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Performance Analysis of DTC Control CSI Fed IM Drives with Stator Resistance Compensation

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Abstract. Current source inverter (CSI) fed drives are employed in high power applications. The conventional CSI drives suffer from drawbacks such as harmonic resonance, unstable operation at low speed ranges, and torque pulsation. CSI fed drives with Direct Torque Control (DTC) has drawn the attention of the motor drives designers because its implementation requires no position sensor. Crucial to the success of this scheme is the estimation of electromagnetic torque and stator flux linkages using the measured stator voltages and currents. The estimation is dependent only on one machine parameter, stator resistance. The variation of the stator resistance, deteriorates the performance of the drive by introducing errors in the estimated flux linkage's magnitude and its position and hence in the electromagnetic torque. Resistance change also skews the torque linearity thus making the motor drive a less than ideal torque amplifier. Parameter compensation using stator current phasor error has been proposed in literature. To obtain the stator current phasor error, the stator current reference is required which is not usually available in direct torque control schemes. An analytical derivation of the stator current phasor reference is derived systematically from the reference electromagnetic torque and flux linkages. The error between the stator current phasor reference and its measured value is a measure of the stator resistance variation from its set value. For the first time, it is demonstrated in this paper that DTC motor drive system can become unstable when the set value of the stator resistance in the controller is higher than the stator resistance in the machine. Hence parameter adaptation is not only important for torque linearity but also for stability of the system is shown in this paper.

Keywords: Current Source Inverter, Direct Torque Control, Stator Estimation.

1. Introduction :

Even though voltage source inverter (VSI)-fed drives are most widely used, current source inverter (CSI)-fed drives find application [1] in high power drives such as fan drives, where fast dynamic response is not needed, because of the following advantages.

- Inherent four quadrant operation
- Reliability

A few control schemes have been proposed to overcome the parameter sensitivity, which restricts the speed control range of the drives. Partial but operating frequency dependent hybrid flux estimator has been proposed for stator resistance tuning [2], which has the problem of convergence and slowness of response. Adjustment of the stator resistance based on the difference between the flux current and its command [3] has the problem in identifying the actual flux current. Finding stator resistance based on the steady state voltage equation [4] has the shortcoming of using direct axis flux linkages which itself has been affected by stator resistance variations.

2. CSI-fed Induction Motor using DTC:

In a direct torque controlled induction motor drive supplied by current source inverter (Figure 1). It is possible to control directly the modulus of the rotor flux linkage space vector through the rectifier voltage, and the electromagnetic torque by the supply frequency of the CSI [5]. The inputs to the optimal switching table are the output of a 3-level hysteresis comparator and the position of the rotor flux-linkage space vector. As a result, the optimal switching table determines the optimum current switching vector of current source inverter. The main features of DTC can be summarized as follows.

DTC operates with closed torque and flux loops but without current controllers.

DTC needs stator flux and torque estimation and, therefore, is not sensitive to rotor parameters.
DTC is inherently a motion-sensor less control method.
DTC has a simple and robust control structure;

However, The performance of DTC strongly depends on the quality of the estimation of the actual stator flux and torque.

The input variables of the proposed algorithm are motor torque T_e* and rotor flux amplitude Ψ_r*, as in the case of basic DTC. Control variables are current components in synchronous reference frame *isd** and *isq** and phase angle between them (Φ*s*). D-axis component *isd** is determined as the output of the PI rotor flux controller, while q-axis component *isq** is calculated from the input variables and motor parameters:

$$i_{sq}^* = \frac{2.L_r}{3.p.L_m.\Psi_r^*} T_e^* \quad (1)$$

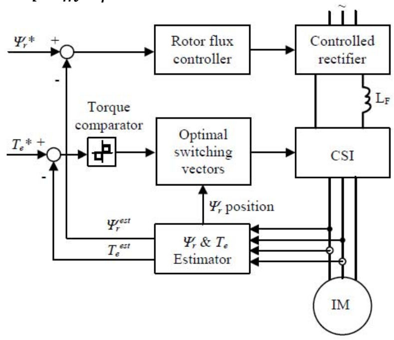

Fig. 1: Basic scheme of DTC CSI-fed IM drive

where *Lr* is rotor inductance, *Lm* is mutual inductance and *p* denotes pair of poles. Phase angle Φ*s* and rectifier reference current are obtained as a result of rectangular to polar coordinate transformation:

