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Design and Implementation of the Control System for a 2 kHz Rotary Fast Tool Servo

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Introduction

This paper presents a summary of the performance of our 2 kHz rotary fast tool servo and an overview of its control systems. We also discuss the loop shaping techniques used to design the power amplifier current control loop and the implementation of that controller in an op-amp circuit.

The design and development of the control system involved a long list of items including: current compensation; tool position compensation; notch filter design and phase stabilizing with an additional pole for a plant with an undamped resonance; adding viscous damping to the fast tool servo; voltage budget for driving real and reactive loads; dealing with unwanted oscillators; ground loops; digital-to-analog converter glitches; electrical noise from the spindle motor switching power supply; and filtering the spindle encoder signal to generate smooth tool tip trajectories. Eventually, all of these topics will be discussed in detail in a Ph.D. thesis that will include this work. For the purposes of this paper, rather than present a diluted discussion that attempts to touch on all of these topics, we will focus on the first item with sufficient detail for providing insight into the design process.

Performance of the 2 kHz Rotary Fast Tool Servo

The use of a fast tool servo with a diamond turning machine for producing non-axisymmetric or textured surfaces on a workpiece is well known. Our fast tool servo has a maximum stroke of 50 μm peak-to-peak (PP) at low frequencies and has demonstrated a 2.5 μm PP stroke at 2 kHz while diamond turning optical quality surfaces. Details on the mechanical design of the fast tool servo were discussed in an earlier paper, so are not repeated here.¹

Figure 1 shows photos of the 2 kHz rotary fast tool servo as well as pictures of a workpiece and measurements of its surface figure. The swingarm has a vertical axis of rotation and carries a diamond tool offset from that axis. Oscillatory rotation of the swingarm causes an in-and-out motion of the tool relative to a workpiece. The workpiece shown in Figure 1 is a 9 mm diameter aluminum cylinder with 250 radial grooves on the face and 250 axial grooves on the outside diameter. Both sets of grooves were machined while the workpiece was rotating at 480 RPM, requiring 2 kHz operation of the fast tool servo. Profiles across either set of grooves are approximately 2.5 μm PP sinusoids with a surface roughness of 12 nm rms along the grooves.

Overview of the Control System

The actuator for the fast tool servo is a commercially available moving magnet galvanometer (Cambridge Technology, Inc. model 6880). This actuator will be referred to as the “motor” in this paper. We designed and built a current control loop having a 30 kHz crossover frequency using analog components to compensate for the motor electrical dynamics, and integrated it with an appropriate power op-amp and linear power supplies. The outer position loop uses a PC-based digital controller (dSPACE GmbH 1103-board) with a sampling rate of 80 kHz to control the tool position. The

¹Richard C. Montesanti and Professor David L. Trumper, “High Bandwidth Short Stroke Rotary Fast Tool Servo”, Proceedings of the American Society for Precision Engineering 2003 Annual Meeting, Volume 30, pages 115–118.

