

Implementation of Switching Instant Optimization on a Two-Level Inverter-Fed Induction Motor Drive Controlled using MPFC

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ABSTRACT-In order to get the high performance of induction motor drives model predictive torque control (MPTC) can be considered as one of the best schemes. When compared with direct torque control (DTC), MPTC is accurate in its voltage vector selection by employing system model directly with the finite switching states. However, it requires tuning work for stator flux weighting factor. This tuning work is time-consuming and tedious process. To solve this problem, this paper proposes model predictive flux control (MPFC) for two-level inverter fed induction motor (IM) drives. This is achieved by converting stator flux magnitude and torque into a single equivalent reference frame of stator flux vector. As only one stator flux is required for error tracking, the usage of weighting factor is eliminated as both have same reference. In this paper proposed MPFC with and without switching instant optimization both are implemented. The results obtained show the effectiveness of both the results.

INDEXING TERMS:Field-oriented control (FOC), Direct torque control (DTC), Model predictive control (MPC), Model predictive torque control (MPTC), Model predictive flux control (MPFC), Induction motor (IM).

1. INTRODUCTION

Field-oriented control (FOC) and direct torque control (DTC) are two methods used effectively for high performance control of induction motor (IM) drives. The direct torque control (DTC) is suitable for variable speed drives especially in cases where torque control is more important than speed control. In DTC, the voltage vector selection is done directly from predefined switching table according to the position of stator flux and error signs of both flux and torque. The advantages of the DTC are fast transient torque response and its simple structure. However, due to the hysteresis characteristics of control algorithm, DTC has two drawbacks. First, variable switching frequency and the second one is high flux and torque ripples. Among the available methods, MPC is recently introduced as an effective alternative to the conventional DTC and has drawn increased attention from both academic and industrial communities.

The application of model predictive control (MPC) in electrical drives includes model predictive current control (MPCC) and model predictive torque control (MPTC). MPTC is performed in stationary frame, hence it is much simpler. In MPTC, torque and stator flux are predicted in accordance to the system model. The obtained optimal output of controller is evaluated based on predefined optimization criteria.

MPTC can directly predict the evolution of variables concerned, such as torque and stator flux. A proper tuning of weighting factor for stator flux is required as both torque and

stator flux have different units and amplitudes. Tuning of weighting factor is still a time-consuming problem as it is done based on empirical procedures. By evaluating the influence of each possible voltage vector, the one minimizing the torque and flux errors is selected as the best voltage vector. By doing this the vector selected from MPTC is more accurate. However, as the number of voltage vectors provided by a two-level inverter is rather limited (only eight vectors), the torque and flux ripples obtained from conventional MPTC are higher than those obtained from a multi-level inverter. To eliminate the usage of weighting factor the relationship between torque and stator flux is investigated.

In order to improve the steady-state performance of MPTC, the sampling frequency has to be high, which requires high computational ability. In MPTC, the voltage vector selected is applied at next control period. For better steady state performance recently, MPTC is implemented with duty cycle control. In this control period is investigated in two parts by investing a non-zero vector and a zero vector. Furthermore, high amount of calculation is required in MPTC due to the predictions of both stator flux and current for each voltage vector. Hence, in MPTC improving the steady torque performance and reducing the control complexity have become important constraints. Even after applying duty cycle control still suffers from tuning of weighting factor as both stator flux and torque are still involved in cost function minimization.

In DTC, it has been known that high torque ripple is partly caused by the application of a single voltage vector during the whole control period. In MPTC, although the selected vector is more accurate and effective in reducing torque and flux errors, applying a single voltage vector during the whole control period still fails to reduce the torque and flux errors to the minimal value. So to eliminate the weighting factor in conventional MPTC, this paper proposes model predictive flux control (MPFC). The torque and flux vector are translated into a single reference new stator flux vector. As both torque and stator flux are now converted into a single reference frame they both have same dimension resulting in the elimination of weighting factor tuning. This makes implementation of MPFC much easier in designing of real time applications compared to MPTC. In this paper, implementation on low and medium two-level IM drives with less designing complexity is achieved with the proposed MPFC. We discussed about two types of MPFC in this paper based on the application of control vector. In the first method, only one voltage vector is applied during one control period. In the second method, to achieve better steady-state performance switching instant of selected voltage vector is investigated, through which torque ripple minimization is achieved by minimizing the tracking of stator flux vector error. Simulation results prove that, better steady-state

performance is achieved for the proposed MPFC especially in the low-speed range.

