Determination of Regulator Parameters and Transient Analysis of Modified Self-commutating CSI-fed IM Drive

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Abstract – In this paper, an attempt has been made to design the current and speed proportional and integral (PI) regulators of self-commutating current source inverter-fed induction motor drive having capacitors at the machine end and to investigate the transient performance of the same for step changes in reference speed. The mathematical model of the complete drive system is developed in closed loop, and the characteristic equations of the systems are derived using perturbation about steady-state operating point in order to develop the characteristic equations. The D-partition technique is used for finding the stable region in the parametric plane. Frequency scanning technique is used to confirm the stability region. Final selection of the regulator parameters is done by comparing the transient response of the current and speed loops for step variations in reference. The performance of the drive is observed analytically through MATLAB simulation.

Keywords: Current source inverter, D-partition technique, Frequency scanning, Induction motor drive, Speed and current regulator, Transient performance

1. Nomenclature

d, q	Direct and quadrature axes	ψ_{ds}	Flux linkages of the d-axis of the stator windings
v_{as}, v_{bs}, v_{cs}	Phase voltages of the PWM inverter	ψ_{qs}	Flux linkages of the q-axis of the stator windings
v_{ds}^e	d-axis stator voltage in synchronously rotating reference frame	ψ_{dr}	Flux linkage of the d-axis of the rotor winding
i_{ds}^e	d-axis stator current in synchronous rotating reference frame	ψ_{qr}	Flux linkage of the q-axis of the rotor winding
v_{qs}^e	q-axis stator voltage in synchronously rotating reference frame	I_{ref}	Reference DC link current
i_{qs}^e	q-axis stator current in synchronously rotating reference frame	V_{inv}	Input voltage of the inverter
i_{dr}^e	d-axis rotor current in synchronously rotating reference frame	V_r	Rectifier output voltage
i_{qr}^e	q-axis rotor current in synchronously rotating reference frame	R_f	Resistance of DC link inductor
i_{as}, i_{bs}, i_{cs}	Line currents of PWM inverter	L_f	Inductance of DC link inductor
I_{DC}	DC link current	R_s	Resistance of stator winding per phase
I_{act}	Active component of stator current	L_s	Self-inductance of stator winding per phase
I_{act}^*	Reference active component of stator current	R_r	Resistance of rotor winding per phase
I_{react}	Reactive component of stator current	L_r	Self inductance of rotor winding per phase
I_{react}^*	Reference reactive component of stator current	L_m	Mutual inductance per phase
ω_e	Synchronous speed of the induction motor	L_l	$L_s L_r - L_m^2$
ω_r	Rotor speed of the induction motor	C	Capacitance per phase
ω_r^*	Reference speed	J	Moment of inertia in kg-m ²
ω_{sl}	Slip speed of the induction motor	B	Viscous friction coefficient
ω_{sl}^*	Reference slip speed	β	Pulse width of PWM rectifier
		V_{LL}	Line-to-line input voltage of the rectifier
		P	Number of poles
		I_c	Instantaneous phase current of capacitor
		V_s	Instantaneous stator phase voltage
		i_{dc}^e	d-axis capacitor current in synchronously rotating reference frame
		i_{qc}^e	q-axis capacitor current in synchronously rotating reference frame
		k_{pi}	Proportional gain of current regulator
		k_i	Integral gain of current regulator
		k_{ps}	Proportional gain of speed regulator
		k_{is}	Integral gain of speed regulator
		p	Differential operator (d/dt) or complex frequency

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<i>Suffix 0</i>	Represents a variable in steady-state conditions
σ	Degree of relative stability
ω	Natural frequency
k_1	Slope of stator active current (I_{act}) vs. slip speed (ω_{sl})
k_2	Slope of stator reactive current vs. slip speed (ω_{sl})
k_{11}	$\frac{\text{Rated value of capacitor current per phase}}{[\text{Rated angular frequency } (\omega_e)]^2}$

2. Introduction

Variable and smooth speed control is the mandatory requirement for any industrial drive. The speed control of induction motors over a wide range can be done with the help of variable frequency VSI or CSI. Due to the controlled current operation of the inverter, slip-regulated CSI is preferred over VSI. The current source at the front end makes the system naturally capable of power regeneration [1]-[4]. In this paper, the closed loop scheme of self-commutating current source inverter-fed induction motor drive employing two PI regulators - one in the outer speed feedback loop and other in the inner current feedback loop, is discussed. The outer speed loop is used to obtain the reference DC link current from the slip speed, whereas the inner current loop regulates the DC link current according to the load torque demand [5]. The design of the feedback loop parameters requires complete analysis and simulation of the entire drive. The parameters for current and speed PI regulators of a self-commutating CSI-fed induction motor drive are designed on the basis of system relative stability, frequency scanning, and transient response of the drive.

