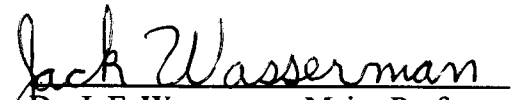
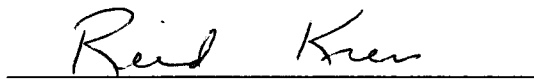


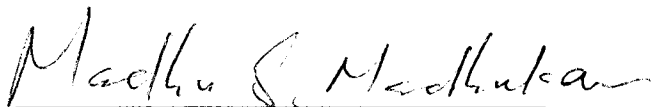
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Date

July 24, 1995

DEVELOPMENT OF A REDUCED COMPLIANCE LOAD CELL

A Thesis

Presented for the

Master of Science

Degree

The University of Tennessee, Knoxville

Joel B. Chesser

August 1995

DEDICATION

Let not the wise man boast of his wisdom,
or the strong man boast of his strength,
or the rich man of his riches,
but let him who boasts boast about this:
that he understands and knows me,
that I am the Lord, who exercises kindness,
justice and righteousness on earth....

[Jeremiah 9:23-24, NIV Bible]

Soli Deo Gloria

ACKNOWLEDGMENTS

The completion of this thesis was dependent on the efforts the individuals listed below. Without their contributions this thesis could not have been completed. I would like to take this opportunity to thank them for their efforts. Chuck Guffey, Oak Ridge National Laboratory (ORNL), provided drafting assistance for the mechanical drawings required for fabrication. John Jansen, Oak Ridge National Laboratory, took the time for patient explanations of control system design and allowed me the use of his system identification software. Ken Thomas, University of Tennessee, Engineering Mechanics shop fabricated the components used in the load cell. My wife, Linda, who kept our son, Phillip, while I was trying to finish my thesis and assisted with the final editing. My parents who encouraged me to make the most of the educational opportunities presented to me.

Finally, I would like to acknowledge the efforts of my committee. Dr. Wasserman originally posed the question in class about how to create a stiff load cell by incorporating a feedback loop, and then allowed me to use that as a thesis topic. Drs. Mhadukar and Kress made themselves available to serve on my committee.

ABSTRACT

Current force sensing technology is dependent on a deformable element to measure force. The deformation of the element can be detected mechanically, electrically or optically. Examples of the current state of the art for force measurement include strain gage or piezoelectric load cells. These are all based on measuring a finite deformation of the load cell. The goal of this thesis is to lay the groundwork for developing a load cell which includes an active feedback loop to compensate for the deformation of the load cell caused by the load. The resulting force sensor will appear to be infinitely stiff over some finite bandwidth to an applied load. The motivation for developing this instrument is to improve the accuracy of mechanical testing of small, stiff specimens. A current problem with testing of such specimens is that compliance of the load cell contributes to errors in determining the mechanical properties such as elastic modulus or stress-strain relations.

The concept was developed, analyzed, and tested. The analysis indicated that significant reductions in compliance could be achieved. The controller designed, a gain stage in series with a pole zero cancellation stage, was calculated to reduce the load cell compliance to $6.2 \mu\text{m/N}$. This is a reduction in compliance by a factor of three over the commercial load cells surveyed. The addition of a lag-lead compensator could further reduce the compliance to $0.01 \mu\text{m/N}$, several orders of magnitude better than the commercial load cells.

The analyzed load cell was then built and tested. The electronics performed as designed. The mechanical plant, however, had significant hysteresis limiting its performance. Several attempts were made to reduce the mechanical hysteresis. The diaphragm flexures were annealed, and slotted with limited improvement. The load cell tested did, however, successfully demonstrate the concept. It exhibited reduced compliance due to the control system action. The compliance was reduced from $333 \mu\text{m/N}$ to $38 \mu\text{m/N}$ by the control system tested, a simple gain stage. The compliance of $38 \mu\text{m/N}$ is of the same order as that of commercial load cells. It was determined by simulation that the mechanical hysteresis limited the load cell accuracy achievable with a more

sophisticated controller, gain followed by pole-zero cancellation. Therefore that controller was not built and tested.

Because of the limitations of the mechanical plant, simulation studies were conducted to determine the potential compliance improvements of the load cell without any mechanical hysteresis. The simulations included the frequency response, noise, and non-linearities of the electronics. Simulation indicated that if the mechanical non-linearities are removed performance is limited by the electronic noise and non-linearities. A problem of the load cell designed is the introduction of systematic errors if load is assumed to be simply proportional to current. A method of compensating for systematic errors caused by displacement and actuator non-linearities was simulated. These simulations indicated that a measurement error of 0.005 N is achievable with the electronics designed in the course of this thesis if the mechanical non-linearities can be eliminated.

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SECTION 1: INTRODUCTION

Problem Statement

Current force sensing technology is dependent on a deformable element to measure force. The deformation of the element can be detected mechanically, electrically or optically. Examples of the current state of the art for force measurement include strain gage or piezoelectric load cells. These are all based on measuring a finite deformation of the load cell. A brief survey of deflection as a function of full load for commercially available load cells is summarized in Table 1. The goal of this thesis is to lay the groundwork for developing a load cell which includes an active feedback loop to compensate for the deformation of the load cell caused by the load. The resulting force sensor will appear to be infinitely stiff over some finite bandwidth to an applied load. The motivation for developing this instrument is to improve the accuracy of mechanical testing of small, stiff specimens. A current problem with testing of such specimens is that compliance of the load cell contributes to errors in determining the mechanical properties such as elastic modulus or stress-strain relations.

Table 1. Comparison of commercial load cells

Manufacturer and model number	Full load g	Deflection at full load mm (in)	Compliance $\mu\text{m/N}$
Entran, ELJ-20G, Semiconductor gage type [1]	20	0.076 (0.0030)	387
Sensotec, #11, semiconductor gage type [2]	50	0.013 (0.0005)	26.5
Sensotec, #11, foil gage type [2]	50	0.025 (0.0010)	51
Omega, LCL113G foil gage bending beam [3]	113	0.025 (0.0010)	22.6

Procedures have been developed to allow correction for the compliance of the test system [4,5]. The compliance correction is determined by testing a series of fibers with successively shorter lengths and extrapolating the results to zero gage length. Such compensation is difficult and can mask behavior of the specimen under test [6,7]. This problem will be accentuated when determining the stress-strain relationship for small specimens.

The test system compliance arises from not only of the load cell, but also the sample grips, crosshead and load frame. This thesis will restrict itself to developing a load cell with reduced compliance. While there might be other possible applications for a zero compliance load cell the application of interest in the thesis is to improving accuracy of tensile testing of carbon and other high modulus fibers.

Two possible approaches to a zero compliance force transducer are: controlling the position of a stiff actuator, and measuring the current or voltage holding the actuator in position to determine load; or mounting a load cell on an actuator, measuring the load cell deflection, and controlling the actuator position.

Design Requirements

Before undertaking to design the load cell it is important to examine its operating requirements. These include maximum forces the load cell is expected to experience, and the control system bandwidth for adequate performance. The design parameters will be defined by examination of published test results from carbon fiber tensile testing. Published results from tensile testing of carbon and other high tensile modulus fibers are summarized in Table 2. From the data in the table design criteria for a load cell will be developed. The key design parameters are: maximum load, specimen elongation at failure, and time required for the load cell to stabilize and accurately measure the loading force. The specimen elongation will be used to determine the maximum allowable deflection of the load cell during testing. Ideally the deflection would be zero. Realistically, a non-zero deflection of the load cell which is a fraction of the fiber

Table 2. Summary of test results from fiber tensile testing

Fiber type	Elastic modulus GPa (Msi)	Strength MPa (ksi)	Notes
PAN-based carbon fiber, T-700 [6]	255 (37)	5000 (725)	
PAN-based carbon fiber, T-40 [6]	290 (42)	5650 (820)	
Thornel-P 9.78 μm dia [8]	381 (55)	2400 (348)	Crosshead speed: 0.125 mm/min Gage length: 25.4 mm
Thornel-P 13.11 μm dia [8]	328 (47)	1880 (273)	see above
Thornel-P [9]	not given	~1600 (~232)	Crosshead speed: 0.5 mm/min Gage length: 25.4 mm
Tungsten 10 μm dia [7]	340 (49)	3300 (478)	Crosshead speed: 0.05 mm/min Gage length: 44.6 mm

elongation is to be expected. The maximum load experienced by a load cell used in testing the specimens in Table 2 is 26 grams. The specimen time to failure, calculated from fiber properties and crosshead speed, ranged from 68 to 501 seconds. The specimen elongation, calculated from fiber properties, ranged from 0.14 to 0.42 mm.

A reasonable estimate of the maximum allowed deflection is 0.005 mm (0.0002"). This is less than 5% of the deflection of the stiffest specimen in Table 2. An estimate for the maximum time for the load cell to stabilize and accurately track the load is 0.05 seconds. The target maximum load for the proposed load transducer is 50 gm. Which is double the maximum load to fracture the strongest specimen in Table 2. The final design requirement is low cost. While there is no explicitly defined cost goal, the system must be easily and inexpensively fabricated.

Design Evolution

The objective of this thesis is to design a load transducer with apparent infinite stiffness or zero compliance ($\text{compliance} = 1/\text{stiffness}$). Several approaches were considered before settling on a final design concept. All of the design concepts were based on the idea of controlling the position of the load cell's load attachment point by an actuator and an active control system. By maintaining the load attachment point fixed in space, the load cell appears to be infinitely stiff to the applied load. The actuator and control system compensate for the deformation of the load cell from the applied forces. Two basic approaches are 1.) to use an actuator to compensate for the deformation of a load cell, as the load cell deforms, the control system compensates for its deformation; and 2.) maintaining a stiff actuator in a null position, and measuring the current or voltage holding the actuator in position to determine load. The first approach allows the calibration of the load cell independently of tuning the control system. The second approach does not allow for independent calibration and tuning. As will be shown, the second approach is, however, mechanically simpler. A variant of the second approach is currently used in null-

balance accelerometers and null-balance torquemeters [10]. For clarity the two approaches to load cell design will be respectively referred to as motion compensation and null balancing

The obvious approach would be to use stepper motor or servo motor micro-positioners as actuators to compensate for the load cell deflection. These positioners are available commercially, produced by several companies. Companies producing these include Klinger, Oriel, Dover, and Melles Griot. They are accurate, capable of sub-micrometer incremental motion, and bi-directional repeatability [11]. Their high cost, however, precludes their use. Commercial micro-positioners cost in excess of \$1000. Custom fabrication would entail significant cost to achieve high precision. The high cost resulted in micro-positioners being rejected from the outset.

Piezoelectric and magneto-restrictive actuators were considered. Use of these actuators for active vibration control has been reported [12,13]. Active vibration control is similar to the application discussed in this paper. Piezoelectric actuators have limited travel, typically limited to tens of micrometers and require high voltage power sources. Magneto-restrictive actuators are more expensive than voice coil actuators. As a result piezoelectric and magneto restrictive actuators were dropped from further consideration.

Finally electromagnetic actuators capable of generating linear motions were examined because they have no backlash and limited hysteresis thus making them easily controlled. If the moving element is supported on springs or flexures, as opposed to bearings, they provide frictionless motion. They are available commercially or can be relatively easily fabricated. The first idea was to use a solenoid as an actuator. They are attractive because they are inexpensive. Savchin and Spitkovskii describe using a solenoid to dynamically load test specimens [14]. The problem with solenoids is their nonlinear response. Voice coil actuators were then considered. Voice coil were more attractive because of improved linearity of force as a function of current simplifying design and calibration. The difference between solenoids and voice coils is that solenoids are singly excited whereas voice coil actuators are multiply excited. Typically one

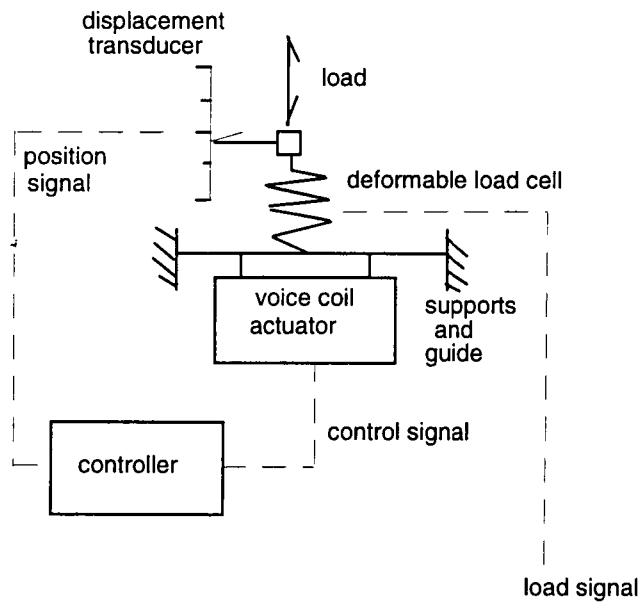
excitation is replaced with a permanent magnet. The use of a moving magnet voice coil actuator to strain TaS₃ "whiskers" is reported by Skove et al [15].

Due to the small motions required to compensate for the load cell deformation, motion "demagnification" was considered. A motion "demagnification" method is reported by Rothberg et al. to strain tin whiskers [16]. They used a micrometer screw to compress a spring. The spring was connected to a pulling rod, which in turn was connected to a cantilever flexure. The specimen was attached to the cantilever flexure. A 1 mm micrometer screw adjustment resulted in an 8 μ m cantilever motion. Motion "demagnification" is attractive because it reduces the actuator control accuracy requirements. Motion "demagnification," however, will not be further considered because of the added expense of fabricating the flexures to perform that function. Supporting the decision to not use motion "demagnification" is Hewlett Packard's use of voice coil actuators to directly produce sub-micrometer motions for component alignment in the 71450A and 71541A optical spectrum analyzers [17].

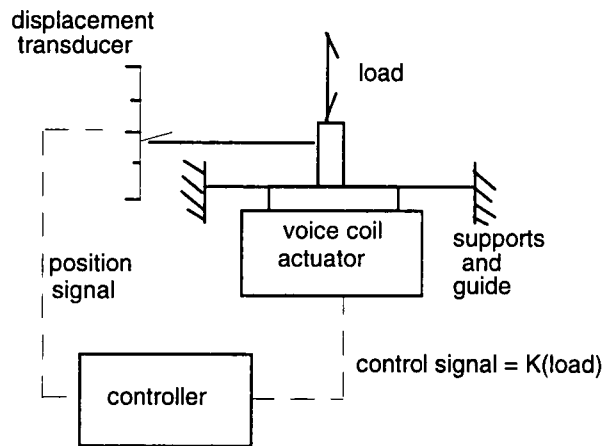
Comparison of Motion Compensation and Null Balancing Schemes

As mentioned earlier there are two possible ways of achieving a stiff load cell. These are: controlling the position of an actuator, and measuring the current or voltage holding the actuator in position to determine load; or mounting a load cell on an actuator, measuring the load cell deflection, and controlling the actuator position. Schematics of the two methods are shown in Figure 1.

The motion compensation scheme is attractive because it simplifies calibration. The load cell can be calibrated independently of tuning the actuator control system. The load cell can be a separate component, tested, and then mounted on the actuator. A potential problem with this scheme is the actuator and displacement sensor are non-collocated. That is, they are separated by a flexible member, the load cell. This can be seen by referring to Figure 1a. This can lead to the system potentially becoming unstable. This is not a sufficient reason to eliminate the design from



a) motion compensation scheme



b) null balancing scheme

Figure 1. Schematic comparison of motion compensation and null balancing schemes.

further consideration as proper design of the control system can compensate for instability. The motion compensation scheme, however, includes one more component, the load cell, than the null balancing approach making the mechanical design more complicated. With the null balancing scheme, actuator and displacement sensor are collocated. That is, they are separated by at most a rigid link. making control somewhat simpler. The combination of the simpler mechanical design, and stable system make the null-balancing design the approach of choice.

SECTION 2: DESIGN AND ANALYSIS

Introduction

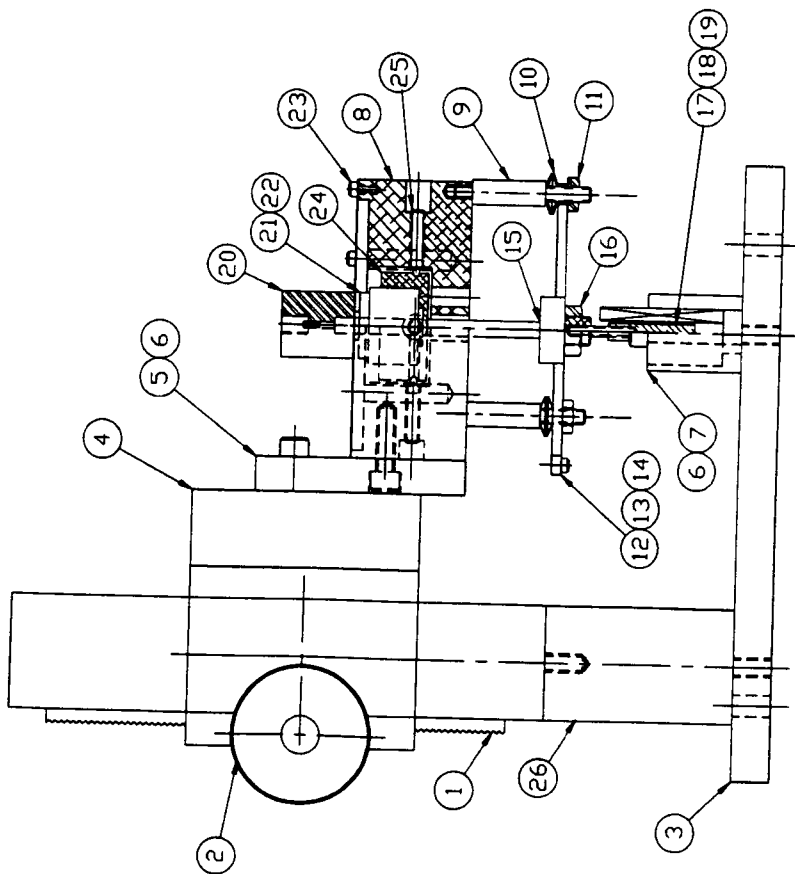
The various aspects of design for a prototype null balancing load cell will be discussed in this section. This will include mechanical, electrical, and controls design. As the load cell to be constructed is a proof of principle prototype, the design will use commercial components wherever possible and it is designed for ease of fabrication, assembly, and calibration. A production instrument would be more compact, however, it would require more custom designed components.

The displacement of the load is monitored by a non-contact displacement sensor. The displacement sensor measures the change in position of the reference surface. The controller acts to keep the reference surface a constant distance from the displacement sensor. The current required to maintain the reference surface in a constant position is proportional to the force applied to the load cell.

Mechanical Design

A proof of principle load cell to test the null-balancing concept was designed. The fabrication drawings can be seen in Figures 2 through 15. The assembly is shown in Figure 2. Two key factors considered in the development of the design were low cost and ease of adjustment. The proof-of-principle design includes features which would not be included in a final design used for actual material specimen testing, some features required for specimen testing, on the other hand, were omitted in this design.

The actuator is not mounted directly to the sensor body but is instead mounted in the actuator mount which "floats" on adjustable spring plungers. This was done to allow adjustment of the actuator centerline with respect to the center post. To reduce fabrication costs for the



PART NO.	QTY.	DESCRIPTION
1	1	NEWPORT #71 POST
2	1	NEWPORT #370-C ROD CLAMP
3	1	BASE PLATE
4	1	NEWPORT #421 STAGE
5	1	ADAPTOR PLATE
6	6	SHCS 1/4-20UNC X 1-1/2 LG.
7	1	LVDT MOUNT
8	1	SENSOR BODY
9	3	LOWER DIAPHRAGM STANDOFF POST
10	6	MORDEX 8DMS-AA-5 BELLEVILLE WASHER
11	3	HEX NUT #10-24UNC
12	1	LOWER DIAPHRAGM
13	1	LOWER DIAPHRAGM MOUNTING RING
14	13	SHCS #4-40UNC X 3/16 LG.
15	1	CENTER POST
16	1	HEX NUT 3/8-16UNC
17	1	LUCAS SCHAEVITZ HR200 LVDT CORE
18	1	LUCAS SCHAEVITZ HR200 ARMATURE
19	1	LVDT CONNECTING ROD
20	1	UPPER DIAPHRAGM RETAINER
21	1	BET LA09-08-000A ACTUATOR
22	2	SHCS #2-56UNC X 1/4 LG.
23	1	UPPER DIAPHRAGM
24	1	ACTUATOR MOUNT
25	4	CARR-LANE CL-30-SSPS-2 SPRING PLUNGER
26	1	SPACER BLOCK

Figure 2. Instrument assembly drawing.

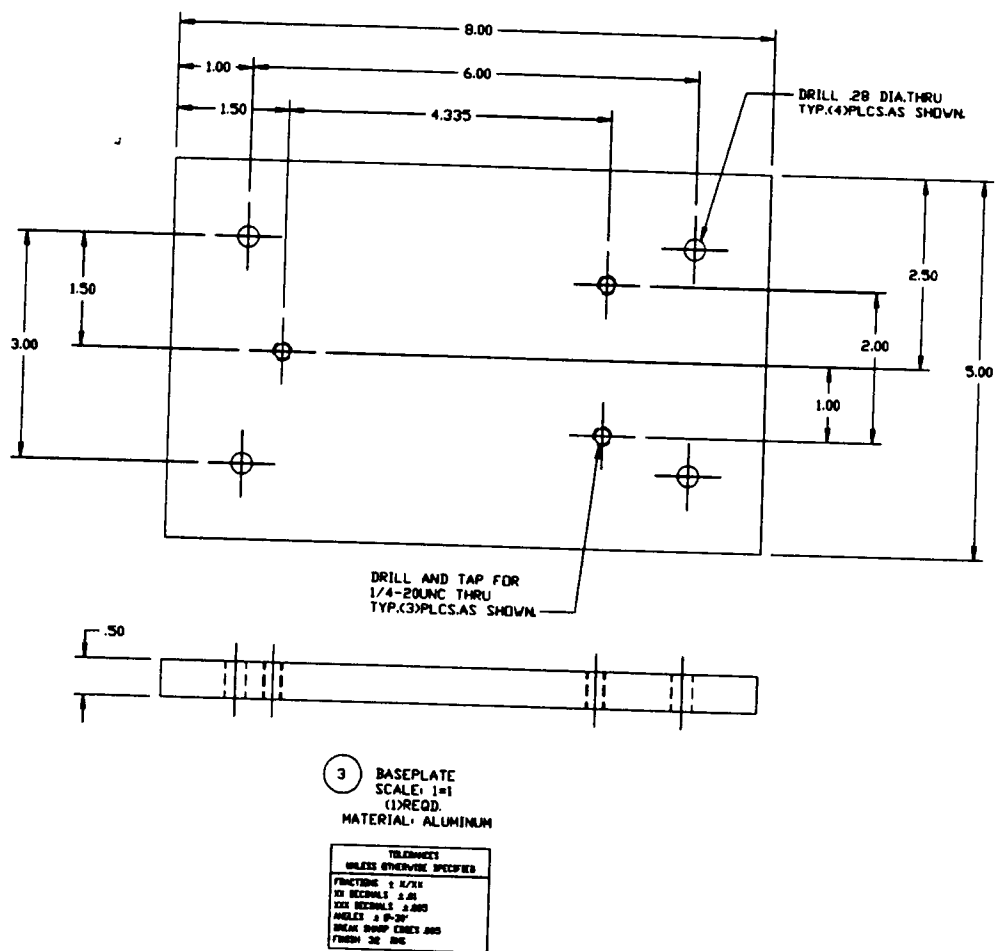
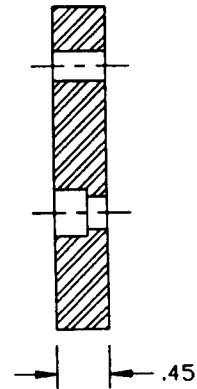
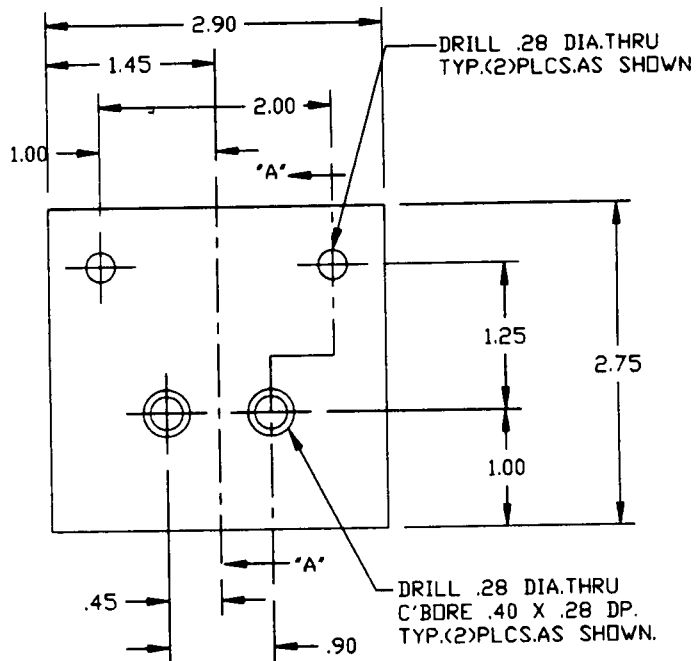


Figure 3. Instrument base plate drawing



SECTION 'A-A'

5 ADAPTOR PLATE
SCALE: 1=1
(1) REQD.
MATERIAL: ALUMINUM

TOLERANCES UNLESS OTHERWISE SPECIFIED	
FRACTIONS	± X/XX
XX DECIMALS	± .01
XXX DECIMALS	± .005
ANGLES	± 0°-30'
BREAK SHARP EDGES	.005
FINISH	32 RMS

Figure 4. Adapter Plate Drawing

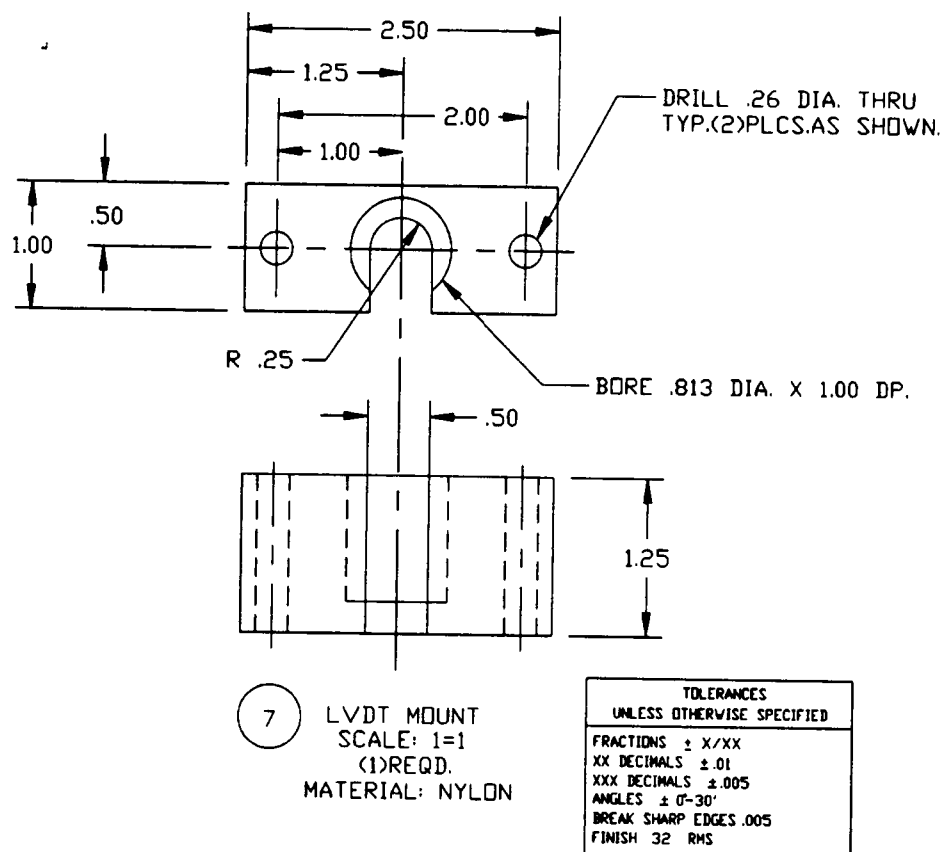


Figure 5. LVDT mount drawing.

