3D ELECTROMAGNETIC SIMULATION TOOL EXPOSURE FOR
UNDERGRADUATE ELECTRICAL ENGINEERS:
INCORPORATION INTO AN ANALOG FILTERS COURSE

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In Partial Fulfillment of the Requirements for the Degree
Master of Science in Electrical Engineering

by
Bobby Bamnang Pheng

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COMMITTEE MEMBERSHIP

TITLE: 3D Electromagnetic Simulation Tool Exposure for
Undergraduate Electrical Engineers: Incorporation into an
Analog Filters Course

AUTHOR: Bobby Bamnang Pheng

DATE SUBMITTED: June 2012

COMMITTEE CHAIR: Dennis Derickson, Department Chair of Electrical
Engineering

COMMITTEE MEMBER: Ahmad Nafisi, Professor of Electrical Engineering

COMMITTEE MEMBER: Bridget Benson, Associate Professor of Electrical
Engineering

ABSTRACT

3D Electromagnetic Simulation Tool Exposure for Undergraduate Electrical Engineers:

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Bobby Pheng

With the growth of wireless communications, comes the need for engineers knowledgeable in 3D electromagnetic (EM) simulation of high-frequency circuits. To give electrical engineering students a better understanding of the behavior of electromagnetic fields, experiments including the use of 3D EM simulation software were proposed. Most students get lost in differential equations, curls, and divergences; this thesis aims to remedy that by exposing them to 3D EM simulation, which may motivate them toward further study in electromagnetics. Also, experience using EMPro is very beneficial for future RF/microwave/antenna engineers, as use of 3D EM simulation is becoming a requirement for this field. 3D EM simulators solve problems where using classical analysis techniques is impractical. Classical EM solutions to simple objects such as boxes, cylinders, and spheres, are widely known; but when the object is more complex, numerical approaches are preferred for their speed.

Currently, Cal Poly does not use 3D electromagnetic simulation in any of its courses. Targeted relevant courses include EE 335/375: EM Fields & Transmission Lines, EE 402: EM Waves, EE 405/445: High-Frequency Amplifier Design, EE 425/455: Analog Filter Design, EE 502: Microwave Engineering, and EE 533: Antennas. As a starting point, EE 425/455 was targeted.

In choosing which filters to investigate, simplicity and cost were the most important factors. For simplicity, transverse electromagnetic (TEM) mode filters were chosen; also, using a trough design for these filters would allow for simple construction and access. Also, a circular waveguide filter was chosen as an alternative to the TEM filters, as the modes are either transverse electric or transverse magnetic. To lower costs, printed circuit board was used to construct the filters, along with brass tubing, semi-rigid coaxial cable, and copper plumbing caps.

From these guidelines, three electronic bandpass filter experiments were investigated: a 1 GHz half-wave coaxial resonator filter, a 2 GHz copper end cap filter, and a tunable 1 GHz quarter-wave coaxial resonator filter. Electric and magnetic field coupling was used to excite the filters. They were then simulated using finite difference time domain (FDTD) simulations in Agilent EMPro. From the simulations, tradeoffs between insertion loss and bandwidth were observed. After, the filters were built and measured using a network analyzer. The quarter-wave filter was incorporated in Cal Poly's EE 455 course during spring 2012. Students completed an EMPro tutorial, simulated the filters, and measured them using network analyzers. Student feedback was mixed, and modifications were made for future implementations.

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1. INTRODUCTION

The field of wireless communications has seen much growth in recent times, and with it, comes the need for engineers knowledgeable in RF/microwave engineering and antenna design. Most electrical engineering curriculums offer courses in these areas, but few offer coursework in 3D EM simulation. This is an important area of RF/microwave engineering, as most if not all engineers in this field use 3D EM software.

The goal of this thesis is to provide Cal Poly electrical engineering students a better understanding of electromagnetic field behavior by incorporating 3D electromagnetic (EM) simulation software into the undergraduate curriculum using Agilent EMPro. This would provide them invaluable 3D EM simulation experience in preparing them for future engineering work.

