# A New Rotor Time Constant Adaptation Method for a VSI Fed Indirect Field Oriented Induction Motor Drive

**S. Palani vel**
Research Student
Department of Electrical Engineering
Indian Institute of Science
Bangalore-12 INDIA

**P. Srinivas**
Student
Department of Electrical Engineering
Indian Institute of Science
Bangalore-12 INDIA

**V. T. Ranganathan**
Associate Professor
Department of Electrical Engineering
Indian Institute of Science
Bangalore-12 INDIA

*Abstract*—**Field Oriented control of an Induction Motor requires the rotor time constant as an essential parameter to calculate both the magnitude and phase of the rotor flux vector. There fore it becomes necessary to adapt the rotor time constant. A simple method based on current measurements alone is proposed. Since this method uses current measurements, which are smoother, compared to the voltage wave forms in a VSI fed drive, the implementation is easy and noise free. The method works during transients also. The stator direct and quadrarture current components are predicted, assuming a perfect value of rotor time constant in the control side. This is compared with the measured currents. A mismatch in this comparison, is an indication of an error in phase of the rotor flux. This information is used to correct the control side rotor time constant.**

## INTRODUCTION

Field oriented control gives an elegant way of achieving high performance control of induction motors. Estimating the magnitude and phase of the rotor flux is very crucial for the implementation of this method. Direct ways of sensing the rotor flux by fabricating suitable hardware around the motor have proved to be inaccurate and impractical at very low speeds of operation. Indirect methods of sensing the rotor flux employ a mathematical model of the rotor, which is excited by measurable machine state variables, like currents and voltages. This model calculates both magnitude and phase of the rotor flux. The accuracy of this method depends very much on accuracy of the model parameters, especially rotor time constant. The accuracy of this parameter depends on the accuracy of rotor resistance and rotor inductance.

The actual motor parameters may change during the operation of the drive, which will introduce inaccuracies in the flux estimation, if left unadapted. The resistance of the stator and rotor windings change with temperature. Due to large thermal time constants, compared to many electrical time constants, these variations are slow. Parameter changes due to over excitation or under excitation of the magnetic circuit happen very fast, on the other hand. This affects mutual inductance of the machine largely, and leakage inductances slightly.

## ROTOR TIME CONSTANT ADAPTATION

One of the earliest methods of adaptation by Luis J Garces[4], is based on the modified reactive power expression. The expression is dependent on the rotor time constant. The expression is evaluated both on the control side and on the motor side in real time. The difference in these two expressions means a mismatch between control side and motor side rotor time constants.

If there is a complete decoupling between d and q axes, a random signal in d-axis current will have zero correlation with the q-axis response. This idea is used by Gabriel and Leonhard[5] for adapting the rotor time constant. A Pseudo Random Binary Sequence (PRBS) is added to the d-axis current reference. A non-zero correlation between the d and q axes control system outputs indicates a cross coupling and this information is used to adapt the rotor time constant. But the undesired effect of the PRBS signal in the performance of the drive has limited the usage of this method.

A method to adapt the rotor time constant in real time using a Kalman filter was proposed by Zai and Lipo[9]. This method is limited to situations where the load conditions are changing slowly and also requires speed at least above 5% of the rated speed.

Li and Venkatesan[13] have used the mismatch between the control side torque feedback and the actual motor torque to adapt the rotor time constant. This scheme, although simple, has large dynamic errors and inaccurate at low speeds. Koyama et al [7] have used a simplified reactive power expression for the adaptation valid only for steady state operation.

Each of the above methods has some limiting factor while almost all of them are not suitable to be applied at or around zero speed conditions. Holtz and Thimm[12] describe a method to adapt all parameters even at zero speed. An analytical model of the machine, with the stator voltages and mechanical speed as inputs, is operated in parallel with the machine in real time. Then, by comparing current trajectories of the model and the machine, all the model parameters are corrected using nonlinear optimization method. But the implementation is sophisticated and requires lot of computational power.

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The present method, apart from being extremely simple, works during transients and at near zero speeds.

### THE PROPOSED METHOD

In a voltage source inverter fed induction motor the field oriented controller output and the d-q axis stator model is represented as shown in the Fig. 1, under coupled conditions. The following notations are used.

