Challenging in Hard Disk Drive Servo System

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Manuscript received January 20, 2013
Revised May 9, 2013

ABSTRACT

In digital world today, digital contents keep increasing every instant. The data storage is therefore crucial to keep this huge amount of digital contents. Currently, the world still relies on hard disk drive for data storage although there are many other methods. The hard disk drive is the marvelous piece of nanotechnology. Each bit of data is magnetically stored and read back in such a tiny area, a few nanometers square, and can be randomly accessed. The secret behind this is the very fast and extremely precise movement of mechanical system. The low level servant to make this happened is that the astonishing servomechanism composing of a carefully design mechanic and a huge pack of control algorithms. This report introduces the fundamental of servo system, focusing on control algorithm, in the current hard disk drive technology and the promising researches for the future.

Keyword: *Hard Disk Drive, Servo, Feedback Control*

1 INTRODUCTION

Digital contents have been substantially produced every moment. They include almost all kinds of data ranging from entertainment sector including, movies, music, TV, picture, e-book, to private corporations that keep huge amount of data every years. Publishing and sharing information through the internet including social network also generates the gigantic amount of digital contents. It was reported, in [1], that, in the year 2009, the amount of digital content was around 800,000 petabytes or around 8×10^{20} bytes and, in the year 2011, was around 1.2 million petabytes, or 1.2 zettabytes. Based on that growth rate, it is projected that by the year 2020, the digital data would be around 44 times the amount in 2009. Imagine that how to store these tremendous amount of data? Currently, the most space and cost effective way to store digital data is a hard disk drive (HDD). The current technology yields the capacity, in term of areal density, of HDD around 600 gigabytes per square inch (Gb/in^2). This areal density will be insufficient soon, and the new target, according to Information Storage Industry Consortium (INSIC), for HDD will be an order of a few terabytes per square inch [2]. This extremely high density pushes the challenge in the HDD servo-mechanical system since each bit of data stored at particular location on the media is accessed by mechanical movement.

There is concern in HDD industry that the solid state memory or drive (SSD) will be soon to replace the HDD. The cases from magnetic tape cassette or film camera are still the good examples for technology changing, and it seems to be the case for HDD to be replaced by SSD, since the advent of tablet personal computers that no longer use an HDD, and a lot of advantages of an SSD that contains no mechanical moving part. However, this concern may not be the case in the near future, since the cost per capacity of an SDD cannot compete with an HDD. It was estimated that the amount of about USD 1 trillion would be needed for the SSD industry to replace the HDD [2]. However, this chasing from an SSD pushes an HDD to achieve higher and higher capacity as well as performance with low cost. Inherently, this also puts a lot of challenges to an HDD servomechanism to achieve such high data density.

This report reviews the HDD servo technology focusing on adopted control approaches for extremely precise position control. The paper is organized as follows. The following section gives some overview on how the HDD work. Then mathematical modeling of HDD is presented in section 3. In section 4, we revisit the control methodologies using in HDD servo system including conventional and modern control. The additional compensators reinforcing the disturbance rejection are discussed in section 5. Section 6 discusses some possible problems toward higher data density and promising technology in the near future. Conclusion is drawn in section 7.

2 HOW HDD WORK

The HDD stores data as magnetic patterns on the media made from glass or aluminum coated with thin layers of magnetic material on both sides. The media spins at high speed from 5400 revolutions per minute (RPM) to 15000 RPM. The magnetic patterns on the media are created or written by strong current-induce magnetic field, and read back by a giant magneto resistive (GMR) sensor. The device that contains read/write head (the actuator to generate the magnetic field and GMR sensor) is called a slider. To access the location on the media, mechanic part that carries (by flying) the read/write head is called suspension that provides suitable space (around 6-10 nm) between a media and a slider, also known as fly height. The suspension works as a spring that pushes against the lifting force from high velocity air flow resulting from high speed spinning media underneath the slider. The suspension is connected to an actuator arm, also known as head gimbal assembly (HGA), rotating about a pivot by torque produced by a voice coil motor (VCM) which is similar concept to a loundspeaker.

The magnetic pattern for each data bit is roughly a rectangle lied on the concentric track. The data location, hence, is accessed by moving the slider across the radial direction of the media to a target track, and waiting for a sector on spinning media. Servo-mechanical system is responsible for this data location access. To achieve extremely high capacity, the rectangular shape of magnetic for data bit must be extremely small. Normally, the length of this rectangle along the circumference direction is shorter than the one along the radial direction that defines track pitch or width. This shorter size of rectangle is determined by the ability to detect the changing in magnetic field, known as kilo flux change per inch (KFCI) from a noisy read back channel. The longer one is determined by the ability of servomechanism to access the track along radial direction. The smaller size of this rectangle yields, of course, the higher areal density, and hence higher capacity. Hence, in servo system, the budget is indicated by number of tracks per inch (TPI). For example, the areal density of 600 Gb/in^2 would require TPI around 300 kTPI resulting in 85 nm track pitch.

