Abstract: This paper presents a new speed control method for the induction motor (IM) following the finite control set-model predictive control (FCS-MPC) strategy. It adopts the cascaded control scheme, which consists of an inner model predictive torque control (MPTC) loop and outer model predictive speed control (MPSC) loop using two individual cost functions. This control approach implements a maximum torque per ampere (MTPA) operation in a wide speed range, which includes flux-increased and flux-limited modes. The MPSC produces the required torque to drive the IM at the reference speed. This torque is taken as the input of the inner MPTC, which in turn generates the optimal finite set of control input voltage. The control signals of the two MPC are constrained with the maximum limits of the system. The state feedback is achieved with a standard Kalman Filter, which estimates the non-measured load torque. The proposed control system is implemented and validated in experimental environment. The behaviour of the control system is evaluated by applying reference and disturbance steps to the system in different operational modes.

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Keywords: FCS-MPC, Induction Motor, MTPA, Torque Control, Speed Control.

1. INTRODUCTION

In last decade, model predictive control (MPC) is mostly used for current and torque control of IM [1] [2]; and the linear PI controller is used in the outer loop for track the speed reference [3], [4]. In comparison to PI controller, MPC shows strong abilities to cope systematically with multivariable systems, uncertainties [5], and incorporation of the system constraints into control design [6].

A cascaded predictive control approach, which regulates the stator currents and the speed of the machine individually via two separate cost functions, has been presented in [6]-[9]. In [6], the torque reference constraint was handled below and above the base speed in a cascaded MPC. However, the system constraints have been calculated off-line. In [8], two cost functions with two different weighting matrices are used that increases the computational burden. Moreover, high ripples embedded in the current and torque waves. The strategy of [9] has not been validated by experimental implementation.

The finite control set-MPC (FCS-MPC) permits to obtain a good transient performance [2], [10]-[12]. However, this method causes high ripples in torque and current waves due to the finite-set of the control inputs. Research efforts have been done for minimizing the torque and current ripples in FCS-MPC using optimization technique for optimizing the time interval in which the selected voltage vector is applied to the inverter [13] [14].

In this paper, a new control method is presented for direct speed control of IM without use of PI controllers. This strategy is based on two MPC controllers to generate the requested control signals i.e. voltage and torque via inner and outer loops, respectively. These control inputs are generated via minimizing a two cascaded cost indices, which include the state errors of current, rotor flux and speed. The inner loop is designed for torque control of the IM for a given reference torque FCS-MPC with optimal modulation index. This scheme adopts two operational modes: flux-increased control (FIC) and flux-limited control (FLC). This scheme is designed under the constraints of rated voltage and rated current at high speeds and high load torques, respectively. The technique for torque control in the inner loop is proposed in [14]. Then, the outer loop for speed control is developed based on another MPC, which considers the mechanical dynamics of the IM. The generated torque reference is constrained by the maximum admissible value at each corresponding speed. In this paper, the constraints in the inner and outer loop systems are calculated on-line. To compensate the external load torque, a standard Kalman Filter is developed to estimate the applied load torque using the measured rotor speed.
2. DESCRIPTION OF THE PLANT

2.1 Electrical Dynamic Model of IM

The stator and rotor voltage equations of an IM in the synchronous reference frame are as follows:

\[ v_{dqs} = R_s i_{dqs} + \frac{d}{dt} \lambda_{dqs} + j\omega_e \lambda_{dqs} \]  
\[ 0 = R_r i_{dqr} + \frac{d}{dt} \lambda_{dqr} + j(\omega_e - \omega_r) \lambda_{dqr} \]

where \( j = \begin{bmatrix} 0 & 1 \\ -1 & 0 \end{bmatrix} \); \( R_s, R_r, L_s, L_r, L_m, \omega_e \), and \( \omega_r \) denote stator resistance, rotor resistance, stator inductance, rotor inductance, mutual inductance, and rotational speeds of the stator current and rotor, respectively. \( \lambda_{dqs} = L_s i_{dqs} + L_m i_{dqr} \) is the stator flux linkage, \( \lambda_{dqr} = L_r i_{dqr} + L_m i_{dqs} \) is the rotor flux linkage. Given the equations presented above, it is possible to represent the motor behavior using four internal states (stator currents and rotor fluxes) and two inputs (stator voltage vectors). The dynamic equations of IM (1)-(2), with rotor flux linkage components \( \lambda_{dqr} \) and stator current components \( i_{dqs} \) as state variables, can be expressed in state space form as

\[ x(t) = A_c(\omega_e(t)) x(t) + B_c u(t) \]  

