ROTARY-AXIAL SPINDLES FOR
PRECISION MACHINING

by

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Abstract

This thesis presents the design, analysis, fabrication, instrumentation, and control of a new type of machine tool spindle. Its primary contributions include the design and experimental demonstration of: two rotary-axial spindle prototypes, MIMO current control for a 1 kW linear power amplifier, sensorless rotary motion feedback, a novel method for increasing ADC resolution, and loop-shaping motion control systems.

Some machining operations, such as face grinding, require rotational and feed motion to remove material. Conventional machine tools accomplish this by attaching a spindle which has thrust and journal support to a feed drive which also has thrust and lateral support systems. In modern machine design, the trend is towards increasing the stiffness of each individual element. However, the inherent serial duplication of support presents a fundamental limitation to stiffness and precision.

The rotary-axial spindle architecture alleviates this problem by discarding the feed drive and spindle thrust bearing, replacing them both with a high force electromagnetic actuator. This provides millimeter range stroke for the spindle shaft, resulting in a single inertial element capable of both rotary and axial motion. This topology has several advantages. It allows kHz range bandwidth and hundreds of N/μm dynamic stiffness, improves acceleration, reduces structural bending moments, and eliminates thermal effects of fluid thrust bearings.

Two prototypes are developed to demonstrate this technology. The first is a small scale rotary-axial spindle. Driven by a four-channel 1 kW linear power amplifier with decoupled current loops, the magnetic thrust bearing can handle 600 N peak axial loads over a 1 mm stroke. A novel method for increasing ADC resolution achieves sub-5 nm RMS positioning noise. Loop shaping compensation of the position loop results in 100 N/μm minimum dynamic stiffness and
2.6 kHz closed loop bandwidth. To control the spindle speed, a sensorless rotary motion feedback algorithm was developed. It produces results equivalent to a 1000 line rotary encoder. The second prototype is a full size machine tool. It demonstrated 6 kN continuous axial load capacity, 340 N/μm minimum dynamic stiffness, 800 Hz bandwidth, and 7 nm RMS positioning noise over a 1.5 mm stroke.
# Table of Contents

Abstract ........................................................................................................................................... ii  
Table of Contents ........................................................................................................................... iv  
List of Tables .................................................................................................................................. viii  
List of Figures ............................................................................................................................... ix  
Acknowledgments ....................................................................................................................... xxii  
Dedication .................................................................................................................................. xxiv  
1 Introduction .............................................................................................................................. 1  
  1.1 Thesis overview .................................................................................................................. 2  
    1.1.1 Machine tool bearings and architectures ................................................................. 4  
    1.1.2 Rotary-axial spindle analysis, design and prototyping ........................................... 5  
    1.1.3 Power electronics .................................................................................................... 6  
    1.1.4 Rotary sensing and control ...................................................................................... 7  
    1.1.5 Axial sensing ........................................................................................................... 8  
    1.1.6 Axial control ........................................................................................................... 9  
    1.1.7 Full scale rotary-axial spindle prototype ............................................................... 10  
2 Machine Tool Bearings and Architectures ........................................................................... 12  
  2.1 Machine tool bearing technology ................................................................................... 13  
  2.2 Typical machine tool configuration ............................................................................. 16  
  2.3 Magnetic thrust bearing spindles .................................................................................. 18  
  2.4 Rotary-axial spindle architecture ................................................................................ 19  
  2.5 Summary ....................................................................................................................... 21  
3 Rotary-axial Spindle Analysis, Design, and Prototyping ...................................................... 22  
  3.1 Electromagnetic thrust actuator/bearing ....................................................................... 23  
    3.1.1 Magnetic bearing selection ................................................................................... 23  
    3.2 Electromagnetic bearing force analysis .................................................................... 26  
    3.2.1 Force calculation with ideal material properties ................................................... 26  
    3.2.2 Finite element analysis .......................................................................................... 33  
  3.3 Rotary-axial spindle design ............................................................................................ 41  
    3.3.1 Magnetic bearing assembly and brushless motor .................................................. 42  
    3.3.2 Aerostatic journal bearing assembly ..................................................................... 44  
    3.3.3 Shaft assembly ...................................................................................................... 46  
    3.3.4 Position sensor assembly ...................................................................................... 46  
    3.4 Structural analysis ....................................................................................................... 47  
    3.4.1 Model simplification and boundary conditions ....................................................... 48  
    3.4.2 Structural analysis and results ............................................................................... 49  
  3.5 Assembly ......................................................................................................................... 50  
  3.6 Summary ......................................................................................................................... 56  
4 Linear Power Amplifier for Magnetic Bearing ....................................................................... 57  
  4.1 Power amplifier requirements ......................................................................................... 58
| 6.5.2  | FPGA-20 bit DAC communication protocol ........................................ 152 |
| 6.6    | Noise modeling .................................................................................. 155 |
| 6.7    | Summary ................................................................................................. 159 |
| 7      | Axial Control and Experimental Results .................................................. 161 |
| 7.1    | System overview .................................................................................. 162 |
| 7.2    | Loop shaping controller design ............................................................... 164 |
| 7.2.1  | Model .................................................................................................. 165 |
| 7.2.2  | System identification and refined model ................................................ 168 |
| 7.2.3  | Refined results ................................................................................... 172 |
| 7.3    | Force measurement ............................................................................... 176 |
| 7.4    | Integrating *Nanozoom* ...................................................................... 177 |
| 7.5    | Further enhancement of axial control loop performance ......................... 182 |
| 7.5.1  | Revised plant model ............................................................................ 182 |
| 7.5.2  | Current saturation ............................................................................... 184 |
| 7.5.3  | Dual controller approach .................................................................... 186 |
| 7.5.4  | Controller improvement ....................................................................... 188 |
| 7.5.5  | Maximizing the axial control loop performance ...................................... 190 |
| 7.6    | Combining rotary and axial control loops ............................................... 195 |
| 7.6.1  | Multi-tasking control implementation .................................................. 195 |
| 7.6.2  | Rotary-axial motion coupling ............................................................... 199 |
| 7.7    | Performance comparison with aerostatic spindle ..................................... 204 |
| 7.8    | Summary ................................................................................................. 205 |
| 8      | Full Scale Rotary-axial Spindle ............................................................. 207 |
| 8.1    | Design .................................................................................................... 209 |
| 8.1.1  | Design goals ........................................................................................ 209 |
| 8.1.2  | Architecture ......................................................................................... 211 |
| 8.1.3  | Magnetic bearing ............................................................................... 211 |
| 8.1.4  | Actuator design for assembly considerations ....................................... 214 |
| 8.1.5  | Brushless DC motor assembly ............................................................... 215 |
| 8.1.6  | Hydrostatic bearing ............................................................................. 215 |
| 8.1.7  | Moving assembly .................................................................................. 216 |
| 8.1.8  | Metrology .............................................................................................. 218 |
| 8.2    | 2 kW power amplifier .......................................................................... 221 |
| 8.2.1  | Power amplifier hardware ..................................................................... 221 |
| 8.2.2  | Coil configurations .............................................................................. 225 |
| 8.2.3  | Parallel-serial coil configuration control ............................................. 226 |
| 8.2.4  | Pure parallel coil configuration control .............................................. 233 |
| 8.3    | Axial motion control ............................................................................ 234 |
| 8.3.1  | Actuator force characterization ............................................................. 234 |
| 8.3.2  | Axial plant identification and controller design .................................... 235 |
| 8.4    | Rotary motion control .......................................................................... 241 |
| 8.4.1  | Sensorless rotary motion feedback for ten HES motor ............................ 241 |
| 8.4.2  | Rotary plant identification and controller design .................................. 241 |
| 8.5    | Hydrostatic bearing losses .................................................................. 247 |
| 8.6    | Rotary-axial motion coupling ............................................................... 249 |
| 8.7    | Summary ................................................................................................. 250 |
Conclusion and Future Work .................................................................................................................. 252
9.1 Conclusions ........................................................................................................................................ 252
  9.1.1 Rotary-axial spindle concept ....................................................................................................... 252
  9.1.2 Design and analysis of rotary-axial spindle ............................................................................... 252
  9.1.3 MIMO current control ............................................................................................................... 253
  9.1.4 Sensorless rotary speed measurement ....................................................................................... 253
  9.1.5 Novel method for increasing ADC resolution .......................................................................... 254
  9.1.6 Axial control and experimental results of first prototype ....................................................... 254
  9.1.7 Full-size prototype .................................................................................................................... 255
9.2 Future work ....................................................................................................................................... 256
  9.2.1 High power linear amplifiers .................................................................................................... 256
  9.2.2 Shaft dynamic study ................................................................................................................ 256
  9.2.3 Improvement of axial error motion through design for assembly ........................................... 257
  9.2.4 Modified sealing designs for hydrostatic bearings ................................................................. 257
  9.2.5 Control loop jitter investigation ............................................................................................... 257
  9.2.6 Cutting tests ................................................................................................................................ 258

References ................................................................................................................................................. 259
List of Tables

Table 3.1: Current to force ratio at increasing axial armature position ........................................ 39
Table 3.2: Position to force ratio at various coil currents ............................................................. 40
Table 4.1: Resistor selections ..................................................................................................... 67
Table 4.2: Parameters for current control circuits ........................................................................ 89
Table 5.1: Rotary control loop parameter list .............................................................................. 105
Table 5.2: HES counting sequence lookup table ........................................................................ 115
Table 5.3: Summary of quadrature signal output sequence ........................................................ 117
Table 7.1: Values of parameters in open loop axial system ............................................................ 167
Table 7.2: Modal parameters for axial vibration modes ................................................................. 171
Table 7.3: Modal parameters for revised motion control plant ..................................................... 184
Table 7.4: Performance comparison between Precitech aerostatic spindle and rotary-axial spindle ........................................................................................................................................ 205
Table 8.1: Full size rotary-axial spindle design requirements ....................................................... 209
Table 8.2: Magnetic bearing design parameters .......................................................................... 214
Table 8.3: Summary of eight channel amplifier stability ............................................................. 231
Table 8.4: Full-size rotary-axial spindle prototype rotary control loop parameter list ....................... 243
List of Figures

Figure 1.1: Rotary-axial spindle prototype. This prototype uses a Precitech SP-75 work holding spindle (generously donated by Precitech) and its aerostatic journal bearings. Its aerostatic thrust bearing was removed and replaced with a high force thrust actuator with 1 mm stroke. A DC brushless motor applies torque to the shaft to produce rotary motion........................................... 3

Figure 1.2: The rotary-axial spindle system. The main components are the rotary-axial spindle, the linear power amplifier, the capacitive probe and its signal conditioning hardware, the sensor signal processing board, the rotary power amplifier, and the dSPACE control unit. ....................... 4

Figure 1.3: Schematic of axial magnetic bearing [3]. Used with permission. ............................... 5

Figure 1.4: Magnetic bearing power amplifier ............................................................................... 7

Figure 1.5: Comparison between interpolated (256x) with adapted HES derived speed feedback signal and optical encoder derived speed feedback (tested on conventional spindle). ................... 8

Figure 1.6: Printed circuit board which implements Nanozoom.................................................... 9

Figure 1.7: Plot of dynamic stiffness of axial control loop.......................................................... 10

Figure 1.8: Full scale rotary-axial. The total length is 1 m and the spindle shaft is 120 mm in diameter......................................................................................................................................... 11

Figure 2.1: Exploded view of a typical rolling element bearing. Photo used with permission [5]. ....................................................................................................................................................... 13

Figure 2.2: Schematic view of a fluid-static bearing. Pressurized fluid is fed to the bearing from an external source. The small gap between the shaft and journal allows pressure to build, thus constraining the shaft. ................................................................................................................... 16

Figure 2.3: Conventional machine tool configuration. Here, the spindle is mounted to a Z axis slide. In the radial direction, the journal bearing and Z axis slide support stiffnesses are duplicated..................................................................................................................................... 17

Figure 2.4: Rotary-axial spindle architecture. In this topology the redundant thrust and radial supports are eliminated. The key benefits of this are an overall increase in stiffness, bandwidth and precision, and reductions in structural bending moments, heat generation, component count and cost. ........................................................................................................................................ 20

Figure 3.1: Cross section view of rotary-axial spindle’s homo-polar normal stress magnetic bearing. A section cut through the magnet ring and armature is also shown. .............................. 25

Figure 3.2: Schematic isolating constant flux from permanent magnet. ........................................ 27

Figure 3.3: Finite element result showing leakage fluxes from stator with centered armature... 29
Figure 3.4: Diagram used for calculating flux and force contributed by excitation currents. The red line indicates the path and positive direction of AC flux flow. .......................................................... 29

Figure 3.5: Simplified reluctance model of magnetic thrust bearing. Each sub-circuit contains a flux source along with the reluctance through which it flows. a) Permanent magnet flux and permanent magnet reluctance. b) Flux in air gap to the right of the armature and right air gap reluctance. c) Flux in air gap to the left of the armature and left air gap reluctance. Reluctance $R_1$ and $R_2$ are variable. ....................................................................................................................... 31

Figure 3.6: Finite element model showing the actuator and boundary condition.......................... 34

Figure 3.7: BH curve for pressed powdered iron with Kenolube® bonding agent used as armature and stator materials. Höganäs reports an initial permeability of 500 [28].......................... 35

Figure 3.8: Hysteresis curve for a typical permanent magnet. The remanence is the flux density remaining in the magnet after the external field is removed. The coercivity is the strength of the external field required to demagnetize the magnet. .......................................................................................... 36

Figure 3.9: Finite element mesh showing the varying element density. The mesh in the air surrounding the actuator can be made coarser to improve analysis speed. ............................................ 38

Figure 3.10: Net force on armature vs. coil current. The force vs. current plot is shown at different axial positions. The data at the bottom of the plot corresponds to the force calculated with the armature in the center position. The data at the top of the plot corresponds to the force calculated with the armature 480 µm away from center......................................................... 39

Figure 3.11: Net force on armature vs. axial position at different coil currents.............................. 40

Figure 3.12: Section view of rotary-axial spindle. The main assemblies are the aerostatic journal bearing assembly, brushless motor assembly, magnetic bearing assembly, and the shaft assembly. Fully assembled, it is approximately 100 mm x 100 mm x 260 mm................................. 42

Figure 3.13: Sectioned and exploded view of magnetic bearing assembly. It can be further broken down into the front, middle, and rear assemblies. The front and rear assemblies each consist of a frame, a set of coils, and a four piece powdered iron stator. The middle assembly frame holds the middle stator portion as well as the twelve magnet segments. The brushless motor stator fits into the front frame................................................................. 43

Figure 3.14: Magnetization direction of permanent magnet arc segments. Manufacturing limitations prevent radial magnetization................................................................. 44

Figure 3.15: Aerostatic journal bearing housing. The journal bearing, taken from a Precitech SP75 spindle, is made of bronze. The internal diameter is approximately 44 mm..................... 45

Figure 3.16: Shaft assembly. Axial force is applied to the armature by the magnetic bearing. The axial position of the assembly is sensed at the target screw. Torque applied to the permanent magnet rotor by the brushless motor stator provides the rotary motion. ........................................ 45
Figure 3.17: Position sensing assembly. a) Probe holder. Two slots are cut to allow for flexing of the clamp once the bolts are tightened. b) Capacitive probe installed in holder. .............................. 47

Figure 3.18: Dominant vibration modes of spindle shaft up to 10 kHz. a) First bending mode. b) Combined axial-torsional mode. c) Second bending mode. d) Pure axial mode. e) Third bending mode ............................................................................................................................................. 49

Figure 3.19: Rotary-axial spindle components designed at UBC ................................................................. 50

Figure 3.20: Coil connection schematic. All the coils are configured into four circuits. Each circuit consists of four coils: one pair of coils connected in parallel, and then connected in series with another pair. The channel current is equal to twice the coil current .................................................. 51

Figure 3.21: Illustration of stator segment and a single coil in frame. A slot, not shown here, is cut in the side of the assembly to allow for the exit of the coil leads. ................................................................. 52

Figure 3.22: Fully assembled front and rear magnetic bearing assemblies. In this photo each assembly has had the stator segments and coils epoxied in (two step process). The surfaces have been precision ground to achieve excellent surface contact. Terminal blocks have been installed for the coil connections ................................................................................................. 52

Figure 3.23: Brushless motor stator supplied by Aerotech. The three Hall Effect sensors are spaced 120° (electrical) apart ......................................................................................................................... 53

Figure 3.24: Fully assembled magnetic bearing middle assembly. The magnet arcs are shown along with the middle stator segments epoxied in to the stainless steel frame. Again, both sides of the assembly are precision ground to achieve excellent contact with the other assemblies ....... 53

Figure 3.25: Fully assembled spindle shaft assembly. The journal bearing shaft is an existing component taken from the Precitech spindle. Once the additional components are mounted to the extension shaft, it is mounted to the journal bearing shaft. Axial force is applied to the magnetic bearing armature to provide feed motion and a torque produced by the brushless motor stator acts on the permanent magnet rotor to rotate the shaft ................................................................. 54

Figure 3.26: Position sensor assembly. Here, the capacitive probe is shown mounted in the probe holder. The additional copper improves the ground connection between the probe and the back shell ......................................................................................................................................... 55

Figure 3.27: Fully assembled rotary axial spindle ......................................................................................... 55

Figure 4.1: Linear power amplifier for magnetic bearing. It consists of two printed circuit boards: one for voltage control and one for current control. A half inch copper plate acts as a heat sink while nine finned copper CPU fans dissipate heat to ambient ................................................................. 58

Figure 4.2: Power amplifier architecture. Each PA52 power device, which is enclosed in a voltage control loop, drives a coil load. The voltage loops receive references from the MIMO current controller ..................................................................................................................... 61
Figure 4.3: Voltage control principle. a) PA52 is enclosed in a non-inverting configuration with $R_2$ reducing the gain of the loop transmission to improve stability. b) Equivalent circuit representation. c) Block diagram for voltage control loop. .......................................................... 63

Figure 4.4: PA52 small signal open loop frequency response [33:3]. Loop transmission gain is shown in red on the left side. ........................................................................................................ 66

Figure 4.5: Voltage control board mounted to copper plate. .......................................................... 68

Figure 4.6: View of bottom side of power amplifier showing nine copper CPU fans which provide air cooling. ....................................................................................................................... 68

Figure 4.7: 15 V step response of power devices. The step response time constant corresponds well with the crossover frequency. ............................................................................................... 69

Figure 4.8: Impedance measurement schematic. ............................................................................. 72

Figure 4.9: Frequency response of self impedances. .................................................................... 73

Figure 4.10: Frequency response of closely coupled impedances. ............................................... 74

Figure 4.11: Frequency response of loosely coupled impedances.................................................. 74

Figure 4.12: Current control block diagram. ................................................................................ 76

Figure 4.13: Feedback stage implementation. a) Summer circuit to generate $i_{1S}$. b) Difference circuit to generate $i_{2S}$. c) Difference circuit to generate $i_{3S}$. d) Difference circuit to generate $i_{4S}$. 79

Figure 4.14: a) schematic of PI control with additional pole for current control. b) controller block diagram assuming ideal op-amp. ............................................................................. 81

Figure 4.15: These summing circuits add the control signals which are fed as reference signals to the power stages. Circuit a) feeds device A. Circuit b) feeds device B. Circuit c) feed device C. Circuit d) feeds device D. ............................................................................................................. 83

Figure 4.16: Current control PCB. The four feedback circuits are located along the top of the board. The four control circuits are situated in the middle of the board. The operational amplifiers summing the control signals for voltage references can be found along the bottom of the board. Common mode reference signal is inputted from the BNC connectors on the right hand side of the board.................................................. 84

Figure 4.17: Schematic of experimental setup for measuring open loop MIMO plant. ............ 85

Figure 4.18: Comparison between experimental measurements of diagonal based plants ($v_j$ to $i_j$, $j = 1$ to 4) and a plant model based on the impedance measurements. The difference at high frequency is due to very slight differences in $Z_0$ and $Z_1$. This is caused by parasitic capacitance in the BNC cables of the measurement device. ........................................................................................................ 86
Figure 4.19: Decoupling results showing the magnitude of the frequency response from a) $v_1$ to $i_{1S}$ through $i_{4S}$ b) $v_2$ to $i_{1S}$ through $i_{4S}$ c) $v_3$ to $i_{1S}$ through $i_{4S}$ d) $v_4$ to $i_{1S}$ through $i_{4S}$ .......................... 87

Figure 4.20: Negative of Loop Transmission for a) $v_1$ to $i_{1S}$ b) $v_2$ to $i_{2S}$ c) $v_3$ to $i_{3S}$ and d) $v_4$ to $i_{4S}$. ....................................................................................................................................................... 90

Figure 4.21: Comparison between simplified and full-order stability ............................................. 92

Figure 4.22: Closed loop current response at various signal amplitudes. The drop off at higher amplitudes is due to saturation of the controller output signals ................................................................. 93

Figure 4.23: 20 mA step response of quad-channel power amplifier .................................................. 94

Figure 4.24: 0.5 A step response showing voltage saturation .......................................................... 95

Figure 5.1: Encoder schematic. Side view shows the required constant air gaps .............................. 95

Figure 5.2: Schematic of drum type encoder. These encoders can tolerate small axial motions. 99

Figure 5.3: Precitech SP75 with custom encoder mounting ............................................................... 101

Figure 5.4: Block diagram of rotary control loop hardware ............................................................ 102

Figure 5.5: Open loop rotary control plant ...................................................................................... 104

Figure 5.6: Open loop frequency response from control effort $u$ [V], to measured spindle speed $\omega$ [RPM]. The model (red) matches very well with the experimental result (blue). The noisy result at high frequency is due to insufficient output amplitude ................. 106

Figure 5.7: Frequency response of filter $A(z^{-1})$. The parameters are $N_r = 205$ and $T_r = 0.0001$ s. Notches occur at the fundamental frequency of 48 Hz, corresponding to 2925 RPM and at each harmonic up to the Nyquist frequency ................................................................. 107

Figure 5.8: Rotary control block diagram ....................................................................................... 108

Figure 5.9: Negative of loop transmission for rotary control loop .................................................. 109

Figure 5.10: Block diagram of rotary motion hardware layout with sensorless feedback. The custom PCB which includes an FPGA chip, takes the place of the rotary encoder. .................. 110

Figure 5.11: Custom PCB which implements sensorless speed feedback algorithm. The PCB is of the four layer variety and is approximately 220 mm by 160 mm. The left side of the board is dedicated to digital electronics for the rotary speed feedback algorithm. The right side of the board is assigned to analog electronics related to the axial sensing which will be discussed in Chapter 6. The PCB is enclosed in an aluminum housing made by Hammond [45]. .................. 111

Figure 5.12: Left: end view of two pole brushless motor, with three Hall Effect sensors. Right: the HES output sequence for one revolution ................................................................. 112
Figure 5.13: Placement of Hall effect sensors for $P$ pole motor................................................ 113

Figure 5.14: HES and quadrature sequences for eight pole, three HES motor. The HES sequence repeats four times per revolution generating twenty four counts per revolution. ...................... 114

Figure 5.15: Flowchart of HES decoding to quadrature encoding algorithm. This figure describes the input-output relationship between the various processes. VHDL coding allows each process to be executed in parallel. ........................................................................................................... 114

Figure 5.16: Comparison between 24 counts (6 lines) per rev HES derived speed feedback and 4000 counts (1000 lines) per rev optical encoder derived speed feedback. The noise produced by the sensorless algorithm has a fundamental harmonic of the spindle speed. It is significantly larger than the noise resulting from the encoder measurement. ................................................. 118

Figure 5.17: Interpolation principle. ........................................................................................... 119

Figure 5.18: Flowchart of HES interpolation algorithm............................................................. 120

Figure 5.19: Frequency response from $u$ to $\omega$. Notch near 30 Hz is caused by flexibility between stator and spindle housing. The algorithm, above 30 Hz, measures the HES motion......... 123

Figure 5.20: Model of sensorless motion feedback phenomenon............................................... 123

Figure 5.21: Comparison between experimental result and model of sensorless feedback algorithm incorporating HES relative motion measurement. ......................................................... 124

Figure 5.22: Comparison between interpolated (256x) HES derived speed feedback and optical encoder derived speed feedback at 5000 RPM. The interpolation reduces the harmonic error in the feedback, but the encoder still provides a more precise signal. ............................................................. 125

Figure 5.23: Flowchart of interpolation algorithm once online calibration has been integrated. 127

Figure 5.24: Flow chart for “on the fly” calibration state machine. Rectangular blocks are states and elliptical blocks are conditions for proceeding to another state................................. 128

Figure 5.25: Comparison between interpolated (256x) with adapted HES derived speed feedback signal and optical encoder derived speed feedback. The precision of the sensorless speed feedback algorithm with interpolation and “on the fly” calibration is as good as that produced by the encoder’s speed measurement................................................................. 129

Figure 5.26: Comparison between model of rotary-axial spindle rotary control plant and experimental frequency response. Low excitation amplitude at high frequencies result in a less than clean identification................................................................. 131

Figure 5.27: Negative of the loop transmission of rotary-axial spindle rotary control loop. The crossover frequency is 10 Hz with 33° phase margin................................................................. 132
Figure 5.28: 100 RPM step response. The time for a half cycle is 0.05 seconds which corresponds to a 10 Hz crossover frequency.

Figure 5.29: Speed regulation noise at 5000 RPM. The RMS error is 0.170 RPM.

Figure 6.1: Axial sensing solution.

Figure 6.2: Block diagram showing the conventional feedback path of an electromechanical system.

Figure 6.3: Data capture demonstrating 15.3 nm resolution when ±10 V range ADC is scaled to represent ±0.5 mm range. Here, the ADC input is grounded through a 50 Ω terminator.

Figure 6.4: Comparison between a ramp signal sampled at a certain sampling rate a), and a ramp signal sampled at a rate four times faster b). The analog signal is shown in blue, and the digital signals in red. Faster sampling ensures the available ADC resolution is maximized.

Figure 6.5: Schematic representation of Nanozoom invention and its interaction with a generic electromechanical system. In conventional digital control systems, the control error computation is performed in the digital domain. Moving this operation to the analog domain and amplifying it improves the resolution $K_z$ times, ensures high resolution over the entire sensor range, and does not add any additional delay to the system.

Figure 6.6: Discretization of a base signal and its amplified counterpart. Discretizing the amplified version makes more effective use of the ADC resolution.

Figure 6.7: Embodiment of Nanozoom for rotary-axial spindle application.

Figure 6.8: Noise on ADC with inputs grounded. The voltage is converted to nanometers and divided by 8. The resulting quantization interval is 1.9 nm which corresponds to 19 bit resolution, as expected.

Figure 6.9: Architecture of Nanozoom portion of rotary-axial spindle PCB.

Figure 6.10: a) Analog electronics schematic. The first OP27 buffers the DAC signal. The second OP27 adjusts the output range. The third OP27 performs the error subtraction and amplification process. b) Block diagram summarizing the arithmetic implemented by the analog circuitry.

Figure 6.11: PCB for rotary-axial spindle sensing. The Nanozoom related analog electronics are located in the upper right hand corner of the board.

Figure 6.12: Nanozoom communication protocol.

Figure 6.13: dSPACE send protocol. After quantizing and adjusting the range of the reference $x_r$ to integers ranging from 0 to $2^{24} - 1$ (24 bit), it is split into three bytes. This information is sandwiched between start and stop bytes. It waits for a “Go” signal from the FPGA to be sent.
Figure 6.14: The finite state machine used by the FPGA to write to and read from the OPB... 152

Figure 6.15: SPI communication sequence [53:10]. The data bits are sent, synchronously with the serial clock, on the SDIO line. Up to four bytes are transmitted in one write process. The first is an instruction byte which contains the register address and the number of bytes to be sent. This is followed by one to three bytes which correspond to the device setup or the output voltage. The low times for the SCLK, $t_9$ and $t_{14}$ are set by the frequency at which the DAC is clocked (2.5 MHz). .......................................................................................................................................... 153

Figure 6.16: Elementary noise model of an operational amplifier. ................................................. 155

Figure 6.17: Noise model for op-amp which buffers the DAC signal............................................. 156

Figure 6.18: Noise model for op-amp which range adjusts the buffered DAC output............... 157

Figure 6.19: Noise model for op-amp which performs error subtraction and amplification steps. ..................................................................................................................................................... 157

Figure 6.20: Plot showing Nanozoom noise, including the contribution from the 16-bit ADC. After conversion to nanometers and rescaling by $1/K_z$, the noise is approximately 3 nm RMS. 159

Figure 7.1: Interaction between components in axial system. Host computer photo is used with the permission of Microsoft [56]. dSPACE DS1103 and connector panel are used with the permission of dSPACE [57, 58].................................................................................................. 164

Figure 7.2: Block diagram of open loop axial system. DAC, amplifier, axial plant, and ADC dynamics are considered.................................................................. 165

Figure 7.3: Bode plot of the open loop system model ignoring flexible modes. The flat region at low frequency is the negative stiffness of the actuator. The response falls off at -40 dB/dec corresponding to a rigid body mode. There is no resonance because the phase is constantly -180°. ..................................................................................................................................................... 167

Figure 7.4: Block diagram of closed loop system. The controller consists of two terms: a loop shaping term $C_{LS}(s)$ and an integrator. The integrator is implemented in parallel to the loop shaping term so that integral windup can be limited. ..................................................................................................................................................... 168

Figure 7.5: Negative of the loop transmission of the model based loop shaping controller $CLS1$. ..................................................................................................................................................... 169

Figure 7.6: Experimentally measured frequency response from $i_r$ to $x_m$. .................................. 169

Figure 7.7: Comparison between experimentally measured frequency response of the open loop system and the model................................................................. 171

Figure 7.8: Bode plots of controller design results. a) the loop shaping controller including integrator. b) the negative loop transmission showing 1 kHz crossover frequency with 38° phase
margin c) the dynamic stiffness showing minimum of 8 N/μm at 145 Hz. d) shows a 1.5 kHz -3 dB closed loop bandwidth.

Figure 7.9: Rotary-axial spindle setup on lab workbench. A 44 Hz resonance is associated with this table.

Figure 7.10: 10 μm step response of the closed loop system. A strong 1 kHz component is evidence that the designed crossover frequency was achieved.

Figure 7.11: a) time domain plot of position regulation noise. The RMS error is 12.8 nm. b) power spectrum density of position regulation error. Implementing Nanozoom should improve this result.

Figure 7.12: The actuator force characteristic. Data was recorded by using the controller to hold the spindle at a constant position. Then, a known load was applied axially and the steady state current delivered by each power device was recorded.

Figure 7.13: Hardware interaction with Nanozoom integrated into system. Host computer photo is used with the permission of Microsoft [56]. dSPACE DS1103 and connector panel are used with the permission of dSPACE [57, 58].

Figure 7.14: Block diagram of axial control loop with Nanozoom saturation block. The error signal saturates when outside the range ±62.5 μm.

Figure 7.15: Instability due to saturation of Nanozoom during the startup process. Due to the saturation of the Nanozoom error, full force is applied to the armature until the spindle is in a linear feedback zone. At this point the shaft is moving too quickly, and the actuator cannot apply enough force to recover.

Figure 7.16: a) time domain plot of position regulation noise. The RMS error is 5.3 nm. b) power spectrum density of position regulation error.

Figure 7.17: Comparison between frequency responses of the measured plants with three screws (blue) and six screws (red) holding the extension shaft to the air bearing shaft.

Figure 7.18: Comparison between measured and modeled motion control plant frequency responses with six screws installed to connect the air bearing shaft and the extension shaft.

Figure 7.19: Negative of loop transmission for an intermediate loop shaping controller/integrator pair. Even though the phase margin indicates a stable system, the startup process is unstable.

Figure 7.20: Unstable startup capture. The left hand vertical axis shows the current command signal (blue), and the right hand vertical axis shows the axial position as measured by the capacitive probe (red). Full load is applied for too long, violently accelerating the spindle. A current command corresponding to full load in the opposite direction is not strong enough to slow the shaft, and so the armature crashes into the stator, creating instability.
Figure 7.21: Dual controller approach as a remedy to current saturation. Here, one robust linear controller is used for startup. Once stabilized, the logic sequence commands a change of input so that a higher bandwidth controller is selected. The use of a single integrator keeps the slower dynamics continuous, and so stability is maintained................................................................. 188

Figure 7.22: Characteristic Bode plots for CLS3 and its respective integrator. a) Full controller frequency response. b) Frequency response of the negative loop transmission c) Closed loop dynamic stiffness frequency response. d) Closed loop bandwidth frequency response. .......... 189

Figure 7.23: 400 nm step response for the closed loop system regulated by CLS3 and its integrator. The 1390 Hz component corresponds with the crossover frequency. The 1695 Hz oscillation is due to the first mode being above unity gain......................................................... 190

Figure 7.24: Varying compliance of 8500 Hz resonance. Excitation amplitudes ranged from 0.0025 A to 0.1 A.............................................................................................................................................. 192

Figure 7.25: Characteristic Bode plots for improved controller. a) Controller frequency response. b) Frequency response of the negative loop transmission c) Dynamic stiffness frequency response. d) Closed loop bandwidth frequency response. .......................................................... 193

Figure 7.26: Position regulation noise achieved with the highest performance controller. The result is 4.6 nm RMS, which is below the capacitive probe noise floor. The DC component of the noise is less than 0.01 picometers........................................................................................................... 194

Figure 7.27: 300 nm step response of closed loop system as regulated by CLS4 and its integrator. A 1560 Hz component, which nearly corresponds to the first crossover frequency, is observed. The large overshoot is undesirable, but is a result of the quest for maximum dynamic stiffness. ............................................................................................................................................. 194

Figure 7.28: Noise induced on the axial control error signal in a) a single-tasking process and b) a multi-tasking process. The controller used to obtain this data was CLS3 and its corresponding integrator. The difference in the low frequency noise between the two situations is approximately three orders of magnitude. ........................................................................................................ 196

Figure 7.29: Time stamp histograms for 10 μs (100 kHz) sample time. a) Time stamp histogram for single-tasking system. b) Time stamp histogram for multi-tasking system. ......................... 197

Figure 7.30: Frequency response of the discrete filter $C_f$ with a sample rate of 100 kHz........... 198

Figure 7.31: Axial control error noise with low pass filter $C_f$ implemented. The improvement at low frequency is approximately two orders of magnitude................................................................. 199

Figure 7.32: Change in current over two rotations of the spindle, at zero speed, when the axial controller regulates the shaft position to a constant value. The change in current is 0.03 A per power device channel.......................................................... 201

Figure 7.33: Comparison between the model predicting an axial motion error caused by a sensor disturbance, the model predicting an axial motion error caused by position dependent force
disturbance, and the experimental result. For this case, the difference between maximum and minimum currents was 0.03 A which corresponds to a static position disturbance of 3.6 μm. Speeds from 1000 RPM to 8000 RPM were tested.

Figure 7.34: Target screw metrology. a) In the ideal case, the sensing surface of the target screw is perfectly perpendicular to both the spindle rotation axis and the capacitive probe. b) In reality, there is a manufacturing tolerance on the perpendicularity. This limitation is the main source of axial error motion with a rotating spindle.

Figure 7.35: Target screw metrology geometry. \(L_b\) is the width of the bolt head, \(e_p\) is the perpendicularity tolerance, \(e_s\) is the offset of the capacitive probe from the spindle shaft, and \(x_d\) is the resulting artifact motion.

Figure 8.1: Prototype and production sized rotary-axial spindles. The full sized machine is approximately 1 m long and weighs about 500 kg. Its force output is more than 25 times greater than that of the prototype.

Figure 8.2: Rotary-axial spindle assembly section. This rotary-axial spindle contains four main assemblies: magnetic bearing assembly, DC motor assembly, hydrostatic bearing assembly, and spindle shaft assembly.

Figure 8.3: Middle assembly. The stator segments are thicker than the permanent magnets, requiring magnet spacers. Pockets are machined out of the spacers and filled with epoxy to prevent the grinding wheel from “loading up” when the assembly is precision ground. The finished assembly is 315 mm in diameter by 27 mm thick.

Figure 8.4: Cutaway section of rear magnetic bearing assembly. The main components are the coil stack, the frame, the inner stator segments, and the outer stator segments. All these components are epoxied into place. The front assembly is similarly designed and assembled.

Figure 8.5: Coils and stator assemblies. a) Each stack consists of 16 dually wound pancake coils, and there are two stacks: one for the front assembly, and one for the rear. Each coil has two layers of 39 turns for a total of 78. Using AWG 19 wire results in the coils being 203 mm in diameter and 2 mm thick. b) The assembled and ground front (right) and rear (left) magnetic bearing assemblies. The assemblies are 315 mm in diameter.

Figure 8.6: Hydrostatic bearing assembly. Water pressurized to at least 1.5 MPa is delivered to the bearing through the supply manifold. After passing through the two cast ceramic bushings, a suction port returns the water to a tank. Pressurized air, forces any air which makes it way passed the suction ports back. Rotary lip seals on either end of the assembly provide a final leak deterrent. A helical cooling jacket removes heat from the bearing housing, and the tank water is cooled, as well.

Figure 8.7: Moving assembly. Mounted to the single piece shaft, is a permanent magnet rotor which is held in place with a locknut. The powdered iron armature is installed and clamped through the flange of the precision ball target. The bump-stop locknut is used to prevent the armature from hitting the stator. Its diameter is 120 mm, its overall length is 865 mm, and its mass is 80 kg.
Figure 8.8: Natural frequencies as predicted by COSMOSWorks [30] finite element solver. The first and second bending modes are at 745 Hz, 1560 Hz, a pure axial mode is at 2890 Hz, an armature bending mode is at 3800 Hz, and there is a mode where the neck of the ball target bends at 5200 Hz.

