Electromagnet Design for a Time Varying Magnetic Flux Leakage System

by

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A thesis submitted in conformity with the requirements for the degree of Master of Applied Science

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Abstract

The aim is to design and build an optimized electromagnet for a time varying magnetic flux leakage (MFL) system that can operate within a frequency range of 0 – 10 Hz. The system will be used for the inspection of fuel storage tank floors. The advantage over DC systems is that the response of each frequency component would help in estimating the remaining plate thickness at a defect location; this is usually the main concern for fuel tanks.

The electromagnet is designed with finite element modeling using COMSOL Multiphysics. Experimental results are compared with simulations and are used to validate the design. A prototype MFL system that utilizes the electromagnet is tested by inspecting steel plate samples with machined defects of different depths. Simulations are then used to demonstrate the ability of AC MFL to differentiate between defects that would be difficult to characterize with DC MFL alone.
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Chapter 1
Introduction

1 Introduction

1.1 Motivation

Fuel storage tank floors are regularly inspected for defects as part of preventative maintenance against potentially disastrous leakage accidents. The floor plates are typically made of low carbon steel, so they are susceptible to corrosion especially on the bottom side that is in contact with the ground. Corrosion on the top side that is in contact with the fuel is also possible because of fuel additives, so a fiberglass coating is usually applied to prevent that. However, defects can still form due to moisture trapped between the protective coating and steel plate. Figure 1.1 presents a few examples of corrosion in storage tanks. Defects in the floor plate can eventually lead to complete failure; therefore, it is essential to periodically assess the state of the tank base.

![Corrosion in Storage Tanks](image)

Figure 1.1: Corrosion in Storage Tanks [1]

The inspection is generally done using non-destructive testing such as the magnetic flux leakage (MFL) method. Steel is a magnetic material; therefore, magnetic testing methods are particularly suitable for this application. MFL involves magnetically saturating a portion of the plate, and the presence of a defect in that region results in flux leakage that is picked up by sensors.
Current systems use a constant (DC) magnetic flux produced by a permanent magnet. Alternatively, the magnetic field may be generated by a constant current electromagnet in which a magnetic core provides a low reluctance path for the flux in and out of the inspected plate. The use of electromagnets is a relatively new concept that is not generally used yet. When the flux encounters a defect, some of the magnetic flux leaks out in a direction normal to the plate due to the associated high magnetic reluctance of the defect. Hall effect sensors are used to detect the leakage signal. Since it relies on abrupt changes in magnetic reluctance, the MFL method is good at detecting pit defects with edges, but might fail to pick up lake shaped defects that involve gradual plate thinning. A schematic showing the basic operation principle of MFL is shown in Figure 1.2.

Contemporary MFL systems are primarily used only as an initial screening tool to locate defects, followed by more detailed ultrasound testing to quantitatively characterize any detected flaws. The two-step process is required because DC MFL signal strength is generally proportional to total material loss, such that individual dimensions of the defect cannot be accurately estimated. In particular, the probability of tank failure is highly dependent on the flaw depth as determined.
from the remaining plate thickness which cannot be determined by DC MFL. In fact, the industrial sponsor, Groupe Mequaltech, has specified that a loss of 20% in plate thickness is the critical defect depth beyond which repairing the tank floor plate is necessary. The standard plate thickness is 0.25 in (6.35 mm); therefore, the critical defect depth is 0.05 in (1.27 mm). Figure 1.3 presents two defects that might result in similar DC MFL signal amplitudes due to their similar volume. However, one is deeper than the other and poses a much higher risk to the integrity of a storage tank floor.

![Figure 1.3: Defects of Similar Volume with Different Depths and Subsequently Different Failure Potential](image)

1.2 Objectives

There exists the possibility of extracting much additional information by moving to a time varying magnetic flux system. The aim of this project is to develop an electromagnet for a time varying MFL system that would be able to provide quantitative evaluation of the detected defect geometry. The electromagnet can magnetize the plate at several frequencies, as opposed to DC MFL that is based on permanent magnets. Magnetic fields at different frequencies have varying skin depths inside the plate under inspection. [4] The advantage over DC systems is that the amplitude and phase response of each frequency component would provide more information about the defect, which in turn helps in flaw characterization. Specifically, it is desired to estimate
the defect depth or remaining plate thickness at a flaw location; this is usually the main concern for fuel tanks. As stated in section 1.1, the decision whether to repair a tank floor plate is based on the remaining plate thickness.

The objective is to design an electromagnet that can magnetically saturate a 0.25 in (6.35 mm) thick low carbon steel plate and that can be powered with current of arbitrary waveform within a bandwidth of 0 - 10 Hz. The electromagnet should also exhibit low background magnetic field at the location of the sensors that pick up the MFL signal. In addition, the magnet should be able to operate with a liftoff of 0.375 in (9.525 mm) between it and the inspected steel plate to account for the presence of a protective coating and wheel clearance. The scanning span must be 12 in (304.8 mm) in the direction normal to the scanning direction. The liftoff and scanning span are specified by the project sponsor. The electromagnet is designed using finite element modeling, and a prototype is developed and tested to validate the design.

1.3 Overview

In the next chapter, chapter 2, background information and a thorough literature review are presented. It is divided into four sections. The first section gives an overview on magnetic properties of materials and a survey of the most popular soft magnetic materials that are used for AC electromagnet cores. The second section discusses electromagnet coils and magnetic circuits. The third section focuses on core and coil losses and cooling methods. The fourth section presents the components of an MFL system.

Chapter 3 contains the details of the electromagnet design process and is divided into five sections. The first section discusses the electromagnet modeling including both magnetic circuit analysis and finite element analysis. The second section focuses on the electromagnet core material and geometry. The third section focuses on the electromagnet coil design, most importantly the
wire gauge and number of turns. The fourth section discusses the electromagnet assembly and cart design. The fifth section presents the details of the electric circuit powering the electromagnet.

In chapter 4, the electromagnet prototype specifications are validated and compared to simulations. It contains four sections. The first section investigates the coil temperature when the electromagnet is in operation. The second section studies the force between the electromagnet and inspected plate. The third section compares the experimental magnetic flux density produced by the prototype to corresponding simulation results. The fourth section includes coil current measurements for the electromagnet in AC operation.

Chapter 5 demonstrates the ability of the prototype to detect defects and consists of two sections. The first section presents the results of a few MFL inspection experiments on steel plate samples. The second section uses a simulation to show the potential advantage of using AC MFL.

Chapter 6 is the conclusion. It includes a summary and conclusions of the key points of the project along with potential ideas for future work.
Chapter 2
Background and Literature Review

2 Background and Literature Review

2.1 AC Electromagnet Core

2.1.1 Magnetic Properties of Materials

All substances can be classified into five categories according to the type of magnetism they exhibit: diamagnetic, paramagnetic, antiferromagnetic, ferromagnetic, and ferrimagnetic. For diamagnetic materials, the resulting magnetic flux density due to an applied magnetic field is less than that of vacuum. On the other hand, it is slightly higher than that of vacuum in paramagnetic and antiferromagnetic materials. Substances that fall in these three categories are usually termed nonmagnetic. As for ferromagnetic and ferrimagnetic materials, they exhibit flux densities that are orders of magnitude higher than that in vacuum for the same magnetic field strength; therefore, they are collectively known as magnetic materials. [6]

The magnetic flux density in a substance due to an applied magnetic field is given by the following equation:

\[ B = \mu_0 (H + M), \quad (2.1) \]

where \( B \) is the magnetic flux density in Tesla (T), \( \mu_0 \) is the magnetic permeability of vacuum (T.m/A), \( H \) is the magnetic field strength in Amperes per meter (A/m), and \( M \) is the magnetization also in Amperes per meter (A/m). [6]

The magnetization \( M \) and magnetic field strength \( H \) are in turn related by the ratio

\[ \chi = \frac{M}{H}, \quad (2.2) \]
where $\chi$ is the magnetic susceptibility and is dimensionless. The susceptibility $\chi$ is zero for vacuum, low and negative for diamagnetic materials, low and positive for paramagnetic and antiferromagnetic materials, and high and positive for ferromagnetic and ferrimagnetic materials. This explains the magnetic behavior exhibited by the different categories, as discussed earlier. When considering electromagnet core design, ferromagnetic and ferrimagnetic materials are of interest, and the others are dismissed because of their relatively weak magnetization.

Using the definition of magnetic susceptibility, the magnetic flux density can be given by

$$B = \mu_0(H + \chi H) = \mu_0(1 + \chi)H. \quad (2.3)$$

The ratio

$$\frac{B}{H} = \mu_0(1 + \chi) \quad (2.4)$$

is defined as the magnetic permeability of the medium $\mu$ (T.m/A), and the ratio

$$\frac{\mu}{\mu_0} = 1 + \chi \quad (2.5)$$

is defined as the relative permeability of the medium $\mu_r$ and is dimensionless. Therefore, the relationship between flux density and field strength is usually presented as

$$B = \mu_0\mu_rH = \mu H. \quad (2.6)$$

The magnetic permeability $\mu$ of a ferromagnetic or ferrimagnetic material is not constant and varies as a function of the magnetic field strength $H$. Therefore, the graph of magnetic flux density $B$ vs. magnetic field strength $H$ for magnetic materials, commonly known as the magnetization curve, is nonlinear as shown in Figure 2.1.
The magnetization curve of magnetic materials also exhibits two phenomena: saturation and hysteresis. Magnetic saturation involves the process that at a high enough magnetic field strength $H$, the relative permeability $\mu_r$ approaches 1. The magnetic flux density $B$ beyond which the slope of the magnetization curve is equal to $\mu_0$ is called the saturation induction $B_s$. On the other hand, hysteresis refers to the fact that after reducing the applied magnetic field $H$ back to zero, the magnetic flux density $B$ does not return to zero as well. The value of the flux density at that point is termed the remanent induction $B_r$. In order to reduce the flux density to zero, it is required to apply a magnetic field $H$ in the opposite direction. The magnitude of that field is called the coercive field $H_c$. An example of a hysteresis loop is shown in Figure 2.2. A material with a high remanent induction and high coercive field is said to have high retentivity and high coercivity respectively. Such a material is classified as magnetically hard and is useful for making permanent magnets. Soft magnetic materials have the opposite properties, which makes them particularly useful for electromagnet cores.
2.1.2 Eddy Currents and Core Laminations

When a substance is subjected to an alternating magnetic field, which is the case for AC electromagnets, an electromotive force (emf) is induced in the material by Faraday’s Law and is given by

\[ \varepsilon = -\frac{d\varphi}{dt}, \]  

(2.7)

where \( \varepsilon \) is the induced emf in volts (V), and \( \varphi \) is the magnetic flux in Weber (Wb) which is equivalent to (T.m\(^2\)). [9] For a uniform cross-section,

\[ \varepsilon = -A \frac{dB}{dt}, \]  

(2.8)

where A is the cross-sectional area of the material in square meters (m\(^2\)), and B is the magnetic flux density in Tesla (T). The induced emf results in current, termed eddy current, that is directly proportional to the emf and inversely proportional to the electrical resistivity of the material.
Therefore, for a given alternating magnetic field, the eddy current is higher for materials with high magnetic permeability and low electrical resistivity. [6]