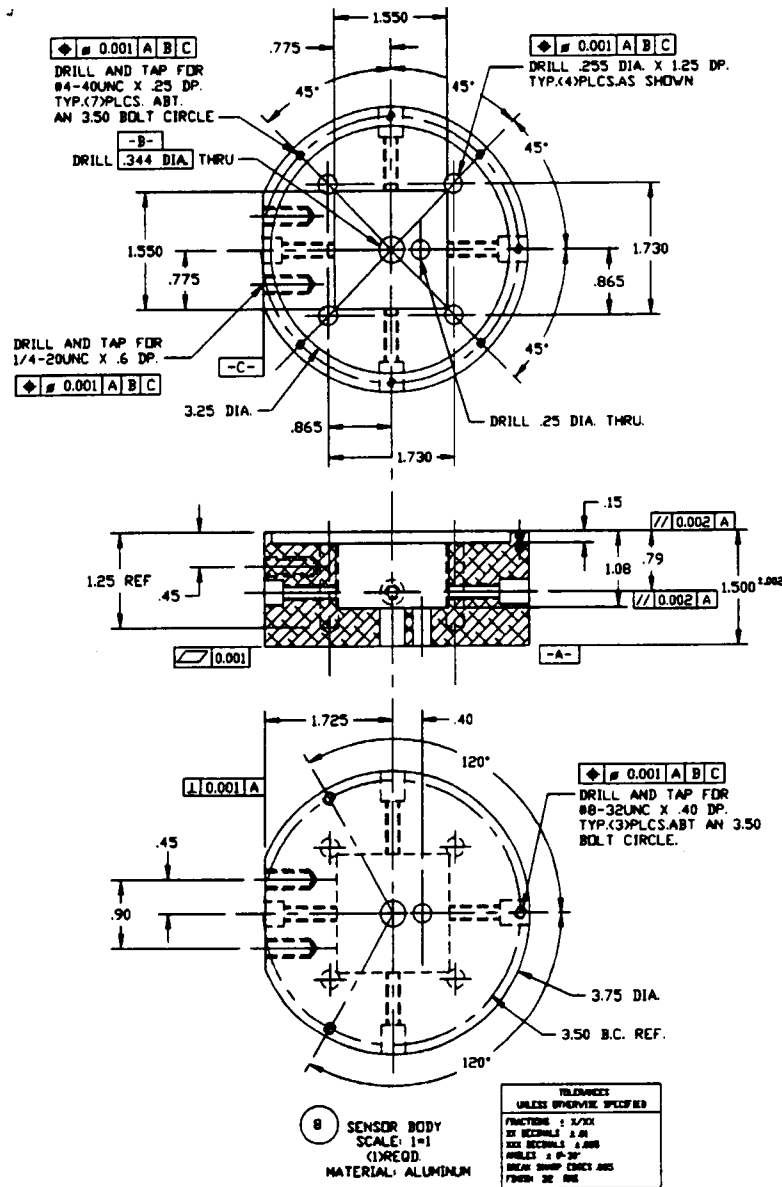


Figure 6. Sensor body drawing.

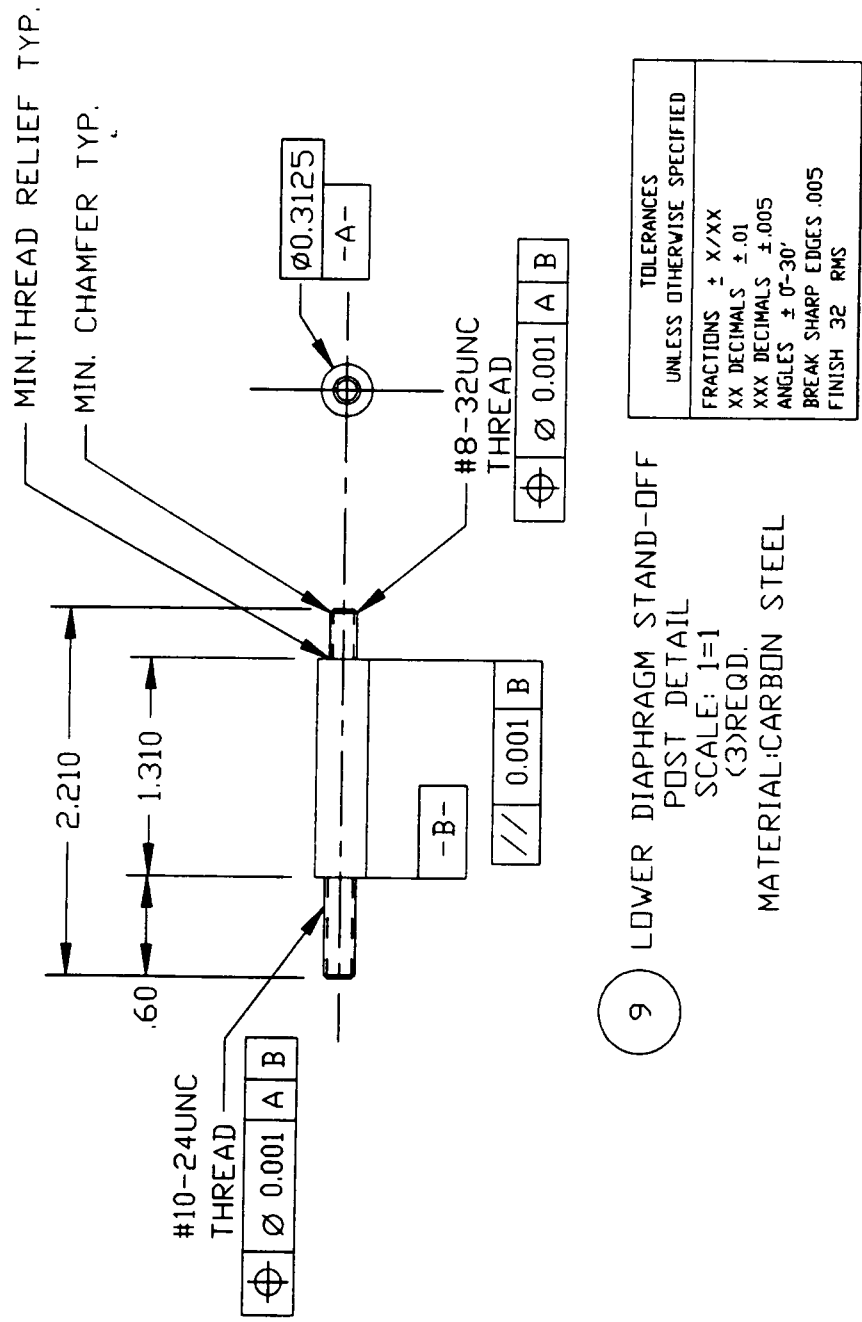
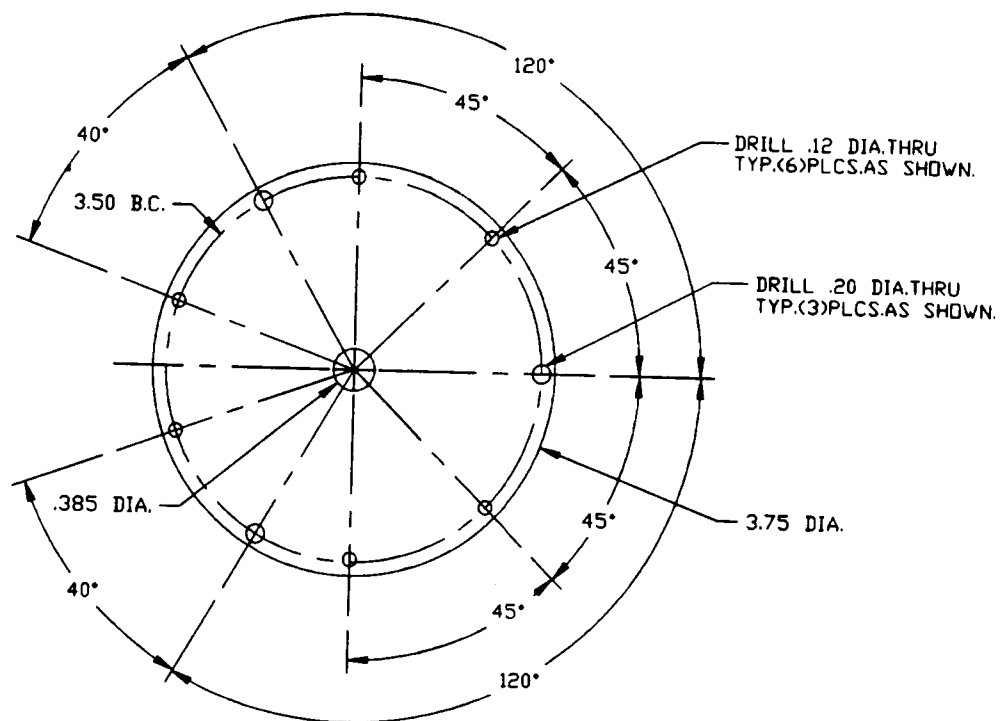


Figure 7. Lower diaphragm stand-off drawing.



12 LOWER DIAPHRAGM
SCALE: 1=1
(1) REQD.
MATERIAL: .005" BRASS SHIM STOCK

TOLERANCES UNLESS OTHERWISE SPECIFIED	
FRACTIONS	± X/XX
XX DECIMALS	± .01
XXX DECIMALS	± .005
ANGLES	± 0°-30'
BREAK SHARP EDGES	.005
FINISH	32 RMS

Figure 8. Lower diaphragm drawing.

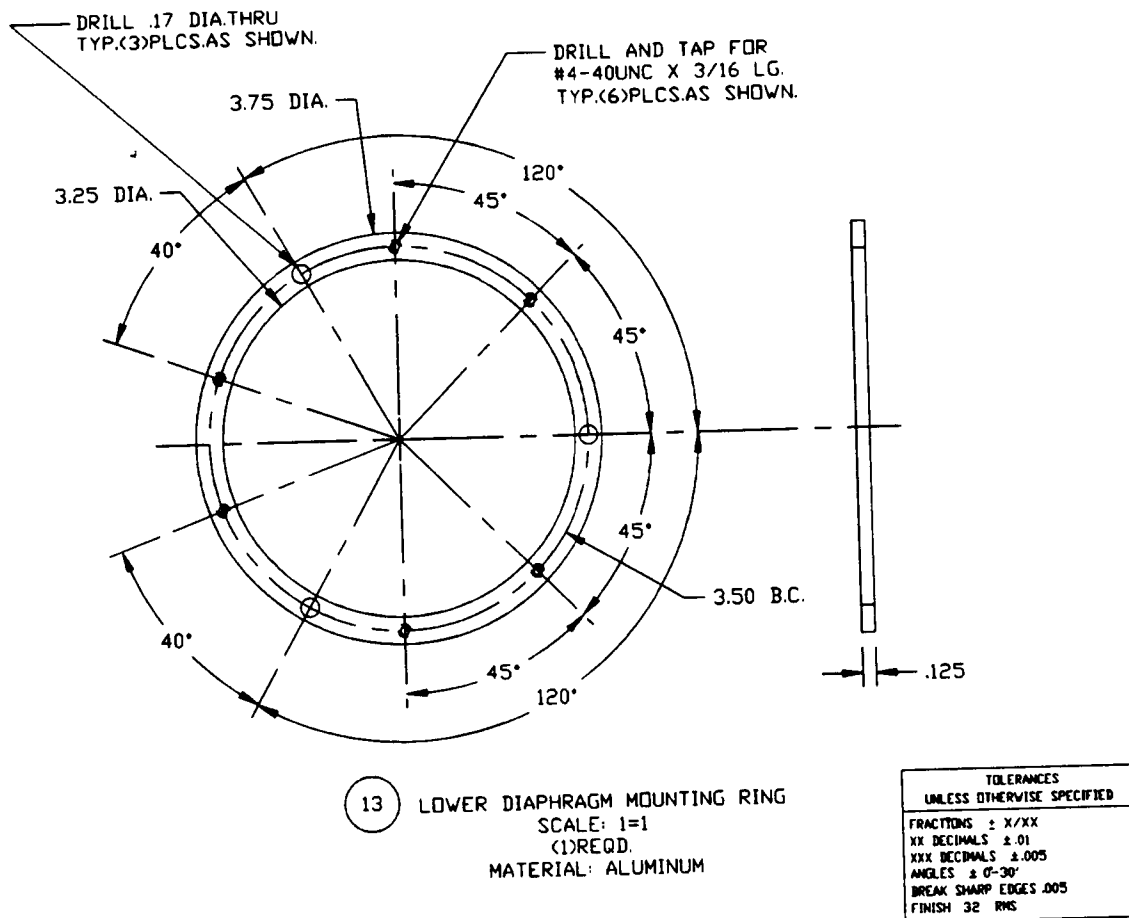


Figure 9. Lower diaphragm mounting ring drawing.

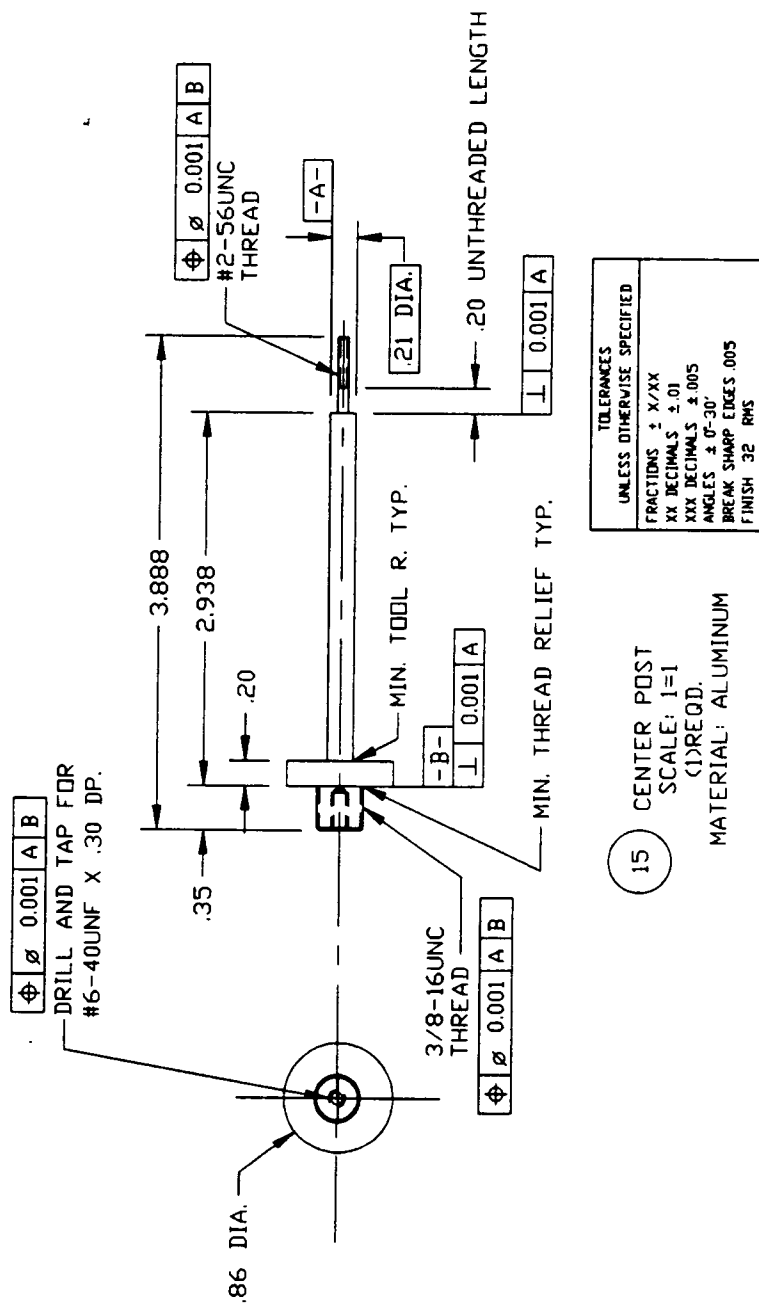


Figure 10. Center post drawing.

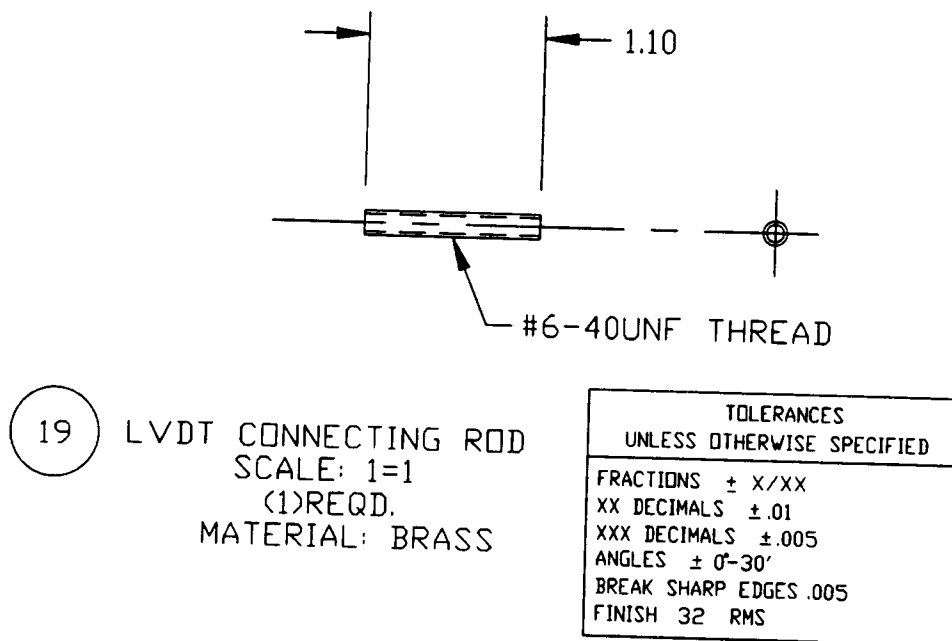
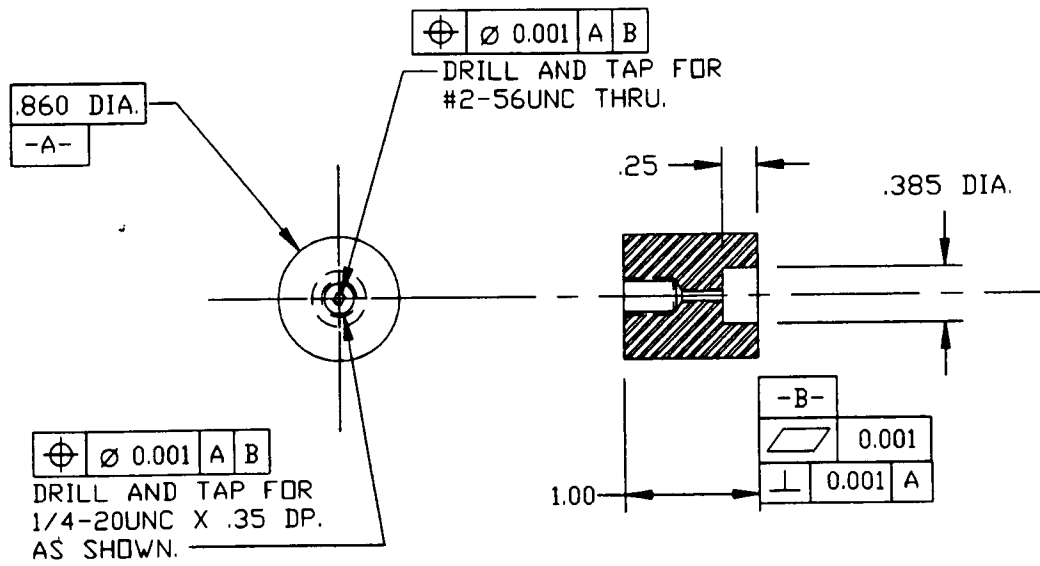


Figure 11. LVDT connecting rod drawing.



(20) UPPER DIAPHRAGM RETAINER
 SCALE: 1=1
 (1)REQD.
 MATERIAL: CARBON STEEL

TOLERANCES UNLESS OTHERWISE SPECIFIED	
FRACTIONS	$\pm$ X/XX
XX DECIMALS	$\pm .01$
XXX DECIMALS	$\pm .005$
ANGLES	$\pm 0^{\circ}-30'$
BREAK SHARP EDGES	.005
FINISH	32 RMS

Figure 12. Upper diaphragm retainer drawing.

