Robust Decoupling Control of Induction Motors by Exact Feedback Linearization

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Abstract. This paper presents an induction motor control system based on the exact feedback linearization approach. The underlying model used for control design is the fifth-order stator-flux model. Linearizing and decoupling transformation and control law are derived. Effects of parameter variations on the control system structure are derived in the frame of this approach. It is shown that the proposed control law does not lose its linearizing and decoupling properties, though some additional feedback connections enter the system structure. PID and PI controllers are designed in the outer loops taking into account a second degree of freedom, introduced by specified prefilters in both loops. Speed control simulation results are presented, confirming the feasibility of the overall control system.

1. Introduction

Due to its reliability, and relatively low cost the induction motor is probably the most widely used electric machine and the choice of many industrial applications. The following reasons, however, present a serious challenge when designing induction motor control algorithms:

- the dynamic behavior of the motor is described by a fifth-order highly coupled and nonlinear dynamical system;
- rotor electric variables i.e. rotor fluxes and currents as well as stator and air-gap fluxes are practically not measured;
- most of its physical parameters may vary significantly while operating the motor – stator and mainly rotor resistance, due to heating, magnetizing induction due to saturation, total moment of inertia of the rotor and the viscous friction coefficient are not easily estimated;
- presence of unknown load torque.

The most renown solution dealing with the complicated motor dynamics is given by the so-called field-oriented control [1,2]. It is based on rewriting motor equations through a nonlinear transformation in a specific coordinate frame, where rotor flux and speed are asymptotically decoupled, i.e. the model looks like that of a separately excited dc motor. The main drawback of this technique is that the decoupling is valid only after the rotor flux is constant. The dynamics of the currents also remain nonlinear.

The control algorithm investigated here is based on input-output feedback linearizing techniques. The advantage of the input-output linearization over the field-oriented control is the fact that, by applying the linearizing transformation, a complete decoupling of the rotor speed and flux is achieved, which enables the optimization of motor performance without degrading the mechanical output regulation.

Control designs, for the induction motor case, based on this approach are found in [2, 6-8, 14-15], all using the rotor-flux model. In [13] input-output linearization based control in current-fed mode, using the stator-flux model is presented.

In this paper, an input-output linearization based control in full (voltage command) mode, using the fifth-order stator-flux induction motor model is presented. The respective nonlinear transformation and control law are derived. The influence of variations in the main motor parameters in the frame of this approach are derived and discussed. A speed control scheme based on the linearized model, realized in a two-degree of freedom frame for both, speed and flux, subsystems is proposed. The outer control loops are realized by output feedback using high-gain PID and PI controllers in speed and flux subsystems respectively. Prefilters, specifying desired time-domain behavior, are added. Simulation results are presented.

2. Dynamic Modeling of the Induction Motor

The induction motor considered here is a three-phase stator, three-phase short-circuited rotor machine. The following considerations are valid for the case of a squirrel-cage rotor, since it is equivalent to a three-phase short-circuited one through a simple transformation. The common assumptions are adopted for the modeling i.e. symmetrical construction, sinusoidal distribution of the field in the air-gap and magnetic circuits' linearity.

Remark. The rotor flux magnitude can be kept away from the saturation zone by an appropriate control action, thus forcing the assumption for linear magnetic circuits.

Writing the equations describing the motor dynamic behavior in the two-phase stator-fixed б-в frame and eliminating rotor fluxes and currents, the following equivalent two-phase stator-flux model is obtained:

$$\begin{aligned}
&\dot{\omega} = n_p J^{-1}(\psi_{S\alpha} i_{S\beta} - \psi_{S\beta} i_{S\alpha}) - cJ^{-1}\omega - J^{-1}\tau_L;\\
&\dot{\psi}_{S\alpha} = -r_S i_{S\alpha} + u_{S\alpha};\\
&\dot{\psi}_{S\beta} = -r_S i_{S\beta} + u_{S\beta};\\
&\dot{i}_{S\alpha} = -\gamma i_{S\alpha} - n_p \omega i_{S\beta} + \zeta\psi_{S\alpha} + n_p(\sigma l_S)^{-1}\omega\psi_{S\beta} + (\sigma l_S)^{-1}u_{S\alpha};\\
&\dot{i}_{S\beta} = -\gamma i_{S\beta} + n_p \omega i_{S\alpha} + \zeta\psi_{S\beta} - n_p(\sigma l_S)^{-1}\omega\psi_{S\alpha} + (\sigma l_S)^{-1}u_{S\beta};\\
&\dot{\theta} = \omega
\end{aligned} \tag{1}$$