In the simple case of a circular cross-section, eddy currents flow in concentric circles. Each eddy current loop produces a magnetic field in the center that is given by

$$ H = \frac{I}{2R}, \quad (2.9) $$

where I is the current in Amperes (A), and R is the radius of the current loop in meters (m). [6] By Lenz’s Law, the direction of the eddy currents is such that they produce a magnetic field that opposes the change in magnetic field that is inducing them. [9] This is the reason for the negative sign for the induced emf in Faraday’s Law. The resultant magnetic field produced by all the eddy current loops is highest in the center and decreases towards the surface. Therefore, the maximum flux density due to an alternating field is not uniform over the cross-section. It is highest on the surface and decreases towards the center. Also, there is a phase difference between the flux densities at points with different distances from the surface. This phenomenon is called the skin effect. [6]

It is always desirable to minimize eddy currents in ac electromagnet cores because they cause a decrease in magnetic flux density and result in power loss that is dissipated as heat. Therefore, ac electromagnet cores are usually made from electrically insulated thin laminations. [6] The eddy current loops are restricted within each lamination, and the smaller cross-sectional area leads to a lower induced emf and lower eddy current. Figure 2.3 highlights the difference between the eddy currents in a solid and laminated core.
Figure 2.3: Eddy Currents in a Solid vs. Laminated Core [6]

The suitable thickness of the laminations is determined by considering the skin effect caused by eddy currents as discussed earlier. The ratio of the amplitude of the magnetic field at a depth \( d \) to that at the surface is given by

\[
\frac{H_d}{H_s} = e^{-(1+j)\frac{d}{\delta}},
\]

(2.10)

where \( \delta \) is the skin depth. [10] The skin depth is the distance where the field amplitude is \( 1/e \) of that at the surface, and it is calculated using

\[
\delta = \frac{1}{\sqrt{\pi \sigma \mu f}},
\]

(2.11)

where \( \sigma \) is the electrical conductivity of the material in Siemens per meter (S/m), \( \mu \) is the magnetic permeability of the material in Henries per meter (H/m), and \( f \) is the frequency of the applied magnetic field in Hertz (Hz). [11] The exponent of the ratio is complex because of the phase difference. The ratio also applies to flux density \( B \) if it is assumed that the permeability \( \mu \) is constant and not a function of the field strength \( H \). Due to the skin effect, if the thickness of the laminations is significantly larger than the skin depth, then the magnetic flux density in the center will be approximately zero. This is a waste of material because the center is not magnetized; therefore, laminations are chosen to be thinner than the skin depth when designing ac
electromagnet cores. Figure 2.4 demonstrates the skin effect for different lamination thickness of silicon steel at 60 Hz.

![Diagram of Magnetic Flux Density Profile](image)

**Figure 2.4: Magnetic Flux Density Profile for Different Lamination Thickness [6]**

### 2.1.3 Survey of Soft Magnetic Materials

As mentioned in the earlier sections, soft magnetic materials are used for electromagnet cores. This is because they are easier to magnetize and demagnetize compared to hard magnetic materials. This section explores the various choices available when selecting a suitable material for electromagnets.

When high efficiency and low power loss are not important, low carbon steel is usually used because it is relatively cheap and widely available. The carbon content is kept as low as practically possible and is normally around 0.03%. Iron carbide precipitates with time and increases core losses in a process termed aging. The maximum relative permeability is 1000 – 2000. The permeability can be increased to around 10,000 by further decreasing the carbon content to 0.01%, and the material is then commonly called high purity iron. [6]
A substantial improvement in magnetic properties can be achieved by using silicon steel, which is otherwise known as electrical steel. Adding silicon increases the electrical resistivity which decreases eddy current losses. It also increases the relative permeability. The main disadvantages are the decrease in saturation induction and the decrease in ductility. [12] The silicon content is normally around 3%. [6] The variation of material properties with silicon percentage is plotted in Figure 2.5.

![Figure 2.5: Properties vs. Silicon Percentage [6]](image)

Grain oriented silicon steel is a special variant of silicon steel which has superior magnetic properties in the rolling direction of the sheet. The relative permeability is higher and the core loss is lower compared to non-oriented silicon steel. Losses are usually around 10 times less than that in low carbon steel. [6] It is important to note that the magnetic properties are highly dependent on the direction of magnetization, so using this material is beneficial only when the magnetization is parallel to the rolling direction. [12] Indeed, the relative permeability normal to the rolling direction is less than 2% of that parallel to it.
Low carbon steel and silicon steel are the most popular options when low cost per pound is required with silicon steel, especially grain-oriented, being the more expensive choice. These materials also have a relatively high saturation induction, so a smaller amount of material is needed for a certain magnetic flux. This decreases the size and weight of the electromagnet. Silicon steel is specifically designed to be magnetized at power frequencies (50Hz or 60Hz), and the recommended operation frequency range is less than 100 Hz. The available lamination thickness range is 0.1 – 1 mm (0.004 – 0.04 in). [6]

If better magnetic properties than those of silicon steel are necessary in terms of permeability, saturation induction, or losses, then other more expensive materials are considered. Nickel-iron alloys are one such material. The nickel content is typically around 50 – 80%. This alloy has a higher relative permeability than that of silicon steel. [13] Also, lower losses are made possible with lamination thicknesses as low as 0.006 mm (0.00025 in), and the electrical resistivity can be increased with the addition of Mo, Cu, or Cr. However, the saturation induction is lower and the price per pound is higher. [6] Alloys containing 50% and 78% Ni are the two standard varieties. The alloy with 50% Ni has a higher saturation induction, whereas the other has a higher relative permeability and lower coercivity. [6] Figure 2.6 shows the maximum permeability of nickel-iron alloys as a function of nickel percentage.
Another class of soft magnetic materials is the cobalt-iron alloys. These alloys have a 10\% higher saturation induction than iron. The cobalt content range is 30 – 50\%. The alloy with 50\% Co has the best magnetic properties, but the price generally increases for higher cobalt content. 2\% V is usually added to increase ductility. [6]

For very high efficiency applications, amorphous and nanocrystalline alloys are used to decrease core losses. It involves high cooling rates from the liquid state, usually 10^5 - 10^6 °C/s. The alloys are composed of Fe, Ni, Co, B, and Si. The saturation induction is 1.5 – 1.9 T, and the resistivity is greater than 10^6 Ω.m. [6]

Bulk metals are not suitable for use in high frequencies, because of the high associated losses. Powder cores made of iron or iron-nickel alloy are designed for high frequencies up to 100 kHz. The powder size is 0.05 – 0.1 mm, and the particles are electrically insulated from each other.
The result is that eddy currents are virtually nonexistent; however, the relative permeability is 10 – 100 which is significantly less than that of the same material in bulk form. [6]

For frequencies in the megahertz range, ferrite cores are used due to their high electrical resistivity. Ferrites are insulators so eddy current losses are not an issue, and cores do not require lamination. The general chemical formula for soft magnetic ferrites is MO₂Fe₂O₄, where M is Mg, Mn, or Ni. The relative permeability of ferrites is significantly less than that of magnetic metals, and the saturation induction is less than a third of that of iron. In fact, zinc ferrite is added to magnetic ferrite to increase the relative permeability and saturation induction. The two main types of ferrites used for electromagnet cores are Mn-Zn ferrites and Ni-Zn ferrites. Mn-Zn ferrites have a permeability of 1000 – 2000, a coercive field of less than 80 A/m, and a resistivity of 0.2 – 1 Ω.m. On the other hand, Ni-Zn ferrites have a permeability of 10 – 1000, a coercive field in the order of hundreds of A/m, and a resistivity of around 1000 Ω.m. The former is suitable for use up to 1 MHz, whereas the latter can be used to frequencies more than 100 MHz. [6] The permeability of different types of ferrite is plotted as a function of frequency in Figure 2.7.

![Figure 2.7: Permeability of Different Ferrites vs. Frequency](image_url)
2.2 AC Electromagnet Coil

2.2.1 Magnetic Field Produced by a Coil

The magnetic field produced by a current in a straight wire is given by

\[ H = \frac{I}{2\pi R} \]  \hspace{1cm} (2.12)

where \( H \) is the magnetic field in Amperes per meter (A/m), \( I \) is the current in Amperes (A), and \( R \) is the distance from the wire axis in meters (m). [9] For a current in a circular loop, the field at the center is given by Equation 2.9. [9] The direction of magnetic field due to current is shown in Figure 2.8.

![Figure 2.8: Direction of Magnetic Field Produced by Current [6]](image)

In order to produce a uniform magnetic field, a winding consisting of multiple loops, called a solenoid, is utilized. The field along the axis of a solenoid is

\[ H = \frac{NI}{L} \left( \frac{L+2x}{2\sqrt{D^2+(L+2x)^2}} + \frac{L-2x}{2\sqrt{D^2+(L-2x)^2}} \right) \]  \hspace{1cm} (2.13)
where \( N \) is the number of turns, \( L \) is the length of the solenoid, \( x \) is the distance from the center, and \( D \) is the diameter of the solenoid. [6] The field exactly at the center is

\[
H = \frac{NI}{L} \left( \frac{L}{\sqrt{D^2 + L^2}} \right).
\] (2.14)

The equation for an infinitely long solenoid

\[
H = \frac{NI}{L} \quad (2.15)
\]

is a valid approximation at a position far from the ends of the coil if the length is significantly larger than the diameter. [6] It is important to note that the field is maximum at the center and decreases by half at either end of the solenoid. Also, the field is approximately uniform in the central half of the coil. Increasing either the number of winding layers or the current will increase the resulting magnetic field. A diagram of the magnetic field produced by a solenoid is shown in Figure 2.9.

![Diagram of the Magnetic Field Produced by a Solenoid](image)

**Figure 2.9: Magnetic Field Produced by a Solenoid [6]**

### 2.2.2 Magnetic Circuits

Electromagnets are made by wrapping a coil around a magnetic core. The core is made of a material with a high relative permeability as discussed earlier. Because of the magnetic core, the
field produced by the coil results in a higher magnetic flux density than that of a solenoid in air. Therefore, the presence of the magnetic core decreases the magnetic field needed to achieve a certain flux density.

When a magnetic core forms a closed loop with no or a few air gaps, a closed magnetic circuit is formed. In the case of a single material with constant permeability, the magnetic flux is given by

\[ \varphi = BA = \mu HA = \mu \frac{NI}{L} A = \frac{NI}{L/\mu A}. \quad (2.16) \]

The product \( NI \) is termed the magnetomotive force (mmf), and the value \( L/\mu A \) is the reluctance \( R \). Therefore,

\[ mmf = \varphi R. \quad (2.17) \]

If air gaps or multiple materials are present, the same equation applies, but the reluctance is the sum of the individual reluctances of all the components. [14] This makes use of Ampere’s Law that when applied to a magnetic circuit gives the following result:

\[ mmf = \oint H \, dl. \quad (2.18) \]

Since the permeability of air is much lower than that of the magnetic core, an air gap increases the total reluctance significantly. In fact, the reluctance increases by a factor of

\[ f = 1 + \left( \frac{g}{l} \right) (\mu_r - 1), \quad (2.19) \]
where \( g \) and \( l \) are the length of the air gap and magnetic core respectively, and \( \mu_r \) is the relative permeability of the core. [6] This leads to a decrease in magnetic flux for the same magnetomotive force. Figure 2.10 presents examples of both closed and open magnetic circuits.