II. DYNAMIC EQUATIONS OF IM

The dynamic equations of IM with stator flux ψ_s and stator currents i_s as state variables can be expressed in stationary frame

$$\dot{x} = Ax + B\bar{u}(1)$$

where $X=[i_s \ \psi_s]^T$ are state variables, $u=us$ is the stator voltage vector, and

$$A = \begin{bmatrix} -\lambda[R_s L_r + R_r L_s] + j\omega_r & \lambda(R_r - jL_r \omega_r) \\ -R_s & 0 \end{bmatrix} (2)$$

$$B = \begin{bmatrix} \lambda L_r \\ 1 \end{bmatrix} (3)$$

R_s, R_r stator resistance, rotor resistance

L_s, L_r, L_m stator inductance, rotor inductance and mutual inductance.

ω_r electrical rotor speed

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

The prediction of the stator current and stator flux at the next sampling instant can be obtained from (1). To ensure good accuracy of the prediction, second-order Euler method is employed.

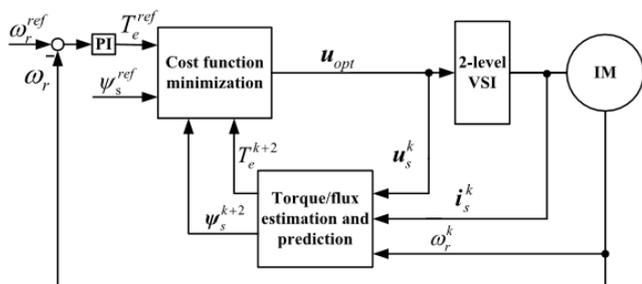


Fig.1. Control diagram of conventional MPTC.

$$\begin{cases} x_p(k+1) = x(k) + T_{sc}(Ax(k) + Bu_s(k)) \\ x(k+1) = x_p(k+1) + \frac{T_{sc}}{2} A(x_p(k+1) - x(k)) \end{cases} (4)$$

where T_{sc} is control period $x_p(k+1)$ is predictor-corrector of state vector;

$$x(k+1) = [i_s(k+1) \ \psi_s(k+1)]^T$$

is predicted state vector for stator current and stator flux. The main difference between $x_p(k+1)$ and $x(k+1)$ is that $x_p(k+1)$ is calculated as a rough approximation of $x(k+1)$ in the first step, which is equivalent to forward Euler integration. The value of $x_p(k+1)$ is refined in the second step to obtain more accurate result of $x(k+1)$, which is equivalent to trapezoidal integration method. It is possible to use more advanced method to obtain accurate discretization of (1), such as Cayley–Hamilton theorem in, but the computation burden will

be significantly increased, so it is not adopted in this paper. The electromagnetic torque can be predicted as

$$T_e(k+1) = \frac{3}{2} N_p \psi_s(k+1) \otimes i_s(k+1) (5)$$

where N_p is the number of pole pairs and \otimes represents cross product.

Omitting the tedious deduction process, the differentiation of torque T_e with respect to time t can be obtained from (1) and (5)

$$\frac{dT_e}{dt} = 1.5N_p \lambda L_m [-\lambda(R_s L_r + R_r L_s) Im(\psi_r^* \psi_s) - \omega_r Re \psi_r^* \psi_s + Im \psi_r^* u_s] (6)$$

where $*$ represents the conjugation of a complex vector. The equation in (6) provides a practical method to calculate the slope of torque, which is useful for the aim of improving torque performance in Section III. For simplicity, the torque slope caused by a zero vector u_0 and a nonzero voltage vector u_i are represented by s_0 and s_i , respectively, which are expressed as

$$s_0 = \left. \frac{dT_e}{dt} \right|_{u_s=0} = 1.5N_p \lambda L_m [-\lambda(R_s L_r + R_r L_s) Im(\psi_r^* \psi_s) - \omega_r Re \psi_r^* \psi_s] (7)$$

$$s_i = \left. \frac{dT_e}{dt} \right|_{u_s=u_p} = s_0 + 1.5N_p \lambda L_m (\psi_r^* u_i) (8)$$