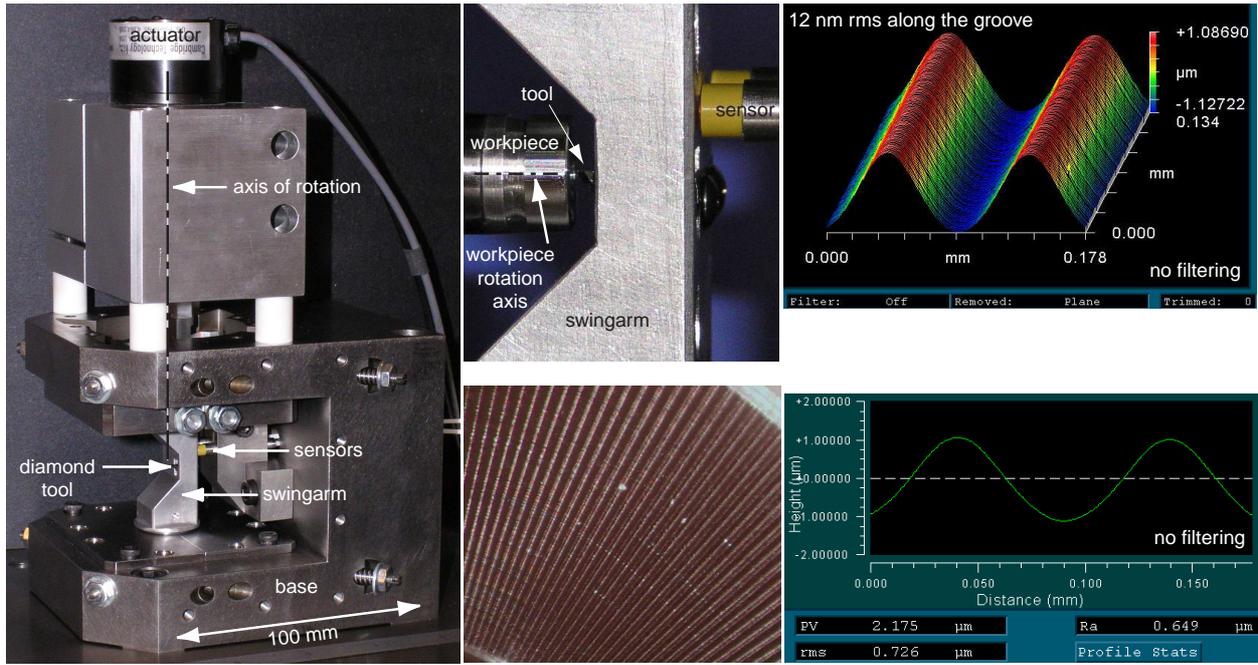


Figure 1: (left) Our 2 kHz rotary fast tool servo. (center, top) Close-up of the tool engaging a workpiece. (center, bottom) Close-up of the face of a machined workpiece with 250 cycles/rev and $2.5 \mu\text{m}$ PP. (right, top) 3-D surface plot from an optical profile of that machined face. (right, bottom) A 2-D slice across the grooves shown in the 3-D plot.

position loop has a crossover frequency of 1 kHz. A pair of eddy current sensors (Kaman Aerospace Corporation model DIT-5200-15N-001) provides position feedback through the differential rotation angle measurement of the swingarm just behind and above the tool. The sensors have a measured noise level on the bench equivalent to a tool motion of 10 nm PP. The 16-bit analog-to-digital converter has a measured noise level of 6 bits PP, which translates into a tool motion of 5 nm PP. In the vicinity of our diamond turning machine with its switching power supply for the spindle motor, after isolating the sensor and machine grounds and adding an analog low-pass filter to the sensor signal, the sensor noise in the digital controller is 15 nm PP. Without forced cooling of the motor we are able to operate at a coil current of ± 8 amps at 2 kHz. The addition of forced cooling would allow doubling the current to achieve the design point of $5 \mu\text{m}$ PP at 2 kHz.

Figure 2 is a block diagram of the tool position control system for the fast tool servo with viscous

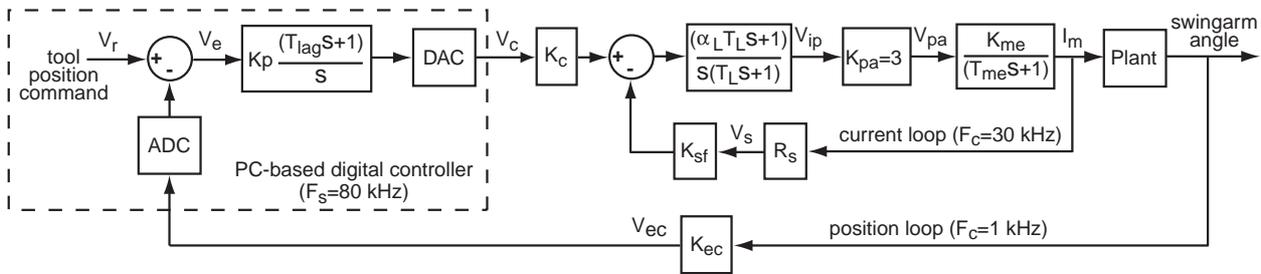


Figure 2: Block diagram of the tool position control system for the fast tool servo with viscous damping.

