ADAPTIVE BACKSTEPPING CONTROLLER DESIGN FOR LINEAR INDUCTION MOTOR POSITION CONTROL

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In this paper, the mover position control of a linear induction motor using an adaptive backstepping control design based on filed orientation is proposed. First, the indirect field oriented control LIM is derived. Then, a novel adaptive backstepping control design technique is investigated to achieve a position and flux tracking objective under parameter uncertainties and disturbance of load torque. The effectiveness of the proposed control scheme is verified by numerical simulation. The numerical validation results of the proposed scheme have presented good performances compared to the conventional backstepping controller.

Keywords: linear induction motor, vector control, adaptive backstepping control

1. Introduction

Nowadays, LIM's are now widely used, in many industrial applications including transportation, conveyor systems, actuators, material handling, pumping of liquid metal, and sliding door closers, etc. with satisfactory performance [1, 2]. The most obvious advantage of linear motor is that it has no gears and requires no mechanical rotary-to-linear converters. The linear electric motors can be classified into the following: D.C. motors, induction motors, synchronous motors and stepping motors, etc. Among these, the LIM has many advantages such as high-starting thrust force, alleviation of gear between motor and the motion devices, reduction of mechanical losses and the size of motion devices, high-speed operation, silence, and so on [1, 2, 3]. The driving principles of the LIM are similar to the traditional rotary induction motor (RIM), but its control

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characteristics are more complicated than the RIM, and the motor parameters are time varying due to the change of operating conditions, such as speed of mover, temperature, and configuration of rail [3, 4].

Field-oriented control (FOC) or vector control [2, 4, 5] of linear induction machine achieves decoupled trust and flux dynamics leading to independent control of the torque and flux as for a separately excited DC motor. This control strategy can provide the same performance as achieved from a separately excited DC machine. This technique can be performed by two basic methods: direct vector control and indirect vector control. Both DFO and IFO solutions have been implemented in industrial drives demonstrating performances suitable for a wide spectrum of technological applications [5, 6, 7]. However, the performance is sensitive to the variation of motor parameters, especially the rotor time-constant, which varies with the temperature and the saturation of the magnetizing inductance. Recently, much attention has been given to the possibility of identifying the changes in motor parameters of LIM while the drive is in normal operation. This stimulated a significant research activity to develop LIM vector control algorithms using nonlinear control theory in order to improve performances, achieving speed (or torque) and flux tracking, or to give a theoretical justification of the existing solutions [1, 6, 7, 8].

Due to new developments in nonlinear control theory, several nonlinear control techniques have been introduced in the last two decades. One of the nonlinear control methods that have been applied to linear induction motor is the backstepping design [8, 9, 10]. Backstepping is a systematic and recursive design methodology for nonlinear feedback control. This approach is based upon a systematic procedure for the design of feedback control strategies suitable for the design of a large class of feedback linearisable nonlinear systems exhibiting constant uncertainty, and it guarantees global regulation and tracking for the class of nonlinear systems transformable into the parametric-strict feedback form. The backstepping design alleviates some limitations of other approaches [8, 9, 10, 11, 12]. It offers a choice of design tools to accommodate uncertainties and nonlinearities and can avoid wasteful cancellations. The idea of backstepping design is to select recursively some appropriate functions of state variables as pseudo-control inputs for lower dimension subsystems of the overall system. Each backstepping stage results into a new pseudo-control design, expressed in terms of the pseudo-control designs from the preceding design stages. When the procedure terminates, a feedback design for the true control input results and achieves the original design objective by virtue of a Lyapunov function, which is formed by summing up the Lyapunov functions associated with each individual design stage [9, 10, 11].

In this paper, an adaptive backstepping control design based on filed orientation is proposed. The proposed controller is applied to achieve a position and flux tracking objective under parameter uncertainties and disturbance of load torque. The reminder of this paper is organized as follows. Section II reviews the principle of the indirect field-oriented control (FOC) of linear induction motor. Section III shows the development of the adaptive backstepping controller design for LIM position control. Section IV gives some simulation results. Finally, some conclusions are drawn in section V.

2. Indirect field-oriented control of the LIM

The primary (mover) of the adopted three-phase LIM is simply a 'cut-open-and-rolled-flat' rotary-motor primary. The secondary usually consists of a sheet conductor using aluminium with an iron back for the return path of the magnetic flux. The primary and secondary form a single sided LIM. Moreover, a simple linear encoder is adopted for the feedback of the mover position.

The dynamic model of the LIM is modified from traditional model of a three-phase, Y-connected induction motor and can be expressed in the *d*-*q* synchronously rotating frame as [1, 8, 13, 14, 15]:

$$\frac{di_{ds}}{dt}=\frac{1}{\sigma L_s}\left(-\left(R_s+\left(\frac{L_m}{L_r}\right)^2 R_r\right)\cdot i_{ds}+\sigma\cdot L_s\cdot\frac{\pi}{h}\cdot v_e\cdot i_{qs}+\frac{L_m\cdot R_r}{L_r^2}\cdot\phi_{dr}+\frac{P\cdot L_m\cdot\pi}{L_r\cdot h}\cdot\phi_{qr}\cdot v_r+V_{ds}\right) \quad (1)$$

$$\frac{di_{qs}}{dt}=\frac{1}{\sigma L_s}\left(-\sigma\cdot L_s\cdot\frac{\pi}{h}\cdot v_e\cdot i_{ds}-\left(R_s+\left(\frac{L_m}{L_r}\right)^2\cdot R_r\right)\cdot i_{qs}-\frac{P\cdot L_m\cdot\pi}{L_r\cdot h}\phi_{dr}\cdot v_r+\frac{L_m\cdot R_r}{L_r^2}\cdot\phi_{qr}+V_{qs}\right) \quad (2)$$

