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Pointing Control System for a High Precision Flight Telescope

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Pointing Control System For a High Precision Flight Telescope

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Abstract

A pointing control system is developed and tested for a flying gimballed telescope. The two-axis pointing system is capable of sub-microradian pointing stability and high accuracy in the presence of large host vehicle jitter. The telescope also has high agility—it is capable of a 50-degree retarget (in both axes simultaneously) in less than 2 seconds. To achieve the design specifications, high-accuracy, high-resolution, two-speed resolvers were used, resulting in gimbal-angle measurements stable to 1.5 microradians. In addition, on-axis inertial angle displacement sensors were mounted on the telescope to provide host-vehicle jitter cancellation. The inertial angle sensors are accurate to about 100 nanoradians, but do not measure low frequency displacements below 2 Hz. The gimbal command signal includes host-vehicle attitude information, which is band-limited. This provides jitter data below 20 Hz, but includes a variable latency between 15 and 25 milliseconds. One of the most challenging aspects of this design was to combine the inertial-angle-sensor data with the less perfect information in the command signal to achieve maximum jitter reduction. The optimum blending of these two signals, along with the feedback compensation were designed using Quantitative Feedback Theory.

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Introduction

The two-axis gimballed telescope is shown below in Figure 1. This project was funded to retrofit the telescope (which was built over ten years ago) to incorporate the latest advances in servo technology and thereby achieve an "order-of-magnitude" improvement in pointing accuracy and line-of-sight stability. There was little documentation on the existing control system. However, input/output characterization determined that it was lightly damped, and limited in disturbance rejection capability. Furthermore, the servo was velocity limited to $2.5^\circ/\text{sec}$, which severely restricted its step response for large re-targeting maneuvers. The telescope was retrofitted with precision resolvers and new (higher torque) motors. On-gimbal Inertial Angle Sensors were also added to enhance inertial pointing stability. A servo system was developed that blends feedback from the resolvers together with the inertial angle sensors to achieve less than one microradian line-of-sight pointing stability, ± 1.5 microradian pointing resolution and ± 30 -microradian accuracy.

This paper will describe in considerable detail, the new servo system including each sub-system, namely: the gimbal (including the motor, and bearing friction), Inertial Angle Sensors (IAS), resolvers (including the resolver-to-digital converters), Inertial Measurement Unit (IMU), feedback compensator, feedback sensor blending compensator, motor power amplifiers, and IAS notch filters. It also contains a brief tutorial on Quantitative Feedback Theory (QFT) which was used throughout the servo design. It further describes the servo modes of operation, namely: step, course, and fine pointing modes. The performance of the refurbished telescope is also presented.

Figure 1. Telescope gimbal

Two separate models were used depending on the operating mode of the telescope. For large re-targeting maneuvers, the telescope is a *Multiple-Input Multiple-Output* (MIMO) servo system with significant non-linear interactions between the two gimbal axes. However, for small maneuvers, gimbal interactions can be ignored—which reduces the model to two de-coupled *Multiple-Input Single-Output* (MISO) systems. The two inputs are: the reference input “R” and the disturbance “D,” while the output is the inertial pointing angle “Y.” (Although, the gimbal angle “ θ ,” is also an output, it will not be controlled independently of “Y.”) Axis de-coupling is possible because both axes of the telescope are balanced—the center of mass is very close to the center of rotation.

A simplified flow graph of the servo system is shown below in Figure 2. While only one axis is shown, with a few parameter substitutions the flow diagrams for both are identical. In Figure 2, and throughout this study, the telescope angle definitions are as follows. The gimbal angle of either axis with respect to its base is referred to as “ θ ,” the inertial gimbal angle is “Y,” and the angular position of the gimbal base is considered a disturbance and is therefore labeled “D.” The relationship between these three angles is: $\theta = Y - D$.

Figure 2. Single-axis (de-coupled) servo model

The “plant” in Figure 2, labeled $P(s)$, models the effect of gimbal inertia “J,” that is: $P(s) = 1/Js^2$. The input to $P(s)$ is torque “ τ ,” while the output is the inertial angle “Y.” The torque input to $P(s)$ comes from two sources namely: 1) $W(s)$ which collectively represents the motor windings, the power amplifiers (which deliver current to the motor), and a feedback compensator, and 2) $F(s)$ which represents gimbal friction. Of course, friction opposes the relative motion and thus carries a negative sign.

The gimbal angle θ is measured by two high-accuracy resolvers (a 1x and a 64x resolver). In Figure 2 above, the resolvers together with the *Resolver-to-Digital* (R/D) converters are lumped together and labeled $Q(s)$ —and for simplicity, will hereafter be referred to as the resolver. The inertial angle “Y” is measured with a Systron Donner *Inertial Angle Sensor*¹ (IAS) and is labeled $H(s)$ in Figure 2. The IAS is a *band-pass* device whose bandwidth is between 2 Hz and 10 kHz—

¹ Harold Morris, “The Inertial Angular Displacement Sensor, Theory and Application,” (Systron Donner Inertial Division), February 18, 1987.

that is; it measures angular displacements between 2 Hz and 10 kHz. The theory of operation, application and evaluation of the IAS is further explained in Reference 1.

Because $H(s)$ rolls-off at low frequencies, another sensor was needed to cancel the low-frequency components of the disturbance "D." A connection to the Inertial Measurement Unit (IMU) on the host vehicle was added to the telescope pointing system to provide low-frequency *feed-forward* cancellation of the base disturbance. Of course, since the IMU is located off-gimbal, a transformation is necessary to convert host attitude data into gimbal coordinates. The frequency response of the IMU together with the coordinate transformation are lumped together in Figure 2, and labeled $B(s)$. The transformed host attitude data are subtracted from the desired pointing angle "R" to produce the commanded angle "C." This calculation, together with the coordinate transformation are computed on a separate processor outside the servo controller system as described in this report, however, these components are shown in Figure 2 for completeness.

Finally, there is the compensator $G(s)$. The function of $G(s)$ is to blend the low-frequency data coming from $B(s)$ and $Q(s)$ with the "high-frequency" feedback data coming from the inertial sensor $H(s)$ —producing a "smooth" transition between the two sensors.

System Identification of Fixed Components

In this section, we identify the *transfer functions* of those components in the servo that are *fixed* in the sense that they are not available for modification by the design process. We also quantify the uncertainties in their parameters. For example, elevation gimbal inertia is estimated at 86 oz-in-s². This estimate was made prior to the telescope being completed so for design purposes elevation inertia was allowed to vary between 75 and 100 oz-in-s². Azimuth gimbal inertia depends on elevation as per Equation 1 below. For an elevation angle of 90 degrees, azimuth inertia is approximated at 142 oz-in-s², for 180° it is estimated at 104 oz-in-s²—that is,

$$J_{az} \approx 50 + 54 \sin^2(\text{elevation}) + 92 \cos^2(\text{elevation}) \text{ oz-in-s}^2. \quad (1)$$

The nominal transfer function of the inertial angle sensor $H(s)$, given by Equation 2 below, has been found to vary slightly with temperature. The manufacturer provided four different transfer functions to represent $H(s)$ at various temperatures. This set of four was used to represent the variation in $H(s)$ as a function of temperature. A Bode plot of this set is shown in Figure 3.

$$H_n(s) \equiv \frac{s^3[s+187][s+7.6]}{[s+201][s+4.2][s+1.1][s+2(0.55)15+15^2]} \quad (2)$$

A Bode plot of the IMU, $B(s)$ was provided from the host vehicle contractor (shown below in Figure 4.) Also shown in Figure 4 is the effect of a variable time delay in the host-vehicle attitude data (which alternates between 15 and 25 milliseconds). From this plot, a nominal transfer function $B_n(s)$ was empirically determined that matched the Bode plot. The transfer function, given by Equation 3 (excluding the time delay), is shown below. For design purposes, the dominant pole at 300 rad/sec was varied $\pm 20\%$, while its damping ratio of 0.6 was varied $\pm 10\%$. A sample space of sixteen variations on $B(s)$, plus the nominal, was used to represent the uncertainty in the IMU.

$$B_n(s) \equiv \frac{300^2(1000)[0.205s+1]}{[0.201s+1][s+2(0.6)300+300^2][s+1000]} \quad (3)$$

Figure 3. Inertial Angle Sensor Frequency Response

Figure 4. Inertial Measurement Unit (IMU) frequency response

The nonlinear behavior of bearing friction was simulated with the Dahl friction model²—whose differential equation is shown below in Equation 4. The telescope manufacturer estimated the Dahl friction model parameters as shown below in Table 1. A simulation of this equation is shown in Figure 5 as a plot of bearing friction versus angle.

$$\frac{dF}{dt} \equiv \text{abs} \left\{ 1 - \text{sign} \left(\frac{d\theta}{dt} \right) \frac{F}{F_{\max}} \right\} (\text{slope}) \frac{d\theta}{dt} \quad (4)$$

Axis	Running Torque $F_{\max}$ (oz-in)			Slope (in-lb./rad.)		
	Min.	Nominal	Max.	Min.	Nominal	Max.
Elevation	0.75	1.50	3.0	30	60	120
Azimuth	3.00	6.00	12.0	230	460	920

Table 1. Dahl friction model parameters

In order to include the nonlinear effect of bearing friction into the design process it was convenient to convert the nonlinear Dahl effect into an equivalent frequency response. This was done using the *describing function technique*. That is, the response of the nonlinear model to various sinusoidal inputs was measured, and the input/output transfer function was derived from these data. Sinusoidal inputs to equation 4 produce outputs with the same fundamental frequency, plus higher harmonics due to the nonlinear nature of Equation 4. Describing function technique tests the nonlinear system at one frequency at a time. The transfer function is determined at each frequency by ignoring the harmonics and computing the input/output ratio for the fundamental.

