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Focus servo performance optimization for an optical disk data storage device

DiMatteo, Joseph Howard, M.S.
The University of Arizona, 1988
FOCUS SERVO PERFORMANCE OPTIMIZATION
FOR AN OPTICAL DISK DATA STORAGE DEVICE

by

Joseph Howard DiMatteo

A Thesis Submitted to the Faculty of the
DEPARTMENT OF ELECTRICAL AND COMPUTER ENGINEERING
In Partial Fulfillment of the Requirements
For the Degree of
MASTER OF SCIENCE
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In the Graduate College
THE UNIVERSITY OF ARIZONA

1988
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ABSTRACT

This thesis concerns a study of the application, and performance optimization, of standard lead-lag compensation techniques to improve the closed loop performance of a focus servo system for an optical data storage device. Only with proper application of these compensation techniques will it be possible to meet the sub-micrometer focus error tolerances while maintaining the stability of the closed loop system. The performance indices used in this optimization study are the Integral of the Squared Error (ISE), the Integral of the Absolute Error (IAE), and the Integral of the Time multiplied by the Absolute Error (ITAE) as defined below, with the error function e(t) being the focus error of the closed loop servo system in response to a step input.

\[
\text{ISE} = \int_{0}^{T} e^2(t) \, dt
\]

\[
\text{IAE} = \int_{0}^{T} |e(t)| \, dt
\]

\[
\text{ITAE} = \int_{0}^{T} t \, |e(t)| \, dt
\]
In the data processing industry, the search for higher performance, lower cost data storage is a never ending one; one of the most popular technologies to emerge in this area over the past several decades is known as optical disk data storage.

1.1 Overview of Optical Data Storage Technology

The term optical disk data storage itself encompasses a broad spectrum of methods that have been developed for the storage and retrieval of digital data but they all have common threads that bind them together. In general, optical storage devices rely on a laser beam focused on a spinning disk to either cause, or detect, a change in the state of the active layer as the basic phenomenon necessary for data storage or retrieval. Typically, an inexpensive semiconductor laser diode is used in these devices that has a wavelength of 830 to 850 nm and a power output in the range of 2 to 40 mW, depending on the application. Through the use of various optical elements the output of the laser is ultimately focused to a spot size of approximately 1 μm at the surface of the active layer[1].

What differentiates the numerous types of optical storage is how they cause, or detect, this change in state and whether they are read only, write once, or erasable devices. Of these three, erasable optical storage holds the most promise for eventually replacing traditional magnetic recording technology because of its architectural compatibility with present data processing systems and its higher bit densities, which at $2 \times 10^8$ bits/cm$^2$
are on the order of five to ten times as large. These densities translate into data capacities of 200 to 500 Mbytes for a 120 mm (5.25") diameter optical disk as compared to 40 to 80 Mbytes for the same diameter magnetic hard disks. The comparison of the cost per Mbyte also favors optical storage with magnetic storage projected to be 2 to 3 times as expensive. Figure 1.1 is a diagram of the major functional elements in an optical storage device [1,2,3].

Within the realm of erasable optical storage, several approaches have proven their feasibility in the laboratory and have emerged as front runners in the race to yield a commercially viable product; these three technologies are known as phase-change, dye polymer, and magneto-optics.

The phase change method makes use of an active layer on the disk that can alternate between a crystalline high reflectivity state and amorphous low reflectivity state when heated by a laser pulse. The difference in reflectivity between the crystalline and amorphous spots on the disk can then be read back as digital data. The attractive feature of erasability comes from the fact that the amorphous to crystalline transition can be reversed by a laser pulse of approximately half the power need for the crystalline to amorphous change [4,5].

Dye polymer recording also relies on the ability to create reflectivity differences between regions on the disk as a means of storing data. The recording surface of a dye polymer disk actually consists of two dye impregnated layers that each absorb different wavelengths of light from two lasers in the recording head. A highly reflective bump is created, represent-
Figure 1.1. Functional Diagram Of An Optical Storage Device
ing a logical 1, when the lower layer is heated by the write laser. The bump is erased by heating the top layer with the erase laser, which forces the disk back to its flat, and less reflective, logical 0 condition [4,6].

Magneto-optical recording is a hybrid technology that combines some of the characteristics of standard magnetic recording with more typical optical storage techniques. In magneto-optics, the laser beam is used to heat up a localized spot on the active layer of the disk, which is usually a thin film of high coercivity magnetic material, past its Curie point. The Curie point, which is the temperature at which the magnetic material loses its magnetic property, is usually between 130° to 150° C. After reaching the Curie point, the dipoles of the magnetic material will align themselves with the flux lines of an externally applied magnetic field. When the laser pulse is turned off, and the active layer cools below the Curie point, the heated spot will lock in its new magnetic orientation [1].

The data created by reversing the magnetic orientation of the bit cells can be read by reflecting the same laser beam, but at much lower power, back off the recorded area and into a photodetector capable of detecting the Kerr effect rotation of the beam between oppositely polarized regions on the disk. Figure 1.2 shows the comparison between how the erasable optical recording techniques discussed here store data on the disk [1.7].

1.2 Need for Position Control Servo Systems

Regardless of the particular optical technology being considered, the ability to keep the laser beam in focus and on the desired track, for reliable data storage and retrieval, is of prime importance. The need for position control servos stems from the fact that the disks themselves have mechani-
Figure 1.2. Comparison Between Erasable Optical Recording Techniques
cal imperfections in both the focus and tracking planes that must be compensated for if the system is to operate properly. Radial runout is the term given to the eccentricity of the tracks on a disk while axial runout is defined as the measure of the warpage of a disk. The performance of these focus and fine tracking servos has a direct relationship to the maximum data transfer rate, data bit error rate (BER), and signal-to-noise ration (SNR) of the optical disk file. The job of the servo engineer is to design control systems that keep the closed loop position errors from these runouts within tolerable limits, which are usually less than 1 μm. Figure 1.3 shows the relationship between these runouts and the focus and fine tracking actuators, as well as the typical range of values in meters for the axial and radial runouts (AR, RR), disk land and groove sizes (L, G), and the laser spot size (S) [8].

Because optical disk technology is relatively new, much of the information on the present state-of-the-art for the position control servos used in these systems is best found in the form of patent disclosures [9-12]. What is known, from these sources and others, is that the servos used are typically analog systems that use some type of lead-lag compensation for stability. In particular, it may be concluded that very little literature on performance optimization for servo systems of this type exists.

1.3 Outline of Thesis

Since the design of the position control servos for the focus and fine tracking actuators are very similar, only the analysis of the focus servo will be included in this thesis to avoid unnecessary redundancy. The major objective of this thesis is to conduct a design study so as to optimize the time
OPTICAL DISK

PROTECTIVE LAYER

ACTIVE LAYER

LAND

GROOVE

SUBSTRATE

ACTUATOR
AND
LENS ASSEMBLY

DISK AXIAL
RUNOUT
(AR)

DISK RADIAL
RUNOUT
(RR)

.5 \times 10^6 \text{m} < S < 1.5 \times 10^6 \text{m}

.4 \times 10^6 \text{m} < L,G < 1.0 \times 10^6 \text{m}

10.0 \times 10^6 \text{m} < RR,AR < 200.0 \times 10^6 \text{m}

Figure 1.3. Radial and Axial Runouts
domain performance of a focus servo system. In general, it is desired to complete as much of the system design and analysis as possible with software models, developed using the Dynamic Simulation Language, because of the inherent flexibility and time savings they afford. The steps taken, and the methods used, in this servo development and optimization are contained in the remaining chapters.

Chapter 2 deals with defining the goals of the system design and the performance criteria. A review of frequency and time domain concepts is also presented with an emphasis on how they apply to general position control servo design. The detailed analysis in both frequency domain and time domain of all the stages in the focus servo development is contained in Chapter 3. Special emphasis is placed on the actuator characterization and how it affects the design of the compensation state. Chapter 4 covers the results of the simulation study for optimization and how they compare with the measured hardware data. The three performance indices, ISE, IAE, and ITAE, are used to optimize the time domain performance of the system.

Finally, Chapter 5 looks at the results and the conclusions drawn from this work and lists some topics for further investigation. In addition, this chapter discusses the contributions of this work with the key ones summarized here.

1. One major contribution of this work is in the practical application of the standard performance indices ISE, IAE, and ITAE to optimize the time domain performance of a closed loop position control servo. This optimization is done by choosing the values of the open loop bandwidth and phase
margin that minimize the loop error in the system in response to a step input.

2. Another contribution of this project is the development of simulation models in software of the focus servo system used in the optimization study.

3. A third significant contribution of this work is the hardware implementation and testing of the optimized servo system that resulted from the software simulation study. This hardware testing on the actual servo system allowed for the validation of the results from the simulation study; this validation is important because the basic goal of this project is to improve the error reduction capability of the real servo system through the optimization study.
CHAPTER 2

GENERAL DESIGN CRITERIA FOR POSITION CONTROL

OPTICAL SERVO SYSTEMS

2.1 Basic System Model and Analysis

The goal of an optical position control servo, as already mentioned in the previous chapter, is to minimize the position error in a stable closed loop system. Stated another way, the goal is for the optical servo actuator to follow, as closely as possible without oscillating, the disk runout to minimize the position error between the two; this definition of the performance objective of an optical servo leads directly to the block diagram of a general closed loop system as shown in Figure 2.1 with the disk runout being the reference input (R) and the actuator displacement being the desired output (X). The position error in the loop (E) is converted from a mechanical displacement into a more easily processed electrical current by the photodiode detector stage (G₁). All the gain and stability compensation in the loop is represented by a single block (G₂) that drives the power amplifier (G₃), which in turn supplies the current to the coil of the focus actuator (G₄). Each of these stages will be discussed in more detail in Chapter 3.

The choice of a block diagram representation for a physical system is not unique, since there are usually several valid ways to model the same system. The decision as to which model to choose should be based on how logically it represents the given components of the system and how simply it lends itself to analysis. The advantage of the representation chosen here is the controlled variable X is fed directly back to the summing junction.
\[
\frac{X}{R} = \frac{G}{1 + GH} = \text{CLOSED LOOP TRANSFER FUNCTION}
\]

FOR H = 1,

POSITION ERROR (E) = DISK RUNOUT(R) - ACTUATOR DISPLACEMENT (X)

Figure 2.1. Block Diagram Of General Position Control Optical Servo
without any modification, given that the feedback transfer function $H$ equals one. This type of unity feedback system is the easiest kind of system model to analyze, and hence is used by most designers whenever possible.

These block diagrams can help a control systems engineer to model an optical servo system, but to analyze, design, and test a servo system, a designer also needs measurable criteria to assess the system performance. There are two basic domains, the frequency and the time, that a designer of control systems can choose between to define the needed performance criteria.

**2.2 Frequency Domain Response Concepts**

Before analyzing frequency domain performance measurements for a control system, a brief review of relevant frequency domain concepts is in order. It is well known that for a system to be unconditionally stable, the open loop transfer function $GH(j\omega)$ can only cross the negative real axis between the origin and $-1 + j0$. This condition for absolute stability is best understood when $GH(j\omega)$ is plotted in polar coordinates, as shown in Figure 2.2, where the point the system becomes unstable, $-1 + j0$, is represented as a magnitude of 1 and a phase of -180 degrees [13]. Analysis of this polar plot leads to a second, more widely used, definition of stability; the phase of the open loop frequency response must not exceed -180 degrees unless the magnitude of the open loop frequency response is less than one for a system to be absolutely stable.

It is also well known that any real system whose open loop frequency response approached unity gain with -180 degrees phase lag has no margin of stability because even the slightest variation in system components
Figure 2.2. Open Loop Transfer Function $GH(j\omega)$ in Polar Coordinates
could push the response into the unstable region. Therefore, it is common to use gain and phase margins as measures of relative stability based on how close the open loop frequency response is to this unity gain, -180 degrees phase point. These gain and phase margins are defined below:

Gain Margin = \frac{1}{|GH(j\omega)|} \text{ where } \angle GH(j\omega) = -180 \text{ degrees} \quad (2.2.1)

Phase Margin = 180 - \angle GH(j\omega) \text{ where } |GH(j\omega)| = 1. \quad (2.2.2)

A third frequency domain performance measure is the open loop bandwidth, which is defined as the width of the frequency band between 0 Hz and the unity gain (0 dB) crossover point of the open loop magnitude response. These three open loop frequency domain performance criteria - gain margin, phase margin, and bandwidth - are graphically shown for a typical position control servo loop in Figure 2.3.