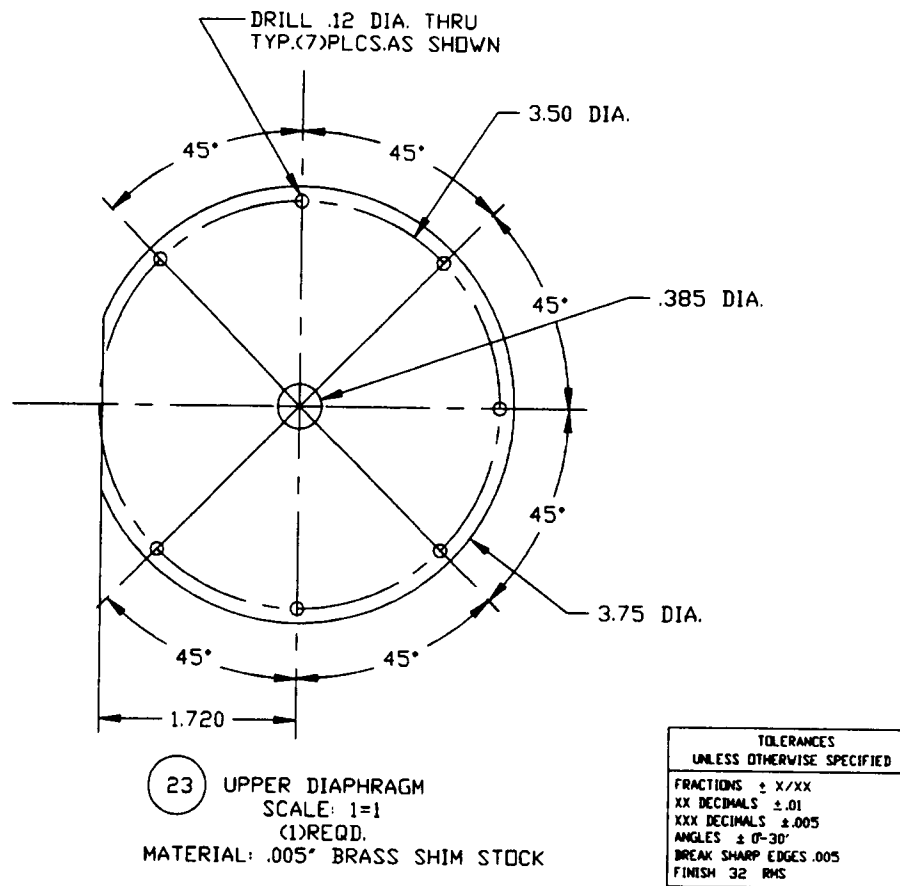


Figure 13. Upper diaphragm drawing.

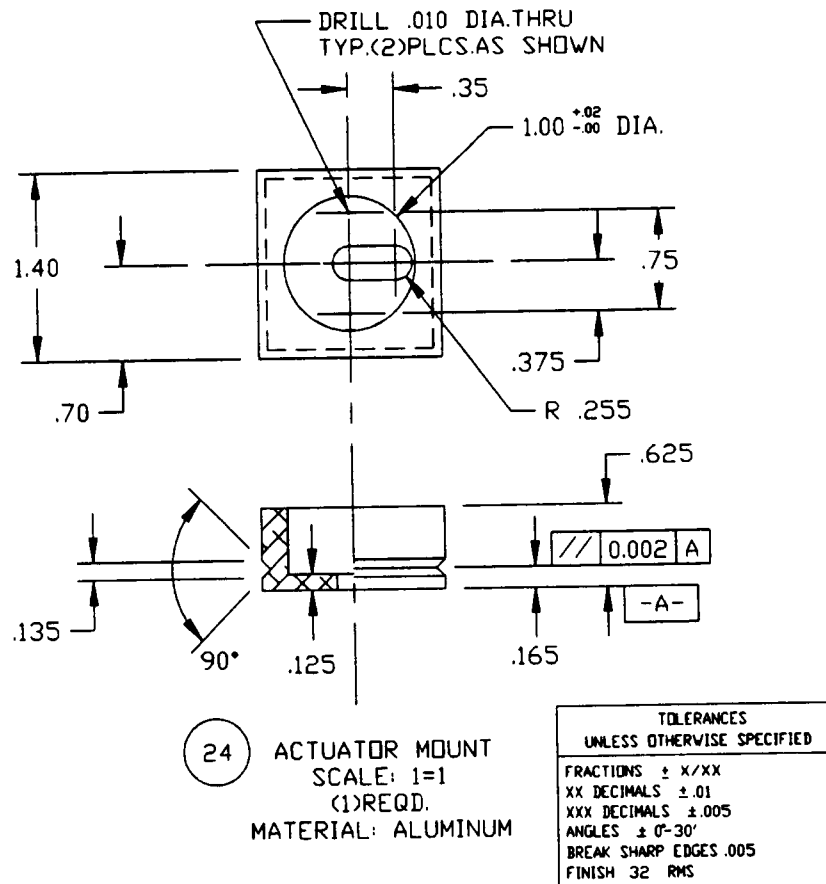
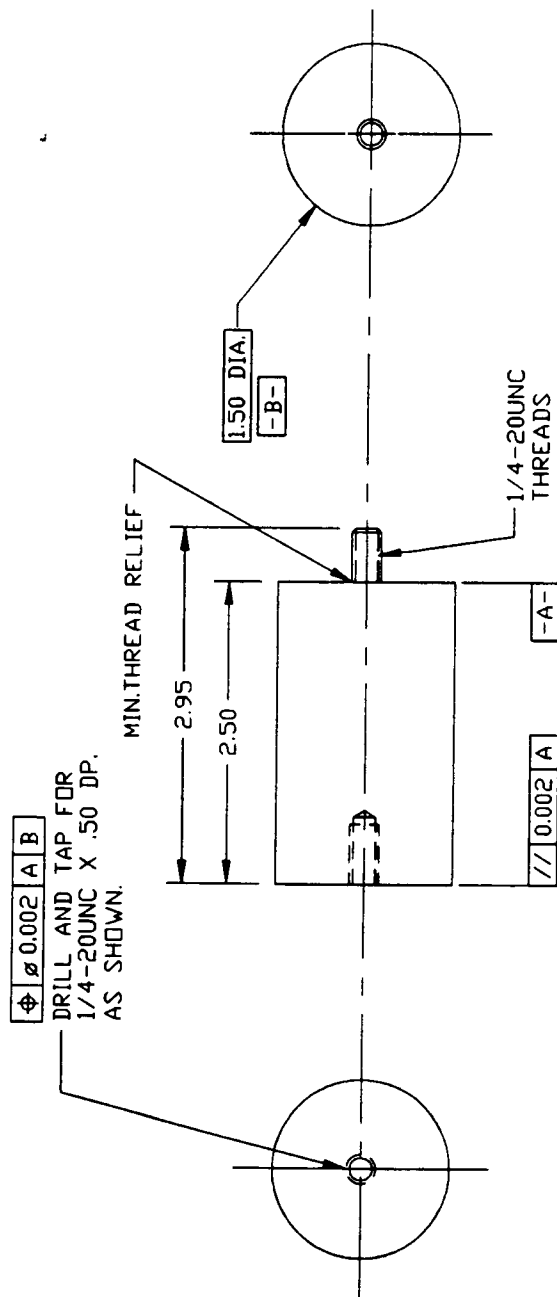


Figure 14. Actuator mount drawing.



TOLERANCES	
UNLESS OTHERWISE SPECIFIED	
FRACTIONS	± X/XX
XX DECIMALS	± .01
XXX DECIMALS	± .005
ANGLES	± 0°-30'
BREAK SHARP EDGES	.005
FINISH	32 RMS

(26) SPACER BLOCK
 SCALE: 1=1
 (1)REQD.
 MATERIAL: CARBON STEEL

Figure 15. Spacer block drawing.

prototype there are no features, such as dowel pins, to positively locate the diaphragms with respect to the sensor body, or the center post with respect to the diaphragms. This made it essential to be able to adjust the actuator position. The resultant air gap between the actuator mount and the sensor body limits the power that can be dissipated safely by the actuator. This limits the current that can be safely applied, and hence the maximum force that can be measured. An improved design would eliminate the need for adjustment and mount the actuator directly to the sensor body improving the heat transfer.

The lower flexure is attached to a mounting ring. The mounting ring is supported by standoff posts. Between the mounting ring and the standoff posts and the mounting ring are belleville washers. These serve as springs to adjust the lower diaphragm position in the axial direction. This adjustment serves to compensate for fabrication tolerance stack-up in the axial direction and to allow adjusting the axial preload on the flexures. The moving masses that are attached to the flexures are given in Table 3.

The sensor body is mounted to two adjustable stages, one with coarse adjustment, and one with fine adjustment. This is to allow the sensor to be moved vertically into position under displacement sensing devices during calibration. This feature would not be required in a production design because a the sensor could be attached to a calibration jig. Being a proof of concept demonstrator, no provisions were made for attaching load grips or for mounting the load cell into a material tester. This remains to be attempted with the next generation model.

Table 3. Moving component masses

Component	Mass (gm)
actuator magnet	8
LVDT core	8
3/8-16 hex nut	9.4
LVDT connecting link	2.1
diaphragm retainer	64.9
center spindle	11.1
total moving mass	103.5

Table 4. Parameter values for the BEI LA-09-08-00A actuator.

Parameter	units	value
continuous stall force	N (oz.)	2.09 (7.5)
force constant (K_f)	N/amp (oz./amp)	1.11 (4.0)
back emf constant (K_e)	volts/(m/sec) (volts/(ft/sec))	1.12 (.34)
resistance @ 25° C (R)	ohms	3.2
electrical time constant	μsec	110
calculated coil inductance (L)	millihenries	0.352
coil assembly weight (m)	kg (oz.)	8.5×10^{-3} (0.3)
maximum actuator travel	mm (in)	1.57 (0.062)

Linear Actuator

The actuator used is a Kimco Magnetics LA09-08-00A. Key electrical and mechanical parameters for the actuator are presented in Table 4. It was selected because its external shape made it easy to incorporate into an instrument design, and because it could generate adequate force. This particular actuator is actually a kinematic inversion of a voice coil actuator. A traditional voice coil design features a fixed magnet and a moving coil. The LA09-08-00A features a moving magnet and a fixed coil. This allows the coil to be mounted in thermal contact with a baseplate improving heat transfer by conduction to the baseplate. The actuator is supplied as two separate pieces, the coil and the magnet, without guides or bearings. It is left to the designer to provide mounts and guides. For this application the coil will be mounted to a base, the magnet mounted to support flexures. The support flexures allow frictionless motion, and constrain the actuator travel.

A simplified analysis of voice coil operation is presented below. A charge moving through a magnetic field causes a force to be applied to the charge resulting in a force to the wire. The force is mutually perpendicular to the conductor and the magnetic field. By Lorentz's law the force can be expressed as

$$\mathbf{F} = i\mathbf{l} \times \mathbf{B} \quad (2.1)$$

Where i is the current flowing through the conductor, l is the length of the conductor, and $\mathbf{B}$ is a vector representing the magnetic flux. Given the geometry for a voice coil actuator which is radially symmetric about its centerline, as shown in figure 16a, equation 2.1 can be rewritten as

$$F(t) = 2\pi r i(t) \sum_{j=1}^N B_i \approx K_f i(t) \quad (2.2)$$

Where r is the coil radius, $i(t)$ is the current as a function of time, N is the number of turns in the coil, and B_i is the magnetic field seen by each coil turn. K_f is the actuator force constant, the ratio of output force to input current. In general B_i is dependent on the relative coil and magnet position [18,19]. For a simplified linear analysis B_i will be considered to be constant, simplifying the equations. This is justified by noting that the coil dimensions are significantly larger than the motion range, hence each coil sees a relatively constant magnetic flux.

As the coil moves through the magnetic field, an opposing emf is induced. By Faraday's law of induction

$$e = -\frac{d\phi}{dt} \quad (2.3)$$

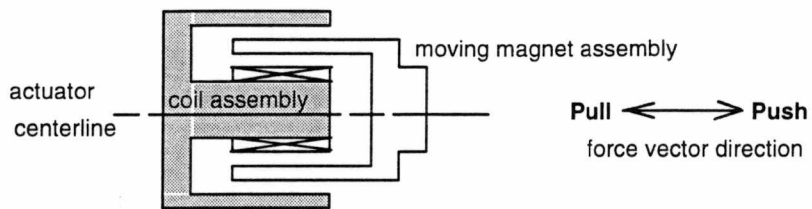
where ϕ is the flux of $\mathbf{B}$. For a coil moving through a magnetic field the back emf generated across the coil is given by,

$$e(t) = 2\pi r \frac{dx}{dt} \sum_{i=1}^N B_i \approx K_e \frac{dx}{dt} \quad (2.4)$$

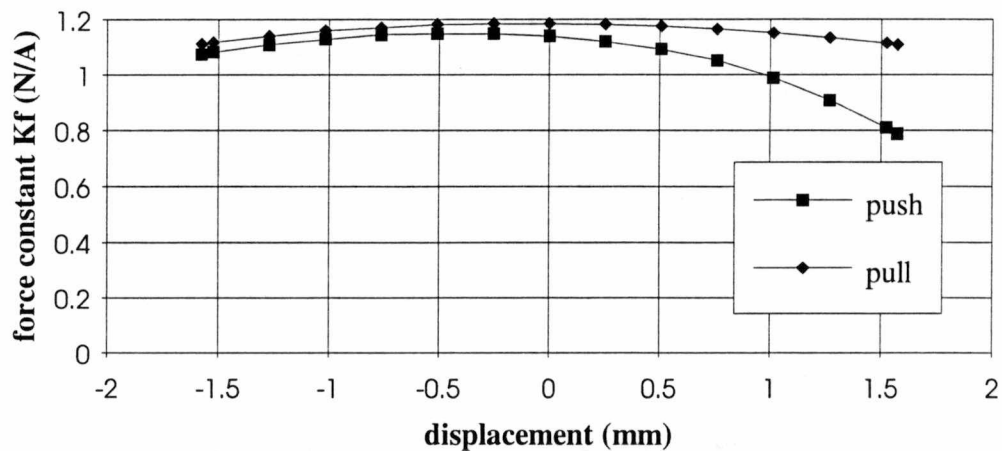
where $\frac{dx}{dt}$ is the relative velocity between the coil and magnet. The coil will be driven by a current source eliminating the effect of the back emf. The current source will adjust its output voltage to supply the commanded current despite the induced emf.

The equation of motion defining the voice coil behavior is given by

$$K_f i(t) = m \frac{d^2x}{dt^2} + c \frac{dx}{dt} + kx \quad (2.5)$$



a) Schematic of moving magnet voice coil actuator. Note force vector direction



b) Actuator force constant, K_f , displacement calibration curves are with actuator current of 500 mA. Refer to figure 16a for force vector direction.

Figure 16. Calibration data for the BEI LA-09-08-00A actuator.

where $K_f i(t)$ is force generated by the coil armature, m is the voice coil armature mass, c is any damping, and kx is the spring force from the coil support flexures.

The preceding was a simplified description of voice coil actuator operation. It treated the coil essentially as an infinite wire and ignored the effect of inductance variation as a function of coil travel. A more thorough analysis would include consideration of the energy stored in the coil magnetic field due to its inductance. Analysis of the voice coil using energy methods reveals a non-linear term proportional to the current squared [20]. The effect of this nonlinearity can be seen in the calibration data provided by BEI shown in Figure 16b. These data will be used later in simulations which will include the nonlinear term.

Equations 2.2, 2.4, and 2.5 can be combined into the block diagram shown in Figure 17. The transfer function for coil position as a function of input current can now be calculated. The transfer function is given by

$$\frac{x(s)}{i(s)} = \frac{K_f}{ms^2 + cs + k} \quad (2.6)$$

Where K_f , K_e , L , and R have been previously given; $m=0.119$ kg, $k=2900$ N/m, and $c=0$. k will be calculated in the next section, m is given in the mechanical design section, and the flexures will be assumed to be undamped, giving $c=0$.

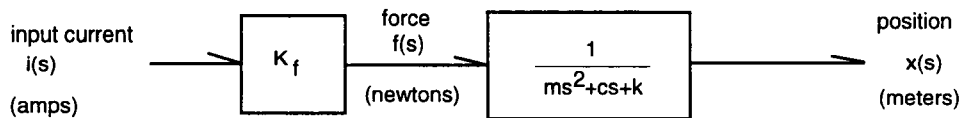


Figure 17. Block diagram of flexure mounted voice coil actuator driven by a current source.

Support Flexures

The actuator and the load attachment point are supported by a pair of flexures. The flexures are two identical thin diaphragms. Their purpose is to guide and limit actuator travel. A pair of flexures is used to ensure linear motion of the magnet relative to the coil in the presence of moments, or off-axis forces applied to the load attachment point. The flexures are compliant in the axial direction and stiff in all other directions. The coil to magnet clearance is only 0.381 mm (0.015"). Should the magnet and coil rub measurement inaccuracies would be introduced due to frictional forces, as well as potentially damaging the actuator.

The other function of the flexures is to keep the reference surface normal to the measurement axis of the displacement sensor. The reference surface determines the plane from which displacement measurements are made. Rotation of this surface would introduce errors into controlling the load position. A single diaphragm was initially considered, however, it was not stiff enough to ensure that the reference surface rotation would not introduce significant error. The reference surface will be defined to be the top of the load attachment "button" shown in Figure 18.

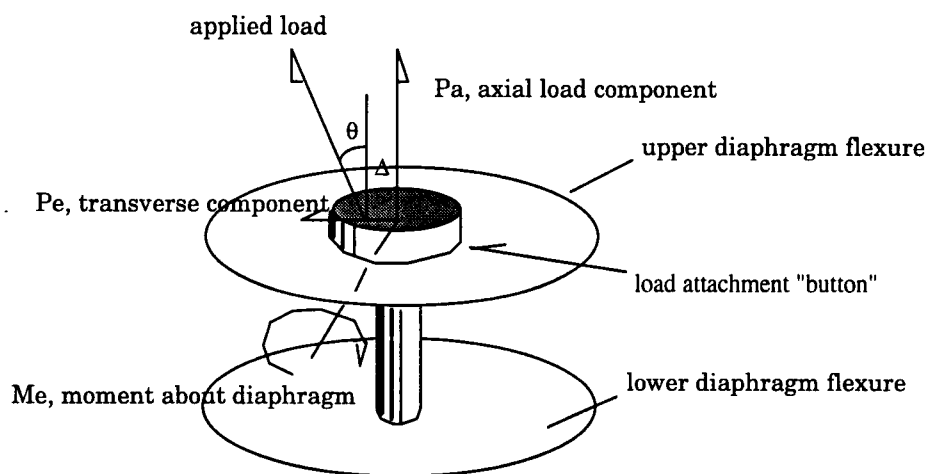


Figure 18. Schematic of a double diaphragm support with an applied load with an offset, Δ , and angular misalignment, θ .

Three loads need to be considered in designing the flexural support. The axial load imposed by the specimen, transverse loading along the plane of the diaphragm, and a moment about the diaphragm plane. The first load is the desired loading from the specimen. The last two arise from errors in mounting the specimen. The loads are illustrated in Figure 18. The transverse component and moment shown in the figure arise as a result of the load offset and angular misalignment.

The analysis will begin with presenting the equations relating the deformation of a diaphragm to applied loads. The deformation, x , of a diaphragm due to an axial load, P_a , is given by

$$x = P_a \delta_a \quad (2.7)$$

where δ_a is given by [21]

$$\delta_a = \frac{3(m^2 - 1)}{4\pi m^2 E t^3} \left[R^2 - r^2 - \frac{4R^2 r^2}{R^2 - r^2} \left(\ln \frac{R}{r} \right)^2 \right] \quad \text{where } m = \frac{1}{\nu} \quad (2.8)$$

δ_a represents the deformation of the diaphragm to a unit load in the axial direction. R is the diaphragm's outer radius, r is the inner radius, t is the thickness, E is the modulus of elasticity, and ν is Poisson's ratio. The deformation of a diaphragm under a transverse load, P_t , is given by

$$Y = P_t \delta_t \quad (2.9)$$

where δ_t , the deformation of the diaphragm to a unit load in the transverse direction, is given by [22] as

$$\delta_t = \frac{(3-\nu)(1+\nu)}{4\pi E t} \left[\ln \frac{R}{r} - \left(\frac{1+\nu}{3-\nu} \right)^2 \cdot \left(\frac{R^2 - r^2}{R^2 + r^2} \right) \right] \quad (2.10)$$

The deformation of a diaphragm due to a moment applied to its center is given by

$$\theta = M \vartheta, \quad (2.11)$$

where ϑ , the deformation of the diaphragm to a unit moment is given by [22] as

$$\vartheta = \frac{3(1-\nu^2)}{\pi Et^3} \left(\ln \frac{R}{r} - \frac{R^2 - r^2}{R^2 + r^2} \right). \quad (2.12)$$

The analysis of the double diaphragm flexure is based on analysis by Malikov et al [22]. A diagram outlining the geometry of the problem is shown in Figure 19. P_e and M_e as shown on the drawing are the loads introduced due to misalignment of the load applied at point A. The element shown by line AB is the rigid element connecting the two diaphragms. This element will be stiff enough, in comparison to the diaphragms, that ignoring its flexibility will introduce minimal error. As pointed out earlier the two diaphragms will be identical, simplifying the analysis. In the analysis deflections will be assumed to be small allowing linear superpositioning and small angle approximations.

The deflection of the flexures due to moment M_e will be considered first. The influence of P_e will be considered separately. The forces and moments P_1, M_1, P_2, M_2 arise from the diaphragm deformation. The problem is statically indeterminate requiring application of geometric constraints to solve. By summing forces in the radial direction we can write

$$P_1 - P_2 = 0 \quad \text{or} \quad P_1 = P_2 \quad (2.13)$$

For that to be true

$$Y_2 = Y_1 \quad (2.14)$$

By summing moments about point A

$$M_e + P_1 d = M_1 + M_2 + P_2 (l + d) \quad (2.15)$$

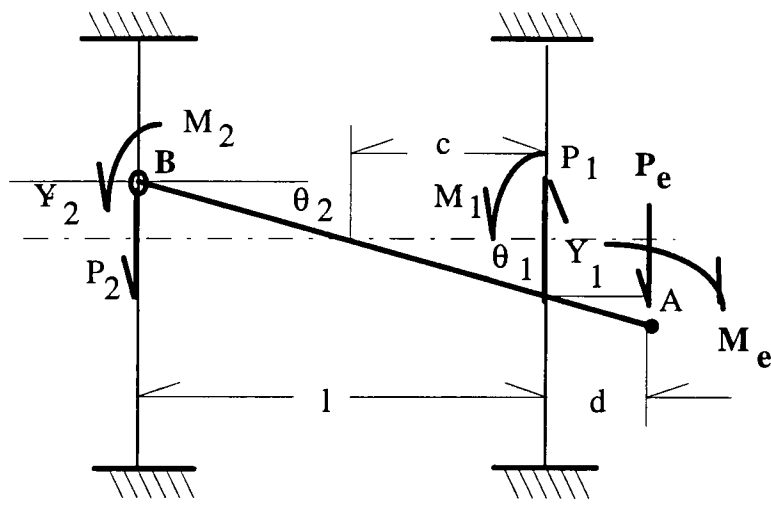
By noting that the center bar is rigid we can state that

$$\theta_1 = \theta_2 \quad (2.16)$$

and therefore

$$M_1 = M_2 \quad (2.17)$$

Because the rotations and deflections are small, the approximation



coordinate system:

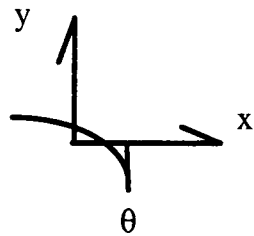


Figure 19. Diagram for analyzing deflections of the support flexures.

$$\theta_1 = Y_1/c \quad (2.18)$$

can be made with good accuracy. Because $Y_1 = Y_2$, $c = \frac{1}{2}l$ allowing equation 2.18 to be rewritten

as

$$\theta_1 = \frac{2Y_1}{l} \quad (2.19)$$

Utilizing the relationships in equations 2.9, 2.11, and 2.19, the moments can be rewritten in terms of the forces

$$M_1 = M_2 = \frac{\theta_1}{\phi} = \frac{2Y_1}{\phi l} = \frac{2P_1\delta_t}{\phi l} \quad (2.20)$$

Using equation 2.20, the summation of moments, equation 2.15 can be rewritten as

$$M_e = \frac{4P_1\delta_t}{\phi l} + P_2(l+d) - P_1d \approx P_1l \quad (2.21)$$

Dropping the term $\frac{4P_1\delta_t}{\phi l}$ in equation 2.21 is justified because it is much smaller than the other terms. With the result of 2.21 it is now possible to state that for a moment loading about the specimen mount

$$Y_{1,2} = \frac{M_e}{l} \delta_t \quad \text{and} \quad \theta_{1,2} = \frac{2}{l} \left(\frac{M_e}{l} \delta_t \right) \quad (2.22)$$

A similar analysis will now be performed for the transverse force loading P_e . By summation of forces we can write

$$P_1 = P_e + P_2 \quad (2.23)$$

and similarly by summation of moments about point A

$$M_1 + M_2 + P_2(l+d) - P_1d = 0 \quad (2.24)$$

Because the bar is assumed to be "infinitely" stiff, there exists the geometric constraint $\theta_1 = \theta_2$, therefore $M_1 = M_2$. As in the previous analysis it is possible to write $\theta_1 = \theta_2 = \frac{Y_1}{c}$. The moments can now be defined in terms of c and Y_1 :

$$M_1 = M_2 = \frac{Y_1}{c \vartheta} \quad (2.25)$$

The problem with the above definition is that c is unknown. This is unlike the previous case of determining the deflection caused by a moment where P_1 and P_2 were of equal values and hence Δ_1 and Δ_2 were equal. Based on similar triangles (see Figure 19) we can write

$$c = \frac{l Y_1}{Y_1 + Y_2} \quad (2.26)$$

By using equations 2.9 and 2.26, equation 2.25 can be rewritten as

$$M_1 = M_2 = \frac{P_1 \delta_t + P_2 \delta_t}{l \vartheta} \quad (2.27)$$

By substituting 2.27 into 2.24 and noting that $P_1 = P_e + P_2$, equation 2.24 can be rewritten as

$$P_2 \left(1 + \frac{4\delta_t}{l\vartheta}\right) = P_e \left(d - \frac{2\delta_t}{l\vartheta}\right) \quad (2.28)$$

The terms containing $\frac{\delta_t}{l\vartheta}$ are small, allowing them to be ignored. So now we can write

$$P_2 = P_e \frac{d}{l}, \quad (2.29)$$

and similarly P_1 is found to be

$$P_1 = P_e \frac{l+d}{l} \quad (2.30)$$

The deflections and rotations of the support flexures can now be determined:

$$Y_1 = P_e \frac{l+d}{l} \delta_t \quad (2.31)$$

$$Y_2 = P_e \frac{d}{l} \delta_t \quad (2.32)$$

$$\theta_{1,2} = P_e \frac{l+2d}{l^2} \delta_i \quad (2.33)$$

The support flexures were designed to be fabricated from 0.127 mm (0.005") thick brass sheet. They have a nominal outer diameter of 80 mm (3.15") and an inner diameter of 25 mm (1"). The diaphragm separation, l , is 57.2 mm (2.25"). The load attachment button height, d , is 25.4 mm (1"). The diaphragms are calculated to have an axial stiffness, $1/\delta_a$, of 2900 N/m, a transverse stiffness, $1/\delta_t$, of 140×10^6 N/m, and a rotational stiffness, $1/\vartheta$, of 102×10^3 Nm/rad. The compliance matrix for the system is given below.

$$\begin{bmatrix} x \text{ (m)} \\ y \text{ (m)} \\ \theta \text{ (rad)} \end{bmatrix} = \begin{bmatrix} 3.45 \times 10^{-4} & 0 & 0 \\ 0 & 1.02 \times 10^{-8} & 1.25 \times 10^{-7} \\ 0 & 2.31 \times 10^{-7} & 4.37 \times 10^{-6} \end{bmatrix} \begin{bmatrix} P_a \text{ (N)} \\ P_e \text{ (N)} \\ M_e \text{ (Nm)} \end{bmatrix} \quad (2.34)$$

The double diaphragm construction achieves stiffness in all but the axial direction. The resultant system natural frequency is 27 Hz (167 rad/sec). These values are for small displacements, on the order of one-half the diaphragm thickness (64 μ m or 0.0025"). For small axial displacements, the load is supported by bending. As the displacement increases, part of load is carried by the diaphragm in direct tension [21] stiffening the diaphragm.

