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High Order Disturbance Observer-Based Control For Suspension Active Magnetic Bearing System

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Abstract: *This paper presents a high order observer based on the sliding mode controller for a suspension active magnetic bearing system (SAMB). The sliding mode surface is constructed by a proportional-integral-derivative (PID) surface. Due to unable to construct the disturbance mathematical model, the measured disturbance are direct feedback and filtered by a proportional-integralgain will be revealed by an observer based on input and output signals. In order to improve the output performance, the lump of the uncertainty is estimated by the sliding mode value, these values are feed forward through a low-pass-filter gain. The output results are achieved by Matlab Simulink. The paper is organized as (i) introduction briefly present about the sliding mode and the suspension active magnetic bearing system. (ii) The suspension active magnetic bearing system model is built. (iii) The proposed method are constructed to the system are presented. (iv) The achieved results are given out. (v) Conclusion will be given in this part.*

Keywords: *Suspension active magnetic bearing (SAMB), sliding mode control (SMC), high order disturbance observer (HODOB), low-pass-filter, proportional –integral filter gain.*

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I. Introduction

This paper is aimed to present the high order disturbance performance by apply it to high nonlinear system as a suspension of the magnetic bearing system. In order to imply a high order disturbance observer for nonlinear, the system output and control input are known [1]. The concept of the HODOB in frequency domain was initially revelled by [2-3] proposed High order mismatched disturbance compensation for motion control systems via a continuous dynamic sliding-mode approach. [4]Proposedhigh order disturbance observer design for linear and nonlinear systems. In order to achieve the desired goals, firstly the sliding mode control (SMC) is mentioned. The SMC was initially developed in mide-1950s, and it was also presented by [5]. Recent few decades the sliding mode control has been received many attention of the scientists. It is a robustness controller, SMC has against ability with output disturbances. The SMC is suitable to construct the controller for high unstable system. Sliding mode control use the difference of the input and output signal to construct the sliding surface. The system state is forced to converge and stabilize on the one pre-determined surface. The velocity and steady-state is depended on how the sliding surface was chosen. There are switching control value and equivalent control value are implemented to converge and stabilize the system state on pre-defined surface, respectively [6]. The chattering values are appeared by high frequency operation of the switching control value. This paper proposes the sliding surface is built by proportional-integral-derivative type. A composed optimization of the proposed method is the lump of the uncertainty is bounded and filtered. The convergence rate of the given output is significantly improved. The controller is applied to control the suspension active magnetic bearing system (SAMB). The SAMB offers many practical applications such as, turbine engines, helicopter and tiltrotor, fly-wheel energy and storage devices, bearingless motor, and vacuum pump, etc.

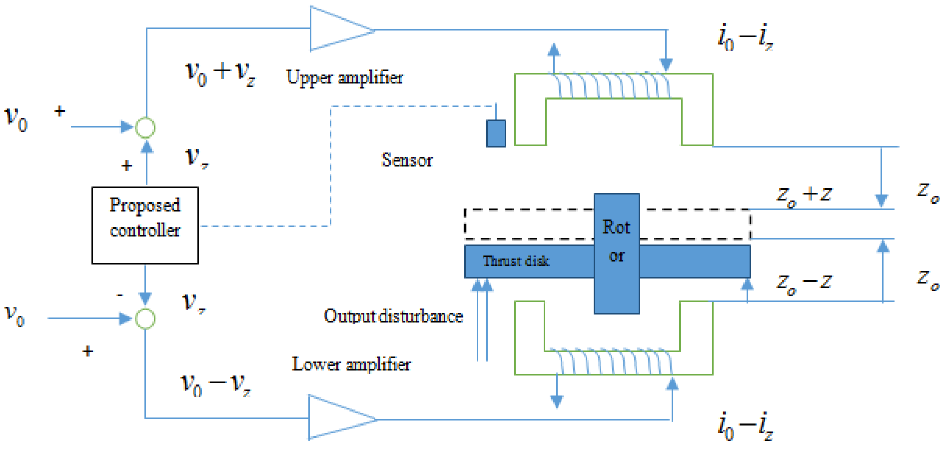


Figure. 1 The structure of the suspension active magnetic bearing system

The Figure. 1 shows the structure of the active magnetic bearing system, there are included rotor is located inside the bearing, and thrust disk is embedded on the rotor. The embedded thrust disk is used to define the rotor position. The original rotor position is refer to z_0, and pre-defined position is $z_0 + z$, and $z_0 - z$ respect to the upper and lower magnet coils, respectively. The SAMB mathematical model is given by the next section.