![Diagram of Magnetic Circuits](image)

**Figure 2.10: Closed and Open Magnetic Circuits [6]**

### 2.2.3 Magnet Wire

Magnet wire refers to the wire used for the windings of electromagnet coils. It is usually made of copper with an insulating coating. [6] The DC current carrying capacity is directly proportional to the cross-sectional area of the wire. For AC, the skin effect that was discussed earlier should be considered if the radius of the wire is larger than the skin depth for a given frequency. The coating is thin and flexible to allow for tight packing and easy winding. Depending on the operating temperature, different types and thickness of insulation is used. For temperatures less than 240 °C, enamel coatings made of organic materials are the most popular. More expensive options that are rated for higher temperatures include fiber glass and ceramics. [6]

### 2.3 AC Electromagnet Losses and Cooling

#### 2.3.1 Coil Losses

The power input to an electromagnet coil is dissipated as heat and is given by

\[
P = I^2 R, \quad (2.20)
\]
where $P$ is the power in watts (W), $I$ is the current in Amperes (A), and $R$ is the resistance of the coil in ohms (Ω). Therefore, increasing the magnetomotive force by increasing either the number of turns of the coil or the current have different effects on the dissipated heat. If the number of turns is increased, the resistance of the coil and the power increase linearly. On the other hand, if the current is increased, the power increases in a square fashion. This can be mitigated by proportionally increasing the cross-sectional area of the coil windings, which decreases the resistance as long as the radius is less than the skin depth at the operating frequency.

### 2.3.2 Core Losses

Core losses are another source of heat dissipation in an AC electromagnet. They represent a power loss and therefore result in a decrease in efficiency. One of the two main types of core loss is caused by eddy currents. The classical eddy current power loss is given by

$$P_e = \frac{(\pi dB f)^2}{6\rho}, \quad (2.21)$$

where $P_e$ is the power loss in watts per cubic meter (W/m$^3$), $d$ is the thickness of the core or lamination in meters (m), $B$ is the maximum magnetic flux density in Tesla (T), $f$ is the frequency of the applied magnetic field in Hertz (Hz), and $\rho$ is the resistivity of the core material in ohm-meters (Ω.m). [15] It is termed classical because the equation only applies if the permeability of the core material is constant and the thickness is significantly smaller than the skin depth. If these conditions are not met, the actual eddy current loss will always be higher than that predicted by this equation. The difference is termed the anomalous loss $P_a$. [6]

The second major type of core loss is caused by hysteresis. The power loss due to hysteresis is given by
\[ P_h = A_h f, \quad (2.22) \]

where \( P_h \) is the power loss in watts per cubic meter (W/m\(^3\)), \( A_h \) is the area of the hysteresis loop of the magnetization curve of the material (T.A/m), and \( f \) is the frequency of the applied magnetic field in Hertz (Hz). Disregarding the anomalous loss \( P_a \), the total core loss would be

\[ P_t = P_e + P_h. \quad (2.23) \]

\( P_e \) and \( P_h \) are proportional to the square of the frequency and the frequency respectively. If the core loss per cycle is considered, \( P_e \) and \( P_h \) would be proportional to the frequency and constant respectively. Therefore, the total loss per cycle is expected to be linear with frequency. However, when the anomalous loss \( P_a \) is added, the linearity no longer holds. [6] Figure 2.11 shows how the different types of power loss change with increasing frequency.

![Figure 2.11: Core Loss per Cycle vs. Frequency](image-url)
2.3.3 Cooling Methods

Due to the significant amount of heat dissipated by electromagnets, various cooling methods have been developed. Using fans to blow air over the coils is the cheapest and easiest method; however, the associated cooling performance is low compared to the others. Better cooling can be achieved by using liquid cooling, where water or a water-based solution circulates in tubes to transfer heat away from the coils and core. For high-power applications, the whole electromagnet is immersed in mineral oil which is an electric insulator that transfers heat efficiently. [6]

2.4 Magnetic Flux Leakage System

2.4.1 Magnetic Flux Production Subsystem

A magnetic flux leakage (MFL) system consists of two subsystems: magnetic flux production and magnetic flux leakage measurement. The flux production subsystem magnetizes the ferromagnetic specimen being inspected to the point of local magnetic saturation. [3] Flux leakage measurement is what ultimately provides information about the presence and size of defects. The focus of this project is on the magnetic flux production subsystem and specifically the design of an electromagnet that serves that purpose. However, since the subsystems are not completely independent, some magnetic flux measurement considerations will factor into the design process.

The utilized magnetic flux can be either constant or of alternating polarity, and an alternating flux can be further classified depending on whether it consists of a single frequency or multiple frequencies. In the case that a constant DC magnetic field is to be used, the options are either to use an electromagnet powered with a constant DC current, or a permanent magnet. [2] Using a permanent magnet eliminates the need for an external power source, such as a battery
module, to produce the magnetic flux. However, an electromagnet provides a practical method to adjust the magnetic flux density by changing the current; an electromagnet can also be turned off when it is desired to disengage the system from a steel test piece.

When an alternating magnetic field is used, the selection of a suitable frequency is governed by the skin effect which dictates the magnetic field penetration in an electrically conductive and magnetically permeable material. The penetration depth decreases with increasing magnetic field frequency, material conductivity, or material permeability. Since the magnetic flux leakage technique is used to inspect an object made of a magnetically permeable material, the skin effect cannot be deemed negligible.

Producing the required magnetic flux is usually done using an electromagnet consisting of a yoke and an excitation coil. The excitation coil provides the magnetomotive force (mmf) that is proportional to the current and number of turns of the coil. The yoke is made of a material with a high magnetic permeability in order to obtain a magnetic flux that is orders of magnitude higher than that produced by a coil in air. The yoke provides a low reluctance path for the magnetic flux between the coil and the object to be inspected.

In order for the MFL non-destructive technique to be effective, the signal-to-noise ratio (SNR) must be as high as practically possible. The magnetic flux leakage in the presence of a defect should be maximized; in addition, direct flux leakage from the yoke or a defect-free specimen to the MFL sensors in the air should be minimal. Therefore, the magnetic flux density in a defect-free object must be high enough to ensure magnetic saturation; this can be determined from the B-H curve of the material. Specifically, the operating point is selected to be where the slope of the curve (i.e., material permeability) decreases most rapidly indicating that magnetic saturation is starting. Further increasing the magnetomotive force at this point might cause the flux
to leak to the air even when no defect is present. Also, the yoke should not be magnetically saturated as that would cause direct leakage from the yoke to the air. This is also determined from the B-H curve of the yoke material. [2]

2.4.2 Magnetic Flux Leakage Measurement Subsystem

The most common instruments used to measure magnetic flux density are hall effect sensors and flux meters. [6] Hall effect sensors make use of the hall effect where an emf is induced in a current carrying conductor in the presence of a magnetic field. The induced emf is given by

\[ e = \frac{R_H IB}{t}, \]  

(2.24)

where \( e \) is the emf in volts (V), \( R_H \) is the hall constant (Ω.m/T) that is material specific, \( I \) is the current in the conductor in Amperes (A), \( B \) is the magnetic flux density in Tesla (T), and \( t \) is the thickness of the conductor in meters (m). [6] Therefore, the magnetic flux density can be calculated by measuring the emf after amplification. The hall effect is larger in semiconductors than in metals, so the hall element, which is 1 mm\(^2\) or less in size, is usually made of InSb. [6] The hall element is connected to a current source and an amplifier, which can be in the form of a control unit with an indicating meter. Hall effect sensors can be used to measure flux densities ranging from less than 0.05 mT to more than 3 T. They are also able to measure alternating fields, and the frequency is limited by the measuring circuit. [6] A diagram of the hall effect is presented in Figure 2.12.

![Figure 2.12: Hall Effect Diagram](image)
Flux meters operate based on Faraday’s Law, where the induced emf in a coil due to an alternating magnetic field is given by Equation 2.7. The integral of the voltage is then given by

$$\int \varepsilon \, dt = -NA\Delta B,$$  \hspace{1cm} (2.25)

where A is the area of the coil in square meters (m\(^2\)) and \(\Delta B\) is the change in flux density in Tesla (T). To calculate the change in flux density, the flux meter uses an operational amplifier with capacitive feedback that outputs the integral of the voltage. It is important to note that unlike hall effect sensors that measure flux density, flux meters measure changes in flux density. The sensitivity of the flux meter is limited by that of the electronic integrator which is typically 1 mV.s. [6]

Other instruments that can be used to measure magnetic flux density include rotating coil magnetometers, magnetoresistance sensors, fluxgate magnetometers, and magnetic potentiometers. For magnetic flux leakage detection and measurement, the hall effect sensor is the most popular choice due to it being a mature technology, and its characteristics are suitable for this application. The sensor is usually connected to a data acquisition device after which signal processing techniques such as noise filtering and amplification can be applied. Eddy currents induced by the relative velocity between the electromagnet in the inspection system and the tested object are a major source of noise when measuring magnetic flux leakage. To reduce the velocity effect, the inspection speed is maintained constant at a low value that does not exceed 100 mm/s. [16] In the case when alternating magnetic fields are used, another source of noise arises, because the variable nature of the magnetic field also induces eddy currents in the studied object.
Chapter 3
Electromagnet Design for the MFL System

3 Electromagnet Design for the MFL System

Before going into the details of the design process, it is important to further explain one of the design specifications mentioned in the objectives in section 1.2. That is the frequency range within which the electromagnet will operate: 0 – 10 Hz. This range is selected based on the skin depth in low carbon steel. It is desired to have various skin depths depending on the frequency; however, the maximum frequency must still have a skin depth of at least 1 – 2 mm inside the 0.25 in (6.35 mm) thick steel plate. This is because the goal of MFL is to detect defects with depths that affect the remaining plate thickness and not surface flaws. Therefore, the maximum frequency is chosen to be 10 Hz. Also, the skin depth decreases with the square root of frequency, so increasing the frequency beyond 10 Hz has only an incremental effect on the skin depth range.

3.1 MFL Electromagnet Modeling

3.1.1 Magnetic Circuit Analysis

The MFL electromagnet along with the steel plate under inspection can be modeled as a magnetic circuit. As discussed in section 2.2.2, the governing law is given by Equation 2.17. The mmf is the product of the current I and number of turns N of the excitation coil of the MFL electromagnet. The equivalent reluctance is the sum of the reluctances of all the components of the circuit, which include the electromagnet core, the inspected plate, and the air gaps between the two previous components. The air gaps represent the liftoff between the core of the electromagnet and the plate including any protective coating on the plate (typically a plastic sheet up to several mm in thickness). Therefore, the reluctance is given by

\[ R = R_{core} + R_{plate} + R_{gap} = \frac{L_c}{\mu_c A_c} + \frac{L_p}{\mu_p A_p} + \frac{L_g}{\mu_g A_g}, \]  

(3.1)
where $L$ is the length, $\mu$ is the magnetic permeability, and $A$ is the cross-sectional area. [14] Figure 3.1 shows a typical diagram of a magnetic circuit involving an electromagnet and a steel plate.

