An electrodynamic vibration absorber

by

Thomas Toby Hansen

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To

C. Harvey Hansen

&

Don O. Blackham

Great men who shall never be forgotten.
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<tr>
<td>(a)</td>
<td>material property of conducting wire, (\frac{\text{turns}}{\text{layer m}})</td>
</tr>
<tr>
<td>(A)</td>
<td>cross-sectional area of a pole piece</td>
</tr>
<tr>
<td>(A_m)</td>
<td>magnet surface area</td>
</tr>
<tr>
<td>(A_{xc})</td>
<td>cross-sectional area of the EVA core</td>
</tr>
<tr>
<td>(A_{xs})</td>
<td>cross-sectional area of the EVA shell</td>
</tr>
<tr>
<td>(B)</td>
<td>magnetic flux density that impinges on a conducting wire (G)</td>
</tr>
<tr>
<td>(B_d)</td>
<td>demagnetized operating flux density of magnet surface closest to the air gap (G)</td>
</tr>
<tr>
<td>(B_c)</td>
<td>flux density at the surface of the EVA core (G)</td>
</tr>
<tr>
<td>(B_g)</td>
<td>air gap flux density (G)</td>
</tr>
<tr>
<td>(B_n)</td>
<td>flux density of the (n)th layer of conducting wire (G)</td>
</tr>
<tr>
<td>(B_o)</td>
<td>flux density of the outermost radial point of the magnet used in Model 1 (G)</td>
</tr>
<tr>
<td>(B_r)</td>
<td>remanence flux density (G)</td>
</tr>
<tr>
<td>(B_s)</td>
<td>saturation flux density (G)</td>
</tr>
<tr>
<td>(BL_1)</td>
<td>product of the magnetic flux and wire length of Model 1 (G m)</td>
</tr>
<tr>
<td>(BL_2)</td>
<td>product of the magnetic flux in the air gap interacting with the wire length of Model 2 (G m)</td>
</tr>
<tr>
<td>(BL_{2,g})</td>
<td>BL contribution by the air gap in Model 2 (G m)</td>
</tr>
<tr>
<td>(BL_{2,f})</td>
<td>BL contribution by the fringing flux density in Model 2 (G m)</td>
</tr>
<tr>
<td>(B_{d1})</td>
<td>operating flux density of Model 1 assuming infinite pole piece permeance (G)</td>
</tr>
<tr>
<td>(B_{d2})</td>
<td>operating flux density of Model 2 assuming infinite pole piece permeance (G)</td>
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<tr>
<td>(B(r))</td>
<td>measured flux along the core (G)</td>
</tr>
<tr>
<td>(B(r_s))</td>
<td>measured flux along the magnet restraining shell (G)</td>
</tr>
<tr>
<td>(B_d(r))</td>
<td>corresponding operating flux density point derived from (B(r)) (G)</td>
</tr>
<tr>
<td>(B_d(r_s))</td>
<td>corresponding operating flux density point derived from (B(r_s)) (G)</td>
</tr>
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BH_{max}  maximum energy product (GOe)
cgs centimeter-gram-second base units
C damping coefficient ($\frac{N m}{s}$)
C_m magnetic damping coefficient
C_p capacitance (Farads)
C/M damping to mass ratio
d absorber conducting wire diameter
d_s mean diameter of a solenoid of conducting wire
dB decibel
DFPR deviation from published results
E induced voltage potential (V)
E_d energy dissipated by the EVA (J)
E_i initial vibrating system energy (J)
Eq equation
E(t) sinusoidal voltage source as a function of time (V)
EVA Electrodynamic Vibration Absorber
F_d damping force of a viscously damped system (N)
F_m damping force provided by the EVA (N)
FFT fast fourier transform
g resistance per meter of conducting wire ($\frac{\Omega}{m}$)
G Gauss, cgs unit for flux density
h_b EVA bottom plate thickness
h_d height of the n^{th} layer of conducting wire used in the EVA
h_t EVA top plate thickness
h_w height of multiple conducting wire turns used in the EVA
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<th>Symbol</th>
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<tr>
<td>$H_c$</td>
<td>coercive magnetizing force (Oe)</td>
</tr>
<tr>
<td>$H_d$</td>
<td>demagnetized operating magnetizing force of the magnet surface closest to the air gap (Oe)</td>
</tr>
<tr>
<td>$H_p$</td>
<td>magnetizing force required to induce a given flux density in a pole piece (Oe)</td>
</tr>
<tr>
<td>$Hz$</td>
<td>Hertz, unit of frequency ($\frac{1}{s}$)</td>
</tr>
<tr>
<td>$H_{ci}$</td>
<td>intrinsic coercive force (Oe)</td>
</tr>
<tr>
<td>$i(t)$</td>
<td>induced current of the EVA as a function of time (Ampere)</td>
</tr>
<tr>
<td>$J$</td>
<td>Joule, si unit of energy ($\frac{Nm}{s}$)</td>
</tr>
<tr>
<td>$K$</td>
<td>equivalent spring stiffness ($\frac{N}{m}$)</td>
</tr>
<tr>
<td>$K'$</td>
<td>dummy variable</td>
</tr>
<tr>
<td>$K_{2}$</td>
<td>inductance equation correction factor for solenoids of finite length</td>
</tr>
<tr>
<td>$K_{f1}$</td>
<td>nonlinear magnetizing force correction factor for Model 1</td>
</tr>
<tr>
<td>$K_{f2}$</td>
<td>nonlinear magnetizing force correction factor for Model 2</td>
</tr>
<tr>
<td>$l$</td>
<td>length a flux line has to travel through a pole piece</td>
</tr>
<tr>
<td>$L$</td>
<td>length of wire that is influenced by the magnetic flux impinging upon it (m)</td>
</tr>
<tr>
<td>$L_c$</td>
<td>EVA core length</td>
</tr>
<tr>
<td>$L_g$</td>
<td>air gap length</td>
</tr>
<tr>
<td>$L_i$</td>
<td>EVA inductance value (H)</td>
</tr>
<tr>
<td>$L_m$</td>
<td>magnet thickness between poles</td>
</tr>
<tr>
<td>$L_s$</td>
<td>EVA shell length</td>
</tr>
<tr>
<td>$L_w$</td>
<td>wire length</td>
</tr>
<tr>
<td>$L_y$</td>
<td>total number of layer of wires of conducting wire used</td>
</tr>
<tr>
<td>$L_m'$</td>
<td>corrected magnet thickness to obtain the operating flux density by assuming infinite pole piece permeance</td>
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<td>$L_{xc}$</td>
<td>length of the air gap between core, top and bottom pole pieces</td>
</tr>
<tr>
<td>$L_{xs}$</td>
<td>length of the air gap between shell, top and bottom pole pieces</td>
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m  meter, unit of length
mm  millimeter, one thousandth of a meter
mmf  magnetomotive force
M  mass
MM  moving mass
MW  moving wire
n  number of wire layers
N  Newton, unit of force \( \left( \frac{Kg \cdot m}{s^2} \right) \)
Nw  number of turns of wire on a solenoid
Oe  Oersted, cgs unit of magnetizing force
P  power dissipated by the EVA
F  fringing flux permeance in Model 2
Pg  cylindrical air gap permeance
Pl  magnet leakage flux permeance in Model 2
Pp  pole piece permeance
Pr  cylindrical return path air gap permeance in Model 2
Pt  total permeance of the magnetic circuit
Pav  average power dissipated by an RLC circuit (watts)
Peq  equivalent permeances of either a parallel or series permeance arrangement
Pgf  air gap permeance of the flux tester
Pg1  air gap permeance of Model 1
Pg2  air gap permeance of Model 2
Pgt  total permeance across an air gap including fringing and leakage
Ptf  total permeance of the flux tester's magnetic circuit
Pt1\infty  total permeance of Model 1 assuming infinite pole piece permeance
\[ P_{t2} \quad \text{total permeance of Model 2 assuming infinite pole piece permeance} \]
\[ P_{1,2} \quad \text{absorber shell permeance} \]
\[ P_{3,4} \quad \text{permeance between the shell and the top and bottom plates} \]
\[ P_5 \quad \text{top plate permeance} \]
\[ P_6 \quad \text{bottom plate permeance} \]
\[ P_{7,8} \quad \text{permeance between the core and the top and bottom plates} \]
\[ P_{9,10} \quad \text{core permeance} \]
\[ Q \quad \text{slope of the linear part of the load line} \]
\[ r \quad \text{core radius} \]
\[ r_n \quad \text{radius of the n\textsuperscript{th} layer of conducting wire} \]
\[ r_s \quad \text{inside radius of the magnet restraining shell} \]
\[ R \quad \text{total EVA resistance} \]
\[ R_l \quad \text{magnetic reluctance} \]
\[ R_s \quad \text{shunt resistor} \]
\[ R_w \quad \text{EVA conducting wire resistance} \]
\[ RLC \quad \text{electrical components of resistor-inductor-capacitor in series} \]
\[ RMS \quad \text{root mean square} \]
\[ s \quad \text{ferromagnetic thickness of the magnetic mass of Model 2} \]
\[ S \quad \text{axial length of a solenoid} \]
\[ s_{id} \quad \text{absorber shell inside diameter} \]
\[ s_{od} \quad \text{absorber shell outside diameter} \]
\[ t \quad \text{time (second)} \]
\[ T \quad \text{Tesla, si unit of flux density} \]
\[ V \quad \text{unit of voltage potential - volts} \]
\[ V_s \quad \text{steady-state velocity of a vibrating system} \]
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<tr>
<td>V(t)</td>
<td>vibrating mass velocity as a function of time</td>
</tr>
<tr>
<td>VEM</td>
<td>visco-elastic materials</td>
</tr>
<tr>
<td>Vrms</td>
<td>root mean square voltage</td>
</tr>
<tr>
<td>W</td>
<td>length of the magnet in Model 2</td>
</tr>
<tr>
<td>wall</td>
<td>non-ferromagnetic wire restraining wall</td>
</tr>
<tr>
<td>x(t)</td>
<td>vibrating mass position as a function of time</td>
</tr>
<tr>
<td>X</td>
<td>distance between conducting wire centers in connecting layers</td>
</tr>
<tr>
<td>X₀</td>
<td>initial zero to peak displacement of a vibrating system</td>
</tr>
<tr>
<td>Xₛ</td>
<td>steady-state maximum displacement</td>
</tr>
<tr>
<td>X(t)</td>
<td>vibrating mass maximum zero to peak displacement as a function of time</td>
</tr>
<tr>
<td></td>
<td>magnitude of the electrical impedance (Ω)</td>
</tr>
<tr>
<td>δ</td>
<td>ferromagnetic thin shell thickness</td>
</tr>
<tr>
<td>δₘ</td>
<td>distance between the core and magnetic mass in Model 2</td>
</tr>
<tr>
<td>µ</td>
<td>permeability defined as the ratio of flux density to magnetizing force</td>
</tr>
<tr>
<td>µₛ</td>
<td>permeability of the bottom plate</td>
</tr>
<tr>
<td>µₙ</td>
<td>permeability of the core</td>
</tr>
<tr>
<td>µₜ</td>
<td>coefficient of kinetic friction</td>
</tr>
<tr>
<td>µ₀</td>
<td>permeability of air (value = 1.0 in cgs units or $4\pi \times 10^{-7}$ (H/m) si units)</td>
</tr>
<tr>
<td>µₜ</td>
<td>permeability of the top plate</td>
</tr>
<tr>
<td>ω</td>
<td>angular frequency (radian/s)</td>
</tr>
<tr>
<td>ωₙ</td>
<td>natural angular frequency of a spring-mass system</td>
</tr>
<tr>
<td>ω₀</td>
<td>resonant angular frequency of an RLC circuit</td>
</tr>
<tr>
<td>Δω</td>
<td>angular frequency bandwidth of one half peak dissipative average power</td>
</tr>
<tr>
<td>Ω</td>
<td>Ohm, unit of resistance</td>
</tr>
<tr>
<td>Φ</td>
<td>flux lines, (Maxwells)</td>
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$\zeta$ damping ratio
INTRODUCTION

Passive vibration damping is widely used today as a means of reducing or eliminating vibration in mechanical systems. Most techniques use elastic elements (springs) in parallel or in series with different types of damping materials to eliminate unwanted vibration. Almost all of the damping elements found in vibration absorbers are made of either air, oil, or viscoelastic materials (VEM). These type of damping methods are widely known to be a very effective means of not only eliminating vibrations but extending the life of mechanical components. Some examples of the use of passive vibration damping are air and hydraulic shocks found in automobiles and the use of VEM to prevent rocket nozzle vibration.

There are limitations to the damping techniques currently available. One limitation is that temperature variations in certain damping materials lack stable damping characteristics due to changes in their material properties [1]. This limitation can be very serious when large temperature variations exist in the environment where these damping materials are to be used. Another limitation of traditional damping techniques is that damping characteristics can be dependent on the natural frequency of the absorber itself [1,2]. In general, maximum damping performance is achieved when an external excitation frequency is the same as the natural frequency of the absorbing system. Damping performance deteriorates markedly as the excitation moves away from the absorbing system's natural frequency [1]. A third limitation arises from an engineering design perspective. Passive damping, using the aforementioned materials, is difficult to model mathematically. Air and oil damped system models depend on fluid viscosity information which is very sensitive to temperature. Additionally, VEM properties are nonlinear and its damping characteristics are not easily modeled [3]. For most vibration applications, damping materials operate in stable thermal environments and the design requirements don't require an exact computational model of the system. These limitations, however, must be overcome when damping is applied in space.

The need for damping is important in space applications. Mechanical vibrations are undesirable in many NASA space systems such as telescope reflectors, communication antennas, and the proposed Space Station Freedom [4]. It has been reported (from modal analysis) that the Space Station Freedom will have 366 vibration modes below 1 Hz and 3000 to 4000 vibration modes below 20 Hz [5]. These modes could easily be excited by a rocket thruster, docking of a space vehicle, or by a meteor strike; thereby creating vibrations that would damage valuable equipment on board [6]. Obviously, some type of damping will be needed to reduce or isolate these potentially problematic vibrations. Traditional vibration
isolation techniques and material-damped vibration absorber devices are the primary approach taken to reduce or isolate space station vibrations. Both options, however, may not be cost effective and may have limited success because of the adverse effect varying environments have on damping elements and materials [1].

Current material damping technology has been applied in the form of a strut to the Hubble Space Telescope [7]. The device uses a fluid based damping material housed in a cylinder. Its success has not been published but there has been a report that the Hubble has had vibration problems caused from thermally induced oscillations. This implies that the strut isn't completely effective in vibration reduction [8]. Damping struts and environmentally controlled housings for VEM are on the forefront of damping technologies being applied in space today. One must be reminded, however, that both of these techniques have their shortcomings. Space presents a harsh environment where the full temperature variation for a spacecraft can be -50°C to 100°C. To get stable damping characteristics over these temperature ranges, environmentally controlled compartments or heating elements will have to be used; this will have the undesirable effects of adding weight and cost associated with requiring more design effort to create the appropriate environment. A second shortcoming is that weight must be added to control a vibrating structure. A study of a 12 meter truss conducted at Wright Patterson Air Force Base showed that a configuration with VEM damping materials weighed 50% more than the undamp ed truss [3]. These shortcomings can be reduced by designing a vibration absorbing device that (1) has more stable damping characteristics, and (2) has a higher damping per unit weight ratio than traditional damped vibration absorbers.

Electric eddy current damping (induced currents by a moving conductor relative to a stationary magnetic flux) has been under limited investigation over the past 30 years. It has the potential of having stable damping characteristics in harsh environments and a high damping to weight ratio. It has been reported that eddy current damping has several advantages such as reliable behavior and high thermal stability [9]. Other studies cite additional eddy current damping benefits such as being non-outgassing, radiation tolerant, and being simply modeled by a linear dashpot dissipative force that is proportional to the velocity [9,10]. The results from such studies are encouraging. Some researchers report reducing vibration amplitudes as much as 90 to 97% [1,11]. Eddy current damping has had success in at least one application in space as a satellite nutation damper where it was successful in stabilizing the spin axis of the SAS-A satellite launched in December 1970 [12]. Eddy current damping is reported to have a good maximum damping to mass ratio (C/M) of 500 s⁻¹ [3]. This compares favorably with the current space strut designs, which have damping to mass ratios of 332 and 614 s⁻¹ [13]. Most
of the work on eddy current damping has focused on using eddy currents to restrain vibration movement with little focus on using the produced current for external thermal dissipation, damping control, or storage.

The potential exists to develop a device that can capitalize on all of the eddy current damping benefits but also can have the ability to export the energy created from the vibration at hand (to be either dissipated or stored elsewhere) and have the ability of providing adjustable damping. The concept is as simple as that for eddy currents - have linear motion of a conductor passing through a magnetic field. The type of devices that use this methodology are formally called electrodynamic devices. Some common devices in this class are loudspeakers, noise generators and shakers. These devices require input of electrical energy to produce vibrations. An electrodynamic vibration absorber, on the other hand, uses the reverse principle by having vibrations produce electrical energy. The electrodynamic absorber has an advantage over an eddy current absorber because most of the energy created can be transported outside the device. This is important for the damping of steady state vibrations. A steady state vibration is a vibration that is continuous for a long period of time even after damping is applied. Transporting energy outside the device is important because energy can build up as heat and destroy the device and its ability to absorb vibrations (as would happen to an eddy current damper subject to steady state vibrations). An electrodynamic vibration absorber has additional advantages over eddy current dampers; the amount of damping they provide can be varied and the energy removed can be stored in battery cells rather than being lost as waste heat like an eddy current device.

In this study the electrodynamic vibration absorber idea is pursued. The main objectives are to characterize design parameters and to determine the potential of electrodynamic damping. This has been done by first analytically modelling the design parameters, building the device and then testing to confirm the analytical modelling.
OBJECTIVES

1. Determine the potential of electrodynamic damping.

2. Characterize design parameters for electric dissipation.
TECHNICAL BACKGROUND

There are many aspects to the mathematical modelling that need to be addressed. The discussion to follow is broken into four parts. The first section looks at modelling the mechanics of electrodynamic damping. This includes derivations for the amount of energy dissipated by the system, damping coefficient and the damping ratio. The second section addresses the electrical equations used in this project. This section includes equations for wire selection, inductance, and circuit resonance. The third section is the magnetic circuit modelling of the absorber. The modelling of this section includes two different methodologies of magnetic circuit theory and special correction factors for the magnets used in this project. The last section is material selection. This section describes why certain materials were chosen or excluded from this project.

Mechanics Derivations

There are two kinds of situations that need to be addressed for a complete discussion on the vibration behavior of the electrodynamic vibration absorber (EVA). The first is the transient response and the second is the steady state response of the absorber. The transient response is modeled for an impulsive excitation of a mechanical system to which the absorber is attached. The steady state response modeling is addressed to determine how much a vibration amplitude is attenuated by the vibration absorber.

The transient response of a vibrating mechanical system can be visualized first with the undamped spring mass system portrayed in Figure 1.

![Figure 1. An undamped mechanical system responding to an impulse](image-url)
The zero to peak relative displacement of the mass $M$ is maximum at time $t = 0$. The response of the mass to the impulse at a later time $t$ is modeled by Eq. 1 as

$$x(t) = X(t)\cos(\omega t)$$

where:

- $\omega = \text{angular frequency} = 2\pi f \ (s^{-1})$
- $f = \text{frequency of the vibration (Hz), and}$
- $X(t) = \text{maximum displacement at time } t$.

The velocity $V(t)$ of the mass for some time $t$ is given by Eq. 2 as

$$V(t) = X(t)\omega \sin(\omega t) + \dot{X}(t)\cos(\omega t)$$

In order to reduce the displacement amplitude of the vibrating mass, damping must be provided to the system. If no damping is applied, the system will never come to rest. Suppose that an electrodynamic vibration absorber is placed in parallel with the spring. This absorber has several turns of conducting wire of total length $L$ wrapped around a cylinder that extends in a plane perpendicular to a radially impinging flux $B$ as shown in Figure 2.

![Electrodynamic Vibration Absorber](image)

Figure 2. Electrodynamic Vibration Absorber

Because the velocity $V(t)$ of the wire is perpendicular to the magnetic flux $B$ there is a force produced on the charged particles (electrons) in the conductor that creates a voltage potential $E$. This relationship is expressed by Eq. 3 as

$$E = BLV(t)$$

Using electric circuit analysis, the current generated in the wire by this motion is,

$$i(t) = \frac{BLV(t)}{Z}$$
where: \[ Z = \text{The magnitude of the real and imaginary impedance of the circuit given by} \]

\[ |Z| = \sqrt{R^2 + (\omega L_i)^2} \]  \hspace{1cm} (5)

where:
\[ R = \text{resistance of the circuit, and} \]
\[ L_i = \text{inductance of the circuit.} \]

The power \( P \) dissipated through the resistance \( R \) is then given by Eq. 6 as

\[ P = [i(t)]^2 R \]  \hspace{1cm} (6)

The total energy dissipated \( E_d \) from time zero to \( T \) is the integral of the power \( P \):

\[ E_d = \int_0^T [i(t)]^2 R \, dt \]  \hspace{1cm} (7)

Substituting Eqs. 2 and 4 into Eqs. 7 yields

\[ E_d = \int_0^T (B L_i X(t) \omega \sin(\omega t) + \frac{[X(t) \cos(\omega t)]^2 R}{|Z|^2}) \, dt \]  \hspace{1cm} (8)

It can be shown that the second term, \( X(t) \cos(\omega t) \), in Eq. 8 can be neglected for small damping (\( \zeta < .20 \)). The resulting analytical solution of the integrand is

\[ E_d = \frac{1}{2} K(X_0)^2 \{ 1 - \exp\left( -\frac{2(B L_i \omega)^2 R}{|Z|^2} \left[ \frac{t}{2} - \frac{\sin(2\omega t)}{4\omega} \right] \right) \} \]  \hspace{1cm} (9)

where \( X_0 \) is the initial zero-to-peak amplitude of the vibration absorber displacement [14]. The assumptions used in developing Eq. 9 are that (1) damping present in the mechanical system is negligible compared with electrodynamic damping, (2) the spring is linear and, (3) the system isn't more than 20% damped. The maximum amplitude as a function of time can be derived from Eq. 9 using energy methods. The expression for the energy in the system equaling the energy in the spring with a maximum amplitude at time \( t \) is given by Eq. 10 as

\[ \frac{1}{2} K(X(t))^2 = E_i - E_d \]  \hspace{1cm} (10)

where:
\[ E_i = \frac{1}{2} K(X_0)^2 = \text{initial energy of the system.} \]

The resulting expression is found by substituting Eq. 9 into Eq. 10,

\[ X(t) = X_0 \exp\left( -\frac{(B L_i \omega)^2 R}{K|Z|^2} \left[ \frac{t}{2} - \frac{\sin(2\omega t)}{4\omega} \right] \right) \]  \hspace{1cm} (11)

The zero to peak maximum displacement amplitude found from numerically integrating Eq. 8 was compared to the value from Eq. 11. These results are shown in Figure 3, where only the real part of the impedance was assumed to be present. As can be seen, the analytical solution
and the numerically integrated solution are equivalent. One may ask, however, why it was said earlier that eddy current damping can be approximated by a dashpot, yet the maximum displacement envelope looks different than those published in literature for a dashpot damper? Included in Figure 3 is the resulting sinusoidal motion of the vibrating mass. In literature the envelope is created by drawing a line through the maximum points of the sinusoid. The resulting maximum displacement envelope for electrodynamic damping can be seen to be of the same shape as would be published in literature. The reason for this 'smooth' curve is due to the sine term in Eq. 11 is zero at half and full cycles. This eliminates the effects of the sinusoidal term from influencing the maximum displacement envelope. Therefore, the statement that electrodynamic damping can be modeled as a dashpot dissipative force still holds.