III. CONVENTIONAL MPTC

The MPTC control diagram considering one-step delay compensation is shown in fig1, where to generate torque reference a PI speed controller is used. From the system model, conventional scheme predicts torque and flux vectors from all the available voltage vector provided by the inverter and from them by the minimization of cost function the best voltage vector is determined consisting of both flux and torque error. The control diagram consisting of A. Flux and torque estimation or prediction B. Cost function minimization C. Machine magnetization

A. State Estimation

To maintain good performance of MPTC the flux and torque estimation should be accurate. A full order observer is adopted and it is having good accuracy over a wide speed range. The error feedback of stator current is introduced, the accuracy of estimation is increased and the observer is more robust against the motor parameter variations. The mathematical model of the observer is based on the IM model in (1), which is expressed as

$$\frac{d\hat{x}}{dt} = A\hat{x} + Bu + G(i_s - \hat{i}_s) (9)$$

Where $\hat{x} = [\hat{i}_s \ \hat{\psi}_s]^T$ are the estimated state variables.

The observer poles of IM drives is proportional to the poles of IM and its factor is greater than one ($k>1$) which produces high imaginary part at high speed and is harmful to the system stability. To remedy these problems the imaginary part of the observer poles should remain unchanged and real part of observer poles should be changed to left in the complex plane

compared to the poles of IM. However, this leads to complicated expressions of observer gains. so to eliminate all those problems a very simple constant gain matrix G is employed to improve stability of the observer and it can be expressed as

$$G = - \left[\begin{array}{c} 2b \\ b/(\lambda L_r) \end{array} \right] \quad (10)$$

where b is negative constant gain. This pole placement method can improve the convergence and stability of observer especially at high speed and is simple to implement.

B. Vector Selection

In conventional MPTC the stator flux and electromagnetic torque at $(k + 1)$ th instant are predicted using predictive model, with the variables of $i_s(k)$ and $\psi_s(k)$ as the initial states. The aim of the cost function is to force both stator flux and torque to track their respective reference value. In other words error between estimated value and reference value of stator flux and torque should be minimized.

$$J = |T_e^{ref} - T_e^{k+1}| + k_\psi \left| |\psi_s^{ref}| - |\psi_s^{k+1}| \right| \quad (11)$$

where k_ψ denotes the weighting factor for stator flux. The tuning of k_ψ is a non-trivial and time consuming process. To assign the same weight on torque and stator-flux, k_ψ is usually chosen as

$$k_\psi = \frac{T_n}{|\psi_{sn}|} \quad (12)$$

where T_n is the rated torque and $|\psi_s|$ is the rated amplitude of stator flux. It should be noted that the weighting factor k_ψ in (12) is only used as a starting point for tuning work and the final practical value of k_ψ may be larger than that in (12).

As can be seen from the first row of matrix A in (2), the prediction of $i_s(k + 1)$ is complicated. As the stator current is not directly used in the cost function of conventional MPTC, the prediction of stator current may be eliminated for the aim of control complexity reduction. To predict the torque at $(k + 1)$ th instant without the prediction of stator current, (5) is replaced by

$$T_e(k + 1) = \frac{3}{2} N_p \psi_s(k + 1) \otimes i_{s\lambda}(k) \quad (13)$$

where $i_{s\lambda}(k)$ is calculated based on the variables at k th instant. In order to reduce this delay, $(k + 2)$ th instant variables are predicted for the evaluation of cost function rather than $(k + 1)$ th instant. The cost function can be represented as (14) after considering the unit step delay can be written from (12) as follows

$$J = |T_e^{ref} - T_e^{k+2}| + k_\psi \left| |\psi_s^{ref}| - |\psi_s^{k+2}| \right| \quad (14)$$

For two-level IM drives, There are eight discrete voltage vectors available: u_0, u_1, \dots, u_7 for a two-level inverter-fed IM drive as shown in fig.3. The best voltage vector is selected after the evaluation of cost function for all the available vectors and the best one is applied at the next control period.

