

Improved Antenna Design for Grain Bin Electromagnetic Imaging

by

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Abstract

Grain such as wheat and canola is usually stored in silos or bins, often for long periods of time. Grain can and regularly does spoil in such bins, thus monitoring the grain quality and quantity inside the bin is crucial. It is possible to monitor grain within metallic grain bins using three-dimensional ElectroMagnetic Imaging (EMI). To acquire data in EMI, multiple antennas must transmit and receive ElectroMagnetic (EM) energy within the imaging bin. These antennas subjected to significant mechanical forces, they must have a low profile.

These antennas should also be easy to model in computational EM software (using a point source). Moreover, since the EMI algorithm uses linear polarization of EM fields, it is common to choose antennas that can detect a linear polarization and reject others, such as antennas that measure magnetic fields normal to the antenna's loop and reject electric fields. This can be done through using Shielded Loop Antennas.

Given their desirable physical and EM characteristics, previous grain bin imaging systems have been designed with Shielded Half Loop Antennas (SHLA). While capable of detecting primarily the tangential magnetic field at the bin wall, existing SHLA suffer from a large reflection coefficient at the antenna terminal, meaning they can be drastically improved by reducing this reflection coefficient, thus reducing noise and receiving more desired signal. Herein, we consider two methods of improving upon this design: matching circuits and improved antennas. For matching techniques, we consider passive and active circuits. Through simulation, we show how these methods can improve antenna performance. For an improved antenna design, we show that loading a SHLA with ferrite material can enhance its performance. To model the behavior of a ferrite-loaded shielded loop antenna, we develop an approximate lumped circuit model which can predict antenna resonance frequency, its reflection coefficient, and its efficiency. This model aids in the design of larger sized ferrite-backed antennas.

An antenna at resonance frequency of 260 MHz is fabricated and tested in a small (labratory-sized) grain bin. Its sensitivity to magnetic field and rejecting electric fields is tested in Gigahertz Transverse EM (GTEM) cell at the Electromagnetic Imaging Laboratory (EIL) of the University of Manitoba, and show that the ferrite-backed antenna increases signal strength while maintaining the rejection of normal-E fields.

The small ferrite loaded shielded half-loop antenna provided a 6 – 18 dB improvement in signal level over existing half-loop antennas over the imaging frequencies of interest (approx 200–600 MHz). Furthermore, experimental laboratory imaging tests with hard-red winter wheat indicate that the

proposed antennas can be modeled more accurately using simple polarized point sources. A full 3D image of a wet-grain target set within a dry-grain background inside the enclosure shows that the improved antennas enhance the quality and accuracy of the reconstructed images, mainly where the antenna performance improvements are prominent.

Based on these promising results, we designed and constructed a scaled up ferrite-backed half-loop antenna suitable for a larger industrial grain bin and at a lower frequency (a resonance of 120 MHz). Preliminary measurements of these antennas show that these larger versions not only can detect magnetic field and reject electric fields, but also increase the signal strength in the grain bin.

To:

Two of history's greatest scientists:

Maxwell

and

Hilbert

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My Publications

- [1] **Mohamadi, Cena. T.**, Mohamad Asefi, Sandeep Thakur, Jitendra Paliwal, and Colin Gilmore. “Improved Resonant Chamber Microwave Imaging Using Ferrite Loaded Antennas”. In: *IEEE Transactions on Instrumentation and Measurement* (2021-Nov: Under review).
- [2] Eungjoo Kim, **Mohamadi, Cena. T.**, Mohammad Asefi, Joe Lovetri, Ian Jeffrey, and Colin Gilmore. “Imaging and Calibration of Electromagnetic Inversion Data With a Single Data Set”. In: *IEEE Open Journal of Antennas and Propagation* 3 (2022), pp. 12–23. DOI: [10.1109/OJAP.2021.3132100](https://doi.org/10.1109/OJAP.2021.3132100).
- [3] **Mohamadi, Cena. T.**, Mohamad Asefi, and Colin Gilmore. “Ferrite Loaded Shielded Half Loop Antenna for Electromagnetic Imaging Inside Metallic Chambers”. In: *Proceedings of the 19th International Symposium on Antenna Technology and Applied ElectroMagnetics (AN-TEM 2021)*. Manitoba, Canada, 2021. URL: <https://antem2021.ieee.ca/>.

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Chapter 1

Introduction

1.1 Scope of Thesis

In ElectroMagnetic Imaging (EMI), ElectroMagnetic (EM) waves are used to reconstruct the properties of a Region Of Interest (ROI) qualitatively and/or quantitatively. With a series of antennas, a field is transmitted into the ROI, and then its scattered fields are measured at multiple locations around it. After these measurements are collected, an inversion algorithm is applied to reconstruct the electrical properties (i.e., complex permittivity [1]) of the ROI. Antennas are an essential part of EMI systems. In particular, they (1) radiate the interrogating energy into the ROI and (2) receive the scattered fields from the ROI.

The topic of this thesis is the design and improvement of antennas for a specific application of EMI: imaging the contents of grain bins. This application of EMI presents many design challenges for antennas compared with traditional antenna design in free-space. These antennas must run in a fully enclosed metallic enclosure and operate successfully both while in air and covered by grain. Additionally, antennas in grain bins face a much more rigorous physical environment than typical antennas as grain moves around them. All of these design constraints must be taken into account, and are discussed as appropriate in this thesis.

1.2 Motivation

In response to the growing world population, there is a need for larger stores to preserve foods for perhaps a more extended period. Among the essential nutrition for humans are wheat, rice, corn, canola, and other grains. These grains are often stored in substantial metallic structures such as the one pictured in Figure 1.1. Grain can, however, spoil due to abiotic and biotic agents [2] and [3]. Humidity of grains, weather conditions of the environment and the temperature of grain are some abiotic elements that may spoil grain. Rodents, birds, microorganisms, worms, and insects are some biotic agents that can spoil and even reduce the amount of the stored grain.

Grain sprouts because it senses that its in the ground and should sprout. This occurs if the grain is in high moisture and temperature. Besides, if moisture content of the grain is too high as 14%, [2], then it will be a suitable place for bacteria, microorganisms and fungi. High percentage of humidity will heat the grain and reduces its storage period. Temperature of the storage bin should be kept

below 30°C [2].

Grain spoiling as the result of high temperature and moisture occurs in a positive-feedback method: when some region of grain starts to spoil, it releases moisture and energy (in the form of heat) from the seed. This excess heat and moisture then causes the neighboring regions to spoil which leads to a positive feedback process.

Monitoring grains is a method to make sure about their quality and healthfulness. Grain should be monitored at least once a month in winter and every other week during other seasons, [2], [4], and [3]. As it is understood from above mentioned factors of grain spoilage, temperature and humidity content are two vital factors that should be monitored.

If we can detect high temperature or high humid content zones in the grain, using some methods like cooling with turning on fans, flip the grain in the bin¹, or emptying the grain into another bin and separating the spoilage, and so on, we can preserve it from destruction, [2] and [4]. So if we can detect the spoiling regions early, considering the mentioned positive-feedback in grain spoilage, the spoilage can be stopped in early stages. Different methods have been employed to monitor wetness and temperature of grains. We will explain them in the next part.

1.2.1 Traditional grain monitoring techniques

One of the simplest method of monitoring grains can be done by using the very basic tools and senses of human [3]: a person measures temperature at different parts of the stored grain by some basic thermostats; visually inspect for rats and insects, etc; smell the grain storage to see if it smells wet or not; etc. This method is very time consuming and not possible in large size grain storage. For example in large bins, it is impossible to measure temperature of every part of the grain storage. Or if there is some spoilage in internal sections of stacked grain, its smell will not easily penetrate outside, therefore the person cannot sense it with smelling.

Using temperature and humidity sensors is a better solution to monitor grain. In this method, depending on the size of the storage bin, these sensors are mounted on some vertically hung cables from the top of the bin, [3] and [4]. However, it has some restrictions. Depending on the size of the bin, there are some limitations in the number of hung cables and therefore number of temperature and humidity sensors. Besides, these sensors can only measure temperature or humidity of grain in their vicinity as grain is an excellent thermal insulator.

Acoustic grain monitoring method has been presented in [5]. The authors have used acoustic tomography and map time of sound travelling in grain to temperature. However, they have not measured or proposed any method to measure humidity content.

1.2.2 EM properties of grain

In [6], the authors have shown the relation between grain temperature and permittivity. Also, in [7], authors have shown the dependency of humidity of grain with its permittivity. Based on those works, from an electromagnetic perspective, high content of humidity in grain means high permittivity grain. EMI is a way of monitoring the permittivity throughout the entire bin. We use this method to monitor the grain in the bin.

¹If the spoiling region in bin is small, one can empty it out of the bottom, and then promptly fill it back in the top. Flipping a bin will often break up the hot/wet region, and spread it out so that it cools off/dries.

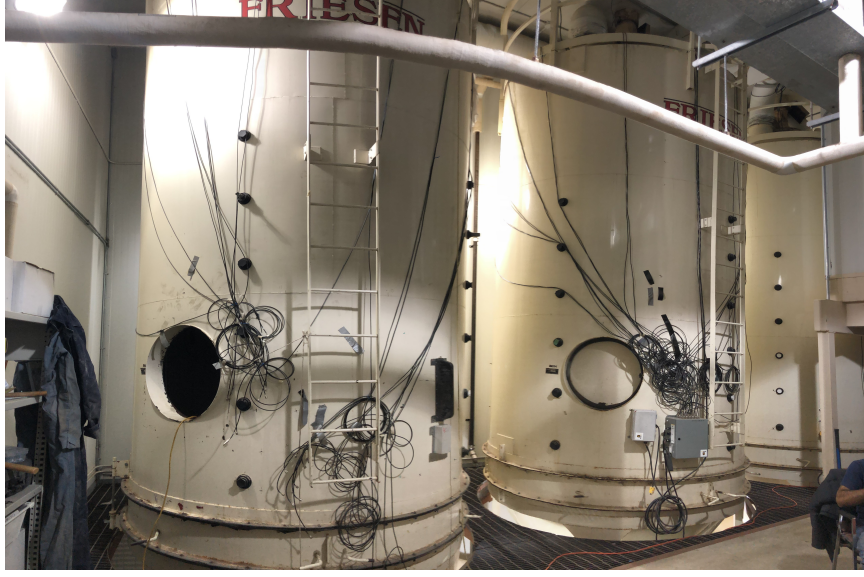


Figure 1.1: Grain bins for storing grains.

1.3 Electromagnetic Imaging System (EMI) of Grain Bin

Using EMI technique, it is possible to determine the relative permittivity of grain stored at differing temperatures and moisture contents. Because EMI does not damage or destroy the grain itself, it is an excellent method of tracking the grain's health². It also helps estimate the amount of grain remaining in the bins.

EMI has many different applications as it offers the ability to see inside of objects and infer important information that is otherwise not available. Applications of EMI include medicine [9, 10], geosciences [11], food science [12], breast cancer detection [13–17].

Monitoring grain bins, through the permittivity of grain inside the bin, is one of the recent applications of EMI that has been developed at the University of Manitoba [18–21]. Grain's complex permittivity has long been used to estimate its moisture content [22]; it is a critical factor for ensuring the safe storage of grain.

Figure 1.2 represents the general schematic of an EMI system. In this figure, the yellow region represents the background media with relative permittivity $\epsilon_b(\vec{r}) = \epsilon'_b(\vec{r}) - j\epsilon''_b(\vec{r})$ (note that this background permittivity can be heterogeneous). In grain bins, the entire bin is actually an ROI. Typically we search for a high-permittivity region inside the grain. This region has been shown in red color and its relative permittivity ($\epsilon_r(\vec{r}) = \epsilon'_r(\vec{r}) - j\epsilon''_r(\vec{r})$) is obtained from EMI. As seen in that figure, EMI consists of two main parts: hardware and software sections. The hardware part is composed of antennas, cables, switches (SW), Vector Network analyzer (VNA), and wireless transceiver. Antennas have been installed around the ROI. For example, in the case of grain bin imaging, they are installed on the bin's walls, so the outer circle represents the walls of a metallic enclosure or bin. Cables transmit and receive EM waves to and from the antenna. SWs can change the antenna mode between transmitter and receiver ones. Received signals by the antennas, sent down the cables and routed through a switch back to the system transceiver (e.g. VNA). These

²Since EMI does not change or destroy the matter under test, this method is called Non-Destructive Test method as well [8].

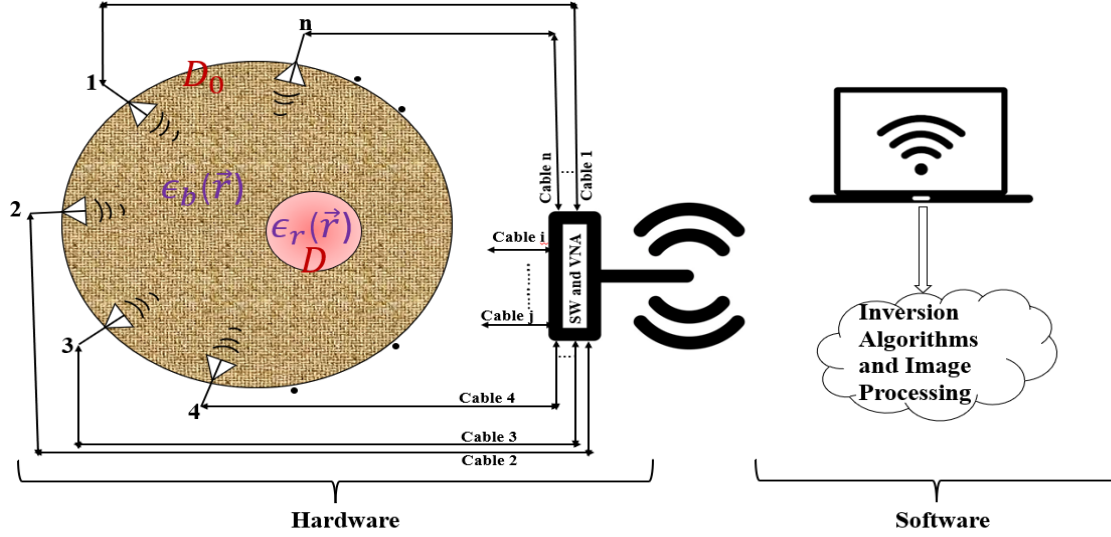


Figure 1.2: Generally, EMI system consists of two main parts: hardware and software. Region D_0 is the heterogeneous background (imaging region) and region D is the desired hot spot (target region) that we look for in EMI.

collected signals are sent to a central computer, through wireless connection, which runs the inversion algorithm, producing the reconstructed ROI based on its desired constitutive parameters. The software section of EMI solves numerically some equations to produce the image of the object of ROI.

In Figure 1.2, D_0 is the heterogeneous background and called imaging region; region D is the target region. Note that $D \subset D_0$.

There have been many developments in different parts of the EMI system, depending on the type of application. Readers can find many of them in literature ([14], [19], [23–25], etc). The software part of EMI that are called inversion algorithms are not the main focus of this work, but will be discussed further in this Section.

1.3.1 EMI's Core Equations

As previously mentioned, EM waves are used to reconstruct an ROI based on its constitutive parameters. Therefore, solving wave equations is the core of the software part of EMI. Maxwell's equations govern electromagnetic waves behavior. Its differential form is represented by the following equations [1]:

$$\vec{\nabla} \cdot \vec{D}(\vec{r}, t) = \vec{Q}_v(\vec{r}, t) \quad (1.1a)$$

$$\vec{\nabla} \cdot \vec{B}(\vec{r}, t) = 0 \quad (1.1b)$$

$$\vec{\nabla} \times \vec{E}(\vec{r}, t) = -\frac{\partial \vec{B}(\vec{r}, t)}{\partial t} \quad (1.1c)$$

$$\vec{\nabla} \times \vec{H}(\vec{r}, t) = \frac{\partial \vec{D}(\vec{r}, t)}{\partial t} + \vec{J}(\vec{r}, t) \quad (1.1d)$$

where, $\vec{r}$ is position vector, t is time, $\vec{D}(\vec{r}, t)$ is the electric flux density, $\vec{Q}_v(\vec{r}, t)$ is electric charge density; $\vec{B}(\vec{r}, t)$ is magnetic flux density; $\vec{E}(\vec{r}, t)$ is electric field intensity; $\vec{H}(\vec{r}, t)$ is magnetic field intensity; and $\vec{J}(\vec{r}, t)$ is electric current density. In (1.1), $\vec{J}(\vec{r}, t)$ and $\vec{Q}_v(\vec{r}, t)$ are sources that generate EM fields. If sources are known, Maxwell's equations along with constitutive relations in medium can be solved to obtain EM fields. In most of the cases at lower frequencies up to microwave frequencies, these constitutive relations can be reduced to simple linear forms:

$$\vec{D}(\vec{r}, t) = \epsilon(\vec{r})\vec{E}(\vec{r}, t) \quad (1.2a)$$

$$\vec{B}(\vec{r}, t) = \mu(\vec{r})\vec{H}(\vec{r}, t) \quad (1.2b)$$

$$\vec{J}(\vec{r}, t) = \sigma(\vec{r})\vec{E}(\vec{r}, t) \quad (1.2c)$$

In (1.2), $\epsilon(\vec{r})$ is the permittivity of the medium and is a complex quantity multiplication of the free space permittivity (ϵ_0), i.e., $\epsilon(\vec{r}) = \epsilon_0(\epsilon'_r(\vec{r}) - j\epsilon''_r(\vec{r}))$; $\mu(\vec{r})$ is the permeability of the medium and is a complex quantity multiplication of the free space permeability (μ_0), i.e., $\mu(\vec{r}) = \mu_0(\mu'_r(\vec{r}) - j\mu''_r(\vec{r}))$; $\sigma(\vec{r})$ is the conductivity of the medium. Permittivity, permeability, and conductivity are scalar quantities in (1.2).

If we consider their time dependence as $e^{j\omega t}$, then $\vec{D}(\vec{r})$ is referred to as the time harmonic (or Phasor) form of electric flux density; its relation with electric flux density is as the following equation, where $\Re$ represents the real part of the function inside parentheses. Similar argument holds true for other EM quantities in (1.1).

$$\vec{D}(\vec{r}, t) = \Re(\vec{D}(\vec{r})e^{j\omega t}) \quad (1.3)$$

$$\vec{\nabla} \cdot \vec{D}(\vec{r}) = \vec{Q}_v(\vec{r}) \quad (1.4a)$$

$$\vec{\nabla} \cdot \vec{B}(\vec{r}) = 0 \quad (1.4b)$$

$$\vec{\nabla} \times \vec{E}(\vec{r}) = -j\omega\vec{B}(\vec{r}) \quad (1.4c)$$

$$\vec{\nabla} \times \vec{H}(\vec{r}) = j\omega\vec{D}(\vec{r}) + \vec{J}(\vec{r}) \quad (1.4d)$$

Finally the time harmonic form of Maxwell's equations will be as (1.4), and linear constitutive relations will be as following equations:

$$\vec{D}(\vec{r}) = \epsilon(\vec{r})\vec{E}(\vec{r}) \quad (1.5a)$$

$$\vec{B}(\vec{r}) = \mu(\vec{r})\vec{H}(\vec{r}) \quad (1.5b)$$

$$\vec{J}(\vec{r}) = \sigma(\vec{r})\vec{E}(\vec{r}) \quad (1.5c)$$

Combining and reorganizing (1.4) and (1.5) will result in complex vector wave equation:

$$\vec{\nabla} \times \vec{\nabla} \times \vec{E}(\vec{r}) - k^2(\vec{r})\vec{E}(\vec{r}) = k^2(\vec{r})\vec{W}(\vec{r}) \quad (1.6)$$

Through solving (1.6), fields are determined. In (1.6) k is called wave number and has the following

relations with permittivity and permeability of medium that waves propagate there:

$$k = \omega\sqrt{\mu\epsilon} \quad (1.7)$$

Also, $\vec{W}(\vec{r})$ represents the source of the equation as follows:

$$\vec{W}(\vec{r}) = -j\omega\mu(\vec{r})\vec{J}(\vec{r}) - \vec{\nabla}\left(\frac{1}{\mu(\vec{r})}\right) \times \vec{\nabla} \times \vec{E}(\vec{r}) \quad (1.8)$$

In grain bin imaging, grain is nonmagnetic material therefore $\mu(\vec{r})$ is a constant value: (μ_0), so the second term in (1.8) vanishes.

1.3.2 Forward and Inverse Problems in EM

Generally, an electromagnetic problem can be modeled as a system (whether linear or nonlinear) with single or multiple inputs and outputs. In some problems the system has a known model. Then the goal is to find the output based on any determined input. This type of problem is called Forward Problem Solving (FPS). However, if we find input or system model based on some available or measured responses in the output of the system, it is called Inverse Problem Solving (IPS).

In the forward electromagnetic problems, system is medium of the propagation of waves with its physical properties (such as ϵ , μ , and σ), boundary conditions, and initial conditions; input can be any types of sources: electric or magnetic current distribution sources. Forward EMI problem finds waves that goes through the media.

IPS uses the response of the medium to one or more input sources to determine the medium based on its physical properties; also, sometimes, the medium of EMI is known and the goal is to find the properties of the source of the electromagnetic fields, like its position or frequency, which is another type of IPS. The first type of IPS is called inverse scattering, and the latter is called inverse source problem solving.

FPSs have stable³ solutions if there is any, otherwise they do not have any solution. However, IPS solution is not unique at all⁴. They have zero or infinite number of solutions. IPSs are ill-posed⁵ thus they are not stable problems. Besides, consider that many different system's models can produce the same response to the same source, therefore IPSs do not have unique solution. One way to find a well suited solution to the problem is adding extra restrictions to the IPS which is called Regularization [8] and [26].

Inverse scattering and source electromagnetic problems are same in many aspects however, the former is nonlinear while the latter is linear. Inverse scattering problems often are solved by their linearized model which results into many unknowns. EMI of grain bin in this thesis is a inverse scattering problem.

1.3.3 Inverse scattering and inverse source problems

Due to the importance of inverse source and scattering problems, we examine these two problems here. To do so and also to focus on the features of scattering problems we use the scalar form of

³Here, *stable* means a small changes in input will not result into huge changes in output.

⁴IPS can have unique solution if and only if we have all outputs of the EMI system based on all possible EMI system, which is impossible in practice.

⁵Ill-posed means a small change in input generates a large change in output.

wave (1.6) which is a scalar Helmholtz equation. The principles of inverse problems can be explained well with a scalar problem, then later it is extended to full-vector solutions, so herein we present the inverse problem with a scalar Helmholtz (1.6).

$$\nabla^2 u(\vec{r}) + k^2(\vec{r})u(\vec{r}) = -\vec{S}(\vec{r}) \quad (1.9)$$

Where u is a scalar field and S is an electromagnetic source. (1.9) is solved on the imaging region (D_0) with its given boundary conditions, Figure 1.2. $\{u, k, S\}$ is a triplet of fields that satisfies (1.9). Having any two of these quantities, the third one is obtained. For example in FPS $\{k, S\}$ is known and (1.9) finds u .

Inverse source problems find S according to the given $\{k, u\}$. In other words, in Figure 1.2, assume that source is located in imaging region and is unknown, then, generated fields by the source are measured by installed antenna on the boundary of D_0 . k is known every where and finding the source is objective of this problem.

Inverse scattering problem solves k for known $\{u, S\}$. An antenna on the boundary of D_0 transmits an electromagnetic fields that interact with target region after reaching there. All reflected back and traveled fields are measured by installed antennas on the boundary of the imaging region. k is known outside but unknown inside the target region. The objective is to find k inside the target region. To solve inverse scattering problem, we set the value of k of data to an arbitrary value like the known value of k of the background material. If we assume that the solution of this problem is u^i , then:

$$\nabla^2 u^i(\vec{r}) + k_b^2(\vec{r})u^i(\vec{r}) = -\vec{S}(\vec{r}) \quad (1.10)$$

Where k_b is the value of the background $\{k\}$. Also we refer to u^i as background or incident field. Total measured fields (u^t) in scattering inverse problems are resulted from incident fields (u^i) by antennas and scattered fields (u^s) by the target region. Therefore after subtracting (1.10) from (1.9), we obtain:

$$\nabla^2 u^s(\vec{r}) + k_b^2(\vec{r})u^s(\vec{r}) = -(k^2(\vec{r}) - k_b^2(\vec{r}))u^t(\vec{r}) \quad (1.11)$$

In (1.11), the right hand term of the equation is called contrast source ($S_c(\vec{r})$) and that equation is called scattering equation. Contrast function of (1.11) is defined as follows:

$$\chi(\vec{r}) = \frac{k^2(\vec{r}) - k_b^2(\vec{r})}{k_b^2(\vec{r})} = \frac{\epsilon_r(\vec{r}) - \epsilon_b(\vec{r})}{\epsilon_b(\vec{r})} \quad (1.12)$$

In (1.12) $\chi(\vec{r})$ is normalized contrast. Also, $\epsilon_r(\vec{r})$ and $\epsilon_b(\vec{r})$ are target and background relative permittivity to free space, respectively. Thus the contrast source is:

$$S_c(\vec{r}) = k_b^2(\vec{r})\chi(\vec{r})u^t(\vec{r}) \quad (1.13)$$

As it is understood from (1.12), contrast source is nonzero only in target region.

Note that the equation for solving inverse source problem is pretty much as same as inverse scattering problem, only we need to find the unknown $S_c(\vec{r})$ inside imaging region. The process of solving for the total permittivity from the measured scattered fields is inverse scattering problem-

solving. This process assumes knowledge of both the incident field in the background permittivity and measurements of the total fields at the receiver points⁶.

1.3.4 Software of EMI of Grain Bin

Practical IPSs do not have analytical solutions. For instance it is impossible to find a closed-form solution for breast cancer electromagnetic imaging. Therefore, numerical techniques are the only methods capable of providing reasonable solutions. This has led to many numerical algorithms and strategies being developed for IPS.

One well-known algorithm is the Born iterative method, which can solve, for instance, an integral form of the Helmholtz equation to determine a material's constitutional parameters [27]. Various optimization methods can be used depending on the type of IPS. Some commonly used methods include the Gauss-Newton algorithm [15], and [24], Newton method [28], and Non-Linear Conjugate Gradient algorithm [24]. Derivative-based algorithms are preferred as they converge much faster and tend to provide better final inversions. In some specific applications the Nelder-Mead technique [29], and Genetic Algorithm [23], are used. Both of these techniques are good nonlinear gradient-free optimization methods. These methods have been used in 3D inversion within grain-bin imaging [18] [19], [20], and [21]. Derivative free algorithms are used for parametric inversion, but for pixel-based full 3D inversion, all grain bin imaging results to date have been obtained with Finite Element Method-Contrast Source Inversion (FEM-CSI) method, which is discussed in the following part.

1.3.5 FEM-CSI Algorithm

Using Maxwell's equations (1.1) the vector form of (1.9) based on scattered electric fields is obtained as follows:

$$\vec{\nabla} \times \vec{\nabla} \times \vec{E}^{sct}(\vec{r}) - k_b^2(\vec{r})\vec{E}^{sct}(\vec{r}) = S_c(\vec{r}) \quad (1.14)$$

As mentioned in previous part, the incident fields are associated with the set of background media $\epsilon_b(\vec{r})$, and we are free to define the background media to best solve the inverse problem. The scattered field is associated with the contrast between the actual media and the selected background. For the grain bins used in this thesis, we choose the background media a mixture of air and a homogeneous grain mass with some height and cone angle. The parameters of this homogeneous grain mass are obtained from a phaseless parametric inversion [30]. Therefore, contrasts (and thus scattered fields) are associated with deviations from homogeneous grain (e.g., higher moisture regions result in scattered fields).

In general, non-linear inverse problems are more successful when the contrast is as small as possible, so choosing, e.g., the background to be entirely free-space, leads to much worse inversion results. To obtain an estimation of the background permittivity, we run the Nelder-Mead algorithm [29]. Once we have obtained the estimate of the background permittivity, the full 3D inversions are obtained via the FEM-CSI method. Using the Contrast Source Inversion (CSI) algorithm with the forward solver implemented in the Finite Element Method (FEM) algorithm, known as FEM-CSI [31]. This algorithm optimizes the following cost function:

⁶IPSs can be solved in integral forms as well. Readers can see [8] and [26].

$$\mathfrak{F}^{CSI}(\vec{\chi}, \vec{w}_j) = \frac{\sum_{j=1}^n \|\vec{E}_j^{sct, meas} - \mathfrak{M}_{D_0, j} \mathfrak{L}[\vec{w}_j]\|_{D_0}^2}{\sum_{j=1}^n \|\vec{E}_j^{sct, meas}\|_{D_0}^2} + \frac{\sum_{j=1}^n \|\vec{\chi} \odot \vec{E}_j^{inc} - \vec{w}_j + \vec{\chi} \odot \mathfrak{M}_{\mathfrak{D}_o, j} \mathfrak{L}[\vec{w}_j]\|_D^2}{\sum_{j=1}^n \|\vec{\chi} \odot \vec{E}_j^{inc}\|_D^2} \quad (1.15)$$

Optimization function is the summation of normalized L_2 norm errors of measurement and forward solver in background or imaging region (D_0) and normalized L_2 norm errors of measurement and forward solver in target region (D). Note that $\mathfrak{L}$ represents the inverse of finite element discrete matrix which depends on the boundary condition, and $\mathfrak{M}$ stands for interpolation matrix operator to convert fields values along edges of meshes of FEM to vector fields at locations of transceivers or points of the imaging region. Moreover, $\odot$ represents the element-wise multiplication of matrices of the same size. The input data to the optimization (1.15) are the scattered fields measured by the antennas. Note that this objective function assumes the measurements of a known field component at the point of each receiver antenna. As this becomes very important in antenna designing and we should design antennas that can measure just one component of EM fields. The CSI algorithm optimizes this functional by applying the conjugate gradient minimization algorithm to the two unknowns, $\chi(\vec{r})$ and $S_c(\vec{r})$ [31]. FEM-CSI has been used in many applications like breast cancer imaging [13] forearm imaging [25] or grain bin imaging [20],[21], [18], and [19].

1.3.6 Hardware of EMI of Grain Bin

The hardware for Grain-bin EMI is similar to that shown in Figure 1.2. The bin's metallic wall forms the outer boundary of the chamber; the wireless data acquisition box is composed of RF switches, a Vector Network Analyzer (VNA), connectors, DC power supply, and some wireless data links to transmit measurements to the central station for algorithm execution.

1.3.7 Review of Previously Used Grain Bin Imaging Antennas

Grain storage bins are metal, this means only tangential H and normal E exist at the boundary. Previous researchers began by attempting to measure normal E with monopoles [32]. In [32], the authors have used two types of monopole antennas: passive monopole and reconfigurable monopole antennas to do 3D microwave imaging from a nylon cylinder inside an air-filled metallic chamber. Twenty-four antennas are installed on the wall of the enclosure. As mentioned, close to the metallic boundaries, the normal electric field is nonzero. Therefore installed monopoles collect the normal electric field.