damping. Our design goal of a position loop closed-loop bandwidth of 2 kHz suggested designing the current compensation to have a loop transmission crossover frequency above 20 kHz, and 30 kHz satisfies this. The constant gain block “ $K_{pa}=3$ ” represents the power amplifier, which is a power op-amp (Apex Microtechnology Corporation PA04) configured for a non-inverting DC gain of 3. The measured gain of the power amplifier is essentially constant up to 1 Mhz, with negligible phase up to 100 kHz and only about -8° of phase at 200 kHz.

A satisfactory mechanical model of the fast tool servo is two lumped rotary inertias joined by a torsional spring. The two inertias are the motor rotor and the swingarm, and the torsional spring is the rigid coupling between them in series with the motor output shaft. Since the flexure blades used to support the swingarm are much more compliant than the rigid coupling and the motor output shaft, to first order they can be ignored and the coupled mode of the motor rotor and the swingarm can be treated as a free mass. Additionally, we have ignored the speed-dependent back-EMF of the motor, providing a further simplification of the dynamic model.

The first resonant mode of the two inertias occurs at approximately 5 kHz. Initially, we experimented with the fast tool servo without using additional viscous damping. This required using either a notch filter on the 5 kHz resonance or an additional pole at 5 kHz to phase stabilize the resonance.² In both cases, achieving a crossover frequency of 1 kHz required using a lead compensator to achieve a magnitude roll-off of -1 decade per decade (dec/dec) of frequency in the vicinity of crossover. A lag compensator was also used, with its pole set at zero to achieve a high desensitivity (controller authority) at low frequencies for rejecting disturbances.

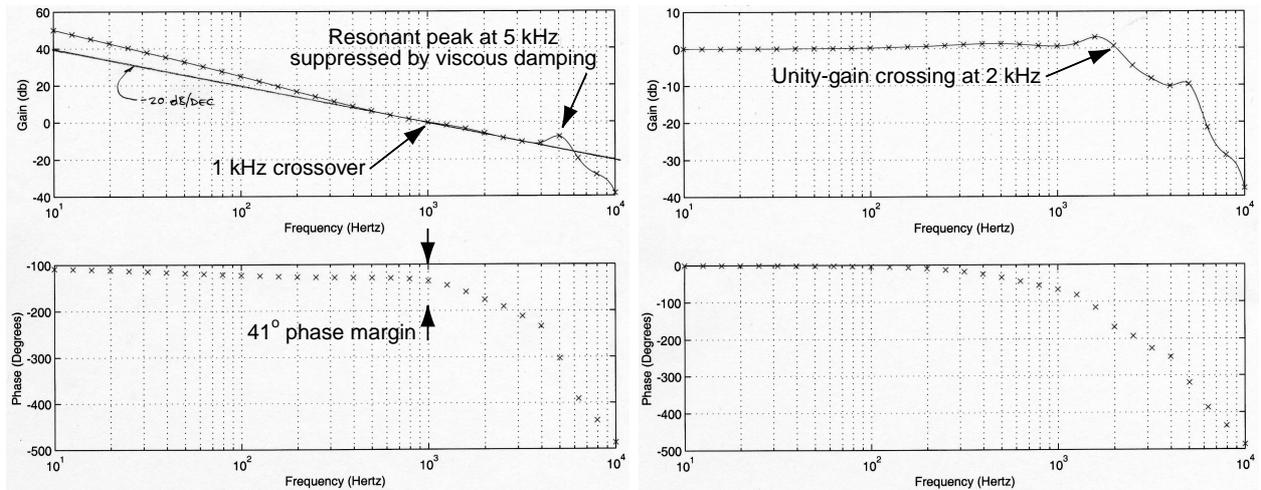


Figure 3: Measured transfer functions for the tool position of the fast tool servo with viscous damping. (left) Negative loop transmission. (right) Closed-loop response.