$$\frac{d\phi_{dr}}{dt}=\frac{L_m\cdot R_r}{L_r}i_{ds}-\frac{R_r}{L_r}\cdot\phi_{dr}+\left(\frac{\pi}{h}\cdot v_e-P\cdot\frac{\pi}{h}\cdot v_r\right)\cdot\phi_{qr} \quad (3)$$

$$\frac{d\phi_{qr}}{dt}=\frac{L_m\cdot R_r}{L_r}\cdot i_{qs}-\left(\frac{\pi}{h}\cdot v_e-P\cdot\frac{\pi}{h}\cdot v_r\right)\cdot\phi_{dr}-\frac{R_r}{L_r}\cdot\phi_{qr} \quad (4)$$

$$F_e=K_f\left(\phi_{dr}\cdot i_{qs}-\phi_{qr}\cdot i_{ds}\right)=M\cdot\dot{v}_r+D\cdot v_r+F_L \quad (5)$$

Where R_s is the winding resistance per phase, R_r is the secondary resistance per phase referred primary, L_m is the magnetizing inductance per phase, L_r is the secondary inductance per phase, L_s is the primary inductance per phase, v_r is the mover linear velocity, *h* is the pole pitch, *P* is the number of pole pairs, ϕ_{dr} and ϕ_{qr} are d-axis and q-axis secondary flux, respectively, i_{ds} and i_{qs} are d-axis and q-axis primary current, respectively, v_{ds} and v_{qs} are d-axis and q-axis primary voltage, respectively, $\tau_r=L_r/R_r$ is the secondary time-constant, $\sigma=1-\left(L_m^2/(L_s L_r)\right)$ is the leakage coefficient, $K_f=3P\pi L_m/(2hL_r)$ is the force constant, F_e is the electromagnetic force, F_L is the external force disturbance, M is the total mass of the moving element and D is the viscous friction and iron-loss coefficient.

The main objective of the vector control of linear induction motors is, as in DC machines, to independently control the electromagnetic force and the flux; this is done by using a d-q rotating reference frame synchronously with the rotor flux space vector [2, 5, 6, 7]. In ideally field-oriented control, the secondary flux linkage axis is forced to align with the d-axis, and it follows that [2, 5, 6]:

$$\phi_{rq} = \frac{d\phi_{qr}}{dt} = 0 \tag{6}$$

$$\phi_{dr} = \phi_r = \text{constant} \tag{7}$$

By use of the indirect field-oriented control technique and with the fact that the electrical time constant is much smaller than the mechanical time constant, the electromagnetic force shown in (5) can be reasonably represented by the following equations:

$$F_e = K_f \cdot i_{qs} \tag{8}$$

$$K_f = \frac{3}{2} P \frac{\pi \cdot L_m^2}{h \cdot L_r} i_{ds} \tag{9}$$

Moreover, using (4) the feedforward slip velocity signal can be estimated using ϕ_{rd} and i_{qs} as follows:

$$v_{sl} = \frac{h \cdot L_m}{\pi \cdot \tau_r} \frac{i_{qs}^*}{\phi_{dr}} \tag{10}$$

3. Adaptive backstepping control of LIM

a. Backstepping technique

Consider the system:

$$\dot{x} = f(x) + g(x)u\,,\ \ f(0) = 0 \tag{11}$$

Where $x \in R^n$ is the state and $u \in R$ is the control input. Let $u_{des} = \alpha(x)$, $a(0) = 0$ be a desired feedback control law, which, if applied to the system in (11), guarantees global boundedness and regulation of $x(t)$ to the equilibrium point $x = 0$ as $t \to \infty$, for all $x(0)$ and $V(x)$ is a control Lyapunov function, where [9, 10] :

$$\frac{\partial V(x)}{\partial x}\left[f(x) + g(x)\alpha(x)\right] < 0\,,\ V(x) > 0 \tag{12}$$

Consider the following cascade system:

$$\dot{x} = f(x) + g(x)y, \quad f(0) = 0 \tag{13}$$

$$\dot{\zeta} = m(x,\zeta) + \beta(x,\zeta)u, \quad h(0) = 0 \tag{14}$$

$$y = h(x) \tag{15}$$

Where for the system in (13), a desired feedback $a(x)$ and a control Lyapunov function $V(x)$ are known. Then, using the nonlinear block backstepping theory in [9, 10, 11, 12], the error between the actual and the desired input for the system in (13) can be defined as $z = y - \alpha$, and an overall control Lyapunov function $V(x,\zeta)$ for the systems in (13) and (14) can be defined by augmenting a quadratic term in the error variable z with $V(x)$:

$$V(x,\zeta) = V(x) + \frac{1}{2} z^2 \tag{16}$$

Taking the derivative of both sides gives:

$$\dot{V}(x,\zeta) = \dot{V}(x) + \frac{1}{2} z\dot{z} \tag{17}$$

From which solving for $u(x,\zeta)$, which renders $\dot{V}(x,\zeta)$ negative definite, yields a feedback control law for the full system in (13-15). One particular choice is [10]:

$$u = \left(\frac{\partial h(\zeta)}{\partial \zeta} \beta(x,\zeta) \right)^{-1} \left\{ -c(y-\alpha) - \frac{\partial h(\zeta)}{\partial \zeta} m(x,\zeta) + \frac{\partial \alpha(x)}{\partial x} \dot{x} - \frac{\partial V(x)}{\partial x} g(x) \right\}, \qquad c > 0 \tag{18}$$

b. Application to linear induction motor

The control objective is that the closed-loop control system is asymptotically stable and the mover position tracking of $d(t)$ to a desired reference signal $d_{ref}(t)$ and , which is assumed to have bounded derivatives up to the third-order.