Qualitative measurement of servo-mechanic system, like other control system, may be separated into transient and steady state. The quality of transient state known as track seeking is measured by settling time that tells how fast the read/write head move from initial track to destination track. The criteria to define steady sate may not be the same as in control text book, e.g., 2% or 5% error. The percentage error used to define the steady state is relative to track width, and is normally different between read and write operation (write operation needs more tighten criteria to prevent data erasure). After the read/write head reach the target track with an acceptable error gap, the read/write head must stay there to perform read or write data. The ability to keep the read/write head at desired track is, in HDD literature, known as track missed registration (TMR). This TMR is measured by the variance of position error signal (PES) in percentage of track width (the expected error should be zero), and normally is presented in 3σ (or the confident of 99.73 per cent), where σ is a standard deviation. For example, 10% TMR for 300 kTPI (85 nm track pitch) means that there is a chance of 0.27 per cent of the head moving away from the track center by more than 8.5 nm. This amazing device is a real nanometer-scale technology. Much detail of HDD operation and servo-mechanism can be found in the books [3], [4].

3 SYSTEM MODELING

3.1 Rigid Body & Resonances

The actuator, VCM, converts the electrical energy (voltage) to mechanical torque driving the arm to move. The conceptual idea could be depicted in Fig. 1, where v, τ, p represent voltage, torque and final position, respectively. The dynamics of the HGA carrying the read/write head to required data location is indeed very complicated.

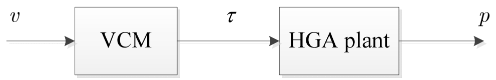

Fig. 1 *Simple concept of HDD plant*

In practice, however, an approximation model is used to design controller. Since HDD servo system is indeed a position control problem that has torque as an input, so the initial candidate model is the double integrators system which is essentially the model of a rigid body HGA plant. The VCM, including power drive, has comparatively very fast dynamics, so it can be modeled as a constant. That is, we can consider that the torque τ changes instantly proportional to input voltage. Hence, the model in frequency domain or transfer function can be described by

$$\frac{P(s)}{V(s)} = \frac{k}{s^2},$$

where $P(s)$, $V(s)$ are the Laplace transform of position and input voltage respectively, and k is a constant depending on many physical factors such as VCM magnet strength and coil resistance, arm length and moment of inertia of HGA, TPI. To further refine the above model to be closed to the real behavior, practically the friction mode will be incorporated, and hence the simple double integrators system can be modified to be of general second order system as follow:

$$\frac{P(s)}{V(s)} = \frac{k\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2}.$$

Normally the natural frequency ω_n and damping factor ζ can be characterized by frequency response testing.

Based on frequency response verification, the above approximate model is normally accurate in low frequency range below 1 kHz. At higher frequency, the vibration, also known as resonance, of mechanical systems will be observed. These resonances may occur at various frequencies from 1.5 kHz to above 60 kHz. The peaks of those resonances may not be too high compare to the gain at DC, but with increasing TPI that requires high bandwidth servo system, they significantly affect the servo system performance. When the read/write head stay on track, and if any resonance mode is exited, the head will be off track and causing high PES. The simple model for resonance can be characterized by general second order system, and then it can be incorporated into above system by series connection as follow:

$$\frac{P(s)}{V(s)} = G(s)$$

$$= \frac{k\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \cdot \prod_{i=1}^{N} \frac{s^2 + 2\zeta_{ni}\omega_{ni}s + \omega_{ni}^2}{s^2 + 2\zeta_{di}\omega_{di}s + \omega_{di}^2}$$

where N is the number of resonance modes to be incorporated into the model. An example of frequency response of the HDD servo-mechanical is shown in Fig. 2, where the solid line represents frequency respond measured from a real drive and the dash line is an approximate model including one resonance mode.

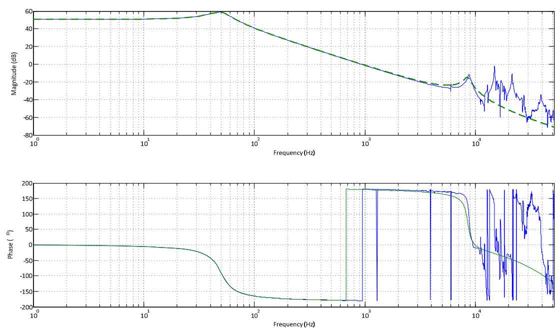

Fig. 2 *Example of HDD plant frequency response.*

3.2 Uncertainties, Disturbances and Noises

Apart from resonances (the major problems in HDD servo-mechanical system) disturbances and noises also cause a lot of problem for a high TPI HDD. Major disturbances can be characterized into two categories namely synchronous and nonsynchronous to spindle speed. For example, if the spindle speed is 5400 RPM, then the first synchronized harmonic is 90 Hz. The synchronous disturbance, known as repeatable run out (RRO) in the HDD literature, is due to imperfect track marks induced from a track writing process that suffers from disturbances during the writing process itself. This causes high order harmonic of an RRO. Low harmonics of an RRO, comparatively high in amplitude, is due to the eccentric track causing by center misalignment during installation of the media.

Nonsynchronous disturbance is normally known as non-repeatable run out (NRRO). These disturbances are from the windage circulating inside the HDD and the vibration induced from outside HDD, for example, loudspeaker, movement from users. The power spectrum of this NRRO spreads across wide range of frequencies.

4 CONVENTIONAL CONTROLLERS

4.1 Seek Controller

During operation, an HDD performs huge amount of track seeking to read and write data in various locations. The seek time or settling time in control literature then significantly contributes to latency or access time. Hence for speed optimization, many HDDs have separated controllers dedicated for transient and steady state response. The track seek controller is generally based on time optimal controller.