![Figure 3. Comparison of numerical integration and the analytical solution to Eq. 9](image)

Now that the absorber response to a transient excitation has been modelled, the response of the absorber to a steady state excitation will be modelled. First, the damping coefficient, $C$, has to be derived so that the damping ratio, $\zeta$, can be implemented to find the absorber steady state displacement amplitude $X_s$. First a new expression for velocity has to be found. This is done simply by replacing $X(t)$ in Eq. 1 with $X_s$ for the steady state displacement amplitude.
Taking the derivative, the steady state velocity for the vibrating system is

\[ V_s(t) = X_s \omega \sin(\omega t) \]  

(12)

The damping coefficient \( C \) is typically used in vibration analysis to quantify the damping force provided by a dashpot damper. The product of the damping coefficient and the velocity of the damping moving element equals force provided by a dashpot damper given in Eq. 13.

\[ F_d = CV_s(t) \]  

(13)

The magnetic force produced by the current carrying conductor in Figure 2 is equal to the damping force generated by the dashpot and is expressed as:

\[ F_m = BLi(t) \]  

(14)

Substituting Eq. 12 into Eq. 4 then into Eq. 14 results in the following expression for electrodynamic damping force:

\[ F_m = \frac{(BL)^2 V_s(t)}{|Z|} \]  

(15)

Equating the magnetic damping force to the dashpot damping force we get the magnetic damping coefficient, \( C_m \), where

\[ C_m = \frac{(BL)^2}{|Z|} \]  

(16)

At any particular frequency, the damping coefficient is constant because all of the quantities in Eq. 16 are independent of frequency. It should be noted that the impedance will not be constant if either its real or imaginary parts change with frequency such as the imaginary impedance \( \omega L_1 \) which is directly proportional to the frequency of the excitation. Although it has been determined that the damping coefficient \( C_m \) is not constant over a range of frequencies, it can still be related to the dimensionless damping ratio \( \zeta \), where the damping ratio also varies with frequency. From basic vibration analysis this is given by

\[ \frac{C_m(\omega)}{M} = 2\zeta \omega_n \]  

(17)

where: \( M \) = mass of the object vibrating, and,

\[ \omega_n = \sqrt{\frac{K}{M}} \] = natural circular vibrating frequency of the system (s\(^{-1}\)).

Manipulating and substituting in Eq. 16 into Eq. 17 the final solution is

\[ \zeta = \frac{(BL)^2}{2\omega_n M|Z|} \]  

(18)

The damping ratio, \( \zeta \), provides a convenient way of determining the steady state displacement amplitude of a vibrating system, as shown in any introductory text on vibrating mechanical systems (which will not be shown here).
Electrical Modelling Equations

This section is to show and derive equations that are needed to model and optimize the damping potential of electrodynamic vibration absorption. The EVA equivalent electrical circuit is shown to aid in this discussion. The equivalent resistance-inductance-capacitance (RLC) circuit of the EVA is shown in Figure 4. From Eq. 14 it can be seen that maximum damping force can be provided by the absorber when the current through the absorber coil is maximum. Maximizing the current requires the impedance to be as small as possible, since current is inversely proportional to the circuit impedance. Theoretically, resistance will be a constant function of frequency and can only be reduced by reducing the resistive load on the absorber. Unlike resistance, the impedance contribution from the inductance (reactance) does increase with increasing frequency (refer back to Eq. 5). To reduce the effect of reactance on current flow through the absorber, a capacitor can be placed in series with the resistor and the inductor of its circuit. The governing equation for the maximum current in the EVA circuit is given in Eq. 19 as

\[ i(t) = \frac{E(t)}{\sqrt{R^2 + (\omega L_i - \frac{1}{\omega C_p})^2}} \] (19)

where:
- \( R = R_s + R_w \) = total real resistance of the circuit,
- \( C_p \) = capacitor,
- \( L_i \) = absorber inductance,
- \( R_s \) = shunt 'load' resistance,
- \( R_w \) = absorber wire resistance and,
- \( E(t) \) = sinusoidal voltage.

Because the inductor phase shifts the voltage phase by 90° and a capacitor shifts the voltage phase -90°, they can be combined to achieve a net reactance contribution of zero. Eq. 19 shows that the net imaginary impedance will be zero only when the driving frequency of the voltage source matches a specific combination of inductance and capacitance values in the circuit. The frequency at which this occurs is called electrical circuit resonance and is given by Eq. 20,

\[ \omega_0 = \frac{1}{\sqrt{C_p L_i}} \] (20)

Eqs. 19 and 20 show that the inductance can be reduced to zero by operating at circuit resonance. This is more apparent when the power dissipated by the absorber is considered.
Figure 4. Electrodynamic Vibration Absorber equivalent electrical (RLC) circuit

The average power equation for an RLC circuit is given by,

$$P_{av} = \frac{V_{rms}^2 R \omega^2}{R^2 \omega^2 + \frac{1}{L_i} \left( \omega^2 - \omega_0^2 \right)^2}$$

(21)

where:  
- $\omega$ = angular frequency of the voltage source (s$^{-1}$)  
- $P_{av}$ = average power of the circuit, and  
- $V_{rms}$ = root-mean square of the sinusoidal voltage source.

Average power dissipated by an RLC circuit due to a sinusoidal voltage source, $V_{rms} \sin(\omega t)$, using Eq. 21 with a circuit resonant frequency of 50 Hz is shown in Figure 5. Referring to Figure 5 it can be seen that there is a term $\Delta \omega$ located inside the average power curve. This is the half-power bandwidth and has an approximate value of,

$$\Delta \omega \approx \frac{R}{L_i}$$

(22)

where:  
- $\Delta \omega$ = The circular frequency bandwidth of one half the peak average power at resonance.

The half power bandwidth expression is a way of quantifying the frequency range over which half or more power can be dissipated when a capacitor is introduced into a series resistance-inductance circuit. For effective damping, the EVA circuit has to resonate or be within the half-power bandwidth at a given vibration frequency.

The circuit analysis just undertaken, presents a catch 22 situation. If the vibration absorber has an appreciable amount of inductance, the magnetic damping coefficient $C_m$ will be small. Therefore, the imaginary impedance of the system has to be minimized (refer to Eq.16). This
can only be achieved by placing a capacitor in series with the absorber's circuit, thereby producing circuit resonance (assuming the necessary capacitor can be found). From Eqs. 20 & 22, it can be seen that the bandwidth of the absorber with a small value of resistance $R$ and a large inductance $L_i$ is small. This implies that the deviation of the capacitance $C_p$ cannot deviate much from the true value of capacitance that produces resonance. Small deviation from

![Graph showing average power vs. frequency with two curves for $R = 20$ ohms and $R = 50$ Ohms.]

Figure 5. Average power of a sinusoidal voltage source in an RLC circuit

the specific capacitance value needed to produce circuit resonance can be relaxed if either the inductance is reduced or circuit resistance is increased. A catch 22 situation is introduced from the fact that the resistance $R$ of the absorber has to be as small as possible for maximum current in the circuit and hence, maximum damping. If, for example, $R$ is 5 $\Omega$ and $L_i$ is .20 H then the resulting half-power bandwidth is 4 Hz. The required capacitance for circuit resonance within the half-power bandwidth at 50 Hz is approximately $50 \mu F \pm 8.6 \%$. For this particular example, a capacitor may be on the market to suffice in getting the absorber circuit to resonate within the half-power bandwidth. This is somewhat encouraging but, this is only for one particular frequency and would not help but hinder current flow at other frequencies. It should be pointed out that lower frequencies require large capacitors to produce resonance. This poses a special problem because a capacitance general rule of thumb is the larger the capacitor, the
lower its voltage tolerance. Lower voltage tolerance is significant to the absorber because even after circuit resonance has been achieved, the capacitor may burn up destroying the circuit or may place a special design constraint on the amount of voltage that can be produced by the absorber limiting its damping potential.

The electrical equations thus derived do not yet complete the discussion on modelling the EVA. The theoretical inductance value $L_i$ needs to be found along with identifying the best size of wire to maximize the damping coefficient and the space the wire requires.

First, we refer to Figure 6 to develop geometric relationships for the EVA wire dimension equations. Since an equilateral triangle is formed by the radii contacting three neighboring circles, the vertical distance between the wire centers $X$ is expressed as

$$X = \frac{\sqrt{3}}{2} d$$  \hspace{1cm} (23)

![Diagram](image)

**Figure 6.** Conducting wire height

The subsequent height of the layered wire is,

- Row 1: $h_w = d$
- Row 2: $h_w = d + \frac{d}{2\sqrt{3}}$
- Row 3: $h_w = d + \frac{d}{2\sqrt{3}} + \frac{d}{2\sqrt{3}}$
For \( n \) number of rows the space needed to house the wire is,

\[
h_w(n) = d\left(1 + \frac{\sqrt{3}}{2}(n-1)\right)
\]  

(24)

Derived from Eq. 24, the distance to the center diameter of an \( n \)th layer is,

\[
h_d(n) = \frac{d}{2}(1 + \sqrt{3}(n-1))
\]  

(25)

The equation used to model the inductance of the EVA is given by Eq. 26. Inductance equations for a solenoid has been derived by several authors in the past. Eq. 26 was chosen because it accounts for the inductance of a finite solenoid with a correction factor based on geometry. The inductance equation, taken from reference [15], is given by Eq. 26

\[
L_i = \frac{\mu_r \pi (N_w)^2 A_s}{S + 4.5d_s}
\]  

(26)

where:
- \( d_s \) = diameter of the mean center of the wire (m),
- \( A_s \) = cross-sectional area of the solenoid (m\(^2\)),
- \( N_w \) = number of turns of wire on a solenoid,
- \( S \) = axial length of the solenoid (m), and
- \( \mu_r \) = relative permeability to that of air, which is \( \mu_0 = 4\pi \times 10^{-7} \) (H/m).

Eq. 26 is significant because it shows how solenoid inductance can be either increased or decreased by adjusting solenoid dimensions or the number turns of wire on the solenoid. To significantly reduce the inductance, the number of turns of wire, \( N \), has to be minimized. Other smaller reductions in inductance can be obtained by keeping the mean diameter as small as possible and the axial length \( S \) as long as possible. The inductance equation presents another catch 22 situation for the optimization of the damping coefficient \( C_m \). In order for \( C_m \) to be large, a long length of wire is required (see Eq. 16). A long length of wire, however, requires many (hundreds perhaps thousands) turns which would increase the solenoid inductance dramatically.

The last part of this discussion concentrates on identifying the most efficient size (gage) of wire to use and the impact that wire gage has on the damping coefficient \( C_m \). The schematic in Figure 7 is used to find the total length of conducting wire needed in the vibration absorber for given dimensions.

The total length of wire, \( L_w \), can be found by summing the individual contributions of each length of wire in the \( n \) layers wrapped around a cylinder. This is expressed in Eq. 27 as

\[
L_w = 2\pi a S \sum_{n=1}^{L_y} r_n
\]  

(27)
where: \( r_n \) = radius to the \( n \)th layer of wire,
\( L_y \) = total number of layers of wire,
\( L_w \) = the total length of wire in the EVA, and
\( a \) = dimensional property of the wire, units are turns per layer meter.

The lowest resistance the EVA can have is from the wire's resistance (short circuited). The lowest resistance also constitutes the largest damping coefficient that a particular EVA design can have. Assuming all of the wire is cutting flux (\( L_w = L \)), the maximum damping coefficient can be expressed as,

\[
C_m = \frac{(BL)^2}{gL}
\]  

where, \( g \) = amount of resistance per meter of wire.

Substituting Eq. 27 into Eq. 28 we obtain

\[
C_m = \frac{B^2}{g} \frac{L_y}{2\pi aS} \sum_{n=1}^{L_y} r_n
\]

Figure 7. Wire wrapped around a cylinder
It can be shown that the flux density, height and radius of the wire adds nothing to a comparison of the maximum damping coefficient potential of a particular size of wire (Appendix C). With this in mind, Eq. 29 reduces to a wire material and size property for each gage of wire:

$$C_m = K' \left( \frac{a}{g} \right)$$  \hspace{1cm} (30)

where:

$$K' = B^2 2 \pi S \sum_{n=1}^{L_y} r_n.$$ 

Eq. 30 is used to find the optimum wire size for the maximum damping coefficient. A comparison of five readily available copper magnet wire gages using Eq. 30 is shown in Table 1. It should be noted that the smaller the wire gage number, the larger the wire diameter. As can be seen from Table 1, the smaller the wire gage, the higher the damping potential. This works well with the inductance equation, Eq. 26, since larger wire requires fewer layers in a given air gap, which lowers inductance. This result is not intuitive. One might falsely think that an increase in the layers of wire in an air gap would correspond to a much higher damping coefficient due to a much higher overall length of wire being cut by magnetic flux. The problem in using smaller diameter wire is that its resistance per unit length is much larger than that of larger diameter wire (compare g in Table 1 for 16 and 22 gage wire). This causes a larger loss in its damping potential than what the smaller wire gains from being able to fit more wire in a given air gap.

<table>
<thead>
<tr>
<th>Wire Gage</th>
<th>a (\text{turns/layer \text{m}})</th>
<th>g (\Omega/\text{m}) 10^{-3}</th>
<th>C_m/K (\text{turns/layer \Omega}) 10^4</th>
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</thead>
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<td>14</td>
<td>583</td>
<td>8.28</td>
<td>7.04</td>
</tr>
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<td>1.14e3</td>
<td>33.2</td>
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<tr>
<td>22</td>
<td>1.42e3</td>
<td>53.2</td>
<td>2.68</td>
</tr>
</tbody>
</table>

**Magnetic Circuit Modelling**

One of the requirements of this project was to make the electrodynamic vibration absorber a passive device. This means that the absorber should be able to dampen vibrations without any energy coming into the device to either monitor or create damping. There are two ways of
generating a magnetic field; one is to generate the flux through electric field windings and the other is to use permanent magnets. Since supplying a current defeats the requirement of no external energy input, permanent magnets are the only passive option.

A device that uses magnetic flux for any particular application should have a magnetic circuit analysis done to maximize the device's potential. A magnetic circuit is used to mathematically model the direction and intensity of magnetic flux lines. One of the best aides for calculating and approximating how a magnetic circuit behaves is a hysteresis loop. A hysteresis loop is a measure of how a magnetic flux density, B, changes in a ferromagnetic material in the presence of an applied magnetizing force, H. A typical hysteresis loop for a toroidal ferromagnetic material is shown in Figure 8. If a magnetically 'virgin' material suddenly has a magnetic field applied to it, a magnetic flux will develop within the material (see the initial curve in the first quadrant). This can be thought of as taking randomly oriented electron 'magnetic domains' and turning them so that they become aligned with the direction of the applied magnetizing force H. After all of the magnetic domains are aligned the material has reached a point where any additional applied magnetic force produces no appreciable change of magnetic flux. The point on the ordinate that corresponds to this point is called the saturation flux density, Bs. If the magnetizing force is reduced to zero after reaching saturation, the

![Hysteresis Loop Diagram](image)

Figure 8. Typical hysteresis loop for a toroidal ferromagnetic material
magnetic flux within the material will trace the upper curve back to the ordinate (zero applied magnetic field) to a point called the remanence flux point \( B_r \). The magnetizing force is reversed at this point and the flux in the material decreases through the second quadrant until it is zero. The magnetizing force at this point is called the coercive force \( H_c \). As the material continues to be subjected to the reversed field, it will again hit saturation (-\( B_s \)) in the third quadrant. The loop is then completed by reversing the direction of the magnetizing field again until it has passed through the third, fourth and first quadrants to reach the positive flux density saturation point.

There are two types of ferromagnetic materials; one is a magnetically soft material and the other is a magnetically hard material. Quantitatively, a magnetically soft material has a coercive force value \( H_c \) less than 200 Oe. A magnetically hard material is defined as having a coercive force value of \( H_c \) greater than 200 Oe. Magnetically soft materials are used as flux transporting materials because they are easily magnetized. Magnetically hard materials are used as flux generators because they are not easily magnetized, however once magnetized, they can easily magnetize soft magnetic materials. In magnetic circuit analysis, the first quadrant is useful in identifying how well a magnetically soft material will transport flux. The second quadrant is useful in determining how much magnetic flux a magnetically hard material will produce.

The amount of flux induced within the magnetic material is dependent upon the magnetizing force. As can be seen from Figure 8, the rate at which flux is produced within the material is variable for different levels of magnetizing force. The measure of the flux in a material for a given amount of magnetizing force is called the permeability \( \mu \) expressed as

\[
\mu = \frac{B}{H}
\]

Permeability is an important quantity for the determination of flux transportation potential since its value gives information on the amount of magnetizing force that has to be supplied to create a given flux within a material.

The hysteresis loop in Figure 8 was from a ferromagnetic material shaped as a toroid. The reason for choosing this shape is that a continuous 'flow' of flux lines will exist through the material without experiencing any discontinuities (air gaps). Earlier, it was said that after the toroidal magnetic material (magnet) reached saturation, the magnetizing force would be brought to zero. The corresponding flux density within the toroid with no magnetizing force applied was said to have a remanence flux density of \( B_r \). So, what happens when a slice is cut out of the magnet introducing an air gap into the flux path? This can be answered by referring to the second quadrant of the demagnetization curve of the magnet shown in Figure 9. The magnet
Figure 9. Self demagnetization of a permanent magnet provides a magnetomotive force (mmf) across the air gap to support a flux in the air gap. When this happens, the magnet's properties are identified from a point on the load line called the operating point of the magnet. This point can be found through simple geometrical means using the area and length of the air gap and magnet. This is expressed by Eq. 32 as

$$\frac{B_d}{H_d} = \frac{L_m}{A_m} P_t$$

where:

- $B_d =$ demagnetized operating flux density of the magnet closest to the air gap (G),
- $H_d =$ demagnetized operating magnetizing force of the magnet closest to the air gap (Oe),
- $L_m =$ magnet length in the direction of magnetization,
- $A_m =$ area of the magnet and,
- $P_t =$ total permeance of the magnetic circuit.

The total permeance of the magnetic circuit, introduced in Eq. 32, is a geometric quantity that takes into account the dimensions and permeabilities of the air gap and the magnetic flux transporting 'pole pieces'. Qualitatively, permeance is a measure of the ease a flux line passes through the magnetic circuit. The total permeance in complicated magnetic circuits can be calculated using a similar adding scheme as that for electrical circuits. In fact, very good
Figure 11. Moskowitz magnetic circuit for Model 1

Figure 12. Moskowitz magnetic circuit for Model 2

The total Moskowitz permeance \( P_{t1\infty} \) for Model 1 is given by

\[
P_{t1\infty} = P_{g1} = \frac{2\pi L_e}{\ln(1 + \frac{L_g}{r})}
\]  

(36)

The assumptions used in Eq. 36 are that the permeance of the pole pieces is infinite, the air gap reluctance between contacting pole pieces is negligible and a repulsing magnet is used at the top and the bottom of the device to prevent leakage. The total Moskowitz permeance \( P_{t2\infty} \) for Model 2 is given by

\[
P_{t2\infty} = 2P_1 + \frac{1}{1 + \frac{1}{P_{g2} + 2P_f} + \frac{1}{P_r}}
\]  

(37)

where:

\[
P_{g2} = \frac{2\pi W}{\ln(1 + \frac{L_g}{r + s + L_m + \delta})} = \text{air gap permeance,}
\]  

(38)

\[
P_f = 3.32(r + s + L_m + \frac{L_g}{2}) = \text{air gap fringing flux,}
\]  

(39)

\[
P_1 = 1.66(r + s + \frac{L_m}{2}) = \text{magnet leakage flux,}
\]  

(40)
\[ P_r = \frac{2\pi W}{\delta_r \ln(1+\frac{r}{\delta_r})} \]  

\( \delta_r \) = length of air gap between core and magnetic mass.

The assumptions used for the total pole piece permeance \( P_{t2oo} \) are that the pole piece permeance is infinite and the reluctance across the contacting pole pieces is negligible. The source for Eqs. 36 - 41 is reference [18]. Now that the total permeance has been obtained the operating flux density point \( B_d \) can now be obtained using Eq. 32.

Referring back to Eq. 32 there is only one way of analytically finding \( H_d \) without having to resort to experimentally finding \( B_d \) from the demagnetization curve of the magnet. Assuming that the operating point of the magnet is in the linear region, from geometry (see Figure 9) the expression for \( H_d \) is,

\[ H_d = \frac{B_r - B_d}{Q} \]  

where: \( Q = \) the linear slope of the load line.

Substituting in the terms for the magnet areas \( A_m \), Eq. 42 into Eq. 32 and entering the expressions of Eqs. 36 and 37 we have final expressions for both designs for the Moskowitz operating flux point \( B_d \),

\[ B_{d1oo} = \frac{B_r}{1 + \frac{Q}{L_m} \left( \frac{2\pi L_c (r + L_g + \delta)}{P_{t1oo}} \right)} \]  

\[ B_{d2oo} = \frac{B_r}{1 + \frac{Q}{L_m} \left( \frac{2\pi W (r + s + L_m)}{P_{t2oo}} \right)} \]  

where: \( B_{d1oo} \) = Moskowitz operating flux density point for Model 1 and,

\( B_{d2oo} \) = Moskowitz operating flux density point for Model 2.