Figure 8.9: New rotary-axial spindle metrology solution using precision ball target and brass bumper. The precision ball target is insensitive to rotational degrees of freedom. If the spindle shaft, ball, and capacitive probe can all be made coaxial, then the effects of shaft rotation on axial sensing will be eliminated. A brass bumper absorbs impacts from the flange of the ball target and the locknut, reducing shocks on the sensor clamp.

Figure 8.10: Ball centering setup. Two capacitive probes are clamped in an “L” bracket to measure the eccentricity. The spindle is rotated using the motor, then the eccentricity is observed, and the target is adjusted by tapping the flange with a hammer. Finally, the bolts are incrementally tightened in a star pattern with a torque wrench.

Figure 8.11: Amplifier power stage. Two voltage control boards each control the voltage output of four power devices. Each 12.7 mm thick copper heat sink has ten Zalman CPU fans [34] attached to it to dissipate 2 kW, total. The current control board spans the voltage stage PCBs. This stacked layout was designed by Kris Smeds.

Figure 8.12: PCB for eight channel current control. This PCB was designed by Kris Smeds.

Figure 8.13: Current control block diagram.

Figure 8.14: Three possible coil configurations. a) The pure parallel configuration. b) The mixed parallel-serial configuration. c) The pure serial configuration.

Figure 8.15: Experimentally measured lumped impedance. The inductance at low frequency is 326 mH and the parasitic capacitance becomes dominant near 7 kHz.

Figure 8.16: Decoupling results for eight channel amplifier.

Figure 8.17: Comparison between –negative of the loop transmission of diagonal entries (blue) and eigenvalues of full order matrix (black).

Figure 8.18: Closed loop response of average current to different common mode reference amplitudes. Voltage level saturation is evident even at small amplitudes.

Figure 8.19: 0.2 A step response. The individual currents do not converge as quickly as the average current response because the bandwidths of the differential controllers are only around 200 Hz. A fully saturated response is already evident at this step size.

Figure 8.20: Large signal step response. The oscillations are caused by the deep level of voltage saturation.
Figure 8.21: First quadrant force characteristic. The bearing negative stiffness and channel current to force ratio are approximately 3.1 N/μm and 1750 N/A, respectively. The 300 N hysteresis is due to lip seal friction.

Figure 8.22: At low frequency, the non-linear flexibility of the lip seals dominates. At 150 Hz, there is a resonance resulting from this stiffness and the shaft mass. From 200 Hz to 600 Hz, the response behaves like a free rigid mass. After this, there exist several natural frequencies, followed by a non-collocated mode at 2800 Hz. The pump supply pressure was 1.8 MPa.

Figure 8.23: Loop shaping controller design results. a) Loop shaping controller. b) Negative of the loop transmission. c) Dynamic stiffness. d) Closed loop frequency response.

Figure 8.24: Hammer impact test at the spindle nose. This result shows a minimum dynamic stiffness of 315 N/μm in the high frequency region.

Figure 8.25: Position regulation noise. a) The RMS error is 7 nm RMS. b) The power spectrum shows increased noise in the low frequency range, which is likely due to a lower integrator corner frequency and non-linearities introduced by the lip seals.

Figure 8.26: Stroke demonstration and tracking performance in air cutting. a) Eighty percent stroke sweep at a feed rate of 240 μm/s. b) The average tracking error is 10.5 nm and the noise about this point is 11.7 nm RMS. Feed forward compensation of the friction will improve this result.

Figure 8.27: Rotary control block diagram. The input to the system is a torque reference, $T_r$, and the output is the speed $\omega$ as measured by the sensorless algorithm.

Figure 8.28: Frequency response of rotary plant from torque command to speed. A rigid body mode corresponding to the spindle shaft inertia is dominant in the low frequency region. Above 30 Hz, the response indicates that the HES motion is being measured instead of that of the rotor.

Figure 8.29: Rotary motion closed loop block diagram.

Figure 8.30: Rotary control loop results. a) Negative loop transmission. b) Dynamic stiffness. c) Closed loop bandwidth.

Figure 8.31: Speed regulation at 3000 RPM. The RMS error is 0.293 RPM. The 115 Hz harmonic is related to the pressure fluctuations from the hydrostatic bearing pump.

Figure 8.32: Hydrostatic bearing losses caused by fluid shear. a) Average torque loss vs. spindle speed. b) Average power loss vs. spindle speed.

Figure 8.33: Axial error motion with ball target metrology system. The result indicates the error motion is below 40 nm at 2500 RPM. A filter was used to remove line and high frequency noise from the raw signal.
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Ad maiorem Dei gloriām
Chapter 1

Introduction

As new consumer and commercial products are continually demanding higher performance for lower costs, engineers often place their attention on the manufacturing process to achieve their design and fiscal goals. Precision manufacturing, specifically, is a hot field because it enables the fabrication of parts with sub-micron and even nanometer tolerances that were an obstacle to increased performance in the past. This is aiding the development of products in many industries which are used each day. In fact, precision machine tools are most frequently found in the factories of semiconductor and optical instrument manufacturers. Ultimately, there is a link between the performance of precision manufacturing equipment and the advancement of consumer goods.

In the design of precision machine tools, the trend is to increase the load capacity, stiffness, and precision to improve the productivity, disturbance rejection characteristics, and achievable tolerances. Generally, there are two approaches to achieving this end. The first focuses on advancing the underlying technologies of individual components such as bearings and actuators. However, at a certain point this method fails to yield further gains because the machine structure becomes the fundamental limitation. Consequently, the second and less conventional method attempts to increase performance through alternative machine architectures.

Since machine tools generally require at least two degrees of freedom (DOF), traditional machine architectures stack single DOF motion stages to meet the particular needs of a given manufacturing process. Herein lays the fundamental limitation. Each stacked component
constitutes a serial stiffness element in the structural loop. Therefore, the component with the smallest stiffness essentially sets the overall stiffness in a given direction. It follows that further improvements to other elements in the structural loop will not affect the machine stiffness.

To counter this, the rotary-axial spindle machine architecture was developed. A rotary-axial spindle is a machine tool whose shaft is radially supported by pressurized fluid bearings and is axially actuated by a high force active magnetic thrust bearing. Thus the shaft can be fed in the thrust direction with mm stroke, yet it is still capable of high speed rotary motion. Such a device can be used in many short stroke manufacturing operations such as diamond turning, meso-scale milling, and face grinding. Since there is only a single stiffness element in a given load path, this topology results in a very stiff structure enabling kHz range bandwidth, nm level axial motion precision, infinite static stiffness, axial dynamic stiffness on the order of hundreds of N/µm, and high axial load capacity.

1.1 Thesis overview

To address the needs next generation precision machining, this thesis presents the design, analysis, control, sensing, instrumentation and experimental testing of a rotary-axial spindle. The prototype, shown in Figure 1.1, has demonstrated 1 mm stroke, 600 N axial load capacity, 2.6 kHz closed loop bandwidth, 4 nm RMS positioning noise, and 100 N/µm minimum dynamic stiffness. It can reach speeds over 10,000 RPM. Additionally, the results of the development of a full scale rotary-axial spindle with hydrostatic journal bearings are presented. The performance is also excellent: 1.5 mm stroke, 6,000 N axial load capacity, 7 nm RMS positioning noise and 340 N/µm minimum dynamic stiffness.
The first chapter in the body of this thesis discusses machine tool bearings and architectures. Next, a chapter of this thesis will be dedicated to the magnetic bearing analysis and precision machine design. The rotary-axial spindle also requires a significant amount of supporting software, electronics, and control to achieve its extreme performance. Therefore, additional chapters in this thesis will be devoted to the magnetic bearing power amplifier, a sensorless rotary motion feedback algorithm, a novel approach to increasing analog to digital converter resolution, and rotary-axial spindle control and experimental testing. A final chapter
will give an overview of the development and testing of a full-size rotary-axial spindle. Figure 1.2 shows the whole rotary-axial spindle system.

**Figure 1.2:** The rotary-axial spindle system. The main components are the rotary-axial spindle, the linear power amplifier, the capacitive probe and its signal conditioning hardware, the sensor signal processing board, the rotary power amplifier, and the dSPACE control unit.

1.1.1 Machine tool bearings and architectures

To better understand the limits of machine tool bearings, the first sections of Chapter 2 survey rolling element, magnetic, and fluid technologies. However, it is often the case that the fundamental limitation with respect to achieving higher stiffness and precision is inherent in the machine structure. So, the conventional stacked stage concept is analyzed and its advantages and
disadvantages are presented. Finally, a new embodiment of the rotary-axial spindle architecture, which addresses the aforementioned limitations, is described and its benefits are discussed.

1.1.2 Rotary-axial spindle analysis, design and prototyping

The first rotary-axial spindle prototype was designed by Dr. Xiaodong Lu in the months preceding May 2007. However, it is still important to discuss this work as it will lay down some fundamental principles which will be useful for the remainder of this thesis, and provide motivation and context for the remaining work described. To that end, this chapter describes the magnetic bearing and mechanical structure. The thrust bearing, shown in Figure 1.3, is an axisymmetric version of the ultra fast tool servo developed by Lu [1, 2].

![Schematic of axial magnetic bearing](image)

**Figure 1.3:** Schematic of axial magnetic bearing [3]. Used with permission.
It has been scaled up to produce significantly more force yet retains its highly linear properties which allow for high bandwidth and exceptional dynamic stiffness. The fundamental principles are described, and the governing equations from [2:85-99] are reiterated. An electromagnetic finite element analysis (FEA) of the actuator is also performed to account for finite permeability, complex geometries and non-linear material properties. A structural analysis of the spindle shaft is performed in order to predict its dynamics. Finally, an overview of the assembly of the magnetic bearing and its integration with the aerostatic journal bearing is given.

1.1.3 Power electronics

In order to control the axial force applied to the spindle shaft, the currents in the magnetic bearing must be precisely controlled with very high bandwidth. Chapter 3 discusses the power amplifier which accomplishes this task. A multi-channel linear amplifier similar to that presented in [1, 2:151-196] is used and is shown in Figure 1.4. In these studies, however, decentralized controllers are used to control the current in each channel. This research investigates the extension of that theory with a multi-input multi-output (MIMO) centralized control scheme. It decouples the load and ensures that the currents remain equal. The current amplifier bandwidth is 50 kHz and the maximum power output is 1 kW. All the details regarding this subsystem are discussed in Chapter 3.
1.1.4 Rotary sensing and control

In order to cancel torque disturbances which are inevitable in machining operations, feedback control of the spindle speed is necessary. Generally, optical encoders are the rotary motion sensor of choice for precision spindle applications. However, they usually require a constant axial air gap between the disc and optical read head. They are unsuitable for rotary-axial spindles. To counter this, Chapter 5 explains how the three Hall Effect sensors (HES) used for commutating the phases of the brushless motor can be used to control the average speed of the spindle. The resulting speed regulation error is equivalent to that achieved with a 1000 line encoder. This result is shown in Figure 1.5. Furthermore, models of the open loop systems of both the Precitech spindle test bed and the rotary-axial spindle are presented and average speed filter design and rotary control are discussed.
### Figure 1.5: Comparison between interpolated (256x) with adapted HES derived speed feedback signal and optical encoder derived speed feedback (tested on conventional spindle).

#### 1.1.5 Axial sensing

The thrust actuator also requires feedback so that the spindle shaft can be precisely positioned along its 1 mm stroke. Using the 16 bit analog to digital converter on our dSPACE control unit this results in 15.3 nm resolution. However, in ultra-precision applications, it is desirable to have nanometer range resolution. To address this, Chapter 6 presents a patent-pending [4] method for increasing the resolution of analog to digital converters without increasing the associated conversion delay. In the current embodiment, it is called Nanozoom. Along with the design of a custom printed circuit board (Figure 1.6), the communication protocols, analytical model of the system, and experimental results are presented. With this technology, 1.9 nm resolution, which corresponds to 19 bits, has been demonstrated.
1.1.6 Axial control

This chapter discusses the axial control, which is the final component in the rotary-axial spindle system. Since the actuation and sensing points between axial and rotary directions are decoupled, a Laplace domain controller can be designed for the axial plant. Modeling, online system identification, and an iterative controller design process are presented. Also, saturations in the system and all other limitations are diagnosed and workarounds are presented where possible. The axial control loop has demonstrated 2.6 kHz bandwidth, infinite static stiffness, 95 N/μm minimum dynamic stiffness below 550 Hz (Figure 1.7), and sub 5 nm RMS position regulation noise.

Figure 1.6: Printed circuit board which implements Nanozoom.
1.1.7 Full scale rotary-axial spindle prototype

Having proven the rotary-axial spindle concept with the prototype, a new full scale prototype was designed, manufactured, instrumented and tested by teams from UBC, the Massachusetts Institute of Technology (MIT), and our sponsor. Shown in Figure 1.8, the machine tool uses a scaled up version of the homo-polar magnetic thrust bearing presented in this thesis along with hydrostatic journal bearings. The hydrostatic bearings provide higher stiffness and load capacity compared to aerostatic designs at the expense of increased heat generation from fluid shear. It produces more than 6 kN of continuous axial force, can be regulated to 7 nm RMS at any point along its 1.5 mm axial stroke, has demonstrated 340 N/μm minimum dynamic stiffness, 800 Hz closed loop bandwidth and can be spun to 3000 RPM. It is driven by a 2 kW linear amplifier with MIMO current control. A summary of the design process, manufacture, control, and experimental results collected to date is given in the final chapter.
Figure 1.8: Full scale rotary-axial. The total length is 1 m and the spindle shaft is 120 mm in diameter.
Chapter 2

Machine Tool Bearings and Architectures

Many machine tools require at least two degrees of freedom (DOF): one translational for feeding, and another rotational for cutting. Common machine tools which use this basic principle are lathes, milling machines, boring machines, and various types of grinders. Although these machines vary in numerous ways, their shared component is the spindle. The performance of the spindle can be characterized in terms of stiffness, load capacity, error motion, and speed. Because the spindle is the heart of the machine tool, it often becomes the focal point of performance enhancement in manufacturing equipment. Specifically, there are two areas which come under increased scrutiny. The first is the spindle’s supporting bearings. The other is the way in which the spindle is mounted in the machine. This chapter presents a literature review of machine tool bearing and architecture technology.

The first section of this chapter discusses the benefits and shortcomings of rolling element bearings, magnetic bearings, and various pressurized fluid bearings. Then, in section 2.2, the typical method of combining the spindle with a set of bearings to achieve cutting and feeding motion is explained. Section 2.3 describes magnetic thrust bearing spindles which are optimized for high acceleration applications and have improved stiffness characteristics compared to stacked stage machines. Section 2.4 presents the rotary-axial spindle architecture for use in
applications requiring mm range stroke, high load capacity and stiffness. Finally, the chapter is summarized in section 2.5.

2.1 Machine tool bearing technology

Bearings, the components which support and constrain the spindle motion, are very important to the machine tool when it comes to stiffness, load capacity and heat generation. Rolling element thrust and journal bearings are the most popular method of spindle constraint, however researchers in both industry and academia are now investigating the benefits of magnetic and pressurized fluid bearings as well.

![Exploded view of a typical rolling element bearing. Photo used with permission [5].](image)

**Figure 2.1:** Exploded view of a typical rolling element bearing. Photo used with permission [5].

Rolling element designs are the most popular type of bearing due to their low cost and high load capacity. Generally, the balls or rollers that act as the rolling elements sit between two surfaces. The relative motion between them allows the element to rotate with little friction. The high strength rolling elements, which are usually made of various grades of steel or ceramics,
can handle the large loads encountered in many machining operations. However, it can be prohibitively expensive to manufacture the rolling elements with sufficient tolerances to provide a consistent rolling contact. Thus, the inconsistencies cause noise and vibration that limit spindle error motions to several microns, even for high-end models. Though tolerable for many conventional manufacturing operations, nanometer level precision is generally unachievable with rolling element bearings.

Magnetic actuators offer another means of motion stage support. Widely used in positioning applications, they are inherently high precision devices and can readily achieve nm range precision because of their high resolution position feedback sensors. Magnetic bearings have commonly been applied in scanning applications such as photo-lithography wafer positioning stages [6], atomic measuring machines [7], electron-beam scanners [8], and nanomachining devices [9]. However, since achieving the highest possible precision is the main goal and the force output is low, the closed loop dynamic stiffness is sacrificed. Magnetic bearings have been applied in other areas of the machine tool as well. Weck et al. demonstrated that a machine tool stage can be magnetically levitated, and achieved several N/μm of radial dynamic stiffness on a 42 kg mass [10]. Denkena et al. presented a design with active magnetic guides for the frictionless feed drive of a spindle [11].

The principles of electromagnetic actuation have also been used to provide journal support. Although additional complexities such as sensors and power amplifiers are required, they allow friction-free motion and are maintenance free. These characteristics make them particularly suitable for high speed applications such as flywheel energy storage systems [12] and turbo machinery [13] where it is only important to maintain separation between stator and rotor. One benefit is that rotor imbalances can be compensated for as demonstrated by Tamisier
et al. [14]. Zhang and his group also developed a five axis active magnetic bearing spindle for electrical discharge machining [15]. In general, these journal bearings still need improvements in terms of load capacity and disturbance rejection before they can be widely used in industrial manufacturing.

Pressurized fluid bearings [16], on the other hand, offer many of the benefits of rolling element and magnetic bearings, namely low friction, high load capacity and stiffness. In fluid-dynamic bearings, the shaft motion sucks fluid into the bearing and around the shaft surface. One drawback of this is that the spindle might have higher friction at low speeds, and under startup and shutdown. This characteristic makes it unsuitable for use in a machine tool which may be constantly starting and stopping. Alternatively, fluid may be constantly supplied from an external pump. These are called fluid-static bearings (Figure 2.2). When air is used as the bearing medium (aerostatic), the motion is essentially frictionless. Speeds up to 200,000 RPM are possible [17] and stiffness between 10 and 100 N/μm is fairly typical [18]. Infinite static stiffness has been achieved using piezoelectric elements to compensate the spindle rotor [19], but the air film still limits the achievable stiffness. Air can be replaced by an incompressible fluid such as water (hydrostatic) or oil to realize stiffness on the order of several hundred N/μm [20]. These bearings exhibit an averaging effect. This means that a spindle shaft and journal bearing manufactured with micron level tolerances can achieve error motions on the order of tens of nanometers. The penalty paid for this boost in performance is that significant cooling is required to compensate for the heat generated by the fluid shear.
Figure 2.2: Schematic view of a fluid-static bearing. Pressurized fluid is fed to the bearing from an external source. The small gap between the shaft and journal allows pressure to build, thus constraining the shaft.

2.2 Typical machine tool configuration

Even though advances have been made in bearing technology, it is often the manner in which they are arranged which limits the overall performance. In fact, even with major breakthroughs in bearing technology, it is often the machine architecture, or layout, which limits the performance. The layout of a typical machine tool with rotary motion for cutting and axial motion for feeding is shown in Figure 2.3. In this stacked axis arrangement, the spindle has its own journal and finite stiffness thrust supports. The spindle is mounted to a Z axis slide with its own lateral support. Furthermore, the Z axis actuator has a finite Z direction stiffness associated with it.
**Figure 2.3:** Conventional machine tool configuration. Here, the spindle is mounted to a Z axis slide. In the radial direction, the journal bearing and Z axis slide support stiffnesses are duplicated.

The noticeable advantages of this stacked configuration are that the rotary and Z direction actuation and sensing are decoupled. Additionally, fairly long strokes are easily attained. The fundamental limitation, however, lies in the duplication of radial and thrust supports. As seen from Figure 2.3, in the radial load path the stiffnesses of the spindle journal bearing and Z axis slide lateral support add in series. The same characteristic is exhibited with the spindle thrust bearing stiffness and Z axis actuator stiffness. The end result is that flexibility (reciprocal of stiffness) in each direction is additive, and so the overall stiffness is decreased. Additionally, a reaction moment is produced from the misaligned actuator feed and axial cutting force reducing the precision further. Moreover, when aerostatic or hydrostatic technologies are implemented as thrust bearings, the fluid shear generates heat that complicates thermal control and reduces
precision in the Z direction. Acceleration in the Z direction is also reduced since the actuator puts both the slide and spindle into motion. The long C-shaped structural loop is very susceptible to thermal errors. Another adverse effect resulting from the increased structural compliance is a reduction in the achievable closed loop Z axis bandwidth. In general, the bandwidth is tied to the control loop’s disturbance rejection capability. Thus, precision during a cutting operation becomes compromised.

2.3 Magnetic thrust bearing spindles

At least two designs [21, 22] have attempted to address some of these disadvantages by eliminating the stacked axes and constructing a single moving mass with two or more degrees of freedom (DOF). This section discusses both the aspects of these spindles which are positive and those which make them unsuitable for precision manufacturing.

Shinshi’s spindle [21] was designed for use in micromachining applications. It used a voice coil motor to provide 10 mm of axial stroke, thus eliminating the need for a Z slide. The actuation coils are mounted in the machine frame, while a permanent magnet is fixed to the spindle shaft. The axial position is measured at a mirror target from the back of the spindle with a laser interferometer. The mass of the moving assembly was 0.26 kg to minimize the load on the actuator. The actuator develops approximately 10 N of force and the position loop has 174 Hz bandwidth. In terms of rotary motion, the combination of contact free support from two aerostatic journal bearings and the air turbine allows for high speed machining. Removing the Z slide has the added benefit of increasing the acceleration capability in that direction. More importantly, the serial duplication of thrust and radial support are gone. For this case where machining forces are low, this is a good concept.
However, many changes would be needed to adapt this device to a typical machine tool whose primary requirements are high load capacity and stiffness. Firstly, Lorentz force motors have relatively low force densities. This limits the achievable feed rate and hampers the controller’s ability to attack high frequency disturbances due to acceleration limitations. To compensate for this, they require dense windings to achieve meaningful force output. So, when current is driven through the winding, heat is dissipated in the windings causing thermal growth and compromising precision. The next concept to look at is the actuation and sensing points (at the permanent magnet and at the shaft end, respectively) in the Z direction. Since the structural loop is lengthy, the likelihood of non-collocated resonances occurring at low frequencies is increased. Non-collocated resonances limit bandwidth and stiffness due to the addition of 180° phase lag to the open loop frequency response. Liebman’s rotary-linear axis [22], which is quite similar to Shinshi’s in architecture, using air bearings and a moving magnet linear motor, suffered from a similar problem. Shaft bending modes in the hundreds of Hz range limited the control bandwidth, and so the dynamic stiffness for grinding operations was insufficient [22:271]. So, there still exists a need for a magnetic bearing spindle with significantly higher force and stiffness.

2.4 Rotary-axial spindle architecture

To address this need, the rotary-axial spindle architecture is presented. This is a fundamental change in precision spindle technology. As illustrated in Figure 2.4, the spindle is radially supported by an externally pressurized fluid journal bearing and axial feed motion is provided by a high force output magnetic thrust bearing with millimeter range actuation. The result is that high speed cutting motion and feed motion are achieved with a single inertial
element. Thus, the requisite two degrees of freedom for grinding, turning, and other meso-scale manufacturing operations are achieved. Simplifying the structure has several important benefits. The first and most obvious is that the redundant bearings are removed. This has a snow ball effect creating a shorter structural loop, a substantial gain in structural stiffness, increased control bandwidth, and increased precision. Since the spindle frame is no longer part of the inertial payload, higher accelerations can also be reached. In particular, the removal of the typically aerostatic/hydrostatic thrust bearing, which usually has a large thrust plate, eliminates heat generated by the fluid shear. Finally, machine production costs are reduced through smaller component counts.

Figure 2.4: Rotary-axial spindle architecture. In this topology the redundant thrust and radial supports are eliminated. The key benefits of this are an overall increase in stiffness, bandwidth and precision, and reductions in structural bending moments, heat generation, component count and cost.
2.5 Summary

This chapter presented a review of machine tool bearings and structures. A survey of machine tool bearings was provided and it was stated that fluid bearings give the best combination of load capacity and stiffness for journal bearings. Their one drawback is the heat generation caused by fluid shear. Next, a conventional stacked stage architecture was described. The fundamental limitation to improved stiffness and precision was identified as the serial duplication of supports. A different machine architecture, which has only a single moving mass, was used in magnetic thrust bearing spindles. These spindles eliminate the serial duplication of supports, but lack sufficient load capacity and dynamic stiffness because they have been optimized for applications requiring higher acceleration. Finally, a rotary-axial spindle machine architecture was proposed. Unlike the magnetic thrust bearing spindles, it employs a high force normal stress magnetic actuator to provide the axial motion. Since the structural loop is short, kHz range bandwidth and several hundred N/μm dynamic stiffness are achievable.
Chapter 3

Rotary-axial Spindle Analysis, Design, and Prototyping

The previous chapter explained the rationale for introducing the rotary-axial spindle architecture. To verify the physical intuition upon which its foundation lies, a prototype rotary-axial spindle has been constructed. The qualitative requirements for this design are very high axial stiffness, high static force, reasonable acceleration levels, and high linearity. The combination of these properties with the rotary-axial spindle architecture should enable kHz range bandwidth, and hundreds of N/μm dynamic stiffness while maintaining nanometer level precision.

In this chapter the analysis, design, and fabrication of the rotary-axial spindle prototype is presented. Section 3.1 explains the operating principles of the magnetic thrust bearing and calculate the force output using a previously presented technique [2:85-99]. This is useful for understanding which parameters have the greatest effect on force output. Additionally, a finite element analysis that simplifies the consideration of effects such as flux leakage and magnetic core saturation is conducted. Section 3.3 describes the rotary-axial spindle’s various assemblies as designed by Dr. Lu and assembled by Irfan Usman. (N.B.: the design was completed prior to May 2007). This step is necessary because it will help the reader understand the machine layout and structure. Section 3.4 presents the results of a structural finite element analysis which is used to predict the vibration modes of the moving assembly. This provides some intuition of the weak
parts of the structure and will help identify the experimentally observed resonances. Section 3.5 shows the fabricated components and how they fit together to form the rotary-axial spindle. A summary is given in section 3.6.

3.1 Electromagnetic thrust actuator/bearing

As mentioned in the introduction, an axis symmetric version of the actuator used in [1] was chosen to provide the feed motion. With the fast tool servo, the design goal was to maximize acceleration for exceptional high frequency performance. However, in this case, the goal is to maximize load capacity. The following subsections describe in more detail the properties of the magnetic bearing, its principles of operation, and the rationale for making this decision. Additionally, the actuator force derivation originally presented by Lu [2:85-96] is reiterated before progressing through a finite element analysis which will account for magnetic saturation and leakage fluxes.

3.1.1 Magnetic bearing selection

Several magnetic bearing technologies are available for use in this new spindle architecture. Lorentz force and solenoid designs are very common in tool servo applications [23, 24] and have also been put to use in magnetic bearing spindles [14, 15, 21, 22]. Lorentz force actuators allow for the possibility of centimeter range strokes and have linear current to force transfer functions, but they cannot produce high forces due to low flux density in the air gap. On the other hand, solenoids (or horseshoe type actuators) can produce very large forces since the flux acts normal to the armature surface. However, the forces generated in this scheme are proportional to current squared and inversely proportional to armature position squared and are
therefore very inefficient at large air gaps [25]. Power inefficient current biasing coils can be used, but these increase the actuator volume. Additionally, in many designs only about half the armature pole area generates force, thus reducing the spatial compactness and increasing the inertial load and the likelihood of armature vibration modes occurring.

The magnetic bearing chosen for the rotary-axial spindle combines the best of both technologies, namely high linearity and high force density. The design is based on Lu’s novel fast tool servo topology [1], although, an axisymmetric form is used. It is shown in Figure 3.1. The four main parts of this bearing are the coils, the permanent magnet ring, and the stator and the armature which are soft magnetic materials. The coils are driven in parallel and produce an excitation flux (red line) which flows through the stator, armature, and axial air gaps only. A constant bias flux (blue lines) is produced by the radially magnetized permanent magnet ring. This flux follows a path which takes it through the radial air gap, into the armature, then splitting through each axial air gap to return to the other permanent magnet pole through the stator. The armature is connected to the shaft so that the electromagnetic force can put the shaft into motion.

The actuator is a reluctance device, like a solenoid. One similarity it shares with solenoid designs is that the flux density is highly concentrated on the armature leading to high force output. One difference is that, in this case, the total air gap is constant, leading to better linearity. Another difference is the inclusion of the permanent magnet ring which not only linearizes the actuator, but also produces additional force.
Figure 3.1: Cross section view of rotary-axial spindle’s homo-polar normal stress magnetic bearing. A section cut through the magnet ring and armature is also shown.
3.2 Electromagnetic bearing force analysis

In this section, the axial force produced by the magnetic bearing is calculated using two methods. The first is via an approach presented by Dr. Lu in his Ph.D. thesis [2:85-99], and the second is through a finite element analysis which accounts for the effects of finite permeability and leakage fluxes.

3.2.1 Force calculation with ideal material properties

There are several steps to calculating the force produced by this type of magnetic bearing. Since there are two flux sources, the coils (alternating flux) and the permanent magnet ring (constant flux), it is best to start by looking at the contribution of each to the total flux using the Maxwell equations. Then, after applying the principle of superposition, a reluctance model of the actuator is created. From there, the force can be calculated using Maxwell stress tensors, or the principles of energy and coenergy. This analysis assumes that the stator and armature are made of a soft magnetic material with permeability much higher than that of air.

To start, the coils and their contributions are eliminated from the analysis, and the flux contribution from the permanent magnet ring is calculated. This situation is shown in Figure 3.2.
The permanent magnet produces a DC flux $\Phi_{PM}$, but leakage reduces this to a value of $\Phi$.

Applying Gauss’ law to the surface $S$ in yields:

$$
\left(\overline{B_1} + \overline{B_2}\right)A - \Phi = 0 \tag{3.1}
$$

Here, $\overline{B_1}$ and $\overline{B_2}$ denote the bias fluxes in right and left hand air gaps respectively and $A$ is the stator pole area. Then, when Ampere’s Law is applied to the contour $C$ and infinite permeability in the stator and armature is assumed, the result is:

$$
-\overline{B_1}(X_o - X) + \overline{B_1}(X_o + X) = 0 \tag{3.2}
$$

where $X_o$ is the total air gap in the axial direction and $X$ is the axial position of the armature in the air gap. When the armature is centered in the air gap, $X$ is zero. The bias flux in each air gap is calculated by combining (3.1) and (3.2).

$$
\overline{B_1} = \frac{\Phi}{2A} \left( \frac{X_o + X}{X_o} \right) \tag{3.3}
$$
\[
\overline{B}_2 = \frac{\Phi}{2A} \left( \frac{X_o - X}{X_o} \right) \tag{3.4}
\]

Also, the following substitution can be made to simplify things:

\[
\overline{B} = \frac{\Phi}{2A} \tag{3.5}
\]

\[
\overline{B}_1 = B \left( \frac{X_o + X}{X_o} \right) \tag{3.6}
\]

\[
\overline{B}_2 = B \left( \frac{X_o - X}{X_o} \right) \tag{3.7}
\]

Pausing to look at equations (3.2), (3.6) it is seen that although the flux in each air gap varies significantly with armature position, the total bias flux remains constant. This is the second of four key features which leads to a linear design.

The next step is to determine the flux leakage from the permanent magnet. Since it is very difficult to model flux linkages, a finite element solver is used to measure the amount of bias flux actually leaving the stator. The program FEMM [26] is used to compute the result. The line, or contour, highlighted in red represents the face of the stator. By integrating along this contour and multiplying by two, a value for \( \Phi \) of 0.002456 Wb is obtained. The multiplication by two accounts for both stator faces. This parameter will be used in the final force calculation.
Figure 3.3: Finite element result showing leakage fluxes from stator with centered armature.

Now, the focus shifts to the flux produced by the coils. In a likewise manner, the permanent magnet ring is momentarily eliminated from the analysis as shown in Figure 3.4.

Figure 3.4: Diagram used for calculating flux and force contributed by excitation currents. The red line indicates the path and positive direction of AC flux flow.
Applying Gauss’ law again to the surface S:

\[ \vec{B}_1 - \vec{B}_2 = 0 \]  

(3.8)

where \( \vec{B}_1 \) and \( \vec{B}_2 \) are the fluxes in the air gaps to the left and right side of the armature respectively. Since they are equal, they can both be denoted as \( \vec{B} \). Now, Ampere’s law is applied to the contour \( C \) in Figure 3.4 resulting in:

\[ \frac{\vec{B}}{\mu_0} (X_o + X) + \frac{\vec{B}}{\mu_0} (X_o - X) = NI_{coil} \]  

(3.9)

where \( I_{coil} \) is the current driven through each coil (the coils are driven in parallel), \( N \) is the total number of turns (both coils) and \( \mu_0 \) is the permeability of free space. Rearranging (3.9) and solving for \( \vec{B} \):

\[ \vec{B} = \frac{NI\mu_0}{2X_o} \]  

(3.10)

The result in equation (3.10) is an expression for the excitation flux density in each air gap. Having derived expressions for both the constant and excitation flux density in each air gap, the two can be summed together using the principle of superposition.

\[ B_1 = \vec{B} + \vec{B}_1 \text{ and } B_2 = \vec{B} - \vec{B}_2 \]  

(3.11) and (3.12)

Now, it can be seen that with the addition of the excitation flux to the bias flux in each air gap, the total flux becomes differential. That is, the excitation and bias fluxes sum in one air gap and subtract in the other. This is the third feature which leads to an actuator with linear characteristics. Now that expressions for magnetic flux densities have been derived for each air gap, the next step is to derive the force. Here, the energy principle is used. It requires an equation summing the system’s total stored energy. Reluctances are the only elements which store energy in this system. Since the stator has been assumed to have infinite permeability, the only
reluctances are the permanent magnet, and the three air gaps (two axial, and one radial). A common expression for the energy stored in a reluctance element is:

\[ W = \frac{1}{2} R \Phi^2 \]  

(3.13)

where \( R \) is the reluctance and \( \Phi \) is the flux flowing through that specific reluctance. So, a simplified reluctance model of the system can now be drawn up as shown in Figure 3.5.

**Figure 3.5:** Simplified reluctance model of magnetic thrust bearing. Each sub-circuit contains a flux source along with the reluctance through which it flows. a) Permanent magnet flux and permanent magnet reluctance. b) Flux in air gap to the right of the armature and right air gap reluctance. c) Flux in air gap to the left of the armature and left air gap reluctance. Reluctance \( R_1 \) and \( R_2 \) are variable.

Lu presents a more complete schematic, but this simplified schematic suffices because all major air gaps and the associated fluxes are included. Each of the three sub-circuits contains a flux source and a reluctance element. All three flux sources have already been calculated and it is not necessary to calculate its reluctance. The radial air gap has been neglected because it contributes nothing to the axial force. But now, expressions for the reluctances of the two axial air gaps need formulating. So, it is best to start with the fundamental reluctance equation [27:33]:

31
In this expression, \( l \) is the length of the path though which the flux flows, \( \mu_r \) is the relative permeability of the medium, and \( A \) is the corresponding cross sectional area of the path. The relative permeability of air is unity. The length of the flux path in the right air gap is the difference between the nominal gap and the armature position. The length in the left air gap is the sum of the nominal gap and the armature position. Making the appropriate substitutions:

\[
R_1 = \frac{X_o - X}{\mu_0 A} \quad \text{and} \quad R_2 = \frac{X_o + X}{\mu_0 A}
\]  
(3.15) and (3.16)

Now, (3.15) and (3.16) and the permanent magnet flux and reluctance can be substituted into (3.13) leaving:

\[
W = \frac{1}{2} R_1 (B_1 A)^2 + \frac{1}{2} R_2 (B_2 A)^2 + \frac{1}{2} R_{PM} \Phi_{PM}^2
\]
(3.17)

By substituting the reluctances from (3.15) and (3.16) into (3.17) it can be noticed that the energy equation is a function of the flux sources \((B_1A, B_2A \text{ and } \Phi_{PM})\) and of the armature position:

\[
W = \frac{X_o - X}{2\mu_0 A} (B_1 A)^2 + \frac{X_o + X}{2\mu_0 A} (B_2 A)^2 + \frac{1}{2} R_{PM} \Phi_{PM}^2
\]
(3.18)

[27:61] also gives a general expression calculating the force from the energy storage equation and if it is modified to suit this case:

\[
F = -\frac{\partial}{\partial X} W(B_1, B_2, \Phi_{PM}, X)|_{B_1, B_2, \Phi_{PM}}
\]
(3.19)

Care must be taken when performing this derivative to keep \( B_1 \) and \( B_2 \) constant even though they are functions of \( X \). But, if this derivative is calculated with this in mind then the end result is:
In (3.20) it is apparent that the two normal stresses (on the right side of the armature and on the left side) oppose each other. This is the fourth and final feature which makes this a hard linearized actuator. Now, expanding (3.20) gives:

\[ F = \frac{A}{2\mu_o} \left( B_1^2 - B_2^2 \right) \]  

(3.20)

This shows that the force output from the actuator is, indeed, linear with respect to armature position and coil current. Generally speaking, the force increases linearly with stator pole area, permanent magnet bias flux, coil turns, and a decrease in nominal air gap.

3.2.2 Finite element analysis

Though the previous analysis proves the linearity and shows the parameter dependencies of the magnetic bearing, the assumption of infinite stator and armature permeability is not very practical. Typically, soft magnetic materials will begin entering a saturated state (when many of the magnetic domains begin aligning) once the flux density in the stator reaches approximately 1 T. The most efficient way to deal with this is to use a finite element solver.