To apply the time optimal control method, the HDD mathematical model can be considered as simple double integrators which can be described by the following differential equation

$$\ddot{p} = ku ,$$

where, again, p is actual position output and u is control effort that has limited power, that is $|u| \leq u_{\max}$. Rather considering the actual desired position, the PES that is the different between desired position and current head position is more suitable to be used to develop a control effort. That is, let $e := p_r - p$, where p_r is desired or referent position. Then, from above equation, the error dynamics is

$$\ddot{e}(t) = -ku(t) .$$

Then the objective is to drive this error to zero in finite time T. Then, we have from optimization theory that the maximum control effort will be used for the first half of travel time and then switch to minimum one for the second half of travel time. The final time T then depends on an initial error $e(0)$, that is

$$T = 2\sqrt{\frac{|e(0)|}{ku_{\max}}} .$$

Then the control effort can be described by

$$u(t) = \begin{cases} \operatorname{sgn}(e(0))u_{\max}, & t \in [0, T/2) \\ -\operatorname{sgn}(e(0))u_{\max}, & t \in [0, T/2] \end{cases},$$

where $\operatorname{sgn}(\bullet)$ is a sign function. That is if the initial error is positive or the desired position is greater than the initial one, then accelerate with maximum rate for half of the total travel time, then later decelerate with maximum rate for the rest of the time. The trajectory of velocity and position error can be depicted in Fig. 3. This is indeed an open-loop bang-bang control which is not robust to disturbances and uncertainties. For example, if there is high friction, then the velocity profile will not follow the ideal one and resulting in position error. Another example is that if the initial velocity of the head is not zero, then applying those control effort would end up with non-zero final velocity. Therefore, the closed-loop form with the same concept will be used instead.

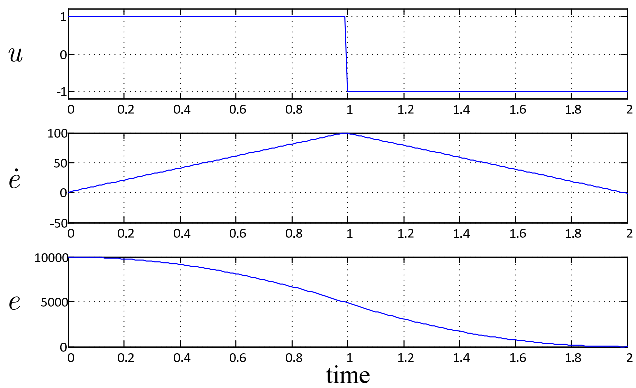

Fig. 3 *Illustration of control effort, velocity and position error trajectory.*

Since the trajectory generated from the open-loop is a time optimal, so we can consider this trajectory as an ideal trajectory that we want. So the close-loop controller will use this velocity trajectory as the referent input. The referent velocity trajectory needs to be a function of error not the time, since the ultimate control objectives are both velocity and error, while the optimal time is considered as the lateral result. Hence, it is better to describe the trajectory of velocity versus error, known as phase plane, as shown in Fig. 4.

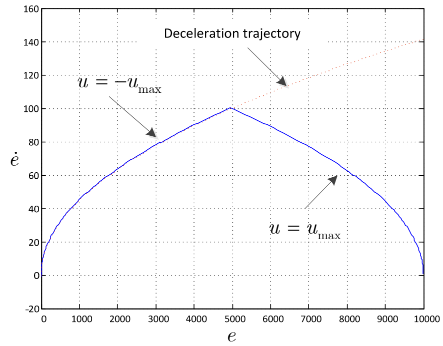

Fig. 4 *Velocity versus position error trajectory.*

From the phase plan in Fig. 4, we can state that if the initial condition on both error and its velocity is under the deceleration trajectory, then drive the system by maximum power and, when the condition of both error and velocity hits the deceleration trajectory, switch to negative maximum power. Thus, to follow this control law, we need to create the deceleration trajectory such that the velocity profile depends on error for all traveling time. The function for this deceleration trajectory can be characterized by

$$v_{ref} = \mathrm{sgn}(e)\sqrt{2ku_{\max}|e|} =: f_t(e),$$

where v_{ref} is referent velocity which is the function of error, i.e., $f_t(e)$. Then the actual velocity will be compared to above reference in order to choose whether positive or negative maximum control effort. This control law can be described by the block diagram in Fig.5

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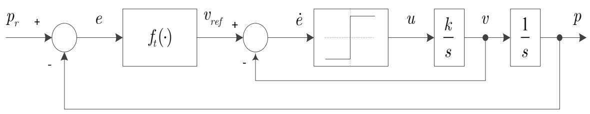


Fig.5 *Closed-loop time optimal controller.*

Note that, this closed-loop time optimal controller needs information on velocity and it can be realized by state estimate.

In practice, the above closed-loop time optimal control needs to be modified to avoid frequently switching when the actual velocity closes to the referent one. The method so called proximate time optimal servomechanism (PTOS), is used instead. The idea for PTOS is simple; that is, when the velocity error $\dot{e}$ closes to zero, then the control effort is proportional to it. This can be realized by simple proportional controller with high gain and saturation as shown in Fig.6.

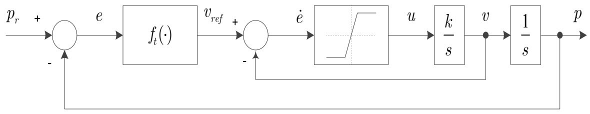


Fig.6 *Approximate time optimal servo-mechanism controller.*

4.2 Track Following Controller

When the head reaches the target track, the track following controller will keep the error variance or TMR as low as possible, so the HDD can perform read/write data on that track correctly. During this track following, disturbances and noises from many sources contribute to the TMR. Therefore, the main purpose of the track following controller is to reject those disturbances. The conventional liner controller is normally used for track following controller.