These monopoles proved unsuitable to the physically rigorous environment inside a grain bin: the E-field normal monopoles were destroyed by moving grain. Besides, reconfigurable antennas need a power supply to turn on and off some of their components which is challenging to be implemented in industrial grain bins.

In [33] similar to [32], the authors introduce another single-polarization measurement technique for EMI. However, they use shielded half-loop antennas. These antennas are mostly sensitive to the magnetic field component normal to the plane of the loop. Hence induced tangential surface current on the antenna is proportional to that component. These types of antennas are collecting only induced currents due to magnetic fields and not electric fields [34]. A scaled version of [33] has been developed and implemented in a big industrial bin at [30].

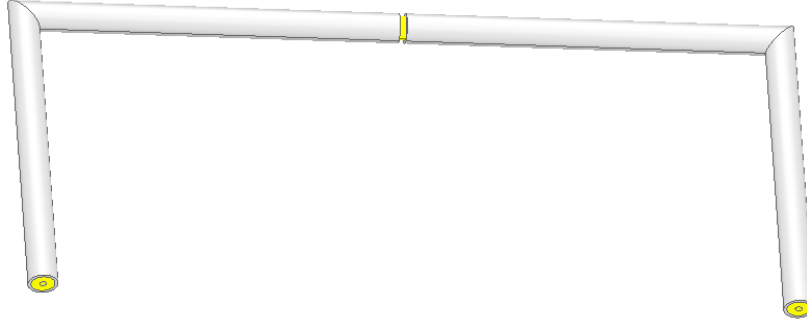


Figure 1.3: Geometry of currently used antenna in grain bin imaging [33] and [30].

1.3.8 Current Grain Bin Imaging Antennas

Current EMI algorithms are developed to work with linear polarization. Also, one measured component of electric or magnetic fields are desired, because of FEM-CSI algorithm. Two different sizes of Shielded Half Loop Antennas (SHLA) are used in current grain bin imaging: smaller size for imaging in lab scale grain bin at 618 MHz [33] and large one for industrial scale grain bin imaging at 90 MHz [30]. Both of these antennas are sensitive to tangential magnetic fields and reject the normal electric field close to the wall's bin. Also, close to metallic walls, tangential component of magnetic field is the only nonzero component of magnetic field. Therefore these antennas are suitable for single polarization EMI algorithm which is used in current FEM-CSI technique at University of Manitoba.

Besides, those types of antennas have small size (sub wavelength sizes), therefore they can be modeled with point source which is another criteria for FEM-CSI algorithm. Figure 1.3, shows the shape of those antennas. Both of them are made of 50Ω coaxial cable in a half loop shape. Authors in [33] use *RG405* coaxial cable with a gap of 0.5 mm in the shield of the coaxial cable at the middle of the loop and two suitable SMA connectors at two ends of the antenna. Dimensions of this antenna are 56 mm × 22 mm.

Presented antennas in [30] are made of *RF195* multi layer coaxial cable with a gap of 1 mm – 2 mm. The gap is just in the outer shield at the middle of the loop. Loop is connected to two suitable SMA connectors at each ends of the loop. Dimensions of this antenna are 400 mm × 100 mm. Both antennas are installed on the metallic walls of grain bins.

1.4 Problem Description

As it is seen from Figure 1.2, EMI uses a number of antennas to send and receive EM waves. If antenna i (where $1 \leq i \leq n$) is in transmitting mode, then all other antennas will be in receiving mode. Received signal by each of those receiver is S_{ji} (where $1 \leq j \leq n$ and $j \neq i$). These S_{ji} values are used as the input of EMI algorithms. Therefore having an antenna with higher level in S_{ji} will result into high Signal to Noise Ratio (SNR) and therefore improve EMI results.

Current antennas in EMI in grain bin imaging can provide imaging of the grain, however they suffer from some poor radiation characteristics. They reflect most of the impressed power to their terminal toward the source, likewise in the receiving mode, most of the incident power is re-radiated

from antenna instead of delivering to the attached circuit to the terminal of the antenna. Also, from the mechanical aspects, those antennas that are supposed to be used in industrial grain bins have a large size.

There can be grain bins of more than 40 m in diameter, and each antenna must be connected to a central VNA by a coaxial cable. Raw S -parameter signal levels in such systems are often between -80 to -100 dB (because of lossy grain, antenna inefficiency, and long cables' losses). Therefore, a more radiating antenna will result in a higher Signal-to-Noise Ratio (SNR) in these large bins.

Operation frequency of the antenna is one of the primary criteria in the design/improvement of antenna. A wide range of frequencies are available for EMI in industrial bins, from 8 MHz to 180 MHz depending on the sized of the bin. For any given bin, the frequency of operation is ultimately determined by both the ability to computationally model the electromagnetic in that bin (i.e., larger bins must use lower frequencies otherwise the inversion problem becomes computationally too large to solve), and the ability to get enough signal through the bin (lower frequencies are better able to penetrate the grain). Within this thesis, we present improved antennas first at a small scale bin (a lab-based test bin that operates in the 200 MHz-600 MHz range.), then present antennas designed to work around 120 MHz. Eventually these could be scaled to the 8 MHz range.

In the process of antenna design/improvement we should consider another critical point as well: mentioned FEM-CSI algorithm which does inversion works for single point sources rather than directive antennas. Moreover, for the sake of compatibility to FEM-CSI algorithm, we are interested in measuring just one component of EM fields. Considering that antennas are installed on the metallic walls of the bin and close to metallic boundaries, only normal component of E and tangential component of H fields exist, we can decide on the type of antenna. We choose a Shielded Loop Antenna (SLA) rather than monopoles. First, it is more compact than monopole. Besides, the only way to get good radiation properties of monopoles is to install them perpendicular to the metallic walls of the bin which will result in a lot of mechanical stress and strain from grain movement.

Finally, we should consider some mechanical limitations in the antenna's size. The new design or improved version cannot be larger than $40\text{ cm} \times 10\text{ cm} \times 10\text{ cm}$ while we try to improve the value of S_{11} of antenna. The bandwidth, efficiency, etc., are not limitations as well, because we do not do wide band imaging.

1.5 Thesis Organization

This study aims to improve the performance of the existing antennas. Several background materials on antennas are discussed in chapter 2. Among these topics are antenna parameters and ferrite materials. Also, in chapter 2 we present reciprocity theorem and use it for the analysis of magnetic probes, which is crucial to understanding shielded loop antenna performance. Having analyzed shielded loop antennas and developed lumped circuit models for shielded loop antennas with and without ferrite material loading, chapter two ends. It is possible to calculate the resonance frequency and efficiency of antennas with these circuits.

The chapter 3, after discussing the characteristics of the current antennas (like their input impedance), introduces some techniques for improving their radiation behavior. The techniques include matching techniques and ferrite loading. We present the characteristic parameters of a new ferrite loaded shielded loop antenna at the end of that chapter.

The chapter 4 describes some test setups and measurement methods to demonstrate that the methods described in chapter 3 and the newly designed antenna at the end of chapter 3 can provide the desired criteria for EMI. Furthermore, we will show how the improvement methods in antennas' structures enhance the signal levels of $S_{ij}(i \neq j)$. Finally, we will compare all these improvements methods and select the best one for the grain bin application in this thesis.

In chapter 5, we discuss imaging setups that use improved antennas. Using a simple target (a bag of wet wheat) which is buried under the dry wheat inside a lab scale bin we perform some tests and find 3D images of the volume of the bin to detect the position and size of the target inside the grain bin using EMI. Through target tracking tests (changing the position of the target inside the bin) we verify that the EMI algorithm does not depend on the position of the target.

Finally chapter 6 concludes by summarizing the important works and points of this thesis and suggesting some applications and other improvement as possible future works.

Chapter 2

Background

This chapter starts by briefly reviewing some of the basic parameters of antennas which are necessary to understand the behaviour of antennas in this thesis, although the readers may be familiar with basic antenna parameters. A thorough explanation of these parameters can be found in [1] and [35–37].

Then we present ferrite materials' properties and structures. These materials later will be loaded to existing antennas' structures to improve their performance. We will mention some important graphs in the data catalogue of ferrite materials.

Additionally, this chapter presents reciprocity theory in EM that is used in analysis of balanced Shielded Loop Antennas (SLA). It should be noted that the grain bin imaging system uses half-loops that are shielded and placed against a metallic wall, making those antennas act as full loops. Accordingly, we will perform an analysis of balanced SLA. The shielded half-loop with ferrite in its core was also analyzed since we propose to add ferrite to these antennas to improve their performance later. For these types of antennas, we present a lumped circuit model based on that analysis. Using a lumped model, we can determine the antenna's resonant frequency and efficiency.

Finally a method to find the sensitivity of SLAs to magnetic fields is introduced: using a special types of wide band chamber that is called Gigahertz Transverse ElectroMagnetic (GTEM) cells. This chamber later is used to measure how the antenna is good in detecting only one component of magnetic field and rejecting electric fields.

2.1 Antenna Parameters

Some electrical and mechanical factors need to be considered when it comes to antenna design. These parameters are radiation pattern, directivity, input impedance, gain, noise figure, bandwidth, polarization, quality factor, metal type in the antenna structure, structural dimensions, weight, budget restriction, mechanical restrictions like the amount of stress and tension, etc. Some of these critical electrical parameters are described in the following sections.

2.1.1 Radiation Pattern

The radiation pattern shows the maximum and minimum directions of power radiated by the antenna. Based on Maxwell's equations, sources can generate fields, electric $\vec{E}(\vec{r}, t)$ or magnetic $\vec{H}(\vec{r}, t)$.

In addition, based on those equations, if one type of produced field is time-dependent, it can generate the other type of field. The Poynting vector associated with those fields determines the radiation distribution as a result of those fields and is as follows [35–37]:

$$\vec{S}(\vec{r}, t) = \vec{E}(\vec{r}, t) \times \vec{H}(\vec{r}, t) \quad (2.1)$$

Equation (2.1) shows that the Poynting vector is orthogonal to electric and magnetic fields. The time average of the Poynting vector is called average radiation density of antenna:

$$S_{av} = \frac{1}{2} \Re(\vec{E}(\vec{r}) \times \vec{H}^*(\vec{r})) \quad (2.2)$$

Where in (2.2), * stands for complex conjugate.

Multiplication of average radiation by $|\vec{r}|^2$ and then integrating over a sphere with radius $|\vec{r}|$ gives radiation intensity of antenna ($U(\theta, \phi)$) that does not depend on distance from antenna ($|\vec{r}|$). Sometimes radiation intensity is used as the radiation pattern of the antenna. In many cases, the radiation pattern is obtained by integrating average radiation over a sphere with radius $|\vec{r}|$ which will result in radiation power (P_{rad}). However, there is no difference between the final results.

The current distribution of the antenna is the main factor determining its radiation pattern. Consequently, most of the antenna parameters which are dependent on the radiation pattern will depend on current distribution on antenna. A simple example of radiation pattern is an isotropic pattern that radiates power equally in all directions in the space. However, it does not exist in practice and it is a theoretical antenna.

2.1.2 Directivity

Directivity shows the antenna's ability to concentrate radiated power in some directions more than others. It is obtained as follows [35–37]:

$$D(\theta, \phi) = 4\pi \frac{U(\theta, \phi)}{P_{rad}} \quad (2.3)$$

If the antenna is isotropic, it radiates all of its input power equally. Therefore, the directivity (D) of an isotropic antenna is considered 1 ($D = 1$). If an antenna concentrates radiations in some direction more than others, it is called the directive antenna, and its directivity is defined as an (2.3). Maximum directivity is shown with D_0 and occurs in the direction with maximum radiation intensity.

2.1.3 Input Impedance

An antenna is not a lumped-element circuit, however we can use its input impedance (or simply its impedance) and model it with a lumped element circuit. This model will be useful in analyzing the complete system which is generally made of electronic parts as well. Definition of an antenna's impedance is defined based on its Poynting vector.

$$Z_A = \frac{\oint S_{av} \cdot d\vec{s}}{|I_0|^2} \quad (2.4)$$

In (2.4), Z_A is antenna impedance, I_0 is the amplitude of the current distribution on the antenna terminal, and integral limits are on the cross-section of the antenna input terminal.

Generally, antenna impedance model can have resistance, capacitance, and inductance terms simultaneously: ($Z_A = R + X_C + X_L = R + \frac{1}{j\omega C} + j\omega L$). Where C and L are the capacitor and inductance associated with the antenna model and R is the resistance of the antenna model. The imaginary part of input impedance of antenna is called reactive input impedance as well, while the real part of input impedance is called resistive input impedance. Resistance, R is composed of all losses in antenna which could be due to radiation (R_{rad}) or ohmic loss (R_{ohm}), therefore $R = R_{rad} + R_{ohm}$. In antennas, radiation loss is the desired loss. For some antennas, one or two terms are often dominant. For example, like the ones in [30] and [32, 33], reactive terms are more significant than the resistance of the antenna. When reactive part of antenna impedance is zero, then it is called resonant antenna. Therefore in such antennas the resistance is the only non-zero term of impedance.

2.1.4 Radiation Efficiency

To combine the effect of all losses of the antenna on antenna radiation behavior, radiation efficiency (e_{rad}) is defined as follows [35–37]:

$$e_{rad} = \frac{R_{rad}}{R_{rad} + R_{all \text{ other losses}}} \quad (2.5)$$

2.1.5 Voltage Standing Wave Ration (VSWR)

Voltage Standing Wave Ration (VSWR) is a way to find the effect of antenna impedance on its radiation efficiency. By connecting a 50Ω generator through a 50Ω transmission line to the terminals of an antenna, depending on antenna impedance, some portion of the generator's power may reflect back to the generator.

$$\Gamma = \frac{Z_A - 50}{Z_A + 50} \quad (2.6)$$

The amount of the reflected power is determined by reflection coefficient (Γ), (2.6). When the connected source is 50Ω , the magnitude of the reflection coefficient of the power is $|\Gamma|^2$; its logarithmic form is shown with S_{11} , i.e. $S_{11} = 20\log_{10}(|\Gamma|)$, and many measurement devices can directly measure this value and display it in dB scale¹.

Standing waves on transmission lines are caused by transmitted or incident voltages and reflected power from antenna terminals. VSWR is a parameter that can be measured on a VNA to determine the amount of reflected power, and it is given as follows:

$$VSWR = \frac{1 + |\Gamma|}{1 - |\Gamma|} \quad (2.7)$$

2.1.6 Gain

There are different types of gains in antenna analysis. The multiplication of maximum directivity (D_0) and radiation efficiency is one type of gain., ($G = D_0 e_{rad}$). This type of gain determines the

¹Sometimes, in literature and books, instead of reflection power coefficient, Return Loss (RL) is a measure that makes more sense with positive sign rather than negative one: $RL = -20\log_{10}(|\Gamma|)$

amount of the radiated power (P_{rad}) from the antenna to the total captured power (P_{cap}) by the antenna from the connected generator, ($G = \frac{P_{rad}}{P_{cap}}$). If the antenna has significant losses, this value cannot be high.

Another type of Gain is Realized Gain, which measures the ratio of radiated power (P_{rad}) to the injected power by the connected generator to the antenna. In other words, if antenna is connected to the source through an unmatched transmission line, or it has been connected directly to an unmatched source, some powers is reflected back at the place of connection. Realized gain considers this reflection as well and its equation is as follows:

$$G_{real} = |\Gamma| \frac{P_{rad}}{P_{cap}} \quad (2.8)$$

Where $|\Gamma|$ is the reflection coefficient at the connection point to the generator, whether through a transmission line or directly to the antenna. Our design will incorporate this type of gain.

2.1.7 Open Circuit Voltage at Antenna Terminal

When an antenna is used in the receiving mode, the open-circuit voltage ($v_{o.c.}$) at the open terminals of the antenna is determined using Faraday's Law:

$$v_{o.c.} = \int_{\mathfrak{X}_a} \vec{\nabla} \times \vec{E}(\vec{r}) \cdot d\vec{s} = -j\omega \int_{\mathfrak{X}_a} \vec{B}(\vec{r}) \cdot d\vec{s} \quad (2.9)$$

Where $\mathfrak{X}_a$ is the cross section of the antenna terminal. By integrating on both sides of (1.4c). After applying Stoke's theorem, [1] and [35], we will obtain the following relation.

$$v_{o.c.} = \oint_{\mathfrak{L}_a} \vec{E}(\vec{r}) \cdot d\vec{l} = -j\omega \int_{\mathfrak{X}_a} \vec{B}(\vec{r}) \cdot d\vec{s} \quad (2.10)$$

Where $\mathfrak{L}_a$ is the curve that inscribes antenna terminal cross section. If the incident electric field upon antenna, i.e. $\vec{E}(\vec{r}) = \vec{E}^{inc}(\vec{r})$, is independent of integration path, (2.10) is converted to:

$$v_{o.c.} = \vec{E}^{inc}(\vec{r}) \cdot \vec{l}_{eff} \quad (2.11)$$

In (2.11), $\vec{l}_{eff}$ is called vector effective length of antenna.

Similarly, we can use the right-hand side of (2.9) for loop antennas, which are used extensively in this thesis.

$$v_{o.c.} = -\frac{d}{dt} \int_{X_a} \vec{B}(\vec{r}, t) \cdot d\vec{s}(\vec{r}, t) \quad (2.12)$$

When the antenna is small compared to wavelength, (2.12), can be written as:

$$v_{o.c.} = -\frac{d}{dt} (\vec{B}(\vec{r}, t) \cdot \vec{s}_{eff}(\vec{r}, t)) \quad (2.13)$$

In (2.13), $\vec{s}_{eff}(\vec{r}, t)$ is called vector effective surface. The phasor form of (2.13) is:

$$v_{o.c.} = -j\omega \vec{B}(\vec{r}) \cdot \vec{s}_{eff}(\vec{r}) \quad (2.14)$$

2.1.8 Noise Figure

Noise Figure (NF) is closely related to antenna radiation efficiency [38]. NF can be calculated using an antenna's equivalent noise temperature (T_A).

$$T_A = e_{rad}T_b + (1 - e_{rad})T_p \quad (2.15a)$$

$$T_b = \frac{\iint T_B(\theta, \phi) D(\theta, \phi) \sin \theta d\theta d\phi}{\iint D(\theta, \phi) \sin \theta d\theta d\phi} \quad (2.15b)$$

This parameter is determined based on brightness temperature (T_b), radiation efficiency (e_{rad}), and physical temperature (T_p) of antenna, according to (2.15). The temperature distribution in the region where the antenna is located is given by $T_B(\theta, \phi)$. In terrestrial applications, it is equal to the antenna's physical temperature, so that the antenna's NF is determined solely by its physical temperature.

2.1.9 Signal-to-Noise-Ratio (SNR)

Signal-to-Noise-Ratio (SNR) is a critical parameter in EMI. At the antenna terminal, the ratio of the power of desired signal to the power of undesired signal (Noise) is called the Signal-to-Noise-Ratio, or SNR. Desired signal power in this definition for EMI consists of the received signal to an antenna which can be from backscattering from a hot spot or traveled wave from a transmitter antenna. Different noise sources contribute to grain bin imaging, such as background temperature, and interference from other radio waves with the same frequency of operation. Using (2.15b), each noise source can be modeled by its equivalent noise temperature. Due to the fact that metals conduct heat and cold well, we cannot control the temperature changing effect on noise levels if the walls of the silo are made of metal. Metallic walls, however, provide good shielding against all other radio waves, which is a significant advantage of such silos.

2.1.10 Bandwidth

The frequency range in which an antenna can operate within some desired specification is the antenna bandwidth. These are typically defined for resonant antennas. An antenna's bandwidth can be determined by its radiation pattern, input impedance, or other factors. Our interpretation of the antenna bandwidth is based on its $VSWR$ (or equivalently S_{11}).

2.1.11 Quality Factor

Assume that d is the diameter of a sphere which inscribe the antenna, then antenna is said to be Electrically Small Antenna (ESA) if and only if $k_0 d \ll 1$, where k_0 is wave number in free space and $k_0 = \frac{\lambda_0}{2\pi}$. Generally, all antennas used in grain bin imaging are ESA.

Quality factor of ESA, Q , is an important parameter that combines bandwidth, efficiency, and size of the antenna. Generally, it is defined as follows [1], [35], and [39, 40]:

$$Q = \frac{\omega_0(W_e + W_m)}{P_{loss}} \quad (2.16)$$

Where ω_0 is resonance angle frequency, W_e and W_m are electric and magnetic stored energy, respectively and P_{loss} is the dissipated power by antenna whether in the form of radiation (P_{rad}), or ohmic loss (P_{ohm}); e_{rad} is radiation efficiency and Q_{rad} is radiation quality factor ($Q_{rad} = \frac{\omega_0(W_e+W_m)}{P_{rad}}$). For the case of small loop antenna which is mainly inductive, (2.16) can be written as follows:

$$Q = \frac{W_m}{P_{rad} + P_{ohm}} = \frac{2\omega_0 \frac{L|I|^2}{4}}{\frac{R_{rad}|I|^2}{2} + \frac{R_{ohm}|I|^2}{2}} = \frac{\omega_0 L}{R_{rad} + R_{ohm}} \quad (2.17)$$

Also, quality factor has the following relationship with bandwidth (BW) [39, 40].

$$Q = \frac{f_0}{BW} \quad (2.18)$$

2.2 Ferrite Materials

One method to improve existing shielded half loop antenna can be through putting ferrite plates in its core, therefore in this part we explain their basic parameters. In many applications in EM engineering, such as antennas, the permeability of ferrites is the most important property. The value of the permeability changes the performance of the EM structure or device. Generally, this quantity is a complex value and depends on the frequency of operation²:

$$\mu = \mu_0 \mu_r = \mu_0(\mu'_r - j\mu''_r) = \mu_0 \mu'_r(1 - j \tan \delta_m) \quad (2.19)$$

where $\tan \delta_m$ is called magnetic tangent loss.

In order to have an efficient small size antenna with a ferrite core, these materials should have high relative permeability, such as $\mu_r = 1000$ to $\mu_r = 2000$. However, this range of big values of relative permeability is generally possible at lower frequencies in the range up to couple MHz . As a result, these materials are usually suitable for using at frequencies below MHz . The higher the frequency, the lower the μ_r of these materials. Moreover, by increasing the frequency, μ''_r and therefore magnetic tangent loss of the material increases which prohibits using ferrite material in higher frequency structure with high efficiency. Still many high frequency structures and systems use ferrites, including switching power supply transformers, interference suppression, filters, waveguides, isolators, phase shifters, antennas, and precision inductance values, among others.

The general ferrite mixture is $MeFe_2O_3$ where Me is one of or a mixture of the following metals: Zinc (Zn), Manganese (Mn), Nickel (Ni), Copper (Cu), Iron (Fe), Cobalt (Co), and Magnesium (Mg) [41]. Manganese and Zinc ($MnZn$) and Nickel and Zinc ($NiZn$) are two important ferrite cores in the desired bandwidth. Ferrites have Curie temperatures (T_C). The temperature at which ferrite loses its magnetic property is this degree. Based on the process of fabrication of ferrites, many categories of ferrite cores differ in their permeability values, the power they can handle, and the suitable frequency range of operating. Some of these materials³ which will be useful for telecommunication systems are 3B1, 3F4, 3F45, eF5, 4A11, 3CD, 4B3, and 4C65. The datasheet of each one of these materials has important graphs like a graph of complex permeability, initial

²Note that both μ'_r and μ''_r , therefore $\tan \delta_m$, are functions of ω , i.e. $\mu'_r(\omega)$, $\mu''_r(\omega)$, and $\tan \delta_m(\omega)$, but for simplicity we do not write ω , clearly.

³These different materials have different frequency range of operation, therefore to choose the suitable material for each application, one should refer to their data catalogue [41].

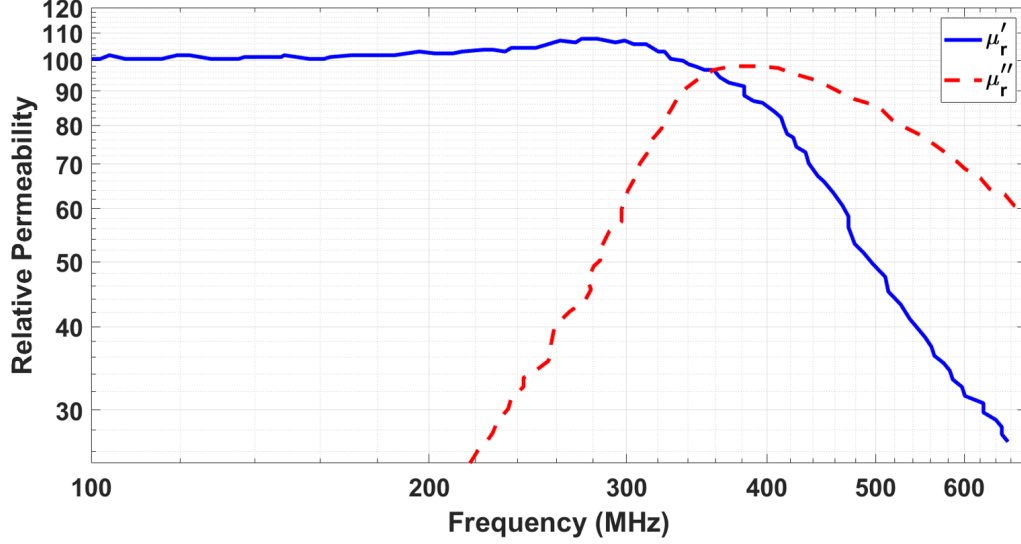


Figure 2.1: Complex permeability of 4C65 based on [41].

permeability, Curie temperature curve, and $B-H$ curve. For example Figure 2.1, shows the complex permeability of 4C65, [41], which we choose for our application.

Apart from permeability, the size and shape of the cores are also important. They come in different sizes and shapes, such as a plate, bar, toroid, E shape, P core, RM core, etc [41]. Plate shape ferrite is used in our antennas.

2.3 Reciprocity Theorem

Before analyzing the shielded loop antenna we need to review reciprocity theorem. The reciprocity theorem states that when we interchange source and measurement device that measures an EM quantity, we will have the same measured values in those two different situations. It has been explained very well in many textbooks such as [1] and [37], however we give an overview here as we will need this theory for most of the analysis in this thesis. Herein, this theory is derived from Maxwell's equations and restated in a lumped element circuit model.

Referring to Figure 2.2, consider two regions with the same geometry and media, but different sources within each region. Sources of $\vec{J}_a$ and $\vec{M}_a$ will produce $\vec{E}_a$ and $\vec{H}_a$. Similarly, sources⁴ of $\vec{J}_b$ and $\vec{M}_b$ will produce $\vec{E}_b$ and $\vec{H}_b$. According to (1.4), there is a relation between each category of sources and produced fields. This can be written as the following equations:

$$\vec{\nabla} \times \vec{E}_a = -j\omega\mu\vec{H}_a - \vec{M}_a \quad (2.20a)$$

$$\vec{\nabla} \times \vec{H}_a = j\omega\epsilon\vec{E}_a + \vec{J}_a \quad (2.20b)$$

$$\vec{\nabla} \times \vec{E}_b = -j\omega\mu\vec{H}_b - \vec{M}_b \quad (2.20c)$$

$$\vec{\nabla} \times \vec{H}_b = j\omega\epsilon\vec{E}_b + \vec{J}_b \quad (2.20d)$$

⁴Note that for simplicity we ignore $\vec{r}$, while we should keep in mind that the phasor forms of these quantities is a function of $\vec{r}$.

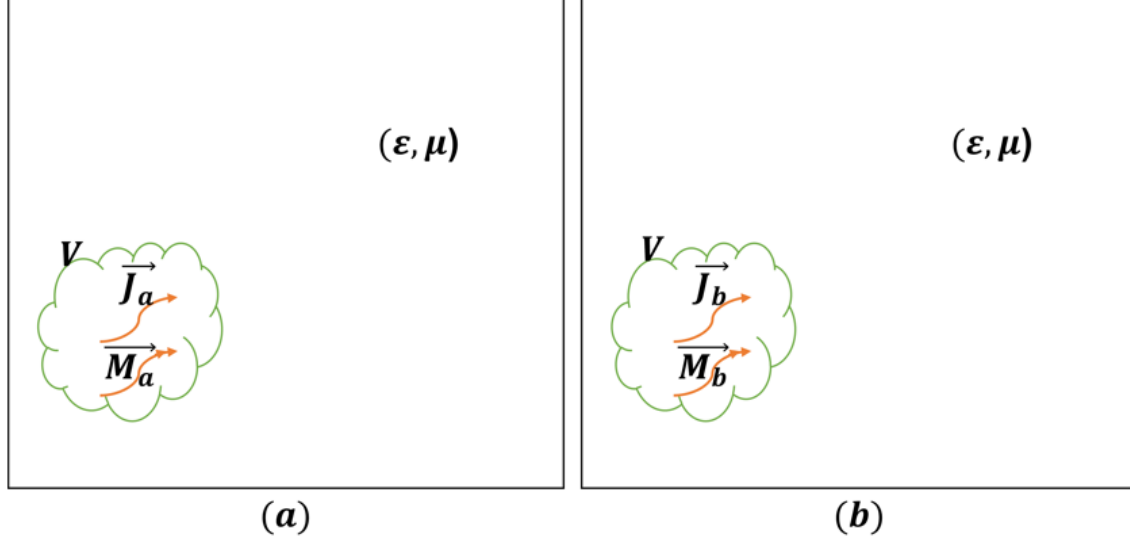


Figure 2.2: Two regions (a) and (b) have same geometry and same media, but they have different sources.

Consider the following vector equation, which is the differential form of Poynting theorem [1]:

$$\vec{\nabla} \cdot (\vec{E} \times \vec{H}) = \vec{E} \cdot \vec{\nabla} \times \vec{H} - \vec{H} \cdot \vec{\nabla} \times \vec{E} \quad (2.21)$$

Multiplying (2.20a) by $\vec{H}_b$, (2.20b) by $\vec{E}_b$, (2.20c) by $\vec{H}_a$ and the last one by $\vec{E}_a$, then using (2.21), integration over source regions, and applying Stoke's theorem, we will have the following relation which is called Lorentz reciprocity theorem.

$$\oint (\vec{E}_a \times \vec{H}_b - \vec{E}_b \times \vec{H}_a) \cdot d\vec{s} = \iiint (-\vec{H}_b \cdot \vec{M}_a + \vec{E}_b \cdot \vec{J}_a + \vec{H}_a \cdot \vec{M}_b - \vec{E}_a \cdot \vec{J}_b) dv \quad (2.22)$$

Where the right-hand side integration is a triple integral over the source regions, and the left-hand side is a double closed integral over the enclosing surface of source regions.

