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A Switching Controller for Piezoelectric Microactuators in Dual-Stage Actuator Hard Disk Drives

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Abstract—Dual-stage actuators (DSAs) with piezoelectric (PZT) microactuators can provide faster track seeking and more accurate track following in hard disk drives (HDDs) than conventional single-stage actuators. However, one of the control challenges of the DSA systems is the PZT actuator saturation. To avoid saturating the PZT actuator, most of the existing methods either carefully design the PZT controllers with small gains or limit the amplitudes of the reference commands. This typically leads to performance conservation because the fast dynamics of the PZT is not fully utilized. Unlike the existing methods, this paper studies a switching controller that optimizes a quadratic performance cost function involving the PZT saturation model explicitly. The controller can not only guarantee the system stability in the presence of saturation but also improve the tracking speed by efficiently allocating the control efforts. Simulation results show the effectiveness of the switching controller with faster disturbance rejection.

I. INTRODUCTION

In recent years, dual-stage actuators (DSAs) with piezoelectric (PZT) microactuators have been adopted in commercial high-performance hard disk drives (HDDs) to meet the ever increasing demands for high-capacity and fast data rate. The mechanism of the DSA structure [1] is simple and of low cost, which makes it feasible in mass production. However, the control design of the DSA systems imposes more challenges than conventional single-stage voice coil motor (VCM) systems mainly because of the PZT actuation redundancy versus single controlled output [the so-called position error signal (PES)]. In other words, for a given desired trajectory, alternative inputs to the two actuators are not unique. Thus, a proper control strategy is required for control allocation in response to external inputs. Otherwise, the two actuators may fight each other and deteriorate the performance instead. One of the control strategies that are popularly used is the decoupled master-slave control [2], where the PZT actuator is aimed at following the position error of the VCM actuator. The other is the PQ design method [3], where the VCM actuator is allocated to response to low frequency components of the system input while the PZT actuator to high frequency components due to its faster dynamics. Moreover, according to the classification of the control tasks in HDDs, control design for track following and settling can be also found in [3]–[5]. In [6], a decoupled track-seeking controller using a three-step design approach is developed to enable high-speed one-track seeking and short-span track-seeking for a dual-stage servo system. Further, short and long-span seeking controls are incorporated in a single control scheme with fast settling time [7], [8].

Although the PZT microactuator has a faster dynamics in response to external inputs, its drawback is the very limited stroke relative to the VCM actuator. In the point of view of control system, this can be regarded as an actuator saturation problem. It is shown that if the controller cannot handle the PZT saturation properly, the system output will have significant oscillations or even lead to potential instability [9]. Therefore, to avoid saturating the PZT actuator, most of the existing controllers attempt to limit the PZT controller gain such that the PZT actuator works in its linear region only. In such methods, the PZT saturation model is not taken into account in the control design process and thus leading to performance conservations such as reduced servo bandwidth of the PZT actuator. In [9], the authors developed a modified decoupled master slave dual-stage control scheme by simply using a nonlinear PZT model with saturation nonlinearity in its observer and showed its improvement on the stability against the saturation. In this paper, we will discuss a switching control scheme [10], [11] for the PZT actuator which can not only guarantee the system stability in the presence of saturation but also improve the tracking performance by efficiently allocating the control efforts.

In our design, we explicitly model the PZT actuator as a saturated actuator whose control problem is then casted as a linear quadratic control problem with input saturation. The solution of the problem eventually leads to a switching controller. Unlike the anti-windup compensator [12] that uses ad-hoc methods to detune the controller with little theoretical guarantee on stability, the switching control scheme can not only guarantee the stability in the presence of saturation, but also optimize a quadratic performance function through properly over-saturating the controller that leads to desired fast convergence of the tracking error.

II. SYSTEM MODEL AND CONTROL STRUCTURE

Fig. 1 shows a typical DSA with a push-pull PZT microactuator in hard disk drives. It consists of a VCM actuator as the primary stage and a PZT actuator as the secondary stage. The PZT is located between the suspension and the E-block, which is moved by the VCM. The two actuators are respectively driven through a PZT amplifier and a VCM driver. The VCM driver has a voltage input limit of ±3.5 V. The PZT actuator has a stroke limit of ±0.5 μm and the PZT amplifier has a voltage input limit of ±1.5 V.
the two actuators are negligible, the DSA plant model of the DSA can be regarded as a decoupled dual-input and single-output system. Moreover, we assume notch filters have been cascaded to the VCM and PZT actuators to actively damp their resonances, respectively (see our work in [8] for details). As such, a control-oriented model of the DSA system can be described by Fig. 2, where the state-space equations of VCM and PZT actuator is given by

\[
\begin{align*}
\Sigma_1 : & \dot{x}_1 = A_1 x_1 + B_1 \sigma(u_1), \quad x_1(0) = 0 \\
\Sigma_2 : & \dot{x}_2 = A_2 x_2 + B_2 \sigma(u_2), \quad x_2(0) = 0 \\
y & = y_1 + y_2 = C_1 x_1 + C_2 x_2
\end{align*}
\]  