Now consider the linear system from the block diagram in Fig.7 , where $D(s)$ and $N(s)$ are disturbances and noises respectively.

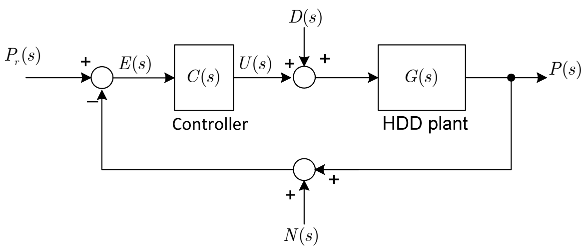
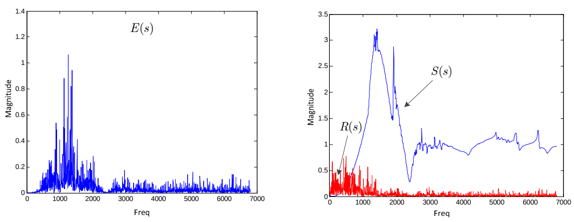


Fig.7 *General linear control system block diagram.*

The main objective for track following controller is to minimize the error variance induced from all inputs, i.e., $P_r(s)$, $D(s)$, $N(s)$. Hence, we will consider the transfer function from those inputs to error; that is

$$E(s) = \frac{1}{1 + C(s)G(s)}\left(P_r(s) - G(s)D(s) - N(s)\right) \\ = S(s)R(s),$$

where $S(s) := \frac{1}{1 + C(s)G(s)}$ is the sensitivity function, and $R(s) := P_r(s) - G(s)D(s) - N(s)$ could be regarded as all sources of disturbance; see Fig.8 for illustrative example. It is obvious from above equation that, to minimize error, we need the gain of open-loop, or $C(s)G(s)$, as high as possible for all range of frequencies to make the sensitivity function $S(s)$ small. Unfortunately, the Bode's integral of sensitivity function theorem states that the integral of sensitivity function for all over frequencies is unity for the stable system with relative degree more than one, or equal to certain constant if the relative degree equals to one. This implies that we cannot achieve small sensitivity for all range of frequency. That is, if some range of frequency yields the sensitivity gain less than one, then there is some other range of frequency such that the sensitivity gain will be greater than one. This constraint is also known as water bed effect; e.g., see Fig.8.



(a) (b)

Fig.8 *Example of power spectrum of error signal (a) and corresponding sensitivity function and TMR source (b).*

The loop shaping method is normally used to design the controller by shaping the open-loop frequency response (hence the sensitivity function) to certain shape. Since disturbances normally have high power spectrum at low frequencies, so we need high gain of open-loop shape to reject these disturbances. Plant dynamic unknown uncertainties, on the other hand, normally contain high frequencies power spectrum, so it is better to keep the open-loop gain low at high frequency. The general idea of loop shaping can be depicted in Fig.9. Hence, with given frequency response of plant $G(s)$, the controller $C(s)$ can be designed to achieve certain frequency response shape of open-loop $C(s)G(s)$ subject to the Bode's integral constrain. The loop shape can be qualitatively characterized by bandwidth, gain margin, phase margin, gain at certain low frequency. The bandwidth in HDD servo system is normally defined at open-loop gain cross over frequency, while gain and phase margin are conventional defined as in system control literature.

The controller, in general, can be separated into two parts to handle low and high frequency ranges separately. The first part, to handle low frequency disturbances, can be the lead/lag compensator or PID controller or any other transfer function that can stabilize the system and provide high gain at DC and low frequency. While at high frequency, the HDD plants suffer a lot of resonances that will limits the bandwidth of the system. To handle these resonances, the narrow band notch filter will be employed to suppress the resonances. Hence, we may write $C(s) = C_c(s)C_n(s)$, where $C_c(s)$, $C_n(s)$ can be regarded as transfer function of compensator and notch filter respectively. The notch filter $C_c(s)$ may composes of many frequencies as necessary to suppress resonances; that is

$$C_n(s) = \prod_i^N k_i \frac{s^2 + 2\zeta_{zi}\omega_{zi} + \omega_{zi}^2}{s^2 + 2\zeta_{pi}\omega_{pi} + \omega_{pi}^2}.$$

The example of loop shaping of HDD is shown in Fig.10 and the corresponding error sensitivity frequency response and Nyquist's plot are shown in Fig.11, in which the gain and phase margin can be read easily.

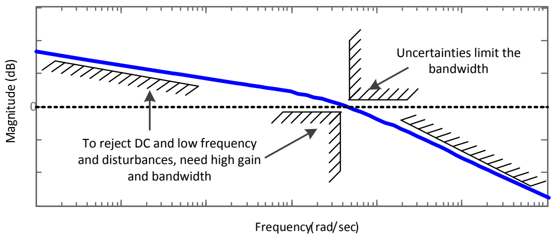
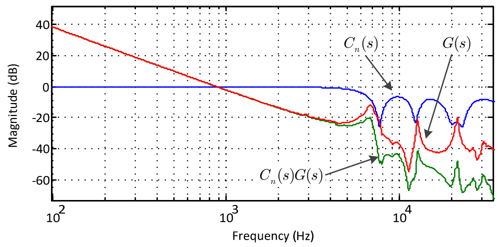
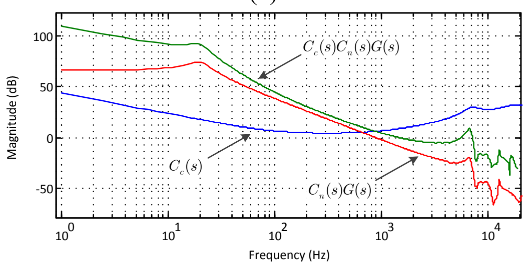


