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On-Line Stator Resistance Tuning of DTC Control CSI Fed IM Drives

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ABSTRACT

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, deteri orates the performance of the drive at low speed operation by introducing errors in the estimated flux linkage's magnitude and its position and hence in the electromagnetic torque. Parameter compensation using stator current phasor error has been proposed in literature. 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 DTCwith CSI fed 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.

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1. INTRODUCTION

Even though voltage source inverter (VSI)-fed drives are most widely used, current source inverter (CSI)-fed drives find application [1] [4] 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**

High performance inverter-fed induction motor drives with no mechanical/optical position sensor[2] are desirable for the emerging market applications. A control scheme, which achieves that for induction motor, is the direct torque controller, also known as direct self-controller.[3] The scheme uses feedback control of electromagnetic torque and stator flux linkages. The electromagnetic torque and stator flux linkages are estimated in stator reference frames using the measured stator voltages and currents. The machine model is dependent on stator resistance only. The direct torque controlled induction motor drives have been studied in different forms starting with Depenbrock **[3]** utilizing the continuous stator flux angle calculation and schemes using the limited information of the stator flux angle being in one of the ***six*** **sixty** degrees sextants [5], **[6].** A number of implementations]7]-[9] of the scheme have been made based on how the currents and voltages are measured. This drive has proved beyond doubt the usefulness and simplicity of the modulation scheme and it is almost a standard feature in many ac drives products.But all of them have precision low speed operational problems due to the errors induced by varying stator resistance [10]-[12] in the flux and its angle calculator that is central to the control scheme. It deteriorates the drive performance particularly at low speeds. Note that at low speeds, the stator resistance voltage drops constitute significant proportion of the applied voltages.

A few control schemes have been proposed to overcome this 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 [13] [14] [15], which has the problem of convergence and slowness of response.

Due to parameter sensitivity, there is a possible instability problem arising in this scheme. This paper **is** concerned with a solution to track the stator resistance R_O that the performance degradation and a possible instability problem can be avoided. Proportional-Integral (PI) adaptive compensator using only the measured stator currents is applied. An analytic expression to evaluate the stator current command from the torque and stator flux linkage commands is derived and presented. This scheme requires two filters and one PI controller apart from the existing electromagnetic torque and stator flux linkages calculator in the drive controller. A signal proportional to stator resistance change is developed using the error between the reference and actual stator current phasor. This error is processed through a PI controller for application in the controller. Stator resistance parameter adaptation results in restoring the\precise and accurate estimation of stator flux linkage magnitude and its position in the controller. Note that this is the key to successful operation at low speeds. **A** set of high bandwidth current and voltage sensors or estimators is required for the implementation of the solution. The performance of the controller is validated by extensive dynamic simulation studies for a wide variety of operating conditions including flux weakening.

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.

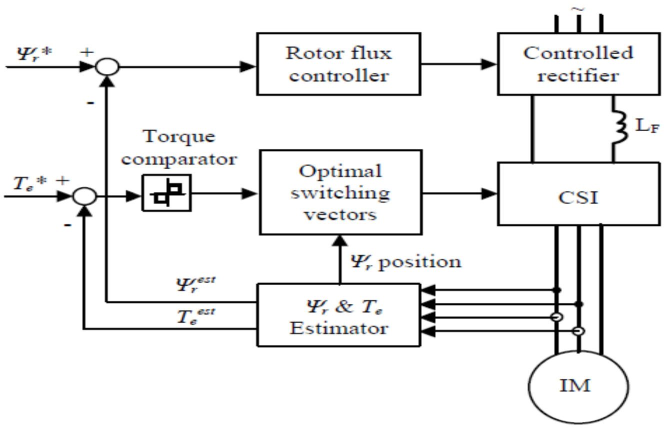

Fig.1. Basic scheme of DTC CSI-fed IM drive

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:

$$\bar{\Psi}_s^s = \int_0^t (\bar{u}_s^s - \bar{R}_s . \bar{i}_s^s)\, dt \quad (1)$$

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

Where $\bar{R}_s$ is the estimated stator resistance value.
Finally, from the estimated stator flux and current vector the motor torque is:

$$T_e = \frac{3}{2} p . (i_{s\beta}\Psi_{s\alpha} - i_{s\alpha}\Psi_{s\beta}) \quad (3)$$

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.

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.

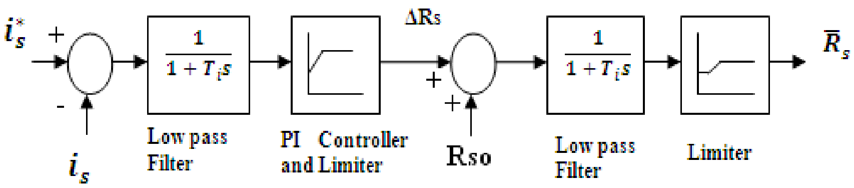

Fig.2. Block diagram schematic of the adaptive stator Resistance Compensator

This technique is based on the principle that the error between the measured stator feedback current phasor magnitude $\mathbf{i}_S$ and its command $\mathbf{i}_S$* 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. 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 $\boldsymbol{i}_s$ is obtained from the q and d axis measured currents as,

