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TP–Based model transformation and gain-scheduling control of electromagnetic actuated clutch systems

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Abstract. This paper suggests two combinations of Tensor Product (TP)–based model transformation and gain-scheduling control leading to two efficient control system structures that are applied to the position control of nonlinear electromagnetic actuated clutch systems. One of the two combinations is included in a cascade control system structure that consists of a TP–based controller in the inner control loop and a gain-scheduling–based controller in the outer control loop. The proposed control system structures were tested on the nonlinear process model and validated by simulation results. A comparative analysis is given.

Keywords: cascade control system, electromagnetic actuated clutch systems, gain-scheduling control, mass position control, TP–based model transformation.

1. Introduction

This paper focuses on the development of two control structures (CSs) made of a combination of Tensor Product (TP)–based model transformation (TPBMT) and gain-scheduling control (GSC), which are applied to the position control of electromagnetic actuated clutch systems (EACSs). Our CSs are designed in the context of popular CS design approaches for EACS given in [1] – [5].

TPBMT is a numerical approach, which starts with Linear Parameter–Varying (LPV) dynamic models and derivates Linear Time-Invariant (LTI) systems as shown [6] – [9]. Representative applications of TPBMT are given in [10]–[15], and a combination of TPBMT and gain-scheduling control is presented in [16]. The combination with sliding mode control reported in [17] proves to be efficient by

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incorporating several theoretical and application results related to sliding mode control given in [18] – [21].

The main contribution of this paper is the design and validation of two CSs for the mass position control of EACS. In the first step a TP-based control structure (TP-CS) is designed for the position control of the EACS using the Parallel Distributed Compensation (PDC) technique, which is suitable for state feedback control, where a general nonlinear design approach is suggested and validated in on an inverted pendulum system application [22]. In the second step, a cascade control structure (GS-TP-CS) is designed for EACS with the TP-MT controller (TP-MT-C) in the inner control loop and the anti-windup Proportional-Integral-Derivative Gain-Scheduling controller (aw-PID-GS-C) in the outer control loop. The proposed CCSs are validated by simulation results. The comparison of these two types of CSs for EACS system, namely the TP-CS and the TP-GS-CS, is also given in this paper by means of a set of system responses and tracking error responses that can be easily compared.

The paper treats these topics: the mathematical model of the EACS is discussed in Section 2. The design of the two CCSs for EACS is briefly described in Section 3. The simulation results that illustrate the CCS performance and the comparative analysis are presented in Section 4. Section 5 highlights the conclusions.

2. Modelling of the controlled process

The process that is considered in this paper, modeled and next controlled, is an electromagnetic actuator clutch systems as part of a clutch system, illustrated in Fig. 1.

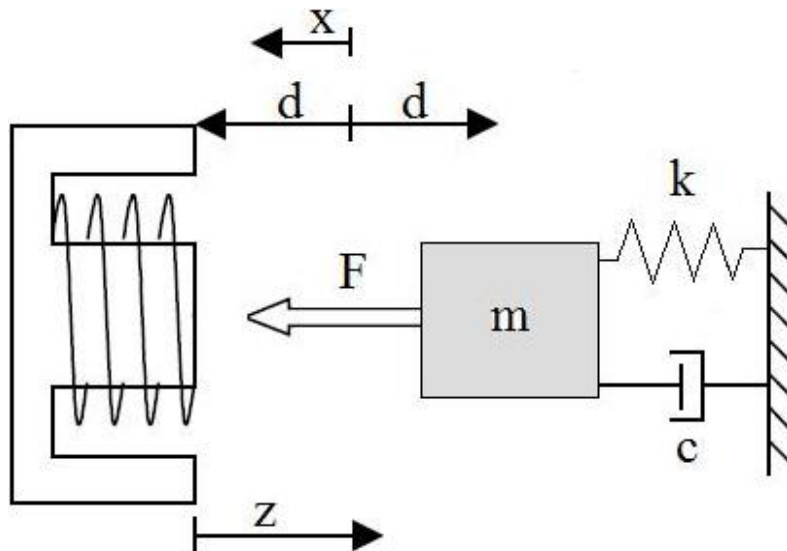


Fig. 1. Schematic structure of the magnetically actuated mass-spring-damper system.

