PROGRESS REPORT

1. Project Title: Retrofit parallel hybrid drive	Ref .No.CiSTUP/RP/09-05/159 Dated 10/12/09
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4. Broad area of Research: Power electronics

Sub Area : (e).Motor speed control

5. Approved Objectives of the Proposal :
6. In the first phase the objective of this project is to design and development of a high dynamic speed control system for Induction motor drive for Electric vehicle applications. A sensorless speed control system is envisaged in the first phase for EV applications.

Date of Start:1-1-2010	Total cost of Project:Rs.5Lakhs
Date of completion: March-2012	Expenditure as on nowRs.5Lakhs

Prof. K.Gopakumar
Investigator,

Prof. L.Umanand
Co Investigator

6.Methodology :

In this project a sensorless vector control scheme for a general purpose induction motor drives using current error space phasor based hysteresis controller is designed, fabricated and tested. In the present work, a new technique for sensorless operation is developed to estimate rotor voltage and hence rotor flux position using the stator current error during zero voltage space vectors. It gives comparable performance with vector control drive using sensors especially at very low speed of operation (less than 1 Hz). Since no voltage sensing is made, dead time effect and loss of accuracy in voltage sensing at low speed are avoided here, with the inherent advantages of current error space phasor based hysteresis controller. But appropriate device on-state drop are compensated to achieve a steady state operation upto less than 1 Hz. Moreover, using a parabolic boundary for current error, the switching frequency of the inverter can be maintained constant for entire operating speed range. Simple sigmaL_s estimation is proposed and the parameter sensitivity of the control scheme to changes in stator resistance, R_s is also investigated in this paper. Extensive experimental results are shown at speeds less than 1 Hz to verify the proposed concept. The same control scheme is further extended from less than 1 Hz to rated 50 Hz six-step operation of the inverter. Here, the magnetic saturation is ignored in control scheme.

CONCEPT OF SPEED AND ROTOR FLUX POSITION ESTIMATION

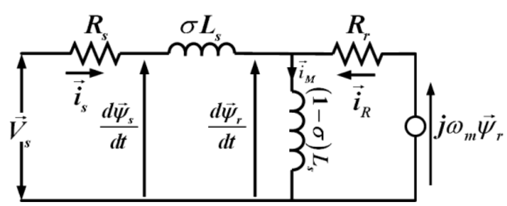

Fig. 1 Space Phasor based equivalent circuit of induction motor with rotor flux as a reference vector

Fig. 1 shows the steady state equivalent circuit of the induction motor. The stator voltage equation in stationary reference frame is,

$$\vec{v}_s = R_s\vec{i}_s + \frac{d\vec{\psi}_s}{dt} \quad (1)$$

And the stator flux equation is

$$\vec{\psi}_s = \sigma L_s\vec{i}_s + a\cdot\vec{\psi}_r \quad (2)$$

where, $\vec{v}_s$ = stator voltage vector = $v_{s\alpha} + jv_{s\beta}$; $\vec{i}_s$ = stator current vector = $i_{s\alpha} + ji_{s\beta}$; $\vec{\psi}_s$ = stator flux vector = $\psi_{s\alpha} + j\psi_{s\beta}$; L_s = stator inductance; $\vec{\psi}_r$ = rotor flux vector = $\psi_{r\alpha} + j\psi_{r\beta}$; σ = leakage inductance co-efficient, $a = L_m/L_r$;

Here, the magnetic circuit is considered to be linear and magnetic saturation effects are not considered.

Here, the actual stator current vector is given as,

$$\vec{i}_s = \vec{i}_s^* + \Delta\vec{i}_s \quad (3)$$

where, $\vec{i}_s^*$ = reference current vector; $\Delta\vec{i}_s$ = current error space vector, [18].

A.Calculation of rotor voltage

Fig. 2(a) and Fig. 2(b) shows power circuit of a two-level voltage source inverter (VSI) fed induction motor (IM) drive and the voltage space phasor structure generated by the inverter. In the voltage controlled space vector PWM based inverter fed induction motor drives, the inverter produces voltage space vector V_1, V_2 ... V_8 of Fig. 2(b).