where: $i_{S\alpha}$, $i_{S\beta}$ – stator currents, $\psi_{S\alpha}$, $\psi_{S\beta}$ – stator fluxes ω – rotor speed, θ – rotor position $u_{S\alpha}$, $u_{S\beta}$ – voltage inputs to the motor, τ_L –load torque, $l_{S(R)}$ –phase stator (rotor) winding inductances, $r_{S(R)}$ –phase stator (rotor) winding resistances, $m_0 = 2/3m$ – mutual inductance, n_p –number of pole-pairs, J –rotor moment of inertia; c –viscous friction coefficient, $\sigma = (l_R l_S - m^2)(l_R l_S)^{-1}$, $\gamma = (l_R r_S + l_S r_R)(\sigma l_R l_S)^{-1}$, $\zeta = r_R(\sigma l_R l_S)^{-1}$

The complete derivation of the model can be found in [1,2,3].

Though the paper deals with speed control applications, a position coordinate is added in the model and the following mathematical analysis is carried out on the sixth-order model obtained for generalization purposes.

The induction motor model (1) is put in the following form:

$$\dot{x} = f(x) + g_1 u_1 + g_2 u_2 + f_\tau \text{, with} \tag{2}$$

$$x = [x_1\ x_2\ x_3\ x_4\ x_5\ x_6]^T = [\omega\ \psi_{S\alpha}\ \psi_{S\beta}\ i_{S\alpha}\ i_{S\beta}\ \theta]^T$$

$$u_1 = u_{S\alpha}\text{, } u_2 = u_{S\beta}\text{.}$$

$$f(x) = \begin{bmatrix} n_p J^{-1}(x_2 x_5 - x_3 x_4) - cJ^{-1} x_1 \\ -r_S x_4 \\ -r_S x_5 \\ -\gamma x_4 - n_p x_1 x_5 + \zeta x_2 + n_p (\sigma l_S)^{-1} x_1 x_3 \\ -\gamma x_5 + n_p x_1 x_4 + \zeta x_3 - n_p (\sigma l_S)^{-1} x_1 x_2 \\ x_1 \end{bmatrix}$$

$$g_1 = \begin{bmatrix} 0 & 1 & 0 & (\sigma l_S)^{-1} & 0 & 0 \end{bmatrix}^T$$

$$g_2 = \begin{bmatrix} 0 & 0 & 1 & 0 & (\sigma l_S)^{-1} & 0 \end{bmatrix}^T$$

$$f_\tau = \begin{bmatrix} -J^{-1}\tau_L & 0 & 0 & 0 & 0 & 0 \end{bmatrix}^T$$

3.Input-output Linearization of the Stator-flux Induction Motor Model

Basically, feedback linearization consists of applying a nonlinear transformation on system variables i.e. expressing them in a new „suitable" coordinate system, which will enable the introduction of a nonlinearities canceling feedback, so that an input-output or state linearization in the new coordinates is achieved. Theoretical foundations and systematic procedures for finding these can be found in [4,5,12].