$$\Phi_s = \arctan\left(\frac{i_{sq}^*}{i_{sd}^*}\right) \quad (2)$$

$$i_{ref} = \sqrt{\left(i_{sd}^*\right)^2 + \left(i_{sq}^*\right)^2} \quad (3)$$

The induction machine stator and rotor flux equations in terms of space vectors, written in a synchronous reference frame, are:

$$\overline{\Psi}_s = L_s.\bar{\imath}_s + L_m.\bar{\imath}_r \quad (5)$$

$$\overline{\Psi}_r = L_r.\bar{\imath}_r + L_m.\bar{\imath}_s \quad (6)$$

Substituting *ir* in (4) and (5) the reference value of the stator flux vector in synchronous frame is determined, knowing the reference rotor flux and reference stator current:

$$\Psi_{s\alpha\beta}^* = \frac{L_m}{L_r}.\Psi_{r\alpha\beta}^* + \frac{L_s.L_r - L_m^2}{L_r}.i_{s\alpha\beta}^* \quad (7)$$

Where:

$$\begin{bmatrix}\Psi_{r\alpha}^*\\ \Psi_{r\beta}^*\end{bmatrix} = \begin{bmatrix}cos\theta_e & -sin\theta_e\\ sin\theta_e & cos\theta_e\end{bmatrix}\begin{bmatrix}\Psi_r^*\\ 0\end{bmatrix}$$

$$\begin{bmatrix}\Psi_{s\alpha}^*\\ \Psi_{s\beta}^*\end{bmatrix} = \begin{bmatrix}cos\theta_e & -sin\theta_e\\ sin\theta_e & cos\theta_e\end{bmatrix}\begin{bmatrix}i_{sd}^*\\ i_{sq}^*\end{bmatrix}$$

and θ*e* is rotor flux angle, *isd** is output of rotor flux controller and *isq** is given in (1).

2.1. Flux and Torque Estimator :

The main feedback signals in DTC algorithm are the estimated flux and torque. They are obtained as outputs of the estimator operating in stator reference frame. This estimator at first performs EMF integration to determine the stator flux vector:

$$\overline{\Psi}_s^s = \int_0^t (\bar{u}_s^s - \bar{R}_s.\bar{\imath}_s^s)\, dt \quad (8)$$

$$\lambda_s = \sqrt{\left(\lambda_{qs}^2 + \lambda_{ds}^2\right)}\ \angle\left(\theta_s = \tan^{-1}\left(\frac{\lambda_{qs}}{\lambda_{ds}}\right)\right) \quad (9)$$

Where $\bar{R}_s$ is the estimated stator resistance value.

Finally, from the estimated stator flux and current vector the motor torque is:

$$T_e = \frac{3}{2} p.\left(i_{s\beta}\Psi_{s\alpha} - i_{s\alpha}\Psi_{s\beta}\right) \tag{10}$$

Where the stator flux and current vectors are given in stationary α-β frame and *p* denotes the number of poles. The scheme uses the feedback control of torque and stator flux linkages, which are computed from the measured motor voltages and currents. The method uses stator reference frame model of the induction motor and the same reference frame is used in the implementation thereby avoiding the trigonometric operations encountered in the coordinate transformations of other reference frames. This is one of the advantages of the control scheme.

The parameter mismatch between the controller and machine also results in a nonlinear characteristic between torque and its reference making it a non ideal torque amplifier. This will have undesirable consequences in a torque drive and to a smaller extent in the speed controlled drive systems. The motor resistance adaptation is essential to overcome instability and to guarantee a linear torque amplifier in the direct torque controlled drive. A stator resistance parameter adaptation scheme is presented in the next section to achieve these objectives.