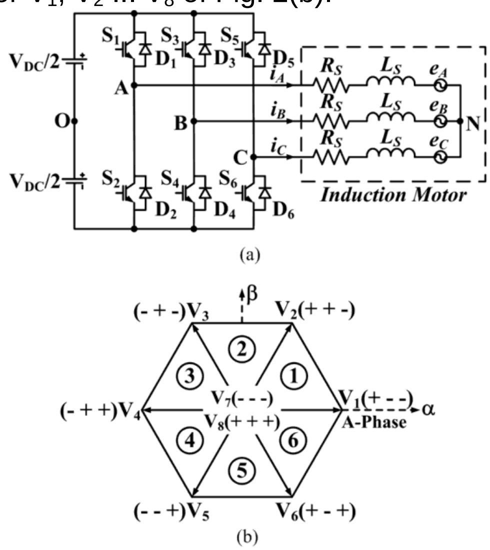

Fig. 2 (a) Power schematic of a three-phase two-level Voltage Source Inverter (VSI) fed induction motor drive, and (b) Voltage space phasor structure of the two-level voltage source inverter

Out of eight space vector, V_1, V_2, ..., V_6 are active voltage vectors and; V_7 and V_8 are virtual zero voltage vectors. Referring (1), the stator space phase voltage equation of induction machine when fed from the inverters is:

$$\vec{v}_K = R_s \vec{i}_s + \frac{d\vec{\psi}_s}{dt} \qquad (4)$$

where, $\vec{v}_K$ is the instantaneous space vector output from inverter. $\vec{v}_K$ can have value of space vectors V_1, V_2 ... V_8 of Fig. 2(b).

The value of stator flux from (2) is put into (4) and it gives,

$$\vec{v}_K = R_s \vec{i}_s + \sigma L_s \frac{d\vec{i}_s}{dt} + \frac{d\left(a \cdot \vec{\psi}_r\right)}{dt} \qquad (5)$$

Calculation based on only zero vector output upto synchronous frequency of 25 Hz

When induction motor is fed with zero voltage space vectors, V_8 and V_7, then all the machine terminals are connected to either the positive-rail or the negative-rail of dc-link for voltage space vector, V_8 and V_7, respectively, which represents a short circuit of the stator windings. Using (5), the stator space vector voltage equations of induction machine when zero voltage vector applied from the inverters is:

$$\vec{v}_K = 0 = R_s \vec{i}_s + \sigma L_s \frac{d\vec{i}_s}{dt} + \frac{d\left(a \cdot \vec{\psi}_r\right)}{dt} \qquad (6)$$

Using (6), the rate change of rotor flux equation (i.e. rotor voltage) of induction motor is as in (7), when it is fed with zero voltage space vectors from inverter.

$$\frac{d\vec{\psi}_r}{dt} = -\frac{1}{a}\left(R_s \vec{i}_s + \sigma L_s \frac{d\vec{i}_s}{dt}\right) = \vec{v}_r^{\sim} \qquad (7)$$

where, $\vec{v}_r^{\sim}$ = calculated rotor voltage vector = $v_{r\alpha}^{\sim} + j v_{r\beta}^{\sim}$

Calculation based on actual vector output above synchronous frequency of 25 Hz

The timing duration of zero vectors is sufficient for calculation of rotor voltage using zero vector output upto synchronous frequency of 25 Hz. For above 25 Hz, rotor voltage calculation is done using actual vector output. Using (5), the rotor voltage space phasor equations of induction machine when actual vector applied through the inverters is:

$$\frac{d\vec{\psi}_r}{dt} = \frac{1}{a}\left(\vec{v}_K - R_s\vec{i}_s - \sigma L_s\frac{d\vec{i}_s}{dt}\right) = \vec{v}_r^{\sim} \qquad (8)$$

B.Rotor flux position stimation

Using either (7) or (8), the rate of change of rotor flux is integrated to get rotor flux as,

$$\vec{\psi}_r^{\sim} = \int \vec{v}_r^{\sim} dt \qquad (9)$$

Where,

$$\vec{v}_r^{\sim} = \begin{cases} -\frac{1}{a}\left(R_s\vec{i}_s + \sigma L_s\frac{d\vec{i}_s}{dt}\right); & \text{For below 25 Hz} \\ \frac{1}{a}\left(\vec{v}_K - R_s\vec{i}_s - \sigma L_s\frac{d\vec{i}_s}{dt}\right); & \text{For above 25 Hz} \end{cases} \qquad (10)$$

and $\vec{\psi}_r^{\sim}$ = estimated rotor flux vector = $\psi_{r\alpha}^{\sim} + j\psi_{r\beta}^{\sim}$;

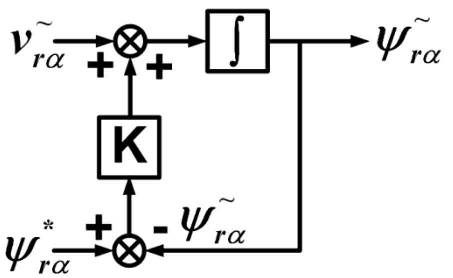

Fig. 3 Block diagram for realization of integration of calculated rotor voltage for estimation of rotor flux.