3. System Description

The CSI-fed induction motor drive consists of a three-phase AC source, a PWM rectifier, a DC link smoothing

reactor, a current-controlled inverter, a three-phase squirrel cage induction motor, and a three-phase capacitor bank, as shown in Fig. 1. A fast-response speed-regulating drive can be realized by incorporating PI regulators in the feedback loop [6]-[8]. Two PI regulators are used - one in the speed feedback loop and the other in the current feedback loop.

The outer speed regulator compares the reference speed and the actual rotor speed and processes the speed error to obtain the reference slip speed, which is used to obtain the reference stator active and reactive currents and hence, the reference DC link current. The inner current PI regulator compares the reference and actual DC link currents and adjusts the turn on period of PWM pulses, hence the output voltage of the converter.

4. Mathematical Modeling

The modeling of the modified CSI-fed induction motor drive is carried out in synchronously rotating reference frame for the following:

- (1) Three-phase PWM rectifier
- (2) Three-phase PWM inverter
- (3) DC link
- (4) Three-phase induction motor with load
- (5) Three-phase capacitor bank

4.1 Three-phase PWM Rectifier

The PWM rectifier output voltage depends on the number of pulses per cycle and their widths. The converter is modeled for 12 numbers of equal pulses per cycle. It leads to two pulses per 60° each of β width. The average output voltage of the PWM rectifier can be expressed as follows:

$$V_r = \frac{3\sqrt{2}}{\pi} V_{LL} (4 \sin \frac{5\pi}{12}) \sin \frac{\beta}{2} \quad (1)$$

Fig. 1. Variable speed modified current source inverter-fed induction motor drive.

Since β is varied from 10% to 90% of $(\pi/6)$ radians, it can be approximated as follows: $\sin(\beta/2) \cong (\beta/2)$

$$V_r = 5.218 V_{LL} (\beta/2) \quad (2)$$

4.2 Three-phase PWM Inverter

The fundamental component of the line currents of the three-phase pulse width modulated inverter i_{as} , i_{bs} , and i_{cs} forms a balanced set of three-phase currents with maximum value as $I_{as(max)}$ and can be expressed as follows:

$$I_{as(max)} = k I_{DC} \quad (3)$$

Where k is obtained through Fourier analysis of inverter line current waveforms, and this is given by the following:

$$k = \frac{\text{maximum value of fundamental inverter line current}}{\text{DC link current } (I_{DC})}$$

The value of k depends on the operating frequency of the inverter and varies from 0.8485 to 0.9970 for variation in operating frequencies from 10 to 50 Hz. Since the inverter output fundamental current peak is taken along the q^e axis of the reference frame, the transformed phase current equations in the q^e - d^e reference frame are as follows:

$$i_{0s}^e = 0; \quad i_{qs}^e = k I_{DC}; \quad i_{ds}^e = 0 \quad (4)$$

Assuming power loss in the inverter to be negligible, i.e., input power = output power, we can derive the following:

$$V_{inv} I_{inv} = v_{as} i_{as} + v_{bs} i_{bs} + v_{cs} i_{cs} = \frac{3}{2} (v_{qs}^e i_{qs}^e + v_{ds}^e i_{ds}^e) \quad (5)$$

Substituting the values of i_{qs}^e , i_{ds}^e , and I_{inv} in terms of I_{DC} , the following equation is obtained:

$$V_{inv} = 1.5 k v_{qs}^e \quad (6)$$

4.3 DC Link

The rectifier output voltage V_r is the sum of the inverter input voltage V_{inv} and DC link voltage, hence

$$V_r = 1.5 k v_{qs}^e + (R_f + pL_f) I_{DC} \quad (7)$$

4.4 Three-phase Induction Motor

The induction motor can be modeled in the q^e - d^e reference frame using the following assumptions:

- (1) The three-phase stator windings of the motor are balanced and sinusoidally distributed in space.
- (2) The DC link current is ripple free.
- (3) The inverter switching transients are ignored.

(4) There is no core loss in the motor.

(5) The motor can be described by fourth-order matrix equation in q^e - d^e reference frame as follows:

$$\begin{bmatrix} v_{qs}^e \\ v_{ds}^e \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} R_s + pL_s & \omega_e L_s & pL_m & \omega_e L_m \\ -\omega_e L_s & R_s + pL_s & -\omega_e L_m & pL_m \\ pL_m & \omega_{sl} L_m & R_r + pL_r & \omega_{sl} L_r \\ -\omega_{sl} L_m & pL_m & -\omega_{sl} L_r & R_r + pL_r \end{bmatrix} \begin{bmatrix} i_{qs}^e \\ i_{ds}^e \\ i_{qr}^e \\ i_{dr}^e \end{bmatrix} \quad (8)$$

The electromagnetic torque equation of the motor is expressed as follows:

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

The equation of motion of the drive is given by the following:

$$T_e = T_l + J \frac{d\omega_r}{dt} + B \omega_r \quad (10)$$

The load torque equation is expressed as

$$T_l = T_L (\omega_r / \omega_{base}) \quad (11)$$

4.5 Three-phase Capacitor Bank

The capacitor current is related to the stator voltage of the induction motor, as shown below:

$$i_c = C \frac{dv_s}{dt} \quad (12)$$

Transforming (12) in the synchronously rotating reference frame q^e - d^e , we have the following:

$$(i_{dc}^e \cos \omega_e t - i_{qc}^e \sin \omega_e t) = C \frac{d}{dt} (v_{ds}^e \cos \omega_e t - v_{qs}^e \sin \omega_e t) \quad (13)$$

Differentiating (13) and comparing the terms on both sides, d -axis and q -axis currents are expressed as follows:

$$i_{dc}^e = C (pv_{ds}^e - \omega_e v_{qs}^e) \quad (14)$$

$$i_{qc}^e = C (pv_{qs}^e + \omega_e v_{ds}^e) \quad (15)$$

5. Design of Regulator Parameters

The transient response of the drive depends on the choice of speed and current regulator parameters, which can be selected on the basis of the stability and response of the drive by applying the D-partition technique [6]. The D-partition method is used to study the effect of two parameters on the control system on stability and transient performance. With regard to stability analysis, the method

provides a possibility of defining the relative stability of the control system with the roots of the characteristic equation lying within a specified p -plane contour of general shape [9]-[10]. The p -plane with two adjustable parameters as coordinate axis is termed as “parametric plane”.

For the determination of regulator parameters by the D-partition technique, the non-linear equations of the system are linearized by the method of small perturbation about the steady-state operating point, and the characteristic equations are developed for both current and speed regulator loops. The D-partition boundaries are plotted in the parametric plane and the probable stable regions are identified in the plane of the proportional and integral gain parameters for both the regulators.

Frequency scanning check is used for further confirmation of the probable stable region obtained from the D-partition boundaries [6]. To apply this technique, a point from the inside of the probable stable region and a point outside the probable stable region of the D-partition boundary are selected for both current and speed regulators. The real and imaginary parts of the characteristic equation of the current and speed regulators are plotted for the points inside and outside the stable region for varying ω from $-\infty$ to $+\infty$. If the origin lies inside the hatched curve, the system is stable, whereas if it is outside the hatched region, the system is unstable.

5.1 Design of Current Regulator

For the design of current regulator parameters (k_{p_i} and k_{i_i}), the speed of the motor is assumed to be constant since the mechanical time constant of the motor is much higher than the electrical time constant; hence, ω_{sl}^* and ω_e are also considered as constant for the design of current regulator parameters. With small perturbation in machine-operating variables around steady-state operating points, (rated point) a linear relation in drive variables is obtained. Perturbing the stator voltages, stator currents, rotor currents, rotor voltages, capacitor currents, stator active currents, stator reactive currents, and rotor speed, the final perturbed equations used for the determination of current and speed characteristic equations are as follows:

$$F_{24}(p)\Delta v_{ds}^e - F_{25}(p)\Delta v_{qs}^e + k F_6(p)\Delta I_{DC} = 0 \quad (16)$$

$$F_{25}(p)\Delta v_{ds}^e + F_{24}(p)\Delta v_{qs}^e - k F_5(p)\Delta I_{DC} = 0 \quad (17)$$

$$\left(k_{p_i} + \frac{k_{i_i}}{p} + R_f + pL_f\right)\Delta I_{DC} + \frac{3}{2}k\Delta v_{qs}^e = 0 \quad (18)$$

$$F_{18}(p)\Delta v_{ds}^e + F_{19}(p)\Delta v_{qs}^e + k F_{15}(p)\Delta I_{DC} + \left[-F_{26}(p)\left(k_{p_s} + \frac{k_{i_s}}{p}\right) - F_{27}(p)\right]\Delta \omega_r = 0 \quad (19)$$

$$[F_{24}(p)\Delta v_{ds}^e - F_{25}(p)\Delta v_{qs}^e] = -[F_7(p)\psi_{dr}^e - F_8(p)\psi_{qr}^e - F_{22}(p)]\Delta \omega_r - k F_6(p)\Delta I_{DC} + F_{22}(p)\Delta \omega_{sl} \quad (20)$$

$$[F_{25}(p)\Delta v_{ds}^e + F_{24}(p)\Delta v_{qs}^e] = -[F_8(p)\psi_{dr}^e + F_7(p)\psi_{qr}^e - F_{23}(p)]\Delta \omega_r + F_{23}(p)\Delta \omega_{sl} + k F_5(p)\Delta I_{DC} \quad (21)$$

$$\left(k_{p_i} + \frac{k_{i_i}}{p}\right) \left[\left\{ \left(k_{p_s} + \frac{k_{i_s}}{p}\right) (-A_1 - A_2) \right\} + A_2 \right] \Delta \omega_r - \left(k_{p_i} + \frac{k_{i_i}}{p} + R_f + pL_f\right) \Delta I_{DC} - \frac{3}{2}k\Delta v_{qs}^e = 0 \quad (22)$$