In order to design the proposed CSs, the following nonlinear state-space model of the EACS is used in [23] – [25]:

$$\begin{aligned}\dot{x}_1 &= x_2, \\ \dot{x}_2 &= -\frac{k}{m}x_1 - \frac{c}{m}x_2 + \frac{k_a x_3^2}{m(k_b + d - x_1)^2}, \\ \dot{x}_3 &= -\frac{R(k_b + d - x_1)}{2k_a}x_3 - \frac{1}{k_b + d - x_1}x_2x_3 + \frac{k_b + d - x_1}{2k_a}u, \\ y &= 1000x_1,\end{aligned}\quad (1)$$

where $u \in [0, 12]$ [V] is the control signal, $x_1 \in [0, 0.004]$ [m] is the mass position, x_2 [m/s] is mechanical subsystem's speed and $x_3 = \lambda$ [V·s] is the magnetic flux; y [m] is the measured mass position, $m=1$ [kg] is the mass, $d=0.004$ [m] is the distance between contact position and spring neutral position, $R=1.2$ [Ω] is the resistance, $c=700$ [N s/m] coefficient of the damper, $k=37500$ [N/m] is stiffness of the spring, $k_a=0.5$ is a constant $k_b=0.375$ is a constant, $i \in [0, 10]$ [A] is the current, and $F \in [0, 150]$ [N] is the external force.

3. Control structures design

In the following paragraphs the design of two CSs for the mass position control of EACS is detailed. The first CS is a TP-based control structure (TP-CS) designed for the position control of the EACS using the Parallel Distributed Compensation (PDC) technique. The second CS combines two CSs into a cascade control structure (TP-GS-CS): the TP-MT controller (TP-MT-C) in the inner control loop and the anti-windup Proportional-Integral-Derivative Gain-Scheduling controller (aw-PID-GS-C) in the outer control loop.

3.1. TP-based control structure design

Starting with the quasi-LPV (qLPV) model of EACS a TP-based control structure, namely TP-CS (Fig. 2), is designed, where u is the control signal, u_{TP} is the output of TP-based controller and r is the reference input.

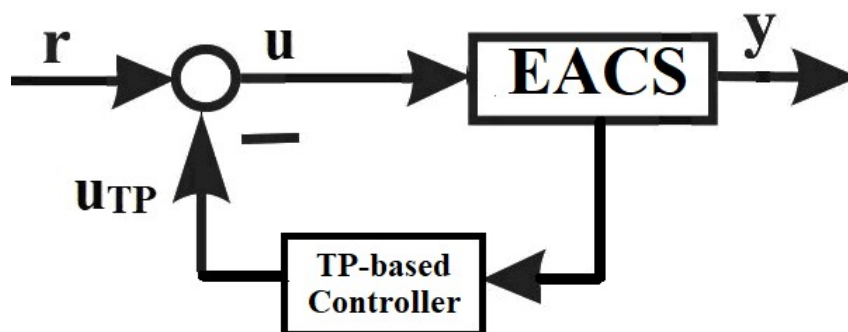


Fig. 2. Block diagram of TP-CS designed for EACS.

The qLPV model of EACS is

$$\begin{aligned} \dot{\mathbf{x}} &= \mathbf{A}(\mathbf{p})\mathbf{x} + \mathbf{b}(\mathbf{p})u, \\ y &= \mathbf{c}^T(\mathbf{p})\mathbf{x}, \end{aligned} \quad (2)$$

where $\mathbf{x} = [x_1 \ x_2 \ x_3]^T \in \mathfrak{R}^{3 \times 1}$ is the process state vector, $\mathbf{p} = p_1 = x_1 \in \mathfrak{R}^{1 \times 1}$ is the parameter vector (which contains the first state variable), y is the controlled output variable and the matrices $\mathbf{A}(\mathbf{p}), \mathbf{b}(\mathbf{p}), \mathbf{c}^T(\mathbf{p})$ are:

$$\mathbf{A}(\mathbf{p}) = \begin{bmatrix} 0 & 1 & 0 \\ a_{21}(\mathbf{p}) & a_{22} & a_{23}(\mathbf{p}) \\ a_{31}(\mathbf{p}) & a_{32}(\mathbf{p}) & a_{33}(\mathbf{p}) \end{bmatrix}, \mathbf{b}(\mathbf{p}) = \begin{bmatrix} 0 \\ 0 \\ k_b/(2k_a) \end{bmatrix}, \mathbf{c}^T(\mathbf{p}) = [1000 \ 0 \ 0], \quad (3)$$

$$\mathbf{A}(\mathbf{p}) \in \mathfrak{R}^{3 \times 3}, \mathbf{b}(\mathbf{p}) \in \mathfrak{R}^{3 \times 1}, \mathbf{c}^T(\mathbf{p}) \in \mathfrak{R}^{1 \times 3}$$