As it is seen from (10), that estimation of rotor voltage uses the induction motor parameters as stator resistance, R_s and leakage inductance, σL_s. These motor parameters are prone to vary during continuous operation of drives. The variation of parameter is needed to be considering for accurate control of drive, where correct value of parameter has to be used to estimate the rotor voltage. But there is no requirement of parameter estimation for general purpose drive where marginal variations in values of parameters do not affect the performance of drive.

The present paper also proposes new estimation method of leakage inductance, σL_s based on stator current error vector in section II.D. For stator resistance adaptation, one of the well proven stator parameters adaptation schemes from literature can be easily implemented in the present scheme, with suitable modifications for stator voltage reconstruction (using the zero vector periods) [7], [12], [20].

For flux estimation, the ideal method is integration, but the disadvantages of open integration have been addressed many times in the literature, [19], [20]. The proposed structure for such a flux estimation scheme is given in Fig. 3. In the block diagram of Fig. 3, if $\psi_{r\alpha}*$ is made zero, the estimation scheme behave like an LPF and introduces phase and magnitude error. However, the estimation scheme behaves like a high-pass filter for any DC drift occurring at the integrator output.

These α- and β- axis rotor flux positions are derived as,

$$\cos\rho = \frac{\psi_{r\alpha}^{\sim}}{\sqrt{\left(\psi_{r\alpha}^{\sim\,2} + \psi_{r\beta}^{\sim\,2}\right)}} \qquad (11)$$

$$\sin\rho = \frac{\psi_{r\beta}^{\sim}}{\sqrt{\left(\psi_{r\alpha}^{\sim\,2} + \psi_{r\beta}^{\sim\,2}\right)}} \tag{12}$$

The values of $\psi_{r\alpha}*$ and $\psi_{r\beta}*$ have to be calculated for decaying effect. The values of $\psi_{r\alpha}*$ and $\psi_{r\beta}*$ can be expressed as $\psi_{r\alpha}* = \psi_r * \cos\rho$ and $\psi_{r\beta}* = \psi_r * \sin\rho$, and the value of ψ_r* is calculated using I_{sd}^* and magnetizing inductance, *M* of the motor parameter as,

$$\psi_r^* = I_{sd}^* \cdot M \tag{13}$$

C.Speed estimation

Speed is estimated using a standard speed estimation method, [3]. To estimate speed, stator angular frequency is calculated as,

$$\omega_s = \frac{\left(v_{r\beta}^{\sim} \cdot \psi_{r\alpha}^{\sim} - v_{r\alpha}^{\sim} \cdot \psi_{r\beta}^{\sim}\right)}{\left(\psi_{r\alpha}^{\sim\,2} + \psi_{r\beta}^{\sim\,2}\right)} \tag{14}$$

where, ω_s = synchronous speed of induction motor

The slip speed is estimated using the current model of induction motor as,

$$\omega_{slip} = \frac{L_m \cdot R_r}{L_r} \frac{\left(i_{s\beta} \cdot \psi_{r\alpha}^{\sim} - i_{s\alpha} \cdot \psi_{r\beta}^{\sim}\right)}{\left(\psi_{r\alpha}^{\sim\,2} + \psi_{r\beta}^{\sim\,2}\right)} \tag{15}$$

Using (14) and (15), the angular speed of the rotor can be easily found out as the difference of synchronous speed and slip speed, which is shown in (16).

$$\omega_m = \frac{2}{P}\left(\omega_s - \omega_{Slip}\right) \tag{16}$$

where, ω_m = rotor speed, ω_{Slip} = slip speed

The estimated speed is used as speed feedback for speed control. One fact about this expression is that this estimated speed is not valid at zero synchronous frequency. According to the explanation given in [1], [2], at zero frequency, the mechanical angular velocity of the rotor does not exert and influence on stator quantities, i.e. the effect of angular velocity of the rotor is not reflected on the stator current. As, in this speed sensorless scheme, only stator voltage and stator currents are used for flux estimation and in turn for speed estimation, the speed estimation of the rotor is nor done at zero synchronous frequency using the above speed estimation algorithm. In the present work, very low speed operation closed to 1 Hz is achieved. Smooth transition from forward direction rotation to reverse direction rotation; and reverse direction rotation to forward direction rotation transition is achievable in work.