Displacement Sensing

Non-contacting displacement sensing was chosen to measure the voice coil position. Strain gages attached to the support flexures were initially considered to measure displacement. Strain gages have been successfully used to measure sub-micrometer motions [17]. Strain gages were, however, rejected because they could be subjected to significant overloads when loads are attached to the sensor. Strain gages could also introduce errors due poor mounting orientation, or a poor bond to the flexure. Capacitive and linear variable displacement transducers (LVDTs)

were compared. An LVDT was selected because of its low cost (\$70), linear response, and the availability of a single chip combined excitation source and conditioning circuit.

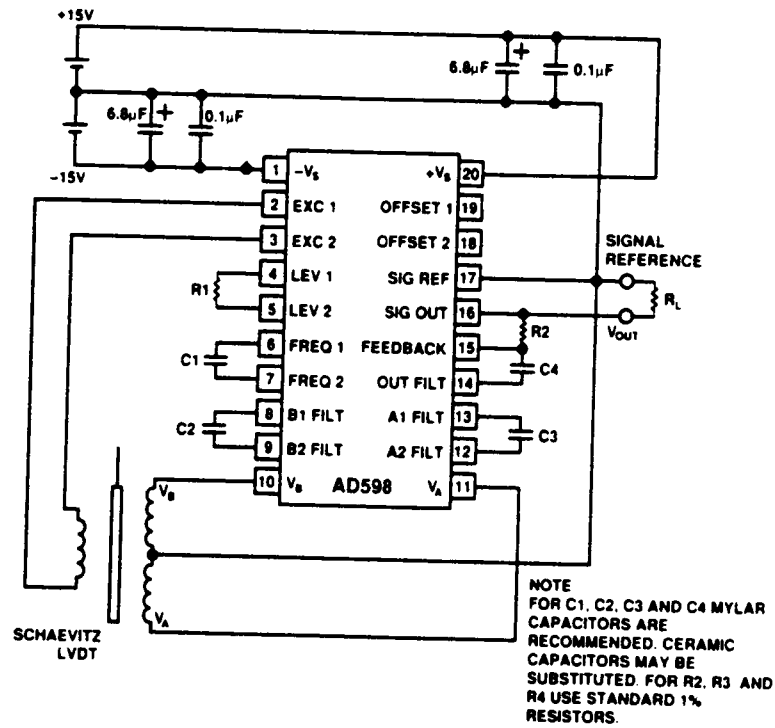
The LVDT selected is a Schaevitz HR-200, a general purpose LVDT with a ± 5.08 mm (± 0.200 in) linear range. This range while significantly larger than that required greatly simplifies mechanical alignment. Excitation and signal conditioning are provided by an Analog Devices AD-598 LVDT signal conditioning chip. Unlike conventional LVDT signal conditioners it uses a ratiometric signal conditioning scheme. The conditioning chip takes the rms amplitude of the signal from each secondary winding. Let V_a and V_b represent the rms voltage of each secondary winding. The following transformation is then performed:

$$V_{out} = G \times \frac{V_a - V_b}{V_a + V_b} \quad (2.35)$$

Where G is the overall signal gain provided by the signal conditioner. This scheme has the advantage that it is insensitive to excitation voltage variations and phase shifts. A complete explanation of the signal conditioner operation is given in the manufacturer data sheets [23].

The complete circuit of the LVDT and signal conditioner is shown in Figure 20. The circuit component values were arrived at by following the design procedures given by Analog Devices in the AD-598 data sheets [23]. The LVDT and signal conditioner combination has a sensitivity of 4 mV/ μ m, a ripple of 0.5 mV rms, and a bandwidth of 3000 Hz. A high bandwidth was chosen to prevent phase shifts from the transducer from adversely affecting the phase margin. An analysis was not performed to see the effect of a noisy displacement signal has on instrument performance. The manufacturer specifications for the AD-598 did not give a noise density information. The only information given was total rms noise content.

The effect of gain variations due to aging or other environmental influences can be compensated by implementing automatic recalibration. With no load attached to the load cell, a known current is sent to the voice coil. The support flexures deflect a set amount from the force



$R1=10\text{ k}\Omega$
 $R2=80\text{ k}\Omega$
 $C1=3.3\text{ nF}$
 $C2, C3, C4=20\text{ nF}$

Offset connections on pins 18 and 19 not used

Figure 20. LVDT circuit schematic.

exerted by the voice coil. The displacement signal is measured and the displacement transducer adjusted. The drift in the zero of the displacement transducer due to environmental conditions can be compensated by re-zeroing before each use. With a micro-processor based instrument this is a simple procedure which can be implemented in software. These ideas are offered as possible improvements for a next generation instrument and will not be implemented in this prototype.

Heat Transfer Analysis

A heat transfer analysis was conducted to ensure that the actuator does not overheat in use. A transient, lumped mass analysis was conducted to determine component temperature as a function of time. Two finite element analyses were conducted to verify if the assumption of uniform internal temperatures required for lumped mass analysis was valid and to ensure that the actuator coil did not have any hot spots. Finally, a finite element stress analysis was conducted to verify thermal deformation of a subassembly. The actuator coil has a maximum rated power of 51 watts and a maximum allowable temperature of 155°C.

For the transient, lumped mass analysis the instrument was broken down into four elements. They are 1) the sensor body; 2) the center post; 3) actuator moving magnet and the upper diaphragm retainer; and 4) the actuator coil and actuator mount subassembly. The overall instrument assembly can be seen in detail in the mechanical design section. The thermal capacity of the air in the internal air gaps was ignored. An initial simplified analysis indicated that the air gap temperature would be within a few degrees of the actuator coil temperature justifying the preceding assumption.

For a lumped mass analysis to have a small error, the Biot number should be small, less than about 0.1. The Biot number is a dimensionless number giving the ratio of internal to external heat transfer resistance. The component with the largest Biot number was the sensor body with a Biot number of 0.005. The convective heat transfer ratio, h_c , was calculated to be

5.4 w/m²K. Radiative heat transfer was included by determining an approximate radiant-heat-coefficient h_r ,

$$h_r = A\epsilon F_t \quad (2.36)$$

where A is the surface area, ϵ is the emissivity, and F_t is given in Figure 21.

The overall heat transfer coefficient h is the sum of h_c and h_r . The equations are derived by writing an energy balance for each lumped mass element:

$$\Delta \text{internal energy during interval } d\theta = \text{heat flow in} - \text{heat flow out} \quad (2.37)$$

The general equation for the temperature of the i^{th} element is

$$C_i \rho_i V_i \frac{dT_i}{d\theta} = \dot{q}_i + \sum_j h_{ij} A_{ij} (T_j - T_i) - \sum_k h_{ik} A_{ik} (T_i - T_k), \quad (2.38)$$

where C_i = specific heat of element i

ρ_i = density of element i ;

V_i = volume of element i ;

$\frac{dT_i}{d\theta}$ = temperature rate of change;

$\dot{q}_i$ = heat generation rate of element i , only the coil has a non-zero value;

$\sum_j h_{ij} A_{ij} (T_j - T_i)$ = the heat transferred from elements j to element i ;

$\sum_k h_{ik} A_{ik} (T_i - T_k)$ = the heat transferred from element i to elements k ;

This gives a system of differential equations that can be solved for element temperatures. A system of four differential equations, one for each element, was programmed in MATLAB™. The resulting temperatures as a function of time are in Figure 22. The initial conditions for the results shown in Figure 22 are a coil power of 4.4 W or a current of 1.17 A and an initial temperature of 294 K (70°F). This current value ensures that the actuator coil does not overheat when steady state conditions are reached. Figure 23 indicates the time to reach the maximum allowable coil temperature as a function of current.

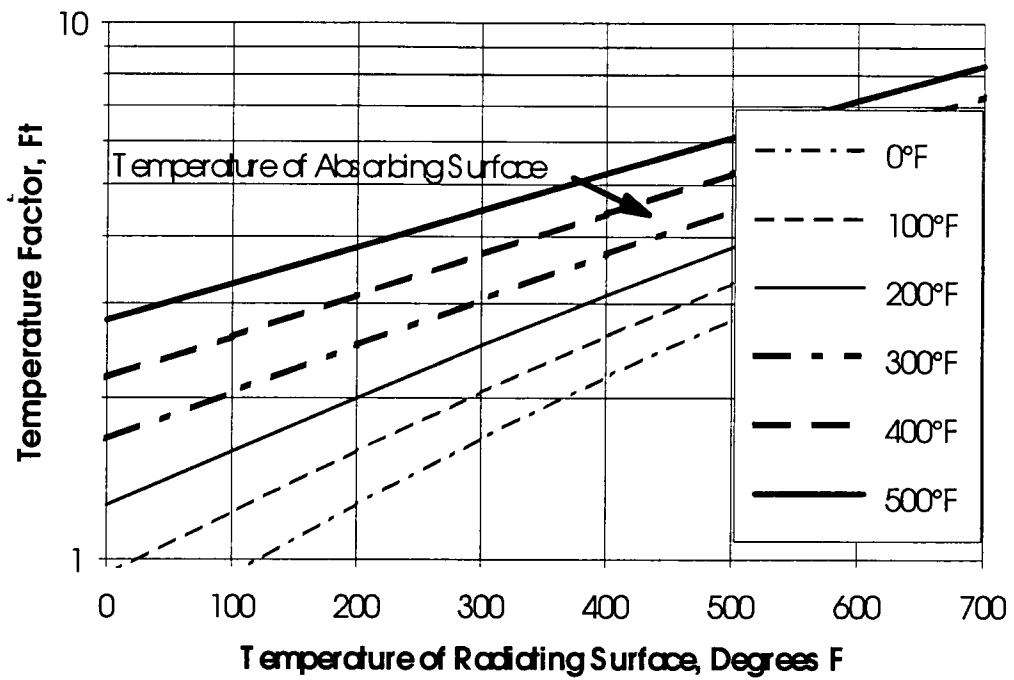


Figure 21. Temperature factor F_t as a function of absorbing and radiating surface temperatures. Adapted from Krieth [24].

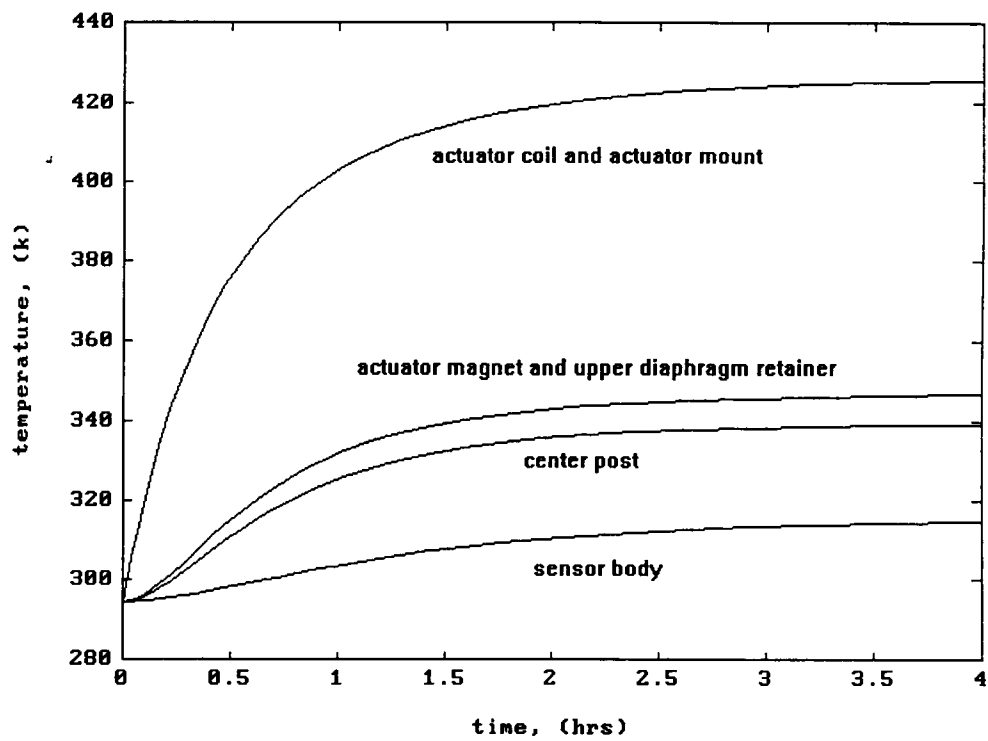


Figure 22. Calculated temperature as a function of time.

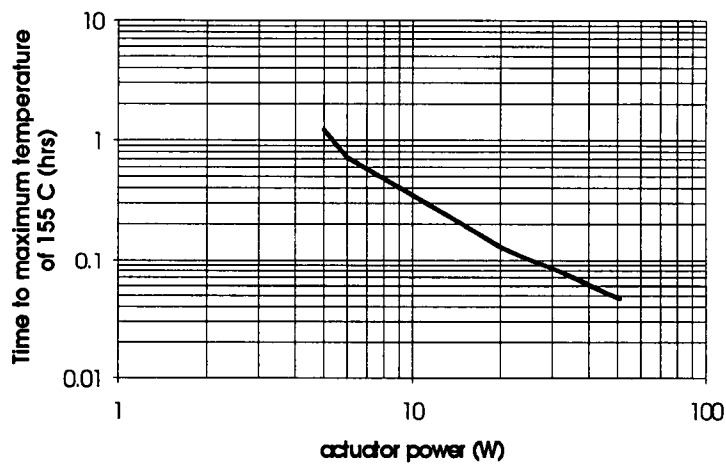
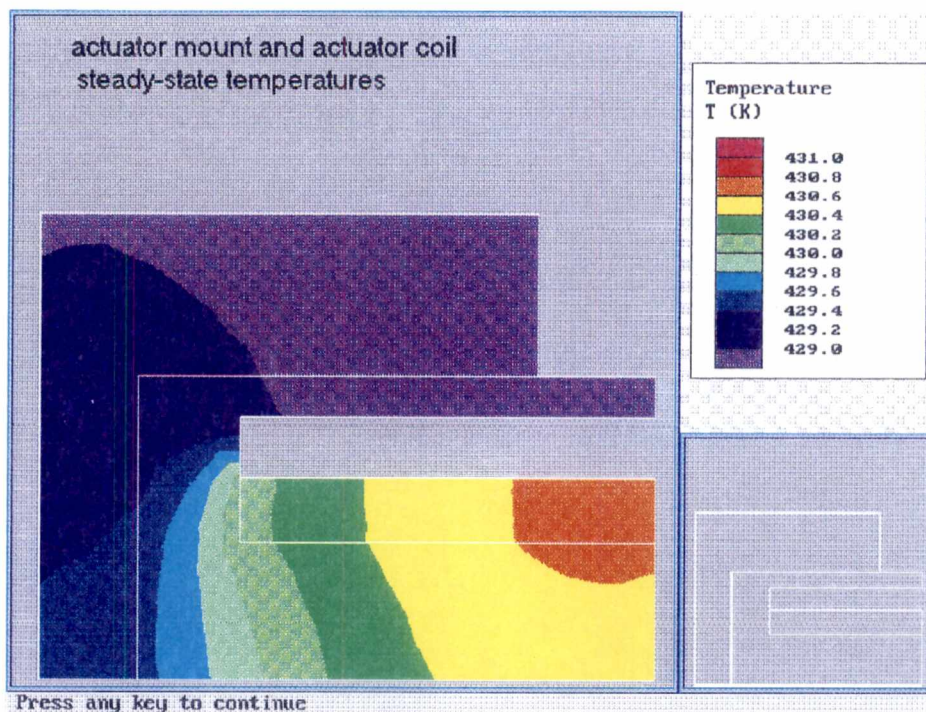


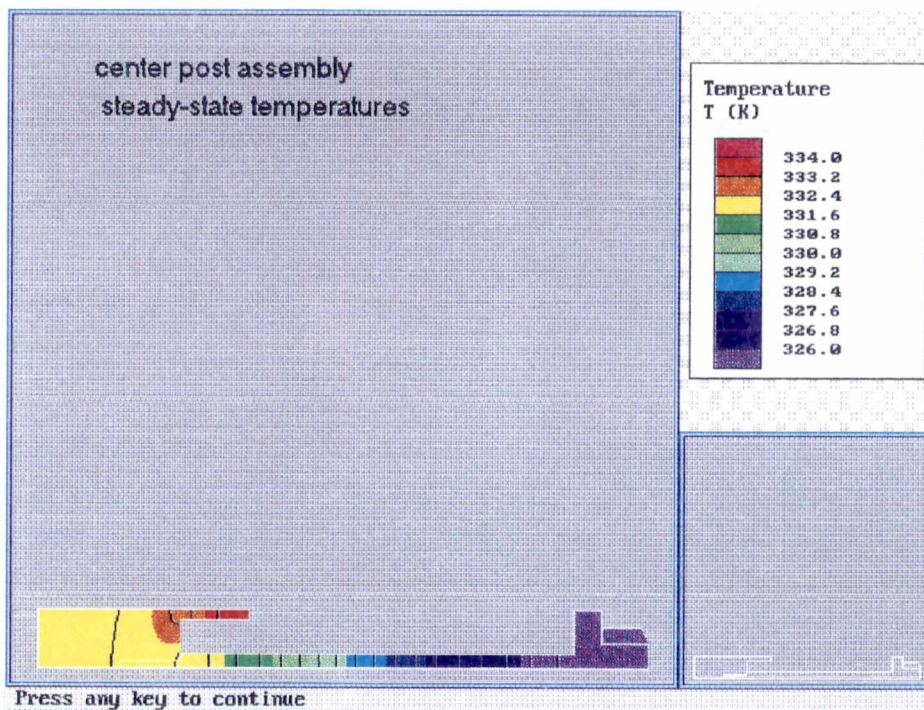
Figure 23. Time to maximum allowable actuator temperature as a function of power.

Following the above analysis a thermal finite element analysis was conducted. The analysis was done using ELCUT a 2-D finite element analysis program for MS-DOS. The two parts analyzed were the actuator mount assembly and the center post assembly. The actuator mount assembly consists of the actuator mount and the fixed section of the actuator. The fixed actuator section in turn consists of the coil windings and the actuator frame. The center post assembly consisted of the center post, upper diaphragm retainer, the actuator moving magnet, and the lower diaphragm retaining nut. The parts were modeled as axisymmetric sections. The boundary conditions for the actuator mount assembly were an internal heat generation of 4.4 W, a surrounding ambient temperature of 420 K, a heat transfer coefficient of $7.5 \text{ W/m}^2 \text{ K}$. The boundary conditions for the center post assembly were an ambient temperature of 420 K for the surfaces adjacent to the actuator coil or within the sensor body, an ambient temperature of 294 K for the surfaces exposed to the atmosphere, and a heat transfer coefficient of $7.5 \text{ W/m}^2 \text{ K}$. The results of the finite element analysis can be seen in Figure 24. The validity of the assumption of uniform temperatures made in the lumped mass analysis can easily be seen. There is no more than a few degrees of temperature variance in the components.

Finally the overall thermal elongation of the center post and upper diaphragm retainer assembly was calculated to be 0.074 mm. A finite element analysis, using ELCUT, of the center post assembly thermal strain indicated an overall deformation of 0.076 mm. The center post was again modeled as an axisymmetric section. These values indicated that some redesign of the LVDT mounting was required. As originally designed the LVDT core was suspended off the bottom of the center post. The large thermal drift of the core position negates the advantage of having a sensitive LVDT. The redesign involved mounting the LVDT off the top of the upper diaphragm retainer. In this position the LVDT can accurately measure the load attachment position of the load cell despite thermal elongation of the center post.



a) actuator mount and actuator coil steady state temperature.



b) center post assembly steady state temperature

Figure 24. Thermal finite element analysis results.

Actuator Power Supply

The initial power supply design considered used an Apex PA-10 op-amp configured as a voltage controlled current supply (VCCS). Calculations, however, showed that using the PA-10 by itself would result in significant output offset current and output current drift. The output current offset would be 7.9 mA, and the output current drift 0.03 mA/°C, due to the PA-10 characteristics. The output current offset is not critical because it can be nulled, drift is another issue. Output current drift in this application is indistinguishable from a change in mechanical loading of the load cell. A drift of 0.03 mA/°C causes the apparent loading of load cell to change at the rate of 0.03 N/°C as the PA-10 warms up.

The final actuator power supply schematic shown in Figure 25 is a VCCS. It uses two op-amps: a precision op-amp, an Analog Devices OP-07, to control the circuit, and a power op-amp, an Apex PA-10, configured as a unity gain buffer to provide the output power. The buffer using the PA-10 has 0 db gain out to 1 MHz, at which point the response rolls off at 20 db/decade.

The DC power supply transfer function is

$$I_o = -\frac{V_{in} R_f}{R_s R_i} \quad \text{given } R_s \ll R_i, R_f \quad (2.39)$$

The resistors R_{cl} , shown in the schematic, limit the current output of the circuit to ± 1 A. Because the load is reactive and in the circuit feedback loop the additional components C_c and R_c are required to stabilize the closed loop. Graphical techniques offer a rapid method for determining closed loop stability for an op-amp circuit. The op-amp open loop gain, A_{ol} , and the inverse of the voltage feedback fraction, $1/\beta$, are plotted together. In our case, A_{ol} is the open loop gain of the OP-07 multiplied by the closed loop gain of the buffer and is shown in Figure 26. To ensure stability and good phase margin the rate of intersection of A_{ol} and $1/\beta$ should be less than 20 db/decade.

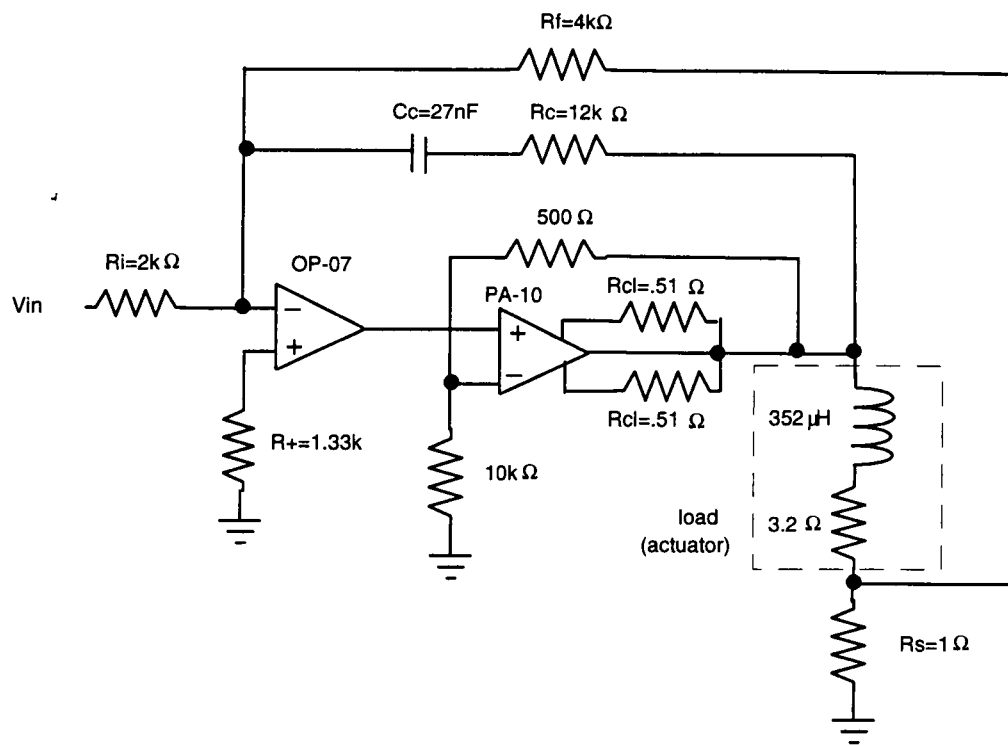


Figure 25. Actuator power supply schematic.