Expressing (16)-(18) in matrix form, the current regulator matrix equation is given as follows:

$$\begin{bmatrix} 0 & 1.5k & \left(k_{p_i} + \frac{k_{i_i}}{p} + R_f + pL_f\right) \\ F_{24}(p) & -F_{25}(p) & k F_6(p) \\ F_{25}(p) & F_{24}(p) & -k F_5(p) \end{bmatrix} \begin{bmatrix} \Delta v_{ds}^e \\ \Delta v_{qs}^e \\ \Delta I_{DC} \end{bmatrix} = 0 \quad (23)$$

Finally, the characteristic equation of the current regulator matrix is expressed as the below equation:

$$k_{p_i}a_4 + k_{i_i}a_3 + a_5 = 0 \quad (24)$$

Separating the real and imaginary parts of (24) and solving them, the values for k_{p_i} and k_{i_i} are given as follows:

$$k_{p_i} = \frac{\begin{vmatrix} -a_{5r} & a_{3r} \\ -a_{5i} & a_{3i} \end{vmatrix}}{\Delta} \quad (25)$$

$$k_{i_i} = \frac{\begin{vmatrix} a_{4r} & -a_{5r} \\ a_{4i} & -a_{5i} \end{vmatrix}}{\Delta} \quad (26)$$

$$\text{Where } \Delta = a_{4r}a_{3i} - a_{4i}a_{3r} \quad (27)$$

All variables in the characteristic equation are the function of p . Substituting $p = (-\sigma + j\omega)$ in (25)–(27) and varying ω from $-\infty$ to $+\infty$ for $\sigma = 0.0$, the D-partition boundary is plotted in the (k_{p_i}, k_{i_i}) plane, as shown in Fig. 2.

The stability of the identified region is confirmed by the frequency scanning technique for the points inside and outside the region. The characteristic vector $D(p) = k_{p_i}a_4 + k_{i_i}a_3 + a_5$ is evaluated at a selected point inside this region ($\sigma = 0.0, k_{p_i} = 0.9, k_{i_i} = 278$) and is plotted in real-imaginary plane, as shown in Fig. 3, for varying ω from $-\infty$ to $+\infty$. The curve of the characteristic vector encloses the origin. Hence, the region is stable and any point in it can be selected.

Another point ($\sigma = 0.0, k_{p_i} = 12, k_{i_i} = 800$) is taken from outside, above the region, and the characteristic vector is plotted as shown in Fig. 4 for varying ω from $-\infty$ to $+\infty$. The origin is not encircled by the curve and hence, the region outside the curve is unstable.

Fig. 2. D-partition boundary of current regulator for $\sigma = 0$.

Fig. 3. Stability investigation for the selected point in the region of Fig. 2 using frequency scanning (stable).

Fig. 4. Stability investigation for the selected point outside the region of Fig. 2 using frequency scanning (unstable).

The $\sigma = 0$ gives the probable stable region. As the value of σ is continuously increased the probable stable region shrinks, and the most probable stable region is achieved. After a certain value of the σ , the stable region vanishes. In order to identify the most stable region, the relative D-partition boundaries are plotted for the different values of σ and varying ω from $-\infty$ to $+\infty$, as shown in Fig. 5. At $\sigma = 10$, the most probable stable region is obtained.

The exact decision on current regulator parameters is made by determining the transient response of the current loop for step variation in reference current from 1.0 to 1.05 of the rated current. The transient response curves for the current loop are plotted for different sets of k_{pi} and k_{ii} values, which were chosen from the most probable stable region of Fig. 5. The parameters, for which the settling time and percentage peak overshoot are moderate, are finally selected as regulator parameters.

The transient response curves are shown in Fig. 6, which correspond to the different sets of k_{pi} and k_{ii} values chosen from the most probable stable region of Fig. 5. The magnified curves are shown in Fig. 7. The settling time and percentage overshoot corresponding to these chosen values of k_{pi} and k_{ii} are shown in Table 1.

The comparison of various transient response curves shows that for $(k_{pi} = 0.9 \text{ and } k_{ii} = 278.0)$, the response is fastest and the percentage peak overshoot is the minimum. Hence, the selected transfer function of current regulator is given by $\left(0.9 + \frac{278}{p}\right)$.

Fig. 5. D-partition boundary of current regulator for different values of σ .

Fig. 6. Transient response of the current loop for different sets of k_{pi} and k_{ii} .