There are several advantages to an engineer of working in the frequency domain to design, analyze, and test control systems using Bode diagrams as tools. First, and foremost, certain test equipment, called dynamic signal analyzers (DSA), have been developed in the last decade that easily allow a designer to directly measure, and display as a Bode diagram, the open and closed loop frequency response data for a system. A common way in which frequency response data is taken by a DSA is to stimulate the system with a swept sine wave source and then measure both the input stimulus and the output response with narrow bandpass filters that track the frequency of the source. Typically, the closed loop frequency response data for a system is actually measured and the open loop frequency response data is then calculated using the waveform manipula-
Figure 2.3. Bode Diagram Of Open Loop Frequency Response For A Typical Position Control Servo
tion capability of the DSA. Figure 2.4 shows the block diagram for the test setup for measuring the closed loop frequency response data and the waveform calculations necessary to convert this data to the open loop frequency response [13].

A second advantage to using Bode diagrams is that they have certain attributes that make them particularly well suited to the design and analysis of control systems. By displaying data using logarithmic units and scales, the effect on the open loop frequency response of adding stages to the control loop can be calculated using addition instead of multiplication. For example, the impact of adding a lead-lag compensator to improve loop stability can readily be seen by graphically combining the compensator response and the original loop response. Finally, the capability to perform the reverse function of estimating a transfer function from a measured frequency response can also be very useful to a designer.

The previous paragraphs emphasize the merits of defining the performance criteria of a control system in terms of the frequency domain parameters such as the open loop bandwidth, phase margin, and gain margin. Unfortunately, however, the basic requirement of a position control optical servo to minimize the position error is inherently a time domain performance criterion. Therefore, a way to map this requirement into the frequency domain must be found.

2.3 Time Domain Response Concepts

Standard time domain performance criteria are usually defined in terms of the step response of a system. Figure 2.5 shows a typical step response along with some of the key measures associated with it. Of special
Figure 2.4. Block Diagram Of The Test Setup Used For Determining The Frequency Response Of A Closed Loop System

\[ \text{MEASURED CLOSED LOOP FREQUENCY RESPONSE} = \frac{Y(j\omega)}{S(j\omega)} = \frac{G_1 G_2 G_3 G_4}{1 + G_1 G_2 G_3 G_4 H(j\omega)} \]

\[ \text{CALCULATED OPEN LOOP FREQUENCY RESPONSE} = \frac{T(j\omega)}{1 - T(j\omega)} \]
Figure 2.5. Typical Time Domain Step Response
interest to us in the present work is how these step response parameters are related to the frequency domain performance criteria discussed previously. For a system whose response characteristics are dominated by a pair of complex poles, which is typical of position control servos with mechanical actuators, the following relationships generally hold. Open loop bandwidth is directly related to the error reduction capability of the system because both the time needed to react to a change in input (i.e., rise time) and the steady state error tend to decrease as the bandwidth increases. On the other hand, open loop gain and phase margins, as has already been established, are most closely identified with the relative stability of the control system. As gain and phase margins are increased the system becomes more stable, which tends to reduce the maximum overshoot and settling time, but the rise time increases indicating a slower response to an input change.

2.4 Performance Indices

Even though the relationships between the standard time and frequency domain performance criteria have been outlined, the fundamental question still remains on how to determine the values for open loop gain margin, phase margin, and bandwidth that result in the minimum possible error for an optical position control servo system. One possible answer to this question lies in the evaluation of time domain performance indices over a range of values for key frequency domain parameters. The usefulness of any performance index depends on how well it converges to some minimum when the parameter being varied approaches the value that maximizes the error reduction capability of the system.
The indices used in this study are all based on the integral of a closed loop error function due to a step input to the system. The general form for this performance integral is

\[ I = \int_{0}^{T} F(e(t), t) \, dt. \quad (2.4.1) \]

The upper integration limit \( T \) should be chosen so that it is greater than the settling time of the error transient \( e(t) \). These performance integrals are ideally suited to an error minimization study because they mathematically "add up" the total position error without regard to its source. As the example of Figure 2.6 shows, the integral of the error function for an underdamped system at some point will equal that for an overdamped system, with the minimum being somewhere in the middle [14].

The three performance indices utilized in this investigation of optimized focus servo performance are the integral of the square of the error, \( ISE \),

\[ ISE = \int_{0}^{T} e^2(t) \, dt \quad (2.4.2) \]

the integral of the absolute error, \( IAE \),

\[ IAE = \int_{0}^{T} |e(t)| \, dt \quad (2.4.3) \]

and the integral of time multiplied by the absolute error, \( ITAE \),

\[ ITAE = \int_{0}^{T} t |e(t)| \, dt. \quad (2.4.4) \]

The frequency domain parameters that are varied to determine the optimal values for these three performance integrals are the open loop phase margin and bandwidth. No attempt will be made to find an optimal setting
Figure 2.6. General Performance Integral Example
for gain or gain margin because these variables usually cannot be adjusted independently of the phase margin and bandwidth. The only requirement placed on the gain margin is that it be sufficient enough to not adversely affect system stability.
CHAPTER 3
FOCUS SERVO ANALYSIS, DESIGN, AND MODELING

Before any of the performance optimization techniques discussed in the previous chapter can be utilized, the servo loop to be studied must be developed and characterized in both the frequency and time domains. These frequency and time domain models then become the foundation that all the DSL software models are built on. The two most important blocks to be analyzed, because they dominate the response of the total control loop, are the actuator and compensation stages. The role of the compensation stage is to counteract any characteristics of the plant being controlled, in this case the focus actuator, that would make the loop unstable and to also supply loop gain for focus error reduction.

3.1 Focus Actuator

Focus and fine tracking actuators for optical disk files are usually based on some form of voice coil motor (VCM), which is essentially a permanent magnet DC motor. The basic principle on which these actuators operate is that when current flows through a wire in a magnetic field, a force is created proportional to the vector product of both quantities where

\[
\text{Force} = F = (ii) \times \vec{B}
\]

and

\[
\vec{i} = \text{current}
\]
\[
1 = \text{length of wire in the B - field}
\]
\[
\vec{B} = \text{magnetic field}
\]

Figure 3.1 illustrates these basic VCM relationships.
$\vec{F} = (\vec{I} \times \vec{B})$

Figure 3.1. Basic Voice Coil Motor Physics

Figure 3.2. Mechanical Focus Actuator Model
This concept is applied to actuators by mounting the wire coil and lens assembly to a chassis that can move independently in the focus and tracking planes on a spring suspension system. Motion is induced along these planes when a current is driven through the appropriate coil. Hence, these actuators are used to transform electrical current into mechanical displacement.

Mechanically, the actuator can be thought of as a damped spring and mass system described by Newton’s second law of motion which states that the summation of the total forces acting on an object equals its mass multiplied by its acceleration.

\[ MA = M \ddot{x}(t) = \Sigma F(t) \]  

(3.1.2)

The actuator mechanical dynamics are shown in Figure 3.2 where

- \( K_s \) = spring constant
- \( K_f \) = friction constant
- \( M \) = mass of actuator
- \( F_i \) = displacement force due to actuator coil current
- \( F_g \) = force due to weight of actuator

\[ = Mg \]

From Figure 3.2 we can write Newton’s second law for a focus actuator to include the forces acting on it as

\[ M \ddot{x}(t) = F_i(t) - K_s x(t) - K_f \dot{x}(t) + F_g(t) \]  

(3.1.3)

which may be rewritten

\[ \ddot{x}(t) = \left( \frac{1}{M} \right) F_i(t) - \left( \frac{K}{M} \right) x(t) - \left( \frac{K_f}{M} \right) \dot{x}(t) + \left( \frac{1}{M} \right) F_g(t) \]  

(3.1.4)

\[ \dot{x}(t) = \int \ddot{x}(t) dt + \dot{x}(0) \]  

(3.1.5)
\[ x(t) = \int x(t) \, dt + x(0). \quad (3.1.6) \]

The frequency domain analysis for this system closely follows that for the time domain with equation (3.1.4) transformed to

\[ X(s)s^2 = \left( \frac{1}{M} \right)F_1(s) - \left( \frac{K}{M} \right)X(s) - \left( \frac{K}{M} \right)X(s)s + \left( \frac{1}{M} \right)F_\gamma(s) \quad (3.1.7) \]

Equation (3.1.7) can be reduced to the form \( X(s)/\Sigma F(s) \) where

\[ X(s)[s^2 + \left( \frac{K}{M} \right)s + \left( \frac{K}{M} \right)] = \left( \frac{1}{M} \right)[\Sigma F(s)] \quad (3.1.8) \]

and, therefore,

\[ \frac{X(s)}{\Sigma F(s)} = \frac{1}{M} \cdot \frac{-\left( \frac{K}{M} \right)s + \left( \frac{K}{M} \right)}{s^2 + \left( \frac{K}{M} \right)s + \left( \frac{K}{M} \right)} \quad (3.1.9) \]

The block diagram models corresponding to (3.1.7) and (3.1.9) are shown in Figure 3.3.

From this analysis, it is readily seen that, mechanically, the actuator acts as a second order complex pole system described by

\[ \frac{X(s)}{\Sigma F(s)} = \frac{K}{s^2 + \omega_n^2 + \omega_n^2} \quad (3.1.10) \]

or

\[ \frac{X(s)}{\Sigma F(s)} = \frac{K}{s^2 + (2\delta\omega_n)s + (\omega_n^2)} \quad (3.1.11) \]

where \( \delta \) is the damping ratio and \( \omega \) is the undamped natural frequency. By equating terms in (3.1.11) and (3.1.9) we get the relationship between the standard parameters, \( \delta \) and \( \omega \), and the parameters of the focus actuator.
Figure 3.3. Frequency Domain Block Diagram For The Mechanical Dynamics Of The Focus Actuator
\( K_s, K_f, \) and \( M \) as

\[
\omega_n = \frac{K_s}{\sqrt{M}}
\]

(3.1.12)

and

\[
\delta = \frac{K_f}{2\omega_n M} = \frac{K_f}{2\sqrt{M K_s}}.
\]

(3.1.13)

The amount by which the actuator magnitude response peaks relative to its DC value is related to the damping ratio by

\[
|G_n| = -20 \log 2\delta \sqrt{1 - \delta^2} \text{ dB}
\]

(3.1.14)

and the exact frequency of peaking is

\[
\omega_{\text{peak}} = \omega_n \sqrt{1 - 2\delta^2}
\]

(3.1.15)

and in the cyclic measure

\[
f_{\text{peak}} = \frac{\omega_{\text{peak}}}{2\pi}.
\]

(3.1.16)

For the actuator used in this study where \( K_s = 380 \text{ N/m}, \)
\( K_f = .358 \text{ N/m/s}, \) and \( M = .00095 \text{ kg}, \) the following values for the frequency domain parameters can be calculated:

\[
\omega_n = 632 \text{ rad/s} \quad f_n = 100.7 \text{ Hz}
\]

\( \delta = .298 \)

\( |G_n|_{\text{peak}} = 4.9 \text{ dB} \)

\( \omega_{\text{peak}} = 572 \text{ rad/s} \quad f_{\text{peak}} = 91.1 \text{ Hz}. \)

So far only the mechanical elements of the focus actuator have been considered in developing its model; to be complete, however, the electrical parameters of the coil must be included in the actuator transfer function. Figure 3.4 gives the electromechanical representation of the focus actuator.
Figure 3.4. Electrical And Mechanical Components Of Focus Actuator Model
where

\[ v_c = \text{voltage across actuator coil} \]

\[ v_b = \text{back emf generated as a function of the linear velocity of the coil} \]

\[ R = \text{coil resistance} \]

\[ L = \text{coil inductance}. \]

The voltage relationships around the loop are

\[ v_c(t) - v_b(t) = i(t)R + L \frac{di}{dt} \]  \hspace{1cm} (3.1.17)

which can be rewritten as

\[ \frac{di}{dt} = \frac{v_c(t) - v_b(t)}{R} - \frac{i(t)}{R} \frac{L}{R} \]  \hspace{1cm} (3.1.18)

The electrical time constant of the actuator coil, \( L/R = \tau \), also defines the location of the electrical pole of the actuator which is given in Hz by

\[ f_c = \frac{(R/L)}{2\pi} \]  \hspace{1cm} (3.1.19)

Since, for the focus actuator used here, \( L = 18 \mu\text{H} \) and \( R = 1.8 \Omega \), then \( f_c = 15.9 \text{ KHz} \). Now, by integrating (3.1.18) we can eliminate the derivative terms from this expression.

\[ di = \left[ \frac{v_c(t) - v_b(t)}{R} - i(t) \right] dt \]  \hspace{1cm} (3.1.20)

\[ i(t) = \int_0^T \left[ \frac{v_c(t) - v_b(t)}{R} - i(t) \right] dt \]  \hspace{1cm} (3.1.21)

\[ = \frac{1}{\tau} \int_0^T \left( \frac{v_c(t) - v_b(t)}{R} - i(t) \right) dt \]  \hspace{1cm} (3.1.22)

The transfer function of the electrical part is derived similarly using the s-plane equivalents for the loop elements of Figure 3.4.
\[ V_c(s) - V_b(s) = I(s) (R + sL) \]  
Therefore,  
\[ I(s) = \left( \frac{1}{R + sL} \right) (V_c(s) - V_b(s)). \]  

The block diagram representation of the electrical part is shown in Figure 3.5 [14,15,16].