Figure 5. Bearing friction versus gimbal angle

² P. R. Dahl, "A Solid Friction Model," *The Aerospace Corporation, Space and Missile Systems Organization Air Force Systems Command*, Report No. TOR-0158 (3107-18)-1, May 1968.

Accordingly, the Dahl equation was simulated at various frequencies, under various conditions. (Four such simulations are shown in Figure 6.) When the data for these simulations were reduced, it was found that the Dahl frequency response varies as a function of both frequency, and input magnitude. The Dahl friction behaves much like a single-order low-pass filter—whose roll-off frequency is a function of not only the Dahl parameters of F_{max} and $Slope$, but also a function of input magnitude. That is, at low frequency and/or large input magnitudes, the bearing friction has a flat frequency response, and the phase shift approaches zero. For small and/or high frequency inputs, the Dahl friction causes a 90° phase shift between input and output, and rolls off at a rate of 20 dB per decade. This is shown more clearly in Figure 7. From these data, it was empirically determined that the describing function equivalent of Equation 4 could be written as shown below in Equation 5. (Note that because the input to the Dahl equation is $d\theta/dt$ instead of θ , Equation 5 must be multiplied by “s” in order to yield the transfer function $F(s)$ with bearing friction as the output and θ as the input.)

$$\left| \frac{F(s)}{s} \right| = \left| \frac{F}{s\dot{\theta}} \right| = \left| \frac{slope}{s+a} \right| \quad \text{Where : } a = \frac{\pi \cdot slope \cdot \text{input magnitude}}{4F_{max}} \quad (5)$$

Figure 6. Dahl bearing friction input/output simulations

Figure 7. Describing function analysis of bearing friction

Figure 8. Describing function analysis of bearing friction with extra phase shift

The raw simulation data (shown in symbols) along with the interpolation of Equation 5 (shown in solid lines) are plotted in Figure 7. Note that the interpolation of Equation 5 perfectly matches the raw magnitude data, but does not match the raw phase data. In order to match Equation 5 to the phase data, four phase lag filters were added as shown below in Equation 6. The resulting phase plot is shown in Figure 8.

$$\text{phase}\left(\frac{F(s)}{s}\right) = \text{phase}\left(\frac{[s + 0.195f][s + 2.5f][s + 6.67f][s + 43f]}{[s + a][s + 0.18f][s + 2f][s + 6f][s + 22f]}\right) \quad (6)$$

$$\text{where : } f = e^{[\ln(a) - 3.9045]}$$

Design Procedure

Applying standard flow graph modeling techniques to Figure 2 yields the transfer function shown below in Equation 7 for the inertial angle "Y." The first term in Equation 7 is the response due to the control reference signal "R," while the second term represents the system response due to the disturbance "D."

$$Y(s) = \frac{RGWP}{1 + P[F + W(GQ + H)]} + \frac{DP[F + GW(Q - B)]}{1 + P[F + W(GQ + H)]} \quad (7)$$

The design goals for this project were: 1) 1.5 microradian pointing resolution, 2) 1 microradian line-of-site pointing stability in the presence of host vehicle attitude disturbance "D" and 3) high agility—that is, *fast* re-targeting capability. Note that the transfer function for the pointing accuracy is represented in the first term of Equation 7, while the stability performance is represented in the second term of Equation 7. The most challenging design goal was the stability criterion.

Resolver-to-Digital (R/D) Converter Design

The gimbal angle θ is measured by two high accuracy resolvers (a 1x and a 64x resolver). The resolver signals are converted into a 22-bit word by combining two 16 bit Resolver-to-Digital converters (R/D)—see Table 2 below. In Figure 2, the resolvers together with the R/D converters are lumped together and labeled Q(s)—and for simplicity, will hereafter be referred to as the resolver.

<u>Resolver</u>	<u>Binary</u>	<u>Degrees</u>
1 x (16-bit)	01 0101 1111 0010 10	123.453
64 x (16-bit)	1111 0010 1010 0111	341.22986 / 64
Composite (22-bit)	01 0101 1111 0010 1010 0111	123.45672

Table 2. Bit relations of the resolver-to-digital converters

For some important military applications radiation hardness is required. Military grade resolver-to-digital converter chips are available from two sources, namely: *ILC Data Device Corporation* (DDC), and *Analog Devices*. Both chips were radiation tested and it was found that the DDC chip was susceptible to radiation induced latch-up and Single-Event Upsets (SEU), while the Analog Devices chip was SEU tolerant and latch-up immune. Therefore, the Analog Devices (AD2S80A R/D) chip was used.

By observing the second term in Equation 7, it is clear that the system response to disturbance "d" can be significantly reduced by designing $Q(s) \approx B(s)$. That is, we want the frequency response of the resolvers to match as closely as possible that of the IMU shown in Figure 4, and Equation 3. Achieving this was a significant challenge.

The R/D converter is a "type II feedback system" (using two integrators) that requires external compensation to achieve the desired closed-loop response. While the standard application notes allow the designer to realize any desired bandwidth, it is not possible to build to the specific frequency response of $B(s)$ without modifying the design procedure detailed in the application notes. It was therefore necessary to obtain a more detailed understanding of the R/D converter chip.

According to the design detail obtained from Analog Devices,³ the converter can be summarized as shown below in Figure 9. Using the design process detailed in the data book, one can select the parameters k , a , and b for a desired tracking rate and bandwidth f_{bw} . This will produce the closed-loop response shown in Equation 8. The parameters k , a , and b , affecting the closed-loop bandwidth f_{bw} of the R/D converter, are selected by the choice of external components. As stated earlier, we want the resolver transfer function $Q(s)$ to closely match that of the IMU $B(s)$, in order to maximize disturbance cancellation. To accomplish this, another pole was added to the compensator, as shown in Figure 10, and the lead-lag filter was adjusted slightly. This produced the desired closed-loop response shown in Equation 3 and Figure 4 (without the variable time delay).

Figure 9. Block diagram of resolver-to-digital converter

$$\frac{\theta_{out}}{\theta_{in}} = \frac{2.4^3 [s_n + 1]}{[s_n + 2.4][s_n^2 + 2(0.7)2.4s_n + 2.4^2]} \quad \text{where } s_n = \frac{2s}{\pi f_{bw}} \quad (8)$$

Figure 10. Modified block diagram of resolver-to-digital converter

The external compensation required to achieve this design is shown in Figure 11, with the extra pole drawn in gray. Assuming $R_3 \ll R_4$, the relationship between the component values and the parameters shown in Figure 10 are as follows: $a = 1/(R_2C_2)$, $b = (C_1 + C_2)/(R_2C_2C_1)$, $c = 1/(R_3C_3)$ and $k = 6.163 \times 10^6 / [R_1C_1(R_3 + R_4)]$. The component values are listed in Table 3 below. For servo design purposes, the gain parameter k , as well as all three capacitors, were varied independently $\pm 20\%$ of their nominal values. This produced a set of sixteen variations on $Q(s)$, plus the nominal, to quantify the uncertainty in the resolver $Q(s)$.

One of the design goals of this project was to improve telescope agility. As mentioned earlier, the telescope was retrofit with larger motors with a maximum torque output of 190 ounce-inches. When making large maneuvers full torque is applied to the gimbal to maximize speed. For step commands greater than 15 degrees, the gimbal exceeds the tracking speed of the 64x Resolver-to-Digital converter, causing the measured gimbal angle to lag behind the true angle. When the error reaches 180° the converter makes a sudden correction to get the angle back on-track. These

³ Mark L. Schirmer, "Using Laplace and Fourier Transform Techniques to Model the Performance of Resolver to Digital Converters," (Analog Devices Inc.: Memory Devices Division, 3 Technology Way, Norwood Massachusetts)

sudden corrections caused a serious problem for the large maneuver servo algorithm, so the 64x R/D converter had to be modified in order to work properly.