Fig. 7. Magnified view of the transient response of the current loop.

Table 1. Settling time and percentage overshoot corresponding to different sets of k_{pi} and k_i values

Sr.	k_{pi}	k_i	Settling time (ms)	Peak overshoot
1	0.9	290	417	0.37%
2	1.0	288	418	0.34%
3	1.0	284	416	0.28%
4	0.9	278	415	0.22%

5.2 Design of Speed Regulator

Expressing (19)-(22) in matrix form, the speed regulator matrix equation is given as follows:

$$\begin{bmatrix}
 F_{18}(p) & F_{19}(p) & -F_{26}(p)\left(k_{ps} + \frac{k_{is}}{p}\right) - F_{27}(p) \\
 F_{24}(p) & -F_{25}(p) & F_7(p)\psi_{dr}^e - F_8(p)\psi_{qr}^e + \left(k_{ps} + \frac{k_{is}}{p} - 1\right)F_{22}(p) \\
 F_{25}(p) & F_{24}(p) & F_8(p)\psi_{dr}^e + F_7(p)\psi_{qr}^e + \left(k_{ps} + \frac{k_{is}}{p} - 1\right)F_{23}(p) \\
 0 & -1.5k & \left(k_{pi} + \frac{k_{ii}}{p}\right)\left[\left(k_{ps} + \frac{k_{is}}{p}\right)(-A_1 - A_2) + A_2\right]
 \end{bmatrix}
 \begin{bmatrix}
 kF_{15}(p) \\
 kF_6(p) \\
 -kF_5(p) \\
 -\left(k_{pi} + \frac{k_{ii}}{p} + R_j + pL_f\right)
 \end{bmatrix}
 \begin{bmatrix}
 \Delta v_{dr}^e \\
 \Delta v_{qr}^e \\
 \Delta \omega_r \\
 \Delta I_{DC}
 \end{bmatrix} = 0 \quad (28)$$

The characteristic equation of the current regulator matrix is expressed as follows:

$$k_{ps} c_{33} + k_{is} c_{31} + c_{34} = 0 \quad (29)$$

Separating the real and imaginary parts of (29) and solving them, the values for k_{ps} and k_{is} are given as the following:

$$k_{ps} = \frac{\begin{vmatrix} -c_{34r} & c_{31r} \\ -c_{34i} & c_{31i} \end{vmatrix}}{\Delta} \quad (30)$$

$$k_{is} = \frac{\begin{vmatrix} c_{33r} & -c_{34r} \\ c_{33i} & -c_{34i} \end{vmatrix}}{\Delta} \quad (31)$$

$$\text{Where } \Delta = c_{33r} c_{31i} - c_{31r} c_{33i} \quad (32)$$

Using (30)-(32) and varying ω from $-\infty$ to $+\infty$, for $\sigma = 0.0$, the D-partition boundary is plotted in the (k_{ps}, k_{is}) plane, as shown in Fig. 8. Any point lying within the shaded region may be considered as a stable point. The stability of the identified region is confirmed by the frequency scanning technique for points inside and outside this region. The characteristic vector $D(p) = k_{ps} c_{33} + k_{is} c_{31} + c_{34}$ is evaluated at a selected point inside the region ($\sigma = 0.0, k_{ps} = 0.23, k_{is} = 0.32$) and is plotted in real-imaginary plane by varying ω from $-\infty$ to $+\infty$, as shown in Fig. 9. The curve of the characteristic vector encloses the origin. Hence, the region is stable and any point in it can be selected.

Another point ($\sigma = 0.0, k_{ps} = 2, k_{is} = 5$) is taken from outside, above the region, and the characteristic vector is plotted, as shown in Fig. 10 for varying ω from $-\infty$ to $+\infty$. The origin is not encircled by the curve and hence, the region outside the curve is unstable. In order to get most stable region, the relative D-partition boundaries are plotted for the different values of σ and varying ω from $-\infty$ to $+\infty$, as shown in Fig. 11. At $\sigma = 2.76$, the most probable stable region is obtained.

The exact decision on speed regulator parameters can be made by determining the transient response of the speed loop for the step variation in the reference speed from 1.0 to 1.05 of the rated speed. The transient response curves for

Fig. 8. D-partition boundary of speed regulator for $\sigma = 0$.

Fig. 9. Stability investigation for the selected point in the region of Fig. 8 using frequency scanning (stable).

Fig. 10. Stability investigation for the selected point outside the region of Fig. 8 using frequency scanning (unstable).

Fig. 11. D-partition boundary of speed regulator for different values of σ .

the speed loop are plotted for the different sets of k_{ps} and k_{is} values chosen from the most probable stable region of Fig. 11. The parameters for which the settling time and percentage peak overshoot are moderate are finally selected as regulator parameters.