The coupling between the electrical and mechanical segments of the actuator is through two constants, \( K_i \) and \( K_v \), defined as:

\( K_i \) = current to force constant

and

\( K_v \) = velocity to voltage constant.

Due to conservation of energy in the actuator air gap, the mechanical power in must be equal to the electrical power out, i.e.

\[ F \dot{x}(t) = v_b i(t). \]  

Also, the actuator force is known to be proportional to coil current, i.e.

\[ F = K_i i(t). \]  

The back emf generated in the actuator coil is likewise known to be proportional to the velocity of the coil in the stationary magnetic field and, hence,

\[ v_b(t) = K_v \dot{x}(t). \]  

By substituting these relationships into (3.1.28) we get

\[ K_i i(t) \dot{x}(t) = K_v x(t)i(t) \]  

and since, in MKS units,

\[ \frac{N m}{s} = 1 \text{ watt} = 1 \text{ V A} \]
Figure 3.5. Frequency Domain Block Diagram For The Electrical Dynamics Of The Focus Actuator
the relationship between $K_i$ and $K_v$ and be expressed as

$$K_i = K_v.$$ \hspace{1cm} \text{(3.1.30)}

All of the previous analysis assumes an ideal actuator when in fact there are physical constraints that must be taken into account when dealing with real systems. One factor that must be considered during the design cycle is that any actuator has a limit on how much power can be dissipated by its coil without doing permanent damage to the windings. This limit on power dissipation is important because it puts an upper bound on the current that can be driven through the coil, which in turn limits the maximum displacement and acceleration of the actuator. It can be shown that maximum acceleration of the actuator can be approximated by

$$x(t) = \frac{i(t) K_i}{M} \max$$ \hspace{1cm} \text{(3.1.31)}

and its displacement by

$$x(t) = \frac{i(t) K_i}{K_s} \max$$ \hspace{1cm} \text{(3.1.32)}

The coil of an actuator must be able to accept enough current so that its maximum acceleration and displacement will be large enough to allow it to follow any expected disk runout. For the actuator used in this study, with $K_i = .2 \text{ N/A}$ and $i(t) = 1.1\text{ A}$, the maximum instantaneous acceleration is $\dot{x}(t) = 231.6 \text{ m/s}^2$ and the maximum low frequency displacement is $x(t) = \pm 579 \text{ \mu m}$.

A second, and more important, design constraint to be dealt with concerns the undesirable effects of high frequency mechanical resonances
that often exist in real actuators. Unlike the transfer function characteristics of the natural frequency (\(\omega_n\)), which can be analytically predicted, an actuator's secondary resonances, if any exist, can be characterized by physically measuring its frequency response, or through finite element modeling. The reason these unwanted resonances are significant is that they can limit the maximum possible gain and bandwidth of the servo loop they are part of, as will be seen in section 3.4.

Even though there are countless variations of these mechanical resonances, they typically can be approximated by a pair of complex poles and zeroes with a transfer function given by

\[
G_R(s) = \frac{s^2 + 2\delta_2\omega_2 s + \omega_2^2}{s^2 + 2\delta_p\omega_p s + \omega_p^2}.
\] (3.1.33)

The time domain model for these resonances can be developed from this transfer function for \(G_R(s)\), recognizing that

\[
G_R(s) = \frac{X(s)}{Y(s)}
\] (3.1.34)

the corresponding differential equation is

\[
x(t) + 2\delta_p\omega_p x(t) + \omega_p^2 x(t) = y(t) + 2\delta_2\omega_2 y(t) + \omega_2^2 y(t)
\] (3.1.35)

which can then be written as a system of first order equations

\[
\dot{x}_1(t) = 2\delta_2\omega_2 y(t) - 2\delta_p\omega_p x(t) + x_2(t)
\] (3.1.36)

\[
\dot{x}_2(t) = \omega_2 y(t) - \omega_p^2 x(t)
\] (3.1.37)

where

\[
x(t) = x_1(t) + y(t)
\] (3.1.38)

The advantage of using these frequency and time domain representations of secondary resonances is that they provide a reasonably simple
and accurate picture of a very complicated part of actuator performance. It should be noted, however, that since these unwanted resonances vary with time and operating environment in a physical actuator, the transfer function model should reflect the worst case conditions expected of the focus actuator to be used in the design. One way to assess the variability of these secondary resonances is by measuring the frequency response of the focus actuator over a period of days. Figures 3.6, 3.7, and 3.8 give the gain and phase response of the focus actuator used in this study measured on three successive days, with a 200 mV sinusoid as the test disturbance. One observation to be made from this data is the presence of a major secondary resonance at a frequency of 5KHz with a maximum phase lag, \( \phi_R \), of 30° to 70° and a resonance magnitude \( |G_R| \), of 5 dB to 15 dB. The secondary resonance transfer function parameters shown here were iteratively selected, using the frequency domain DSL actuator model, to match the measured worst case secondary resonance at 5KHz of \( \phi_R \) equal to 70° and \( |G_R| \) equal to 15 dB. From this measured data, the parameters of the transfer functions \( G_R(s) \) can be determined as, \( \delta_z = .008 \), \( \omega_z = 31,400 \) rad/s (5000 Hz), \( \delta_p = .010 \), and \( \omega_p = 31,800 \) rad/s.

A DSL frequency response plot for the total focus actuator, including the 5 KHz secondary resonance, is shown in Figure 3.9. A comparison of the measured actuator response versus the simulated response illustrates the good correlation between the two; Table 3.1 summarizes the results of the measured, simulated, and calculated values of several key actuator performance parameters. The complete frequency domain focus actuator block diagram is shown in Figure 3.10.
Figure 3.6: Measured Focus Actuator Frequency Response. Day 1

X = 105.2 Hz  
Y = 57.7812 dB

HFREQ RESP
60.0

FOCUS ACTUATOR
50 k

FXD Y 10 Log Hz

Phase
Deg

-360

200 MV DIS. 000 MV OFF 50 k
Figure 3.7: Measured Focus Actuator Frequency Response. Day 2

X = 105.2 Hz
Y = 56.5007 dB

FREQ RESP
60.0

dB

V

EU1
-40.0

Fxd Y 10 Log Hz

FOCUS ACTUATOR

50k

Phase

Deg

Fxd Y 10 Log Hz

FREQ RESP
0.0

-360

200MV DIS. 000MV OFF 50k
Figure 3.8: Measured Focus Actuator Frequency Response. Day 3
Figure 3.9. Simulated DSL Focus Actuator Frequency Response
Figure 3.10. Focus Actuator Frequency Domain Block Diagram
<table>
<thead>
<tr>
<th>Parameter</th>
<th>CALCULATED RESULTS</th>
<th>DSL SIMULATION RESULTS</th>
<th>MEASURED RESULTS</th>
</tr>
</thead>
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<td>100.7Hz</td>
<td>DAY 1 DAY 2 DAY 3</td>
</tr>
<tr>
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<td></td>
<td>105.2Hz 103.0Hz 100.8Hz</td>
<td></td>
</tr>
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<td>$f_c$</td>
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<td>19.5KHz 20.0KHz 18.0KHz</td>
<td></td>
</tr>
<tr>
<td>$G_n$</td>
<td>4.9dB</td>
<td>7.2dB 6.5dB 6.2dB</td>
<td></td>
</tr>
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<td></td>
</tr>
<tr>
<td>$\phi_R$</td>
<td>---</td>
<td>70 70 70 30</td>
<td></td>
</tr>
</tbody>
</table>

Table 3.1. Summary Of Calculated, Simulated, And Measured Actuator Parameter Values
3.2 Power Amplifier Stage

The basic function of the power amplifier stage is to convert the control signal in the servo loop into the driving function, in this case the coil current, required by the focus actuator. There are two fundamental ways in which the actuator coil current can be generated using power operational amplifiers; one is to drive the actuator coil with a constant voltage source and the other is to drive it with a constant current source.

As the name implies, a constant voltage driver simply sets up a constant voltage across the actuator coil in response to the reference control signal \( V_{\text{ref}} \). Figure 3.11 illustrates one possible implementation of a constant voltage coil driver along with its frequency domain block diagram. The coil current can be solved for by observing that:

\[
V_d(s) = V_e(s) + I(s)R_s = [I(s)(R + sL) + V_b(s)] + I(s)R_s \quad (3.2.2)
\]

Since \( V_d(s) = V_{\text{ref}}(s) \), (3.2.2) can be rearranged to give

\[
\frac{V(s) - V_{\text{ref}}(s)}{R_s + (R + sL)} = \frac{I(s)}{R_s + (R + sL)} \quad (3.2.3)
\]

From (3.2.3) it can be seen that as the frequency of the control signal \( V_{\text{ref}} \) increases, the coil current \( I \) decreases and hence, the actuator response rolls off proportional to its electrical pole.

A constant current driver, by contrast, supplies a constant current to the actuator coil regardless of the frequency of the control signal input, within the limits of the power amplifier itself. Figure 3.12 shows a non-
Figure 3.11. Actuator Coil Driven By A Constant Voltage Power Amplifier Stage
inverting constant current source power amplifier stage implementation
and its associated block diagram. Similarly, the relationship between the
coil current and $V_{ref}$ can be developed from (3.2.3) except that here

$$V_d(s) = (V(s) - I(s)R_s)K_a$$ (3.2.4)

where $K_a$ equals the open loop gain of the power amp. Combining equations
(3.2.2) and (3.2.4) gives

$$(V(s) - I(s)R_s)K_a = [I(s)(R + sL) + V_b(s)] + I(s)R_s .$$ (3.2.5)

Solving (3.2.5) for the coil current results in

$$I(s) = \frac{(V(s)K_a - V_b(s))}{R_sK_a + R_s + (R + sL)}$$ (3.2.6)

Here again, the coil current is dependent on the frequency of $V(s)$ but
if $K_a >> 1$ then $I(s)$ can be approximated by

$$I(s) = \frac{V(s)}{R_s}$$ (3.2.7)

Because the constant current source driver virtually eliminates the effects
of the actuator electrical pole on system performance it is the preferred
power amplifier configuration and is therefore, the type used in this design.

### 3.3 Photodetector And Preamplifier Stage

A key element in any position control servo system is the transducer
that converts the position error in the loop to a control signal the system
can process, in this case from focus error in meters to electrical current.
The transducer used in this design is a four quadrant photodetector whose
Figure 3.12. Actuator Coil Driven By A Constant Current Power Amplifier Stage
output current from each quadrant varies as a function of the position error in the loop. Figure 3.13 illustrates how the shape of the reflected beam on the quadrant detector is distorted as the beam incident on the disk surface moves in and out of focus. The focus error signal (FES) is formed by the preamplifier stage by summing the output current from the four quadrants of the detector according to the relationship

\[ \text{FES} = (I_a + I_c) - (I_b + I_d). \]  

(3.3.1)

Because of the extremely small magnitude of the variables in question, it is very difficult, due to instrumentation limitations, to directly measure the defocus to FES current relationship. Figure 3.14 does, however, show the output of a software model used to simulate this relationship for the detector used in this design.

Finally, Figure 3.15 is a simplified diagram of the quadrant photodetector and preamplifier stage. The first stage of the preamplifier consists of four transconductance amplifiers that convert the current out of each of the photodetector quadrants into a corresponding voltage that can then be combined into a voltage proportional to the FES by a differential summing amplifier. This FES voltage signal forms the basic control signal that is then processed by the remainder of the servo electronics.