![Magnetic Circuit of Electromagnet with Inspected Plate][17]

There are multiple reasons as to why this analysis might not yield accurate results. First, it is difficult to assign a value for the cross-sectional area of the air gaps $A_g$. The magnetic flux in the air gaps is not confined to a volume of high permeability as in the rest of the circuit, so the actual cross-sectional area is larger than that of the electromagnet core. The magnetic field lines are said to “bulge out” in a manner termed “fringing” field lines. This effect becomes more pronounced for larger air gaps; therefore, the corresponding cross-sectional area increases when the air gap length is increased. [14]

The second issue is that the magnetic circuit analysis assumes that the magnetic flux is completely within the closed magnetic circuit loop, defined by the magnet iron core, steel test...
plate, and the narrow gaps between them. However, the magnetic permeability of the surrounding medium (air) is not zero and is in fact equal to $\mu_0$. Therefore, a portion of the magnetic lines are outside the closed loop. [14] This flux is termed the leakage flux and is very difficult to estimate analytically – finite element simulations are therefore used.

The third and perhaps most significant factor is the nonlinear magnetic behavior of the core and plate. Since both are made of ferromagnetic materials, their respective magnetic permeabilities are not constant, but are functions of the applied magnetomotive force. Assigning constant values for $\mu_c$ and $\mu_p$ in the calculation of the reluctance would lead to very inaccurate results for magnetic flux density unless the core and plate are sufficiently far from magnetic saturation. Another nonlinear effect is magnetic hysteresis, which causes a deviation from the governing law given by Equation 2.17. The magnetic flux is no longer dependent on only the current value of mmf, but also on the magnetization history as well.

3.1.2 Finite Element Analysis

Since magnetic circuit analysis involves many assumptions and approximations, finite element analysis is usually conducted. Therefore, COMSOL Multiphysics 5.3 is used to perform finite element simulations of the MFL system. The relevant physics is part of the AC/DC module that can compute electric and magnetic fields in systems where the wavelength is significantly larger than the dimensions of the studied geometry. In the case of the MFL system, it is required to model static and low frequency (0 – 10 Hz) magnetic fields, so the corresponding electromagnetic wavelength is greater than $3 \times 10^7$ m. Specifically, the Magnetic Fields physics interface is used as it is able to compute magnetic field and induced current distributions in and around coils, conductors, and magnets. It does that by solving Maxwell’s equations expressed in terms of the magnetic vector potential and the coil scalar electric potential as the dependent
variables. Two types of studies are considered. A stationary study is conducted when computing DC magnetic fields resulting from DC coil current, whereas a time dependent study is used for the AC case.

The model geometry, shown in Figure 3.3, consists of all the components of the MFL system including the electromagnet core, the coils, and the steel plate being inspected. An air domain is set up surrounding the entire system, and an infinite element air domain layer is added around the exterior of the model. The infinite element domain is treated as infinitely large by the software; therefore, the simulation solution is the same as that resulting from extending the dimensions of the original air domain. This is done in order to eliminate the need to increase the size of the air domain beyond the vicinity of the MFL system. [18]

When selecting the materials for the domains in the study, it is important to identify which material properties are required for the simulation. When solving for magnetic fields, the important properties are the electrical conductivity and the relative permeability. Since the steel plate and electromagnet core are made of magnetic materials, their relative permeabilities are not constants, so the constitutive relation between flux density and field strength within these two domains is given by the characteristic B-H curve of the material. This takes into account the nonlinear relationship between the magnetic flux density and the magnetic field in the plate and core. The electrical conductivity is also needed to account for the skin effect for AC magnetic fields. The material used to model the inspected plate is 1020 low carbon steel which has a conductivity of 8.41 MS/m. The B-H curve of 1020 steel is plotted in Figure 3.2. As for the coil domain, it takes as inputs the number of turns, the wire cross section area, and the excitation current or voltage.
Although the problem is 3D in nature, it can be reduced to 2D by assuming that the electromagnet and plate are infinitely wide. This assumption is valid in the center of the MFL system and significantly far from the edges. In both the 3D and the 2D cases, the model can be reduced by using planes of symmetry and imposing the proper boundary conditions. [19] The planes of symmetry are defined relative to the length, width and height of the model. The width is the direction normal to the 2D model plane. The height is normal to the inspected plate surface, while the length is parallel to the scanning direction.

The 3D geometry can be cut along a plane normal to the width with a magnetic insulation boundary condition. It can then be cut along a plane normal to the length with a perfect magnetic conductor boundary condition. The magnetic insulation and perfect magnetic conductor boundary conditions correspond to parallel and normal magnetic field constraints respectively. The resulting
3D geometry is a quarter of the original. For the 2D geometry, the cut is made along a line normal to the length with a perfect magnetic conductor boundary condition. This results in a 50% reduction in the size of the problem. It is noted that the boundary condition on the infinite element domain does not affect the solution since the boundary is modeled as being infinitely far from the MFL system. [18] Therefore, any of the two previously mentioned conditions would work.

![Figure 3.3: 2D and 3D COMSOL Models](image)

### 3.2 Electromagnet Core Design

#### 3.2.1 Material Selection

In order to select a suitable material for the MFL electromagnet core, several material properties must be considered. Since the core should perform the role of a low reluctance path for the magnetic field in and out of the inspected plate, its material would ideally have the highest practical magnetic permeability. The low reluctance path is required to minimize the magnetomotive force needed to attain a certain magnetic flux. This leads to a decrease in the current and number of coil turns required for the electromagnet.
Another factor to consider is the saturation induction of the core material. When the electromagnet is powered, the core should ideally be operating in a region of maximum permeability which is typically at a level of magnetization well below that of its magnetic saturation region. In other words, for any core material and desired magnetic flux, the core thickness should be high enough to result in a magnetic flux density lower than the saturation induction. Therefore, using a material with a higher saturation induction will allow the use of a thinner core.

Since the electromagnet may be powered by an alternating current (AC) profile, minimizing core loss resulting from hysteresis and eddy current is also important. To reduce hysteresis loss, the core material should have low retentivity and coercivity. As for eddy current loss, it is significantly decreased by using a material with high resistivity or using laminations in order to limit current amplitudes. The operating frequency range of the electromagnet is a key factor when choosing the core material because core loss is a function of frequency. As stated previously, for the MFL system, the range of operating frequencies is approximately $0 - 10$ Hz.

In section 2.1.3, various soft magnetic materials were introduced. They included low carbon steel, silicon steel, nickel-iron alloys, cobalt-iron alloys, amorphous cores, powder cores, and ferrite cores. The relative magnetic permeability of powder cores and ferrite cores is significantly less than that of the other choices. It is only justifiable to use them when core losses need to be kept low at frequencies in the kHz and MHz regions. Since the MFL electromagnet is to be operated at a much lower frequency range, these two material categories can be ruled out.

Of the remaining choices, low carbon steel is not suitable for AC applications because of the high associated core losses as discussed in section 2.1.3. Because of their relatively high price per pound, nickel-iron alloys, cobalt-iron alloys, and amorphous cores are usually used for
applications where the properties of silicon steel are inadequate. Nickel-iron alloys have the highest relative magnetic permeability, but it comes at the expense of reduced saturation induction compared to silicon steel. Cobalt-iron alloys have the highest saturation induction, but the relative magnetic permeability is lower than that of silicon steel. Amorphous cores are only used for special applications where even core lamination does not achieve the desired efficiency at the operating frequency. Figure 3.4 is a typical selection matrix for soft magnetic materials.

![Figure 3.4: Typical Selection Matrix for Soft Magnetic Materials](image)

Therefore, silicon steel is the most appropriate material category for the MFL electromagnet core as it provides a good balance between magnetic permeability, saturation induction, core losses, and cost. As discussed in section 2.1.3, silicon steel, which is also known as electrical steel, can be either grain oriented or non grain oriented. Grain oriented electrical steel has a higher relative permeability and lower losses in the rolling direction, but is also more expensive. Both types also come in different grades and lamination thicknesses. More expensive grades have higher relative permeability and less core loss, but the saturation induction is lower. As for core laminations, thinner ones result in lower losses, but are costlier. Figure 3.5 presents the properties of the various available electrical steel grades.
Non grain oriented electrical steel is a more economical option than its grain oriented counterpart. The non oriented grade with highest magnetic permeability and lowest losses is selected. Therefore, the core material of choice is M15 non grain oriented electrical steel (2.7% Si). This is typically supplied in 0.014 in (0.35 mm) thick laminations that are coated with insulation for use in transformers as part of electrical transmission & distribution systems. Unlike grain oriented electrical steel, the magnetic properties of this material are the same in all directions. It has a maximum relative permeability of 8000, and its saturation induction is 2.01 T. Also, its
density is 7.65 g/cm³ and its electrical conductivity is 2 MS/m. [21] Using the relative magnetic permeability and electrical conductivity, the skin depth at 10 Hz is calculated to be 1.26 mm which is significantly higher than the 0.35 mm lamination thickness. Therefore, the skin effect is negligible in the laminated core of the MFL system, where the maximum frequency is 10 Hz. The B-H curve of the material is plotted in Figure 3.6.

![B-H Curve for M15 Non Grain Oriented Electrical Steel](image)

**Figure 3.6: B-H Curve for M15 Non Grain Oriented Electrical Steel**

### 3.2.2 Core Geometry

The electromagnet core plus a portion of the inspected steel plate should form a complete magnetic circuit. Therefore, an inverted U-shaped core is the logical choice, with the magnetic poles at the ends of both sides or legs of the electromagnet. As for the required thickness of the core, it can be calculated using the saturation induction of the 1020 low carbon steel plate and the maximum permeability induction of the M15 non grain oriented electrical steel core. Assuming
that the magnetic flux loss is negligible in the magnetic circuit, the magnetic flux in the core and the plate should be equal. In other words,

\[ \int B_{\text{core}} \, dA_{\text{core}} = \int B_{\text{plate}} \, dA_{\text{plate}}, \]  

(3.2)

where \( B \) is the magnetic flux density and \( A \) is the cross sectional area. The width of the cross sectional area of the plate that is magnetized is larger than that of the core, but is difficult to calculate. Also, the flux density in the plate is maximum at the center and decreases towards the edges of the electromagnet. This decrease is also difficult to estimate. To simplify the calculation of the flux in the plate, the maximum flux density is used instead of the average, and this is compensated for by taking the width of the plate to be equal to the core which is an underestimation. This simplifies the relation to

\[ t_{\text{core}} = \frac{B_{\text{plate, max}}}{B_{\text{core}}} \, t_{\text{plate}}, \]  

(3.3)

where \( t \) is the thickness. It is required for the flux density in the steel plate to be in the magnetic saturation region and that of the electromagnet core to be in the region of maximum magnetic permeability. Therefore, from the B-H curves of the two materials plotted in Figure 3.7, it can be deduced that the flux density in the plate is around 1.6 T, while the flux density in the core is around 0.8 T. In other words, the ratio of the maximum flux density in the plate to the flux density in the core is 2. The thickness of the inspected plate is 0.25 in (6.35 mm); therefore, the thickness of the electromagnet core should be 0.5 in (12.7 mm).
The other core dimensions include the length, width, and height. All three are defined similarly to those in the COMSOL model discussed in the previous section. The width of the core is 12 in (304.8 mm) and was specified by the end user as the intended tank floor scanning span. The length and height of the electromagnet, on the other hand, are determined using finite element simulations with the goal of minimizing the magnetic flux density at the location of the hall effect sensors when no defect is present in the inspected steel plate (this in effect minimizes the background noise of the system). A parametric sweep study is performed where the magnet length and height are varied between 4 – 24 in (101.6 – 609.6 mm) and 2.5 – 12.5 in (63.5 – 317.5 mm) respectively. For each combination, the coil placement is also changed where it is either entirely wound on the top part or each half is wound on the sides of the electromagnet.
Figure 3.8: Background Flux Density (T) for Different Coil Configurations: Top vs. Sides Respectively

Figure 3.8 presents the simulation results for a single combination of length and width. The background magnetic flux density at the sensor location (vicinity of the coordinates [0, -1.75]) is much lower when the coil is wound on the electromagnet legs rather than the top. This is true for all other dimension combinations as well. Of the magnet dimension combinations that exhibit low background magnetic flux density, it is optimal to choose the smallest ones. That results in the lowest weight and amount of material needed for the core. More importantly, it decreases the coil current and amount of power needed because a smaller portion of the inspected steel plate needs to be magnetized when the length of the electromagnet is decreased. Based on the simulations, the
optimal length and height are 12 in (304.8 mm) and 4.5 in (114.3 mm) respectively. Figure 3.9 shows the background field for the chosen dimensions.