It should be noted here that there is an underlying assumption used in both of these equations. This underlying assumption is that the demagnetized operating magnetizing force of the magnet \( H_d \) is constant through the length (thickness) of the magnet. This implies that the flux density inside the magnet is constant as well. A constant coercive force is present within rectangular or square magnets but is not necessarily true for cylindrical magnets. Cylindrical magnets of large radii will essentially have a constant coercive force through the thickness. For smaller cylinders this is not a valid assumption. So, quantitatively when can a constant coercive force assumption be used? The following discussion addresses this issue and comes up with an answer.
Figure 13 shows the internal flux density and magnetizing force of a magnet attached to a cylinder of finite radius. An expression can be found that takes into account the non-constant magnetizing force within the magnet by integrating the magnetizing force within the magnet. The integral of the magnetomotive force within the magnet is equal to the mmf needed to supply a magnetic field in the air gap (denoted by $L_g$ in Figure 10). Knowing that the magnitude of the magnetizing force (Oe) in the air gap is equal to the magnitude of the flux density (G),

$$\int H_d \, dL_m = \int B_g \, dL_g$$

where: $B_g$ = air gap flux density.

Two functions that express $H_d$ and $B_g$ in terms of flux density can be derived from simple magnetic circuit analysis. The (very simple) magnetic circuit shown in Figure 11 is used here for the non-constant magnetizing force derivation. From Model 1 (see Figure 10), it can be shown that the functions to express $H_d$ and $B_g$ are,

$$H(L_m) = \frac{1}{Q} \left[ B_r - \frac{r + L_g + \delta}{r + L_g + \delta + L_m} B_d \right]$$

$$B_g(L_g) = \frac{B_c r}{r + L_g}$$

where: $B_c = B_d \left( \frac{r + L_g + \delta}{r} \right)$ = flux density located at the surface of the core.

Figure 13. Flux density and magnetizing force inside a cylindrical magnet of Model 1
\( H_p \) that is needed to sustain the pole piece flux density is found. The ratio of the magnetizing force \( H_p \) to the demagnetized operating magnetizing force \( H_d \), is the percent increase in the thickness of the magnet required to achieve the calculated flux density \( B_d \). This is expressed as,

\[
L_{m'} = L_m \left(1 + \frac{H_p}{H_d}\right)
\]

where:

\( L_{m'} \) = the magnet thickness needed to obtain the calculated value \( B_d \).

The second method of analyzing magnetic circuits, the finite pole piece permeance method, is not radically different from the Moskowitz method. The basic idea is that the permeance is not infinite, which means that the pole pieces have a finite permeance that must be included in the total permeance calculation. Unlike the Moskowitz method, the finite permeability method does not have a correction factor that uses the first quadrant of the hysteresis loop. This method does, however, need the permeability value at a given level of flux density from the first quadrant of the hysteresis loop. The total magnetic circuit for Model 1 is given in Figure 17 and Figure 18 is the total magnetic circuit of Model 2. As can be seen from these figures, the total permeance will require extensive calculations due to the large number of permeance paths. The reward of such calculation may be improved accuracy for magnetic flux density predictions. The total permeance of these magnetic circuits are,

\[
P_{t1} = \frac{1}{\frac{1}{P_{g1}} + \frac{1}{P_{r1} + P_{r2}}} \]

\[
P_{t2} = 2P_1 + \frac{1}{\frac{1}{P_{g2} + 2P_f} + \frac{1}{P_{r1} + P_{r2}} + \frac{1}{P_r}} \]

where:

\[
P_{r1} = \frac{1}{\frac{1}{P_1} + \frac{1}{P_3} + \frac{1}{P_5} + \frac{1}{P_7} + \frac{1}{P_9}},
\]

\[
P_{r2} = \frac{1}{\frac{1}{P_2} + \frac{1}{P_4} + \frac{1}{P_6} + \frac{1}{P_8} + \frac{1}{P_{10}}},
\]

\[
P_1 = P_2 = \frac{\mu S A_s}{L_s} = \text{shell permeance},
\]

\[
P_3 = P_4 = \frac{\mu S A_{xs}}{L_{xs}} = \text{permeance between the top/bottom plates and the shell},
\]

\[
P_5 = \frac{2\pi \mu_{th} t}{\ln \left( \frac{s_{od} + s_{od}}{4r} \right)} = \text{top plate permeance},
\]
\[ P_6 = \frac{2\pi \mu_B h_b}{\ln\left(\frac{s_{id} + s_{od}}{4r}\right)} = \text{bottom plate permeance,} \]

\[ P_7 = P_8 = \frac{\mu_0 A_{xc}}{L_{xc}} = \text{permeance between the top/bottom plates and the core,} \]

\[ P_9 = P_{10} = \frac{\mu_c A_c}{L_c} = \text{core permeance,} \]

\[ \mu_0 = \text{Permeability in air (value = 1.0 in cgs units),} \]

\[ \mu_B = \text{Permeability of the bottom plate,} \]

\[ \mu_c = \text{Permeability of the core,} \]

\[ \mu_t = \text{Permeability of the top plate,} \]

\[ A_{xs} = \text{cross-sectional area of the shell,} \]

\[ L_{xs} = \text{length of the air gap between shell pole pieces,} \]

\[ A_{xc} = \text{cross-sectional area of the core,} \]

\[ L_{xc} = \text{length of the air gap between core pole pieces,} \]

\[ s_{id} = \text{inside diameter of the absorber shell, and} \]

\[ s_{od} = \text{outside diameter of the absorber shell.} \]

---

Figure 17. Model 1 total magnetic circuit
Figure 18. Model 2 total magnetic circuit

The operating flux density for each model is obtained by manipulating Eqs. 32, 49 and 51 is

\[ B_{d1} = \frac{B_T}{K_f1 + \frac{Q}{L_m} \left( \frac{2\pi L_g (r+L_g+\delta)}{P_{t1}} \right)} \]  \hspace{1cm} (56)

\[ B_{d2} = \frac{B_T}{K_f2 + \frac{Q}{L_m} \left( \frac{2\pi W(r+s+L_m)}{P_{t2}} \right)} \]  \hspace{1cm} (57)

Eqs. 56 and 57 represent the operating flux densities with the respective correction factors for each model and the pole piece permeances.

To this point all three 'technical background' sections have been divorced from one another, describing only certain aspects of the total electrodynamic vibration absorber phenomenon. In Eq. 3 the term BL denotes the flux density B that impinges on a length of wire L to create voltage. The wire length L was addressed briefly in the second section and the flux density B was addressed exhaustively in this section. Now, for analysis convenience, a BL product term will be derived so that each absorber model's damping coefficient \( C_m \) and damping force \( F_m \) can be easily found. The derivations will begin by first referring to Figure 19. The flux lines of Model 1 penetrate the conducting wire as shown. A simple relationship for the flux density in the wire at layer \( n \) can be related to the operating flux density by assuming all of the flux that comes out of the magnet goes through the wire expressed as,

\[ 2\pi(r+L_g+\delta)SB_d = 2\pi[r_n^d+\frac{d}{2}(1+\sqrt{3(n-1)})]SB_n \]
Manipulating this equation, the flux density at layer $n$ of the wire is,

$$B_n = \frac{B_d(r+L_g+\delta)}{r_n^d\left(1+\sqrt{3(n-1)}\right)}$$

Using Eqs. 25 and 27 we get the BL product for the $n$th layer of wire this is,

$$(BL)_n = \frac{B_d(r+L_g+\delta)}{d}\left[2\pi a(r_n^d\left(1+\sqrt{3(n-1)}\right))\right]$$

Manipulating this equation results in,

$$BL_1 = \sum_{n=1}^{Ly} 2\pi a(r+L_g+\delta)SB_d$$

Solving for the total number of layers $Ly$ by assuming that each layer of wire has an equal number of turns, the final expression for the Model 1 BL product is:

$$BL_1 = 2\pi a(Ly)[r+L_g+\delta]SB_d$$  \hspace{1cm} (58)$$

From Eq. 58 it can be seen that there are several different variables that can be manipulated to increase the BL product. These include the core radius, the height of the solenoid and the flux density.

![Diagram](image-url)

Figure 19. Model 1 flux impinging on conducting wire
The BL product can be developed in a similar fashion for Model 2. The derivation isn’t as straight forward for Model 2 as it was for Model 1. What complicates the derivation is that all of the flux coming from the magnetic face must pass across the air gap to cut the conductor at a right angle to maximize the voltage produced. Figure 20 is a drawing of the magnetic mass that shows the various flux paths. The BL product derivation will be broken into two parts for Model 2. The first part will find the BL product contribution from the flux in the air gap ($P_{g2}$) and the second part will find the contribution from the fringing flux. The ratio of permeances is the percentage of the flux from the magnet that goes into the air gap. This is expressed in the flux balance of the air gap as,

$$2\pi(r+s+L_m)WB_d\left(\frac{P_{g2}}{P_{g2}+2P_f+2P_l}\right) = 2\pi a W[r+s+L_m+\delta+\text{wall}+\frac{d}{2}(1+\sqrt{3}(n-1))]B_{n,g}$$

where: \(\text{wall} = \) a non-ferromagnetic wire restraining wall.

From this equation we find the flux density at the \(n^{th}\) layer of wire in the air gap is

$$B_{n,g} = \frac{B_d(P_{g2}/(P_{g2}+2P_f+2P_l))(r+s+L_m)}{[r+s+L_m+\delta+\text{wall}+\frac{d}{2}(1+\sqrt{3}(n-1))]$$

![Figure 20. Model 2 flux paths impinging on conducting wire](image-url)
Again using Eqs. 25 and 27 the BL product of the nth layer is

\[(BL)_{n,g} = \frac{B_{d}(P_{g2}^{2} + 2P_{f} + 2P_{l})}{[r+s+L_{m} + \delta + \text{wall} + \frac{d}{2}(1 + \sqrt{3}(n-1))]} (2\pi aW[r+s+L_{m} + \delta + \text{wall} + \frac{d}{2}(1 + \sqrt{3}(n-1))])\]

Summing over all layers of wire in the air gap and rearranging the above expression, the result is expressed by Eq. 59 as

\[(BL)_{2,g} = 2\pi aW(L_{y})B_{d}[r+s+L_{m}][\frac{P_{g2}^{2}}{P_{g2}^{2} + 2P_{f} + 2P_{l}}]\]  \( (59) \)

To give an approximation of the BL product contribution from the fringing flux, the assumption is made that between the two fringing paths a total equivalent flux cutting potential of one half the height of one path is perpendicularly cutting the conductor. Using this assumption and knowing that the arc height of the fringing flux is just the air gap length, \( L_{g} \), the contribution from the fringing flux is given by:

\[(BL)_{2,f} = 2\pi aL_{g}(L_{y})B_{d}[\frac{P_{f}}{P_{g2}^{2} + 2P_{f} + 2P_{l}}](r+s+L_{m})\]  \( (60) \)

Adding Eqs. 59 and 60 the total BL product of Model 2 is

\[BL_{2} = \frac{2\pi a(L_{y})B_{d}(r+s+L_{m})}{P_{g2}^{2} + 2P_{f} + 2P_{l}} [P_{g2}W+P_{f}L_{g}]\]  \( (61) \)

Material Selection

The discussion of magnetic circuits used soft magnetic materials (flux transporters) and hard magnetic materials (flux generators). This section presents a survey of both hard and soft magnetic materials describing what materials are available and what the best materials are to fulfill the needs of the electrodynamic vibration absorber on the basis of cost, availability and magnetic properties.

Whichever methodology one takes to model these magnetic circuits, a soft magnetic material has to have a low coercive force with high permeability so that there won't be an appreciable reduction in flux density in the air gap. A low magnetization force and high permeability go hand in hand with one another. If a material has a high permeability (>1000 G/Oe) at a high flux density (> 7000 G), the amount of magnetizing force required to provide this flux within the material will in general be low (< 30 Oe). There are many soft magnetic materials available on the market that have high permeability at various flux densities. In order to minimize the weight of the EVA, the pole pieces have to be able to carry a large amount of flux (i.e. a flux density >12,000 G). Therefore, one of the requirements of the EVA is that the
saturation flux density has to be high. Applying this requirement to soft magnetic materials commonly available, one finds that the number of suitable materials drops dramatically because of low saturation flux density values ($B_s < 10,000$ G). The remaining materials, which will be addressed here, are magnetic iron (pure iron), low carbon steel (< .25% C), silicon iron and cobalt steels. Among this list there is no mention of high carbon steels. Even though it appears that high carbon steels are 'very ferromagnetic' they do not carry flux well at high flux densities. The reason for this is that as the carbon content or impurities increase in steel, the maximum permeability and flux saturation decrease [19]. A comparison of a low carbon steel (1018) to that of a high carbon steel (1095) is depicted in the hysteresis loops of Figures 21 and 22. As can be seen from these two figures, there is a considerable difference in their permeabilities and saturation flux densities. One common bond pure iron, low carbon steels, silicon iron and cobalt steels have is that their magnetic properties are sensitive to internal stresses. Internal stresses have a detrimental effect on the permeability of a soft magnetic material. To relieve the internal stresses caused from machining or cold working, a material is annealed. By annealing these materials, their permeability increases and their hysteresis loops are narrowed, producing the desired performance required by the vibration absorber [19]. An example of how annealing affects magnetic properties is seen by comparing Figures 21 and 23. These two figures represent the same 1018 steel material. Figure 21 is for a cold rolled sample, and Figure 23 was annealed at $875^\circ$ C for 1 Hour per inch in diameter of specimen, furnace cooled and then machined to size. As can be seen, there is quite a large effect from the annealing process, with the permeability increasing by almost a factor of 2.

There are four candidates of soft magnetic materials thus far with potential for use in the EVA. These are, again, magnetic iron, low carbon steel, silicon iron and cobalt steels. The composition of materials within these headings does vary some and will give variable magnetic properties. Some of the best magnetic properties from these groups for the vibration absorber are magnetic iron with a composition of 99.63% pure iron, Carpenter Silicon Core Iron B-FM which has a high composition of iron with 2.5% silicon and Carpenter Hiperco 50 (cobalt steel) with a composition of 49% iron, 49% cobalt and 2% vanadium (i.e. vanadium permendur). Table 3 is a survey of these materials, after annealing, of their maximum permeability, saturation flux density, and hardness properties along with their availability and cost for small orders [20,21,22]. From this table it can be seen that for maximum permeability and flux saturation, Hiperco 50 is far superior than the rest. Also, one should note the dramatic difference that increased carbon content has on the maximum permeability.
Figure 21. 1018 Steel Hysteresis Loop

Figure 22. 1095 Steel Hysteresis Loop (note abscissa values twice that of Figure 21)
1020 Steel is only 0.2 wt % carbon yet its maximum permeability is three times less than the maximum permeability of magnetic iron. To machine a device out of these materials, magnetic iron is the best material presented in the table since it is the softest and therefore would be the easiest to work with. The clear winner in availability and cost is the low carbon 1020 Steel. The other three materials all have high costs because few companies stock or sell these materials in small quantities. Now that all of the materials have been compared, there are truly no clear winners. Each material has its advantages and disadvantages. The only weighing factor that decided the 'winner' was the cost because funds were limited for this project.

Table 3. Soft magnetic materials survey

<table>
<thead>
<tr>
<th>Material</th>
<th>$\mu_{\text{max}}$</th>
<th>$B_s$(G)</th>
<th>Hardness ($R_B$)</th>
<th>Availability</th>
<th>Cost ($/LB$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1020 Steel</td>
<td>1529</td>
<td>18000</td>
<td>72</td>
<td>High</td>
<td>.50</td>
</tr>
<tr>
<td>Magnetic Iron</td>
<td>5000</td>
<td>~20000</td>
<td>45</td>
<td>Low</td>
<td>5.00</td>
</tr>
<tr>
<td>Silicon Iron B</td>
<td>5000</td>
<td>20600</td>
<td>90</td>
<td>Low</td>
<td>10.00</td>
</tr>
<tr>
<td>Hiperco 50</td>
<td>10000</td>
<td>23000</td>
<td>84</td>
<td>Low</td>
<td>70.00</td>
</tr>
</tbody>
</table>
Now that the soft magnetic material considerations for the use in the EVA have been discussed, it is time to turn the discussion on the best hard magnetic materials for the absorber. There are several hard magnetic materials on the market that fall under three main types of materials (magnets) that could be used in the vibration absorber. The first are Ceramic magnets that have barium, strontium and iron for their main chemical components. The second are Alnico magnets whose main chemical components (as the name suggests) are aluminum, nickel and cobalt. The third type of magnets are made of rare earth compounds of neodymium-iron-boron or samarium-cobalt. The best way to describe these materials in a quantitative manner is to draw the typical demagnetization curves for all three types of magnets. These are shown in Figure 24. In this figure a new magnetic quantity called \((BH)_{\text{max}}\) is introduced. This quantity is commonly referred to as the maximum energy product, which in cgs units is quantified by GOE (gauss-ortsteds). The maximum energy product is the maximum amount of flux and magnetizing force that can be produced simultaneously by a magnet. It is arguably the measure of the quality of a magnet for the EVA since a high flux and a magnetizing force to sustain that flux in the air gap is desired. There is another new quantity presented in Figure 24 called the intrinsic coercive force, \(H_{\text{cj}}\). The intrinsic coercive force is a measure of how demagnetizable a magnet is; the larger the value of \(H_{\text{cj}}\) the more resistant the magnet is to demagnetization.

From Figure 24, the intrinsic coercive force lines for Rare Earth and Ceramic magnets are

![Figure 24. Demagnetization curves of common hard magnetic materials](image-url)
larger than the load line. The Alnico magnets have an intrinsic coercive force with essentially the same value as the load line. Rollin Parker classifies magnets into two categories, type I and type II [23]. A type I magnet is one in which \( H_{ci} < B_r \) and conversely a type II magnet is one where \( H_{ci} > B_r \). The Rare Earth magnets are type II and Ceramic magnets are or very close to being type II. Alnico's are all type I magnets which classifies them as being very susceptible to demagnetization.

Before a survey is undertaken to compare the magnets presented here, a simplification of the survey will be undertaken to eliminate the Alnico magnets from consideration for the EVA. To visualize why Alnico magnets should be eliminated, Figure 25 was made to give an example using an Alnico magnet in a hypothetical vibration absorber design. For a given design, the operating point of the Alnico magnet is found from the intersection point of the load line and the slope \( B_d/H_d \) dictated by Eq. 32 (Line 1). If this particular magnet were to be 'closed circuited' (i.e. have no air gaps) the magnet would not follow the load line back to \( B_r \) but would follow the recoil Line A to a new point denoted as \( B'_r \). This new load line is a result of an irreversible process in the magnet where the magnet cannot 'rebound' back to its original position. The failure to rebound comes only when the operating point of the magnet drops below the 'knee' of the load line. Above the knee the magnet will follow the original load line.

Figure 25. Alnico magnet in a hypothetical Electrodynaminc Vibration Absorber design
back to its original position time and time again. It should be said here that if the magnet does not operate below the operating point at the intersection of Line A with the load line, the magnet will continue to rebound time and time again along Line A for slopes greater than \( B_d/H_d \).

From Figure 25, it can be seen that slope of the load line above the knee and the slope of the recoil Line A are the same. This was not drawn by accident but because of the actual behavior of Alnico magnets. The slope \( Q \) is essentially the same for all slopes of the recoil lines below the knee. The problem that Alnico magnets have in their application in the EVA is that the magnets will have to be exposed to an equivalent magnetic circuit of air before they are placed in the magnetic circuit. This means that the new load line slope will be much lower than expected for the design as shown by Line C in Figure 25. Line 2 is the slope of the line dictated by Eq. 32 for a magnet exposed in air. If this new load line is not taken into account in the original design calculations (Line 1) the resulting operating flux density point will be \( B'_d \) with a new remanence flux value of \( B'_r \). This new value could possibly result in a huge reduction in the flux density compared with the originally calculated flux density value of \( B_d \).

This phenomenon does not eliminate Alnico magnets from consideration completely. Alnico magnets do have a large \( B_r \) value (12600 G) compared with Ceramic magnets (3800 G); therefore an Alnico magnet may still have enough energy left to compete strongly with Ceramic magnet performance. The eliminating factor comes from the fact that the induced current in the coil of the vibration absorber produces its own magnetizing force that assists in the further demagnetization of the magnet. What this means is that the induced current could drop the recoil line further below Line C, forcing the magnet to produce less flux and hence have less potential for the EVA to provide damping. This puts a ceiling on how much damping the absorber can produce and is definitely undesirable.