If we consider a point that is not in source region, from (2.20), we can prove that $\vec{\nabla} \cdot (\vec{E}_a \times \vec{H}_b - \vec{E}_b \times \vec{H}_a) = 0$ [1], therefore Lorentz reciprocity is written as (2.23).

$$\iiint_V (\vec{E}_b \cdot \vec{J}_a - \vec{H}_b \cdot \vec{M}_a) dv = \iiint_V (\vec{E}_a \cdot \vec{J}_b - \vec{H}_a \cdot \vec{M}_b) dv \quad (2.23)$$

(2.23) is a type of Ramsey's relation. Ramsey's formula, $\langle a, b \rangle = \langle b, a \rangle$, represents the relation between fields (first element) and sources (second element) of each problem. For instance in the case of (2.23), $\langle a, b \rangle = \langle \{\vec{E}_b, \vec{H}_b\}, \{\vec{J}_a, \vec{M}_a\} \rangle$ and $\langle b, a \rangle = \langle \{\vec{E}_a, \vec{H}_a\}, \{\vec{J}_b, \vec{M}_b\} \rangle$. In practice impressed sources are electric types and there is no practical magnetic radiating source, therefore $\vec{M}_a = \vec{M}_b = 0$. Referring back to the definition of open circuit voltage at the terminal of an antenna ((2.9) and (2.10)), we can represent (2.23) in terms of voltages and currents as follows:

$$\vec{v}_b \cdot \vec{i}_a = \vec{v}_a \cdot \vec{i}_b \quad (2.24)$$

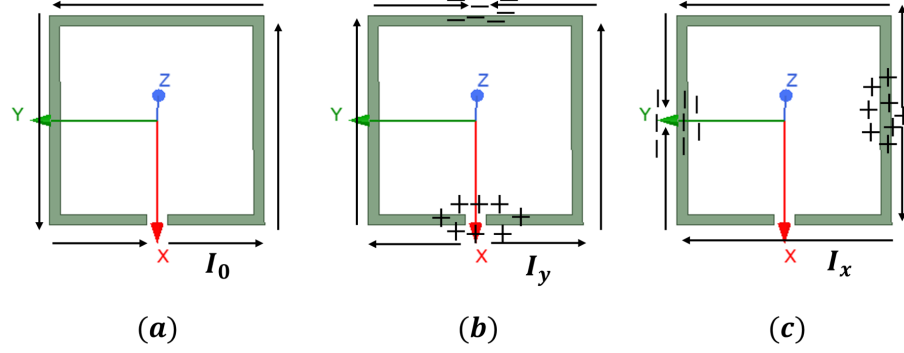


Figure 2.3: First three modes of current and charge distribution on the loop: (a) magnetic mode, (b) E_x mode, and (c) E_y mode.

(2.24) is the reciprocity theorem in lumped circuits. In Fundamental Electric Circuit courses this is called Tellegen's theorem [42].

In summary, based on the reciprocity theorem, we can swap the positions of source and observation and still get the same measured EM values. This is used in the following sections when we analyze balanced shielded loop antennas: sometimes in the transmitting mode and sometimes in the receiving mode; only for the sake of simplicity and convenience we choose the more straightforward mode to analyze.

2.4 Magnetic Probes

To know about shielded loop antennas, first, we should learn about the magnetic probe. Authors in [43] have explained and implemented a magnetic probe. Here, we explain magnetic probe with more details. Also, there are good probe analysis in [44, 45]. To begin analysis, similar to [43], consider a wire loop antenna with side length (a) that is very thin compared to the wavelength, Figure 2.3. Thus, current flows only in the line of the direction of the loop [46, 47]. Exact analysis of thin wire loop antennas can be found in [48]. By breaking the current into first three phase sequences⁵ we can extract some relations between incident magnetic fields and the voltage at the loop's terminal. Current distribution with one phase changing in x direction, and current distribution with one phase changing in y direction, respectively, Figure 2.3, correspond to electric dipole excitation. We call these modes electric x -dipole, and y -dipole modes, as they behave similar to the chosen corresponding names. Also, uniform current distribution is related to the magnetic incident field upon the loop. We call this mode magnetic current mode. Total induced open circuit voltage ($v_{o.c.}$) in the loop is calculated from (1.4c), by integrating from both sides.

$$v_{o.c.} = \oint_{\mathcal{L}_a} \vec{E}(\vec{r}) \cdot d\vec{l} = -j\omega \int_{\mathcal{X}_a} \vec{B}(\vec{r}) \cdot d\vec{s} \quad (2.25)$$

Total fields at the surface of the loop are composed of incident (B^i and E^i) and reradiated (B^r and E^r) fields, such that total fields (B^t and E^t) are $B^t = B^i + B^r$ and $E^t = E^i + E^r$. Ohm's law presents the relation between electric current density and electric fields, i.e. $\vec{J}(\vec{r}) = \sigma \vec{E}^t(\vec{r})$. Assuming linear

⁵See [43] for more information.

media, recalling relation of magnetic vector potential with magnetic field, ($\vec{B}(\vec{r}) = \mu \vec{\nabla} \times \vec{A}(\vec{r})$), and also using Stoke's theorem, [1], we will obtain (2.26).

$$-j\omega \int_{\mathfrak{X}_a} \vec{B}^i(\vec{r}) \cdot d\vec{s} = j\omega \int_{\mathfrak{X}_a} \vec{B}^r(\vec{r}) \cdot d\vec{s} + \oint_{\mathfrak{X}_a} \vec{E}^t(\vec{r}) \cdot d\vec{s} = j\omega \int_{\mathfrak{X}_a} \mu \vec{\nabla} \times \vec{A}(\vec{r}) \cdot d\vec{s} + \oint_{\mathfrak{X}_a} \vec{E}^t(\vec{r}) \cdot d\vec{s} \quad (2.26a)$$

$$= j\omega \oint_{\mathfrak{L}_a} \vec{A}(\vec{r}) \cdot d\vec{l} + \oint_{\mathfrak{X}_a} \vec{E}^t(\vec{r}) \cdot d\vec{s} \quad (2.26b)$$

We know that $\vec{J}(\vec{r}) = \frac{\vec{i}(\vec{r})}{X_{loop}}$, where X_{loop} is the width of the metal path of the square loop, which magnetic mode current $\vec{i}(\vec{r}) = I_0$ flows on it. Where $k = \omega \sqrt{\mu_0 \epsilon_0}$ is the wave number of free space. We know that the relation of $\vec{A}$ is as follows:

$$\vec{A}(\vec{r}) = \frac{\mu_0}{4\pi R} \int \frac{\vec{J}(\vec{r}') e^{-jkR}}{R} d\vec{l}' \quad (2.27)$$

Where R is the distance between source and observation points. By organizing (2.26) based on incident and reradiated fields, along with mentioned relation for magnetic potential vector, open circuit voltage at the terminal of the loop we will obtain according to the following equation, which is the same as (3) in [43].

$$-j\omega \int_{\mathfrak{X}_a} \vec{B}^i(\vec{r}) \cdot d\vec{s} = j\omega \oint_{\mathfrak{L}_a} \oint_{\mathfrak{L}'_a} \frac{\mu_0 I(\vec{r}')}{4\pi} \frac{e^{-jkR}}{R} d\vec{l}' \cdot d\vec{l} + \oint_{\mathfrak{L}_a} \vec{i}(\vec{r}) Z^{PUL} \cdot d\vec{l} \quad (2.28)$$

where $Z^{PUL} = \frac{1}{\sigma X_{loop}}$ is per unit length internal impedance of the loop (assuming that σ is the conductivity of the metallic path in the loop) [43], and $\mathfrak{L}'_a$ is the curve path which closes surface current distribution.

By plugging I_0 in the right-hand side of (2.28), a relation between normal incident field and magnetic current mode is obtained and magnetic sensitivity (K_B) is defined as follows [43].

$$K_B = \frac{Y_o h_b}{\lambda} \quad (2.29a)$$

$$h_b = -jkS \quad (2.29b)$$

$$Y_o = j\omega \oint_{\mathfrak{L}_a} \oint_{\mathfrak{L}'_a} \frac{\mu_0}{4\pi} \frac{e^{-jkR}}{R} d\vec{l}' \cdot d\vec{l} + \oint_{\mathfrak{L}_a} Z^{PUL} \cdot d\vec{l} \quad (2.29c)$$

Where h_b is called magnetic effective length, a similar concept to electrical effective length in straight antennas; Y_o is the input impedance of the loop that carries a constant current distribution. Finally, if we consider the appropriate direction of magnetic field to the loops of Figure 2.3, magnetic current mode, magnetic sensitivity, and incident field are related through (2.30) [43].

$$K_B = \frac{I_0}{\lambda c B_z^{inc}} \quad (2.30)$$

where c is speed of light ($\approx 3 \times 10^8 \frac{m}{s}$).

Now consider Figure 2.3b. With this type of current distribution, the determined charge will

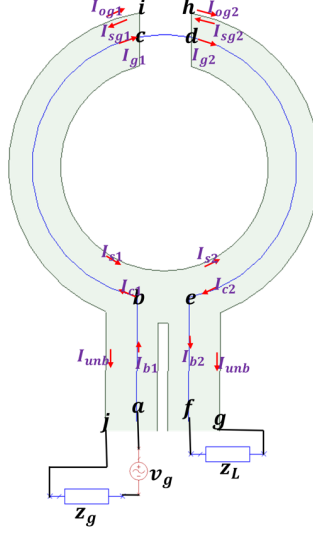


Figure 2.4: Balanced shielded loop antenna with determined current in each section of the antenna.

distribute as it has been shown in that figure and will result into only E_x fields. Recalling reciprocity, this type of current corresponds to incident $\hat{y}$ electric component. Similarly we can justify relation between current and incident field in (2.3c). Hence we can go through the earlier procedure, derive and define similar relation to (2.30) for these bipolar electric currents and electric sensitivity (K_E) [43]. Again, we emphasize that loop is very small and thin.

$$K_E^y = \frac{I_y}{\lambda E_y^{inc}} \quad (2.31a)$$

$$K_E^x = \frac{I_x}{\lambda E_x^{inc}} \quad (2.31b)$$

In [43], authors have purposed several methods to measure magnetic and dipole modes currents.

2.5 Balanced Shielded Loop Antenna

When the length of the loop is comparable with wavelength, all of those mentioned current modes will appear on the loop. Since we do EMI in different frequency ranges antenna length may be in the order of some wavelengths. So, if one is interested in only measuring the magnetic field, then some way must be found to eliminate the measurement of mentioned electric dipoles on the loop. Balanced Shielded Loop Antennas, or simply Shielded Loop Antennas (SLA) are an antenna that can measure only magnetic mode current, equivalently only normal magnetic field to the surface of the loop. A simple shielded loop antenna made of a coaxial cable has been shown in Figure 2.4, with current flowing in each antenna section.

This antenna is made of two coaxial cables; one consists abc core and ji shield, the other one has def core and gh shield. As it is seen one cable is connected to a generator with voltage v_g and impedance z_g , and the other cable is attached to a load z_L . The gap is implemented at cd , balanced current I_{b1} at point b is converted to core current I_{c1} ; this current induces the counter phase current

in its shield (I_{s1}). Core current at gap point (c) becomes I_{g1} ; if gap is small, then $I_{sg1} = I_{sg2}$. Because of discontinuity at the shield, some portion of I_{s1} at point i , i.e. I_{og1} , flows through outer surface of the shield as well [49, 50]. The same scenario holds for the right side of the balanced shielded loop in Figure 2.4.

Finally, by applying Kirchhoff's Current Law (KCL) at each determined points by (a,...,h) in Figure 2.4, relation between currents is determined [49]. It is found that $I_{og1} = -I_{og2}$, when current and loop are balanced. Therefore using a balanced loop which has a small gap at its middle part will result into cancellation of x-direction dipole mode (2.3). When we compare figures 2.3 and 2.4, at least one of those dipole modes is canceled out (depending on orientation of coordinate system at the antenna's plane) which will enhance antenna capability in detecting magnetic mode alone.

For instance consider a right hand rectangular coordinate system with $\hat{z}$ out of the plane of the plane of antenna in Figure 2.4, along with an incident far field with incident magnetic field normal to the surface of loop and polarization $\hat{y}$. The loop can cancel out induced current as the result of $\vec{E}^{inc}$ and only induced current at the terminal of antenna will be originated from $\vec{B}^{inc}$; if we rotate that loop a dozen degrees in its plane, its output current is no longer only from the magnetic field but also from the electric incident field as well. Or if we rotate the plane of the loop as much as 90° , then it will detect the current due to dipole and magnetic part will be cancelled out. Because the balanced shielded loop can detect magnetic induced current separately from other dipoles, this type of antenna is also used as a direction finder: direction of the maximum magnetic field (equal to the normal magnetic field to the antenna's surface).

2.6 Lumped Model of Balanced Shielded Loop Antenna

A lumped model of the shielded loop antennas is useful in estimating their operation frequency, bandwidth, and efficiency. The exact analysis of balanced shielded loop antenna can be found in [50]. Here we represent a simple explanation for this antenna and derive the complete lumped model of shielded loop antenna in Figure 2.4. This antenna is the combination of two transmission lines and an antenna. The outer shield is the antenna part that its analysis is the same as small loop antennas in [35]. The input impedance of this antenna is the result of two observed impedance by each transmission line. To calculate each of those impedance we can use (2.32) with z_L or z_g for each transmission line as Z_L .

$$Z_{in} = Z_0 \frac{Z_L + jZ_0 \tan(\beta l)}{Z_0 + jZ_L \tan(\beta l)} \quad (2.32)$$

Further, a loop antenna is modeled by its inductance, radiation resistance, and ohmic resistance. Accordingly, the balanced shielded loop antenna's equivalent circuit is shown in Figure 2.5. According to [35], values of each lumped element can be obtained from (2.33).

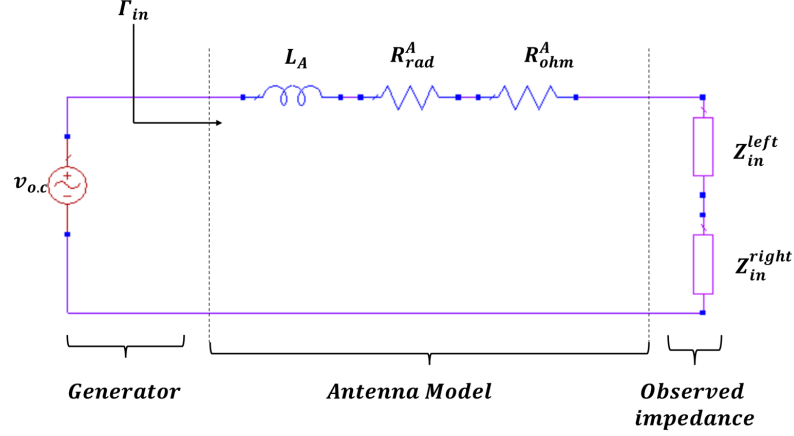


Figure 2.5: Lumped model of balanced shielded loop antenna.

$$L_A = \mu_0 a \left(\ln\left(\frac{8a}{b}\right) - 2 \right) \quad (2.33a)$$

$$R_{rad}^A = 31200 \left(\frac{S}{\lambda^2} \right)^2 \quad (2.33b)$$

$$R_{ohm}^A = \frac{b}{a} R_s = \frac{b}{a} \sqrt{\frac{\omega \mu_0}{2\sigma}} \quad (2.33c)$$

$$Z_{in}^{left} = Z_0 \frac{z_g + jZ_0 \tan(\beta l)}{Z_0 + jz_g \tan(\beta l)} \quad (2.33d)$$

$$Z_{in}^{right} = Z_0 \frac{z_L + jZ_0 \tan(\beta l)}{Z_0 + jz_L \tan(\beta l)} \quad (2.33e)$$

$$v_{o.c.} = \vec{l}_{eff} \cdot \vec{E}^{inc} = -j\omega \vec{B}^{inc} \cdot \vec{S}_{eff} \quad (2.33f)$$

Where a is the loop radius; b is the wire radius; Z_0 is the impedance of the coaxial transmission line. In this structure, coaxial cables are the same type. Once we have lumped element values, we can find the antenna's reflection coefficient and other parameters.

2.6.1 Lumped Model of Balanced Shielded Loop Antenna Loaded with Ferrite

Ferrite is an anisotropic material used in many RF devices and systems, such as circulators and isolators. Depending on its external magnetic field bias, ferrite can propagate some polarizations of waves and prevent others [51]. However, there is no external magnetic field bias for our purpose. We start the analysis of ferrite-loaded shielded loop antenna by developing an approximate lumped circuit model. Figure 2.6 is the same as Figure 2.4, only with a thin ferrite plate that has been located in its center.

We model ferrite material by deriving its inductance and resistance. Since the plate and antenna dimensions are significantly smaller than the wavelength, we use quasi-static approximation and calculate the inductance and resistance of this material in a method similar to magnetostatics and electrostatics. These calculations are similar to usual self and mutual inductance calculations that

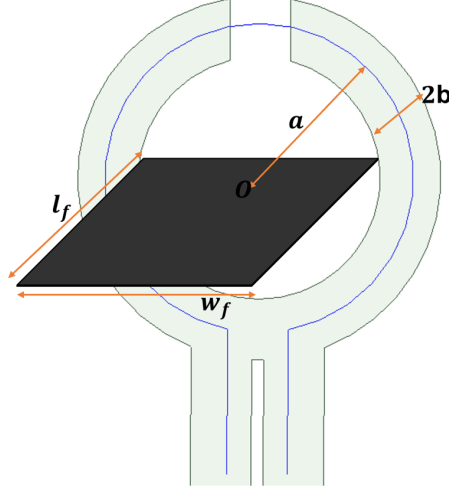


Figure 2.6: Balanced shielded loop antenna with a thin plate of ferrite (black color) in its center.

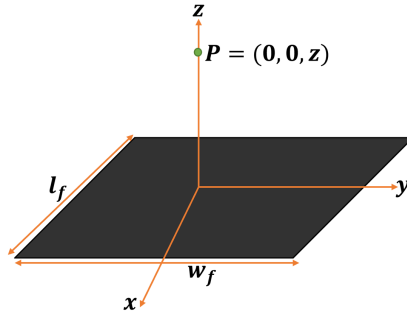


Figure 2.7: Symmetric observation point to calculate self inductance of the plate.

can be found in many engineering electromagnetic books like [52, 53].

The added inductance is calculated based on the self-inductance of the ferrite plate (L_P) and the mutual inductance between ferrite plate and antenna loop (M). General calculation of the self-inductance of the plate is not easy. However, considering that the ferrite plate is located at the center of the loop and it has been expanded symmetrically from both sides of the loop, we calculate the total magnetic field of the plate at a point that is located on the symmetric axis of the plate (point P in Figure 2.7). Consider a uniform magnetic current distribution on the surface of the plate⁶ K_0 , ($K_0 = \frac{I_0}{w_f}$). After that we use magnetic stored energy relation to calculate the inductance of the plate (L_P) and its resistance (R_P). Finally it should be considered that L_P is obtained by $\{\lim_{z \rightarrow 0}\}$ in calculating self-inductance of the plate from (2.35).

⁶Note that I_0 is the constant flowing current on the magnetic plate in the direction of $\hat{x}$. Also, according to Figure 2.7, w_f is the width of the plate.

$$H = \frac{I_0 l_f}{2\pi w_f} \quad (2.34a)$$

$$\frac{1}{2} L I_0^2 = \frac{1}{2} \mu |H|^2 w_f l_f t_f \quad (2.34b)$$

$$\mu = \mu_0 (\mu'_r - j\mu''_r) \quad (2.34c)$$

$$L_P = \mu_0 \mu'_r \left(\frac{l_f}{2\pi w_f} \ln \left| \frac{w_f + \sqrt{l_f^2 + w_f^2}}{l_f} \right| \right)^2 w_f l_f t_f \quad (2.34d)$$

$$R_P = \omega \mu_0 \mu''_r \left(\frac{l_f}{2\pi w_f} \ln \left| \frac{w_f + \sqrt{l_f^2 + w_f^2}}{l_f} \right| \right)^2 w_f l_f t_f \quad (2.34e)$$

In the above equations, H is the magnetic intensity produced by the flowing current (I_0) on the ferrite plate with permeability μ , width w_f , thickness t_f , and length l_f , Figure 2.7. Note that μ'_r and μ''_r are relative real and imaginary part of permeability of ferrite plate, respectively. Also, $\omega = 2\pi f$ is the angular frequency. We can calculate the mutual inductance (M) and resistance (R_m) between the loop and the ferrite plate in the same scenario. General calculation is not easy, and numerical methods do it; however, we can use an approximation of mutual inductance for our purposes. Magnetic field at the center of the circular loop is $H = \frac{I_0}{2a}$, then using (2.35), and going through the same procedure we will obtain mutual values as follows:

$$M = \mu_0 \mu'_r \left(\frac{l_f w_f}{2a} \right) \quad (2.35a)$$

$$R_m = \omega \mu_0 \mu''_r \left(\frac{l_f w_f}{2a} \right) \quad (2.35b)$$

Also, we can calculate the capacitance in the gap, by using quasi-static approximation method, parallel plate capacitor in the gap (C_g):

$$C_g = \epsilon_0 \frac{\pi(r_s^2 - r_{ins}^2)}{g} \quad (2.36)$$

where r_s and r_{ins} are radius of shield and insulator, respectively; g is the gap's length. The final lumped model of the loaded shielded loop antenna has been represented in Figure 2.8.

From the lumped model we can calculate resonant frequency of the antenna. When transmission line instead of short ended is opened, then that piece of transmission line's impedance is calculated through $Z_{in}^{right} = -jZ_o \cot \beta l$. However, when it is short circuited, its impedance is calculated through $Z_{in}^{right} = jZ_o \tan \beta l$. To verify the developed model, we have tuned resonant frequency by changing ferrite width (w_f). This has been presented in Figure 2.9. Note that in this figure length of the ferrite is kept constant 100 mm. As it is seen from Figure 2.9, when width of the ferrite plate is close to zero, resonant frequency closes to 2 GHz, which is very close to the experimental findings; without ferrite plate, resonance frequency of the antenna is almost 1.8 GHz. As noted earlier, several parts of the model were calculated using very rough estimations. It predicts the antenna's behavior somewhat well, however.

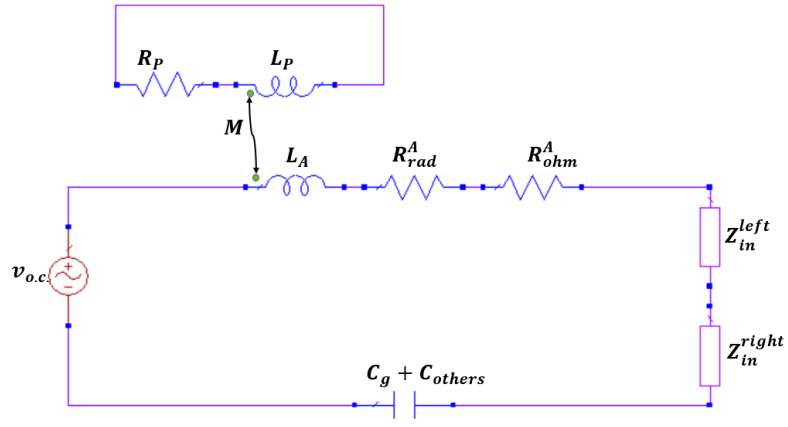


Figure 2.8: Lumped model of ferrite loaded balanced shielded loop antenna.

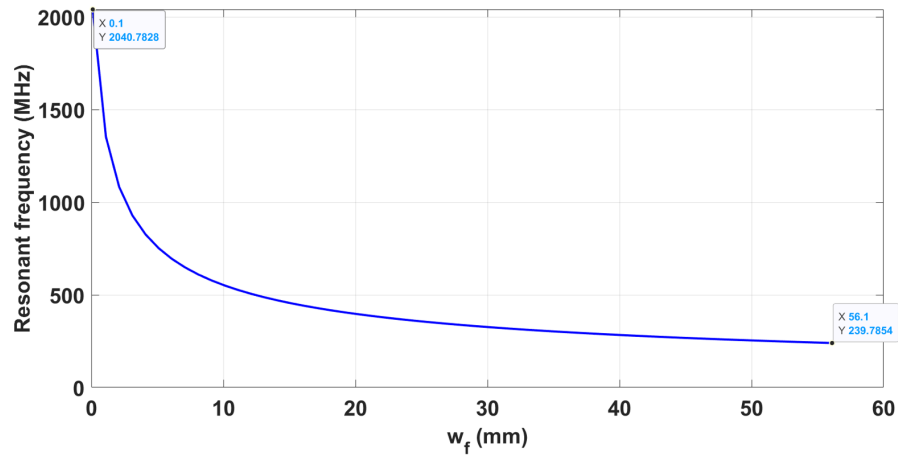


Figure 2.9: Tuning resonance frequency by changing ferrite width.

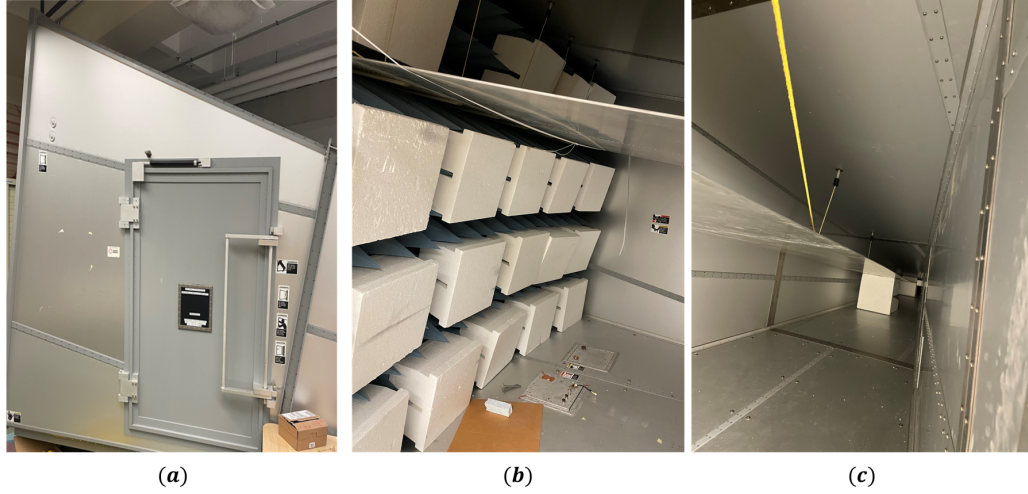


Figure 2.10: Gigahertz Transverse Electromagnetic (GTEM) cells: (a) outside of rectangular base pyramid chamber, (b) terminating end of GTEM which has been filled with electromagnetic absorbers, and (c) inner tapered conductor along with top and bottom metallic walls.

2.7 Method of Finding Sensitivity of SLA to Magnetic Mode Currents

SLAs can measure magnetic mode current, which is proportional to the normal magnetic field to the surface of the antenna if they design appropriately. Indeed, detecting magnetic mode current is one of the primary goals in using SLAs. The magnetic field is sensed most strongly when it is perpendicular to the shielded loop's surface. With Gigahertz Transverse ElectroMagnetic (GTEM) cells, we can find how well the shielded loop antenna can measure the magnetic field while rejecting the dipole modes of electric fields.

2.7.1 Gigahertz Transverse ElectroMagnetic (GTEM) cells

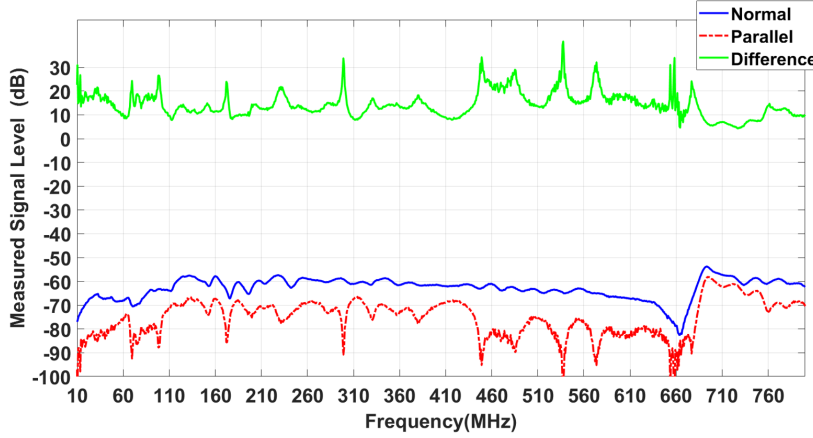
Gigahertz Transverse Electromagnetic cells⁷ are rectangular based pyramid chamber that can produce TEM waves. As it is seen from Figure 2.10, they are made of a tapered rectangular conductor at the middle of top and bottom metallic walls. Therefore they are some types of stripline. A $50\ \Omega$ source that connected to one end of this chamber, generates TEM waves. At the other end of the chamber some absorber are installed which behaves like a $50\ \Omega$ load at low frequencies, while they attenuate higher frequencies. Therefore this chamber can be considered a chamber that works from DC, which is impossible in anechoic chamber. In other words with GTEM cell we can do far field tests from DC to microwave frequency ranges.

Considering the produced TEM fields by GTEM cell, we changed the antenna position to normal and parallel to the generated magnetic field by the GTEM cell⁸ to measure the received signal with two orthogonal directions⁹ to the antenna surface. The difference between the two situations estimates the antenna's ability to reject the normal electric field. Figure 2.11 shows the magnetic

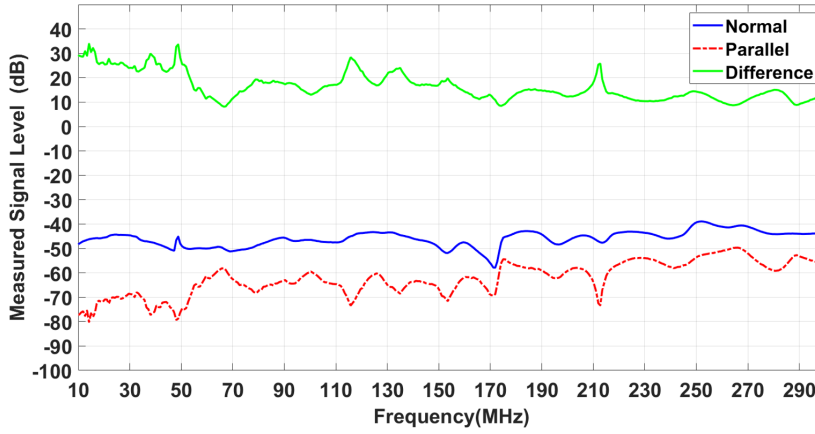
⁷We used GTEM with series number 5300 manufactured by Electro Mechanics Company- Texas Austin.

⁸Magnetic field is circulating around inner tapered conductor.

⁹These two orthogonal directions are different from the polarization of antennas described in the previous section.



(a)



(b)

Figure 2.11: Received signal in two orthogonal situations inside GTEM cell by (a) shielded loop antenna in small bin [33] and (b) shielded loop antenna in large industrial bin [30]

mode current sensitivity of antennas in available small [33] and large [30] grain bin imaging. Note that the normal incident is the desired position in both plots, and the parallel incident is undesirable.

In plots of Figure 2.11, amount of received signal in each position is not essential, but their difference determines how well the antenna can reject the normal component of electric field. For instance for small antenna the difference between desired and undesired position of antenna at 260 MHz is 12 dB. As a result, in the desired mode, the antenna receives about 15 times more power than in the undesired mode. Or in the large antenna, the desired signal is almost 79 times bigger than the received signal when the antenna is in undesired position. In subsequent chapters, we present some antennas that enhance this quantity.