- $i'_{sd}$, $i'_{sq}$ - d-q axis currents in the actual motor.
- $i_{sd}$, $i_{sq}$ - d-q axis currents measured in the control side.
- $U'_{sd}$, $U'_{sq}$ - d-q axis voltages in the actual motor.
- $U_{sd}$, $U_{sq}$ - d-q axis current controller outputs in the control side.
- $U^{*\prime}_{sd}$, $U^{*\prime}_{sq}$ - d-q axis voltages in the actual motor after coordinate transformation and before subtracting the de-coupling terms.
- $U^*_{sd}$, $U^*_{sq}$ - d-q axis voltage references in the control side after adding the de-coupling terms to the current controller outputs.
- $U_{sa}$, $U_{sb}$ - a-b ( stationary ) axis voltages.
- $D'_{sd}$, $D'_{sq}$ - de-coupling terms in the actual motor.
- $D_{sd}$, $D_{sq}$ - de-coupling terms in the control side.
- $i'_{mR}$ - actual rotor flux magnetizing current in the motor.
- $i_{mR}$ - estimated rotor flux magnetizing current in the control side.
- $\omega'_1$ - actual speed of the rotor field oriented reference frame in the motor.
- $\omega_1$ - estimated speed of the rotor field oriented reference frame in the control side.
- $\sigma$ - total leakage coefficient of the motor.
- $L_s$, $L_r$ - stator and rotor total magnetizing inductance.
- $R_s$, $R_r$ - stator and rotor resistance.
- $T_s$, $T'_r$ - the actual stator and rotor time constants.
- $T_r$ - the estimated rotor time constant.
- $\rho$ - the estimated angle of rotor flux in the control side.
- $\rho'$ - the actual angle of rotor flux in the motor.
- $i_{s1}, i_{s2}, i_{s3}$ - the three phase currents iat the motor terminals.

The stator voltage equations in the rotor field oriented reference frame is as follows

$$U'_{sd} = U^{*\prime}_{sd} - D'_{sd} = U^{*\prime}_{sd} - \left[(1-\sigma)L_s\tfrac{d}{dt}(i'_{mR}) - \omega'_1\sigma L_s i'_{sq}\right] ..(Ia)$$
$$U'_{sq} = U^{*\prime}_{sq} - D'_{sq} = U^{*\prime}_{sq} - \left[\omega'_1(1-\sigma)L_s i'_{mR} + \omega'_1\sigma L_s i'_{sd}\right] ....(Ib)$$

To ensure perfect decoupling, $D_{sd}$ & $D_{sq}$ will be added with $U_{sd}$ & $U_{sq}$ the corresponding controller outputs, respectively. Therefore, on the controller side,

$$U^*_{sd} = U_{sd} + D_{sd} = U_{sd} + \left[(1-\sigma)L_s\tfrac{d}{dt}(i_{mR}) - \omega_1\sigma L_s i_{sq}\right] ..(IIa)$$
$$U^*_{sq} = U_{sq} + D_{sq} = U_{sq} + [\omega_1(1-\sigma)L_s i_{mR} + \omega_1\sigma L_s i_{sd}] .....(IIb)$$

These d-q reference frame quantities, $U^*_{sd}$ & $U^*_{sq}$ are transformed into stationary reference frame quantities $U_{sa}$ & $U_{sb}$ by a coordinate transformation as follows.

$$U_{sa} = U^*_{sd}\cos\rho - U^*_{sq}\sin\rho ...........1a$$
$$U_{sb} = U^*_{sd}\sin\rho + U^*_{sq}\cos\rho ..........1b$$

These a-b axis voltage references are converted to three phase voltage references and are applied to the inverter. Taking the inverter gain to be unity for analysis purposes, the motor side d-q axis voltages can be found as follows.

$$U^{*\prime}_{sd} = U_{sa}\cos\rho' + U_{sb}\sin\rho' ..........2a$$
$$U^{*\prime}_{sq} = -U_{sa}\sin\rho' + U_{sb}\cos\rho' ..........2b$$

When the control side motor model parameters match with those of the actual motor, these two coordinate transformations cancel each other. Then the following relations hold.

$$U^*_{sd} = U^{*\prime}_{sd} \qquad \rho = \rho' \qquad U^*_{sq} = U^{*\prime}_{sq}$$
$$D_{sd} = D'_{sd} \qquad D_{sq} = D'_{sq}$$

Under non-ideal conditions, the control side motor parameters will not match with those of the motor. In particular, if the rotor time constant values on control side and motor side differ, then it can be shown that it will result in an error of the estimated phase of the flux. Then, all the above expressions are not valid.

