Design and Simulation of Field oriented control of VSI fed induction motor drive

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ABSTRACT

Field orientated control has emerged as a powerful tool for controlling induction motor drives to achieve high performance and good dynamic response. In this paper, a new control scheme is developed for control of the VSI fed induction motor drive by deriving flux signals from stator voltages and currents. In this scheme stator currents are controlled for obtaining desired torque response and the field weakening control is also considered in order to allow operation of variable-speed induction motor drives at high speeds. Simulation results of the proposed control scheme are presented.

Key words: VSI, induction motor drive, field weakening, current control.

INTRODUCTION

Variable-speed ac drive employing induction motors and static inverters has been developed in recent years to meet high standards of performance. In the past, dc motors were extensively used in variable speed applications because their flux and torque can be easily controlled. However, dc motors have certain disadvantages owing to the existence of the commutator. To overcome these obstacles, the induction motor has simple and rugged structure, low maintainability, and economy etc. but its control schemes are quite complex since the ac motor is highly coupled nonlinear multivariable control unit as compared to the dc motor with its much simpler decoupled control structure. Therefore, some simplifications are required before the induction motor variable-speed drive can become an economical alternative to dc drive [1], [2]. In a drive system the crucial variables are the flux, the torque developed and the motor speed, in dc motor drive system, the flux and the torque are related to the field and armature currents by simple direct relations. Hence, the flux and torque can be directly and independently controlled. However, in case of induction motor, no such direct relations are there and it is considered to be a complex dynamic system. A tuning procedure for high performance applications of field-oriented controller was developed [3] and the over modulation PWM techniques for inverters for field oriented control of ac drives are investigated [4], [5]. The new design of controllers for current and speed were proposed [6], [7] but more complexity is involved in the design process of the controllers.

In this paper, the flux and torque relations for an induction motor are presented in a simple format using the method of field orientation. The current controller, speed controller and flux controller are designed accurately with less complexity. The operation of the induction motor when fed from current controlled voltage source PWM inverter is evaluated and simulation results of the proposed control scheme are presented.

Dynamic model of the Induction motor

The dynamic performance of an induction motor as a control unit can be described by a set of nonlinear differential equations linking the stator and rotor currents and voltages with the mechanical quantities
torque, speed, and angular position. The advantage of rotational symmetry of an induction motor is considered by a vectorial representation in three phase or two-phase coordinates. Considerable simplification results without noticeably affecting the validity of control model if the stator and rotor windings are assumed to produce sinusoidal magnetodynamic motive force (MMF) waves in the air gap of the machine, disregarding spatial harmonics of the windings as well as slot and end effects. Additional motor inaccuracies are the assumption of infinite permeability of the iron core and neglecting of iron losses and eddy currents in the conductors.

The symmetry of the motor construction gives rise to complex currents and voltage vectors, defined in a plane perpendicular to the motor axis:

\[ i_s(t) = i_{S1}(t) + i_{S2}(t)e^{j2\pi/3} + i_{S3}(t)e^{j4\pi/3} \]  
\[ i_r(t) = i_{R1}(t) + i_{R2}(t)e^{j2\pi/3} + i_{R3}(t)e^{j4\pi/3} \]  

(1a)  

(1b)

The values \(i_{Sp}(t)\) and \(i_{Rp}(t)\) are the stator phase currents and the rotor phase currents respectively and they are arbitrary functions of time, as long as

\[ \sum_{1}^{3} i_{Sp} = 0, \quad \sum_{1}^{3} i_{Rp} = 0 \]

holds as dictated by the isolated neutrals of the windings. The instantaneous magnitudes and angles of the vectors indicate the amplitude and direction of the pertinent MMF waves.

The vector of the stator terminal voltages is defined accordingly:

\[ v_s(t) = v_{S1}(t) + v_{S2}(t)e^{j2\pi/3} + v_{S3}(t)e^{j4\pi/3} \]  

(2)

The stator and rotor inductances per phase are made equal by considering suitable turns ratio,

\[ L_s = (1+\sigma_s)L_h, \quad L_R = (1+\sigma_s)L_h \]  

(3)

With

\[ \sigma = 1-1/(1+\sigma_s)(1+\sigma_R) \]  

(4)

being the total leakage factor, \(R_s\) and \(R_r\) are the winding resistances per phase and \(L_h\) is the main inductance.

With \(J\) as the total inertia of the drive, \(T_e\) and \(T_L\) the driving and load torque respectively, \(\omega\) the angular velocity, and \(\epsilon\) the angle of rotation, the equations of the two-pole induction machine with shorted rotor windings may be written as follow:

\[ R_s i_s + L_s d_i_s + L_h d(i_R e^{j\epsilon}) = v_s \]  
\[ R_i i_s + L_i d_i_R + L_h d(i_se^{j\epsilon}) = 0 \]  
\[ Jd\omega/dt = T_e - T_L = 2/3 L_h I_m[i_s(i_R e^{j\epsilon})^*] - T_L \]
Equations (5)-(8) constitute a dynamic extension of the well-known stationary model of the induction machine in synchronously rotating reference frame operating with constant speed with the sinusoidal stator currents described by constant complex phasors,