For the feedback path through the load

$$\beta_{load} = \left(\frac{1\Omega}{352 \times 10^{-6} j\omega + 4.2\Omega} \right) \left(\frac{2k\Omega}{2k\Omega + 4k\Omega} \right). \quad (2.40)$$

As can be seen in Figure 26 the rate of closure at the intersection of A_{ol} and $1/\beta_{load}$ is 40 db/decade which indicates instability. To ensure stability, compensation will be required so the values for C_f and R_f need to be determined. Apex makes the following recommendations for determining the compensation required:

1. The compensation feedback curve, $1/\beta_{comp}$, should intersect with A_{ol} 20 db lower than the point of intersection of $1/\beta_{load}$ with A_{ol} .
2. The compensation feedback should have a corner occurring at a frequency at least an order of magnitude lower than the intersection of the compensation feedback with A_{ol} [25].

Using the guidelines above $1/\beta_{comp}$ is set to 20 db and f_{break} is to 500 Hz. The component values are now calculated, $R_f=12k\Omega$ and $C_f=27nF$, giving

$$\beta_{comp} = \frac{1.33k\Omega}{\frac{1}{27nFj\omega} + 13.33k\Omega} \quad (2.41)$$

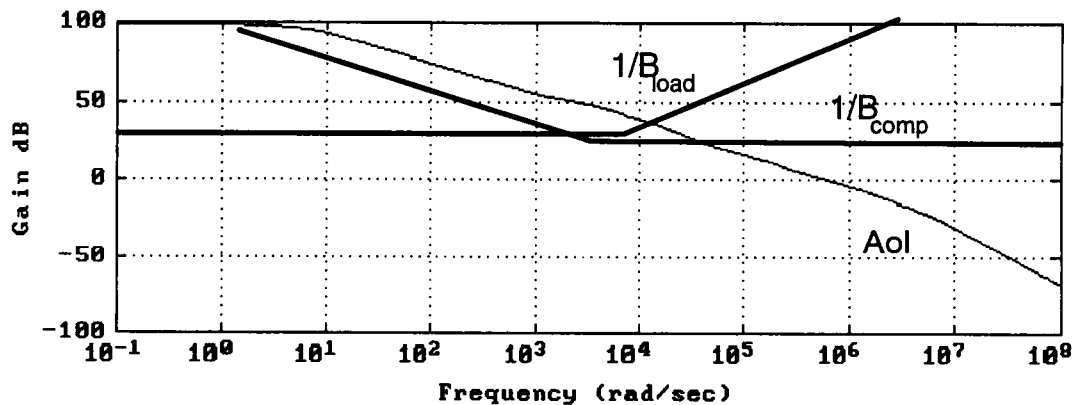


Figure 26. Open loop frequency response characteristics of cascaded op-amps and feedback paths.

The plot of $1/\beta_{\text{comp}}$ is shown in Figure 26. The rate of closure is now 20 db leading to a stable response. The closed loop response of the circuit is shown in figure 27. The circuit was determined to have a 50 db gain margin and 80° phase margin.

The load cell accuracy is affected by offset and drift voltages and currents. As pointed out earlier, the PA-10 by itself did not have suitable characteristics for a precision circuit. The OP-07 was selected because of its low offset and offset drift. Typical offset characteristics for the OP-07 are listed: offset voltage, $V_{\text{os}}=10 \mu\text{V}$; offset current, $I_{\text{os}}=0.3 \text{ nA}$; bias current, $I_{\text{b}}=0.7 \text{ pA}$; offset voltage drift, $\text{TCV}_{\text{os}}=0.2 \mu\text{V}/^\circ\text{C}$; offset current drift, $\text{TCI}_{\text{os}}=5 \text{ pA}/^\circ\text{C}$; and bias current drift, $\text{TCI}_{\text{b}}=8 \text{ pA}/^\circ\text{C}$.

The output voltage offset as a function of the op-amp offset and bias voltage and current is given by

$$V_o = \frac{1}{\beta_{\text{dc}}} (V_{\text{os}} + I_{\text{b}}(R_+ - R_-) + I_{\text{os}}R_-) \quad (2.42)$$

where R_+ and R_- are Thevenin equivalent dc resistances associated with the input terminals, and β_{dc} is the dc feedback. For this circuit $\beta_{\text{dc}}=0.079$ and $R_-=1.33\text{k}\Omega$. As can be seen from the equation above output offset can be minimized by setting $R_+=R_-$. The output offset is now calculated to be $10.4 \mu\text{V}$. Because the force is determined by measuring the current through the actuator this is calculated to be a zero offset error of 0.004 N . This error is not very significant because it can be corrected for by calibration, a more serious error is output drift. This can be calculated to be $2.6 \mu\text{V}/^\circ\text{C}$ corresponding to a measurement drift of $0.001 \text{ N}/^\circ\text{C}$. This circuit design gives an order of magnitude improvement in drift error over the initially analyzed circuit, the PA-10 alone serving as a VCCS.

The offset and drift can be reduced further by using a chopper stabilized op-amp, such as the Maxim MAX-430 with $V_{\text{os}}=5 \mu\text{V}$ and $\text{TCV}_{\text{os}}=0.05 \mu\text{V}/^\circ\text{C}$. This would reduce the measurement drift to $5 \times 10^{-5} \text{ N}/^\circ\text{C}$. Another step that could be taken to further reduce drift is to

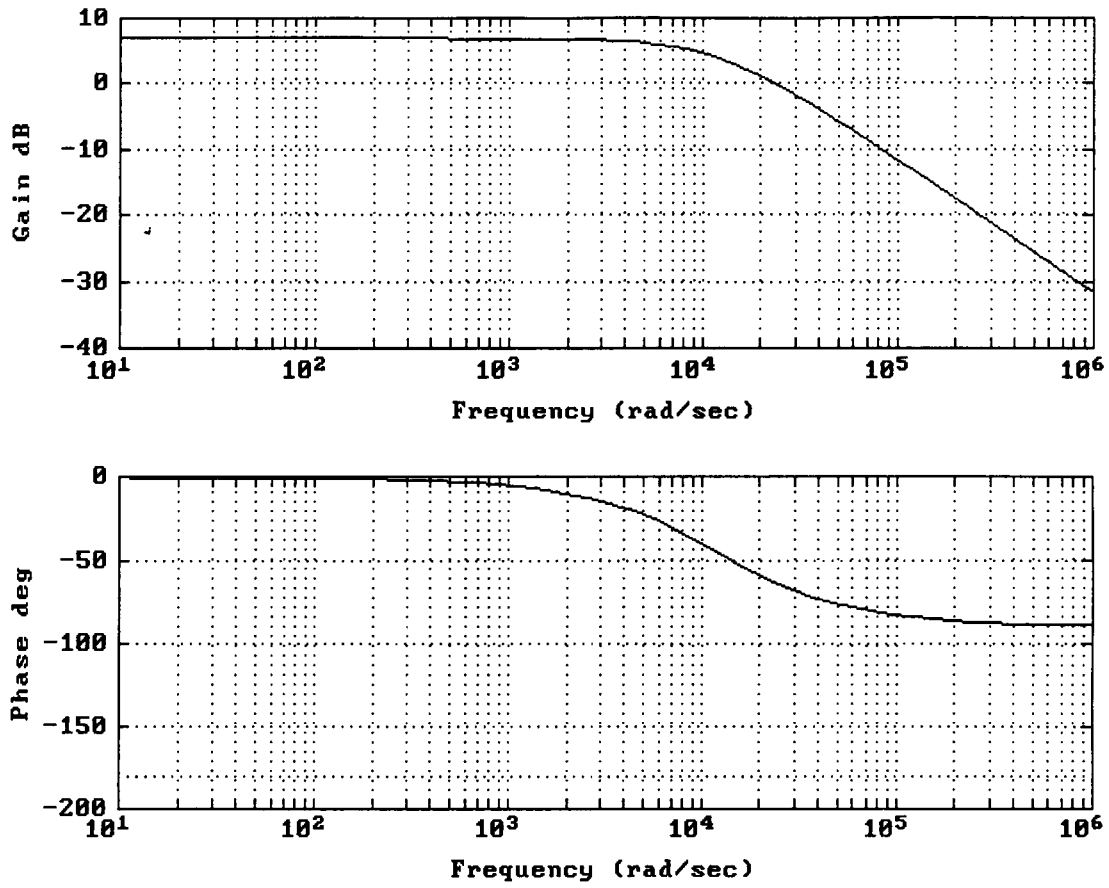


Figure 27. Power supply transfer function. Gain is in units of I_{out}/V_{in} .

place the precision op-amp in a temperature controlled "oven." These steps are pointed out for completeness and will not be attempted in the course of this thesis.

Another source of error is noise in the op-amp circuitry. The input referred noise spectral density for the OP-07 and PA-10 can be found in the manufacturer data sheets in the appendices. This data will be used to calculate the output power noise spectral density and rms noise. Op-amp noise is modeled by treating the op-amp as being noiseless and having an equivalent input referred noise source. The equivalent noise model for the actuator power supply is shown in Figure 28.

Given the frequency response function of a system $H(\omega)$ and an arbitrary input with spectral density $S(\omega)$ it is possible to determine the system's output spectral density, $S_o(\omega)$ by the following relationship

$$S_o(\omega) = |H(\omega)|^2 S(\omega) \quad (2.43)$$

The rms output over a given frequency range is given by

$$V_{o,rms} = \sqrt{\int_{f_1}^{f_2} |H(\omega)|^2 S(\omega) d\omega} \quad (2.44)$$

By examining Figure 28 it can be seen that the noise introduced by the PA-10 can be ignored because of the low gain applied to that noise signal. A_{OP-07} is very large and $\beta_{feedback}$ is relatively constant, therefore the gain, $H(\omega)$, on the OP-07 input referred noise is approximately $1/\beta_{feedback} = 12.6$. For the OP-07 the following noise information is given by the manufacturer:

1. for 0.1 to 10 Hz input noise voltage is $0.6 \mu V_{p-p}$
2. at 1 kHz the input noise spectral density is $11 \text{ nV}/\sqrt{\text{Hz}}$.

From this information the noise spectrum can be approximated as

$$S(\omega) = \frac{1.9 \times 10^{-15}}{\omega} + 1.2 \times 10^{-16} \frac{V^2}{\text{Hz}} \quad (2.45)$$

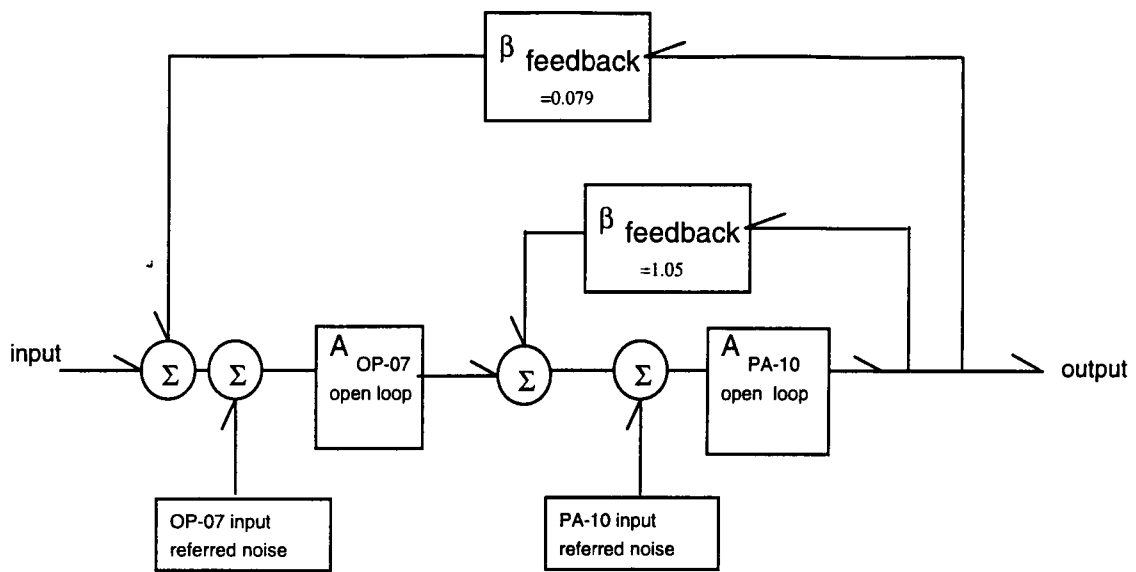


Figure 28. Equivalent noise model for the power supply.

By substituting the value for $H(\omega)$ and equation 2.42 into 2.41, the rms noise voltage, for the frequency range of 0.01 to 5000 Hz, is calculated to be 0.01 mV rms. This leads to a noise current of 0.03 mA rms through the actuator. The Johnson (thermal) noise and input noise current were ignored in the calculation because of the relatively low circuit impedances. The standard deviation of the force measurement error due to op-amp noise is calculated to be 33×10^{-6} N. The measurement error from offset voltages is 0.004 N and from drift is 0.001 N/°C. It can be seen that the dominant error source from the power supply is due to offset and drift.

Control System Design

The control system integrates all the previously discussed components into an operating instrument. The block diagram model of the load cell is shown in Figure 29. The ideal requirements of the control system are 1) to maintain zero displacement under applied loads thus resulting in zero compliance and 2) provide an output signal proportional to the applied load. Because this is a proof of concept instrument a simple analog controller will be designed. The transfer function and physical realization of the controller block meeting the above requirements will be determined in this section. The open loop response of the plant, the series connection of the mechanical response, transducer, power supply, and actuator gain, is shown in Figure 30. The system's mechanical resonance is at 167 rad/sec (27 Hz). A damping value of $c=1.7$ (i.e. $\xi=0.05$) was assumed. Notice the large phase shift associated with the mechanical resonance. The large phase shift around 10000 rad/sec is due to the power supply and transducer response.

A simple controller for this system is an amplifier in series with a bridged-tee circuit for pole-zero cancellation of the mechanical resonance. A bridged-tee circuit was designed with a complex pole pair occurring at 100 rad/sec. Because of the difficulties associated with maintaining exact pole cancellation the cancellation network was deliberately designed to not exactly cancel the mechanical resonance. The difficulties causing inexact cancellation include

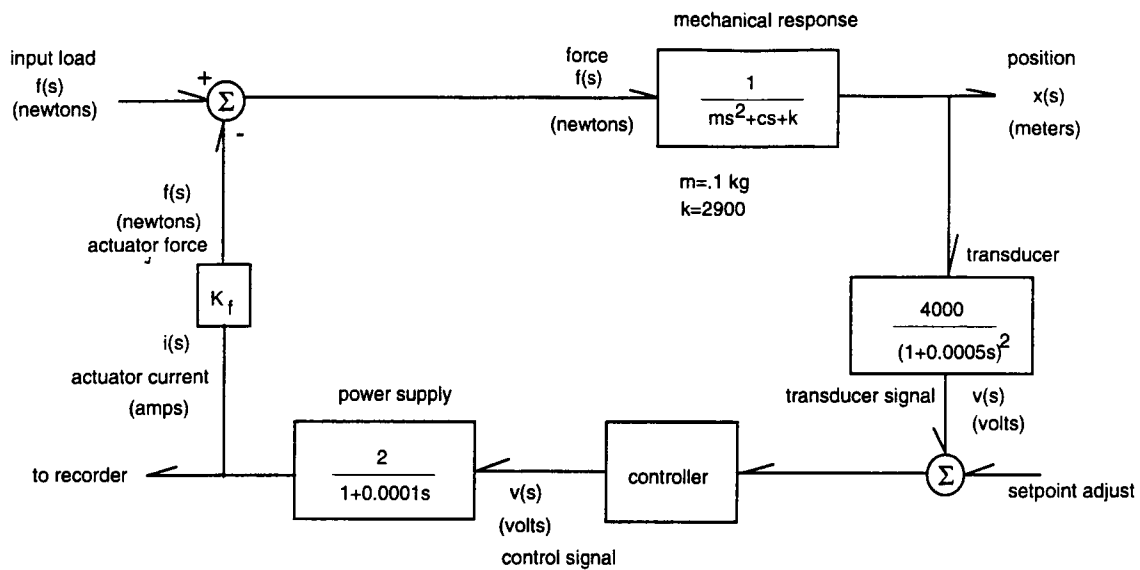


Figure 29. Block diagram of load cell.

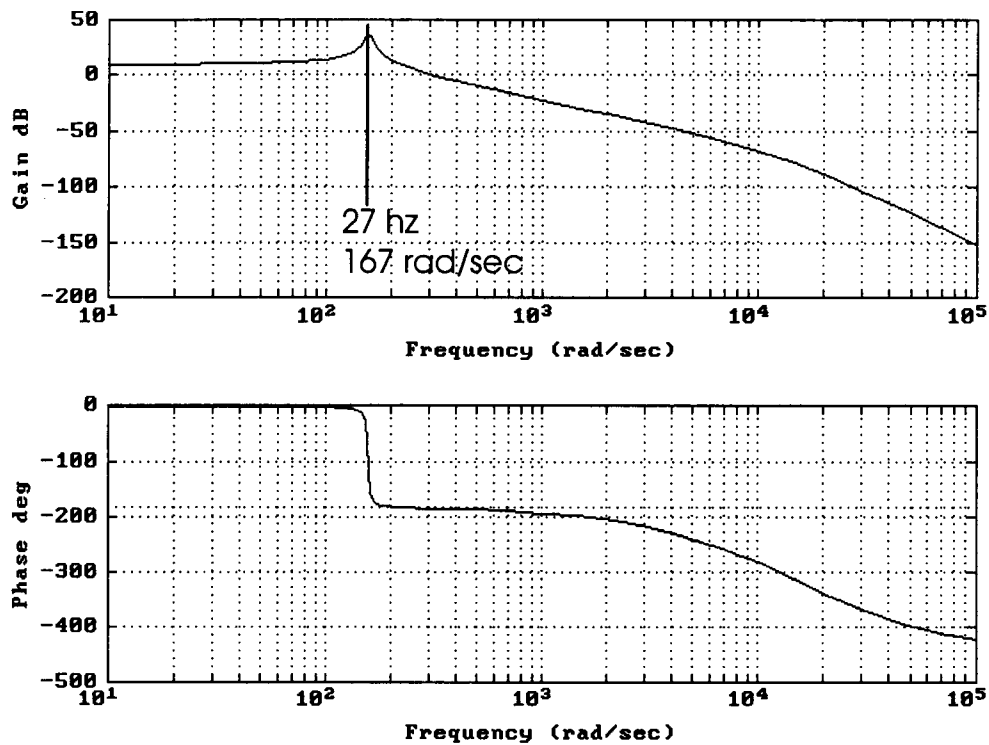


Figure 30. Plant open loop response without controller

plant aging and plant uncertainty. The cancellation network is shown in Figure 31. The compensation network's transfer function is

$$\frac{V_{out}(s)}{V_{in}(s)} = \frac{1 + 2R_1Cs + C^2R_1R_2s^2}{1 + C(R_2 + 2R_1)s + C^2R_1R_2s^2} \quad (2.46)$$

The open loop response resulting from the addition of a bridged tee compensator is shown in Figure 32, along with the uncompensated response for comparison. The responses in Figure 32 are shown in a log magnitude versus phase plot because it allows the gain and phase margins to easily be seen. The uncompensated system has negative gain and phase margins. The addition of the compensator increases the gain margin to better than 35 db and the phase margin to better than 90°. The loop gain can be increased by 25 db and still leave adequate gain and phase margins. The system was determined to be stable using the Nyquist criteria.

Having devised a stable system we now examine its closed loop response. Although quantitative criteria have not been set, the desired response is minimal displacement and accurate tracking of the applied force by the actuator current. The actuator current is the measured signal which is to be proportional to applied force. The relation between current and applied force is the calibration factor. The static closed loop compliance was calculated to be 6.2 um/N. The static closed loop calibration factor was calculated to be 0.9 N/A. The closed loop frequency responses of compliance and calibration factor are shown in Figure 33. The time response of the load cell to unit ramp load, force (N)=time (sec), is shown in Figure 34. A unit ramp load was chosen because it is most representative of the loading applied in the intended application of tensile testing. Figure 34 shows the load cell displacement, actuator current, and measurement error as a function of time. The effect of noise sources has not been taken into account. The measurement error shown in Figure 34c is due solely to system dynamics without noise considerations. The solid line in figure 34c plots the measurement error assuming that

$$\text{input load} = \text{actuator force} = K_f \times \text{current} \quad (2.47).$$

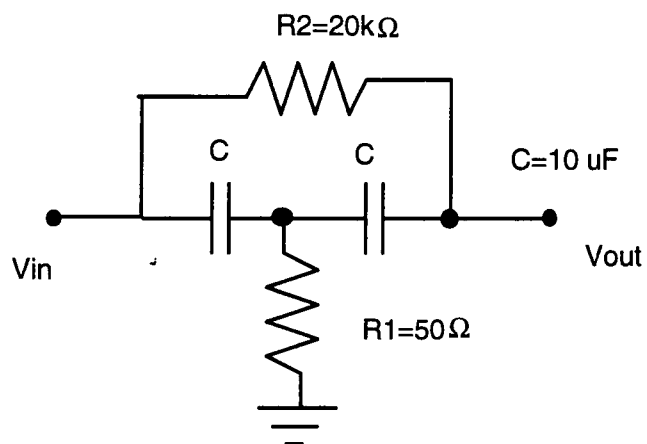


Figure 31. Bridged-tee compensation network

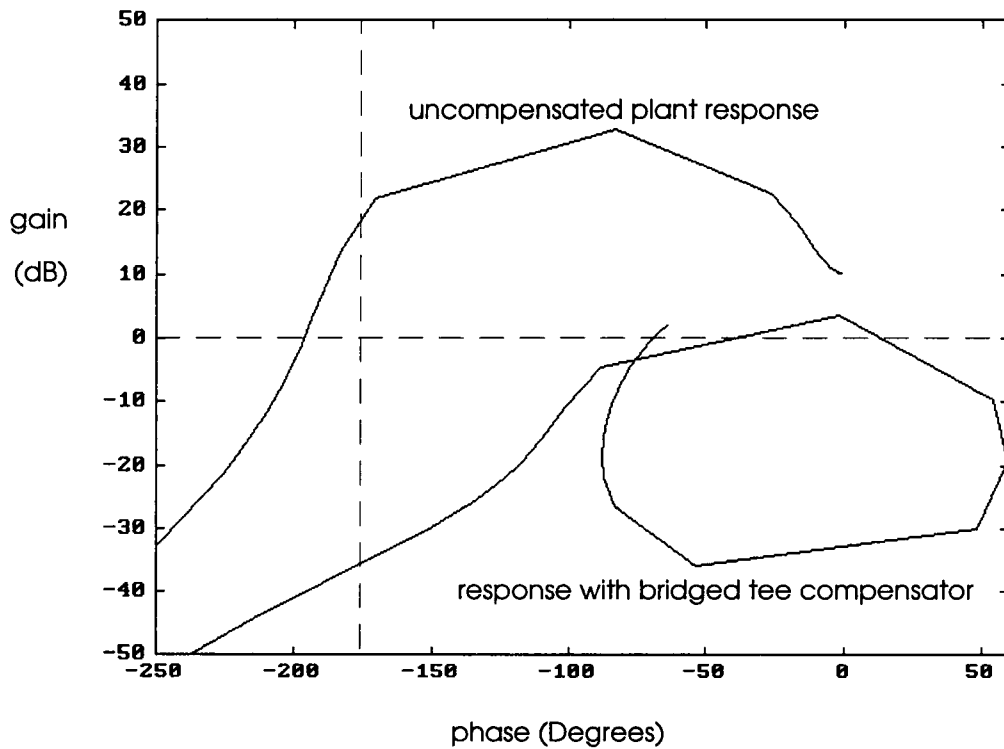
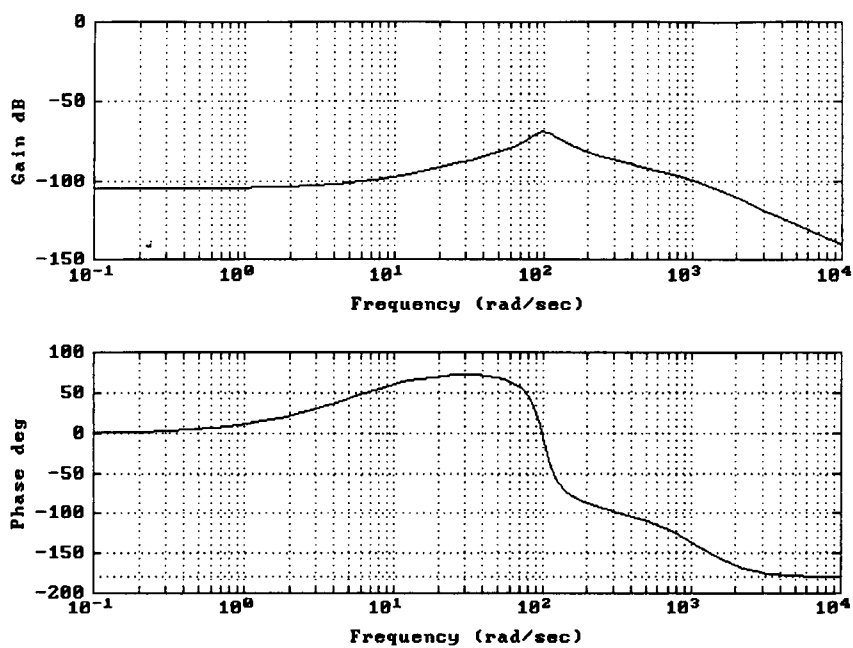
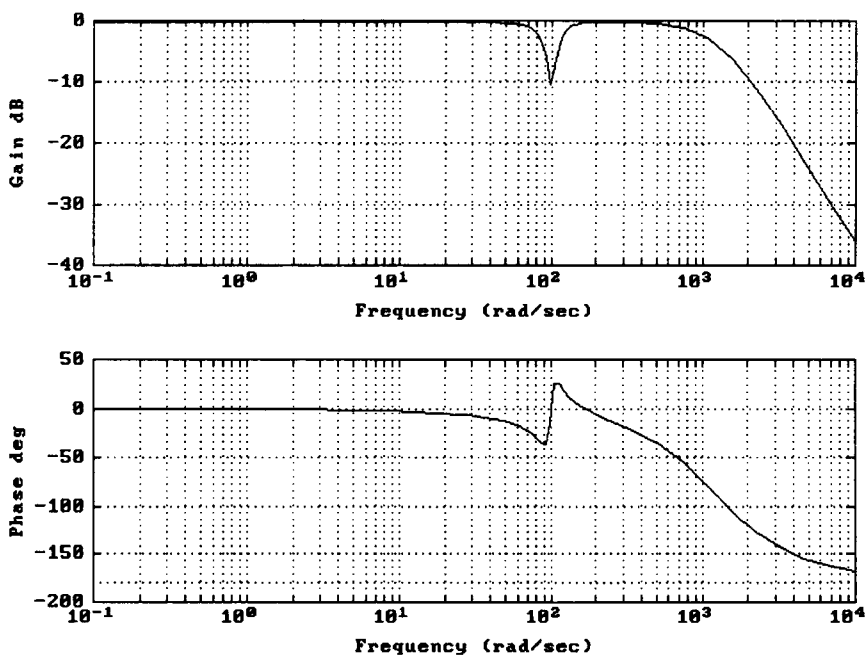


Figure 32. Plant open loop response with and without compensator.