![Figure 3.9: Background Magnetic Flux Density (T) Corresponding to the Chosen Core Dimensions](image)

3.3 Electromagnet Coil Design

3.3.1 Material, Gauge, and Number of Turns

The property of main importance when choosing a suitable material for electromagnet coils is electrical conductivity. There are four metals that have significantly higher electrical conductivity than the others. [22] In order of decreasing conductivity, they are silver, copper, gold, and aluminum, as shown in Figure 3.10. These materials have the added advantage of being nonmagnetic, so the skin effect is not as prominent as when using iron for example. Since silver and gold are relatively expensive compared to copper and aluminum, only the latter two are considered for electromagnet coils in the form of magnet wire.
The electrical conductivity of copper is 59% higher than that of aluminum. Therefore, for the same rated current, the wire of an aluminum coil should have a cross sectional area that is of the order of 59% larger than that of a copper coil. Hence, copper is the preferred choice for most applications where minimizing size and amount of material is preferred. However, the density of copper is 8.96 g/cm³ compared to 2.7 g/cm³ for aluminum, so the larger amount of aluminum needed to compensate for its lower conductivity would still weigh less than a similar current rated copper coil. The price of copper is also higher; hence aluminum is a suitable choice for large scale applications where weight and cost savings are important. For the MFL system, space limitations and winding difficulty are the main factors; these both worsen with increasing winding cross sectional area. Added to the fact that copper winding is more readily available than its aluminum counterpart, copper magnet wire is chosen.

Copper magnet wire is available as single conductor wire or litz wire. Litz wire consists of multiple strands that are insulated, bundled, and twisted together. It is used at frequencies in the kHz region to reduce the skin and proximity effects. The skin effect causes an exponential decrease in the current density in a wire with increasing distance from the surface; the proximity effect refers to the altered current distribution in a wire due to eddy currents induced by neighboring wires with AC current such as in a coil. [23] Both effects increase the effective resistance of a wire and become significant with increasing frequency, as demonstrated in Figure 3.11 and Figure 3.12.

![Figure 3.10: Electrical Conductivity of Metals [22]](image)
For the MFL system, the maximum operating frequency is 10 Hz, at which the skin depth in copper is 20.55 mm. Because the radius of single conductor copper wire for our system is significantly less than 20.55 mm, litz wire is not required.

Figure 3.11: Ratio of AC/DC Resistance Due to Skin Effect

Figure 3.12: Ratio of AC/DC Resistance Due to Proximity Effect [25]
Single conductor magnet wire comes in many forms including round, square, and rectangular cross sections. Square and rectangular wires are used to increase the fill factor, which is the ratio of the cross sectional area of the conducting wires to the total cross sectional area taken by the coil. They allow denser packing of windings with smaller air gaps and better heat conductivity between the coil layers. For round magnet wire, the highest theoretical fill factor is 90% and is achieved using orthocyclic winding where wires in each layer sit in the grooves between the wires in the layer below. Since square and rectangular wire are relatively expensive, round magnet wire is the usual choice unless space efficiency becomes an issue. Also, round wire is available in a much wider size range; therefore, round magnet wire is selected.

In order to determine the cross sectional area and number of turns of wire required for the electromagnet coil, the maximum magnetomotive force that needs to provided by the MFL electromagnet should be calculated. According to the design requirements for our magnet, this corresponds to the magnetomotive force that magnetically saturates a 0.25 in (6.35 mm) thick 1020 steel plate with a liftoff of 0.375 in (9.525 mm). The liftoff accounts for wheel clearance and any protective coating applied on fuel tank floors. Using COMSOL Multiphysics, a finite element simulation was conducted with a parametric sweep where the magnetomotive force was varied between 1000 – 10000 Ampere-turns in steps of 1000. The results in Figure 3.13 show that the increase in magnetic flux density becomes marginal when the magnetomotive force surpasses 6000 Ampere-turns, an indication of the onset of magnetic saturation in the plate.
Figure 3.13: Magnetic Flux Density in the Inspected Plate vs Magnetomotive Force

There is a wide range of cross sectional areas available for magnet wire based on the American Wire Gauge (AWG) system. As a rule of thumb, the recommended current carrying capacity of bundled copper wire, such as in a coil, is 700 circular mils per ampere. [26] A circular mil is the area of a circle with a diameter of 0.001 in. In metric units, the current carrying capacity corresponds to a value of 3 A/mm². This is very conservative compared to a wire in free air because in coil form, heat transfer via convection to the surrounding air is very limited. Exceeding the recommended current carrying capacity can lead to temperatures that exceed the thermal rating of the magnet wire insulation coating.

With the various wire cross sectional areas, there is a wide range of possible combinations of current and number of turns that can achieve the desired magnetomotive force of 6000 Ampere-turns. A large area allows operation at a high current and low number of turns. This corresponds
to low resistance and inductance, because resistance is proportional to length and inversely proportional to area while inductance is proportional to the ratio of number of turns to current. The circuit for such a system would require low voltage levels. However, thicker wire is more difficult to wind because the bending stiffness or flexural rigidity is proportional to the fourth power of the cross section radius. Also, high current circuitry is challenging because the power loss and heat dissipation are proportional to the square of the current.

Using thin wires, on the other hand, would require a higher number of turns because of current magnitude limitations. Both the wire resistance and inductance are relatively high, so it is required to connect several sections of the coil in parallel to lower the required voltage. This makes the process of connecting the coil to the power supply more difficult and increases the chance of faulty wiring. Also, thinner wire can take less tension when straightening and winding.

The choice of wire is a balance among all the mentioned factors. Power circuitry for non-industrial use is typically rated for less than 10 A without cooling. This rules out the use of wire that is thicker than 12 AWG. In order to minimize the number of turns required, the thickest wire within that range is used, namely 14 AWG. Its diameter is 1.628 mm, and its cross sectional area is 2.08 mm$^2$ which corresponds to a current carrying capacity of around 6 A. This allows a coil design with around 1000 turns.

There are different standard thermal classes of insulation coatings available, ranging from 105 to 240°C as shown in Figure 3.14. [27] The most common insulation coating materials are polyurethane rated at 155°C and polyester overcoated with polyamide-imide rated at 200°C. The insulation coating rated at 200°C is chosen because 14 AWG magnet wire and larger is generally only available with high thermal rating. Its thickness is 0.081 mm and its properties and specifications are based on the MW35C magnet wire insulation standard.
3.3.2 Cooling

As discussed in section 2.3.3, cooling is an essential design consideration because of the heat dissipated by electromagnets. DC fans are the easiest and most cost-effective cooling method. The most important factor when choosing a cooling fan is its maximum flow rate that is directly related to the convection heat transfer coefficient. The flow rate is in turn directly related to fan size and speed. Typical fan sizes include 80, 92, 120, 140, and 200 mm. The larger the fan, the less speed is required to achieve the desired flow rate. For a given fan, the flow rate is inversely proportional to the static pressure. A fan with higher power operates at a higher combination of flow rate and static pressure.

In order for the fans not to be obstructed by the top part of the electromagnet, they must be smaller than 4 in (101.6 mm). The largest suitable standard fan size is 92 mm. Although a higher

<table>
<thead>
<tr>
<th>Insulation Type</th>
<th>Thermal Class</th>
<th>Diameter (mm)</th>
<th>Diameter (in.)</th>
<th>AWG Wire Size Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>Polyurethane</td>
<td>120°C, 130°C, or 155°C</td>
<td>0.08 to 1.00</td>
<td>0.0031 to 0.0394</td>
<td>20-18</td>
</tr>
<tr>
<td>Polyester</td>
<td>&gt;155°C</td>
<td>0.08 to 1.6</td>
<td>0.031 to 0.063</td>
<td>20-14</td>
</tr>
<tr>
<td>Polyester-imide</td>
<td>180°C</td>
<td>0.1 to 1.00</td>
<td>0.0039 to 0.0394</td>
<td>38-18</td>
</tr>
<tr>
<td>Polyamideimide</td>
<td>220°C</td>
<td>0.1 to 1.6</td>
<td>0.0394 to 0.063</td>
<td>38-14</td>
</tr>
<tr>
<td>Self-bonding polyurethane</td>
<td>130°C</td>
<td>0.08 to 1.2</td>
<td>0.0031 to 0.0472</td>
<td>20-16</td>
</tr>
<tr>
<td>Self-bonding Polyester-imide</td>
<td>180°C</td>
<td>0.1 to 0.8</td>
<td>0.0039 to 0.031</td>
<td>38-20</td>
</tr>
<tr>
<td>Polyurethane overcoated with polyamide</td>
<td>130°C or 155°C</td>
<td>0.08 to 1.6</td>
<td>0.0031 to 0.063</td>
<td>20-14</td>
</tr>
<tr>
<td>Polyester overcoated with polyamide</td>
<td>&gt;155°C or 180°C</td>
<td>0.1 to 1.6</td>
<td>0.0039 to 0.063</td>
<td>38-14</td>
</tr>
<tr>
<td>Polyester-imide overcoated with polyamide-imide</td>
<td>200°C</td>
<td>0.1 to 1.6</td>
<td>0.039 to 0.063</td>
<td>38-14</td>
</tr>
</tbody>
</table>

Figure 3.14: Table of Magnet Wire Insulation Types [27]
rated fan speed is always beneficial in terms of increasing flow rate, high noise levels and power consumption can become prohibitive. Therefore, 92 mm 12 V fans with a speed of 4000 RPM were selected. The resulting noise level and rated power consumption are 43 dB and 6.96 W respectively. The maximum flow rate and static pressure are 2.54 m$^3$/min and 100 Pa respectively.