Now that one group of magnets has been eliminated from consideration altogether, a comparison can be initiated. The magnets that will be compared within the categories of Rare Earth and Ceramic magnets are the best of each category. These comparisons will consider the subcategory of flexible and cast arc magnets [24,25]. Table 4 presents such a comparison on the basis of analogous categories found in Table 3. As can be seen from Table 4, the hard Rare Earth arc magnet is far more superior to any other magnet in the table on the basis of magnetic properties. It should be pointed out that it is near impossible to obtain an arc magnet with this material. Even though the relative cost for NdFeB 37 is only listed as 28 times the cost per unit weight of a Ceramic magnet, its actual cost after manufacturing may be as much as 100 times. The reason for this huge increase in cost is because Rare Earth arc magnets are not readily available and would require special tooling to grind them to specification. Taking the
Table 4. Hard magnetic materials survey

<table>
<thead>
<tr>
<th>Material</th>
<th>B_r (G)</th>
<th>H_c (Oe)</th>
<th>Hardness</th>
<th>Availability</th>
<th>Rel. Cost^2</th>
</tr>
</thead>
<tbody>
<tr>
<td>NdFeB 37 Arc</td>
<td>12400</td>
<td>11800</td>
<td>R_c 58</td>
<td>Near Impossible</td>
<td>28</td>
</tr>
<tr>
<td>Neoflex</td>
<td>5450</td>
<td>4900</td>
<td>Shore D 55</td>
<td>Moderate</td>
<td>17</td>
</tr>
<tr>
<td>Arnox 4000 Arc</td>
<td>4000</td>
<td>3700</td>
<td>R_c 58</td>
<td>High</td>
<td>1</td>
</tr>
<tr>
<td>High Flex 3</td>
<td>2650</td>
<td>2200</td>
<td>---</td>
<td>Moderate</td>
<td>1</td>
</tr>
</tbody>
</table>

^2Relative cost by weight to Ceramic magnets.

availability and cost into consideration, Rare Earth arc magnets are eliminated with the remaining three magnets still able candidates for use in the vibration absorber. The High Flex 3 flexible magnet has several attractive qualities for use in the absorber but has a fatal quality of having a very low B_r value. As has been mentioned before, the flux density value for generating voltage has to be as large as possible. Since the damping coefficient is proportional to the flux density squared, the resulting flux produced by a High Flux 3 magnet would be several times less than the flux from either Neoflex, the Rare Earth flexible magnet or the Arnox 4000 Arc Ceramic magnet. The last two candidates, Neoflex and Arnox 4000 have some what similar magnetic properties but Neoflex has slightly better coercive force and remanence flux values. This does not eliminate the Arnox 4000 magnets from contention. What does put the final selection towards the Neoflex material is that it can be very easy to work with and versatile for many different designs. The Shore D hardness number of 55 is a way of quantifying Neoflex as having the hardness of a firm rubber material. With Neoflex having properties like rubber, it has the advantages of rubber in that it can be cut and shaped to fit various configurations. Arnox 4000 magnets or any ceramic arc magnets in general have fixed shapes and are very hard to machine. The hardness value of Arnox 4000 is not given nor can it be found in literature. The value reported in Table 4 is an experimentally derived value for an Arnox 8 magnet. During the test it was apparent that the magnet had little tolerance for localized plastic deformation by the flakes that appeared on the specimen caused by cracking. What this leads one to conclude is that this material would be very difficult to machine, therefore the only way to get a Ceramic arc magnet to size in a particular design is to design the absorber around the arc magnet. Use of a material that dictates the size or shape of the device is not a very practical way of optimizing a design.
FINAL DESIGNS

This section is devoted to explaining the final designs in detail. The final designs were a result of taking into account all of the aforementioned modelling considerations. Particular emphasis was placed on each design's maximum damping ratio $C_m$, BL product, damping to mass ratio $C_m/M$, maximum amplitude, inductance, coefficient of kinetic friction $\mu_k$, weight, availability of materials and cost. Computer programs for Models 1 and 2 were developed to calculate input dimensions and materials of each type of design. These programs were not optimization programs because of the overwhelming number of design considerations. The code of these programs are found in the appendix. The following designs were developed and built in reverse order from the model numbers, i.e. Model 2 is Design MM (moving magnet) was built first and Model 1 is Design MW (moving wire) was built second. Figure 26 is an assembly drawing of Design MM and below it in Table 5 is the theoretical design values using the Moskowitz magnetic circuit methodology. Figure 27 is an assembly drawing of Design MW and following it in Table 6 is the theoretical design values using the finite permeability magnetic circuit methodology. The difference in using a different magnetic circuit method for the theoretical values in Design MW will be explained later in the results and discussion section. Due to a more efficient use of soft magnetic material (lighter weight) than in Model 1, Model 2 was fabricated first. Efficient use of material was based on magnetic flux paths and densities. The goal in efficient use of soft magnetic material is to always carry as much flux as possible with the least amount of material. In the center of the core and the shell of Model 1, the flux density of the soft magnetic material small. This comes from the positioning of the magnets (refer back to Figure 10). As the flux moves away from the center of Model 1, the flux density increases until maximum flux density is reached at the top and bottom. As can be seen from this example, the soft magnetic material in Model 1 is not used efficiently. In contrast, Model 2 efficiently uses almost all of the soft magnetic material because virtually all of the soft magnetic material has a high flux density.

Design MM

Design MM, shown in Figure 26, has many features that need explaining. The 1018 Annealed Steel was selected based on both its availability and cost. For magnetic purposes Magnetic Iron would have been the next best, but, because of limited funding resources the cheaper low carbon steel was selected. Neoflex was selected because of its ability of supplying
a large amount of flux to the air gap and it allowed for design flexibility. Hardwood bearings were selected for their low coefficient of kinetic friction and low thermal expansion as compared to its plastic competitors. An added bonus for the selection of hardwood bearings was their high wear capacity. The wire used in Design MM was, unfortunately, 18 Gage magnet wire. This type of wire was selected because of funding constraints and, an error was made in the initial wire analysis which later suggested that 14 gage wire was the best to use. Brass shimstock was used to protect and constrain the wire from moving. It also had the property of being non-magnetic which was needed so that the flux coming from the magnet was not interfered with as it traveled through the wire. The final major feature of Design MM is the use of linear bearings. Linear bearings were used to constrain any rotation of the magnetic mass and to provide alignment for the transfer of motion from outside the absorber to the magnetic mass.

Figure 26. Design MM - moving magnet relative to stationary wire

<table>
<thead>
<tr>
<th>Table 5. Design MM theoretical design values</th>
</tr>
</thead>
<tbody>
<tr>
<td>C_m (kg/s)</td>
</tr>
<tr>
<td>------------------</td>
</tr>
<tr>
<td>110</td>
</tr>
</tbody>
</table>

$^3$Solenoid encompassed in air, hence $\mu_r = \mu_0$. 
face by placing like poles at the ends against those of the magnet. This action results in forcing the flux to be uniform in the air gap and to force flux into the core. The wire used in Design MW was 18 Gage wire. The use of this wire was not due to funding constraints but solely due to error in the wire analysis. If the appropriate analysis would have been taken place (as that presented in this document), 14 or 16 gage wire would have been purchased and implemented into this design to increasing its damping potential.

The theoretical values of Design MW are presented in Table 6. As can be seen from Table 6, the values are essentially the same as Design MM. The BL product is smaller than Design MM, but because the resistance of the wire is also smaller, the resulting damping coefficient was a little larger than for Design MM. The damping to mass ratio is nearly the same for both absorber designs but it should be pointed out that there is a difference in maximum

![Diagram](image)

Figure 27. Design MW - moving wire relative to stationary magnet

<table>
<thead>
<tr>
<th>Table 6. Design MW theoretical design values</th>
</tr>
</thead>
<tbody>
<tr>
<td>$C_m$ (kg/s)</td>
</tr>
<tr>
<td>-------------</td>
</tr>
<tr>
<td>129</td>
</tr>
</tbody>
</table>

$^4$Solenoid encompassed in air, hence $\mu_r = \mu_o$. 
displacement amplitudes. If in Design MW an equivalent maximum amplitude were designed, the weight and the mass of Design MW would be much greater. The reason for this is that if the amplitude of the absorber is increased, the solenoid length would also be increased. More magnet surface area would have to be added. A greater magnet surface area produces more flux which requires more soft magnetic material. The maximum amplitude was reduced for Design MW because the testing methodology was changed from using a spring-mass-damper system to just attaching the device to a shaker. The change in methodology also meant there was a change in maximum amplitude. The shaker only has a maximum amplitude of 2.54 cm (1.0 inch). The spring-mass-damper system, on the other hand, has a maximum amplitude only limited to the spring's ability to stretch or compress. The last Design MW parameter was inductance. The reduced value in inductance of Design MW was purposely designed to be less than that for Design MM. The reason for doing this was because the impedance contribution from the inductance of Design MM was thought to be very significant.
EXPERIMENTAL DETAILS

Equipment List

1. Hall Generator (Gauss Probe) F.W. Bell
   Accuracy: ± 1.5%, Range 0 - 10,000 G

2. Magnetic flux analyzer Walker Scientific Model MG-3A
   Range: 1, 3, 10, 30, 100, 300, 1K, 3K, 10K & 30KG
   Accuracy: ± 0.1%

3. Hall Probe Walker Scientific Model HP - 13S
   ± 1% Linearity of Reading to 10,000 G

4. Solartron Schlumberger 1260 Impedance/Gain-Phase Analyzer

5. Norland 3001/DMX Data Acquisition System
   Options: Four Channel
   Accuracy: ± 0.75 % from DC to 60 KHz
   Sensitivity: Ten ranges from ±100 mV to ±100 V
   Frequency Response: Flatness within 0.5 % from DC to 10 KHz

6. Wavetek 4MHz Sweep/Function Generator

7. PCB Power Amplifier Model 480D06

8. PCB 308B Accelerometer
   Sensitivity 99.9 mV/g, S/N 2381

9. PCB 302A02 Accelerometer
   Sensitivity 10.04 mV/g, S/N 8625

10. Unholtz-Dickie Vibration Testing System
    Model: TA250-206 Shaker SN 502
    Shaker Stroke: 1 inch
    Max. Free Table Acceleration: 60 g
    Frequency Response: ±0.5 dB - 5000 Hz

11. Macintosh SI with LabView data acquisition software

Flux Test

The flux tests of both devices were conducted at the Ames Laboratory. The tests used two different devices that reported flux density values according to the Hall effect principle. A Hall
probe was used in Design MM located approximately in the center of the device. The probe was attached on the inside of the shell as demonstrated in Figure 28. The magnetic mass of Design MM was moved at approximately 1/8 inch increments starting from the bottom and finishing when no appreciable flux (< 100 G) could be measured. Flux testing Design MW involved using a Gauss probe. This probe was entered into the device from the bottom and moved at 10 mm increments along the core and the shell closest to the magnets until the probe was fully extended. The flux test for Design MW is also shown in Figure 28.

Impedance Analysis

The impedance analysis was conducted on each design using a Solaratron Impedance Analyzer. The wire leads from Design MM were connected to the clips of the analyzer. Then, a computer program was set to sweep and measure impedance values of integer frequencies ranging from 1 - 100 Hz. A test consisting of one sweep of the test range was done with the magnetic mass positioned at the bottom, middle and top of the absorber. The same procedure was then used to test Design MW. An additional test was done with the solenoid of wire of Design MM being encompassed in air with no absorber components near by.
Voltage Measurements

The voltage measurements were done using PCB accelerometers, a Wavetek function generator, an Unholtz-Dickie Electrodynamic shaker and a Norland data acquisition system. The test setup is shown in Figure 29. Each absorber design was connected to the test stand and the shaker head. An accelerometer was placed on the top plate holding the absorber and on the shaker head. The output of each accelerometer was then fed into power amplifiers. The wire leads of the vibration absorber were connected to the Norland data acquisition system to read the open circuit voltage of the device. The driving frequency of the shaker was placed at integer frequencies, at an arbitrary shaker amplitude, through the range of 1 - 20 Hz and then through the individual frequencies of 30, 40, 50 and 100 Hz. At each frequency, the data acquisition rate was adjusted to give the best frequency resolution. The voltage outputs were read from both accelerometers and the vibration absorber at each frequency. To give an average response, the output voltage root mean square (RMS) value was taken from the accelerometer on the shaker head and the absorber. The top plate accelerometer maximum voltage output was found by taking the fast fourier transform (FFT) of the signal. This was done instead of taking the RMS value because multiple frequency components were present in the signal. Taking the RMS value of a signal with multiple frequency components produces erroneous amplitude values. This happens because voltage readings from other frequencies add to a desired frequency falsely giving a voltage value much larger than is present at that frequency. The last component of information needed was obtained by taking the FFT of the vibration absorber's output to find at what frequency the absorber was being driven.

![Diagram of voltage measurement experimental apparatus](image-url)
RESULTS AND DISCUSSION

This section summarizes the results of the tests done on both designs of the EVA. First to be discussed is the impedance measurements that reveal the damping potential of this type of damping. The next section will analyze the magnetic flux measurements and will contain the answer of what magnetic circuit method is the most appropriate for modelling both vibration absorber designs. This section will also introduce and discuss a flux testing device specifically made for this project to better understand the flux test results. The last section will compare theoretical to experimental results for the amount of voltage generated by each design.

Electrical Impedance

It was originally thought that the resistance and the inductance would be linear quantities. For the solenoids encompassed in air (air solenoids) of Designs A and B, this is approximately true. When the soft magnetic material and magnets are assembled with the solenoids, the impedance becomes nonlinear. Figure 30 gives the results of the impedance test for Design MM with the solenoid encompassed in air. The impedance from the inductance (reactance)

![Impedance Graph](image)

Figure 30. Design MM solenoid in air impedance plot
in Figure 30 is linear as expected. The resistance is slightly non-linear because it gradually increases with frequency. This implies that there may be an imaginary term present in the inductance that could add to the real 'resistance' part. The increase in resistance over this frequency range is small and can substantiate the earlier claim that the impedance is approximately linear. A plot of the impedance of Design MM after the absorber has been assembled with the soft magnetic material and the magnet is shown in Figure 31. The reactance has increased by approximately four times over that of the air solenoid. This increase was expected since a high permeability material had been placed inside and outside the absorber. What is unusual about the reactance is that between 1 and 10 Hz it is nonlinear. The apparent nonlinearity is probably caused from electromagnetic interactions between the solenoid and either the magnetic flux from the magnet or the soft magnetic material. This type of interaction may be the cause of the dramatic increase in the resistance. For this particular

![Graph showing impedance vs frequency](image)

**Figure 31.** Design MM impedance with soft magnetic material and magnet assembly

range of frequencies, resistance was the major contributor of the impedance. This result was definitely not expected. The theoretical section described a method of how the inductance contribution could be nullified by electrical resonance using a capacitor in series with the absorber. The resistance cannot be reduced as easily. Since the resistance increases
dramatically, as is seen in Figure 31, the maximum damping coefficient will be very limited. The maximum damping coefficients for the frequency range of 1 to 100 Hz are shown in Figure 32. Figure 32 shows that the maximum damping coefficient is dramatically reduced by the increase in resistance. This indicates that the potential of Design MM to dampen vibrations is very limited. Remembering that the damping to mass ratio is important (especially for space applications), it goes from about 11 at 1 Hz to a dismal value of 2 at 100 Hz. What the value of 2 means for damping applications at 100 Hz, is that for every two pounds of damping force needed to dampen a vibrating structure, one pound of vibration absorber has to be added to the structure. Space applications are the only applications identified at this time that would benefit from magnetic types of damping. With the damping to mass ratio being so small, an eddy current damper could be used with much more effectiveness than the EVA [3].

An increase in resistance has been shown to limit the maximum damping coefficient of Design MM. An explanation of the cause of the increasing resistance phenomenon has not been discussed. The resistance could have increased by an electromagnetic interaction from either the flux produced by the magnet or from the soft magnetic material. To further explain this phenomenon, an impedance test was conducted where the magnetic mass was taken out of

![Graph showing the relationship between Maximum Damping Coefficient (kg/s) and Frequency (Hz).](Image)

Figure 32. Design MM maximum damping coefficient vs. frequency
Design MM leaving just the housing and solenoid. Figure 33 is a plot of the resulting impedance measurements compared against those presented earlier for impedance of the total assembly of Design MM. The impedance with no magnetic mass is essentially the same as that with a magnetic mass. This clearly indicates that the increase in resistance or reactance with frequency is not due to an electromagnetic interaction between the solenoid and the magnetic flux from the magnetic mass magnets. What is surprising are the results of the reactance in Figure 33. Virtually no flux is present in the pole pieces. The lack of flux present in the pole pieces means that the magnetic domains in the pole pieces are free to interact with the magnetic flux coming from the current in the solenoid. When an interaction of this kind takes place in the presence of a reduced exterior magnetic field, there should be an increase in reactance. The results shown imply that there is no increase in reactance. It appears that there is some 'steady-state' or constant electrical permeability value associated with the housing material used where the reactance remains the same value whether or not a magnetic field is present. This statement should not be confused with the permeability associated with the ratio of the magnetic flux to the magnetizing force inside the magnetic material. The electrical permeability, as it is meant here, is the permeability value used in Eq. 26.

![Graph](image)

*Figure 33. Design MM impedance with no magnetic mass*
The impedance values for Design MW are similar in form of those in Design MM. Figure 34 is the impedance of Design MW’s solenoid wrapped around an aluminum cylinder in air. The resistance is seen to increase more rapidly with frequency than did the resistance of Design MM (53% increase vs. 25% increase of Design MM). One would guess that the resistance from the much smaller solenoid would, if anything, have a smaller increase with frequency because it contains less wire. The answer may be found from the aluminum cylinder to which the wire is attached. Aluminum has a very low resistivity value which means that eddy currents are easily induced when a time varying flux is produced in the surrounding wire. Eddy currents produce a magnetic flux that opposes the flux that created them. What this means electrically, is that the current in the coil experiences a force contrary to its motion effectively increasing the resistance of the wire. This may explain the increase observed in resistance in Design MM when the steel is present around the solenoid. Steel has similar electrical properties to that of aluminum. The solenoid in Design MM is surrounded by conducting steel implying that this may be the cause of its increase in resistance.

The reactance in Figure 34 has essentially the same form as that of Design MM. It appears that an eddy current effect causes the resistance to change but does not have any effect on the inductance. This can be seen by the linearity of the reactance in Figure 34. What is important

![Impedance plot](image.png)

Figure 34. Design MW solenoid in air wrapped around an aluminum cylinder impedance plot
about this result is, that if eddy currents did effect the reactance of Design MW then, the value of the reactance should show some sort of nonlinear behavior; behavior that is obviously not seen in Figure 34.

The results of the impedance test of Design MW after assembly are shown in Figure 35. The resistance is seen to increase dramatically after assembly and is the major contributor to the total magnitude of the impedance. Comparing the increase of the resistance of Design MW to Design MM over the 100 Hz range, one finds that the resistance of Design MM has increased more than Design MW (approx. 450% increase vs. a 350% increase). The reason for this increase is probably due to the fact that the solenoid in Design MM is surrounded by steel over a larger amount of area than Design MW. Using the eddy current explanation, the added area of a good conductor close to the solenoid in Design MM should induce greater resistance than for Design MW.

The reactance of Design MW is a small contributor to the overall impedance of the device. This implies that electrical resonance would have a small effect in reducing the impedance of Design MW. This is shown in Figure 36 with the maximum damping ratio plotted against the absorber excitation frequency with and without circuit resonance. The damping coefficient does not increase appreciably with the elimination of reactance. Not shown in Figure 36, is the

![Graph showing impedance vs. frequency]

Figure 35. Design MW impedance with soft magnetic material and magnet assembly
maximum damping coefficient with constant resistance. Referring back to Table 6, the maximum value was reported to be 129 kg/s. Comparing this value with the values in Figure 36, it can be seen that the changing resistance considerably limits the damping potential of this design. This implies that Design MW has essentially the same limited damping potential as Design MM on either a maximum damping coefficient or a damping to mass ratio basis.

![Graph showing damping coefficient vs. frequency]

Figure 36. Design MW maximum damping coefficient vs. frequency

Magnetic Flux Measurements

This section will compare the experimental magnetic flux measurements against the theoretical modelling values in order to determine which magnetic circuit method is the most accurate. Included will be the theory and results of a flux testing device that was used to confirm magnetic flux values given by vendors for the magnetic materials purchased.

Table 7 was constructed using the theoretical results obtained by computer programs and by experimentally measuring the flux in the air gaps of both designs. The measured values were derived from the experimental values using a calculation scheme found in Appendix F. The theoretical values have taken into account the two different types of neoflex used in the project. The magnetic material used in Design MM had a specified $B_r$ value of 6200 G and the
magnetic material in Design MW was specified to have a $B_T$ value of 5450 G. The Moskowitz values that appear in Table 7 have been adjusted using a form of Eq. 53 given here as

$$B_{d'} = B_d (1 - \frac{H_p}{H_d})$$

(62)

where: $B_{d'}$ = the corrected Moskowitz flux density value.
Instead of adding an equivalent magnet thickness to achieve the calculated value $B_d$, the pole piece to demagnetized operating magnetizing force ratio is subtracted to give a new reduced equivalent thickness. Since the operating point $B_d$ is proportional to the thickness of the magnet, Eq. 62 is a valid means of correcting the calculated flux density.

Table 7. Design MM measured vs. Moskowitz and finite pole piece flux density values

<table>
<thead>
<tr>
<th>Design</th>
<th>Measured, $B_d$(G)</th>
<th>Moskowitz, $B_d$(G)</th>
<th>% Error</th>
<th>Finite, $B_d$(G)</th>
<th>% Error</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>$2.01 \times 10^3$</td>
<td>$3.00 \times 10^3$</td>
<td>33.0</td>
<td>$2.66 \times 10^3$</td>
<td>24.4</td>
</tr>
</tbody>
</table>

Referring to Table 7, it can be seen that the measured and theoretical values are in large disagreement. The finite pole piece permeance flux density values agree better with the measured values than do the Moskowitz infinite pole piece permeance flux density values. The better agreement of the finite method values appears to indicate that it is the best method for magnetic circuit modelling. Any conclusions drawn to state that the finite method is the better of the two modelling methods is premature at this point because of the large errors associated with its value for Design MM. The errors could have come from two possibilities. The first is that an unspecified foreign material may have been used in the magnetic circuit. The second is that the magnetic material used in Design MM could have a lower $B_T$ value than specified. The first possibility was explored by obtaining hysterisis loops and conducting Rockwell B hardness tests of all of the soft magnetic materials used in both absorbers. The materials used in the top, bottom, and shell pieces for each design had hysterisis loops similar to that found in Figure 23 for 1018 Annealed Steel. The core piece in Design MW also was similar to that of 1018 Annealed Steel but there was a foreign material found in the core of Design MM. Figure 37 is the hysterisis loop of the material used in the core of Design MM. Comparing this hysterisis loop to that of Figure 23, it is quite clear that the core material was not 1018 Steel. The permeability is considerably less and the coercive force is larger than 1018 Steel. Adding this new information in to the theoretical analysis will improve the theoretical predictions.
The hysteresis loop finding was confirmed by hardness tests. Table 8 gives the results of the hardness tests conducted on the soft magnetic material components. The hardness tests clearly show that the core in Design MM is foreign. Published literature hardness values of annealed carbon steels shows that the hardness values for low carbon 1018/1020 steels have a Rockwell hardness value near 70 [22]. The hardness value of 93 for the core of Design MM corresponds well with the published values of a 1055 steel. Therefore, the steel used in the core of Design MM was not a low carbon 1018 steel but a medium carbon steel with hardness values near those of 1055 steel.