Defining, the error angle, $\delta = \rho' - \rho$, $U^*_{sd}$ & $U^*_{sq}$ on the control side, can be related to $U^{*\prime}_{sd}$ & $U^{*\prime}_{sq}$ on the motor side as follows ( Refer Fig. 3 ).

$$U^{*\prime}_{sd} = U^*_{sd}\cos\delta + U^*_{sq}\sin\delta ................3a$$
$$U^{*\prime}_{sq} = -U^*_{sd}\sin\delta + U^*_{sq}\cos\delta ..............3b$$

These d-q axis currents appear as three phase currents at the motor terminals, after coordinate transformation by angle $\rho'$ and a-b to three phase conversion, inside the motor. If $\rho$ is used to convert them back to d-q axis, $i_{sd}$ & $i_{sq}$ on the control side will be related to $i'_{sd}$ & $i'_{sq}$ on the motor side as follows.

$$i'_{sd} = i_{sd}\cos\delta + i_{sq}\sin\delta .........6a$$
$$i'_{sq} = -i_{sd}\sin\delta + i_{sq}\cos\delta .........6b$$

Substituting (3) and (6) in (Ia),

$$U'_{sd} = U^{*\prime}_{sd} - D'_{sd} = U^*_{sd}\cos\delta + U^*_{sq}\sin\delta - D'_{sd}$$
$$U'_{sd} = (U_{sd} + D_{sd})\cos\delta + (U_{sq} + D_{sq})\sin\delta$$
$$+\omega'_1\sigma L_s(-i_{sd}\sin\delta + i_{sq}\cos\delta) - (1-\sigma)L_s\tfrac{d}{dt}(i'_{mr})$$

The above equation relates the actual motor side Usd in the motor to the control side quantities, through error angle, and actual motor imr $\left| i'_{mR} \right|$.

Since the stator parameters are accessible for measurement, unlike the rotor parameters, the stator parameters are assumed to be known accurately. When the machine T'r is changing, due to either a change in R'r or L'r, the slip frequency error starts increasing. The slip frequency on the control side is $\omega_2 = \frac{i_{sq}}{T_r i_{mR}}$, while on the motor side it is $\omega'_2 = \frac{i'_{sq}}{T'_r i'_{mR}}$. Assuming no error in speed feedback, the error in flux angle is only due to an error in slip frequency. But, after the current loop delay, $i_{sq}$ control loop responds to the new operating conditions, and the slip frequency is again matched with the actual value, automatically, even without any adaptation of the parameter. Then the error angle settles with a non zero value, momentarily.

$$\frac{d}{dt}(i'_{mr}) = L^{-1}\left[\frac{s}{1+sT'_r}(i'_{sd})\right]$$

$$\frac{s}{1+sT'_r}(i'_{sd}) = \frac{s}{1+sT'_r}[i_{sd}\cos\delta + i_{sq}\sin\delta]$$

At this point, the following observations can be made. In a typical drive with reasonably fast current control loops, it is true that $\omega_1 \approx \omega'_1$. This implies, $\frac{d\delta}{dt} = \frac{d\rho'}{dt} - \frac{d\rho}{dt} = \omega'_1 - \omega_1 \approx 0$. This means that $\delta$ can be taken to be independent of time. Therefore,

$$\frac{s}{1+sT'_r}[i_{sd}\cos\delta + i_{sq}\sin\delta]$$

$$= \frac{s(1+sT_r)}{\left[1+sT'_r\right]}\left[i_{mR}\cos\delta + \frac{i_{sq}}{1+sT_r}\sin\delta\right] \quad ...8$$

Substituting (8) in the expression for $U'_{sd}$, and reorganizing the terms,

$$U'_{sd} = \left[U_{sd} - (1-\sigma)L_s\left\{\left(\frac{1+sT_r}{1+sT'_r}s\right)[i_{mR}] - (s)i_{mR}\right\}\right]\cos\delta$$

$$+ \left[U_{sq} - (1-\sigma)L_s\left\{\frac{s}{1+sT'_r}(i_{sq}(t))\right\} + \omega_1(1-\sigma)L_s i_{mR}\right]\sin\delta$$

The above equation is in the following form.