This analysis uses FEMM. It is a freeware finite element tool that can be used for solving magnetostatic and electrostatic problems. It was chosen because it allows for easy definition of geometry and material properties. The software provides a scripting tool for quickly running many different scenarios. Moreover, the actuator’s geometry could be taken advantage of with axisymmetric mode. To conduct a finite element analysis with this software, the actuator geometry must first be defined. Next, material and coil properties are assigned. After generating a mesh, studies can be run and their results can be post-processed.
Figure 3.6 shows the geometry of the finite element model. Here, only half the actuator is modeled, taking advantage of the actuator symmetry. The section of the stator shown is approximately 72 mm wide by 31 mm tall and its outer radius is 45 mm. The armature is 25 mm in diameter by 12.7 mm wide. The coil pockets are approximately 20 mm wide by 16 mm tall. The width of the magnet segment is 13.9 mm, making the total axial air gap 1.2 mm. A boundary condition must be provided, as well, so a sphere (seen as a semicircle in axisymmetric 2D) of 125 mm diameter is drawn. These dimensions were determined by Dr. Lu during the initial design phase.

**Figure 3.6:** Finite element model showing the actuator and boundary condition.

Having defined the geometry, the material properties are assigned for the stator, armature, magnet ring, and coils. A pressed powdered iron, a soft magnetic material, bonded with polymers was chosen as the material for the stator and armature. Since this pressed
powdered iron has very low conductivity, eddy current generation in the moving assembly is greatly reduced. This is advantageous since there is no precision reducing Joule heating induced by eddy current flow. The BH curve, which describes the magnetic flux density in a material in the presence of an externally applied magnetic field, for this pressed powdered iron is shown in Figure 3.7. The manufacturer, Höganäs, reports an initial relative permeability (the slope in the linear region close to the origin) of 500 [28]. Additionally, the curve shows a pronounced saturation effect above 1.0 T.

![Figure 3.7: BH curve for pressed powdered iron with Kenolube® bonding agent used as armature and stator materials. Höganas reports an initial permeability of 500 [28].](image-url)
Figure 3.8: Hysteresis curve for a typical permanent magnet. The remanence is the flux density remaining in the magnet after the external field is removed. The coercivity is the strength of the external field required to demagnetize the magnet.

Next, the properties of the permanent magnet are defined. Since the relative permeability of the magnet is close to unity, only the magnet coercivity, $H_c$ needs to be specified. Coercivity is measure of the external field required to completely demagnetize a magnet. This is demonstrated on the BH curve for a typical permanent magnet shown in Figure 3.8. A short computation is required to obtain this parameter. Using the governing equation for the magnetic field in a permanent magnet [2:313]:

$$\vec{B} = \mu_0 (\vec{M} + \vec{H})$$

Another permanent magnet property that appears in this figure is the permanent magnet remanence. This is the flux density remaining in the magnet when the external field is removed. Using the remanence:

$$B_r = \mu_0 (M + 0)$$

$$M = \frac{B_r}{\mu_0}$$
Knowing that the flux density is zero when the applied field is equal to the coercivity, and reusing (3.23) gives:

\[ 0 = \mu_0(M + H_c) \]  \hspace{1cm} (3.25)

Substituting the result from (3.24):

\[ H_c = -\frac{B_r}{\mu_0} \]  \hspace{1cm} (3.26)

Since N44SH grade magnets are employed, \( B_r \) is 1.3 T, making \( H_c = -1,034,507 \) A/m. The last set of properties belongs to the coils which are made from AWG 19 solid copper. Each set of coils has 224 turns. For the purposes of this analysis, the current in each coil can be considered equal.

The final step before running the analysis is to generate a mesh for the model. It is important to find a balance between speed and accuracy. So, a fine mesh should be used in areas such as the air gaps around the armature, in the armature itself, in the stator, and permanent magnet ring. In order to capture the effects of leakage fluxes around the permanent magnet, a fine mesh in the coil pockets is also used. A coarse mesh in the air surrounding the actuator, where the flux density is orders of magnitude lower than in the areas of interest, will help speed the simulations. The final mesh is shown in Figure 3.9.
Now that the geometry, materials properties, and mesh have been defined for this analysis, the simulation can be run. With everything being ideal, the bearing generates no force in the radial direction. All focus is placed on axial force production. Specifically, it is necessary to look at the force at different armature positions along the stroke and at different coil currents. So, using a Lua script, which allows several different studies to be run consecutively, the armature is set at various positions starting in the center of the air gap, and moving in 40 µm increments to 480 µm away from center. Then, at each position the coil currents are set to various currents from -1.25 A to 1.25 A in 0.25 A increments. In total, there are 143 tests. At each position/current test point the solver reports the net axial force on the armature as calculated by Maxwell stress tensors. The simulation results are plotted in Figure 3.10 and Figure 3.11.
Figure 3.10: Net force on armature vs. coil current. The force vs. current plot is shown at different axial positions. The data at the bottom of the plot corresponds to the force calculated with the armature in the center position. The data at the top of the plot corresponds to the force calculated with the armature 480 µm away from center.

Table 3.1: Current to force ratio at increasing axial armature position.

<table>
<thead>
<tr>
<th>X [µm]</th>
<th>k_i [N/A]</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>376</td>
</tr>
<tr>
<td>40</td>
<td>376</td>
</tr>
<tr>
<td>80</td>
<td>374</td>
</tr>
<tr>
<td>120</td>
<td>372</td>
</tr>
<tr>
<td>160</td>
<td>369</td>
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<td>200</td>
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<td>343</td>
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<tr>
<td>400</td>
<td>337</td>
</tr>
<tr>
<td>440</td>
<td>330</td>
</tr>
<tr>
<td>480</td>
<td>323</td>
</tr>
</tbody>
</table>
A first glance at the results shows evidence of reasonable linearity. Looking a little deeper, the linearity is very good near the armature center position and at lower current levels. This makes sense. When the actuator is in the center, the bias flux density is evenly distributed through each half of the stator thus avoiding magnetic saturation. Also, if the coil currents are
low, the applied field is also small, thus introducing only small additions to the bias flux. Conversely, as the actuator position is moved closer to one extreme, a larger portion of the bias flux is directed to the respective half of the stator. Consequently, portions of the powdered iron core reach saturation even without the application of currents. Still, it seems logical to use the least squares method to fit linear equations to each set of data. The slope of each line is the important factor because then the actuator force can be expressed as a function of $X$ and $I$, similar to (3.21):

$$F(X, I) = k_x X + k_i I_{coil}$$

(3.27)

The position and current to force ratios are $k_x$ and $k_i$ respectively. These slopes are presented in Table 3.1 and Table 3.2. The data in these tables are also consistent with the preceding thought process. The current to force ratios shown in Table 3.1 are quite similar for approximately the middle 50% of the actuator stroke, but begins to fall off once the bias flux begins saturating the stator. Similarly, the position to force ratios are fairly close for small excitation currents but diverge once the current increases. However, the maximum non-linearity is approximately 14% for both parameters. For the purposes of modeling, the values of $k_x$ and $k_i$ at the center point and zero current points can be used: 1.54 N/µm and 376 N/A.

3.3 Rotary-axial spindle design

This section presents the rotary-axial spindle design. The machine has four main assemblies: the magnetic thrust bearing assembly, the aerostatic journal bearing assembly, the shaft assembly, and the position sensor assembly. The following sections describe each assembly, some of the key materials, and the main components.
3.3.1 Magnetic bearing assembly and brushless motor

An exploded view of the magnetic bearing as described in the previous sections is shown in Figure 3.13. It can be further broken down into three subassemblies: front, middle, and rear. The front and rear assemblies are comprised of four powdered iron segments, a set of coils which sit in the stator slot. These pieces fit into a nonmagnetic stainless steel frame. Powdered iron, a soft magnetic material, is chosen for the stator to minimize eddy current effects. As its name suggests, it is produced by applying pressure to a mixture of iron powder and polymer-based bonding agent. It is necessary to have the stator segmented because it is difficult to obtain powdered iron blanks large enough to produce a whole front or rear stator. A 20 mm x 16 mm pocket is machined out of the stator segments to make room for coils which are made from AWG 19 solid copper.

The second crucial part of the magnetic bearing, the magnet ring, lies in the middle of the assembly. Due to difficulties in radially magnetizing a ring, it is constructed from twelve
segments of Neodymium-Iron-Boron (NdFeB) rare earth magnet. Each segment is linearly, not radially (see Figure 3.14) magnetized to 1.3 T remanence. The magnets are surrounded by a thin segment of powdered iron which provides a connection in the magnetic circuit between the two halves of the stator and the outer diameter of the permanent magnet. The brushless motor can apply a maximum of 4.8 Nm of torque and can spin to 15,000 RPM. The motor stator also has three Hall Effect sensors which are used for commutating the motor phases.

**Figure 3.13:** Sectioned and exploded view of magnetic bearing assembly. It can be further broken down into the front, middle, and rear assemblies. The front and rear assemblies each consist of a frame, a set of coils, and a four piece powdered iron stator. The middle assembly frame holds the middle stator portion as well as the twelve magnet segments. The brushless motor stator fits into the front frame.
3.3.2 Aerostatic journal bearing assembly

The rotary-axial spindle architecture requires a pressurized fluid bearing to provide radial support. So, the bronze journal from a Precitech SP75 work holding spindle is used. A solid model of this bearing is shown in Figure 3.15. The outer diameter of the spindle shaft is 44 mm. When supplied with dry, filtered air, pressurized to 690 kPa (100 psi), the bearing has a radial load capacity of 180 N and a stiffness of 22 N/µm [18]. It can support shaft speeds up to 15,000 RPM.
Figure 3.15: Aerostatic journal bearing housing. The journal bearing, taken from a Precitech SP75 spindle, is made of bronze. The internal diameter is approximately 44 mm.

Figure 3.16: Shaft assembly. Axial force is applied to the armature by the magnetic bearing. The axial position of the assembly is sensed at the target screw. Torque applied to the permanent magnet rotor by the brushless motor stator provides the rotary motion.
3.3.3 Shaft assembly

The shaft assembly, shown in Figure 3.16, has thirteen components. The magnetic stainless steel spindle shaft was originally fitted to the Precitech spindle. The gap between this shaft and the bronze journal is 10 µm. The remaining pieces mount to the extension shaft which is fastened with six bolts to the spindle shaft. An 8 pole permanent magnet rotor, again taken from the Precitech spindle, is pressed onto the extension shaft. Torque is applied to the rotor by the brushless motor stator to provide the cutting motion. A shaft spacer with a ¾”-16 ID thread twists onto the extension shaft and locks the motor rotor in place. Next, the magnetic bearing armature slides over the extension shaft with a slight clearance fit and butts up to the shaft spacer. It is made of the same pressed powdered iron as described in earlier sections. The electromagnetic force acts on the armature to provide the feed motion. The armature is secured with a locknut. Finally, a 5/16” hex head screw, the head of which has been ground, acts as a target surface for axial position sensing. The extension shaft, shaft spacer, locknut, and target screw are all made of non-magnetic stainless steel so as not to disrupt the path of the fluxes generated by the permanent magnets and coils. The target screw and armature are located close together so as to keep the structural loop short. This minimizes the chance of a non-collocated resonance occurring at lower frequencies.

3.3.4 Position sensor assembly

The position sensor assembly consists of the probe holder, a capacitive probe sensor, and two screws used for tightening the holder. The probe holder, shown in Figure 3.17a, has a finely reamed hole through which the capacitive probe passes. Then, slots are cut in the probe holder so that when the locking screws are tightened the holder is deformed thus applying a clamping force along the body of the capacitive probe. In comparison to setscrew based clamping methods
which have low stiffness and loosen easily, this method gives excellent clamping stiffness and force. The capacitive probe, shown in the probe holder in Figure 3.17b, is an ADE 6503 with 1 mm range, 20 kHz bandwidth, and 5.5 nm RMS noise. Ideally, the sensing target (the shaft assembly) should have a good electrical connection to the probe casing. In the present case, however, this is not possible as the shaft is levitated by the magnetic bearing and aerostatic journal bearing. According to the capacitive probe user manual, the capacitance generated between the spindle shaft and journal bearing surfaces should be sufficient [29]. The assembly bolts into the back of the rear magnetic bearing assembly so that the probe can sense the position of the target screw.

![Figure 3.17: Position sensing assembly. a) Probe holder. Two slots are cut to allow for flexing of the clamp once the bolts are tightened. b) Capacitive probe installed in holder.](image)

3.4 Structural analysis

Now that the moving assembly has been designed, a simple analysis can be performed to predict the vibration modes of the structure. To do this, the frequency analysis option in COSMOSWorks® of SolidWorks® [30] is used. It approximates the natural frequencies and enables the user to visualize the mode shapes at these frequencies. Meshing is performed
automatically. Since the solid model has already been generated, it only needs to be simplified in order to speed the progress of the solver. After this, the boundary conditions can be set, the solver can be run, and the results can be analyzed.

3.4.1 Model simplification and boundary conditions

The first step in setting up a frequency analysis is to generate a good representative model. A model exactly matching the structure can be used. However, this extends the computation time because a finer mesh is required to model certain features. A stripped down model with chamfers, small holes, and other fine details removed, significantly reduces the number of elements in the model. This in turn shortens the computation time without losing much accuracy. Replacing socket head fastener models with items neglecting hex sockets also makes a difference.

Another important point is to ensure that the parts in the assembly are mated properly. In the present model, the key point is at the bolted joint between the extension and spindle shafts. Specifically, the stiffness of the joint needs to be captured. In the bolted joint, the fasteners provide the preload, but it is the clamped members which provide the dominant stiffness. To capture this, the mating faces of the spindle shaft and extension shaft are made coincident.

Next, the boundary conditions are set. In this case, the only boundary condition to set is related to the aerostatic bearing. Knowing that the static stiffness of the bearing is 22 N/µm, a spring of this stiffness can be distributed across the spindle shaft in the radial direction. The spindle is left free in the axial direction.
3.4.2 Structural analysis and results

The finite element analysis is completed in only a few minutes, and the qualitative mode shapes and approximate natural frequencies up to 10 kHz are shown in Figure 3.18.

![Figure 3.18: Dominant vibration modes of spindle shaft up to 10 kHz. a) First bending mode. b) Combined axial-torsional mode. c) Second bending mode. d) Pure axial mode. e) Third bending mode.](image)

A first look at each of the modes in Figure 3.18 seems to show that the weak point of the structure is the spindle and extension shaft joint. At higher frequencies, the slender neck of the extension shaft becomes flexible. The first mode is a simple bending mode. Next, at 4200 Hz, a coupled torsional-axial mode occurs. The next mode is the 2<sup>nd</sup> bending mode at 6400 Hz. The structure exhibits a pure axial mode near 8700 Hz. A 3<sup>rd</sup> bending mode occurs at 9840 Hz. If the components are well-aligned, and all threads are tightened well, then it is likely that only the pure axial modes will be seen in the frequency response of the structure. In either case, the limitations of the designed structure are now apparent.
3.5 Assembly

This section briefly describes the assembly procedure for the rotary-axial spindle. The components designed at UBC are shown in Figure 3.19. The coils are dually wound “pancake” coils of 28 turns each. There are sixteen coils in total, connected in a parallel-serial configuration as shown in Figure 3.20. This results in four circuits and takes advantage of the four channel power amplifier described in the next chapter. Each circuit has four “pancakes.” Two pancakes connected in parallel are joined in series to another pair of pancakes which are also connected in parallel. This forms one circuit. The relationship between the total current in each circuit, or channel, and the coil current is:

\[ I_{CH} = 2I_{coil} \]  

(3.28)

This is also demonstrated on channel A in Figure 3.20.

Figure 3.19: Rotary-axial spindle components designed at UBC.
Figure 3.20: Coil connection schematic. All the coils are configured into four circuits. Each circuit consists of four coils: one pair of coils connected in parallel, and then connected in series with another pair. The channel current is equal to twice the coil current.

The front and rear assemblies are assembled as follows. The four stator segments are first epoxied into the frame, after which the coil assemblies are inserted (Figure 3.21). The coils are also epoxied into the frame. The excess epoxy is removed using a milling process. The final step is to have the parts ground by a precision grinding machine to achieve flatness tolerances on the order of 10 µm. This ensures good contact between the mating surfaces of the stator. The front and rear assemblies are shown in Figure 3.22.
Figure 3.21: Illustration of stator segment and a single coil in frame. A slot, not shown here, is cut in the side of the assembly to allow for the exit of the coil leads.

Figure 3.22: Fully assembled front and rear magnetic bearing assemblies. In this photo each assembly has had the stator segments and coils epoxied in (two step process). The surfaces have been precision ground to achieve excellent surface contact. Terminal blocks have been installed for the coil connections.
Figure 3.23: Brushless motor stator supplied by Aerotech. The three Hall Effect sensors are spaced 120° (electrical) apart.

Figure 3.24: Fully assembled magnetic bearing middle assembly. The magnet arcs are shown along with the middle stator segments epoxied in to the stainless steel frame. Again, both sides of the assembly are precision ground to achieve excellent contact with the other assemblies.
The assembly of the magnetic bearing’s middle assembly is by far the most challenging and dangerous part of the process due to the strong magnetic fields produced by the magnets. Again, a two step epoxy process is employed and the surfaces are ground in a manner similar to the front and rear assemblies.

The spindle shaft assembly, shown in Figure 3.25, is fairly simple to put together. However, care must be taken with the precision ground stainless steel shaft. Also, a dial indicator should be used to ensure that the armature is well centered to the rotation axis of the spindle. The effort is repaid with improved dynamic balance and a better aligned axial force.

**Figure 3.25:** Fully assembled spindle shaft assembly. The journal bearing shaft is an existing component taken from the Precitech spindle. Once the additional components are mounted to the extension shaft, it is mounted to the journal bearing shaft. Axial force is applied to the magnetic bearing armature to provide feed motion and a torque produced by the brushless motor stator acts on the permanent magnet rotor to rotate the shaft.
**Figure 3.26:** Position sensor assembly. Here, the capacitive probe is shown mounted in the probe holder. The additional copper improves the ground connection between the probe and the back shell.

**Figure 3.27:** Fully assembled rotary axial spindle.
A photo of the ADE supplied capacitive probe displacement sensor is shown mounted in the probe holder in Figure 3.26. The photo also shows additional copper strands which provide an improved connection between the probe and the back shell. This effect is a marked improvement in capacitive probe noise should the set screws holding the back shell to the probe become loose. The final step is to install the assembly into the back of the rear assembly using four Phillips head screws.

3.6 Summary

In this chapter, the magnetic bearing, machine design, structural analysis, and assembly were reviewed. Section 3.1 discussed the rationale for choosing the up scaled fast tool servo as the magnetic bearing and section 3.2 presented the axial force calculation using the existing analytical approach along with a finite element analysis. The main components and materials are discussed in section 3.3. The vibration modes of the shaft assembly are predicted in section 3.4. Section 3.5 presents a brief overview of the rotary-axial spindle assembly process.
Chapter 4

Linear Power Amplifier for Magnetic Bearing

In order to control the axial position of the armature, currents must be driven through the magnetic bearing coils to generate force that balances the negative spring effect of the bias fluxes as well as disturbance forces. To accomplish this, a multichannel, high power, linear power amplifier is used. The specifications are 47 kHz bandwidth and 1 kW power output. A photo of the complete power amplifier is shown in Figure 4.1.

The power amplifier operation principle is as follows. Four PA52 power devices drive four separate coil circuits in parallel. The voltage output for each device is controlled individually and each of the four currents is measured using current sensing resistors. Now, due to mutual inductance of the coils, these currents are tightly coupled and the system cannot be controlled as four separate single input single output systems (SISO). So, a decoupling multi-input, multi-output (MIMO) current controller is used. More specifically, a feedback matrix takes the current feedback signals and performs a transformation of variables to average and differential currents. These four transformed variables are subtracted from four reference commands to produce an error signal which is regulated to zero by four analog controllers. The controllers produce four control signals which are retransformed and are appropriately given as references to the four power device voltage loops, thus closing the current loop. This MIMO theory was developed by Lu in his Ph.D. work [2:190-193], but he did not test it to its full extent.
The first section of this chapter briefly discusses power amplifier architectures and provides rationale for using a linear amplifier. Following this, some of the requirements for this power amplifier are given. Sections 4.3 and 4.4 present and analyze the voltage and current control principles. The control theory verification and overall performance will be presented in section 4.5 before the chapter is closed with a summary.

Figure 4.1: Linear power amplifier for magnetic bearing. It consists of two printed circuit boards: one for voltage control and one for current control. A half inch copper plate acts as a heat sink while nine finned copper CPU fans dissipate heat to ambient.

4.1 Power amplifier requirements

This section qualitatively reviews the power amplifier requirements. The overall goal of the rotary-axial spindle is to achieve kHz range axial motion control bandwidth so that very high axial dynamic stiffness can be reached. Therefore, the first requirement for the amplifier is that it have bandwidth of at least several tens of kHz. This ensures that the current control loop contributes very little phase lag to the compensated motion control loop transmission. A fast
acting amplifier is also needed to attack high frequency disturbances that are unavoidable in most manufacturing operations. In order to drive high frequency currents however, high voltages are required since the coil impedance is primarily inductive. It is difficult to predict the exact impedance of the coil circuits. But, a voltage supply requirement higher than that required for Lu’s fast tool servo can be expected. This is due to the quadratic relationship between coil turns and inductance. Based on this, an amplifier capable of outputting at least 50 to 100 V is needed.

Because the rotary-axial spindle application demands high static force, the amplifier’s low frequency current driving capability is also important. For basic operation (i.e. armature liftoff from rest against the stator), a sufficient force from the coils is required to balance the force produced by the permanent magnet as shown in (4.1).

\[ k_x I_{\text{coil}} = k_i X \]  

(4.1)

This liftoff force can be calculated and then related to the current using the value of \( k_x \) at an extreme armature position and \( k_i \) at maximum current from Table 3.1 (1.3 N/μm) and Table 3.2 (324 N/A). Knowing that half the nominal actuator gap is 625 μm, the minimum predicted DC current required for liftoff and position holding is 2.5 A per coil. In the mixed serial-parallel coil configuration, the relationship between coil current and power device current is:

\[ I_{ch} = 2I_{\text{coil}} \]  

(4.2)

This translates into 5 A per channel, or a 20 A total current requirement. (N.B.: from this point on, all currents referred to will be channel current unless otherwise stated). So, assuming a maximum voltage output of 50 to 100 V, and an inductive load, the dissipative power requirement is 1 to 2 kW.

At this point the search for an amplifier to drive the magnetic bearing can begin. In terms of architecture, amplifiers are either switching or linear. Switch mode amplifiers operate by
applying a high frequency duty cycled voltage to the load. In a pulse width modulation (PWM) topology, the duty cycle is proportional to the average output voltage. Modern PWM amplifiers switch the duty cycle from 18 to 50 kHz and can easily produce high power levels with high efficiency. However, they have several drawbacks. Because of the switching frequency in the tens of kHz range, the overall bandwidth is only a few kHz. Also, the transistors which are constantly being switched produce large amounts of noise which seriously degrades the performance of the capacitive probe sensor. Finally, the output current ripple will result in a force ripple at the switching frequency that acts on the armature, significantly reducing motion precision.

On the other hand, linear amplifier topologies, in theory, can result in drives with high bandwidth and “clean” output waveforms. Due to their fundamental limitation when it comes to energy inefficiency, however, their power dissipation levels are relatively low. In fact, a commercial solution producing tens of kHz bandwidth and greater than 1 kW dissipation are very difficult to acquire. Lu, however, designed a linear amplifier with 100 kHz bandwidth and 1 kW power dissipation for his fast tool servo [1, 2:151-196]. The maximum output voltage can range from ±30 to ±100 V, and the current output varies accordingly. In this research, the power stage from Lu’s work will be used.

4.2 Power amplifier architecture

This section gives an overview of the power amplifier system which is shown in Figure 4.2. The architecture can be broken down into three subsystems: the load, the power stage, and the control stage. The load consists of four coil circuits which can be modeled as pure inductors. However, the load is highly coupled due to mutual inductance: current in one coil induces a
current in another and vice versa. The coils are driven by four PA52 power devices. Each power device is enclosed in its own voltage control loop. A current sensing resistor, placed in each circuit, is the last element in power flow path. It provides current feedback to the current control system. The final stage is the current control system. The current control plant can be characterized as a MIMO system with voltage reference to each power device as input and current through each coil circuit as an output. Because of the mutual inductance, it cannot be treated as four separate SISO problems. So, four feedback signals are synthesized and are subtracted from four reference commands. Then four individual controllers are designed independently and their signals are distributed to the power stage reference input according to the decoupling scheme, thus closing the current control loop.

**Figure 4.2:** Power amplifier architecture. Each PA52 power device, which is enclosed in a voltage control loop, drives a coil load. The voltage loops receive references from the MIMO current controller.
Cascaded control architectures such as these have been used before in a variety of applications. For example, in the control of a machine tool feed drive, it is quite common to control the speed of the drive in a sub-loop and have the position controller produce speed commands to regulate the position. Liebman also used the cascaded voltage-current loop approach to tackle his SISO power amplifier problem [22:236-244]. Lu also used cascaded loops to control a MIMO problem similar to this case, but only regulated the average value of the currents, allowing them to vary. Here, an extension of that approach, which ensures all currents are equal, is presented.

4.3 Voltage control

This section covers the voltage control principle and analysis. Following sections describe the component selection process. A section discussing the implementation and the experimental results follows.

4.3.1 Principle and analysis

The voltage control stage is centered on the PA52 by Apex Precision Power. It is a power op-amp which offers 50 V/μs slew rate, ±100 V voltage output, can source up to 40 A, and can dissipate up to 400 W. No other commercially available amplifier has higher performance. Due to very high open loop gain and voltage supply limitations, the output voltage must be controlled. To do this, a modification of the non-inverting configuration [31] is employed. A schematic of this circuit is shown in Figure 4.3a. If the PA52 is assumed to be an ideal op-amp, that is, it has infinite open loop gain, resistors $R_1$ and $R_2$ become useless. The closed loop transfer function can then be modeled as a simple gain:
Figure 4.3: Voltage control principle. a) PA52 is enclosed in a non-inverting configuration with \( R_2 \) reducing the gain of the loop transmission to improve stability. b) Equivalent circuit representation. c) Block diagram for voltage control loop.
In reality, the op-amp is not ideal and has finite gain as shown in Figure 4.4. So, a non-ideal analysis is needed. The output voltage is expressed as:

\[ V_{\text{out}} = G_o(V_+ - V_-) \]  

(4.4)

where \( G_o \) is the open loop gain of the op-amp, \( V_+ \) and \( V_- \) are the voltages at the amplifier input terminals. In order to derive expressions for \( V_+ \) and \( V_- \), it is helpful to construct an equivalent circuit before moving forward. First, the op-amp is removed and the input terminals are labeled as shown in the left hand side of Figure 4.3b. Next, a Thévenin equivalent circuit is constructed with \( R_3 \), \( R_4 \), and \( V_{\text{out}} \). The potential from \( V_- \) to ground is used as the “terminals” for the equivalent circuit. Now, all sources should be grounded and the equivalent resistance between the terminals is the Thévenin resistance. This resistance is the parallel combination of \( R_3 \) and \( R_4 \). The voltage at \( V_- \) from \( V_{\text{out}} \) is:

\[ V_- = \frac{R_3}{R_3 + R_4} V_{\text{out}} \]  

(4.5)

which is calculated using a voltage divider. This is also the Thévenin voltage. Putting this together, the complete equivalent circuit is shown in the right hand side of Figure 4.3b. Since this is a linear system, the principle of superposition can be applied as the contributions from \( V_{\text{in}} \), and \( V_{\text{out}} \) to the differential voltage between \( V_+ \) and \( V_- \) are calculated. These contributions are easily handled using Ohm’s law and voltage division. The result is the block diagram shown in Figure 4.3c. If \( R_3/(R_3+R_4) \) is moved from the feedback to the feed forward loop, then the block diagram shown in Figure 4.3d results.

Here it is easier to see the effects of each resistor. \( R_3 \) and \( R_4 \) set the static gain of the system, which is equal to the ideal case. It is also apparent that \( R_1 \) and \( R_2 \) can be used to tune system dynamics and stability within the limitations set by the open loop gain.
4.3.2 Design

Having analyzed the voltage loop, it can now be designed. A logical starting point is with the closed loop gain. It will be set to fifteen so that up to 100 V can easily be outputted without requiring large swings from the OP27 based controllers which have slew rates of 2.8 V/μs [32:1]. However, slew rate saturation of the OP27 is only a problem when large signals are driven above 100 kHz [2:181]. An arbitrary value is chosen for one of the resistors and the other selected using (4.3). 4 kΩ and 56 kΩ make a good pair. However, care should be taken to ensure that the power limits of typically ¼ W resistors are not exceeded. Assuming a maximum PA52 output of 100 V then the current flowing through the resistors is:

\[ I = \frac{100 \, V}{60 \, k\Omega} = 1.6 \, mA \]  

(4.6)

And so the corresponding power dissipated at each resistor is:

\[ P_{56k} = 1.6 \, mA \times 56 \, k\Omega = 155 \, mW \]

(4.7)

\[ P_{4k} = 1.6 \, mA \times 4 \, k\Omega = 11 \, mW \]

(4.8)

The selections are acceptable for the ¼ W rating.

Before \( R_1 \) and \( R_2 \) can be selected, unity gain crossover frequency and desired stability margin should be chosen. If the loop was closed without \( R_1 \) and \( R_2 \), then the crossover frequency and phase margin would be about 660 kHz and 60° respectively. The stability of the voltage loop is paramount, so a fast dynamic response can be sacrificed to improve the stability. If the open loop gain is reduced by 30 dB, then the crossover frequency is halved to 330 kHz and the phase margin is raised to 80°. This is an improved scenario. Now that the total loop attenuation is known, an arbitrary selection in the kΩ range can be made for either \( R_1 \) or \( R_2 \), and the other can
be solved for using the information from Figure 4.3d. Selecting $R_1 = 3.3 \, k\Omega$ and $R_2 = 6.8 \, k\Omega$ sets the overall attenuation at 29.7 dB which is sufficiently close to the desired value.

![Figure 4.4: PA52 small signal open loop frequency response [33:3]. Loop transmission gain is shown in red on the left side.](image)

4.3.3 Implementation

To realize the circuit described in the previous section, an essentially identical setup to that of Lu [2:155-167] is used. A photo of the voltage stage printed circuit board (PCB) is shown in Figure 4.5. The circuitry described in the previous sections is replicated four times so that four power devices can be driven in parallel. With the exception of resistors $R_3$ and $R_4$, the assembly process was simply a matter of populating the PCB and following the assembly procedure outlined in Dr. Lu’s thesis.

Before soldering in the voltage feedback resistors $R_3$ and $R_4$, it is important to note that if the voltages applied to the load are different, then the currents flowing in the coils will also be different. So, it is important to match the closed loop gains of the four power devices as closely
as possible. To do this, sixteen 4 kΩ and sixteen 56 kΩ resistors with 0.1% tolerance were selected at random and their exact values were measured with a four lead multimeter. Then, a matrix of the gains that could be implemented with every combination was constructed. After scanning the 256 possibilities to find a few groups of four which had identical gains to two or three decimal places, the group with the lowest standard deviation was soldered into the board. The final values are given in Table 4.1. The standard deviation of the four values for $k_v$ is 0.2444.

**Table 4.1: Resistor selections**

<table>
<thead>
<tr>
<th>$R_3$ [kΩ]</th>
<th>$R_4$ [kΩ]</th>
<th>$k_v$ [V/V]</th>
</tr>
</thead>
<tbody>
<tr>
<td>4.3005</td>
<td>56.146</td>
<td>15.0204</td>
</tr>
<tr>
<td>4.0123</td>
<td>56.254</td>
<td>15.0204</td>
</tr>
<tr>
<td>4.0091</td>
<td>56.215</td>
<td>15.0208</td>
</tr>
<tr>
<td>3.9982</td>
<td>56.058</td>
<td>15.0208</td>
</tr>
</tbody>
</table>

The half inch copper heat sink was machined at the department’s machine shop since it required the placement of nearly 90 holes, and four pockets. Nine Zalman 7000B-CU copper finned fans [34] that were purchased from NCIX\(^1\) were installed on the backside of the copper plate. A view showing this side of the amplifier is shown in Figure 4.6. The whole assembly weighs approximately 25 kg and stands on four 5/8”-18 threaded rods.

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\(^1\) [www.ncix.com](http://www.ncix.com)
Figure 4.5: Voltage control board mounted to copper plate.

Figure 4.6: View of bottom side of power amplifier showing nine copper CPU fans which provide air cooling.
4.3.4 Experimental results

Now that the amplifier has been assembled, it can be tested to verify the design and performance. For a system which is approximately first order, much can be learned from the step response. Figure 4.7 shows the voltage loop response to a 1 V step input.

![Figure 4.7: 15 V step response of power devices. The step response time constant corresponds well with the crossover frequency.](image)

The first observation is that the closed loop gain is indeed 15 V/V. Additionally, the time spent matching gain resistors was well spent as the response amplitudes are nearly identical. Next, the bandwidth can be verified by looking for the characteristic time constant. The relationship between bandwidth and time constant for a first order system is well known to be:

\[ f_{bw} = \frac{1}{2\pi \tau} \]  

(4.9)
The 0.48 μs response corresponds to approximately 330 kHz bandwidth which matches very well with the design. Since the slew rate limit of the PA52 is 50 V/μs it is unlikely that larger steps will show significant slew rate saturation.

4.4 Current control method

Now that the power stage of the amplifier is complete, currents can readily be driven through the magnetic bearing coils. It is possible to simply apply a voltage to the coils without any additional control. However, since the coil impedance is primarily inductive, it will add -20 dB/dec slope to the open loop transfer function, and thus an extra 90° of phase lag. This makes it impossible to stabilize a position control plant which already has -180° phase lag while in rigid body motion. So, if the current is controlled through the magnetic bearing coils, then the force is also controlled. Thus, the issues involved with only voltage control are alleviated. The first part of this section discusses the MIMO coupled plant to be controlled. Then the control methodology is presented. Finally, the implementation of the current control system and the experimental results is covered.

4.4.1 Load

As explained earlier, the channel (one coil circuit) currents are highly coupled. In general, the input to output relationship can be written as:

\[
\begin{bmatrix}
I_A \\
I_B \\
I_C \\
I_D
\end{bmatrix} =
\begin{bmatrix}
Z_{aa} & Z_{ab} & Z_{ac} & Z_{ad} \\
Z_{ba} & Z_{bb} & Z_{bc} & Z_{bd} \\
Z_{ca} & Z_{cb} & Z_{cc} & Z_{cd} \\
Z_{da} & Z_{db} & Z_{dc} & Z_{dd}
\end{bmatrix}^{-1}
\begin{bmatrix}
V_A \\
V_B \\
V_C \\
V_D
\end{bmatrix}
\]  (4.10)
where \( V_A \) to \( V_D \) are the voltages produced by the power devices, and \( I_A \) to \( I_D \) are the corresponding currents. Each channel has a self-impedance, a close coupling with the adjacent coil circuit, and loose coupling with the two channels in the opposite housing. So, a simplifying assumption can be made. Grouping these three types of impedances accordingly leaves:

\[
[Z] = \begin{bmatrix}
Z_0 & Z_1 & Z_2 & Z_2 \\
Z_1 & Z_0 & Z_2 & Z_2 \\
Z_2 & Z_2 & Z_0 & Z_1 \\
Z_2 & Z_2 & Z_1 & Z_0 \\
\end{bmatrix}
\]

where the diagonal element \( Z_0 \) represents the self-impedance, \( Z_1 \) represents the closely coupled impedance, and \( Z_2 \) represents the loosely coupled impedance. This assumption can now be verified using a signal generator, a set of four voltage probes, and a current probe to measure the impedance matrix. The general procedure is as follows. First, three of the four channels are open circuited. Then, a voltage probe is connected from the PA52 output to ground on the circuit that is still intact. Next, for each of the open circuited channels, a voltage probe is connected to the positive terminal and the reference clip is connected to the negative terminal. This setup is shown in Figure 4.8.
Finally, a frequency sweep is performed using the voltage reference at the input to the PA52. The corresponding amplitudes and phase lags of the four voltages and current signal are measured. The impedances can then be calculated using the following:

$$Z_{xy}(j\omega) = \frac{I_x(j\omega)}{V_y(j\omega)}$$  \hspace{1cm} (4.12)

All sixteen impedances were calculated from 1 kHz to 400 kHz. It is a tedious task to measure sixteen frequency responses with sufficient frequency resolution. So, Dr. Lu’s Thunderstorm real time computer [2:197-226] was used as a dynamic signal analyzer with 1 MHz sampling frequency. Figure 4.9 shows the four self impedances, Figure 4.10 shows the four closely coupled impedances, and Figure 4.11 shows the eight loosely coupled impedances. Also included on each plot, respectively, are $Z_0$, $Z_1$, and $Z_2$. To calculate these values, a geometric mean of the magnitudes is taken, and an arithmetic mean is used to compute the angle [2:183]. It is more intuitive to use a geometric mean to calculate the magnitude and arithmetic mean to calculate the phase since magnitudes naturally multiply and angles naturally add.
Analyzing the Bode plots draws the conclusion that the assumption was valid and that the impedances can be lumped into the three groups. As expected, impedances are primarily inductive, though parasitic capacitance is evident above 100 kHz. There are two reasons for this. The first is that the pancake coils act as parallel plates. The second is related to capacitive coupling from the twisted coil leads. The overall effect imposes a fundamental bandwidth limitation.