![Graph](image)

Figure 37. Hysteresis loop of the core material in Design MM

<table>
<thead>
<tr>
<th>Design</th>
<th>Shell 5</th>
<th>Core</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>64.4</td>
<td>93.8</td>
</tr>
<tr>
<td>B</td>
<td>75.1</td>
<td>69.5</td>
</tr>
</tbody>
</table>

5 Top and bottom plates were made from the same material.
Table 9 is a new comparison of the measured and theoretical flux density values adjusting for the use of the foreign material in the core in Design MM. There is a dramatic reduction in the error of the finite permeance values. The error is cut in half from what it was without considering the foreign material. The reason for this is that the permeability value used before was 745 and now it is 159 for the core. This dramatic improvement was not realized in the Moskowitz calculations. The reason for this is that the amount of magnetizing force required for the core pole piece only went from 25 Oe for 1018 Annealed Steel to 55 Oe for the foreign 1055 Steel. The demagnetized magnetizing force $H_d$ is approximately $2.66 \times 10^3$ Oe. The total effect of an increase of 30 Oe had little effect on reducing the predicted operating flux density value of Design MM. Despite the improvement in the error, the resulting errors associated with each design are still excessive.

Table 9. Design MM measured to theoretical flux density values with material adjustment

<table>
<thead>
<tr>
<th>Design</th>
<th>Measured, $B_d(G)$</th>
<th>Moskowitz, $B_d(G)$</th>
<th>% Error</th>
<th>Finite, $B_d(G)$</th>
<th>% Error</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>$2.01 \times 10^3$</td>
<td>$2.96 \times 10^3$</td>
<td>32.1</td>
<td>$2.30 \times 10^3$</td>
<td>12.6</td>
</tr>
</tbody>
</table>

Another possibility is that the magnetic material purchased was not the material specified. A flux testing device was built to measure the $B_r$ value for each magnetic material used in both designs. This was done by designing and building a device that had high magnetic circuit calculability and could give good experimental values. The resulting flux testing device (flux tester) is shown in Figure 38. The methodology in using the device was to first measure the air gap flux density and then work back through the magnetic circuit equations to find the $B_r$ value. The equations for the flux tester were developed using a form of Eq. 32. The operating point $B_d$ and remanence flux density $B_r$ of the magnet are expressed as

$$B_d = B_g \left( \frac{P_{gt}}{P_{gf}} \right)$$

$$B_r = B_d (1 + \frac{Q_A m}{L_m P_{tf}})$$

where:

- $P_{gf}$ = air gap permeance of the flux tester,
- $P_{gt}$ = total permeance across the air gap including fringing and leakage, and
- $P_{tf}$ = total permeance of the magnetic circuit.

The additional expressions for the permeances, magnetic circuit and error analysis for the flux tester are found in Appendix D.
There are some important features that should be noted about the flux tester. The flux tester can only test materials where the load line slope $Q$ is approximately known and the magnet used in the test has to be operated in its linear region of the load line. This means that almost all of the Alnico magnets cannot be tested, but nearly all Ceramic and Rare Earth magnets can be tested by this device. One feature of the flux tester is the circular shape of the steel pole piece inhibits airborn flux paths making it easier to calculate the permeance contribution of the pole pieces. The two paths the flux can take through the pole piece is advantageous for flux testing because the permeance is doubled allowing for a significant decrease in reluctance losses. The flux tester is very sensitive to the air gap length. If the air gap length varies by more than .02 inch, the accuracy of any value calculated by Eq. 64 will be questionable. The flux tester is also sensitive to the slope $Q$ of the load line. For a given Rare Earth magnet, the slope $Q$ will be approximately equal to the ratio of $B_T$ to $H_C$. The error in the slope can be estimated by knowing that $B_T$ can vary by $\pm 5\%$ and $H_C$ can vary by $\pm 7\%$. Variations in the $B_T$ value are caused by discrepancies in magnet chemical composition, while variations in the $H_C$ value are caused by differences in heat treatment [26].

The flux tester was calibrated against a sample of Arnox 8 Ceramic permanent magnets. Table 10 is a summary of the calibration and test results for each type of magnet used in both designs and subsequent 'guessed' materials. The acronym of DFPR stands for 'deviation from published results'. It was found that the flux tester was 8% higher than published values of the

![Diagram](image)

**Figure 38. $B_T$ Flux Tester**
Arnox 8 Ceramic magnets. The published values have a tolerance of ±5% which means that the flux tester is either good to within 2% or off by 13%. These results indicate that the flux tester reads the actual value higher than the true value. The rest of the values in Table 10 should be analyzed with this in mind. The next magnet after Arnox 8, PM 80, is the magnet specified for Design MM. As can be seen from the test, the resulting $B_T$ value is 9% lower than published. Realizing that the flux tester is reading flux densities higher than the true value, this indicates that the magnet used in Design MM was not the magnet specified for the design. The next value in Table 10, PM 70, is a lower grade flexible Rare Earth magnet sold by the same vendor. The values of this magnet were used to see if there was a better correlation between measured and published values. The result using PM 70 was better than that of PM 80 but is still lower than published values. Using the same argument as that used for PM 80, PM 70 is still not the correct material. The last chance of correlating experimental values to published values for the flexible Rare Earth magnet was to try another known lower grade magnet called Neoflex. The Neoflex values for Design MM (MM) correlated very well to the experimental results. Since the new measured value is higher than the published result, a conclusion can be drawn that the magnet used in Design MM is the lower grade flexible rare earth material Neoflex. What this means is that the error between theoretical and experimental flux density values in Design MM will improve dramatically with this new information. Table 10 suggests that the specified Neoflex was used in Design MW (MW). The published and experimental values are almost equal. Since the flux tester does report high flux density values, the actual value could be a few percent lower but would probably not affect the theoretical to experimental values for Design MW to any appreciable extent.

The final theoretical versus experimental values for Design MM of the operating point $B_d$ using all of the new information uncovered about the magnetic hard and soft materials are presented in Table 11. The theoretical values assumed that the magnet used in Design MM was

<table>
<thead>
<tr>
<th>Magnet</th>
<th>Published $B_T$</th>
<th>Deviation</th>
<th>Measured $B_T$</th>
<th>Deviation</th>
<th>DFPR</th>
</tr>
</thead>
<tbody>
<tr>
<td>Arnox 8</td>
<td>3850 G</td>
<td>±192</td>
<td>4190 G</td>
<td>±75</td>
<td>+8.11%</td>
</tr>
<tr>
<td>PM 80</td>
<td>6200 G</td>
<td>±310</td>
<td>5640 G</td>
<td>±125</td>
<td>-9.03%</td>
</tr>
<tr>
<td>PM 70</td>
<td>6000 G</td>
<td>±300</td>
<td>5650 G</td>
<td>±125</td>
<td>-5.83%</td>
</tr>
<tr>
<td>Neoflex, MM</td>
<td>5450 G</td>
<td>±272</td>
<td>5570 G</td>
<td>±120</td>
<td>+2.15%</td>
</tr>
<tr>
<td>Neoflex, MW</td>
<td>5450 G</td>
<td>±272</td>
<td>5440 G</td>
<td>±333</td>
<td>-0.16%</td>
</tr>
</tbody>
</table>
Neoflex. As can be seen from the table, the error for the finite flux density is very acceptable. Comparing the two magnetic circuit methods, it can be seen clearly that the finite pole piece permeance is the best method for predicting the magnetic circuit behavior of the vibration absorber. The reason for the failure of the Moskowitz method may come from the fact that this method does not take into account the length of the pole pieces; just the magnetizing force required to sustain a magnetic flux in them. Implied by Eq. 45, the mmf produced by the magnet will be reduced the longer a pole piece gets since sustaining a flux in a longer pole piece requires more mmf from the magnet. This lost mmf will not be present to support the magnetic flux in the air gap and therefore will result in a reduced magnetic flux density. With the failure of the Moskowitz method and the success of the finite permeance method, Design MW was modelled by the finite method.

Table 11. Design MM measured to theoretical flux density values with total adjustment

<table>
<thead>
<tr>
<th>Design</th>
<th>Measured, B_d(G)</th>
<th>Moskowitz, B_d(G)</th>
<th>% Error</th>
<th>Finite, B_d(G)</th>
<th>% Error</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>2.01 x 10^3</td>
<td>2.74 x 10^3</td>
<td>26.6</td>
<td>2.10 x 10^3</td>
<td>4.28</td>
</tr>
</tbody>
</table>

A comparison of the measured to theoretical flux density values for Design MW are shown in Table 12. As can be seen, the finite permeance model is in better agreement with the theoretical values than the Moskowitz method. The error is greater for this design than the other design. The reason for this may be from deviation in the B_r value which the vendor suggests can deviate ± 5%. Another reason is that there may have been a larger air gap between the pole pieces than was predicted (.001 in.). This is a very possible because there is a good chance that either one or both of the contacting surfaces between the top and bottom plates does not fit squarely with the core. If this happened, the permeance of an introduced air gap would lower the flux density in the air gap. As can be seen from Table 12, the calculated value is higher than the measured value, indicating that this may have happened. Another option for the error is the fact that there is a non-uniform flux density in the core and the shell causing the permeability to vary as a function of distance from the center of the device. This other cause of error has been eliminated because the permeance of the shell and the core was calculated by guessing the flux density at the core surface and numerically integrating the permeance from the midpoint of the absorber to the top and bottom of the device. By doing this, the change in permeability with flux density was taken into account, enhancing the accuracy of permeance in the core and the shell. The Moskowitz method shown in Table 12 is
Figure 39. Design MM theoretical to experimental RMS voltage generating results

Figure 40. Design MM experimental to theoretical error results
voltage source was taken and came back as an incorrect value. Despite knowing this deficiency in the data acquisition system, the values from the top accelerometer, since they were small relative to the bottom, were taken to give an approximate value of the relative displacement. Figure 40 shows that there is a ±10% discrepancy near 50 Hz. This comes exciting a the test stand resonant mode. At these frequencies, the amplitude of the shaker is small (< 1 mm zero to peak) due to the increase of acceleration needed to sustain the vibration. With such a small driving amplitude, the top plate amplitude becomes significant contributing appreciably to the relative displacement. Subtracting the two outputs results in a relative amplitude that is more subject to error than if a single large output were been used with a stationary test stand.

Figure 41 is a graph of the voltage measurements from Design MW. The experimental and theoretical results are appreciably off. One reason for the lack of correlation could come from the equipment giving incorrect measurement values. The experimental equipment has been checked to see if this is the cause for the poor correlation. The equipment was found to be operating correctly, eliminating this as a viable source of explaining the error. Another reason for the lack of correlation could come from the absorber being accidentally grounded to the Norland. After checking the system with an ohm meter, it was found that the absorber was not grounded to the Norland; eliminating another another possible source of the error present. Exhausting the possible options for a reason for this error, solicits the statement that the lack of agreement is unknown. Although from Figure 42 there is some hope of an explanation for the lack of agreement. This plot looks suspiciously similar to the increase in resistance of the solenoid. From theoretical considerations, the resistance of the absorber should not change any of the voltage measurements. The reason for this is that the open circuit voltage was taken. This was done by connecting a 970 kΩ Norland machine in series with the absorber. The theoretical model should hold regardless of what type of impedance the absorber has but, there may be some reason why the theoretical model is incorrect.

What confuses the issue more is another test was done on Design MW where the bottom plate was detached from the absorber. Shown in Figure 43 is the original results of the bottom plate on adjusted relative to the test done with the bottom plate off. The voltage generated by the absorber with the bottom plate off is higher than if it were on. This is totally backwards from what should be observed. It was hypothesized that the flux density is higher with the bottom off (somehow) thus creating more voltage. A flux test was done on the absorber with its bottom off and it was found that the flux density was lower (as expected) than the bottom
Figure 41. Design MW theoretical to experimental RMS voltage generating results

Figure 42. Design MW experimental to theoretical error results
Figure 43. Relative voltage generation with the bottom plate on and off Design MW on. This result eliminates the possibility of larger flux density creating more voltage. As for explaining the phenomenon, there are no other explanations that can be explored at this time. The reason is that the two other variables that are a function of generated voltage, frequency and the length of the wire, are the essentially constant for both the bottom plate on or off.

Damping Measurements

Measurements to illustrate that the EVA provides damping to a mechanical system were conducted. Design MW was used for this test with its bottom off. The reason for using this design as opposed to Design MM was because of the friction in Design MM was large. The friction prevented motion of the magnetic mass when it was attached to low energy vibrations from a transient responding spring-mass system. The bottom of Design MW was left off because it provided more damping than when it was on. The spring-mass-EVA damper system that was tested is depicted in Figure 44. The damping was found by pulling the mass towards the EVA and then letting it go with the absorber's circuit opened and closed. The mass motion would then be detected by the accelerometer located on the mass and recorded by a LabView data acquisition system. The open circuit results are shown in Figure 45 and the closed circuit
results are shown in Figure 46. The initial damping of the system in Figure 45 is from air and friction in the system. To give an equivalent viscous damping value for the open circuit, the damping ratio was found to be $\zeta \approx 11\%$. This value was found by taking the logarithmic decrement of the first two major humps in the response curve. Damping can be seen to occur after the circuit was closed in Figure 46. The resulting damping ratio increased to a value of $\approx 23\%$.

![Damping experimental apparatus](image)

**Figure 44.** Damping experimental apparatus

![Graph showing damping with circuit open](image)

**Figure 45.** Design MW damping with circuit open
Figure 46. Design MW damping with circuit closed
CONCLUSIONS AND RECOMMENDATIONS

This section offers the conclusions in this study and sums up the objectives that were reached. Following the conclusions, are the recommendations for further study and suggestions about pitfalls uncovered in the development of the vibration absorbers.

The resistance and reactance of the air solenoid of Design MM were close to being linear as expected. The linearity of these two quantities suddenly became skewed after assembly of the magnet and the soft magnetic material with the solenoid. The reactance increased as expected since highly permeable material was introduced inside and outside the solenoid. What was surprising was that the resistance dramatically increased with frequency. There were two possible causes for this phenomenon to occur. The first was that the resistance increased because of electromagnetic interactions between the solenoid and the flux from the magnet. The other was that eddy currents developed in the steel components produced a counter-force against the current induced in the wire resulting in an increase in resistance. It was found that the magnetic flux produced by the magnet did not increase either the resistance or the reactance. After analyzing the behavior of the air solenoid of Design MW, the answer was found. The increase in the resistance in the air solenoid in Design MW was much greater than Design MM. The reason for this was because the wire was wrapped around a good electrically conducting aluminum cylinder. The time varying flux in the wire produced eddy currents that opposed the current in the coil resulting in an increase in resistance. It was deduced that the resistance of both designs increased after assembly because eddy currents in the steel were induced.

The increase in the resistance with frequency in both designs significantly reduced the potential for electrodynamic damping frequencies greater than 5 Hz. The damping potential was limited because the resistance cannot be changed by adding any electrical components to the circuit to reduce it. The increase in resistance also lowered the damping to mass ratio to such an extent that already existing eddy current dampers could be used with much more effectiveness than electrodynamic damping.

The initial theoretical versus experimental flux density values had large discrepancies for Design MM. These discrepancies were hypothesized to be coming from two sources. The first source was that an unspecified soft magnetic material could have been used. The second source was that the magnetic materials used may not have met specifications. It was found, that there was an unspecified material used in the core of Design MM. This finding resulted in a 50% decrease of error in the finite permeance magnetic circuit modelling method values. A flux testing device was constructed to check the $B_r$ values of both magnets used in the vibration
absorbers. It was found that the magnetic material in Design MM was not the material specified in the design. Incorporating this new information into the theoretical calculations for Design MM improved the error to within 4% of the experimental value. Due to this success in predicting the magnetic circuit behavior of Design MM and the lack of predictability of the Moskowitz infinite permeability method, the finite permeability method was found to be the best method for predicting the magnetic circuit behavior of the vibration absorbers. It was deduced that the failure of the Moskowitz infinite pole piece permeance method to predict the magnetic circuit behavior of both absorbers was due to its inability to take into account the length of the pole pieces. The experimental to theoretical values of Design MW were found to be in good agreement with the measured values using the finite pole piece permeance method. The Moskowitz method had an error larger than the finite method but predicted the behavior of the magnetic circuit with much less error than that for Design MM. The reason for this was probably due to the simplicity of the circuit, shorter pole piece length and the use of low carbon steels.

Good correlation was found between the voltage generating experimental results and the theoretical calculations for Design MM. There was, however, some large error found in these results below 5 Hz. It was deduced that this error came from the shaker and vibrations from the test stand that interfered with producing a pure sinusoid. The reason for this error was that the signal from the accelerometer contained multiple frequency components. The resulting accelerometer RMS voltage value was reported higher than the true value. This larger voltage output corresponded to a larger resultant amplitude that caused the theoretical calculations to be higher than the true value, thus causing the discrepancy between theoretical and experimental results below 5 Hz.

The Design MW voltage generating experimental to theoretical calculations were in poor agreement. There were two possible causes for the discrepancy. The first was that the experimental equipment was not reporting correct values, and the second was that the vibration absorber may have been grounded to the data acquisition system. Both possibilities were checked and it was found that the equipment was working properly and the vibration absorber was not grounded to the data acquisition system. It was also found that the voltage generated was larger when the bottom plate was off Design MW. The flux density was hypothesized to be larger with the bottom off Design MW causing the increase. After testing the flux density of Design MW with the bottom plate off, it was found that it was less than when the bottom plate was on eliminating the hypothesis as a possible source of error. The reason for lack of correlation and for the increase in voltage generation when the bottom plate is off, to date,
remains a mystery.

The Electrodynamic Vibration Absorber has been a worthwhile project to pursue. The information that was obtained in this study is useful to a certain extent. The extent of this information is to advise those who are attempting to pursue a similar scheme of damping to consider the pitfalls discovered by this research. These are the resistance of a solenoid does increase with frequency and that an appreciable amount of weight has to be added to the absorber to transport the flux. These considerations reduced the maximum damping coefficient and the damping to mass ratio to such an extent that electrodynamic damping potential is very limited. It is suggested here that current eddy current damping technology could be used with much more effectiveness. It is recommended that this type of electrodynamic damping could be effective, however, if field windings were used to create a dense flux without the need to use pole pieces. Doing this would diminish the eddy current 'resistance increase' effect and the weight of the device itself would be reduced thus eliminating the pitfalls of this type of damping. It should be pointed out that to create a magnetic field flux using windings, energy would have to supplied from an outside source thus eliminating the property of being a passive damper.

Most all of the electrodynamic damping phenomenon has be discussed and explained in this document. As was noted at the end of the results and discussion section, an explanation has yet to be found for the large discrepancy between the experimental and theoretical voltage values in Design MW and the lack of magnetic circuit correlation. It is recommended that an explanation for this be pursued because valuable information may be found in the process. To date, an electrical engineering graduate student is modelling both vibration absorber designs and may be able to explain the discrepancy sometime in the future.

There has been one pitfall that has been indirectly addressed in this document. This is the difficulty of finding good steel and magnet vendors. I found it to be very disturbing that all of the local steel vendors that I dealt with did not know what kind of steel they were selling. The parties include the ERI machine shop who gave me the highest rate of confidence in what I was buying but for an university establishment I found it strange they didn't 'know' exactly what they were selling me. The magnet vendors that I dealt with apparently had similar problems as the steel vendors. It seemed to me that if you were a small order, you were small time. I understand that in the real world people need to make a 'buck' but it does not justify the time I spent building a flux testing device, conducting hardness tests and hysteresis loops to confirm that the products that I received were not what I had specified. My advice to those people out there that are looking for specific materials is to beware of the vendors with whom they deal.
As a side note to this, I found the quickest way to check if a vendor supplied a specified metal was to hardness test it.
REFERENCES


14. The methodology of solving this equation was first developed by Dave Hall.


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I would like to thank both the College of Engineering and the University Research Institute for their personal and project financial support. In addition, I would like to thank the Iowa Space Consortium for the personal financial support and the trip to a vibration conference a year ago February (San Diego in the middle of winter was nice!).

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I would also like to thank two individuals who had a significant impact on this research. Mike Devine of the Ames Lab has always been there to help me with any magnetic measurements that I needed. If it wasn't for the assistance I had from Mike, this project would have been a complete and utter wash out. So thank you Mike and good luck in your pursuit in becoming Dr. Devine. The second individual I would like to thank is Dave Hall. Dave is a man of a million questions and suggestions. Some of those questions and suggestions had a significant impact on this project. One of them was obvious, he came up with the methodology of modelling the transient response of a vibrating system. Another was the way of approaching analytical models. Dave taught me to 'always find the analytical solution if possible'. So to Dave thank you (sometimes) for your million questions and suggestions, and for the analytical approach philosophy that I now have for modelling systems.

My thanks go to several other individuals that aided me in this project. I would like to thank Hitendra Patel (H.P.) for allowing me to use his Impedance Analyzer and for the day of his time it took to run the tests. I wish to thank Joe Schlesselman also, for his many helpful suggestions and work on getting the data acquisition system ready for my tests. I would also like to thank Gay Scandrett, Jim Deutremont and Tom Elliot for their patience and assistance throughout the project.

Special thanks goes toward Jim Roozee at Arnold Engineering for all of the information on magnets and for the free Ceramic magnet samples that he gave to me for use in the project. Its nice to know that there is at least one competent person/company in the magnetics industry who treats graduate students with respect.

