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Nonlinear Controllers for Saturated AC Motors

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Abstract— **Two passivity-based controllers and an input-output linearization controller for current-fed saturated induction motors are presented. Advantages and disadvantages of the controllers are discussed. Finally, experimental results obtained with one of the passivity-based controllers are presented.**

I. INTRODUCTION

In recent years, passivity-based [1, 2] and input-output linearization methods [3, 4, 5] have proven to be effective for nonlinear control of induction motors.

Following the work in [1, 2], Gökdere *et al.* [6] designed, and also experimentally implemented, a current-command passivity-based controller that forces an induction motor to track time-varying speed/position and flux trajectories (The passivity-based controller in [6] is equivalent to the current-fed induction motor version of the passivity-based controller in [1, 2]). Gökdere *et al.* [7] have also incorporated magnetic saturation effects into the current-command passivity-based control of induction motors. Implementation of a current-command input-output linearization controller based on the saturated magnetic model of the induction motor can be found in [4, 5].

In this paper, a brief comparative study of two passivity-based controllers and an input-output linearization controller based on the saturated magnetic model of current-fed induction motors is presented. An interesting comparative study can also be found in [8].

II. CURRENT-COMMAND PASSIVITY-BASED CONTROLLERS FOR SATURATED INDUCTION MOTORS

In one type of passivity-based control of induction motors, $\mathbf{i}_{dq}^*$ (the commanded value of the stator current vector) and f (the angular position of the reference frame with respect to the fixed stator reference frame) are considered as control inputs. Specifically, $\mathbf{i}_{dq}^*$ is given by

$$\mathbf{i}_{dq}^* = \begin{bmatrix} f_m^{-1}(l_d^*) + \dfrac{L_r}{R_r M}\dfrac{d l_d^*}{dt} \\ \dfrac{n_{ph}L_r}{2n_p M}\dfrac{t^*}{l_d^*} \end{bmatrix} \tag{1}$$

and f is chosen as the solution of

$$\frac{df}{dt} = \frac{n_{ph}R_r}{2n_p}\frac{t^*}{l_d^{*2}} + n_p w \tag{2}$$

where $f_m^{-1}(\cdot)$ is the inverse of the magnetization curve function of the induction motor, l_d^* is the magnitude of reference rotor flux vector and t^* is the reference torque.

Passivity-based controller (1)-(2) is a saturated version of the passivity-based controller in [6]. A block diagram for passivity-based controller (1)-(2) is given in Fig. 1.

Passivity-based controller (1)-(2) can be redefined in the rotor reference frame as [7]

$$\mathbf{i}_{rdq}^* = \begin{bmatrix} \cos a & -\sin a \\ \sin a & \cos a \end{bmatrix} \begin{bmatrix} f_m^{-1}(b) + \dfrac{L_r}{R_r M}\dfrac{db}{dt} \\ \dfrac{n_{ph}L_r}{2n_p M}\dfrac{t^*}{b} \end{bmatrix} \tag{3}$$

where a is the solution of

Fig.1. The block diagram for the current-command passivity-based control of saturated induction motors.

$$\frac{da}{dt} = \frac{n_{ph}R_r}{2n_p}\frac{t^*}{b^2} \tag{4}$$

In (3) and (4), b is the magnitude of the reference rotor flux vector. A block diagram for passivity-based controller (3)-(4) is given in [7].

Current-command passivity-based controllers (1)-(2) and (3)-(4) are exactly equivalent to each other. However, the latter requires more computational effort since $\cos a$ and $\sin a$ have to be calculated in addition to the *d-q* (direct-quadrature) coordinate transformation (It is assumed that the current-command controllers are implemented in the corresponding reference frames).

Finally, current-command passivity-based controller (1)-(2) is an indirect (strictly feedforward) field-oriented controller [6].

III. CURRENT-COMMAND INPUT-OUTPUT LINEARIZATION CONTROLLER FOR SATURATED INDUCTION MOTORS

In the field-oriented coordinates, an input-output linearization controller based on the saturated magnetic model of an induction motor can given by [4, 5]

$$\mathbf{i}^*_{rdq} = \begin{bmatrix} K_{lI}\int_0^t \left(l^*_{rd} - l_{rd}\right)dt + K_{lP}\left(l^*_{rd} - l_{rd}\right) \\ \dfrac{n_{ph}L_r}{2n_pM}\dfrac{t^*}{l_{rd}} \end{bmatrix} \tag{5}$$

where l_{rd} is the *d* component (or magnitude) of the rotor flux vector. Since the input-output linearization controller requires the knowledge of the rotor flux, the following flux estimator is used to estimate the rotor flux and transformation angle r [4, 5]:

$$\frac{d\hat{l}_{rd}}{dt} = \frac{MR_r}{L_r}\left(-f_m^{-1}(\hat{l}_{rd}) + \hat{i}_{rd}\right) \tag{6}$$

$$\frac{d\hat{r}}{dt} = n_p w + \frac{MR_r}{L_r}\frac{\hat{i}_{rq}}{\hat{l}_{rd}} \tag{7}$$

Note that the q component of the input-output linearization controller (5) is similar to the q component of the passivity-based controller (1).