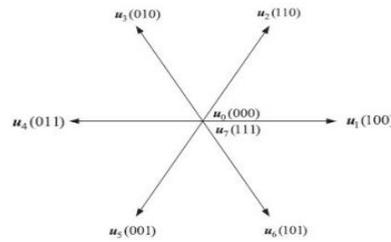


Fig. 2. Voltage vector and corresponding switching states of two level inverter.

IV. PROPOSED MPFC

A. Principle of Basic MPFC

To achieve simultaneous control of torque and stator flux a proper weighting factor for stator flux is necessary as torque and stator flux are combined into a single cost function. Tuning of weighting factor is not only a time-consuming process but also lacking of proper theoretical design it results in a nontrivial tuning work. In this we design a new stator flux vector reference based on IM model in equivalence with the original torque and stator flux vectors. Now both quantities are in single reference weighting factor is eliminated in proposed MPFC.

When the rotor speed is below rated speed, the magnitude of stator flux reference ψ_s^{ref} is set to the rated value ψ_s^{ref} in this paper,

$$|\psi_s^{ref}| = \psi_s^{ref} \quad (15)$$

The torque can be expressed as a cross product of stator -flux and rotor flux ,which is expressed as,

$$T_e = \frac{3}{2} p \lambda L_m (\psi_r \otimes \psi_s) \quad (16)$$

From (16), if the rotor-flux ψ_r is already known, the references of torque and stator-flux should satisfy with the following equation

$$T_e^{ref} = \frac{3}{2} p \lambda L_m (\psi_r \otimes \psi_s^{ref}) \quad (17)$$

Based on (15) and (16), stator flux reference ψ_s^{ref} is expressed by ψ_s^{ref} and T_e^{ref} as follows

$$\psi_s^{ref} = \psi_s^{ref} \cdot \exp(j \cdot \angle \psi_s^{ref}) \quad (18)$$

$$\psi_s^{ref} = \angle \psi_r + \arcsin \left(\frac{T_e^{ref}}{\frac{3}{2} p \lambda L_m |\psi_r^{ref}| |\psi_s^{ref}|} \right) \quad (19)$$

Similarly, the stator flux vector reference ψ_s^{ref} at $(k+2)$ th instant has to be determined for the delay compensation in the conventional MPTC. To know the value of ψ_s^{ref} we first need to know the information of T_e^{ref} and ψ_r at $(k+2)$ th instant. If the sampling is very much higher than the band-width of outer speed control loops (10) then the future reference of torque can be considered to be approximately equal to the value of present state of reference. The value of ψ_r at $(k+2)$ th instant can be predicted by the following procedure. The rotor

flux at (k+1)th instant is predicted first from (k+1)th stator flux and stator current as follows,

$$\psi_r^{k+1} = \frac{L_r}{L_m} \psi_s^{k+1} - \frac{1}{\lambda L_m} i_s^{k+1} \quad (20)$$

where ψ_s^{k+1} and i_s^{k+1} are predicted from (4).

ψ_r^{k+1} represent rotor flux at (k+1)th instant and ψ_s^{k+1} and i_s^{k+1} represent stator flux and current at (k+1)th instant. Now using the value of ψ_r^{k+1} we should obtain the value of ψ_r^{k+2} with the help of (9) and (19) represented by following equation

$$\psi_r^{k+2} = \psi_r^{k+1} + T_{sc} \left(R_r \frac{L_m}{L_r} i_s^{k+1} - \left(\frac{R_r}{L_r} - j \omega_r \right) \psi_r^{k+1} \right) \quad (21)$$

Based on the predicted value of ψ_r^{k+2} from (21), the phase angle reference of ψ_s^{ref} at (k+2)th instant is predicted as follows

$$\angle \psi_s^{ref} = \angle \psi_r^{k+2} + \arcsin \left(\frac{T_e^{ref}}{\frac{3}{2} P \lambda L_m |\psi_r^{k+2}| |\psi_s^{ref}|} \right) \quad (22)$$