Lastly, I would like to thank the people in my life that have been very supportive of me throughout my college career. These are my parents Tom and Annette, grandparents Margaret
Hansen and Ruth Black, and my sister Stacy. It's through their wisdom that I have been able to go this far in school and through their support that has been my backbone each day.
APPENDIX A: COMPUTER PROGRAM FOR DESIGN MM

*Design MM.f - This program is used to provide information about
*the designed electromagnetic plunger it gives the infinite pole
*piece prediction and the finite pole piece prediction.
*
*Variable List.
*a = How many turns per inch layer for a particular gage of wire.
*b = Dummy variable
*d = The diameter of a particular gage of wire.
*e = Young's Modulus for the hardened Thompson shafts.
*h = Dummy variable
*j = Dummy variable
*l = Magnet length
*m = Absorber mass
*n = # of turns
*p = Dummy variable
*q = Dummy variable
*r = Core radius
*s = Soft magnetic cylinder between magnet and core.
*t = Magnet thickness
*v = Maximum absorber r.m.s. voltage.
*w = Circular Frequency
*x = The maximum displacement.
*ba = Outer cylinder flux
*bc = The flux density of the core.
*bd = The outside magnet flux.
*bi = The limiting flux value inside the magnets.
*bl = The product of cutting flux and length of wire influenced.
*bs = The flux density of the shell.
*bx = Flux at maximum energy product.
*c8,c9 & 117 are used to calculate the bending stress in the aluminum plate.
*hd = The outside magnet coercive force.
*hi = The inside magnet coercive force.
*hx = Coercive force at maximum energy product.
*h1 = Dummy variable
*fn = Absorber natural frequency
*lg = Air gap length
*lk = Distance between pole pieces.
*l2 = Total length of wire (m).
*l3 = Length of the solenoid (m).
*lr = Inside core return path length.
*mb = Maximum bending moment applied to the aluminum plate.
*mc = Core permeability at a given flux density.
*ms = Shell permeability.
*mt = Top & Bottom Plate permeability
*mu = The relative permeability factor.
*mx = The maximum mass load to achieve a given damping.
*pa = Connection across shell piece permeance.
*pb = Bottom plate permeance
*pc = Core permeance
*pd = Connection across core piece permeance.
*pf = Fringing permeance
*pg = Air gap permeance
*ph = Shell permeance
*pl = Leakage permeance
*po = The pressure exerted by the mag. material and inertia.
*pr = Inside core return permeance
*ps = The poison's ratio for aluminum.
*pt = Top plate permeance
*rd = Dummy variable
*r1 = Resistance of all windings.
*tb = Bottom plate thickness
*tt = Top plate thickness
*tp = The aluminum top or bottom thickness.
*wt = The weight of the absorber.
*xm = The maximum amplitude the absorber can handle.
*z1 & z2 = Circuit impedance magnitude.
*ans = Dummy variable
*blr = The bl product squared divided by the wire resistance.
*bmn = The flux density half way into the windings.
*cap = The capacitance needed to achieve circuit resonance.
*cur = Current generated with inductance included.
*pow = Power generated with inductance included.
*del = The thickness of the steel shell surrounding the magnet.
*dia = The diameter of the solenoid (m).
*ind = Inductance caused by the coil.
*imx = The maximum instantaneous current allowed by the absorber.
*len = Solenoid length (in.)
*lay = The number of layers of copper wire.
*frb = Inside core bearing return path length.
*mag = The type of hard magnetic material used to produce flux.
*cr = Critical load of the rods.
*pbt = The total permeance in the air gap.
*prb = Bearing return permeance
*pr1 = Dummy variable
*pr2 = Dummy variable
*res = The required resistance needed for a given damping ratio.
*srt = The maximum tensile stress in the rods.
*sid = Shell inside diameter.
*sod = Shell outside diameter.
*g16, g18, g20 & g22 = Wire resistance of specific gages of wire.
*sg16, sg18, sg20 & sg22 = Wire weight/length
*vol = The volume of soft magnetic material.
*vol2 = The volume of the magnetic material.
*vol3 = The volume of the soft magnetic inner ring.
*drod = The hardened shaft diameter.
*fmag = The maximum instantaneous magnetic force.
*fmax = The total combined force from both fmag and the inertia force.
*gage = Dummy variable
*inrt = The inertial of the rods.
*lrod = The hardened shaft length.
*mag1 = Dummy variable
*imax = The max. r.m.s. current that can be achieved at given conditions.
*pmax = The max. r.m.s. power that can be achieved at given conditions.
*ptot = Total magnetic circuit permeance
*sgmt = The maximum tensile stress in the moving mass.
*wall = The wall thickness of the material that holds the wire into place.
*zeta = The average damping ratio.
*finrt = Inertia force
*k safe = The buckling safety factor of the rods.
*iloss = The % current loss due to the inductance.
*ploss = The % power loss due to the inductance.
*romag = Density of neoflex
*rostel = Steel density
*sgbnd = Maximum bending stress
*
integer i, k, lay
real a, c, d, g, h, j, l, m, n, p, r, s, t, w, x, y
real ba, bc, bd, bi, br, bs, bx, hd, hi, hx, hl, lg, l2, l3, mu
real lr, ir, mc, ms, mt, pa, pb, pc, pd, pf, pg, ph, pi, pl, pr
real pt, tb, tt, g16, g18, g20, g22, blr, bmn, sid, sod, vol2
real bl, lw, xm, del, dia, ind, len, wall, zeta, rostel, romag
real r1, wt, vol, pr1, pr2, prb, pgt, sg16, sg18, sg20, sg22, vol3
real rd, lr, ptot, cap(20), fn(20), res(20), imax(20), pmax(20)
real v(20), mx(20), cur(20), pow(20), iloss(20), ploss(20)
real imaxm(20), pmaxm(20), vm(20), curm(20), powm(20)
real ilossm(20), plossm(20), z1(20), z2(20)
real e, c8, c9, 117, mb, po, ps, tp, fmag, fmax, lrod
real imx, pcr, sgt, sgm, droid, irnt, ksafe, finrt, sgbnd
character*2 ans, gage
character*15 mag
character*10 mag1
*

k = 0
pi = 3.1415926
*

*Wire resistance per meter
 g16 = .01318
 g18 = .02095
 g20 = .03324
 g22 = .05315

*Wire weight (lb.) per meter
 sg16 = .02610
 sg18 = .01647
 sg20 = .01042
 sg22 = .006539

*Steel and Neoflex Density (lb./in^3)
 romag = .215
 rostel = .284

*Mechanical Properties of aluminum
ps = .33

*Mechanical Properties of the Steel Rods.
e = 30.0e6

*

Print*, 'This is the EM damper dimension optimization program.'

*

*Loading the magnetic properties of the magnet to be used.

*

Print*, 'What is the magnetic material?'
read(*,'(a10)') mag
if (mag .eq. 'ceramic 5') then
    br = 3800
    bx = 1900
    hx = 1800
else if (mag .eq. 'ceramic 7') then
    br = 3400
    bx = 1700
    hx = 1700
else if (mag .eq. 'ceramic 8') then
    br = 3850
    bx = 1900
    hx = 1900
else if (mag .eq. 'ceramic 9') then
    br = 3800
    bx = 1900
    hx = 1900
else if (mag .eq. 'ceramic 10') then
    br = 4100
    bx = 2000
    hx = 2000
else if (mag .eq. 'hf2') then
    br = 2450
    bx = 1225
    hx = 1225
else if (mag .eq. 'hf3') then
    br = 2650
    bx = 1325
    hx = 1325
else if (mag .eq. 'pm80') then
    br = 6200
    bx = 3100
    hx = 2600
else if (mag .eq. 'neoflex') then
    br = 5450
    bx = 2725
    hx = 2450
else
    Print*, 'What is Br?'
    read*, br
    Print*, 'What is Bx?'
    read*, bx
Print*, 'What is Hx?'
read*, hx
end if

Print*, 'What is the material of the soft magnetic core?'
read(*',(a12))', mag
Print*, 'What is the permeability of the core?'
read*, mc
Print*, 'What is the permeability of the shell?'
read*, ms
Print*, 'What is the permeability of the top & bottom?'
read*, mt

*Asking for design conditions.
*
1 Print*, 'What is the radius of the core (in.)?'
read*, r
Print*, 'What is the ring thickness (in.)?'
read*, s
rd = r+s
Print*, 'What is the magnet thickness (in.)?'
read*, t
Print*, 'What is the magnet outer shell thickness (in.)?'
read*, del
Print*, 'What is the wire shell thickness (in.)?'
read*, wall
Print*, 'What is the shell inside diameter (in.)?'
read*, sid
Print*, 'What is the shell outside diameter (in.)?'
read*, sod
Print*, 'What is the average inner core return path length (in.)?'
read*, lr
Print*, 'What is the average bearing return path length (in.)?'
read*, lrb
Print*, 'What is the av. gap length between pole pieces (in.)?'
read*, lk
Print*, 'What is the top plate thickness (in.)?'
read*, tt
Print*, 'What is the bottom plate thickness (in.)?'
read*, tb
Print*, 'What is the magnet height (in.)?'
read*, l
Print*, 'What is the bearing height (in.)?'
read*, y
Print*, 'What is the magnetic mass height (in.)?'
read*, lw
tp = (lw-l)/2.0
Print*, 'What is the diameter of the rods (in.)?'
read*, drod
Print*, 'What is the length of the rods (in.)?'
read*, lrod
Print*, 'How many rods?'
read*, nrod
Print*, 'What is the length of the solenoid (in.)?'
read*, len
I3 = .0254*len

* Asking about wire type and absorber parameters.

Print*, 'What is the wire gage (16,18,20 or 22)?'
read*, (a2) gage
Print 3, 'How many layers of', gage, ', gage wire?'
format (a,1x,a,1x,a)
read*, lay
Print*, 'What is the desired damping ratio?'
read*, zeta
Print*, 'What is the electrical permeability of the core?'
read*, mu

if (gage .eq. '16') then
  a = 18.6
  d = .0539
  g = sg16
  n = a*lay*len
  lg = d*(1.0+sqrt(3.0)*(lay-1.0)/2.0) + wall
else if (gage .eq. '18') then
  a = 23.2
  d = .0431
  g = sg18
  n = a*lay*len
  lg = d*(1.0+sqrt(3.0)*(lay-1.0)/2.0) + wall
else if (gage .eq. '20') then
  a = 28.9
  d = .0346
  g = sg20
  n = a*lay*len
  g = d*(1.0+sqrt(3.0)*(lay-1.0)/2.0) + wall
else
  a = 36.2
  d = .0278
  g = sg22
  n = a*lay*len
  lg = d*(1.0+sqrt(3.0)*(lay-1.0)/2.0) + wall
end if

* Calculating the optimum conditions for a given radius and length.
* This procedure uses the equations found on pp. 30 - 50 in the new
* lab book. Note: bi < br which satisfies the constraint on P.123
* in the old lab notebook.

p = (br-bx)/hx
I2 = 0.0
\[ lg = \frac{sid}{2.0} - (rd + t + del) \]

\[
\begin{align*}
p_a &= (\pi/4.0*(sod**2-sid**2))/lk \\
p_b &= 2.0*\pi*ib*mt/(\log((\pi+sid)/(4.0*r))) \\
p_c &= (mc*\pi*r**2)/(\text{len}/2.0) \\
p_d &= (\pi*r**2)/lk \\
p_f &= 3.32*(rd+t+del+lg/2.0) \\
p_g &= 2.0*\pi*l/(\log(1.0+lg/(rd+t+del))) \\
p_h &= (ms*\pi/4.0*(sod**2-sid**2))/(\text{len}/2.0) \\
p_l &= 1.66*(rd+t/2) \\
p_r &= 2.0*\pi*(l-y)/\log(1.0+lr/r) \\
p_{rb} &= 2.0*\pi*y/\log(1.0+lr/r) \\
p_t &= 2.0*\pi*tt*mt/(\log((\pi+sid)/(4.0*r))) \\
p_{r1} &= 1.0/(1.0/ph+1.0/pa+1.0/pb+1.0/pd+1.0/pc) \\
p_{r2} &= 1.0/(1.0/ph+1.0/pa+1.0/pt+1.0/pd+1.0/pc) \\
p_{gt} &= pg+2.0*pf+2.0*pl \\
\end{align*}
\]

if (k.eq.1) then
    \[
p_{r1} = 1.0e5 \\
p_{r2} = 1.0e5
\]
end if

\[
ptot = 2.0*pl+1.0/(1.0/(pg+2.0*pf)+1.0/(pr1+pr2)+1.0/(pr+prb))
\]

\[
\begin{align*}
b_d &= br*\pi/((rd+t)*\log(1.0+t/rd)+p*(2.0*\pi*l/ptot)) \\
b_c &= ((rd+t)*l/r**2)*((pg+2.0*pf)/pgt)*bd \\
b_s &= (4.0*(rd+t)*l/(sod**2-sid**2))*((pg+2.0*pf)/pgt)*bd \\
b_i &= bd*(rd+t)/rd \\
h_d &= (br-bd)/p \\
h_i &= (br-bi)/p \\
h &= rd+t+del \\
h_1 &= rd+t \\
b_a &= bd*(rd+t)*(pg/pgt)/(rd+t+del+lg) \\
bmn &= bd*(rd+t)*(pg/pgt)/(rd+t+del+wall+lg-wa)/2.0 \\
dia &= 0.254*(2.0*(rd+t+del+wall+lg-wa)/2.0)
\end{align*}
\]

*Finding the \( b \times l \) factor and the total absorber wire length.

\[
bl = 2.0*\pi*a*lay*(bd*1.0e-4)*(r+s+t)*(0.254)* \\
+((pg**1+pf**1*lg)/(pg+2.0*pf+2.0*pl)
\]

* do 6 i = 1, lay
    \[
    j = (d/2.0)*(1.0+sqrt(3.0))*(i-1)
    \]
    \[
    l2 = l2 + 2.0*\pi*0.254*a*len*(rd+t+del+wall+j)
\]

continue

* if (gage .eq. '16') then
    \[
l_1 = l2**g16
\]
else if (gage .eq. '18') then
    \[
l_1 = l2**g18
\]
else if (gage .eq. '20') then
    \[
l_1 = l2**g20
\]
else
r1 = l2*g22
end if

blr = bl**2/r1

*Finding the total weight of the absorber.
vol = pi*(len**2 + len*(sod**2-sid**2))/4.0 + r*sod**2/4.0
vol2 = pi*((r+s+t)**2 - (r+s)**2)*1
vol3 = pi*(((r+s)**2 - r**2)*1)
m = .453515*(rostel*vol3 + romag*vol2)
wt = rostel*(vol+vol3) + romag*vol2 + g*l2

*Computing the inductance.
c = (1.256637e-6*(n**2)*(pi/4.0)*dia**2)/(13+.45*dia)
ind = mu*c

*Computing the maximum amplitude.
x = (l3-.0254*lw)/2.0
xm = (len-lw)/2.0

*Calculating the electrical properties for a given damping ratio.
do 95 i = 1, 20
   fn(i) = real(i)
   w = 2.0*pi*fn(i)
   mx(i) = (bl)**2/(4.0*pi*r1*fn(i)*zeta) - m
   if (mx(i) .le. 0) mx(i) = 0.00
   res(i) = (bl)**2/(4.0*pi*zeta*m*fn(i))
   imax(i) = .70710678*(bl*x*w/res(i))
   pmax(i) = res(i)*(imax(i))**2
   v(i) = imax(i)*res(i)
   if (res(i) .lt. r1) then
      res(i) = 0.0000
      imax(i) = 1e9
      pmax(i) = 1e9
   end if
*Computing the maximum r.m.s. current values with the inductance.
z1(i) = sqrt((res(i))**2+(ind*w)**2)
cur(i) = v(i)/z1(i)
pow(i) = res(i)*(cur(i))**2
iloss(i) = (1.0-(cur(i)/imax(i)))*100.0
plloss(i) = (1.0-(pow(i)/pmax(i)))*100.0
*Calculating the capacitance required for circuit resonance.
cap(i) = 1.0e6/(ind*(2.0*pi*fn(i))**2)
*
*Calculating the maximum peak voltage, current and power
*assuming the resistance in the circuit is from the wire only.
v(i) = bl*x*w
iimaxm(i) = vm(i)/r1
pimaxm(i) = r1*(iimaxm(i))**2
*Computing the maximum peak current values with the inductance.
z2(i) = sqrt(r1**2+(ind*w)**2)
curm(i) = vm(i)/z2(i)  
powm(i) = r1*(curm(i))**2  
13  !lossm(i) = (1.0-(curm(i)/imxm(i)))*100.0  
  plossm(i) = (1.0-(powm(i)/pmaxm(i)))*100.0
95  continue

*Doing stress analysis on the moving mass. The moving mass is
*assumed to have an aluminum plate on top and the rods are Thompson
*harden shafts.

s1 = rd  
s2 = rd+t+del  
if (imxm(20) .gt. 25.0) then  
  imx = 25.0
else  
  imx = imxm(20)
end if

fmag = (1.0/4.448)*imx*b1  
finrt = (1.0/4.448)*m*x*(2.0*pi*fn(20))**2  
fmax = sqrt(fmag**2+finrt**2)  
inrt = pi/64.0*(dod**4)  
sgt = fmax/(nrod*(pi/4.0)*dod**2)  
sgmt = fmax/(pi*((r+wall)**2-r**2))  
prc = (pi**2*e**inrt/irod**2)/((pi/4.0)*dod**2)  
ksafe = prc/sgt  
po = fmax/(pi*(s2**2-s1**2))  
c8 = .5*(1.0+ps+(1.0-ps)*(s1/s2)**2)  
c9 = (s1/s2)*((1.0+ps)/2.0*log(s2/s1)+(1.0-ps)/4.0*  
  + (1.0-(s1/s2)**2))  
117 = .25*(1.0-(1.0-ps)/4.0*(1.0-(s1/s2)**4)-  
  + (s1/s2)**2*(1.0+(1.0+ps)*log(s2/s1)))  
mb = -po*s2**2/c8*(c9/(2.0*s1*s2)*(s2**2-s1**2) - 117)  
sgbdn = 6.0*mb/t**2

*Printing the optimum design results.

*Print*  
Print 25, 'Magnetic Material,'mag,'Thickness,'t,'(in.),'  
  'Ring,'s,'(in.)  
25  format(a,1x,a,a,f6.4,1x,a,1x,a,f5.3,1x,a)

*Print*  
Print*, 'Soft magnetic core properties made of,'mag  
Print 30, 'Core Flux,'bc,'(G),'Core Radius,'r,'(in.),'  
  'Air Gap,'lg,'(in.)  
30  format(a,f6.0,1x,a,3x,a,f5.3,1x,a,3x,a,f6.4,1x,a)

*Print*  
Print*, 'Absorber Shell Dimensions (in.)'  
Print 35, 'Length','len','I.D.','sid','O.D.','sod','Weight','  
  wt','(Lbs.)'
format(a,f5.2,3x,a,1x,f6.3,3x,a,1x,f6.3,3x,a,f6.2,1x,a)
Print*

Print*, 'Wire Holding Shell Dimensions (in.)'
Print 37, 'Inside Radius',h,'Wall Thickness',wall
format(a,1x,f6.3,3x,a,f5.3)
Print*

Print*, 'Magnet Holding Shell Dimensions (in.)'
Print 39, 'Inside Radius',h1,'Wall Thickness',del
format(a,1x,f6.3,3x,a,f5.3)
Print*

Print*, 'Main Air Gap Flux Distribution (G)'
Print 40, 'Flux at inner shell',ba,
+    'Mean Flux',bmn,'Shell Flux Density',bs
format(a,f6.0,3x,a,f6.0,3x,a,f6.0)
Print*

Print*, 'Magnet Operating Properties'
Print 42, 'Bi',bi,'(G)',hi,'(Oe)',Bd,'bd','(G)',
+    'Hd',hd,'(Oe)
format(a,f6.0,1x,a,3x,a,f6.0,1x,a,3x,a,f6.0,1x,a,3x,a,f6.0,1x,a)
Print*

Print*, 'Absorber Mass Properties'
Print 44, 'Max. Amplitude',xm,'(in.)',
+    'Head Mass',m,'(kg)','Height',l,'(in.)
format(a,f5.2,1x,a,2x,a,f5.3,1x,a,2x,a,1x,f5.3,1x,a)
Print*

Print*, 'Wire Properties'
Print 45, 'Length',l2,'(m)','Turns',n,'Wire Resistance',
+    r1,'(ohms)
format(a,1x,f6.2,1x,a,3x,a,f6.1,3x,a,f5.2,1x,a)
Print*

Print*, 'Magnet Circuit Air Gap & Total Permeance (in.)'
Print 46, 'Air gap',pg,'Fringing',2.0*pf,'Leakage',2.0*pl,
+    'Total',pbt
format(a,f6.2,2x,a,f5.2,2x,a,f5.2,2x,a,f6.2)
Print*

Print*, 'Magnet Circuit Return Path Permeance (in.)'
Print 48, 'Shell',ph,'Across Top',pa,'Top',pt,'Bottom',pb
format(a,f6.1,2x,a,f6.1,2x,a,f6.1,2x,a,f6.1)
Print*
Print 49, 'Across Core',pd,'Core',pc,'Final Return',pr
format(a,f6.1,2x,a,f6.1,2x,a,f6.1)
Print*
Print*, 'Inductance Properties'
Print 50, 'Electrical Mu', mu, 'Inductance', 'ind', '(henry)'
format(a,f5.0,3x,a,f10.7,1x,a)
Print*
*
Print*, 'Damping Properties'
Print 55, 'Zeta', zeta, 'B*L Product', bl, '(T*m)',
+ '(B*L)^2/ R', blr, '((T*m)^2/ohm)'
format(a,f4.2,3x,a,f6.2,1x,a,2x,a,1x,f6.2,1x,a)
Print*
*
Print*, 'Magnetic Mass Stress Analysis (Maximum Values)'
Print 56, 'Max. Current', 'imx', '(amps)', 'Mag. Force', 'fmag',
+ ' (lbs)', 'Inertia Force', 'finrt', '(lbs)'
format(a,f5.2,a,2x,a,f6.2,a,2x,a,f6.2,a)
Print 58, 'Total Force', 'fmax', '(lbs)', 'Tensile Stress',
+ ' sgmt', '(psi)', 'Bending Stress', 'sgbnd', '(psi)'
format(a,f6.1,a,1x,a,f6.1,a,1x,a,f7.1,a)
Print*
*
Print*, 'Rod Stress Analysis'
Print 60, 'Rod Buckling Stress', 'pcr', '(psi)', 'Tensile Stress',
+ ' sgt', '(psi)', 'Safety Factor', 'ksafe'
format(a,f7.1,a,1x,a,f6.1,a,1x,a,f5.1)
Print*
*
Print 62, 'Electrical Properties (rms) at for a zeta of', zeta
format(a,f3.2)
Print*, 'freq. resist. icalc imax lloss pcalc pmax lloss'
+ ' vmax max. mass impedance'
Print*, '(hz) (ohms) (amps) (amps) (%) (W) (W) (%)',
+ ' (volts) (kg) (ohms)'
do 68 i = 1, 20
    Print 64, fn(i), res(i), cur(i), imax(i), lloss(i), pow(i), pmax(i),
+ lloss(i), v(i), mx(i), z1(i)
format(f3.0,2x,f7.2,2x,f5.2,3x,f5.2,2x,f5.2,1x,f6.1,1x,f6.1,
+ 2x,f5.2,3x,f6.2,4x,f6.2,5x,f6.2)
68 continue
Print*
*
Print*, 'Maximum peak voltage, current and power.'
Print*, 'freq. icalc imax lloss pcalc pmax lloss'
+ ' vmax impedance'
Print*, '(hz) (amps) (amps) (%) (W) (W) (%)',
+ ' (volts) (ohms)'
do 70 i = 1, 20
    Print 72, fn(i), curm(i), imaxm(i), llossm(i), powm(i), pmaxm(i),
+ llossm(i), vm(i), z2(i)
72 format(f3.0,3x,f5.2,3x,f5.2,3x,f5.2,3x,f5.2,3x,f6.1,2x,f6.1,
+ 1x,f6.2,2x,f6.2,3x,f6.2)
70 continue
Print*

Print*, 'See the required capacitance for circuit resonance?'
read(*,'(a1)') ans
if ((ans .eq. 'y') .or. (ans .eq. 'Y')) then
  Print*, 'freq capacitance (mcF)'
  do 76 i = 1, 20
      Print 74, fn(i), cap(i)
  continue
  format(2x,f3.0,6x,f10.2)
end if

if (k.eq.0) then
  Print*
  Print*, 'Moskowitz infinite pole piece assumption values.'
  k = k+1
  goto 5
end if

Print*, 'Do it again, same values but different magnet height?'
read(*,'(a1)') ans
if ((ans .eq. 'y') .or. (ans .eq. 'Y')) then
  Print*, 'What is the new magnet height (in.)?'
  read*, l
  Print*, 'What is the length of the solenoid (in.)?'
  read*, len
  l3 = .0254*len
  goto 5
end if

Print*, 'Do it again with different parameters?'
read(*,'(a1)') ans
if ((ans .eq. 'y') .or. (ans .eq. 'Y')) goto 1

Stop

End
APPENDIX B: COMPUTER PROGRAM FOR DESIGN MW

This section is devoted to give a glance at the original computer program to give the design parameters of Design MW. The reason for just giving a 'glance' instead of the whole program is due to redundancy between programs. The variables that were loaded, stress, electric circuit and weight analysis are essentially the same for the Design MW computer program as that for the computer program for Design MM. There are some unique property of the Design MW computer program. The permeance of the core and the shell were calculated by numerically integrating Eq. 35. This was done because the flux density starts at zero at the middle of the absorber and then increases with distance to a maximum value at the top and bottom of the absorber. Permeability, as it has been shown, is a function of flux density. What has been done was the permeability data of a 1018 Annealed 1020 Steel sample was loaded into a graphing software package. The graphing package fitted a ninth order polynomial function to the data. This function was loaded into the computer program and then used to calculate the permeabilities of both the shell and the core.

