

On tracking control problem for polysolenoid motor model predictive approach

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ABSTRACT

The Polysolenoid Linear Motor (PLM) have been playing a crucial role in many industrial aspects due to its functions, in which a straight motion is provided directly without mediate mechanical actuators. Recently, with several commons on mathematic model, some control methods for PLM based on Rotational Motor have been applied, but position, velocity and current constraints which are important in real systems have been ignored. In this paper, position tracking control problem for PLM was considered under state-independent disturbances via min-max model predictive control. The proposed controller forces tracking position errors converge to small region of origin and satisfies state including position, velocity and currents constraints. Further, a numerical simulation was implemented to validate the performance of the proposed controller.

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1. INTRODUCTION

Linear Motor transmission systems are widely applied to provide directed straight motions in which, mechanical actuators are eliminated, resulting in better performance of motion systems. Generally, Polysolenoid Linear Motor (PLM) has a durable structure, operations according to electromagnetic phenomenon with principles as shown in [1-12] and various applications such as CNC Lathe [13], sliding door [14]. Without the need of any gear box for motion transformation, the PLM system becomes sensitive due to external impacts such as frictional force, end – effect, changed load and non-sine of flux. These effects encounter both in the longitudinal and in the transversal direction, which is along with saturation in supplied voltage, make good control performance from the linear drive a difficult task.

There are several researches taking into account the position control of PLM in presence of external disturbances. The authors in [15] presented a control design method to regulate velocity based on PI – self-tuning combining with appropriate estimation technique at slow velocity zone, but if load is changed, PI–self-tuning controller will be not efficient. In order to overcome changed load, model reference control method based on Lyapunov stability theory was employed in [16]. Additionally, the compensation approaches were proposed in research [17] on which, the frictional force were estimated by Lugrie and Stribeck friction model respectively. In [18], the advantage of that the sliding mode control applied in Linear Motor is that real position value tracks set point. However, the disadvantages of this method are finding sliding surface and chattering. In the view of nonlinear systems, the study in [19] apply linearization method to PLM system but this method is restricted by uncertain parameter and disturbances. It is clear that

the previous researches do not mention position, velocity and currents constraints as well as impact of external disturbance which is important properties of the control systems.

2. DYNAMIC MODEL

Polysolenoid linear motor is constructed according to electromagnetic induction as shown in Figure 1.

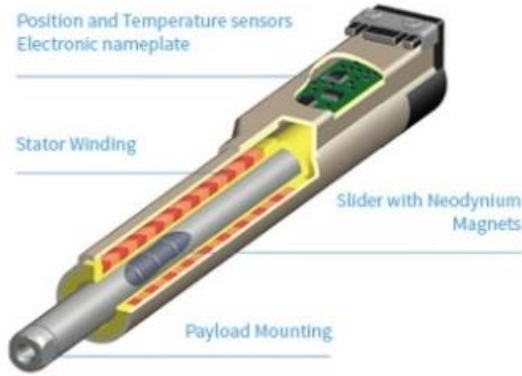


Figure 1. Composition of polysolenoid linear motor

Let us consider a dynamic model of PLM in [20-24]:

$$\begin{aligned} \frac{di_{sd}}{dt} &= -\frac{R_s}{L_{sd}}i_{sd} + \left(\frac{2\pi p}{\tau}v\right)\frac{L_{sq}}{L_{sd}}i_{sq} + \frac{u_{sd}}{L_{sd}}, \\ \frac{di_{sq}}{dt} &= -\frac{R_s}{L_{sq}}i_{sq} - \left(\frac{2\pi}{\tau}v\right)\frac{L_{sd}}{L_{sq}}i_{sd} - \left(\frac{2\pi p}{p\tau}v\right)\frac{\psi_p}{L_{sq}} + \frac{u_{sq}}{L_{sq}}, \\ \frac{dv}{dt} &= \frac{2\pi p}{\tau}(\psi_p + (L_{sd} - L_{sq})i_{sd})i_{sq} - \frac{1}{m}F_l, \\ \frac{dx}{dt} &= v. \end{aligned} \quad (1)$$