Fig.9 *Typical desired open-loop shape.*



(a)

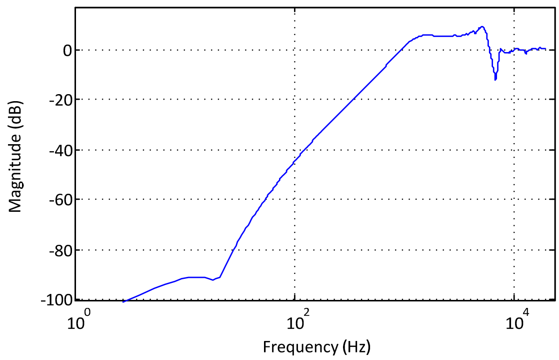


(b)

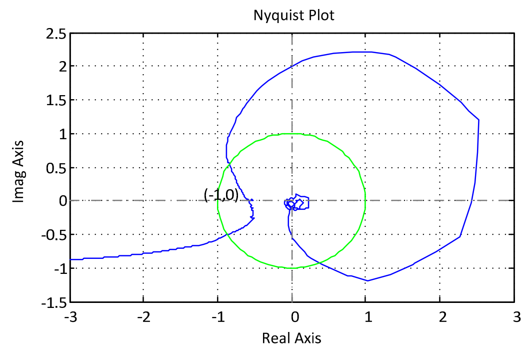
Fig.10 *Illustrative example of loop shaping, (a) using notch filter to suppress resonances at high frequency, (b) using lead/lag controller to shape the low frequency response.*

4.3 Modern Control Approaches

To achieve higher track densities, the classical control based switch controllers between PTOS based track seeking and loop shaping based track following may require considerably effort on tuning the parameters especially with large parameter variations induced from large production volume of drives. Many modern control approaches, hence, have been explored to overcome this issue. These modern control approaches are mainly based on state space design composing of controller and state estimate. There have been also researches considering multirate sampling scheme that can improve high frequency resonances as well as disturbance rejection. For example, the recent research [5] applies robust H_∞ control with multirate setting, where the sampling rate of control signal applying to actuator is faster than a measurement input. The resulting controller's parameter, rather constant from steady state solution, is then periodic; see also [6]. The same approach based on H_∞ setting is adopted for irregular sampling time system [7]. Another example of the work based on multirate system can be found in [8]. In this paper, the authors adopted multirate sampling in both track seeking and track following controllers that update the control signal faster than the measurement. The track seeking is designed based on dead-beat control; that is the error would be zero in one step time (assuming no saturation constraint). The track following controller is designed based on disturbance rejection by including the disturbance model inside the state estimate.



(a)



(b)

Fig.11 *(a) Error sensitivity response, and (b) Nyquis's plot for open-loop frequency response.*

The novel concept of blending control from [9] where several optimal control objectives are blended into one uniform controller is explored in [10] for track following controller to handle the multi-frequency disturbances.

To reduce a chattering when the controller is switched from track seeking to track following, the scheme known as initial value compensation (IVC) is normally used; see., e.g. [11], [12], [13], [14]. That is, the internal states of the track following controller will be initialized by some values rather than zero. The effect is, indeed, the same as feedforward command which is the same as taking the Laplace transform of differential equation with non-zero initial condition. This initial value can be determined, for example, by using optimal control technique (e.g., see [12]) with certain cost function reflecting the oscillation of PES when switch to track following controller.

Also, there have been researches that deal with unified controller for both track seeking and track following to handle a chattering issue. Most of them, in fact, combine the non-linear controller based on saturation as seen in track seeking and linear part into a unify one; see , e.g., [15], [16], [17], [18]. For example, [15], [16] proposed the unified controller composing of linear controller with saturation for track seeking purpose and nonlinear gain for track following. Both controllers are active simultaneously, but, with large error (track seeking phase), the linear controller with saturation contributes more on the control effort.

5 ADD-ON COMPENSATOR

5.1 Disturbance Rejection

The main compensator together with the notch filter is, practically, insufficient to reject disturbances. The controller can be further improved by characterizing the disturbance and then rejecting them. As discuss earlier, disturbances in HDD can be classified into synchronous or RRO and non-synchronous or NRRO. The method to handle both RRO and NRRO is based on the same principle, known as internal model principle (IMP), but different in detail. The IMP states that, to reject disturbances, the controller must contain the model of those disturbances, or more specifically, the poles of controller must contain the poles of disturbances. Using an integral controller to follow step input (or reject the step input from steady state error) is the simple example of the IMP.

For RRO, it is relatively easy to model, since the frequencies are synchronize to the spindle speed. Hence the RRO can be modeled by the sum of sinusoidal signal, i.e., Fourier series, as follow:

$$rro(t) = \sum_{i=1}^{\infty} a_i \cos(\omega_i t) + b_i \sin(\omega_i t),$$

and the corresponding frequency domain is

$$RRO(s) = \sum_{i=1}^{\infty} \frac{a_i s + b_i \omega_i}{s^2 + \omega_i^2}.$$

The RRO has high power spectrum at low and certain harmonics, so, in practice, only finite harmonics need to be rejected. Therefore, to apply IMP, the above RRO model needs to be included in the controller whether in cascade or parallel structure. In practice, the parallel form as in above model is easier to implement, since the required harmonics can be added naturally. The parallel structure can be described in Fig.12, where $C_{rro}(s) = \frac{\beta(s+\rho)}{s^2 + \omega_i^2}$, β, ρ are constant.