2.8 Conclusion

In this chapter after reviewing some characteristic performance parameters of antennas we examined ferrite material properties. Reciprocity theorem was stated and used in explanations of shielded loop

antennas. Developing a lumped circuit model for shield loop antennas with and without ferrite was done. Finally, we introduced a method to find the sensitivity of shielded loop antennas to magnetic fields and rejecting electric fields, i.e. GTEM cell.

Chapter 3

Matching and Ferrite Loading to Improve Antenna's Performance

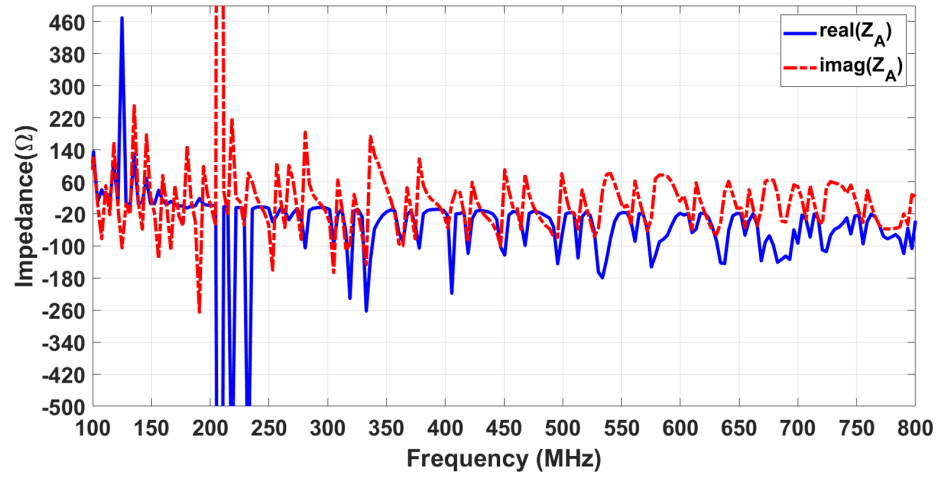
In this chapter we present some methods to improve radiation characteristics properties of mentioned antenna in grain bin imaging. First, we present magnitude of reflection coefficients and input impedance of small and large antennas at [33] and [30] respectively. As it has been mentioned in the first chapter size of the small shielded loop antenna is 56 mm×22 mm and the large shielded loop antenna is 400 mm×100 mm. We use these types of plots as our starting points of improvement. Then some improvement methods and ideas are stated. The methods are categorized as matching and ferrite material loading. Matching techniques are widely used in ESA matching. Also, based on the mentioned properties of ferrite in the previous chapter, here we use it in the structure of antenna to achieve a resonance at the antenna's frequency. There have been several suggestions, which have been designed and simulated. Finally we present a new compact antenna which is the scaled version of small Shielded Half Loop Antenna (SHLA) in [33] and can be used in industrial big bins.

3.0.1 Existing Antennas' Performance Parameters

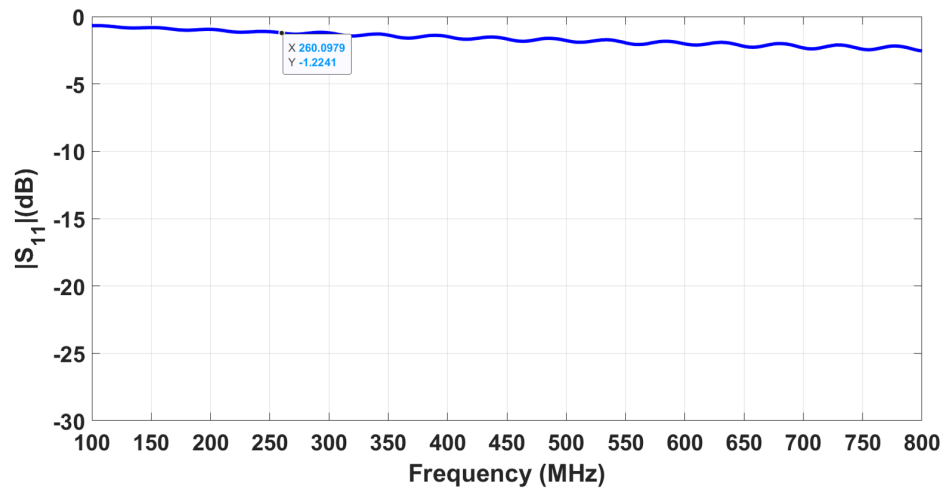
As mentioned, antenna's impedance plays an important role in its performance and connecting to the remaining parts of EMI system. So, we use impedance plots as the starting point for improving the antenna performance (e.g. it gives us information on how to change the impedance to get a better match). To know impedance of small [33] and large size [30] antennas, we installed each antenna on the wall of an appropriately sized bin and their S-parameter and impedance have been measured by VNA. Figure 3.1, represents the measured S_{11} and impedance of antenna in [33]. Also, Figure 3.2, represents the measured S_{11} and impedance of antenna in [30]. Note that small and large antenna refer to those ones that have been installed inside small ([33]) and large size ([30]) grain bins, respectively. Of course, the the size of the antenna in small grain bin is smaller than the size of antenna in large one.

As it is seen from impedance plots in these figures, small and large size antennas have some resonance and antiresonance frequencies¹. Furthermore, at the resonance frequency, they do not

¹Resonance and antiresonances occur when reactive part of antenna's impedance is zero. If we show reactive impedance with X_A , then resonant points are zero crossing point with positive slope at the frequency of zero crossing (ω_0), i.e. $\frac{dX_A}{d\omega}|_{\omega_0} > 0$. However, in antiresonant point rate of the change of reactive part at zero crossing frequency

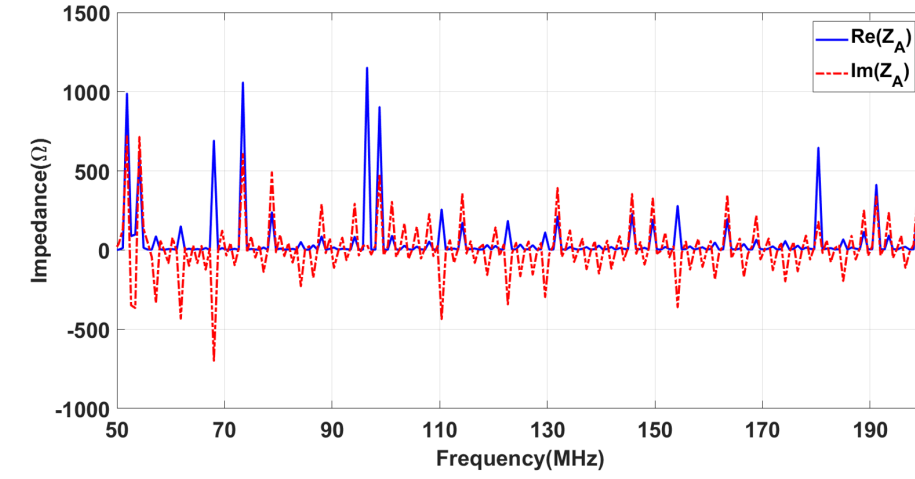


(a)

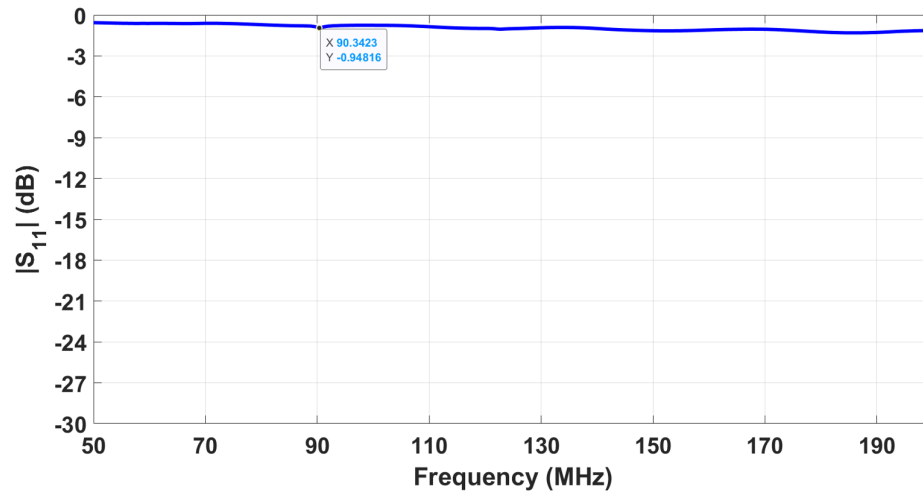


(b)

Figure 3.1: (a) Measured impedance and (b) Measured reflection coefficient of reported SHLA in [33]; size of that antenna is 56 mm $\times$ 22 mm.



(a)



(b)

Figure 3.2: (a) Measured impedance and (b) Measured reflection coefficient of reported SHLA in [30]; size of that antenna is 400 mm×100 mm.

have enough radiation resistance. So, at none of those frequencies we expect to have a well matched antenna to 50^{Ω} , which is the impedance of excitation source. This is confirmed by looking at reflection coefficient plots for each of those antennas. The corresponding values of S_{11} at the reported imaging frequency in each paper (618 MHz for the small and 89.9 MHz for the large antennas), have been determined on plots of figures 3.1 and (3.2).

3.1 Matching Technique Using Passive Components

One possible method of improving the performance of the existing SHLA antennas is through the use of matching. At minimum, a better matched antenna would mean less power reflected back to the switch and source. Using matching, you can transfer the maximum amount of power from a generator to a load. Applications include circuits and antennas. If the load impedance is Z_L and the source's impedance is Z_g , then the matching condition means $Z_L = Z_g^*$. The antenna does not radiate at its maximum power if the impedance of the source connected to the antenna is not matched with the antenna's impedance (Z_A). Furthermore, when the antenna is connected to a non-matched transmission line, the reflected power may damage the receiver and/or transmitter. To resolve these issues, adding a matching circuit between antenna and generator or antenna and load can be very helpful. Nevertheless, it should be noted that the matching circuit does not enhance gain, directivity, or any other characteristic of the antenna. When the antenna is connected to a load through a matching circuit, it does not receive more power; instead, it will receive the same amount of power as in the case without the matching circuit. However, the matching circuit will deliver an increased proportion of the absorbed power (equal to the case without the matching circuit) to the connected load. Various matching techniques have been developed for different circuits or antennas, which are detailed in the following parts.

3.1.1 Passive Lumped Element Circuits

If the frequency of circuit or antenna is not too high, usually under 100 MHz, matching circuit with lumped component is one possible solution. There are different types of matching circuits. Table (3.1), shows three important types of matching circuits: L, T, and Π matching circuits. It is impossible to match elements for broadband applications with this type of matching mechanism, because the lumped elements have high-quality factors limiting the bandwidth. Additionally, to use these matching circuits, we need to opt for a resonance or anti-resonance frequency of the antenna since they do not compensate for reactive impedance.

Matching formulas are easy and can be found in many RF electronic books like [38]. They are briefly mentioned, however. The general matching formula is the same as the conjugate matching one. Using the conjugate matching formula, we will have the following values for L-type matching (series (X_s) and parallel (X_p) component).

$$X_p = \pm \sqrt{\frac{R_g}{R_L - R_g}} \quad (3.1a)$$

$$X_s = \mp \sqrt{R_g(R_L - R_g)} \quad (3.1b)$$

is negative, i.e., $\frac{dX_A}{d\omega}|_{\omega_0} < 0$.

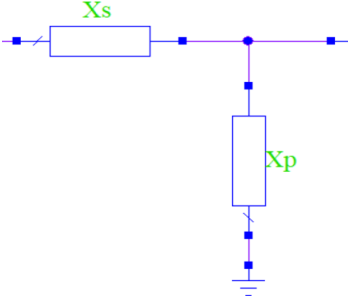
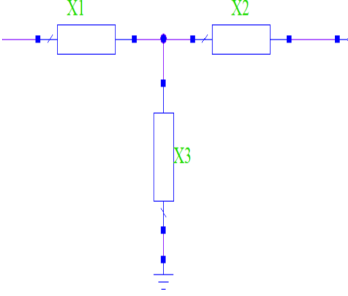
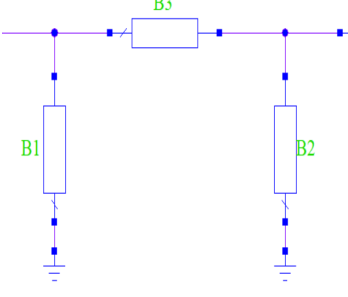
Matching Type	Circuit Schematic
L-Matching	
T-Matching	
Π-Matching	

Table 3.1: Different matching circuits' schematic with lumped elements.

These values were calculated assuming the load resistor (R_L) is connected to the left port of the matching network, and the source resistor (R_g) is attached to its right port. According to (3.1), always, X_s and X_p are different types of elements.

In T-Type matching, the same method and assumption are used to find the values of the parallel and series components.

$$X_3 = \mp R_L \sqrt{\frac{R_g}{R_L - R_g}} \quad (3.2a)$$

$$X_1 = \pm R_g \sqrt{\frac{R_g}{R_L - R_g}} - X_3 \quad (3.2b)$$

$$X_2 = \pm R_L \sqrt{\frac{R_g}{R_L - R_g}} - X_3 \quad (3.2c)$$

As in (3.2), parallel and series components are of different types in T-Type matching. Using the

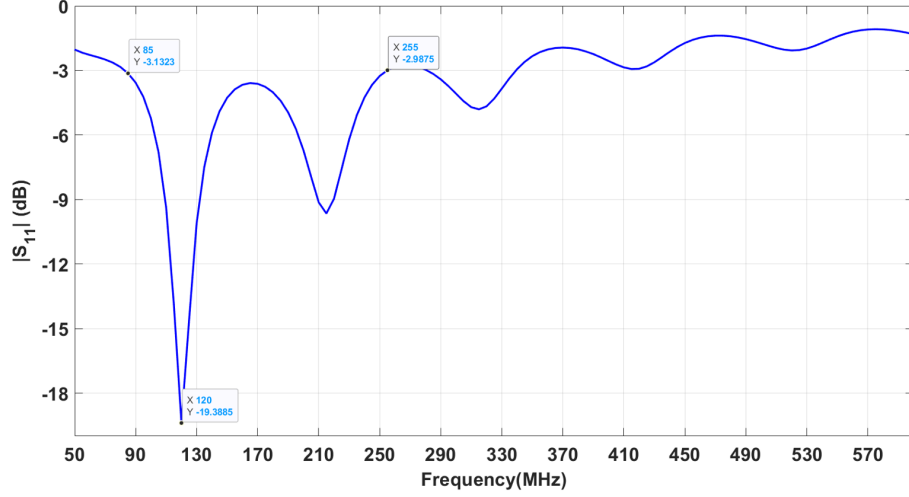


Figure 3.3: ADS simulation results for L-type matching to SHLA in [33]

same assumptions and method, values of Π -type matching circuits are found:

$$B_3 = \frac{1}{X_3} \quad (3.3a)$$

$$B_1 = \frac{1}{X_1} \quad (3.3b)$$

$$B_2 = \frac{1}{X_2} \quad (3.3c)$$

In the following of this sections we present implementations of matching circuits and simulation results for the discussed antennas in the previous chapters, i.e. small [33] and large [30] size SHLAs. For simulating the matching circuits we use Keysight- Advanced Design System (ADS). Since both mentioned SHLAs are available, we measure their S -parameters, then import them to ADS to use in matching designs. The values of elements in matching circuits have been modified based on standard components available in the market.

Implementing L-type Matching

We try to tune the small size SHLA, [33], by implementing a L-type matching circuit, to be suitable for using in large size bins. By selecting X_s a capacitor with 22 pF and X_p an inductance of 300 nH, a good matching point occurs at 120 MHz. Consequently, it is possible to use this antenna in the larger bins, but it will not absorb enough power, because of its limited size and effective area of the loop (see (2.13)). Keysight Advanced Design System (ADS) simulated results are shown in Figure 3.3.

Figure (3.3) shows a bandwidth of 170 MHz for $S_{11} \leq -3$ dB, which is in the desired criteria range. By choosing a 2 nH inductance for X_s and 100 pF capacitor for X_p , large SHLA in [30] is matched at 120 MHz. However, as it is seen from ADS simulation result in Figure (3.4), the match is very narrow, as we predicted earlier. Other types of matching circuits will produce similar results, so their results will not be presented here.

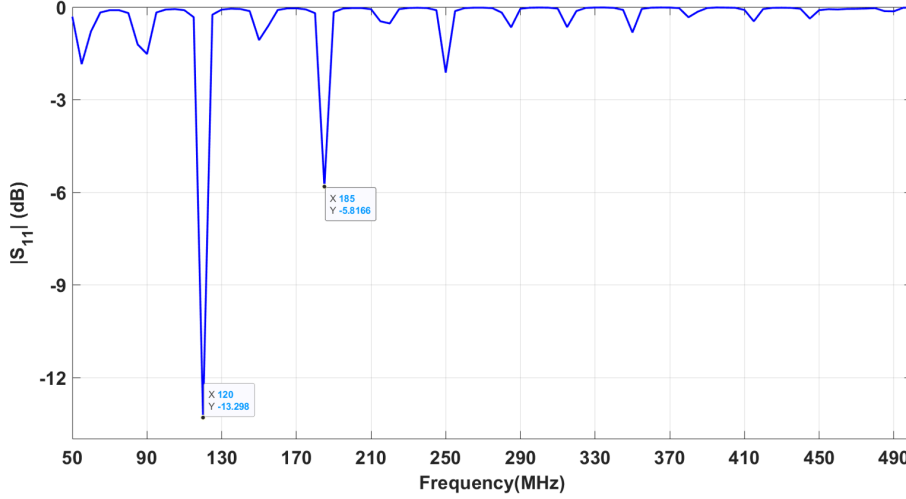


Figure 3.4: ADS simulation results for L-type matching to SHLA in [30]. Note that antenna is installed in the bin.

3.1.2 Transmission Line Matching

When a transmission line is used for matching circuits, it is usually done at high frequencies. Radio Frequency Integrated Circuits (RFICs) are the major application of them. (2.32) is the main idea behind matching through a transmission line, where Z_L is the load impedance that has been connected to the end of the transmission line with impedance Z_0 ; Z_g is the impedance of generator that transmits power to the line; and Z_{in} is the total input impedance from the generator side, which is loaded by the transmission line and the attached load to its end. Since the lengths of transmission lines (l) are small in matching circuits, they are considered lossless, and waves propagate with a constant phase propagation² β .

In our application size of the implemented transmission line at our desired frequencies is too long, therefore it is not practically feasible. Also, we should not neglect that this technique likewise the lumped element matching technique is a narrow band one.

3.2 Matching Technique Using Active Circuits (Non-Foster Methods)

Based on the Foster's paper [54], reactance ($X(\omega)$) or susceptance ($B(\omega)$) of every passive two-terminal element has a positive slope versus the frequency (ω), i.e.:

$$\frac{\partial X(\omega)}{\partial(\omega)} > 0 \quad (3.4a)$$

$$\frac{\partial B(\omega)}{\partial(\omega)} > 0 \quad (3.4b)$$

²Propagation constant or β is the speed of changing of phase in the lossless medium and is related to wavelength λ by $\beta = \frac{2\pi}{\lambda}$ [1] and [35].

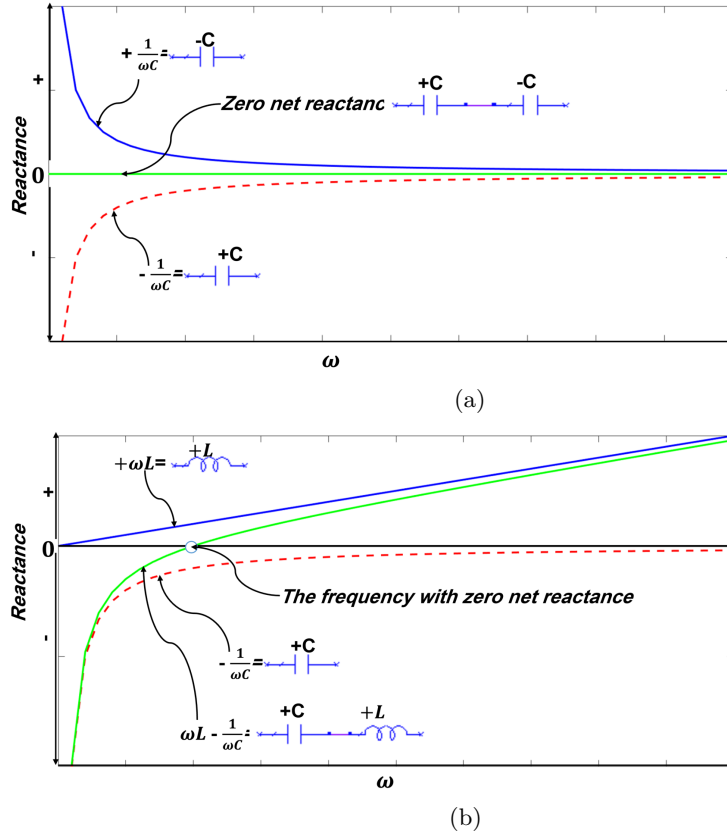


Figure 3.5: Comparison between (a) Non-Foster matching and (b) conventional matching techniques. In (a) we have a zero reactance in all frequencies (wide band), while in (b) zero net reactance occurs in only one point (narrow band).

These types of elements are called Foster elements. However, if an element's reactance or susceptance does not satisfy (3.4), then it is called Non-Foster element.

One of the important applications of Non-Foster is in wide band matching for ESAs [55–57]. This matching technique can overcome the limitation of gain-bandwidth³ and result in more efficiency. The idea behind the Non-Foster matching technique is straightforward: if the imaginary part of the impedance of antenna is a positive capacitor (inductor), it use a negative capacitor (inductor) with the same value to cancel out the reactive part. As it is seen from Figure 3.5a, this method provides a more wideband matching that does not suffer from limitations in the quality factor of the antenna and earlier mentioned matching circuits. However, creating a negative capacitor and/or inductor requires an active (powered) circuit.

For the sake of comparison we have shown the traditional matching technique in Figure 3.5b, too. As it is seen in that figure, in the conventional matching circuit, if the imaginary part is capacitive (inductive), it is possible to use an inductor (capacitance) to make a resonance at the desired frequency, and in this way, the impedance will be purely resistive only in that frequency, which makes it very narrow band.

Natural elements like resistors, capacitors, and inductors show the Foster behaviour, (3.4). Non-

³Based on this limitation, we cannot increase the bandwidth of antenna and its gain simultaneously.

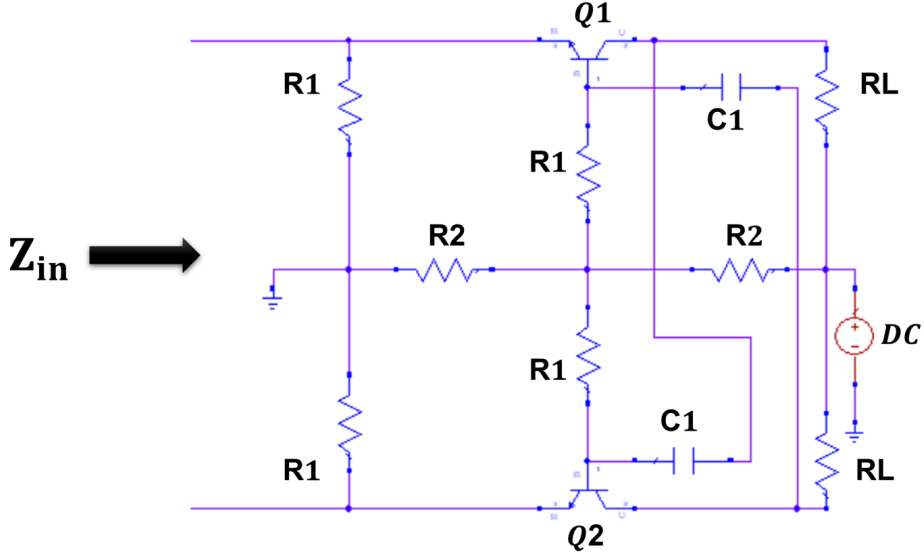


Figure 3.6: Ideal Linvill's Negative Impedance Converter [58]. Two resistors R_L are the desired positive values loads.

Foster elements are made of active circuits which consists of some transistors like Bipolar Junction Transistor (BJT), Field Effect Transistor (FET), or Operational Amplifier (Op-Amp). Using these types of circuits we can make some circuits that are called Negative Impedance Converter (NIC) [58]. The non-Foster technique can be designed to make a negative resistor, capacitor, inductor, or a combination of them. As a drawback of this active matching technique, it requires a power supply. Therefore its implementation in the industrial bins is not easy, but it can be implemented in the small grain bin, i.e. [33].

3.2.1 Negative Impedance Converter Circuits (NICs)

First NIC circuits were introduced and made of BJTs by Linvill [58], in 1953. The ideal Linvill's NIC has been shown in Figure 3.6. By using the BJT signal model and going through the math, we obtain the following impedance of NIC as $Z = -\gamma Z_{load}$, where γ is a coefficient which obtains through the BJT's signal model of Figure 3.6. In the following explanations, we first analyze three critical categories of NICs, then implement some of them in the mentioned small SHLA [33].

Analysis of NIC with BJT

To analyze the BJT NIC circuit, we use Figure (3.7), which just shows the transistor signal model without bias components. Note that in this figure source v_{in} is not part of the circuit. We have plugged it to calculate the observed impedance from source side ($z_{in} = \frac{v_{in}}{i_{in}}$) based on the flowing current through the source (i_{in}). This technique in fundamental circuit theories is called Thevenin circuit method [42]. We can deduce the following relations from this model:

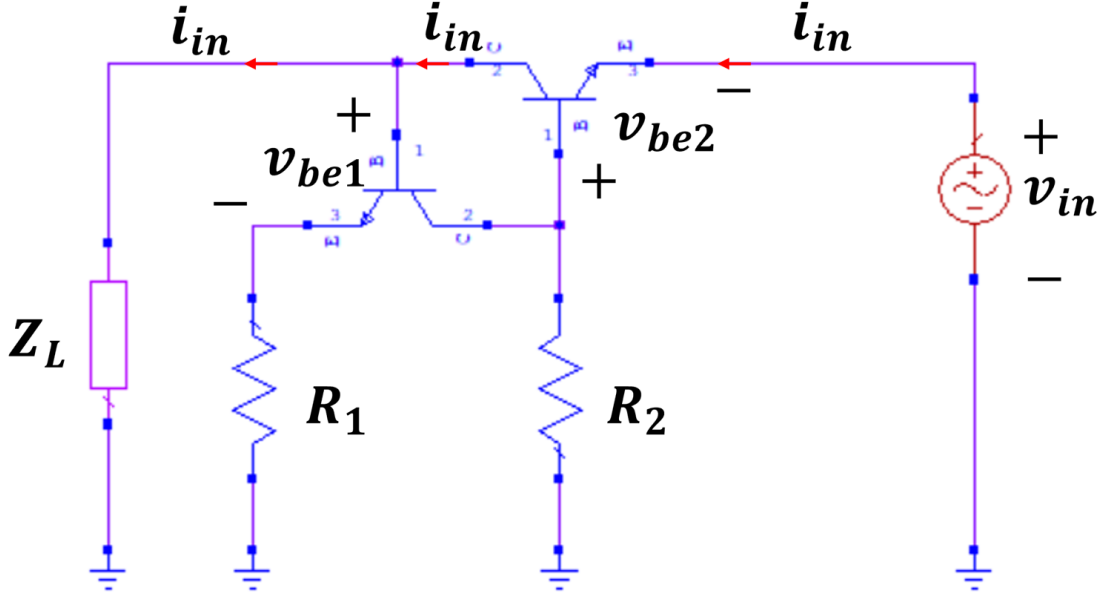


Figure 3.7: Model of analysis of NIC with BJT. Note that we use v_{in} to calculate input impedance of the circuit (z_{in}) based on its flowing current (i_{in}).

$$v_{be} = -\frac{i_{in}}{g_m} \quad (3.5a)$$

$$i_{in}Z_L + v_{be1} + g_mv_{be1}R_1 = 0 \quad (3.5b)$$

$$-g_mv_{be1}R_2 + v_{be2} + v_{in} = 0 \quad (3.5c)$$

$$z_{in} = \frac{v_{in}}{i_{in}} = -\frac{R_2}{\frac{1}{g_m} + R_1}Z_L - \frac{1}{g_m} \approx -\frac{R_2}{R_1}Z_L \quad (3.5d)$$

Where quantities in (3.5) are according to Figure 3.7. Also, note that those two transistors are similar with the same value of conductance (g_m). As it is seen from calculations of (3.5), the configuration in Figure 3.7 for big values of conductance of BJTs can produce a negative impedance with Coefficient $-\frac{R_2}{R_1}$.

Analysis of NIC with FET

To analyze the FET NIC circuit, we use Figure 3.8, which shows a NIC made of two FETs; the signal model only has been shown. Note that in this figure source v_{in} is not part of the circuit and similar to the analysis for BJT in Figure 3.7, we plug it to use Thevenin circuit model and calculate input impedance of the circuit.

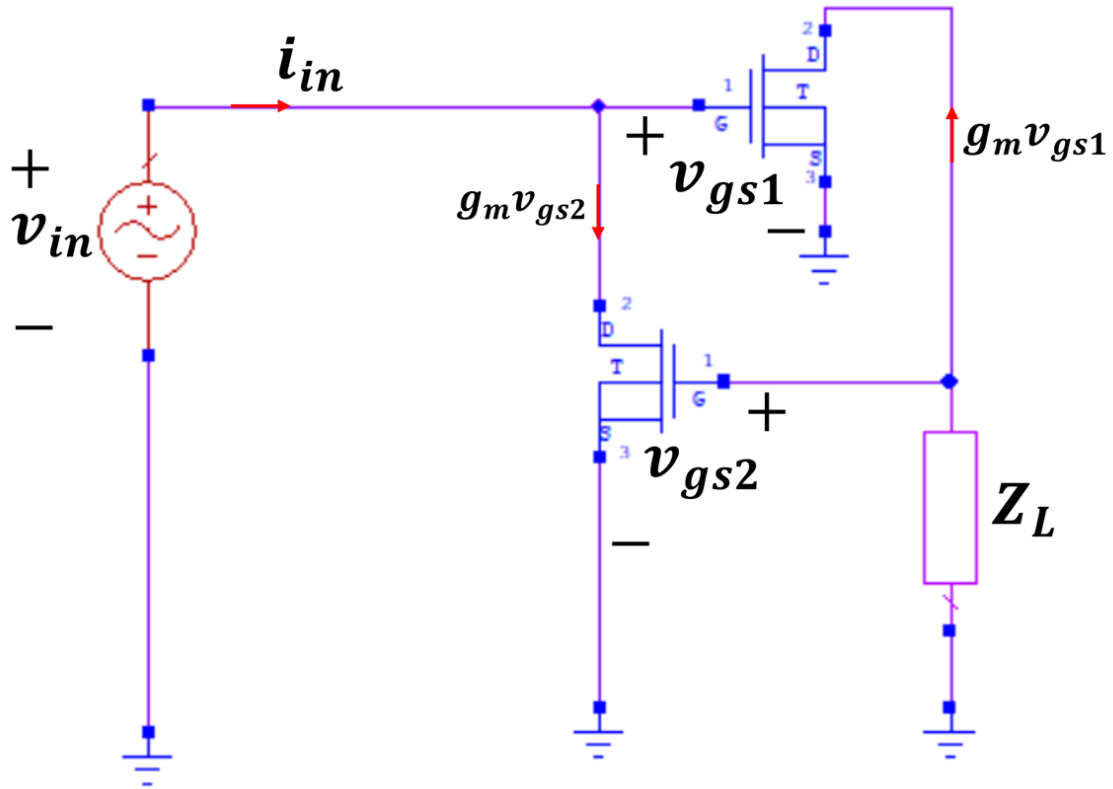


Figure 3.8: Model of analysis of NIC with FET. Note that we use v_{in} to calculate input impedance of the circuit (z_{in}) based on its flowing current (i_{in}).