For the induction motor case, by choosing the output functions as the rotor position and stator flux square respectively

$$y_1 = h_1(x) = x_6\text{, } y_2 = h_2(x) = x_2^2 + x_3^2 \tag{3}$$

and applying the following coordinate transformation:

$$\begin{aligned} {}^1\xi_1(x) &= h_1(x) = x_6; \\ {}^1\xi_2(x) &= L_f h_1(x) = x_1; \\ {}^1\xi_3(x) &= L_f{}^2 h_1(x) = n_p J^{-1}(x_2 x_5 - x_3 x_4) - cJ^{-1} x_1; \\ {}^2\xi_1(x) &= h_2(x) = x_2^2 + x_3^2; \\ \chi_1(x) &= x_2 - \sigma l_S x_4; \\ \chi_2(x) &= x_3 - \sigma l_S x_5 \end{aligned} \tag{4}$$

where $L_f h$ denotes the Lie derivative of the scalar function h with respect to (or along) the vector field f and represents a scalar function defined by $L_f h = \frac{\partial h}{\partial x} f$ (iteratively $L_f{}^n h = L_f L_f{}^{n-1} h$),

the system (2) is transformed in the following normal form:

$$\begin{aligned} {}^1\dot{\xi}_1 &= L_f h_1; \\ {}^1\dot{\xi}_2 &= L_f{}^2 h_1 + L_{f_\tau} L_f h_1; \\ {}^1\dot{\xi}_3 &= L_f{}^3 h_1 + L_{g_1} L_f{}^2 h_1 . u_1 + L_{g_2} L_f{}^2 h_1 . u_2 + L_{f_\tau} L_f{}^2 h_1; \\ {}^2\dot{\xi}_1 &= L_f h_2 + L_{g_1} h_2 . u_1 + L_{g_2} h_2 . u_2 \end{aligned} \tag{5}$$

$$\begin{aligned} \dot{\chi}_1 &= L_f \chi_1; \\ \dot{\chi}_2 &= L_f \chi_2. \end{aligned} \tag{6}$$

For notation simplification purposes (x) is omitted in the expressions that follow in the paper. The corresponding Lie derivatives are given by:

$$\begin{aligned} L_f{}^3 h_1 &= -n_p J^{-1}(cJ^{-1} + \gamma)(x_2 x_5 - x_3 x_4) + c^2 J^{-2} x_1 + \\ &+ n_p^2 J^{-1} x_1 (x_2 x_4 + x_3 x_5) - n_p^2 J^{-1} (\sigma l_S)^{-1} x_1 (x_2^2 + x_3^2); \\ L_{f_\tau} L_f h_1 &= -J^{-1}\tau_L; \\ L_{f_\tau} L_f{}^2 h_1 &= cJ^{-2}\tau_L; \\ L_f h_2 &= -2r_S(x_2 x_4 + x_3 x_5); \\ L_{g_1} L_f{}^2 h_1 &= n_p J^{-1}(x_5 - (\sigma l_S)^{-1} x_3); \\ L_{g_2} L_f{}^2 h_1 &= -n_p J^{-1}(x_4 - (\sigma l_S)^{-1} x_2); \\ L_{g_1} h_2 &= 2x_2\text{, } L_{g_2} h_2 = 2x_3; \end{aligned}$$

$$\begin{aligned} L_f \chi_1 &= \sigma l_S(\gamma x_4 + n_p x_1 x_5 - \zeta x_2 - n_p (\sigma l_S)^{-1} x_1 x_3) - r_S x_4; \\ L_f \chi_2 &= \sigma l_S(\gamma x_5 - n_p x_1 x_4 - \zeta x_3 + n_p (\sigma l_S)^{-1} x_1 x_2) - r_S x_5 \end{aligned}$$

Choosing the linearizing control law in the form:

$$\begin{bmatrix} u_1 \\ u_2 \end{bmatrix} = A(x)^{-1} \begin{bmatrix} v_1 - L_f{}^3 h_1 \\ v_2 - L_f h_2 \end{bmatrix} \tag{7}$$

the linearized system is put in the following form:

$$
\begin{aligned}
&{}^1\dot{\xi}_1 = {}^1\xi_2; \\
&{}^1\dot{\xi}_2 = {}^1\xi_3 - J^{-1}\tau_L; \\
(8)\quad &{}^1\dot{\xi}_3 = v_1 + cJ^{-2}\tau_L; \\
&{}^2\dot{\xi}_1 = v_2;
\end{aligned}
$$

$$
(9)\quad \begin{aligned}
\dot{\chi}_1 &= L_f \chi_1; \\
\dot{\chi}_2 &= L_f \chi_2.
\end{aligned}
$$