\[ i_{s1}(t) = \sqrt{2}/2[I_s e^{j\omega_l t} + I_s e^{-j\omega_l t}], \quad I_s = I_s e^{j\alpha}, \quad I_s = I_s e^{j\alpha} \]

\[ i_{s2}(t) = \sqrt{2}/2[I_s e^{j(\omega_l t - 2\pi/3)} + I_s e^{-j(\omega_l t - 2\pi/3)}], \]

\[ i_{s3}(t) = \sqrt{2}/2[I_s e^{j(\omega_l t - 4\pi/3)} + I_s e^{-j(\omega_l t - 4\pi/3)}], \]

the current vector defined in (1a),

\[ i_s(t) = 3\sqrt{2}/2 I_s e^{j\omega_l t} \]

Exhibits constant magnitude and angular velocity, indicating a constant circumferential MMF wave moving at constant speed. By also introducing phasor representation for the sinusoidal stator voltages and rotor currents, equations (5)-(8) can be written as follows

\[ (R_s+j\omega_l L_s)I_s+j\omega_l L_s I_R = V_s, \]

\[ (R_r/s+j\omega_l L_r)I_r+j\omega_l L_r I_s = 0, \]

\[ T_{el} = 3L_{h}\Im[I_s I_r^*] = \text{Constant}, \]

\[ \omega = \text{Constant}, \]

where \( s=(\omega_l-\omega)/\omega_l \) represents the rotor slip. This corresponds to the well-known single-phase equivalent circuit in stationary condition.
Control strategy

An ac induction motor has a more complex control structure than a dc motor. This is due to the complex coupling structure between all the control quantities such as flux, electromagnetic torque and stator currents etc.

It has been shown by “Blaschke” that these problems may be overcome by field orientation, essentially reducing the control dynamics of an ac motor to those of a separately excited compensated dc motor. This is achieved by defining a time varying vector, which corresponds to a sinusoidal flux wave moving in the airgap of the machine. When referring the MMF wave of the stator currents described by \(i_S(t)\) to this flux wave, it realizes that only the quadrature component of \(i_S(t)\) is contributing to the torque, where as the direct axis component affects the magnitude of the flux. Hence, the stator current vector is defined in a frame of reference defined by the time-varying field or in field coordinates. This indicates a close correspondence to a dc machine with the direct component of the stator current vector being analogous to the field current and the quadrature component to the armature current. The flux-oriented frame of reference may also be compared to the d-q co-ordinates commonly used with synchronous machines; however, now the reference is no longer determined by rotor position. The concept of field orientation thus provides a clue to a very effective method of controlling ac motors.

A modified magnetizing current vector representing the flux reference is defined as

\[
i_{mR}(t) = i_S(t) + (1+\sigma)I_R(t)e^{j\rho} \quad \text{---- (14)}\]

Under stationary conditions this is the magnetizing current inducing the voltage \(V_{mr}\).

The instantaneous angular velocity of this vector is the stator frequency

\[
d\rho/dt = \omega_l(t) \quad \text{------------------(15)}
\]

With the vector \(i_{mR}(t)\), the direct and quadrature components may be defined according to fig.1

\[
i_S(t)e^{j\rho} = i_{sd} + j i_{sq} \quad \text{------------- (16)}
\]

\[
v_S(t)e^{j\rho} = v_{sd} + j v_{sq} \quad \text{-------- (17)}
\]

At constant speed and torque the transformed variables are constant except for ripple caused by inverter operation, using the equations (14)-(17) in the motor model and separating real and imaginary parts results in six real differential equations for the transformed variables with \(TS = L_S/R_R\) and \(TR = L_R/R_R\)

\[
\sigma_S T_S i_{sd}/dt + i_{sd} = v_{sd}/R_S + \omega_l \sigma T_SI_S i_{sq} - (1-\sigma)T_S dM_t/dt \quad \text{---------(18)}
\]

\[
\sigma_S T_S i_{sq}/dt + i_{sq} = v_{sq}/R_S - \omega_l \sigma T_SI_S i_{sd} - (1-\sigma)\omega_l T_S i_{mR} \quad \text{(19)}
\]
Equations (18) and (19) describe the transition from the field oriented voltage components $v_{Sd}$ and $v_{Sq}$ to the current components $i_{Sd}$ and $i_{Sq}$ with small leakage time constants and some interactions, which can be balanced by suitable control.
The slow dynamics in the direct axis are governed by (20), with $i_{sd}$ acting as control input. The angular position of the flux vector is determined by the (21) with $i_{sq}$ serving as a convenient input in order to control the torque (22) as the product of $i_{mR}$ and $i_{sq}$ analogous to the control of dc motor, it is the best policy to maintain $i_{mR}$ at its maximum level, limited either by iron saturation or, above base speed, the ceiling voltage of the inverter. This also removes some of the nonlinear interactions.

It is noted that the control scheme could be simplified by employing quasi-impressed stator currents obtained through fast current control of the inverter [4],[6]-[7]. The transformed machine equations are kept in transfer function form in the fig 2, which also shows the transformation in two steps from the three terminal voltages $v_{S1}$, $v_{S2}$, and $v_{S3}$ to the field-oriented voltages $v_{Sd}$ and $v_{Sq}$.