Figure 4.9: Frequency response of self impedances.
Figure 4.10: Frequency response of closely coupled impedances.

Figure 4.11: Frequency response of loosely coupled impedances.
4.4.2 Decoupling and control principle

This section explains the decoupling scheme needed to control the strongly coupled power amplifier system. Decoupling the plant implies diagonalizing it. To diagonalize an arbitrary $n \times n$ matrix $[A]$, the following linear transformation is applied:

$$[D] = [B]^{-1} [A] [B]$$

(4.13)

where $[B]$ is a matrix containing $n$ linearly independent eigenvectors of $[A]$. To diagonalize a matrix it is necessary for it to have $n$ linearly independent eigenvectors. However, it is sufficient to find $n$ distinct eigenvalues [35].

To decouple, or diagonalize the system, the inverse of $[Z]$ must be computed. MATLAB’s Symbolic Math Toolbox [36] is used to hasten the process. This aid can also calculate the eigenvalues and eigenvectors. The matrix of the eigenvalues of $[Z]^{-1}$ is denoted as $[P]$:

$$[P] = \begin{bmatrix}
\frac{1}{Z_0 + Z_1 + 2Z_2} & 0 & 0 & 0 \\
0 & \frac{1}{Z_0 - Z_1} & 0 & 0 \\
0 & 0 & \frac{1}{Z_0 - Z_1} & 0 \\
0 & 0 & 0 & \frac{1}{Z_0 + Z_1 - 2Z_2}
\end{bmatrix}$$

(4.14)

There are only three distinct eigenvalues. However, MATLAB calculates four linearly independent eigenvectors, which means that the impedance matrix is indeed diagonalizable. The eigenvectors can be concatenated into a matrix $[V]$ which has inverse $[W]$:

$$[V] = \begin{bmatrix}
1 & -1 & 0 & -1 \\
1 & 1 & 0 & -1 \\
1 & 0 & -1 & 1 \\
1 & 0 & 1 & 1
\end{bmatrix}$$

(4.15)
The product of \([V], [Z]^{-1}\), and \([W]\) is the matrix of eigenvalues (4.14) which is diagonal. Each entry in this matrix becomes the plant of a SISO control problem. To control these plants, a diagonal control matrix \([C]\) can be designed:

\[
[C] = \begin{bmatrix}
C_1(s) & 0 & 0 & 0 \\
0 & C_2(s) & 0 & 0 \\
0 & 0 & C_3(s) & 0 \\
0 & 0 & 0 & C_4(s)
\end{bmatrix}
\]

This closes the current loop. A block diagram of the current loop is shown in Figure 4.12.

---

\(\{i_r\}\) represents the reference vector, \(\{v_i\}\) is the vector of control signals produced by \([C]\), \(\{v_r\}\) is the set of reference voltages that are amplified by the power stage, \(\{V_o\}\) is the set of power stage voltages, \(\{I_o\}\) is the set of channel currents, and \(\{i_S\}\) is the transformed feedbacks that are subtracted from the reference signals. \([G_v]\) is a diagonal matrix of power stage frequency responses and \(R_s\) is the current sensing resistor gain. For clarity, the nomenclature is given below:
\[
\{v_i\} = \{v_1, v_2, v_3, v_4\}^T \tag{4.18}
\]
\[
\{v_r\} = \{v_{ra}, v_{rb}, v_{rc}, v_{rd}\}^T \tag{4.19}
\]
\[
\{V_o\} = \{V_A, V_B, V_C, V_D\}^T \tag{4.20}
\]
\[
\{I_o\} = \{I_A, I_B, I_C, I_D\}^T \tag{4.21}
\]
\[
\{i_S\} = \{i_{1S}, i_{2S}, i_{3S}, i_{4S}\}^T \tag{4.22}
\]

Since the system is linear and decoupled, any fundamental stability criterion such as Nyquist or phase margin based stability will suffice.

4.4.3 Physical interpretation

Putting the engineer’s hat back on allows for the derivation of some physical intuition. To start, look at the product of feedback matrix \([W]\) with \(\{I_o\}\) and \(R_s\). The first output is proportional to the average value of currents. This is called the common mode current. The next output is proportional to the difference between \(I_B\) and \(I_A\) (front actuator currents) and the third is proportional to the difference between \(I_D\) and \(I_C\) (rear actuator currents). The fourth output is proportional to the difference between front and rear currents: \(I_D + I_C - I_B - I_A\). These three quantities are called differential currents. So, if the first feedback is controlled, then the average value of the current (and consequently, the actuator force) is controlled. And if a zero reference is given to the second controller, \(I_A\) and \(I_B\) will be equal. Similarly, if the third feedback is controlled and commanded to be zero, \(I_C\) and \(I_D\) will be equal. If the same is done for the fourth feedback, \(I_A\) and \(I_B\) will be maintained equal to \(I_C\) and \(I_D\). Thus, all currents are regulated to the same value. The relationships between the feedback sums and actual channel currents are described in (4.23). The reference commands are given in (4.24).
In summary, this feedback matrix \([W]\) performs a change of variables from physical quantities to common mode and differential sums. The role of the matrix \([V]\) is to retransform the control signals \(\{v_i\}\) to the physical domain of PA52 reference voltages.

Physical sense can also be made of the eigenvalues. The first eigenvalue, representing the plant for common mode current control, is the reciprocal of the sum of a single self-impedance, and a single closely coupled impedance, and two loosely coupled impedances. In other words, this is the load seen by one power device when the others are open circuited. The second eigenvalue can be thought of as the difference between the impedance of the coil circuits in the front stator housing. The third eigenvalue can be thought of as the difference between the impedances in the front housing and the rear. This matches the intuition developed in the previous paragraphs.

4.4.4 Implementation

Due to very fast dynamics and the simplicity of system, digital control is not warranted nor is it practical. It is much easier and more cost effective to implement control system in the analog domain. Feedback summing, control, and control signal summing stages are implemented with operational amplifier based circuits on a PCB. The schematics showing the implementation of the feedback stage are shown in Figure 4.13.
Figure 4.13: Feedback stage implementation. a) Summer circuit to generate $i_{1S}$. b) Difference circuit to generate $i_{2S}$. c) Difference circuit to generate $i_{3S}$. d) Difference circuit to generate $i_{4S}$.

To generate $i_{1S}$, an inverting summer circuit [37] is used, while difference circuits are used to synthesize the other three feedbacks. Here, an OP37 op-amp is used instead of the popular OP27 because the OP37 is optimized for gains of five or greater (Figure 4.13b and Figure 4.13c) [38:1]. The final relationship from output current to sensing variables is given as:
\[
\begin{bmatrix}
i_{1s} \\
i_{2s} \\
i_{3s} \\
i_{4s}
\end{bmatrix} = R_s \begin{bmatrix}
-2.5 & -2.5 & -2.5 & -2.5 \\
-5 & 5 & 0 & 0 \\
0 & 0 & -5 & 5 \\
-2.5 & -2.5 & 2.5 & 2.5
\end{bmatrix} \begin{bmatrix}
I_A \\
I_B \\
I_C \\
I_D
\end{bmatrix}
\]  
(4.25)

This is essentially a scaled version of \([W]\). The first row is inverted to allow for the use of the simple inverting configuration. Additionally, the whole matrix is scaled by a factor of five to compensate for the current sensing resistor gain and by a factor of two to maximize the amount of current that can be commanded. Since the digital to analog converter (DAC) on our dSPACE control box has limits of ±10 V, the gain of two will enable an amplifier input range of ±5 A per channel. Since the gains are applied across the whole matrix, it does not affect the decoupling.

The next block in the loop is the controller matrix. According to the model, the plants to be controlled have simple scaled integrator transfer functions, so a proportional-integral (PI) control with low pass filter to attenuate parasitic effects is sufficient. A common topology for this controller is taken from [2:170]. Schematics and a block diagram are shown in Figure 4.14.
This implementation is used for each of the four controllers. $R_c$ and $C_c$ perform the integral action. $R_b$ can be changed to adjust the proportional loop gain. $C_d$ contributes an additional pole for high frequency roll off. From the block diagram, an expression for the controller can be formed as:

$$C_x(s) = \frac{1}{R_b C_d} \times \frac{1 + R_c \left( C_c + C_d \right) s}{1 + R_c C_d s} \quad (4.26)$$

The “back to back” Zener diodes clamp OP27’s output. This also limits the PA52 output. This prevents the PA52 from hitting the power supply rails, $V_s$. The saturation voltage is the sum of the Zener voltage $v_z$ plus the forward voltage drop $v_f$ of the diode. The diodes are chosen such that:
\[ v_z + v_f < \frac{V_s}{k_v} \]  

(4.27)

where \( k_v \) is the voltage gain of the power stage. A block diagram for this circuit can be formed and is shown in Figure 4.14b. Since \( V_s = \pm 50 \) V, \( k_v = 15 \) V/V, and the typical forward drop for a diode is 0.7 V, a Zener diode with \( v_z < 2.6 \) V is acceptable. The selection of the resistor and capacitor values is discussed in the next section.

The final element in the current control loop is the control signal summing stage, which is illustrated in Figure 4.15. Here, voltage signals from four controllers are summed and distributed to the voltage stage reference inputs, implementing the matrix \([V]\). Here, the summer and difference topologies, based on the OP27 which has better unity gain performance when compared to the OP37 [38:11], are used.
Figure 4.15: These summing circuits add the control signals which are fed as reference signals to the power stages. Circuit a) feeds device A. Circuit b) feeds device B. Circuit c) feed device C. Circuit d) feeds device D.

Again, Zener diodes are used to cap output voltage, using the criteria given in (4.28). The relationship from \( \{v_i\} \) to \( \{v_r\} \) is given below:

\[
\begin{bmatrix}
    v_{RA} \\
    v_{RB} \\
    v_{RC} \\
    v_{RD}
\end{bmatrix} =
\begin{bmatrix}
    -1 & -1 & 0 & -1 \\
    -1 & 1 & 0 & -1 \\
    -1 & 0 & -1 & 1 \\
    -1 & 0 & 1 & 1
\end{bmatrix} \begin{bmatrix}
    v_1 \\
    v_2 \\
    v_3 \\
    v_4
\end{bmatrix}
\]

(4.28)
This is a direct implementation of $\mathcal{V}$ except that the sign of the first column is switched to match the first row of the feedback matrix, thus positive feedback is eliminated.

Also, it is important to make sure that the op-amp based controllers can supply sufficient current to drive the control signal summing op-amps. For example, $v_I$ must drive four parallel 5.1 k$\Omega$ resistors which constitute a 1.2 k$\Omega$ load for this particular OP27. However, the output swing vs. load plot in the OP27 datasheet [32:11] shows that with the 1.2 k$\Omega$ load, the op-amp is capable of driving ±13 V. Since $v_I$ is capped to < 3.3 V the current output capability will not be exceeded.

The above circuits are implemented on a PCB which is shown in Figure 4.16. This is a four layer PCB (two signal layers, two planes). Two 96 contact male connectors on the bottom of the board attach it to mating connectors on the voltage stage board. The current feedback signals pass from the voltage stage to the current stage, and the voltage reference signals pass from the current stage to the voltage stage.

![Current control PCB](image)

**Figure 4.16**: Current control PCB. The four feedback circuits are located along the top of the board. The four control circuits are situated in the middle of the board. The operational amplifiers summing the control signals for voltage references can be found along the bottom of the board. Common mode reference signal is inputted from the BNC connectors on the right hand side of the board.
4.4.5 Controller design

Having synthesized and implemented a decoupling scheme for the system, the next step is controller design. First, system identification must be performed to verify the model (4.14). To do this, the sixteen frequency responses which make up the MIMO open loop plant must be measured. The inputs are \( v_1 \) through \( v_4 \) and the corresponding outputs are \( i_{1S} \) through \( i_{4S} \). This matrix is denoted \([P']\). To perform this measurement, swept sine waves are inputted at \( v_1 \) while the other three inputs are grounded. Then, the response is measured at \( i_{1S} \) through \( i_{4S} \). This setup is shown in Figure 4.17. The process is repeated for all four inputs.

![Figure 4.17: Schematic of experimental setup for measuring open loop MIMO plant.](image)

First, the frequency responses of the diagonal elements of \([P']\), \( v_1 \) to \( i_{1S} \), \( v_2 \) to \( i_{2S} \), \( v_3 \) to \( i_{3S} \), and \( v_4 \) to \( i_{4S} \), were measured. These are the plants that are to be controlled. The measurements,
shown in Figure 4.18, were compared to the plants which are derived from the impedance measurements, and (4.14).

![Figure 4.18: Comparison between experimental measurements of diagonal based plants (\(v_j\) to \(i_{JS}\), \(j = 1\) to \(4\)) and a plant model based on the impedance measurements. The difference at high frequency is due to very slight differences in \(Z_0\) and \(Z_1\). This is caused by parasitic capacitance in the BNC cables of the measurement device.]

The agreement is exceptional at low frequencies where the load is purely inductive. In Figure 4.18a this agreement nearly extends to 100 kHz, 20 kHz in Figure 4.18b and Figure 4.18c, and to 100 kHz in Figure 4.18d. Parasitic capacitance from the measurement cables is the cause
of the anomalies. Performing impedance measurements with the Thunderstorm computer approximately 4 m away and right beside the power amplifier confirmed this.

Furthermore, the off-diagonal elements should be measured to quantify the success of the decoupling method. Again, the Thunderstorm DSA was used to perform the measurements and the results are shown in Figure 4.19.

**Figure 4.19:** Decoupling results showing the magnitude of the frequency response from a) $v_1$ to $i_{1S}$ through $i_{4S}$ b) $v_2$ to $i_{1S}$ through $i_{4S}$ c) $v_3$ to $i_{1S}$ through $i_{4S}$ d) $v_4$ to $i_{1S}$ through $i_{4S}$
In the ideal case, the off-diagonal entries of \([P']\) are zero, which corresponds to perfect decoupling. In reality, these entries are expected to be non-zero but small compared to the diagonal entries of \([P']\). Looking at the measurements shows mixed results. Frequency responses with input \(v_2\) and \(v_3\) show very good decoupling (Figure 4.19b Figure 4.19c). Also, the common mode output, \(i_{1S}\), is well decoupled from the inputs to all three differential modes \((v_2, v_3, \text{ and } v_4)\). This means that the average value of the four currents is relatively unaffected when one of these differential controllers produces a control effort. In all cases, the diagonal term is the dominant response, though only up to 70 kHz for the common mode plant \((v_1 \text{ to } i_{1S})\).

Even though the decoupling is not perfect, the system will still work well. For example, if \(C_1(s)\) outputs a control effort to track common mode current command, the results in Figure 4.19a indicate that the common mode response will be comparable to the difference between \(I_A\) and \(I_B\), and between \(I_C\) and \(I_D\), therefore errors build at the input to controllers \(C_2(s)\) and \(C_3(s)\), but not at \(C_4(s)\). When \(C_2(s)\) and \(C_3(s)\) output a control signal to regulate the error to zero, it will not affect the common mode current. In this sense, the decoupling, though indirect, is effective.

It is not by luck that things work out in this way. Returning to (4.14), the second and third entries in the matrix are the reciprocal of \(Z_0 - Z_1\). However, in the low frequency region, these two impedances are primarily inductive, and vary by only 1% on average. So, the reciprocal of the difference is quite large, making the terms dominant compared to the others.

Now that there is confidence in the decoupling theory, the four controllers can be designed based on the diagonal entries using a loop shaping approach. The metric used for judging stability will again be the phase margin of the negative of the loop transmission. With the chosen PI, low pass filter topology, the general process is as follows. First, the unity gain crossover frequency, \(f_c\), is selected where the plant measurement is dominant compared to the
other three outputs from the same input. Also, there must be sufficient phase margin for stability’s sake. Next, the pole for the low pass filter, \( f_p \), is set such that any parasitic capacitance is well attenuated. Finally, the integrator frequency, \( f_i \), is set at least ten times smaller than the crossover frequency. Then, select components which realize the designed frequencies. Table 4.2 gives the parameters used in this implementation.

### Table 4.2: Parameters for current control circuits

<table>
<thead>
<tr>
<th>( C_x(s) )</th>
<th>( f_c ) [kHz]</th>
<th>( f_l ) [kHz]</th>
<th>( f_p ) [kHz]</th>
<th>( R_a ) [kΩ]</th>
<th>( R_b ) [kΩ]</th>
<th>( R_c ) [kΩ]</th>
<th>( C_c ) [nF]</th>
<th>( C_d ) [pF]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>40</td>
<td>6.4</td>
<td>45</td>
<td>0.51</td>
<td>0.51</td>
<td>25</td>
<td>1</td>
<td>140</td>
</tr>
<tr>
<td>2</td>
<td>120</td>
<td>6</td>
<td>930</td>
<td>12</td>
<td>12</td>
<td>3.9</td>
<td>6.8</td>
<td>44</td>
</tr>
<tr>
<td>3</td>
<td>120</td>
<td>6</td>
<td>930</td>
<td>12</td>
<td>12</td>
<td>3.9</td>
<td>6.8</td>
<td>44</td>
</tr>
<tr>
<td>4</td>
<td>107</td>
<td>0.5</td>
<td>624</td>
<td>3.9</td>
<td>3.9</td>
<td>51</td>
<td>6.8</td>
<td>5</td>
</tr>
</tbody>
</table>

The resulting negative of the loop transmission Bode plots for each loop are shown in Figure 4.20. The phase margin for the common mode control loop is 50° at a crossover frequency of 40 kHz. The closed loop current bandwidth can be expected to be slightly higher than this. For the first two differential modes, calculating the difference between the front and rear currents, the crossover frequencies are 120 kHz with 90° of phase margin. The third differential loop has crossover frequency 107 kHz with nearly 90° of phase margin.
If the decoupling was perfect, then the job would be complete. However, as was shown in Figure 4.19, the decoupling is imperfect, yet controllers have been designed and stability has been assessed assuming that it is. In Lu’s study on the MIMO power amplifier problem, he provides a proof that rationalizes the use of the generalized Nyquist stability criteria [2:190:193]. To apply the theory to this case, the product of \([C]\) and \([P']\) is computed, and then the eigenvalues are solved for. Since the experimental data is discrete in the frequency domain. So, for the \(k^{th}\) frequency point of a set of \(N\) points:

\[
[G]_k = [C]_k [P]_k
\]  

(4.29)
which is a four by four matrix. Then the loci of the eigenvalues are calculated, solving for $\lambda$ in
the following equation:

$$\text{det}(G_k - \lambda I) = 0$$  \hspace{1cm} (4.30)

In the end, there are $N$ sets of four eigenvalues which are assembled into the matrix $[\Lambda]$:

$$[\Lambda] = 
\begin{bmatrix}
\lambda_{11} & \lambda_{12} & \lambda_{13} & \lambda_{14} \\
\lambda_{21} & \lambda_{22} & \lambda_{23} & \lambda_{24} \\
\vdots & \vdots & \vdots & \vdots \\
\lambda_{(N-1)1} & \lambda_{(N-1)2} & \lambda_{(N-1)3} & \lambda_{(N-1)4} \\
\lambda_{N1} & \lambda_{N2} & \lambda_{N3} & \lambda_{N4} \\
\end{bmatrix}$$  \hspace{1cm} (4.31)

Bode plots are produced, one for each column of $[\Lambda]$. Furthermore, each column corresponds to one of the current control mode negative loop transmissions. Therefore, a direct comparison can be made between the stability judged by looking only at the diagonal elements of $[P']$ and the stability judged by looking at all the elements of $[P']$. Practically speaking, however, MATLAB’s eigenvalue solver does not return the eigenvalues in any particular order because they are just roots of a characteristic polynomial. The result is a set of four Bode plots which are discontinuous. However, D’Errico wrote a MATLAB script, `eigenshuffle`, which consistently sorts the eigenvalues [39]. The end result is shown in Figure 4.21. The results are essentially identical. It is safe to conclude that the stability can be judged solely on the phase margin of the diagonal elements of $[P']$. 

91
4.5 Performance results

This section verifies the designed performance of the power amplifier. To quantify the performance, the speed of the response and the functionality of the MIMO control system are observed. For these tests, the power supply rails were set to 50 V, and the PA52 booster circuits [2:163:167] were not implemented. (N.B.: these boost circuits allow the small signal stages of the PA52 to drive rail to rail). Figure 4.22 shows the response of the system common mode current to swept sine waves ranging from 50 mA to 0.5 A, peak to peak. The small signal -3 dB

**Figure 4.21:** Comparison between simplified and full-order stability.
bandwidth is 47 kHz which is slightly higher than the 40 kHz crossover frequency, as expected. As the signal amplitude increases, the effective bandwidth is reduced. This is directly related to the voltage supply limitations and high coil impedance. However, since this is not a fast tool servo application, it should not pose any problems. If higher large signal bandwidth is required, the boost circuit as discussed in Lu’s work [2:163:167] can be implemented but this will only yield marginal improvements. Alternatively, much higher voltage supply rails should be used, thus necessitating an amplifier with much higher power dissipation capability.

![Figure 4.22:](image.png)

**Figure 4.22:** Closed loop current response at various signal amplitudes. The drop off at higher amplitudes is due to saturation of the controller output signals.

The next set of verification tests are step responses. The 20 mA step response shown in Figure 4.23 shows good linearity, and an approximate peak time of 19 μs for the slowest responding current. Another observation is that the four currents converge to an equivalent value in approximately half the time that it takes the common mode to settle. This is logical since the
differential control loops have bandwidths approximately two to three times higher than the common mode loop. Also, this shows that the MIMO system is performing the function of equalizing the currents. The overshoots of 25% to 50% are also respectable considering the 35% damping ratio observed in the closed loop frequency response of the common mode.

The 0.5 A step response shown in Figure 4.24 demonstrates the voltage saturation issue. The amplifier cannot drive right up to the voltage rails due to the protecting Zener diodes whose Zener and forward voltage drops will change slightly depending on the amount of current being driven through them by the OP27s. In either case, the high current output pins of the PA52 cannot make full swings without the aid of the boost circuit [33:4] which is not implemented here.

![Figure 4.23](image.png)

**Figure 4.23:** 20 mA step response of quad-channel power amplifier.
Figure 4.24: 0.5 A step response showing voltage saturation.

4.6 Summary

This chapter presented the control, implementation, and experimental testing of a 1 kW linear power amplifier. Section 4.1 briefly discussed the requirements for this amplifier and the reasons for choosing a linear topology. Section 4.2 explained the amplifier architecture. Section 4.3 described the power device voltage control principle and its implementation. Section 4.4 derived a method for controlling the currents through the highly coupled magnetic bearing coils. After explaining the decoupling and control principles, stability criteria, and implementation, the
experimental results were presented. The decoupling, though imperfect and indirect, results in a
stable system with equalized currents. The overall bandwidth of the common mode current is 47
kHz. Voltage saturation becomes an issue at high frequency and large current step changes due
to high coil impedance.
Chapter 5

Rotary Motion Sensing and Control

In spindle applications such as milling, grinding or axis symmetric turning, precision rotary motion control is not necessary. The only requirements are that high rotational speeds be maintained and that cutting torque disturbances are rejected. This warrants feedback control. However, a rotary encoder, the most common means of feedback in a precision spindle, cannot be used in this application. To address this, a “sensorless” speed feedback algorithm is developed to generate the speed feedback, and its efficacy is experimentally verified in this chapter.

In spindle speed control, it is the average speed that is regulated. When measuring average speed there is an inversely proportional relationship between sensor bandwidth and averaging period. A longer averaging period results in a more precise speed measurement but sensing bandwidth is reduced. This reduces motion control bandwidth and disturbance rejection performance. However, angular resolution is also a factor. When it is increased, the speed measurement is refined for a given averaging period. Consequently, the averaging period can be shortened to increase the sensor bandwidth. Here, the goal is to design a sensorless speed measurement algorithm with high resolution so that a sufficiently high measurement bandwidth can be achieved.

The principle of the sensorless speed algorithm, designed by Professor Lu, is as follows. The Hall effect sensors (HES) are used to provide a coarse indication of the permanent magnet rotor position. The HES output digital signals which can be encoded into a quadrature signal, much like an encoder. Assuming the spindle speed is constant, then the time between two HES
edges triggering is equal. This information can be used to generate additional pulses between the HES edges, thus increasing the angular resolution. Finally, due to manufacturing tolerances and stray fields, an “on the fly” calibration technique correctly locates the HES edges in the spatial domain. This makes the final speed measurement as good as that produced by a 1000 line per revolution optical encoder.

In the remaining parts of this chapter, beginning with section 5.1, a review of existing speed measurement technology is given. Section 5.2, describes and models the Precitech SP-75 spindle, which is used for evaluating the algorithm’s performance. Section 5.4, introduces the hardware on which the sensorless speed algorithm is implemented. Following this, sections 5.5 through 5.7 explain in more detail the various elements of the speed sensing algorithm while providing comparisons to the results obtained from the test bed. Finally in section 5.8, the algorithm is implemented on the rotary-axial spindle and the results are reported. The chapter is closed with a summary.

![Diagram](image)

**Figure 5.1:** Encoder schematic. Side view shows the required constant air gaps.
5.1 Existing rotary motion sensing techniques

In this section a review of some of the traditional methods of spindle rotary motion control is given. Optical encoders are the most common precision rotary motion feedback sensors. The disc type shown in Figure 5.1 has grating marks along the perimeter. As the disc rotates, the gratings pass between the photo diode and the read head producing a pair of digital signals which are shifted 90° from each other. Very high angular resolutions are possible. This technology has one limitation with respect to the rotary-axial spindle application. It requires a fixed gap between the photo diode and disc, and photo detector and disc of approximately 100 μm. Obviously, this is not acceptable since the rotary-axial spindle will have 1 mm axial stroke. Drum type encoders (Figure 5.2) offer another potential rotary motion sensing because they tolerate small axial motions. However, the installation of such an encoder would add shaft length, thus lowering the frequency of the first bending mode. This is an undesirable effect.

![Figure 5.2: Schematic of drum type encoder. These encoders can tolerate small axial motions.](image)

Tachometers, which are essentially DC motors with an open circuit stator winding, offer another potential rotary sensing solution. Since the winding is open circuited, the voltage across the terminals is the motor’s back EMF, and so it is proportional to speed. One common problem with tachometers is a noisy signal at low speeds. This is not really an issue in this application,
though, because only high speed motion is considered. However, the sensor does add shaft length which reduces the shaft bending mode frequencies. Additionally, care must be taken to prevent interaction with the magnetic bearing and brushless DC motor. Continuing along the same train of thought, the back electromotive force of the brushless DC motor can be used as a feedback since it is proportional to speed. However, this would require measurement of the phase voltages, and additional filtering to remove noise generated by the pulse width modulation (PWM) amplifier.

Jones and Lang [40, 41] introduced another potential rotary motion measurement method. They developed a sensorless non-linear observer which estimates the phase currents and the rotor speed and angle. The rotor angle measurement can be used to commutate the motor, and the speed can be used for motion control. The main issue with this is that a well defined model of the system must exist. Liebman found that errors in model parameters of only 10% can drive the observer unstable [22:281].

Since none of these existing methods suit our application, a new speed feedback method will be developed.

5.2 Test bed

The sensorless speed algorithm will be tested on the SP-75 spindle (Figure 5.3), which was donated by Precitech\(^2\). This will provide a good testing platform since the brushless DC motor on this spindle will be used on the rotary-axial spindle. Also, a 1000 line per revolution encoder has been installed on the tail of the spindle to compare our sensorless results.

\(^2\) www.precitech.com
Figure 5.4 shows a block diagram describing the interaction between the various hardware elements in the rotary control loop. The three-phase brushless DC motor is driven by an Aerotech BA10 PWM amplifier (donated by Aerotech\(^3\)) with 20 kHz switching frequency [42]. The HES, which are mounted to the stator, are fed back to drive in order to commutate the motor. A rotary encoder measures the rotation of the shaft and outputs quadrature signals which are then fed to the dSPACE DS1103 real time computer. Here the control routines can be executed in real time, at a deterministic rate. Programs with several discrete transfer functions, gain blocks, arithmetic, and analog to digital and digital to analog conversion can easily be

\(^3\) www.aerotech.com
executed with 100 kHz sampling rate. The output of the motion controller is a voltage which is proportional to the average torque applied to the permanent magnet rotor.

![Block diagram of rotary control loop hardware.](image)

**Figure 5.4:** Block diagram of rotary control loop hardware.

5.3 Modeling and rotary speed control with optical encoder feedback

This section deals with four topics related to the rotary speed control of the SP75 spindle. First, a brief overview of rotary control of brushless DC motors will be given. Next, the system will be modeled and compared to an experimental measurement. Finally, the speed feedback will be synthesized and a controller will be designed to regulate the average spindle speed.

5.3.1 Overview of brushless DC motor control

Here it is helpful to provide a brief overview on the control of DC brushless motors which is the rotary actuator chosen to provide the rotary motion for the rotary-axial spindle. A DC brushless motor is essentially a synchronous motor with a permanent magnet rotor. To rotate
the shaft, a rotating magnetic field must be produced in the stator and it must be phased correctly with respect to the poles of the permanent magnet(s) on the rotor.

In general, this external field is produced by driving sinusoidal currents of the same frequency through the multiple windings of the motor stator. In this case, there are three phases to control. It can be difficult to command these phase currents directly. But, Park’s transformation [27:268], which transforms the three currents and views them from a rotating reference frame, reduces the problem to just two: direct current $i_d$, and quadrature current $i_q$. Further reduction can be had in the case of a round, that is, non-salient, rotor. The rotor in this application meets this criterion. The result is that $i_d$ also drops out, leaving only $i_q$. In this case, the motor torque $T_m$ is:

$$T_m = \frac{3}{2} k_i i_q$$

(5.1)

where $k_i$ is the motor’s torque constant. The factor of 1.5 is a result of Park’s transformation. As with most rotary drives set to operate in torque mode, the BA10 takes an input that is proportional to $i_q$, and performs the necessary calculations to produce the three phase currents based on the rotor position information obtained from the HESs. Practically speaking, this means that instead of commanding three alternating currents, the motor torque can be controlled directly from $i_q$, without modulation.

5.3.2 Modeling

Before proceeding to the controller design section, the system must first be modeled. A block diagram of the open loop rotary plant is shown in Figure 5.5. The control effort, $u$, is scaled by the reciprocal of $k_d$, the digital to analog converter (DAC) gain to provide a one to one relationship between the input and output. At this point, a delay is usually modeled to account for
the zero order hold of the DAC. However, it will be omitted since the sampling frequency of 10 kHz is much higher than the anticipated motion bandwidth of 10 Hz. The DAC output voltage, $i_r$, is sent to the rotary amplifier as the quadrature current reference. This signal is amplified by $k_{vr}$ resulting in $i_q$, the quadrature current. To measure $k_{vr}$, the HESs were disconnected from the motor to disable commutation, and the output current in a given phase was measured with a current probe. The result was 0.4756 A/V.

The next block in the diagram is the torque constant $k_t$. Aerotech specifies it to be 0.105 Nm/A. The shaft can be considered a rigid body, so the transfer function from motor torque $T_m$ to output angle $\theta$ is:

$$
\frac{P_r(s)}{T_m(s)} = \frac{1}{J_{shaft}s^2}
$$

(5.2)

Where $J_{shaft}$ is the spindle shaft inertia, and $s$ is the Laplace variable. A solid model was constructed in SolidWorks\textsuperscript{4} to calculate the shaft inertia using the mass properties tool. The resulting inertia is 0.00072 kg-m\textsuperscript{2}. For now, the rotor angle is measured by the optical encoder.

---

\textsuperscript{4} www.solidworks.com
This measurement method results in two effects. The first is the addition of encoder error caused by, among other things, encoder disc and spindle shaft misalignment and errors in encoder grating placement [2:256]. The end result is that a measurement error, $\theta_n$, is superimposed on the true angle measurement $\theta$. This is modeled with the summation of $\theta_n$ and $\theta$. Additionally, since this optical encoder is a digital sensor, its output is discretized into 1000 lines/rev corresponding to an angular resolution of $\frac{1}{4}$ line or 0.09°. This is modeled with the quantization block. The signal presented to the digital controller is $\theta_m$ and has units of lines. To obtain a speed measurement, $\theta_m$ must be differentiated. To do this in the digital domain, $\theta_m$ is subtracted from its value at the previous sampling interval, and scaled by the sampling interval $T_s$, which is 0.0001 s. This creates the intermediate speed signal $\omega_{\text{int}}$ which has units of lines/s. Multiplying $\omega_{\text{int}}$ by both 60 and $k_e$ converts this quantity to revolutions per minute (RPM). The gain $k_e$ depends on the resolution of the measurement device. In this case, $k_e$ is 1000 lines/rev. At this point, the open loop model of the system is complete. Each parameter in the open loop system is given in Table 5.1.

### Table 5.1: Rotary control loop parameter list.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>$k_d$</td>
<td>0.1</td>
<td>V/V</td>
</tr>
<tr>
<td>$k_{\text{fr}}$</td>
<td>0.4756</td>
<td>A/V</td>
</tr>
<tr>
<td>$k_i$</td>
<td>0.105</td>
<td>Nm/A</td>
</tr>
<tr>
<td>$k_e$</td>
<td>1000</td>
<td>lines/rev</td>
</tr>
<tr>
<td>$T_s$</td>
<td>0.0001</td>
<td>s</td>
</tr>
<tr>
<td>$J_{\text{shaft}}$</td>
<td>0.00072</td>
<td>kg-m^2</td>
</tr>
</tbody>
</table>

To verify it, the dSPACE DSA tool, written by Lilienkamp [43], is used to perform a swept sine wave test. The excitation was given at $u$ and $\omega$ was taken as the output. The resulting frequency response is shown in Figure 5.6. With the exception of a few degrees of phase lag
above 10 Hz, the model matches very well with the experimental result. This can be attributed to the zero order hold delay of the DAC and the finite bandwidth of the measurement tool. Otherwise, it is very accurate.

![Graph](image.png)

**Figure 5.6:** Open loop frequency response from control effort $u [V]$, to measured spindle speed $\omega [\text{RPM}]$. The model (red) matches very well with the experimental result (blue). The noisy result at high frequency is due to insufficient output amplitude.

### 5.3.3 Average speed feedback synthesis

To measure the average spindle speed, two feedback filters were embedded in the feedback loop. The first filter, $F(s)$, operates on the speed, $\omega$. It is a four pole filter of the form:

$$F(s) = \left( \frac{\omega_{bw}^2}{s^2 + \omega_{bw}s + \omega_{bw}^2} \right)^2$$

with $\omega_{bw}$ equal to 1250 rad/s (200 Hz). This filter attenuates the effects of quantization error and has one other purpose that will be discussed in section 5.6.3. It is implemented now so that a fair comparison between the sensorless algorithm and the encoder can be made.
The second filter, \( A(z^{-1}) \), operates on the output of \( F(s) \) and is the filter which calculates the average speed, and thus sets the measurement bandwidth. It has the form:

\[
A(z^{-1}) = \frac{1}{N_r} \times \frac{1-z^{-N_r}}{1-z^{-1}}
\]  

(5.4)

The averaging period is the product of the sampling period \( T_s \) and the number of delay taps, \( N_r \). It also has repeating zeros at frequencies:

\[
f_{\text{notch}} = \frac{1}{T_s N_r}, \frac{2}{T_s N_r}, \ldots
\]  

(5.5)

The frequency response of the filter is shown in Figure 5.7 for \( N_r = 205 \) and \( T_s = 0.0001 \text{s} \). For a given sampling rate, \( N_r \) is selected to place the first notch at the intended operating spindle speed. For this example, 2925 RPM was chosen. This is done to eliminate the encoder error which occurs at harmonics of the spindle speed.

![Figure 5.7: Frequency response of filter \( A(z^{-1}) \). The parameters are \( N_r = 205 \) and \( T_s = 0.0001 \text{s} \). Notches occur at the fundamental frequency of 48 Hz, corresponding to 2925 RPM and at each harmonic up to the Nyquist frequency.](image)
5.3.4 Rotary speed control

A block diagram of the rotary speed control loop is shown in Figure 5.8. The output of $A(z^{-1})$ is the average speed $\omega_{\text{avg}}$ and this is the variable that will be controlled.

![Rotary control block diagram](Image)

**Figure 5.8:** Rotary control block diagram

Since the plant, from $u$ to $\omega$ is a simple -20 dB/dec system, an equally simple PI controller can be used to regulate the speed reference $\omega_{\text{ref}}$. Using a parallel implementation to limit windup, the controller is designed as:

$$C_r(s) = k_p \left(1 + \frac{k_{\text{int}}}{s}\right)$$

(5.6)

A 10 Hz unity gain crossover frequency for the negative loop transmission was chosen, comfortably below the bandwidth of both feedback filters. The integrator corner frequency is set to 1 Hz. So $k_{\text{int}} = 2\pi$ and $k_p = 0.061$ to reduce the loop gain to unity at 10 Hz. The negative of the loop transmission for the controller, plant, and feedback filters is shown in Figure 5.9. The phase margin at 10 Hz is 32°.
Figure 5.9: Negative of loop transmission for rotary control loop.

5.4 Sensorless feedback hardware

Additional hardware is required for the implementation of the sensorless feedback algorithm. The system is designed around a field-programmable gate array (FPGA) chip since it offers a high I/O count, high clock speeds, and a lot of flexibility in terms of the different applications that it can be used for. The FPGA chip executes code written in VHDL, a hardware description language. An illustration showing how this hardware interacts with other devices in the system is shown in Figure 5.10.
Figure 5.10: Block diagram of rotary motion hardware layout with sensorless feedback. The custom PCB which includes an FPGA chip, takes the place of the rotary encoder.