II. Mathematical Modelling Of The SAMB System

According to the Figure. 1, when the coils are supplied the currents the magnetic field surrounds the coils is presented as follows

$\oint Hds = i$ (1)

where H is magnetic field, i is the supplied current of the magnet coil, and ds is part of the space where the magnetic flux go through, then the Eq. (1) is rewri

$$H = \frac{i}{2\pi r} \qquad (2)$$

[7-11].This device is referred from our system [7], it is shown in figure below.

where $\vec{B} = \mu \vec{H}$

The Lorentz force acting on an electric charge Q is presented as

$\vec{f} = Q(\vec{v} * \vec{B} + \vec{E})$ (3)

where f is magnetic force is generated from the electrical charge $\vec{Q}$ and the $\vec{Q}$ is moving with the velocity called $\vec{v}$ in the magnetic fux density, and $\vec{E}$ is electric field.

Because the electric field $\vec{E} << \vec{v} * \vec{B}$, so the Lorentz force is approximated as

$\vec{f} = Q(\vec{v} * \vec{B})$ (4)

The mathematical model of the SAMB is built from the literature [8]as

$$F_1 = \frac{B_+^2 S}{2\mu_0} \quad (5)$$

$$F_2 = \frac{B_-^2 S}{2\mu_0} \quad (6)$$

where B_+, and B_- are the upper component and the lower component of magnetic density vector. The magnetic force of the SAMB is modelled as

$F_z = F_1 - F_2 = K(\frac{(i_0 + i_z)^2}{(z_0 - z)^2} - \frac{(i_0 - i_z)^2}{(z_0 + z)^2})$ (7)

Using Taylor expansion of the Eq. (7) will leads to

$F_z = k_p z + k_i i_z$ (8)

With k_p, k_i are the amplification factors of the rotor position and the magnet coil currents, respectively.Where

$$k_p = \left.\frac{\partial F_z(z, i_z)}{\partial z_z}\right|_{z=0, i_z=0} = (\frac{2k(i_0 + i_z)^2}{(z_0 - z)^3} - \frac{-2k(i_0 + i_z)^2}{(z_0 - z)^3}) = 4k\frac{i^2}{z^3}$$

$$k_i = \left.\frac{\partial F_z(z, i_z)}{\partial i_z}\right|_{z=0, i_z=0} = (\frac{2k(i_0 + i_z)}{(z_0 - z)^2} - \frac{-2k(i_0 + i_z)}{(z_0 - z)^2}) = 4k\frac{i}{z^2}.$$

Follows the Newton II law:

$$m\ddot{z} = F_{z(t)} - f_{dt_{z(t)}}$$

where m is the mass of the inside rotor, F_z is the Lorentz force, and $f_{dt_{z(t)}}$ is the unexpected outputdisturbance.The Eq. (8) can be written as

$$m\ddot{z}(t) = -c\dot{z} + k_p.z(t) + k_i.i_{z(t)} - f_{dt_{z(t)}} \quad (9)$$

or

$$\ddot{z}(t) = \frac{1}{m}(-c\dot{z} + k_p.z(t) + k_i.i_{z(t)} - f_{dt_{z(t)}}) \quad (10)$$

denotes, $A = -c / m, B = k_p / m, C = k_i / m, \gamma = 1 / m.$

Then the Eq. (10) can be re-written as:

$\ddot{z}(t) = A\dot{z}(t) + Bz(t) + Ci_z(t) - \gamma.f_{dx(t)}$ (11) or,

$\ddot{z}(t) = A\dot{z}(t) + Bz(t) + Ci_z(t) + d$ (12)

where the output disturbance term are $d = -\gamma.f_{dz(t)}$.