Figure 11. Schematic diagram of resolver-to-digital external compensation

R_1	R_2	R_3	R_4	C_1	C_2	C_3
8.25k Ω	205k Ω	23.7k Ω	681k Ω	2200pF	1 μ F	0.1 μ F

Table 3. Component values for Figure 11

Of the two axes, elevation has the least inertia, as well as the largest range of travel, and thus will potentially see the highest angular velocities. Assuming a maximum step command of 100°, maximum torque of 190 in-oz and a gimbal inertia of 86 oz-in-s², the elevation gimbal would reach a maximum angular velocity of about 0.3 revolutions-per-second. Thus, the 64x R/D converter must have a maximum tracking speed of about 20 RPS—which corresponds to a 300 Hz bandwidth (see Equation 8). To implement the higher bandwidth, the standard design procedure was followed. Adding C4 and R6 and replacing R4 with R5 thus modified the circuit in Figure 11. An analog switch was also added. (In Figure 12, $C_4 = 0.039\mu$ F, $R_5 = 635$ k Ω , and $R_6 = 47.5$ k Ω .)

Figure 12. Schematic diagram of 64x resolver-to-digital external compensation

When making large maneuvers the normally-closed switch shown in Figure 12 is opened—effectively changing C_1 from $1\mu\text{F}$ to $0.038\mu\text{F}$, while the normally-open switch is closed—bypassing the extra pole (shown in dashed lines in Figure 11). After the maneuver is completed, the switches return to their default positions which returns the resolver-to-digital converter to the frequency response required for maximum jitter cancellation. Note that the two switches shown in Figure 12 are from the same (Double-Pole-Single-Throw) DPST analog switch, so that only one control line is used to switch both. This circuit solved the above-mentioned problem for large maneuvers.

Design of Feedback Sensor Blending Compensator $G(s)$

The most challenging aspect of this design was that of the compensator $G(s)$. As previously mentioned, the function of $G(s)$ is to blend the low-frequency data coming from $B(s)$ and $Q(s)$ with the *high-frequency* feedback data coming from the inertial sensor $H(s)$. The effective feedback sensor is the composite $G(s)Q(s) + H(s)$. Since $H(s)$ rolls-off for frequencies below 2 Hz, we want to use the resolver feedback data from $Q(s)$ for all frequencies below 2 Hz, and use the inertial angle sensor $H(s)$ for all frequencies above 2 Hz. Therefore, ideally $G(s)$ should cut-off $Q(s)$ *sharply* at 1 Hz as shown in Figure 13, however, this presents a significant problem. Factoring the equation $G(s)Q(s) + H(s)$, reveals zeros in the right-half of the s -plane—resulting in unstable poles in the closed-loop system, (since according to root-locus analysis, closed-loop poles “seek” the open-loop zeros at high gain).

This can be more easily seen by studying the root-locus plot of $G(s)Q(s) + H(s)$. The root-locus for $G(s)$ as proposed in Figure 13 (a second-order low-pass at 1 Hz) is shown in Figure 14. Using the root-locus technique, it is possible to modify $G(s)$ to cause $G(s)Q(s) + H(s)$ to have all zeros (and poles) in the left-half plane. However, this proved to be quite difficult when taking into account the divergence in $H(s)$ and $Q(s)$ due to temperature and component variations. That is, finding a compensator that meets this specification (no right-half-plane poles and zeros) is relatively simple for each combination of $H(s)$ and $Q(s)$ individually. But finding *one* compensator that satisfied these criteria for *all* combinations of H and Q in the set proved to be quite impractical using the root-locus technique.

Figure 13. Bode plot of $H(s)$ and desired $G(s)Q(s)$

Figure 14. Root-locus of $G(s)Q(s) + H(s)$ as proposed in Figure 13

It was therefore necessary to find another method. Dr. Isaac Horowitz, a consultant to Sandia on this project, suggested a simple technique that proved to be very effective. As it turned out, a lesser-known design technique called *Quantitative Feedback Theory* (QFT) is ideally suited for such a design challenge. Over thirty years ago Dr. Horowitz pointed out that while the sole function of feedback control is to reduce system variation and nonlinearity, at that time there was no design technique available to *quantitatively* tailor the feedback design to the amount of uncertainty and/or nonlinearity in the system. He submitted that if a system is perfectly Linear and Time-Invariant (or LTI), there is no need for feedback control, that LTI systems can be more easily controlled open-loop. Furthermore, he insisted that the more variable and the more nonlinear a system, the more feedback is needed for effective control—thus the *amount* of feedback used (measured in units of gain X bandwidth) should be commensurate with the amount of uncertainty and/or nonlinearity to be controlled using feedback.⁴⁻⁷

⁴ Isaac Horowitz, "Fundamental Theory of Linear Feedback Control Systems," *Transaction of Institute of Radio Engineers on Automatic Control*, AC-4, December 5, 1959; See also: Horowitz, *Synthesis of Feedback Systems*, Academic Press, 1963

⁵ Horowitz, "Synthesis of Feedback Systems With Nonlinear Time-varying Uncertain Plants to Satisfy Quantitative Performance Specifications," *Proceedings of IEEE*, vol. 64, No. 1, (1976)

⁶ Horowitz, "A survey of Quantitative Feedback Theory," Invited Paper, *International Journal of Control*, vol. 53, 1991, pp. 255-291.

⁷ Horowitz, *Quantitative Feedback Design Theory*, vol. 1, QFT Publications, 1993, (4470 Grinnell Ave., Bolder, Co., 80303).

Quantitative Feedback Theory greatly simplifies the task of designing control systems with non-linear and/or time-varying behaviors to quantitative performance specifications. Typical QFT designs specify *upper* and *lower* performance bounds on the closed-loop frequency response. For example, in our case the design specifications on $G(s)Q(s) + H(s)$ were quite simple: First we needed a *smooth transition* between the $Q(s)$ and $H(s)$ feedback sensors (that is, there should be no significant *peaking* over the desired operating range). Secondly the composite sensor $GQ + H$ must be minimum phase (no right-half-plane zeros). To be more precise on the first specification, *smooth transition* was defined as $-2.3 \leq |GQ + H| \leq 2.3$ dB for frequencies less than 100 Hz. Here ± 2.3 dB was used for the upper and lower bounds which correspond to damping ratio of about 0.42—that is, the composite feedback sensor $GQ + H$ should have no poles or zeros with damping ratios less than 0.42. Designing to both of these specifications simultaneously, while taking into account plant uncertainty, turned out to be relatively simple using QFT.

With QFT, compensator design is performed using the Nichols chart. The Nichols chart is a convenient plotting technique that *quantitatively* relates the open-loop frequency response to that of the closed-loop. The open-loop frequency response is plotted as phase (in degrees) versus magnitude (in decibels)—as opposed to the more familiar *Bode* format of magnitude versus frequency and phase versus frequency. A sample Nichols plot is shown in Figure 15 for a simple second-order system $L(s)$. The following short example is given to familiarize the reader with some of the basics of QFT design.

Typically a system starts out with high gain at low frequency, and thus begins high in the Nichols chart. As the frequency is increased, the magnitude typically decreases as does the phase so that the system moves down and to the left on the Nichols chart with increasing frequency. The *contour lines* on the Nichols chart relate the open-loop to the closed-loop frequency response—assuming a closed-loop topology of $L/(1+L)$. For example, the open and closed-loop magnitude and phase for the five discrete points plotted in circles in Figure 15 are summarized in Table 4 below. Compare Figure 15 and Table 4 with the closed-loop Bode plot shown in Figure 16 (which is also marked with circles corresponding to the points in Table 4).

Frequency (Hz)	Open Loop		Closed Loop	
	Magnitude (dB)	Phase	Magnitude (dB)	Phase
1	35	-91°	0.0	-1°
23	9	-110°	0.5	-20°
33	6	-117°	1.0	-30°
88	-6	-144°	-3.0	-120°
140	-13	-156°	-12.0	-150°

Table 4. Open vs. closed-loop magnitude and phase for Nichols plot in Figure 15

The critical point in the Nichols Chart (-180° and 0 dB) is surrounded by closed-loop magnitude contours, which go to infinity at the critical point. These contours correspond to phase and gain margins as used in Bode plot design. In this simple example, there is 50 degrees of phase margin with infinite gain margin. The highest valued contour to which the Nichols plot of $L(s)$ approaches is the 2-dB contour. Accordingly, the closed-loop response (Figure 16) has slightly less than 2 dB of peaking—corresponding to a second-order response with a damping ratio of about 0.45.