3.4 Mechanical Design of Prototype

3.4.1 Electromagnet Assembly

Since the magnetic core is made from laminations, the mechanical design of the assembly is important for system integrity. As discussed in section 3.2.2, the core dimensions are 12 x 12 x 4.5 in (304.8 x 304.8 x 114.3 mm), and it has a thickness of 0.5 in (12.7 mm). The laminations are 0.014 in (0.35 mm) thick and are to be stacked in the thickness dimension. Therefore, the number of laminations along the core thickness is 36. The electrical steel laminations are obtained in sheets that are cut by wire EDM to 180 rectangles of size 12 x 4 in (304.8 x 101.6 mm). That is the exact size of the sides of the core, each consisting of 36 sheets. As for the 12 x 12 in (304.8 x 304.8 mm) top, 108 sheets are stacked in 3 adjacent stacks of 36 each. With a density of 7.65 g/cm$^3$, the mass of the magnetic core is approximately 15 kg.

A frame is designed to hold all the laminations together. Since AC magnetic fields are featured in the MFL system, the frame cannot be made of electrical conductors and magnetic materials, both of which can cause interference and field distortion due to eddy currents. Therefore, plastics are the suitable choice. The factors to determine the type of plastic to be selected are the tensile strength and machinability. Plastics that have high tensile strength include acrylic, polycarbonate, acetal, and nylon as shown in Figure 3.15. [28] There are other high strength plastics; however, they are used for high temperature applications and are more expensive. To minimize the cost of machining, acetal is selected for its superior machinability rating compared
to the other three. In fact, acetal has the highest machinability among all plastics. [28] Therefore, acetal bars are cut to size and are assembled together along with the laminations using bolts and nuts. Acetal has a density of 1.41 g/cm³, and the mass of the frame is approximately 4 kg.

<table>
<thead>
<tr>
<th>Clear</th>
<th>Temperature Range</th>
<th>-450°F</th>
<th>0°F</th>
<th>550°F</th>
<th>Tensile Strength, psi</th>
<th>Impact Strength, ft-lbs/in.</th>
<th>Machinability, % (Based on Acetal = 100%)</th>
<th>Faster cutting, smoother finish</th>
<th>Relative Cost</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cellulose</td>
<td>Easy to form</td>
<td></td>
<td></td>
<td></td>
<td>6,300</td>
<td>4.5</td>
<td>Not Rated</td>
<td>$</td>
<td></td>
</tr>
<tr>
<td>PETG</td>
<td>Easy to form</td>
<td></td>
<td></td>
<td></td>
<td>7,100</td>
<td>1.7</td>
<td>Not Rated</td>
<td>$</td>
<td></td>
</tr>
<tr>
<td>Acrylic</td>
<td>Scratch and UV resistant</td>
<td></td>
<td></td>
<td></td>
<td>10,000</td>
<td>0.3</td>
<td>65%</td>
<td>$$</td>
<td></td>
</tr>
<tr>
<td>Polycarbonate</td>
<td>Impact resistant</td>
<td></td>
<td></td>
<td></td>
<td>8,900</td>
<td>5+</td>
<td>50%</td>
<td>$$</td>
<td></td>
</tr>
<tr>
<td>Multipurpose</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>$</td>
<td></td>
</tr>
<tr>
<td>Polypropylene</td>
<td>Economical</td>
<td></td>
<td></td>
<td></td>
<td>4,100</td>
<td>0.9</td>
<td>40%</td>
<td>$</td>
<td></td>
</tr>
<tr>
<td>Polystyrene</td>
<td>Economical, easy to form</td>
<td></td>
<td></td>
<td></td>
<td>2,400</td>
<td>1.5</td>
<td>56%</td>
<td>$</td>
<td></td>
</tr>
<tr>
<td>Polyethylene (LDPE)</td>
<td>Flexible</td>
<td></td>
<td></td>
<td></td>
<td>1,400</td>
<td>5+</td>
<td>Not Rated</td>
<td>$</td>
<td></td>
</tr>
<tr>
<td>Polyethylene (HDPE)</td>
<td>Moisture resistant</td>
<td></td>
<td></td>
<td></td>
<td>3,800</td>
<td>1.1</td>
<td>55%</td>
<td>$</td>
<td></td>
</tr>
<tr>
<td>UHMW, VHMW, Polyethylene</td>
<td>Slippery</td>
<td></td>
<td></td>
<td></td>
<td>4,300</td>
<td>5+</td>
<td>70%</td>
<td>$$</td>
<td></td>
</tr>
<tr>
<td>PVC, CPVC</td>
<td>Chemical resistant</td>
<td></td>
<td></td>
<td></td>
<td>7,000</td>
<td>0.7</td>
<td>50%</td>
<td>$$</td>
<td></td>
</tr>
<tr>
<td>AERIS</td>
<td>Impact resistant, easy to form</td>
<td></td>
<td></td>
<td></td>
<td>4,200</td>
<td>5+</td>
<td>85%</td>
<td>$$</td>
<td></td>
</tr>
<tr>
<td>Acetal</td>
<td>Easy to machine</td>
<td></td>
<td></td>
<td></td>
<td>10,000</td>
<td>1.0</td>
<td>100%</td>
<td>$$</td>
<td></td>
</tr>
<tr>
<td>Nylon</td>
<td>Wear resistant</td>
<td></td>
<td></td>
<td></td>
<td>11,200</td>
<td>0.6</td>
<td>75%</td>
<td>$$</td>
<td></td>
</tr>
<tr>
<td>Polyester</td>
<td>Moisture resistant</td>
<td></td>
<td></td>
<td></td>
<td>11,600</td>
<td>0.7</td>
<td>60%</td>
<td>$$</td>
<td></td>
</tr>
<tr>
<td>Noryl PPO</td>
<td>Electrical insulating</td>
<td></td>
<td></td>
<td></td>
<td>9,200</td>
<td>3.5</td>
<td>56%</td>
<td>$$</td>
<td></td>
</tr>
</tbody>
</table>

**Figure 3.15: Tensile Strength and Machinability of Plastics [28]**

As for the coil, it is divided equally and symmetrically among the two sides of the electromagnet, as mentioned in the previous section. In order to have flexibility and greater control
of the overall impedance of the coil, it is beneficial to further divide the coil to allow for multiple parallel connection arrangements. The impedance of a coil decreases when several groups are connected in parallel instead of having all the turns in series. This is beneficial when operating at higher frequencies because the coil impedance increases with frequency, and a mechanism is needed to keep the required voltage to within the range of the power source. Therefore, each side of the electromagnet consists of 12 coils of 40 turns each wound concentrically. This results in 480 turns on each side or 960 turns in total.

Before starting the winding process, polyester straps are used to cover the sharp corners of the core that could potentially damage the magnet wire insulation. Also, to aid in the winding process, the magnet wire is passed through a wire straightener tool after unwinding from the spool. The tool consists of three nylon cylinders in a triangular configuration, where the wire is forced in the gap between them. In order to monitor the temperature of the coils, type K thermocouples are placed between each coil layer. The mass of the coils is approximately 12 kg, which brings up the total mass of the electromagnet to 31 kg. The final electromagnet design is shown in Figure 3.16.

Figure 3.16: Final Electromagnet Prototype
3.4.2 Cart for System Mobility

In order to move the MFL system while inspecting a steel plate, a cart with wheels is designed. The maximum force applied on the cart corresponds to the dc magnetic saturation case, which was previously established as a magnetomotive force of 6000 Ampere-turns. Using a finite element simulation, the magnetic force between the electromagnet and the inspected plate with a liftoff of 0.375 in (9.525 mm) is 471 N. Combined with the weight of the electromagnet, the total force is 775 N. Similar to the electromagnet frame, the cart cannot be made of magnetic or conductive materials to avoid magnetic field interference due to eddy currents. It is important to use as thin a sheet as possible to achieve the low liftoff. Therefore, the strongest insulator material should be selected, which is the composite material garolite that is made of epoxy resin with fiberglass reinforcement. The most common grade of garolite is G10/FR4, which has a tensile strength of 240 MPa. [29]

The size of the sheet is 18 x 36 in (457.2 x 914.4 mm), and the wheels are placed along the longer side. In other words, the scanning direction is parallel to the longer side. The width of the sheet is decreased as much as possible to just fit the electromagnet. This is done to decrease the deflection of the sheet between the wheel supported sides. The sheet is significantly longer than the electromagnet to accommodate the power circuitry. As for the thickness of the sheet, it is desired to minimize the sum of thickness and deflection. Using basic beam theory, the deflection of the sheet at maximum load is calculated for different sheet thickness ranging from 0.125 in (3.175 mm) to 0.375 in (9.525 mm). A factor of safety of 1.3 is applied to the load to account for variation in material properties. The results in Figure 3.17 show that the minimum sum of thickness and deflection occurs at a sheet thickness of 0.1875 in (4.7625 mm).
Figure 3.17: Sum of Sheet Thickness and Deflection vs. Sheet Thickness

Six edge-mount bracket casters are used to support the setup. They have 2 in (50.8 mm) polypropylene wheels and are mounted so as to have a 0.1875 in (4.7625 mm) clearance between the bottom of the sheet and the inspected plate. This results in the overall liftoff of 0.375 in (9.525 mm) that is incorporated in the MFL system design. The cart is shown in Figure 3.18.
3.5 Electric Circuit

3.5.1 Relationship Between Voltage, Current, and Magnetic Flux Density

In this section, the design of the circuit powering the electromagnet coil is discussed. The first thing to note is that the total resistance of all the coils in series is 6.24 Ω. Since the coil is divided into 24 groups, the overall resistance can be decreased by connecting them in parallel. The inductance is dependent on the magnetic permeability of the core and plate, so it is not constant. The inductance of the coil decreases as the magnetomotive force is increased as shown in Figure 3.19. At the onset of magnetic saturation in the plate, the total inductance, calculated using finite element simulation, is 0.7 H.
To estimate the power required for our system, there are two things that need to be considered. Using the number of turns of coil as 960 turns, the magnetic flux density in the inspected plate as a function of the magnetomotive force is revisited in terms of current. Using finite element simulation, the coil current required to achieve DC saturation in the plate is around 6 A as shown in Figure 3.20. With a coil resistance of 6.24 Ω, the power consumption in DC operation is 225 W.

Figure 3.19: Coil Inductance vs. Current
The second factor to consider is the reactance of the coil during AC operation. The reactance of the coil increases with frequency and is equal to \(2\pi f L\), where \(f\) is the frequency in Hertz and \(L\) is the inductance in Henry. The impedance of the coil is then given by

\[
Z = \sqrt{R^2 + (2\pi f L)^2}, \quad (3.4)
\]

where \(Z\) and \(R\) are the impedance and resistance respectively. [9] It is plotted in Figure 3.21. Therefore, for the same current amplitude, the required voltage and power increase with frequency.
When operating at 10 Hz, which is the maximum intended operation frequency of the electromagnet, the impedance is 44.4 ohms. This is an increase by a factor of 7.1 compared to DC resistance. If the coils are to be operated at the same rms current as the DC case, 6 A, the power required would be 1600 W. To account for this scenario, a power supply of 2200 W is chosen for the MFL system. The power supply voltage is selected to be 12 V in order to maximize compatibility with power circuitry that is usually designed for either 12 V, 24 V, or 48 V. The choice of voltage would also help in easily transitioning to using widely available 12 V batteries for increased portability of the MFL system.