The matrix $A(x)$, called „decoupling" is given by:

$$
A(x) = \begin{bmatrix} L_{g_1} L_f{}^2 h_1 & L_{g_2} L_f{}^2 h_1 \\ L_{g_1} h_2 & L_{g_2} h_2 \end{bmatrix} =
$$

$$
(10)\quad = \begin{bmatrix} n_p J^{-1}(x_5 - (\sigma l_S)^{-1} x_3) & -n_p J^{-1}(x_4 - (\sigma l_S)^{-1} x_2) \\ 2x_2 & 2x_3 \end{bmatrix}
$$

The decoupling matrix invertibility condition is given by:

$$
\det A(x) = -2n_p J^{-1}(\sigma l_S)^{-1}(x_2^2 + x_3^2) + 2n_p J^{-1}(x_2 x_4 + x_3 x_5) \neq 0
$$

which is satisfied during normal operaton of the motor, though it can be monitored and kept away from zero by setting appropriate reference values of the stator flux magnitude.

By choosing the control law as (7), the dynamics of the original nonlinear system are decomposed into two parts: a linear input-output map, given by (8) and a nonlinear, unobservable through the outputs, internal part (9) described in χ_1 and χ_2 variables. These are chosen in the form (4) in order to obtain $L_{g_i}\chi_j = 0,\ i = 1,2;\ j = 1,2$ thus rendering them independent from the inputs and on the other hand ensuring the validity of the transformation (4), [4,5]. This choice is possible since g_1 and g_2 are constant vector fields, thus forming an involutive set.

The stability properties of these internal dynamics represent a general limitation of feedback linearization control. According to their definition χ_1 and χ_2 can be considered as normalized rotor flux values, thus their boundedness is guaranteed by the control action.

In *figure 1 it* is shown the resulting linear decoupled input-output system. It is seen, that the problem of controlling mechanical output is rendered to controlling a triple integrator (for position) or double integrator (for speed) and a single integrator for the flux loop. As seen both subsystems are exactly decoupled.

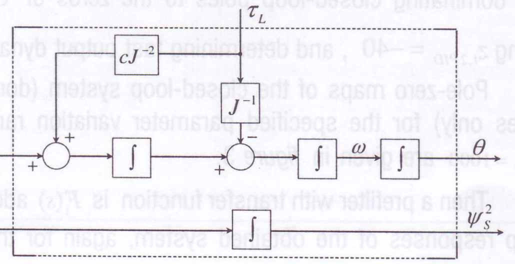

Figure 1. Input-output linearized system

Effects of Variations in Motor Parameter Values

To study the effects of variations in motor parameter values, let assume uncertainties on these parameters, formalized in the following form:

$$
\overline{J} = J^{-1},\ \overline{J}_p = k\overline{J},\ c_p = c + \triangle c,\ r_{Sp} = r_S + \triangle r_S,\ r_{Rp} = r_R + \triangle r_R
$$

System (2) can be rewritten as:

$$
(11)\quad \dot{x} = f(x) + g_1 u_1 + g_2 u_2 + f_S(x) + f_R(x) + f_m(x) + f_\tau
$$

with:

$$
f_R(x) = \triangle r_R \begin{bmatrix} 0 \\ 0 \\ 0 \\ (\sigma l_R l_S)^{-1} x_2 - (\sigma l_R)^{-1} x_4 \\ (\sigma l_R l_S)^{-1} x_3 - (\sigma l_R)^{-1} x_5 \\ 0 \end{bmatrix} \qquad f_S(x) = \triangle r_S \begin{bmatrix} 0 \\ -x_4 \\ -x_5 \\ -(\sigma l_S)^{-1} x_4 \\ -(\sigma l_S)^{-1} x_5 \\ 0 \end{bmatrix}
$$

$$
f_m(x) = \left[(k-1)(n_p \overline{J}(x_2 x_5 - x_3 x_4) - c\overline{J}x_1) - \triangle ck\overline{J}x_1 \quad 0 \quad 0 \quad 0 \quad 0 \quad 0 \right]^T
$$

$$
f_\tau = \left[-k\overline{J}\tau_L \quad 0 \quad 0 \quad 0 \quad 0 \quad 0 \right]^T.
$$