The input-output linearization method requires a real-time division by the rotor flux state variable, which is a computational disadvantage.

$$\frac{d\left(\hat{t}_L / J\right)}{dt} = 0 + l_3\left(q - \hat{q}\right) \tag{10}$$

In the passivity-based methods, the discrete values of $\frac{1}{l^*_d}$ (or $\frac{1}{b}$) have to be stored in the memory of the processor in advance to prevent real-time division by l^*_d (or b). This means the passivity-based methods require more memory storage. In the input-output linearization method, implementing the flux estimator requires very little memory (only a couple of lines of programming code).

IV. EXPERIMENTAL RESULTS

Current-command passivity-based controller (1)-(2) was tested on an experimental setup which consisted of (i) a 3-phase, 6-pole, 1-Hp, squirrel cage induction motor, (ii) a Motorola DSP96002 (floating point processor) ADS system, (iii) a data acquisition board, and (iv) three 20 kHz PWM amplifiers (± 150 Volts and ± 10 Amperes). The parameters of the induction motor are listed in Table I. The position measurements were obtained through a 2880 pulses per revolution (resolution of $\frac{2p}{2880}$ radians) line encoder.

TABLE I

INDUCTION MOTOR PARAMETERS

M	0.225 H
L_r	0.244 H
L_s	0.244 H
R_r	2.1 Ω
R_s	1.85 Ω
J	0.0185 N-m-s^2
f	0.0 N-m/rad/sec

Fig. 2 shows the magnetization curve of the induction motor, which was determined by Novotnak [5].

In the experiment, the requirement was to maintain a constant speed regardless of an unknown load torque, which was assumed to be constant. A separately excited DC generator was coupled to the induction motor to act as a load torque.

The speed was estimated using the following speed estimator:

$$\frac{d\hat{q}}{dt} = \hat{w} + l_1\left(q - \hat{q}\right) \tag{8}$$

$$\frac{d\hat{w}}{dt} = \frac{2n_pM}{n_{ph}JL_r} l^*_d i_q - \frac{f}{J}\hat{w} - \frac{\hat{t}_L}{J} + l_2\left(q - \hat{q}\right) \tag{9}$$

The speed estimator models the load torque as a constant. Though the estimate of the load torque is computed in real-time, it is not used in the feedback controller.

Speed estimator (8)-(10) is a modified version of the speed estimator in [9]. Namely, we replaced the estimated flux with reference flux.

In the experiment, the motor was first brought up to a constant speed of 60 radians per second and then the DC generator was activated by closing a switch on the armature circuit of the DC generator at 1.5 seconds (The armature current was dumped into a high wattage resistor bank). Figures 3-5 show the results obtained with the passivity-based controller (1)-(2). Fig. 3 is a plot of estimated speed $\hat{w}$ and reference speed w^*. Note that the speed initially drops when the load torque was applied to the induction motor at 1.5 seconds. But, the feedback controller quickly responds and resumes the tracking. Fig. 4 shows the applied load torque whose value was computed by first measuring i_a, the armature current of DC generator, and then using $t_L = K_t i_a$ where K_t is the torque constant of DC generator.

In the experiment, the flux reference was chosen to be constant at 0.7 Webers since the speed was constant [5]. The magnitudes of the estimated rotor flux vector and reference rotor flux vector are given in Fig. 5. The rotor flux vector was estimated off-line by solving the following equation:

$$\frac{d\hat{l}_{dq}}{dt} + w_s \mathbf{J}\hat{l}_{dq} + \left(\frac{R_r M}{L_r} \frac{f_m^{-1}\left(\left|\hat{l}_{dq}\right|\right)}{\left|\hat{l}_{dq}\right|} \right)\hat{l}_{dq} = \left(\frac{R_r M}{L_r} \right)\mathbf{i}_{dq} \tag{11}$$

where $w_s = \frac{R_r M}{L_r}\frac{i_q^*}{l_d^*}$. The components of $\mathbf{i}_{dq}$ and i_q^* were collected from the experiment.

Fig. 2. Magnetization curve of the induction motor.

Fig. 3. Estimated (solid) and reference (dashed) speeds in radians per second versus time in seconds.

Fig. 4. Load torque in N-m versus time in seconds.

Fig. 5. Estimated (solid) and reference (dashed) fluxes in Webers versus time in seconds.

In addition to these experimental results, the same fast point-to-point move as in [4] and [7] was carried out using the passivity-based controller (1)-(2) (Again, in [4] and [7], the input-output linearization controller (5) and the passivity-based controller (3)-(4) are implemented, respectively). It was observed that the use of passivity-based controllers (1)-(2) and (3)-(4) results in reduced position tracking errors as compared with input-output linearization controller (5).

V. CONCLUSIONS

In this paper, two passivity-based controllers and an input-output linearization controller for current-fed saturated induction motors were presented. Advantages and disadvantages of these controllers were discussed briefly. Finally, experimental results obtained with one of the passivity-based controllers were presented.

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