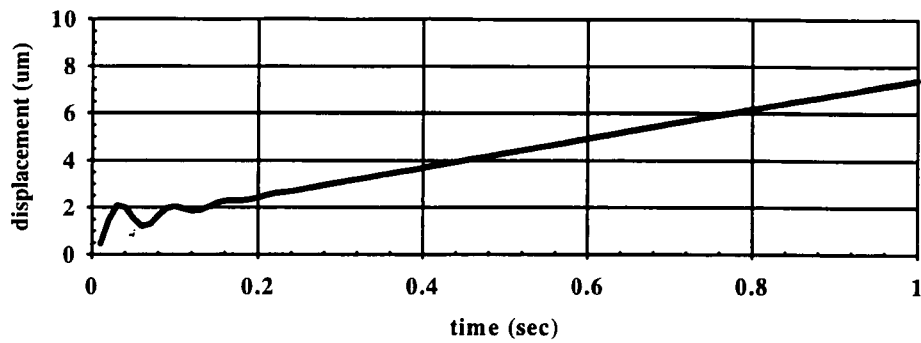


a) output displacement as a function of input load

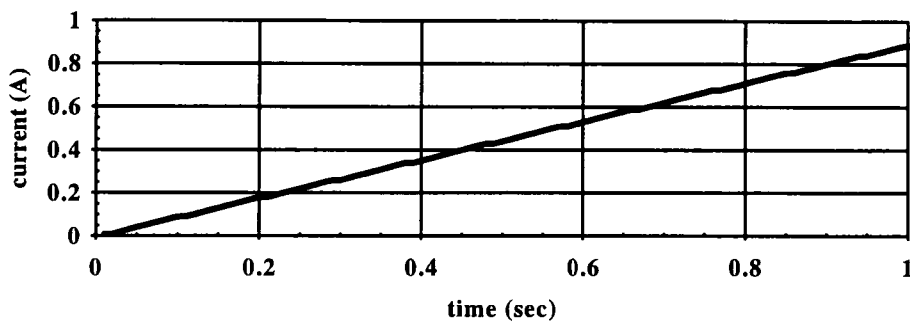


b) calibration factor

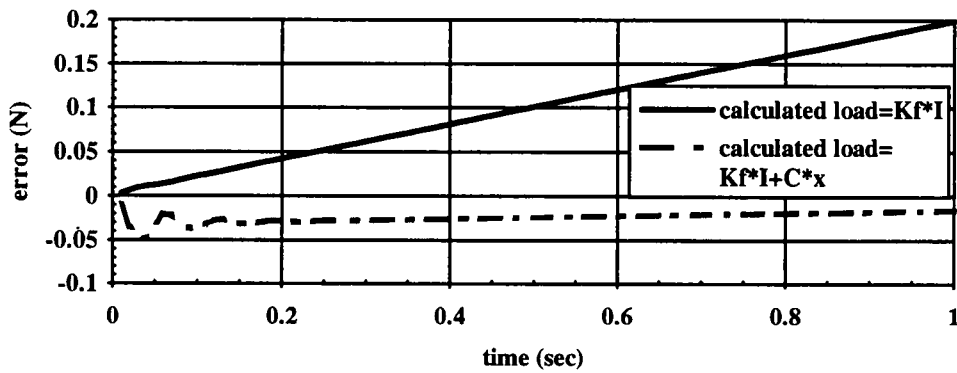
Figure 33. Load cell closed loop frequency response.



a) load cell displacement at load attachment point.



b) load cell actuator current



c) measurement error

Figure 34. Time response of load cell to ramp load.

This assumption is not correct. For static equilibrium of the load cell's moving mass, the force balance is given as

$$\begin{aligned} \text{input force} &= \text{actuator force} + \text{spring force} \\ &= K_f \times I + k \times x \end{aligned} \quad (2.48)$$

The dashed line in Figure 34c plots the measurement error if equation 2.47 is used to determine the applied load.

Can the load compliance be further reduced? By adding a lag-lead compensator in series with the gain and pole-zero cancellation already in the controller it can be. Although it will not be implemented, this solution is of interest for future developments. Over a limited bandwidth the lag-lead compensator acts as an integrator. An integral controller tries to force the integral of the error to zero, minimizing static errors. A true integrator is not used because integrators can have difficulties with saturation. The controller now becomes

$$\begin{aligned} &\text{gain.....lag- lead....pole- zero cancellation} \\ 10 \times \frac{s+12}{s+0.012} \times \frac{s^2+10s+10000}{s^2+2010s+10000} \end{aligned} \quad (2.49)$$

The resulting system compliance is now 0.01 um/N.

SECTION 3: EXPERIMENTAL AND SIMULATION RESULTS

Introduction

This section presents test results of the completed load cell. The actual load cell deviated from the ideal linear system. The electronics in the system had near ideal behavior: low noise and good linearity. The mechanical system, however, had significant hysteresis. The mechanical hysteresis was the largest source of discrepancies between the ideal system analyzed in the previous section and the actual system. Test results of the individual sub-systems are presented then the results of the entire assembled instrument. DC electrical measurements were made using a BK 388A multimeter. Dynamic electrical measurements were made using an HP 3562A dynamic signal analyzer.

Because the mechanical system had significant deviations from an ideal system, the results of a numerical simulation are presented. The simulation includes noise and non-linearities of the electronics and the actuator but has an ideal mechanical plant. This models the best performance that could have been achieved without any mechanical non-linearities.

Power Amplifier Calibration

The output current of the power amplifier as a function of input voltage is plotted in Figure 35. As can be seen in close-up plot there is some hysteresis. The power supply was designed to have a transfer ratio of $-2A_{\text{out}}/V_{\text{in}}$. The achieved transfer ratio was 1.996. The power supply was designed to limit the current at ± 1 A to prevent overheating the actuator. As built it limited the current to +0.8 A and -1 A. Adjusting the current limit is a matter of changing the current limiting resistors. They are shown in Figure 25 in the Design and Analysis section. They were not changed because the as built current limit would not affect testing of the instrument.

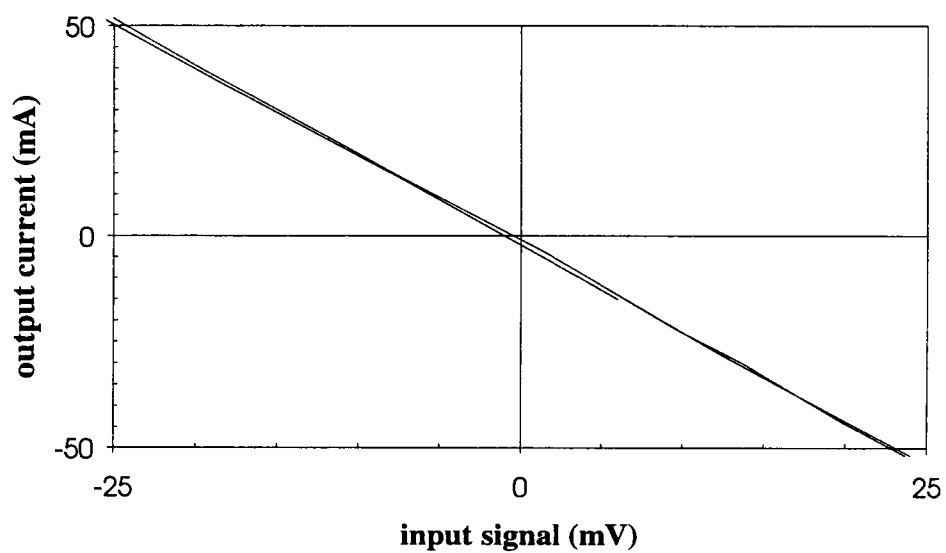
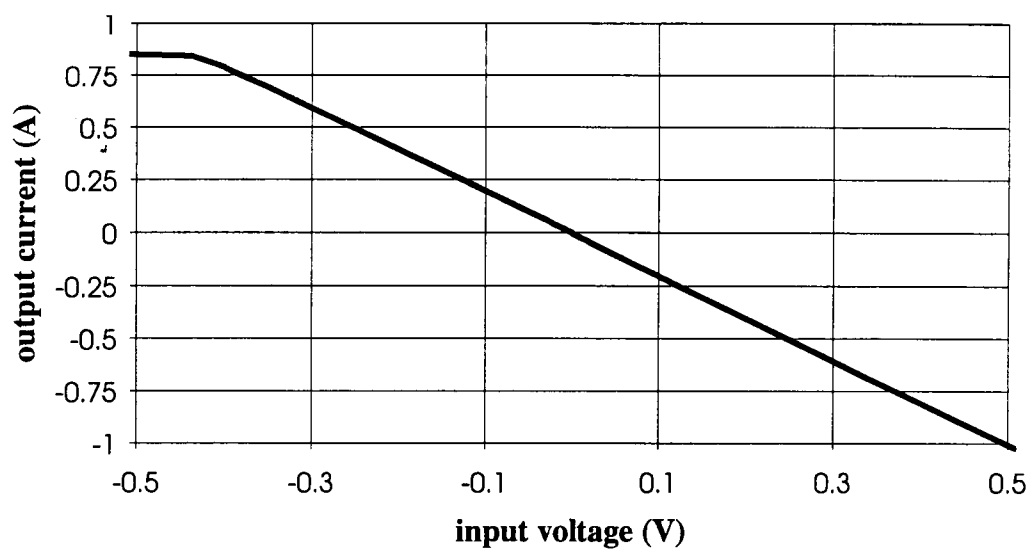


Figure 35. Power supply DC transfer function.

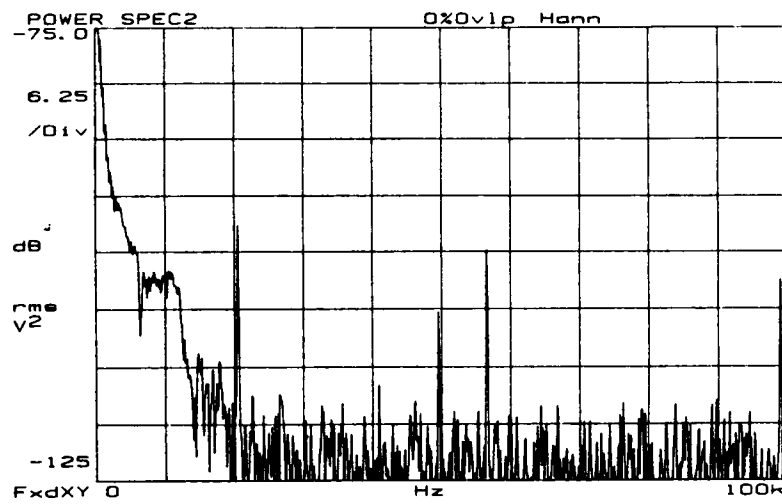
Figure 36 gives the noise power spectral density and the noise distribution over a 5 kHz bandwidth. The measured rms noise current through the actuator was 0.5 mA as compared to a calculated value of 0.003 mA. The reasons the estimated noise and measured noise levels do not correspond are 1) the noise calculation gives at a best an order of magnitude estimate [26] and 2) shielding and isolation were not built into the circuit allowing it to pick up external noise. The bandwidth of the amplifier as designed was 1500 Hz. Initial testing showed it was considerably lower. The resistor R_f was changed from 2 k Ω to 2.8 k Ω slightly increasing the amplifier's bandwidth to 300 Hz. Figure 37 shows the frequency response function of the power supply with the 2.8 k Ω resistor. The actual bandwidth of 300 Hz does not match the as designed bandwidth because of stray capacitances, inductances, and resistances present in real components.

LVDT Calibration

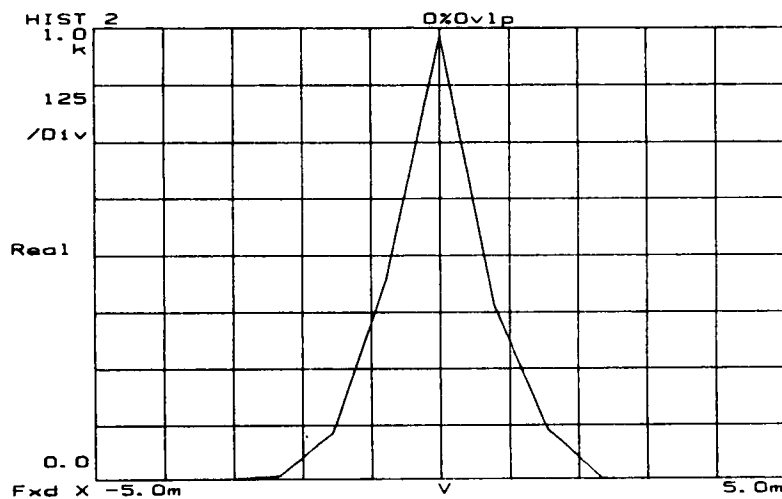
The LVDT calibration, voltage output as a function of input displacement is shown in Figure 38. A close-up of the calibration curve shows the LVDT hysteresis. The LVDT has a range of 5000 μm , it was only calibrated over a range of 2500 μm because its full range was not required. The calibration constant is 2770 V/m. The noise spectrum and noise distribution histogram are shown in Fig 39. Noise spikes can be seen at multiples of 10 kHz. The LVDT excitation frequency was set at 10 kHz and the noise spikes are excitation signal leakage.

Mechanical System Testing

The assembled instrument is shown in Figure 40. The mechanical performance of the instrument was compromised by the non-linear performance of the flexures. The mechanical flexures as initially fabricated had a significant curvature, rather than being flat. The flexures are fabricated of brass sheet. The curvature was probably due to residual stresses in the sheet from the rolling operation to produce the sheet stock. The instrument was assembled and tested. The flexures rather than deforming elastically when loaded would pop audibly and move to the



a) noise power spectral density



b) noise amplitude distribution

Figure 36. Experimentally measured power supply noise characteristics.

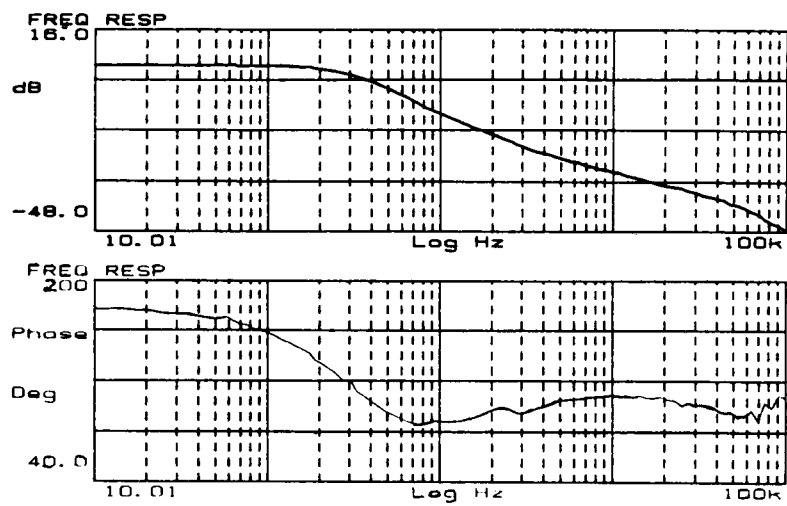


Figure 37. Experimentally measured power supply frequency response.

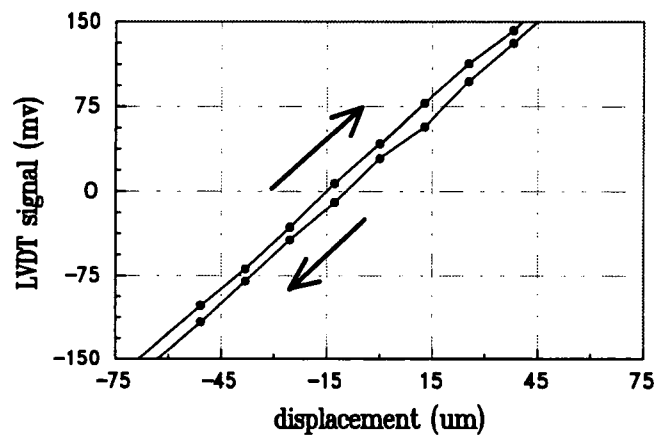
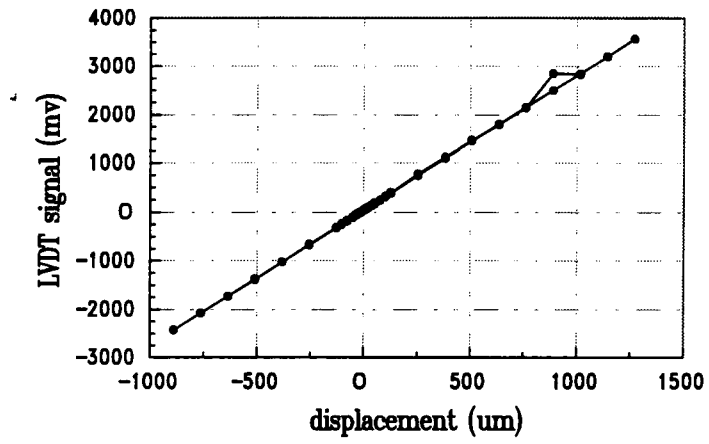
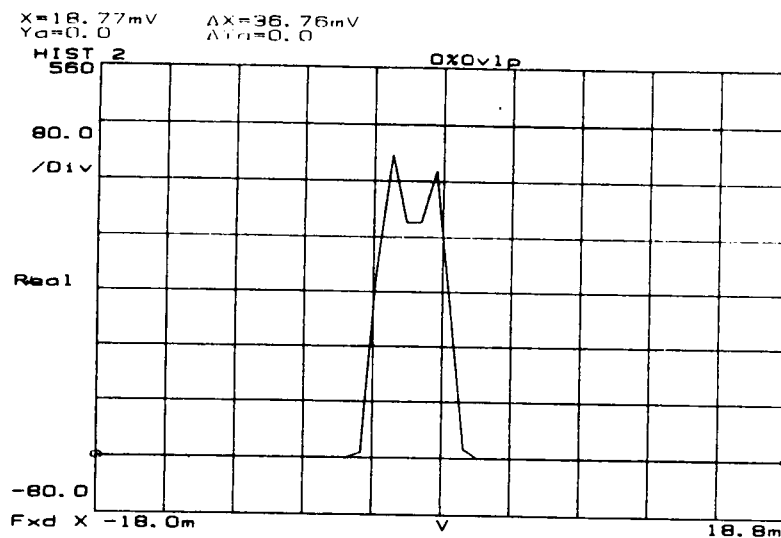
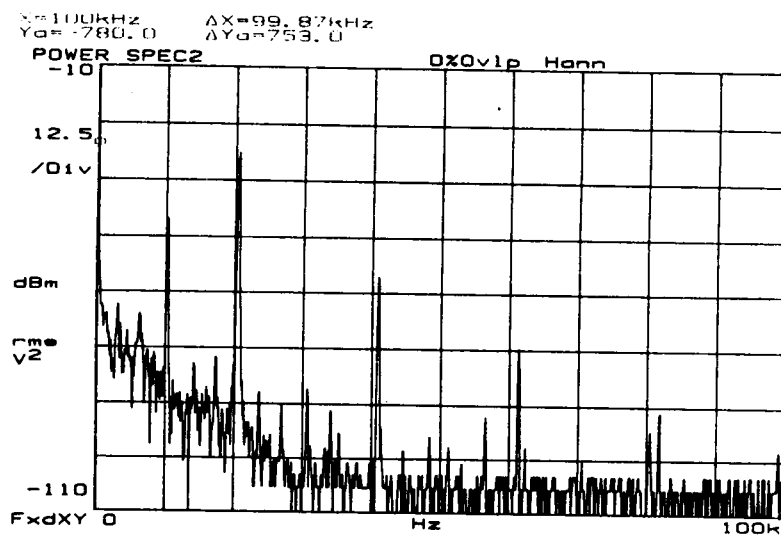


Figure 38. LVDT DC transfer function.