With variables defined as in (4), expressions (5) and (6) take the following form.

$$
\begin{aligned}
&{}^1\dot{\xi}_1 = L_f h_1; \\
(12)\quad &{}^1\dot{\xi}_2 = L_f{}^2 h_1 + L_{f_m} L_f h_1 + L_{f_\tau} L_f h_1; \\
&{}^1\dot{\xi}_3 = L_f{}^3 h_1 + L_{f_R} L_f{}^2 h_1 + L_{f_S} L_f{}^2 h_1 + L_{f_m} L_f{}^2 h_1 + \\
&\qquad + L_{f_\tau} L_f{}^2 h_1 + L_{g_1} L_f{}^2 h_1 . u_1 + L_{g_2} L_f{}^2 h_1 . u_2; \\
&{}^2\dot{\xi}_1 = L_f h_2 + L_{f_S} h_2 + L_{g_1} h_2 . u_1 + L_{g_2} h_2 . u_2;
\end{aligned}
$$

$$
(13)\quad \begin{aligned}
\dot{\chi}_1 &= L_f \chi_1 + L_{f_R} \chi_1; \\
\dot{\chi}_2 &= L_f \chi_2 + L_{f_R} \chi_2
\end{aligned}
$$

where all Lie derivatives that are equal to zero are omitted and the present ones are given by the following expressions:

$$
L_{f_m} L_f h_1 = (k-1)(n_p \overline{J}(x_2 x_5 - x_3 x_4) - c\overline{J}x_1) - \triangle ck\overline{J}x_1 =
$$

$$
= (k-1)\,{}^1\xi_3 - \triangle ck\overline{J}\,{}^1\xi_2, \qquad L_{f_\tau} L_f h_1 = -k\overline{J}\tau_L;
$$

$$
L_{f_R} L_f{}^2 h_1 = \triangle r_R n_p \overline{J}(\sigma l_R)^{-1}(x_3 x_4 - x_2 x_5) = -\triangle r_R (\sigma l_R)^{-1}({}^1\xi_3 + c\overline{J}\,{}^1\xi_2);
$$

$$
L_{f_S} L_f{}^2 h_1 = \triangle r_S n_p \overline{J}(\sigma l_S)^{-1}(x_3 x_4 - x_2 x_5) = -\triangle r_S (\sigma l_S)^{-1}({}^1\xi_3 + c\overline{J}\,{}^1\xi_2);
$$

$$
L_{f_m} L_f{}^2 h_1 = -c\overline{J}(k-1)(n_p \overline{J}(x_2 x_5 - x_3 x_4) - c\overline{J}x_1) + \triangle cck\overline{J}^2 x_1 =
$$

$$
= -c\overline{J}(k-1)\,{}^1\xi_3 + \triangle cck\overline{J}^2\,{}^1\xi_2, \qquad L_{f_\tau} L_f{}^2 h_1 = ck\overline{J}^2\tau_L;
$$

$$
L_{f_S} h_2 = -2\triangle r_S (x_2 x_4 + x_3 x_5) = \triangle r_S / r_S . L_f h_2
$$

$$
L_{f_R} \chi_1 = \triangle r_R l_S l_R^{-1} x_4 - \triangle r_R l_R^{-1} x_2,\ L_{f_R} \chi_2 = \triangle r_R l_S l_R^{-1} x_5 - \triangle r_R l_R^{-1} x_3
$$