The inputs to the sensorless feedback hardware are the three HESs which are pulled up through 10 kΩ resistors, filtered, and then buffered. Then, the signals are fed to FPGA input pins, where they are processed by the algorithm. The FPGA chip has two sets of outputs. The first is the set of HES signals which are fed to optical isolation integrated circuits (IC). These are outputted to the drive for commutating the motor phases. This is done to avoid noise issues produced by coupling with the switching amplifier. The second set of outputs is the A, B, and index quadrature signals which are synthesized by the sensorless feedback algorithm. These are sent to a differential driver which output the differential quadrature signals to dSPACE.

Undergraduate students Yoyo Au and Henry Tsin made an expansion card which connects to a Project Spartan 3 development board [44] for initial testing. This model, however, did not include the optical isolation feature. I then implemented a more permanent solution on a custom PCB shown in Figure 5.11.
Figure 5.11: Custom PCB which implements sensorless speed feedback algorithm. The PCB is of the four layer variety and is approximately 220 mm by 160 mm. The left side of the board is dedicated to digital electronics for the rotary speed feedback algorithm. The right side of the board is assigned to analog electronics related to the axial sensing which will be discussed in Chapter 6. The PCB is enclosed in an aluminum housing made by Hammond [45].

The PCB is a four layer design and is approximately 220 mm by 160 mm. Digital circuitry for the sensorless feedback algorithm are implemented on the left side of the board, while the far right is reserved for sensitive analog electronics which are used for axial position sensing. The latter will be discussed in Chapter 6. The FPGA chip is a Xilinx XC3S1000-4FTG256I. An onboard crystal oscillator produces a 150 MHz clock which latches flip flops in the FPGA. The board can handle systems with up to ten HESs. It also has two differential encoder output ports.
5.5 Hall Effect sensor position feedback

Now that the spindle’s dynamics are understood, the sensorless speed measurement algorithm can be developed. The algorithm contains three subroutines, and this section describes the first.

5.5.1 Principle

In brushless DC motor designs, it is very common to use HES to detect rotor position in order to commutate the motor phases. The most simplistic setup is shown in Figure 5.12 which shows a two pole permanent magnet rotor and three HESs placed around the circumference. Each sensor turns on under the south pole and off under the north pole. This produces six edges per revolution.

**Figure 5.12:** Left: end view of two pole brushless motor, with three Hall Effect sensors. Right: the HES output sequence for one revolution.
Figure 5.13 shows a more general case where a $P$ pole motor has $n$ HESs distributed over one set of adjacent pole pairs. Since the HES sequence repeats for each pole pair, there are $M$ counts per revolution, where $M$ is:

$$M = \frac{P}{2} \times n \times 2$$  \hspace{1cm} (5.7)

The reasoning is that there are $P/2$ pole pairs, and therefore $P/2$ sequences of $n$ HESs, each with two edges. If $M$ is divisible by four, then it can be assembled into a set of quadrature signals. In this specific case, the permanent magnet rotor has eight poles, and three HESs, so it generates 24 counts per revolution. Since 24 is divisible by four, it can be decoded into quadrature as shown in Figure 5.14.
5.5.2 HES decoding and quadrature encoding algorithm

A flow chart showing the algorithm which decodes the HES sequence and encodes it into quadrature is shown in Figure 5.15. There are three main processes in this particular algorithm: digital filtering, HES decoding, and quadrature encoding. The input-output relationship is serial, but the processes are executed in parallel.

Figure 5.14: HES and quadrature sequences for eight pole, three HES motor. The HES sequence repeats four times per revolution generating twenty four counts per revolution.

Figure 5.15: Flowchart of HES decoding to quadrature encoding algorithm. This figure describes the input-output relationship between the various processes. VHDL coding allows each process to be executed in parallel.
The digital filters ensure that noise on the HES from the switching amplifier do not affect the logic level detected by the FPGA. The filter latches fifteen consecutive samples at 150 MHz. If all fifteen bits have the same logic value, then this value is assigned to the filter output.

The next step to be described is the HES decoding routine. In this process, there are three subtasks to complete. A counter, $k_{\text{decoder}}$, running from 0 to 25 (26 counts) at the FPGA clock speed controls the process. The first task is to latch in the incoming three incoming HES signals, while storing the previous value, when $k_{\text{decoder}} = 0$. Next, when $k_{\text{decoder}} = 1$, the algorithm enters into a lookup table which contains all possible combinations of current and previous HES signals that could lead to a change in rotary position. This information is shown in Table 5.2.

Table 5.2: HES counting sequence lookup table.

<table>
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<tr>
<th></th>
<th>HES 1</th>
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<th>HES 2</th>
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<th>HES 3</th>
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<tbody>
<tr>
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<td>Previous</td>
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<td>Previous</td>
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<td>Current</td>
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<td>H</td>
</tr>
</tbody>
</table>

Once it has been determined whether to increment or decrement the rotary position, an increment flag or a decrement flag is raised accordingly. The final step in this subroutine is to update the position tracker which occurs when the $k_{\text{decoder}} = 2$. If the increment flag is set, then the position counter is incremented. In a similar fashion, it is decremented when the decrement flag is set. If both flags remain unset at the end of the cycle, the position tracker is not altered.
Finally, care must be taken to ensure that the position counter loops appropriately when incrementing past 23 ($M-1$) and decrements past zero. Although the decoding process is complete when $k_{decoder} = 2$, it continues counting to 25 to match the length of the quadrature decoding process.

The third and final step in this process is the quadrature construction. This process works in parallel with the HES decoder and has its own internal process counter, $k_{quad}$ which also counts from 0 to 25 so that it remains synchronous with the HES encoding process. Its single input is the angular position produced by the HES decoder. There is one important point to consider when constructing this quadrature signal. It cannot be assumed that the position counter will increase incrementally. For example, it is possible that the counting sequence could be 0, 1, 3, 6, etc, instead of 0, 1, 2, 3, etc. This is especially true once the interpolation algorithm presented in section 5.6 is implemented. However, the quadrature output sequence must still represent each count. With this in mind, two additional variables are used. The first is the past position which stores the previous value outputted from the HES decoder. The second is an intermediate position which counts up to the actual position as determined by the HES decoder. The intermediate position is initialized with the value stored in the past position.

With this in mind, it is easier to understand the quadrature construction process. The first step is to latch the current and past positions, which happens when $k_{quad} = 0$. When $k_{quad} = 1$, the direction of spindle rotation is judged to be positive or negative by comparing the current and past position values. When $k_{quad} = 2$, the current position as calculated by the HES decoder, is compared against the intermediate position. If they are equal, then the correct quadrature signal has already been updated to the correct position. If not, the algorithm must continue and an increment or decrement flag is raised accordingly. In the next step, when $k_{quad} = 3$, the A and B
quadrature signals are assigned based on the current estimated position, the estimated position during the previous cycle, and the direction of rotation. This is summarized in Table 5.3. Then the estimated position is incremented or decremented appropriately. The final step in the process triggers the index once per revolution. This is done when $k_{quad} = 25$.

Table 5.3: Summary of quadrature signal output sequence.

<table>
<thead>
<tr>
<th>Previous</th>
<th>New</th>
<th>Direction of rotation</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>0</td>
<td>0</td>
<td>1</td>
</tr>
<tr>
<td>1</td>
<td>0</td>
<td>1</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
<td>0</td>
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<td>0</td>
<td>1</td>
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<td>0</td>
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<td>0</td>
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<tr>
<td>0</td>
<td>1</td>
<td>1</td>
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<tr>
<td>1</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>1</td>
<td>0</td>
<td>0</td>
</tr>
</tbody>
</table>

5.5.3 Results

Having developed a preliminary sensorless feedback algorithm, it can now be tested. Since the plant is identical, the same controller can be used providing $k_e$ is adjusted to 6 lines/rev. To make a comparison between the optical encoder and the sensorless algorithm, data is recorded for two separate cases. The first is when the encoder measurement is used as the controlled feedback, and the second is when the output of the sensorless algorithm is used as the feedback. The result is shown in Figure 5.16. Several observations can be made from this data. First, the rotational speed can definitely be measured and it can be regulated in a stable manner. The quantization noise, however, is quite large and is unacceptable compared to the rotary encoder. To improve this, the angular resolution must be improved.
Figure 5.16: Comparison between 24 counts (6 lines) per rev HES derived speed feedback and 4000 counts (1000 lines) per rev optical encoder derived speed feedback. The noise produced by the sensorless algorithm has a fundamental harmonic of the spindle speed. It is significantly larger than the noise resulting from the encoder measurement.

5.6 Temporal interpolation for increased counts per revolution

In this section an addition to the HES decoding routine is introduced which aims to increase the angular resolution of the sensorless algorithm. The key is to assume that the shaft is spinning at a constant speed over short temporal lengths. In this way, interpolation between the HES signals can be used to generate significantly higher angular resolution. The interpolation algorithm can simply be inserted between the HES decoding and quadrature encoding.

5.6.1 Interpolation principle

Since this is a constant speed application, the angular resolution can be increased through interpolation. The interpolation principle relies on the assumption that the spindle speed is constant. In one respect, this is true since the stiffness of an inertial element (such as the rotating
shaft) tends to infinity at high frequency. As long as the speed is measured at short time scales, the assumption will hold. Another important fact is that the spindle speed is regulated to a constant speed by a controller, thus keeping the speed constant over long time scales.

The governing equation, therefore, is very simple. Shown in Figure 5.17, the angular distance, or linewidth $L_i$, between adjacent HES edges divided by the time between successive edge detection, $T_i$, is constant. Based on this, an intermediate linewidth, $L_{int}$, can be defined if the corresponding timing information, $T_{int}$, is known:

$$L_{int} = T_{int} \frac{L_i}{T_i}$$  \hspace{1cm} (5.8)

![Figure 5.17: Interpolation principle.](image)

5.6.2 Interpolation algorithm

To perform interpolation, the angular distance between each count, the linewidhts, need to be measured and stored. The time between HES edges needs to be calculated and stored as well. The original algorithm structure can be retained, but now it is necessary to add the
interpolation algorithm between the HES decoding and quadrature encoding routines. In terms of execution order, the interpolation process executes in series with HES decoding. The input-output relationship is shown in Figure 5.18.

![Flowchart of HES interpolation algorithm.](image)

**Figure 5.18:** Flowchart of HES interpolation algorithm.

Again, a counter is used to control the process flow. Since HES decoding and interpolation are executed serially, $k_{decoder}$ is lengthened to fifty. In order to keep the quadrature encoding in sync, its cycle counter is also extended to fifty. The first step in the interpolation process is to start a timer which begins once a HES edge is detected and continues until the next HES edge is detected. This value is stored in a table. Here, an absolute minimum is placed on the measurable speed:

$$n_{\text{min}} = \frac{60 \times f_{CLK}}{50 \times M \times 2^b}$$  \hspace{1cm} (5.9)$$

where $b$ represents the number of bits in the timer, and $f_{CLK}$ is the clock speed timing the entire program. In this implementation, a 16-bit counter and 150 MHz clock are used, so the lowest speed that can be measured is 114 RPM. This step only consumes a single tick $k_{decoder}$ counter.

Next, a predefined linewidth map, which is a table containing the distance between each HES edge, is loaded. There are $M$ rows, each corresponding to a HES count. The first entry is zero. The second, third and $M^{th}$ entries are $K$, $2K$ and $(M-1)K$. Here, $K$ represents the number of interpolated counts that are generated between HES edges. At this point, the algorithm assumes
that the HES edges trigger precisely every spatial 15° increment. Next, the algorithm enters into these two tables and retrieves the appropriate data: the previous linewidth \( L_i \) and corresponding time \( T_i \). \( T_{int} \) is taken from the timer which calculates \( T_i \). Next, the interpolation arithmetic is performed. Since VHDL only has addition and subtraction as native operations, two third party intellectual property (IP) [46, 47] cores are employed to compute \( L_{int} \). (N.B.: IP cores are neatly packaged blocks of code which perform complex operations). One IP core performs the multiplication of \( L_i \) and \( T_{int} \), and another performs the division between the resulting product and \( T_i \).

A long wait period follows the start of the arithmetic because the operations have internal pipeline delays [46:3, 47:2]. Finally, the interpolation result is added or subtracted to (depending on the direction of motion) the coarse measurement found in the linewidth map. The final resolution is \( M \times K \) counts/rev. With \( K = 256 \), the result is 6144 counts/rev (1536 lines/rev) which provides a resolution 50% higher than the optical encoder. The goal of significantly increasing the angular resolution of the feedback has been achieved.

However, the number of interpolations cannot be increased indefinitely because there is a fundamental limitation with regards to maximum measurable speed. This limitation can be calculated as follows:

\[
n_{\text{max}} = \frac{60 \times f_{\text{CLK}}}{50 \times M \times K} \quad (5.10)
\]

where \( n_{\text{max}} \) is the top speed in RPM, and 50 is the length of the counter, \( k_{\text{decoder}} \). In this case, a 150 MHz clock, 24 counts/rev from the HESs, and 256 interpolations, give a top measurable speed of approximately 30,000 RPM. This is acceptable for low to medium speed applications such as grinding or turning. However, the Xilinx Spartan 3 FPGA chosen for this application uses a low cost technology. Virtex 5, the fastest mature FPGA technology available on the
market can operate at clock speeds of 550 MHz\(^5\). So, in theory, the algorithm will work at speeds over 100,000 RPM which is high enough for all but the highest speed drilling processes. The practicality of using this algorithm at such high speed needs to be experimentally verified.

5.6.3 Experimental results

Now that the resolution has been improved, a cleaner frequency response can be obtained. A frequency response from \(u\) to \(\omega\) is shown in Figure 5.19 with \(k_e = 1536\) lines/rev. A result very similar if not identical to Figure 5.6 would be expected. However, this is not the case. At low frequency there is a rigid body mode corresponding to the shaft inertia, but at 30 Hz there is a very lightly damped zero which is followed by a +20 dB/dec slope. The reason for this is that there is a very slight interference fit between the motor stator and the spindle housing, and no set screws, clamping mechanisms or epoxies to increase the stiffness of the joint. A diagram illustrating this situation is shown in Figure 5.20.

\(^5\) http://www.xilinx.com/products/virtex5/lx.htm
Figure 5.19: Frequency response from $u$ to $\omega$. Notch near 30 Hz is caused by flexibility between stator and spindle housing. The algorithm, above 30 Hz, measures the HES motion.

So, when a torque, $T$, is applied to the rotor inertia $J_{\text{shaft}}$, an equal and opposite force is applied to the stator inertia, $J_{\text{stator}}$. Since the HESs are mounted to the stator, they actually measure the relative motion between the stator and the rotor. This can be expressed as:

$$\frac{\dot{\theta}_1(s) - \dot{\theta}_2(s)}{T(s)} = s \left( \frac{1}{J_{\text{shaft}}s^2} + \frac{1}{k_\theta + J_{\text{stator}}s^2} \right)$$

(5.11)
However, at low frequency the joint stiffness is much larger than the inertial stiffness of $J_{states} s^2$ leaving:

$$\frac{\dot{\theta}_1(s) - \dot{\theta}_2(s)}{T(s)} = s \left( \frac{1}{J_{shaft} s^2} + \frac{1}{k_\theta} \right)$$

(5.12)

This can be simplified to:

$$\frac{\theta_1(s)}{T(s)} = \frac{k_\theta + J_{shaft} s^2}{k_\theta J_{shaft} s}$$

(5.13)

Using the shaft inertia from Table 5.1 and the 30 Hz anti-resonance gives $k_\theta$ equal to 20.7 Nm/rad. If (5.13) is substituted for $P_r(s)$, then a direct comparison can be made between the model and experimental result (Figure 5.21). The agreement is very good.

![Figure 5.21: Comparison between experimental result and model of sensorless feedback algorithm incorporating HES relative motion measurement.](image)

In section 5.3.3, while discussing the feedback filters, the reader was asked to accept the details of the filter design without explanation. Now, the rationale for the use of the fourth-order
filter with cutoff of 100 Hz is clear. The response above 30 Hz must be significantly attenuated to eliminate the effects of the erroneous motion measurement. Since this filter has been in use all along, no changes to the PI controller are necessary. The resulting speed regulation noise at 5000 RPM is shown in Figure 5.22 compared to that achieved with the optical encoder feedback. The noise level is significantly reduced due to the increased resolution, but the optical encoder still produces a better result. In the next section an “on the fly” calibration process is used to improve the result even further.

![Figure 5.22: Comparison between interpolated (256x) HES derived speed feedback and optical encoder derived speed feedback at 5000 RPM. The interpolation reduces the harmonic error in the feedback, but the encoder still provides a more precise signal.](image)

5.7 Real time calibration of HES edges

In the previous discussion on interpolation, two assumptions were made. The first was that the spindle is always at a constant speed. This is a reasonable and valid assumption. The
second was that the HES edges occur at equally spaced intervals. Using the intuition of a
precision engineer, it is fairly obvious that this is an invalid assumption. It is impossible to place
the HES on the stator very accurately. It is also impossible to have the magnets on the rotor equal
in magnetic strength, equal in size, and equal in placement. Additionally, the HES themselves
have some limitations which could cause them to trigger slightly differently from each other.
Some form of calibration is needed so that the precise location of the HES edges can be known.
A self-calibration technique is best so that the linewidth maps can be constantly updated. In this
section, an “on-the-fly” self-calibration technique which spatially locates the HES edges in the
spatial domain and makes the precision of the sensorless speed measurement equivalent to that
produced by an optical encoder in constant speed applications.

5.7.1 “On the fly” self-calibration principle

The self-calibration principle is fairly simple. First, the constant speed condition must be
confirmed. Then, using this fact, the ratio of the time at any given point in a rotation, \( t \), to time
per revolution, \( T_{pr} \), is equal to the ratio of the swept angle, \( \theta \), to one complete revolution, \( 2\pi \).

\[
\frac{\theta}{2\pi} = \frac{t}{T_{pr}} \tag{5.14}
\]

This can be extended to this specific application. The ratio of the total linewidth up to the \( i \)th HES
\( (L_i^{\mu}) \) to total counts \( (M \times K) \) is equal to the ratio of partial time per revolution up to the \( i \)th HES
edge \( (\Sigma T_i) \) to time per revolution \( (T_{pr}) \).

\[
\frac{L_i^{\mu}}{M \times K} = \frac{\Sigma T_i}{T_{pr}} \tag{5.15}
\]

Then, solving for \( L_i^{\mu} \):
By repeating this procedure for each HES edge, the linewidth table can be accurately filled.

5.7.2 “On the fly” calibration algorithm

The “on the fly” calibration algorithm operates independently of other processes. Figure 5.23 shows the input-output relationship between the interpolation algorithm and the newly-implemented online calibration algorithm. The calibration algorithm generates an “adapted” linewidth map which is used in place of the default map relied upon in the interpolation process. The map is updated whenever the spindle is judged to be in a steady state.

![Flowchart of interpolation algorithm once online calibration has been integrated.](image)

Figure 5.23: Flowchart of interpolation algorithm once online calibration has been integrated.

The adaptation routine takes the form of the state machine shown in Figure 5.24. It begins in the wait state, where the routine pauses for some time before latching the table storing the time between HES edges, which was generated in the interpolation algorithm. From here, the algorithm advances one state and checks that the spindle is at constant speed by calculating $T_{pr}$ for two cycles and comparing. $T_{pr}$ is computed by summing the elements in the aforementioned time table. To account for the quantization error introduced by the timer, a 0.05% variation in $T_{pr}$ is permitted from one cycle to the next. If the speed is judged to be variable, then the algorithm
returns to the wait state and starts over. If the speed is constant, then it proceeds to check if the speed is reasonable. For this particular application, 300 to 12,000 RPM represents a reasonable range. Again, it returns to the wait state if the speed is outside this range, or moves forward if it is acceptable.

![Flow chart](image)

**Figure 5.24:** Flow chart for “on the fly” calibration state machine. Rectangular blocks are states and elliptical blocks are conditions for proceeding to another state.

Now, the routine enters the first of two calibration states. In the first state, the partial turn per revolution, $\Sigma T_i$, is calculated. Then the algorithm shifts to the second calibration state where it pauses while (5.16) is computed by an additional set of multiplication and division IP cores. A check is carried out to make sure the result is within ±10% of a nominal HES linewidth. If the
result is out of tolerance, the algorithm jumps back to the wait state, ignoring the calculation. If it is within tolerance, it loops back to the first calibration state to begin the process of calibrating the next linewidth. When the linewidth table has been filled, all 24 HES edges have been correctly located, and so the routine can return to the wait state.

5.7.3 Experimental result

When both interpolation and adaptation of the HES signals are implemented, the resulting speed regulation is quite impressive. In fact, it is just as good as the benchmark optical encoder. A comparison between the two is shown in Figure 5.25.

![Figure 5.25: Comparison between interpolated (256x) with adapted HES derived speed feedback signal and optical encoder derived speed feedback. The precision of the sensorless speed feedback algorithm with interpolation and “on the fly” calibration is as good as that produced by the encoder’s speed measurement.](image)
5.8 Rotary control of rotary-axial spindle

This section discusses the rotary control of the rotary-axial spindle. For this spindle, the axial control loop must be operating and stable before the shaft can be rotated due to the open loop bi-stable nature of the thrust bearing. This will be discussed in detail in Chapter 7. The rotary control, though, is nearly identical to that of the Precitech spindle. There are only two possible differences. First, there is the difference in shaft inertia, but this is easy to compensate for with the proportional gain. Second, there is a possible difference in the stiffness of connection between the motor stator and spindle housing, which may affect the motion control bandwidth.

Using the SolidWorks mass properties calculator, the inertia of the spindle shaft is computed to be 0.000468 kg-m$^2$. There is no simple way to go about estimating the joint stiffness, so a swept sine test is performed to identify the new system. A Bode plot of the measurement from $u$ to $\omega$ is shown in Figure 5.26. One noticeable difference is that the frequency at which the zero occurs is actually higher, at 40 Hz, than it was on the Precitech spindle (30 Hz). Again, the joint stiffness can be calculated using this frequency and the spindle shaft inertia. The resulting $k_\theta$ is 29.5 Nm/rad. The modeled frequency response from $u$ to $\omega$ is also plotted in Figure 5.26. Again, the model matches quite well with the experimental response.

This slight shift in zero frequency is not significant enough to change the feedback filter parameters or to alter the crossover frequency. To maintain the crossover frequency at 10 Hz, the proportional gain is decreased to 0.0418 to compensate for the decrease in shaft inertia. The integrator corner frequency remains at 1 Hz. The resulting negative loop transmission, the product of every element in the loop, is shown in Figure 5.27. The phase margin at the unity gain crossover frequency is 33°. This should be sufficient for stable operation. To verify the controller
design, a 100 RPM step response was captured (Figure 5.28) about a 2000 RPM operating point. The time from initial reaction to the peak of the response, $t_p$, is 0.05 s which corresponds to a 10 Hz frequency. This matches nicely with the designed crossover frequency. With this controller regulating the speed at 5000 RPM, the error noise about the set point is only 0.170 RPM RMS. Speeds as low as 1000 RPM and as high as 8000 RPM were also tested with similar results.

![Figure 5.26: Comparison between model of rotary-axial spindle rotary control plant and experimental frequency response. Low excitation amplitude at high frequencies result in a less than clean identification.](image-url)
**Figure 5.27:** Negative of the loop transmission of rotary-axial spindle rotary control loop. The crossover frequency is 10 Hz with 33° phase margin.

**Figure 5.28:** 100 RPM step response. The time for a half cycle is 0.05 seconds which corresponds to a 10 Hz crossover frequency.
Figure 5.29: Speed regulation noise at 5000 RPM. The RMS error is 0.170 RPM.

5.9 Summary

Since this is a spindle application, precision angular motion is not necessary. Only good average speed control is required. In average speed measurements, there is a fundamental tradeoff between speed regulation noise and measurement bandwidth. To simultaneously achieve a high bandwidth and low noise measurement, a high resolution angular position sensor must be used.

Most applications would use an optical encoder with several thousand counts per revolution to provide position feedback. However, an encoder is not acceptable for use in the rotary-axial spindle due to the incompatibility between the spindle’s 1 mm axial motion and the 100 μm constant air gap requirement for an optical encoder.

To solve this dilemma, a sensorless position feedback system was developed by Dr. Lu specifically for this application. It uses the 24 counts per revolution generated by the
combination of the eight pole permanent magnet rotor and three HES to generate a quadrature signal. Since this results in only 15° of angular resolution, the resolution was improved further by implementing time based interpolation between the HES signals. In total, the interpolation increases the resolution to 6144 counts/rev. However, this is still not optimum as the algorithm assumes that the three HES change their digital output at precise 15° spatial intervals. This is not the case due to manufacturing tolerances and non-idealities of the HES themselves. So, an “on the fly” calibration occurs whereby the precise location of each HES signal edge is identified.

In order to test the algorithm, a donor Precitech SP75 spindle with a 4000 count/rev encoder was used. First, it was modeled in order to identify all the parameters in the system. Then, the average spindle speed was calculated using the encoder to measure the motion, and using averaging and low pass filters to synthesize the average speed feedback.

The speed regulation noise of the 24 count/rev HES decoding algorithm was significantly worse than the result achieved with the encoder based feedback. Once the interpolation algorithm was implemented on top of the HES decoding routine, the results improved dramatically, and it was now at least competitive with the encoder. When the “on the fly” calibration routine was implemented, it produced a result which was just as good as the 4000 count/rev encoder.

Finally, the sensorless rotary motion feedback algorithm was tested on the rotary-axial spindle. The proportional gain of the PI controller was modified slightly to compensate for the difference in the inertia between the two spindle shafts. The spindle was successfully run at speeds up to 8000 RPM.
Chapter 6
Axial Motion Sensing

This chapter discusses the axial sensing method for the rotary-axial spindle. A capacitive probe sensor measures the motion at the target screw surface and provides an analog signal which can be used for feedback. However, when the feedback is compensated with a digital controller, the analog to digital (A/D) conversion process introduces quantization error to the measurement which often buries the signal of interest. Thus, the analog to digital converter (ADC) resolution becomes a limiting factor in high precision applications. This chapter summarizes the modeling, development and experimental testing of a patent pending [4] method of increasing ADC resolution without any of the usual drawbacks.

For any given ADC, two quantities generally define its performance: resolution and conversion delay. The resolution, if larger than the analog noise on the sensor signal, limits precision. The conversion delay, which adds frequency dependent phase lag to the feedback signal, can impose a limit on achievable motion control bandwidth. There is a tradeoff between these two properties. Modern technology allows for sixteen bit converters with sampling rates into the MHz range, and sub microsecond delays [48]. However, some select applications require a higher bit count. The problem is that above 16 bits, commercially available ADCs have sampling rates that are limited to tens of kHz due to long digital filters which must be used to attenuate noise inherent in the conversion process. So, for systems with kHz range bandwidth, 16 bits is the current limit.
To jump this hurdle, Dr. Lu and I have invented a technology called Nanozoom. In conventional digital control systems, the subtraction of the feedback from the reference takes place in the digital domain. This invention, however, performs this subtraction in the analog domain. So, if the system is stable, this error is always small. The next step is to amplify this analog error signal before passing it to a high speed ADC. This enhances the resolution by the amplification factor, without introducing additional delay from the ADC. Practically speaking, Nanozoom allows a system to have both high resolution and low conversion delay, which helps achieve higher control bandwidth.

The remaining sections of this chapter show how the spindle shaft’s axial position is sensed and converted to the analog domain and the obstacles that this presents. Next, section 6.2, looks at other attempts to increase ADC resolution and the benefits and shortcomings of each approach. Section 6.3 describes the “Nanozoom” principle and in section 6.4 explains how it is implemented on a custom PCB. Section 6.5 explains the communication protocols. Section 6.6 presents an electrical noise model, and compares this model to an experimental measurement. The chapter is closed with a summary.

6.1 Conventional sensing and analog to digital conversion

The axial sensing requirements that have been set for the system are 1 mm range with sub 10 nm resolution. This corresponds to a dynamic range of over 100 dB, which is quite challenging to achieve. There are two options to begin with: capacitive probe technology and laser interferometers. The ADE 6503 capacitive probe position sensor and 6810 gauge module produce 5.5 nm RMS noise over 1 mm range with 20 kHz bandwidth. A good laser interferometer system can easily surpass the probe in terms of resolution, range and bandwidth,
but can cost approximately ten times more than the ADE setup. Based on this, the capacitive probe sensor is chosen.

Figure 6.1 shows the target screw which fastens into mating threads on the end of the spindle extension shaft. A grinding operation was performed on the target screw face to achieve good perpendicularity between the axis of the thread body and the face. This was done with the intent of decoupling the axial sensing from the rotary motion. When the probe assembly is installed on the rear housing, it looks axially at the target screw surface.

![Figure 6.1: Axial sensing solution.](image)
The probe output is an analog signal which needs to be digitized for use in a digital control system. In the general case, the analog signal from the sensor is fed to a $b$-bit ADC, as shown in Figure 6.2.

![Figure 6.2: Block diagram showing the conventional feedback path of an electromechanical system.](image)

If the sensor range is $R$, then the resolution or quantization interval $q$ is:

$$q = \frac{R}{2^b} \quad (6.1)$$

In this case, with $R = 1$ mm, and $b = 16$, the resulting resolution is 15.3 nm. This is confirmed by the experimentally collected data shown in Figure 6.3.

For some ultra precision applications such as silicon wafer grinding or mirror polishing, it is desirable to have sub 10 nanometer resolution. To attain this, 18 bit or higher ADCs are required. Since an ADC with this bit count is generally unavailable in a packaged digital controller, the goal is to increase the resolution of the existing 16 bit device.
Figure 6.3: Data capture demonstrating 15.3 nm resolution when ±10 V range ADC is scaled to represent ±0.5 mm range. Here, the ADC input is grounded through a 50 Ω terminator.

6.2 Increasing ADC resolution

There are a number of ways in which an analog sensor signal can be digitized at a higher resolution. The first is to ensure that the sampling rate is fast enough. Figure 6.4 shows a comparison between two identical ramp signals which are sampled at different rates. Nyquist’s sampling theorem states that a signal should be sampled at a frequency of at least twice the signal bandwidth to avoid aliasing. However, this does not necessarily utilize the full resolution unless a much faster sampling rate is used. Faster sampling, as shown in Figure 6.4b, ensures that each incremental bit is captured. In other words, this technique does not increase the resolution, but only maximizes the available resolution. A good rule of thumb is to sample five to ten times faster than the sensor bandwidth.
Figure 6.4: Comparison between a ramp signal sampled at a certain sampling rate a), and a ramp signal sampled at a rate four times faster b). The analog signal is shown in blue, and the digital signals in red. Faster sampling ensures the available ADC resolution is maximized.

Another technique for achieving higher resolution is oversampling and averaging. The first step in this process is to sample the analog signal with a high speed but low resolution ADC at a significantly higher rate than the control loop execution rate. The ratio between the two rates, must be an integer, and it is called the oversampling ratio (OSR). A moving average filter of length OSR is applied to the signal, and is then re-sampled at the lower rate. The “effective” resolution is improved by a factor of the OSR, thus improving quantization noise. This process, however, has three shortcomings. It consumes additional computational and memory resources, it imposes a limit on the achievable system bandwidth due to the filter phase lag, and high frequency motion information is lost due to filter attenuation. Overschie et al [49] developed a custom circuit board to perform the analog to digital conversion and sampling processes outside of the digital controller, thus sparing processor resources. This addresses the first issue, but not the second and third.
Sigma delta-modulation is another over-sampling analog to digital conversion technique. However, like oversampling and averaging, some quantization error remains. One major drawback to this approach is the requirement of a digital filter on the output. These FIR filters typically contain hundreds of taps creating a significant pipeline delay. A good example is the Analog Devices AD7760 24-bit ADC [50]. One of its features allows the user to select various filtering schemes to achieve the desired noise level. Although it can achieve microsecond delays without any filtering, light to moderate filtering introduces delays on the order of tens of microseconds. Reducing the length of the filter to decrease delay generally results in a dynamic range comparable to a good high speed 16-bit ADC. Additionally, a rather complicated interface between the AD7760 and the digital I/O on our dSPACE DS1103 would need to be designed.

One method which improves the resolution of an ADC without increasing its delay is to amplify the analog signal before conversion. By increasing the signal amplitude, the quantization intervals are reduced providing higher resolution. However, there is one major downside; the sensor range is reduced by twice the amplification factor. This places a considerable limit on the motion range of the device in question. Guery et al proposed using a programmable variable gain amplifier to ensure the amplifier gain would reduce as the signal level increased [51]. This fixes the input saturation problem, but the result is variable resolution across the range of motion-high resolution for small signals and low resolution for signals closer to the input limits.

All of these solutions come with drawbacks which are more or less incompatible with our application. The next section describes the particular invention which overcomes these shortcomings.
6.3 Nanozoom principle

This section explains the principles of operation of the invention, *Nanozoom*. Its aim is threefold:

1. To increase ADC resolution.
2. To maintain the high resolution over the entire sensor range.
3. To perform tasks 1) and 2) without increasing the A/D conversion delay.

Figure 6.5 shows a schematic of how the technology is most typically used in a feedback control system. The electromechanical system is provided with a control signal $u$ from a DAC. This signal, $u$, is generated by the digital controller, $C(z^{-1})$ which operates on the high resolution error signal, $e_{NZ}$. To synthesize $e_{NZ}$, a high resolution reference signal is generated which should be quantized at a bit count equal to or higher than the desired feedback resolution. Now in the analog domain, the sensor output signal $x$, which is a measure of the target’s position, is subtracted from the reference $x_{ra}$ to produce the analog error signal, $e_a$. This accomplishes the second task. Next, $e_a$ is amplified by $K_z$, the zooming gain, which is the key to achieving the first aim. The amplification result, $e_{ak}$ is then converted back into the digital domain, by a high speed $b$-bit ADC. This accomplishes the third task. Finally, the digital representation of $e_{ak}$ is rescaled by the reciprocal of $K_z$ to recover the error which has resolution:

$$b_{NZ} = b + \log_2 K_z$$  \hspace{1cm} (6.2)

Accordingly, the quantization interval is:

$$q_{NZ} = \frac{R}{2^b K_z}$$  \hspace{1cm} (6.3)

where $R$ is measurement range.
Figure 6.5: Schematic representation of Nanozoom invention and its interaction with a generic electromechanical system. In conventional digital control systems, the control error computation is performed in the digital domain. Moving this operation to the analog domain and amplifying it improves the resolution $K_z$ times, ensures high resolution over the entire sensor range, and does not add any additional delay to the system.

Now, a description of how each of these three aims is achieved will be given, beginning with the first item: the increase in ADC resolution. Signal amplification makes more effective use of the available ADC bits as demonstrated in Figure 6.6. The application of the gain can also be seen as decreasing the quantization interval. This has a more powerful effect when the signal of interest is smaller than 1 least significant bit (LSB) of the ADC. The key to maintaining this high resolution over the entire range of motion is to move the error subtraction to the analog domain. The fundamental principle here is that in a stable system, the error is regulated very close to zero. It is right to assume that the error will always be small in steady state, so when the signal is amplified it will never reach the input limits of the ADC. Therefore, no matter where the target is in its range of motion, the error can always be viewed in high resolution. Lastly, by using a high speed device with 16-bits or lower to perform the A/D conversion, higher resolution...
is achieved for the delay cost of the lower resolution process. There is but one small drawback. The amplification process limits the error range:

\[ R_{NZ} = \pm \frac{R}{2K_z} \quad (6.4) \]

This is only an issue in transient states, such as step changes in the position reference larger than \( R_{NZ} \). This is explored further in Chapter 7.

Figure 6.6: Discretization of a base signal and its amplified counterpart. Discretizing the amplified version makes more effective use of the ADC resolution.

6.4 Nanozoom Hardware

A custom PCB was designed to implement the Nanozoom concept. This section describes the PCB architecture and reviews some points of the circuit design.
6.4.1 Nanozoom for rotary-axial spindle

Figure 6.7 shows a block diagram illustrating the main components in the embodiment of Nanozoom that will be used for the rotary-axial spindle application. To generate the high resolution reference signal, $x_r$ is quantized and transformed into a 24 bit integer and is then transmitted via RS232, which is available on our dSPACE control box, to an FPGA chip. Upon receiving the position word, the FPGA transmits the position information to a 20 bit DAC, which outputs a low noise analog signal $x_{ra}$. The capacitive probe signal is then subtracted from $x_{ra}$, which generates the analog control error $e_a$. The subtraction result $e_a$ is scaled by a factor of eight, and is then digitized by a high speed 16 bit ADC on the dSPACE unit. According to (6.2), this should give 19 bit resolution. The ADC output is then rescaled to produce the high resolution error, $e_{NZ}$. A data capture of $e_{NZ}$ with the ADC inputs grounded is plotted in Figure 6.8. This result proves the theory behind the resolution enhancement.

![Figure 6.7: Embodiment of Nanozoom for rotary-axial spindle application.](image)
Figure 6.8: Noise on ADC with inputs grounded. The voltage is converted to nanometers and divided by 8. The resulting quantization interval is 1.9 nm which corresponds to 19 bit resolution, as expected.