Based on the model, we can derive the following relations:

$$v_{gs1} = v_{in} \quad (3.6a)$$

$$i_{in} = g_m v_{gs2} \quad (3.6b)$$

$$v_{gs2} = -g_m v_{in} Z_L \quad (3.6c)$$

$$z_{in} = \frac{v_{in}}{i_{in}} = -\frac{1}{g_m^2 Z_L} = -\frac{1}{g_m^2} Y_L \quad (3.6d)$$

Where quantities in (3.6) are according to Figure 3.8. Also, note that those two transistors are similar with the same value of conductance (g_m). As it is seen from calculations (3.6), configuration in Figure 3.8 produces a negative impedance with coefficient $-\frac{1}{g_m^2}$.

Analysis of NIC with Op-Amp

With operational amplifiers, negative impedance can be achieved in some configurations. Figure 3.9 shows one of the easy Op-Amp NIC circuits. This circuit produce a negative impedance at its input ports. Note that similar to previous NIC circuits, here v_{in} is not part of the circuit. Considering the quantities in Figure 3.9, we can find the coefficient of the NIC with Op Amp as follows:

$$v_{in} = v_1 \quad (3.7a)$$

$$i_{in} = \frac{i_{in}}{Z_L} = \frac{v_{in} - v_o}{R_1} \quad (3.7b)$$

$$v_o = \frac{v_{in}}{Z_L} (R_2 + Z_L) \quad (3.7c)$$

$$z_{in} = \frac{v_{in}}{i_{in}} = -\frac{R_1}{R_2} Z_L \quad (3.7d)$$

This circuit produce a negative impedance at its input ports with coefficient $-\frac{R_1}{R_2}$, according to (3.7). Note that the quantities in these equations are based on Figure 3.9.

Design and Simulation of Non-Foster Matching Technique

For choosing appropriate NIC configurations, we should consider desired frequency range. By recalling from RF designs [38], there are some limitations in using transistors and Op-Amps. Their bias point, input impedance, forward gain, backward gain, output impedance, stability, maximum transconductance, noise figure, power dissipation, saturation points, cut-off frequency, and so on should all be considered.

The BJT with the serial number *BFP740ESD* from Infineon Technology was selected. Their data catalog can be found on their website [59]. The schematic of the important part of the designed circuit can be found in Figure 3.10. In the above example, the values of C_2 , C_3 , L_4 , and L_6 do not reflect the true values. These values were set in simulations, and later they can be modified in the experiment. In addition, as it can be seen, this circuit is similar to the circuit in Figure 3.7.

We use ADS to simulate this circuit. It starts by choosing $L_4 = L_6 = 0$ and tuning $C_2 = C_3$ to get a good response at the desired frequency range. Simulations have been run in three different values of 25 pF, 50 pF, and 75 pF. Results in Figure 3.11 show that this type of matching is very wideband compared with the previous methods of matching.

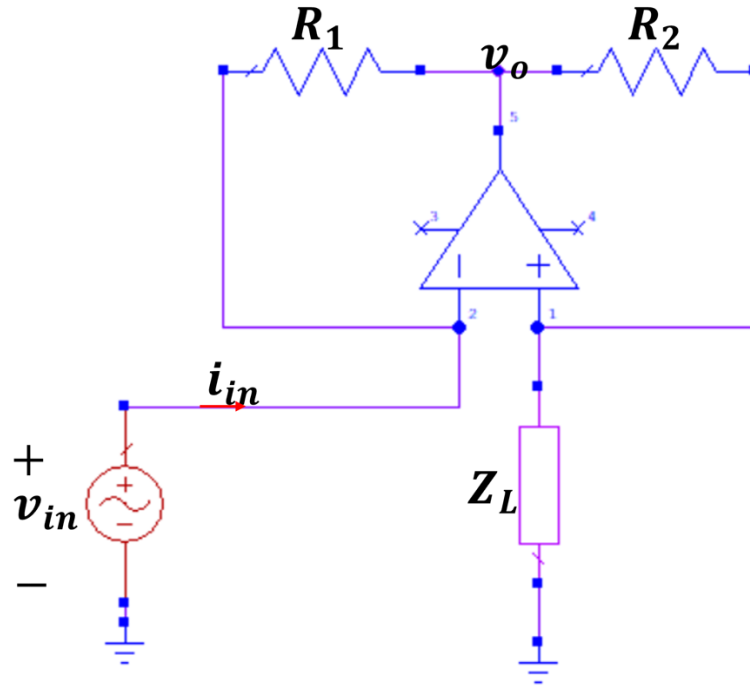


Figure 3.9: Model of analysis of NIC with op-Amp. Note that we use v_{in} to calculate input impedance of the circuit (z_{in}) based on its flowing current (i_{in}).

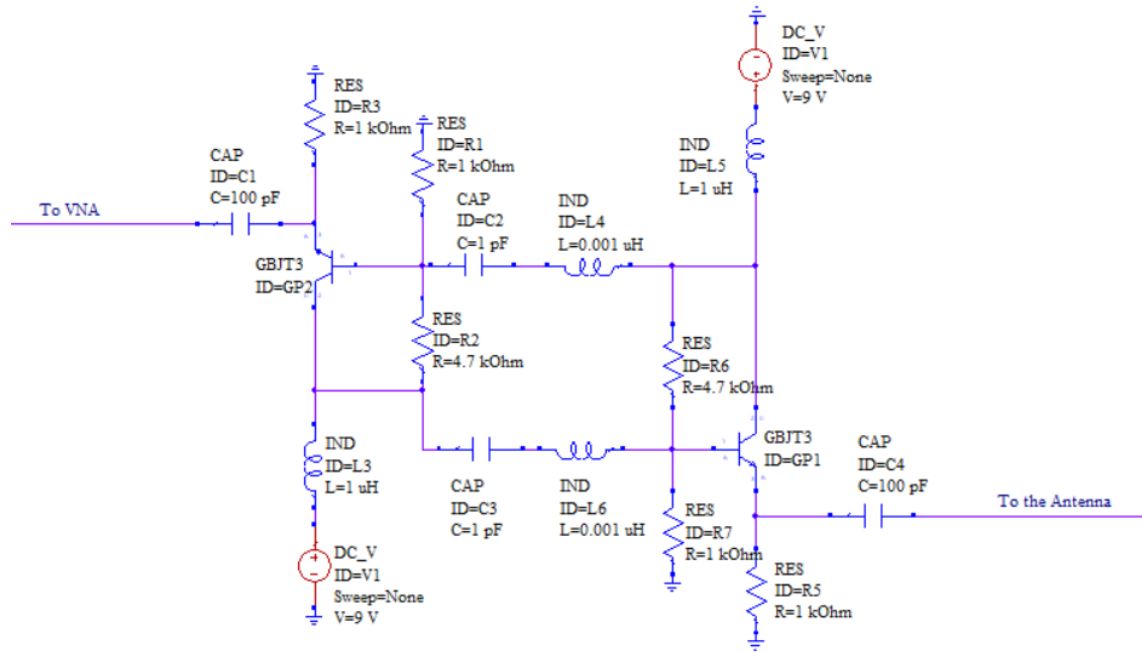


Figure 3.10: Schematic of the designed NIC with BJT.

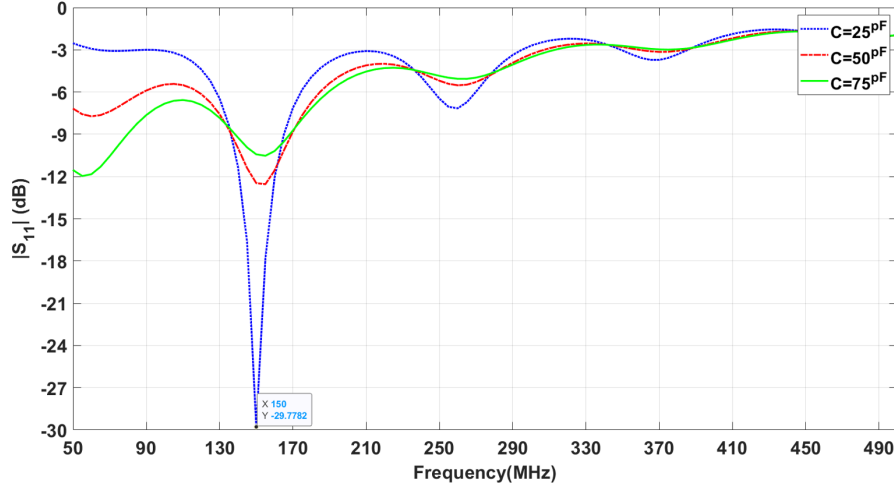


Figure 3.11: Simulation results of the designed NIC with three different values of capacitance for small SHLA, i.e. [33] (size of that antenna is 56 mm×22 mm). $L_4 = L_6 = 0$ in all cases.

3.3 Matching Technique through Loading with Ferrite Material

Here we introduce a method to improve EMI by loading existing antennas with ferrite materials. Since ferrite constitutional parameters depend on frequency of operation, therefore they need to be modeled properly, before using in the antenna structure. Therefore, first we introduce a suitable model for ferrite, then implement it in the antenna structure.

3.3.1 Ferrite Modeling

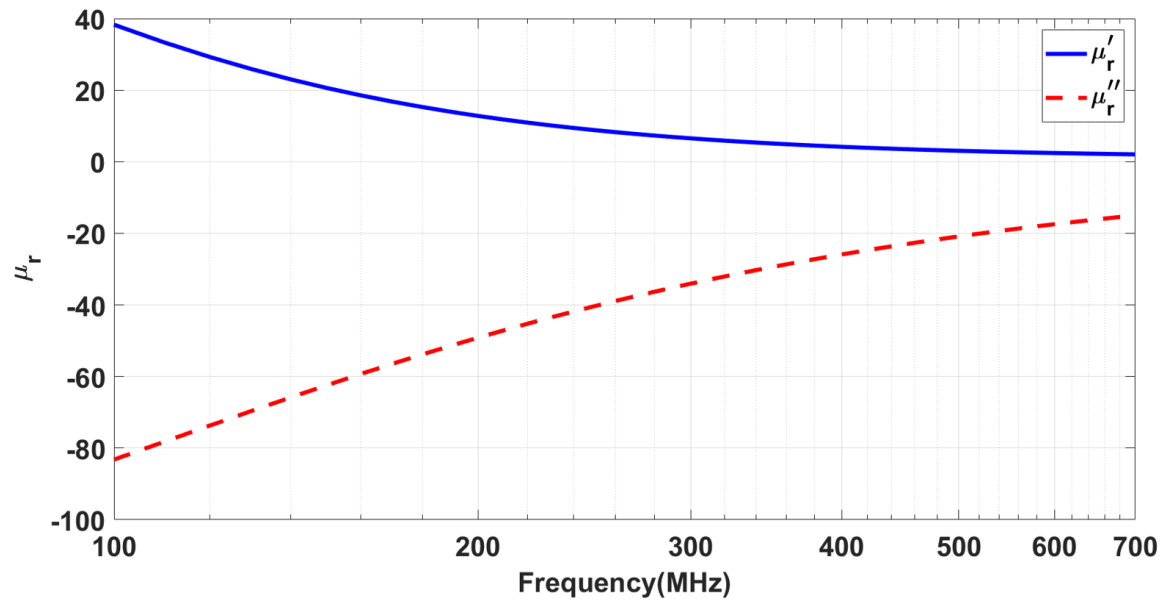
At microwave frequencies, ferrites are dispersive solid ceramics with low conductivity and relative permittivity between 10 and 15 [41] and [60]. Unlike non-magnetic media, both permeability (μ) and permittivity (ϵ) have to be considered when using ferrites in antenna design. Their dispersion can be modeled by a 2nd order Debye model, [60], for both permittivity and permeability:

$$\epsilon_r = \epsilon_\infty + \sum_{n=1}^N \frac{\epsilon_{sn} - \epsilon_\infty}{1 + j\omega\tau_n^e} \quad (3.8a)$$

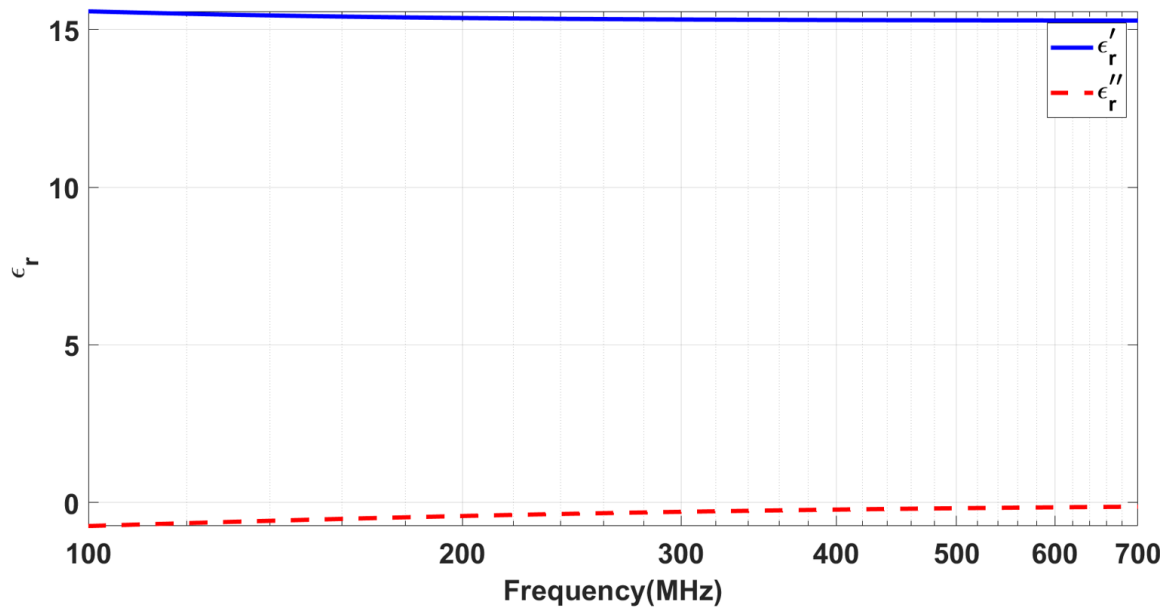
$$\mu_r = \mu_\infty + \sum_{n=1}^N \frac{\epsilon_{sn} - \epsilon_\infty}{1 + j\omega\tau_n^h} \quad (3.8b)$$

The parameters for the second-order Debye model for permittivity and permeability of the ferrite material in the antenna design are presented in Table (3.2).

Using the parameters in Table (3.2) results in permittivity and permeability over the desired maximum design frequency range are shown in Figure 3.12.



(a)



(b)

Figure 3.12: Debye model of the ferrite: (a) relative permeability and (b) relative permittivity.

ϵ_∞	15.271	μ_∞	1
ϵ_{s1}	24.835	μ_{s1}	271.227
ϵ_{s2}	16.75	μ_{s2}	104.46
ϵ_{s3}	-	μ_{s3}	-22.282
$\tau_1^e(s)$	1.042×10^{-7}	$\tau_1^h(Sec.)$	1.142×10^{-8}
$\tau_2^e(s)$	3.168×10^{-9}	$\tau_2^h(Sec.)$	2.364×10^{-9}
$\tau_3^e(s)$	-	$\tau_3^h(Sec.)$	2.532×10^{-8}
$\sigma_e(s/m)$	2.206×10^{-4}	$\sigma_h(s/m)$	0

Table 3.2: Debye's second order ferrite model's parameters [60]

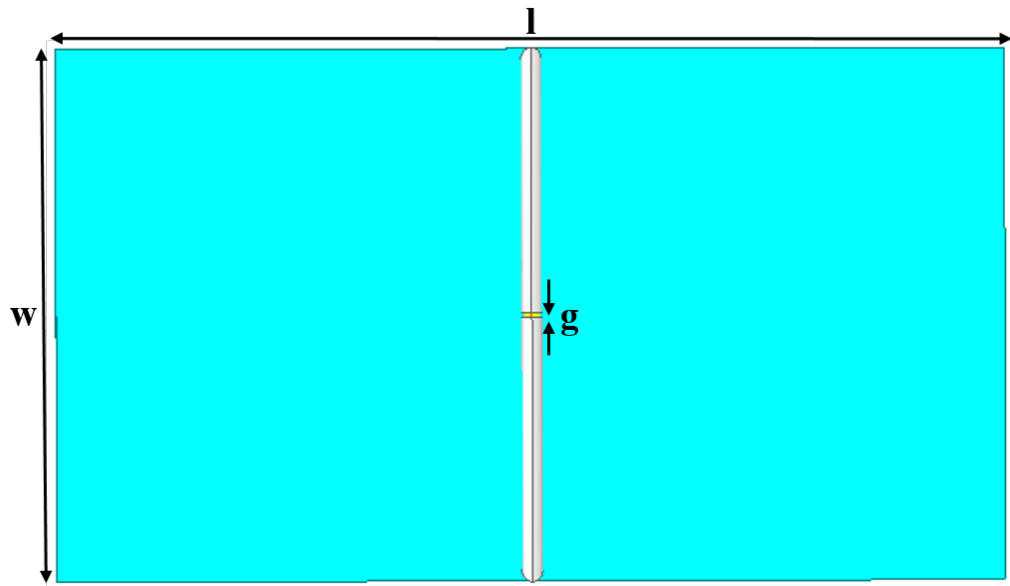
3.3.2 Small Ferrite Loaded Antenna (FLA) Simulation

Through a design process that carefully considered cost, mechanical stability (ferrite is very brittle, and grain bins are harsh environments), and ferrite availability, we attempted multiple configurations of ferrite media in the half-loop antenna. These possible configurations were simulated, but we finally settled on the design shown in Figure 3.13 (where the ferrite material is shown in aqua color). As this antenna must be mounted in a cavity containing grain, and the movement of grain can produce large forces (and rip antennas off the bin wall), we have mounted ferrite plates directly to the bin walls in order to prevent extra protrusions into the bin (the protrusions increase the forces on the antenna). With this design, the antenna remains electrically small [40] and [39], with a resonance occurring at approximately 260 MHz.

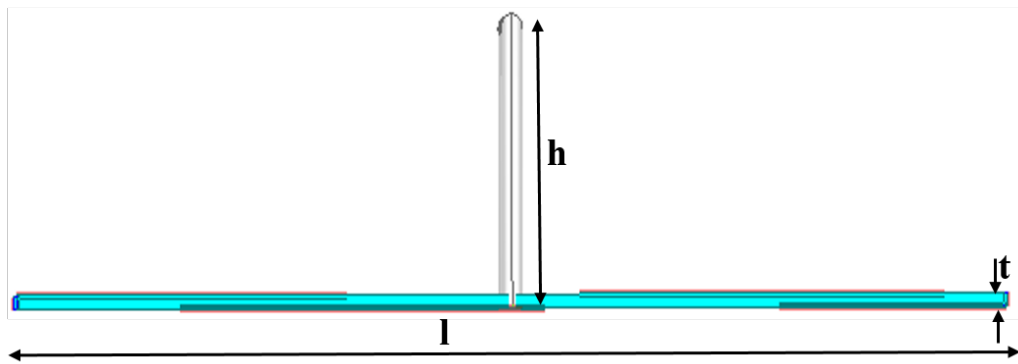
The proposed design is verified by simulating in Ansys Electromagnetic Desktop Suite (HFSS) 2021. To do so, we use a 50Ω excitation port to feed antenna. It is worth noting that the simulation model included a metallic back-plane in order to simulate the enclosure's wall instead of the full resonant metallic enclosure (to reduce the computation cost of the simulations) and a homogeneous region of the air surrounding the antenna. Removal of the bin in the simulation ignores some important effects, but is suitable for an initial analysis and provides similar S_{11} values (see later in the thesis). The plot of the reflection coefficient and input impedance of the antenna (i.e., S_{11}) from the simulation are shown in Figure 3.14. As it is seen from Figure 3.14a, antenna has a resonance at 260 MHz. Also, from (3.14b) for a reflection coefficient with criteria $S_{11} \leq -10$ dB, the design has a bandwidth of 60 MHz, centered at 260 MHz. At this frequency range, the inclusion of the ferrite has created a well-matched antenna.

3.3.3 Radiation Pattern of Small FLA

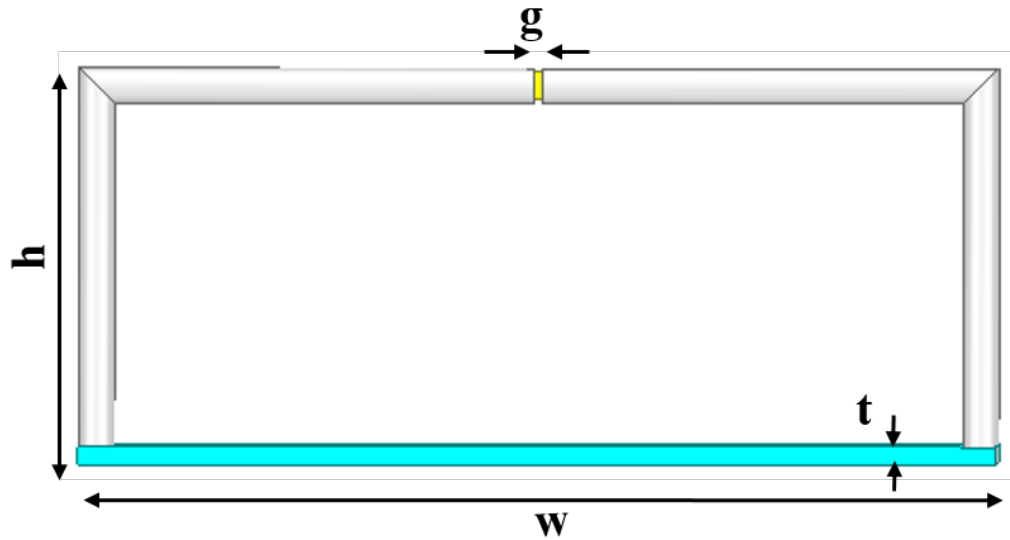
The antenna pattern has an important effect on how the antennas couple together, consequently affecting EMI results. Figure 3.15 shows the pattern of simulated FLA in HFSS. As it is seen, it has an Omni-directional pattern with a null broadside to the loop. The importance of having the omni-directional radiation pattern is that the source acts as a magnetic dipole, which is the desired criteria as mentioned in the first chapter. Besides, this pattern is as same as the pattern of the antenna in [33]. Additionally, there is no difference between the pattern of [33] and [30].



(a)

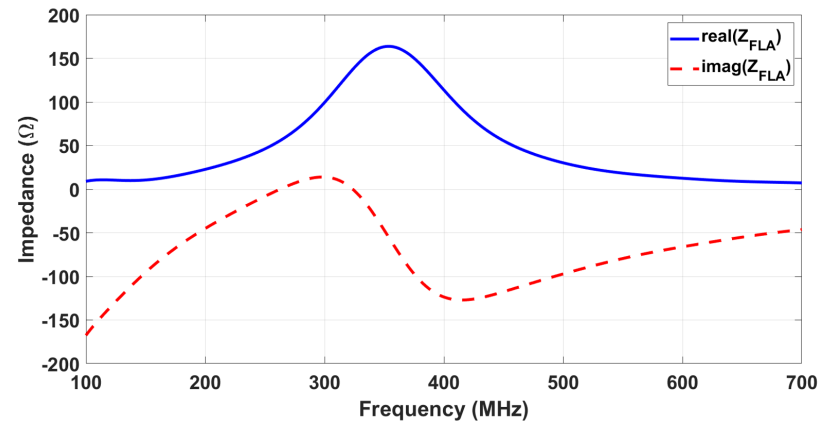


(b)

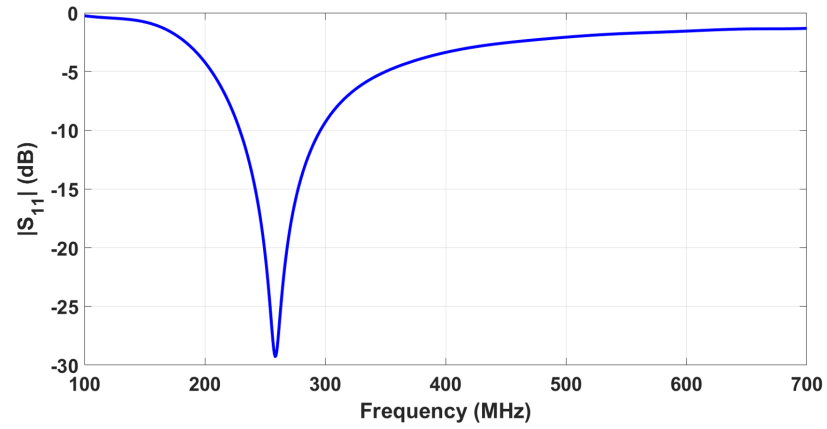


(c)

Figure 3.13: FLA structure where aqua color is ferrite and other parts are shielded coaxial cable RG405. (a) top view, (b) side view, and (c) front view. Dimensions in mm are, $l = 100$, $w = 56$, $h = 22$, $t = 1.1$, and $g = 0.5$.



(a)



(b)

Figure 3.14: Simulation results of small Ferrite Loaded Antenna (size of that antenna is 56 mm $\times$ 22 mm). (FLA) [33]: (a)input impedance and (b) reflection coefficient of the antenna.

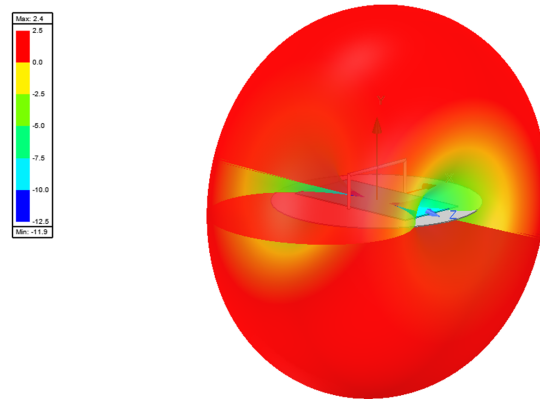


Figure 3.15: Radiation pattern of FLA

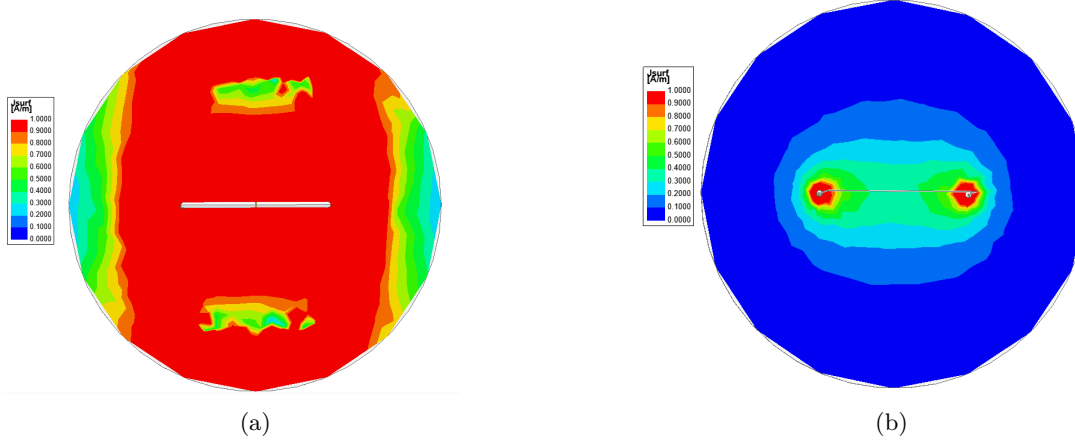


Figure 3.16: Distribution of current density at 260 MHz: (a) Ferrite Loaded Antenna (FLA) (b) None-Ferrite Loaded Antenna (NFLA).

3.3.4 Effect of ferrite on Current Distribution in Small FLA

The core idea of using ferrite inside a loop antenna is that the ferrite's high permeability creates stronger magnetic displacement currents, $\vec{M}(\vec{r}) = j\omega\mu\vec{H}(\vec{r})$, that ultimately strengthen the antenna's ability to create magnetic fields for a given input current (and vice-versa). To further emphasize this improvement, the surface current density distribution for both the FLA and the None-Ferrite Loaded Antenna (NFLA) at 260 MHz are shown in Figure 3.16. These images clearly show the increased surface current caused by the presence of the ferrite medium. A small loop antenna designed this way can be modelled by a simple magnetic dipole (i.e. a point magnetic current $\vec{M}(\vec{r})\delta(\vec{r} - \vec{r}_j)$, where j means antenna number j) [33].

3.3.5 Scaling Small FLA for Lower Frequencies

We did some simulation for the available large size SHLA [30], that is loaded with ferrite. Results show that if we want to load large SHLA with ferrite, we should decrease its size. Therefore a new antenna with new size and dimensions needs to be fabricated. Figure 3.17 shows the new FLA from the top, front, and bottom views, along with the new dimensions. Ferrite plates are outlined in aqua color. In this new structure, two ferrite plates with a thickness of 1.1 mm are used. An RG402 half-loop antenna is connected to the ground plane through two SMA connectors compatible with RG402 cable. Antenna ports can be connected to sources or loads; it is a symmetric antenna. The gap length (g) and height of the ground structure (w_g) are not critical. The metal ground is chosen to have a thickness of 1 mm to be strong enough to withstand tensions and stresses in the grain bin. Two magnet bars are used to attach this metallic part to the bin's wall in the final fabrication. Moreover, we can cover the bottom plate of the ground structure with ferrite materials; however, the antenna's performance does not improve that much, so we have used two ferrite plates to reduce the net cost of antenna fabrication.

To understand the degree of miniaturization in this newly developed FLA, the antennas that are used in large grain bins for EMI are shown in Figure 3.18. Old antennas are large, as you can see in the picture.