Where i_{sd} , i_{sq} , v , x stand for current, velocity and position respectively. In addition, the constant parameter of PLM includes: R_s , L_{sd} , L_{sq} , p , τ , ψ_p , m as resistance, inductor, pole pair, pole step, flux and mass of rotor. The input voltage is presented as u_{sd} , u_{sq} and F_c is unmeasured external force. Suppose that, F_l can be classify into finite set as $W = \{F_{1c}, F_{2c}, \dots, F_{Nc}\}$. It is worth to note that, the model (1) is similar to permanent magnet rotation synchronization motor model. Hence, let us assume that, L_{sd} and L_{sq} has the approximate similar values and the term $(L_{sd} - L_{sq})i_{sd}i_{sq}$ can be ignored in the third equation of (1) that leads to linear relationship between current i_{sq} and position-velocity.

3. CONTROL DESIGN

In this paper, let us separate dynamic model (1) into current subsystem and position subsystem. As aforementioned, position subsystem can be considered as a linear time invariant system under external disturbances and then, the position controller is designed with the aid of existed min-max model predictive control theory [25]. On the other hand, the cross-current compensation method between is used to transform first of two equation in (6) into linear form. Moreover, the current controller is based CCS model predictive control which can solve current constraint problem.

3.1. Control of current subsystem

By applying decoupling control law as follow:

$$u_{sd} = -\left(\frac{2\pi p}{\tau}v\right)L_{sq}i_{sq} + u_1, u_{sq} = \left(\frac{2\pi p}{\tau}v\right)L_{sd}i_{sd} + \left(\frac{2\pi p}{p\tau}v\right)\psi_p + u_2. \quad (1)$$

The current subsystem is transformed to linear system:

$$\frac{di_{sd}}{dt} = -\frac{R_s}{L_{sd}}i_{sd} + \frac{u_1}{L_{sd}}, \frac{di_{sq}}{dt} = -\frac{R_s}{L_{sq}}i_{sq} + \frac{u_2}{L_{sq}} \quad (2)$$

To design CCS-MPC in current closed loop, from (3), by using simple manner, let us obtain the discrete time predictive model as follows:

$$i_{dq}^{est}(k+i+1) = \Phi i_{dq}^{est}(k+i) + H\bar{u}_{dq}(k+i), \forall i = 0, 1, \dots, N_c - 1 \quad (3)$$

Where N_c is a prediction horizon, $i_{dq}^{est}(k+i+1) = [i_d^{est}(k+i+1), i_q^{est}(k+i+1)]^T$ is $(i+1)^{th}$. Estimated current vector on dq -coordinate and $i_{dq}^{est}(k) = i_{dq}(k) = [i_d(k), i_q(k)]^T$ at sample time k . The predictive control input vector is presented as $\bar{u}_{dq}(k+i) = [\bar{u}_1(k+i), \bar{u}_2(k+i)]^T$, $\bar{u}_1(k) = u_1(k)$, $\bar{u}_2(k) = u_2(k)$. With T_s stand for sampling time, the state and input matrix Φ, H are:

$$\Phi = \begin{bmatrix} e^{-\frac{R_s T_s}{L_{sd}}} & 0 \\ 0 & e^{-\frac{R_s T_s}{L_{sq}}} \end{bmatrix}, H = \begin{bmatrix} \frac{1}{L_{sd}} \\ \frac{1}{L_{sq}} \end{bmatrix} \quad (4)$$

The control task is to find the sequence of control input vector $\bar{u}_{dq}(k), \bar{u}_{dq}(k+1), \dots, \bar{u}_{dq}(k+N-1)$, which minimize the following cost function:

$$J = \sum_{i=1}^N \left(i_{dq}^{ref} - i_{dq}^{est}(k+i+1) \right)^T Q_1 \left(i_{dq}^{ref} - i_{dq}^{est}(k+i|k) \right) \quad (6)$$