3.5.2 H-Bridge Circuit and PWM Control
In order to have the ability to control the magnitude and direction of the applied voltage, h-bridge circuits are used. An h-bridge circuit consists of four transistors, where depending on their digital voltage input, the voltage of a power supply or battery can be output in either polarity.

The h-bridge circuits are powered by a power supply or battery and controlled using digital signals from a microcontroller. The digital signals include ones that determine the output voltage magnitude and others that determine the output voltage direction. The signals that control the direction are simple digital signals that are switched between high and low states. The voltage magnitude, on the other hand, is controlled by a PWM signal. A PWM signal is a pulse train whose duty cycle determines the percentage of time within a period where the transistors of the h-bridge are turned on or conducting. The signal is also digital and its duty cycle is simply the ratio of the time at high state to the time at low state in one period. The output of the h-bridge is directly proportional to the duty cycle. At 0% duty cycle, the output is 0V, while at 100% duty cycle, the output is equal to the supplied voltage. By varying the PWM duty cycle, between 0 and 100, the output can be set to any voltage less or equal to that of the power supply or battery. A few examples of PWM signals are given in Figure 3.22.
The h-bridges used for the MFL system are the VNH2SP30 h-bridge motor drivers. This h-bridge uses 5 V logic and is capable of operation at 12 V with a continuous current rating of 14 A with no additional cooling. It supports PWM signal frequencies up to 20 kHz. Multiple h-bridges are used to allow independent control of coil groups placed in parallel. This also limits the current handled by a single h-bridge to the previously mentioned intended coil current of 6 A, which is well below its current rating.

The selected microcontroller board is the Arduino Mega 2560, that is programmed using the Arduino programming language. It features the ATmega2560 microcontroller that uses 5 V logic and a 16 MHz clock frequency. Its PWM output properties are configurable, with the main properties being the frequency of the PWM signal and the duty cycle resolution.

When operating in DC, the duty cycle of the PWM signal is kept constant to achieve the required constant current. Therefore, all that is required of the PWM frequency is for it to be
significantly higher than the cutoff frequency of the MFL system. The cutoff frequency is the ratio of resistance to inductance, which is around 8.9 Hz. However, for the AC case, the duty cycle should be updated continuously at around 100 times per period to achieve the different levels of instantaneous current in a smooth fashion, and the output voltage should be switched every half period. An example is shown in Figure 3.23. Operating the electromagnet at the specified maximum frequency of 10 Hz would then require updating the duty cycle at a rate of 1 kHz. Therefore, the PWM frequency should be higher than 1 kHz.

The duty cycle resolution refers to the number of discrete steps between 0% and 100% duty cycle. This determines the smallest voltage increment the system can achieve. A better resolution allows finer control of the current. For a given microcontroller, the PWM frequency and the duty cycle resolution are inversely proportional and their product is half the clock frequency. For example, with a clock frequency of 16 MHz, the Arduino will have a duty cycle resolution of 8000 for a PWM frequency of 1 kHz.

Since the PWM frequency is required to be higher than 1 kHz to meet the sampling rate requirements and lower than 20 kHz to be compatible with the h-bridges, a frequency of 8 kHz is selected. The resulting duty cycle resolution is 1000, which is equivalent to 0.1%.
The electromagnet coil resistance increases with temperature, so current feedback control is used to maintain the desired current. This is implemented using the ACS 712 current sensor. It has a sensitivity of 100 mV/A, a range of -20 to 20 A, and a frequency bandwidth of 80 kHz. The sensor output is read using the 10-bit ADC of the Arduino, which results in a resolution of around 50 mA. Once the current drops below the intended magnitude due to an increase in resistance, the PWM duty cycle is proportionally increased to compensate.
4 Validation of Electromagnet Design

Before starting to inspect steel plate samples and collecting MFL results, it is important to validate the specifications of the electromagnet experimentally. This ensures that all the properties of the prototype are as designed and intended. The properties to be investigated are the coil temperature, the force between the electromagnet and the inspected plate, the magnetic flux density produced by the electromagnet, and the AC coil current waveform. The experimental results are also compared to the finite element simulations that were used in the design process.

4.1 Coil Temperature

In section 3.4.1, it was mentioned that thermocouples are placed between the coil layers in order to monitor the temperature during operation if needed. Each side of the electromagnet has 12 coil groups that are wound concentrically. Therefore, for each side, there are 12 thermocouples, one for each coil group. The importance of measuring the temperature stems from the fact that the coil resistance increases with temperature because the resistivity of copper increases. This results in a drop in current if no feedback control is implemented. Therefore, current sensors are used to measure the coil current as well.

In order to investigate the increase in coil temperature and the resulting decrease in current, two experiments are conducted. In the first one, the electromagnet is run in DC mode at an initial current of 6 A and no cooling fans are used. In the second experiment, the electromagnet is run at the same current, but the cooling system is implemented. The cooling system consists of four fans, where two fans are used for each side. Both experiments are run without feedback control in order to monitor the current drop with time. Temperature and current readings were taken every minute.
The temperature of the 12 coil layers are not equal, so the maximum temperature is used to monitor the change with time. It is noted in Figure 4.1 that when cooling is implemented, the maximum coil temperature decreases from 103 °C to 59 °C. This is a substantial drop of 44 °C. In both cases, the maximum temperature is below the rated temperature of the magnet wire insulation coating (200 °C). Being below the rated temperature is not satisfactory, because the power needed to achieve a certain current in the coil increases with resistance which itself increases with temperature. Therefore, the lower the temperature, the more electrically efficient the magnet would be.
As mentioned earlier, both experiments start with a coil current of 6 A. Figure 4.2 shows that the current in the coil with no cooling decreases to 4.86 A by the time it reaches 200 minutes of operation, which is a drop of 19%. On the other hand, when fans are used, the coil current decreases to a final value of 5.64 A, which is a drop of only 6%. Since the current drop is not completely eliminated, current feedback control is still necessary to compensate for the increase in coil resistance.
The final steady state temperature of all the 12 coil layers is measured in order to plot the temperature profile. The coil layers are labeled from 1 to 12 starting with the innermost layer in contact with the core and ending with the outermost layer in contact with the surrounding air. The temperature profiles in Figure 4.3 show that the maximum coil temperature corresponds to the central coil layers. As expected, the temperature of the outermost coil (coil 12) in contact with air is the lowest. The lower temperature as you move inwards from the center is due to the magnetic core acting as a heat sink. From the temperature profile, the average coil temperature is calculated to be 94 °C and 51 °C for the uncooled and cooled cases respectively. This is a decrease of 43 °C and explains why the current drop is significantly lower when cooling is implemented.
4.2 Force Between the Electromagnet and Inspected Plate

In this section, the magnetic force between the electromagnet and inspected plate is measured experimentally and compared with finite element simulations. The experiment is conducted by placing a scale between the electromagnet and plate and varying the current in steps of 1 A up to a value of 6 A. Instead of having the electromagnet on a cart, it is put on a scale with a 36 x 12 x 0.25 in (914.4 x 304.8 x 6.35 mm) steel plate underneath. The thickness of the scale results in a 0.8 in (20 mm) liftoff between the electromagnet and plate. This is more than double the designed liftoff for the prototype. Since that is the only liftoff at which the force measurements can be taken, the comparison between experimental results and finite element simulations is done with a 0.8 in (20 mm) liftoff. By showing that the simulations are valid, the cart design would be validated because it is based on the magnetic forces calculated using simulations.

![Magnetic Force vs. Current](image)

Figure 4.4: Magnetic Force Between the Electromagnet and Plate at a liftoff of 0.8 in
Figure 4.4 shows that the experimental results are in good agreement with the simulations, and the maximum difference is within 9%. This error can be attributed to the difference between the material properties of the steel plate and that of steel in the COMSOL software. The B-H curve of an actual steel plate cannot be expected to exactly match the one used in the finite element simulations.

The design of the cart was done for a liftoff of 0.375 in (9.525 mm), but the experiment could only be done with a liftoff of 0.8 in. Since the experimental results and corresponding simulations are in agreement, it shows that the magnetic force in the COMSOL model matches that of the built electromagnet. Since the cart design was based on the force calculated by COMSOL simulations, it is validated.

![Figure 4.5: Magnetic Force Between the Electromagnet and Plate at a liftoff of 0.375 in](image-url)
In order to calculate the magnetic force for the designed liftoff of 0.375 in (9.525 mm), finite element simulations are run with a current sweep between 1 A and 6 A. This is the liftoff at which the prototype is designed to operate when the electromagnet is placed on its cart. The results are shown in Figure 4.5.

4.3 Magnetic Flux Density Validation for the Electromagnet in DC

The magnetic flux density caused by the electromagnet inside the steel plate cannot be measured experimentally. This is the variable of interest because it is important for the electromagnet to be able to magnetically saturate the inspected plate. Therefore, the COMSOL model is compared to experimental results for magnetic flux density measured in air. After showing that the finite element simulation is valid, the electromagnet design would be validated because the steel plate in the COMSOL model exhibits magnetic saturation.

Two experiments are conducted. In the first, the flux density produced by the electromagnet is measured without the presence of a steel plate. A gaugemeter (AlphaLab Magnetometer) is used, and its probe is placed at a distance of 0.8 in (20 mm) from one of the legs of the electromagnet. The legs of the electromagnet are essentially the poles. The coil current is varied between 1 A and 6 A in steps of 1 A. A finite element simulation is run with no steel plate in order to compare with the experimental results. The experimental and simulation results in Figure 4.6 agree to within 3%. This is an indication that the COMSOL software model is properly simulating the electromagnet.
In the second experiment, the magnetic flux density is measured using the same procedure of the first test. However, a 0.25 in (6.35 mm) thick steel plate is placed below the electromagnet at a liftoff of 0.8 in (20 mm). The gaussmeter probe is placed on the surface of the steel plate under the scale. Figure 4.7 shows that the margin of error is less than 4%. Since the magnetic flux density inside the steel plate cannot be measured, this is the only indication of the accuracy of the magnetic circuit model in COMSOL. By verifying that the COMSOL model is reliable, it can be deduced that, since the simulations show that the steel plate reaches magnetic saturation, the built electromagnet is able to magnetically saturated the inspected plate.
A few experiments are run to test the electromagnet in AC. Specifically, it is required to test whether the electromagnet is capable of operation up to a frequency of 10 Hz. First, a 1 Hz voltage with an amplitude of 12 V is applied to the coil with four groups in parallel. The equivalent would be applying a 48 V signal to the whole coil where all connections are in series. The electromagnet is placed on the cart with a steel plate underneath. The coil current was measured using a current sensor. This is also simulated in COMSOL for comparison. The data was recorded for 5 seconds. Both the experimental and simulation results in Figure 4.8 exhibit a current amplitude of around 6 A.
A similar experiment is conducted at a frequency of 10 Hz. The voltage amplitude and coil configuration are kept the same, but the data is collected for 0.5 s only. As shown in Figure 4.9, the current amplitude is much less than that at 1 Hz due to the increased coil impedance. The experimental signal is distorted especially at the peaks. The difference in amplitude and phase is caused by the error between the actual and simulated inductance. However, the operation of the electromagnet using AC up to 10 Hz is fully functional in principle.
Figure 4.9: Coil Current vs. Time – 10 Hz
Chapter 5  
MFL System Demonstration

5 MFL System Demonstration

5.1 Magnetic Flux Leakage Measurements for Different Size Defects

A series of experiments are conducted to test the applicability of the prototype magnet and inspection system for characterizing defects of different sizes and shapes in steel plates. Four 0.25 in (6.35 mm) thick low carbon steel plates of length 36 in (914.4 mm) and width 12 in (304.8 mm) are used. 30 mm wide defects of different depths are machined across the width of the plates. The defect depths are 1.12 mm (17% of plate thickness), 1.945 mm (30%), 3.04 mm (47%), and 4.02 mm (63%). In order to read the magnetic flux leakage signals, an A1324 hall effect sensor is placed on the cart halfway between the electromagnet legs resulting at a liftoff of 0.375 in (9.525 mm). The sensor has a sensitivity of 5 mV/G. A NI USB-6002 data acquisition device is used to read the sensor output. Since the MFL signal is antisymmetric for DC and symmetric for AC with respect to the center of the defect, only half of the signal is required for analysis.