Another view point about above RRO compensator in parallel structure is that the transfer function $C_{rro}(s)$ is indeed a peak filter that amplifies input signal only at frequency ω_i, and hence its output is a sinusoidal signal. It is worth noting that this additional control effort from $C_{rro}(s)$ forces the head movement to follow the RRO so that there is no residue error induced from this RRO left. Note that most high power harmonic RROs are very certain; that is, their magnitudes and phases are almost constant at certain position or data track. Therefore after certain transient time, the magnitude and phase of the output from $C_{rro}(s)$ will be constant too. So, $C_{rro}(s)$ seems to generate the sinusoidal signal with constant magnitude and phase that looks like a *feedforward* signal. During transient, the magnitude and phase of this sinusoidal signal are adapted in such a way that no corresponding harmonic is observed in error signal, and then they reach steady state optimum values. This view point sometimes leads to the so called adaptive feedforward compensator (AFC) technique [19], [20], [21], [22]; even though, it is essentially a feedback controller.

The AFC concept, in fact, comes from different aspect from the IMP. The idea is from the assumption that the RRO harmonic has constant magnitude and phase. Therefore, we can create the feedforward sinusoidal signal from

$$u_i = A\cos(\omega_i t + \varphi)$$
$$= a_i \cos(\omega_i t) + b_i \sin(\omega_i t).$$

Note that, we do not know exactly the value of magnitude, A, and phase, φ, or a_i and b_i. However, we expect that, if we inject this signal with correct magnitude and phase into control effort, the error would be free of this harmonic. If, initially we chose incorrect values of magnitude and phase, the error will contain high power of this harmonic. Then we can use adaptive technique to adapt a_i and b_i by monitoring the power of this harmonic. More precisely, a_i and b_i can be adapted by:

$$a_i = \int \alpha \cos(\omega_i t) e(t) dt,$$
$$b_i = \int \alpha \sin(\omega_i t) e(t) dt,$$

where α is constant considered as learning rate in adaptive literature. It is worth noting that above equation is similar to finding the coefficient of the Fourier series. It is also worth noting that the above equation is somehow finding cross correlation between two signal $\cos(\omega_i t)$ and $e(t)$. If these two signals are correlated so the integral will not be zero, and it will be used to adapt the value a_i until there is no correlation between those two, and then the integral will be zero. Then the error signal contains no more considered harmonic. It is straight forward to verify that the transfer function from $e(t)$ to u_i is $C_{rro}(s)$. In fact, from Fourier's series view point, it obvious that u_i is a certain harmonic of $e(t)$.

The similar idea also can be applied for the non-synchronous disturbance or NRRO. In this case, NRRO contains broad spectrum of frequencies, but may concentrate in particular band of frequency. Then to apply the IMP, the disturbances can be approximately modeled by certain narrow-band frequencies by using band pass filter; that is, $C_{nrro}(s) = \dfrac{\beta(s+\rho)}{s^2 + 2\zeta\omega + \omega^2}$.

Note that the structure is similar to RRO compensator but with an addition small damping factor. The NRRO compensator can be many narrow-band pass filters in parallel structure similar to the RRO one. However, extensively use of this compensator will affect the overall loop shape, particularly phase lost at low frequency. Hence, in practice, the NRRO spectrum must be analyzed from the error signal to locate the dominant frequencies. The paper [23] proposes the approach to find the optimum value of the zero of the transfer function $C_{nrro}(s)$ that is based on the phase shift of the system at considered frequency, and also proposes the method to adaptively locate the peak frequency.

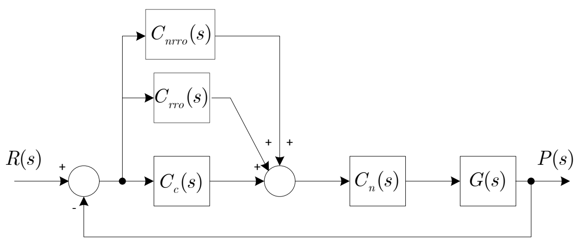


Fig.12 *Typical parallel structure for add-on RRO and NRRO compensator.*

The IMP can also be adopted by state space based design. That is the control effort can be design based on state feedback together with state estimate. The disturbance can be modeled by white Gaussian noise passing through certain band pass filter yielding the so called colored noise, and then the model of this filter, or noise shaping filter, can be incorporated with the model of the system in the state estimate design. This technique is well known in optimal control literature; see, e.g., [24], [25], [26], and has been applied for HDD servo system [27], [10].

