ZEIT 4500 Engineering Project

Project Summary

Conformal antennas in seawater

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Conformal Antennas in Seawater
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A conformal antenna is a flat radio antenna that is designed to conform to a pre-determined surface or shape. Antennas of this type have found applications in aircraft for reducing aerodynamic drag. Electromagnetic propagation in underwater environments is becoming an area of increasing interest. Conformal antennas could have applications in improving the hydrodynamic properties of submersible vehicles and surface vessels that utilise underwater electromagnetic links.

1. Introduction

1.1 Motivation
Electromagnetic waves propagating through seawater undergo high, frequency dependent attenuation which restricts the range that signals can be transmitted. The attenuation is even greater when the signal passes through an air-water interface due to the large mismatch in the intrinsic impedances of air and water [1]. If the transmitter and receiver are both immersed in the water, avoiding the interface loss, and if they will be subject to flowing water, then the size and shape of the antenna must be considered. Here the conformal antenna has a distinct advantage since it will ideally have no protrusions from the surface of the device or vessel.

Even with both antennas submerged, there is still significant attenuation depending on the frequency. This means that a trade-off must be made between range and data rate depending on the application. Lasers can achieve wide bandwidth communications but this is only under conditions of good visibility. So the applications for radio antennas still exists for coastal and shallow water areas where visibility tends to be poor [1].

Figure 1 shows the frequency dependent attenuation for EM radiation in different types of water. Here a value of $80\varepsilon_0$ is assumed for the electrical permittivity just to provide a general idea. It is interesting to note that the attenuation increases linearly with frequency up to a certain cut off. Once past this frequency, the attenuation is constant. Unfortunately for seawater, the attenuation is already much too high before it becomes constant.

These plots were obtained using the magnitude of the electric field vector component of a propagating electromagnetic wave given by:

$$\vec{E} = E_0 e^{\gamma z}$$

(1)

Where $\gamma = \sqrt{j\omega \mu \sigma - \omega^2 \mu \varepsilon}$, $\varepsilon$ is the permittivity, and $\mu$ is the permeability of the material in farads per meter and henrys per meter respectively, $\sigma$ is the conductivity of the material in Siemens per meter and $\omega$ is the angular frequency.

Here the wave is propagating in the $z$-direction so $z$ and $E_0$ were simply set to one to
provide the attenuation per meter.

1.2 Aims
The primary focus of this project was to model and investigate the behaviour of a conformal antenna in salt water so that informed decisions could be made about the application of such antennas. The antenna consists of dielectric disc mounted on to the hull of the submerged vehicle or structure with an electrode in the middle which is exposed to the water. Figure 2 shows a cross section diagram of this type of antenna with electric field lines that show what will be the likely shape of the current distribution.

The modelling method was a Finite Difference Time Domain (FDTD) algorithm implemented using MATLAB. The FDTD modelling was more time consuming than using a commercial software package since the method needed to be validated before it could be used to make predictions. This was done by simulating some simple transmission line problems and comparing the results to analytical solutions.

The experimental component tested the predictions made by the FDTD model. A suitable conformal antenna was designed, constructed and characterised by measuring the electric field around it in a small salt water tank. The results from the characterisation were then compared to predictions made by the model.

2. FDTD Implementation
The FDTD method is a powerful tool for solving Maxwell’s equations. It uses updating equations, derived from scalar representations of Maxwell’s equations, which are discrete in time and space. This discrete nature allows us to construct a three dimensional grid which represents the problem geometry. This was first done by Yee in 1966 [4] so one cell of this grid is known as a Yee cell [2].

There are two main motivations behind developing a model as opposed to making use of existing software. Firstly, developing a model is an excellent way to learn about how Maxwell’s equations operate on a fundamental level. Any mistakes made or difficulties encountered will provide insight into how the equations behave when quantised. It will also provide a better understanding of simulation in general rather than simply learning how to use a particular piece of software. The second motivation is to produce a model, tailored specifically to the conformal antenna problem, which anyone can use without concern for licenses or access to other software.

2.1 FDTD Theory
The basic theory behind the FDTD method is the updating equations. To derive these equations and construct a FDTD algorithm, the starting point is Maxwell’s time domain equations.

\[
\nabla \times \vec{E} = -\frac{\partial \vec{B}}{\partial t} - \vec{M}
\]

\[
\nabla \times \vec{H} = \frac{\partial \vec{D}}{\partial t} + \vec{J}
\]

\[
\nabla \cdot \vec{D} = \rho_e
\]

\[
\nabla \cdot \vec{B} = \rho_m
\]
Where $\vec{E}$ is the electric field strength vector in volts per meter, $\vec{D}$ is the electric displacement vector in coulombs per square meter, $\vec{H}$ is the magnetic field strength vector in amperes per meter, $\vec{B}$ is the magnetic flux density vector in webers per square meter, $\vec{J}$ is the electric current density vector in amperes per square meter, $\vec{M}$ is the magnetic current density vector in volts per square meter, $\rho_e$ is the electric charge density in coulombs per cubic meter, and $\rho_m$ is the magnetic charge density in webers per cubic meter [2].