Figure 15. Nichols plot of a simple second-order system

Figure 16. Bode plot of closed-loop system in Figure 15

When designing feedback for an uncertain system, parameter variation can be represented with a *set* of transfer functions (or plants) that span the range of variation encountered in the system. To continue with the above example, suppose that our system (which consists of a gain k , an integrator, and a single pole at frequency ω) contained the following uncertainty: $10^5 \leq k \leq 3 \times 10^5$ and $300 \leq \omega \leq 600$ (with k varying independently of ω). A family of Bode Plots is shown in Figure 17 that represent this uncertain system. When this *set* of plants is plotted on the Nichols chart, it becomes difficult to interpret. So, rather than representing the phase and gain at each frequency with a single point on the Nichols chart, each frequency will carry with it a *set* of points, which collectively define an *area* on the chart. These *areas* are referred to as “*plant templates*.” The more uncertainty in a system, the larger the templates—which typically increase with frequency. (See Figure 18 for the plant templates of the system in Figure 17.)

Rather than keeping track of each plant separately, the QFT design process *groups* all plant combinations in the set together, and identifies the entire template by a *nominal plant*. The plant with the lowest gain in the set is typically chosen as the nominal plant (marked with asterisks in Figure 18)—however, this is strictly by convention. The choice of the nominal plant can be arbitrary. In our example, the nominal plant is marked with circles in Figure 18. Note that the highest valued contour to which the templates approach is the 6 dB contour. Accordingly, the closed-loop response (Figure 19) has slightly less than 6 dB of peaking—corresponding to a second-order system with a damping ratio of 0.26. Also, note that the closed-loop contour lines in Figure 18 *diverge* at the top of the Nichols chart. This is a graphic illustration of the benefit of feedback. That is, when a system has large gain (and is thus high up in the Nichols chart) it can accommodate large amounts of open-loop uncertainty. Compare the open-loop to the closed-loop uncertainty of Figures 17 and 19. Note that as the gain decreases at higher frequencies, the amount of plant variation rejected through feedback is reduced. This is indicated on the Nichols chart by closed-loop contours approaching the open-loop grid for gains less than 0dB.

A QFT design generally contains five basic steps. 1) Identify the plant using Laplace transforms. 2) Quantify and represent plant uncertainties and/or nonlinearities with a *set* of plants. 3) Create a set of plant templates by sampling the set of plants at several discrete frequencies. 4) Use these templates to create Nichols chart design “bounds” relative to the nominal plant. These bounds effectively translate the design requirements for the entire set of plants onto the nominal plant. That is, if the compensated nominal plant satisfies these bounds, then all plants in the set are guaranteed to meet or exceed the design requirements. 5) The feedback design can then proceed using “classical” design techniques to shape the nominal loop, and by so doing, cause the entire set of plants to meet the quantitative performance specifications. (A QFT toolbox available from *The Mathworks Inc.*, can be used to calculate the bounds in step 4—typically the most tedious step in a QFT design. The toolbox is also highly useful in executing step 5.)

With these basic concepts introduced, we can continue describing the design process used for the design. The above example was a simple demonstration on how a Nichols chart can be used for any feedback system whose closed-loop response is $L/(1+L)$. However, similar “*Nichols charts*” can be created that simplify the design of other topologies such as $GQ + H$ as in our case.

As stated earlier, the design specifications for $GQ + H$ were: 1) $-2.3 \leq |GQ + H| \leq 2.3$ dB for frequencies less than 100 Hz, and 2) the composite sensor $GQ + H$ must be minimum phase and stable (no right-half-plane zeros or poles). Designing to the first requirement is straightforward using QFT. However, meeting the second criterion required some problem manipulation. As pointed out by Dr. Horowitz, the sensor equation simply needed to be *rearranged* to fit the classical QFT approach as follows: $GQ + H = H(1 + GQ/H) = H(1 + L)$, where $L = GQ/H$.

Figure 17. Bode plot of a second-order system with uncertainty

Figure 18. Nichols plot of uncertain system in Figure 17

Figure 19. Bode plot of closed-loop system of Figures 17 & 18

If $H(s)$ and $Q(s)$ are both stable and minimum phase, (which is true in our case) and if $1 + L$ is also minimum phase and stable, then $GQ + H$ is also. To guarantee that $1 + L$ is minimum phase and stable, we need to satisfy the Nyquist criterion—that the number of critical-point encirclements of $L(s)$ is equal to the number of right-half-plane zeros minus the number of right-half-plane poles in $L(s)$. Since $L(s)$ has neither right-half plane poles nor zeros, we can simply design $L(s)$ to have no encirclements of the critical point. On the Nichols chart, the critical point is found at 0 dB and -180 degrees. Thus if $L(s)$ crosses 0 dB to the right of this critical point, then we will have no encirclements, and our system will be stable and minimum phase. Ideally, of course, we would want to have some phase margin when $L(s)$ crosses 0 dB, so we will stay a considerable distance away from the critical point—specifically, we will stay outside the 2.3 dB Nichols chart contour.

Figure 20 shows the Nichols plot of both the nominal plant Q_n/H_n (top) and the compensated nominal plant $GL_n(s) = GQ_n/H_n$ (bottom). While Figure 20 shows only the nominal, each point carries with it a *set of plants* (or template) that define an area of uncertainty. When using QFT the closed-loop contours of the Nichols chart are replaced with *bounds* on $L_n(s)$. These bounds can be obtained for a set of frequencies by moving the plant templates around the Nichols chart to determine the regions on the chart where the open-loop variation can be accommodated into the closed-loop performance specification. These bounds are with respect to the nominal plant—that is, if the *nominal* plant satisfies the bounds at all frequencies, then *all* plants within the set will meet or exceed the design specifications for the system. In our case, the performance bounds were defined as $-2.3 \leq |GQ + H| \leq 2.3$ with respect to the nominal plant Q_n/H_n .

At each frequency where bounds were calculated, there are two bounds plotted in Figure 20. The dashed lines represent the upper bound of $|GQ + H| \leq 2.3$, while the solid lines represent the lower bound $-2.3 \leq |GQ + H|$. As long as the compensated nominal plant $GL_n(s)$ lies below the dashed lines, and above the solid lines at each frequency, then $-2.3 \leq |GQ + H| \leq 2.3$ for all combinations of Q and H in the set. Note that the uncompensated plant violates these bounds, thus a lead-lag compensator $G(s)$ was added to meet the design specifications at all frequencies— $G(s)$ is given below in Equation 9. (The Quantitative Feedback Theory toolbox from *The Mathworks Inc.* was used in this design to calculate the bounds, and to shape the compensators—which greatly simplified the design process.)

$$G(s) = \frac{27[s+4]}{[s+9][s+12]} \quad (9)$$

Figure 21 shows a Bode plot of $GQ(s)$ along with $H(s)$, and a simulated disturbance $D(s)$ as measured by $B(s)$. This sample disturbance is taken as the absolute worst case that could be expected on the host vehicle. Note the large spike at 4.5 Hz. Figure 22 shows the Bode plot of the composite sensor $G(s)Q(s) + H(s)$.

Redesign of Feedback Sensor Blending Compensator $G(s)$

The simulated host vehicle pointing error data $d(t)$ is shown in Figure 23, which reveals a ± 6 microradian jitter at 4.5 Hz. The host vehicle attitude data from $B(s)$ is sampled at 40 Hz, and contains a variable latency. Every other sample is expected to have a maximum latency of 15 milliseconds, with the remaining samples having a latency of 25 milliseconds. Since this delay is not inside the feedback loop, it will not affect stability, but it severely restricts the disturbance rejection performance. Detailed simulations revealed that because of this variable latency, attenuation of the 4.5 Hz jitter was quite difficult. (See disturbance attenuation simulations in Figures 24 and 25.) The impasse is due to the fact that the 4.5 Hz disturbance is measured by both the inertial angle sensors $H(s)$ as well as IMU $B(s)$, but is corrupted in $B(s)$ by the variable latency. The delay becomes more significant at higher frequency, causing the two sensors $B(s)$ and $H(s)$ produce conflicting information on $D(s)$ at 4.5 Hz, which seriously limits disturbance attenuation.

Figure 20. Nichols plots of Q_n/H_n and GQ_n/H_n with bounds on $GQ + H$.

Figure 21. Bode plot of $G(s)Q(s)$ and $H(s)$ along with $D(s)B(s)$.