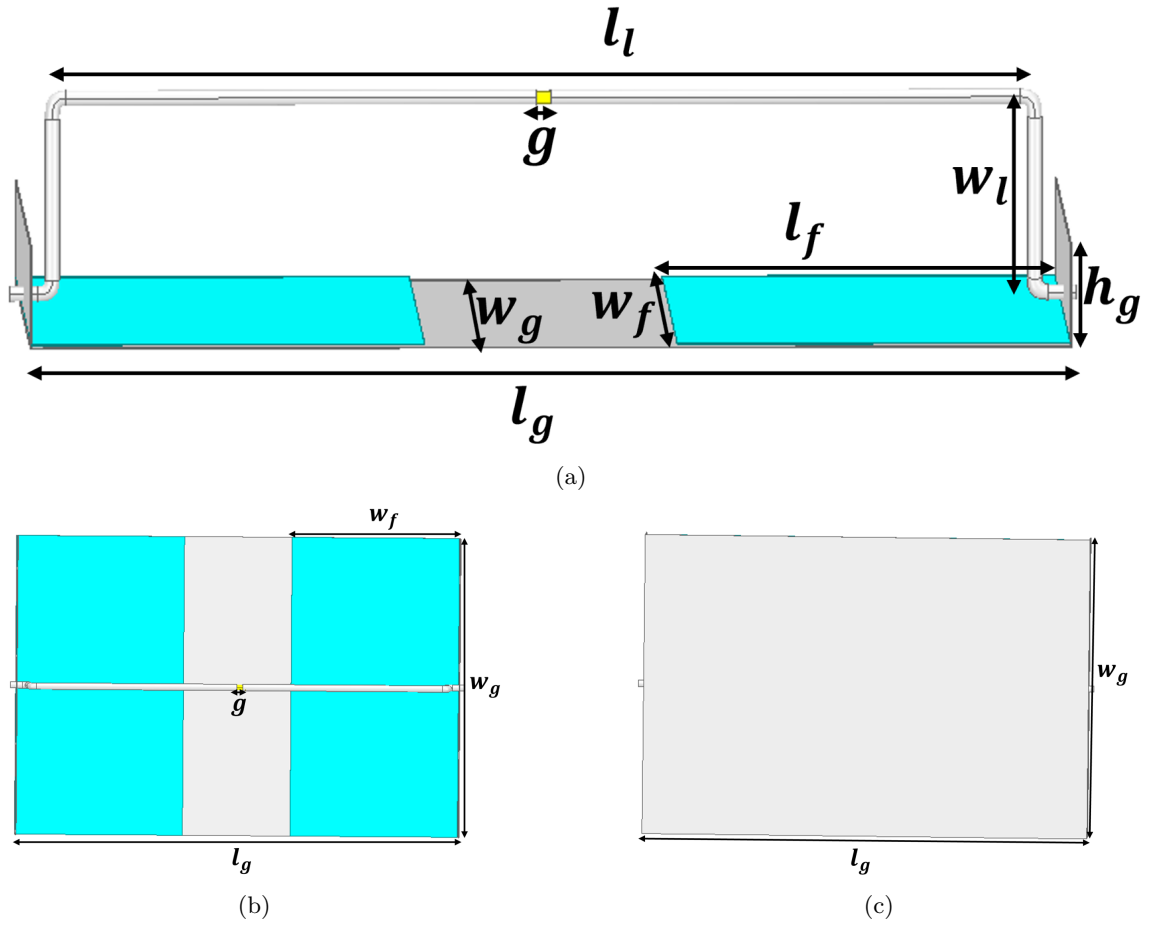


Figure 3.17: (a) front, (b) top, and (c) bottom views of new designed FLA. Aqua color areas present ferrite plates. Dimensions in mm are: $w_g = 56$; $l_g = 160$; $h_g = 10$; $w_f = 56$; $l_f = 100$; $l_l = 140$; $w_l = 30$.

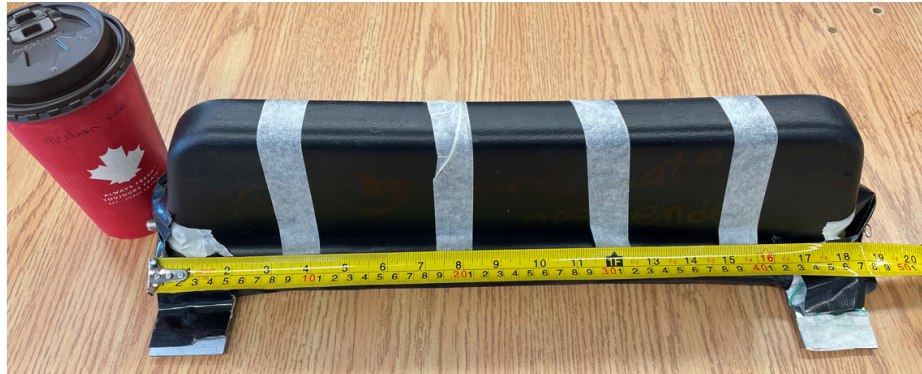


Figure 3.18: Large antennas, [30], with size 400 mm×100 mm that are used in EMI inside large metallic grain bins.

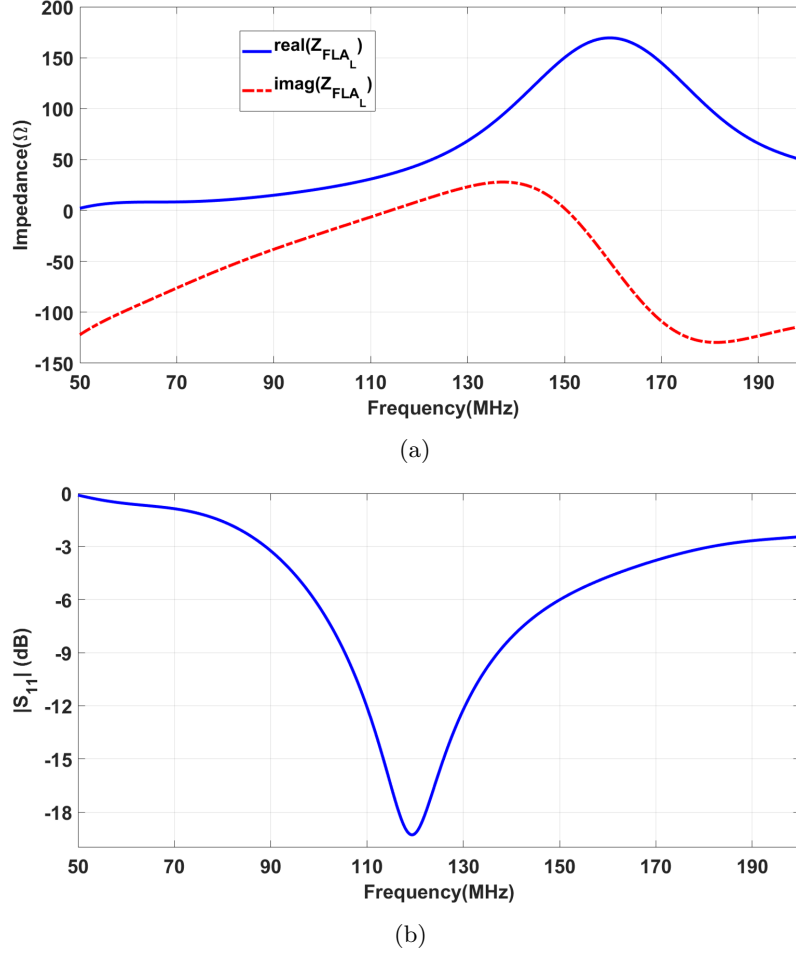


Figure 3.19: Simulation results of the scaled large Ferrite Loaded Antenna (FLA_L) [33]: (a) input impedance and (b) reflection coefficient of the antenna. Note that subscript L refers to size of the antenna which is large in this figure.

Using the mentioned dispersive model of ferrite (3.8), we simulated the new antenna in HFSS. Figure 3.19 shows the simulation results of reflection coefficient and input impedance of the scaled FLA. According to these plots, the antenna can be used at a low frequency (120 MHz). Given the wavelength at this frequency and the size of the new FLA, this new antenna is a miniaturized resonant antenna.

3.4 Conclusion

In this chapter some techniques to improve small and large SHLAs, [33] and [30] respectively, were introduced and designed. Those designs were verified in simulation software. We implement matching circuit with lumped element component for small and large SHLAs. Results show that small ferrite loaded SHLA has a good bandwidth but unfortunately it does not absorb enough signal because of the small effective surface area and size of the antenna in lower frequencies like 120 MHz. However, it is well enough to be used in higher frequencies like 260 MHz. Also, matching of the large SHLA with lumped element result into very narrow band matching in low frequencies in

the order of 100 MHz. Therefore none of these matching circuit can be a good candidate for final implementation in EMI system. We examined implementing of transmission line matching circuit as well. However, the size of the needed lines are too big to be implemented at lower frequencies.

Non-Foster matching techniques was the other introduced improvement method. We implemented it for small SHLA and simulation results shows a wide bandwidth. Besides this method increases total gain and efficiency of the antenna as well. However, active circuits that produce NIC does need a power supply. This feature makes it challenging to use Non-Foster matching in large industrial bin. We can implement it in small laboratory scale EMI grain bins.

Finally, we loaded small SHLA with ferrite material to improve its performance. Simulations in HFSS shows good improvements in antenna's performance. However, this method cannot be implemented in the large SHLA. Therefore we scaled up the size of small FLA to design a new large FLA which is suitable for lower frequencies. Then we verify the new design in HFSS and obtained promising results at 120 MHz.

Initial tests of these ferrite loaded and new scaled antennas are presented in the next chapter to make sure about their correct performance and verify obtained simulation results and designs in this chapter.

Chapter 4

Ferrite Loaded Vs. Matched Antennas

Ferrite materials provide good matching. Additionally, the size of the antenna is compact. Therefore, they are good material to be used in the structure of antennas. In this chapter, we present three important measurements to verify that ferrite loaded antennas, whether small or large sizes, can be good candidate for EMI in their corresponding bins. First we present the measurements of antennas' parameters, i.e. input impedance and reflection coefficients, after installing on the walls of bin. Then we introduce an experimental setup to find the level of improvement in signal of ferrite loaded SHLAs in comparison with their counterparts without ferrite. Finally, measurements in GTEM cell will be presented to show that ferrite will not deteriorate the SHLAs to absorb magnetic fields and reject electric fields.

Note that for simplicity, in this chapter, we use FLA for Ferrite Loaded shielded half loop Antenna (the improved version of existing antennas) and NFLA for Shielded Half Loop Antenna without ferrite (the existing version of antennas in EMI in grain bin). We have two types of NFLA: small size which is used in laboratory grain bin [33] and large size which is used in industrial bins [30]. Therefore we do the following measurements for both types of these antennas. At the end of this chapter, we will compare all presented improvement techniques, whether matching or ferrite loading, and choose the best one to be used in EMI inside grain bins.

4.1 Measuring Antenna's Parameters

Measuring antennas' parameters is simple: just install each type of antenna on the bin's wall, seal the bin¹, and use a cable that connects antenna to VNA we can measure those parameters. For example Figure 4.1a shows the small FLA installed on the small bin's wall and Figure 4.1b) shows the large FLA installed on the industrial bin's wall. In the small FLA, ferrite plate has been fixed on the walls but in large FLA it has been located inside the plastic box². Then we fill the bin with grain (here we use red hard winter wheat). The antenna is buried under the grain. After closing the

¹Bin is a barrel with diameter of 28.5 cm and height of 70 cm, 0.25λ and 0.61λ at 260 MHz, respectively.

²Note the neither plastic box in large FLA, nor the plastic frame in the small FLA does not affect the antennas' behaviour.

lid of the bin, we connect Vector Network Analyzer (VNA) to the antenna terminal and measure its impedance and reflection coefficient.

4.1.1 Small FLA's Parameters

Measurement results of reflection coefficients of small FLA's have been presented in Figure 4.2. As it is seen from Figure 4.2, measurement and simulation S_{11} are very well matched. For comparison purposes, the measured reflection coefficient for the NFLA is also graphed on the same plot; moreover, this plots verifies that the ferrite loading has lowered antenna's reflection coefficient, which is an improvement.

The differences in S_{11} are evident between the manufactured small FLA and NFLA antennas: most of the incoming energy is reflected from the small NFLA for all the design frequencies (approx. 200 MHz-600 MHz). For the small FLA, there are minor differences between simulated and measurement. The differences can be attributed to the fact that the manufactured antenna was placed inside the metallic bin (where the simulation was in free-space³), the physical antenna has imperfections in its shape, and the simulation uses Perfect Electric Conductor (PEC) while the actual antenna is made of copper. Given these differences, there is a good match between simulation and measurement.

Measurement results of input impedance of small FLA's have been presented in Figure 4.3. As it is verified from Figure 4.3, there is a good resonance at 260 MHz, i.e. reactive part of antenna's impedance is zero. Also, real part of input impedance is very close to 50 Ω .

4.1.2 Large FLA's Parameters

Design process of large FLA for EMI inside large bin⁴ has been presented in section (3.3.5). We use that designed antenna to do our measurements. Measurement results of reflection coefficients of large FLA's have been presented in Figure 4.4. As it is seen from Figure 4.4, measurement and simulation S_{11} are very well matched. For comparison purposes, the measured reflection coefficient for the large NFLA is also graphed on the same plot.

The differences in S_{11} are evident between the manufactured large FLA and NFLA antennas: most of the incoming energy is reflected from the large NFLA for all the design frequencies (approx. 80 MHz-145 MHz). For the large FLA, there are minor differences between simulated and measurement. The differences can be attributed to the same reasons for small FLA.

Measurement results of input impedance of large FLA's have been presented in Figure 4.5. As it is verified from Figure 4.5, there is a good resonance at 120 MHz, i.e. reactive part of antenna's impedance is zero.

³Since the SHLA has been already fabricated, we installed and measured its parameter performance in real world without simulating the big structure of the bin and antenna in simulator.

⁴The large bin has almost a big cylindrical shape with radius of about 5 m and height of 7 m, 2λ and 2.8λ at 120 MHz, respectively.



(a)



(b)

Figure 4.1: Installed FLA in the test grain bins: (a) small FLA in small bin and (b) large FLAs in large bin.

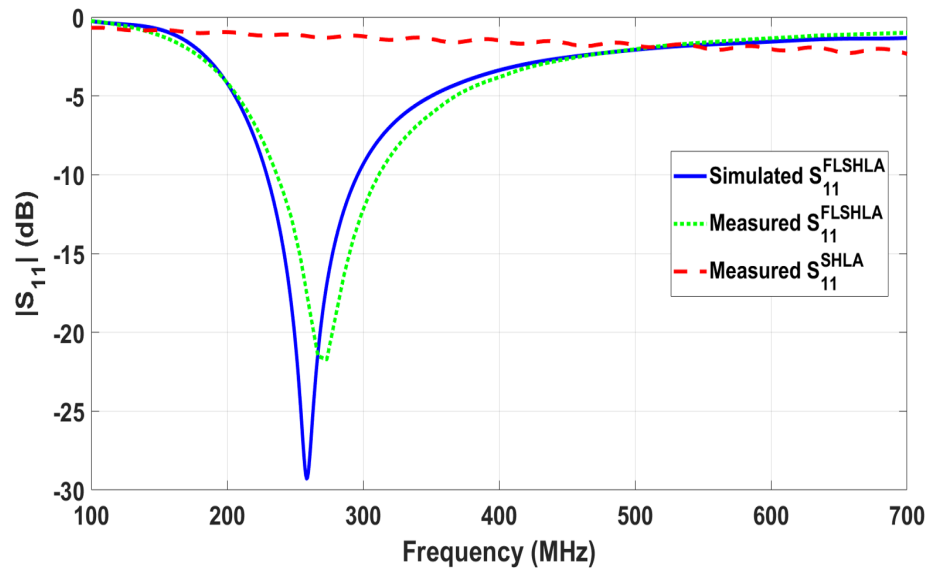


Figure 4.2: Comparison of measured and simulated reflection coefficients of small FLA with the measured reflection coefficient of small NFLA.

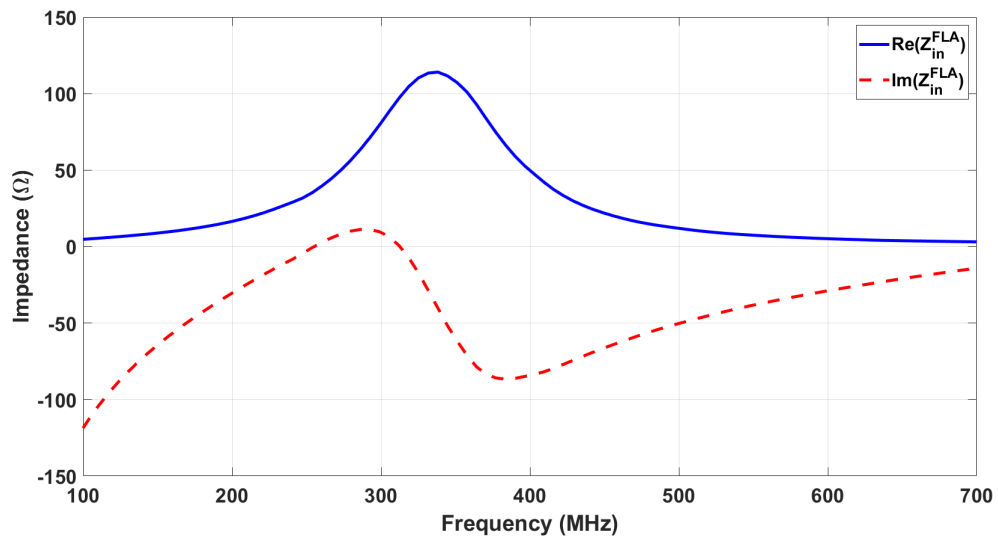


Figure 4.3: Measured impedance of small FLA.

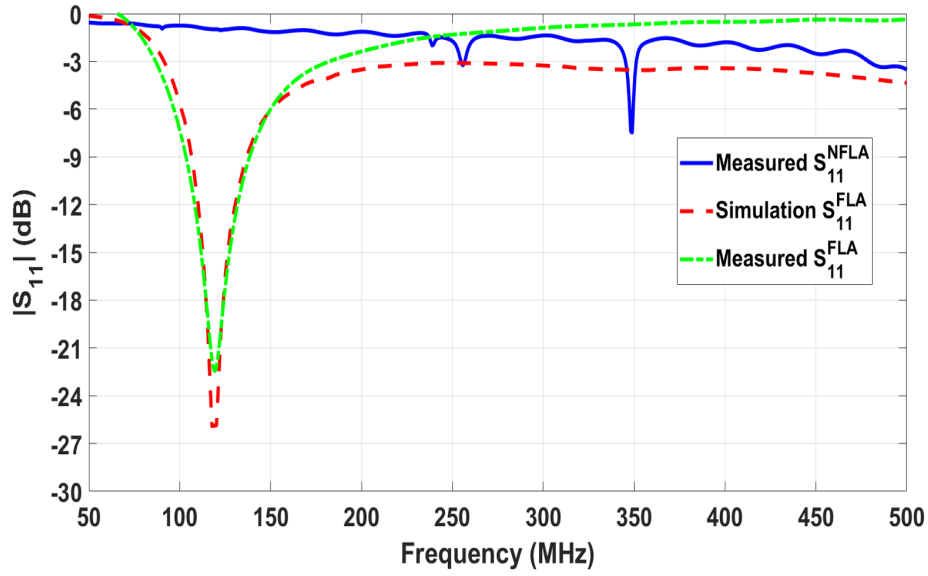


Figure 4.4: Comparison of measured and simulated reflection coefficients of large FLA with the measured reflection coefficient of large NFLA.

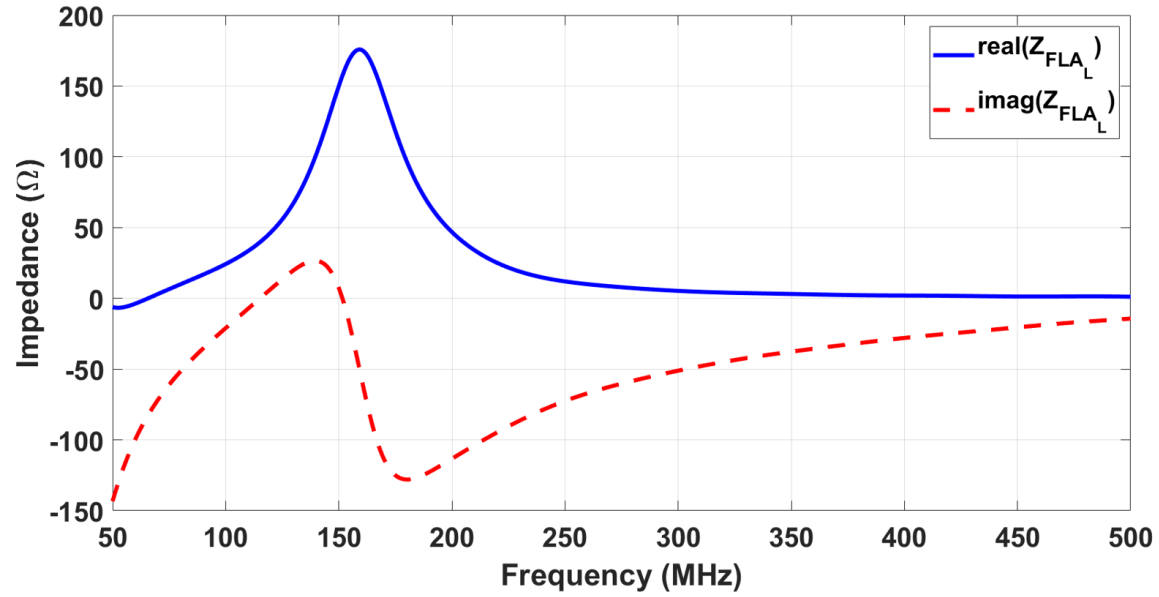


Figure 4.5: Measured impedance of large FLA.

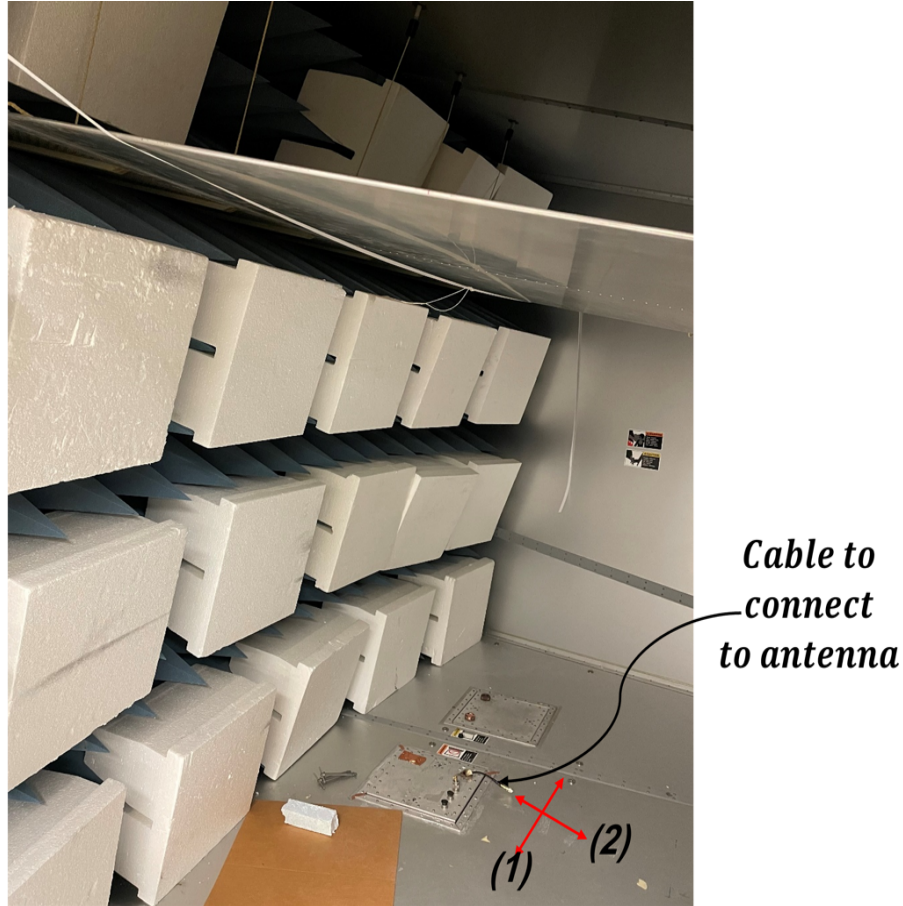


Figure 4.6: To measure sensitivity of antennas to magnetic field and rejecting electric field, we connect the antenna to a VNA through a cable and put it in two different shown orthogonal positions (position (1): parallel and (2): normal). Bottom metallic plate of GTEM cell behaves like the metallic wall of the bin, that antenna is installed on it.

4.2 Comparing Antenna's Sensitivity to Normal Magnetic Field Measurement for FLA and NFLA

We know from previous chapters that we need to have an antenna that can reject electric field and absorb normal magnetic field to the loop surface. Therefore we measure the sensitivity of each type of antenna to normal magnetic field and rejecting electric fields in GTEM cell. By putting each type of antennas in two orthogonal positions inside the GTEM cell (Figure 4.6) we measure their sensitivity to normal magnetic field and rejecting electric fields. Then the difference between these two situations is used as an indication of capabilities of each antenna in rejecting electric fields and accepting normal magnetic fields to the surface of the antenna. Based on Figure 4.6, position (2) is the normal (desired) position, which magnetic field is perpendicular to that direction, and antenna is placed parallel to it for having the most magnetic field detection. While position (1) is the parallel (undesired) position; antenna in this direction has the least sensitivity to magnetic fields.

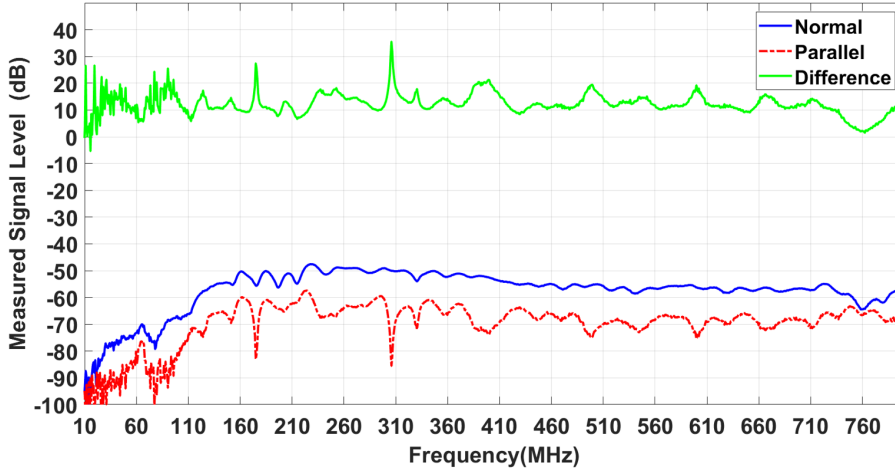


Figure 4.7: Desired (normal) and undesired (parallel) levels of received signals along with their difference for small FLA.

4.2.1 Magnetic Sensitivity of Small FLA

Results which have been presented in Figure 4.7, show that loading with ferrite material enhances the ability of the antenna in detecting magnetic mode currents. As it is seen from Figure 4.7, small FLA is twice sensitive to normal magnetic field in compare to the case without ferrite (Figure 2.11). As can be seen from the fact that there the blue line in Figure 2.11, is significantly higher than the equivalent blue line in the non-ferrite loop, shown in Figure 2.11(a).

4.2.2 Magnetic Sensitivity of Large FLA

In order to test the large FLA's ability to detect magnetic mode current, we performed the mentioned measurement inside the GTEM cell. Figure 4.8 shows this measured quantity for large FLA. As it is seen from figures (4.8) and (2.11), antennas designed with ferrite are almost as sensitive to magnetic currents as those without ferrite.

4.3 Comparing the Signal Level of FLA with the One of NFLA

The main goal of the proposed FLA antennas is to increase the signal strength - in practice, the $S_{ji}(\text{for } j \neq i)$ - above those of the existing NFLA antennas. So, in the next measurement we want to know the level of the improvement in signal level by FLA in comparison with NFLA. We examine this enhancement for small and large FLA. To do so, similar to Figure 4.9, we install one antenna on the bin's wall as the transmitter and connected it to the first port of VNA. Also, in the opposite side of that installed antenna, we installed another antenna of the same type and connected it to the second port of VNA. Then we buried them under wheat and closed the bin's lid; next we measured S_{21} . This experiment is done for small FLA and then is repeated for small NFLA. We try to keep the level of wheat, cone angle of the wheat, and every other conditions in both experiments the same as possible. Measuring the increasing of signal level for large FLA inside large bin is done inside

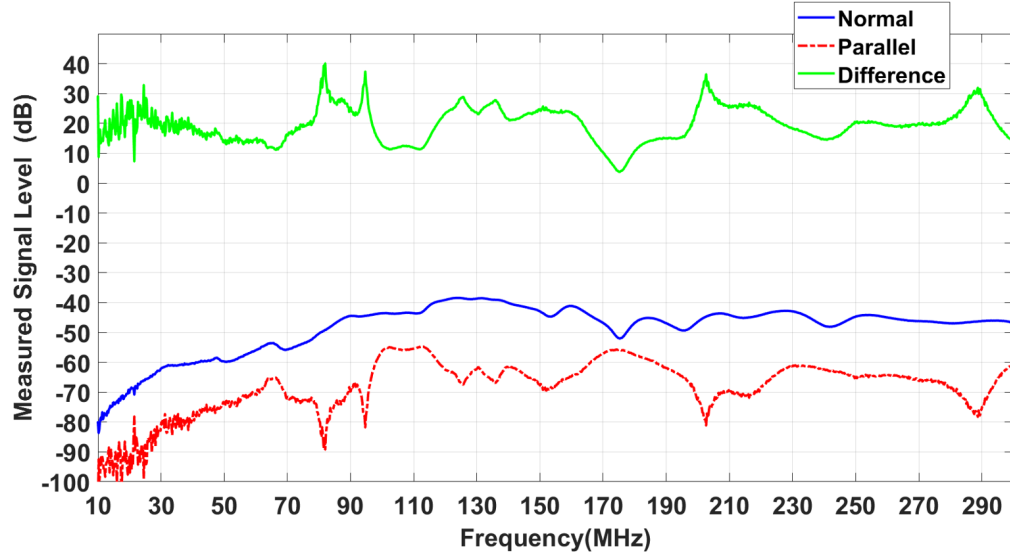


Figure 4.8: Desired (normal) and undesired (parallel) levels of received signals along with their difference for large FLA.

GTEM cell for some reasons that will be explained in next section.

4.3.1 Improvement Levels of Small FLA

To show the higher S_{21} signal levels with the small FLA, we measured S_{21}^{NFLA} for two NFLA (original) antennas with the test bin partially full. We next used FLA and re-took the measurement, resulting in S_{21}^{FLA} .