$$U'_{sd} = \bar{U}_{sd}\cos\delta + \bar{U}_{sq}\sin\delta$$

Applying a first order lag to both LHS & RHS,

$$L^{-1}\left[\frac{1}{R_s}\frac{1}{1+s\sigma T_s}\right]U'_{sd}(t) = L^{-1}\left[\frac{1}{R_s}\frac{1}{1+s\sigma T_s}\right]\bar{U}_{sd}(t)\cos\delta$$

$$+ \quad L^{-1}\left[\frac{1}{R_s}\frac{1}{1+s\sigma T_s}\right]\bar{U}_{sq}(t)\sin\delta$$

$i'_{sd}(t) = \bar{i}_{sd}(t)\cos\delta + \bar{i}_{sq}(t)\sin\delta$, where $i'_{sd}$ is the actual d axis motor current.

The above equation can be rewritten as follows.

$$i_{sd}(t)\cos\delta + i_{sq}(t)\sin\delta = \bar{i}_{sd}(t)\cos\delta + \bar{i}_{sq}(t)\sin\delta$$

After reorganizing the terms,

$$\tilde{i}_{sd}\cos\delta + \tilde{i}_{sq}\sin\delta$$

$$= \left[\bar{i}_{sd} - i_{sd}\right]\cos\delta + \left[\bar{i}_{sq} - i_{sq}\right]\sin\delta = 0 \quad .....9$$

The above equation represents a triangle, whose sides are as shown in Fig. 2.

This implies, if $\tilde{i}_{sd}$ & $\tilde{i}_{sq}$ has the same sign,

$$\sin\delta = \frac{-\tilde{i}_{sd}}{\tilde{i}^2_{sd}+\tilde{i}^2_{sq}} \Rightarrow \delta = \arcsin\left[\frac{-\tilde{i}_{sd}}{\tilde{i}^2_{sd}+\tilde{i}^2_{sq}}\right] \quad ....10$$

For small values of $\delta$,

$$\delta = \frac{-\tilde{i}_{sd}}{\tilde{i}^2_{sd}+\tilde{i}^2_{sq}} \quad ..........11$$

So $\delta \equiv 0$, if and only if $\tilde{i}_{sd} \equiv 0$.

And signum $\left(-\tilde{i}_{sd}\right)$ gives the signum($\delta$), if signum $\left(\tilde{i}_{sq}\right)$ is positive, which can be used for the direction of correction in the adaptation loop.

This method will work during transients.

**THE IMPLEMENTATION OF THE ADAPTATION LOOP.**

*1. Calculation of* $\bar{U}_{sd}$

The error for the adaptation loop is formed as follows.

$$-\tilde{i}_{sd} = i_{sd} - \bar{i}_{sd} = i_{sd} - \frac{1}{R_s}\frac{1}{1+s\sigma T_s}\bar{U}_{sd}$$

In evaluating the above expression, except $\bar{U}_{sd}$, other variables and parameters are known. The nature of $\bar{U}_{sd}$ is studied by considering three cases as follows.

*Case I*: During base speed operation, it is known that

$$\frac{d}{dt}[i_{mR}(t)] \equiv L^{-1}\left[\frac{1+sT_r}{1+sT'_r}\right]\left[\frac{d}{dt}(i_{mR}(t))\right] \equiv 0$$

$$\Rightarrow \bar{U}_{sd} \equiv U_{sd}$$

*Case II*: During field weakening, i.e. $\frac{di_{mR}}{dt} \neq 0$,

When $T_r \approx T'_r$, $L^{-1}\left[\frac{1+sT_r}{1+sT'_r}\right]\left[\frac{di_{mR}(t)}{dt}\right] \approx \frac{di_{mR}(t)}{dt}$.

In this case also, as a momentary approximation,

$$\bar{U}_{sd} = U_{sd}.$$

When $T_r \neq T'_r$,

$$L^{-1}\left[\frac{1+sT_r}{1+sT'_r}\right]\left[\frac{di_{mR}(t)}{dt}\right] - \frac{di_{mR}(t)}{dt} \neq 0$$

But, as the adaptation loop proceeds, and Tr approaches Tr', it can be easily seen that,

$$\underset{T_r \to T'_r}{Lt} \quad L^{-1}\left[\frac{1+sT_r}{1+sT'_r}\right]\left(\frac{di_{mR}(t)}{dt}\right) - \frac{di_{mR}(t)}{dt} = 0$$

This implies, the error caused by neglecting this term will be zero after settling of the loop. Moreover it can be seen from the above expression the term will have non zero value only during transients of the $\frac{di_{mR}}{dt}$. So, neglecting this term affects the rate of settling, only during field weakening, and when $\frac{d^2 i_{mR}}{dt^2} \neq 0$.