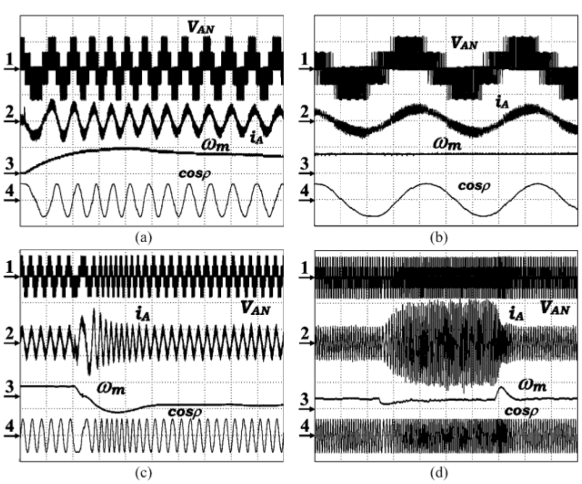

Fig. 1(a) Result of drive from standstill to 5.50 Hz operation, (b) Result of drive at steady state for 5.50 Hz operation, (c) Result of drive during speed reversal from 5.50 Hz to -5.50 Hz operation, (d) Result of drive during loading at 5.50 Hz operation: (1) Phase voltage, VAN; (2) Motor phase current, iA; (3) Speed of motor, ωm, (3d) Magnitude of estimated rotor flux, ψr; (4) Rotor flux position, cosρ. x-axis: (a) 200 ms/div, (b) 40 ms/div, (c) 200 ms/div, 2.0 s/div; y-axis: (1a), (1b): 200.0 V/div, (1c), (1d): 300.0 V/div; y-axis (2): 2.00 A/div, (2d): 4.00 A/div; y-axis (3): 240 rpm/div, (3d): 1.66 wb/div; y-axis (4): 1.6

unit/div.

Conclusion

The present sensorless vector control scheme eliminates requirements of terminal voltage sensing for entire speed operation upto six-step mode. For frequencies of operation less than 25 Hz, the rotor voltage and hence rotor flux position is computed during the inverter zero voltage space vector using steady state model for the induction motor. For above 25 Hz, active vector period and steady state model of induction motor is used. Since no terminal voltage sensing is involved, dead time effects will not create problem in rotor flux sensing at low frequency operation. But appropriate device on-state drop are compensated to achieve a steady state operation upto less than 1 Hz. Since a constant switching frequency current error space phasor based hysteresis controller is used for PWM control, a smooth transition in six-step mode is possible in the proposed scheme. Also the proposed adaptation technique for leakage inductance, σLs has been proposed exhibits very small error in estimation. Experimental results of steady state as well as transients are shown for the proposed drive starting from 0.75 Hz to 50 Hz operation with quick speed reversal. Whole control scheme is implemented on DSP platform. Experimental results show that proposed sensorless drive, with good dynamic performance and smooth operation extending upto six-step mode, can be considered as a replacement to conventional general purpose V/f control drive. However, simulation analysis for the effect of stator resistance, Rs variation on system reveals that control scheme can work satisfactorily above 5 Hz operation with marginal variation in speed error, torque develop and angle error upto resistance variation upto 20% of actual value. Resistance adaptation scheme is needed while speed operation below 5 Hz operation. For that many well proven stator parameter adaptation schemes are available in literature using only stator current sensing and can be considered with suitable modifications, for implementation in the present scheme.

Appendix

Induction motor parameters: 3.7 kW, 3Φ, 4 Poles, 415 V, 50 Hz, 1445 rpm, $R_s = 4.8\ \Omega$, $R_r = 3.8\ \Omega$, $L_s = 0.5632$ H, $L_r = 0.577$ H, and $M = 0.546$ H, Moment of inertia, $J = 0.1$ kg-m^2.

Now the project is nearly completed.
We have two publications form this

1.Fast direct torque control of an open-end induction motor drive using 12-sided polygonal voltage space vectors- IEEE Transactions on Industrial Electronics.

2.Direct torque control scheme of scheme of IM drive with 12-sided polygonal voltage space vectors. EPE(European Power Electronics Association)-2011 conference -UK

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