### 3.4 Compensation Stage

If the current characteristics of the plant, in this case the focus actuator, that the control system is being designed for do not allow the closed loop system to meet desired performance criteria for stability and error reduction, the designer has two choices: either the plant itself must be modified so that the system meets the requirements or compensation stages should be added to force the system to perform acceptably. Since
Figure 3.13. Relationship Between Focus Condition And Quadrant Detector Output
Figure 3.14. Detailed Defocus To FES Relationship
Figure 3.15. Photodetector And Preamplifier Stage Diagram
the response of the focus actuator is inherent in its design, the logical alternative is to add compensation stages to achieve the desired system performance. Lead-lag compensation techniques are used in this design because of their well documented characteristics and applicability to this system [13,14,15].

Given that the focus actuator's response is dominated by a low frequency complex pole pair, as was seen in section 3.1, the need for a phase lead compensator is clear. If the system is to be stable, the -180 degree phase lag from the actuator must be reduced at the zero db magnitude crossover point. The phase lead compensator used to reduce the open loop phase lag of the system is shown in figure 3.16. Besides reducing the phase lag in the system, a phase lead compensator also increases the open loop bandwidth which, in turn, increases the error reduction capability of the closed loop system.

One practical consideration that must be taken into account when using phase lead compensators to maintain stability and increase bandwidth in a real system is that they act as a high frequency gain amplifier where the maximum gain boost is given by

\[ G_{\text{max}} = 20 \log_{10} \alpha \]  \hspace{1cm} (3.4.1)

where \( \alpha \) is the constant

\[ \alpha = \frac{R_1(R_1 + R_2)}{R_1R_2} \]  \hspace{1cm} (3.4.2)

This additional gain will, therefore, amplify the effects of any high frequency noise or mechanical resonances in the system, which can then saturate the amplifiers in the loop. Since the amount of phase lead the compensator can
\[ F_z = \frac{1}{2\pi R_1 C} \quad F_{\text{max}} = \sqrt{F_z F_p} \quad F_p = \frac{1}{2\pi R_2 C} \]

Figure 3.16. Phase Lead Compensator
supply, \( \phi_{\text{max}} \), also increases as a function of \( \alpha \), where

\[
\phi_{\text{max}} = \sin^{-1}\left(\frac{\alpha - 1}{\alpha + 1}\right)
\]

(3.4.2)

it was decided to use the minimum amount of phase lead necessary in this system to meet the performance criteria.

The choice of the lag compensator employed in this system is based on two main considerations: one is to eliminate any steady state offsets in the loop and, second, to increase the low frequency gain to help compensate for the large magnitude of the low frequency disk runout components. As figure 3.17 illustrates, the majority of the runout for a typical disk is contained in the fundamental and low order harmonics, which is representative of the disks used in this study. The lag compensator used to meet these considerations, called a low frequency integrator, is shown in Figure 3.18. The low frequency integrator differs from a classical integrator in that it has a zero \( (F_0) \) in addition to the traditional pole at the origin; this zero is needed to cancel the phase lag effect of the integrator pole at higher frequencies so that it does not interfere with the phase lead compensator near the zero dB crossover.

So far in this chapter each of the principal components of the servo loop have been analyzed individually; the next chapter will deal with the optimization of this focus servo system in terms of the open loop bandwidth and phase margin, and consequently the final values for the key compensation poles and zeroes.
Figure 3.17. Disk Axial Runout Representations
$$F_z = \frac{1}{2\pi RC}$$

Figure 3.18. Low Frequency Integrator (Lag) Compensator
CHAPTER 4
OPTIMIZATION OF SYSTEM PERFORMANCE

4.1 Optimization Strategy

Given the control system design and the choice of compensation, as outlined in the previous chapters, the next step is to develop a methodology for selecting the optimal settings for key system parameters. As has already been discussed, three performance indices, ISE, IAE, and ITAE, will be used in determining the optimal values for the open loop bandwidth and the phase margin. Typically, the selection of these parameters is based solely on trial and error, or the past experience of the designer. The performance optimization strategy employed here brings discipline to this crucial step of the final system design.

The overall goal of this strategy, which is outlined in the steps below, is to use software models of the focus servo system to perform the actual optimization study (because of the time savings it affords), followed by hardware testing of the optimized compensation values for validation and comparison purposes.

1. Develop frequency and time domain DSL simulation models based on the analysis performed in Chapter 3.

2. Using the simulation models developed, perform the optimization study based on the performance indices, with and without the 5KHz actuator resonance included in the model, for a step input disturbance. The reason for analyzing the system both with and without the 5KHz resonance is to determine what impact it has on system performance.
3. Implement the optimized compensation values determined from the simulation study on the real hardware system and then make the following measurements:

- The ISE, IAE, and ITAE for the optimal bandwidth versus phase margin.
- The open loop and closed loop frequency responses.
- The peak-to-peak (P-P) and RMS focus error at the preamplifier output for two different disks over a range of rotational speeds (2000, 3000, 4000, 5000, 6000 RPM). This range of rotational disk speeds was chosen because it is representative of the rotational speeds that the disks ultimately will have to spin at to achieve the data transfer rates required of an optical data storage product. It should be noted, however, that the data transfer rate is dependent on other factors besides disk speed, such as the type of recording code used when writing data on the disk.

In all cases the measured data will be compared to the simulated results from the optimization study and the correlation between the two will be discussed.

4.2 Simulation Software Overview

All of the simulation models used in this thesis were written using Dynamic Simulation Language (DSL/VS), an IBM FORTRAN-based continuous systems simulation language, running in a VM/CMS interactive environment. DSL/VS has numerous features that make it particularly well suited to the simulation of control systems. One of the most important
of these features is the availability of function blocks that reduce the coding of frequently needed system elements, input functions, and special features to one line DSL statements. For instance, in the time domain, the performance indices, as given by questions (2.4.2), (2.4.3), and (2.4.4), are represented in DSL by

\[ ISE = \text{INTGRL}(0.0, \text{ERRSqd}) \]  
\[ IAE = \text{INTGRAL}(0.0, \text{ABSERR}) \]  
\[ ITAE = \text{INTGRAL}(0.0, \text{TABSER}) \]  

In the frequency domain, displaying complex transfer functions is made easy by using the DSL functions GAIN and PHASE. The use of these functions is illustrated by the following statements that compute the open loop gain and phase where

\[ \text{MAGOL} = 20.0 \times \text{GAIN(OL)} \]  
\[ \text{PHASOL} = \text{RADEG} \times \text{PHASE}(0.0, \text{GOL}) \]  

and GOL is the overall focus servo open loop transfer function. Appendix A gives the source code for several of the key DSL simulation programs written for this study [16].

There are many reasons that simulation has become so prevalent in the design process in recent years, with one of the major ones being the ability to help reduce the risk and time involved in the product development cycle. Typically, as was the case in this project, the availability of hardware prototypes to do development work is very limited in the early phases of a project, while at the same time the demand for their use by the different groups involved is heavy. The use of simulation models for the various components of the product, like the focus servo, allows the multiple design
efforts involved in the overall project to proceed in a parallel fashion with the hardware prototypes reserved mainly for final testing and system integration.

Another advantage of incorporating simulation as an integral part of the development process is that it gives the designer a tool to do "what if" types of analysis that would not be possible, or practical, to perform on a hardware prototype. One example of this is the ability to calculate the performance indices for a focus servo system without the 5KHz resonance in the actuator, something that is not possible on the real hardware system used in this study.

4.3 Performance Optimization Simulation Results

In order to calculate the performance indices chosen for this study it is necessary to apply a step disturbance to the system and then measure the error resulting from this input. Since we are interested in reducing the position error in the loop caused by the disk axial runout, it would be ideal to use a step axial runout as the test disturbance input. Unfortunately, it is not feasible to physically generate a step input using a real disk, even though it would be easy to do so in the simulation model. Another possibility would be to input a step voltage as the test disturbance, a very close approximation of which can be physically created using a modern function generator. Therefore, in the interest of compatibility between the simulation model and the hardware prototype, a .5 volt step input will be used as the test disturbance for both cases when calculating the performance indices. Figure 4.1 gives a diagram of this test setup; a key thing to note from this figure is that the disk runout input must be equal to zero
Figure 4.1. Test Setup For Measuring Performance Indices
when measuring the error in the loop due to the step voltage input.

As was already discussed, the reason for calculating the values of performance indices for a simulated system with and without the 5KHz resonance in the focus actuator is to see what effect it has on the selection of the optimal values for the open loop bandwidth and the phase margin. The ranges chosen for both the bandwidth and phase margin, 2 to 10KHz and 20 to 60 degrees respectively, are based on practical limits that can realistically be implemented in an actual system. Also, the performance indices are calculated over a 10 ms interval to give the error transient adequate time to settle.

The results of the simulation for the focus servo system without the 5KHz resonance in the actuator are graphically summarized in Figures 4.2, 4.3, and 4.4. The important thing to note from these results is that, without any secondary actuator resonances, optimization with all three indices resulted in 10KHz and 50 to 55 degrees as the optimal values for the bandwidth and phase margin. It is also apparent that the minimum values for all three indices are asymptotically approaching a limit as the bandwidth is increased. Therefore, it is logical to conclude that even in an "ideal" system of this type (one without secondary actuator resonances, noise considerations, etc.) there would be little practical advantage in setting the bandwidth excessively high. The detailed data for this case is given in Appendix B.

The analysis of the focus servo system with the 5KHz actuator resonance included in the model produced dramatically different results compared to the previous simulation. Figures 4.5, 4.6, and 4.7 show the
INTEGRAL OF THE ERROR SQUARED (10E-5 V**2*S)

Figure 4.2. ISE For A System Without A 5KHz Actuator Resonance For 2, 4, 6, 8, And 10KHz Bandwidths
INTEGRAL OF THE ABSOLUTE ERROR (10E-5 V*S)

Figure 4.3. IAE For A System Without A 5KHz Actuator Resonance For 2, 4, 6, 8, and 10KHz Bandwidths
Figure 4.4. ITAE For A System Without A 5KHz Actuator Resonance For 2, 4, 6, 8, And 10KHz Bandwidths
INTEGRAL OF THE ERROR SQUARED (10E-5 V**2*S)

Figure 4.5. ISE For A System With A 5KHz Actuator Resonance For 2, 4, 6, 8, And 10KHz Bandwidths
Figure 4.6. IAE For A System With a 5KHz Actuator Resonance For 2, 4, 6, 8, And 10KHz Bandwidths
INTEGRAL OF THE TIME * ABSOLUTE ERROR (10E-8 V*S**2)

Figure 4.7. ITAE For A System With A 5KHz Actuator Resonance For 2, 4, 6, 8, And 10KHz Bandwidths
results from the simulations for 2, 4, 6, 8, and 10KHz bandwidth values. From these graphs it can now be seen that the performance indices do not approach an asymptotic limit as the bandwidth is increased; in fact, the optimal bandwidth would now appear to be somewhere between 2 and 4KHz. For bandwidths greater than 6KHz, the system becomes unstable which causes the performance indices to grow unbounded. These unbounded results are plotted as straight lines across the top of the graphs. It should also be noted that the graphics system used to plot this data, and all succeeding data like it, automatically connects the discrete data points with continuous lines to make interpretation of the data easier; Appendix C contains the detailed tabular data for this analysis should any questions arise from its graphical representation.

To more narrowly define the optimal bandwidth, the previous simulation was repeated, except that the bandwidths used for this run were 1, 2, 3, 4, and 5KHz. In this case, the results of the analysis with each performance index, as summarized here, differed as to where they achieved their minimum, or optimal, value.

<table>
<thead>
<tr>
<th>Performance Index</th>
<th>Minimum Value Location</th>
</tr>
</thead>
<tbody>
<tr>
<td>ISE</td>
<td>5KHz and 60 Degrees</td>
</tr>
<tr>
<td>IAE</td>
<td>4KHz and 45 Degrees</td>
</tr>
<tr>
<td>ITAE</td>
<td>3KHz and 45 Degrees</td>
</tr>
</tbody>
</table>

Table 4.1. Optimum Values of Performance Measures.

This data is graphically plotted in Figures 4.8, 4.9, 4.10 while the original simulation output is contained in Appendix D.