The model of the SAMB is presented by a nonlinear model. Subsequently, the paper will show the controller for controlling SAMB system. The proposed method employed PID type to construct the sliding mode surface. Due to unable to directly estimate the output disturbance, the paper used the input control signal and output signal to model the disturbance mathematical model.

III. Proposed approach

- Sliding mode controller.

The sliding mode control has several advantage as low sensitivity to the system, and robustness against to the expected external disturbances [12]. The sliding mode control are included an equivalent control value, and a switching control value. This paper propose the sliding surface are

$s(t) = k_d \cdot \dot{e}(t) + k_p \cdot e(t) + k_i \cdot \int_0^t e(\tau) \cdot d\tau$ (13)

refer $e(t) = z_r - z_m$ where z_r is the reference distance value, and z_m is the measured distance. k_p, k_i, k_d, are positive value and it should be chosen such that the real parts of the roots then sliding surface is satisfied Lyapunov law as

$\dot{V}(t) = \dot{s}(t) \cdot s(t) < 0$ (14)

Combining Eq. (12) and Eq. (13), the current is calculated as

$$I_{cref}(t) = \frac{1}{k_d \cdot C}\Big[\ddot{z}_r(t) - \big(A \cdot \dot{z}_m(t) + B \cdot z_m(t) + d\big) + k_p \cdot \dot{e}(t)$$

$$+k_i \cdot e(t) + k \cdot sign\left(s\left(t\right)\right)\Big] \qquad (15)$$

The chattering value is appeared in the switching control part, in order to reduce these value, this research used saturation function to replace the sign function. Where

$sat\left(s\right) = sign\left(s\right) \cdot \min\left\{1, \left|s\right|\right\}$ or

$$sat\left(s\right) = \begin{cases} 1 \quad if \quad s > \varepsilon \\ \frac{1}{\varepsilon} s \quad if \quad s \in \left[-\varepsilon; \varepsilon\right] \\ -1 \quad if \quad s < -\varepsilon \end{cases} (16)$$

Then Eq. (15) can be written as

$$I_{cref}\left(t\right) = \frac{1}{k_d \cdot C}\Big[\ddot{z}_r\left(t\right) - \left(A \cdot \dot{z}_m\left(t\right) + B \cdot z_m\left(t\right) + d\right) + k_p \cdot \dot{e}(t)$$

$$+k_i \cdot e(t) + k \cdot sat\left(s\left(t\right)\right)\Big] (17)$$

The term of $k \cdot sat\left(s\left(t\right)\right)$ is called switching control part, there are chattering appeared. Due to unable directly to estimate the uncertainty of the system [13], this paper proposes the uncertainty estimator based on the sliding mode surface. Firstly, the system from Eq. (9) need to be modified as

$$\dot{x} = (G_{1n} + \Delta G_1) \cdot x + (G_{2n} + \Delta G_2) \cdot i + G_3 d \ (18)$$

or

$$\dot{x} = (G_{1n}) \cdot x + (G_{2n}) \cdot i + L + G_3 d \ (19)$$

where lump of the uncertainty is $L = \Delta G_1 \cdot x + \Delta G_2 \cdot i.$ basically, the derivative of the sliding mode surface is

$$\dot{s}_{in} = -k \cdot sat(s_{in}) \qquad (20)$$

Or

$$\dot{s}_{in} = -L - u_L(\mathrm{t}) - k \cdot sat(s_{in}) \ (21)$$

where $L = -u_L(\mathrm{t}).$ The lump of uncertainty can be observed by a filter. A suggestion of strictly proper low-pass-filter is $G_f(\mathrm{s}),$ the lump of the uncertainty can be approximated by