Also, most students are discouraged from concentrating in electromagnetics after seeing Maxwell's equations for the first time, as they are put off by the many differential equations, integral equations, divergences, and curls. EMPro allows for the visualization of electric and magnetic fields; by looking at the behavior of these fields, students would gain a better understanding of how Maxwell's equations work, and be shown that EM research does not just consist of curls and divergences.

1.1 Overview of Relevant Cal Poly EE Coursework & Targeting of EE 425/455

In choosing a course to implement the software, a list of relevant Cal Poly EE courses was analyzed. It includes EE 335/375: EM Fields & Transmission Lines, EE 402: EM Waves, EE 405/445: High-Frequency Amplifier Design, EE 425/455: Analog Filter Design, EE 502: Microwave Engineering, and EE 533: Antennas.

EE 335/375 and EE 402 are the most attractive courses for the software, as all electrical engineering students are required to complete them. Also, EE 533 could greatly benefit from the software, as various antenna parameters are easily viewed in simulation. As a starting point, EE 425/455 was chosen to implement the software because of the current resources available: computers with EMPro, multiple network analyzers, and multiple soldering stations. Simple-to-construct filters could be simulated in EMPro then built and measured using network analyzers. This provides students both simulation and hands-on measurement experience.

1.2 What Do 3D EM Simulators Do?

3D EM simulators essentially analyze the 3D EM effects of structures. Usually, basic structures (cylinders, boxes, and spheres) are easily analyzed using Maxwell's equations. What happens if the structure is a combination of all of these shapes? What if different parts of the structure have different permittivities and/or permeabilities? In these cases, standard analytical problem-solving approaches become impractical because of the amount of time required to solve the problem.

These problems are solved using computational electromagnetics (CEM). CEM includes numerous methods in solving Maxwell's equations over these irregularly-shaped structures. Although each method has its own advantages and disadvantages, they all essentially do the same thing. First, the irregularly-shaped structure is broken down into many regularly-shaped pieces, which can be analyzed using approximations to Maxwell's equations. The electric and/or magnetic fields are solved for in each of the smaller pieces and combined to give a solution for the entire structure. In doing this, an initial approximation is made in terms of sampling the volume. To solve Maxwell's equations,

approximations also have to be made to the equations, either in integral or differential form. These approximations lead to errors, which can be minimized by increasing solution times.

The different methods are usually labeled by what type of Maxwell's equations approximation they use. The three most popular methods include an integral equation solver (method of moments) and two differential equation solvers (finite element method and finite difference time domain). The two differential equation solvers will be discussed in later sections.

Specifically, EM simulators are great for finding the modes of a rectangular cavity resonator, finding the amount of unwanted EM radiation being generated from an integrated circuit, or finding the directivity and radiation pattern of an antenna.

1.3 Selection of EMPro over Other 3D EM Simulators

Several simulators were considered, including AWR Microwave Office, Ansys HFSS, and Agilent EMPro. The decision to use EMPro for this thesis came down to several factors, the most important being cost. Cal Poly currently uses Agilent ADS for many of its courses; because of this, a discounted 50-seat EMPro license was bundled with the existing ADS license. Also, licenses for HFSS and Microwave Office were much more expensive than EMPro.

In addition to its simulation capabilities, EMPro includes a very powerful 3D modeling environment. Existing models can also be imported from other popular packages such as AutoCAD or Solidworks. Python scripting is also available for further automation.

Another benefit of using EMPro involves its integration with ADS designs. For example, FEM and/or FDTD simulations of an existing ADS microstrip filter can be completed in EMPro by importing the model from ADS. Also, a component simulated in EMPro can be imported into ADS; the component parameters can then be simulated with an existing circuit.

EMPro includes two 3D EM simulators: the finite element method (FEM) and the finite difference time domain (FDTD) method. Both methods are used to approximate and solve partial differential equations. A brief summary of how these methods work is included in the following sections followed by a comparison of both techniques. Ultimately, the FDTD method is chosen mainly because of its shorter simulation times.