where T indicates matrix transposition and the elements of the matrices are:

$$\begin{aligned} a_{21} &= -\frac{k}{m} + \frac{2k_a x_3^2}{m(k_b + d - \mathbf{p})^3}, a_{22} = -\frac{c}{m}, a_{23} = \frac{2k_a x_3}{m(k_b + d - \mathbf{p})^2}, \\ a_{31} &= \frac{Rx_3 - u}{2k_a} - \frac{x_2 x_3}{(k_b + d - \mathbf{p})^2}, a_{32} = -\frac{x_3}{k_b + d - \mathbf{p}}, a_{33} = -\frac{x_2}{k_b + d - \mathbf{p}} - \frac{R(k_b + d - \mathbf{p})}{2k_a}. \end{aligned} \quad (4)$$

Using the notation: $\mathbf{S}(\mathbf{p}) = [\mathbf{A}(\mathbf{p}) \ \mathbf{b}(\mathbf{p})] \in \mathfrak{R}^{3 \times 4}$, the qLPV model is obtained as follows:

$$\begin{aligned} \dot{\mathbf{x}} &= \mathbf{S}(\mathbf{p})[\mathbf{x}^T \ u]^T, \\ y &= \mathbf{c}^T(\mathbf{p})\mathbf{x}, \end{aligned} \quad (5)$$

which aggregates a set of LTI models in terms of this process model:

$$\dot{\mathbf{x}} = \mathbf{S}(\mathbf{p}(t)) \underset{n=1}{\otimes}^N \mathbf{w}_n(\mathbf{p}_n) [\mathbf{x}^T \ u]^T = \sum_{i=1}^I w_{1,i}(p_1) \mathbf{S}_i [\mathbf{x}^T \ u]^T, \quad (6)$$

$$y = \mathbf{c}^T(\mathbf{p})\mathbf{x},$$

where $\mathbf{w}_n(\mathbf{p}_n)$ are the weighting functions, N is tensor's dimension, \mathbf{S} is the $(N+2)$ -dimensional coefficient tensor of the model given in (3) and $I=3$ is the number of singular values of the tensor \mathbf{S} obtained after applying HOSVD to \mathbf{S} .

The LTI system matrices are $\mathbf{S}_i = [\mathbf{A}_i \ \mathbf{b}_i]$, where the matrices \mathbf{A}_i and \mathbf{b}_i belong to the state-space system model

$$\begin{aligned} \dot{\mathbf{x}} &= \sum_{i=1}^I w_{1,i}(p_1) (\mathbf{A}_i \mathbf{x} + \mathbf{b}_i u), \\ y &= \mathbf{c}^T(\mathbf{p})\mathbf{x}, \end{aligned} \quad (7)$$

as a particular model of the general ones given in [22], and the matrix $\mathbf{c}^T(\mathbf{p})$ is constant in this application, i.e. $\mathbf{c}^T(\mathbf{p}) = \mathbf{c}^T$.

The asymptotic stabilization of the CS is the first control objective. It is achieved by the PDC technique, which determines one LTI state feedback gain matrix for each LTI vertex system of the convex TP model. The asymptotic stability of the closed-loop control system is equivalent to the existence of $\mathbf{X} = \mathbf{P}^{-1} > 0$ (where \mathbf{P}

is a positive definite matrix) and \mathbf{M}_i that satisfy the following Linear Matrix Inequalities (LMIs) [8]:

$$\begin{aligned} & -\mathbf{X}\mathbf{A}_i^T - \mathbf{A}_i\mathbf{X} + \mathbf{M}_i^T\mathbf{X}_i^T + \mathbf{b}_i\mathbf{M}_i > 0, \\ & -\mathbf{X}\mathbf{A}_i^T - \mathbf{A}_i\mathbf{X} - \mathbf{X}\mathbf{A}_s^T - \mathbf{A}_s\mathbf{X} + \mathbf{M}_s^T\mathbf{b}_i^T \\ & + \mathbf{b}_i\mathbf{M}_s + \mathbf{M}_i^T\mathbf{B}_s^T + \mathbf{B}_s\mathbf{M}_i \geq 0, \quad i=1\dots3 \end{aligned} \quad (8)$$

in the Lyapunov-type formulation. However, Popov-type problem setting and expression of inequality-type stability conditions [26] can also be formulated and derived. The LTI state feedback gain matrices \mathbf{F}_i are then computed as:

$$\mathbf{F}_i = \mathbf{M}_i\mathbf{X}^{-1}. \quad (9)$$

Finally, the application of PDC technique to EACS results in the following state feedback control law:

$$u = r - u_{TP} = r - \left[\sum_{i=1}^3 w_{i,i}(p_1)\mathbf{F}_i \right] \mathbf{x}. \quad (10)$$

where u_{TP} is produced by the TP-based controller.