$\hat{L} = L^{-1}\left\{G_f(\mathrm{s})\right\} \times \left(-u_L(\mathrm{t}) - k \cdot sat(s_{in}) - \dot{s}_{in}\right)$(22)where $L^{-1}\left\{\bullet\right\}$ is the inverse Laplace transfer function, and $\times$ sign is convolution operator. Substituting $\hat{L} = -u_L(\mathrm{t})$ to Eq. (22) yields

$$u_L(\mathrm{t}) = L^{-1}\left\{\frac{G_f(\mathrm{s})}{1 - G_f(\mathrm{s})}\right\} \times \left(k \cdot sat(s_{in}) + \dot{s}_{in}\right) \qquad (23)$$

The control signal is improve as

$$V_c(\mathrm{t}) = V_{0c} + L^{-1}\left\{\frac{G_f(\mathrm{s})}{1 - G_f(\mathrm{s})}\right\} \times \left(k \cdot sat(s_{in}) + \dot{s}_{in}\right) (24)$$

Then choose $G_f(\mathrm{s}) = \frac{1}{1 + Ts}$ as a filter, then

$$V_c(\mathrm{t}) = V_{SMC} + \left(\frac{k}{T} \cdot sat(s) + k \cdot \int_0^t sat(s(\mathrm{t}))\,\mathrm{dt}\right) (25)$$

The stability of the uncertainty estimation as following

$$s \cdot s(\mathrm{s}) = -\mathrm{k} \cdot s(\mathrm{s}) - \left[1 - G_f(\mathrm{s})\right] \cdot L(\mathrm{s}) \ (26)$$

or

$$s\left(s\right) = k_d \cdot s \cdot e(s) + k_p \cdot e(s) + k_i \cdot \frac{1}{s} \cdot e(s) \ (27)$$

Substituting Eq. (27) to Eq. (26) yields

$$e(\mathrm{s}) = \frac{s}{s+k} \cdot \frac{G_f(\mathrm{s}) - 1}{k_d \cdot \mathrm{s}^2 + \mathrm{k}_p \cdot \mathrm{s} + \mathrm{k}_i} \cdot L(\mathrm{s}) \qquad (28)$$

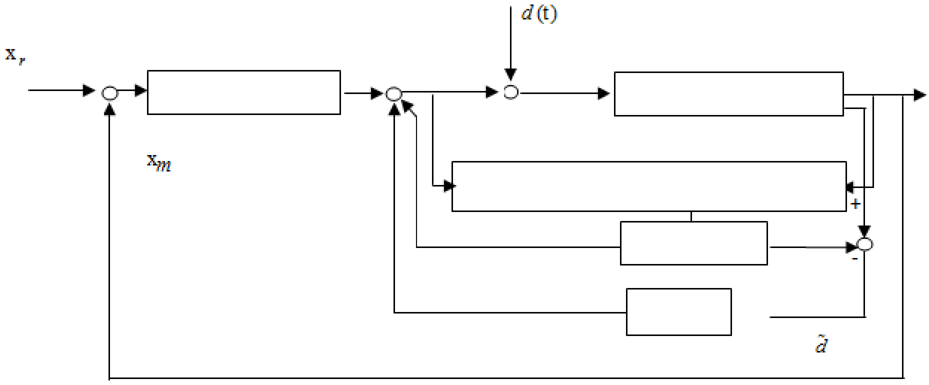
On the time domain the distance tracking error are

$$\lim_{t\to\infty}(e(t)) = \lim_{s\to 0} s\cdot\frac{s}{s+k}\cdot\frac{G_f(s)-1}{k_d\cdot s^2+k_p\cdot s+k_i}\cdot L(s) \quad (29)$$

$$= 0$$

High order disturbance observer design.