The difference in the results for each index is expected when one
Figure 4.8. ISE For A System With A 5KHz Actuator Resonance For 1, 2, 3, 4, And 5KHz Bandwidths
Figure 4.9. IAE For A System With A 5KHz Actuator Resonance For 1, 2, 3, 4, And 5KH Bandwidths
Figure 4.10. ITAE For A System With A 5KHz Actuator Resonance For 1, 2, 3, 4, And 5KHz Bandwidths
considers how each is computed and how the nature of the response of this system affects this computation. The major difference between these indices is how they weight each portion of the error based on its location in the time interval used to compute the index. By squaring the loop error, the ISE index tends to weight the initial error more heavily because the initial transients tend to be the largest in magnitude. On the other hand, the ITAE index multiplies the absolute error by time, which linearly increases the impact of the errors that occur further away from the origin. The IAE index, by contrast, does not inherently weight any portion of the error more than another. Therefore, as the bandwidth of this system is increased towards the point where the system goes unstable, the initial transient error is reduced while the ringing increases. This leads to the results seen in this simulation - the ISE index picks a higher bandwidth system as being optimal, while the ITAE index selects a lower bandwidth system as optimal, and the IAE index falls somewhere in between [14].

More importantly the ISE index can actually choose an unstable system as the optimal one. Figure 4.11 shows the simulated system error in response to a .5V step input for the bandwidth and phase margin values, 5KHz and 60 degrees, which gave the minimum ISE value even though the ringing in the response is clearly growing. The fact that the ringing in the closed loop error is increasing in magnitude indicates that the system is unstable for this bandwidth and phase calculation.

It must be noted, however, that the previous discussion of the results and relative merits of each performance indice is based solely on the 10 millisecond time interval chosen to compute the index for this study. For
Figure 4.11. The Closed Loop System Error In Response To A Step Input For 1, 2, 3, 4, And 5KHz Bandwidths
longer time intervals it is reasonable to expect the results of the indices would converge to the same choice for an optimal system. As an example, the ISE index would not have calculated the 5KHz bandwidth as the minimum value if the time interval had been long enough to include in the computation more of the growing oscillation in the system at this bandwidth. The point here is the choice of the computation interval is a very important consideration and in most cases the longest time interval possible will provide the best results. Unfortunately, practical considerations like computer simulation runtimes and test equipment capabilities are usually the limiting factors when selecting the maximum time interval to be used.

Given the previous analysis of the simulation results, as well as the stated goal of keeping the system bandwidth at the minimum necessary for an optimal system, the logical selection of an optimal bandwidth and phase margin would be those determined by the ITAE index. The next section will look at the performance of the hardware prototype set to these optimal values determined from the simulation study. In addition, because of the important effect it has on overall system performance, the hardware system will be tested to determine its maximum stable bandwidth.

4.4 Comparison of Hardware and Software Simulation Results

Up to now, this project has dealt with the theoretical optimization of the focus servo through the use of simulation models. For this work to be of use in the design of a practical system, it must accurately reflect the performance of the system when it is implemented in hardware. This section will look at the performance of the focus servo hardware that is
based on the analysis in Chapters 3 and 4 and compare the measured results with the simulation results.

The first step, before measuring the performance indices of the hardware system, is to verify that the frequency response of the system is close to the simulated response, especially in the region where the bandwidth, phase margin, and gain margin are measured. These measurements are necessary to confirm that the system performs as expected and will be stable enough to allow further testing. The results of these frequency response measurements are shown in Figures 4.12 to 4.23, along with the corresponding simulation data, for two cases. The first set of data is for the optimal bandwidth and phase margin settings determined from the simulation study while the second set is for the maximum stable bandwidth setting determined through hardware testing. The open loop frequency response calculated from the measured closed loop frequency response is only shown between 1KHz and 10KHz because the dynamic range of the DSA used was insufficient to be able to accurately make the calculation over the full frequency range of 10Hz to 100KHz shown for the simulation results.

The correlation between the hardware measurements and simulation results for the frequency response of the system at the optimal and maximum bandwidth settings is very good, as Figures 4.12 to 4.23 clearly illustrate. Furthermore, both the hardware and simulation data agree that the maximum stable open loop bandwidth is between 4 and 5KHz. It should be noted that the hardware prototype was very near instability at the 4KHz point and the system would at times move into the unstable region and
Figure 4.12. Simulated Open Loop Frequency Response For 3KHz Bandwidth And 45° Phase Margin
Figure 4.13. Hardware Open Loop Frequency Response For 3KHz Bandwidth And 45° Phase Margin
Figure 4.14. Simulated Closed Loop Frequency Response For 3KHz Bandwidth And 45° Phase Margin
Figure 4.15. Hardware Closed Loop Frequency Response For 3KHz Bandwidth And 45° Phase Margin
Figure 4.16. Simulated Compensation Stage Frequency Response For 3KHz Bandwidth And 45° Phase Margin
Figure 4.17. Hardware Compensation Stage Frequency Response For 3KHz Bandwidth And 45° Phase Margin
Figure 4.18. Simulated Open Loop Frequency Response For 4KHz Bandwidth And 45° Phase Margin
Figure 4.19. Hardware Open Loop Frequency Response For 4KHz Bandwidth And 45° Phase Margin
Figure 4.20. Simulated Closed Loop Frequency Response For 4KHz Bandwidth And 45° Phase Margin
Figure 4.21. Hardware Closed Loop Frequency Response For 4KHz Bandwidth And 45° Phase Margin
Figure 4.22. Simulated Compensation Stage Frequency Response For 4KHz Bandwidth And 45° Phase Margin
Figure 4.23. Hardware Compensation Stage Frequency Response For 4KHz Bandwidth And 45° Phase Margin
oscillate. One possible explanation for this may be that the gain margin decrease that occurred, from 10dB to 6dB, as the bandwidth increased from 3KHz to 4KHz does not allow a sufficient margin for the normal parameter fluctuations in the system. In any case, the maximum practical operational bandwidth for the system is approximately 3KHz, which lends even more credibility to the choice of the ITAE results as the optimal system values.

The next step is to measure the performance indices for the 3KHz bandwidth setting on the hardware system for comparison with the simulated data. The simulated and measured data are plotted for all three indices in Figures 4.24, 4.25, and 4.26. In all three cases the hardware results showed greater sensitivity to changes in the phase margin than did the simulation results; there is no known reason for the disparity but the results do converge to a great extent in the region between 45 and 60 degrees phase margin. Appendix E contains the actual performance index data as measured and calculated using the waveform math capabilities of a Tektronix 7854 storage oscilloscope.

Ultimately, the focus servo performance criterion that matters most is how well the system keeps the laser spot focused on a spinning optical disk that has a finite amount of axial runout. Minimizing the focus error in this situation is, after all, the purpose of doing the performance optimization study.

The final step in this study, then is to measure the focus error for the optimized hardware prototype, due to the runout of the spinning disk, for comparison against the same data generated using the simulation model.
INTEGRAL OF THE ERROR SQUARED \( (10^{-5} \times V^2 \times S) \)

Figure 4.24. Measured Versus Simulated ISE For A 3KHz Open Loop Bandwidth
Figure 4.25. Measured Versus Simulated IAE For A 3KHz
Open Loop Bandwidth
INTEGRAL OF THE TIME * ABSOLUTE ERROR (10E-8 V*S**2)

PHASE MARGIN IN DEGREES
MEASURED VERSUS SIMULATION DATA FOR 3 KHZ BW

Figure 4.26. Measured Versus Simulated ITAE For A 3KHz Open-Loop Bandwidth
Figure 4.27. A Comparison Of The Measured And Simulated FES For The Disk 2B139 With A 3KHz Bandwidth And 45° Phase Margin
Figure 4.28. A Comparison Of the Measured And Simulated FES For The Disk 2B138 With A 3KHz Bandwidth And 45° Phase Margin
Figure 4.29. A Comparison Of The Measured And Simulated FES At The Preamplifier Stage Output.
Since the focus error cannot be measured directly in the physical system, the data was taken at the point nearest to where the positional error was generated, which in this case is the output of the preamplifier stage. The focus error was taken in both peak-to-peak (P-P) and RMS form to give both a "worst case" and "average" perspective on the data. This error, both measured and simulated, is summarized in Figures 4.27 and 4.28 for two disks spinning at 2000, 3000, 4000, 5000, and 6000 RPM. In this case, the simulated and measured focus error results are very close to each other, especially for the RMS values. An example of the actual measured and simulated focus error data is shown in Figure 4.29; the complete set of this data can be found in Appendix F.
CHAPTER 5
CONCLUSIONS

The major emphasis, and hence the principal contribution, of this work is in the practical application of the standard performance indices ISE, IAE, and ITAE, to optimize the time domain performance of a closed loop position control servo. The optimization strategy itself was unique in that the performance indices, which are inherently time domain functions, were calculated over a range of the frequency domain parameters, open loop bandwidth and phase margin, instead of time domain parameters. The reason this approach is so useful is that it bridges the gap between the time domain, or "real" world, and the more easily analyzed, and physically measured, frequency domain. In the broader sense, this work is significant in that not only did it thoroughly explore the theoretical ramifications of using these indices to optimize the error reduction capability of a system but it also applied the results to improving the performance of an actual position control focus servo system, as well as validating the simulation results.

Some of the other key contributions of this project that allowed the optimization study to be undertaken are summarized as follows.

1. The analysis, characterization, and modeling of the focus actuator including the secondary mechanical resonances.

2. The design, modeling, and implementation of the servo gain and compensation stages.

3. Development of both time and frequency domain software models used in the simulation study.
4. The hardware implementation and testing of the optimized servo system that resulted from the software simulation study. As is discussed in Chapter 4, the correlation between the software simulation results and the hardware implementation results is very good.

Several conclusions can be drawn from this work and one of the most important is that the choice of a performance index to be used in an optimization study must take into consideration the fact that each index can result in dramatically different optimum parameter values from the same error data. This was illustrated very clearly with the different optimal open loop bandwidth values calculated by each of the indices in the simulation results as summarized in Figure 5.1, in this case 3KHz for the ITAE, 4KHz for the IAE, and 5KHz for the ISE. Even though this discrepancy in results stemming from the indices is well understood [1], the serious practical implications become obvious when one considers that the results measured with the hardware system indicate that not only is 3KHz the true optimal band width for the system but the system was unstable at 4KHz and 5KHz bandwidths. The reason the ITAE simulation was more accurate in predicting the true optimum bandwidth value for this system is that it weighted the ringing in the focus error after the initial transient more heavily than did the other indices. In general, the ITAE index is more sensitive to the oscillations that can occur in second order, or greater, systems as they near instability and is therefore preferable to the other indices in the analysis of these types of systems. Since most physical control systems are of second order or greater, the ITAE index would appear to be the most useful for practical studies. Figure 5.1 also illustrates that
Figure 5.1. Comparison Of Normalized Performance Indices
the selectivity of the ITAE index is better than that for the IAE and ISE indices, although not nearly as much as previous work would have suggested [14].

One question that should be addressed as a result of this work is whether the rigorous analysis and simulation phase of the optimization study was worth the effort? Would it have been more productive to skip the software development stage and have done the optimization study directly on the hardware prototype? The answer based on the results of this study is emphatically no, even if one ignores the practical problems associated with considerations such as the availability of the prototype for testing, the tedious nature of measuring the performance indices on the hardware, etc. The software model allowed for the testing of options that could not have been explored on the physical system because they either could have potentially damaged the hardware or were not physically realizable at all. One example of this is the calculation of the performance indices for a system without a 5KHz resonance in the actuator to determine the effect this imperfection had on overall system performance, something that would have been impossible to investigate otherwise.

As interesting and successful as this project has been in the use of various performance indices to optimize the design of an optical disk focus servo, many more possible avenues of research were raised by this work. One logical follow on would be to examine adaptive control techniques for potential application to this system to allow it to continually optimize its performance under dynamically varying conditions. Another area that holds tremendous potential for improving the performance is to use
feedforward compensation to reduce the effect of the repetitive disk axial runout. Also, with the continued improvement of digital signal processors (DSP), digital implementation of the control electronics becomes more attractive.