Now, we use the adaptive backstepping techniques to achieve the stability and position tracking objectives.

Step 1:

For the control objective, the position tracking control, we regard the velocity v_r as the “control” variable (called virtual control in [10, 11, 12]). Define the position tracking error signal

$$e_1(t) = d_{ref}(t) - d(t) \tag{19}$$

Then its time derivative is

$$\dot{e}_1(t) = \dot{d}_{ref}(t) - \dot{d}(t) = \dot{d}_{ref}(t) - v(t) \tag{20}$$

Using the simple Lyapunov function

$$V_1 = \frac{1}{2} e_1^2 \tag{21}$$

We can obtain a proportional (P-) feedback with feedforward of the desired reference velocity [16, 17]

$$v(t) = k_1 e_1(t) + \dot{d}_{ref}(t) \tag{22}$$

Step 2:

Define another error signal between the velocity and the “desired velocity”

$$e_2(t) = v_{ref}(t) - v(t) = \left[k_1 e_1(t) + \dot{d}_{ref}(t)\right] - v(t) \tag{23}$$

So, the equation (22) can be expressed as

$$\dot{e}_1(t) = -k_1 e_1(t) + e_2(t) \tag{24}$$

Its time derivative can be writing as follows

$$\dot{e}_2(t) = k_1\left[-k_1 e_1(t) + e_2(t)\right] + \ddot{d}_{ref}(t) - Fv(t) - \frac{K_T}{M} i_{qs}(t) - \Gamma \tag{25}$$

We could use the Lyapunov function

$$V_2 = \frac{1}{2} e_1^2 + \frac{1}{2} e_2^2 \tag{26}$$

The derivative of V_2 along the trajectory of the error dynamical equations is

$$\dot{V}_2 = e_1 \dot{e}_1 + e_2 \dot{e}_2 \tag{27}$$

$$\dot{V}_2 = -k_1 e_1^2 - k_2 e_2^2 + e_2\left[(1 - k_1^2) e_1 + (k_1 + k_2) e_2 + \ddot{d}_{ref} - Fv - \Gamma\right]$$

Where $k_2 > 0$ is a design constant, $F = \frac{f_c}{M}$ and $\Gamma = -\frac{F_L}{M}$

If we choose $i_{qs}(t)$ as

$$i_{qsref} = \frac{M}{K_T}\left[(1 - k_1^2) e_1 + (k_1 + k_2) e_2 + \ddot{\theta}_{ref} - Fv - \frac{K_T}{J} i_{1q} - \Gamma\right] \tag{28}$$

Then we could get

$$\dot{V}_2 = -k_1 e_1^2 - k_2 e_2^2 < 0$$

So the “control” $i_{qs}(t)$ in (28) is asymptotically stabilizing.

Since the parameters M, F and Γ are unknown, we need to use their estimates $\hat{M}(t)$, $\hat{F}(t)$, $\hat{\Gamma}(t)$ in (28), that is,

$$\hat{i}_{qsref} = \frac{\hat{M}}{K_T}\left[(1 - k_1^2) e_1 + (k_1 + k_2) e_2 + \ddot{\theta}_{ref} - \hat{F}v - \hat{\Gamma}\right] \tag{29}$$

Step 3:

Now, we define the final error signal,

$$e_3(t) = \hat{i}_{qsref} - i_{qs}(t)$$

$$e_3(t) = \frac{\hat{J}}{K_T}\phi_1(t) - i_{qs}(t) \tag{30}$$

Where $\phi_1(t)$ is a known signal. Using this definition, we can express the dynamical equation (25) as,

$$\dot{e}_2(t) = -e_1(t) - k_2 e_2(t) + \frac{K_T}{M} e_3(t) - \frac{\tilde{M}}{M}\phi_1(t) + \tilde{F}\omega_r(t) + \tilde{\Gamma} \tag{31}$$

and compute the derivative equation for $e_3(t)$ as

$$\dot{e}_3(t) = \dot{\hat{i}}_{qsref} - \dot{i}_{qs}(t)$$

$$\dot{e}_3(t) = \phi_2(t) - bv_{1q}(t) + \frac{\tilde{J}}{J}\phi_3(t) + \tilde{F}\phi_4(t)\omega_r(t) + \tilde{\Gamma}\phi_4(t) \tag{32}$$

Where, $\tilde{M} = \hat{M} - M, \tilde{F} = \hat{F} - F, \tilde{\Gamma} = \hat{\Gamma} - \Gamma$ are the parameter estimation errors, and ϕ_1, ϕ_2, ϕ_3, ϕ_4 are known signal expressed by the following expressions [16, 17]

$$\phi_1(t) = \left[(1 - k_1^2)e_1 + (k_1 + k_2)e_2 + \ddot{\theta}_{ref} - \hat{F}\omega_r - \hat{\Gamma}\right]$$

$$\phi_2(t) = \frac{\dot{\hat{J}}}{K_T}\phi_1(t) + \frac{\hat{J}}{K_T}\Big[(1 - k_1^2)(-k_1 e_1 + e_2) + (k_1 + k_2)(-e_1 - k_2 e_2)$$
$$+ \dddot{\theta}_{ref} - \dot{\hat{F}}\omega_r - \hat{F}^2\omega_r - \dot{\hat{\Gamma}} - \hat{F}\hat{\Gamma}\Big] - a_1\omega_r - a_2 i_{1q}$$
$$+ (k_1 + k_2)e_3 - \hat{F}i_{1q}$$

$$\phi_3(t) = (k_1 + k_2)e_3 - \frac{\hat{J}}{K_T}(k_1 + k_2)\phi_1 - \hat{F}i_{1q}$$

$$\phi_4(t) = \frac{\hat{J}}{K_T}(k_1 + k_2 + \hat{F})$$

Step 4:

Now, we add terms concerning e_3 and $\tilde{M}$, $\tilde{F}$ and $\tilde{\Gamma}$ to V_2 to form the following Lyapunov function

$$V_e = \frac{1}{2}\left[e_1^2 + e_2^2 + e_3^2 + \frac{1}{J\gamma_1}\tilde{J}^2 + \frac{1}{\gamma_2}\tilde{F}^2 + \frac{1}{\gamma_3}\tilde{\Gamma}^2\right]$$

Where $\gamma_i(i = 1, 2, 3)$ are positive design constants of adaptive gains.

Using equations (23), (31), and (32), we can compute the derivative of V_e along the trajectory of error dynamical equations as

$$\dot{V}_e = e_1\dot{e}_1 + e_2\dot{e}_2 + e_3\dot{e}_3 + \frac{\tilde{M}}{M\gamma_1}\dot{\hat{M}} + \frac{1}{\gamma_2}\tilde{F}\dot{\hat{F}} + \frac{1}{\gamma_3}\tilde{\Gamma}\dot{\hat{\Gamma}}$$

$$= -k_1 e_1^2 - k_2 e_2^2 - k_3 e_3^2 + \frac{K_T}{J} e_2 e_3 + e_3\left[k_3 e_3 + \phi_2 - b v_{1q}\right]$$

$$+ \frac{\tilde{J}}{J}\left[-\phi_1 e_2 + e_3\phi_3 + \frac{1}{\gamma_1}\dot{\hat{J}}\right] + \tilde{F}\left[e_2\omega_r + e_3\phi_4\omega_r + \frac{1}{\gamma_2}\dot{\hat{F}}\right]$$

$$+ \tilde{\Gamma}\left[e_2 + e_3\phi_4 + \frac{1}{\gamma_3}\dot{\hat{\Gamma}}\right]$$

Therefore, if we choose the control law as

$$v_{qs} = \frac{1}{b}\left[k_3 e_3 + \phi_2\right] \tag{33}$$

And the update laws as

$$\begin{aligned} \dot{\hat{J}} &= -\gamma_1\left[-e_2\phi_1 + e_3\phi_3\right] \\ \dot{\hat{F}} &= -\gamma_2\left[e_2\omega_r + e_3\phi_4\omega_r\right] \\ \dot{\hat{\Gamma}} &= -\gamma_3\left[e_2 + e_3\phi_4\right] \end{aligned} \tag{34}$$

Then we get

$$\dot{V}_e = -k_1 e_1^2 - k_2 e_2^2 - k_3 e_3^2 + \frac{K_T}{M} e_2 e_3 \le 0 \tag{35}$$

For sufficiently large $k_2, k_3 > 0$. With the control law (33), the dynamical equation (32) can be written as

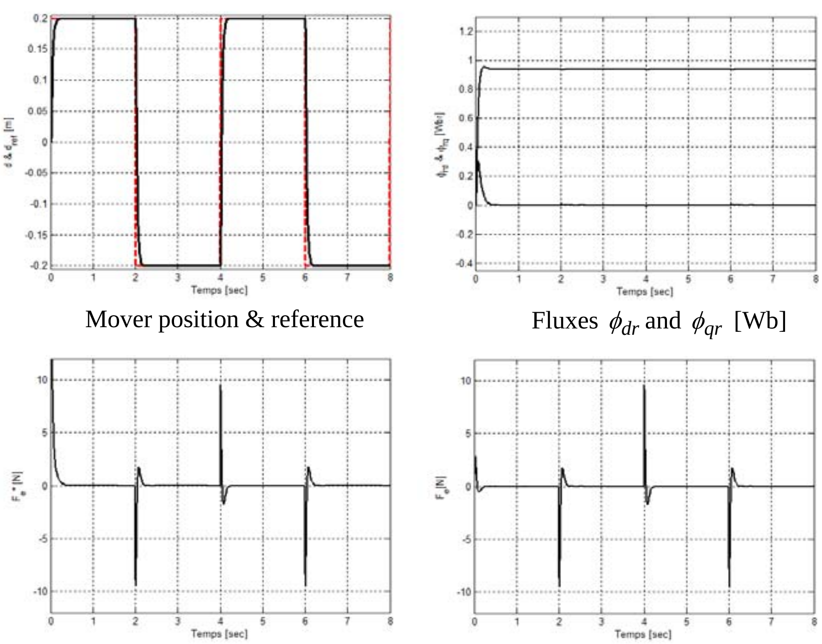
$$\dot{e}_3(t) = -k_3 e_3(t) + \frac{\tilde{J}}{J}\phi_3(t) + \tilde{F}\phi_4(t)\omega_r(t) + \tilde{\Gamma}\phi_4(t) \tag{36}$$

4. Simulation results

We demonstrate the effectiveness of the proposed control scheme for position control of the linear induction motor.

First, we present the simulated results of the proposed adaptive backstepping control system for periodic square, sinusoidal and triangular inputs. The parameter used in simulation are chosen as $k_1 = 18$, $k_2 = k_3 = 12$, $\gamma_2 = 0.025$, $\gamma_2 = 0.0036$, $\gamma_3 = 0.006$.