$$= \sqrt{(i_{qs}^2 + i_{ds}^2)} \quad (4)$$

The stator command current phasor magnitude $\boldsymbol{i}_S^*$ 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 $\boldsymbol{\lambda}_S^*$. The flux linkages, rotor equations, and torque are given by,

$$\lambda_{qs}^e = L_s i_{qs}^e + L_m i_{qr}^e, \lambda_{ds}^e = L_s i_{ds}^e + L_m i_{dr}^e \quad (5)$$

$$\lambda_{qr}^e = L_r i_{qr}^e + L_m i_{qs}^e, \lambda_{dr}^e = L_r i_{dr}^e + L_m i_{ds}^e \quad (6)$$

$$R_r i_{qr}^e + \omega_{sl}\lambda_{dr}^e + p\lambda_{qr}^e = 0 \quad (7)$$

$$R_r i_{qr}^e - \omega_{sl}\lambda_{qr}^e + p\lambda_{dr}^e = 0 \quad (8)$$

$$T_e = \frac{3}{2}\frac{P}{2}(i_{qs}^e\lambda_{ds}^e - i_{ds}^e\lambda_{qs}^e) \quad (9)$$

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($\omega_s - \omega_r$), 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,

$$\lambda_{qs}^{e} = 0, p\lambda_{qs}^{e} = 0, \lambda_{ds}^{e} \quad (10)$$

Substituting (10) in (5) to (9), the resulting dynamic equations are,the flux linkages, rotor equations, and torque are given by,

$$0 = L_s i_{qs}^{e} + L_m i_{qr}^{e}, \lambda_s = L_s i_{ds}^{e} + L_m i_{dr}^{e} \quad (11)$$

$$\lambda_{qr}^{e} = L_r i_{qr}^{e} + L_m i_{qs}^{e}, \lambda_{dr}^{e} = L_r i_{dr}^{e} + L_m i_{ds}^{e} \quad (12)$$

$$R_r i_{qr}^{e} + \omega_{sl}\lambda_{dr}^{e} + p\lambda_{qr}^{e} = 0 \quad (13)$$

$$R_r i_{qr}^{e} - \omega_{sl}\lambda_{qr}^{e} + p\lambda_{dr}^{e} = 0 \quad (14)$$

$$T_e = \frac{3}{2}\frac{p}{2} i_{qs}^{e}\lambda_s \quad (15)$$

Then the q-axis current command is directly obtained from (15) Using the torque command Te* and stator flux linkage command λ_S*, as,

$$i_{qs}^{e*} = \frac{2}{3}\frac{2}{p}\frac{\lambda_e^*}{T_e^*} \quad (16)$$

Because λ_S, is a constant, the following relations are derived from (11) as,

$$L_s p i_{qs}^{e} = -L_m p i_{qr}^{e} \quad (17)$$

$$L_s p i_{ds}^{e} = -L_m p i_{dr}^{e} \quad (18)$$

In steady state, $pi_{ds}^{e} = pi_{qs}^{e} = 0$. Therefore, all rotor variables in (13) can be removed using (11), (12), (17) and (25). The resultant rotor q axis equation that is a function of stator variables only is given by,

$$-\frac{R_r L_s}{L_m} i_{qs}^{e*} + \omega_{sl}^{*} L_m i_{ds}^{e*}\left(1 - \frac{L_s L_r}{L_m^2}\right) + \omega_{sl}^{*}\frac{L_r}{L_m}\lambda_s^{*} = 0 \quad (19)$$

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

$$-\frac{R_r L_s}{L_m} i_{ds}^{e*} - \omega_{sl}^{*} L_m i_{qs}^{e*}\left(1 - \frac{L_s L_r}{L_m^2}\right) + \frac{R_r}{L_m}\lambda_s^{*} = 0 \quad (20)$$

Because λ_s^* and i_{qs}^{e*} are known values, ω_{sl}^* and i_{ds}^{e*} are found by solving (19) and (20) simultaneously, and are given by,

$$L_s\left(i_{ds}^{e*}\right)^2 - \lambda_s^{*}\left(1 - \frac{L_s L_r}{L_m^2 - L_s L_r}\right) i_{ds}^{e*} + L_s\left(i_{qs}^{e*}\right)^2 - \frac{\left(\lambda_s^{*}\right)^2 L_r}{L_m^2 - L_s L_r} = 0 \quad (21)$$