Estimation of the unknown disturbance has been interest topic in recent decades [12]. Many article published the performance of this topic research [14-17]. The disturbance observer (DOB) were introduced early in 1980s[18]. Since the day the DOB was appeared, many article applied it successfully. In process control [18], in mechatronic [19-21]. Our proposed a disturbance observer based system states for linear system [11], there are proportional-integral filter gains are applied to force system state converge on the stable region by choosing suitable parameter. In this paper the high order disturbance observer will be appeared, the structure of the controller are the disturbance will be filtered by a proportional-integral gain, and the disturbance error is approximated by a low-pass-filter gain. The system model from Eq. (19) will be modify by



As mentioned above, the lump of uncertainty is included in the term of the disturbance error. This paper proposes the filter gain of the disturbance by separate the disturbance and lump of the uncertainty as

$$u_{\tilde{d}} = u_{lump} + u_{d-\hat{d}} \quad (31)$$

$$\begin{cases} \dot{y} = F^+\cdot(G_{1n})\cdot x + F^+\cdot(G_{2n})\cdot i + F^+\cdot L + \hat{d} \\ \hat{d} = diag\{r\}\cdot\{F^+\cdot x + z\} \end{cases} \quad (32)$$

The generalized disturbance error are

$$\tilde{d} = d - \hat{d} \quad (33)$$

In the presence of a lump of the uncertainty in Eq. (30), it may be shown as

$$\tilde{d} = F^+\cdot L + d - \hat{d} \quad (34)$$

Taking derivative of Eq. (33) leads to

$$\dot{\tilde{d}}(t) = \frac{d}{dt}\cdot\left(F^+\cdot L(t) + d(t)\right) - \dot{\hat{d}}(t) \quad (35)$$

$$F^+\cdot\dot{x} = F^+\cdot(G_{1n})\cdot x + F^+\cdot(G_{2n})\cdot i + F^+\cdot L + F^+\cdot G_3\cdot d$$

$$= F^+\cdot(G_{1n})\cdot x + F^+\cdot(G_{2n})\cdot i + F^+\cdot L + d \quad (30)$$

where $F^+\cdot G_3 = I$, the disturbance observer system are

The Eq. (34) shows that the $\dot{\hat{d}}(t) \to \frac{d}{dt}\cdot\left(F^+\cdot L(t) + d(t)\right)$ depending on the filter gain. The $\tilde{d}(t)$ is depending on the matched uncertainty term. By choosing suitable $diag\{r\}$ to reality application of the proposed method to system as where u_{lump}, and $u_{d-\hat{d}}$ are the feedback value of uncertainty and disturbance value, the output of filter gain are

$$V_C(\mathrm{t}) = V_{SMC} + \left(\frac{k}{T} \cdot sat(s) + k \cdot \int_0^t sat(s(\mathrm{t}))\,\mathrm{dt} \right) + k_P \cdot \hat{d} + k_I \int_0^t \hat{d}(\tau) \cdot d\tau$$

$$+ L^{-1}\left\{ G_f(s) \right\} \times \left(d - \hat{d} \right) \quad (36)$$

By choosing a suitable low-pass-filter gain for disturbance error part the lump of disturbance error are

$$u_{d-\hat{d}} = \frac{1}{T} \cdot e^{-\frac{1}{T} \cdot t} \quad (37)$$

$u_{d-\hat{d}} \to 0$ when $t \to \infty$ with a suitable time constant of the filter gain. When the observer of disturbance is constructed a proportional-integral the disturbance are

$$\hat{d} = k_P \cdot diag\{r\} \cdot \left\{ F^+ \cdot x + z \right\} + k_I \cdot \int_0^t diag\{r\} \cdot \left\{ F^+ \cdot x + z \right\} \cdot dt \quad (38)$$

Taking derivative of the disturbance leads to

$$\dot{\hat{d}} = k_P \cdot diag\{r\} \cdot \left\{ F^+ \cdot \dot{x} + \dot{z} \right\} + k_I \cdot diag\{r\} \cdot \left\{ F^+ \cdot x + z \right\} \quad (39)$$

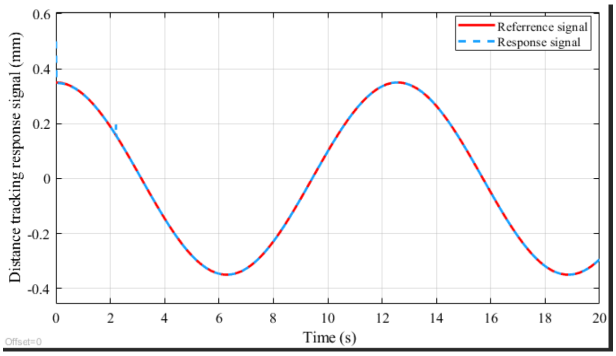
In order to $\hat{d} \to d$ the k_I, and k_P are reasonable choose. Then, the performance of the proposed method are below.