6.4.2 PCB architecture and analog circuit design

The main goal of this design is to minimize noise on the analog side so that we can maximize our effective resolution. This Nanozoom implementation shares the board with the rotary sensing hardware. The FPGA has plenty of I/O to handle both applications. Figure 6.9 shows how the Nanozoom portion is laid out. Sensorless rotary motion feedback related circuitry is not shown here. The digital and analog circuits are powered by separate voltage regulators. A combination of RS232 drivers and buffers convert the RS232 signal from the dSPACE universal asynchronous receiver/transceiver (UART) to the 3.3 V level required by the FPGA and vice versa, since the communication is bidirectional. A serial peripheral interface (SPI) bus connects the FPGA to the 20 bit DAC, with a buffer in between, so that the internal registers in the DAC can be written to. The buffer, with its high impedance input, is used so that no current is sourced.
from the DAC. A Texas Instruments DAC1220E, which has 0 to 5 V output that can be quantized at 20 bit resolution was chosen. It straddles a split in the ground plane on the circuit board. The purpose of this split is to prevent noise generated by the fast switching digital circuitry from affecting the extremely sensitive analog electronics. However, there is a small break in the plane so that the analog and digital circuits are not floating relative to each other. A sequence of three op-amps perform buffering, DAC output range adjustment, error subtraction, and amplification. This circuitry is shown as “Nanozoom electronics” in Figure 6.9.

Figure 6.9: Architecture of Nanozoom portion of rotary-axial spindle PCB.
Figure 6.10: a) Analog electronics schematic. The first OP27 buffers the DAC signal. The second OP27 adjusts the output range. The third OP27 performs the error subtraction and amplification process. b) Block diagram summarizing the arithmetic implemented by the analog circuitry.

Since the analog circuit has further implications in terms of achievable noise levels, it is worth discussing here in greater detail. A schematic of the circuit is shown in Figure 6.10a. The OP27 op-amps used here are EZ models which are optimized for low noise applications. The first device, which receives its input from the 20 bit DAC, is set in a follower configuration as a unity gain buffer. A 1 kΩ resistor is placed in the feedback loop, however, on recommendations from the manufacturer [32:14]. Since the DAC’s output ranges from 0 to 5 V, it needs to be adjusted to match the ±10 V range of the capacitive probe sensor. So, the buffered DAC output is subtracted from a 2.5 V reference and amplified by four. A low pass filter with 220 kHz
bandwidth is added to attenuate high frequency noise. The final OP27 performs the error subtraction and amplification processes. Since the buffered DAC output was inverted by the scale adjustment amplifier, a simple inverting configuration can be used to subtract the sensor signal from the reference. Back to back Zener diodes are used to limit $V_{NZ}$ to ±10 V to protect dSPACE’s ADC inputs. The block diagram from the $V_{DAC}$ to $V_{NZ}$ is shown in Figure 6.10b. The unpopulated PCB is shown in Figure 6.11.

![Figure 6.11: PCB for rotary-axial spindle sensing. The Nanozoom related analog electronics are located in the upper right hand corner of the board.](image)
6.5 Nanozoom communication

This embodiment of Nanozoom requires two different communication protocols. A UART transfers data between dSPACE and the FPGA. A SPI two wire system, controlled by the FPGA, is used to program the output registers of the DAC. The following sections describe each protocol. This is summarized in Figure 6.12.

![Diagram of Nanozoom communication protocol]

**Figure 6.12**: Nanozoom communication protocol.

6.5.1 dSPACE-FPGA communication protocol

The reference position for the axial control loop, which is commanded by the user, must be communicated to the FPGA. Twenty bits of the dSPACE digital I/O could be put to use, but instead the RS232 protocol is a better choice since it will consume fewer resources and only requires three wires: read, transmit, and common. Handshaking is used to ensure that data is not corrupted. The communication protocol from the dSPACE point of view is shown in Figure 6.13. The position reference is quantized at 24 bit resolution (3 bytes), and its range is adjusted from $\pm 500 \, \mu\text{m}$ (1mm) to integers from 0 to $2^{24} - 1$. After being separated into three data bytes, the information is sandwiched between a start byte and a stop byte which are two distinct characters. At this point, the reference signal has been prepared, and the transmitter waits for a “go” signal in the form of a byte of data from the FPGA.
Figure 6.13: dSPACE send protocol. After quantizing and adjusting the range of the reference $x_r$ to integers ranging from 0 to $2^{24} - 1$ (24 bit), it is split into three bytes. This information is sandwiched between start and stop bytes. It waits for a “Go” signal from the FPGA to be sent.

The communication task from the FPGA point of view is slightly more complicated since there is no neatly packaged serial communication interface by default. However, the Spartan 3 family does have a feature called the On-chip Peripheral Bus (OPB) and there is an IP core called OPB UART Lite [52] which interfaces to it. This IP core has several layers of abstraction. For this application, the layers are unraveled until the OPB can be accessed directly through the control and status registers. These registers allow writes and reads to and from the transmitter’s first in first out (FIFO) data queues. In this way, communication can be initiated with the dSPACE UART.
At the indicated abstraction level, a finite state machine is used to interface with the OPB registers. A chart depicting the flow from state to state is shown in Figure 6.14. The process begins by clearing any data in the read FIFO queue. Next, the “go” byte is sent to dSPACE. If no data was found in the FIFO after a certain time spent waiting, the state machine returns to the beginning and clears the FIFO. If a read process was performed, the data is acquired from the FIFO. Next, a check is performed to ensure it matches the start byte-three bytes of position data-stop byte format. If there is no match, the receive FIFO is cleared and the data is discarded. If it does correspond to the correct sequence, the data is latched and is ready to be sent to the DAC via the SPI.

6.5.2 FPGA-20 bit DAC communication protocol

To operate the 20 bit DAC, its internal setup and output registers must be written to. This is easily accomplished using the FPGA chip. The general process is as follows. First, transmit the command byte which contains the register address and the number of bytes that will follow.
Then, one, two, or three bytes of data are written to the register. The data transfer sequence is shown in Figure 6.15. Each bit is sent, synchronously with the serial clock (SCLK), on the SDIO line.

Figure 6.15: SPI communication sequence [53:10]. The data bits are sent, synchronously with the serial clock, on the SDIO line. Up to four bytes are transmitted in one write process. The first is an instruction byte which contains the register address and the number of bytes to be sent. This is followed by one to three bytes which correspond to the device setup or the output voltage. The low times for the SCLK, $t_9$ and $t_{14}$ are set by the frequency at which the DAC is clocked (2.5 MHz).

Now, the information sent to the DAC can be further specified. Upon power up, the DAC must be checked to make sure it is in a good working state. The DAC provides a means for performing this action through a reset procedure which can be triggered when a specific pattern is detected on the SCLK. Next, the DAC operating mode must be specified. For example, it can run in 16 bit or 20 bit mode, and the bits written to the voltage output registers can be in straight binary or two’s complement formats. There are also three different output filter settings. In order to achieve the lowest output noise, it should run in 20 bit mode with the lowest filter cut off frequency. Finally, bits corresponding to the desired output voltage must be written into the output registers. In summary, there are three states: a reset state, a setup state, and an output state. Wait periods are also necessary before the reset state, and both before and after the setup state.

The wait periods and reset state are easy to implement in VHDL. The waits are accomplished by letting counters run up to their specified limits, then advancing to the next state.
Executing the reset procedure is simply a matter of setting the output level of the SCLK in sync with a timer according to the specifications on the datasheet [53:5].

Writing to the registers is slightly more involved than the previous tasks. A 15 bit, 50 MHz counter is started. It is helpful to know that the frequency at which each bit changes value is:

$$f_{bit} = \frac{50 \text{ MHz}}{2^n}$$

where $n$ is the bit of interest, with the LSB meaning that $n = 0$. So, the LSB changes value at 25 MHz, bit 1 changes at 12.5 MHz, and so on. The most significant byte consists of bits fifteen down to seven of this counter. Bit seven changes its value at a rate of 380 kHz. Equating this bit to SCLK during the transmission time, results in a 50% duty cycle square wave with frequency 195 kHz. It must be set to zero for $t_9$ and $t_{14}$. This takes care of the SCLK.

The next step is to handle the shifting of data bits. Referring back to Figure 6.15, there is a gap between each byte that is transmitted. One easy way to create the gaps while maintaining simplicity in the algorithm is to begin shifting the bits of the second, third, or fourth byte using a slower clock. However, synchronicity with the SCLK must be maintained. Any adjacent pair of bits in the counter will always count through the following sequence: 00, 01, 10, 11 (or in decimal: 0, 1, 2, 3). If this pair is chosen from the same process counter, then it remains in sync with the SCLK. Choosing bits fourteen and fifteen from the fifteen bit counter satisfy the requirements placed on $t_9$ and $t_{14}$. This process can be used for both the setup and output states.
6.6 Noise modeling

The goal of Nanozoom was to increase resolution for improvement quantization noise, and thus precision. However, if the noise generated is higher than the quantization noise found in the conventional solution, then the purpose of the invention is defeated. A model should be created to predict the output noise.

Noise in op-amp circuits is generally grouped into five categories [54:2]. Shot noise which is caused by undesirable current flow. Thermal disturbance of charge carriers cause thermal noise. There is flicker noise which is present in all active components. Burst noise is believed to be caused by imperfections in semiconductors. Avalanche noise, related to semiconductor phenomenon, produces current pulses which are like those generated by shot noise. Manufacturers generally lump these five types of noise into two parameters: the input noise voltage ($n_v$) and current ($n_i$) densities, which are given as RMS values in units of V/$\sqrt{\text{Hz}}$ and A/$\sqrt{\text{Hz}}$. Texas Instruments, another manufacturer of the OP27, gives a basic op-amp noise model which integrates these two noise sources with an ideal op-amp [54:10]. The schematic of this circuit is shown in Figure 6.16. Two current noise sources and one voltage noise source is included. Thermal noise from passive components is ignored in this model.

![Figure 6.16: Elementary noise model of an operational amplifier.](image-url)
To analyze the noise, each op-amp is isolated from Figure 6.10a, the principle of superposition is applied, and ideal op-amp theory to calculate the contribution from each source. Since the noise is quantified in units of $A/\sqrt{Hz}$ and $V/\sqrt{Hz}$, the current and voltage densities must be scaled by the square root of the op-amp bandwidth. This bandwidth is calculated based on the OP27’s 8 MHz gain bandwidth product and the closed loop gain of the amplifier. Figure 6.17 shows the noise model for the unity gain buffer. Figure 6.18 shows the noise model for the DAC output range adjusting op-amp. Figure 6.19 shows the noise model for the op-amp which performs the error subtraction and amplification. The contribution from each noise source in each schematic is given immediately below each figure.

\[V_{n_{buf}} = 8000 \times \sqrt{8 \text{MHz}} \times n_i [V] \quad (6.6)\]

\[n_{buf} = \sqrt{8 \text{MHz}} \times n_i [V] \quad (6.7)\]
Figure 6.18: Noise model for op-amp which range adjusts the buffered DAC output.

\[ n_{\text{scale}}^{i} = 4,000\Omega \sqrt{220\text{kHz} \times n_{i} \times 2[V]} \]  
\[ n_{\text{scale}}^{v} = 5\sqrt{220\text{kHz} \times n_{v}[V]} \]

Figure 6.19: Noise model for op-amp which performs error subtraction and amplification steps.

\[ n_{\text{NZ}}^{i} = 40,000\Omega \sqrt{1\text{MHz} \times n_{i}[V]} \]  
\[ n_{\text{buf}}^{v} = 16\sqrt{1\text{MHz} \times n_{v}[V]} \]

The datasheet for the Analog Devices OP27EZ states that typical values for \( n_{i} \) and \( n_{v} \) are 0.4 pA/√Hz and 3 μV/√Hz respectively. There are two additional noise sources in the system:
the 20 bit DAC \( (n_{DAC}) \) and dSPACE’s 16 bit ADC \( (n_{ADC}) \). From the DAC datasheet, the output noise is estimated to be 8.5 \( \mu V \) RMS [53:5], assuming that the PCB was laid out properly. The design guidelines provided in the datasheet were followed, so this is, more or less, a valid assumption. An attempt to measure this value was made, but the noise floor of the lab oscilloscope is approximately 100 times larger than the expected measurement. So, the estimated value was used in the calculation. The ADC noise is taken from the result plotted in Figure 6.8.

This translates to 208 \( \mu V \) RMS.

When a system contains several noise sources, they are not summed together in the regular fashion. They are added as follows:

\[
n_T = \sqrt{\sum n_j^2}
\]

(6.12)

where \( n_j \) is a noise source. Substituting \( n_{DAC}, n_{ADC}, \) and (6.6) through (6.11) into (6.12) gives:

\[
n_T = \frac{1 \times 10^6 [nm]}{20[V]} \frac{1}{K} \left[ \begin{array}{c}
32 \left( n_{DAC}^2 + n_{buf}^2 + n_{buf}^2 \right) \\
+ 8 \left( n_{scale}^2 + n_{scale}^2 \right) \\
+ \left( n_{NZ}^2 + n_{NZ}^2 \right) + n_{ADC}^2
\end{array} \right]^{\frac{1}{2}} [V] 
\]

(6.13)

This predicted value is lower than the noise of our sensor (5.5 nm RMS) and is much lower than the 10 nm RMS quantization noise (Figure 6.3) using the conventional data acquisition method.

Figure 6.20 shows a plot of the system noise for the complete Nanozoom system, with the capacitive probe input grounded with a 50 \( \Omega \) terminator. The result is 3.0 nm RMS which is very close to the predicted result reported in (6.13). Additionally, this corresponds to a dynamic range of over 300,000 which far surpasses the limits of existing high speed data acquisition systems.
Figure 6.20: Plot showing Nanozoom noise, including the contribution from the 16-bit ADC. After conversion to nanometers and rescaling by $1/K_z$, the noise is approximately 3 nm RMS.

6.7 Summary

In order to achieve nanometer range resolution for the rotary-axial spindle, a patent pending method [4] for improving the resolution of a feedback control system was presented. This resolution increase is realized over the entire input range, and does not contribute additional delay to the system. This goal is unachievable with current technology.

The conventional method of passing a sensor signal directly to an ADC can introduce limitations due to the converter’s quantization noise. Currently, resolution is improved through high big count ADCs, oversampling and averaging, or amplifying the feedback signal. All three of these approaches, however, bring drawbacks which include adding phase lag to the system, or have variable resolution over the converter input range. *Nanozoom* technology addresses all these issues.
The operating principle of Nanozoom is as follows. The feedback sensor’s signal is subtracted from a high resolution position reference signal, creating an analog control error. This error is then amplified before being passed to a high speed ADC. The converted result is then rescaled in the digital domain by the reciprocal of the applied gain to recover the error at an improved resolution.

One embodiment of the invention was implemented on a customized PCB. The high resolution reference, generated in the digital domain, is transferred by RS232 to an FPGA chip on the PCB. The FPGA chip programs a 20 bit DAC through an SPI protocol. The DAC output signal is processed by a series of low noise operational amplifiers. Aspects of the PCB design were discussed along with explanations of the RS232 and SPI communication protocols.

In order to predict if the noise level in the presented invention is an improvement upon existing technology, a noise model of the analog system is developed. The model predicted a full system noise result of 2.8 nm RMS. A measurement of 3.0 nm RMS indicates good agreement with the experimental results. This invention represents a tenfold improvement in system noise compared to the conventional solution. The dynamic range of over 300,000 in a high speed data acquisition system far surpasses any commercially available product.
Chapter 7

Axial Control and Experimental Results

This chapter traces the development of loop shaping controllers to regulate the axial position of the spindle shaft. Since the system is linear from the input current command to the output axial position, linear control laws can be used for compensation. Specifically, a loop shaping controller design process will be used to progressively extract higher performance levels from the system until the maximum is reached. Along the way, physical phenomena will be explained and workarounds for existing limitations will be implemented.

Loop shaping controller design is a classical frequency domain control framework for linear systems which requires the knowledge of the plant to be controlled. For most systems, it involves designing lead-lag compensators, low pass and notch filters, for example, to “shape” the system loop transmission as desired, within the physical limitations of the system. The theory is not reproduced here; readers unfamiliar with this topic may read a NASA-produced tutorial [55].

The controller design process used can be described as follows. First, the system should be modeled, and that model should be verified. Once this has been completed, a controller is designed and implemented. If the controller makes the system unstable, a mistake has been made in one of the first three steps. Once the system is stabilized by the controller, the performance is checked. If it is unacceptable, the limitation must be found. If the limitation can be bypassed, the
workaround is implemented. Then the controller is tuned to advance the performance. This iterative process is repeated until physical limitations halt meaningful improvement.

The system performance is measured in terms of four quantities: bandwidth, dynamic stiffness, phase margin, and position regulation noise. Bandwidth is a measure of how well the system tracks position reference inputs. Dynamic stiffness gauges the disturbance rejection characteristics of the system. Stability is quantified using phase margin criteria, a subset of the Nyquist stability laws. Position regulation noise, the final quantity, indicates which vibration modes are excited at a constant position and also quantifies the electrical noise in the system.

In the remaining sections of this chapter, further details on the axial control and experimental results are presented. Section 7.1 describes the interaction between various elements in the axial control system. In section 7.2, the initial stages of the controller design process, which include modeling of the axial control plant are set forth. Sections 7.3 discusses the experimental measurement of the actuator force. Section 7.4 explains the integration of Nanozoom into the system. In section 7.5, the steps taken to maximize the performance of the axial control loop are thoroughly discussed. Section 7.6 covers the implementation of rotary and axial control. Finally, the chapter is closed with a summary.

7.1 System overview

Figure 7.1 shows the interaction between each element in the axial control system. A host computer provides a user interface and gives read and write access to variables on the real time computer, which is also called the target. Communication between the target computer, a dSPACE DS1103, and the host is through a fiber optic cable which enables high speed data transfer. Interfacing with the target is an I/O box which gives the user access to several 16 bit
high speed ADCs and DACs. The control algorithm, which runs on the target, outputs a control signal through a DAC. This signal is the current reference command, and is sent to the power amplifier. The power amplifier drives current through the magnetic bearing coils, which in turn produces force on the armature. Axial motion of the spindle shaft results. Using conventional feedback, this motion is measured by the capacitive probe and is then fed back to the dSPACE target computer through an ADC, thus closing the loop.
Figure 7.1: Interaction between components in axial system. Host computer photo is used with the permission of Microsoft [56]. dSPACE DS1103 and connector panel are used with the permission of dSPACE [57, 58].

7.2 Loop shaping controller design

This section presents the first iteration of the controller design process. First, a model of the system is developed. After, the results are shown for the controller.
7.2.1 Model

The open loop block diagram for the axial system is shown in Figure 7.2. The input to the open loop is the current command, $i_r$, which is the signal produced by the control law. This signal is scaled by the reciprocal of the amplifier and DAC gains, $k_a$ and $k_d$, so that there is a one-to-one relationship between the control effort and the average output channel current, $i_{avg}$. The DAC is modeled with gain, zero order hold, and saturation elements. The 16 bit DAC quantization is ignored since its resolution is sufficiently high enough and in the feed forward loop. The DAC outputs a voltage to the current amplifier. Since the amplifier’s measured bandwidth (Figure 4.22) is much higher than the anticipated bandwidth, it is modeled as a gain, $k_a$. The force produced by the actuator is proportional to $i_{avg}$ through the gain $k_i^{ch}$. This gain is a modified current to force ratio related to $k_i$ by:

$$k_i^{ch} = \frac{k_i}{2}$$  \hspace{1cm} (7.1)

Scaling by half accounts for the difference between coil current and channel current.

**Figure 7.2:** Block diagram of open loop axial system. DAC, amplifier, axial plant, and ADC dynamics are considered.
The transfer function from total force (actuator plus disturbance) $F$ to actuator position is $x$:

$$
\frac{x(s)}{F(s)} = \frac{10^6}{ms^2 - k_x} + \sum_i \frac{\alpha_i + \beta_i s}{s^2 + 2\zeta_i \omega_i s + \omega_i^2} \tag{7.2}
$$

The first term on the right hand side models the rigid body motion of the shaft mass, $m$, and the negative stiffness of the actuator, $k_x$. The negative spring effect can be best explained by comparing it to a compression spring. When a compression spring is pushed from its equilibrium position, it pushes back, providing a reaction force. When a negative spring is perturbed from its equilibrium point, it reacts in the same direction, making the system unstable. Thus, there are two right hand plane poles in the denominator of the first term of the transfer function. The factor of $10^6$ in the numerator converts the transfer function output, $x$, which is the axial position of the target screw, from meters to microns. The second term in the right hand side represents the vibration modes of the structure. Although a finite element analysis was performed, the parameters in this term are difficult to predict. This portion of the model will be addressed in section 7.2.2.

Since the capacitive probe has 20 kHz bandwidth, which is likely much higher than the expected unity gain crossover frequency, it can be modeled as the gain $k_s$. Sensor noise, or other similar disturbances are modeled as the input $x_d$. The next element in the feedback loop is the ADC which is modeled with an input saturation, delay, quantizer, and gain. The delay, $T_d$, is a lumped term which takes into account the zero order hold delay, analog to digital conversion delay and any other extraneous phase lag. The final step is to rescale the feedback by the reciprocals of $k_{adc}$ and $k_s$ so that the controlled variable is the measured axial position, $x_m$, in microns. Table 7.1 shows the values of each parameter in the model. The values for $k_{ch}^i$, $k_s$, and $T_d$ are initial estimations. These will be better defined once the system is identified.
Table 7.1: Values of parameters in open loop axial system.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>$k_d$</td>
<td>10</td>
<td>V/V</td>
</tr>
<tr>
<td>$v_{s\text{at}}$</td>
<td>±10</td>
<td>V</td>
</tr>
<tr>
<td>$k_a$</td>
<td>0.5</td>
<td>A/V</td>
</tr>
<tr>
<td>$k_i^{ch}$</td>
<td>162</td>
<td>N/A</td>
</tr>
<tr>
<td>$m$</td>
<td>1.85</td>
<td>kg</td>
</tr>
<tr>
<td>$k_x^*$</td>
<td>$1.3 \times 10^6$</td>
<td>N/m</td>
</tr>
<tr>
<td>$k_a$</td>
<td>0.02</td>
<td>V/μm</td>
</tr>
<tr>
<td>$T_d^*$</td>
<td>0</td>
<td>s</td>
</tr>
<tr>
<td>$k_{\text{adc}}$</td>
<td>0.1</td>
<td>V/V</td>
</tr>
</tbody>
</table>

*Initial parameter estimation

The complete open loop transfer function, from $i_r$ to $x_m$ is:

$$P_a(s) = \frac{x_m(s)}{i_r(s)} = k_i^{ch} e^{-T_d s} \frac{x(s)}{F(s)}$$  \hspace{1cm} (7.3)

A Bode plot of the open loop system, neglecting structural modes is shown in Figure 7.3.

Figure 7.3: Bode plot of the open loop system model ignoring flexible modes. The flat region at low frequency is the negative stiffness of the actuator. The response falls off at -40 dB/dec corresponding to a rigid body mode. There is no resonance because the phase is constantly -180°.
7.2.2 System identification and refined model

Since the open-loop system is unstable, the loop must be closed before it can be identified. A closed loop block diagram is shown in Figure 7.4. The controller consists of two terms, a loop shaping controller $C_{LS}(s)$ and an integrator. These are implemented in parallel so that the integrator state can be limited, so as to prevent windup.

Figure 7.4: Block diagram of closed loop system. The controller consists of two terms: a loop shaping term $C_{LS}(s)$ and an integrator. The integrator is implemented in parallel to the loop shaping term so that integral windup can be limited.

The loop shaping term contains a lead-lag compensator and proportional gain of the form:

$$C_{LS}^1(s) = \frac{a s}{\omega_c s + 1}$$

(7.4)

where $a$ is related to the amount of phase added and $\omega_c$ is the radian frequency about which the compensator is centered. The integrator corner frequency is $\omega_i$. Setting $a$ to 5, $\omega_c$ to 1885 rad/s (300 Hz), $k_p$ to 0.0125, and $\omega_i$ to 189 rad/s (30 Hz) gives an intended unity gain crossover frequency of 300 Hz with $60^\circ$ phase margin as shown by the loop transmission in Figure 7.5. The experimental implementation of this controller gave a stable result and so the plant could be experimentally measured. The experimentally measured frequency response is shown in Figure 7.6. It was obtained using the dSPACE DSA swept sine wave tool [43].
Figure 7.5: Negative of the loop transmission of the model based loop shaping controller $C_{LS}$.

Figure 7.6: Experimentally measured frequency response from $i_r$ to $x_m$.

As expected, the response exhibits the negative spring effect up to 100 Hz, shifting to a rigid mode at 120 Hz. Using the first term of the model derived in (7.2), $k_x$ can be identified as:

$$k_x = m\omega_n^2$$  \hspace{1cm} (7.5)
where $\omega_{mx}$ is the mass-negative stiffness corner frequency in rad/s. With $m = 1.8$ kg and $\omega_{mx} = 754$ rad/s, $k_x$ equals $1.04 \times 10^6$ N/m. Knowing this and the approximate value of the open loop transfer function below the corner frequency, $k_i^{ch}$ can be solved for:

$$k_i^{ch} = \frac{k_x}{1 \times 10^6} \times \frac{x_m(j\omega)}{i_r(j\omega)} \bigg|_{\omega = \omega_0}$$  \hspace{1cm} (7.6)

This yields $k_i^{ch} = 110$ N/A. These experimentally measured values of $k_i$ and $k_i^{ch}$ are approximately 23% and 31% lower, respectively, than the values predicted by the finite element analysis. There is some additional phase lag beyond -180° caused by the DAC zero order hold, capacitive probe dynamics, and ADC delay. So, $T_d$ can now be identified as 25 $\mu$s through trial and error process. This corresponds to a two to three sample delay. Then a set of collocated vibration modes exists with the first at 1700 Hz, and another mode at 3200 Hz. The lower frequency modes are lightly excited modes which are caused by flexible joints in the machine structure. The resonance at 3200 Hz is an axial mode of the shaft.

To model these vibrations, the variables in the second term of (7.2) must be identified. This can be accomplished using the modal analysis tool in CutPRO, a commercially available software package used for modeling and simulating machine tool systems [59]. This software uses non-linear least square fitting to calculate the residues, natural frequencies, and damping ratios of the transfer function. The process for translating the residues to $\alpha_i$ and $\beta_i$ from (7.2) is given in [60] and is as follows. The transfer function from force to position for a mechanical structure is:

$$\frac{x(s)}{f(s)} = \sum \frac{r_i}{s - s_i} + \frac{r_i^*}{s - s_i^*}$$ \hspace{1cm} (7.7)

where $r_i$ is the residue which is represented as:
\[ r = \sigma + j\nu \quad (7.8) \]

\[ \alpha = 2(\zeta\omega_n\sigma - \omega_n\nu) \quad (7.9) \]

\[ \beta = 2\sigma \quad (7.10) \]

Assuming proportional damping where the residues are purely imaginary, the \( \beta_i \) terms are zero. The values of \( \omega_{ni}, \zeta_i, \alpha_i \) identified for our plant are summarized in Table 7.2. The full model from \( i_r \) to \( x_m \) plotted against the experimental measurement is shown in Figure 7.7. The agreement is excellent up to about 4 kHz where the experimental measurement is poor.

Table 7.2: Modal parameters for axial vibration modes.

<table>
<thead>
<tr>
<th>Mode</th>
<th>( \omega_n ) [rad/s]</th>
<th>( \zeta )</th>
<th>( \alpha ) [( \mu m/N\cdot s^2 )]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>10,680</td>
<td>0.047</td>
<td>19,400</td>
</tr>
<tr>
<td>2</td>
<td>14,700</td>
<td>0.00825</td>
<td>3,750</td>
</tr>
<tr>
<td>3</td>
<td>17,090</td>
<td>0.02</td>
<td>10,880</td>
</tr>
<tr>
<td>4</td>
<td>20,260</td>
<td>0.0136</td>
<td>55,260</td>
</tr>
</tbody>
</table>

Figure 7.7: Comparison between experimentally measured frequency response of the open loop system and the model.
7.2.3 Refined results

Based on the experimentally measured plant, several iterations of the controller design process were performed. The goal in this stage was to reach the highest possible crossover frequency possible while maintaining stability. The refined controller, designated $C_{LS}^2$, consists of one lead compensator centered at 900 Hz ($\alpha = 4$) and another at 1100 Hz ($\alpha = 4$) to increase the phase margin at the desired crossover frequency. Two low pass filters with cutoff 1400 Hz are included to attenuate the vibration modes. The proportional gain shifts the loop crossover to 1000 Hz. Lumped together, the refined loop shaping term is:

$$C_{LS}^2(s) = \frac{0.0606\left(7.07 \times 10^{-4} s + 1\right)
\left(5.79 \times 10^{-4} s + 1\right)}{
\left(4.42 \times 10^{-5} s + 1\right)\left(3.62 \times 10^{-5} s + 1\right)\left(1.14 \times 10^{-4} s + 1\right)^2} \quad (7.11)$$

The integrator corner frequency was also increased to 90 Hz. The loop shaping term is digitized using a matched pole method to improve the execution time, and the results are shown in Figure 7.8. The control loop is implemented at a 100 kHz sampling rate.
Figure 7.8: Bode plots of controller design results. a) the loop shaping controller including integrator. b) the negative loop transmission showing 1 kHz crossover frequency with 38° phase margin c) the dynamic stiffness showing minimum of 8 N/μm at 145 Hz. d) shows a 1.5 kHz -3 dB closed loop bandwidth.

The frequency response of the overall controller (loop shaping and integrator) is shown in Figure 7.8a. A crossover frequency of 1000 Hz with 38° phase margin was achieved. This is close to the limit set by the resonance and subsequent drop in the phase after 3300 Hz. A structural change is needed to further increase the crossover frequency. The closed-loop dynamic stiffness, or the reciprocal of the closed loop motion response to a force disturbance, can be calculated as follows:
Due to the integrator, the steady state stiffness tends to infinity, but in reality, this specification is limited by the load capacity of the actuator and the stiffness of the shaft. At high frequency, the stiffness also goes to infinity due to the shaft mass. The minimum dynamic stiffness achieved by $C_{LS}^2$ is 8 N/μm at 145 Hz. The -3 dB closed-loop bandwidth is 1500 Hz. A 10 μm step response is shown in Figure 7.10. Some saturation is evident at this larger step size, but a strong 1 kHz component is observable which coincides with the designed crossover frequency.

The RMS position regulation error is 12.8 nm RMS as shown in Figure 7.11. In the power spectrum density, there is a strong component at 44 Hz, harmonics of the 60 Hz line frequency as well as some noise from a high frequency structural mode above 20 kHz. The component at 44 Hz is from a resonance associated with the work bench (Figure 7.9) where the rotary-axial spindle sits. This conclusion can be made because the 44 Hz disturbance is absent when the spindle is mounted on a granite table. The overall noise performance can be improved when Nanozoom is integrated into the system (section 7.4).
**Figure 7.9:** Rotary-axial spindle setup on lab workbench. A 44 Hz resonance is associated with this table.

**Figure 7.10:** 10 μm step response of the closed loop system. A strong 1 kHz component is evidence that the designed crossover frequency was achieved.
Figure 7.11: a) time domain plot of position regulation noise. The RMS error is 12.8 nm. b) power spectrum density of position regulation error. Implementing Nanozoom should improve this result.

7.3 Force measurement

Since the actuator force is both a function of armature position and current, it is easier to measure the force characteristic when the position can be finely controlled. So, to obtain this measurement, the position loop is closed, a known load is applied, and the average current supplied by the power amplifier is recorded. The result of this test is shown in Figure 7.12. The gradients between force and current, and force and position, lead to the identification of the coefficients $k_l$ and $k_i$ which are 240 N/A and $1.04 \times 10^6$ N/m. The corresponding value of $k_i^{ch}$ is 120 N/A. This result matches quite well with the results obtained from the frequency response plotted in Figure 7.6. Since the power amplifier can deliver a continuous average current of 2.6 A and peak current of 5 A per channel, this translates into a continuous load capacity of 300 N, or a peak force of 600 N. This represents a 60% increase in load capacity compared to the 180 N capacity of the aerostatic spindle. Considering that the thrust plate in the aerostatic spindle is
125% (75 mm in diameter) larger by area than the armature of the rotary-axial spindle (50 mm), this is a promising result. Depending on the location of the armature in the stroke, up to 1100 N unidirectional force can be produced.

One question that has yet to be definitively answered is why the force is 30% lower than expected. The most likely cause is that the permeability of the powdered iron is lower than specified. However, this result has not yet been confirmed.

![Figure 7.12: The actuator force characteristic. Data was recorded by using the controller to hold the spindle at a constant position. Then, a known load was applied axially and the steady state current delivered by each power device was recorded.](image)

7.4 Integrating Nanozoom

The next step in increasing the performance of the rotary-axial spindle is to integrate Nanozoom into system. A pictorial block diagram is shown in Figure 7.13. The Nanozoom PCB
receives the capacitive probe output voltage, and the position reference via RS232, then synthesizes the analog error signal which is fed back to the dSPACE 16 bit ADC. Otherwise, the system remains setup in a similar manner.

From a control point of view, Nanozoom only brings one additional element to the loop, a saturation block. This is illustrated in Figure 7.14. When the analog error signal is multiplied by the zooming gain $K_z = 8$, its range is adjusted from $\pm 10$ V to $\pm 80$ V. Naturally, the ADC input range remains at $\pm 10$ V. So, at some point the error will saturate. For the $\pm 500$ $\mu$m range sensor the saturation limits $R$, are:

$$\frac{R}{K_z} = \frac{\pm 500 \mu m}{8} = 62.5 \mu m$$

(7.13)
Figure 7.13: Hardware interaction with Nanozoom integrated into system. Host computer photo is used with the permission of Microsoft [56]. dSPACE DS1103 and connector panel are used with the permission of dSPACE [57, 58].
Figure 7.14: Block diagram of axial control loop with Nanozoom saturation block. The error signal saturates when outside the range ±62.5 μm.

Linear controllers can handle small non-linearities. As it turns out, the saturation imposed by Nanozoom cannot be handled in this case. The initial implementation of Nanozoom and $C_{LS}^2$ gave an unstable result, as shown in Figure 7.15, during the startup process. When the axial loop is driven in an unstable manner, it makes large noise as it bounces off the stator several hundred times per second.

Figure 7.15: Instability due to saturation of Nanozoom during the startup process. Due to the saturation of the Nanozoom error, full force is applied to the armature until the spindle is in a linear feedback zone. At this point the shaft is moving too quickly, and the actuator cannot apply enough force to recover.
When the initial reference command is the center position and the shaft is at \( x = -500 \) \( \mu m \), the startup error is 500 \( \mu m \). However, the reading seen by the controller is only 62.5 \( \mu m \). So, full force is applied until the shaft is in the linear feedback region. Once this occurs, however, it is too late. The spindle shaft is moving too quickly, and the actuator cannot apply enough counteractive force to recover. Consequently, the armature strikes the opposite side of the stator and the process repeats until the control loop is broken and the reference current is set to zero by the RMS current protector. (NB: the RMS current protector monitors the RMS current being supplied by each channel. The absolute maximum current which can be driven through each coil circuit constantly, is 2.6 A. If this is violated, the control loop is broken).

There are at least two solutions to this problem. The first is to set the initial reference position close enough to the actual shaft position such that the startup error is inside the linear feedback range. However, since not enough constant current can be driven through the coils to balance the force produced by the bias fluxes, the axial position cannot be held constant near the extremes of the stroke. It can, however, be moved through the extremities of the stroke by following a ramp reference command. This option was not tested. Another solution is to use a combination of conventional and Nanozoom derived feedback. Since the saturation problem is non-existent with conventional feedback, a logic based switch can intelligently select the conventional feedback when the error range is outside ±62.5 \( \mu m \), and Nanozoom feedback once the error is inside the zone. The only downside to this is that a second ADC is required.

Figure 7.16 shows the position regulation noise with \( C_{ES}^2 \) and Nanozoom implemented. The RMS error is 5.3 nm RMS which is essentially the noise floor of the capacitive probe. This result was extremely difficult to achieve as it requires that the system is free of ground loops, and that all connections are terminated properly.
Figure 7.16: a) time domain plot of position regulation noise. The RMS error is 5.3 nm. b) power spectrum density of position regulation error.

7.5 Further enhancement of axial control loop performance

This section focuses on improving the bandwidth and stiffness of the axial control loop. To move past the roadblock found with $C_{2s}$ and its integrator, the spindle was disassembled and three additional fasteners were installed (for a total of six) in the joint between the journal bearing shaft and the extension shaft. This should enhance the stiffness significantly, paving the way for a higher performance controller. Initially, the axial plant must be re-measured and the appropriate changes can be made to the model. After this, progress can be made improving the controller.

7.5.1 Revised plant model

After installing the additional fasteners, the axial plant was re-measured and is plotted in Figure 7.17, overlapping the response obtained with three screws. No change is evident up to 3300 Hz, but after this, it can be observed that there is no drop off in the phase. This is an
improvement. Next, the frequency response increases to a resonance at 6 kHz, before a very flexible non-collocated mode at 8500 Hz. This is a high ordered armature bending mode which is dependent on the preload developed by the locknut.

\begin{figure}[h]
\centering
\includegraphics[width=\textwidth]{figure7.17}
\caption{Comparison between frequency responses of the measured plants with three screws (blue) and six screws (red) holding the extension shaft to the air bearing shaft.}
\end{figure}

Since this mode is at the high end of the frequency range of interest, it is adequate to model it with a collocated mode as long as the additional 180° of phase lag is accounted for. The modal parameters for the new plant measurement are given in Table 7.3. In this case, proportional damping is not assumed in order to get a better fit. Therefore, the residues are complex and $\beta$ is non-zero.
### Table 7.3: Modal parameters for revised motion control plant

<table>
<thead>
<tr>
<th>Mode</th>
<th>$\omega_n$ [rad/s]</th>
<th>$\zeta$</th>
<th>$\alpha$ [(\mu m/N-s^2)]</th>
<th>$\beta$ [(\mu m/N-s)]</th>
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<td>13,380</td>
<td>0.24</td>
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<td>41,250</td>
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</tr>
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<td>0.0085</td>
<td>-381,560</td>
<td>-20.4</td>
</tr>
</tbody>
</table>

**Figure 7.18:** Comparison between measured and modeled motion control plant frequency responses with six screws installed to connect the air bearing shaft and the extension shaft.