The two designs above have low phase margins of 13° and 23° , respectively. To overcome this, we added viscous damping to the fast tool servo mechanism. The viscous damper is a circular plate attached to the bottom of the swingarm and captured in a tight-fitting reservoir of heavy grease. With this damper, the 5 kHz resonance peak magnitude is reduced by 20 dB. Figure 3 shows the measured transfer functions for the tool position loop after viscous damping was added to the fast tool servo. The loop transmission crossover frequency is 1 kHz with a phase margin of 41° . Note in

²Phase stabilizing with an additional pole is best understood by considering a Nyquist plot of the negative loop transmission. The additional pole causes the previous encirclement of the -1 point to swing away from that point.

the closed-loop response that the phase angle changes fairly smoothly up to the unity-gain crossing frequency of 2 kHz, which will make feedforward compensation viable for improving the performance of the fast tool servo.

Current Control Loop

Referring back to Figure 2, the current control loop is the portion of the block diagram connecting the command voltage (V_c) to the motor current (I_m). As was mentioned earlier, the constant gain block $K_{pa}=3$ represents the power amplifier. The motor electrical dynamics are in the block after the power amplifier. In this case, the coil inductance and resistance combine with the current sense resistor (R_s) to create a time constant of $T_{me} = 2.55(10)^{-4}$ (sec) and a DC gain of $K_{me} = 0.91$ (amps/volt). The compensation dynamics for this loop are in the block preceding the power amplifier and consist of a lead transfer function cascaded with a pure integrator. More will be said about tuning the compensator dynamics later in this section. The constant gain term of the compensator was put on the input side of the summing junction in the form of the gains K_c and K_{sf} to make the assignment of resistor values for the compensation op-amp circuit more straightforward.

The remainder of this section of the paper will use the Bode magnitude plot of the negative loop transmission of the current control loop to illustrate the loop shaping techniques used to design the current compensation, and then describe the mapping from the block diagram to the op-amp circuit used to realize it.