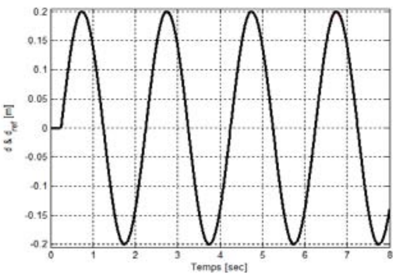
The position responses of the mover, electromagnetic force, d-flux, q-flux and the control effort are shown in Figs. 1, 2 and 3. From the simulated results, the proposed adaptive backstepping controller can track periodic step, sinusoidal and triangular inputs precisely. Next, the simulated results of the proposed adaptive backstepping control system for periodic step, sinusoidal and triangular inputs with load force disturbances (constant, sinusoidal and triangular load force) are shown in Figs. 4, 5, 6, 7 and 8. From simulated results, the tracking responses of the proposed controller are insensitive to load force application (the controller reject the external disturbance without overshoot and with a minimum response time). Fig.9 shows error position for adaptive backstepping control of LIM. A comparison between the proposed controller (adaptive backstepping) and the conventional backstepping is shown in Fig. 10 and 11 for step, sinusoidal and triangular reference signal (error position) for different variation of the total mass. In Figs. 10 and 11, it can be observed that the position response of the adaptive backstepping controller present better tracking characteristics, have minor insensitive to the mass variation and is more robust than the conventional backstepping controller. Fig. 12 shows the values of the estimated parameters M , F and Γ .



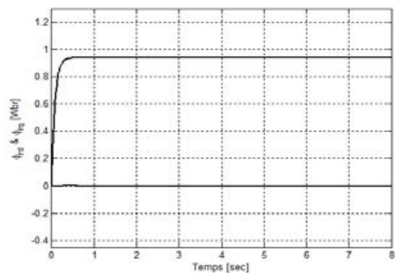
Mover position & reference — Fluxes ϕ_{dr} and ϕ_{qr} [Wb]

Control effort — Electromagnetic force [N]

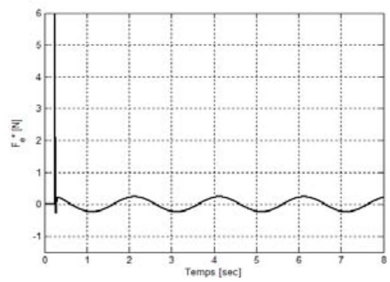
Fig. 1. Simulated results of adaptive backstepping controller for LIM position control (Step reference change)



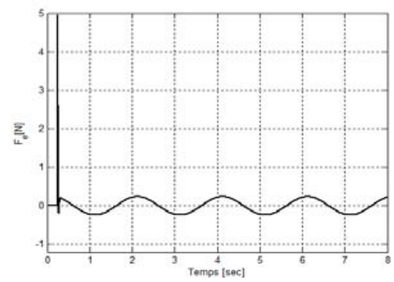
Mover position & reference



Fluxes ϕ_{dr} and ϕ_{qr} [Wb]

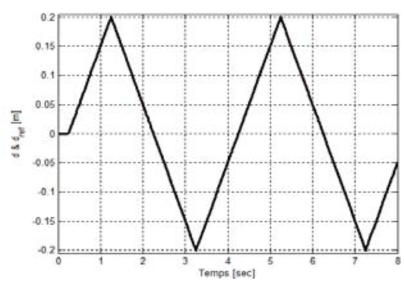


Control effort [N]

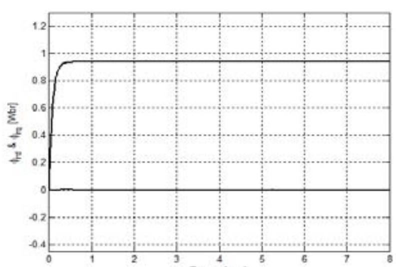


Electromagnetic force [N]

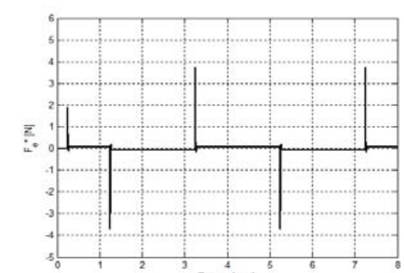
Fig. 2. Simulated results of adaptive backstepping controller for LIM position control (Sinusoidal reference change)



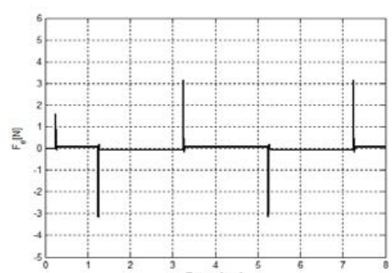
Mover position & reference [m]



Fluxes ϕ_{rd} and ϕ_{rq} [Wb]

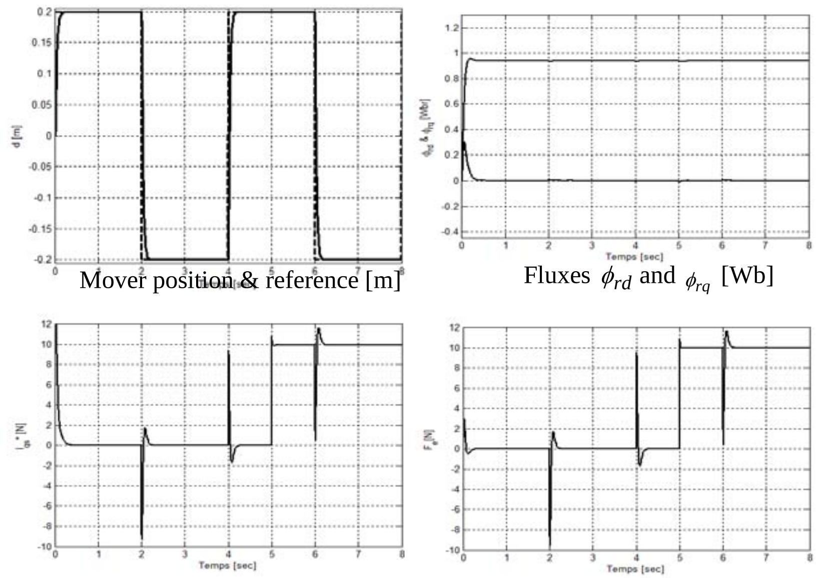


Control effort [N]