Figure 22. Bode plot of $G(s)Q(s) + H(s)$

Figure 23. Simulated host vehicle attitude data $d(t)$

Figure 24. Simulated disturbance attenuation results (without attitude latency)

Figure 25. Disturbance attenuation with host vehicle attitude latency

Figure 26. Improved disturbance attenuation results with latency

Ideally, we would have rolled-off $B(s)$ *sooner* so it would not see the 4.5 Hz signal (as shown in Figure 13). However, as it turns out, a certain amount of over-lap between the two sensors is needed in order to satisfy the minimum phase requirement. Thus, another approach was needed to further attenuate $D(s)$. To accomplish this, $G(s)$ was modified (as shown in Equation 10) to purposely violate the upper bound $|GQ + H| \leq 2.3$ for frequencies local to 4.5 Hz. The idea was to allow the 4.5 Hz jitter to come through the feedback sensor with a higher gain so that it could be acted upon more aggressively than its neighboring frequencies. In an effort to emphasize the uncorrupted information coming from $H(s)$ with respect to $B(s)$, a gain of 3 was added to $H(s)$.

$$G(s) = \frac{42[s + 4]}{[s + 14][s + 12]} \quad (10)$$

This significantly improved the simulated disturbance attenuation at 4.5 Hz, (Figure 26). However, when the servo was implemented on the actual telescope, large structural resonances in the inertial angle sensor mounts that attach the *Inertial Angle Sensors* to the telescope were discovered. The lowest of these resonances were too close to the system bandwidth to effectively notch-out, and so it became necessary to sacrifice performance in order to avoid exciting the structural resonances. (These resonances will be discussed in detail, later in this report.) The gain on $H(s)$ was thus lowered from 3 to 1.5. The Bode plot of the composite sensor $GQ + H$ in Figure 27 includes the gain of 1.5 on the inertial angle sensor $H(s)$.

Design of Feedback Compensator $W(s)$

As mentioned in the system overview, $W(s)$ collectively represents the motor windings, the power amplifiers (which deliver current to the motor), and a feedback compensator. The power amplifiers (whose design will be detailed later in this report) were designed as current drivers with current feedback from the motor windings. This effectively removed the motor winding dynamics from the design process, and the bandwidth of these amplifiers was large enough to be ignored with respect to the design of $W(s)$.

The driving factor in the design of $W(s)$ was that of disturbance attenuation. The desired pointing stability is ± 1 microradian. Simulated host vehicle pointing error data is shown in Figure 28.

Discounting the initial transient, the worst-case host-vehicle attitude data contains disturbances of about ± 250 microradians at low frequency and ± 6 microradians at 4.5 Hz. Of the two axes, the azimuth disturbance was the most challenging to attenuate, and since the host attitude is coupled to the telescope gimbals differently depending on gimbal angles, the azimuth disturbance was used for the design of both axes. For design purposes, it was assumed that the disturbance $D(s)$ would be no greater than the bound shown in Figure 29—derived by taking the Fourier transform of the azimuth data shown in Figure 28. To achieve a factor-of-two design margin, the system was designed to attenuate this disturbance down to ± 0.5 microradian. To accomplish this, an attenuation of 55 dB is needed at low frequencies and 22 dB is needed at the higher frequencies. These requirements are represented by the disturbance attenuation specification (W_d) shown in Figure 30. Also shown in Figure 30 is the effect of $W_d(s)$ on the disturbance $D(s)$.

The only other quantitative specification imposed on the design of $W(s)$ was the stability criterion—that the closed-loop system must have less than 2.3 dB of peaking. Converting these design specifications into bounds on the Nichols chart is a relatively straightforward task using the Quantitative Feedback Theory Toolbox.

Figure 27. Bode plot of composite sensor $G(s)Q(s) + 1.5 H(s)$

Figure 28. Simulated host vehicle attitude data (worst-case)

Figure 29. Fourier transform of $D(s)$ with maximum disturbance limits

Figure 30. Disturbance attenuation specification with attenuated disturbance

The small-signal output equation (7) was made to match the *General Bound Manager* format of the QFT Toolbox (*QFT Toolbox manual* [1994], pp. 6-7) as follows:

$$Y(s) = \frac{R(s)C(s)W(s)}{K(s) + L(s)W(s)} + \frac{A(s) + E(s)W(s)}{K(s) + L(s)W(s)}, \quad (11)$$

where: $A(s) = D(s)P(s)F(s)$; $C(s) = G(s)P(s)$; $E(s) = D(s)C(s)[Q(s) - B(s)]$; $K(s) = 1 + P(s)F(s)$; $L(s) = P(s)[G(s)Q(s) + H(s)]$.

The stability criterion is given by Equation 12 and the disturbance attenuation specification is given by Equation 13. Nichols chart stability bounds were calculated from Equation 12 with respect to $L_n(s)$ using the QFT toolbox. These bounds are shown in Figure 31. Similarly, disturbance attenuation bounds were calculated from Equation 13 with respect to $L(s)$ also using the QFT toolbox. These bounds are shown in Figure 32. Note that the disturbance attenuation bounds are more restrictive, in this case, than the stability bounds. Figure 33 shows the intersection of the stability bounds with the disturbance attenuation bounds—which closely matches the disturbance attenuation bounds since they are the more restrictive of the two. Also shown in Figure 33 is the Nichols plot of $L_n(s) = P_n(s)[G(s)Q_n(s) + H_n(s)]$. (Only the azimuth axis is shown in Figures 31 through 33 because the elevation axis looks almost identical to azimuth.) Note that without compensation the system is unstable since it crosses zero dB to the left of the stability bounds.

$$\left| \frac{C(s)W(s)}{K(s) + L(s)W(s)} \right| \leq 2.3 \text{ dB} \quad (12)$$

$$\left| \frac{A(s) + E(s)W(s)}{K(s) + L(s)W(s)} \right| \leq W_d(s) \quad (13)$$

To stabilize the servo loop, a lead-lag compensator $W(s)$ given by Equation 14 is included in Figure 34. Note that the servo system is now stable, and meets the bounds at all frequencies. This is indicated by the fact that the plot of $L_n(s)W(s)$ —(Figure 34) falls above (or to the right) of all the solid bounds and below (or to the right) of all dashed bounds. Bode plots of the closed-loop system are shown in Figures 35 and 36. Figure 35 is the Bode plot of Equation 12 and shows the closed-loop response to the command signal $R(s)$, while Figure 36 shows the closed-loop disturbance attenuation (Equation 13). Also plotted in Figures 35 and 36 are the design criteria—which are met at all frequencies. Figure 37 is a Bode plot of the denominator of equation 13, which is the disturbance attenuation due to feedback. Note that the servo system cannot cancel jitter above 20 Hz. Thus, the *high-frequency* disturbance attenuation shown in Figure 36 is due to the fact that the plant ($1/Js^2$) rolls off at 40 dB per decade. In other words, the low frequency disturbance attenuation is a result of the feedback while at frequencies above 20 Hz we rely on the inertia of the telescope for jitter reduction.

$$W(s) = \frac{k \times 10^6 b[s+a]}{a[s+b]} \quad \text{where,} \quad \begin{matrix} \text{Azimuth} & \text{Elevation} \\ \begin{bmatrix} a \\ b \\ k \end{bmatrix} & = \begin{bmatrix} 500 & 170 \\ 4400 & 2800 \\ 9 & 2 \end{bmatrix} \end{matrix} \quad (14)$$

Figure 31. Nichols chart stability bounds calculated from Equation 12

Figure 32. Disturbance attenuation bounds calculated from Equation 13

Figure 33. Combined bounds with Nichols plot of $P(s)[G(s)Q(s) + H(s)]$

Figure 34. Nichols plot of $W(s)P(s) [G(s)Q(s) + H(s)]$ with bounds

Figure 35. Bode plot of closed-loop servo response (Equation 12)

Figure 36. Disturbance rejection Bode plot (Equation 13)

Figure 37. Disturbance rejection due to feedback (denominator of Equation 13)

Design of the Power Amplifiers

The power amplifiers were designed to provide current to the motor windings and to have large enough bandwidth to be effectively ignored in the servo loop design of $W(s)$. Thus a bandwidth of no less than 200 Hz was needed. The schematic for the power driver is shown in Figure 38. The op-amp boosts the power to the motor windings from ± 15 volts up to a maximum of ± 25 volts at ± 1 amp. The motor current is measured across a 0.1Ω resistor in the return path. The measured current is subtracted from the input control signal and fed back into the power amplifier to close the loop. The motor has redundant windings that are normally used in parallel. Should one of the windings fail, it can be switched out with latching mechanical relays. When operating with only one winding, the feedback system automatically adjusts the current flowing in the remaining coil to make up the difference.*

* The *maximum* power is cut roughly in half when one winding is switched out. However, as long as the requested motor current is less than 0.4 amps, the amplifier adjusts the motor voltage so that the current (and thus the torque) remains the same regardless of the number of windings used. With only one winding energized, motor resistance is about 61Ω , which limits the maximum current to about 0.41 amp due to the 25 volt maximum output of the amplifiers.