The three-phase inverter shown in fig.2 supplies three independently controlled voltages, usually pulse width modulated (PWM) square waves. The control dynamics are approximated by a small delay $T_D$, which is determined by the clock frequency. It contains an inverse coordinates transformation by modulating the dc output quantities of the controllers with $\sin \rho$, $\cos \rho$ in order to generate the ac voltage references for the inverters. The nonlinear interactions in the motor are reduced by maintaining $i_{mR}$ constant at its maximum level. It is seen that the motor controls its stator frequency through feedback from the flux signal; therefore, no pullout effect exists. If the motor is overloaded or if the speed reference is changed too quickly for the motor to follow, the speed controller will saturate and the motor generates constant maximum torque. Thus the drive can operate safely in all four quadrants of the speed-torque plane, as is the case with a dc motor supplied from a four-quadrant converter.

For the field oriented control method it is necessary to know both the magnitude and angle of the flux vector, this vector is obtained from the stator variables as follows:

\[
\psi_{mr}(t) = (1+\sigma_R) \psi_m(t) - \sigma_R L_h i_{s}(t) \equiv L_h i_{mr}(t)
\]

The flux computation is essentially a rectangular to polar conversion as shown in fig.2. $i_{mr}$ and $\rho$ are magnitude and phase angle of the flux vector. The angle information is used to demodulate the stator currents into direct and quadrature components $i_{sd}$ and $i_{sq}$. These signals directly relate to the magnitude of the field and the torque, which enables the control to be decoupled into two loops, one for the field control and one for the torque control. The outputs from the controllers are modulated by $\sin \rho$, $\cos \rho$ in order to produce stator voltage references for the inverter.

**Simulation results**

The proposed control scheme is designed and simulated using matlab /simulink software. The simulink representation of the block diagram with main components is shown in fig 3. To demonstrate the proposed control scheme, simulation results of the proposed control scheme are shown in fig.4. The current commands in the synchronously rotating frame denoted by $i_{sd}^*$ and $i_{sq}^*$, must be transformed into three phase domain to yield the reference current commands, denoted by $i_{sa}^*$, $i_{sb}^*$, $i_{sc}^*$ using the coordinate translator and sin/cos generator. The actual stator currents of the induction motor, $i_{sa}$, $i_{sb}$, $i_{sc}$ are generated from the PWM inverter by comparing the current commands with three independent two level hysteresis comparator. In the simulation, a dc link voltage of 100v is fed to the proposed control scheme inverter. The dead band of current controller in the proposed control scheme is 0.3A. The proportional and integral of speed controller, $K_p$ and $K_i$ in this control scheme are 2.5 and 0.8 respectively. The control scheme is simulated for 2sec. It is observed from the results that the rotor flux is constant and speed is also constant due to constant values of rotor flux. In the simulation the speed has taken 0.45-sec time to reach the final speed, after that the speed and rotor flux will remain constant even the load torque changes. From the result it can also be noticed that the speed response of the proposed control scheme is fast and also the current and the torque ripple of the proposed control scheme are reduced considerably.
Fig 3. Simulink representation of the block diagram showing main components.
Fig 4(a) Rotor flux vs time
Fig 4 Simulations results: Rotor flux vs time and isd vs time.

Fig 4(c) isd vs time

Fig.4 Simulations results: Rotor flux vs time and isd vs time.

Fig 4(b) speed vs time
Fig 4 Simulations results: Speed vs time and isq vs time.

Fig 4(d) ids vs time

Fig.4 Simulations results: Speed vs time and isq vs time.

Fig 4(e) torque vs time
Fig 4 Simulations results: Torque vs time and isa vs time.

Fig 4(f) isa vs time

Fig.4 Simulations results: Torque vs time and isa vs time.
CONCLUSIONS:
The field oriented control for VSI fed induction motor is designed and simulated. The field-oriented control method has been shown to be an effective means for controlling an ac induction motor. The control scheme is functioning with a flux feedback signal based on stator voltages and currents. The proposed control is simple to implement. It has been observed that the transients in the torque are less and also variations in speed are minimal.

REFERENCES

Nomenclature
Rs = Stator winding resistance
Rr = referred rotor winding resistance
Xm = stator magnetizing reactance
Xls = stator leakage reactance
Xlr' = referred rotor leakage reactance
d, q = direct axis and quadrature axis quantities
ε = rotor position angle
σ = Rotor leakage factor
S, R = Stator and Rotor quantities

Appendix
Motor Data:
20 hp
V_{rated} = 220 v
I_{rated} = 45.8 amps
Frequency = 60 Hz
Slip = 0.0287
N_{rated} = 1748 rpm
Rated r.m.s phase current = 49.68 amps

r_s = 0.1062 Ω
r_r = 0.0764 Ω
\dot{x}_{ls} = x_{lr}' = 0.2145 Ω
x_m = 5.8334 Ω
Rotor inertia, J = 2.8 Kg-m^2
Inertia constant, H = 3.33 Kg-m^2/s