Electromagnetic force [N]

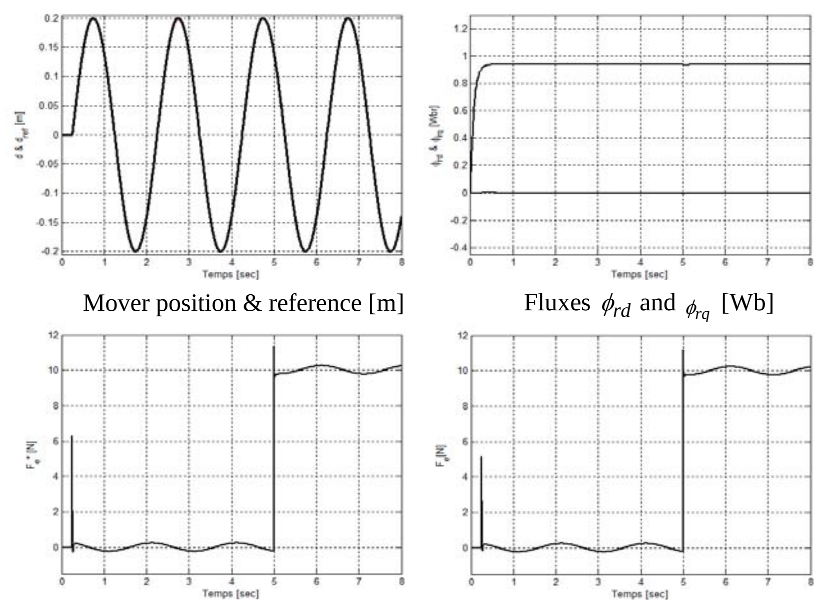
Fig. 3. Simulated results of adaptive backstepping controller for LIM position control (Triangular reference change)



Mover position & reference [m] Fluxes ϕ_{rd} and ϕ_{rq} [Wb]

Control effort [N] Electromagnetic force [N]

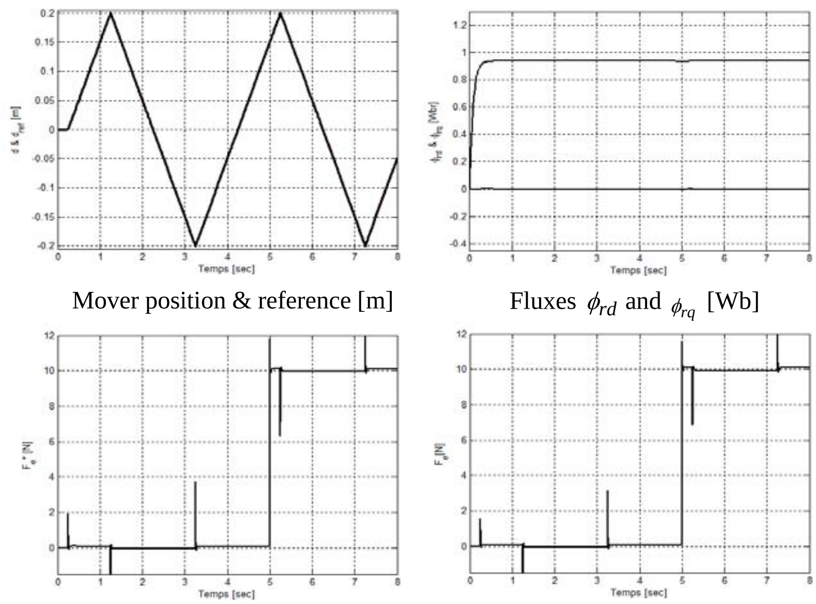
Fig. 4. Simulated results of adaptive backstepping controller
for LIM position control with load force variation:
Constant load force 10N occurring at 5sec.



Mover position & reference [m] Fluxes ϕ_{rd} and ϕ_{rq} [Wb]

Control effort [n] Electromagnetic force [N]

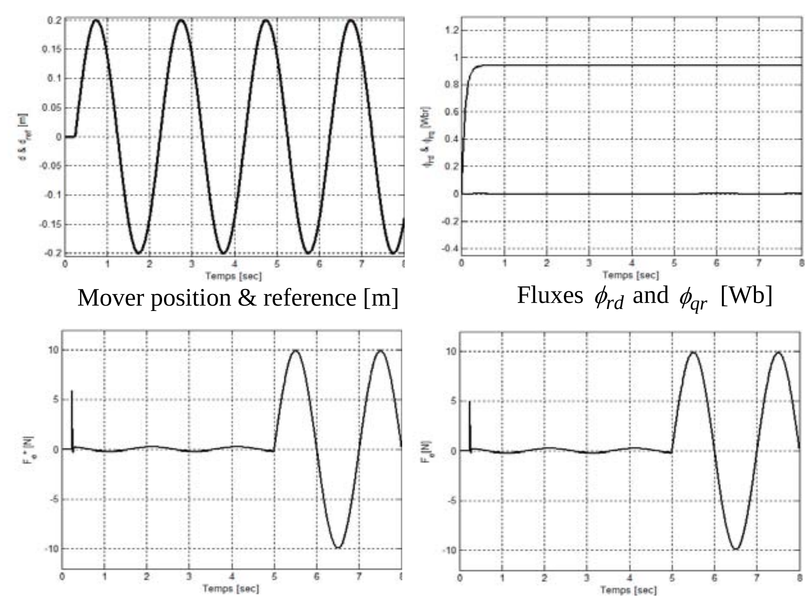
Fig. 5. Simulated results of adaptive backstepping controller for
LIM position control with load force variation:
Constant load force 10N occurring at 5sec.