(1)

where the state \( x_1 = [y_1 \; \dot{y}_1]^T \), \( x_2 = [y_2 \; \dot{y}_2]^T \),

\[
A_1 = \begin{bmatrix} 0 & 1 \\ 0 & 0 \end{bmatrix}, \quad B_1 = \begin{bmatrix} 0 \\ b_1 \end{bmatrix}, \quad C_1 = \begin{bmatrix} 1 & 0 \end{bmatrix},
\]

\[
A_2 = \begin{bmatrix} 0 & a_1 \\ a_2 & 0 \end{bmatrix}, \quad B_2 = \begin{bmatrix} 0 \\ b_2 \end{bmatrix}, \quad C_2 = \begin{bmatrix} 1 & 0 \end{bmatrix},
\]

and the saturation function \( \sigma(u_i) \) (i = 1, 2) is defined as

\[
\sigma(u_i) = \text{sgn}(u_i) \min \{ \bar{u}_i, |u_i| \}
\]

(2)

where \( \bar{u}_i \) is the saturation level of the \( i \)th control input. The DSA model parameters in (2)-(2) are given by

\[
b_1 = 1.7 \times 10^6, \quad a_1 = -10^6, \quad a_2 = -3.1 \times 10^4, \\
b_2 = 4.3 \times 10^6, \quad \bar{u}_1 = 3 \text{ V}, \quad \bar{u}_2 = 1.25 \text{ V}.
\]

As mentioned earlier, the control strategies for the coordination of the two actuators are not unique as the DSA system is a dual-input single-output system. Here, we use the decoupled master-slave control structure because it offers the benefit that the overall stability of the DSA loop can be guaranteed by independently stabilized VCM loop and PZT loop. This control structure is shown in Fig. 3, where the system input is the disturbance \( d \). The control objective is to regulate the position error signal \( y_e \) swiftly and as small as possible.

From Fig. 3, it is easy to find that the VCM position error \( y_{1e} \) equals to

\[
y_{1e} = y_2 - y_e = d - y_1 
\]

(3)

and when combined with the VCM feedback controller \( C_1 \), we have

\[
y_{1e} = \frac{1}{1 + G_1 C_1} d.
\]

(4)

We can see that the VCM loop is decoupled from the PZT loop and \( C_1 \) can be designed using any conventional methods and is thus not detailed here because it is not the primary goal of this paper. Typically, due to the relatively slow dynamic characteristics of the VCM, we can only expect \( |y_{1e}| \leq \epsilon, \quad (\epsilon > 0) \) for any persistent bounded disturbance input. It follows that the PZT control effort \( u_2 \) should be designed to compensate for the residual VCM position error, i.e., driving \( y_2 = y_{1e} \). If this can be achieved, it immediately implies that \( y_e = y_2 - y_{1e} = 0 \).

To formulate the control problem of the PZT actuator, define \( u_1 = y_e \) and \( w_1 = y_2 - y_{1e} \), then

\[
w_2 = a_1 w_1 + a_2 w_2 + b_2 (\sigma(u_2) - f),
\]

(5)

where \( f = (\bar{y}_{1e} - a_2 \bar{y}_{1e} - a_1 y_{1e})/b_2 \). We observe that the coupling signal \( f(t) \) is dominated by the VCM position error dynamics. Intuitively, it is only when the VCM drives \( f(t) \) to converge to a small region within the PZT’s control limit that the PZT can make some meaningful control effort. In view of this, we assume that the VCM control loop is well designed such that \( |f(t)| \leq f_0, \forall t \geq 0 \) and \( f_0 < u_2 \). Under this circumstance, we can introduce a feedforward input in \( u_2 \) as follows:

\[
u_2 = u + f
\]

(6)
such that \( f(t) \) is compensated. Then, we can formulate the PZT control design as a regulation control problem with input saturation and the system model (5) can be rewritten as follows:

\[
\dot{x} = Ax + B\sigma(u), \quad x(0) = x_0
\]

where \( x = [w_1 \ w_2]^T, \ A = A_2, \ B = B_2, \) and \( \sigma(\cdot) \) with saturation level equal to 1. In the next section, we will discuss a switching control design for the PZT control input \( u \).