Where $Q_1 = [\lambda_d 0; 0 1]$ is positive matrix, i_q^{ref} is reference input vector from position controller which is present in the next section, λ_d represent proportional ratio between d -current error $|i_d^{ref} - i_d|$ and q -current error $|i_q^{ref} - i_q|$. In order to simplify the optimization (6), let us assume that the reference input vector i_q^{ref} is constant in the prediction horizon time. The assumption is common in practical experiment due to the fact that current loop transition response is considerably faster than its position loop. Moreover, the predictive control input normally subjected to the linear constraint $A_{con}\bar{u}_{dq}(k+i) < B_{con}$. The current controller frequently operate in small sample T_s , thereby the one step horizon $N_c = 1$ was take into account for the control design. And then, by selecting optimal variable $\bar{u}_{dq}(k) = u_{dq}(k)$, the minimization (6) can be rewritten as:

$$\begin{aligned} \min_{\bar{u}_{dq}(k)} J &= \bar{u}_{dq}(k)^T (H^T Q H) \bar{u}_{dq}(k) + 2(\Phi i_{dq}(k) - i_{dq}^{ref})^T Q_2 H \bar{u}_{dq}(k) + \\ \text{C.s.t. } &A_{con} \bar{u}_{dq}(k) < B_{con}, \forall k = 1, 2, \dots, N-1 \end{aligned} \quad (5)$$

3.2. Control of position subsystem

The dynamic of position subsystem is significantly slower than current subsystems. Hence, in the control design of position, we assumed that the desired current equals to actual current. By setting $i_{sq} = u + m\tau(2\pi p\psi_p)^{-1} \ddot{x}_r$, the last two equation of (1) can be rewritten as discrete state space model of tracking errors as:

$$\begin{aligned} z_{k+1} &= A_d z_k + B_d u_k + D_d d_k, \\ y_k &= C z_k. \end{aligned} \quad (6)$$

Where: $e_x = x - x_r, e_v = v - v_r, z = [e_x, e_v]^T, d_k = F_l(kT_w), u_k = i_{sq}(k)$

$$A_d = \begin{bmatrix} 1 & T_w \\ 0 & 1 \end{bmatrix}, B_d = \begin{bmatrix} \frac{T_w^2}{2} \\ 2\pi p(m\tau)^{-1} T_w \end{bmatrix}, D_d = \begin{bmatrix} \frac{T_w^2}{2} \\ -m^{-1} T_w \end{bmatrix}, C = [1, 0],$$

The control input u_k is designed to force error vector z_k converge to small region centered at origin. Moreover, z_k, u_k satisfies the following constraints:

$$z_k \in Z, u_k \in U \text{ and } z_k \rightarrow Z_0 \in Z, \quad (7)$$

where $Z_0 \triangleq \{(z_1, z_2) \in \mathbb{R}^2 | e_{12} v_{max} 0 v_{min} 0 x_{max} 0 x_{min}\}$
 $Z_0 \in Z \triangleq \{(z_1, z_2) \in \mathbb{R}^2 | e_{12} v_{max} v_{min} x_{max} x_{min}\}$
 $U \triangleq (U_{min}^{max} \in \mathbb{R}.)$

To achieve control objective (9), min-max model predictive control proposed in [6] was applied. The operation of the position controller is divided into two modes: an “inner” and an “outer” controller. The inner controller is activated when the state is in the robust control invariant set Z_0 , and its role is to keep the error state in Z_0 under external disturbance d_k . The inner controller is linear feedback $u_k = Kz_k$ and it is important in the construction of the control robust invariant set Z_0 which is selected based on $(A_d + B_d K)^s = 0$, s is positive integer number. To be specific, let us select $Z_0 = \sum_1^s (A_d + B_d K)^s D_d W$ where, W is disturbance set.