*Case III*: During steady state speed operation in field weakened region.

$$L^{-1}\left[\frac{1+sT_r}{1+sT'_r}\right]\left(\frac{di_{mR}(t)}{dt}\right) \equiv \frac{di_{mR}(t)}{dt} \equiv 0$$

$$\Rightarrow \bar{U}_{sd} \equiv U_{sd}$$

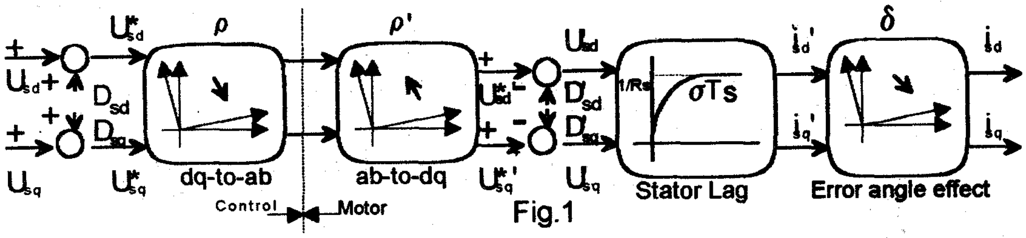

Fig.1

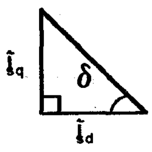
Fig.2

Therefore, for implementation purposes, $\bar{U}_{sd}$ can always be taken as $U_{sd}$.

So, $-\tilde{i}_{sd} = i_{sd} - \bar{i}_{sd} = i_{sd} - \frac{1}{R_s}\frac{1}{1+s\sigma T_s}U_{sd}$

*2. Inferring the signum of* $\tilde{\bar{i}}_{sq}$

To infer the signum of $\tilde{i}_{sq}$, we examine the following expression.

$$\tilde{i}_{sq} = \left(\bar{i}_{sq} - i_{sq}\right)$$

$$= \frac{1}{R_s}\frac{1}{1+s\sigma T_s}\left[U_{sq} + (1-\sigma)L_s\omega_1 i_{mR} - (1-\sigma)L_s\left\{\frac{s}{1+sT_r}\right\} i_{sq}\right] - i_{sq}$$

$$= \frac{1}{Rs}\frac{1}{1+s\sigma T_s}(U_{sq}) + \frac{1}{Rs}\frac{1}{1+s\sigma T_s}[(1-\sigma)L_s\omega_1 i_{mR}.$$

$$-\frac{1}{Rs}\frac{1}{1+s\sigma T_s}\frac{s}{1+sT_r}[(1-\sigma)L_s i_{sq}] - i_{sq}$$

From the above expression, following observations can be made.

- $\frac{1}{Rs}\frac{1}{1+s\sigma T_s}[U_{sq}] \approx i_{sq}$
- In steady state,
  i.e. $\underset{s\to 0}{\mathrm{Lt}} \frac{s}{(1+s\sigma T_s)(1+sT_r)}[(1-\sigma)T_s i_{sq}] = 0$
- Even during transients, the above approximation is found to introduce a very small error for speeds above 5%.
- $i_{sq}$ is one of the fastest state variables. There fore, it either catches the command or hits the limits and stays there. So, most of the time, it is constant. So, the observation is valid for most of the time.

So the signum of the above expression is dependent only on the term $\frac{1}{Rs}\frac{1}{1+s\sigma T_s}[(1-\sigma)T_s\omega_1 i_{mR}]$, whose signum in turn depends only on $\omega_1$. This implies, the signum of $\left[\frac{\omega_1}{1+s\sigma T_s}\right]$., decides the direction of correction. If we are adapting the parameter very slowly, we can take the signum of $\omega$ itself for the direction of correction.

This gives the necessary information about the error angle.