### III. SWITCHING CONTROL DESIGN

In this section, we first introduce the fundamental theory of a linear quadratic saturation control design. Next, a switching controller is developed based on the saturation control design method, which offers performance improvement. Finally, we present the design result of the switching controller with application to the PZT actuator.

#### A. Saturation Control Design

Consider the system in (7), we first introduce the following quadratic cost function

\[
J(x_0, u) = \int_0^{\infty} (x^T Q x + r \sigma(u)^2) dt \tag{8}
\]

for some \( Q = Q^T > 0 \) and \( r > 0 \) with \((A, B)\) being controllable. Ideally, we aim to seek an optimal linear state feedback

\[
u = K x
\]

for each given initial state \( x_0 \) such that \( J(x_0, u) \) is minimized. It is well-known that if the control is not saturated, the optimal solution to \( K \) is given by

\[
K = -r^{-1} B^T P_0 \tag{10}
\]

where \( P_0 = P_0^T > 0 \) is the solution to the following Ricatti equation

\[
A^T P_0 + P_0 A + Q - r^{-1} P_0 BB^T P_0 = 0 \tag{11}
\]

Moreover, the minimal cost is given by \( x_0^T P_0 x_0 \).

However, in the presence of saturation, the optimal \( K \) is difficult to give. To overcome this difficulty, we parameterize the controller by using an optimal sector bound [10]. More specifically, define the level of over-saturation \( \rho \geq 0 \) such that the control input \( u \) is restricted to be

\[
|u| \leq 1 + \rho \tag{12}
\]

It is easy to verify that for any \( u \) constrained by (12), \( \sigma(u) \) lies in the following sector bound

\[
\sigma(u) = \rho_1 u + \delta(u) \tag{13}
\]

\[
|\delta(u)| \leq \rho_2 u, \ \forall |u| \leq 1 + \rho \tag{14}
\]

where

\[
\rho_1 = \frac{2 + \rho}{2(1 + \rho)}, \quad \rho_2 = \frac{\rho}{2(1 + \rho)} \tag{15}
\]

Here, \( \rho_1 \) is the optimal value so that \( \delta(u) \) has the smallest sector to bound the nonlinearity cause by the saturation.

Now, we give some analysis on the design of a control gain \( K \) to minimize the worst-case cost for all \( \delta(\cdot) \) satisfying the sector bound (14). For a given \( \rho > 0 \), consider the Lyapunov function candidate

\[
V(x) = x^T P_\rho x, \quad P_\rho = P_\rho^T > 0 \tag{16}
\]

and define

\[
\Omega_\rho = A^T P_\rho + P_\rho A + Q - r^{-1} P_\rho BB^T P_\rho, \quad u^* = -r^{-1} B^T P_\rho x \tag{17}
\]

Given any initial state \( x_0 \) and any \( \delta(\cdot) \) satisfying (14), it is easy to verify that

\[
J(x_0, u, T) = \int_0^T (x^T Q x + r \sigma(u)^2) dt = V(x_0) - V(x(T)) + \int_0^T (\dot{V}(x) + r^2 \sigma(u)^2) dt \leq V(x_0) + \int_0^T g(x, u, \delta(u)) dt \tag{19}
\]

where

\[
g(x, u, \delta(u)) = x^T \Omega_\rho x + r (\rho_1 u + \delta(u) - u^*)^2 \tag{19}
\]

This implies that if \( g(x, u, \delta(u)) \leq 0 \) for all \( x \in \mathbb{R}^n \) and \( \delta(\cdot) \) satisfying (14), then

\[
J(x_0, u) \leq V(x_0) \tag{20}
\]

From the analysis above, we formulate the following relaxed optimal control problem:

**Pl:** For a given \( \rho \geq 0 \), design \( P_\rho \) and \( u \) to minimize \( V(x_0) \) subject to \( g(x, u, \delta(u)) \leq 0 \) for all \( x \in \mathbb{R}^n \) and \( \delta(\cdot) \) satisfying (14). Moreover, determine the largest invariant set \( X_\rho \) characterized by an ellipsoid of the form

\[
X_\rho = \{ x : x^T P_\rho x \leq \rho_3^2 \}, \quad \mu_\rho > 0 \tag{21}
\]

such that if \( x_0 \in X_\rho, x(t) \in X_\rho \) and \( |u(t)| \leq 1 + \rho \) for all \( t \geq 0 \), we have \( J(x_0, u) \leq V(x_0) \).