Applying the same linearizing control law (7), the input-output system is put in the following form:

$$
\begin{aligned}
&{}^{1}\dot{\xi}_1 = {}^{1}\xi_2; \\
&{}^{1}\dot{\xi}_2 = k\,{}^{1}\xi_3 - k_3\,{}^{1}\xi_2 - k\bar{J}\tau_L; \\
&{}^{1}\dot{\xi}_3 = k_1\,{}^{1}\xi_3 - k_2\,{}^{1}\xi_2 + v_1 + ck\bar{J}^2\tau_L; \\
&{}^{2}\dot{\xi}_1 = k_4 L_f h_2 + v_2
\end{aligned}
\tag{14}
$$

where:

$$
\begin{aligned}
&k_1 = \Delta r_R(\sigma l_R)^{-1} + \Delta r_S(\sigma l_S)^{-1} + c\bar{J}(k-1); \\
&k_2 = c\bar{J}(\Delta r_R(\sigma l_R)^{-1} + \Delta r_S(\sigma l_S)^{-1} + \Delta ck\bar{J}); \\
&k_3 = \Delta ck\bar{J},\quad k_4 = \Delta r_S / r_S.
\end{aligned}
\tag{15}
$$

Figure 2 visualizes the structure of the resulting input-output system in presence of uncertainties.

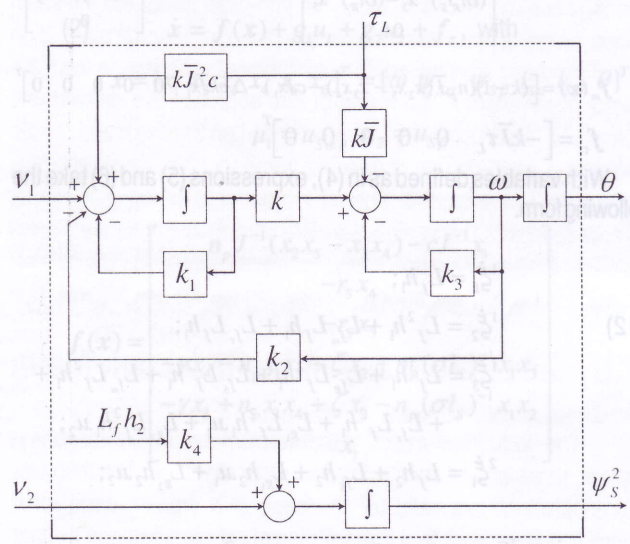

Figure 2. Perturbed input-output system

The following conclusions can be made as a result of the carried out mathematical analysis:

- the mechanical subsystem remains not affected by the flux one;
- it preserves its linear properties, though some additional feedback connections (either positive or negative depending on the direction of parameter values variations) and disturbances appear in its structure;
- the integration properties of the mechanical subsystem, except for the inherent from its physical structure one, are lost;
- flux subsystem also remains linear;
- stator resistance variation induces a state dependent disturbance signal in the flux subsystem, which restores the coupling with the mechanical subsystem.

Following these considerations it can be concluded that the linearizing and the more important decoupling property of the designed control law are not affected by parameter variations i.e. these properties of the control system are rendered robust to parameter variations.

4.Speed Control Based on Input-output Linearization

In this section, a speed control scheme for the input-output feedback linearized induction motor is proposed. Based on the carried out uncertainty analysis, the following transfer function describes the speed subsystem of the linearized motor in the general case:

$$
\frac{\omega(s)}{v_1(s)} = \frac{k}{s^2 + (k_1 + k_3)s + k_1 k_3 + kk_2}.
\tag{16}
$$

The motor parameters used for simulation purposes are chosen as:
$r_S = 20{,}13\Omega$, $r_R = 13\Omega$, $l_S = 1.05H$, $l_R = 1.33H$, $m = 0.957H$, $J = 0{,}0005Nms^2$, $c = 0{,}00014Nms$, $n_p = 2$.

Assuming parametric variations in the following ranges $\Delta r_R = (-0{,}2 \div 0{,}5)r_R$, $\Delta c = (0 \div 1)c$, $J_p = (1 \div 2)J / k = (1 \div 0{,}5)/$ the resulting transfer function is specified by:

$$
\frac{\omega(s)}{v_1(s)} = \frac{(0{,}5 \div 1)}{s^2 + (-5{,}82 \div 14{,}48)s + (-3.1 \div 8)}.
$$

Though the coefficients of the characteristic polynomial are not independent, as seen from (15) and (16), quite different dynamic behavior is still possible in presence of uncertainties, ranging from that of stable to that of unstable second-order transfer functions.