*Thomas T. Hansen
*
*Design MW.f - Program calculates both the finite and infinite pole piece magnetic circuits.
*Variable List.
*l = Magnet length
*n = # of turns
*r = Core radius
*t = Magnet thickness
*y = Dummy variable
*a1-a9 = Coefficients for the shell permeability equation.
*as = The shell cross-sectional area.
*ba = The flux at the core.
*bd = The operating flux of the magnet.
*bs = Operating soft magnetic material flux.
*hd = The operating coercive force of the magnet.
*lg = Air gap length
*lk = Distance between pole pieces.
*l2 = Total length of wire (m).
*l3 = Length of the wire solenoid (m).
*ms = Shell permeability.
*mt = Top & Bottom Plate permeability
*pa = Connection across shell piece permeance.
*pb = Bottom plate permeance
*pc = Core permeance
*pd = Connection across core piece permeance.
*pg = Air gap permeance
*ph = Shell permeance
integer i, lay
   a, c, d, g, j, l, m, n, p, r, s, t, w, te
real bc, bd, bl, bs, dl, hd, lg, l2, l3, rc
real dt, lk, lw, mt, pa, pb, pc, pd, pg, ph, pt, tb, tt
real g16, g18, g20, g22, blr, bmn, sid, sod, tol
real rs, dia, ind, len, wall
real mu, pi, r1, pr1, pr2, bout, hout, len2
double precision a0, a1, a2, a3, a4, a5, a6, a7, a8, a9
double precision k, q, y, as, ac, ba, bo, dy
character*2 ans, gage
character*10 mag
character*15 mag1

a0 = 229.48467104
a1 = 0.83062316038
a2 = -0.00032376115322
a3 = 1.4499493037d-07
a4 = -4.6992047943d-11
a5 = 8.5739991535d-15
a6 = -8.9269804112d-19
a7 = 5.3100089428d-23
a8 = -1.6834870082d-27
a9 = 2.2093278056d-32

Print*, 'What is the material of the soft magnetic core?'
read*(*,(a12)*) mag1
Print*, 'What is the permeability of the top/bottom?'
read*, mt
Print*, 'What is the guessed core flux, ba?'
read*, ba
Print*, 'What is the radius of the core?'
read*, r
Print*, 'What is the wire shell thickness (in.)?'
read*, wall
Print*, 'What is the ferromagnetic shell thickness (in.)?'
read*, dl
Print*, 'What is the tolerance between the shells (in.)?'
read*, tol
Print*, 'What is the magnet thickness (in.)?'
read*, t
Print*, 'What is the magnet coating thickness (in.)?'
read*, dt
Print*, 'How many layers of coating?'
read*, lay
dt = lay*dt
Print*, 'What is the wire zone height (in.)?'
read*, l
l3 = .0254*l
Print*, 'What is the moving mass height (in.)?'
read*, lw
tp = (lw-l)/2.0
Print*, 'What is the top plate thickness (in.)?'
read*, tt
Print*, 'What is the bottom plate thickness (in.)?'
read*, tb
Print*, 'What is the end repulsing magnet thickness (in.)?'
read*, te
Print*, 'What is the av. gap length between pole pieces (in.)?'
read*, lk
Print*, 'What is the diameter of the rods (in.)?'
read*, drod
Print*, 'What is the length of the rods (in.)?'
read*, lrod
Print*, 'How many rods?'
read*, nrod
Print*, 'What is the length of the solenoid (in.)?'
read*, len
len2 = len + 2.0*te

* Asking about wire type and absorber parameters.
Print*, 'What is the wire gage?'
read*(,'(a2)') gage
Print 3, 'How many layers of', gage, ', gage wire?'
format (a,1x,a,1x,a)
read*, lay
Print*, 'What is the desired damping ratio?'
read*, zeta
Print*, 'What is the absorber mass (kg)?'
read*, m
Print*, 'What is the electrical permeability?'
read*, mu
Calculating the optimum conditions by iterating bd.

\[
\begin{align*}
  s &= r + lg \\
  bo &= ba * r / (r + lg + dl + t + dt) \\
  sid &= 2.0 * (r + lg + dl + t + dt) \\
  sod &= \sqrt{(2.0 * r)^{2} + sid^{2}} \\
  k &= 4.0 * ba * r * len / (sod^{2} - sid^{2}) \\
  q &= 4.0 * bo * sid / (sod^{2} - sid^{2}) \\
  pa &= (pi / 4.0 * (sod^{2} - sid^{2})) / lk \\
  pb &= 2.0 * pi * tb * mt / (log((sid + sod) / (4.0 * r))) \\
  y &= 0.0 \\
  rc &= 0.0 \\
  rs &= 0.0 \\
  ac &= pi * r^{2} \\
  as &= pi / 4.0 * (sod^{2} - sid^{2}) \\
  dy &= len / 2000 \\
  y &= y + dy \\
  rc &= ac * (a0 + a1 * (k - q * y) + a2 * (k - q * y)^{2} + \\
       a3 * (k - q * y)^{3} + a4 * (k - q * y)^{4} + a5 * (k - q * y)^{5} + \\
       a6 * (k - q * y)^{6} + a7 * (k - q * y)^{7} + a8 * (k - q * y)^{8} + \\
       a9 * (k - q * y)^{9}) / (as * (a0 + a1 * (k - q * y) + a2 * (k - q * y)^{2} + \\
       a3 * (k - q * y)^{3} + a4 * (k - q * y)^{4} + a5 * (k - q * y)^{5} + \\
       a6 * (k - q * y)^{6} + a7 * (k - q * y)^{7} + a8 * (k - q * y)^{8} + \\
       a9 * (k - q * y)^{9})) \\
  rs &= dy / (as * (a0 + a1 * (k - q * y) + a2 * (k - q * y)^{2} + \\
       a3 * (k - q * y)^{3} + a4 * (k - q * y)^{4} + a5 * (k - q * y)^{5} + \\
       a6 * (k - q * y)^{6} + a7 * (k - q * y)^{7} + a8 * (k - q * y)^{8} + \\
       a9 * (k - q * y)^{9})) \\
  pc &= 1.0 / rc \\
  ph &= 1.0 / rs \\
  pd &= (pi * r^{2}) / lk \\
  pg &= 2.0 * pi * len / log(1.0 + lg / r) \\
  pt &= 2.0 * pi * tt * mt / (log((sid + sod) / (4.0 * r))) \\
  pr1 &= 1.0 / (1.0 / ph + 1.0 / pa + 1.0 / pb + 1.0 / pd + 1.0 / pc) \\
  pr2 &= 1.0 / (1.0 / ph + 1.0 / pa + 1.0 / pt + 1.0 / pd + 1.0 / pc) \\
  if (k1, eq, 1) then \\
    pr1 &= 1.0 e5 \\
    pr2 &= 1.0 e5 \\
  end if \\
  ptot &= 1.0 / (1.0 / pg + 1.0 / (pr1 + pr2)) \\
  bd &= br * t / (r + lg + dl) * (log(1.0 + (t + dt / 2.0) / (r + lg + dl)) + \\
       p * 2.0 * pi * len / p + ptot)) \\
  hd &= (br - bd) / p \\
  ba &= bd * (1.0 + lg / r) \\
    bc &= ba * (len / t) \\
  bs &= 4.0 * r^{2} * len * ba / (sod^{2} - sid^{2}) \\
  bmn &= bd * (2.0 + lg / r) / 2.0 \\
  dia &= 0.0254 * (2.0 * (r + wall + (lg - (wall + tol)) / 2.0)) \\
  bout &= ba * r / (r + lg + dl + t + dt)
\end{align*}
\]
hout = (br-bout)/p

* 12 = 0.0
  do 5 i = 1, lay
    j = (d/2.0)*(1.0+sqrt(3.0))*(i-1))
    12 = 12 + 2.0*pi*(.0254)*a*i*(r+wall+j)
  5 continue

* Finding the b*l factor and the total absorber wire length.
  bl = 2.0*pi*(r*.0254)*a*lay*(ba*1.0e-4)
  blr = bl**2/r1

* Printing the optimum design results.

* Print*, 'Soft magnetic core properties made of:',mag1
Print 30, 'Core Flux','bc', '(G)', 'Core Radius', 'r', '(in.)',
+ 'Air Gap', 'lg', '(in.)'
30 format(a,f6.0,1x,a,3x,a,f5.3,1x,a,3x,a,a,f6.4,1x,a)
Print*

* Print*, 'Magnet Holding Shell Dimensions (in.)'
Print 37, 'Inside Radius', 's', 'Tolerance', 'tol'
37 format(a,1x,f6.3,3x,a,f5.3)
Print*

* Print*, 'Main Air Gap Flux Distribution (G)'
Print 40, 'Flux at core', 'ba',
+ 'Mean Flux', 'bmn', 'Shell Flux Density', 'bs'
40 format(a,f6.0,3x,a,f6.0,3x,a,f6.0)
Print*

* Print*, 'Magnet Circuit Air Gap & Total Permeance (in.)'
Print 46, 'Air gap', 'pg', 'Total', 'ptot'
46 format(a,f6.2,3x,a,f6.2)
Print*

* Print*, 'Magnet Circuit Return Path Permeance (in.)'
Print 48, 'Shell', 'ph', 'Across Top', 'pa', 'Top', 'pt', 'Bottom', 'pb'
48 format(a,f6.1,2x,a,f6.1,2x,a,f6.1,2x,a,f6.1)
Print*
Print 49, 'Across Core', 'pd', 'Core', 'pc'
49 format(a,f6.1,2x,a,f6.1,2x,a,f6.1)
Print*

* Print*, 'Damping Properties'
Print 55, 'Zeta', 'zeta', 'B*L Product', 'bl', '(T*m)',
+ '(B*L)^2/ R', 'blr', '((T*m)^2/ohm)'
55 format(a,f4.2,3x,a,f6.2,1x,a,2x,a,1x,f6.2,1x,a)

Stop
End
APPENDIX C: WIRE SELECTION PROGRAM

The following program was made to find the optimum wire for the vibration absorber. This program reads input from the user and calculates a range of damping coefficients for each gage of wire. To get an equivalent amount of wire to make an appropriate comparison, the dimensions of 14 gage wire were used to determine the size of the air gap. The other smaller gages of wire were 'fit' in this air gap by determining the maximum integer number of turns that could exist in the air gap created by Ly layers of 14 gage wire. The values in Table 12 show that 14 gage wire is the most optimum to use in the EVA. A constant flux value of 3000 G was used in all of the calculations for the use of comparison only. Assuming a constant value is valid because the air gap is essentially the same for all wires used in the comparison.

*Thomas T. Hansen
*
*C Factor.f - Is a program that calculates the (bl)^2/r factor for wire.
*
*Variable List.
*a = How many turns per meter layer for a particular gage of wire.
*b = Flux density
*d = The diameter of a particular gage of wire.
*m = Dummy variable
*r = Dummy variable
*s = Solenoid length (in.)
*cc = Dummy variable
*ro = Initial radius
*lay = Number of layers of 14 gage wire
*
    integer i, j, k, l(100), lay
    real a(10), c(100,100), d(10), g(10), l(100)
    real b, f, r, s, cc, pi, ro
    character*2 ans, gage
*
    f = .0254
    pi = 3.1415926
*
*Wires: 1 = 14 gage, 2 = 16 gage, 3 = 18 gage, 4 = 20 gage, 5 = 22 gage.
    a(1) = 583.0
    a(2) = 732.0
    a(3) = 913.0
    a(4) = 1140.0
    a(5) = 1420.0
*Wire diameter in inches to meters.
    d(1) = .0675*f
    d(2) = .0539*f
    d(3) = .0431*f
d(4) = .0346*f
d(5) = .0276*f

*Resistance ohm per meter.
g(1) = .00828
g(2) = .01318
g(3) = .02095
g(4) = .03324
g(5) = .05315

* Print*, 'The optimum C value based on wire type and geometry.'
Print*

* Asking for initial conditions.
Print*, 'What is the radius of the wire holding shell(in.)?'
read*, ro
ro = ro*f
Print*, 'What is the flux density (G),'
read*, b
b = b/1.0e4
Print*, 'What is the length of the solenoid (in.)?'
read*, s
s = s*f
Print*, 'How many layers of 14 gage wire (<100)?'
read*, lay

* Breaking the air gap into smaller increments to find the
* number of layers of each gage of wire can fit into the air gap.
cc = 0.0

* do 10 i = 1, lay
    l(i) = d(1)*(1+sqrt(3.0)/2.0*(i-1))
l(1) = l
    do 15 j = 2, 5
        l(j) = l + int((2.0/sqrt(3.0))*(l(i)/d(j)-1.0))
    15 continue
do 20 k = 1, 5
do 25 m = 1, l(k)
r = ro + d(k)/2.0*(1+sqrt(3.0))*(m-1)
cc = cc + 2.0*pi*r*a(k)*s*b**2/g(k)
25 continue
c(k,i) = cc
cc = 0.0
20 continue
10 continue

* Printing the values in column form to load into kaleidaGraph
Print*
Print*, 'lg(cm) 14G 16G 18G 20G 22G'
Print*, '-----------------------------------------'
do 30 i = 1, lay
Print 28, 100*lg(i), c(1,i), c(2,i), c(3,i), c(4,i), c(5,i)
28 format(f5.1,2x,f7.1,3x,f7.1,3x,f7.1,3x,f7.1,3x,f7.1,3x,f7.1,3x,f7.1)
30 continue
* Print*, 'Would you like to see the displacements in a column?'
read(*,'(a1)') ans
if ((ans .eq. 'y') .or. (ans .eq. 'Y')) then
Print*, 'Air gap length'
do 40 i = 1, lay
  Print*, lg(i)
40 continue
do 50 i = 1, 5
  if (i.eq.1) then
    gage = '14'
  else if (i.eq.2) then
    gage = '16'
  else if (i.eq.3) then
    gage = '18'
  else if (i.eq.4) then
    gage = '20'
  else
    gage = '22'
  end if
* Print 45, gage,'gage'
format (a,1x,a)
do 60 j = 1, lay
  Print 55, c(i,j)
55 format (f7.2)
60 continue
print*
50 continue
end if
*
Print*, 'Would you like to see the displacements in a column?'
read(*,'(a1)') ans
if ((ans .eq. 'y') .or. (ans .eq. 'Y')) goto 1
*
pause
stop
end

Table 13. Optimum damping coefficient sample data

<table>
<thead>
<tr>
<th>Lg(cm)</th>
<th>14 Gage$^6$</th>
<th>16 Gage$^6$</th>
<th>18 Gage$^6$</th>
<th>20 Gage$^6$</th>
<th>22 Gage$^6$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.1</td>
<td>787.1</td>
<td>700.4</td>
<td>765.1</td>
<td>706.1</td>
<td>723.3</td>
</tr>
<tr>
<td>3.1</td>
<td>3024</td>
<td>2940</td>
<td>2824</td>
<td>2873</td>
<td>2776</td>
</tr>
<tr>
<td>5.1</td>
<td>5891</td>
<td>5703</td>
<td>5668</td>
<td>5552</td>
<td>5446</td>
</tr>
<tr>
<td>7.2</td>
<td>9830</td>
<td>9680</td>
<td>9489</td>
<td>9236</td>
<td>9067</td>
</tr>
<tr>
<td>9.1</td>
<td>14280</td>
<td>13990</td>
<td>13710</td>
<td>13560</td>
<td>13210</td>
</tr>
</tbody>
</table>

$^6$Units of kg/s
APPENDIX D: $B_r$ FLUX TESTING DEVICE MAGNETIC CIRCUIT & ERROR ANALYSIS

This section analyzes the magnetic circuit and derives the equations used for error analysis of the $B_r$ flux testing device. The flux tester is shown with dimensional variables in Figure 47.

Figure 47. $B_r$ Flux Testing Device with dimensions

Figure 48. $B_r$ Flux Testing Device magnetic circuit
and the magnetic circuit is shown in Figure 48. The magnetic circuit looks quite complicated suggesting that it could be subject to a large amount of error. Several of the permeances shown in Figure 48 can be easily calculated and end up in the final analysis as having a very small effect on the total permeance. Referring to both Figures 47 and 48 the expressions of the permeances are given by,

\[ P_1 = P_4 = P_5 = P_8 = 0.263H_1 \]  \hspace{1cm} (65)  
\[ P_2 = P_3 = P_6 = P_7 = 0.263H_4 \]  \hspace{1cm} (66)  
\[ P_{11} = P_{12} = \frac{\mu_5(H_3H_4)}{L_5} \]  \hspace{1cm} (67)  
\[ P_9 = \frac{\mu_1(H_1H_4)}{L_1} \]  \hspace{1cm} (68)  
\[ P_{10} = \frac{\mu_2(H_2H_4)}{L_2} \]  \hspace{1cm} (69)  
\[ P_{13} = \frac{\mu_2(H_2H_4)}{L_3} \]  \hspace{1cm} (70)  
\[ P_{14} = \frac{\mu_1(H_1H_4)}{L_4} \]  \hspace{1cm} (71)  
\[ P_{gf} = \frac{H_1H_4}{L_g} \]  \hspace{1cm} (72)  

where: \( \mu_1 \) = permeability in the pole pieces of lengths \( L_1 \) and \( L_4 \),  
\( \mu_2 \) = permeability in the pole pieces of lengths \( L_2 \) and \( L_3 \), and  
\( \mu_5 \) = permeability in the pole piece of length \( L_5 \).

The full expressions for the permeances found in Eqs. 63 and 64 are,

\[ P_{gt} = P_{gf} + P_5 + P_6 + P_7 + P_8 \]  \hspace{1cm} (73)  
\[ P_{tf} = P_5 + P_6 + P_7 + P_8 + \frac{1}{P_9} + \frac{1}{P_{10}} + \frac{1}{(P_{11} + P_{12})} + \frac{1}{P_{13}} + \frac{1}{P_{14}} \]  \hspace{1cm} (74)  

For the device that was built, the permeances of \( P_{gf} \), \( P_{11} \) and \( P_{12} \) are the most significant permeance path contributors in the prediction of the \( B_t \) value. The other permeances are relatively small but their inclusion helps to obtain a more accurate result. It should be pointed out that the areas in the calculation of the pole piece permeances are rectangles. This was done to give the best approximation to the actual cross-sectional area for which the flux lines must pass through. These areas and the approximated lengths are under and over estimated to not only simplify the model but to give an 'average' pole piece permeance estimate. The pole piece permeances were calculated by using the same polynomial function used in Design MW. This was done by using the measured air gap flux density and the ratio of areas to give the flux
values inside each pole piece.