The solution to the above problem is given by the following Theorem:

**Theorem 1** [10]: Consider the system in (7) and the cost function in (8). For a given level of over-saturation \( \rho \geq 0 \), suppose the equation

\[
A^T P_\rho + P_\rho A + Q - r^{-1} (1 - \rho_0^2) P_\rho BB^T P_\rho = 0 \tag{22}
\]

where

\[
\rho_0 = \frac{\rho_2}{\rho_1} = \frac{\rho}{2 + \rho} \tag{23}
\]

has a solution \( P_\rho = P_\rho^T > 0 \). Then the optimal feedback control law \( K_\rho \) for the relaxed optimal control problem **Pl** is given by

\[
K_\rho = -r^{-1} B^T P_\rho \tag{24}
\]

and the associated invariant set \( X_\rho \) is bounded by

\[
\mu_\rho = r \frac{\rho}{(1 - \rho_0) \sqrt{B^T P_\rho B}} \tag{25}
\]
Remark 1: If \( \rho = 0 \), the Riccati equation (22) and the control law (24) recover the results in (11) and (10) for optimal control without saturation. The associated invariant set is given by
\[
X_0 = \{x : x^T P_0 x \leq \mu_0^2\}, \quad \mu_0 = \frac{r}{\sqrt{B^T P_0 B}}
\tag{26}
\]

Remark 2: Taking \( \rho \to \infty \) (or equivalently, \( \rho_0 \to 1 \)) and solving for \( P_\rho \) in (22) gives the largest invariant set as
\[
X_\infty = \{x : x^T P_\rho x < \mu_\infty^2\}
\]
\[
\mu_\infty = \frac{r}{(1 - \rho_0)\sqrt{B^T P_\rho B}}, \quad \rho_0 \to 1
\tag{27}
\]
Note that the solvability of \( P_\rho \) for any \( \rho > 0 \) is guaranteed by the controllability of \((A, B)\) and positive definiteness of \( Q \).

Remark 3: Despite that the invariant set enlarges when \( \rho \) increases, it can be seen that the upper bound of the performance cost in (20) is as well larger. This implies that the saturated controller can bring a good benefit when \( \rho \) is not close to 0 and not too large. Generally, \( \rho \) can be selected as the minimal one satisfying \( x_0 \in X_\rho \).

1) Property of the control law: The proposed controller in Theorem 1 has two nice properties, i.e., the nesting property of \( X_\rho \) and monotonicity of \( P_\rho \). More specifically, define
\[
S_\rho = (1 - \rho_0)P_\rho
\tag{28}
\]
We can rewrite the Riccati equation in (22) as
\[
A^T S_\rho + S_\rho A + (1 - \rho_0)Q - r^{-1}(1 + \rho_0)S_\rho BB^T S_\rho = 0
\tag{29}
\]
and the invariant set can be expressed as
\[
X_\rho = \{x : x^T S_\rho x \leq \frac{\rho^2}{B^T S_\rho B}\}
\tag{30}
\]

Lemma 1 [10]: The solution \( S_\rho \) to (29) is monotonically decreasing in \( \rho > 0 \), i.e., for a sufficiently small \( \epsilon > 0 \), \( S_\rho > S_{\rho+\epsilon} \), if \( 0 \leq \rho < \rho + \epsilon \). Consequently, \( X_\rho \) are nested in the following sense:
\[
X_\rho \subset X_{\rho+\epsilon}, \quad \forall 0 \leq \rho < \rho + \epsilon
\tag{31}
\]
Moreover, the solution \( P_\rho \) to the Riccati equation in (22) is monotonically increasing in \( \rho > 0 \). That is,
\[
P_\rho < P_{\rho+\epsilon}, \quad \forall 0 \leq \rho < \rho + \epsilon
\tag{32}
\]

B. Switching Control

Thanks to the nesting property of \( X_\rho \) and monotonicity of \( P_\rho \), we can apply Theorem 1 to design a sequence of control gains \( K_i \), based on which a nested switching control can be developed to improve the performance. More specifically, choose a sequence of over-saturation bounds \( 0 = \rho_0 < \rho_1 < \cdots < \rho_N \) and solve the corresponding Lyapunov matrices \( P_i \), invariant sets \( X_i \) and controller gains \( K_i, i = 0, 1, \cdots, N \). We then construct the nested switching control law by selecting the control gain \( K_i \) when \( x \in X_i \) and \( x \notin X_{i+1} \) (unless \( i = 0 \)). The following result shows the advantage of the nested switching control in the performance improvement.