In the first experiment, the electromagnet is powered with 6 A DC, and the prototype is moved on the steel plates over the defects one at a time with the defects on the top or nearside with respect to the inspection system. The same experiment is then repeated with the defects on the bottom or far side. As shown in Figure 5.1 and Figure 5.2, the MFL signal amplitude increases with defect depth for both cases. When comparing the nearside and far side MFL signals, the far side signal is marginally lower than its nearside counterpart. However, the difference is not significant compared to the amplitude differences resulting from varying defect depth. It can be concluded that the MFL signal amplitude for our inspection configuration is much more sensitive to defect depth than to whether the defect is on the nearside or farside.
Figure 5.1: Experimental DC MFL Signal for Different Defect Depths – Nearside Defects

Figure 5.2: Experimental DC MFL Signal for Different Defect Depths – Farside Defects
A finite element study is run to simulate the DC MFL case. The same current of 6 A DC is used with the same sensor liftoff, and the same defect depths are simulated one at a time. The data is collected from the center of the defect to a distance of 50 mm from the center. Although the absolute value of the MFL signal peak is different than that measured experimentally, the increase with defect depth exhibits a similar trend as shown in Figure 5.3 and Figure 5.4. Again, the MFL signal is affected mostly by the defect depth and not whether the defect is on the nearside or farside.

![Magnetic Flux Leakage vs. Distance From Defect Center](image)

**Figure 5.3: Simulated DC MFL Signal for Different Defect Depths – Nearside Defects**
Another set of very similar experiments are run with the electromagnet powered by a 5 Hz current with an amplitude of 2 A. This is done to verify the ability of the prototype to run AC MFL. All four plates are tested, and again, both the nearside and farside cases are investigated. Since, the MFL signal is sinusoidal, the measured peak to peak amplitude of the cycles is used for plotting the results shown in Figure 5.5 and Figure 5.6. As is the case for DC, the MFL signal amplitude increases with defect depth.
Figure 5.5: Experimental AC MFL Signal for Different Defect Depths – Nearside Defects

Figure 5.6: Experimental AC MFL Signal for Different Defect Depths – Farside Defects
5.2 Simulation Demonstrating the Potential Advantage of Using AC

To conclude our simulations, it is desired to demonstrate the potential advantage of using AC MFL. The simulation focuses on two defect sizes. The first defect is 30 mm wide and has a depth of 30% of the plate thickness. The second defect is only 5 mm wide and has a depth of 40% of the plate thickness. Both defects are on the farside. The simulations show that by using a coil current of 6 A DC, both defects result in a MFL signal of amplitude 79 Gauss. Therefore, by using DC MFL, it is not possible to differentiate between the two defects of different depth, which is the aim.

By switching to AC excitation and using multiple frequencies one at a time, additional information can be gained that allows one to characterize the defects more accurately. The frequencies used are 1, 2, 4, and 8 Hz, and all had an amplitude of 6 A. The MFL signal amplitude for both defects at each of the utilized frequencies is plotted in Figure 5.7. For DC as well as 1 and 2 Hz, the two defects cannot be distinguished using the MFL signal amplitude. However, at 4 and 8 Hz, the higher MFL signal amplitude for the 40% defect depth is distinguishable from that of the 30% deep defect. Since the simulation is run for the farside case, the MFL signal for both defects decreases with frequency, because for higher frequencies only the top part of the plate is magnetized due to the skin effect. As the frequency increases, the magnetic flux density in the bottom part of the plate, where the defects are located, decreases. This decreases the MFL signal caused by the defects.

By studying the response of defects to multiple frequencies of magnetization, more information can be collected that might enable us to differentiate between defects that would otherwise produce the same signal in DC. Since each frequency has a unique skin depth inside the
plate, defects with different depth will respond differently to a frequency sweep applied by the electromagnet.

![Magnetic Flux Leakage Amplitude vs. Frequency](image)

**Figure 5.7: Simulated MFL Amplitude vs. Magnetization Frequency**
Chapter 6
Conclusion

6 Conclusion

6.1 Summary and Review

An electromagnet was designed and built to serve as part of a time varying magnetic flux leakage system. It is able to magnetize a steel plate at various frequencies using current waveforms of arbitrary shape with a bandwidth between 0 and 10 Hz. The design of the electromagnet core and coil was optimized with the aid of finite element modeling. To power the electromagnet, an electric circuit was designed that could produce waveforms that are digitally specified on a computer. The performance of the electromagnet in terms of magnetic flux density was measured experimentally and compared to the design characteristics. A prototype MFL system that utilizes the new magnet was tested by using it to inspect steel plate samples with machined defects. The results were compared to simulations that were then used to demonstrate the ability of AC MFL to differentiate between defects that would be difficult to characterize with DC MFL alone.

6.2 Conclusions

The objective was to design an electromagnet that is optimized for time varying MFL. Specifically, it should produce magnetic fields within a frequency range of 0 – 10 Hz. It should be able to magnetically saturate a 0.25 in (6.35 mm) thick low carbon steel plate at a liftoff of 0.375 in (9.525 mm). Also, the background magnetic flux density in the region of the hall effect sensors that measure the MFL signal should be minimized.

The core material was selected such as to minimize eddy current and hysteresis losses within the frequency range. The material is also magnetic to lower the power needed to magnetize the steel plate. The core geometry, including the dimensions, was designed based on COMSOL
finite element simulations in order to minimize the background field as mentioned previously. This also led to the coils being wound around the sides of the core rather than the top. The thickness of the core was designed for it to be operating at maximum permeability when the steel plate is magnetically saturated.

The coil number of turns and current were chosen so as to achieve the required magnetomotive force of around 6000 Ampere-turns to magnetically saturate the steel plate at a liftoff of 0.375 in (9.525 mm). This liftoff accounts for typical protective fiberglass coatings on tank floors and wheel clearance. The coil is cooled using fans. The temperature is important because coil resistance increases with temperature. Besides not exceeding the temperature rating of the magnetic insulation of the magnet wire (200 °C), which can be achieved without cooling, the coil temperature is kept as low as possible to decrease the required power.

The electromagnet core required assembly because it was made of laminations to reduce eddy currents. A cart was also designed based on the force between the electromagnet and inspected plate calculated using finite element analysis.

To produce the magnetic fields at the desired frequencies, the electromagnet is powered by an h-bridge circuit with PWM input from an Arduino microcontroller. The power supply was sized so as to account for the increase in coil impedance with frequency. Current feedback control was implemented using current sensors to compensate for the increase in coil resistance with temperature.

Experiments were run to show that the desired specifications were met. The performance of the cooling system was studied by comparing the coil temperature with and without fans.
Thermocouples were used to read the temperature. The large drop in temperature when fans are used demonstrates the importance of cooling.

The force as a function of current was experimentally measured with a scale and compared to finite element simulations. The results matched to within a few percentage points showing that designing the cart based on simulations is valid.

The magnetic flux density produced by the electromagnet was measured with a gaussmeter at different coil current levels. That was compared with corresponding COMSOL results. Since the values were consistent, the finite element model is accurate and can be used to calculate the magnetic flux density inside the plate, which cannot be measured experimentally. Therefore, by using the current needed to magnetically saturate the plate in the model, the prototype should also be able to do the same.

The electric circuit operation and ability to produce AC current at frequencies up to 10 Hz was investigated by measuring the coil current experimentally. The experiment was conducted for frequencies of 1 Hz and 10 Hz. Although, some distortion was present, the current amplitudes pointed to a very similar coil inductance value to that calculated using COMSOL. By measuring the current experimentally, it was verified that the current waveform was as intended.

The electromagnet was used for MFL inspection of 0.25 in (6.35 mm) thick steel plate samples with machined defects of width 30 mm in the scanning direction. The depths were 17%, 30%, 47%, and 63% of the plate thickness. The MFL signals were not masked by the background signal, indicating the success of the core geometry design in producing a region of low background field at the sensor location. Both DC and AC MFL were tested for nearside and farside defects,
and the increase of MFL signal amplitude with defect depth was observed, which is consistent with simulations.

Simulations were then used to demonstrate how AC MFL can be used to distinguish between defects that would otherwise be indistinguishable with DC MFL. The studied defects were 30 mm and 5 mm wide with depths of 30% and 40% respectively. Using DC MFL, both resulted in the same signal amplitude. By running MFL at 4 Hz and 8 Hz for both defects, a difference in the signal amplitude between the defects was detected.

6.3 Future Work

The aim of this project was to build the electromagnet for a time varying MFL system prototype. The completion of the prototype opens up a lot of opportunities for future work. First of all, more MFL tests can be conducted with steel plate samples that contain defects of different widths not just different depths. That would provide more insight into the effects of varying the defect dimensions on the MFL signal. Also, the use of different magnetization frequencies simultaneously and studying the resulting frequency response can be an interesting prospect in terms of reducing defect characterization time. For this project, the focus was mainly on MFL signal amplitude, but valuable information can be collected from the phase response of different frequencies since the phase of the signal, and not just the amplitude, is affected by the location or depth of a flaw.

Some improvements can be made to the prototype hardware as well. A motorized cart can be used to control the scanning speed of the prototype. Moving the prototype at a constant speed while scanning will eliminate any effects the change of speed has on the MFL signal. Controlling the speed will also allow us to record the MFL signal along with the corresponding location with respect to the defect instead of recording with respect to time. Another possible hardware
modification that would tackle the issue of the signal being affected by velocity changes is the use of other magnetic sensors that are less sensitive to velocity changes than hall effect sensors.

The ultimate goal is to develop an algorithm that uses the MFL signals from different frequencies to quantitatively characterize a defect, where the parameter of interest is primarily the defect depth. If the algorithm is capable of reliably estimating the remaining plate thickness, the reliance on slow and expensive ultrasound testing as a backup inspection method for fuel storage tank floors can be drastically reduced. This would result in a much faster inspection and lower downtime for the tanks.
References


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