The second control objective is to constrain the control value. Let us assume that $\|\mathbf{x}(0)\|_2 \leq \phi$, where the initial state vector $\mathbf{x}(0)$ is unknown, but the upper bound ϕ is known. The constraint $|u| \leq \mu$ is enforced at all time moments if the following LMIs are satisfied [8]:

$$\phi^2\mathbf{I} \leq \mathbf{X}, \quad (11)$$

$$\begin{pmatrix} \mathbf{X} & \mathbf{M}_i^T \\ \mathbf{M}_i & \mu^2\mathbf{I} \end{pmatrix} \geq 0, \quad (12)$$

which guarantee that the second objective is achieved.

In this paper $\phi = 5.32 > 0$ and $\mu = 100$. The matrices \mathbf{X} and \mathbf{M}_i are computed using the YalmipR2015 solver. The solutions are next substituted in (9) leading to the following values of LTI state feedback gain matrices:

$$\begin{aligned} \mathbf{F}_1 &= [0.0093 \quad -0.0012 \quad -0.0003], \\ \mathbf{F}_2 &= [-0.0005 \quad -0.0009 \quad 0.00017], \\ \mathbf{F}_3 &= [-1.214 \quad -1.1749 \quad -1.36] \end{aligned} \quad (13)$$

3.2. The cascade control structure design

Due to the process nonlinearities, an aw-PID-GS-CS, Fig. 3, is employed in order to ensure the improvement of the CS performance such as the zero steady-state control error (or the tracking error), the alleviation of overshoot, and the bumpless switching between controllers [27].

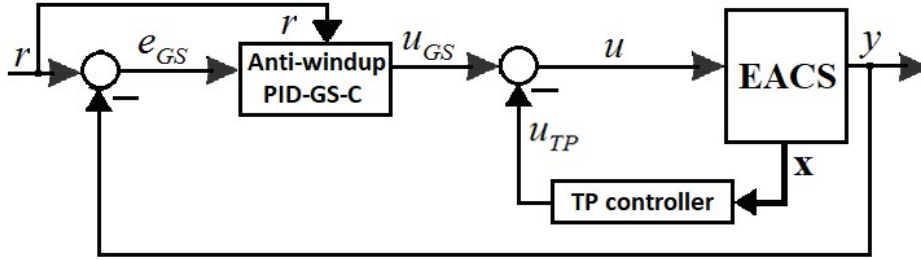


Fig. 3. Block diagram of CCS.

The design of this CS starts with a simple least-squares-based experimental approximation of the inner control loop (TP-CS) resulting in four third-order benchmark-type closed-loop transfer functions (t.f.s):

$$H_p^{(j)}(s) = \frac{k_p^{(j)}}{(1+T_1^{(j)}s)(1+T_2^{(j)}s)(1+T_3^{(j)}s)}, j = \overline{1,4}, \quad (14)$$

where $k_p^{(1)} = 0.34$, $k_p^{(2)} = 0.42$, $k_p^{(3)} = 0.61$, $k_p^{(4)} = 0.96$ are the inner control loop gains, $T_1^{(1)} = 0.065$, $T_1^{(2)} = 0.078$, $T_1^{(3)} = 0.085$, $T_1^{(4)} = 0.1$ are the large time constants, $T_2^{(1)} = 0.0156$, $T_2^{(2)} = 0.0162$, $T_2^{(3)} = 0.0165$, $T_2^{(4)} = 0.0168$, $T_3^{(1)} = T_3^{(2)} = T_3^{(3)} = T_3^{(4)} = 0.00166$ are the small time constants.