5.2 Dual State Actuator

The limitation of the conventional servomechanism based on a VCM actuator is that the bandwidth which is limited by the resonances. Apart from well design in mechanic that can push the bandwidth up to 2 kHz, another approach used in today technology is that the dual state actuator (DSA) servomechanism. In DSA configuration, a micro actuator (MA) is mounted at a suspension or a slider level providing additional movement to a VCM at the location closed to read/write head. There are three configurations proposed for this DSA system; that is, Microelectromechanical systems (MEMS), piezoelectric transducer (PZT) and electromagnetic based [28], but the last one is difficult to manufacture. The MEMS based DSA is introduced by the IBM [29], while the piezoelectric based is introduced by Hutchinson technology [30]. The final position of a DSA system can be considered as the sum of each position or movement from a VCM and an MA as depicted in Fig. 13. There was much research, in the late 1990s, on the mechanical level design for those DSA based on both MEMS and PZT [28], [29], [30], [31], [32], [33]. The paper [34] did some comparison between MEMS and PZT based DSA by using μ-synthesis robust control technique. The MEMS based DSA has advantage on the relative position (the displacement of an MA itself relative to the displacement induced from VCM) availability that is useful in designing the multiple inputs multiple outputs (MIMO) controller. The PZT, on the other hand, does not provide the relative displacement but it has relatively higher resonance mode (around 5kHz compare with 2kHz for MEMS) and then easy to handle. Hence, from low to middle frequency range, a PZT is well behave and then easily to model and predict the relative displacement that is practically used in controller design such as decoupling method, e.g., see [33], [35].

In a DSA servo system, the controller becomes multiple inputs single output system. It can be seen from Fig. *13*, that the control loop of a VCM and a PZT will mutually affect each other. This make the control loop design complicated. The most widely used controller configuration is so called decoupling method. That is, one of the actuator control loop must know the relative position of another, so that it can compensate the total position. Practically, the VCM control loop is regarded as the main loop, and then the relative position from an MA will be fed to it. It can also be another way around, in theory. However, since the VCM provides large position or span, so it is suitable to be used as a main control loop while an MA loop can be considered as an add-on component that can be turned on and off. The decouple configuration, then, can be described in Fig. 14. Mathematically, it is obvious that the sensitivity function of the decouple loop can be described by

$$S(s) = \frac{1}{1 + C_{vcm}(s)G_{vcm}(s)} \bullet \frac{1}{1 + C_{ma}(s)G_{ma}(s)}.$$

Then the overall sensitivity can be considered as the product of two independent sensitivity functions from a VCM and an MA. This can simplify the servo loop design procedure by separately designing each loop shape; that is, the original loop shape from a VCM only can be improved by an MA loop shape without any change in the original VCM loop shape. Another advantage of this decouple configuration is that an MA loop can be disable without any change in a VMC loop, but with some performance dropping. Note that, as mentioned earlier, to obtain an MA relative position, the estimate is normally used.

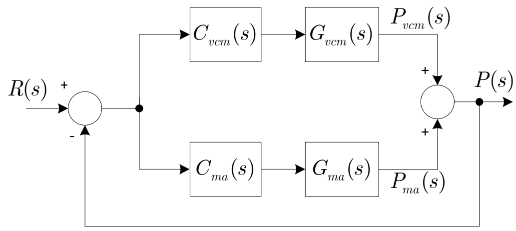


Fig. 13 *A diagram of DSA servo system*

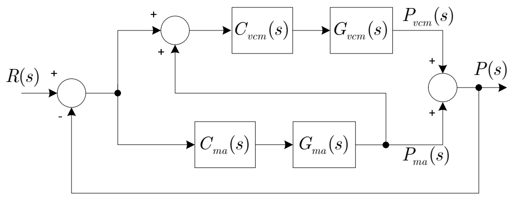


Fig. 14 *Decouple control scheme*

The decoupling method mentioned above is practically easy to implement and to design or tune. It is, however, not an optimal one. There has been much research on attempting to improve DSA servomechanism. For example, the slightly modification to the conventional decoupling method by applying certain filter called zero phase error tracking (ZPETC) inside the MA loop can improve the phase margin to the overall DSA loop [36]. To make the DSA servomechanism more effective, there have been many studies on applying modern control based on MIMO design such as optimal, robust control; see [37, 38, 39, 40]. Particularly, the paper [40] studied various control methods ranging from conventional SISO loop shaping to robust controller based on H_2, $\mathrm{H}_2/\mathrm{H}_\infty$ and μ-synthesis for the MEMS based DSA.

Although an MA has short span, it is fast response. It can, then, also be used to improve track seeking controller as well, in particular for short seek within an MA range happening very often during HDD operation. This can reduce a significant latency time. Some non-linear control technique such as sliding mode control is adopted for DSA; see e.g. [41], [42]. Feedforward technique is also applied to improve short seek controller in DSA HDD, [43]. Additional to the conventional decouple controller, the different feedforward commands for VCM and PZT are fed to VCM and PZT loop independently. A unify approach for both track seeking and track following controller is also proposed in [44, 45] based on decouple scheme and taking saturation nonlinearity of both actuators into account. The PTOS is adopted in VCM loop for fast track seeking, and the compensator in PZT loop is design to dominate the whole system during track following. That is, when position error is small, the servo loop is more sensitive to the PZT to which its movement relatively more contributes.

Apart from using a PZT as an actuator, there were some attempts to use a PZT as a sensor to detect vibration from both arm resonances and windage for active vibration rejection (AVR) scheme. The approaches include using a PZT as a sensor only to detect the structure resonances in AVR, e.g., [46], [47], using one strip of PZT as a sensor and another as an actuator, e.g., [48], [49], and using a PZT as an actuator and sensor , e.g.,[50], [51], [52], [53], [54]. There was also a report on using a PZT as both a sensor and actuator to reduce the disk flutter. In this case, however, a PZT is mounted on the cover at the position near the edge of platters, [55].