The most important part in designing the flux tester was based on error analysis. The error analysis was done by taking the partial derivative with respect to each variable in Eqs. 63 and 64 and multiplying that variable with the error in that variable. The total error of a particular quantity is the square root of the squares of each product of the partial derivative and the error. The total error using this method in the value of $B_T$ is expressed in Eq. 75 as

$$
\partial B_T = \sqrt{[(1 + \frac{Q_A m}{L_m P_{tf}})\partial B_d]^2 + [(\frac{B_d A m}{L_m P_{tf}})\partial Q]^2 + [(\frac{B_d Q A m}{L_m P_{tf}})\partial A_m]^2 + [(\frac{B_d Q A m}{\frac{1}{2} L_m P_{tf}})\partial L_m]^2 + [(\frac{B_d Q A m}{\frac{1}{2} L_m P_{tf}})\partial P_{tf}]^2}
$$

where:

$\partial = \text{error in the particular variable.}$

The error terms in Eq. 75 have been calculated via partial differentiation these are expressed as,

$$
\partial Q = \sqrt{[\frac{1}{H_c} \partial B_{Tp}]^2 + [(\frac{B_m}{H_c})\partial H_c]^2}
$$

$$
\partial P_{gf} = \sqrt{[(\frac{H_2}{L_g})\partial H_1]^2 + [(\frac{H_1}{L_g})\partial H_2]^2 + [(\frac{H_1 H_2}{L_g^2})\partial L_g]^2}
$$

$$
\partial B_d = \sqrt{[(\frac{B_g}{P_{gf}})\partial B_g]^2 + [(\frac{B_g}{P_{gf}})\partial P_{gt}]^2 + [(\frac{B_g}{P_{gf}})\partial P_{tf}]^2}
$$

where:

$B_{Tp} = \text{published } B_T \text{ value and,}$

$\partial P_{gf} = \partial P_{gt} = \partial P_{tf}.$

The error in the permeance of the air gap is assumed to be the same for the total permeance of the air gap and the total magnetic circuit. For the device that was built, this assumption is true. Referring to Eq. 75, it can't be easily seen but the terms that contribute the most error are $Q$, $B_d$, and $P_{tf}$. The error contribution of $Q$ cannot be curbed in anyway and is an inherent property of magnetic materials in general but, $B_d$ and $P_{tf}$ can be designed. Earlier in the results and discussion section it was mentioned a variation of .02 inches in the air gap will make the results questionable, this can be seen in Eq. 77. The air gap was specified to be large enough so that the error in the air gap length won't be accentuated. The catch 22 of making the air gap too large is seen in Eq. 78. When the air gap increases the permeance of the air gap decreases. As the air gap permeance is decreased then, the error in the operating point $B_d$ starts to increase which in turn increases the error in $B_T$. 
APPENDIX E: COMPUTER PROGRAM FOR THE BR FLUX TESTING DEVICE

* Thomas T. Hansen
* 
* Flux Tester - This program evaluates the br testing device.
* 
* Variable List.
* a0-a10 = Coefficients for the ninth order permeability polynomial.
* k = Dummy array
* p = Load line slope
* t = Magnet thickness
* w = Magnet width
* bd = The operating flux of the magnet.
* br = Remanence flux value of the material.
* hc = The coercive force point at the zero B intersection of the magnet.
* mu, mu2, mu3 = The relative permeability factor.
* bp, bp2, bp3 = The flux densities in various parts of the magnetic circuit.
* dp = Deviation in the slope of the load line.
* dt = Deviation in the thickness of the magnet.
* tt = The height of the magnet.
* wd = The permanent length of the flux tester’s magnet/measurement gap.
* dag = Deviation in the area of air gap.
* dam = Deviation in the area of the magnet.
* dbd = Deviation in the operating point.
* dbr = Deviation in the remanence flux value.
* dhc = Deviation in the coercive force value.
* dlg = Deviation in the air gap.
* mag = The type of hard magnetic material used to produce flux.
* p1, pp3 etc. see PP. 48-49 in the second Thomas T. Hansen lab book.
* 
real n, t, w, br, hc, pi, tt
real wd, l(10), area(10)
real p1, p3, pg, pg1, plm, pp1, pp2, pp3, pp4, pp5
real bd, bg, am, lg, pt, pgt, p, dp, dl, lm
real dbd, dbg, dbrr, dam, dgl, dlm, dhc, dmu, da, dt, dpt1
real k(10), dpp(20), j(20), dw, dpt, dpdk, dtt, dpt2
double precision mu, mu2, mu3, mud, bp, bp2, bp3
double precision a0, a1, a2, a3, a4, a5, a6, a7, a8, a9
character*2 ans
character*10 mag
* 
* Loading in the permeability 9th order coefficients as a function of flux density.
* a0 = 229.48467104
* a1 = 0.83062316038
* a2 = -0.00032376115322
* a3 = 1.4499493037d-07
* a4 = -4.6992047943d-11
* a5 = 8.5739915355d-15
* a6 = -8.9269084112d-19
* a7 = 5.3100089428d-23
a8 = -1.6834870082d-27
a9 = 2.2093278056d-32
pi = 3.1415926

*Loading in the measurements of the flux tester.
*Area(5) is the arc area.
    wd = .8015
    dmu = 50
    l(1) = .364
    l(2) = .298
    l(3) = 1.150
    l(4) = 1.136
    l(5) = 5.89
    da = sqrt(.010**2+.005**2)
    area(1) = 1.018
    area(2) = 1.019
    area(3) = 1.516
    area(4) = 1.515
    area(5) = 1.214

*Loading the magnetic properties of the magnet to be used. The
*following values are approximations of the linear slope of the
*load line. These values are from Arnold Engineering demag. curves.
    ans = 'n'
    Print*, 'What is the magnetic material?'
    read(*,'(a10)') mag
    if (mag .eq. 'arnox 5') then
        br = 3900
        hc = 3550
    else if (mag .eq. 'arnox 7') then
        br = 3450
        hc = 3200
    else if (mag .eq. 'arnox 8') then
        br = 3850
        hc = 3600
    else if (mag .eq. 'arnox 8h') then
        br = 3800
        hc = 3600
    else if (mag .eq. 'arnox 4000') then
        br = 4000
        hc = 3900
    else if (mag .eq. 'pm80') then
        br = 6200
        hc = 5200
    else if (mag .eq. 'pm70') then
        br = 6000
        hc = 5000
    else if (mag .eq. 'neoflex') then
        br = 5450
        hc = 4900
    else
Print*, 'What is Br?'
read*, br
Print*, 'What is Hc?'
read*, hc
end if

* 

p = br/hc

* 

Print*, 'What is the average air gap flux (G)?'
read*, bg
Print*, 'What is the magnet height, T (in.)?'
read*, tt
Print*, 'What is the magnet width, W (in.)?'
read*, w
Print*, 'What is the magnet thickness (in.)?'
read*, t
Print*, 'Are there any non-magnetic laminations on the material?'
read(*, '(al)') ans
if ((ans .eq. 'y') .or. (ans .eq. 'Y')) then
    Print*, 'What is the thickness of each lamination (in.)?'
    read*, lm
    Print*, 'How many layers?'
    read*, n
else
    lm = 1.0e-6
end if

* 

lg = wd-t

* 

t = t-n*lm

* 

Print*, 'Finding the error analysis values.'
Print*, 'What is dLg (in.)?'
read*, dlg
Print*, 'What is dt, magnet (in.)?'
read*, dt
Print*, 'What is dT (in.)?'
read*, dtt
Print*, 'What is dW (in.)?'
read*, dw
if ((ans .eq. 'y') .or. (ans .eq. 'Y')) then
    Print*, 'What,' s dlm?
    read*, dlm
else
    dlm = 0.0
end if

am = tt*w
dam = sqrt(dtt**2+dw**2)
DBG = .01*bg

* 

*Doing the permeance calculations for the magnetic circuit.
p1 = .263*tt
p3 = .263*w
pg = am/lg
pg1 = pg+(2.0*p1+2.0*p3)
gt = pg1+(2.0*p1+2.0*p3)
plm = am/lm

*Calculating the pole piece permeability.

bp = bg*pg1/pg
bp2 = bp*area(2)/(2.0*area(5))
bp3 = bp*area(2)/area(3)
mu = a0+a1*bp+a2*(bp**2)+a3*(bp**3)+a4*(bp**4)+
+ a5*(bp**5)+a6*(bp**6)+a7*(bp**7)+a8*(bp**8)+a9*(bp**9)
mu2 = a0+a1*bp2+a2*(bp2**2)+a3*(bp2**3)+a4*(bp2**4)+
+ a5*(bp2**5)+a6*(bp2**6)+a7*(bp2**7)+a8*(bp2**8)+a9*(bp2**9)
mu3 = a0+a1*bp3+a2*(bp3**2)+a3*(bp3**3)+a4*(bp3**4)+
+ a5*(bp3**5)+a6*(bp3**6)+a7*(bp3**7)+a8*(bp3**8)+a9*(bp3**9)

*Calculating the pole piece permeance.

pp1 = mu*area(1)/l(1)
pp2 = mu*area(2)/l(2)
pp3 = mu*area(3)/l(3)
pp4 = mu*area(4)/l(4)
pp5 = mu*area(5)/l(5)
pt1 = 1.0/(1.0/pp1+1.0/pp4+1.0/(2.0*pp5)+
+ 1.0/pp2+1.0/pp3+1.0/pg1+n/plm)
pt = pt1+(2.0*p1+2.0*p3)

*Error analysis - This is an approximation to the maximum error due to
*the pole piece permeances. This method assumes p2 & p3 errors are
*negligible but finds the lowest total permeance assuming that the pole
*pieces are all at the lowest possible permeances from their calculated
*value. By using pg1 (pg+2*p1+2*p3) - pt1(lowest) we find the maximum
*deviation since pg1 is the maximum permeance the circuit can have since
*the pole piece permeance would be infinite.

k(1) = pp1
k(2) = pp2
k(3) = pp3
k(4) = pp4
k(5) = 2.0*pp5

do 5 i = 1, 5
  if ((i.eq.1).or.(i.eq.2)) then
    dl = .020
    mud = mu
  else if ((i.eq.3).or.(i.eq.4)) then
    dl = .050
    mud = mu3
  else
    dl = .100
    mud = mu2
  end if

*Finding the pole piece error.

dpp(i) = sqrt((area(i)*dmu/l(i)))**2+(mud*da/l(i))**2+
+ (mud*area(i)*dl/(l(i)**2))**2)
\begin{align*}
  j(i) &= k(i) - dpp(i) \\
  \text{if } (i \text{ eq.5}) j(i) &= k(i) - 2.0*dpp(i) \\
  \text{continue}
\end{align*}

*Finding the lowest air gap permeance.
\begin{align*}
  \text{dpp}(6) &= \sqrt{((w*\text{dtt/lg})**2+(tt*\text{dw/lg})**2+(tt*\text{w*dl/g/lg**2)**2})} \\
  \text{dpp}(7) &= \sqrt{((w*\text{dtt/Im})**2+(tt*\text{dw/Im})**2+(tt*\text{w*dl/Im/Im**2)**2})} \\
  j(6) &= \text{pg}1 - \text{dpp}(6) \\
  \text{if } ((\text{ans .ne. 'y')} \text{ .and. (ans .ne. 'Y')}) \text{ dpp}(7) &= 0.0 \\
  j(7) &= \text{plm} - \text{dpp}(7)
\end{align*}

*Finding the maximum magnetic circuit permeance error.
\begin{align*}
  \text{dpdk} &= 1.0/(1.0/j(1)+1.0/j(4)+1.0/j(5)+ \nonumber \\
  &\quad 1.0/j(2)+1.0/j(3)+1.0/j(6)+n/j(7)) \\
  \text{dpt1} &= \text{pt1} - \text{dpdk} \\
  \text{dpt2} &= \text{pg}1 - \text{pt1} \\
  \text{if } ((\text{dpt1.gt.dpt2}) \text{ then} \\
  \text{dpt} &= \text{dpt1} \\
  \text{else} \\
  \text{dpt} &= \text{dpt2} \\
  \text{end if}
\end{align*}

*Error analysis of measured quantities.
\begin{align*}
  \text{dbr} &= .05*\text{br} \\
  \text{dch} &= .07*\text{hc} \\
  \text{dp} &= \sqrt{((\text{dbr/\text{hc}})**2+(\text{br*\text{dch/\text{hc}}**2)**2})}
\end{align*}

*Calculating the final results:
\begin{align*}
  \text{bd} &= \text{bg}*(\text{pg/pg}) \\
  \text{br} &= \text{bd}*(1.0+p*am/(t*pt)) \\
  \text{dbd} &= \sqrt{((\text{pg/pg})*db/pg)**2+(\text{bg*dpp}(6)/pg)**2+ \nonumber \\
  &\quad (\text{bg*pg}*dpp(6)/pg)**2)} \\
  \text{dbr} &= \sqrt{((1.0+p*am/(t*pt))*dbd)**2+(\text{bd*am*dp}/(t*pt))**2+ \nonumber \\
  &\quad (\text{bd}*p*am*dp/(t*pt)**2+((\text{bd}*p*am*dp}/(t*pt**2)**2+ \nonumber \\
  &\quad (\text{bd}*p*am*dp}/(t*pt**2)**2)}
\end{align*}

*Printing the results.
Print*
Print 10, 'Magnetic Material:',mag,'Slope:',p
\begin{align*}
  \text{format}(a,1x,a,a,f6.4) \\
  \text{Print*}
\end{align*}

*\text{Print*}, 'Calculated Flux Values From Measurement'
Print 15, 'br',br,'br','dbr','br','bd', 'bd', 'dbd', 'dbd'
\begin{align*}
  \text{format}(a,1x,f6.1,2x,a,1x,f6.1,2x,a,1x,f6.1,2x,a,1x,f6.1) \\
  \text{Print*}
\end{align*}

*\text{Print*}, 'Magnetic Circuit Permeance Values'
Print 20, 'pg',pg,'pg1',pg1,'pt1','pt1','pgt','pgt','pgt','plm','plm', \nonumber + 'ptot','pt'
\begin{align*}
  \text{format}(a,1x,f5.2,2x,a,1x,f5.2,2x,a,1x,f5.2,2x,a,1x,f5.2,2x,a,1x,f6.2, \nonumber \\
  + 2x,a,1x,f6.2,2x,a,1x,f6.2) \\
  \text{Print*}
\end{align*}
Print*, 'Error Values'
Print 25, 'dP', 'dp', 'dPg', ',dpp(6)', 'dLg', ',dlg', 'dt', ',dt,
 + 'dAm', ',dam', 'dplm', ',dpp(7)', 'dPt', ',dpt
format(a,1x,f5.3,2x,a,1x,f5.3,2x,a,1x,f4.3,2x,a,1x,f4.3,
 + 2x,a,1x,f4.3,2x,a,1x,f6.2,2x,a,1x,f5.2)
Print*
 *
Print*, 'Permeability & Inside Flux Density Values'
Print 30, 'bp', 'bp', 'mu', ',mu', 'bp2', ',bp2', 'mu2',
 + 'bp3', 'bp3', 'mu3', ',mu3
format(a,1x,f6.1,2x,a,1x,f6.1,2x,a,1x,f6.1,2x,a,1x,f6.1,
 + 2x,a,1x,f6.1,2x,a,1x,f6.1)
Print*
 *
Print*, 'Table of pt1 permeance and errors'
Print*, 'PPx  Permeance   Error'
do 35 i = 1, 5
 Print 40, i, k(i), dpp(i)
format(I2,5x,f7.2,4x,f7.3)
35 continue
 *
Print*
Print 45, 'Lowest possible Pt1 permeance:', dpdk
format(a,1x,f6.2)
Print*
 *
Print*, 'Do it again with different parameters?'
read(*, '(a1)') ans
Print*
if ((ans .eq. 'y') .or. (ans .eq. 'Y')) goto 1
 *
Stop
End
APPENDIX F: EXPERIMENTAL FLUX MEASUREMENTS

This appendix was made to understand how the flux measurements were manipulated to give values shown in this paper. Flux measurements were conducted on Design MM by gluing the hall probe on the inside of the steel shell and moving the magnetic mass over its position. The flux plot of this test is shown in Figure 49. The flux density points at the ends of Figure

![Flux Density Graph](image)

Figure 49. Design MM flux density measurement

49 were assumed to be zero. The area under the curve was calculated by an integration routine found in a graphing package called KalaiedaGraph. The total flux lines, \( \Phi \), coming from contributions from the fringing and air gap permeances is equal to the product of the area under the curve in Figure 51 and the circumference of the shell. This is expressed as

\[
\Phi = \pi S_{id}(BY)
\]  \hspace{1cm} (79)

where: \( BY \) = area under the measured flux density curve.

The flux lines are all assumed to be coming from the face of the magnet. To relate the flux lines from the magnet to those measured, the area of the magnet, permeance ratio and the operating flux density \( B_d \) are equated to Eq. 79. Manipulating the expression yields,

\[
B_d = \frac{\Phi_{P_{gt}}}{2\pi(r+s+L_m)W(P_{g2}+2P_f)}
\]  \hspace{1cm} (80)
The tabulated experimental value reported in Tables 10, 11, and 12 was obtained by using Eq. 80.

A similar method was used to obtain the experimental operating flux density of Design MW. This was done by using the averaged experimental values shown in Table 14. Also shown in Table 14, are the corresponding operating flux density values calculated by the following equations,

\[
B_d (r) = \frac{r}{r + L_g + \delta} B (r) \quad (81)
\]

\[
B_d (r_s) = \frac{r_s}{r + L_g + \delta} B (r_s) \quad (82)
\]

where:  

- \( r_s \) = inside radius of the magnet restraining shell,
- \( B (r) \) = measured flux along the core,
- \( B (r_s) \) = measured flux along the magnet restraining shell,
- \( B_d (r) \) = corresponding operating flux density point derived from \( B (r) \), and
- \( B_d (r_s) \) = corresponding operating flux density point derived from \( B (r_s) \).

Table 14. Design MW averaged experimental flux density values (G)

<table>
<thead>
<tr>
<th>Condition</th>
<th>B (r)</th>
<th>B (r_s)</th>
<th>B_d (r)</th>
<th>B_d (r_s)</th>
<th>% Leakage</th>
</tr>
</thead>
<tbody>
<tr>
<td>Open Bottom</td>
<td>1603</td>
<td>1306</td>
<td>1190</td>
<td>1263</td>
<td>5.78</td>
</tr>
<tr>
<td>Closed Bottom</td>
<td>1854</td>
<td>1607</td>
<td>1376</td>
<td>1554</td>
<td>11.4</td>
</tr>
</tbody>
</table>

Table 14 has another category associated with quantifying the percentage of flux that does not reach the core. For some reason, the leakage flux is higher with the bottom on the device than when it is off. There was the possibility that the repulsion magnets on the top and bottom plates were accidently reversed. This could have encouraged more flux to not reach the core by leaving the air gap. The magnets on the top and bottom were checked, however, and it was found that they were in fact in repulsion eliminating this as a possible source for the leakage.
APPENDIX G: MACHINE DRAWINGS FOR DESIGN MM

Detail A (Top Plate)

Detail B (Magnetic Assembly)

Detail C (Housing)

Detail D (Bottom Plate)

Detail E (Base Plate)

Detail G (Bearing Restraining Plate)

1018 Steel
Annealed

Magnetic
Material

Wire

Polo - B

Hardwood

Bearings

Iowa State University Dept. of Aerospace Engineering & Engineering Mechanics

Scale: Full
Advisor: Alison Platau PhD
Title: Electromagnetic Absorber – Generic Overview
Date: 21 Oct. '91
Rev. Date:
Checked By:
Drawing Number: 2B
Dr. By: T. Hansen
Rev. By:
Rev. Checked By:
Revision Number:
- All Dimensions In Inches
1/8" Aluminum Plate

24 Gage Sheet Metal

Detail #3 (Mass & Wood Bearing Assembly)

Detail #4 (Rod Mount)

Detail #2 (Bottom Aluminum Plate)

Bottom View

- All Dimensions In Inches

Iowa State University Dept. of Aerospace Engineering & Engineering Mechanics

<table>
<thead>
<tr>
<th>Scale: Full</th>
<th>Advisor: Alison Flatau PhD</th>
<th>Title: Electromagnetic Vibration Absorber - Detail B</th>
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<td>Rev. Date:</td>
<td>Checked By:</td>
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<tr>
<td>Dr. By: T. Hansen</td>
<td>Rev. By:</td>
<td>Rev. Checked By:</td>
</tr>
<tr>
<td></td>
<td></td>
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</tr>
</tbody>
</table>
Outside radius of a .030 wide, .060 deep groove.
Groove Dimension Tolerances -

- All Dimensions In Inches

Iowa State University Dept. of Aerospace Engineering & Engineering Mechanics

<table>
<thead>
<tr>
<th>Scale: Full</th>
<th>Advisor: Allison Flapan PhD</th>
<th>Title: EM Absorber -- Detail #1 Top Aluminum Plate</th>
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<tr>
<td>Date: 21 Oct. '91</td>
<td>Rev. Date:</td>
<td>Checked By:</td>
</tr>
<tr>
<td>Dr. By: T. Hansen</td>
<td>Rev. By:</td>
<td>Rev. Checked By:</td>
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<td>Drawing Number: 1B</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Revision Number:</td>
<td></td>
</tr>
</tbody>
</table>
Outside radius of a .030 wide, .060 deep groove.
Groove Dimension Tolerances - +.010
-.010

- All Dimensions In Inches
Bottom View

Note: Top is same as bottom without rod holes

* Inside radius of bearing housing. Bearing itself is .973 ± .002 (Press Fit)

- All Dimensions In Inches
Picture Doubled is Size.
Picture is Axisymmetric.

- All Dimensions in inches.

** This region is to be heated & air cooled to remove hardness.
18 AWG Copper Magnet Wire

- All Dimensions in inches.

.005 Brass Shimstock

.1870/.1865 Hardened Rod

#25 Drill Thread 10-24

Counterbore for a 10-24 Flat Head Screw

Iowa State University Dept. of Aerospace Engineering & Engineering Mechanics

Scale: Full
Advisor: Allison Plattan PhD
Title: Electromagnetic Vibration Absorber - - Detail C Housing
Date: 21 Oct. '91
Rev. Date:
Checked By:
Drawing Number: 1
Dr. By: T. Hansen
Rev. By:
Rev. Checked By:
Revision Number:
- All Dimensions In Inches
- All Dimensions in Inches
- All Dimensions in Inches
- All Holes Thru Drill
Detail #B
(Mass & Wood Bearing Assembly)

1/16" Aluminum Plate

Detail #A
(Bottom Aluminum Plate)

Bottom View

- All Dimensions In Inches

* Tolerance +/- 1.0
- All Dimensions In Inches

* Tolerance +/- 1.0°
Bottom View

- All Dimensions In Inches

* Tolerance +/- .005

* Inside radius of bearing housing. Bearing outside radius is 1.177 +.002 (Press Fit)
- All Dimensions in inches.
* Tolerance +/- 1.0°

- All Dimensions in Inches
- All Dimensions In Inches

Iowa State University Dept. of Aerospace Engineering & Engineering Mechanics

Scale: None  Advisor: Alison Plaut PhD  Title: EM Vibration Absorber - Detail 5 Base Plate

Date: Feb. 11, 1992  Rev. Date:  Checked By:  Drawing Number: 8

Dr. By: T. Hansen  Rev. By:  Rev. Checked By:  Revision Number:
- All Dimensions In Inches
- All Holes Thru Drill
Diameter Before Threads = +0.005
(Threads are UNC 6-32)

.750 .100
+.025
.500 -.025

4.000 +0.050
-.050

.1870/.1865 Totherwise Hardened Shaft

.375 +0.050
-.000

+.020
-.000

.250 +.025 Thread 1/4 - 20

4.000 +.050
-.050

1/4 Diameter Aluminum Shaft

.500 +.025 Thread 1/4 - 20

.125 +.010
-.010

.080 +.010
-.010

This Groove is made for wire-in, wire-out transport.

++ This region is to be heated & air cooled to remove hardness.

- All Dimensions in inches.