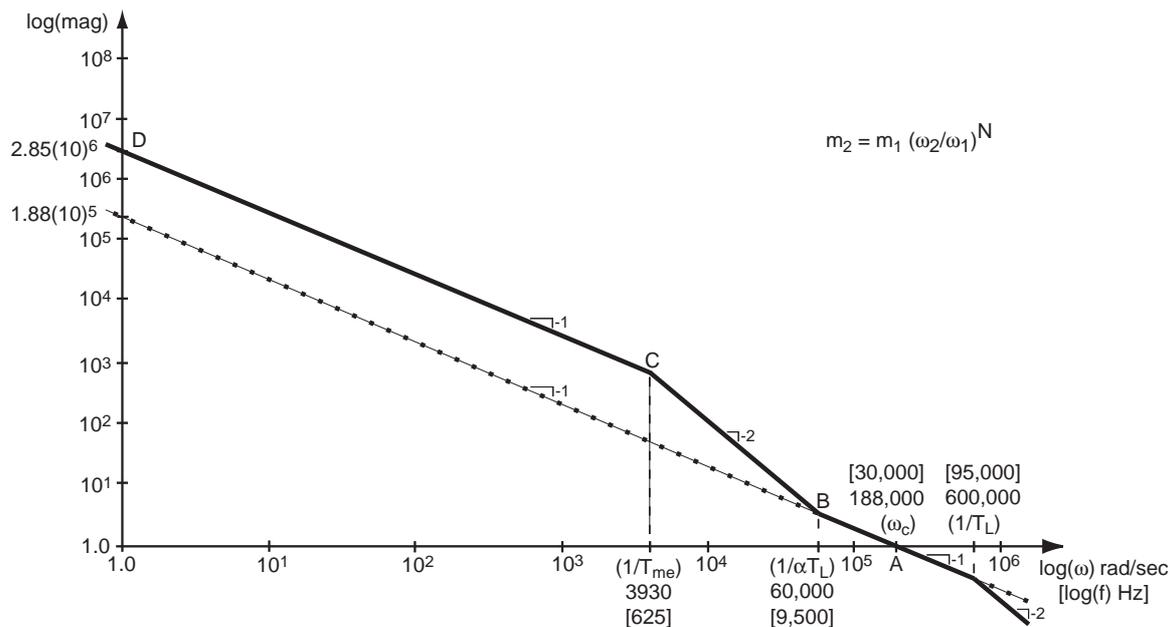


Figure 4: Straight-line approximation of the Bode magnitude plot for our current control loop transmission (upper, solid curve), and for an alternate compensation (lower, dotted line).

The solid upper curve in Figure 4 is a straight-line approximation of the Bode magnitude plot for the current control loop transmission used in our fast tool servo. Starting at the low frequency end, the pure integrator provides the current loop with a high gain for rejecting low frequency disturbances. As we move to the right towards higher frequencies along the magnitude curve we encounter the pole from the motor electrical dynamics ($1/T_{me}$), which causes the slope to decrease from -1 (dec/dec) to -2 (dec/dec). Before reaching the desired crossover frequency (ω_c) we encounter the lead zero ($1/\alpha T_L$), which brings the slope back up to -1 (dec/dec) so that we can pass through crossover

with a positive phase margin. After crossover we reach the lead pole ($1/T_L$), which rolls off the gain at -2 (dec/dec) to avoid high frequency noise. More will be said about high frequency noise later in this section. Regarding the phase angle, which is not shown in the figure, in the case of our current compensation circuit the calculated angle stays above -135° for frequencies below and around crossover. The measured closed-loop performance of the current loop has a -3 dB point at 30 kHz and negligible phase up to 4 kHz. If there was a larger spread between the motor electrical pole and the lead zero, then greater attention would need to be paid to the possibility of conditional stability of the loop in the event of any saturation of the power elements.

Designing the current control loop compensation is accomplished by first assigning values to the lead transfer function, and then moving backwards along the magnitude plot from the crossover point to determine the product of the constant gain terms in the loop transmission. Choosing $\alpha = 10$ for the lead transfer function provides a good balance between a maximum phase advance of 55° and limiting the increase in gain on the higher frequency signals in the compensation network to a factor of 10. The time constant (T_L) is picked by setting the geometric mean of the lead zero and pole equal to the crossover frequency so that the maximum phase advance occurs at crossover. To determine the product of the constant gain terms in the loop transmission we start at crossover and use the similar triangles relationship shown in the upper-right corner of Figure 4 to move backwards along the magnitude plot until reaching the frequency 1 (rad/sec). Since the transfer functions for the current loop in Figure 2 are expressed in Bode format, then the product of the constant gain terms in the loop transmission is equal to the magnitude at point “D”. As an interesting aside, the product of the constant gain terms found using this graphical method agreed within 1% of the value found using a more rigorous approach with a MATLAB script.

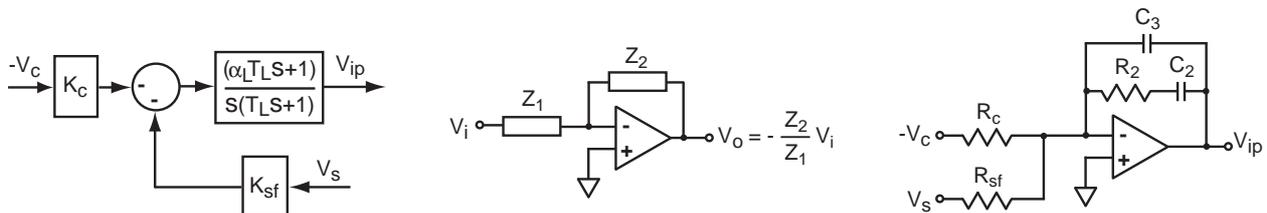


Figure 5: Mapping of the current compensation block diagram (left) to an op-amp circuit (right). The circuit with the generic complex impedances (center) provides the mapping.