2.2. Adaptive Stator Resistance Compensation:

A block diagram schematic of the applied stator resistance compensation scheme is shown in Fig. 2 and its incorporation in the drive schematic is given in Fig. 3.

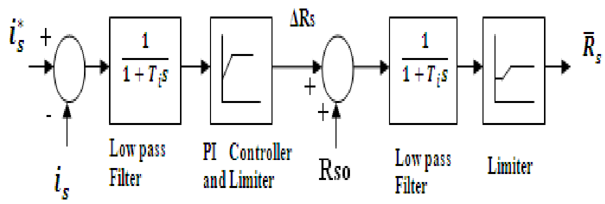

Fig.2. Block diagram schematic of the adaptive stator Resistance Compensator

This technique is based on the principle that the error between the measured stator feedback current phasor magnitude i_S and its command $i_S{}^*$ is proportional to the stator resistance variation which is mainly caused by the motor temperature and to a smaller extent by the varying stator frequency. The incremental value of stator resistance for correction is obtained through a PI controller and limiter. The current error goes through a low pass filter, which has very low cutoff frequency in order to remove high frequency components contained in the stator feedback current. This low pass filter does not generate any adverse effect on the stator resistance adaptation if the filter time constant is chosen to be smaller than that of the adaptation time constant. This incremental stator resistance, ΔR_S, is continuously added to the previously estimated stator resistance, R_{SO} . The final estimated value $\bar{R}_s$ is obtained as the output of another low pass filter and limiter. This low pass filter is necessary for a smooth variation of the estimated resistance value. This final signal is the updated stator resistance and can be used directly in the controller. The above algorithm requires the stator current phasor command, which is a function of the commanded torque and commanded stator flux linkages. An analytic procedure to evaluate the stator current command from the torque and stator flux linkages commands is presented in the following The stator feedback current phasor magnitude $i s$ is obtained from the q and d axis measured currents as,

$$i_s = \sqrt{\left(i_{qs}^2 + i_{ds}^2\right)} \tag{11}$$

The stator command current phasor magnitude i_S^* is derived from the dynamic equations of the induction motor in the synchronously rotating reference frame using the torque command T_C^* and stator flux linkage command λ_S^* . The flux linkages, rotor equations, and torque are given by,

$$\lambda_{qs}^e = L_s i_{qs}^e + L_m i_{qr}^e, \lambda_{ds}^e = L_s i_{ds}^e + L_m i_{dr}^e \tag{12}$$

$$\lambda_{qr}^e = L_r i_{qr}^e + L_m i_{qs}^e, \lambda_{dr}^e = L_r i_{dr}^e + L_m i_{ds}^e \tag{13}$$

$$R_r i_{qr}^e + \omega_{sl}\lambda_{dr}^e + p\lambda_{qr}^e = 0 \tag{14}$$

$$R_r i_{qr}^e - \omega_{sl}\lambda_{qr}^e + p\lambda_{dr}^e = 0 \tag{15}$$

$$T_e = \frac{3}{2}\frac{p}{2}\left(i_{qs}^e\lambda_{ds}^e - i_{ds}^e\lambda_{qs}^e\right) \tag{16}$$

Where p is differential operator, λ_{qs}^e, λ_{ds}^e are q-d axis stator flux linkages,λ_{qr}^e&λ_{qr}^e are q-d axis rotor flux linkages i_{qs}^e, i_{ds}^e are q-d axis stator currents, $i_{dr}^e, i_{qr}^e$are q-d axis rotor currents, ω_{sl} is the slip speed given by(ω_s - ω_r), and P is the number of poles. The resultant stator flux linkage λs is assumed to be on the direct axis. This step is to reduce the number of variables in the equations by one. Moreover, it corresponds with the reality that the stator flux linkages is a single resultant phasor. Hence aligning the &axis with stator flux linkage phasor,

$$\lambda_{qs}^{e} = 0, p\lambda_{qs}^{e} = 0, \lambda_{ds}^{e} \quad (17)$$

Substituting (17) in (12) to (16), the resulting dynamic equations are,

$$0 = L_s i_{qs}^{e} + L_m i_{qr}^{e}, \lambda_s = L_s i_{ds}^{e} + L_m i_{dr}^{e} \quad (18)$$