Using pole-zero cancellation four aw-PID-Cs were developed for the TP-CS with the t.f.s extended with a first-order lag filter:

$$H_C^{(j)}(s) = \frac{k_c^{(j)}}{s} (1+T_{cl}^{(j)}s)(1+T_{c2}^{(j)}s) \frac{1}{1+T_{fd}^{(j)}s}, j = \overline{1,4}, \quad (15)$$

with the tuning parameters $k_c^{(j)}$, $T_{cl}^{(j)}$, $T_{c2}^{(j)}$ and $T_{fd}^{(j)}$ obtained by the Modulus Optimum method, which is similar to the Symmetrical Optimum one and its extended version [28]:

$$k_c^{(j)} = \frac{1}{2 \cdot k_p^{(j)} \cdot T_3^{(j)}}, T_{cl}^{(j)} = T_1^{(j)}, T_{c2}^{(j)} = T_2^{(j)}, T_{fd}^{(j)} = 0.1 \cdot T_1^{(j)}, j = \overline{1,4}. \quad (16)$$

The output produced by the aw-PID-GS controller, u_{GS} , is expressed as:

$$u_{GS} = k_p(t)e_{GS}(t) + k_i(t) \int_0^t e_{GS}(\tau) d\tau + k_d(t) \dot{e}_{GS}(t), \quad (17)$$

with the tuning parameters, k_p , k_i and k_d obtained as follows [29]:

$$\begin{aligned} k_p(t) &= k_{p\max} - (k_{p\max} - k_{p\min}) \exp[-(\alpha(t) | e(t) |)], \\ k_d(t) &= k_{d\max} - (k_{d\max} - k_{d\min}) \exp[-(\alpha(t) | e(t) |)], \\ k_i(t) &= (1 - \alpha(t)) k_{i\max}. \end{aligned} \quad (18)$$

where the parameters $k_{p\max}$, $k_{p\min}$, $k_{d\max}$, $k_{d\min}$, $k_{i\max}$ and $k_{i\min}$ are computed using the tuning parameters of the aw-PID-Cs:

$$\begin{aligned}
 k_{p_{\max}} &= \max[k_c^{(j)} \cdot (T_{c1}^{(j)} + T_{c2}^{(j)})], k_{p_{\min}} = \min[k_c^{(j)} \cdot (T_{c1}^{(j)} + T_{c2}^{(j)})], j = \overline{1,4} \\
 k_{d_{\max}} &= \max(k_c^{(j)} \cdot T_{c1}^{(j)} \cdot T_{c2}^{(j)}), k_{d_{\min}} = \min(k_c^{(j)} \cdot T_{c1}^{(j)} \cdot T_{c2}^{(j)}), j = \overline{1,4} \\
 k_{i_{\max}} &= \max(k_c^{(j)}), k_{i_{\min}} = \min(k_c^{(j)}), j = \overline{1,4}.
 \end{aligned}
 \tag{19}$$

and the parameter $0 \leq \alpha(t) \leq 1$ is chosen to ensure a smooth and continuous variation of the switching between aw-PID-Cs [29]:

$$\alpha(t) = \tanh(\eta\beta(t)) = \frac{\exp(2\eta\beta(t)) - 1}{\exp(2\eta\beta(t)) + 1},
 \tag{20}$$

The parameter η determines the rate at which $\alpha(t)$ changes between 0 and 1, and it is set to ensure a certain dynamics of the variation of $\alpha(t)$. The parameter $\beta(t)$ is set in terms of [30]:

$$\beta(t) = \begin{cases} 1, & |e(t)| > \xi, \\ 0, & |e(t)| < \xi, \end{cases}, \quad \xi = 0.9 \cdot r(t).
 \tag{21}$$

To design of aw-PID-GS-CS, the following conditions can be taken into account [29]: when $|e(t)|$ is large, $k_{p_{\max}}$ and $k_{i_{\min}}$ are activated to produce a large control signal and to reduce the undesirable oscillation and overshoot; the parameters, $k_{p_{\max}}$ and $k_{i_{\min}}$ are activated to obtain a small value of $|e(t)|$ and to reduce the high overshoot.

Finally, the control law u of the CCS is computed by combining output variable of TP-based controller, u_{TP} in (10), with the output variable of the aw-PID-GS controller, u_{GS} :

$$u = u_{GS} - u_{TP} = k_p(t)e_{GS}(t) + k_i(t) \int_0^t e_{GS}(\tau) d\tau + k_d(t)\dot{e}_{GS}(t) - \left[\sum_{i=1}^3 w_{1,i}(p_1) \mathbf{F}_i \right] \mathbf{x}.
 \tag{22}$$

4. Simulation results

This section is dedicated to test the CSs discussed and designed in the previous sections, namely the TP-CS and the GS-TP-CS.

The following testing scenario was considered and performed: a staircase reference input acting for 10 s was applied to the nonlinear EACS and the corresponding outputs versus time for both the designed CSs are illustrated in Fig. 4. Zero initial conditions were used in the digital simulations that led to the simulation results presented in this paper.