IV. An illustrative Example

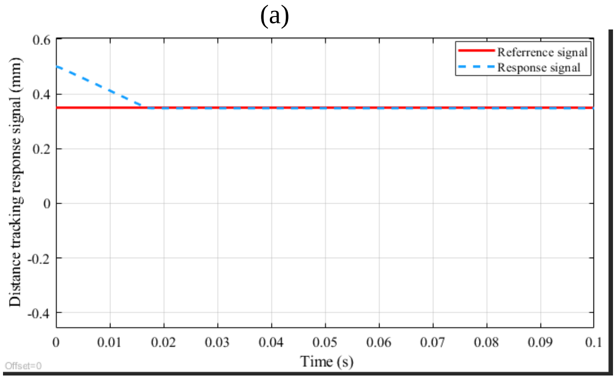
In order to emphasize the powerful of the proposed technics, this part describes the given output of the proposed method under two cases

Case 1: The results without using the disturbance error feedback.

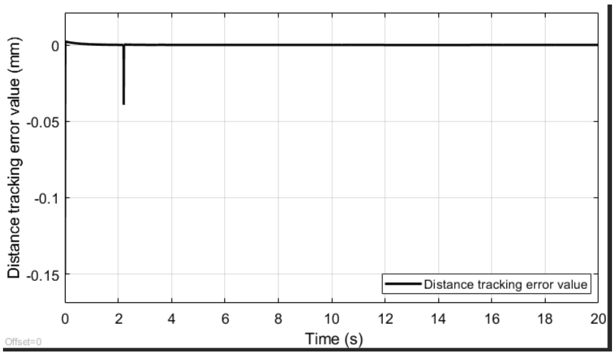
In this case, the proposed controller is sliding mode control and high order disturbance observer with a proportional-integral filter gain. The given out results are good track with the sinusoidal signal. The given output signal of this case are presented as



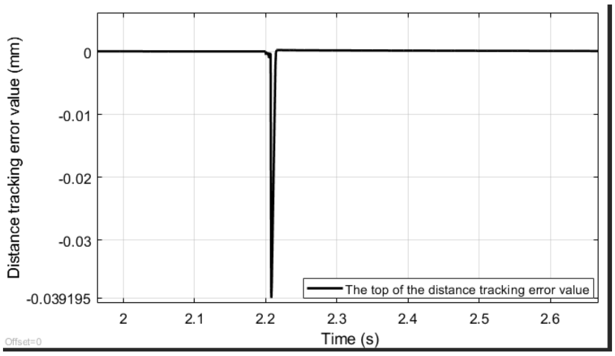
(a)



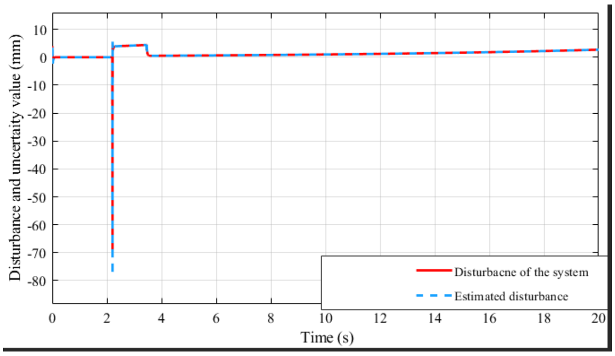
(b)



(c)