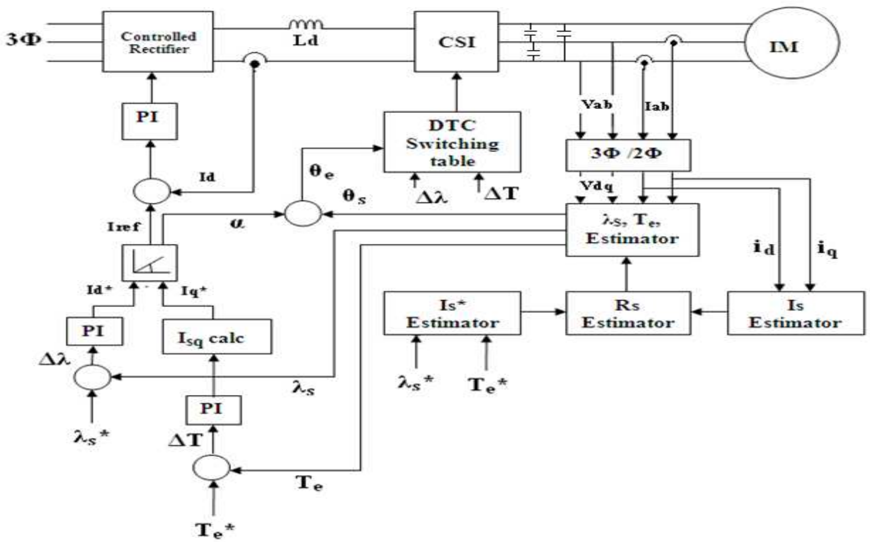

Fig.3. Proposed DTC scheme for CSI fed IM drive with Stator resistance compensation.

3. SIMULATION RESULTS:

The Proposed DTC scheme for CSI fed IM drive with Stator resistance compensation shown in fig(3). Dynamic simulations are performed to validate the performance of the torque controlled drive system with the compensation technique. The induction motor d drive system output current, torqueand speed for with compensation and without compensation are shown in fig 4,5 and6. In the implementation of the drive algorithm, only six non zero current vectors are used and three zero current 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),5(a),6(a) and Fig.4.(b),5(b),6(b) are Stator current, Electromagnetic torque and rotor speed for without and with stator resistance compensation respectively.

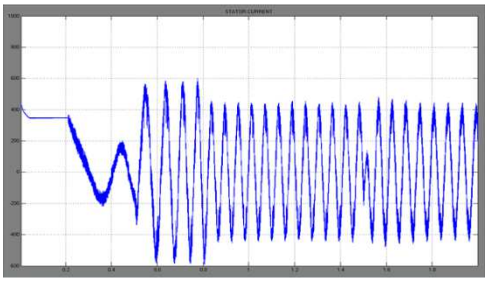

Fig.4.a.Stator Current I_A without stator Resistance Compensation

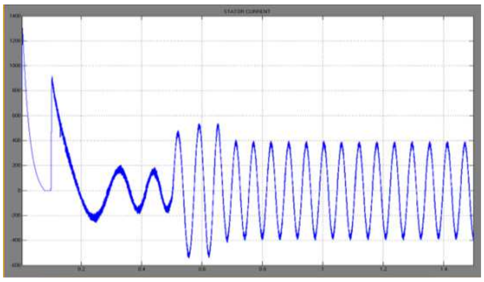

Fig 4.b. Stator Current I_A with stator Resistance Compensation

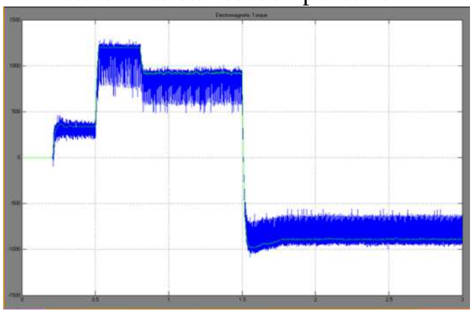

Fig 5.a.Torque T_e without stator Resistance Compensation

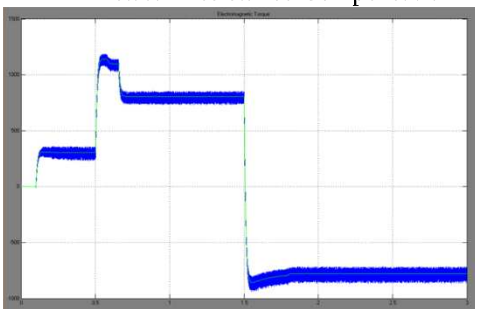

Fig 5.b.Torque Te with stator Resistance compensation

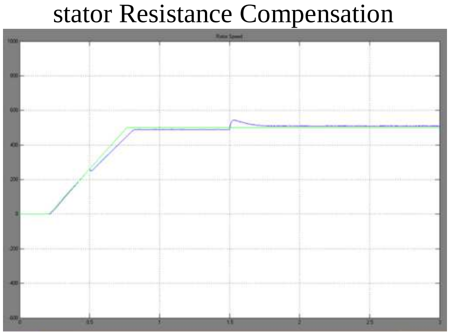

Fig 6.a.Rotor Speed ω without stator Resistance Compensation

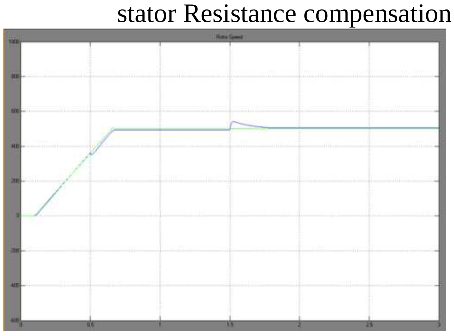

Fig 6.b.Rotor Speed ω with stator Resistance Compensation

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 variation 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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