Final stator flux reference ψ_s^{ref} can be calculated using (18) and (21) which is equivalent to original stator flux amplitude reference and torque reference. After calculating that we can derive a cost function similar to (14) which can force stator flux ψ_s to track its reference, which is expressed as

$$J_1 = |\psi_s^{ref} - \psi_s^{k+2}| \quad (23)$$

From equation (23) it is clear that cost function minimization does not require weighting factor. The prediction of stator flux at (k+2)th instant is as follows,

$$\psi_r^{k+2} = \psi_r^{k+1} + (u_s^{k+1} - R_s i_s^{k+1}) T_{sc} \quad (24)$$

where ψ_s^{k+1} and i_s^{k+1} are predicted from (9) to compensate the influence of unit step delay. To summarize the process of the proposed MPFC, first we have to measure dc-link voltage, rotor speed at k th instant and stator current. Now estimate the stator flux and predict i_s^{k+1} and ψ_s^{k+1} with the help of their initial states. The desired phase angle ψ_s^{ref} using (22) after predicting the values of ψ_r^{k+2} from (19) and ψ_s^{k+2} from (24). Final new stator flux reference in accordance to (18) and it is now substituted into (23) to select the best voltage vector minimizing.

B. Improved MPFC with Switching Instant Optimization

From equation (23) it is clear that the weighting factor of stator flux is eliminated by transforming the torque and stator flux references into a single reference frame so that they both have same unit i.e., into an equivalent reference of stator flux vector. With this practicability of MPFC is improved as offline tuning work is eliminated. However, applying selected voltage vector during the whole control period is still being considerable torque ripple and current harmonics which is similar to conventional MPTC.

In the improved MPFC with switching instant optimization, to improve the steady-state performance of the scheme further the switching instant of the selected voltage vector is being optimized. The old voltage vector during the last control period will be applied first in proposed MPFC followed by the selected voltage vector. This is different from the MPTC with duty cycle control which applies the selected voltage vector at the beginning of next control period. The stator flux at the end of next control period can be represented with the following equation

$$\psi_s^{k+2} = \psi_s^{k+1} + f_{old}^k t_{opt} + f_i^{k+1} (T_{sc} - T_{opt}) \quad (25)$$

where f_{old} and f_i^{k+1} ($i \in \{0, 1, \dots, 7\}$) are slopes of stator flux vector u_{old} applies at previous period and a selected voltage vector u_{si}^{k+1} , respectively. t_{opt} is the optimal duration of u_{old} . The stator flux slope u_{si}^{k+1} is assumed to be constant for a short period of time and can be calculated as follows

$$f_i^{k+1} = \frac{d\psi_{s1}}{dt} = u_{si}^{k+1} - R_s i_s^{k+1} \quad (26)$$

On substituting (25) into (23), the cost function is considered as a function of t_{opt} , namely $J_1 = J_1(t_{opt})$. Minimizing (23) is equivalent to solving the following equation:

$$\frac{dJ_1(t_{opt})}{dt_{opt}} = 0 \quad (27)$$

On combining (25) to (27), the optimal switching instant is finally derived as

$$t_{opt} = \frac{(\psi_s^{ref} - \psi_s^{k+1} - f_i^{k+1} T_{sc}) \odot (f_{old} - f_i^{k+1})}{|f_{old} - f_i^{k+1}|^2} \quad (28)$$

where \odot represents the dot product of two complex vectors. t_{opt} is limited to the range of $[0, T_{sc}]$ for the aim of protection in practical application.

The optimal switching instant t_{opt} is calculated for each feasible voltage vector ψ_s^{k+1} ($i \in \{1, 2, \dots, 7\}$) according to (28). After determination of t_{opt} for a given voltage vector, stator flux at (k+2)th instant is predicted as follows

$$\psi_{si}^{k+2} = \psi_s^{k+1} + (u_{old} - u_{si}^{k+1}) t_i + (u_{si}^{k+1} - R_s i_s) T_{sc} \quad (29)$$

The value of ψ_{si}^{k+2} is evaluated by the cost function from the equation (23) to determine the best voltage vector and its optimal duration.

At the end of next control period, by following the procedure mentioned above the minimal tracking of stator flux vector is achieved. However, there is high possibility of having high deviation of stator flux at switching instant may occur. We can observe this from the two possible stator flux trajectories shown in Fig. 3.