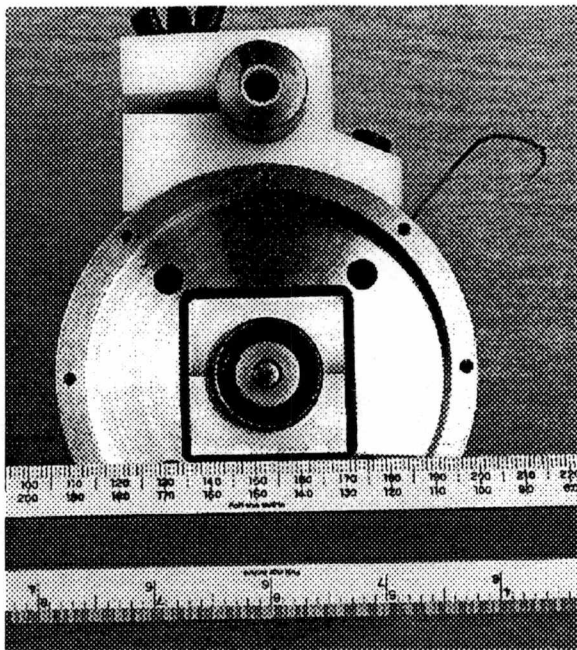


a) noise amplitude distribution histogram, over a 5 kHz bandwidth

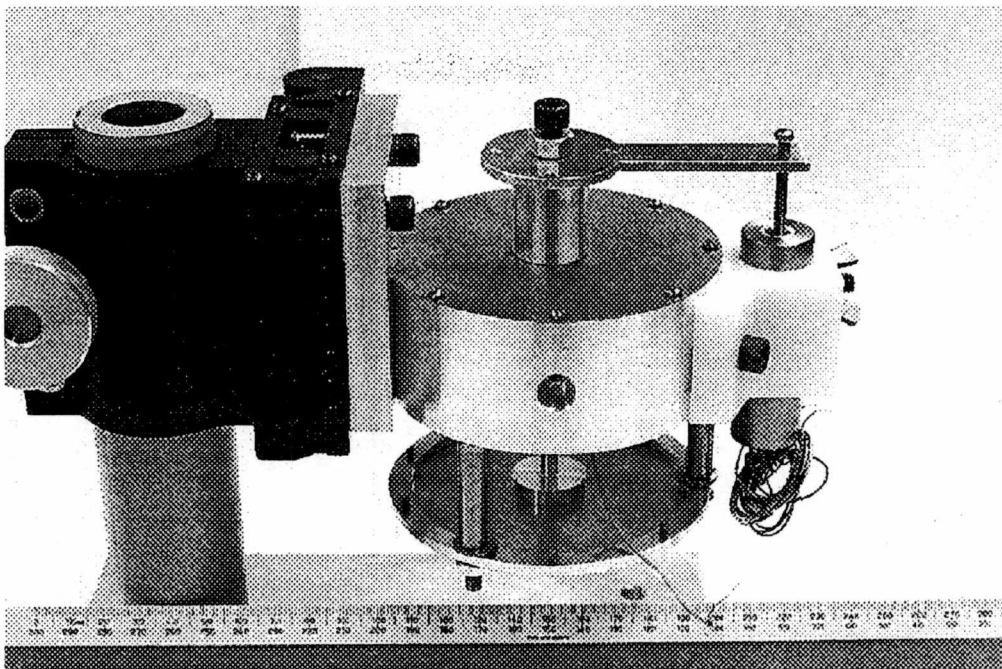


b) noise power spectral density

Figure 39. Experimentally measured LVDT noise characteristics.



a) view looking into sensor body. Upper flexure is not yet mounted. Visible in the center of the picture are the actuator and the center post.



b) view of assembled instrument. Note the LVDT mounted on the side.

Figure 40. Photographs of instrument.

extreme position of their motion range. The flexures were at this point unusable. To remove the curvature they were oven annealed at 260 °C for one hour.

The instrument was reassembled and tested. The deflection as a function of load for the annealed diaphragm was measured. The diaphragm performance was improved by annealing but still exhibited considerable hysteresis. Gibbs suggests cutting slots into instrument diaphragms to improve their linearity[27]. This was attempted. Unfortunately, it did not improve performance. The slotted diaphragm is shown in figure 41. The plot of the load-deflection curves comparing unslotted and slotted diaphragms is shown in figure 42. Slotting the diaphragms reduced the stiffness slightly. The mechanical resonant frequency with unslotted diaphragms was 37 Hz. The calculated resonant frequency was 25 Hz. The discrepancy arises due to the flexures stiffening for large deformations (i.e. motions larger than one-half the diaphragm thickness). Slotting the diaphragms lowered the resonant frequency to 25 Hz.

The final attempt to improve flexure performance was to apply a dither signal to the actuator. The reasoning was that if the hysteresis was caused by stick-slip friction at the bolted interfaces maintaining the flexures in continual motion would minimize the effects of stick-slip. This idea was tested by applying a 125 Hz sine wave to the actuator as the dither signal. The dither signal frequency was chosen to be higher than the natural frequency to avoid causing large displacements of the load attachment point. The sine wave signal introduced a varying load on the flexures. A DC offset was added to the signal to simulate the effect of a constant load on the flexures. The effective load applied was determined by calibration data for the actuator. The results of this test are shown in Figure 43. Applying a dither signal did not improve linearity of the flexures. Although the results are not presented here, a 40 Hz dither signal was also tested. It also resulted in no improvement in performance.

The effect of the hysteresis was to result in a plant whose response varied with excitation signal amplitude. The plant response was determined by exciting the power amplifier signal input with a sinusoidal voltage signal and measuring the output signal from the LVDT signal

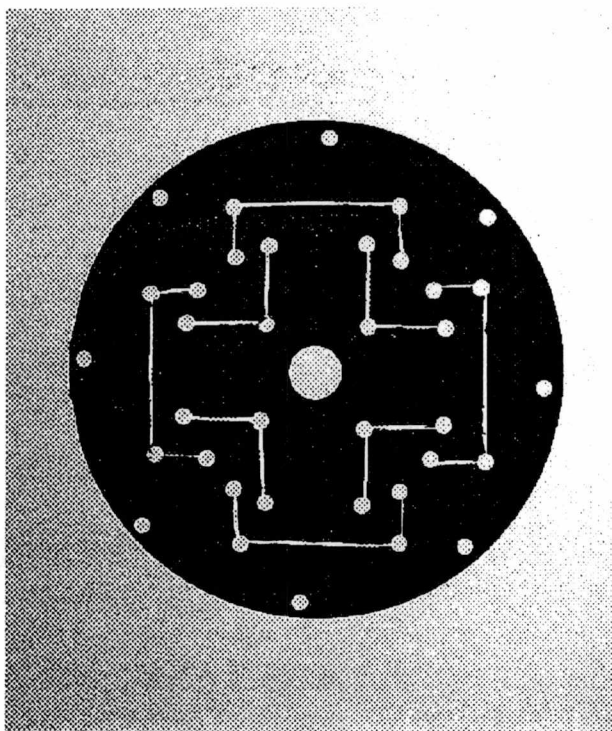


Figure 41. Photograph of slotted flexure diaphragm.

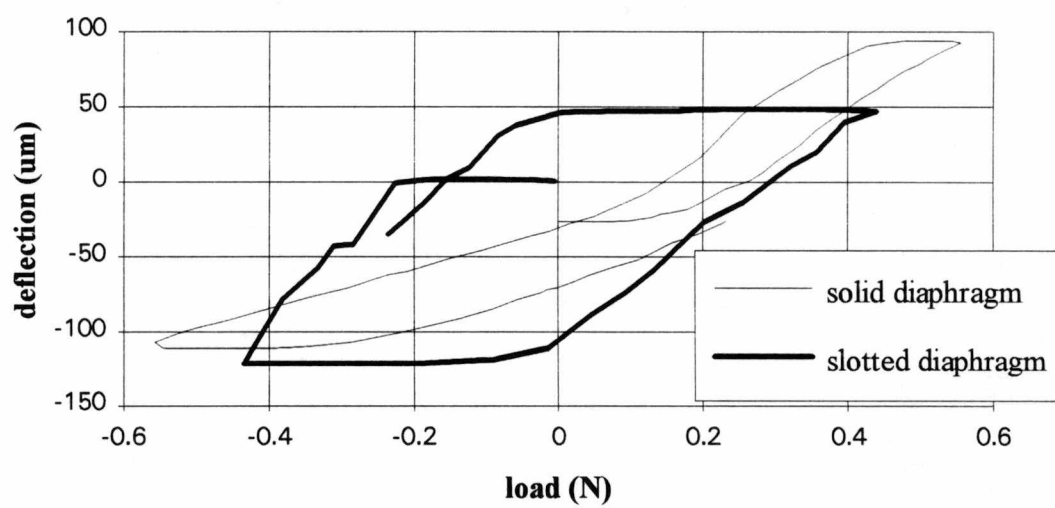


Figure 42. Comparison of response for slotted and unslotted diaphragms.

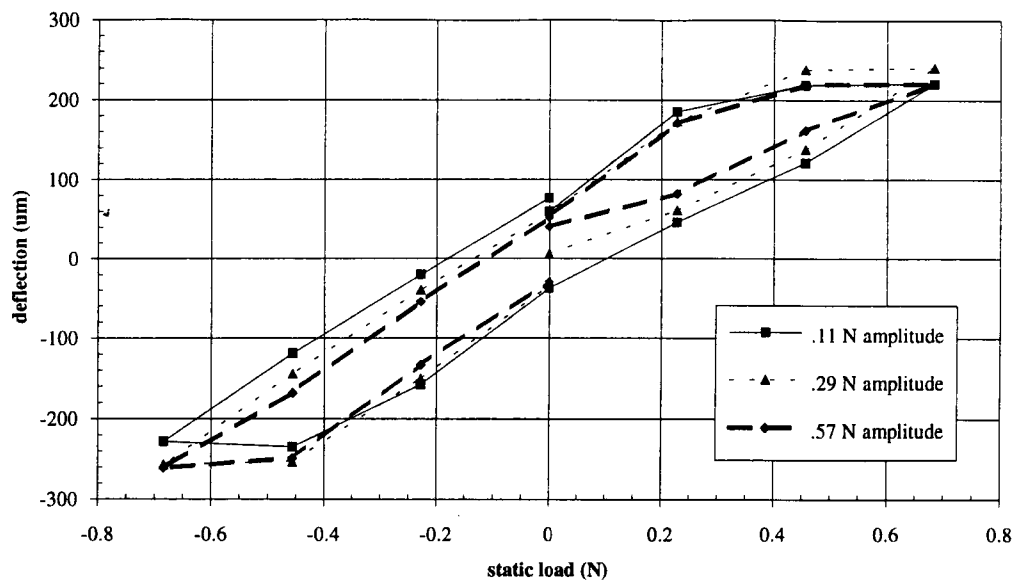


Figure 43. Effect of dither on flexure linearity. 125 Hz dither signal applied.

conditioner. Figure 44 shows the plant response for three different excitation amplitudes (145 mV, 170 mV, and 250 mV excitation amplitude). Note the changes in plant response with excitation level. The frequency response function (FRF) was determined by using the signal analyzer to excite the system and measure its response. The frequency response testing input was the power amplifier and the output was the LVDT signal. The illustrated FRFs are for the power amplifier, mechanical plant, and LVDT in series. At low frequencies (<100 Hz), the electronics do not significantly affect the FRF other than by shifting the gain. The electronic dynamics are essentially invisible at low frequencies.

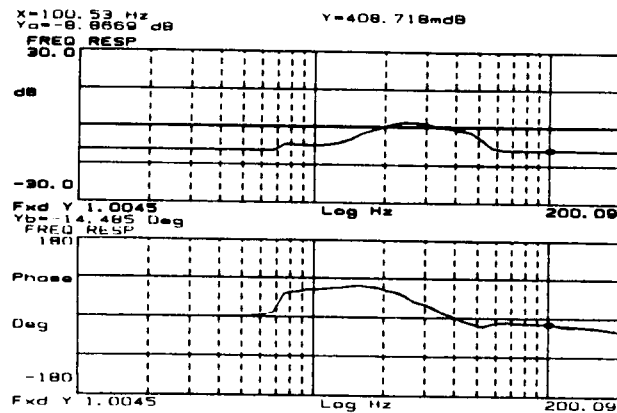
A transfer function was fitted to the data from FRF for the 250 mV excitation. This FRF was used for fitting because it most closely modeled the response of a second order spring mass system: 180° phase shift at resonance. This implied that this FRF was the least "contaminated" by nonlinear effects. The fitted transfer function is given by

$$\frac{2.92 \times 10^4}{s^2 + 19.2s + 2.21 \times 10^4} \quad (3.1)$$

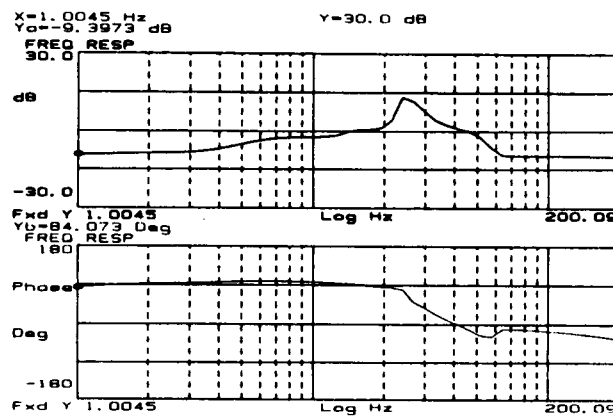
the discrete data points used fit the transfer function and the fitted transfer function are shown in Figure 45. The fitted transfer function will be used later for simulation. The electronic gains will be factored out to give the response of just the mechanical plant.

Complete Load Cell Testing

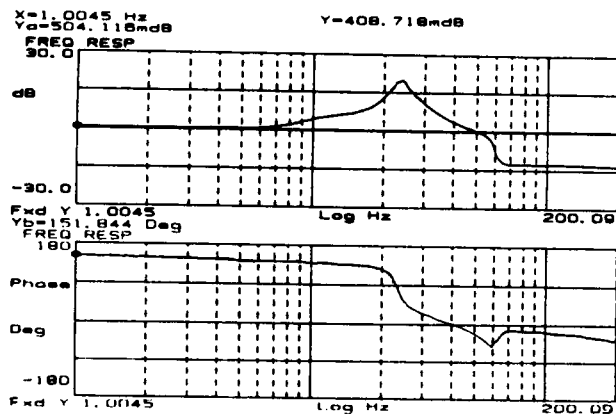
The test results of the completed load cell are discussed in this section. Due to the significant mechanical non-linearity initial analysis and simulation were conducted to determine if the instrument could still work. The main concern was that the instrument would limit cycle and be unable to operate stably. Numerical simulations were done using a block diagram simulation language, VisSim. A system is modeled in VisSim by graphically selecting and then connecting blocks. The blocks can represent a variety of linear and non-linear operators. A block diagram used to simulate the load cell is shown in figure 46.



a) input voltage 145 mV



b) input voltage 170 mV



c) input voltage 250 mV

Figure 44. Open loop frequency response as a function of input voltage.

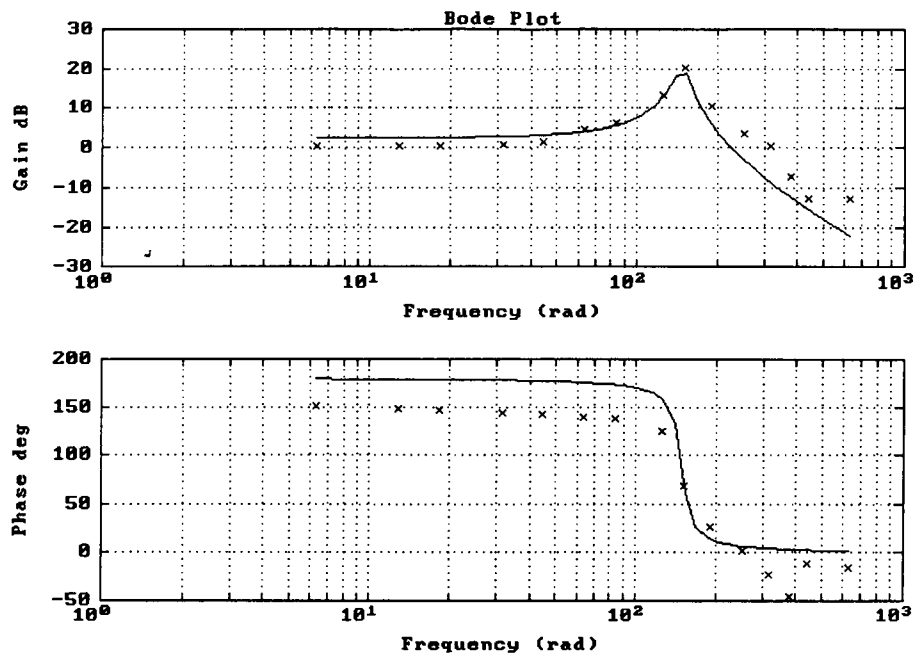


Figure 45. Comparison of experimental and fitted frequency response.

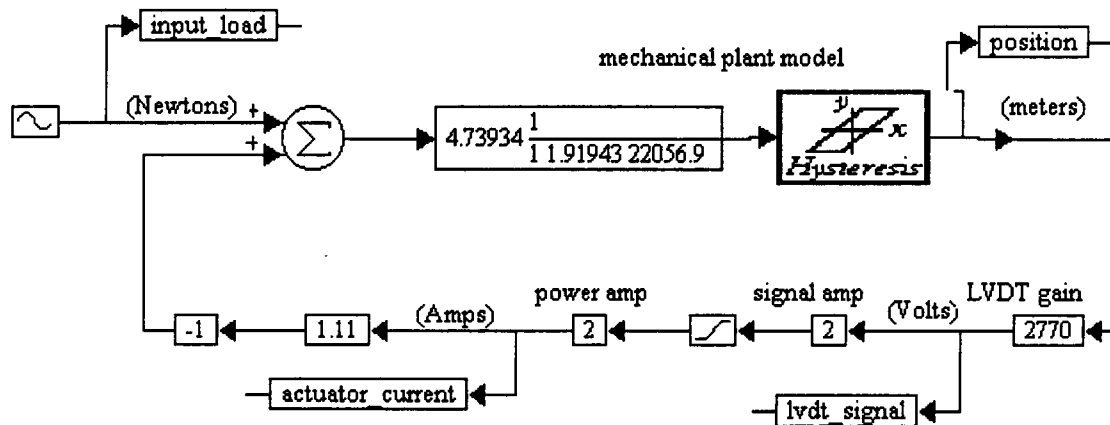


Figure 46. VisSim block diagram simulation of the load cell. The mechanical hysteresis and signal amplifier clipping non-linearities are included in this simulation.

To close the feedback loop in the instrument a signal amplifier was placed between the output of the LVDT signal conditioner and the input of the power amplifier. The signal amplifier has adjustable gain. The initial testing and simulation did not include the pole-zero cancellation network designed in the design and analysis section. Because of loading effects the output of the signal amplifier output clipped at ± 300 mV. The saturation could have been removed by changing the resistor values on the input to the power amplifier. This was not done because the simulation indicated that eliminating the clipping would increase the limit cycle amplitude. This was also verified experimentally, though data was not collected because the limit cycle amplitude was large enough to damage the instrument making testing impractical.

Table 5 compares limit cycle amplitude and frequency as a function of signal amplifier gain. The table presents both experimental and numerical simulation data. Both the simulation and experimental data show that the limit cycle amplitude and frequency are independent of the signal amplifier gain and that there is a gain setting below which the system will operate stably. The simulation and experimental data however are not in agreement with each other on the limit cycle amplitude and frequency. The reason for this is that an empirical model of the mechanical plant is used in the simulation. This model does not fully capture the dynamics of the real mechanical plant. Unmodeled dynamics include stick-slip and spring hardening. Spring hardening at large deflections would raise the natural frequency and minimize displacement. The simulation predicts that during limit cycling position as a function of time is a sine wave. Figure 47 shows test data of position as function of time during limit cycling. Notice the significant distortion from a sine wave. For comparison, the behavior of the system was examined using describing function methods [20, 28]. Considering only the clipping nonlinearity, determined by simulation to be the dominant nonlinearity, a limit cycle frequency of 40 Hz was predicted by describing function methods.

Following the limit cycle testing the load cell was tested to determine its measurement accuracy and the ability of the control system to reduce its compliance the load cell was loaded

Table 5. Limit cycle amplitude and frequency as a function of signal amplifier gain.

signal amplifier gain linear (dB)	simulation:		experimental:	
	amplitude μm	frequency Hz	amplitude μm	frequency Hz
0.3 (-10.5)	—	not simulated	—	no limit cycling
1.3 (2.3)	—	no limit cycling	—	not tested
1.4 (2.9)	1083	23.4	—	not tested
2.6 (8.3)	1083	23.4	113	34.5
3.9 (11.8)	1083	23.4	75	35.3
15 (23.5)	1083	23.4	75	36.0
49 (33.8)	1083	23.4	83	37.0
122 (41.7)	1083	23.4	109	34.4

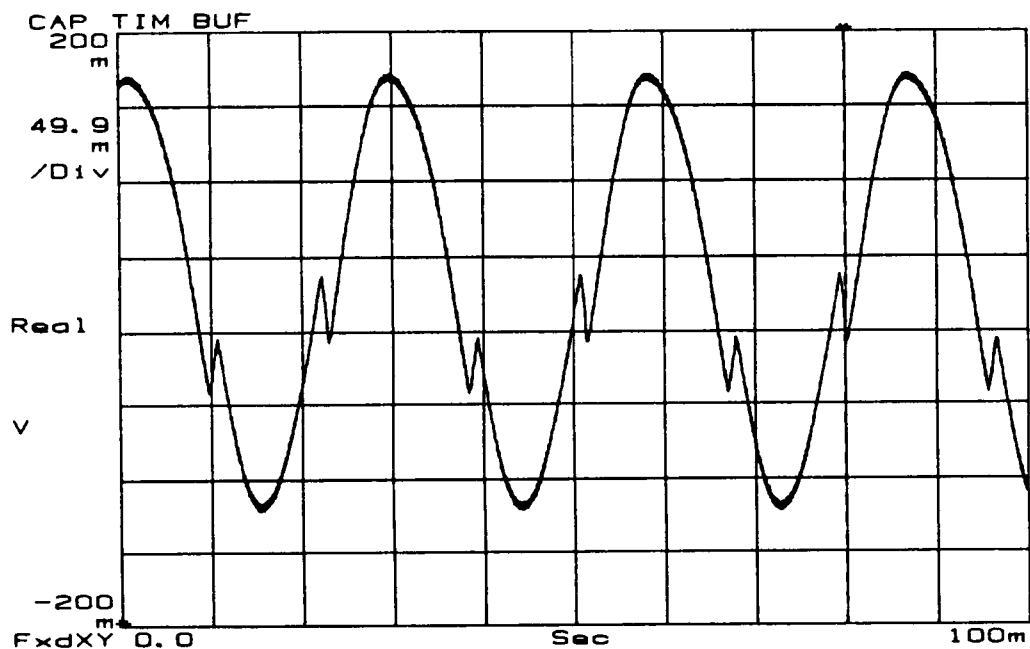


Figure 47. Experimentally observed limit cycling.

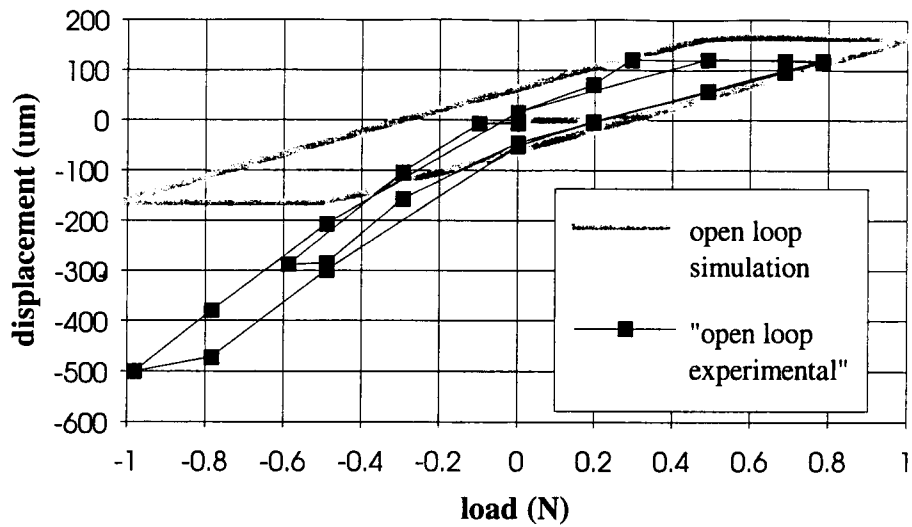
with the signal amplifier disconnected, that is with the controller turned off. The deflection as a function of load was recorded. The nominal load cell compliance without the control system was $333 \mu\text{m/N}$. The signal amplifier was reconnected and its gain adjusted to 2 (6 dB). The nominal load cell compliance with the control loop closed was $38 \mu\text{m/N}$. The control system resulted in almost a ten-fold reduction in compliance. Both the open loop and closed loop cases were simulated using VisSim. The experimental and simulation results of deflection as a function of load are shown in figure 48.

The actuator current as a function of load was also determined experimentally and by simulation. Ideally the current should be linearly proportional to load. The current being measured to determine load. Due to the substantial mechanical hysteresis, the current as a function of load also showed substantial hysteresis. The experimental and simulation results are shown in Figure 49. The nominal calibration factor is 1.4 N/amp .

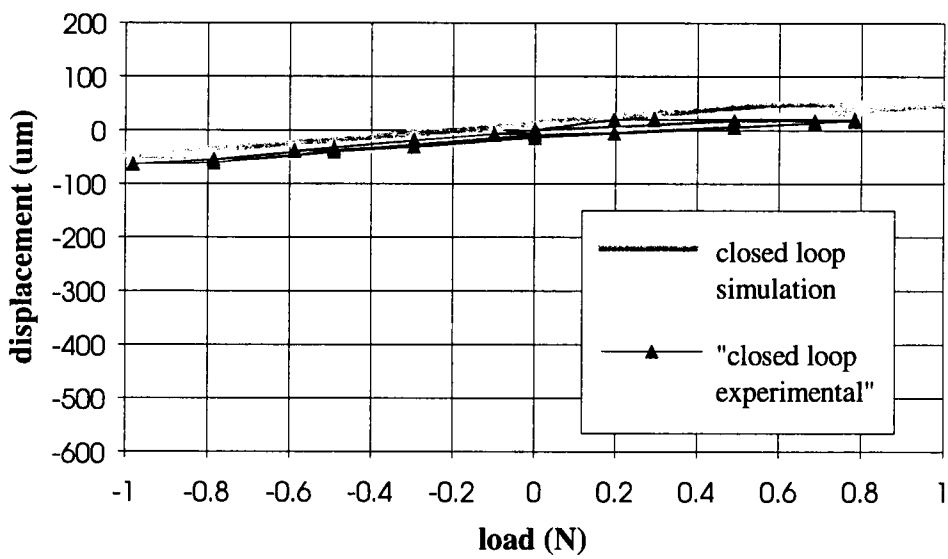
The testing to this point did not include the pole-zero cancellation controller initially designed. The pole-zero cancellation was simulated to determine if including it would improve the instrument performance. The following results were determined by simulation. The addition of the pole-controller reduced compliance to $16 \mu\text{m/N}$. This is because the pole-zero cancellation allows raising the signal amplifier gain. The nominal calibration factor with pole-zero cancellation is 1.15 N/A . Actuator current as a function of load, however, still showed considerable hysteresis. Because pole-zero cancellation did not improve accuracy, the circuitry to test it was not built and tested experimentally.