Mover position & reference [m] Fluxes ϕ_{rd} and ϕ_{rq} [Wb]

Control effort [N] Electromagnetic force [N]

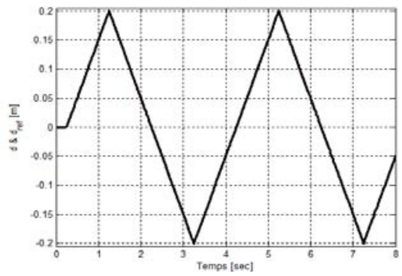
Fig. 6. Simulated results of adaptive backstepping controller for LIM position control with constant load force *10N*



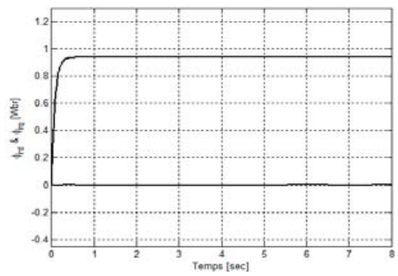
Mover position & reference [m] Fluxes ϕ_{rd} and ϕ_{qr} [Wb]

Control effort [N] Electromagnetic force [N]

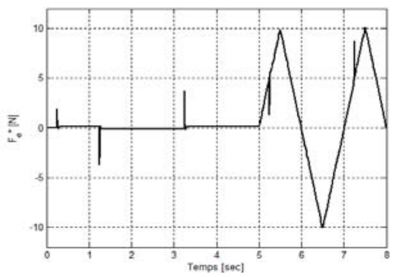
Fig. 7. Simulated results of adaptive backstepping controller for LIM position control with load force variation: Sinusoidal load force *10N* occurring at 5s.



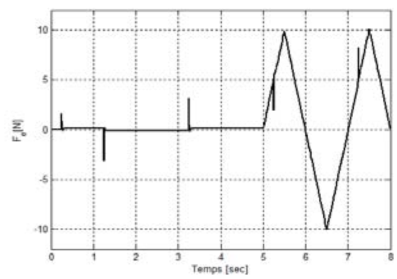
Mover position & reference [m]



Fluxes ϕ_{dr} and ϕ_{qr} [Wb]



Control effort [N]



Electromagnetic force [N]

Fig. 8. Simulated results of adaptive backstepping controller for LIM position control with load force variation: Triangular load force *10N* occurring at 5s.

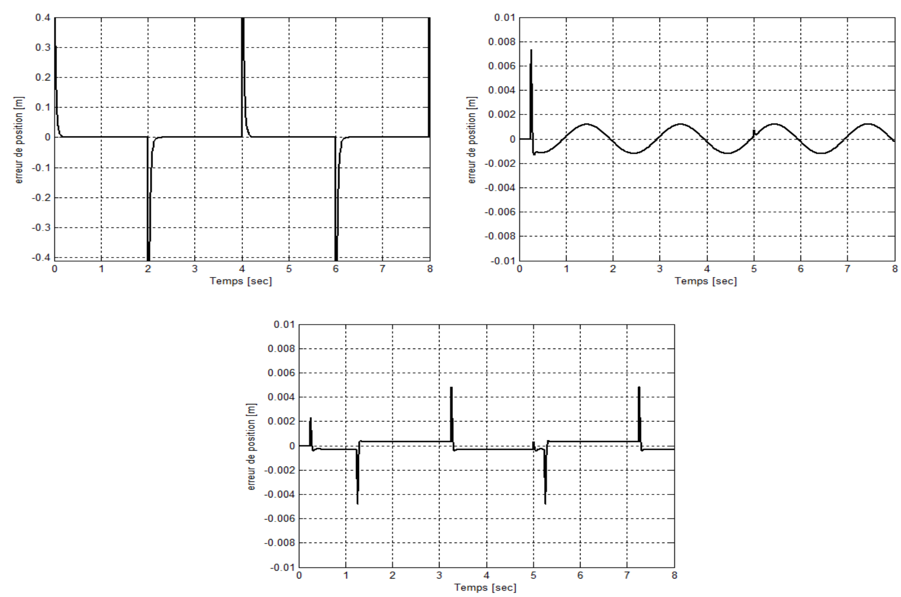


Fig. 9. Simulated results of the adaptive backstepping control for LIM error tracking