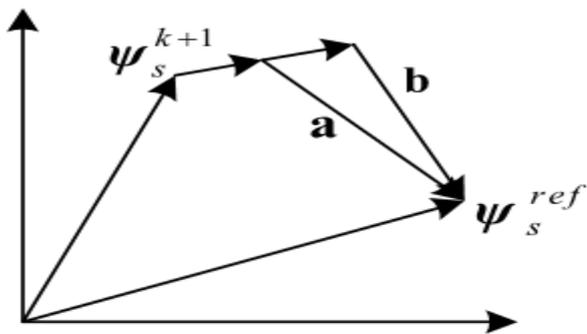


Fig. 3. Two trajectories of stator flux during one control period.

Although both stator flux trajectories (shown as a and b in Fig 3) can reach the stator flux reference at the end of control period, the trajectory a is preferred because it is smoother than b and also has small deviation from the reference vector during the whole control period.

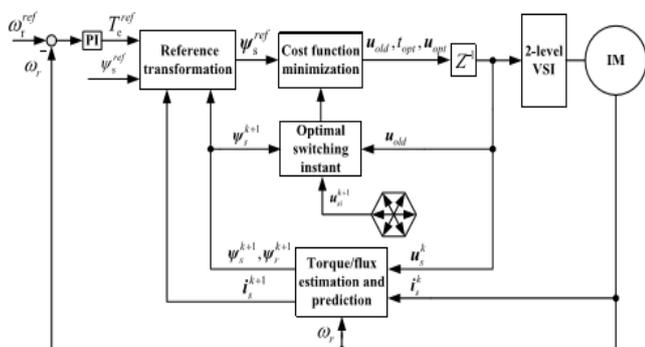


Fig.4. Control diagram of proposed MPFC with switching instant optimization.

To prevent high deviation of stator flux during the control period, a penalty on deviation at the switching instant is added in cost function (23), namely

$$j_2 = |\psi_s^{ref} - \psi_s^{k+2}| + |\psi_s^{ref} - |\psi_s^t|| \quad (30)$$

where stator flux ψ_s^t is the stator flux vector at the optimal switching instant and it can be obtained as follows:

$$\psi_s^t = \psi_s^{k+1} + u_{old}t; \quad (31)$$

Hence, after ψ_{si}^{k+2} and ψ_s^t are obtained from (29) and (31) for a given voltage vector u_{si}^{k+1} are evaluated from the cost function represented in (30). The best voltage vector u_{opt} with its optimal switching instant t_{opt} minimizing is selected among all the voltage vectors together with their optimal switching instants while evaluating with their cost function and applied in the next control period. The overall control diagram of proposed MPFC is shown in Fig.4.

TABLE I

MACHINE AND CONTROL PARAMETERS

| | | |
|--------------------------|--------------|----------------|
| DC-bus voltage | U_{dc} | 540 V |
| Rated power | P_N | 2.2 kW |
| Rated voltage | U_N | 380 V |
| Rated frequency | f_N | 50 Hz |
| Number of pole pairs | N_p | 2 |
| Stator resistance | R_s | 3.126 Ω |
| Rotor resistance | R_r | 1.879 Ω |
| Mutual inductance | L_m | 0.221 H |
| Stator inductance | L_s | 0.230 H |
| Rotor inductance | L_r | 0.230 H |
| Sampling period | T_{sc} | 50 μs |
| Flux amplitude reference | $ \psi_s^* $ | 0.91 Wb |

IV. SIMULATION RESULTS

The proposed MPFC is simulated in the environment of MATLAB/Simulink to validate its effectiveness. The machine and control parameters are listed in Table I. For convenience, the basic MPFC without flux weighting factor is referred as method I and the improved MPFC with switching instant optimization is referred as method II in the following text respectively.

Fig.5. and Fig.6.shows the starting response from standstill to 1500r/min for both methods. The stator flux is first established using pre-excitation and during the acceleration stage the torque is limited to 120% rated value (16.8N.m). At t=0.4sec an external load with rated value (14N.m) is applied suddenly to the machine. From top to bottom, the curves shown in Fig.5 and Fig.6. are speed, torque, stator flux and stator current respectively. It is clearly seen that the proposed MPFC works well over a wide speed range and exhibits strong robustness against load disturbance. Similarly dynamic responses can be observed in method II, but with much lower torque and current ripples.