7.5.2 Current saturation

After trying to make marginal improvements to $C_L^2z$, it was not possible to stabilize the system with a crossover frequency targeted above 1 kHz. The negative of the loop transmission for a loop shaping controller/integrator pair is shown in Figure 7.19. Even though the phase margin is more than sufficient at 36°, the loop is destabilized upon startup. This stability issue is
different than that discussed in section 7.4. However, a snapshot of the startup process (Figure 7.20) can also be used to diagnose the problem.

![Graph](image)

**Figure 7.19:** Negative of loop transmission for an intermediate loop shaping controller/integrator pair. Even though the phase margin indicates a stable system, the startup process is unstable.

Initially, with the reference position set to zero, the startup error is 500 μm. So, the controller demands a very large current. This, however, is limited by the saturation block which sets the input limits of the power amplifier. From the plot, it can be seen that the maximum current, and consequently full actuation force, is applied for more than 3 ms. This accelerates the spindle shaft quite violently. By the time the control signal switches to the opposite limit, the combination of the inertial force and magnetic force is too great to be overcome by the commanded force alone. The end result is that the armature crashes into the stator, bounces off, and the process repeats in an unstable manner. In summary, current level saturation is driving the system unstable upon startup. This is a fundamental obstruction to stiffer controllers.
Figure 7.20: Unstable startup capture. The left hand vertical axis shows the current command signal (blue), and the right hand vertical axis shows the axial position as measured by the capacitive probe (red). Full load is applied for too long, violently accelerating the spindle. A current command corresponding to full load in the opposite direction is not strong enough to slow the shaft, and so the armature crashes into the stator, creating instability.

7.5.3 Dual controller approach

There are a number of solutions to the current saturation problem caused by high bandwidth controllers. A radical path would be to use substitute materials or more coil turns to generate significantly more force per amp of current. A more practical approach is to make use of the fact that the lower bandwidth controller starts up in a stable manner. Then, two controllers can be used: a less aggressive controller which is very robust, and a high performance controller which only tracks smooth reference signals, but has very high stiffness. To do this, a switching mechanism needs to be developed. Furthermore, a direct switch cannot be made between each controller because the high performance controller will, in most cases, have a much higher
integrator gain. So, when the control effort is switched, there may be a large discontinuity in the control signal, which may cause instability.

Several potential solutions were developed, each with advantages and disadvantages. The general principle for the route that was chosen is as follows. The loop shaping dynamics, in general are quite fast, so it can quickly recover to a stable state. The integrator, on the other hand, cannot. This term needs to be kept continuous. Knowing this, the architecture shown in Figure 7.21 is implemented. The main components are two switches, some controller selection logic, two loop shaping controllers, and one integrator. Each switch selects either Input 1, or Input 2, based on the output of the controller select logic, $C_{sel}$. If Input 1 is selected, then the low bandwidth loop shaping controller $C_{LBW}$ and its integrator gain $\omega_{i\,LBW}$ is active. If Input 2 is selected, then the high bandwidth loop shaping controller $C_{HBW}$ and its integrator gain $\omega_{i\,HBW}$ is active. The path through the first switch implements the integrated term, and the path through the second switch implements the feed through term. So, even though the integrator gains change, the integrator term is kept continuous because its output cannot change instantaneously, even with a step change in input. Since $C_{LS}^2$ and its respective integrator did not experience any difficulties with respect to stability, it was chosen as the low bandwidth controller. The controller logic consists of a set-reset latch which is set when the error and its RMS value is small. It resets when the error becomes very large, thus re-enabling the more robust of the two controllers.
Figure 7.21: Dual controller approach as a remedy to current saturation. Here, one robust linear controller is used for startup. Once stabilized, the logic sequence commands a change of input so that a higher bandwidth controller is selected. The use of a single integrator keeps the slower dynamics continuous, and so stability is maintained.

7.5.4 Controller improvement

A higher bandwidth controller with increased performance was designed to more thoroughly test the proposed dual controller approach. The four characteristic Bode plots for the controller $C_{LS}^3$ and its integrator are shown in Figure 7.22. The controller form is similar to that given in section 7.2.3, except here, two notch filters are added at 3345 Hz and 8500 Hz to attenuate the resonances in the loop transmission. One of the low pass filters is also removed to improve the phase margin. The value of each pole and zero in the controller, and the proportional gain were slightly modified to achieve one crossover frequency at 1400 Hz, and a second at 1830 Hz.
The final controller transfer function before discretization is:

$$C_{LS}^3(s) = \frac{0.346(3.18 \times 10^{-4}s + 1)(1.99 \times 10^{-4}s + 1)}{(1.99 \times 10^{-5}s + 1)(3.18 \times 10^{-5}s + 1)(3.0 \times 10^{-5}s + 1)} \times \frac{s^2 + 2488s + 4.3 \times 10^8}{s^2 + 8294s + 4.3 \times 10^8} \left(\frac{s^2 + 317.8s + 2.81 \times 10^9}{s^2 + 1.06 \times 10^5s + 2.81 \times 10^9}\right)$$  \hspace{1cm} (7.14)

The corner frequency of the integrator was also shifted slightly to 180 Hz. All of this results in 44° and 38° of phase margin at the first and second crossover frequencies, respectively. The low frequency dynamic stiffness is significantly enhanced to 50 N/μm at 340 Hz. The -3 dB
bandwidth is 2.5 kHz. This controller was successfully switched to after the system was initially stabilized by the lower bandwidth controller.

A 400 nm step reference (Figure 7.23) tracked with the improved controller demonstrates the increased performance. The peak time of 0.36 ms confirms the designed crossover frequency, though the first vibration mode near 1700 Hz is excited, superimposing a ripple on the response. However, the position regulation noise decreases to 4.4 nm RMS because the noise on the capacitive probe is tracked as a reference command.

![Figure 7.23: 400 nm step response for the closed loop system regulated by $C_{ls}^3$ and its integrator. The 1390 Hz component corresponds with the crossover frequency. The 1695 Hz oscillation is due to the first mode being above unity gain.](image)

7.5.5 Maximizing the axial control loop performance

No fundamental limitations were met when the controller presented in the previous section was designed. It stands to reason that gains can still be had, especially in terms of closed
loop dynamic stiffness. This performance metric is directly related to the integrator corner frequency, since the controller provides the low frequency stiffness, and the integrator dominates the low frequency controller response. The main challenge is to increase the proportional and integral gains as much as possible while maintaining sufficient phase margin at the crossover frequencies, and while keeping resonances below unity gain in the loop transmission.

After some time tuning, the limitation of the system was reached with the following loop shaping controller:

\[
C_{ls}^4(s) = \frac{0.26(4.44 \times 10^{-4}s + 1)(2.27 \times 10^{-4}s + 1)}{(1.99 \times 10^{-5}s + 1)(2.53 \times 10^{-5}s + 1)(3.0 \times 10^{-5}s + 1)}
\times \frac{(s^2 + 2488s + 4.3 \times 10^8)}{(s^2 + 8294s + 4.3 \times 10^8)} \times \frac{(s^2 + 752s + 1.41 \times 10^9)}{(s^2 + 3760s + 1.41 \times 10^9)}
\times \frac{(s^2 + 317.8s + 2.81 \times 10^9)}{(s^2 + 1.06 \times 10^5s + 2.81 \times 10^9)} \times \frac{(s^2 + 1480s + 5.49 \times 10^9)}{(s^2 + 7410s + 5.49 \times 10^9)}
\]

(7.15)

It consists of two phase compensators, a low pass filter, and four notch filters. The integrator was pushed all the way to 750 Hz, which is more than four times higher than in the previous iteration. Again, there are two crossover frequencies. The first, at 1530 Hz has 43° phase margin and the second crosses at 1950 Hz with 36° of phase margin. At 540 Hz the dynamic stiffness is 100 N/μm which is a minimum in the low frequency region. The integrator provides infinite static stiffness. But in the high frequency region the resonance around 8500 Hz results in a dynamic stiffness global minimum. A variety of impulse hammer and swept sine tests shows that its value varies from 10 to 50 N/μm depending on the amplitude of the disturbance. An example of this is shown in Figure 7.24 where the compliance at the 8500 Hz resonance was measured with swept sine tests of varying excitation amplitude. This variation is due to small non-linearities in the aerostatic bearing. The -3 dB bandwidth was improved slightly to 2.6 kHz. The controller, stability, stiffness and bandwidth results are summarized in the Bode plots in Figure 7.25. Even
though this controller is extremely aggressive, the position regulation noise does not degrade, as 4.6 nm RMS is achieved (Figure 7.26). Transforming the data to the frequency domain shows that the static component is smaller than 0.01 picometers. A step response for the complete closed loop system is shown in Figure 7.27.

**Figure 7.24:** Varying compliance of 8500 Hz resonance. Excitation amplitudes ranged from 0.0025 A to 0.1 A.
Figure 7.25: Characteristic Bode plots for improved controller. a) Controller frequency response. b) Frequency response of the negative loop transmission. c) Dynamic stiffness frequency response. d) Closed loop bandwidth frequency response.
Figure 7.26: Position regulation noise achieved with the highest performance controller. The result is 4.6 nm RMS, which is below the capacitive probe noise floor. The DC component of the noise is less than 0.01 picometers.

Figure 7.27: 300 nm step response of closed loop system as regulated by $C_{LS}$ and its integrator. A 1560 Hz component, which nearly corresponds to the first crossover frequency, is observed. The large overshoot is undesirable, but is a result of the quest for maximum dynamic stiffness.
7.6 Combining rotary and axial control loops

In this section, combined rotary and axial control is investigated in two areas. First, the issue of implementing two separate control loops is dealt with. Next, the performance of the metrology solution with respect to combined rotary-axial motion is investigated.

7.6.1 Multi-tasking control implementation

The dSPACE DS1103 can handle the control law computation, one D/A conversion, and two A/D conversions, logic, and the RS232 engine for Nanozoom at a 100 kHz control rate. In this setup, there is just one sampling time, so it is called a single-tasking system. It will be difficult, however, for this single tasking system to handle the additional computational load of the rotary control loop at this sample rate. As shown in Chapter 5, 10 kHz is sufficient for running the rotary control loop. Similarly, it is not necessary to execute the Nanozoom communication protocol at such a high rate. 10 kHz is also sufficient. dSPACE provides the option of running two processes simultaneously at different rates, using interrupts. This is called a multi-tasking system.

Using a multi-tasking implementation, the rotary loop is operated at 10 kHz and the axial loop is executed at 100 kHz with the interrupt timer. The axial loop is assigned a higher priority to prevent overruns. With this approach, however, a problem arises. The interrupts generate a significant amount of low frequency noise on the axial control error signal as is evident when comparing the frequency spectrum in Figure 7.28a to Figure 7.28b. This phenomenon is only observable with the loop shaping controller $C_{Ls}^3$ and its corresponding integrator, but not with $C_{Ls}^2$ and its integrator.
Figure 7.28: Noise induced on the axial control error signal in a) a single-tasking process and b) a multi-tasking process. The controller used to obtain this data was $C_{ls}$ and its corresponding integrator. The difference in the low frequency noise between the two situations is approximately three orders of magnitude.

It was determined that the source of this noise is interrupt jitter. Jitter is defined as the uncertainty of a clock period. To quantify the jitter in the system, an experiment was performed for two cases: first the single-tasking setup, and then the multi-tasking setup. In each case, a nominal 10 μs control rate was specified, and one million data samples were collected for each case. The average and standard deviation of the set of time stamps was calculated for both cases. In the single tasking setup, the average sampling period was 10.000 μs with a standard deviation of 14 ns. In the multi-tasking environment, the actual average sampling time was 9.976 μs with a standard deviation of 121 ns. The experimental data is summarized with histograms in Figure 7.29. So, in the multi-tasking case, there is significant uncertainty in the sampling period.
Uncertainty in the sampling period can lead to errors in the DAC [61], the ADC [62], and the digital controller [63]. Some researchers [64] have attempted to tackle this problem from a task scheduling point of view. However, dSPACE does not allow such low level access to the processor. Lincoln developed a method to compensate for this type of problem in motion control loops [65], but it requires knowledge of the time stamps, which in this case, are not available. Knowing that this noise issue was not a problem with $C_{LS2}C_{LS}^2$, which has an additional low pass filter and lower proportional gain compared to $C_{LS}^3$ (i.e. $C_{LS}^2$ has more high frequency attenuation) a solution was developed based on appending additional low pass filters to $C_{LS}^3$. Conventional single and multi-pole low pass filters did improve the noise performance, but they also add phase lag to the loop transmission which decreases the loop transmission phase margin. A better solution is to include the low pass filter $C_f$ to the digitized loop shaping controller.

\[
C_f(z) = \frac{1 + z}{2}
\]  

(7.16)
This is essentially a low pass filter which falls to zero at the Nyquist frequency of the axial control loop. Its frequency response is plotted in Figure 7.30.

![Frequency response of the discrete filter $C_f$ with a sample rate of 100 kHz.](image)

**Figure 7.30:** Frequency response of the discrete filter $C_f$ with a sample rate of 100 kHz.

This filter holds an advantage over single and multi-pole low pass filters in that it actually adds phase to the system. Since $C_f$ is non-causal, it cannot be implemented in a separate block and placed in series with the loop shaping controller. It must be included in the controller transfer function. Care was taken to ensure that there is at least one more pole than zero in the controller before the introduction of $C_f$ to maintain causality. The experimental result with $C_f$ implemented is shown in Figure 7.31. The resulting axial control error noise at low frequency is approximately 100 times improved and the RMS value is 4.8 nm. But, since the noise performance was not fully recovered compared to the single-tasking case, a model should be produced, in the future, to better understand the control loop jitter problem.
7.6.2 Rotary-axial motion coupling

In this section, rotary-axial motion coupling is investigated. Steps were taken in the design process to remove possible coupling between the motor and the thrust bearing, and also between the two degrees of freedom. However, upon operating the motor with the axial control loop closed, it was noticed that the positioning error increased significantly with spindle speed. The error signal was sinusoidal with the fundamental harmonic coinciding with the spindle speed. So, either a force disturbance, or a sensing disturbance, or a combination of the two was perturbing the system.

This problem was diagnosed by making use of the fact that the sensor disturbance to axial position transfer function will be distinct from the force disturbance to axial position transfer function. The resulting tracking error from a sensor disturbance $x_d$ is:
\[
\frac{x(s)}{x_d(s)} = \frac{1}{1 + \left(1 + \frac{\omega_n}{s}\right)C_{LS}(s)P_a(s)}
\] (7.17)

The closed loop force disturbance to position transfer function is the reciprocal of (7.12):

\[
\frac{x(s)}{F_d(s)} = \frac{1}{k_i^{ch} + \left(1 + \frac{\omega_n}{s}\right)C_{LS}(s)P_a(s)}
\] (7.18)

To make a comparison between these two transfer functions, the static disturbances, \(F_d\) and \(x_d\) must first be estimated. This can be done by rotating the spindle by hand while regulating the shaft position to a constant value. Then the shaft is incrementally rotated in 30° steps by hand. The average current supplied by the axial power amplifier is recorded at each angular position. Figure 7.32 shows the experimental result for this test. Based on this, the difference between the maximum and minimum currents, \(i_{max}\) and \(i_{min}\), is 0.03 A per channel. Now, a static force balance can be used to identify \(F_d\) and \(x_d\):

\[
F_d = k_i^{ch} (i_{max} - i_{min})
\] (7.19)

\[
x_d = \frac{k_i^{ch}}{k_x} (i_{max} - i_{min})
\] (7.20)

Making the substitutions of \(k_i^{ch} = 120\) N/A and \(k_x = 1.04 \times 10^6\) N/m gives \(F_d = 3.6\) N and \(x_d = 3.46\) μm. The next step is to substitute these results into (7.17) and (7.18). Now, \(x\) can be evaluated across a range of frequencies for both cases.
Figure 7.32: Change in current over two rotations of the spindle, at zero speed, when the axial controller regulates the shaft position to a constant value. The change in current is 0.03 A per power device channel.

To determine which, if either, of the two hypotheses is correct, the spindle was run at speeds from 1000 RPM to 8000 RPM and the axial error motion was measured. Figure 7.33 shows the experimental result compared with the two hypothesized models. The experimental data clearly indicates that the error at high speeds results from a disturbance acting on the position sensor.
Figure 7.33: Comparison between the model predicting an axial motion error caused by a sensor disturbance, the model predicting an axial motion error caused by position dependent force disturbance, and the experimental result. For this case, the difference between maximum and minimum currents was 0.03 A which corresponds to a static position disturbance of 3.6 μm. Speeds from 1000 RPM to 8000 RPM were tested.

Now that the type of error has been classified, it is easier to narrow the search for the root cause. The answer can be found if the problem is looked at from a metrology point of view. Figure 7.34a shows the ideal case where the target screw surface is perfectly perpendicular to both the axis of spindle rotation and to the capacitive probe. In reality, there is a manufacturing limitation on the achievable perpendicularity of the target face relative to the body of the fastener, and this results in a once per revolution error. This situation is illustrated in Figure 7.34b. The geometry of the problem is shown in Figure 7.35. The perpendicularity tolerance \( e_p \) was specified to be 12.7 μm (0.0005 in). The length of the bolt face, \( L_b \), is 11.1 mm. With the probe offset \( e_s \) from the shaft axis, and using the small angle approximation, the artifact error \( x_d \) (which is the same as the position disturbance mentioned above) is:
Using the 3.46 μm result from above, the probe offset can be estimated as 1.5 mm. Attempts were made to reduce this offset using shims, but only mild improvements were made.

Figure 7.34: Target screw metrology. a) In the ideal case, the sensing surface of the target screw is perfectly perpendicular to both the spindle rotation axis and the capacitive probe. b) In reality, there is a manufacturing tolerance on the perpendicularity. This limitation is the main source of axial error motion with a rotating spindle.
Figure 7.35: Target screw metrology geometry. $L_b$ is the width of the bolt head, $e_p$ is the perpendicularly tolerance, $e_s$ is the offset of the capacitive probe from the spindle shaft, and $x_d$ is the resulting artifact motion.

7.7 Performance comparison with aerostatic spindle

The results produced by replacing the aerostatic thrust bearing of the Precitech spindle with a high output electromagnetic actuator are a significant step forward in precision spindle technology. Table 7.4 summarizes some of the main performance parameters. The data in the table shows that the prototype rotary-axial spindle outperforms the standard spindle in spatial compactness, stiffness, and load capacity. It falls short in the area of axial error motion, but has the advantage of mm range axial stroke. The axial error motion issue will be addressed in section 8.1.8.
Table 7.4: Performance comparison between Precitech aerostatic spindle and rotary-axial spindle.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Aerostatic Spindle</th>
<th>Rotary-axial Spindle</th>
</tr>
</thead>
<tbody>
<tr>
<td>Thrust plate diameter</td>
<td>Ø75 mm</td>
<td>Ø50 mm</td>
</tr>
<tr>
<td>Thrust plate thickness</td>
<td>12.5 mm</td>
<td>12.5 mm</td>
</tr>
<tr>
<td>Axial static stiffness</td>
<td>70 N/μm</td>
<td>Infinite</td>
</tr>
<tr>
<td>Axial dynamic stiffness</td>
<td>25 N/μm</td>
<td>100 N/μm</td>
</tr>
<tr>
<td>Axial load capacity</td>
<td>180 N continuous, 360 N max</td>
<td>300 N continuous, 600 N max</td>
</tr>
<tr>
<td>Axial error motion</td>
<td>50 nm</td>
<td>&lt; 5 nm (static), speed dependent, see Figure 7.33</td>
</tr>
<tr>
<td>Axial motion stroke</td>
<td>N/A</td>
<td>1 mm</td>
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7.8 Summary

This chapter focused on the axial control and verification of the performance of the rotary-axial spindle. In the beginning, the open loop system was modeled. The model was updated to include vibration modes after a frequency sweep was performed. The first loop shaping controller achieved a crossover frequency of 1 kHz and was used for initial experiments.

Next, the actuator force characteristic was measured. The load capacity of the prototype is lower than expected, possibly due to un-modeled material properties. After this, Nanozoom was integrated into the system, which caused destabilizing feedback limit saturation. A workaround for this non-linearity was presented. In the end, 4.6 nm RMS position regulation noise was achieved. The effects of combined rotary-axial motion were also investigated. A target screw-based metrology solution is insufficient for ultra-precision applications due to manufacturing limitations of the target. It was also discovered that using interrupts to run two control loops at different rates decreases noise performance, but low pass filtering solves the problem.
In the final sections of the chapter, the performance of the axial control loop was improved. Initially, this was not possible due to current level saturation. However, a dual controller approach which maintains stability while switching from the startup controller to the high performance controller bypassed this issue and allowed further progress to be made. In the end 100 N/μm minimum dynamic stiffness, 2.6 kHz closed loop bandwidth, and 4.6 nm RMS position regulation noise was achieved. The experimental results indicate that the rotary-axial spindle is a significant step forward in precision spindle technology.
Chapter 8
Full Scale Rotary-axial Spindle

With the fantastic performance demonstrated by the first rotary-axial spindle prototype described in Chapters 3 through 7, the sponsors of this project commissioned UBC to lead the design, fabrication, and assembly of a production scale machine tool for use in silicon wafer grinding. The research presented in the preceding chapters is to be used as a guide. The overall machine architecture remains the same: two fluid bearings at the tool end of the spindle provide journal support while a high force magnetic bearing controls the axial position of the shaft. A hydrostatic design which offers significantly higher stiffness and load capacity replaces the aerostatic bearing used in the first prototype, and the thrust bearing has been resized, increasing the continuous force output 20 times.

Since this is an enormous undertaking, it was accomplished in collaboration with Dr. Alex Slocum and Gerald Rothenhofer from Massachusetts Institute of Technology (MIT), a team of support engineers from our sponsor, UBC undergraduate students Jeff Abeysekera, Thomas Huryn, Benito Moyls, Kris Smeds, and Irfan Usman, along with Dr. Lu and Matthew Paone. The MIT team designed the hydrostatic journal bearings. The crew of undergraduate students at UBC, supervised by Dr. Lu and Matthew Paone, designed, fabricated and tested the magnetic bearing, a 2 kW linear power amplifier, the Nanozoom PCB, and communication software. A photo of the first prototype and full sized prototype machines are shown in Figure 8.1.

The design stage began was from September 2007 to January 2009, some experimental work is still necessary. This chapter presents an overview of the results achieved to date. In
section 8.1, the machine requirements and its design are described. Section 8.2 discusses the 2 kW linear power amplifier and how the MIMO control scheme is extended to eight channel operation. The next sections, 8.3 and 8.4 discuss axial and rotary control. The hydrostatic bearing losses are measured in section 8.5. Section 8.6 presents the rotary-axial motion control results. Finally, the chapter is summarized in section 8.7.

Figure 8.1: Prototype and production sized rotary-axial spindles. The full sized machine is approximately 1 m long and weighs about 500 kg. Its force output is more than 25 times greater than that of the prototype.
8.1 Design

In this section, certain aspects of the full-scale rotary-axial spindle’s design are discussed. To begin, the design goals as given to us by the sponsor are presented. Next, the machine architecture is explained as it pertains to the spindle assemblies. Then, there will be a short discussion on the magnetic bearing which is capable of producing more than 6 kN of force. After this, the hydrostatic bearing and spindle shaft assembly is presented. Finally, the new metrology solution that should alleviate the rotary-axial motion coupling issue described in section 7.6.2 is presented in section 8.1.8.

8.1.1 Design goals

With silicon wafer grinding chosen as the intended end use, the machine specification in terms of stroke, axial load capacity, stiffness, precision, and cutting speed were chosen. These specifications are summarized in Table 8.1.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Unit</th>
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<td>Axial force</td>
<td>6000</td>
<td>N</td>
</tr>
<tr>
<td>Static axial stiffness</td>
<td>6000</td>
<td>N/μm</td>
</tr>
<tr>
<td>Axial precision</td>
<td>&lt; 10</td>
<td>nm</td>
</tr>
<tr>
<td>Axial stroke</td>
<td>1.5</td>
<td>mm</td>
</tr>
<tr>
<td>Maximum cutting speed</td>
<td>3000</td>
<td>RPM</td>
</tr>
</tbody>
</table>

Table 8.1: Full size rotary-axial spindle design requirements

These requirements present significant challenges. Firstly, the large axial force required will necessitate a large number of coil turns to produce a high current to force ratio. The associated problem is that the current state of the art power amplifiers, the PA52, only allow ±100 V supply, thus voltage level saturation may make it difficult to drive current at high
frequency. Moreover, total amplifier power dissipation limitations may be troublesome. The next challenge is related to the axial precision. Maintaining sub-10 nm resolution over 1.5 mm stroke requires a dynamic range of at least 150,000. This was easily achieved in the calm of the UBC laboratory where the prototype rotary-axial spindle was developed. However, this will not be easily achieved in a system which has heavy duty switch mode power supplies and amplifiers, and which also resides in an industrial environment with other noisy three phase devices such as pumps, cranes, and the like. Decoupling the rotary and axial motions to below 10 nm will produce additional fabrication and assembly challenges.

Figure 8.2: Rotary-axial spindle assembly section. This rotary-axial spindle contains four main assemblies: magnetic bearing assembly, DC motor assembly, hydrostatic bearing assembly, and spindle shaft assembly.
8.1.2 Architecture

A cross section view of the machine, which can be separated into four subassemblies, is shown in Figure 8.2. The magnetic bearing, which itself consists of three subassemblies, sits at the top end of the machine, and drives the armature which is mated to the spindle shaft. Atop the spindle shaft is a precision ball which acts as the sensing target. A permanent magnet rotor is fitted to the shaft, and coupled to a commercial DC brushless motor stator. At the bottom of the spindle lies the hydrostatic journal bearing assembly.

8.1.3 Magnetic bearing

The main difference between the actuator design criteria for grinding and that for fast tool servos is that here, the goal is to maximize the static force, and in fast tool servos acceleration becomes the focus. With this in mind, the magnetic bearing presented in Chapter 3 is scaled up so that more than 6000 N of force can be produced continuously. There are three subassemblies: the middle assembly which contains the permanent magnet ring and two stator assemblies (front and rear) which each contain a stack of coils.

A cutaway solid model and photo of the fully assembled permanent magnet ring is shown in Figure 8.3. Aside from the size increase, there are only three slight differences. The first is that 18 linearly magnetized permanent magnet segments are used, instead of 12. One effect that this will have is that there will be six additional discontinuities in the bias flux distribution in the radial air gap, but this effect should be negligible. The second change is that the middle stator segments are wider than the permanent magnet ring, allowing the armature to also be thicker than the magnets, and so flux leaking from the armature is reduced. To fill the space, two polycarbonate spacers with epoxy filled pockets sandwich the ring of magnets. Epoxy is easier to grind than polycarbonate, hence the fills. This represents the third variation.
Figure 8.3: Middle assembly. The stator segments are thicker than the permanent magnets, requiring magnet spacers. Pockets are machined out of the spacers and filled with epoxy to prevent the grinding wheel from “loading up” when the assembly is precision ground. The finished assembly is 315 mm in diameter by 27 mm thick.

Figure 8.4: Cutaway section of rear magnetic bearing assembly. The main components are the coil stack, the frame, the inner stator segments, and the outer stator segments. All these components are epoxied into place. The front assembly is similarly designed and assembled.

The front and rear magnetic bearing assemblies are described together, since the components pertaining to the actuator are mirror images of each other. A cutout section of the
rear assembly is shown in Figure 8.4. The twelve outer and twelve inner stator segments, made of powdered iron, are first epoxied into the AISI 304 frame. Next, the sixteen dually wound coils are epoxied into the stator. For the prototype rotary-axial spindle, the coils were wound by hand. Due to the number of turns and the aspect ratio, it is too difficult to do the same. To accomplish this task, I worked with a local winding shop, Beaver Electrical Machinery Ltd.\(^6\), to develop a manufacturing process for these coils. After the coils are secured in place, the excess epoxy is milled off, and the surfaces are ground flat. Two stacks of coils are shown in Figure 8.5a, and the finished front and rear actuator assemblies are pictured in Figure 8.5b. The characteristic parameters are shown in Table 8.2.

Figure 8.5: Coils and stator assemblies. a) Each stack consists of 16 dually wound pancake coils, and there are two stacks: one for the front assembly, and one for the rear. Each coil has two layers of 39 turns for a total of 78. Using AWG 19 wire results in the coils being 203 mm in diameter and 2 mm thick. b) The assembled and ground front (right) and rear (left) magnetic bearing assemblies. The assemblies are 315 mm in diameter.

\(^6\) www.beaverelectrical.com
Table 8.2: Magnetic bearing design parameters.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Permanent magnet remanence</td>
<td>1.3</td>
<td>T</td>
</tr>
<tr>
<td>Permanent magnet outer diameter</td>
<td>129</td>
<td>mm</td>
</tr>
<tr>
<td>Permanent magnet inner diameter</td>
<td>205</td>
<td>mm</td>
</tr>
<tr>
<td>Permanent magnet thickness</td>
<td>20</td>
<td>mm</td>
</tr>
<tr>
<td>Coil turns</td>
<td>78</td>
<td>Turns</td>
</tr>
<tr>
<td>Number of coils</td>
<td>32</td>
<td></td>
</tr>
<tr>
<td>Coil inner diameter</td>
<td>125</td>
<td>mm</td>
</tr>
<tr>
<td>Coil outer diameter</td>
<td>203</td>
<td>mm</td>
</tr>
<tr>
<td>Coil thickness</td>
<td>2</td>
<td>mm</td>
</tr>
<tr>
<td>Inner stator segment pole area</td>
<td>7,464</td>
<td>mm²</td>
</tr>
<tr>
<td>Outer stator segment pole area</td>
<td>12,240</td>
<td>mm²</td>
</tr>
<tr>
<td>Total air gap</td>
<td>2</td>
<td>mm</td>
</tr>
<tr>
<td>Armature diameter</td>
<td>126</td>
<td>mm</td>
</tr>
<tr>
<td>Armature thickness</td>
<td>25</td>
<td>mm</td>
</tr>
</tbody>
</table>

8.1.4 Actuator design for assembly considerations

Due to potential clamping forces on the order of thousands of Newtons generated when the middle assembly gets close to another soft magnetic material, extreme caution must be taken during the installation of the three magnetic bearing subassemblies. For this reason, design for assembly considerations were made. Backtracking to Figure 8.3 and Figure 8.5b, various holes can be seen situated around the perimeter of the stainless steel frames. Some of these are through holes and others are threaded. These holes can be used to lower the middle and rear assemblies onto the front assembly using threaded rods. This allows the assembly process to happen in a safe and controlled manner.
8.1.5 Brushless DC motor assembly

The brushless DC motor used on the machine is a commercial unit built by Motion Control Systems. The stator has three phases, commutated by ten HES, and the rotor has eighteen permanent magnet poles. Coupled to an AX6000 PWM amplifier, it can continuously produce 70 Nm of torque up to 3000 RPM. To install the stator into the housing, the stator must be cooled, or the housing must be heated due to an interference fit. To dissipate heat generated in the motor windings, a cooling jacket is integrated into the motor housing. The amplifier gain is 10.7 A/V, while the torque constant is 0.67 Nm/A.

8.1.6 Hydrostatic bearing

The hydrostatic bearing was designed at MIT by Dr. Slocum and Gerald Rothenhofer following the principles and guidelines presented in [20]. This technology is used because, water, being essentially incompressible, provides a much stiffer medium than does air. The tradeoff is that a considerable amount of heat is generated at the bearing interface. If left unchecked, the spindle shaft will eventually expand into the bearing surface, potentially causing a catastrophic failure. Taking this into account, the design shown in Figure 8.6 was produced.

Pressurized water (greater than 1.5 MPa) is forced into the bearing through the supply manifold and into the two cast ceramic hydrostatic bushings which are 170 mm long by 120 mm in diameter, each. These bushings accept a 120 mm (nominal) shaft, which is cylindrically ground to provide 10 μm of radial clearance. This setup has been tested to give 490 N/μm of radial stiffness [66]. After water clears the bushings, it is ideally sucked up through the vacuum manifold. If some water happens to pass this point, externally supplied pressurized air should

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7 www.motioncontrol.org
force it back into the suction ports. Rotary lip seals at each end of the assembly act as a final barrier to possible leaks. Finally, in order to dissipate the heat generated by the shearing of the water, the bearing housing is wrapped in a cooling jacket through which chilled water flows. Additionally, the tank from which the bearing water is drawn is also regulated at a constant temperature by a chiller.

![Hydrostatic bearing assembly diagram](image)

**Figure 8.6:** Hydrostatic bearing assembly. Water pressurized to at least 1.5 MPa is delivered to the bearing through the supply manifold. After passing through the two cast ceramic bushings, a suction port returns the water to a tank. Pressurized air, forces any air which makes it way passed the suction ports back. Rotary lip seals on either end of the assembly provide a final leak deterrent. A helical cooling jacket removes heat from the bearing housing, and the tank water is cooled, as well.

8.1.7 Moving assembly

The spindle shaft design is crucial since its dynamics can be a limiting factor when it comes to maximizing bandwidth and dynamic stiffness, as was demonstrated by the prototype
rotary-axial spindle. However, the basic shaft design was more or less optimized in the previous work. The fully assembled shaft is pictured in Figure 8.7. The major shaft diameter is 120 mm. In this case, the armature is not fastened to the shaft with a locknut, but is clamped axially by bolts which pass through the ball target flange and into a shoulder on the spindle shaft. This will prevent variation in the armature bending mode, as was possible on the prototype when the locknut loosened. Finite element analysis predicted the first and second bending modes at 745 Hz, 1560 Hz, a pure axial mode at 2890 Hz, an armature bending mode at 3800 Hz, and a mode where the neck of the ball target bends at 5200 Hz (Figure 8.8). The combined static stiffness of the shaft and housing is 12.5 kN/μm [67], assuming infinite static control stiffness.

Figure 8.7: Moving assembly. Mounted to the single piece shaft, is a permanent magnet rotor which is held in place with a locknut. The powdered iron armature is installed and clamped through the flange of the precision ball target. The bump-stop locknut is used to prevent the armature from hitting the stator. Its diameter is 120 mm, its overall length is 865 mm, and its mass is 80 kg.
Figure 8.8: Natural frequencies as predicted by COSMOSWorks [30] finite element solver. The first and second bending modes are at 745 Hz, 1560 Hz, a pure axial mode is at 2890 Hz, an armature bending mode is at 3800 Hz, and there is a mode where the neck of the ball target bends at 5200 Hz.

8.1.8 Metrology

One conclusion drawn from the previous chapter was that manufacturing limitations prevent the use of a flat surface as the sensing target. A new concept revolves around a sphere, or a ball target. A spherical target mounted to a spindle shaft, is not sensitive to rotational degrees of freedom when its axial position is measured with a probe, in contrast with a plane. The probe only sees three translational degrees of freedom. From this, it follows that if the shaft, the ball target, and the capacitive probe can be perfectly aligned, making them coaxial, then the sensor will see no rotationally induced artifact motion.
Another finding was that when the axial control loop was deactivated and the armature collided with the stator due to the negative spring effect, it was possible for the capacitive probe to shift axially, therefore posing a calibration problem. One potential solution is to provide a softer landing pad for the armature which would reduce the shock load applied to the sensor clamp.

![Figure 8.9: New rotary-axial spindle metrology solution using precision ball target and brass bumper. The precision ball target is insensitive to rotational degrees of freedom. If the spindle shaft, ball, and capacitive probe can all be made coaxial, then the effects of shaft rotation on axial sensing will be eliminated. A brass bumper absorbs impacts from the flange of the ball target and the locknut, reducing shocks on the sensor clamp.](image)

Based on these principles, a metrology system was designed as shown in Figure 8.9. The ball target was manufactured by Professional Instruments\(^8\), and the ball itself has a nominal roundness of 10 nm. Bolts pass through the ball target’s mounting flange, securing it and the armature to the spindle shaft. The bolt holes are opened wider than standard, leaving room to center the ball target to the spindle shaft. A brass bumper was also integrated into the setup to

\(^8\) www.airbearings.com
absorb axial impacts when the axial controller is deactivated. In one direction, the bumper is hit by the target’s flange. In the other, it is hit by the locknut which threads on to the neck of the ball target. Since there is a small gap between the bumper and the probe clamp, the sensor is removed from the force loop, which is highly desirable in any metrology solution.