Lemma 2: Suppose the switching controller above is applied to the system in (7) with \( x_0 \in X_N \). Let \( t_i \) be the time instance \( K_i \) is switched on, \( i = 0, 1, \cdots, N \). Particularly, \( t_N = 0 \). Then the cost of the switching control is bounded by
\[
J(x_0, u) \leq x_0^T P_N x_0 - \sum_{i=0}^{N-1} x_i^T(t_i)(P_{i+1} - P_i)x(t_i)< x_0^T P_N x_0.
\tag{33}
\]

Proof: According to the proof of Theorem 1 [10], we have
\[
x^T Q x + r \sigma(K_i x)^2 \leq -\frac{d}{dt} V_i(x) = -\frac{d}{dt}(x^T P_i x)
\tag{34}
\]
along the trajectory of \( x(t) \), where \( t \in [t_{i+1}, t_i] \). Following the monotonicity of \( P_i \) and integrating the inequality above yields (33).

From the theorem above, we can clearly see the advantage of the switching control by means of the negative term in (33) that decreases the cost gradually. In what follows, we will discuss the application of this switching control scheme to the design of the PZT controller for improved tracking performance.

C. Controller Design for PZT Actuator

Our main purpose here is to use the switching controller for the PZT to expedite the convergence of the position error \( (y_\rho) \) at the presence of the output disturbance \( d \). Particularly, we suppose that the disturbance is a shock wave with an amplitude larger than the PZT’s stroke. This scenario generally occurs when the HDDs are used in mobile environment. If the servo controller cannot compensate for the shock disturbance quickly within a small time frame, the read/write head has to wait for a few more revolutions until the head is regulated to the desired sector. This obviously decreases the data throughput. To improve this situation, it is intuitive to inject the maximum control input to the PZT to expedite the convergence of the position error \( y_\rho \) for the PZT to achieve the fastest acceleration at the initial stage. Then the control input should be gradually decreased (by applying a relatively small control gain \( K_i \) ) to achieve appropriate robustness when the position error approaches zero. Such a control strategy would impose some conditions on \( Q, r \) and \( \rho_i \). The switching controller design results for the PZT are given as follows.

First, we design the control gain \( K_0 \) (i.e., \( \rho_0 = 0 \)). Under this circumstance, the position error \( y_\rho \) is close to zero and the control input \( u \) is not saturated. It is straightforward to verify that the closed-loop system is a linear system and can be expressed as
\[
\dot{x} = (A - BB^T P_0)x
\tag{35}
\]
Clearly, we can select \( Q \) and \( r \) (hence corresponding to a unique solution of \( P_0 \)) such that the dominated poles of \( A - BB^T P_0 \) lead to desired characteristics. More specifically, we first set \( r = 1 \) without loss of generality since the performance cost can be normalized as \( J/r \). Then, let the
desired poles for the closed-loop system matrix $A - BB^T \bar{P}_0$ be $(-\zeta \pm j\sqrt{1 - \zeta^2}) \omega$, where $\zeta$ represents the damping ratio and $\omega$ the natural frequency. We then parameterize $Q$ as

$$Q = \begin{bmatrix} q_1 & 0 \\ 0 & q_2 \end{bmatrix}$$

(36)

with

$$q_1 = \frac{1}{b_2^2} (\omega^4 - a_1^2)$$

(37)

$$q_2 = \frac{1}{b_2^2} (4 \zeta^2 \omega^2 - 2 \omega^2 - 2a_1 - a_2^2).$$

(38)

Clearly, to guarantee $Q > 0$ requires $\omega$ and $\zeta$ satisfying the conditions

$$\omega > \sqrt{|a_1|}$$

(39)

$$1 \geq \zeta^2 > \frac{(2a_1 + a_2^2)}{4\omega^2} + 0.5.$$  

(40)

Substituting $Q$ and $r = 1$ into (22) solves $P_\rho$ as

$$P_\rho = \begin{bmatrix} p_1 & p_2 & p_3 \\ p_2 & p_3 & p_3 \end{bmatrix}$$

(41)

with

$$p_1 = b_2^2 (1 - \rho_0^2) p_2 p_3 - a_2 p_2 - a_1 p_3$$

(42)

$$p_2 = \frac{a_1 + \sqrt{(1 - \rho_0^2) \omega^4 - \rho_0^2 a_2^2}}{b_2^2 (1 - \rho_0^2)}$$