$$\lambda_{qr}^{e} = L_r i_{qr}^{e} + L_m i_{qs}^{e}, \lambda_{dr}^{e} = L_r i_{dr}^{e} + L_m i_{ds}^{e} \quad (19)$$

$$R_r i_{qr}^{e} + \omega_{sl}\lambda_{dr}^{e} + p\lambda_{qr}^{e} = 0 \quad (20)$$

$$R_r i_{qr}^{e} - \omega_{sl}\lambda_{qr}^{e} + p\lambda_{dr}^{e} = 0 \quad (21)$$

$$T_e = \frac{3}{2}\frac{p}{2} i_{qs}^{e}\lambda_s \quad (22)$$

Then the q-axis current command is directly obtained from (22) Using the torque command Te* and stator flux linkage command λ_S*, as,

$$i_{qs}^{e*} = \frac{2}{3}\frac{2}{p}\frac{\lambda_e^*}{T_e^*} \quad (23)$$

Because λ_S, is a constant, the following relations are derived from (18) as,

$$L_s p i_{qs}^{e} = -L_m p i_{qr}^{e} \quad (24)$$

$$L_s p i_{ds}^{e} = -L_m p i_{dr}^{e} \quad (25)$$

In steady state, $pi_{ds}^{e} = pi_{qs}^{e} = 0$. Therefore, all rotor variables in (20) can be removed using (18), (19), (24) and (25). The resultant rotor q axis equation that is a function of stator variables only is given by,

$$-\frac{R_r L_s}{L_m} i_{qs}^{e*} + \omega_{sl}^* L_m i_{ds}^{e*}\left(1 - \frac{L_s L_r}{L_m^2}\right) + \omega_{sl}^* \frac{L_r}{L_m}\lambda_s^* = 0 \quad (26)$$

Similarly rotor d axis equation can be derived in the steady state using (18), (19), and (21) as,

$$-\frac{R_r L_s}{L_m} i_{ds}^{e*} - \omega_{sl}^* L_m i_{qs}^{e*}\left(1 - \frac{L_s L_r}{L_m^2}\right) + \frac{R_r}{L_m}\lambda_s^* = 0 \quad (27)$$

Because λ_s^* and i_{qs}^{e*} are known values, ω_{sl}^* and i_{ds}^{e*} are found by solving (26) and (27) simultaneously, and are given by,

$$L_s\left(i_{ds}^{e*}\right)^2 - \lambda_s^*\left(1 - \frac{L_s L_r}{L_m^2 - L_s L_r}\right) i_{ds}^{e*} + L_s\left(i_{qs}^{e*}\right)^2 - \frac{\left(\lambda_s^*\right)^2 L_r}{L_m^2 - L_s L_r} = 0 \quad (28)$$

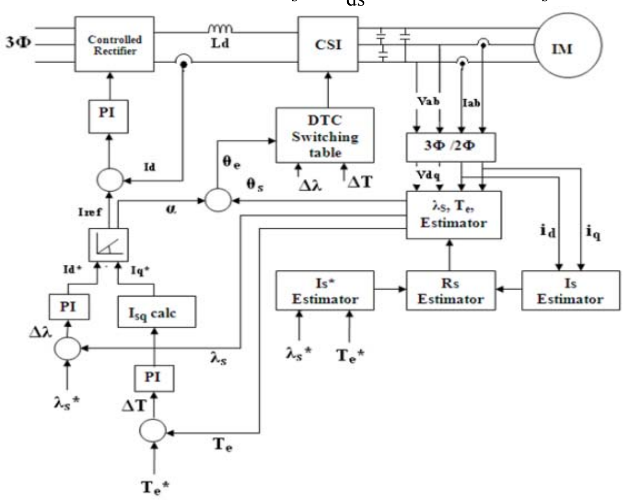

Fig.3. Proposed DTC scheme for CSI fed IM drive with Stator resistance compensation.