The matrices \( A_c \) and \( B_c \) are state space matrices, given by

\[ A_c(\omega_e(t)) = \begin{bmatrix} \frac{1}{\alpha} (R_s + \beta^2 R_r) + j \omega_e & \frac{\beta R_r}{\alpha \omega_r} + j \frac{\beta}{\alpha} \omega_r \\ \beta R_r & -\frac{R_r}{\omega_r} \end{bmatrix} \]

\[ B_c = \begin{bmatrix} 1 \\ 0 \end{bmatrix} \]

where \( x(t) = [i_{dqs} \quad i_{dqr} \quad \lambda_{dqs} \quad \lambda_{dqr}]^T \) are the internal states, \( u(t) = [v_{dqs} \quad v_{dqr}]^T \) are the inputs, \( \alpha = L_s - i^2_{dqs}/L_r \) is the stator transient inductance, \( \beta = L_m/L_r \) is the rotor coupling factor. It is worth noting that the matrix \( A_c \) depends on the instantaneous value of mechanical speed \( \omega_r \). Hence, a discrete time-varying model that can be updated at every sampling interval with the new measured value of \( \omega_r \) is obtained. The discrete state space model of the IM is described using the Euler method with the sampling period \( T_s \). The prediction of the stator current and rotor flux at the next sampling instant can be obtained as:

\[ x[k+1] = Ax[k] + Bu[k] \]  

It can be rewritten as

\[ \begin{bmatrix} i_s[k+1] \\ \lambda_r[k+1] \end{bmatrix} = \begin{bmatrix} A_{11} & A_{12} \\ A_{21} & A_{22} \end{bmatrix} \begin{bmatrix} i_s[k] \\ \lambda_r[k] \end{bmatrix} + \begin{bmatrix} B_{11} \\ 0 \end{bmatrix} \begin{bmatrix} v_{dqs}[k] \\ v_{dqr}[k] \end{bmatrix} \]

where \( A = I_4 \times 4 + A_c T_s \) and \( B_{11} = B_c T_s \). Then, the electromagnetic torque of the IM in terms of the stator current, rotor flux and number of poles \( P \) is given as:

\[ T_e = \frac{3 P L_m}{2 L_r} (\lambda_{dqs} - \lambda_{dqr}) \]

2.2 Mechanical Dynamic Model of the Outer-Loop

The dynamic relationship that links the inner and outer loops is described by

\[ \dot{\omega} = (-B \omega + \omega_T - \omega_L)/J \]

where \( J \), \( B \), and \( \omega_T \) are inertia and friction coefficients, and load torque, respectively. The state form in time domain of the dynamic system is as

\[ \omega_r(k+1) = (1 - \frac{B_m T_s}{J})\omega_r(k) + \frac{p T_s}{J} \omega_L(k) - \frac{p T_s}{J} \omega_T(k) \]

where \( \omega(k) \) is the electrical speed, \( T_s \) is the sampling time of speed loop, and \( p \) is the number of pole pairs. Due to the discrete nature of digital control platforms, (9) is discretized by Euler approximation. The discrete mechanical state space can be described as

\[ x_m[k+1] = A_m x_m[k] + B_m u_m[k] + E_m d_m[k] \]

where \( x_m[k] = \omega_r[k] \) is the state variable, \( u_m[k] \) is the input control variable, and the disturbance term \( d_m \)

![Figure 1 IM drive employing the inner/outer-MPC for torque/speed control scheme](image_url)
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