Finally, any discussion of further research would be incomplete without mentioning the fact that the optimization of this servo design could have been accomplished using a state variable model approach, as opposed to the frequency domain model approach chosen. The state variable technique has many advantages, most notably its ability to handle time varying, nonlinear, and multivariable systems; the author's decision to utilize the frequency domain techniques documented in previous chapters was based on several factors. It is however recognized that an optimization study of this system based on a state model would be a very interesting exercise.
APPENDIX A

SOURCE CODE FOR DSL SIMULATION MODELS

Note: These source code listings have been edited to eliminate redundant code for the sake of brevity. They are included only for purposes of illustration.
TITLE  FOCUS SERVO FREQUENCY RESPONSE MODEL

* FILE NAME:  FSFRQR1  DSL A1  (RESIDES IN VM2)

* DATE CODED:  NOVEMBER 28, 1985

* REVISED ON:  JANUARY 9, 1986

* WRITTEN BY:  JOSEPH H. DIMATTEO

* DESCRIPTION:  THIS PROGRAM WILL SIMULATE THE FREQUENCY RESPONSE OF THE
* FOCUS SERVO FOR A GIVEN OPEN LOOP BANDWIDTH AND PHASE
* MARGIN INCLUDING THE 5 KHZ MECHANICAL RESONANCE IN THE
* ACTUATOR.

* THE FOLLOWING ARE THE DEFINITIONS OF THE PARAMETERS:

* M:  FOCUS ACTUATOR'S MASS (KG)
* G:  ACCELERATION DUE TO GRAVITY (M/SEC**2)
* KI:  FOCUS ACTUATOR's CURRENT TO FORCE CONSTANT (NT/AMP)
* KV:  FOCUS ACTUATOR'S VELOCITY TO VOLTAGE CONSTANT (V/M/S)
* N:  FOCUS COIL RESISTANCE (OHMS)
* L:  FOCUS COIL INDUCTANCE (HENRYS)
* KS:  FOCUS SPRING CONSTANT (NT/M)
* KF:  FOCUS MECH. DAMPING CONSTANT (NT/M/SEC)
* KD:  DETECTOR GAIN CONSTANT (V/M)
* KP:  PREAmp GAIN CONSTANT
* KG:  LOOP GAIN ADJUSTMENT
* KA:  OPEN LOOP GAIN OF POWER AMPLIFIER
* RS:  CURRENT SENSE RESISTOR (OHMS)
* RPL1:  PHASE LEAD COMPENSATOR RESISTOR (OHMS)
* RPL2:  "   "   "   CAPACITOR (FARADS)
* CPL:  "   "   "   "   CAPACITOR (FARADS)
* RINT:  LOW FREQUENCY INTEGRATOR RESISTOR (OHMS)
* CINT:  "   "   "   CAPACITOR (FARADS)
* DCZ1:  MECHANICAL RESONANCE ZERoES DAMPING COEFFICIENT
* DCPI:  "   "   "   POLES  "   "
* WZ1:  "   "   "   ZERoES
* WPI:  "   "   "   POLES

* INITIAL SEGMENT

* FIXED  FLAG1,FLAG2

* COMPLEX S,G1,G2,G3,G4,G5,G6,G7,G8,G9,G10,G11,G20,G21,G22,G23,G24, ...
  H1,H2,H20,GCL,GOL

* CONTRL FNTIM=4.0
\[ W = 10 \cdot e^{(T I M E + 1.8)} \]

\[ F = W / (2. \cdot \pi) \]

\[ S = \text{CMPLX}(0., \omega) \]

\[ WZPL = FZPL / (2. \cdot \pi) \]

\[ WPPL = FPPL / (2. \cdot \pi) \]

\[ WZINT = FZINT / (2. \cdot \pi) \]

\[ D W = 10. \cdot e^{(T I M E + 1.8)} \]

\[ F = W / (2. \cdot \pi) \]

\[ S = \text{CMPLX}(0., \omega) \]

\[ WZPL = FZPL / (2. \cdot \pi) \]

\[ WPPL = FPPL / (2. \cdot \pi) \]

\[ WZINT = FZINT / (2. \cdot \pi) \]

The following are the S-domain transfer functions for each block in the frequency domain model.

The detector and preamp stages

\[ G_1 = K_D \]

\[ G_2 = K_P \]

The variable gain stage

\[ G_3 = K_G \cdot (7.5) \]

The double pole lowpass filter (not used)

\[ G_4 = 1.0 \]

The phase lead compensator

\[ G_5 = (S + WZPL) / (S + WPPL) \cdot (WPPL / WZPL) \]

The low frequency integrator

\[ G_6 = (S + WZINT) / S \]

The overall compensation stage

\[ G_7 = G_4 \cdot G_5 \cdot G_6 \]

\[ G_8 = G_7 \cdot K_A \]

The actuator model

\[ G_8 = 1.0 / (R + S \cdot \omega) \]

\[ G_9 = K_I \]

\[ G_{10} = (1./\omega)^2 + (K_F / M)^2 + (K_S / M) \]

\[ G_{11} = (S^2) / (2. \cdot DCP1 \cdot WP1 + WZ1^2) \]
H1=S*K
H2=R
G20=G1*G2*G3*G4*G5*G6
H20=H2/(G9*G10*G11)
G22=G21/(1.+G21*H20)
G23=(G7*G22)/(1.+G7*G22*H20)

* The overall closed loop transfer function

GCL=(G20*G23)/(1.+G20*G23)

* The overall open loop transfer function

GOL=G20*G23

MAGOL=20.*GAIN(GOL)
PHASOL=RADEG*PHASE(0.0,GOL)
MAGCL=20.*GAIN(GCL)
PHASCL=RADEG*PHASE(0.0,GCL)
MAGCMP=20.*GAIN(G24)
PHSCMP=RADEG*PHASE(0.0,G24)

TERMINAL SEGMENT

IF(KSIM.LT.4) CALL RERUN
PRINT FPFL,FPPL,FZINT,CPL,RPL2,RINT,KG
PRINT 0.1,F,W,MAGOL,PHASOL,MAGCL,PHASCL
SAVE 0.001,F,W,MAGOL,PHASOL,MAGCL,PHASCL,MAGCMP,PHSCMP
SAVE 0.001,F,W,MAGOL,PHASOL

GRAPH (G1,DE=GA3277,PO=0,1) F(AX=LOG,NI=4,LO=10, TI=2.5, UN=‘HZ’), ...
MAGOL(NI=8,TI=.75,UN=DB,LI=1, RU=*)
GRAPH (G2,DE=GA3277,OV) F(AX=LOG,NI=4,LO=10, TI=2.5, PO=0.0,7, UN=H(Z), ...
PHASOL(PO=10.0,1,TI=.75, UN=DEG, LI=4,NI=8, RU=*)
LABEL (G1) FOCUS SERVO FREQUENCY RESPONSE - OPEN LOOP
GRAPH (G3,DE=GA3277,PO=0,1) F(AX=LOG,NI=4,LO=10, TI=2.5, UN=’HZ’), ...
MAGCL(NI=8,TI=.75,UN=DB,LI=1, RU=*)
GRAPH (G4,DE=GA3277,OV) F(AX=LOG,NI=4,LO=10, TI=2.5, PO=0.0,7, UN=H(Z), ...
PHASCL(PO=10.0,1,TI=.75, UN=DEG, LI=4,NI=8, RU=*)
LABEL (G3) FOCUS SERVO FREQUENCY RESPONSE - CLOSED LOOP
GRAPH (G5,DE=GA3277,PO=0,1) F(AX=LOG,NI=4,LO=10, TI=2.5, UN=’HZ’), ...
MAGCMP(NI=8,TI=.75,UN=DB,LI=1, RU=*)
GRAPH (G6,DE=GA3277,OV) F(AX=LOG,NI=4,LO=10, TI=2.5, PO=0.0,7, UN=H(Z), ...
PHSCMP(PO=10.0,1,TI=.75, UN=DEG, LI=4,NI=8, RU=*)
LABEL (G5) FOCUS SERVO COMPENSATION NETWORK FREQUENCY RESPONSE
LABEL (G1,G3,G5) OPEN LOOP PARAMETERS: BW=03000. PHASE MARGIN=45 DEGREES
LABEL (G1,G3,G5) FZINT=0.170 KHz  FZPL=1.182 KHz  FPPL=07.016 KHz
END
STOP
TITLE  FOCUS SERVO TIME RESPONSE MODEL-STEP RESPONSE VERSION  
FILE NAME:  FSSTPR2 DSL A1 (RESIDES IN VM2)  
DATE CODED:  NOVEMBER 11, 1986  
REVISED ON:  APRIL 12, 1986  
WRITTEN BY:  JOSEPH H. DIMATTEO  
DESCRIPTION:  THIS PROGRAM WILL SIMULATE THE TIME DOMAIN STEP RESPONSE  
OF THE FOCUS SERVO WITH AN ACTUATOR HAVING A 5KHZ  
MECHANICAL RESONANCE FOR OPEN LOOP SERVO BANDWIDTHS OF  
1, 2, 3, 4, AND 5KHZ.  

THE FOLLOWING ARE THE DEFINITIONS OF THE PARAMETERS:  
M:  FOCUS ACTUATOR'S MASS (KG)  
G:  ACCELERATION DUE TO GRAVITY (M/SEC**2)  
KI:  FOCUS ACTUATOR'S CURRENT TO FORCE CONSTANT (NT/AMP)  
KV:  FOCUS ACTUATOR'S VELOCITY TO VOLTAGE CONSTANT (V/M/S)  
R:  FOCUS COIL RESISTANCE (OHMS)  
L:  FOCUS COIL INDUCTANCE (HENRYS)  
KS:  FOCUS SPRING CONSTANT (NT/M)  
KF:  FOCUS MECH. DAMPING CONSTANT (NT/M/SEC)  
RS:  POWER AMPLIFIER CURRENT SENSE RESISTOR (OHMS)  
KA:  POWER AMPLIFIER OPEN LOOP GAIN  
FPPL: PHASE LEAD COMPENSATOR POLE (HZ)  
FZPL:  " " ZERO (HZ)  
FZINT:  LOW FREQUENCY INTEGRATOR ZERO (HZ)  
DCZ1: MECHANICAL RESONANCE ZERO DAMPING COEFFICIENT  
DZ1:  " " POLE " "  
WP1:  " " POLE  

INITIAL SEGMENT  
CTRL FINTIM=10.0D-3,DELT=1.0D-5,DELPRT=1.0D-5,DELPRT=10.0D-3  
METHOD KSFX  
LENGTH 1500  
FIXED FLAG1,FLAG2,FLAG3,FLAG4,FLAG5,FLAG6,FLAG7,FLAG8,FLAG9,FLAG10  
ARRAY KG1(5),FZPL1(5),FPPL1(5),FZINT1(5)  
ARRAY KG2(5),FZPL2(5),FPPL2(5),FZINT2(5)  
ARRAY KG3(5),FZPL3(5),FPPL3(5),FZINT3(5)  
ARRAY KG5(5),FZPL5(5),FPPL5(5),FZINT5(5)
ARRAY KG6(5),FZPL6(5),FPPL6(5),FZINT6(5)
ARRAY KG7(5),FZPL7(5),FPPL7(5),FZINT7(5)
ARRAY KG8(5),FZPL8(5),FPPL8(5),FZINT8(5)
ARRAY KG9(5),FZPL9(5),FPPL9(5),FZINT9(5)
ARRAY A(3),B(3)

A(1) = 1.0
A(2) = 2.0 * DCZ1 * WZ1
A(3) = WZ1 ** 2.0
B(1) = 1.0
B(2) = 2.0 * DCPL1 * WP1
B(3) = WP1 ** 2.0

CONST
M = 0.00095, KS = 380., KF = 2.0, KV = 2.0, L = 18.0D-6, R = 1.8, ...
RS = 1.0, K = 1.0D5, G = 9.806, KD = 17.0D3, RG1 = 09.0D3, RF1 = 100.0D3, ...
RG2 = 10.0D3, RF2 = 27.6D3, KD = 1.4673D6, N2 = 0.0, N2 = 1.0, ...
FESUM2 = 0.0, X0 = 0.0, ...
FESJ2 = 0.0, ...
DCZ1 = 0.008, DGPI = 0.01D0, WP1 = 031400., WZ1 = 032500., ...
FLAG1 = 0, FLAG2 = 0, FLAG3 = 0, FLAG4 = 0, FLAG5 = 0, FLAG6 = 0, FLAG7 = 0, ...
FLAG8 = 0, FLAG9 = 1, FLAG10 = 0