5.3 PZT Estimate

Many HDD servo systems rely on good state estimation on the read/write head movement, and most of them use the Kalman filter for a state estimate, especially in decoupling technique that requires the knowledge of PZT movement [56], [57], [44]. The standard Kalman filter can handle some parameter variation as well as noises [58]. However, when there is significant change in PZT gain, the fidelity of the state estimate suffers from this nonlinear effect. The PZT gain normally can be calibrated before using. It may also need to be on-line tracking for such an undesired event that its gain drops significantly. The paper [59], for example, proposed the adaptive based PZT gain tracking and fault detection.

From the properties of the Kalman filter, in ideal case, the estimation error should be completely random and is not correlated to any signal, and then the Kalman filter will be optimal in least mean square error sense. But most of the cases, we normally have suboptimal Kalman filter, since uncertainties, process noise, and measurement noise may not be completely random and the covariance of those noses using in the Kalman filter design are normally not exactly matched with the real one. So when the PZT gain varies, we expect to see that this estimation error will not be completely random and should somehow relate to the variation of the PZT gain. So by analyzing the estimation error signal, the signature of the PZT gain variation could be observed, and can be used to adjust the model. This technique has been used in adaptive Kalman filter to make it close to optimal filter [6], [7].

6 FUTURE TECHNOLOGY

6.1 Low Flying Height Effect

To achieve high areal density, the requirement in very low flying height is crucial; e.g., to achieve 100 Gbits per inch square would require around 6 nm flying height, or even lower than 5 nm for 1Tbits per inch square [60], [61]. Such a low flying height causes the problem of head-disk interface (HDI) that makes the media damage. This HDI issue is observed in many operation phase of the HDD operation including track following and during load the read/write head from a ramp to a media. The latter issue is observed frequently in high volume production, since it is an effect from just slightly miss-alignment installation of an HGA. So it is very important to detect the HDI during this load/unload phase since it is the first time that the head will be loaded to fly over the media.

Currently there are many approaches to detect the HDI during load/unload phase including monitoring the spindle's speed, voice coil motor's (VCM) current. It may be, however, not fast or sensitive enough to detect this HDI problem. Another method is to use additional sensor such as acoustic emission (AE) sensor, but this comes with additional cost.

When a slider contacts a media, it vibrates at certain frequencies depending on an air bearing surface [60]. This vibration may be observed from PES if its frequency component is relatively high enough compare to other resonance modes. Much computation, e.g., Fourier transform, will be required to extract this certain range of resonance frequency. However, the PES monitoring is limited by the nature of sampling in which the high frequency (above 100 kHz) HDI mode may not be detected. Beside the PES is suitable to use during track following phase for the slider is staying on track, and the PES is settle in small value, but during load/unload phase, the PES will rapidly change with very large value that is not suitable to be used. The HDI mode vibration can also be detected from AE sensor [62]. Another sensor that can be used to detect the HDI mode resonance is the laser doppler vibrometer (LDV) [63], [64]. This sensor provides information of the movement of the slider. However, both AE and LDV sensors are normally used in research or testing in a development phase.

Also, when a slider touches a media, there is an adhesion and friction force between the slider and media [65]. This friction can be measured by patching the strain gage at the tip of the slider. Again, this method requires additional sensor and is normally used in testing but not in normal operation. The friction may slightly affect spindle speed and may be detected from spindle back electromotive force (EMF).

There are also techniques that can detect the HDI mode resonance from the available signal with no additional sensor, that is thermal asperity (TA) detection and variable gain amplifier (VGA) [62]. The TA is an event when the read/write touches the media at high speed that causing some heat at read/write head. This heat will affect a magneto resistive of a read head and will be observed in abnormal read back magnetic signal. Another method to detect HDI mode resonance is that using VGA. When a slider flies close to a media, read back magnetic signal will be strong and then a pre-amplify gain will be adapted such that the read back signal is in the normal operation range. Hence the VGA is somehow correlated to the flying height of the slider.

In DSA system, a PZT can behave as actuator and sensor in the same time and can be used as self-sensing for smart structure; see e.g. [66]. During the load/unload phase, a PZT is not used. So it is possible to use this PZT as a sensor similar to AE sensor during load/unload, to detect an HDI with no additional cost.

6.2 Patterned Media

In current technology, in order to increase data density, the size of magnetism domain per data bit needs to be reduced. When the magnetism domain bit size is too small, the magnetization will be thermally unstable; that is increasing in temperature causes the magnetic polarity change, known as superparamagnetism [67]. Hence to maintain the thermal stability, the magnetism domain for each data bit must compose of many magnetic grains. Another way to reduce the size of magnetism domain with small amount of magnetic grain is to group a few magnetic grains together and separate them from others. These patterned magnetic groups or islands can hold there magnetic polarity well, since there are some spaces (non-magnetize material) among them. This future technology is referred to bit patterned media (BPM); see, e.g., [68].

By adopting this BPM, the bit size will be smaller, and hence it requires more precise servomechanism to access such dense data. This makes the servo system more challenge in the near future; e.g., for BPM the TPI would be above million tracks per inch [69]. During the past decade, there were some researches in servomechanism for this BPM; e.g., see [69], [70], [71].

7 CONCLUSION

The nano-scale mechanical positioning in an HDD cannot be achieved without complex control theories. This article provides brief detail on how control theories are applied in an HDD servo system from classical to modern control methodologies. The major characteristic of the movement of mechanic part such as double integrators and resonances is discussed and then comes the control algorithm to make it move fast and precise in the present of uncertainties, disturbances and noises. Various control methods are revisited as well as some reviews on advance control algorithms that have been studied and tested in the real HDDs. The promising research toward higher bit density is also reviewed.

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