Additional Simulations

A question remaining at this point is what would the system performance have been had the mechanical plant not had hysteresis? This was determined by simulation. An ideal mechanical plant and real electronics were modeled. The transfer function determined experimentally was used to model the mechanical plant. The hysteresis block previously included in the mechanical



a) Experimental and simulation results of load cell compliance without controller.



b) Experimental and simulation results of load cell compliance with controller.

Figure 48. Displacement of load attachment point as a function of applied load.

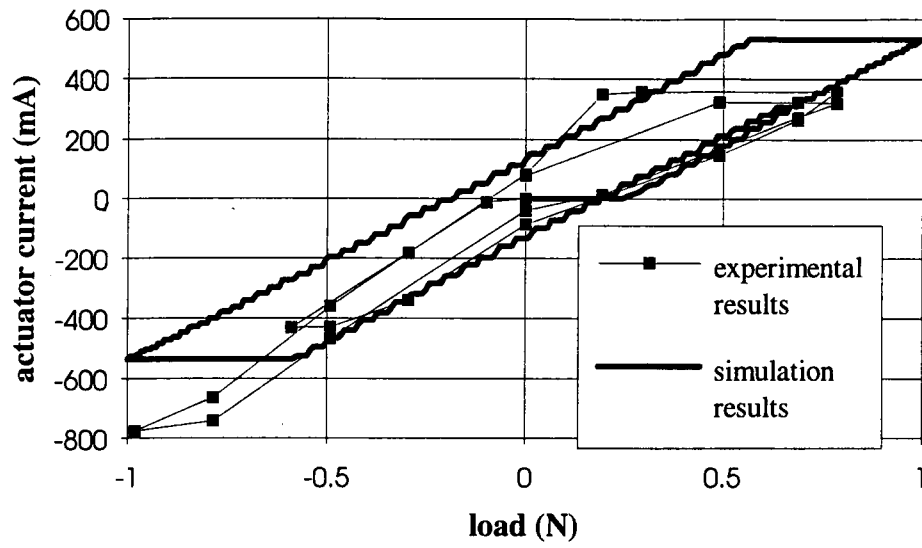


Figure 49. Actuator current as a function of load, experimental and simulation results.

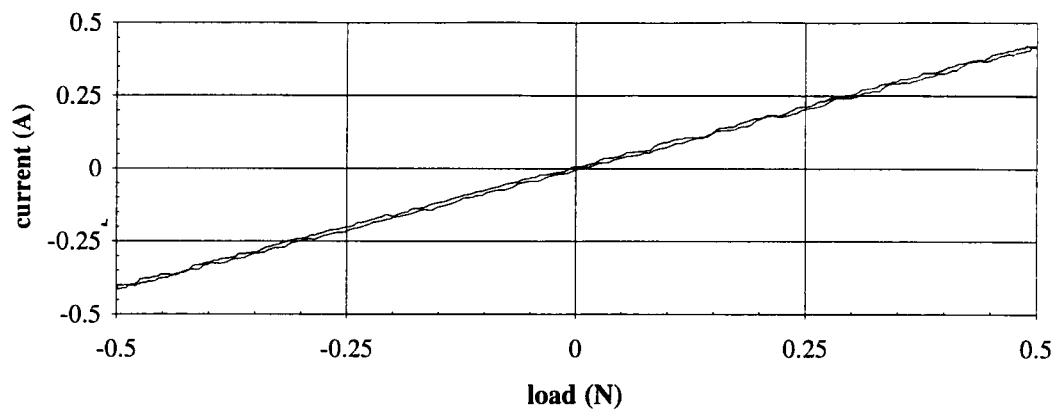
plant model was removed. The LVDT model included hysteresis and noise previously determined experimentally. The power amplifier included hysteresis, noise, and frequency response previously determined experimentally. The actuator nonlinearity as a function of current is included. The actuator nonlinearity as a function of displacement, however, was not included because the deflections are small. The controller used in the simulation included gain and pole-zero cancellation. Deflection as a function of load for the simulated plant is shown in Figure 50. The nominal compliance is 8 $\mu\text{m/N}$. Current as a function of load is also shown in Figure 50. The nominal calibration factor is 1.25 A/N. A ramp load was then applied to idealized load cell. The ramp response is shown in Figure 51. The measurement error shown in figure 51 is given by

$$\text{measurement error} = \text{actual load} - K_f \times \text{current} \quad (3.2)$$

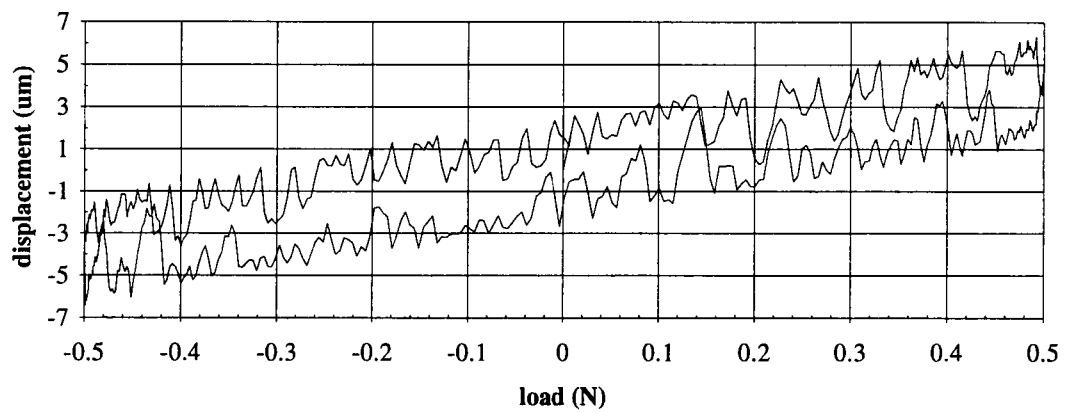
Compare this plots to the results obtained for a noiseless system without nonlinearities presented in the control design section.

As was noted earlier in the control system design section, if actuator current is assumed to be proportional load the measurement error will be systematically related to displacement. Some brief consideration reveals the source of the error. For equilibrium of the load attachment position, the applied load is counteracted by the actuator force, which is a function of current, and the spring force, which is proportional to the deflection of the support flexures. This reveals two avenues for reducing the measurement error. The first is to stiffen the system by making modifications to the controller. The other is to incorporate an error correction scheme based on the measured current and displacement to compensate for displacement and actuator nonlinearities.

The system was stiffened by adding a lag-lead compensator in series with the gain and pole-zero cancellation already in the controller. Over a limited bandwidth the lag-lead compensator acts as an integrator. An integral controller tries to force the integral of the error to zero,



a) current as a function of applied load



b) displacement as a function of applied load

Figure 50. Simulation results for a load cell without mechanical non-linearities.

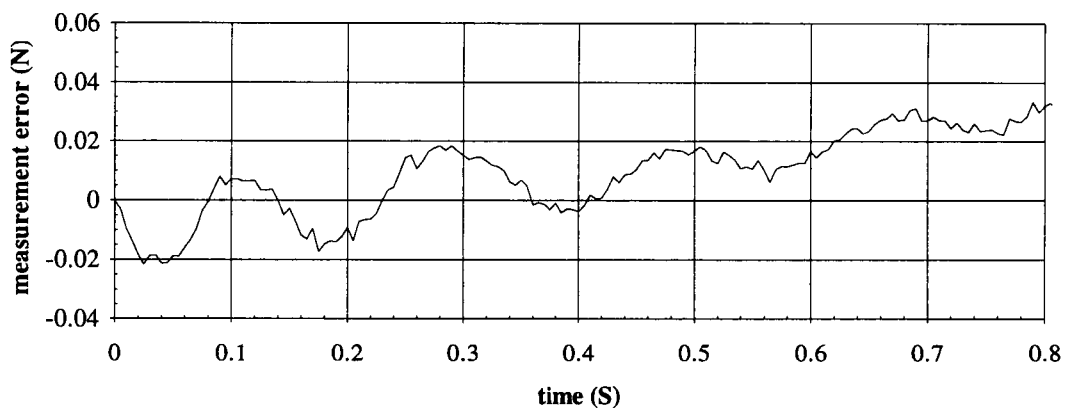
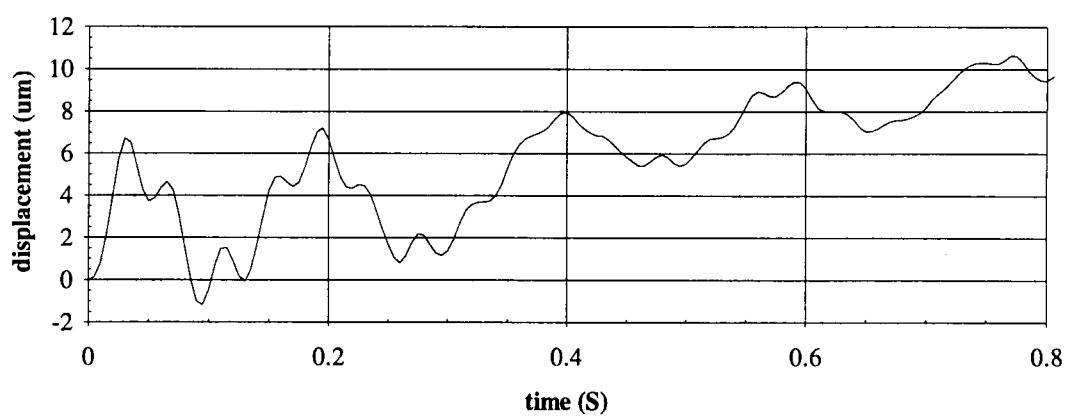
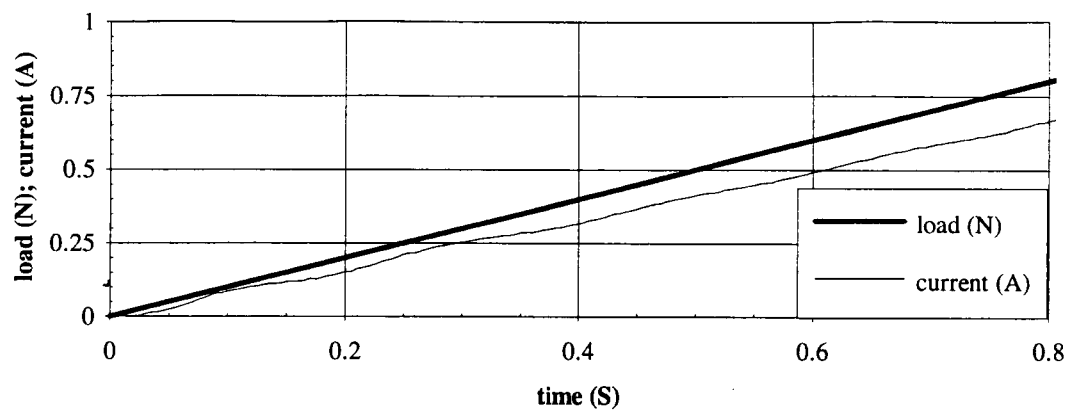


Figure 51. Simulation results of a load cell without mechanical non-linearities to a ramp load.

minimizing static errors. A true integrator was not used because integrators can have difficulties with saturation. The controller used in this simulation is

$$\text{gain} \dots \text{lag} - \text{lead} \dots \text{pole} - \text{zero cancellation} \\ 10 \times \frac{s+12}{s+0.012} \times \frac{s^2+10s+10000}{s^2+2010s+10000} \quad (3.3)$$

A correction scheme was applied to compensate for errors from any remaining displacement and for the non-linearity in the actuator. The equation for measured load is now

$$\begin{aligned} \text{measured load (N)} = \\ \text{current (A)} \times \text{calibration factor (N/A)} + \text{displacement (M)} \\ \times \text{correction term (N/M)} + \text{current}^2 \times \text{correction} \\ \text{term (N/A}^2\text{)} \end{aligned} \quad (3.4)$$

The resulting system displacement and measurement error in response to a unit ramp load are shown in Figure 52. It can be seen that the system stiffness has been improved. The systematic change in displacement as a function of applied load has largely been removed. The compliance is now approximately 1.25 $\mu\text{m/N}$. There is some "ringing" with an amplitude of about 5 μm . This could be removed by some additional tuning of the control system or by adding mechanical damping. Figure 51 shows the improvement in measurement error by adding correction terms. Even with the improved controller there is still a substantial measurement error if the error correction is not applied. Application of the correction terms largely removes any systematic error.

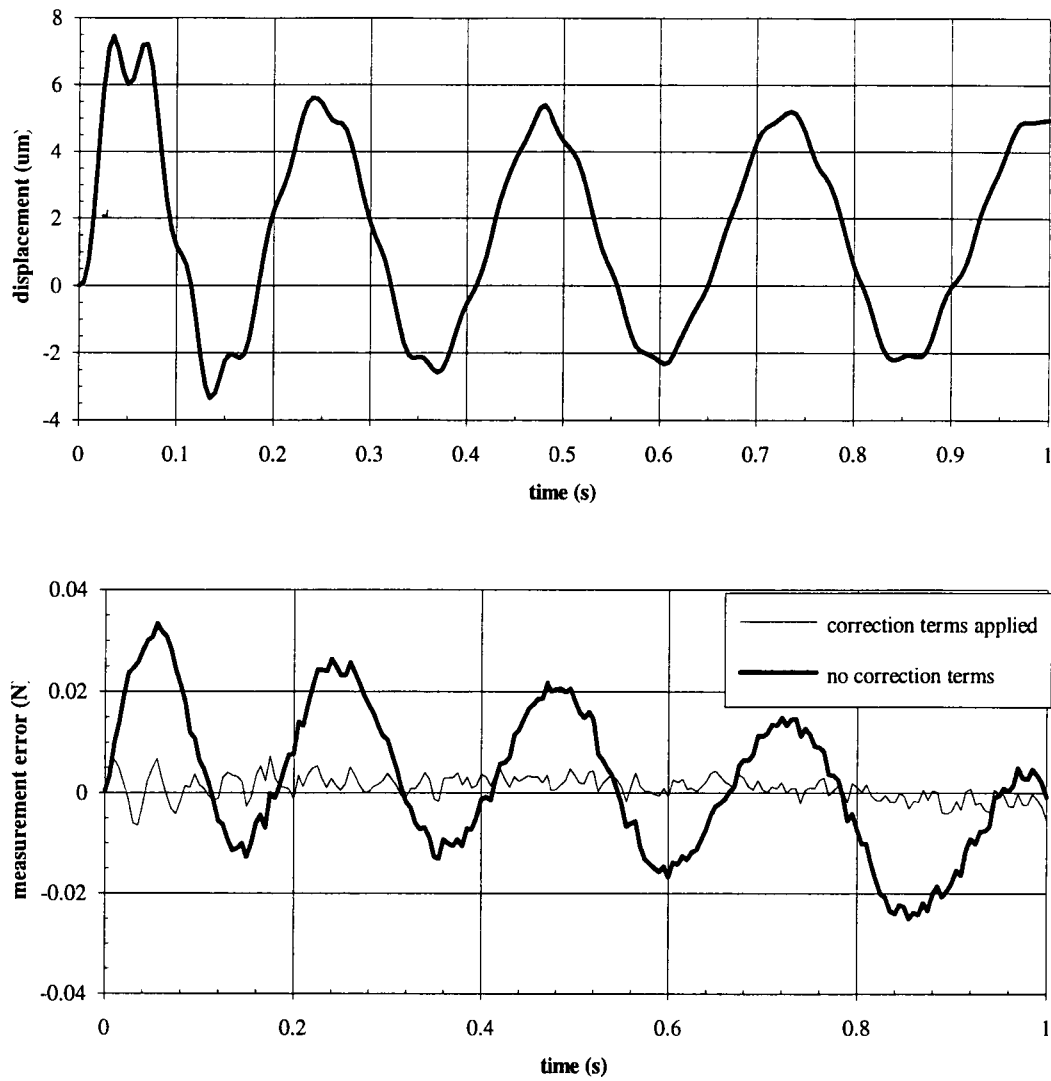


Figure 52. Simulated response of load cell without mechanical non-linearities with improved controller to a unit ramp load. The controller consists of gain, pole-zero cancellation, and lag-lead terms in series as given in equation 3.3.

SECTION 4: SUMMARY AND FUTURE IMPROVEMENTS

Summary

The goal of this thesis project was to demonstrate a proof of concept zero compliance load cell. The concept was developed, analyzed, and tested. The analysis indicated that significant reductions in compliance could be achieved. The controller designed, gain in series with pole zero cancellation, was calculated to reduce the load cell compliance to $6.2 \mu\text{m/N}$. This is a reduction in compliance by a factor of three over the best load cell listed in Table 1. The addition of a lag-lead compensator could further reduce the compliance to $0.01 \mu\text{m/N}$, several orders of magnitude better than the commercial load cells.

The analyzed load cell was then built and tested. The electronics performed as designed. The mechanical plant, however, had significant hysteresis limiting its performance. Several attempts were made to reduce the mechanical hysteresis. The diaphragm flexures were annealed, and slotted with limited improvement. The load cell tested did, however, successfully demonstrate the concept. It exhibited reduced compliance due to the control system action. The compliance was reduced from $333 \mu\text{m/N}$ to $38 \mu\text{m/N}$ by the control system tested, a simple gain stage. The compliance of $38 \mu\text{m/N}$ is of the same order as that of commercial load cells. It was determined by simulation that the mechanical hysteresis limited the load cell accuracy achievable with a more sophisticated controller, gain followed by pole-zero cancellation. Therefore that controller was not built and tested.

Because of the limitations of the mechanical plant simulation studies were conducted to determine the potential compliance improvements of the load cell without any mechanical hysteresis. The simulations included the frequency response, noise, and nonlinearities of the electronics. The mechanical plant in the simulation was the experimentally derived transfer function given in equation 3.1. The gain with pole-zero cancellation controller yielded a

compliance of 8 $\mu\text{m/N}$ in the simulation, comparable to the analytically obtained compliance of 6.2 $\mu\text{m/N}$. Adding a lag-lead compensator to the controller improved the compliance to 1.25 $\mu\text{m/N}$, two orders of magnitude worse than the analytically calculated 0.01 $\mu\text{m/N}$, indicating that the ultimate performance is limited by the electronic noise and nonlinearities. A problem of the load cell is the introduction of systematic errors if load is assumed to be simply proportional to current. A method of compensating for systematic errors caused by displacement and actuator nonlinearities was simulated. The simulations indicated that a measurement error of 0.005 N is achievable with the electronics designed in the course of this thesis if the mechanical nonlinearities can be eliminated.

Future Improvements

Several improvements suggest themselves for future development of a zero compliance load cell. The first is to raise the controller gain and flexure stiffness. Consider the simplified block diagram in Figure 53. If the limiting case of $t \rightarrow \infty$ is considered the position for a unit step is

$$\lim_{t \rightarrow \infty} x(t) = \frac{1}{k + G} \quad (4.1)$$

as the stiffness and loop gain increase the load cell stiffness increases. Based on the experience gained in the development of the load cell previously described, monolithic machined flexures are suggested. Although they have lower initial fabrication cost, diaphragm flexures fabricated from sheet metal present a number of practical problems. Monolithic flexures have the advantage of being inherently stiffer.

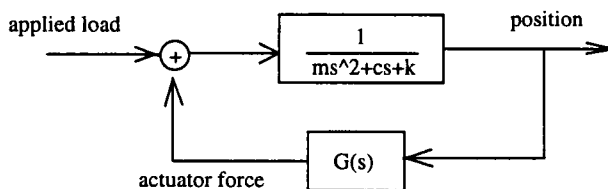


Figure 53. Simplified load cell block diagram.

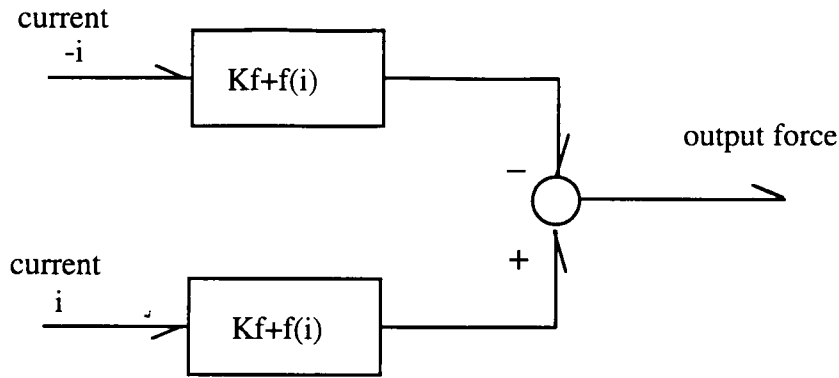


Figure 54. Block diagram of actuators mounted in a push-pull configuration.

To limit the effect of actuator hysteresis, two matched actuators should be used in a push-pull configuration as shown in block diagram representation in figure 54. The output force of each actuator is equal to K_f times the actuator current plus some non-linear term represented by $f(i)$, a function of current. The resultant output force is given by

$$\text{Force} = 2K_f \times I - f(i) + f(i) = 2K_f \times I \quad (4.2)$$

The non-linear terms have canceled.

To prevent damage to the load cell, overload limit stops should be incorporated. A schematic representation of the physical layout is shown in figure 55. The load cell designed in the thesis used a moving magnet actuator to improve heat transfer. Moving magnet actuators have higher hysteresis than moving coil actuators. A change to moving coil actuators is suggested. To improve heat transfer in the actuator air gap the use of a ferrofluid is suggested. They are manufactured by Ferrofluidics Corporation of Nashua, New Hampshire. Their typical application is to improve heat transfer in loudspeakers. Additionally, heat transfer surfaces should be explicitly designed into the load cell to avoid over-heating.

Several changes can be incorporated into the electronics. Changing the precision op-amp in the power supply to a chopper type op-amp, and enclosing the op-amp in a temperature controlled "oven" would reduce the instrument drift. The electronics built in the thesis did not incorporate any noise shielding features which would improve the accuracy of the electronics.

Finally the most significant change in the electronics would be to switch to a micro-processor based controller as opposed to analog. This would allow automated calibration prior to use.

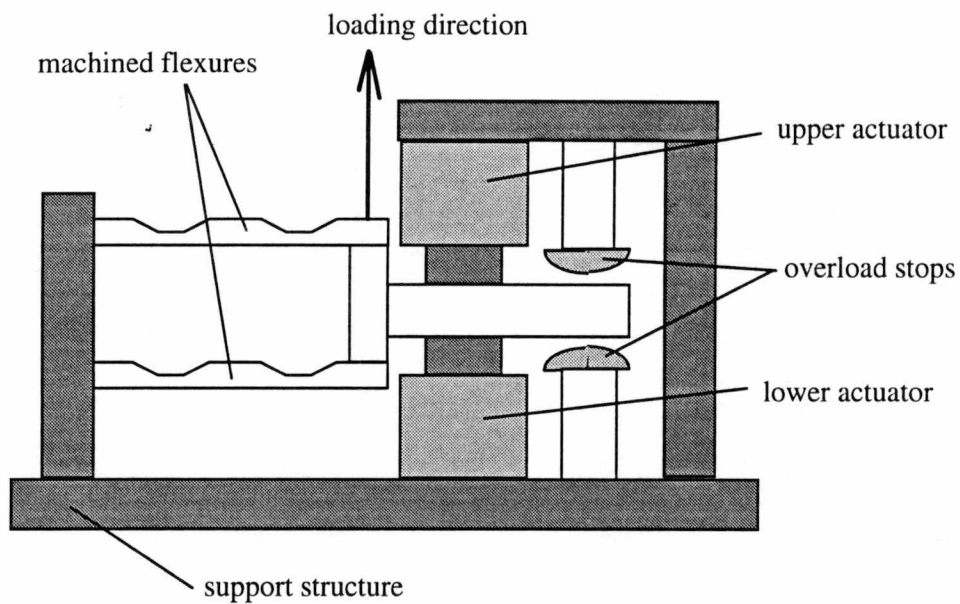


Figure 53. Schematic of an improved reduced compliance load cell.

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