* Program operation is controlled using the flags defined below (1 = flag on, 0 = flag off):
*
* FLAG1 - USE THE 20 DEGREE PHASE MARGIN COMPENSATION DATA
* FLAG2 - USE THE 25 DEGREE PHASE MARGIN COMPENSATION DATA
* FLAG3 - USE THE 30 DEGREE PHASE MARGIN COMPENSATION DATA
* FLAG4 - USE THE 35 DEGREE PHASE MARGIN COMPENSATION DATA
* FLAG5 - USE THE 40 DEGREE PHASE MARGIN COMPENSATION DATA
* FLAG6 - USE THE 45 DEGREE PHASE MARGIN COMPENSATION DATA
* FLAG7 - USE THE 50 DEGREE PHASE MARGIN COMPENSATION DATA
* FLAG8 - USE THE 55 DEGREE PHASE MARGIN COMPENSATION DATA
* FLAG9 - USE THE 60 DEGREE PHASE MARGIN COMPENSATION DATA
* FLAG10 - USE CONSTANT CURRENT SOURCE POWER AMP CONFIGURATION

* The focus error signal is developed using a four quadrant photodetector assembly numbered as shown at left. The following curves give the current out of each quadrant for a given focus error.
* * *
** DETECTOR QUADRANT 1 DEFOCUS-TO-CURRENT CURVE, SLED 2 **
* AFGEN KD1=-.3000E-04, 0.0000019059, ...
   -.2000E-04, 0.0000029889, ...
   -.1000E-04, 0.0000077537, ...
   .0001E-12, 0.0000170810, ...
   .1000E-05, 0.0000168250, ...
   .1000E-04, 0.0000140340, ...
   .2000E-04, 0.0000121750, ...
   .3000E-04, 0.0000059218
* * *
** DETECTOR QUADRANT 2 DEFOCUS-TO-CURRENT CURVE, SLED 2 **
* AFGEN KD2=-.3000E-04, 0.0000086315, ...
   -.2000E-04, 0.0000113260, ...
   -.1000E-04, 0.0000147220, ...
   .0001E-12, 0.0000161880, ...
   .1000E-04, 0.00000670815, ...
   .2000E-04, 0.0000030223, ...
   .3000E-04, 0.0000024924
* * *
** DETECTOR QUADRANT 3 DEFOCUS-TO-CURRENT CURVE, SLED 2 **
* AFGEN KD3=-.3000E-04, 0.0000019088, ...
   -.2000E-04, 0.0000031839, ...
   -.1000E-04, 0.0000058003, ...
   .0001E-12, 0.0000153110, ...
   .1000E-04, 0.0000145840, ...
   .2000E-04, 0.0000126660, ...
   .3000E-04, 0.0000102360
* * *
** DETECTOR QUADRANT 4 DEFOCUS-TO-CURRENT CURVE, SLED 2 **
* AFGEN KD4=-.3000E-04, 0.0000086315, ...
   -.2000E-04, 0.0000113260, ...
   -.1000E-04, 0.0000147240, ...
   .0001E-12, 0.0000161880, ...
   .1000E-04, 0.0000070830, ...
   .2000E-04, 0.0000030223, ...
   .3000E-04, 0.0000024918
* * *
** USE THE COMPENSATION VALUES FOR 20 DEGREES PHASE MARGIN IF **
** FLAG1 EQUALS ONE **
* TABLE KG1(1)=0.019,KG1(2)=0.045, ...
   KG1(3)=0.100,KG1(4)=0.190, ...
   KG1(5)=0.275, ...
   FZPL(1)=0.675D3, ...
   FZPL(2)=1.359D3, ...
   FZPL(3)=2.023D3, ...
   FZPL(4)=2.698D3, ...
   FZPL(5)=3.373D3,
FPPL1(1)=01.483D3, ...
FPPL1(2)=02.965D3, ...
FPPL1(3)=04.448D3, ...
FPPL1(4)=05.930D3, ...
FPPL1(5)=07.413D3, ...
FZINT1(1)=0.110D3, ...
FZINT1(2)=0.150D3, ...
FZINT1(3)=0.170D3, ...
FZINT1(4)=0.210D3, ...
FZINT1(5)=0.250D3

* USE THE COMPENSATION VALUES FOR 25 DEGREES PHASE MARGIN IF
* FLAG2 EQUALS ONE

TABLE KG2(1)=0.01, KG2(2)=0.04, ...
KG2(3)=0.09, KG2(4)=0.170, ...
KG2(5)=0.250, ...
FZPL2(1)=0.613D3, ...
FZPL2(2)=1.226D3, ...
FZPL2(3)=1.838D3, ...
FZPL2(4)=2.451D3, ...
FZPL2(5)=3.064D3, ...
FPPL2(1)=01.632D3, ...
FPPL2(2)=03.264D3, ...
FPPL2(3)=04.896D3, ...
FPPL2(4)=06.527D3, ...
FPPL2(5)=08.159D3, ...
FZINT2(1)=0.110D3, ...
FZINT2(2)=0.150D3, ...
FZINT2(3)=0.170D3, ...
FZINT2(4)=0.210D3, ...
FZINT2(5)=0.250D3

* USE THE COMPENSATION VALUES FOR 30 DEGREES PHASE MARGIN IF
* FLAG3 EQUALS ONE

TABLE KG3(1)=0.01, KG3(2)=0.040, ...
KG3(3)=0.08, KG3(4)=0.155, ...
KG3(5)=0.23, ...
FZPL3(1)=0.554D3, ...
FZPL3(2)=1.109D3, ...
FZPL3(3)=1.662D3, ...
FZPL3(4)=2.217D3, ...
FZPL3(5)=3.064D3, ...
FPPL3(1)=01.804D3, ...
FPPL3(2)=03.608D3, ...
FPPL3(3)=05.412D3, ...
FPPL3(4)=07.216D3, ...
FPPL3(5)=09.020D3, ...
FZINT3(1)=0.110D3, ...
FZINT3(2)=0.150D3, ...
FZINT3(3)=0.170D3, ...
FZINT3(4)=0.210D3, ...
FZINT3(5)=0.250D3
USE THE COMPENSATION VALUES FOR 60 DEGREES PHASE MARGIN IF
FLAG9 EQUALS ONE

TABLE KG9(1)=0.005, KG9(2)=0.02, ...
    KG9(3)=0.04, KG9(4)=0.07, ...
    KG9(5)=0.105, ...
    FZPL9(1)=0.249D3, ...
    FZPL9(2)=0.499D3, ...
    FZPL9(3)=0.748D3, ...
    FZPL9(4)=0.997D3, ...
    FZPL9(5)=1.247D3, ...
    FPPL9(1)=0.04010D3, ...
    FPPL9(2)=0.08021D3, ...
    FPPL9(3)=0.12032D3, ...
    FPPL9(4)=0.16043D3, ...
    FPPL9(5)=0.20054D3, ...
    FZINT9(1)=0.0110D3, ...
    FZINT9(2)=0.150D3, ...
    FZINT9(3)=0.170D3, ...
    FZINT9(4)=0.210D3, ...
    FZINT9(5)=0.250D3

TAU=L/R
TAU2=L/(R+RS)

DYNAMIC SEGMENT

CONVERT THE 20 DEGREE PHASE MARGIN COMPENSATION POLE/ZERO
LOCATIONS TO RADIANS IF FLAG1 EQUALS ONE.

IF(FLAG1.EQ.1) THEN
    WPPL=2.0*PI*FPPL1(KSIM)
    WZPL=2.0*PI*FZPL1(KSIM)
    WZINT=2.0*PI*FZINT1(KSIM)
ELSE
ENDIF

CONVERT THE 25 DEGREE PHASE MARGIN COMPENSATION POLE/ZERO
LOCATIONS TO RADIANS IF FLAG2 EQUALS ONE.

IF(FLAG2.EQ.1) THEN
    WPPL2=2.0*PI*FPPL2(KSIM)
    WZPL2=2.0*PI*FZPL2(KSIM)
    WZINT2=2.0*PI*FZINT2(KSIM)
ELSE
ENDIF

CONVERT THE 60 DEGREE PHASE MARGIN COMPENSATION POLE/ZERO LOCATIONS TO RADIANS IF FLAG9 EQUALS ONE.

IF(FLAG9.EQ.1) THEN
WPPL=2.0*PI*FPPL9(KSIM)
WZPL=2.0*PI*FZPL9(KSIM)
WZINT=2.0*PI*FZINT9(KSIM)
ELSE
ENDIF

CALCULATE THE PERFORMANCE INDICES: ISE, IAE, ITAE

ABSERR=ABS(VT)
TABSER=TIME*ABSERR
ERRSQD=VT*VT
ISE=INTEGRAL(0.0,ERRSQD)
IAE=INTEGRAL(0.0,ABSERR)
ITAE=INTEGRAL(0.0,TABSER)

INPUT THE TEST VOLTAGE DISTURBANCE
FOCDIS=0.5*STEP(0.0D-02)

INPUT THE VIBRATION DISTURBANCE
VIBDIS=0.0D-12

INPUT THE DISK RUNOUT
RUNOUT=0.0D-12
OFFSET1=0.0D-12

DERIVATIVE SEGMENT

LOOP MODEL

Focus error is equal to the disk runout minus the actuator
displacement.

\[ \text{FESL} = \text{RUNOUT-XLIM} \]

\[ \text{FES} = \text{LIMIT}(-4.0D-6, 4.0D-6, \text{FESL}) \]

**Detector and preamp stage**

ID1 = NLFGEN(KD1, FES)

ID2 = NLFGEN(KD2, FES)

ID3 = NLFGEN(KD3, FES)

ID4 = NLFGEN(KD4, FES)

VD1 = RD * ID1

VD2 = RD * ID2

VD3 = RD * ID3

VD4 = RD * ID4

N = (RF1/RG1) * ((VD1 + VB3) - (VD2 + VD4))

D = (RF2/RG2) * (VD1 + VD2 + VD3 + VD4)

VP = 10.0 * (N/D)

\[ \text{VP} = \text{KD} \times \text{FES} \]

**Gain and offset stage**

USE THE APPROPRIATE GAIN SETTINGS FOR THE CHOSEN PHASE MARGIN

NOSORT

\[ \text{IF} (\text{FLAG1} \neq 1) \text{ THEN} \]

\[ \text{VG} = (\text{KG1(KSIM)} \times \text{VP} \times 7.5) + \text{OFSET1} \]

ELSE

ENDIF

\[ \text{IF} (\text{FLAG9} \neq 1) \text{ THEN} \]

\[ \text{VG} = (\text{KG9(KSIM)} \times \text{VP} \times 7.5) + \text{OFSET1} \]

ELSE

ENDIF

SORT

**Test disturbance and measurement stage**

VT = VG + VCDIS

**Lowpass filter stage**

VLP = VT * (WPPL/WZPL)

\[ \text{IF} (\text{FLAG1} \neq 1) \text{ THEN} \]

\[ \text{VG} = (\text{KG1(KSIM)} \times \text{VP} \times 7.5) + \text{OFSET1} \]

ELSE

ENDIF

\[ \text{IF} (\text{FLAG9} \neq 1) \text{ THEN} \]

\[ \text{VG} = (\text{KG9(KSIM)} \times \text{VP} \times 7.5) + \text{OFSET1} \]

ELSE

ENDIF

SORT

**Test disturbance and measurement stage**

VT = VG + VCDIS

**Lowpass filter stage**

VLP = VT * (WPPL/WZPL)
Phase lead compensator stage

\[ VPLL = \text{ZEROPL}(0.0, WZPL, WPPL, VLP) \]
\[ VPLL = \text{LIMIT}(-12.0, 12.0, VPLL) \]

Low frequency integrator stage

\[ VREFL = VPL \]
\[ VREF = \text{ZEROPL}(0.0, WZINT, 0.0, VPL) \]
\[ VREF = \text{LIMIT}(-12.0, 12.0, VREF) \]

If FLAG4 is set, then set up the power amplifier as a constant current source. Otherwise set it up as a constant voltage source.