When deriving the FDTD equations only the curl equations need to be considered because the divergence equations can be satisfied by the FDTD updating equations [3]. Using the relationships between $\vec{E}$ and $\vec{D}, \vec{H}$ and $\vec{B}$ the curl equations can be rewritten as

$$\nabla \times \vec{E} = -\mu \frac{\partial \vec{H}}{\partial t} - \sigma^m \vec{H} - \vec{M}_i$$  \hspace{1cm} (3.1)

$$\nabla \times \vec{H} = -\varepsilon \frac{\partial \vec{E}}{\partial t} - \sigma^e \vec{E} - \vec{J}_i$$ \hspace{1cm} (3.2)

These two vector equations can now be decomposed into six scalar equations in Cartesian coordinates. Only the x-component of the electric and magnetic field strength is shown here.

$$\frac{\partial E_x}{\partial t} = \frac{1}{\varepsilon_x} \left( \frac{\partial H_z}{\partial y} - \frac{\partial H_y}{\partial z} - \sigma_x^e E_x - j_{ix} \right)$$ \hspace{1cm} (4.1)

$$\frac{\partial H_x}{\partial t} = \frac{1}{\mu_x} \left( \frac{\partial E_y}{\partial z} - \frac{\partial E_z}{\partial y} - \sigma_x^m H_x - M_{ix} \right)$$ \hspace{1cm} (4.2)

These equations now need to be approximated to a form that is discrete in time and space and applied to a Yee cell. Figure 2 shows the positions of the field components in four adjacent Yee cells. Only the field components that contribute to $\partial E_x / \partial t$ are shown since the picture gets very crowded otherwise. Note the four magnetic field vectors surrounding the electric field vector, simulating Ampere’s law.

![Figure 2: Field components in four adjacent Yee cells.](source: [2])

With these positions the discrete approximation to equation 4.1 can now be written. The final result is equation 5, the updating equation for the x-component of the electric field.

![Figure 3: Field components around E_x](source: [2])
\[ E_{x}^{n+1}(i,j,k) = \frac{2\varepsilon_x(i,j,k) - \Delta t\sigma_x(i,j,k)}{2\varepsilon_x(i,j,k) + \Delta t\sigma_x(i,j,k)} E_x^n(i,j,k) \]
\[ + \frac{2\Delta t}{2\varepsilon_x(i,j,k) + \Delta t\sigma_x(i,j,k)\Delta y} \left( H_z^{n+\frac{1}{2}}(i,j,k) - H_z^{n+\frac{1}{2}}(i,j-1,k) \right) \]
\[ - \frac{2\Delta t}{2\varepsilon_x(i,j,k) + \Delta t\sigma_x(i,j,k)\Delta z} \left( H_y^{n+\frac{1}{2}}(i,j,k) - H_y^{n+\frac{1}{2}}(i,j,k-1) \right) \]
\[ - \frac{2\Delta t}{2\varepsilon_x(i,j,k) + \Delta t\sigma_x(i,j,k)\Delta x} I_x^{n+\frac{1}{2}}(i,j,k). \] (5)

Where \( \Delta x, \Delta y, \) and \( \Delta z \) are the Yee cell dimensions shown in figure 1 and \( \Delta t \) is the discrete time step. The updating equations for the \( y \) and \( z \)-components of the electric field have exactly the same form however the magnetic field equations are slightly different because the magnetic field vectors are offset from the electric field vectors by half a cell. This means that they are also offset in time by \( \frac{\Delta t}{2} \).

### 2.2 FDTD Validation

With guidance from a textbook [2] on the subject, the FDTD method was then implemented using MATLAB. Some simple one dimensional simulations were performed first then the simulation was expanded to three dimensions. Two different transmission line problems were modelled. Both problems have straightforward analytical solutions or approximations. The first was an air filled coax cable with a square cross section. Being square meant that the surface boundaries of the cable would align perfectly with the edges of the Yee cells thus giving the simulation the best chance for accurate results. The next transmission line was also air filled coax but with a circular cross section. This allowed the effects of imperfect geometry to be observed since the cubic Yee cells could never exactly match the curved surfaces of the cable.