Figure 4.10 shows the $|S_{21}|$ comparisons of FLA with NFLA over the designed frequency ranges. This plot shows that the small FLA receives up to 19 dB more S_{21} signal than small NFLAs, with the strongest signal improvement occurring at the resonance frequency of the antenna.

4.3.2 Improvement Levels of Large FLA

Measuring the level of signal improvement using the scenario of small bin is not easy to be used in the big bin⁵. Therefore, we used GTEM cell to measure the level of signal improvement. Measurement steps are as follows:

- First we put the large FLA (or NFLA) in the desired position (normal), like the one that has been shown in Figure 4.6.
- We run a measurement and measure S_{21} , where antenna has been connected to the second port of VNA and the transmitter of GTEM cell is connected to the first port of VNA..

After having those measurements for S_{21} of large FLA and NFLA, we can compare the level of improvement in the signal level in the GTEM cell⁶. This comparison and improvement in the signal

⁵Filling and emptying the industrial grain bins needs special protocols and rules to be followed. It is often takes too time and is not safe to do alone.

⁶Note that GTEM cell is filled with air, therefore this comparison gives us a good idea about the level of improvement in FLA compare to NFLA. However, in real grain bin, grain has loss and improvement level may change into another value.

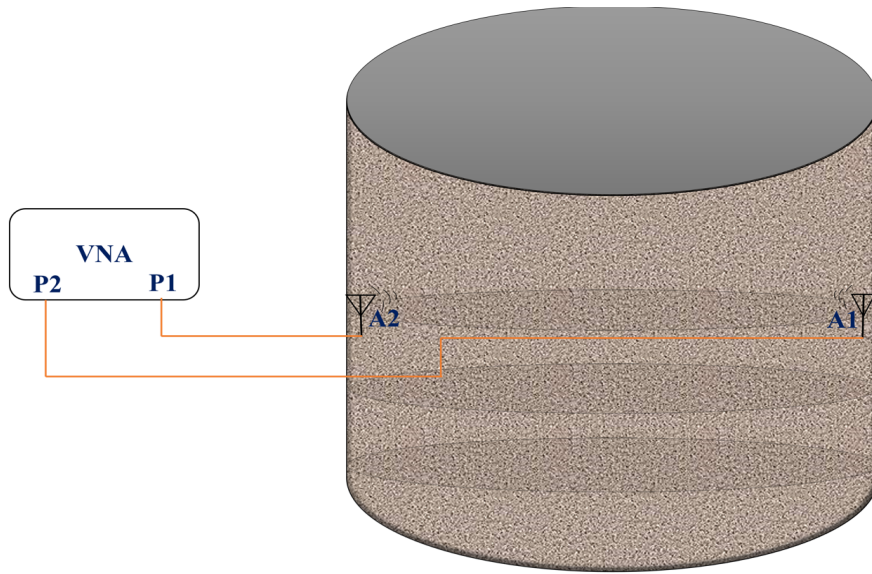


Figure 4.9: To measure received power by an antenna from a same type of antenna as a transmitter, after installing them on bin's wall and burring them under wheat, we connect each antenna to VNA (port1: P1, port2: P2) through low loss cables to measure S_{21} (or equivalently S_{12}).

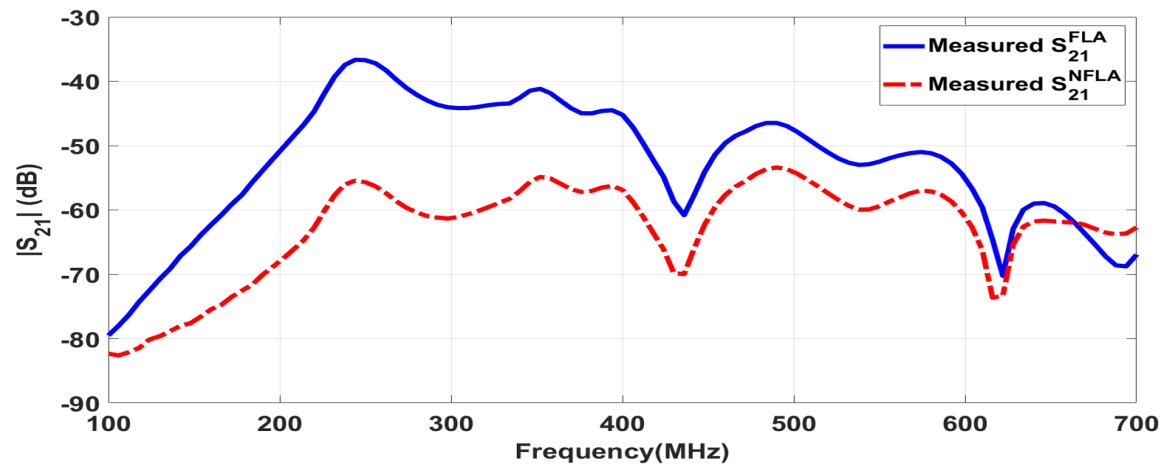


Figure 4.10: Comparison of small FLA and NFLA performance in the scaled grain bin.

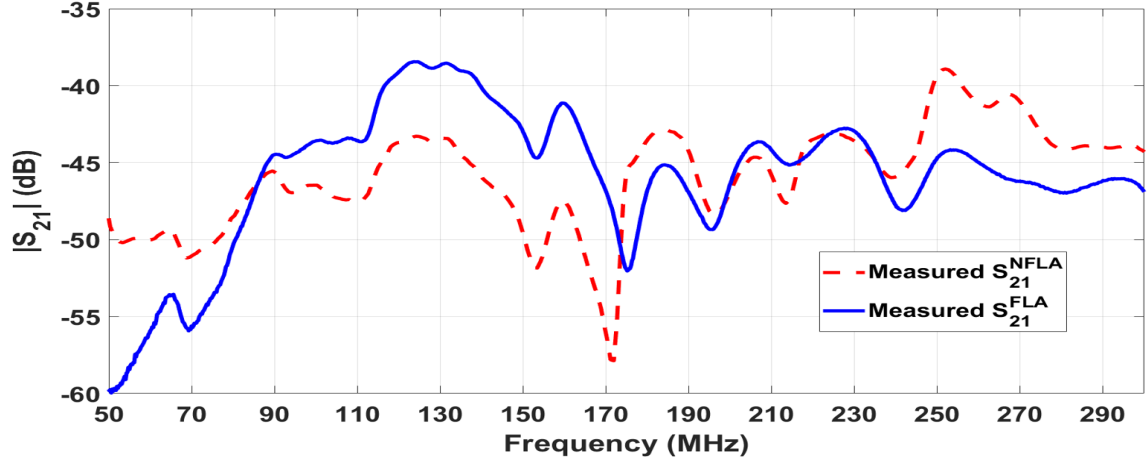


Figure 4.11: Comparison of signal level of large FLA and NFLA performance in the GTEM cell.

level for large FLA and NFLA has been shown in Figure 4.11. As it is seen from this figure, FLA has almost 5 dB more signal than NFLA, in the desired frequency (120 MHz).

4.4 Comparing Matching and Ferrite Loading Methods

We overveiwed and simulated some matching circuits to enhance antenna performance in EMI (chapter 3). The table (4.4) compares those different explained matching methods. We compare different methods from high to low priority based on: bandwidth, DC power supply requirements, size, weight, and easiness in fabrication. According to chapter 3, a lumped element or transmission line matching technique has a poor bandwidth. Even though non-foster circuits are wideband, they require a DC power supply, making them unaffordable in many cases, such as industrial silos' EMI.

In the current chapter, we investigated, and fabricated small and large FLA. Test results showed that ferrite loading would result into antenna enhancement.

In the table (4.2), we highlight small/ large FLA's cons and pros. From high to low priority, we consider its sensitivity to the normal magnetic field, radiation pattern, radiation efficiency, bandwidth, size, weight, and easiness in fabrication. Remember that ferrite is very lossy at the frequency used in this study. For this reason, FLAs are not expected to be highly efficient. However, efficiency was not a critical factor in our experiments. Besides, efficiency and SNR, which is another critical parameter in EMI, for FLA are better than NFLA. The fabrication process of FLA is easy and can be done with primary laboratory tools.

After comparing tables 4.4 and 4.2, we can choose among those techniques. We chose the best method for this study: FLA. In the next chapter we will use small FLA to do EMI inside a lab scale grain bin and represent its results. The large FLA, later, can be used in big industrial grain bin.

4.5 Conclusion

We fabricated the simulated improved ferrite loaded antenna (small FLA) and scaled ferrite loaded antenna (large FLA) from last chapter. Measurements of parameters of those FLA verified its correct performance. Lowering the reflection coefficients at the input of FLA makes it more radiative antenna

Advantages and Disadvantages of stated matching techniques in chapter 3.		
Matching Technique	Good (✓)	Bad (✗)
1. Lumped element matching		
• BW	-	✗
• DC power supply	✓	-
• Dimensions	✓	-
• Weight	✓	-
• Easy fabrication	✓	-
2. Transmission line matching		
• BW	-	✗
• DC power supply	✓	-
• Dimensions	✓	-
• Weight	✓	-
• Easy fabrication	✓	-
3. Non-foster matching		
• BW	✓	-
• DC power supply	-	✗
• Dimensions	✓	-
• Weight	✓	-
• Easy fabrication	-	✗

Table 4.1: Comparison of different matching techniques which were introduced in chapter 3.

Advantages and Disadvantages of small and large FLA.		
Design	Good (✓)	Bad (✗)
Small/ large FLA		
• magnetic current mode sensitivity	✓	-
• Radiation pattern	✓	-
• Radiation efficiency	-	✗
• BW	✓	-
• Dimensions	✓	-
• Weight	✓	-
• Easy fabrication	✓	-

Table 4.2: Pros and cons of loading antennas with ferrite materials.

in different modes: transmitter/receiver modes. Additionally, its input impedance is almost $50\ \Omega$, which result into delivering most of the absorbed power by the antenna to the connected load/circuit to its terminals.

Besides to check how those antennas well reject electric fields and absorb magnetic fields, we performed some tests in GTEM cell. Results of those experiments show that FLA can reject electric fields even in some frequencies better than NFLA. Therefore, it is assured that received signal by these types of antennas is proportional to just one component of EM fields, i.e. normal magnetic component of magnetic field, which is an important criteria for input signals to FEM-CSI algorithm.

Next, we performed another test to measure increasing in the signal levels and we showed that small and large FLA can increase the amount of received signal at least 15 dB and 5 dB, respectively, in comparison with their non-ferrite counterparts.

Finally, we compared different introduced methods in chapter 3, i.e. matching techniques, and ferrite loading technique which were used in this chapter, and chose to use ferrite loaded antennas because of their better performance and features according to tables 4.4 and 4.2. In the following chapter we implement small FLA in real EMI system in the laboratory scale.

Chapter 5

Electromagnetic Imaging Results

Imaging results of FLA¹ inside a laboratory scale grain bin are presented in this chapter. FLAs must ultimately be able to provide useful images inside the resonant enclosure. Several experiments have been performed in the small-scale grain bin with FLA and NFLA to verify that these design parameters are met. They are described below.

5.1 Imaging Steps

The EMI inside the grain bin involves several steps that are outlined here.

5.1.1 Small scaled Grain Bin Measurement Setup

The scaled grain bin has been depicted in Figure(5.1). The bin has a depth of approximately 70 cm and a radius of 28.5 cm, respectively 0.25λ and 0.6λ at 260 MHz. The total volume of the bin is approximately 190 liters. Previous results with this scaled grain bin have been presented in [33] and [32]. We installed either 24 FLA or NFLA antennas, as shown in Figure 5.1. Each antenna was connected to the hardware imaging system that contains a Vector Network Analyzer (VNA), amplifiers, control computers, and a 24×2 port switch. Antennas were connected to the GraivViz system through 1.5 m long coaxial cables. They were staggered (with a 15° shift per layer) in six layers. This arrangement of antennas was chosen to maximize the spread of antennas on the inner wall of the enclosure and ensure proper field sampling.

5.1.2 Grain Description

We used two treatments of hard red spring wheat as our test grain. The bulk of the wheat was very dry (7.5% moisture content, wet basis). This grain was harvested in approximately 2018 and stored in a stable indoor environment. Thus it is assumed to be homogeneous. From the available literature, the approximate relative permittivity of the hard red spring wheat at ≈ 260 MHz is $\epsilon_r = 3 - j0.3$, [61]. It is worth noting that this permittivity is approximate as the density of the

¹In this chapter FLA or NFLA purely mean small FLA or small NFLA respectively, unless otherwise has been mentioned clearly.

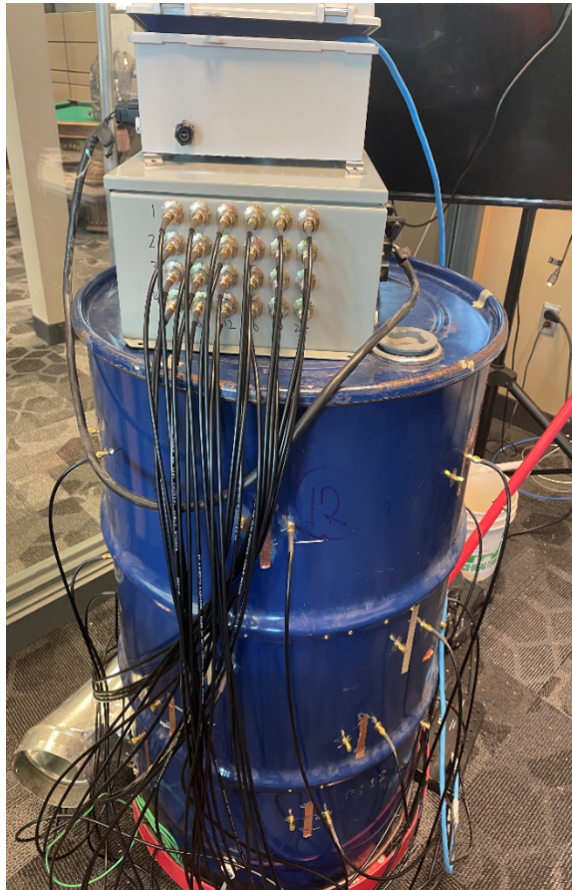
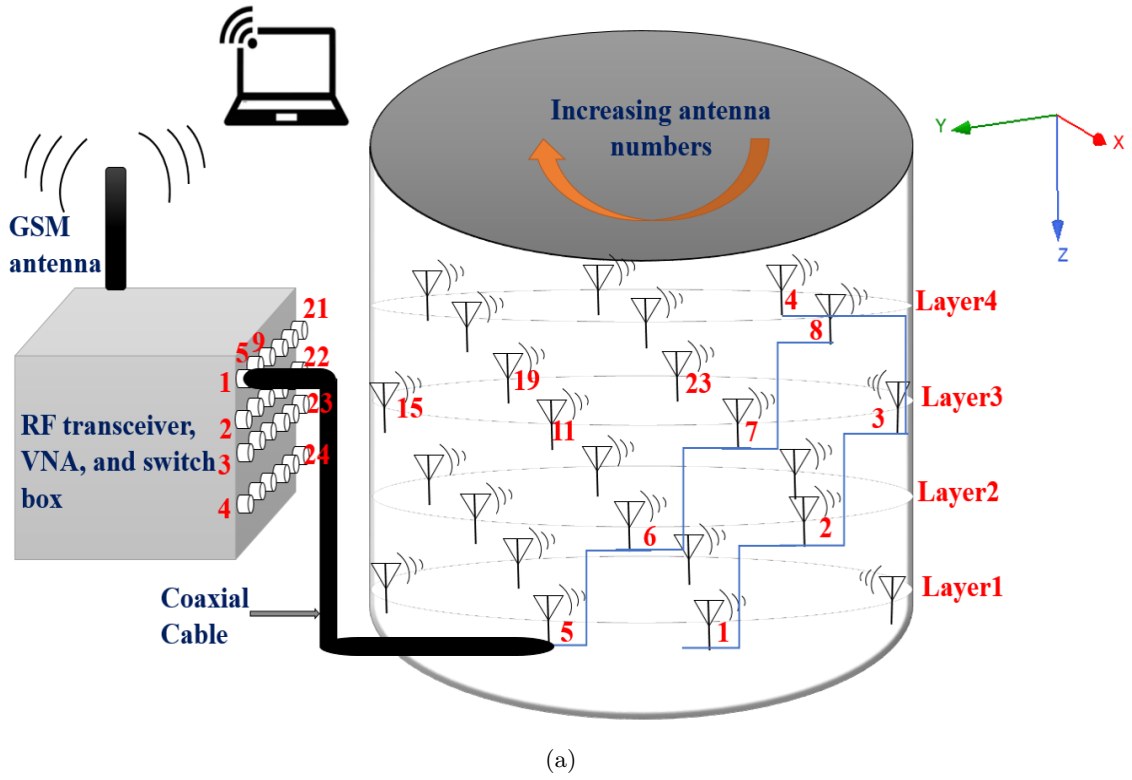


Figure 5.1: (a) block diagram and (b) real laboratory system used in microwave cavity imaging



Figure 5.2: Target close to FLA 15 while it is in the bucket before covering with wheat.

grain can have significant effects on the permittivity (we did not directly measure the permittivity of the grain)².

We next created a wet-grain target by placing 1410 grams of dry wheat mixed with 210 grams of water inside two plastic bags. A photo of the wet-grain target in position is shown in Figure 5.2.

This target was allowed to sit at standard refrigeration temperatures for at least one week to allow the water to be absorbed by the grain. Assuming all of the water was absorbed over this time, the wet-grain target had a moisture content of 19.5% and would have an approximate permittivity of $\epsilon_r = 4.5 - j0.6$ at ≈ 260 MHz, [61]. At room temperatures, such high-moisture wheat could only be stored for a couple of days without spoilage [62]. While still malleable (and thus not purely in the shape of the rectangular cuboid), the approximate dimensions of the wet-grain target were 7.5 cm $\times$ 11 cm $\times$ 21 cm; these values at 260 MHz are 0.06λ , 0.09λ , and 0.18λ , respectively.

Next, we removed the dry grain inside of the bucket and placed the wet-grain target inside (see Figure 5.2). After re-filling the bucket back to level with a dry grain (not pictured), we then took a total S -parameter measurement, which has been labeled S^{tot} .

Finally, note that we repeated each of the steps above for both the proposed FLA and NFLA.

5.1.3 Inversion Process

Standard inversion work-flow for imaging inside of grain bins has been previously outlined in [30] [32] and [33] and is as follows:

1. Obtain $\epsilon_b(r)$: Using raw S^{inc} (the measurements without the wet grain target), estimate the background permittivity, $\epsilon_b(r)$ through phaseless parametric inversion [30]. This algorithm outputs the grain height, cone angle, and average permittivity. The background estimation algorithm uses the magnitude data and a gradient-free optimization to find the best average permittivity, grain height, and cone-angle.

²Temperature also affects the permittivity of grain, but [61] was at 25° C. Our grain was approximately at room temperature ($\approx 21^\circ$ C), which has relatively minor effects on permittivity.

2. Generate Incident Fields: Using the inhomogeneous air/grin background permittivity, $\epsilon_b(r)$, as the input to a FEM forward solver, the incident magnetic for the numerical model, called H^{bkg} , is obtained.
3. Calibrate Scattered Fields: We generate calibrated scattered fields from the incident magnetic fields, and the total/incident S -parameter measurements via the equation:

$$H_{t_x, r_x}^{sct} = \frac{H_{t_x, r_x}^{bkg}}{S_{t_x, r_x}^{inc}} (S_{t_x, r_x}^{tot} - S_{t_x, r_x}^{inc}) \quad (5.1)$$

where t_x, r_x denote the transmitter and the receiver index.

4. Full Inversion: Finally, we pass the calibrated scattered fields from (5.1), along with the background permittivity $\epsilon_b(\vec{r})$, to the FEM-CSI algorithm described in chapter 1.

For this work, We choose to run the FEM-CSI independently at 10 frequencies in two ranges: 240, 244, 250, 260, 270, 280 MHz, and 540, 550, 560, 570 MHz. Frequencies as the ≈ 250 MHz group was close to the FLA resonance, while the higher frequencies are chosen as this is where the NFLAs have been previously used successfully [32], and [33].

5.1.4 Parametric Inversion Results

The objective function for the parametric inversion function, (1.15), is L_2 norm based on the magnitudes of the data. As a way of determining how successful the parametric inversion was at computing the correct result, we use the Pearson correlation coefficient between the magnitude of the raw data, $|S^{inc}|$ and the fields generated from the computational model $|H_{bkg}|$.

The correlation coefficient is insensitive to scaling differences between the fields and the measured S -parameters and can be used as an intuitive metric to compare how well the computational model fits the data: 100% correlation is a perfect match between the real world and the model. Any deviation from 100% means the computational model is not a perfect fit. A low correlation coefficient for a given antenna would mean that it is not a good fit for the computational model (e.g., if the antenna did not act like a magnetic dipole, the correlation coefficient would be low).

For S^{inc} data sets taken with both FLAs and NFLAs, Figure 5.3 displays the correlation coefficients at each frequency between $|S^{inc}|$ and $|H_{bkg}|$. In every case, the correlation coefficient for FLAs is higher than for the NFLAs. The FLA correlations are also higher near the lower, resonant frequency band. A correlation coefficient of 1 would indicate an ideal computational model of bin.

5.1.5 Height, Angle, and Background Permittivity Detection

The height, cone angle, and real permittivity estimated by the parametric inversion for both FLA and NFLA S^{inc} data sets is shown in figures (5.4), (5.5), and (5.6). The heights are within 3 cm of the expected result, with errors both above and below the expected values. Note that 1 cm at 250 MHz is approximately $\frac{1}{70}^{th}$ of a wavelength in the grain, or $\frac{1}{30}^{th}$ of wavelength at 550 MHz (in grain). Thus, in terms of wavelength these errors are small. Further, the discretization of the mesh is done with tetrahedral-side lengths being approximately $\frac{1}{10}^{th}$ of a wavelength at 550 MHz (about 3 cm). Thus, the observed errors in grain height are on the order of the discretization errors of the mesh.

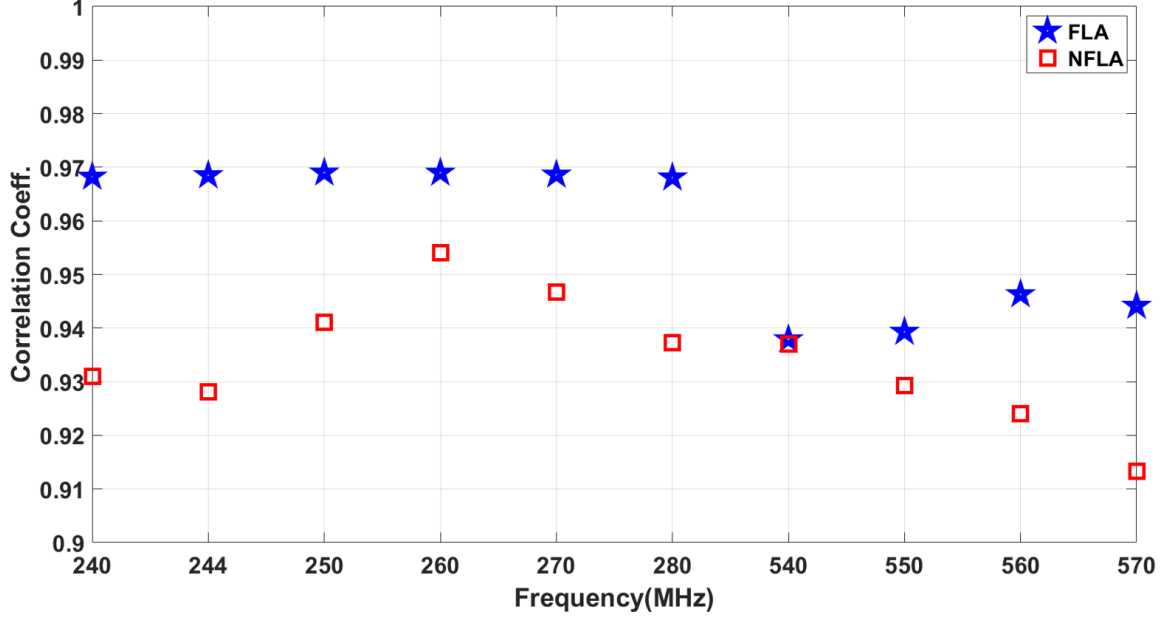


Figure 5.3: Correlation coefficients between $|S^{inc}|$ and $|H_{bkg}|$ for FLA and NFLA. Data sets have been taken with no target in the bin. Note that the x -axis (frequency) division is not uniform.

Errors appear larger in the angle of repose of the grain, shown in Figure 5.5. We have assumed the true repose angle is 0 degree, but the repeated positive repose angle for both data sets likely means that our manual flattening process was leaving a small peak of grain. These errors are larger than the discretization errors inherent in the super-position of the cone-angle onto our unstructured mesh. For example, considering the FLA data set, the average angle is 10 degrees (rounded to the nearest degree), this represents a peak height of 5 cm in the center of the bin (or about 1.5 tetrahedral edge lengths). Thus, unlike the height error, the error in angle of repose is slightly larger than the discretization error.

The values for the real permittivity at the chosen frequencies have been shown in Figure 5.6. The real permittivity is ultimately used to detect the moisture of grain [61], and the imaginary permittivity is much more variable on grain variety/density, etc. [61]. Therefore, we present only the real permittivity results (although the imaginary part of the permittivity is reconstructed in the inversion algorithm). While many factors (most importantly density of the wheat) can affect the permittivity of the wheat, the obtained values are compared to the literature values. Using the permittivity values in [61] for wheat at 7.5% at 25°C, we have added the ‘Nelson’s’ model points to Figure 5.6. It is worth noting that [61] does not have data above 300 MHz. Comparing $Re(\epsilon_b)$ in plots of Figure 5.6, the results from the FLAs follow the expected value in our lower frequency band, while at the upper frequency band they are very similar. On the whole, the FLAs are likely to provide a more accurate estimate of real permittivity than the NFLAs.

5.2 Full 3D Imaging results

We followed the outlined imaging procedure, and collected the total-field data S^{tot} with the wet-grain target buried in the grain mass for both FLA and NFLA. Then the data as per (5.1), is

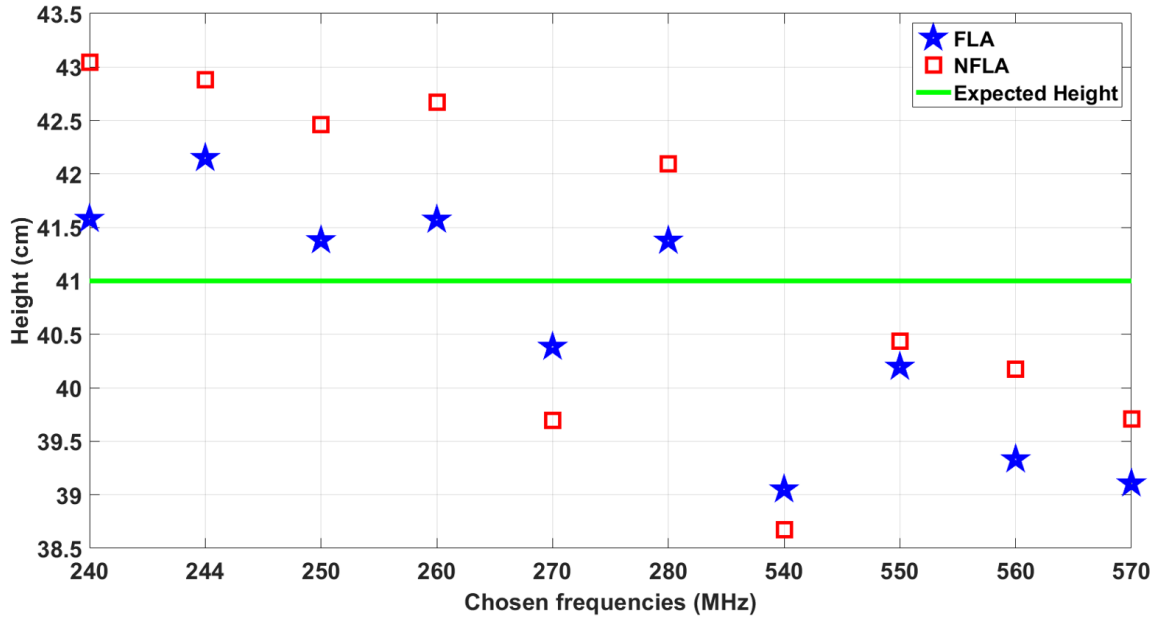


Figure 5.4: Comparing height estimation at different frequencies for FLA and NFLA. Note that the x -axis (frequency) division is not uniform.

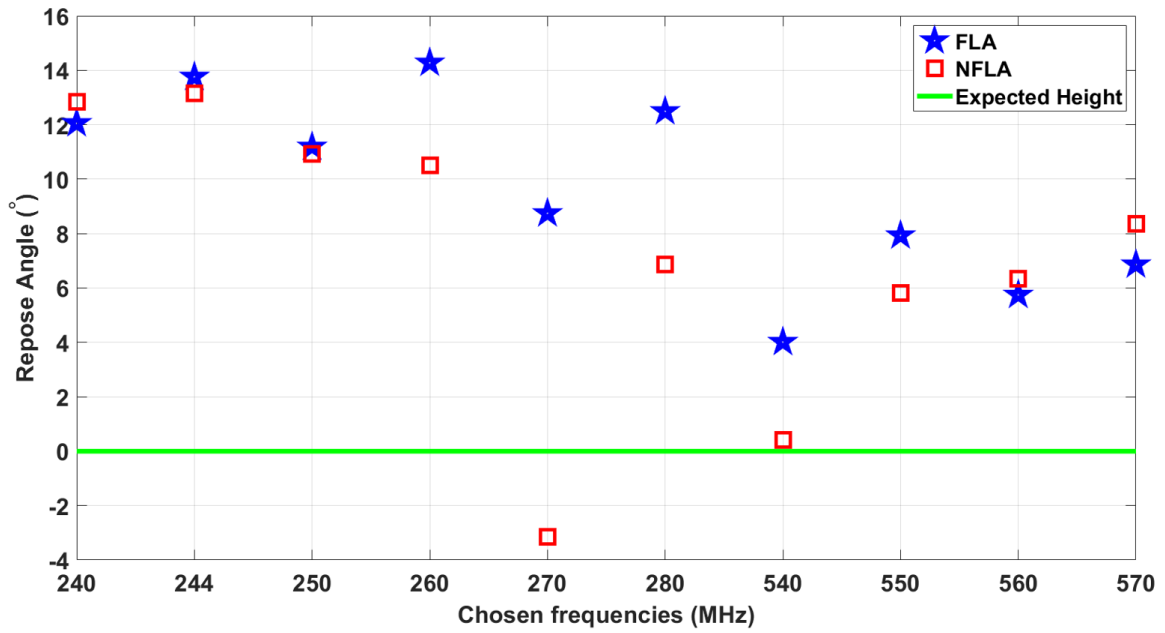


Figure 5.5: Comparing position angle estimation at different frequencies for FLA and NFLA. Note that the x -axis (frequency) division is not uniform.