Figure 39 shows the Bode plot of the power amp. The motor winding inductance was measured at about 39 mH, but due to mutual inductance, this varies between 29 and 49 mH as a function of gimbal angle. The motor resistance is constant with angle, but varies from one winding to the next between 58 and 65 ohms. Additional motor winding uncertainty was added to take into account the effect of operating on only one of the redundant windings. Figure 40 shows the Nichols plot of the open-loop current driver (which includes the power amp, the motor windings, and the current feedback measurement). The uncertainty of the motor windings is included in Figure 40. The closed-loop amplifier was designed to have a maximum peaking of 3 dB (corresponding to a damping ratio of about 0.38), and to roll-off past 200 Hz. Each of the open-loop plots shown in Figure 40 is marked with a circle at $f = 200$ Hz. Note that at 200 Hz, all the plants in the set are above 0 dB—thus meeting the bandwidth requirement. Also, note that all plants lie outside the 3-dB contour—thus meeting the stability requirement.

Figure 41 shows the Bode plot of the closed-loop power amplifier. Note that there exists significant phase lag at 100 Hz that does slightly affect the design of $W(s)$ as mentioned above. Accordingly, the frequency response of the power amplifiers was included in Figures 33 through 37 to ensure that the design contained sufficient phase margin. Also, note that the current monitor circuit has an over-all gain of 10. Thus, the closed-loop gain of the power amplifier is 0.1 (or -20 dB)—that is, 10 volts input to the power amplifier results in 1 amp of current delivered to the motor. For this reason, the 3-dB design criterion is shown at -17 dB (-20 + 3 dB) in Figure 41.

Figure 38. Schematic diagram of power amplifier

Figure 39. Bode plot of power amplifier (does not include motor winding)

Figure 40. Nichols plot of power amplifier loop (including motor winding)

Figure 41. Bode plot of closed-loop power amplifier

The amplifiers are powered by a supply with ± 30 volts and ± 2 amps. During nominal tracking operations, (when the servo is not making large maneuvers) very little power is needed to move the gimbal—whose maximum bearing friction is anticipated at about 9 oz-in. Assuming a worse case condition of operating under a single winding, the motor would have 95 oz-in/amp. Thus to overcome bearing friction the power amplifiers would have to provide a minimum of 95 mA. Assuming a maximum motor winding resistance of 65 ohms, we would only need 6.2 volts to move the gimbal. To avoid wasting power under normal conditions when large currents are not needed, a circuit was built to switch the amplifier power supply between 30 and 12 volts. (Twelve volts was chosen to allow sufficient overhead voltage for the power amplifiers.) This circuit is shown below in Figure 42.

Figure 42. Schematic diagram of amplifier power switching circuit

Design of the Coarse Mode Servo System

The feedback system described above was designed for the low-speed tracking required to observe a fixed point on the ground as the host vehicle travels along the flight path. This is referred to as the *fine* mode. While the telescope *appears* to be staring at a single point, it is actually tracking very slowly. In this mode, the telescope is expected to point to within 30 microradians of the target, while canceling out the host vehicle attitude jitter. Since the inertial angle sensors saturate at about 100 microradians, it is necessary for the pointing error to be less than 0.02 degrees (380 microradians) before the fine mode can be engaged. Thus, a *coarse* mode of operation was also needed to bring the telescope error into the capture range of the fine mode. The coarse mode could also be used as a back up in the event that one of the inertial angle sensors (which are non-redundant) should fail. The design of the coarse mode will be described in the following section.

Removing the inertial angle sensor from the block diagram of Figure 2 simplifies the feedback equation significantly. Figure 43 shows the simplified diagram while Equation 15 shows the feedback equation of Figure 43. In Figure 43 and Equation 15, $H(s)$ (inertial angle sensor) has been removed and the compensator $G(s)$ has been set to unity.

Figure 43. Coarse mode block diagram

$$Y(s) = \frac{RWP}{1 + P[F + WQ]} + \frac{DP[F + W(Q - B)]}{1 + P[F + WQ]} \quad (15)$$

The design criteria were significantly relaxed for the coarse mode. There was no disturbance attenuation specification, and the stability criterion was the same as for the fine mode—less than 2.3 dB peaking in the closed-loop system. The other design requirement was that the steady-state error must be less than 0.02 degrees to allow the fine mode to switch on.

Because of the nonlinear behavior of the bearing friction, it is not possible to use the final value theorem to calculate steady-state error. Instead, we used a more simple approach to ensure that the servo gain was large enough to satisfy this requirement with a factor-of-4 error margin. With a steady-state gain of about 10^5 , an error as small as 100 microradians would result in about 10 oz-in of torque applied to the gimbal. Assuming that the maximum bearing friction is 9 oz-in, we should have enough torque to overcome the bearing torque and move the gimbal towards the control angle—allowing the fine mode to switch in. (An error of 90 microradians would produce just enough torque to counter-act the bearing friction, but not enough to move the gimbal. The fine mode minimum capture error of 380 microradians would produce 38 oz-inches of torque—four times more torque than what is needed to overcome 9 oz-in of bearing torque.) Thus, a coarse mode steady-state gain of 10^5 should be sufficient to ensure that the telescope pointing error can reach the fine mode capture angle of 0.02 degrees—with a gain margin of 12 dB.

The same design procedure that was used to design the fine mode was also used to design the coarse mode. Figure 44 shows the Nichols plot of $P[F + WQ]$, along with the bounds used in the design of $W(s)$. (Again, only elevation is shown in Figure 44 since there are only subtle differences between the two axes.) Note that since the main design specification was that of stability, all of the frequency bounds encircle the critical point. Figure 45 shows the Bode plot of the open-loop system, while the closed-loop response is shown in Figure 46. Note that due to the bearing friction, there is about 1 dB of low-frequency uncertainty in the closed-loop system. Figure 47 shows what the closed-loop system would be if the friction were reduced to zero. Finally, Figure 48 shows the Bode plot of the coarse-mode disturbance attenuation. The lead-lag compensator $W(s)$ is shown below in Equation 16.

$$W(s) = \frac{10^5 a [s + 16]}{16 [s + a]} \quad \text{where, } a = \begin{bmatrix} 2800 \\ 4400 \end{bmatrix} \begin{matrix} \text{(elevation)} \\ \text{(azimuth)} \end{matrix} \quad (16)$$

Figure 44. Nichols plot of coarse mode design $P[F + WQ]$ with bounds

Figure 45. Open-loop frequency response of $P[F + WQ]$

Figure 46. Closed-loop frequency response—coarse mode

Figure 47. Closed-loop frequency response without bearing friction

Figure 48. Closed-loop disturbance attenuation—coarse mode

Design of the Large-Step Algorithm

The capture angle for the coarse mode is about 1 degree, and thus a third servo mode was needed for large maneuvers with a steady-state error of less than 1 degree in order for the coarse mode to kick in. Because our microcontroller lacked the computing horsepower to use a time-optimal approach to large retarget maneuvers, a simple-minded approach was used. This was done with an algorithm that works as follows:

1. Whenever a pointing error of more than 0.3° is calculated, the servo is switched over to the *step* mode.
2. The initial pointing error is calculated and stored.
3. New pointing errors are constantly calculated and compared to the initial pointing error. If the new error is larger than the stored error (which would occur if the telescope were being pushed off target) the new error overwrites the stored value.
4. Full positive torque is applied to the gimbal to accelerate the gimbal towards the commanded angle, as long as the pointing error is more than half of the initial error.
5. As soon as the pointing error is less than half of the initial error, full negative torque is applied to start decelerating the gimbal.
6. Full deceleration torque is applied until gimbal velocity goes to zero, or changes sign.
7. The pointing error is then checked, if less than 0.3° , the servo system is switched back into the coarse mode. Otherwise, the algorithm loops back to step 1.
8. Once the coarse mode has stabilized the error to less than 0.02° , the servo is switched over to the fine mode.

This algorithm was tested extensively and found quite effective for rapid retarget maneuvers. In the absence of gimbal friction, this simple algorithm arrives at the commanded angle in just one pass through the above loop. However, when friction is included in the model, the gimbal takes less energy to decelerate than to accelerate—causing the gimbal to come to a stop slightly before arriving at the target angle. For smaller retarget maneuvers this effect is negligible and the gimbal arrives in just one pass. The largest retarget maneuver requires only two passes through the loop. Interaction between the two gimbals (which becomes significant only for larger retarget maneuvers) also adds to the large retarget error. However, the second pass through the above loop (being a much smaller maneuver) is sufficient to correct for these errors. See examples of small and large retarget maneuvers in Figures 49 through 51.