First the geometry of the cable needed to be programmed. To get an idea for the appropriate dimensions, the following equations were used to calculate the characteristic impedance, \( Z_0 \). Figures 4 and 5 show the cross sections for the square line and circular line respectively.

\[ Z_0 = \frac{\eta_0}{\sqrt{\varepsilon_r}} \left[ \frac{1}{4\left(\frac{D-d}{D\cdot d}-0.558\right)} \right] \] (6)

Source: [4]

\[ Z_0 = \frac{n}{2\pi} \ln \left( \frac{D}{d} \right) \] (7)

Source: [6]

Where \( \eta_0 = \sqrt{\frac{\mu_0}{\varepsilon_0}} \) and \( \varepsilon_r \) is the relative permittivity of the dielectric, air in this case. \( d \) and \( D \) are the widths of the inner and outer conductors respectively.

\[ V_{meas} = V_{in} \frac{Z_0}{Z_0 + Z_{int}} \] (8)

Figure 4: Cross section of square coax cable.

Figure 5: Comparison between 1mm and 0.25mm cells
Figure 6 shows the side view of the complete geometry of the cable. The cable is entirely PEC and it is capped at both ends with PEC plates. The voltage is sampled between the inner conductor and the outer conductor at a position 5mm from the source. A simple voltage divider can be used as an equivalent circuit to predict the voltage we should measure (Equation 8). Equations 6 and 7 were substituted into the voltage divider equation. Setting $Z_{int}$ to 50Ω, $D$ to 40mm and $V_{in}$ to 1V, the inner conductor diameter, $d$, was set to produce a characteristic impedance of as close to 50Ω as possible.

With some idea of what to expect, the simulation could now be run using the calculated dimensions. The simulation used a unit step voltage to excite the system. Figure 7 shows the results from a square cable simulation comparing models using mesh cells with 1mm sides and cells with 0.5mm sides i.e. 8 times the amount of cells.

The first thing we notice is that the measured voltage does not stay high for the entire time. This is because once the voltage step propagates to the end of the line, it will be reflected back causing destructive interference along the cable. This is what we would expect from a system that is composed entirely of PEC since eventually, all the voltage potentials will be zero.

The second thing we notice is the oscillations as the step reaches it’s maximum. This could either be an effect of the discrete nature of the simulation or a real transient effect that is being simulated. It can be seen that as the oscillations settle, the voltage agrees very well with the predicted value, marked on the figure as a horizontal red line. If the oscillations were a product of discretisation the reduction in cell size would have produced an obvious change. It can be seen that while the smaller cells produced a smoother curve, the frequency and amplitude of the oscillations remain the same, indicating that they are a transient effect that is being simulated.

Next a Gaussian unit pulse was used to excite the system. This had the advantage of eliminating the oscillations and providing a single clear peak to use as a measure of the accuracy of the simulation. Figure 8 shows the results for the square coax simulation.
The figure shows the peak of the Gaussian pulse and the red line is the predicted DC voltage. For the square coax case, decreasing the size of the cells resulted in a voltage increase from 0.493V to 0.497V, an error margin decrease of 0.8% (1.2% down to 0.4%).

The same procedure as the square coax was followed for the circular coax, the only differences being the circular cross section of the cable. Figure 9 shows the results for the circular coax simulation.

The figure shows that the reduction in cell size resulted in the peak voltage going from 0.495V to 0.502, an error margin decrease of 1.5% (2.3% down to 0.8%). This is in line with expectations, there was an increase in the accuracy but the simulation will always be less accurate than the square geometry due to the approximation of curved surfaces.
2.3 FDTD Antenna Simulations

With the FDTD method validated, a basic simulation was set up that would resemble the geometry of a conformal antenna. Construction of the antenna had not begun at this stage so this simulation did not match the real situation. A DC source was used so that there were no phase or propagation effects to be concerned with and this resulted in an electrostatic type solution. Contour plots of the x and y components of the electric field were made and these are shown in Figure 9. This simulation was simply a qualitative exercise and it was decided not to operate the antenna above 5MHz so that propagation effects would not become a significant factor anyway. A 5MHz signal has a wavelength of 705mm in salt water and the measurements were only to be taken within a distance of 250mm from the antenna.

![Figure 9: Simulated electric field contours using a DC source.](image)

3 Antenna Characterisation

The experimental part of the project began with the construction of the antenna. This consisted of a galvanised steel ground plate which acted as the hull of a submerged structure. The dielectric disc was simply adhesive plastic 30cm in diameter. A hole was drilled through the centre of the disc that would allow the core and Teflon insulation of a coax cable to pass through it. A screw was welded onto the rear side of plate close to the hole and the shielding of the coax cable was attached to this. On the front of the plate, a metal dome was soldered onto the core of the coax cable. All the gaps were then sealed up with silicon to ensure that there would be no unwanted current paths once the antenna was submerged in salt water. Figure 10 is a diagram of the antenna electrode.