Figure 5 shows the mapping from the current compensation block diagram to an op-amp circuit. One way to understand this mapping is to consider qualitatively what we want each element in the compensation transfer function to do. At low frequencies we want the compensation to act like an integrator. At mid frequencies, just before and around crossover, we want the zero to compensate for the integrator and make the gain constant. At high frequencies (above crossover) we want the pole to roll off the gain. The middle circuit with the generic complex impedances in Figure 5 shows the ideal op-amp relationship between the input and output, which is a valid assumption because of the high speed op-amp that we used.³ Referring to the actual op-amp circuit on the right in Figure 5, for the moment ignore the capacitor C_3 . At low frequencies the capacitor C_2 lets very little current through the feedback path, so its impedance dominates over that of the resistor R_2 and we have an integrator. At mid frequencies the impedance of the capacitor C_2 becomes negligible compared to that of R_2 , so the gain approaches a constant value of $(-R_2/R_c)$ or $(-R_2/R_{sf})$, depending on which

³The current compensation circuit uses a unity-gain stable op-amp with a 3.5 MHz gain-bandwidth product (Burr-Brown OPA602).

input is being considered. Now add the capacitor C_3 , whose value is chosen so that its impedance is high at low and mid frequencies, and low at high frequencies. At low and mid frequencies C_3 looks like an open circuit and acts as if it were not there, and at high frequencies C_3 looks like a short and the gain approaches zero. The inverting terminal of an op-amp configured as shown in Figure 5 acts as a summing point for the individual input signals acting through their respective impedance (Z_1), so superposition applies. To obtain the passive component values, the ratio of complex impedances (Z_2/Z_1) is formed for each input in terms of the component variables, and then terms are matched with the coefficients for the corresponding transfer function in the block diagram. The values of the gains K_{sf} and K_c are set so that a ± 10 amps swing in the motor current (I_m) causes the same voltage swing at the inverting terminal as does a ± 10 volt swing in the command voltage (V_c).

The lower, dotted line with a constant slope of -1 (dec/dec) in Figure 4 represents an alternate current compensation network that might be attractive to use. In this case, the compensation zero cancels the stable pole from the motor electrical dynamics and a free integrator is used to provide high gain at low frequencies. The circuit topology for this compensation is the same as the one shown in Figure 5 after omitting the capacitor C_3 . Finding the product of the constant gain terms in the loop transmission and assigning values to the analog component are accomplished by the same methods described above. The alternate compensation provides a constant phase angle of -90° at all frequencies, and therefore a conservative phase margin of 90° . Note that the gain at frequencies below the motor pole is approximately a factor of ten less than that of the more aggressive compensation that we used (described earlier). Therefore, the more aggressive compensation is better at rejecting disturbances in the current control loop. The difference in gain between the two compensators is in the op-amp circuit, and a factor of ten increase requires paying closer attention to the signal levels in the current loop. Also, note that the alternate compensation needs an additional high frequency pole above the crossover frequency to roll off its gain to avoid high frequency noise. This is done by including the capacitor C_3 shown in Figure 5.

With regard to high frequency noise and the need to roll off the gain of the compensator after crossover, considering the internal signal V_{ip} of the current loop in Figure 2 provides the necessary insight. Note that the op-amp feedback compensation does not include the pole from the motor electrical dynamics. Without the high frequency pole in the compensation, once we reach the frequency of the compensator's zero the gain of the op-amp circuit would stay at a constant (high) value for all frequencies above it, making the circuit sensitive to any high frequency noise. This is unnecessary, and easily mitigated by placing a high frequency pole in the compensation.

Summary

This paper presented a summary of the performance of our 2 kHz rotary fast tool servo and an overview of its control system. A discussion on the loop shaping techniques used to design the current control loop and how to map its block diagram to an op-amp circuit was presented to provide insight into the design process.

Future work will include adding Adaptive Feedforward Cancellation compensation for improving the performance of the fast tool servo when using it to generate multiple frequency sinusoids on a workpiece, and packaging the electronics for more rugged use.

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