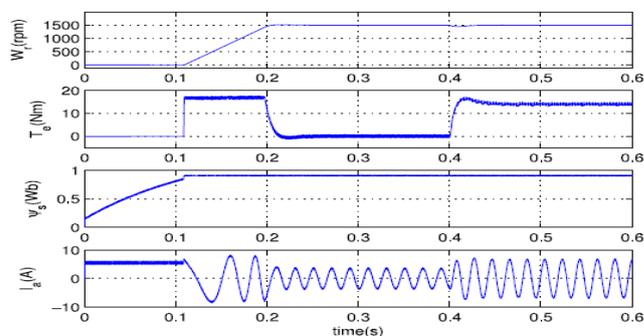


Fig.5. Simulated starting responses from standstill to 1500r/min for method I

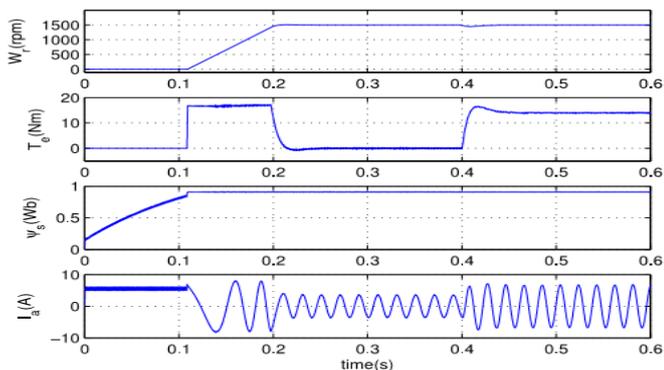


Fig.6. Simulated starting responses from standstill to 1500r/min for method II

A more detailed steady-state wave-form of torque and stator flux with 100% rated torque are observed in Fig.7 and Fig.8. It can be observed clearly from the result that both torque and flux ripples in method II are much lower than those in method I. With this we can confirm the effectiveness of method II with switching instant optimization.

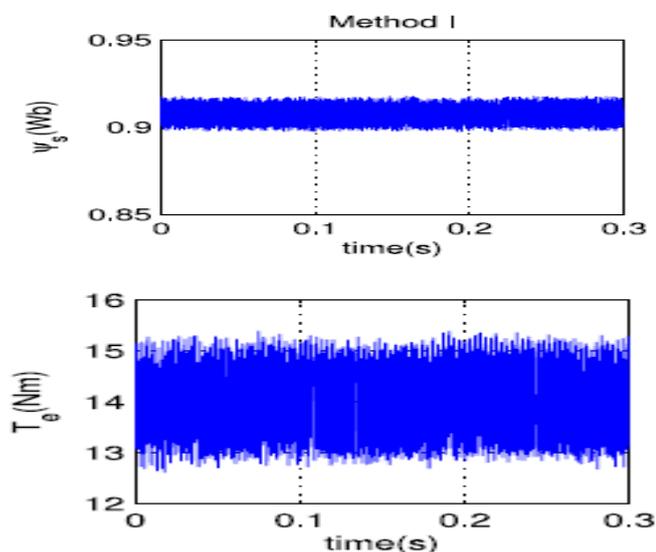


Fig. 7. Simulated steady state performance of method I

To know the influence of machine parameters variations on performance of system observe Fig.9. and Fig.10. Simulated responses of method II at low and high speeds when stator and rotor resistances are increased by 50% and 100% respectively are represented in Fig.9 and Fig.10.

