applying LQR

It is also observed that stability is also analyzed after applying the controller. The results show that the margin of stability also increases by incorporating the controller. The stability tests are carried out using h-infinity definition and Lyapunov's test for positive definiteness.

CONCLUSION

The simulation of LQR controlled induction motor drive is successfully implemented in this paper. The application of the controller and its response improvement contributing to the stability enhancement is studied. This paper can be further extended by comparing this controller performance with the existing controlling methods like PI and so on. This paper can be further extended to other types of drives also enhancing the performance.

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APPENDIX

Induction Motor Parameters

Un (V)	= 440 (Stator line voltage)
P_n (hp)	= 100 (Nominal output power)
R_s (Ω)	= 0.095
R_r (Ω)	= 0.075
L_s (mH)	= 16.5
L_r (mH)	= 16.4
L_m (mH)	= 16
f (Hz)	= 60
p	= 2
J ($kg.m^2$)	= 5