One of the most challenging aspects of the entire project is centering the ball target to the spindle shaft. It is fairly easy to achieve sub-micron eccentricity tolerance initially. However, it is difficult to maintain it once the bolts are tightened. The assembly becomes further complicated as the bolts become tighter because the increasing friction between the armature and ball target flange makes it very difficult to make adjustments to the ball position. After several weeks of attempting to make the two components concentric to submicron levels, a limit of approximately 2 μm peak to valley was reached. The six bolts were tightened to a conservatively estimated 6 kN preload. The general ball target installation procedure is as follows. First, a simple thrust plate is mounted to the nose of the spindle to provide axial support. Before installing the target, lithium grease is applied to the bottom of the target’s flange and to the corresponding surface on the armature so that adjustments can still be made while the bolt preload is increased. An “L” bracket, with two capacitive probes installed in it, perpendicular to each other, is bolted to the spindle housing. This allows for the measurement of the eccentricity of the target’s axis with respect to the spindle’s (Figure 8.10). Next, the spindle shaft is rotated using the motor. Once the eccentricity is observed, the target can be adjusted accordingly, and the bolts can be tightened slightly. This approach, however, is far from ideal. Design for assembly principles should be applied here.
8.2 2 kW power amplifier

This section discusses the linear eight channel 2 kW power amplifier which drives current through the magnetic bearing. It is based on the same architecture presented in Chapter 4. However, some modifications have been made to handle the increased power output. In the first part of this section the modified hardware is shown. Afterwards, the current control results and performance are presented.

8.2.1 Power amplifier hardware

The two main pieces of hardware are the power stage (Figure 8.11) and the current control board (Figure 8.12). The power stage is essentially two four channel power amplifiers

Figure 8.10: Ball centering setup. Two capacitive probes are clamped in an “L” bracket to measure the eccentricity. The spindle is rotated using the motor, then the eccentricity is observed, and the target is adjusted by tapping the flange with a hammer. Finally, the bolts are incrementally tightened in a star pattern with a torque wrench.
which are driven in parallel to increase the overall power dissipation capability from 1 kW to 2 kW. Water cooling solutions were investigated, but concerns of leaks and difficulty in achieving an even temperature distribution across the plate led to the abandonment of this approach. So, the air cooling system was carried over, and an additional fan was added per plate.

Figure 8.11: Amplifier power stage. Two voltage control boards each control the voltage output of four power devices. Each 12.7 mm thick copper heat sink has ten Zalman CPU fans [34] attached to it to dissipate 2 kW, total. The current control board spans the voltage stage PCBs. This stacked layout was designed by Kris Smeds.

All components on the board are rated for ±100 V operation. The feedback resistors of the power stages were chosen such that the closed loop gain, $k_v$, was set to five. The input
circuitry to the voltage stage is designed to reduce the bandwidth to 200 kHz, thereby increasing stability. Another action that was taken to enhance stability was the removal of several large electrolytic capacitors that are used to supply high frequency current. Since capacitors can discharge current instantaneously, they can, in certain circumstances, unload huge currents which can destroy the fuses. This is the reasoning for their removal.

Since this is an eight channel amplifier, the MIMO system has eight input voltages and eight output currents. The current control board, shown in Figure 8.12 and designed by Kris Smeds, implements the theory, presented in section 4.4. The option of adding a small plastic capacitor to the feedback resistor of the voltage reference summing stages was incorporated to attenuate the inevitable parasitic capacitance that is sure to result from the coils that act as large parallel plates.

![Figure 8.12: PCB for eight channel current control. This PCB was designed by Kris Smeds.](image-url)
The current control block diagram is repeated in Figure 8.13.

\[
\begin{bmatrix}
Z_0 & Z_1 & Z_1 & Z_2 & Z_2 & Z_2 & Z_2 \\
Z_1 & Z_0 & Z_1 & Z_2 & Z_2 & Z_2 & Z_2 \\
Z_1 & Z_1 & Z_0 & Z_1 & Z_2 & Z_2 & Z_2 \\
Z_2 & Z_2 & Z_2 & Z_0 & Z_2 & Z_2 & Z_2 \\
Z_2 & Z_2 & Z_2 & Z_2 & Z_0 & Z_2 & Z_2 \\
Z_2 & Z_2 & Z_2 & Z_2 & Z_2 & Z_0 & Z_0 \\
\end{bmatrix} (8.1)
\]

Where \( Z_0 \) is the load self-impedance, \( Z_1 \) is the closely coupled impedance, and \( Z_2 \) is the loosely coupled impedance. The input decoupling matrix \([V]\) is:

\[
\begin{bmatrix}
1 & -1 & -1 & 0 & -1 & 0 & 0 & 0 \\
1 & -1 & -1 & 0 & 1 & 0 & 0 & 0 \\
1 & -1 & 1 & 0 & 0 & -1 & 0 & 0 \\
1 & -1 & 1 & 0 & 0 & 1 & 0 & 0 \\
1 & 1 & 0 & -1 & 0 & 0 & -1 & 0 \\
1 & 1 & 0 & -1 & 0 & 0 & 1 & 0 \\
1 & 1 & 0 & 1 & 0 & 0 & 0 & -1 \\
1 & 1 & 0 & 1 & 0 & 0 & 0 & 1 \\
\end{bmatrix} (8.2)
\]

The synthesized feedback, which includes the feedback matrix \([W]\) is:
The current sensing resistor, \( R_s \), has 0.2 \( \Omega \) resistance. The next step is to measure the eight by eight plant matrix, design controllers, and assess the stability of the system.

### 8.2.2 Coil configurations

In Chapter 4, only one possible coil configuration was investigated. However, the 32 coils used in the full-sized actuator can be arranged into three basic configurations. These are shown in Figure 8.14. The first is a pure parallel mode in which four coils are connected in parallel. The second is a mixed parallel-serial mode and is used on the prototype rotary-axial spindle. In this configuration, one pair of coils connected in parallel, are serially connected to another pair of coils which are also connected in parallel. The third is a pure serial mode.

Given an amplifier without power limitations, there would be no advantage in using one configuration over another with regards to load capacity and driving high frequency current, which are the two characteristics of interest. High force is required to achieve fast feed rates during cutting and a certain amount of high frequency current driving capability is required to avoid introducing non-linearities into the motion loop while attacking high frequency disturbances. Obviously, the power amplifier has limitations on both the maximum voltage and current outputs. Force production is related to the number of ampere-turns that can be generated, and the ability to drive high frequency current is related to the load inductance.

\[
\begin{bmatrix}
1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 \\
-1 & -1 & -1 & -1 & 0 & 0 & 0 & 0 \\
-2 & -2 & 2 & 2 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & -2 & -2 & 2 & 2 \\
-4 & 4 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & -4 & 4 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & -4 & 4 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & -4 & 4
\end{bmatrix}
\begin{bmatrix}
I_d \\
I_s \\
\vdots \\
I_{H_k}
\end{bmatrix} = \frac{1}{4}
\begin{bmatrix}
1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 \\
-1 & -1 & -1 & -1 & 0 & 0 & 0 & 0 \\
-2 & -2 & 2 & 2 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & -2 & -2 & 2 & 2 \\
-4 & 4 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & -4 & 4 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & -4 & 4 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & -4 & 4
\end{bmatrix}
For a given current $i_{ch}$ delivered by a power device, pure serial mode delivers the highest force, parallel-serial mode delivers half this amount, and pure parallel mode delivers one quarter. This is due to the total number of ampere-turns. On the other hand, the parallel configuration results in the lowest inductance, the parallel-serial mode inductance is four times larger, and the pure serial mode is sixteen times larger. So far, the mixed configuration provides a good balance. It is always possible to switch to either parallel mode if voltage saturation becomes an issue or to serial mode if the load capacity is insufficient.

### 8.2.3 Parallel-serial coil configuration control

In this section, the current control results for the mixed parallel-serial coil configuration are presented. A first measurement of the common mode current control plant made it
immediately clear that two phenomena which had only mild effects on the four channel amplifier would now become significant: parasitic capacitance and voltage level saturation. The frequency at which the parasitic capacitance became dominant was now 7 kHz, and the lumped inductance seen by each power device \((Z_0 + 3Z_1 + 4Z_2)\) was a rather large 326 mH. This measurement is shown in Figure 8.15.

![Experimentally measured lumped impedance. The inductance at low frequency is 326 mH and the parasitic capacitance becomes dominant near 7 kHz.](image)

**Figure 8.15:** Experimentally measured lumped impedance. The inductance at low frequency is 326 mH and the parasitic capacitance becomes dominant near 7 kHz.

By comparison, the corresponding values for the 1 kW amplifier described in Chapter 4 were 130 kHz and 4 mH. When looking at the inductance, this is a difference of nearly two orders of magnitude. The parasitic capacitance will limit the achievable bandwidth of the common mode current loop, and the large inductance, a consequence of the quest for large load capacities, will seriously limit the ability to drive even hundreds of mA faster than a few hundred Hz. Tests showed that the parasitic capacitance does not change significantly for any given coil
combination. So, it was decided to stay with the parallel-serial mode so that a higher load capacity could be achieved.

After analyzing the load, the decoupled current control plant was identified. This required measuring 64 frequency responses. The decoupling results shown in Figure 8.16, though not as good as achieved with the four channel amplifier, are acceptable. The common mode diagonal entry ($v_1$ to $i_{1S}$) was dominant up to only 155 Hz, and the response from $v_2$ to $i_{2S}$ was only dominant to 200 Hz. However, the other six diagonal entries were dominant across a 50 Hz to 40 kHz range. Though, the effect is even stronger below 1 kHz. Additionally, all differential inputs ($v_2$ to $v_8$) are extremely well decoupled from the average current output, $i_{1S}$. This means that the differential inputs have no affect on the dynamic response of the average current. So, the approach to take in the controller design is to keep the bandwidths of the differential modes on the order of a few hundred Hz to keep the control frequencies in the region of stronger decoupling, and design the bandwidth of the common mode controller as high as possible.

The analog controllers themselves retain the same PI plus low pass filter topology. However, the option of placing an additional capacitor across the feedback resistor of the voltage reference summing stage to help attenuate the parasitic capacitance was taken up. This pole is set at 5 kHz. The resulting negative loop transmissions are compared with the eigenvalues in Figure 8.17. The agreement is acceptable, though not as good as previously achieved. This is related to the decoupling results. A summary of the unity gain crossover frequencies and corresponding phase margins for each negative loop transmission and eigenvalue locus is given in Table 8.3.
Figure 8.16: Decoupling results for eight channel amplifier.
Figure 8.17: Comparison between negative of the loop transmission of diagonal entries (blue) and eigenvalues of full order matrix (black).
Table 8.3: Summary of eight channel amplifier stability

<table>
<thead>
<tr>
<th>Loop with output</th>
<th>Neg. Loop Transmission Criteria</th>
<th></th>
<th>Eigenvalue Criteria</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>X-over [Hz]</td>
<td>PM [deg]</td>
<td>X-over [Hz]</td>
</tr>
<tr>
<td>$I_{1S}$</td>
<td>3700</td>
<td>53</td>
<td>3700</td>
</tr>
<tr>
<td>$I_{2S}$</td>
<td>174</td>
<td>80</td>
<td>325</td>
</tr>
<tr>
<td>$I_{3S}$</td>
<td>244</td>
<td>60</td>
<td>195</td>
</tr>
<tr>
<td>$I_{4S}$</td>
<td>244</td>
<td>60</td>
<td>180</td>
</tr>
<tr>
<td>$I_{5S}$</td>
<td>174</td>
<td>80</td>
<td>200</td>
</tr>
<tr>
<td>$I_{6S}$</td>
<td>174</td>
<td>83</td>
<td>195</td>
</tr>
<tr>
<td>$I_{7S}$</td>
<td>174</td>
<td>83</td>
<td>138</td>
</tr>
<tr>
<td>$I_{8S}$</td>
<td>174</td>
<td>80</td>
<td>220</td>
</tr>
</tbody>
</table>

Now, the performance of the MIMO control system can be evaluated. The power supply voltage rails are set to ±80 V. This is the highest possible achievable output with the combination of a voltage gain of five, while powering the current control board with ±15 V. The current limit is conservatively set at ±2.5 A/channel to limit the overall power dissipation. A series of closed loop responses from the reference average current to the actual average current are shown in Figure 8.18. A fully linear response can be achieved when the input signal is 0.02 A peak to peak or smaller. The corresponding -3 dB bandwidth is about 4 kHz. Above this amplitude there is significant voltage level saturation. This is demonstrated further in the step responses plotted in Figure 8.19 and Figure 8.20.
**Figure 8.18:** Closed loop response of average current to different common mode reference amplitudes. Voltage level saturation is evident even at small amplitudes.

**Figure 8.19:** 0.2 A step response. The individual currents do not converge as quickly as the average current response because the bandwidths of the differential controllers are only around 200 Hz. A fully saturated response is already evident at this step size.
8.2.4 Pure parallel coil configuration control

Since there was such extreme saturation evident in the dynamic response of the average current, it was decided to switch to the pure parallel coil configuration to attempt to improve the high frequency performance. Although the inductance drops by a factor of four, the current to force ratio is cut in half. Thus, the overall ability to produce dynamic forces is amplified by a factor of two. However, to match the same static load capacity with that of the parallel-serial configuration, the current supply limits must be doubled to ±5 A/channel. Unfortunately, with the voltage supply at ±80 V, this would exceed the PA52 400 W power dissipation limit. For now, this compromise will be conceded. The proportional gains of the current controllers are easily adjusted to compensate for the change in inductance.
8.3 Axial motion control

This section deals with the axial control of the full scale rotary-axial spindle. First, the actuator force characteristic is identified. Then, the axial control plant, controller, and performance are described. The power amplifier is set to drive a load in the pure parallel configuration.

8.3.1 Actuator force characterization

The actuator was identified using a load cell to measure the force, a capacitive probe to measure the axial position, and threaded rods to slide the spindle shaft across the stroke. The result is plotted in Figure 8.21. From here, the magnetic bearing negative stiffness and current per channel to force ratio are identified as 3.14 N/μm and 1750 N/A, respectively. Another interesting phenomenon is the fairly consistent 300 N hysteresis which comes from friction caused by relative motion between the spindle shaft and the lip seals. Most importantly, it shows that by driving at least 2 A per channel in parallel mode, then the target of 6,000 N is achieved. If the voltage supply rails are lowered, then the full 5 A/channel can be driven through the load and the actuator should be able to output 10-12 kN. This is a promising result.
Figure 8.21: First quadrant force characteristic. The bearing negative stiffness and channel current to force ratio are approximately 3.1 N/μm and 1750 N/A, respectively. The 300 N hysteresis is due to lip seal friction.

8.3.2 Axial plant identification and controller design

After stabilizing the axial control plant using a controller that was designed based on the actuator identification, the open loop frequency response of the axial plant was measured using swept sine waves from current command to shaft position. The result is shown in Figure 8.22. At low frequency, there is a static stiffness exhibited which is caused by the flexibility of the lip seals. This is only visible at small excitation amplitudes where the friction does not come into play. The stiffness is estimated at 92 N/μm. At 150 Hz, there is a resonance which occurs due to the interaction between the lip seal and shaft mass. Following this, from 200 Hz to 600 Hz, the
response behaves as a free rigid body. Above 1 kHz, there exists a pair of collocated bending modes which is followed by a non-collocated resonance at 2800 Hz. This resonance limits the achievable bandwidth. The phase drops off rapidly after 2 kHz due to additional resonances and a 2nd order roll off of the actuator bandwidth which is caused by parasitic capacitance.

Figure 8.22: At low frequency, the non-linear flexibility of the lip seals dominates. At 150 Hz, there is a resonance resulting from this stiffness and the shaft mass. From 200 Hz to 600 Hz, the response behaves like a free rigid mass. After this, there exist several natural frequencies, followed by a non-collocated mode at 2800 Hz. The pump supply pressure was 1.8 MPa.

Based on this, a loop shaping controller was designed and its frequency response is plotted in Figure 8.23a. It consists of an integrator which has effective range up to 200 Hz which provides infinite static stiffness, three phase compensators to increase the phase at the intended crossover frequency, notch filters at 1700 Hz, 2900 Hz, 4000 Hz, and 5000 Hz to attenuate resonances, and a pair of low pass filters with poles at 1300 Hz and 2400 Hz to roll off the high
frequency response. Keeping the integrator separate, the loop shaping controller transfer function is:

\[
C_{LS}^2 = \frac{(s + 880)(s + 1000)(s + 1400)}{(s + 8168)(s + 15080)(s + 21990)(s + 25130)(s + 28270)} \times \frac{s^2 + 214s + 1.14e8}{s^2 + 2.14e4s + 1.14e8} \times \frac{s^2 + 364s + 3.32e8}{s^2 + 3.64e4s + 3.32e8} \times \frac{s^2 + 5030s + 6.32e8}{s^2 + 2.51e4s + 6.32e8} \times \frac{s^2 + 3770s + 9.87e8}{s^2 + 6.28e4s + 9.87e8}
\] (8.4)

All this results in a 680 Hz unity gain crossover frequency with a phase margin of 31° as demonstrated in Figure 8.23b. Figure 8.23c demonstrates an impressive low frequency dynamic stiffness of 340 N/μm. The minimum dynamic stiffness at 2800 Hz in this test is still greater than 100 N/μm. Impact hammer tests at the spindle nose have shown this mode to have stiffness greater than 300 N/μm (Figure 8.24). The -3 dB bandwidth is estimated at 800 Hz as shown in Figure 8.23d. For initial startup, a less aggressive controller is needed. The transfer function for this startup controller is:

\[
C_{LS}^{start} = \frac{(s + 540)(s + 942)}{(s + 3770)(s + 4400)(s + 6600)} \times \frac{s^2 + 3770s + 3.55e8}{s^2 + 3.77e4s + 3.55e8}
\] (8.5)

\(C_{LS}^{start}\) operates in series with an integrator that has a 30 Hz corner frequency. Its crossover frequency is 180 Hz with 60° phase margin. The whole control loop is executed at 100 kHz on the dSPACE DS1103.
**Figure 8.23:** Loop shaping controller design results. a) Loop shaping controller. b) Negative of the loop transmission. c) Dynamic stiffness. d) Closed loop frequency response.
Without rotation, the position regulation noise resulting from the controller given in (8.4) and its integrator is 7 nm RMS (Figure 8.25a), with the help of Nanozoom. This corresponds to a dynamic range of 215,000. This result was very challenging to achieve since there are several additional electrical and mechanical disturbances in the system such as the switch mode DC power supplies, pressure fluctuations resulting from the hydrostatic bearing pump, and line noise from a pair of chillers. However, the power spectrum of the noise (Figure 8.25b) shows that there are low frequency components which are not insignificant. This is likely due to an integrator corner frequency of only 200 Hz, and the non-linearity introduced by the lip seals.
Figure 8.25: Position regulation noise. a) The RMS error is 7 nm RMS. b) The power spectrum shows increased noise in the low frequency range, which is likely due to a lower integrator corner frequency and non-linearities introduced by the lip seals.

Figure 8.26: Stroke demonstration and tracking performance in air cutting. a) Eighty percent stroke sweep at a feed rate of 240 μm/s. b) The average tracking error is 10.5 nm and the noise about this point is 11.7 nm RMS. Feed forward compensation of the friction will improve this result.

Figure 8.26a and Figure 8.26b show a demonstration of the spindle travelling across 80% of its designed stroke, and the error during one portion of the axial feed. At a 240 μm/s feed rate,
a 10.5 nm steady state error exists, and the RMS error about the mean is 11.7 nm. The steady state error is caused by the lip seal friction. If a constant current reference corresponding to the friction force is fed forward, then the error can be compensated. However, this has not been attempted as of the time of writing.

8.4 Rotary motion control

This section describes the rotary controller design for the full size rotary-axial spindle prototype. After describing modifications made to the sensorless rotary motion feedback algorithm, the rotary plant identification and controller results are presented.

8.4.1 Sensorless rotary motion feedback for ten HES motor

Some modifications were made to the rotary motion feedback algorithm because the chosen brushless motor has ten HES, instead of three. So, certain limitations are reached in the VHDL code and IP cores. The ten HESs produce 180 counts/rev, and 128 interpolations are made between each count. The result is 23,040 counts/rev. Using a 16 bit timing counter and taking advantage of the 150 MHz clock on the PCB, this makes the minimum and maximum measurable speeds approximately 15 RPM and 7800 RPM. This covers the desired speed range well.

8.4.2 Rotary plant identification and controller design

A slightly different approach was taken to the rotary speed control of the full-size prototype compared to the first prototype. Since the rotary amplifier can be operated in torque mode, it is desirable to have the transfer function from open loop input to torque output as unity. This means that the input to the open loop system is in fact a torque reference, which gives the
controller output a more intuitive physical meaning. Figure 8.27 illustrates how this is accomplished. The input to the system is the torque command $T_r$. This signal is scaled by 2/3 and by the reciprocal of the torque constant, $k_t$ to produce the quadrature current, $i_q$. The quadrature current is further divided by the rotary amplifier gain $k_a$ leaving the current reference $i_r$. Next, the DAC gain $k_d$ is cancelled by its reciprocal. When the output of the DAC, which is also $i_r$, is fed to the rotary drive, it is amplified to produce the quadrature current $i_q$. The DC motor then applies a torque $T_m$ to the shaft inertia, $J_{shaft}$ which results in angular motion of the shaft, $\theta$. The output motion is measured by the sensorless rotary motion algorithm, resulting in the quantized angular position signal, $\theta_m$. To generate the speed, signal $\omega_{int}$, $\theta_m$ is digitally differentiated over the sampling interval $T_s$. A final scaling by 60 and $k_e$ converts the speed to RPM. The parameters in the open loop model are given in Table 8.4.

**Figure 8.27:** Rotary control block diagram. The input to the system is a torque reference, $T_r$, and the output is the speed $\omega$ as measured by the sensorless algorithm.
Table 8.4: Full-size rotary-axial spindle prototype rotary control loop parameter list

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>$k_t$</td>
<td>0.44</td>
<td>Nm/A</td>
</tr>
<tr>
<td>$k_a$</td>
<td>10.7</td>
<td>A/V</td>
</tr>
<tr>
<td>$k_d$</td>
<td>0.1</td>
<td>V/V</td>
</tr>
<tr>
<td>$k_e$</td>
<td>5760</td>
<td>lines/rev</td>
</tr>
<tr>
<td>$T_s$</td>
<td>0.0001</td>
<td>s</td>
</tr>
<tr>
<td>$J_{shaft}$</td>
<td>0.187</td>
<td>kg-m$^2$</td>
</tr>
</tbody>
</table>

Figure 8.28 shows the frequency response of the open loop rotary control plant. At low frequency, the response indicates that the spindle rotor is in rigid body motion. The rotor inertia can be extracted from this portion of the curve. It is 0.187 kg-m$^2$. However, at 30 Hz, a zero occurs, and the response tends upward with +20 dB/dec slope. Although this phenomenon was observed on the first prototype rotary-axial spindle, this is a slightly strange result. In the full-size prototype, the stator is shrunk fit into the machine frame, and the HESs are mounted on the stator coils. In theory, this should create a very stiff joint and a solid backbone for mounting the HESs. It would be reasonable to expect the notch to occur at a higher frequency, if at all. On the other hand, the inertias in this system are much larger than on the prototype rotary-axial spindle. One plausible explanation is that the flexibility results from the finite stiffness of the motor windings themselves.
**Figure 8.28:** Frequency response of rotary plant from torque command to speed. A rigid body mode corresponding to the spindle shaft inertia is dominant in the low frequency region. Above 30 Hz, the response indicates that the HES motion is being measured instead of that of the rotor.

The feedback filters which attenuate this effect, and the PI speed regulator $C_r$, can now be designed. These items are placed in the control loop as shown in Figure 8.29. Seeing that the zero frequency does not vary appreciably from the case of the first spindle prototype, the same feedback filters are used and are repeated here:

\[
F(s) = \left(\frac{\omega_{bw}^2}{s^2 + \omega_{bw}s + \omega_{bw}^2}\right)^2
\]

(8.6)

\[
A(z^{-1}) = \frac{1}{N_r} \times \frac{1-z^{-N_r}}{1-z^{-1}}
\]

(8.7)

where $\omega_{bw}$ is 1250 rad/s and $N_r$ is 205. The proportional gain is adjusted to set the crossover frequency at 10 Hz. The integrator which has its corner frequency ten times lower than crossover will eliminate the effects of the lip seal friction. The negative of the loop transmission for this
plant, filter, and controller combination is shown in Figure 8.30a. The minimum dynamic stiffness to torque disturbances is 0.56 Nm/RPM and the closed loop bandwidth is 14 Hz. These quantities could both be improved considerably if the stiffness of the HES mounting point could be enhanced.

![Rotary motion closed loop block diagram](image)

**Figure 8.29:** Rotary motion closed loop block diagram.
Figure 8.30: Rotary control loop results. a) Negative loop transmission. b) Dynamic stiffness. c) Closed loop bandwidth.

The spindle was run at speeds from 200 RPM up to 3000 RPM without dynamic balancing. A short time scale capture of the spindle running at 3000 RPM is shown in Figure 8.31 which shows a harmonic at approximately 115 Hz. This disturbance is likely caused by the lip seal resonance, and limits the RMS performance of the rotary speed loop to 0.293 RPM.
Figure 8.31: Speed regulation at 3000 RPM. The RMS error is 0.293 RPM. The 115 Hz harmonic is related to the pressure fluctuations from the hydrostatic bearing pump.

8.5 Hydrostatic bearing losses

Although hydrostatic bearings are superior to other bearing technologies when it comes to stiffness, one of their major drawbacks is heat generation from viscous losses. In hydrostatic bearings shear forces are produced when there is relative motion between the bearing medium (water) and the spindle shaft. This is known as viscous friction, and it is speed dependent. Moreover, the static friction generated by the lip seals adds to the heat loss. Using a lever arm and a force gauge, the static friction torque was measured as 8 Nm. To check this result the friction torque was estimated using the hysteresis force measured in Figure 8.21. Half of this force, approximately 130 N to 150 N, accounts for the friction force. Then, using the shaft radius \( R \) as the moment arm, the friction torque \( T_f \) can be calculated as:

\[
T_f = R \times F_f
\]  

(8.8)
The spindle shaft has a radius of 60 mm. This places $T_f$ somewhere between 7.8 Nm and 9 Nm. This is consistent with the lever arm-force gauge based measurement.

The viscous friction properties of the hydrostatic bearing were characterized by recording the steady state torque at a steady state speed. This result is shown in Figure 8.32a. As expected, the friction torque is approximately a linear function of the spindle speed. Using a least square fit, the viscous damping coefficient is estimated as 0.0063 Nm/RPM. The product of the friction torque and spindle speed gives the power loss. Figure 8.32b shows that in the 2500 RPM to 3000 RPM range, more than 5 kW of cooling are required. Not providing adequate cooling not only affects the precision of the manufacturing operation, but it also is a major reliability concern. For example, the shaft and bearing may experience thermal expansion. Eventually, they may grow into each other which could potentially destroy the spindle. It is, therefore, crucial that the spindle housing be kept at a stable temperature by an effective water cooling system.

$\text{Figure 8.32: Hydrostatic bearing losses caused by fluid shear. a) Average torque loss vs. spindle speed. b) Average power loss vs. spindle speed.}$
8.6 Rotary-axial motion coupling

Now that both the axial and rotary control loops have been compensated, the capacitive probe can be adjusted in such a way that the axial error motion of the spindle shaft is minimized, if not eliminated. Compared to centering the ball target this task is slightly easier because the probe clamp need not be fastened into place with such a large preload. It is still a challenging task, nonetheless. The method for finding the optimum position is as follows. First, both control loops are activated. An axial controller with low bandwidth is used so that the artifact motion is not reduced by the controller. Next, with the spindle rotating, the axial position is monitored as the axial probe holder is laterally adjusted with light taps from a rubber hammer. After tightening the probe holder bolts slightly, the process is repeated until the clamp is secured and the error motion is acceptable. Figure 8.33 shows that after proper lateral adjustment of the probe holder, the axial error motion at 2500 RPM is only 40 nm peak to valley. This is a significant improvement over the target screw based metrology. Making the target centering a more deterministic process is sure to yield improved results.
Figure 8.33: Axial error motion with ball target metrology system. The result indicates the error motion is below 40 nm at 2500 RPM. A filter was used to remove line and high frequency noise from the raw signal.

8.7 Summary

This chapter presented the results of a collaborative effort to design, fabricate, assemble, and test a full sized rotary-axial spindle prototype. In section 8.1, an overview was given of the magnetic bearing design, the hydrostatic journal bearing, and a new metrology solution based on a spherical target. Next, in section 8.2, the eight channel 2 kW linear amplifier was discussed. After showing that the decoupling theory is effective on the eight input eight output impedance load, it was shown that the Nyquist phase margin criteria is sufficient for assessing the stability of the MIMO control system. It was also found that the $\pm 80$ V supply rails are insufficient for linear operation above a few hundred Hertz with the mixed parallel-serial coil configuration. In section 8.3, the force characteristic of the magnetic bearing was presented. With the coils in a pure parallel configuration the actuator can produce 6 kN, but in the parallel-serial mode, the
axial load capacity can potentially be raised to 12 kN. After this, the axial motion control plant was measured and a loop shaping controller was designed. It results in impressive performance: 340 N/μm minimum dynamic stiffness, 800 Hz closed loop bandwidth, and 7 nm RMS positioning noise with Nanozoom. Next, in section 8.4, the rotary control loop was compensated. Since the brushless motor used for this machine has 10 HES, the sensorless feedback resolution can be improved to 23,040 counts/rev. However, pressure fluctuations from the hydrostatic bearing pump limit the speed regulation noise to under 0.3 RPM RMS. In section 8.5, hydrostatic bearing losses were identified. More than 5 kW cooling capacity is needed above 2500 RPM. Finally, section 8.6 showed that the spherical target based metrology is a significant improvement over the screw based design in rotary-axial spindle applications. At 2500 RPM, the error motion is only 40 nm.
Chapter 9
Conclusions and Future Work

This thesis presents the foundation for significant improvements to precision spindle technology. This chapter summarizes the primary conclusions, and recommends areas for future work.

9.1 Conclusions

Over the course of this research, several important findings were made. They are summarized in this section.

9.1.1 Rotary-axial spindle concept

A new spindle concept was presented (section 2.4) which addresses the need for higher stiffness and load capacity in manufacturing operations requiring mm range feed stroke. This architecture sidesteps the fundamental limitation to improved stiffness found in stacked stage layouts by removing the feed stage and spindle thrust bearing, and replacing it with a novel high force magnetic thrust bearing with mm range stroke.

9.1.2 Design and analysis of rotary-axial spindle

Chapter 3 presented the analysis of the magnetic bearing and its design and integration with the rest of the rotary-axial spindle. The permanent magnet biased actuator has a fully linear force characteristic and mm range stroke. The key features which lead to the linearity are constant bias flux, constant total air gap, and differential fluxes and subtractive forces on each
side of the armature. A dynamic study was performed on the shaft assembly in order to predict its natural frequencies. However, a comparison between the results of this study and experimental frequency sweeps did not provide much agreement. It seems, therefore, that experimental methods such as impact hammer testing are a better means of determining the mode shapes of the structure. One other lesson learned was that design for assembly principles should be applied to the magnetic bearing installation for a safer, more controlled assembly process.

9.1.3 MIMO current control

In Chapter 4, a current control law for multi-channel power amplifiers with coupled loads was designed and tested. Due to the simplicity of the algorithm, an analog implementation using operational amplifiers is preferred over digital control. The effectiveness of the decoupling scheme and stability criteria were experimentally verified. It was determined that if the decoupling is sufficient, then perfect decoupling can be assumed for the purposes of controller design and stability assessment. It was also discovered that the combination of finite voltage supply rails and large load impedance introduce significant slew rate saturation of the output current for fast large signal commands.

9.1.4 Sensorless rotary speed measurement

Chapter 5 presented a high resolution sensorless rotary motion feedback algorithm for spindle speed control. It proved to be as effective as a 1000 line/rev optical encoder when regulating the average spindle speed. Three processes must be executed to obtain this result. Firstly, a coarse position measurement must be acquired using the Hall Effect sensors. Next, constant speed is assumed and interpolation is applied. Finally, the locations of the Hall Effect
sensor edges are correctly placed in the spatial domain through a calibration process. The downside of this approach is that a limitation is placed on the achievable speed control bandwidth. This limitation depends on the stiffness of the member to which the Hall Effect sensors are mounted.

9.1.5 Novel method for increasing ADC resolution

To obtain sub-10 nm motion resolution, a novel method for increasing the resolution of an analog to digital converter over the entire input range, without increasing the conversion delay, was presented in Chapter 6. The dynamic range was improved from 100,000 to 300,000 corresponding to a 3 nm RMS noise floor. The key to achieving this result was to move the control error subtraction process to the analog domain. A noise model of the analog electronics required to implement this invention was developed. It was shown that neglecting thermal noise generated by resistors produces a sufficiently accurate result.

9.1.6 Axial control and experimental results of first prototype

Chapter 7 presented the thorough exploration of the axial control system. The potential of the rotary-axial spindle was demonstrated by achieving infinite axial static stiffness, 100 N/μm minimum axial dynamic stiffness in the low frequency region, sub-5 nm RMS positioning noise, dynamic range over 220,000, and 2.6 kHz closed loop bandwidth. The key to achieving this performance is through the iterative design of loop shaping controllers. This form of control law allows great flexibility and the use of notch filters which attenuate the controller’s response at structural modes. One disadvantage of this approach is that there is limited immunity to non-linearities. This characteristic was brought to light when current level and error range saturations
were encountered. Switching between controllers of low and high stiffness and between conventional and high resolution feedback paths provided a solution to these two problems.

Two additional phenomena were discovered while investigating combined rotary and axial control. The first was additional noise introduced by sampling rate jitter. A solution, in the form of a discrete low pass filter, was developed to attenuate most of this noise. The second issue was related to the metrology solution. It was discovered that limitations on the achievable perpendicularity between the spindle shaft and face of the sensing target introduced a sinusoidal axial motion error which was periodic with spindle speed. With the given design, this could not be compensated for.

9.1.7 Full-size prototype

Based on the discussions in Chapters 2–7, a full size rotary-axial spindle was designed, manufactured and tested (Chapter 8). It demonstrated infinite axial static stiffness, 340 N/μm minimum axial dynamic stiffness in the low frequency region, 7 nm RMS positioning noise, 1.5 mm stroke, and 800 Hz closed loop bandwidth. Although a 6 kN thrust capacity was measured, many coil turns were required resulting in an extremely large inductance left to be handled by the power amplifier. This introduced significant voltage level saturation to the current control loops, which could pose problems for attacking high frequency disturbance forces. From a metrology point of view, a new precision ball target solution showed that the axial error motion at a shaft speed of 2500 RPM was only 40 nm. This is a significant improvement over screw-based metrology. Finally, the viscous friction characteristics of the hydrostatic journal bearing were identified. By recording the average motor torque and spindle speed, it was estimated that 7 kW cooling capacity is required at an operational speed of 3000 RPM.
9.2 Future work

Future development of the rotary-axial spindle can proceed in many different directions. Keeping in mind that the ultimate goal is to utilize the technology in silicon wafer grinding, and based on the conclusions discussed in the previous section, potential research and development of the rotary-axial spindle are proposed.

9.2.1 High power linear amplifiers

In both the four and eight channel power amplifier, voltage level saturation was evident, especially in the case of the latter. Fuller use of the PA52 amplifier can be made with water cooling, but since this device is already the state of the art, a “homemade” design may be required to make significant gains. Such a power device should output several hundred volts, should potentially dissipate 1 kW continuous reactive power, and should be capable of working in parallel with similar devices. Significant challenges would be presented in this research in the areas of amplifier compensation, manufacturing, cooling, and safety. Switch mode topologies should not be eliminated outright.

9.2.2 Shaft dynamic study

In both the prototype and full scale rotary-axial spindles, the shaft dynamics became the roadblock to even more extreme performance in the areas of bandwidth and dynamic stiffness. Other shaft architectures could be investigated. One example could have journal bearings at each end of the shaft with the magnetic bearing in the center. A study could be performed on the spindle shaft dynamics with the aim of constructing a rotary-axial spindle shaft design framework. The framework would focus on achieving the highest possible natural frequencies,
while keeping the mass down in order to reduce the load on the power amplifier and actuator. Such an analysis would surely make use of finite element methods.

9.2.3 Improvement of axial error motion through design for assembly

The ball target based metrology was a significant improvement over the target screw based design. So far, the limitation of its success has been in the assembly method and the patience of the installer. Any future iterations of the rotary-axial spindle should include a design for assembly study on the installation of the ball target. The biggest challenge here is to develop an installation method which allows the technician or engineer to make fine and controlled adjustments to the assembly while increasing the torque on the clamping bolts.

9.2.4 Modified sealing designs for hydrostatic bearings

The lip seals used as a last resort for keeping water inside the hydrostatic bearing proved to be less than ideal due to their high friction and low service life. Since friction is the enemy of the precision engineer, non-contact sealing designs such as labyrinth seals, for example, should be investigated and tested.

9.2.5 Control loop jitter investigation

In section 7.6.1 a noise disturbance in the system was uncovered and it was attributed to the implementation of the controller in a multi-tasking controller environment. A solution was developed which reduces the additional noise levels, but it is by no means optimal. If a mathematical model of the problem could be created, then the problem can be better understood. From there, new solutions can be proposed and tested.
9.2.6 Cutting tests

In the not too distant future, the rotary-axial spindle should be used in a manufacturing operation. So far, it has not done so. Such a test would require the installation of a stiff grinding wheel and the retuning of both the rotary and axial control loops. This may also warrant the investigation of adaptive feed forward cancellation of disturbance forces \[2:234-254\]. Additionally, the chatter stability theory of silicon wafer grinding should be developed so that chatter free machining can be achieved. The silicon wafer grinding process is an abrasive process, but in some ways similar to plunge milling. The literature review could begin in these areas.
References