(43)

$$p_3 = \frac{a_2 + \sqrt{a_2^2 + b_2^2 (1 - \rho_0^2) [(2p_2 + q_2)}{b_2^2 (1 - \rho_0^2)}$$

(44)

Accordingly, the resulting control gain with respect to a given $\rho$ is obtained by

$$K_\rho = \frac{b_2}{p_1} [p_2 \ p_3]$$

(45)

To this end, it is easy to verify that the closed-loop system characteristic polynomial for $\rho = 0$ is given by

$$\Delta_0(s) = |sI - A - BK_0|$$

$$= s^2 + (b_2 p_3 - a_2)s + (b_2^2 p_2 - a_1)$$

$$= s^2 + 2\zeta \omega s + \omega^2$$

(46)

which yields the poles as specified initially.

Based on the above analytic results, we choose $\omega = 2\pi 6000$, $\zeta = 0.8$, which leads to the following controller:

$$K_0 = [-0.9796 \ 0.0001],$$

$$P_0 = \begin{bmatrix} 2.9262 \times 10^{-4} & 2.2781 \times 10^{-9} \\ 2.2781 \times 10^{-9} & 1.5521 \times 10^{-13} \end{bmatrix},$$

$$\mu_0 = 3.4845 \times 10^{-5}.$$ 

Second, we design the saturated controllers $K_1$ by choosing $\rho = 5$. It follows that

$$K_1 = [-1.8278 \ 0.0001],$$

$$P_1 = \begin{bmatrix} 3.0335 \times 10^{-4} & 2.4795 \times 10^{-9} \\ 2.4795 \times 10^{-9} & 1.8370 \times 10^{-13} \end{bmatrix},$$

$$\mu_1 = 3.6066 \times 10^{-4}.$$ 

Finally, combining the above two controllers leads to the switching controller as follows:

- $u = 0$ when $[y_e \ \dot{y}][y_e \ \dot{y}]^T > \mu_1^2$;
- Controller gain $K_1$ is switched on when $[y_e \ \dot{y}][y_e \ \dot{y}]^T \leq \mu_1^2$ and $[y_e \ \dot{y}][y_e \ \dot{y}]^T > \mu_1^2$;
- Controller gain $K_0$ is switched on when $[y_e \ \dot{y}][y_e \ \dot{y}]^T \leq \mu_0^2$.

Note that the control algorithm ensures that there is only one controller active at any time instance.

IV. SIMULATION RESULTS

Simulation is carried out to verify the switching controller. The VCM actuator is simply controlled using a lead-lag controller as follows:

$$C_1(s) = \frac{0.2s^2 + 5236s + 3.298 \times 10^6}{s^2 + 24551s + 1.223 \times 10^8}. $$

(47)

We assume that the disturbance signal is with Fig. 4 and the displacement of $y_2$ can be observed. Then, the simulation results for the designed switching controller are shown in Figs. 5 and 6. We can see that position error of the DSA converges to zero much faster than that of the VCM (see Fig. 6). To clearly see the benefit of the switching scheme, we compare the performance with the non-switching case (i.e., with only the controller $K_1$). The result is shown in Fig. 7, which indicates a relatively slow settling time instance at 0.5 ms achieved by the non-switching controller as compared to the switching case at 0.35 ms shown in Fig. 6. Thus, we can see that although the controller $K_1$ guarantees the stability of the system in the presence of saturation, it cannot improve the performance without switching to $K_0$ that is specifically designed to work around the origin. On the other side, we simulate the result using a proportional controller for the PZT actuator. This is a conventional method [9] by simply ignoring the saturation in the design. Fig. 8 shows that in our specific case when the PZT is saturated, the position error of the DSA achieved by the conventional controller contains significant oscillations and thus results in tedious settling time. Therefore, we conclude that the switching controller can offer the benefit of guaranteed stability through $K_1$ and fast convergence through $K_0$.

V. CONCLUSIONS

This paper developed a switching controller for the PZT actuator in DSA systems for HDDs. The advantage of the
switching control scheme lies in that the actuator saturation nonlinearity is explicitly considered in the design process such that the closed-loop system stability can be guaranteed in the presence of saturation, meanwhile faster convergence is offered through switching the controllers that optimize a quadratic cost function. The simulation results have shown that the new control scheme can provide faster and more accurate disturbance rejection capability. In future work, we will implement this promising controller on a real experimental platform to evaluate its practical performance.

REFERENCES