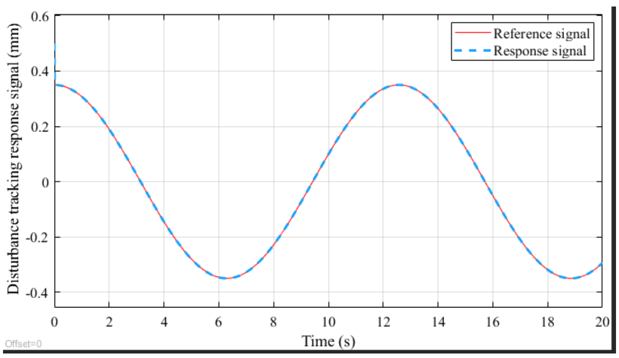
(d)



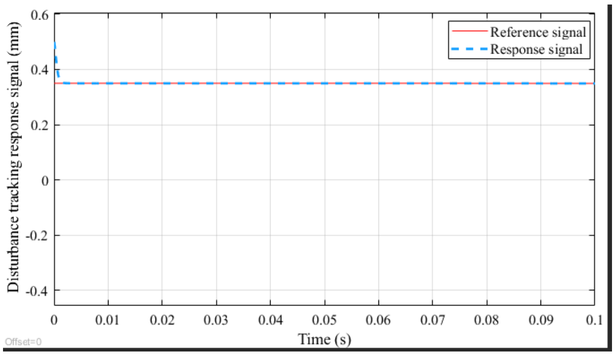
(e)

Figure. 3 The proposed control method without using the disturbance error feedback: (a) The distance response signal in first 20s, (b) The distance response signal at first 0.1s, (c) The distance tracking error value, (d) The top value of the distance tracking error value, (e) The disturbance response signal.

Case 2: The results while the full disturbance and disturbance error are used.

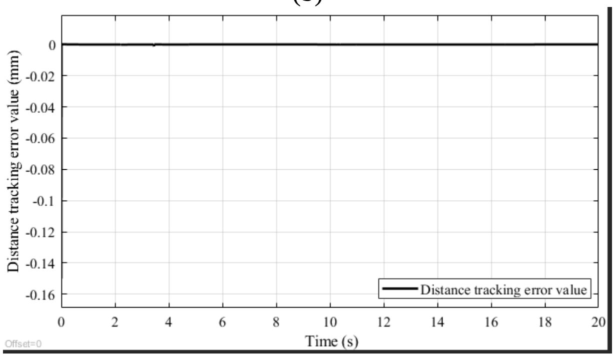


(a)



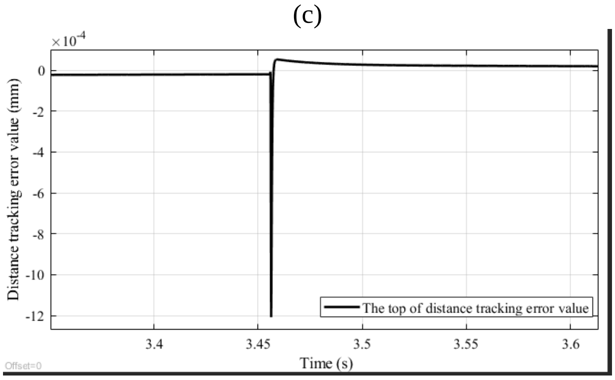
Reference signal
Response signal
Disturbance tracking response signal (mm)
0.6
0.4
0.2
0
-0.2
-0.4
0
0.01
0.02
0.03
0.04
0.05
0.06
0.07
0.08
0.09
0.1
Time (s)

(b)



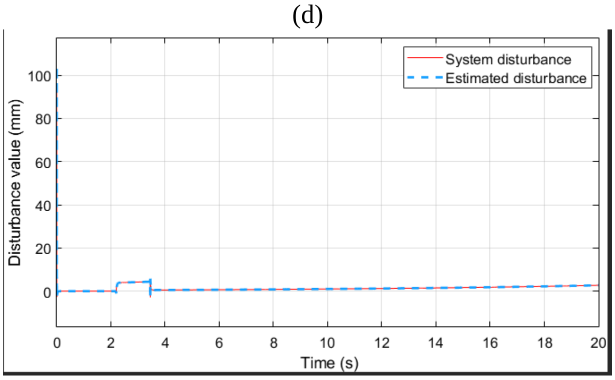
Distance tracking error value (mm)
0
-0.02
-0.04
-0.06
-0.08
-0.1
-0.12
-0.14
-0.16
Distance tracking error value
0
2
4
6
8
10
12
14
16
18
20
Time (s)