![Figure 10: Conformal antenna electrode components.](image)

The antenna was then placed in a square tank with dimensions approximately 1mx1m×0.7m and the tank was filled with water.

Figure 11 is a photo of the front of the antenna submerged in water. The edge of the dielectric disc has been outlined with a green marker.

![Figure 11: Conformal antenna submerged in salt water.](image)
Once the antenna had been constructed, a method for measuring the electric field needed to be established. A simple dipole probe was made using a small Printed Circuit Board. The ends of the dipole were 1cm apart and were left exposed so that they could conduct when immersed in salt water. This is because in salt water, at the frequencies this antenna would be operating at, the contribution to the electric field from conduction currents are much stronger than the contribution from displacement currents. Figure 12 is a photo of the end of the probe.

The tracks of the PCB probe were soldered to a shielded twin wire cable and a hollow plastic rod was used to hold the probe and cable in place. The end of the rod was then sealed up with silicon as can be seen in Figure 12. Figure 13 is a photo of the complete apparatus with a mount for the probe’s rod which allowed it to be moved easily but remain held in place when required.

The probe cable was connected to a spectrum analyser through a balun purchased from Jaycar so that the differential signal from the probe could be balanced with reference to the analyser’s ground. The antenna itself was driven directly from a signal generator at frequencies of 2MHz and 5MHz. The coordinate system that was used for the measurements had the origin at the antenna electrode with the ground plate lying in the y-z plane and the bore sight of the antenna as the x-axis. Measurements of the x and y components of the electric field were measured by aligning the probe parallel to the respective axes. Figures 14 and 15 are contour plots of the measurements for 2MHz and 5MHz respectively.
The results show that the electric field exhibits the expected pattern shown in Figure 2 and it even resembles the electrostatic simulation shown in figure 9. The x components are strongest at the centre of the disc and attenuate over distance with the attenuation being greater when moving directly away from the x-axis. The y-components however are weaker at the centre and increase in strength when moving away from the x-axis before decreasing again. For the 5MHz source, the fields are weaker than for the 2MHz source as would be expected from Figure 1. The third plot on the right was produced using MATLAB’s `quiver` function to combine the x and y components and it shows the ‘umbrella’ like electric field vectors.

When looking at the y components, both 2MHz and 5MHz results show some asymmetry in the pattern. The fields appear stronger in the lower half of the plot. It may be an error in the measurements but since both results exhibit the effect it is probably a real measurement. It is likely caused by the screw source that the source shielding is connected to since it was approximately 2cm from the centre. This would cause the current distribution to favour the side of the screw since this is the shortest path.

Figures 16 and 17 show comparisons between simulated and measured data for the 2MHz and 5MHz sources respectively. This was data taken in a line parallel to the ground plate at a distance of 200mm. The reason why there are so few simulated data points is because the method for extracting data from the simulation is somewhat crude. A field at a point in the volume has to be sampled for the whole simulation run with a typical run going for 500,000 time steps. So the points to be sampled had to be carefully selected.
The plots show that the basic features of the field patterns agree with the measurements with a clear peak at the centre for the x components and a minimum at the centre for the y components. The Simulations for both the x and y component with the 2MHz source show a steeper attenuation with distance than what was measured. This was also the case for the x component with the 5MHz source. The reasons for this are as yet unknown and need further investigation however the y component with the 5MHz source showed a remarkably similar trend indicating that the simulations and measurements can agree very well under the right conditions.

4 Recommendations

One of the greatest limiting factors of the FDTD simulation used in this project was the time it took to run a simulation of adequate length. Since the antenna is electrically small, it meant that the maximum cell size was quite small. This resulted in over 2 million cells needed to simulate the whole volume. The perfect solution in this situation would be adaptive meshing where small cells can be used in one area and large ones in another. The implementation of adaptive meshing would dramatically reduce the number of required cells and hence the time needed for simulations requiring a high number of time steps.

Another limiting factor was the amount of data that could be extracted from the simulation. The implementation of a more efficient extraction method would be a major benefit and would reduce the number of repeated simulation runs. It would also allow for more instructive simulated results such as animations.

The FDTD code that was put together for this project could have other applications in future projects. This project has demonstrated that it works well for small confined geometries, such as the transmission line problems used for validation, and so could be useful in studying waveguides or meta-materials.

With respect to experimentation, the method used for taking measurements could be improved. A problem was encountered which involved a common mode signal being introduced into the probe. The method that was used to reduce its effects involved taking twice the number of measurements and even so, this common mode signal may have been responsible for some of the discrepancies between the simulated and measured data. This common mode problem could be further investigated an eliminated to provide more reliable measurements. Additionally, further experiments could be conducted on the conformal antenna, such as a broader frequency range, different disc sizes, different electrode surface areas, or using different signal waveforms such as pulses or chirps.
References