The transient response curves are shown in Fig. 12, which corresponds to different sets of k_{ps} and k_{is} values chosen from the most probable stable region of Fig. 11. The magnified curves are shown in Fig. 13. The settling time and percentage overshoot corresponding to these chosen values of k_{ps} and k_{is} are shown in Table 2.

The transient response curves show that for ($k_{ps} = 0.23$ and $k_{is} = 0.32$), the percentage peak overshoot is within the permissible tolerance limit and the settling time is near the fastest. Hence, the transfer function of the speed regulator is given by $\left(0.23 + \frac{0.32}{p}\right)$.

Fig. 12. Transient response of the speed loop for different sets of k_{ps} and k_{is} .

Fig. 13. Magnified view of the transient response of the speed loop.

Table 2. Settling time and percentage overshoot corresponding to various values of k_{ps} and k_{is}

Sr.	k_{ps}	k_{is}	Settling time (s)	Peak overshoot
1	0.28	0.45	2.2	1.41%
2	0.25	0.40	2.2	1.23%
3	0.23	0.32	2.4	0.94%
4	0.22	0.24	2.7	0.62%

6. Performance Investigation of the Drive

Having designed the parameters of both current and speed regulators, the performance of the drive is investigated through MATLAB simulation by implementing the designed values of regulator parameters and analyzing the current and speed transient responses of the drive for different step variations in reference speed. The current and speed transient responses of the drive for step changes in reference speed each after an interval of 10 seconds are shown in Figs. 14-15.

The effectiveness of both regulators can be demonstrated by considering the current and speed transient responses of the drive separately for each alteration in reference speed. Figs. (16)-(19) show the current transient responses of the drive separately for step changes in reference speed from 200 rad/s to 250 rad/s, 250 rad/s to 298.2999 rad/s, 298.2999 rad/s to 250 rad/s, and 250 rad/s to 200 rad/s, respectively. Figs. (20)-(23) show the speed transient responses of the drive separately for step changes in reference speed from 200 rad/s to 250 rad/s, 250 rad/s to 298.2999 rad/s, 298.2999 rad/s to 250 rad/s, and 250 rad/s to 200 rad/s, respectively.

Fig. 14. Current transient response of the drive for step changes in reference speed each after an interval of 10 seconds.

Fig. 15. Speed transient response of the drive for step changes in reference speed each after an interval of 10 seconds.

These response curves show that the settling time of the drive for the designed values of regulator parameters and for the different step variations in reference speed are 3.7, 2.8, 3.6, and 3.1 seconds, respectively. This shows the effectiveness of the closed loop operation of the drive at the designed speed and current regulator parameters.

Fig. 16. Current transient response of the drive for step changes in reference speed from 200 rad/s to 250 rad/s.

Fig. 17. Current transient response of the drive for step changes in reference speed from 250 rad/s to 298.2999 rad/s.

Fig. 18. Current transient response of the drive for step changes in reference speed from 298.2999 rad/s to 250 rad/s.

Fig. 19. Current transient response of the drive for step changes in reference speed from 250 rad/s to 200 rad/s.

Fig. 20. Speed transient response of the drive for step changes in reference speed from 200 rad/s to 250 rad/s.

Fig. 21. Speed transient response of the drive for step changes in reference speed from 250 rad/s to 298.2999 rad/s.

Fig. 22. Speed transient response of the drive for step changes in reference speed 298.2999 rad/s to 250 rad/s.

Fig. 23. Speed transient response of the drive for step changes in reference speed from 250 rad/s to 200 rad/s.

7. Conclusions

The D-partitioning technique was applied to determine the current and speed regulator parameters. The characteristic equations needed for the determination of the parameters was obtained by perturbing the machine variables around the rated conditions. The most probable region was obtained by plotting the parametric plane for the different values of σ . The frequency scanning check was applied to further confirm the stable region. The final selection of regulator parameters was carried out by determining the transient response of the current and speed loops of the drive. With the regulator parameters obtained by the D-partitioning technique, the effectiveness of the closed loop operation of the drive at designed speed and current regulator parameters was analyzed for different step variations in the reference speed.