NoSort

\[ \text{IF}(\text{FLAG10} = \text{EQ} 1) \text{THEN} \]
Sort

\[ VPA = VREFL - VS \]
\[ VPA^2 = VPA^2 \times KA \]
\[ VPA^2L = \text{LIMIT}(-12.0, 12.0, VPA^2) \]
\[ VC = VPA^2L - VS \]
NoSort

Else
Sort
\[ VC = VREFL - VS \]
NoSort

EndIf

Sort

Actuator stage

\[ VB = KV \times XD0T \]
\[ VC2 = VC - VB \]
\[ IR = VC2 / R \]
\[ IL = \text{REALPL}(0.0, TAU, IR) \]
\[ ILIM = \text{LIMIT}(-1.1, 1.1, IL) \]
\[ VS = ILIM \times RS \]
\[ FI = ILIM \times KI \]
\[ FTOTAL = FI + VIBDIS \]
\[ AF = FTOTAL / M \]
\[ X2DOT = AF + G - ((KF / M) \times XD0T) - ((KS / M) \times XLIM) \]
\[ XD0T = \text{INTGRL}(XD0TIC, X2DOT) \]
\[ X = \text{INTGRL}(XIC, XD0T) \]
\[ Y = \text{TRNFR}(2.2, Y0, A, 8, 1.0 \times X) \]
\[ XLIM = \text{LIMIT}(-400D-6, 400D-6, Y) \]
IF(KSIM.LT.5) CALL RERUN

SAVE (SUBSAV) FES,X2DOT,XDOT,X,RUNOUT,VG,VP2A,IL,VRFL,VC,...
IAE,ITAE,VP,VT,VPL,VP2D,ISE,ABSRK,ERRSQD
PRINT ISE,IAE,ITAE

GRAPH (G20/SUBSAV,DE=GA3277,PO=0,1,RU=1) ...
TIME(LO=0.000,SC=0.0005,NI=10,UN=SECONDS),...
VG(UN=VOLTS,LI=1,SC=0.2,LO=-1.4)
LABEL (G20) GAIN AND OFFSET STAGE OUTPUT VOLTAGE

GRAPH (G7/SUBSAV,DE=GA3277,PO=0,1,RU=1) ...
TIME(LO=0.000,SC=0.0005,NI=10,UN=SECONDS),...
ISE(UN='VOLTS**2*SECONDS',LI=1,SC=.000015,LO=0.0)
LABEL (G7) INTEGRAL OF THE ERROR SQUARED (ISE) INDEX

END
STOP
APPENDIX B
FOCUS SERVO CLOSED LOOP ERROR IN RESPONSE TO A STEP INPUT FOR 2KHz TO 10KHz OPEN LOOP BANDWIDTH WITHOUT 5KHz SYSTEM RESONANCE.
TEST SIGNAL STAGE OUTPUT VOLTAGE
PHASE MARGIN=45 DEGREES, OPEN LOOP BANDWIDTH=2, 4, 6, 8, 10 KHZ
FOCUS SERVO TIME DOMAIN RESPONSE FOR A .5 VOLT STEP INPUT AT T=0
INTEGRAL OF THE ERROR SQUARED (ISE) INDEX
PHASE MARGIN=45 DEGREES OPEN LOOP BANDWIDTH=2, 4, 6, 8, 10 KHZ
FOCUS SERVO TIME DOMAIN RESPONSE FOR A .5 VOLT STEP INPUT AT T=0
INTEGRAL OF THE ABSOLUTE ERROR (IAE) INDEX
PHASE MARGIN=45 DEGREES  OPEN LOOP BANDWIDTH=2,4,6,8,10 KHZ
FOCUS SERVO TIME DOMAIN RESPONSE FOR A .5 VOLT STEP INPUT AT T=0
INTEGRAL OF THE TIME * ABSOLUTE ERROR (ITAE) INDEX
PHASE MARGIN=45 DEGREES OPEN LOOP BANDWIDTH=2, 4, 6, 8, 10 KHz
FOCUS SERVO TIME DOMAIN RESPONSE FOR A .5 VOLT STEP INPUT AT T=0
APPENDIX C

FOCUS SERVO CLOSED LOOP ERROR IN RESPONSE TO A STEP INPUT FOR 2KHz TO 10KHz OPEN LOOP BANDWIDTH WITH 5KHz SYSTEM RESONANCE.
TEST SIGNAL STAGE OUTPUT VOLTAGE
PHASE MARGIN=45  OPEN LOOP BANDWIDTH=2,4,6,8,10 KHZ  5KHZ RESONANCE
FOCUS SERVO TIME DOMAIN RESPONSE FOR A .5 VOLT STEP INPUT AT T=0
INTEGRAL OF THE ERROR SQUARED (I^2SE) INDEX

PHASE MARGIN=45° OPEN LOOP BANDWIDTH=2, 4, 6, 8, 10 kHz 5kHz RESONANCE
FOCUS SERVO TIME DOMAIN RESPONSE FOR 0.5 VOLT STEP INPUT AT T=0
INTEGRAL OF THE ABSOLUTE ERROR (IAE) INDEX
PHASE MARGIN=45
OPEN LOOP BANDWIDTH=2, 4, 6, 8, 10 KHZ
5 KHZ RESONANCE
FOCUS SERVO TIME DOMAIN RESPONSE FOR A .5 VOLT STEP INPUT AT T=0
INTEGRAL OF THE TIME * ABSOLUTE ERROR (ITAE) INDEX
PHASE MARGIN=45° OPEN LOOP BANDWIDTH=2.4.6.8.10 KHZ 5KHZ RESONANCE
FOCUS SERVO TIME DOMAIN RESPONSE FOR A .5 VOLT STEP INPUT AT T=0
APPENDIX D
FOCUS SERVO CLOSED LOOP ERROR IN RESPONSE TO A STEP INPUT FOR 1KHz TO 5KHz OPEN LOOP BANDWIDTH WITH 5KHz SYSTEM RESONANCE.
TEST SIGNAL STAGE OUTPUT VOLTAGE
PHASE MARGIN=45° OPEN LOOP BANDWIDTH=1.2.3,4.5 KHZ 5 KHZ RESONANCE
FOCUS SERVO TIME DOMAIN RESPONSE FOR A .5 VOLT STEP INPUT AT T=0
INTEGRAL OF THE ERROR SQUARED (ISE) INDEX
PHASE MARGIN=45, OPEN LOOP BANDWIDTH=1, 2, 3, 4, 5 KHZ, 5 KHZ RESONANCE
FOCUS SERVO TIME DOMAIN RESPONSE FOR A .5 VOLT STEP INPUT AT T=0.
INTEGRAL OF THE ABSOLUTE ERROR [IAE] INDEX
PHASE MARGIN=45° OPEN LOOP BANDWIDTH=1, 2, 3, 4, 5 kHz 5 kHz RESONANCE
FOCUS SERVO TIME DOMAIN RESPONSE FOR A 5 VOLT STEP INPUT AT T=0
INTEGRAL OF THE TIME * ABSOLUTE ERROR (ITAE) INDEX
PHASE MARGIN=45 OPEN LOOP BANDWIDTH=1,2,3,4,5 KBHz RESONANCE
FOCUS SERVO TIME DOMAIN RESPONSE FOR A .5 VOLT STEP INPUT AT T=0
APPENDIX E

THE MEASURED RESPONSE OF THE HARDWARE PROTOTYPE SYSTEM WITH A 3kHz OPEN LOOP BANDWIDTH
Measured Performance Indices With 3KHz Open Loop Bandwidth From 20° To 60° Phase Margin

1. 60°

\[ e(t) \]

\[ ISE \]
Measured Performance Indices With 3KHz Open Loop Bandwidth From 20° To 60° Phase Margin

1. 60° (cont.)
Measured Performance Indices With 3KHz Open Loop Bandwidth From 20° To 60° Phase Margin

2. 55°
Measured Performance Indices With 3KHz Open Loop Bandwidth From 20° To 60° Phase Margin

2. 55° (cont.)
Measured Performance Indices With 3KHz Open Loop Bandwidth From 20° To 60° Phase Margin

3. 50°
Measured Performance Indices With 3KHz Open Loop Bandwidth From 20° To 60° Phase Margin

3. 50° (cont.)
Measured Performance Indices With 3KHz Open Loop Bandwidth From 20° To 60° Phase Margin

4. 45°
Measured Performance Indices With 3KHz Open Loop Bandwidth From 20° To 60° Phase Margin

4. 45° (cont.)
Measured Performance Indices With 3KHz Open Loop Bandwidth From 20° To 60° Phase Margin

5. 40°
Measured Performance Indices With 3KHz Open Loop Bandwidth From 20° To 60° Phase Margin

5. 40° (cont.)
Measured Performance Indices With 3KHz Open Loop Bandwidth From 20° To 60° Phase Margin

6. 35°
Measured Performance Indices With 3KHz Open Loop Bandwidth From 20° To 60° Phase Margin

6. 35° (cont.)

IAE

ITAE
Measured Performance Indices With 3KHz Open Loop Bandwidth From 20° To 60° Phase Margin

7. 30°
Measured Performance Indices With 3KHz Open Loop Bandwidth From 20° To 60° Phase Margin

7. 30° (cont.)
Measured Performance Indices With 3KHz Open Loop Bandwidth From 20° To 60° Phase Margin

8. 25°
Measured Performance Indices With 3KHz Open Loop Bandwidth From 20° To 60° Phase Margin

8. 25° (cont.)
Measured Performance Indices With 3KHz Open Loop Bandwidth From 20° To 60° Phase Margin

9. 20°
Measured Performance Indices With 3KHz Open Loop Bandwidth From 20° To 60° Phase Margin

9. 20° (cont.)
APPENDIX F

MEASURED VERSUS SIMULATED FOCUS ERROR DATA (FES) DATA
FES Measured At The Output Of the Preamplifier Stage For Disk 2B139 Spinning At 2000, 3000, 4000, 5000, And 6000 RPM.

1. 2000 RPM

![Graph for 2000 RPM]

2. 3000 RPM

![Graph for 3000 RPM]
FES Measured At The Output Of the Preamplifier Stage For Disk 2B139 Spinning At 2000, 3000, 4000, 5000, And 6000 RPM.

3. 4000 RPM

![Graph for 4000 RPM]

4. 5000 RPM

![Graph for 5000 RPM]
FES Measured At The Output Of the Preamplifier Stage For Disk 2B139 Spinning At 2000, 3000, 4000, 5000, And 6000 RPM.

5. 6000 RPM
FES measured at the output of the preamplifier stage for disk 2E138 spinning at 2000, 3000, 4000, 5000, and 6000 RPM.

1. 2000 RPM

2. 3000 RPM
FES Measured At The Output Of the Preamplifier Stage For Disk 2E138 Spinning At 2000, 3000, 4000, 5000, And 6000 RPM.

3. 4000 RPM

4. 5000 RPM
FES Measured At The Output Of the Preamplifier Stage For Disk 2E138 Spinning At 2000, 3000, 4000, 5000, And 6000 RPM.

5. 6000 RPM
FOCUS ERROR IN METERS
RPM=2000  BW=3000 HZ  PHASE MARGIN=45 DEGREES
FOCUS SERVO TIME DOMAIN RESPONSE FOR DISK 28139
FOCUS ERROR IN METERS
RPM=3000  BW=3000 Hz  PHASE MARGIN=45 DEGREES
FOCUS SERVO TIME DOMAIN RESPONSE FOR DISK 2B139
FOCUS ERROR IN METERS

RPM=4000  BW=3000 Hz  PHASE MARGIN=45 DEGREES

FOCUS SERVO TIME DOMAIN RESPONSE FOR DISK 28139
FOCUS ERROR IN METERS
RPM=5000  BW=3000  HZ  PHASE MARGIN=45 DEGREES
FOCUS SERVO TIME DOMAIN RESPONSE FOR DISK 26139
FOCUS ERROR IN METERS

RPM=6000  BW=3000 Hz  PHASE MARGIN=45 DEGREES

FOCUS SERVO TIME DOMAIN RESPONSE FOR DISK 28139
FOCUS ERROR IN METERS

RPM=2000  BW=3000 Hz  PHASE MARGIN=45 DEGREES

FOCUS SERVO TIME DOMAIN RESPONSE FOR DISK 2E138
FOCUS ERROR IN METERS
RPM = 3000  BU = 3000  Hz  PHASE MARGIN = 45 DEGREES
FOCUS SERVO TIME DOMAIN RESPONSE FOR DISK ZE139
FOCUS ERROR IN METERS
RPM=5000  BW=3000 Hz  PHASE MARGIN=45 DEGREES
FOCUS SERVO TIME DOMAIN RESPONSE FOR DISK 2E138
FOCUS ERROR IN METERS
RPM-4000  BW-3000  HZ  PHASE MARGIN-45 DEGREES
FOCUS SERVO TIME DOMAIN RESPONSE FOR DISK 2E138
FOCUS ERROR IN METERS
RPM=6000  BW=3000 Hz PHASE MARGIN=45 DEGREES
FOCUS SERVO TIME DOMAIN RESPONSE FOR DISK 2E138
REFERENCES


9. U.S. Patent 4,439,848

10. U.S. Patent 4,446,546

11. U.S. Patent 4,541,084

12. U.S. Patent 4,544,837