3. Simulation Results

Dynamic simulations are performed to validate the performance of the torque controlled drive system with the compensation technique. The induction motor and drive system details used in the simulation are given in the appendix. In the implementation of the drive algorithm, only six nonzero voltage vectors are used and three zero voltage vectors are excluded. It is noted that this has no impact on the basic Performance of the system. Because the stator resistance voltage drop is a considerable portion of the applied voltage at low speed, the performance of the system at low speed due to variation in the stator resistance deteriorates much more compared with that in the high speed range. Fig.4.(a),(b) and Fig.5.(a),(b) are Stator current, and Electromagnetic torque for without and with stator resistance compensation respectively.

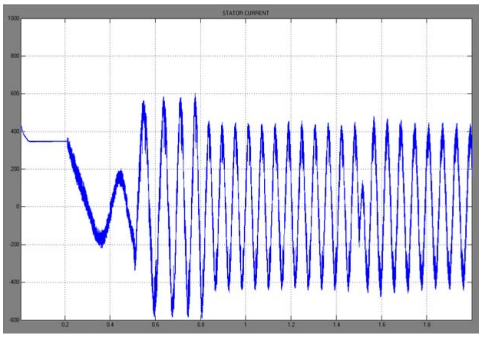

Fig.4.a.Stator Current Ia without Stator Resistance Compensation

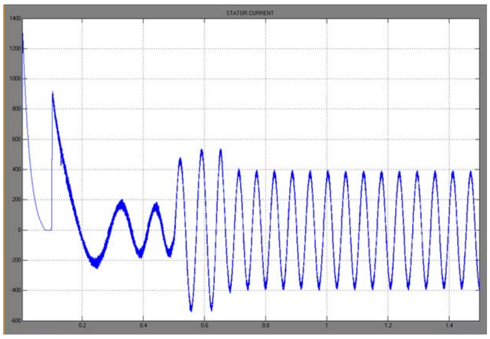

Fig 5.a.Stator Current Ia with Stator Resistance Compensation

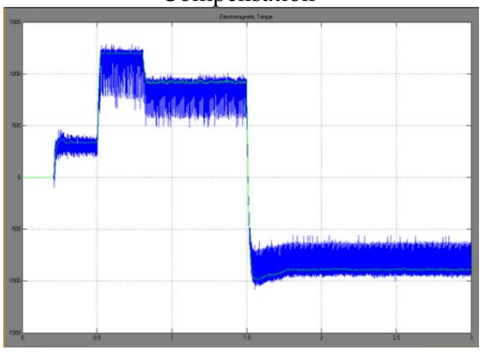

Fig 4.b.Torque Te without Stator Resistance Compensation

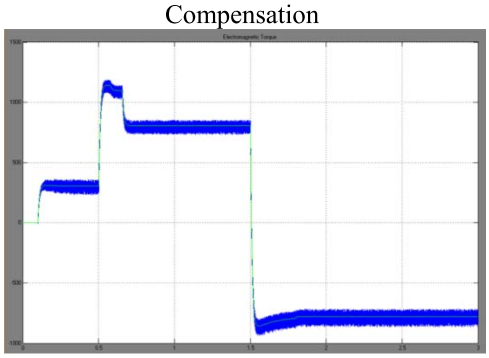

Fig 5.b.Torque Te with Stator Resistance Compensation

4. Conclusion:

The following are considered to be the original contributions of this study:

An adaptive stator resistance compensation scheme is applied to eliminate the stator resistance parameter sensitivity using only the existing stator current feedback. It is simple to implement and its realization is indirectly dependent on stator inductances. A procedure for finding the stator current phasor command from the torque and stator flux linkage commands is derived to realize the adaptation scheme and is independent of stator and rotor resistances of induction motor. The scheme is verified with dynamic simulation for various operating conditions including flux-weakening mode and even in the face of rapid changes in the stator resistance such as step changes and simultaneous variations of torque. The PI controller in the stator resistance adaptation loop gives a good performance and the design and implementation of a PI controller is easier, this study demonstrates the sufficiency of the PI controller for parameter compensation. It is also possible that the PI controller performance is good because the derived stator current phasor reference is independent of stator and rotor resistances.

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