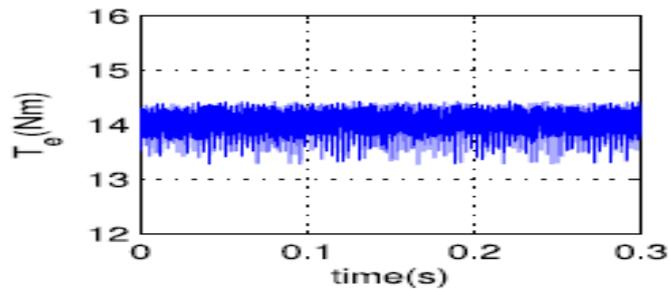
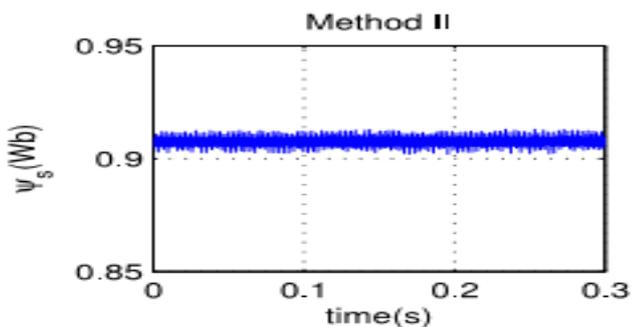


Fig.8. Simulated steady state performance of method II

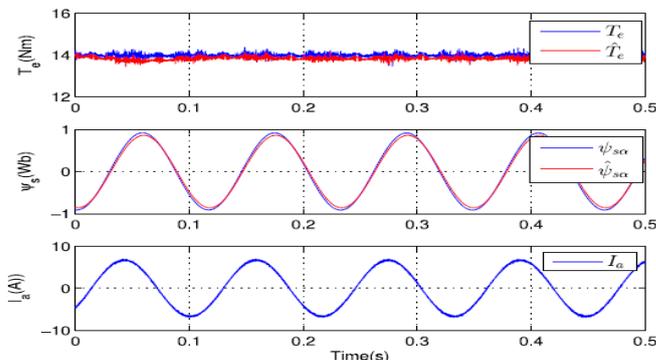


Fig. 9. With mismatched machine parameters of method II with both stator and rotor resistances are increased by 50% at 150r/min

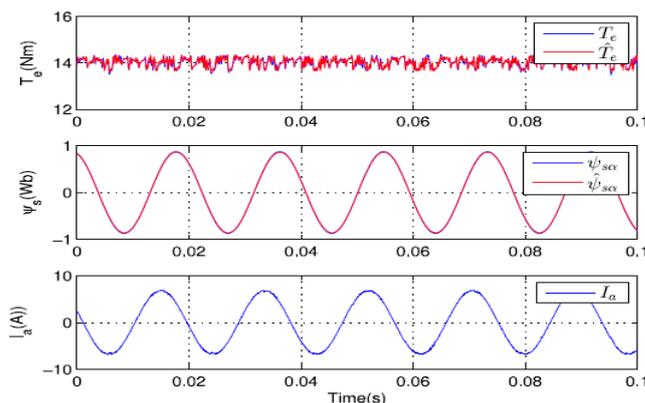


Fig. 10. With mismatched machine parameters of method II with both stator and rotor resistances are increased by 100% at 1500r/min

From the simulated results it is clear that system works well at both low and high speeds, even if the stator and rotor resistances vary significantly from their actual value. Current is very sinusoidal in shape. Only very minor tracking error of stator flux can be observed in low-speed operation and the flux error at high speed is negligible as we use only a closed-loop full-order observer is employed. The simulation results prove that the proposed method has some robustness against machine parameter variations. Hence, method II is more effective in terms of torque, stator flux and current ripples. This shows us that method II would be more helpful in applications requiring higher steady state performance, while in applications where low switching frequency is major concern method I can be preferred.

V. CONCLUSION

In two-level inverter fed IM we have limited voltage vectors and it can be applied to conventional MPTC then it gives high torque ripples at steady state conditions. To improve the steady state performance, torque ripples and to eliminate the requirement of tuning of stator flux weighting factor we introduce switching instant optimization in MPFC. This is achieved by a proper investigation about the stator flux and torque relation-ship in a two-level IM drive. A new stator flux reference vector is formulated using original torque and flux references in which both stator flux and torque are brought together into a single unit. Thus eliminating tuning of weighting factor is achieved. We have two different methods in this paper. One is mainly related to the low switching frequency and second method is entirely on steady-state performance improvement without having impact on torque or flux ripples and current harmonics. By observing the performance with much lower sampling frequency proposed MPFC gives better performance and also improves the practicability of the system.

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