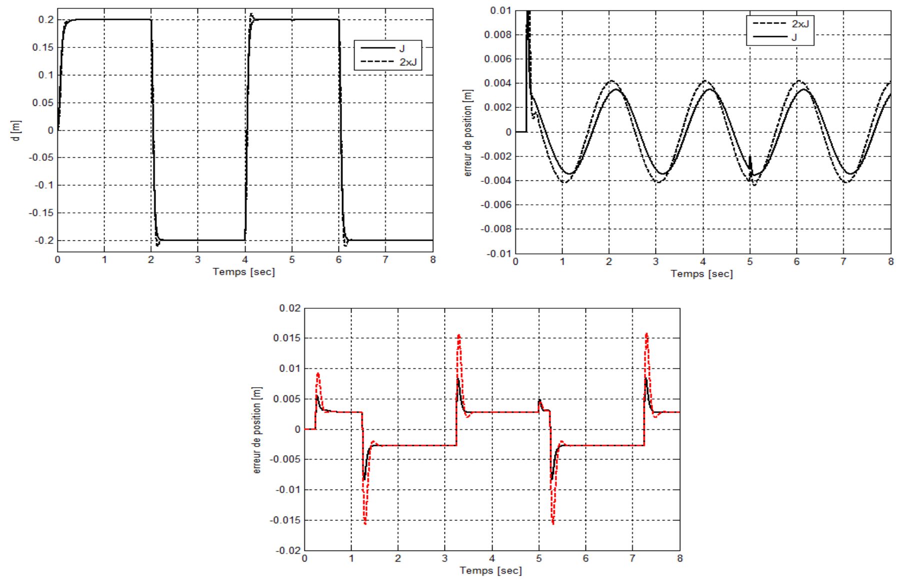


Fig. 10. Simulated results of the a conventional backstepping control for LIM error tracking with mass value variation

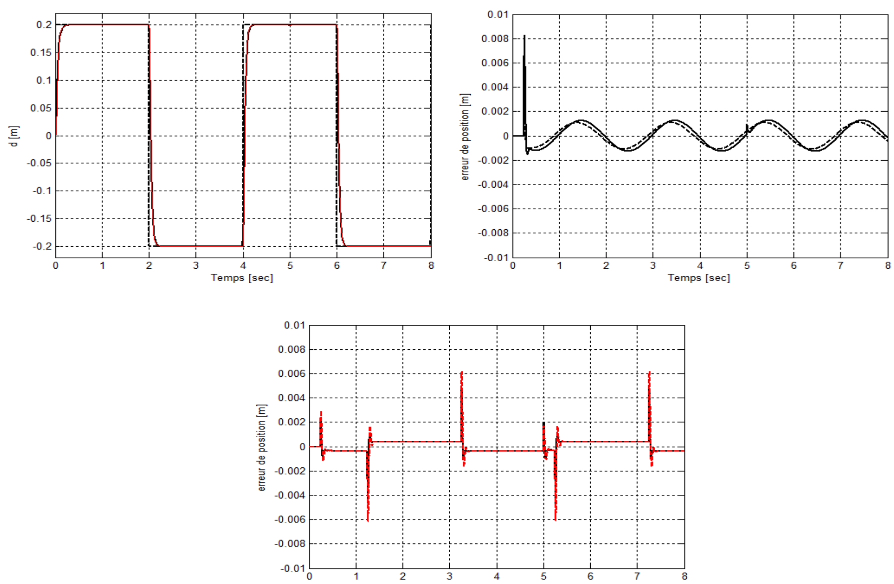


Fig. 11. Simulated results of the a conventional backstepping control for LIM error tracking with mass value variation

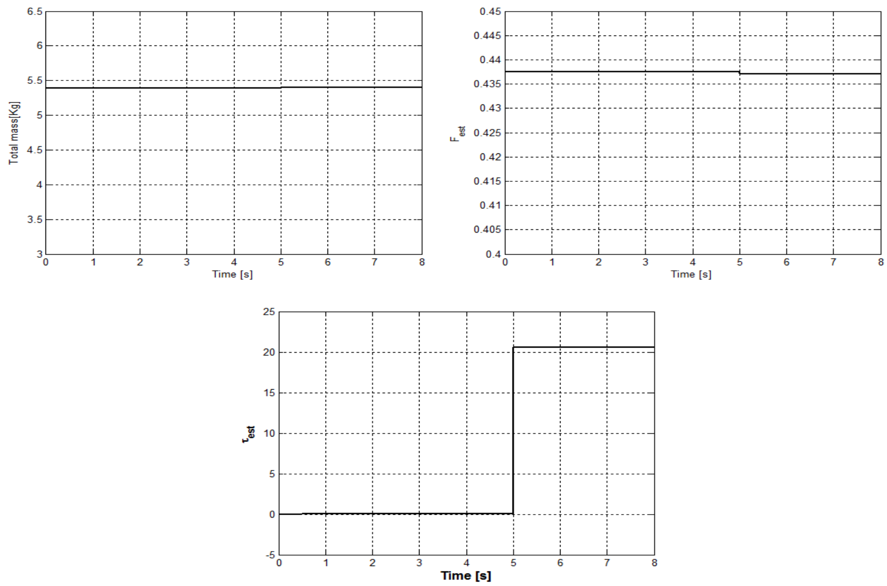


Fig. 12. Estimated parameters $\hat{M}$, Γ and $\hat{F}$ in nominal case

5. Conclusions

This paper has demonstrated the applications of a nonlinear adaptive control system to the periodic motion control of a LIM. First, an adaptive backstepping controller for position control of LIM was designed. Moreover, a novel adaptive backstepping control design technique is investigated to achieve a position and flux tracking objective under parameter uncertainties and disturbance of load torque. The control dynamics of the proposed hierarchical structure has been investigated by numerical simulation. Simulation results have shown that the proposed adaptive backstepping controller has presented satisfactory performances (no overshoot, minimal rise time, best disturbance rejection) for time-varying external force disturbances and total mass variation. Finally, the proposed controller provides drive robustness improvement.

Appendix

Table 1

Linear induction motor parameters

ϕ_{2s} [Wb]	0.9378	L_s [H]	0.1078
R_s [Ω]	0.34	f_n [Hz]	50
R_r [Ω]	0.195	M [kg]	5.47
L_r [H]	0.1078	D [Nm.s/rd]	2.36
L_m [H]	0.1042	p	2

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