The desired dynamic behavior of the speed response is specified by the following transfer function:

$$
F_1(s) = \frac{1600}{s^2 + 80s + 1600}
$$

The above given design problem is approached by a two-degree of freedom control system. The speed control loop is realized with high-gain PID-controller with transfer function given by:

$$
G_{PID}(s) = k_{PID}\frac{s^2 + 80s + 1600}{s(0{,}001s + 1)},
$$

with the following goals:

- ensuring zero steady-state error;
- reducing the effect of parameter variations by moving the dominating closed-loop poles to the zeros of $G_{PID}(s)$, being $z_{1,2PID} = -40$, and determining fast output dynamics.

Pole-zero maps of the closed-loop system (dominating poles only) for the specified parameter variation range and $k_{PID} = 1000$ are given in *figure* 3.

Then a prefilter with transfer function is $F_1(s)$ added. The step responses of the obtained system, again for the entire

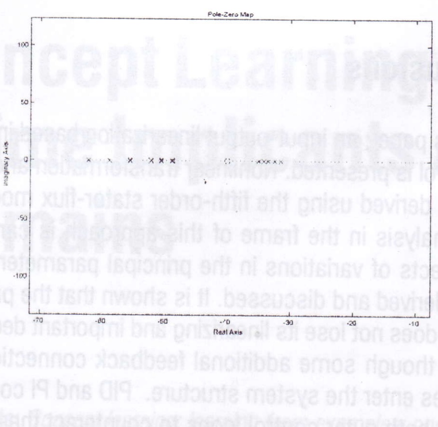

Figure 3. Pole-zero maps

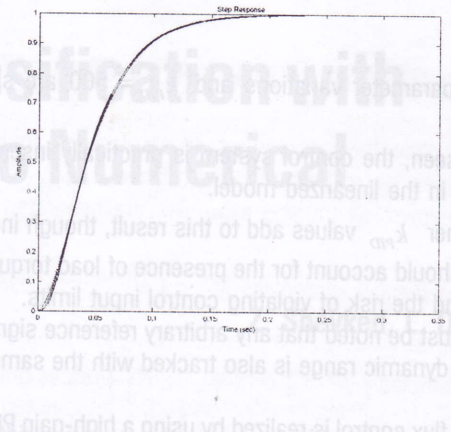

Figure 4. Step responses

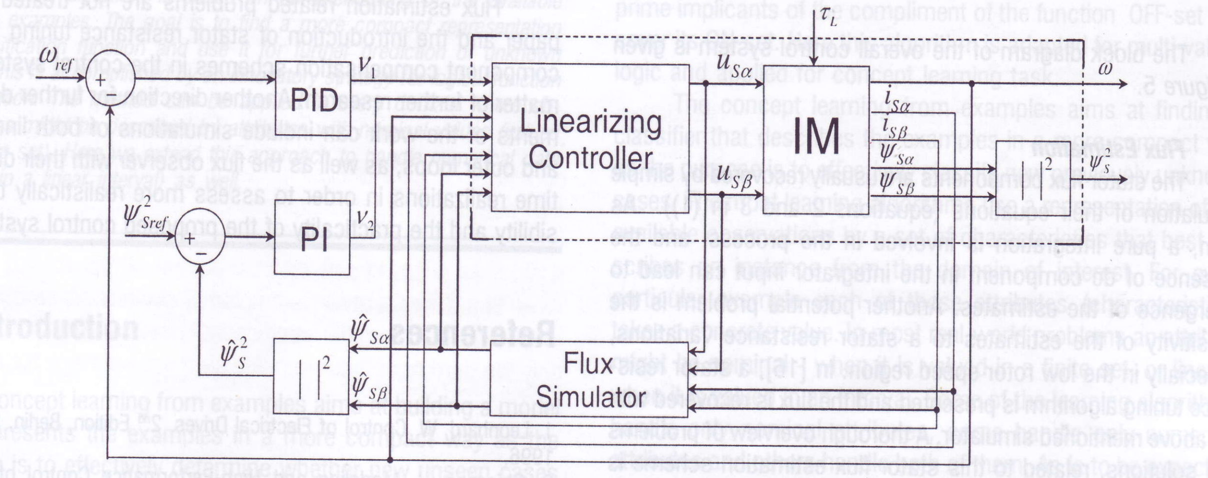

Figure 5. Overall control system

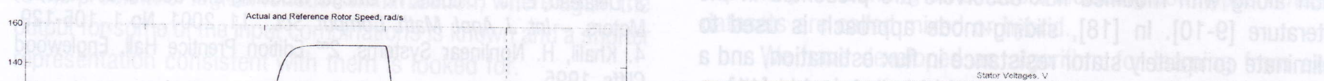

Figure 6. Transient responses(time in seconds on the x-axis)

range of parameter variations and k_{PID} = 500 are shown in *figure 4*.

As seen, the control system is practically insensitive to variations in the linearized model.

Higher k_{PID} values add to this result, though increasing its value should account for the presence of load torque disturbances and the risk of violating control input limits.

It must be noted that any arbitrary reference signal in the prefilter's dynamic range is also tracked with the same performance.

The flux control is realized by using a high-gain PI controller and a prefilter designed with the same idea, specified by:

$$G_{PI}(s) = k_{PI}\frac{s+40}{s},\ F_2(s) = \frac{40}{s+40}.$$

The block diagram of the overall control system is given in *figure* 5.

Flux Estimation

The stator flux components are usually recovered by simple simulation of their equations (equations 2 and 3 in (1)). As seen, a pure integration is involved in the process, and the presence of dc component in the integrator input can lead to divergence of the estimates. Another potential problem is the sensitivity of the estimates to a stator resistance variations, especially in the low rotor speed region. In [16], a stator resistance tuning algorithm is presented and the flux is recovered with the above mentioned simulator. A thorough overview of problems and solutions, related to this stator flux estimation scheme is given in [17]. Different algorithms for stator resistance estimation along with modified flux observers are presented in the literature [9-10]. In [18], sliding-mode approach is used to eliminate completely stator resistance in flux estimation, and a method for motor parameters identification is found in [11].