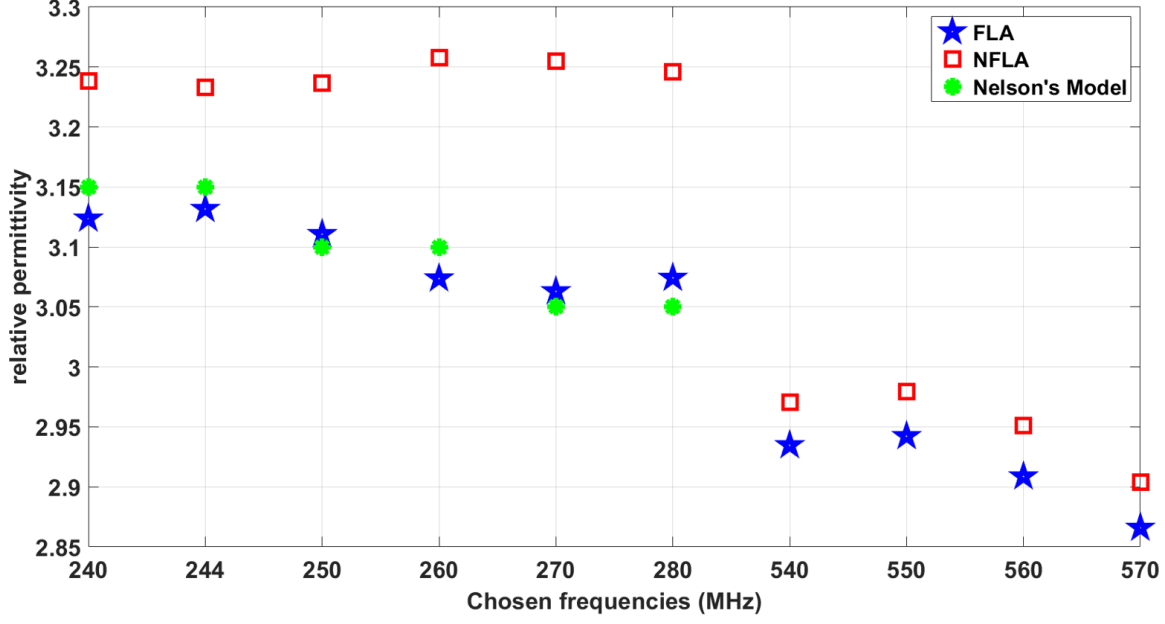


Figure 5.6: Comparing ϵ_r estimation at different frequencies for FLA and NFLA. Note that the x -axis (frequency) division is not uniform.

calibrated and the FEM-CSI inversion algorithm is ran.

For each data set, We present full 3D results for two different frequencies: 260 MHz and 570 MHz³. As mentioned earlier only real permittivity⁴ results are used to eventually measure the moisture content of the grain. All inversion results are presented on the same color scale, and this color scale is shown in Figure 5.7. This common color scale allows the direct comparison of imaging results between the FLAs and NFLAs. For the 3D threshold images, we use a threshold of $Re(\epsilon_r) > 3.6$, which is 90% of the maximum color scale of 4. Note that according to [61] at 260 MHz $Re(\epsilon_r) \approx 4.7$.

5.2.1 Imaging Results with FLA

We first present the 2D and 3D of $Re(\epsilon(\mathbf{r}))$ reconstructions of the grain with the wet target. Figure 5.7 represents the reconstructed 2D images of ROI for FLA. Additionally, 3D reconstructed ROI has been shown in Figure 5.8 along with some metrics in Table 5.1.

For the FLA imaging results in figures 5.7 and 5.8, the target center was measured about $(x, y, z) = (0.00 \text{ m}, -0.14 \text{ m}, -0.57 \text{ m})$. Based on Figure 5.7, we can estimate size and position of target; these values have been presented in Table 5.1. In this table, dimensions of the target at each frequency are d_x , d_y , and d_z where $|e_x|$, $|e_y|$, and $|e_z|$ are corresponding relative error to true dimensions in each direction and have been represented in percentage format. Moreover, p_x , p_y , and p_z are the coordinates for the center of the target. We note that all errors must be taken as approximate as the actual target was not a rectangular cuboid.

³Note that other frequencies in the two ranges provided similar results, and the results for only these two frequencies are presented in order to save space.

⁴We do not use imaginary part of relativity because it does not converge very well in reconstructing the hot spot. Also, as it has been mentioned in literature like [7], the relationship between humidity content of wheat and imaginary part of its permittivity changes a lot from one sample of wheat to another sample of wheat. Grain moisture, for these reasons is measured commercially by measuring real permittivity of the grain at various frequencies. Thus to matching other commercial grain moisture meters we only consider real part of permittivity.

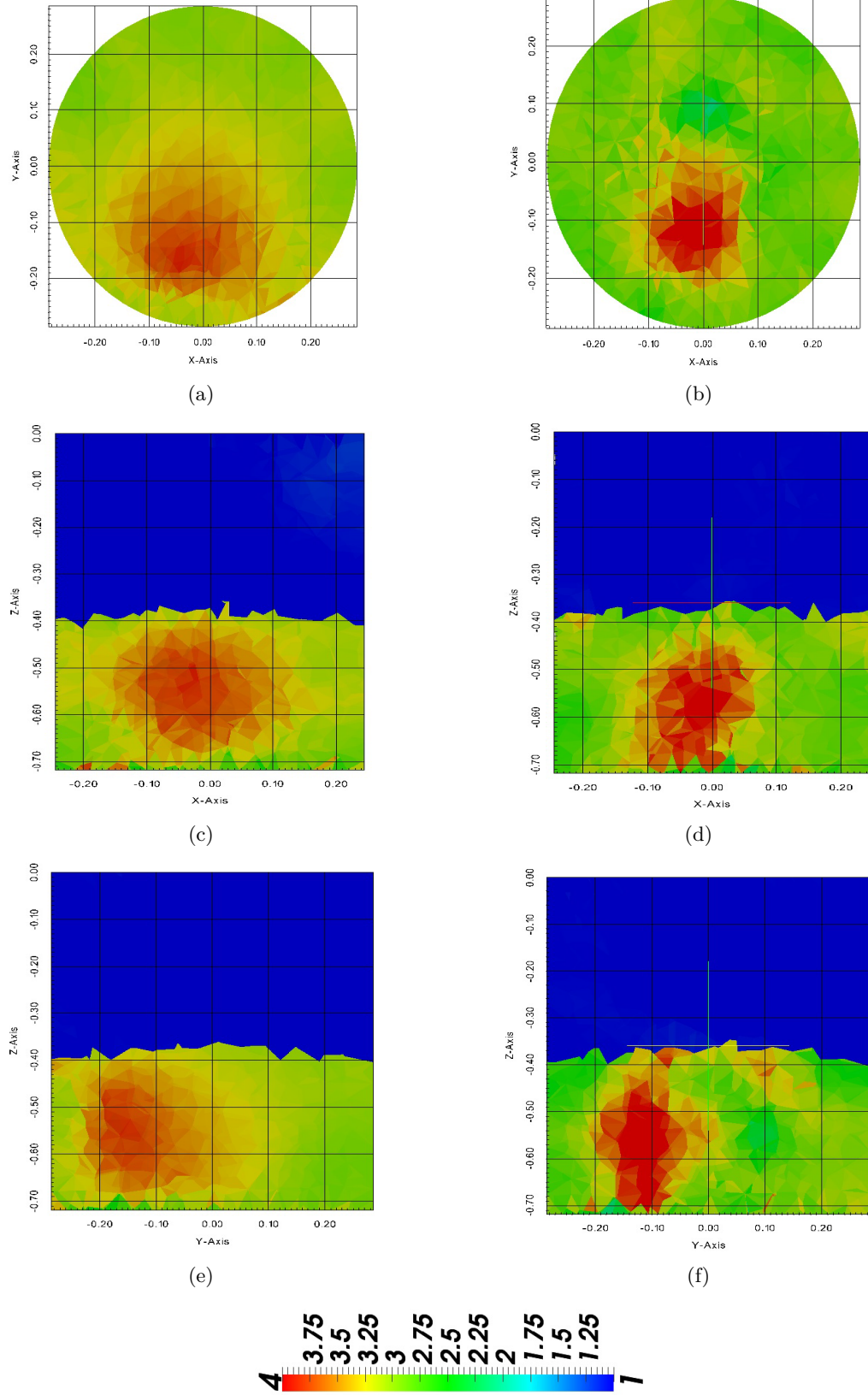


Figure 5.7: Full 3D $Re(\epsilon_r)$ image when using FLAs from different views. Left column is the reconstruction of 260 MHz data, right column is 570 MHz. (a)-(b) show the $x-y$ plane, (c)-(d) show the $x-z$ plane and (e)-(f) show the $y-z$ plane.

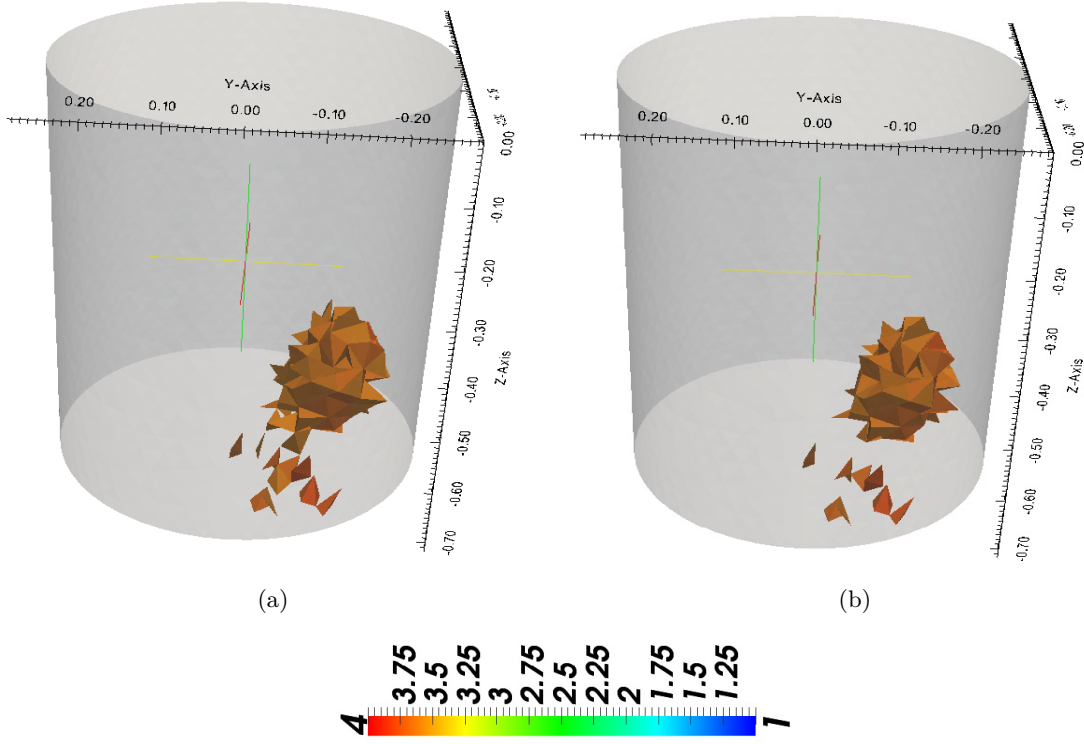


Figure 5.8: Full 3D $Re(\epsilon_r)$ image when using FLAs: (a) reconstruction of 260 MHz data, and (b) reconstruction of 570 MHz; they show 3D threshold images with all cells $Re(\epsilon_r) > 3.6$ solid.

-	260[MHz]	570[MHz]
$d_x[cm]$	12.5	11
% $ e_x $	67	47
$d_y[cm]$	10	12.5
% $ e_y $	9	14
$d_z[cm]$	20	27.5
% $ e_z $	4	31
$p_x[cm]$	-1	0
$p_y[cm]$	-14	-14
$p_z[cm]$	-58	-57

Table 5.1: FLA Results Target Metrics. Target's size (d), absolute size errors $|e|$, and position (p), obtained by imaging with FLAs while target is close to antenna 11.

-	260[MHz]	570[MHz]
$d_x[cm]$	27	11
% $ e_x $	260	47
$d_y[cm]$	16	12
% $ e_y $	45	9
$d_z[cm]$	35	16
% $ e_z $	67	24
$p_x[cm]$	0	0
$p_y[cm]$	-18	-13
$p_z[cm]$	-60	-56

Table 5.2: NFLA Imaging Results Target Metrics. Target's size and position, obtained by imaging with NFLAs, while target is close to antenna 3. Absolute value of percentage error in dimension detection has been represented as well.

To calculate the dimensions d_i and center, p_i , of the target we visually marked the red zones in each direction (x, y, z) . We then calculated the distance between each mark, and averaged them over each plane. For instance, from Figure 5.7, we obtain 12 cm and 13 cm for length of target along x -direction from xy and xz side views at 260 MHz, respectively. We then present the obtained dimension as $d_x = 12.5$ cm. Absolute percent errors, $\%e_i$ in each dimension are based on the approximate target dimensions of 7.5 cm $\times$ 11 cm $\times$ 21 cm. We again emphasize that at 250 MHz, 1 cm is about $\frac{1}{70}^{th}$ of a wavelength in grain.

5.2.2 Imaging Results with NFLA

Reconstructions of the same target with the NFLAs have been shown in Figure 5.9 and Table 5.2. Dimensions and positions of the reconstructed object has been done in a similar way to FLA's case.

NFLA does not perform well at lower frequencies, as seen both from table 5.2 and Figure 5.9. In addition, high frequency results are not that different than obtained high frequency results with FLA. As mentioned earlier we expected this because of dispersive behavior of ferrite at high frequencies. In table 5.2, dimensions' estimations are represented.

5.2.3 Target Tracking

As a final experiment, we performed a series of target tracking experiments with the FLAs, where we placed the targets in four different positions. We performed the full imaging experiment process (i.e., incident field measurement, bury the target, total field measurement, run the parametric optimization on the incident field, calibrate, then run FEM-CSI). The goal of these experiments is to show that the target is properly located in the inversion images and to determine if the target's position caused large differences in the inversion images (e.g., perhaps the target is reconstructed more accurately depending on its position). We moved the target to positions close to antennas 7, 15, 3, and 19. Figure 5.11 presents the threshold images for this experiment. Note that, to save

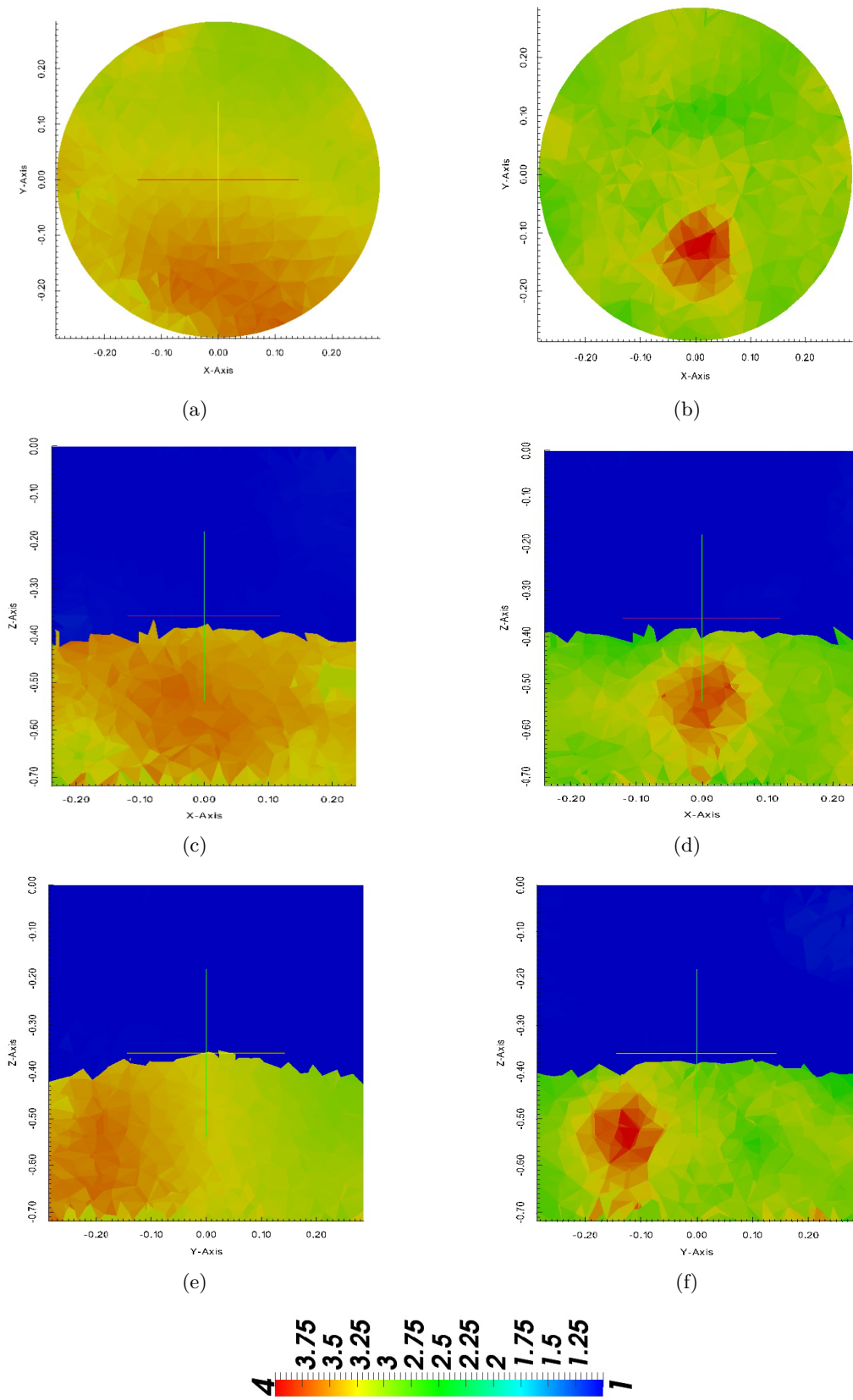


Figure 5.9: Full 3D $Re(\epsilon_r)$ image when using NFLAs from different views. Left column is the reconstruction of 260 MHz data, right column is 570 MHz. (a)-(b) show the $x-y$ plane, (c)-(d) show the $x-z$ plane and (e)-(f) show the $y-z$ plane.

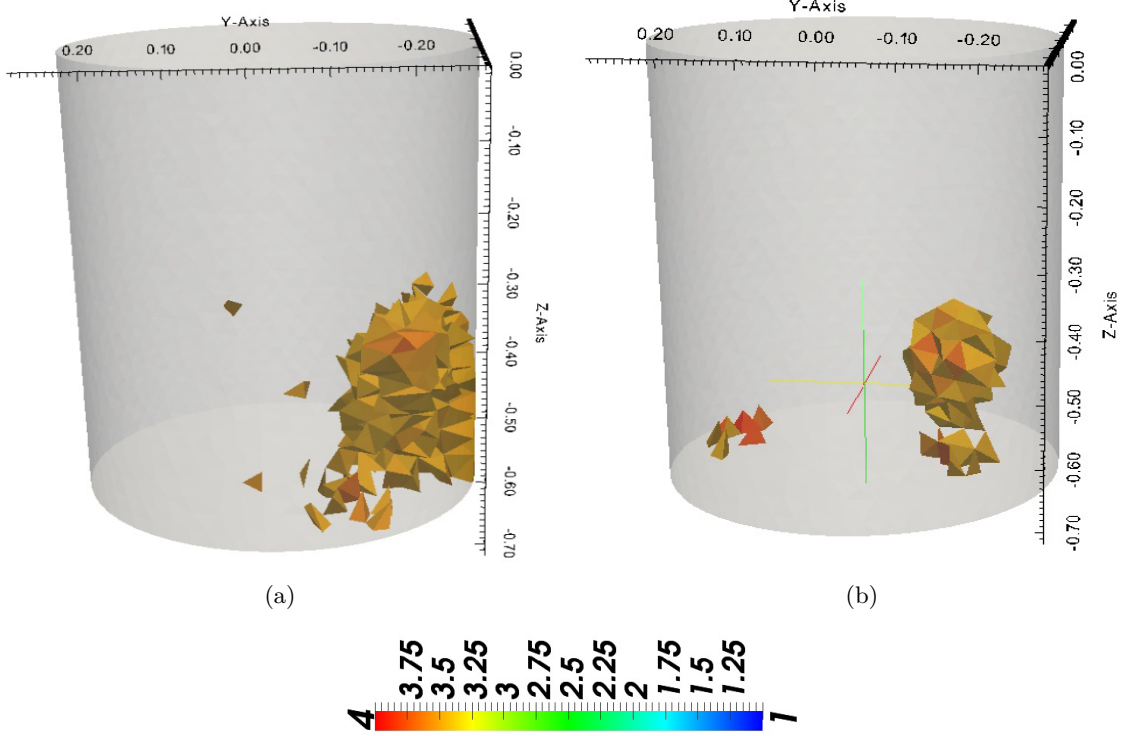


Figure 5.10: Full 3D $Re(\epsilon_r)$ image when using NFLAs: (a) reconstruction of 260 MHz data, and (b) reconstruction of 570 MHz; they show 3D threshold images with all cells $Re(\epsilon_r) > 3.6$ solid.

space, we present two different positions of the target in the same image (but would like to be clear that each image represents two different experiments of imaging a single target).

The target tracking experiments support the claim that inversion results represent the actual target position, and the reconstructed target size and position are not highly dependent on the target position.

5.3 Conclusion

We implemented small FLAs in EMI inside a small grain bin in the laboratory. Imaging results shows that FLAs can do imaging better than NFLA at lower frequencies. However, there is not big difference between imaging results of FLA and NFLA at higher frequencies, as we expected from our earlier designs based on the behavior of ferrite in the structure of the antenna.

We did some target tracking experiments by changing the position of the target, we wheat, and showed that EMI by FLA is not sensitive to the position of the target and it can detect target positions and dimensions, well.

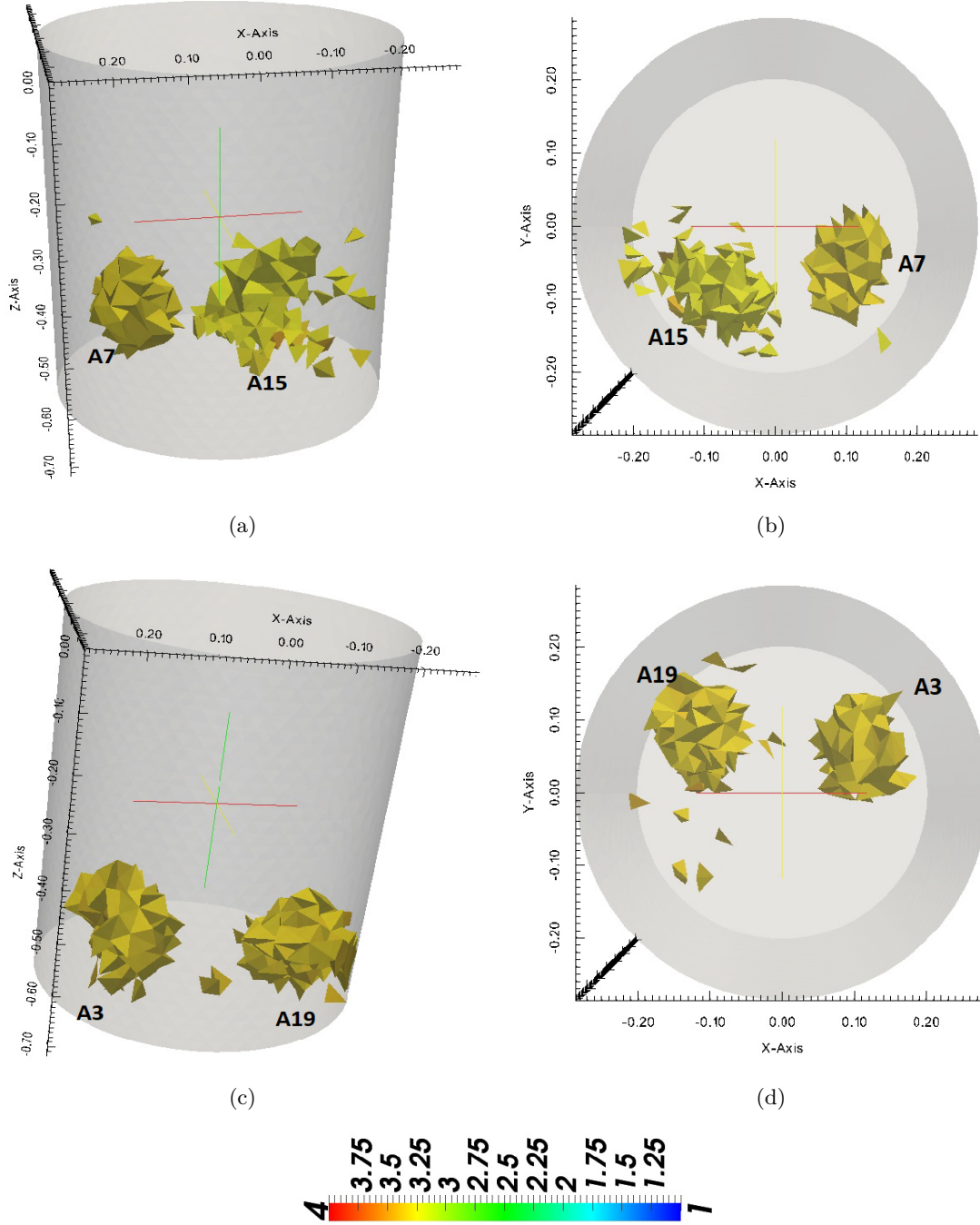


Figure 5.11: Target tracking at 260 MHz with FLAs, where A# represents target's location close to that antennas in 3^{rd} layer. In each image, we have superimposed two target positions to save space (i.e. each target position represents a separate experiment with a single wet-grain region present). (a) and (c) show perspective view of target and (b)/(d) show top view of target's position according to its distance to closest antenna in 3^{rd} layer. Threshold criteria have been chosen 90% of maximum scaled relative permittivity, i.e. 3.6.

Chapter 6

Conclusion and Future Works

In this thesis, we developed an approximate lumped model circuit for ferrite-loaded shielded loop antennas that can predict their resonance frequency and reflection coefficient. The relations for some of those components in the circuit model were derived using quasi-static approximation, although values of some of those elements, like inductance of the loop, were the same as the available ones in literature like [1]. We can also use the developed lumped model to optimize antenna efficiency by tuning it based on its radiation efficiency.

Then, we introduced some techniques like matching circuits (chapter 3) and ferrite material loading to the antenna (chapter 3 and 4) to improve its performance. Next, we designed a large FLA to be used inside large grain bins for EMI.

After comparing all of these methods, we chose the best method for this study: ferrite loading method. We presented all imaging results using this antenna in chapter 5.

Experiments with small FLAs at smaller grain bins show that they can provide up to 18 *dB* more signal than the NFLAs while maintaining (or even improving on) the essential characteristics of the NFLAs. In general, the FLA is a better fit to the assumptions made in the computational model used in the inversion process (e.g., assumption of a point magnetic source¹ and no coupling between antennas). This better fit is shown by the higher correlation coefficients between the computational fields and the raw data and the images of the wet-grain target, where the imaging performance measures (target size accuracy) were improved with the NFLA (particularly around the resonant frequency of the FLAs).

Future work will focus on re-designing this antenna concept in full-scale industrial grain bins ranging from 3-40 meters in diameter.

6.1 Future Possible Improvements in Antennas for EMI

It seems unlikely that further efforts into passive matching techniques will provide good results in the grain bin (particularly as the grain covers the antenna and shifts resonance frequencies). Nevertheless, there might be ways to improve the non-Foster technique (i.e. the use of active elements for antenna matching) for grain bin imaging.

¹This has been shown in correlation coefficient plots in the previous chapter. It has been shown that ferrite loading would increase the accuracy of the point source model in imaging, at least in the scaled (small) grain bin.

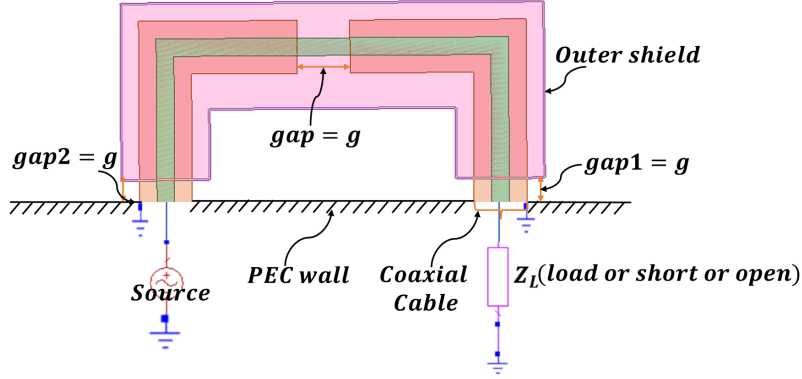


Figure 6.1: Purposed multi shielded half loop antenna without ferrite. By inserting a ferrite plate into its core, a ferrite loaded multi shielded half loop antenna is made. Ferrite has not been shown in the picture.

It is very challenging to attach a DC power supply to the bin in an industrial grain bin. However, by choosing low power consumption components for that matching circuit, we can send power to those circuits wirelessly. In other words, Wireless Power Transfer (WPT) may be added to the system. WPT can be designed in low frequency or high frequency, depending on the required power level for components of the circuits. The power transmitter can be hooked up somewhere close to the bin.

When it comes to using ferrite materials in the structure of antennas, one should remember that ferrite is very lossy at high frequencies, including the frequencies used in this study therefore still there is work to improve the efficiency of FLAs for future researches by optimizing the ferrite shape and size in the structure of antenna.

As another improvement in antenna design, we can design a multi-shielded loop with implemented gaps at different shields. Positions of gaps in each shield should be similar to positions of zero current in Figure 2.3. In this way, we can cancel out both I_x and I_y dipole modes, and the loop will be purely sensitive to magnetic current mode. See Figure 6.1 for a better understanding of this design. As it is seen from this figure, we have the usual coaxial cable with the gap at the center of the loop antenna (as same as Figure 3.1). However, in this new antenna, we have an extra metallic shield with two gaps at two ends. Size of the gap can be as same as the size of the middle gap or different than that. Also, by inserting ferrite plate into the core of the loop (similar to the mentioned small and large FLAs in the previous chapters), we will have another antenna. The latter should be better than the first one, when we consider the analysis about FLA in this thesis. For further improvement, we can add more shields to the antenna by inspecting the zero current point on the antenna in higher order current modes to make sure that higher order electric dipoles are cancelled out.

6.2 Potential Applications for Introduced Improvement Methods

Researchers at EIL research grain bin imaging, as well as breast cancer detection and forearm imaging. These types of imaging could be one of the applications of the designed FLAs in this thesis (after scaling them to the desired frequencies). The non-foster matching method can be used in

breast cancer imaging because DC power is not restricted, there.

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