The outer controller works when the error state is outside the invariant set Z_0 and steers the system state to the invariant set. For the outer controller, we use min-max model predictive control, and consider a fixed horizon formulation. In this section, the selected quadratic cost function as:

$$L(z, \underline{u}, \underline{d}) = \sum_{i=0}^{N_W-1} (z_{k+i}^T Q z_{k+i} + u_{k+i}^T R u_{k+i}),$$

$$Q \geq 0, R > 0. \tag{8}$$

Where $\underline{z} = [z_k, z_{k+1}, \dots, z_{k+N-1}]^T, \underline{u} = [u_k, u_{k+1}, \dots, u_{k+N-1}]^T, \underline{d} = [d_{k+1}, d_{k+2}, \dots, d_{k+N-1}]^T$. Herein, sequence vector \underline{u} is chosen to minimize cost function (10) in the worst case where, \underline{d} is maximum point of (10). The following algorithm summarize the operation of position controller.

Algorithm 1: (Position Controller)

Data: \mathbf{z}_k

If $z_k \in Z_0$, set $u_k = Kz_k$. Otherwise, find the solution of (11) and set \mathbf{u}_k to the first control in the optimal sequence \underline{u} .

4. NUMERICAL SIMULATION

In this section, we simulate the position tracking of whole system under state, input constraint and external disturbance in following table. We use the same current controller and two different prediction horizons of position controller to compare quality of each controller. The parameter of polysolenoid linear motor and controller show in Table 1. As can be seen in Figure 2, both position and velocity error stay in state constraints region and converges to small ball centered at origin. The current is satisfying input constraint under time varying external forces.

Table 1. The parameter of polysolenoid linear motor and controller

Parameter	Value	Controller parameter	Value
Pole pair	1	$e_{0x} min_{0x} max$	0.2 (mm)
Pole step	20 (mm)	$e_{0v} min_{0v} max$	1.5 (m/s)
Rotor mass	0.17 (kg)	λ_d	10
Phase coil Resistance	10.3 (Ω)	K	[-300 -5]
d-axis inductance	1.4 (mH)	U	[-50, 50]
q-axis inductance	1.4 (mH)		
Flux	0.035 (Wb)		

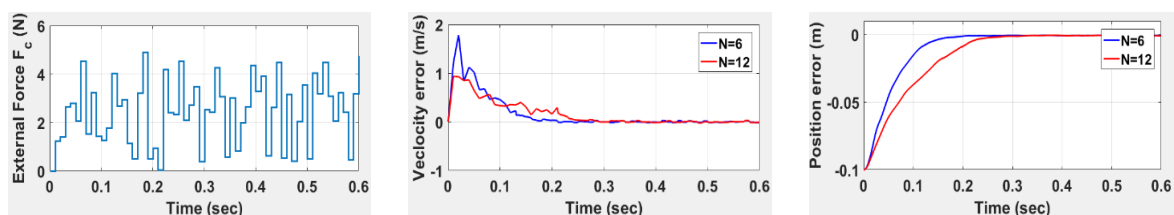


Figure 2. The performance of proposed controller (continue)

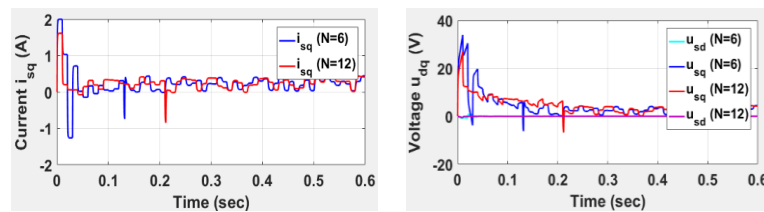


Figure 2. The performance of proposed controller

5. CONCLUSION

We also illustrated the impact of prediction horizon N_w on performance of the system. If N_w is small, the dynamic closed loop system is fast, settling time is small but input constraints may not hold. We have to choose N is sufficiently large to satisfy constraints. The proposed two cascade loops based on MPC with sufficiently small prediction horizon (N_c) of the current loop to reduce calculation load of microprocessors due to small sampling time of the current loop. When considering the current response is ideal, the system performance is totally depending on min-max MPC of the outer loop.

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