The error angle information is used for correcting the control side rotor time constant. This error angle information should be sign corrected with the signum[$i_{sq}$] before it is used to correct Tr. An example VSI fed vector controlled induction motor drive scheme is shown in Fig. 4. The current loop controller output becomes the d-axis voltage command reference after the limiter, which is meant to take care of the inverter voltage output rating limitation. This is added with the decoupling terms and applied to the inverter as instantaneous voltage command reference. The same d-axis voltage command reference is also applied to a first order lag, identical to the stator lag. The output of the first order lag will be a prediction of isd. This current is compared with the measured machine current, as shown in Fig. 5. The error is sign corrected to derive the information of error angle. This should again be sign corrected, before applying to the adaptive controller. The output of the adaptive controller adds a correction term to the cold value of the controller side rotor time constant. The adaptive controller time constant is designed to be large to avoid unwanted nonlinear interaction with the dynamics of the drive during adaptation. The adapted time constant affects the slip frequency and in turn corrects the error angle by suitably increasing or decreasing the frequency of the supply voltage of the motor. The factor $\left[\frac{i_{sq}}{i_{mR}}\right]$ acts as a sensitivity factor for the correction. When designing the adaptive controller gain this factor should be taken into account. The design should be made for the worst case (largest value) of this factor. This value typically depends on the desired maximum speed.

*3. Improvement of adaptation performance in low speed and heavy loading conditions.*

The adaptation performance in very low speed and heavy loading conditions can be done as follows.

At very low speeds, using $\omega_1$ alone for knowing the signum $\left[\tilde{i}_{sq}\right]$ is not accurate. Instead, signum $\left[\frac{1}{1+s\sigma T_s}\left\{\omega_1 i_{mR} - \frac{s}{1+sT_r}(i_{sq})\right\}\right]$ can be used. This requires more logic in the implementation, but gives better adaptation response at very low speeds.

## THE EXPERIMENTAL RESULTS.

The adaptation method is tested on two drives. The first system is a 40 hp, 51 Amps, 4 pole, wound rotor machine. This vector control drive has been implemented using TMS320C50 DSP processor. The adaptation is made tc correct $\frac{1}{T_r}$, which can be easily achieved by reversing th signum of the $\tilde{i}_{sd}$. This drive has a tacho-generate feedback of the speed. A three phase resistance is connecte or shorted in the rotor circuit, to effect a change in the rotc resistance. This changes rotor time constant and the adaptation method was made to adapt for this change. The results are presented in the following the figures.

A step change in speed is issued to the drive, from 40% of the rated speed to 90% of the rated speed. The resulting speed response, along with the step speed command is shown in Fig. 6.4. The noise in the wave forms are mostly due to measurement. This response, with the adaptation enabled, has a settling time of 2 seconds, while it takes around 2.4 seconds to settle without adaptation with added rotor resistance. The next wave form shows the response of the adaptation, i.e. the correction in the $\frac{1}{T_r}$, along with the $\bar{i}_{sd}$. The adaptation when the Rr is increased from 0.1ohms to 0.45ohms is shown in Fig. 6.5. The increase in $\bar{i}_{sd}$ for a step change in the Rr should be noted. This increase causes the $\frac{1}{T_r}$ to be increased. Finally the predicted and actual $i_{sd}$ s become equal, which means the settling of the adaptation loop. The adaptation response is fast in the field weakening region. It is mainly because of the sensitivity factor $\frac{i_{sq}}{i_{mr}}$ in the adaptation loop. This factor increases in the field weakening region, which contributes for the faster adaptation response.

The second drive is 550 Watts, 1.5 Amps, 4 pole squirrel cage machine. The vector control is implemented using ADSP2101 DSP processor. The rotor position of the motor is measured with an incremental encoder. The performance of the drive is shown here with the exact value of the rotor resistance. The exact rotor resistance value is obtained by the following method.

If the rotor time constant value is not correct on the control side, then for a change in the speed within the base operating region, the direct axis loop gets disturbed. The amount of disturbance in the direct axis can be a criteria to obtain the correct value of the rotor time constant. The direct axis current will usually be very noisy. So, the disturbance in the $i_{mr}$ can be used as an indication of the discrepancy. Fig. 6.3 shows the disturbances in for various values of rotor time constant. The minimum disturbance in $i_{mr}$ means the best possible value of rotor time constant.