(c)



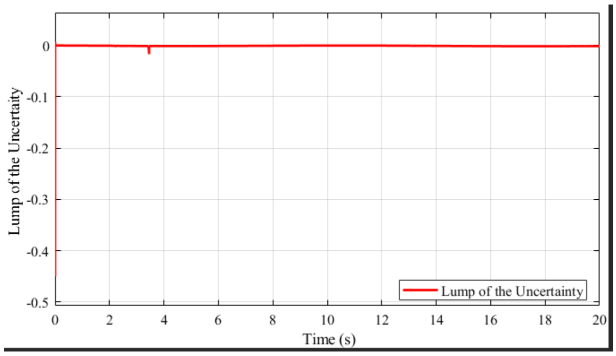
$\times 10^{-4}$
Distance tracking error value (mm)
0
-2
-4
-6
-8
-10
-12
The top of distance tracking error value
3.4
3.45
3.5
3.55
3.6
Time (s)

(d)



System disturbance
Estimated disturbance
Disturbance value (mm)
100
80
60
40
20
0
0
2
4
6
8
10
12
14
16
18
20
Time (s)

(e)



(f)

Figure. 4 The proposed control method without using the disturbance error feedback: (a) The distance response signal in first 20s, (b) The distance response signal at first 0.1s, (c) The distance tracking error value, (d) The top value of the distance tracking error value, (e) The disturbance response signal, (f) The lump of Uncertainty value.

In comparison of the two cases give out briefly understanding knowledge that, the proposed controller is good at tracking flexible signal, it has able to against the uncertainty, and good estimated disturbance is given out. The improvement of the disturbance error feedback through two terms are significantly performed at Figure. (4). the settling time is improved from $16.725(ms)$ to $2.715(ms)$, distance tracking error value has been improved from $2.12 \cdot 10^{-5}(mm)$ to $1.86 \cdot 10^{-5}(mm)$, and the top of the disturbance error tracking value also been reduced from $0.03915(mm)$ to $1.2 \cdot 10^{-3}(mm)$ by applying the disturbance feedback method.

V. Conclusion

The proposed controller is good at tracking the flexible reference input signal. In the case 1 of the paper, the disturbance observer with a proportional filter gain given a good performance, but not very well smoothly. It has been improve by equipped a disturbance error feedback control. There are included a term of estimated disturbance and a term of uncertainty value, these values are filter by a low-pass-filter to achieve the desired goals. This result is better much more than our previous published paper [7]. Every given output of this proposed method has been improved significantly.

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Appendix

The system parameters of suspension AMB system, and controller parameters are given by tables below.

Table. 1 System parameters.

Parameter	Description	Value	Unit
m	Mass of rotor	2.565	kg
k_{ai}	Current stiffness of Electromagneticforce of Thrustdisk AMB	40	N/A
k_{ap}	Position stiffness of electromagnetic force of thrust disk AMB	25200	N/m
z_0	Nominal air gap where thrust diskis centred	1	mm
r	Thrust disk mass	0.38	kg
c	Damping constant	0.001	$N\cdot s/m$
v_0	Reference voltage	1.4	V
i_0	Amplifier range	0.5	A/V

Table. 2 The controller parameters

Parameter	Description	Value
k_p	Proportional coefficient	3000
k_i	Integral coefficient	5000
k_d	Derivative coefficient	1
k	Positive gain	30
T	Low-pass-filter time	0.001
k'_p	PI filter gain	0.9
k'_i	PI filter gain	0.01
$diag(r)$	Observer gain	$diag(1,500)$

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