These problems are not in the scope of this paper and it is assumed that flux components are estimated accurately, i.e stator resistance value is known.

Simulation Results

Figure 6, visualizes some transient responses of the designed control system. In this particular simulation, parameter deviations and controller gains are set to $\Delta r_R = 0{,}5r_R$, $J_p = 1{,}5J$ and $k_{PID} = k_{PI} = 500$ respectively. At $t = 0{,}4s$. a $2Nm$ load torque, unknown to the controller is applied. According to the proposed control design ω_{ref} and ψ^2_{Sref}, seen in *figure* 5, are generated by the following scheme $\omega_{ref}(s) = F_1(s)R_1(s)$, $\psi^2_{Sref}(s) = F_2(s)R_2(s)$, where $R_1(s)$ and $R_2(s)$ represent series of step functions, given by the respective reference values.

As seen the outputs are able to track without significant error the desired references and no coupling is present.

5. Conclusions

In this paper, an input-output linearization based induction motor control is presented. Nonlinear transformation and a control law are derived using the fifth-order stator-flux model. Uncertainty analysis in the frame of this approach is carried out and the effects of variations in the principal parameters of the motor are derived and discussed. It is shown that the proposed control law does not lose its linearizing and important decoupling properties, though some additional feedback connections and disturbances enter the system structure. PID and PI controllers are used in both outer control loops to counteract these structural changes. Both control systems are given second degree of freedom by a respective prefilter specifying desired output dynamics. Simulation results are presented.

Flux estimation related problems are not treated in the paper and the introduction of stator resistance tuning and dc component compensation schemes in the control system is a matter of further research. Another direction for further developments of the work can include simulations of both linearizing and outer loops, as well as the flux observer with their discrete-time realizations in order to assess more realistically the feasibility and the practicality of the proposed control system.

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