The performance of the drive is shown in the base speed region, in Fig. 6.1. Fig. 6.1a shows $i_{mr}$ & $i_{sd}$, while Fig. 6.1b shows the speed and $i_{sq}$ response for a reversal of speed command from -35.8% to +35.8% of the rated speed, in the base speed region. The wave forms are shown for tuned condition. The coupling between direct and quadrature axis can be observed to be minimal. The noise in the speed wave form is mainly due to the differentiation done to obtain the speed from the encoder position feedback. The rotor time constant adaptation method can be verified in this machine by deliberately assuming a wrong value on the control side. This activates the adaptation algorithm to correct the control side value to the correct value. The adaptation is made to work very slow to observe the adaptation response during repeated reversals in the speed, for every six seconds. The correction in $\frac{1}{T_r}$ is shown in Fig. 6.2. The method is found to work very well in the field weakening region and at low speeds. The method does not work at zero stator frequency, because the flux vector stops rotating, and the loop corrects

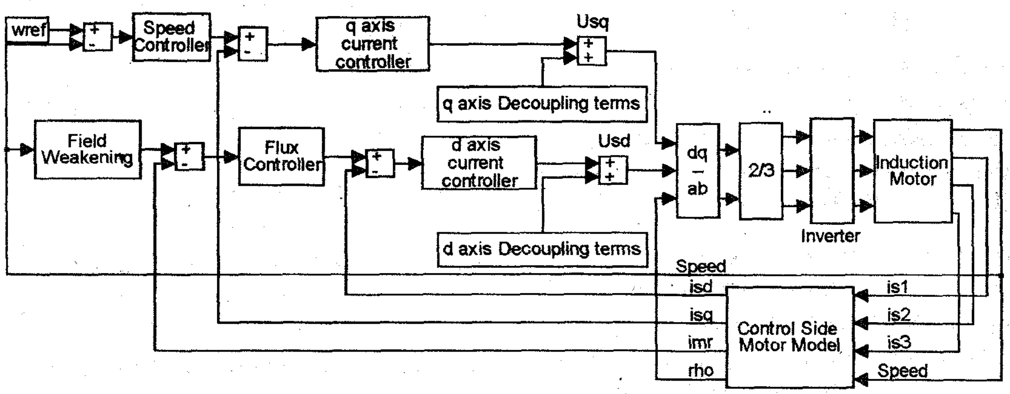


Fig. 4 A. Typical Indirect Vector Control Drive with VSI

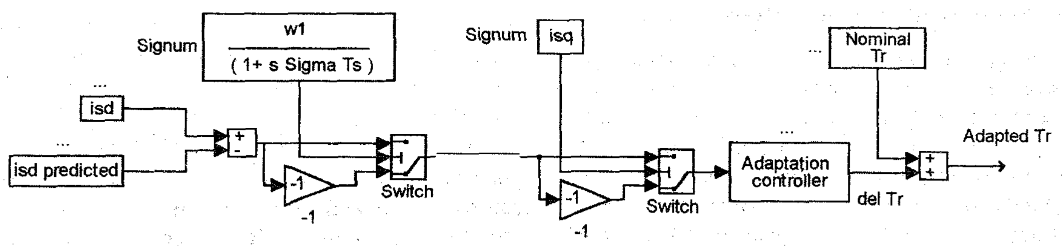


Fig. 5. The proposed Adaptation Scheme

only the rotor time constant which is a factor in the slip frequency.

## CONCLUSIONS

From the above results, the effectiveness of this simple scheme of adaptation can be seen. The main drawback of this scheme is that it assumes that the stator resistance, gain of the inverter and stator time constant are known accurately. When the stator inductance is not known perfectly, this scheme can be left to act only in steady state. In steady state, this scheme assumes a much more simple form as follows.

$$-\tilde{i}_{sd} = \frac{U_{sd}}{R_s} - i_{sd} \qquad \text{.........12}$$

Of course, the scheme again depends on the stator resistance and gain of the inverter. Stator resistance can be meaured periodically. If the gain of the inverter is varying due to dc link voltage, the variation can be modelled.