Appendix

Name plate ratings of induction motor

1 HP, three-phase, 400 V, 50 Hz, four-pole, 1425 r.p.m., star connected

Induction motor parameters

$$R_s = 3.520 \, \Omega \quad R_r = 2.780 \, \Omega \quad L_s = 0.165 \, \text{H} \\ L_r = 0.165 \, \text{H} \quad L_m = 0.150 \, \text{H} \quad J = 0.01289 \, \text{kg-m}^2$$

DC link parameters

$$R_f = 0.250 \, \Omega, \quad L_f = 0.040 \, \text{H}$$

Functional values and constants

$$F_1(p) = \frac{Z_r(p)}{[(Z_r(p))^2 + \omega_{sl}^2 L_r^2]}$$

$$F_2(p) = \frac{-\omega_{sl} L_r}{[(Z_r(p))^2 + \omega_{sl}^2 L_r^2]}$$

$$F_3(p) = p F_1(p) - \omega_{sl} F_2(p)$$

$$F_4(p) = \omega_{sl} F_1(p) + p F_2(p)$$

$$F_5(p) = Z_s(p) - p L_m^2 F_3(p) + L_m^2 F_4(p) \omega_e$$

$$F_6(p) = \omega_e L_s - \omega_e L_m^2 F_3(p) - p L_m^2 F_4(p)$$

$$F_7(p) = L_m [p F_2(p) + \omega_e F_1(p)]$$

$$F_8(p) = L_m [-p F_1(p) + \omega_e F_2(p)]$$

$$F_9(p) = -L_m F_3(p)$$

$$F_{10}(p) = -L_m F_4(p)$$

$$F_{11}(p) = L_m F_4(p)$$

$$F_{12}(p) = -L_m F_3(p)$$

$$F_{13}(p) = [F_2(p) \psi_{dr}^e + F_1(p) \psi_{qr}^e]$$

$$F_{14}(p) = [-F_1(p) \psi_{dr}^e + F_2(p) \psi_{qr}^e]$$

$$F_{15}(p) = \frac{3}{2} \cdot \frac{P}{2} \cdot L_m [i_{dr}^e + F_{11}(p) i_{qs}^e - F_{12}(p) i_{ds}^e]$$

$$F_{16}(p) = \frac{3}{2} \cdot \frac{P}{2} \cdot L_m [-i_{qr}^e + F_9(p) i_{qs}^e - F_{10}(p) i_{ds}^e]$$

$$F_{17}(p) = \frac{3}{2} \cdot \frac{P}{2} \cdot L_m [F_{13}(p) i_{qs}^e - F_{14}(p) i_{ds}^e]$$

$$F_{18}(p) = -C [F_{15}(p) \omega_e + p F_{16}(p)]$$

$$F_{19}(p) = -C [p F_{15}(p) - \omega_e F_{16}(p)]$$

$$F_{20}(p) = -C [F_{15}(p) v_{ds}^e - F_{16}(p) v_{qs}^e]$$

$$F_{21}(p) = F_{17}(p) + J \frac{2}{P} \cdot p$$

$$F_{22}(p) = [-\psi_{qs}^e + C F_5(p) v_{qs}^e + C F_6(p) v_{ds}^e + F_7(p) \psi_{dr}^e - F_8(p) \psi_{qr}^e]$$

$$F_{23}(p) = [\psi_{ds}^e - C F_5(p) v_{ds}^e + C F_6(p) v_{qs}^e + F_8(p) \psi_{dr}^e + F_7(p) \psi_{qr}^e]$$

$$F_{24}(p) = 1 + C [p F_5(p) - \omega_e F_6(p)]$$

$$F_{25}(p) = C [\omega_e F_5(p) + p F_6(p)]$$

$$A_1 = \frac{I_{act}^* \times k_1 \times 2}{k^2 I_{ref}}$$

$$A_2 = \frac{-(I_{react}^* - I_c) \times k_{12} \times 2}{k^2 I_{ref}}$$

$$a_1 = [F_{24}(p) F_5(p) + F_6(p) F_{25}(p)]$$

$$a_2 = R_f + p L_f$$

$$a_3 = [(F_{24}(p))^2 + (F_{25}(p))^2]$$

$$a_4 = p a_3$$

$$a_5 = p(a_2 a_3 + 1.5 k^2 a_1)$$

$$c_1 = F_{18}(p)$$

$$c_2 = F_{19}(p)$$

$$c_3 = -F_{26}(p)$$

$$c_4 = -F_{27}(p)$$

$$c_5 = k F_{15}(p)$$

$$c_6 = F_{24}(p)$$

$$c_7 = F_{25}(p)$$

$$c_8 = F_{22}(p)$$

$$c_9 = F_7(p) \psi_{dr}^e - F_8(p) \psi_{qr}^e$$

$$c_{10} = k F_6(p)$$

$$c_{11} = F_{25}(p)$$

$$c_{12} = F_{24}(p)$$

$$c_{13} = F_{23}(p)$$

$$c_{14} = F_8(p) \psi_{dr}^e + F_7(p) \psi_{qr}^e$$

$$c_{15} = -k F_5(p)$$

$$c_{16} = 0$$

$$c_{17} = -1.5k$$

$$c_{18} = -\left(k_{ps} + \frac{k_{is}}{p}\right) (A_1 + A_2)$$

$$c_{19} = \left(k_{ps} + \frac{k_{is}}{p}\right) A_2$$

$$c_{20} = -\left(k_{ps} + \frac{k_{is}}{p} + R_f + p L_f\right)$$

$$c_{21} = c_8 - c_9$$

$$c_{22} = c_{13} - c_{14}$$

$$c_{23} = c_3 (c_7 (c_{11} c_{20} - c_{15} c_{16}) - c_6 (c_{12} c_{20} - c_{15} c_{17}) + c_{10} (c_{12} c_{16} - c_{11} c_{17}))$$