During step changes, it is necessary to *cage* the Inertial Angle Sensors (IAS) to avoid saturation. This is done by feeding the IAS output back into the sensor allowing an internal servo system to keep the sensor at the null position. The IAS cage mode requires external compensation for stability that was defined by the manufacturer (Systron Donner)—see Figure 52. To uncage the IAS sensor one simply needs to disconnect the feedback signal into the IAS. In order to utilize just one analog switch to cage/uncage both sensors it was necessary to use a slightly different approach on one of the sensors. The input to the sensor is a relatively low impedance (about 10 k Ω), so simply outputting zero volts into the elevation compensation network realizes the same effect as disconnecting the feedback signal to uncage the sensor. This allowed the normally-open half of the switch to uncage the azimuth sensor while at the same time, the normally-closed switch uncages elevation (Figure 52).

Figure 49. Maximum travel step response

Figure 50. Ten-degree step response

Figure 51. One-degree step response

Figure 52. Inertial Angle Sensor (IAS) caging circuit

Design of the Motor Commutation Algorithm

The motors used in this system are two-phase, 14 pole, brushless DC motors made by Vernitron Motion Control. Of course, brushless motors require a commutation algorithm. Energizing a single phase with a DC current produces a torque profile that is sinusoidal with gimbal angle. Each motor has two windings 90° out-of-phase from one another, so that together they produce a sine-cosine pair of torque profiles as shown in Figure 53. Figure 54 shows the absolute value of these torque profiles, along with the commutated motor torque which is plotted in '+' marks. The commutation algorithm switches the motor current between the two phases depending on the gimbal angle as shown in Figure 54. In order to apply a positive torque over the entire 360° degrees it is also necessary for the commutation algorithm to invert the current when in the negative half of the current profile (this is also shown in Figure 54). Since this is a 14-pole motor (7 pole pairs), the pattern shown in Figure 54 would be repeated seven times per revolution. Thus the degrees shown in Figures 53 and 54 are *electrical* degrees not *mechanical* degrees—(to convert electrical to mechanical degrees one would divide electrical degrees by 7.)

Each time the commutator switches from one phase to the next, a small disturbance is caused due to the sudden change. At small current levels, the disturbance is negligible. However, at full current a barely audible "click" is heard at each commutation point. If the commutation were performed exactly as shown in Figure 54 and the gimbal were to be commanded precisely to one of these commutation points, the system would oscillate between the two adjoining phases, which would increase telescope jitter at those angles. To avoid this problem, a commutation hysteresis of 0.1° degrees was added to the algorithm. For example, suppose the telescope is moving towards a commutation point, it must pass that point by 0.1° before the commutator will switch to the next phase. Should the gimbal be commanded back the other direction, the current will not switch back until the gimbal has passed the commutation point by 0.1° in the opposite direction. This worked well in practice, and no commutation-related problems were seen on the telescope. The motor commutation circuit (including the current inverter) is shown below in Figure 55.

Figure 53. Sine and cosine torque profiles of the two-phase motor

Figure 54. Motor commutation of the two-phase motor

Figure 55. Motor commutation schematic

Design of the Inertial Angle Sensor Notch Filters

The system was first tested on a single-axis gimbal with the same resolver, *Inertial Angle Sensor* (IAS) and roughly the same inertia, torque and bearing friction as the actual telescope. This provided an ideal test-bed for debugging servo hardware and software. After eliminating all software errors the system worked perfectly well. However, once implemented on the actual telescope structural resonances on the IAS mounts caused the telescope to oscillate in the (high-gain) fine mode. The single-axis test gimbal did not exhibit this phenomenon since the IAS was mounted rigidly at the center of the gimbal, while the makeshift telescope mounts, being retrofits, were significantly more compliant (see Figure 1). The open-loop frequency response of the fine mode servo system was measured using a network analyzer, and is shown below in Figure 56. This revealed several destabilizing resonances starting at about 200 Hz.

Figure 56. Open-loop frequency response showing IAS resonances

To eliminate this problem, a notch-filter board was added as a pre-filter for the IAS signals. The board provided a maximum of six, second-order active-notch filters, together with a third-order low-pass filter for each axis. The notch filter schematic is shown in Figure 57 along with the frequency response, while the low-pass filter is shown in Figure 58. The transfer function of the notch filter is shown in Equation 17, while that of the low-pass filters is shown in Equation 18. Note that for elevation, ω_3 is infinite. This is realized simply by removing R_1 and making R_2 zero. The specific notch frequency ω for each filter is given in table 5, while the specific values of R and C used to achieve each notch are not listed in this report.

$$N(s) = \frac{s^2 + \omega^2}{s^2 + 2\zeta\omega s + \omega^2} \quad \text{where, } \omega = \frac{1}{RC}, \quad \zeta = \frac{2R_1}{R_1 + R_2} = \frac{2k}{11k} \quad (17)$$

Figure 57. Notch filter schematic and frequency response

$$LP(s) = \frac{\omega_1^2 \omega_2 [s + \omega_3]}{\omega_3 [s^2 + \omega_1 s + \omega_1^2] [s + \omega_2]}$$

$$\begin{bmatrix} \omega_1 \\ \omega_2 \\ \omega_3 \end{bmatrix} = \begin{bmatrix} 5000 & 1500 \\ 10000 & 10000 \\ 1500 & \infty \end{bmatrix} \quad (18)$$

$$\omega_1 = \frac{1}{RC}, \quad \omega_2 = \frac{1}{100k \cdot 1000pF}, \quad \omega_3 = \frac{R_1}{R_2} \omega_1$$

Figure 58. Low-pass schematic and frequency response

The frequency responses of Figure 56 provided a general idea of what notch filters would be needed. However, the process used to select the number of filters used as well as their frequencies (described below) was not based on this preliminary data—since it was gathered before the telescope was entirely assembled, and would most likely be different than the final configuration. Also, since each notch filter added unwanted phase lag, (see Figure 57) it was important to not use more notches than absolutely necessary to avoid eroding the phase margin. When the notch-filter board was designed, it was unclear how many filters would be needed, or what frequencies would be notched-out. Each filter block was thus designed to be "stuffed" as needed, and the board was designed with jumpers to bypass unused filter blocks.

The notch filters were selected as follows. The fine mode was engaged on the telescope while the dominant oscillation frequency for each axis was measured. A notch filter was then built to eliminate that frequency and inserted in front of the IAS signal. The fine mode was again enabled, while the next most dominant mode was identified. This continued until all oscillations were eliminated. The low-pass filter was then added to roll-off all high frequency resonances. The azimuth low-pass filter includes a zero at 1500 rad/sec needed to restore phase margin that was lost due to lower frequency notches. See Table 5 below for the list of notch filters used for each axis. Also, note that the elevation axis needed one less notch filter than did azimuth.

Azimuth	264 Hz	335 Hz	406 Hz	504 Hz	2.5 kHz	5 kHz
Elevation		335 Hz	555 Hz	635 Hz	720 Hz	875 Hz

Table 5. Notch filter locations

Hardware Implementation of Compensators

The compensator $G(s)$ was implemented as shown below in Figure 59. The switches in Figure 59 are shown in the fine-mode position and Equation 19 shows the input/output equation for the first stage (also in the fine mode). Note that when in the fine mode Equation 19 has a gain of about 0.5. The output of this compensator is fed into a summing amplifier that includes a gain of 2 on the output of the first stage to bring the over-all gain back to unity. The summing amplifier also subtracts the inertial angle sensor signal (with a gain of 3) from that of the compensated resolver error. As previously noted, the IAS gain was eventually lowered from 3 to 1.5. Since the circuit boards were already built when this change was made, the simplest way to lower this gain without effecting $G(s)$ was to lower the gain on the IAS buffer from unity to 0.5—as shown in Figure 59.

When in the coarse mode, the normally-open switch is closed—shorting R_3 to zero, while the normally-closed switches are opened—converting the $G(s)$ compensator into a unity-gain buffer, and switching the IAS signal out of the circuit. As was done previously, a single DPST analog switch was used to switch $G(s)$ between modes and another was used to switch the IAS signal out of the loop. The normally-open pole of this switch (not shown) was used to lower the D/A converter gain—which provides the "Resolver Error" signal to $G(s)$.

$$\frac{v_{out}}{v_{in}}(s) = \frac{-R_4[sC(R_1 + R_3) + 1]}{R_1(sCR_4 + 1)(sCR_3 + 1)} \quad \text{Where, } R_4 = \frac{R_1 R_2}{R_1 + R_2} \quad (19)$$

(The values of C and R_1 through R_3 are given in Figure 59.)