## ACKNOWLEDGMENT

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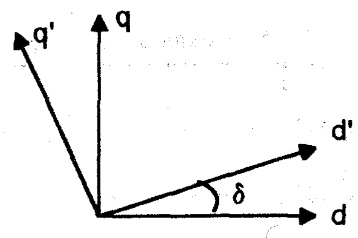


Fig. 3


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Fig. 6.1a Currents during a speed reversal

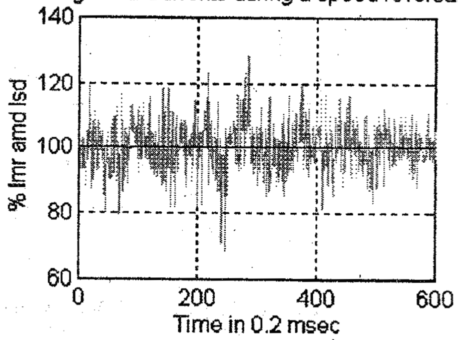



Fig. 6.1b Speed & Current during a speed reversal

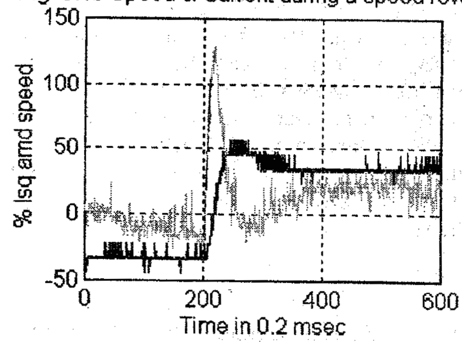



Fig. 6.2 Adaptation of One By Tr

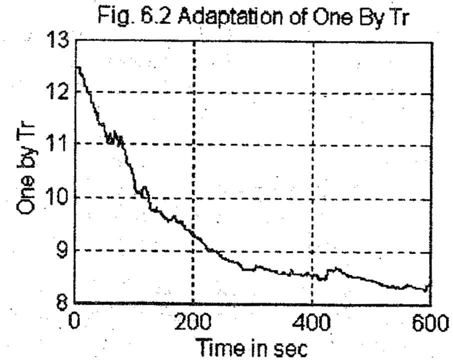
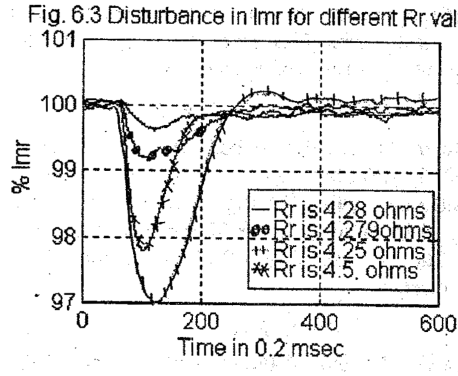



Fig. 6.3 Disturbance in Imr for different Rr values



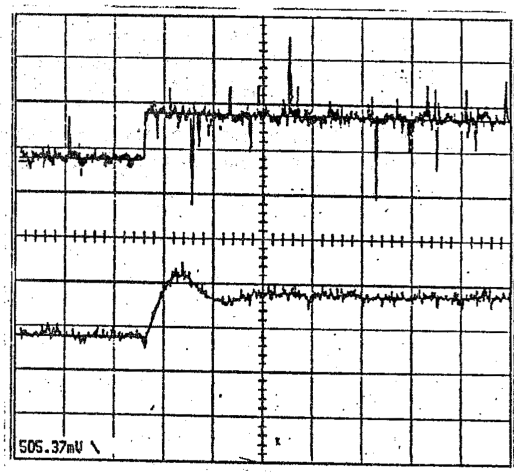




Fig. 6.4 Speed reference ( Top Trace )
and Speed response ( Bottom Trace )
Scale : 20 rad/sec per division

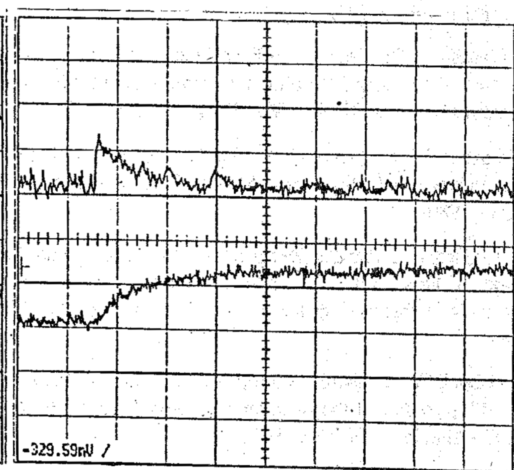



Fig. 6.5 $\bar{i}_{sd}$ and One By Tr during a step change in one by Tr.
Scales : $\bar{i}_{sd}$ ( Top trace ) 30.75 A per division.
One By Tr ( Bottom trace ) 7.11 $sec^{-1}$ per div.