$$c_{24} = c_4 (c_7 (c_{11} c_{22} - c_{15} c_{16}) - c_6 (c_{12} c_{20} - c_{15} c_{17}) + c_{10} (c_{12} c_{16} - c_{11} c_{17}))$$

$$c_{25} = -c_2 (c_{21} (c_{11} c_{20} - c_{15} c_{16}) - c_6 (c_{12} c_{20} - c_{15} c_{19}) + c_{10} (c_{12} c_{16} - c_{11} c_{19}))$$

$$c_{26} = -c_2 (c_9 (c_{11} c_{20} - c_{15} c_{16}) - c_6 (c_{14} c_{20} - c_{15} c_{18}) + c_{10} (c_{14} c_{16} - c_{11} c_{18}))$$

$$c_{27} = c_1 (c_{21} (c_{12} c_{20} - c_{15} c_{17}) - c_7 (c_{22} c_{20} - c_{15} c_{19}) + c_{10} (c_{22} c_{17} - c_{12} c_{19}))$$

$$c_{28} = c_1 (c_9 (c_{12} c_{20} - c_{15} c_{17}) - c_7 (c_{14} c_{20} - c_{15} c_{18}) + c_{10} (c_{14} c_{17} - c_{12} c_{18}))$$

$$c_{29} = -c_5 (c_{21} (c_{12} c_{16} - c_{11} c_{17}) - c_7 (c_{22} c_{16} - c_{11} c_{19}) + c_{10} (c_{22} c_{17} - c_{12} c_{19}))$$

$$c_{30} = -c_3 (c_9 (c_{12} c_{16} - c_{11} c_{17}) - c_7 (c_{14} c_{16} - c_{11} c_{18}) + c_{10} (c_{14} c_{17} - c_{12} c_{18}))$$

$$c_{31} = c_{23} + c_{26} + c_{28} + c_{30}$$

$$c_{32} = c_{24} + c_{25} + c_{27} + c_{29}$$

$$c_{33} = p c_{31}$$

$$c_{34} = p c_{32}$$

References

- [1] P. Agarwal and V. K. Verma, "Performance Evaluation of Current Source Inverter- Fed Induction Motor Drive", Journals of Institution of Engineers, Vol.72, pp.209-217, February 1992.
- [2] P. N. Enjeti, P. D. Ziogas and J. F. Lindsay, "Pro-

- grammed PWM technique to eliminate harmonics: A critical evaluation", IEEE Transactions on Industry Applications, Vol. 26, No. 2, pp. 302-315, March/April 1990.
- [3] S. R. Bowes and R. I. Bullough, "Optimal PWM Microprocessor Controlled Current-Source Inverter Drives", IEEE proceedings, Vol. 135, Pt. B, No. 2, pp. 59-75, March 1988.
- [4] Y. Xiao, B. Wu, S. Rizzo and R. Sotuden, "A Novel Power Factor Control Scheme for High Power GTO Current Source Converter", Conference Record IEEE-IAS, pp.865-869, 1996.
- [5] Pramod Agarwal, V. K. Verma and A. K. Pandey, "Performance Evaluation of a Self-Commutating CSI-Fed Induction Motor Drive for Different Operating Conditions, IETE Journal of Research, Vol. 54, Issue 4, July/Aug. 2008.
- [6] P. Agarwal and V. K. Verma, "Parameter Coordination of Microcomputer Controlled CSI Fed Induction Motor Drive", IE (I) Journal – EL, Vol. 88, December 2007.
- [7] A. K. Pandey, Pramod Agarwal and V. K. Verma, "Optimal Capacitor Selection for Modified Self Commutated CSI-fed Induction Motor Drive", IEEE ISIE 2006, Montreal, Quebec, Canada, pp. 1166-1171, July 9-12, 2006.
- [8] Pramod Agarwal, A. K. Pandey and V. K. Verma, "Performance Investigation of Modified Self-Commutated CSI-fed Induction Motor Drive, Asian Power Electronics Journal, Vol. 3, No. 1, pp. 21-29, Sept 2009.
- [9] J. Nanda, "Analysis of Steady State Stability of a Two Machine System by the D-Composition Technique", IEEE Transactions of Power Apparatus and Systems, Vol. 90, pp. 1848-1855, July / Aug. 1971.

- [10] M. A. Aizerman, "Theory of Automatic Control", Chapter -3, Pergamon Press, New York, 1963.

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