The aw-PID-GS-C parameters $k_{p_{\max}} = 71.4$, $k_{p_{\min}} = 36$, $k_{d_{\max}} = 0.53$, $k_{d_{\min}} = 0.32$, $k_{i_{\max}} = 886$ and $k_{i_{\min}} = 314$ are determined using (21) and several experiments have been conducted such that to get the best values in the context of a compromise to tradeoff and overshoot.

In order to highlight the dynamic and steady-state performance of the two CSs designed in the previous sections, two performance indices were recorded and computed, namely the tracking error and the mean square error.

The tracking errors were computed as the difference between the reference input and the measured output of the TP-CS, y_{TP} , and as the difference between the reference input and the measured output of the GS-TP-CS, y_{GS-TP} .

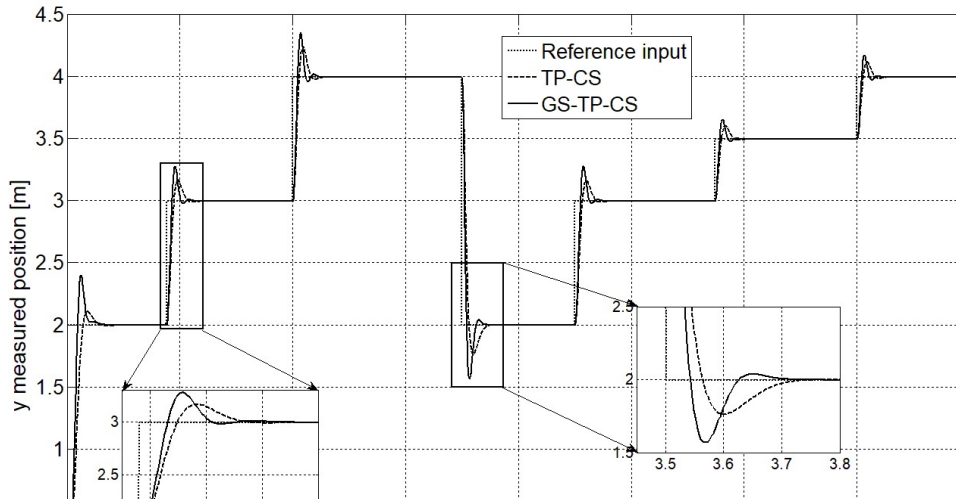


Fig. 4. Mass position vs time in the proposed CS with staircase reference signal.

The mean square error (MSE) was then computed for all the two CS as:

$$\text{MSE}_C = \frac{1}{N} \sum_{t_d=1}^N (r(t_d) - y_C(t_d))^2, \quad (23)$$

where the subscript $C \in \{\text{TP-CS}, \text{GS-TP-CS}\}$ indicates the designed CS, $r(t_d)$ is the reference at time moment $t_d = 1 \dots N$, and $N=3334$ is the number of samples for staircase change inputs. The results are: $\text{MSE}_{\text{TP-CS}} = 0.0876$ and $\text{MSE}_{\text{GS-TP-CS}} = 0.0746$.

The simulation results prove that both the designed CSs ensure zero steady state control error and the reference is well tracked.

In case of GS-TP-CS the overshoot as a measure of the tracking error is bigger than in case of TP-CS, but the rise time and settling time are smaller. The smallest value of MSE was obtained in case of GS-TP-CS.



Fig. 5. Tracking error vs time in the proposed CS with staircase reference signal.

5. Conclusion

The paper presented the design and the validation of two CSs using a combination of Tensor Product (TP)-based model transformation and gain-scheduling control for mass position control of EACS, namely TP-CS and GS-TP-CS. The GS-TP-CS consists of an inner TP-based control loop and an outer gain-scheduling controller-based control loop.

The two CS structures were tested on a time horizon of 10 s applying a staircase type reference input. The simulation results prove that zero steady-state control error was ensured in case of both CSs. The best performance in terms of rise time, settling time and MSE is achieved in case of GS-TP-CS. However, a fair comparison should account for the complexity of the CS structures, and the cascade one is more complicated, hence it is normal to exhibit better over the non-cascade one. In addition, the number of tuning parameters and the simplicity of design procedure are important factors in the comparison.

Future research will be focused on the development of hybrid control techniques including TPBMT fuzzy control and gain-scheduling TPBMT nonlinear control using specific features of several nonlinear techniques outlined in [31]–[36] and various applications [37]–[42], aiming the further CS performance improvement.

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