Figure 59. Implementation of compensator G(s)

The compensator $W(s)$ implementation is shown below in Figure 60 and Equation 20. Again, the switches are shown in the fine mode position. When in the coarse mode the normally-closed switch is opened which moves the zero in Equation 14 from 500 or 170 rad/sec down to 16 rad/sec as shown in Equation 16. With R_1 in the circuit, the gain of the R-C network is drastically reduced, so the normally-open switch is closed to increase the amplifier gain and make up some of the difference. Because $W(s)$ is a lead-lag network, it is expected to respond to step inputs with large overshoots. Since the coarse mode has significantly more phase lead than that of the fine mode, a lower over-all gain was used in coarse mode to allow more "head-room" in the amplifier to follow these transient signals.

$$\frac{v_{out}(s)}{v_{in}(s)} = \frac{R_3(R_4 + R_5 + R_6)(sCR + 1)}{(R_4 + R_5)[R + R_3(sCR + 1)]} \quad \text{Where, } R = R_1 + R_2 \quad (20)$$

Figure 60. Implementation of compensator W(s)

As described above, when a large maneuver is performed on the telescope, full current is applied to start the gimbal accelerating in the desired direction. When the gimbal arrives halfway to the commanded angle, full reverse current is applied to start decelerating. When in this mode the $G(s)$ compensator is switched to the coarse mode configuration—a unity gain buffer, while the $W(s)$ compensator is switched to the fine mode to minimize overshoot transients.

A current limit circuit was also incorporated to disable either motor should the motor push against a physical stop with full torque for more than one second. While software stops were coded into the servo computer (which reject commands too close to the gimbal stops), the ability to override these “soft stops” was also provided. Thus under normal conditions this current limit circuit will never be used, however, it was added as a precautionary measure.

The current-limit circuit is shown below in Figure 61 and works as follows: The first stage in the circuit is a “leaky integrator” which takes as its input the output of the circuit shown in Figure 60—which goes directly to the motors through the commutator (Figure 55) and power amplifiers (Figure 38). The second stage in this circuit is a full-wave precision rectifier, and the last stage is simply a comparator. When a large maneuver is performed the requested acceleration current is integrated over time, and the output of the rectifier circuit approaches the comparator switch point. The instant the gimbal begins to decelerate (which is less than one second for the largest possible maneuvers) the integrator output begins to fall and the rectifier output moves away from the comparator switch-point. Should something go drastically wrong, and the controller does not begin decelerating after about one second, the comparator switches which then disables the motor drive using the “motor enable” switch shown in Figure 55.

Figure 61. Current-limit circuit

The rectifier circuit has a gain of 0.5 and so the comparator switch point is set at 2.9 volts—roughly half the maximum output voltage of the integrator. The comparator includes a 10MΩ resistor to provide *positive* feedback to the amplifier. This produces a small amount of hysteresis to keep the comparator from *chattering*. The two back-to-back Zener diodes at the input to the first amplifier block low current commands from being summed in the integrator. Only control signals larger than ± 6 volts (corresponding to ± 0.6 amps) will be integrated. The 10MΩ resistor in the feedback path allows the integrating capacitor to slowly “bleed off” any residual charge.

The back-to-back Zeners in the feedback path keep the amplifier from saturating and allow the integrator to instantly respond to a reversal in the requested current. Should the bearing friction increase to the point that full current is needed for more than one second to accelerate toward a commanded angle, and the current limiter disables current to a motor while accelerating, the gimbal would simply *coast* toward the commanded angle until it reached the half-way point at which time full current would be applied in the opposite polarity. Because the integrating amplifier is not allowed to saturate it will respond *instantly* and re-enable the motor—allowing the gimbal to come to a stop without overshoot.

Because of friction and back-electromotive-force of the motor, more current is required to accelerate the gimbal than to decelerate. If a number of step commands were executed in series, the integrator would slowly build up a bias on the capacitor. Thus at the end of each step command the normally open switch across the integrating capacitor is momentarily closed to quickly short-out the residual charge on the capacitor. This switch also provides a means to disable the current-limit circuit. In the event that the bearing friction increases to the point that significantly more current is needed to move the gimbal, the current-limit can be disabled to prevent it from constantly disabling the motor under such conditions.

Implementation Results

Figures 49 through 51 show the large step responses. Figures 62 through 65 show the performance of the coarse and fine modes as well as the transition between the three modes. Figure 62 shows a step response of 0.3 degrees in both axes simultaneously. The top graph in Figure 62 shows the gimbal angles in degrees. The second graph shows the output of the *Inertial Angle Sensor* (IAS) in microradians, while the bottom graph shows a close-up of the gimbal angles. The pointing system starts out in the big-step mode, and exits the big-step mode to begin coarse mode control at about $t = 0.3$ seconds. During this first segment, the IAS is "caged" to keep it out of saturation. While caged, the sensor behaves like an accelerometer. The sensor remains caged until about $t = 0.75$ seconds, to allow for the fluid inside the IAS to stabilize.* After about one second, the telescope is switched into the fine mode, which causes a small disturbance to the IAS signal. The higher gain of the fine mode allows the steady-state pointing error to be reduced down to a single bit (1.5 microradians).

The length of time that the IAS sensor is caged is determined as follows. The IAS is caged once a gimbal error greater than 0.02 degrees (the fine mode capture angle) is detected. Once caged it remains caged until the error has been less than 0.02 for 0.5 seconds. Once both axes have met this criterion the sensors are uncaged and the controller waits another 0.26 seconds, before switching to the fine mode. Should the error exceed 0.02 degrees on either axis during this segment the timer is reset, the sensors are re-caged and the controller must wait another 0.5 seconds to uncage the IAS's. However, unless another retarget command has been received, the system usually switches to the fine mode control in just one timing sequence of 0.76 seconds.

Figure 63 shows a step response small enough to avoid the big-step algorithm. In this test the controller stayed in the coarse mode until about $t = 1$ second, then the fine mode was switched on. The sensors were uncaged at about $t = 0.7$ seconds. Figure 64 shows an even smaller step response that was executed entirely in the fine mode. Here the error remained just under 0.02 degrees, allowing the sensors to remain uncaged, and the controller to stay in the fine mode. Note that, the fine mode was optimized for maximum disturbance rejection—not minimum response time. Thus, the step response is significantly more "sluggish" than either the big-step mode or the coarse mode, however, because the sensors did not need to be caged, the pointing error stabilizes sooner than in the other modes (at about 0.6 seconds). Figure 65 shows actual in-flight pointing performance with typical host vehicle base motion disturbances. The azimuth pointing jitter as measured by the IAS's over this time interval is 295 nanoradians (rms) while the elevation pointing jitter is 385 nanoradians (rms). Note that both axes comfortably exceed the design goal of 1 microradian pointing stability.

* Since the sensors are caged prior to $t = 0.75$ seconds, the IAS data is not scaled to units of microradians until after $t = 0.75$ seconds. For this reason the data is shown in units of volts—which is equivalent to microradians once the sensors are uncaged.

Figure 62. Step response of 0.3 degrees in both axes simultaneously

Figure 63. Coarse mode step response (both axes)

Figure 64. Fine mode step response

Figure 65. In-flight tracking performance

Conclusions

Quantitative Feedback Theory (QFT) provided a simple, yet powerful design tool which resulted a solution that exceed the performance goal of 1 microradian pointing stability in spite of large host vehicle pointing errors. One of the most challenging aspects of this project was the blending of feedback and feed-forward sensors to achieve "optimum" performance. These feedback sensors included the gimbal resolvers and Inertial Angle Sensors (IAS), whose signals were blended with the variably-delayed host vehicle attitude data coming from the Inertial Measurement Unit (IMU). The blending of these sensors was greatly simplified using QFT. Using this technique, design trade-off options could be easily weighed against each other in terms of cost versus performance benefit.

Another significant challenge was overcoming the structural resonances in the telescope. The performance goal of 1 microradian pointing stability resulted in a relatively high gain/bandwidth controller. Because of the large bandwidth, several structural resonances were excited, causing pointing performance degradation. Several notch filters were added to the controller to eliminate the possibility of exciting these structural resonances. The extra phase lag introduced by these notch filters decreased the phase margin of the closed-loop system, which in turn further degraded pointing performance. Ultimately the "optimum" trade-off between gain and phase margin was obtained to produce a controller that exceeded the design goals without exciting structural resonances.

Because a suitable radiation-hardened floating-point CPU was not available to implement the control algorithms, all components of the feedback compensator and notch filters were implemented in analog hardware. Thus, the controller modifications, which were necessary to avoid exciting the structural resonances, were more difficult to create. Because of the short design cycle, it was not possible to fabricate new printed circuit boards in time to meet the delivery schedule. Therefore, all modifications had to be carefully reworked on the original flight hardware. This provided another significant challenge.

Yet despite these obstacles, the pointing control system was delivered on time, with measured in-flight performance that meets the design goal with significant margin

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