1.4 Possible EE 425/455 Filters & Implementation during Spring 2012

In deciding which filter the students would simulate and build, three designs were considered: a half-wave coaxial resonator filter, a tunable quarter-wave coaxial resonator filter, and a tunable copper end cap filter. Classical analysis of the electromagnetic fields in a coaxial resonator is presented in later sections as well as coupling analysis. Other filters are considered in the conclusion section for future work.

All three filters were simulated in EMPro and measured using a network analyzer. For each filter, various simulations were completed that involved changing the dimensions of certain parts of the filters. This shows what students would expect to see when they tune the filters themselves. All three filters were tuned based on S21 insertion loss and bandwidth, although other parameters could be considered.

Ultimately, the quarter-wave filter was chosen since it was tunable, cheaper to construct than the other two, simple to build, and was less memory intensive to simulate.

It was implemented into EE 455 during spring 2012, and students were given a short EMPro tutorial, then built and measured the filters. Detailed results and feedback are discussed in later sections.

1.5 Provided Educational Materials

Several educational materials are enclosed in the appendices for future sections of EE 425/455. Included is an EMPro tutorial in appendix A, a lecture material set on electromagnetic simulation in Appendix B, and a laboratory material set in Appendix C. These materials provide quick overviews on computational electromagnetics and how to use EMPro.

2. EMPRO SIMULATORS

2.1 The Finite Element Method (FEM)

EMPro includes two simulators, one of which uses the finite element method (FEM) to solve simulations. The basic idea of the FEM involves dividing the simulation domain into elements, generating governing equations for these elements, relating all of the elements, and solving the resultant system of equations [1].

To further explain how the FEM works, Laplace's equation is solved for a simple 2D example, but the analysis can be applied to other EM partial differential equations and 3D models. The partial differential equation being solved for is Laplace's equation for voltage:

$$\nabla^2 V = 0 \quad (2.1)$$

Solving this equation will result in finding unknown voltages along the example geometry. The example consists of a quadrilateral divided into two triangles as shown in Figure 1 below. Nodes are assigned numbers globally (red/outside) and locally (blue/inside). Elements are assigned as (1) and (2). Global nodes are assigned randomly, but local nodes are labeled in a counter clockwise manner. This eases matrix organization later on.

Figure 1: FEM 2D Example

The first step involves finding the potential in each element, then these potentials are interrelated among elements. The potential in a single 2D triangle element changes linearly with position and is approximated as

$$V_e(x,y)=a+bx+cy=[1 \quad x \quad y] \begin{bmatrix} a \\ b \\ c \end{bmatrix} \quad (2.2)$$

Where the subscript 'e' represents a single element. This assumes linear potential variation in the element. The total potential for the quadrilateral is the sum of all elemental potentials. Next, a set of governing equations is defined using the potentials at each node:

$$\begin{bmatrix} V_{e1} \\ V_{e2} \\ V_{e3} \end{bmatrix} = \begin{bmatrix} 1 & x_1 & y_1 \\ 1 & x_2 & y_2 \\ 1 & x_3 & y_3 \end{bmatrix} \begin{bmatrix} a \\ b \\ c \end{bmatrix} \quad (2.3)$$

V_{e1} , V_{e2} , and V_{e3} are the node voltages, and they are located at x_i , y_i for $i = 1, 2$, and 3 .

The a , b , and c unknowns are then solved for and inserted into equation (2.2):

$$V_e(x,y)=[1 \quad x \quad y] \frac{1}{2A} \begin{bmatrix} x_2y_3-x_3y_2 & x_3y_1-x_1y_3 & x_1y_2-x_2y_1 \\ y_2-y_3 & y_3-y_1 & y_1-y_2 \\ x_3-x_2 & x_1-x_3 & x_2-x_1 \end{bmatrix} \begin{bmatrix} V_{e1} \\ V_{e2} \\ V_{e3} \end{bmatrix} \quad (2.4)$$

$2A$ is the determinant, and it can be shown that A is the area of the triangle element.

Equation (2.4) can be represented as

$$V_e(x,y)=\sum_{i=1}^3 \alpha_i(x,y)V_{ei} \quad (2.5)$$

where

$$\alpha_1=\frac{1}{2A} [(x_2y_3-x_3y_2)+(y_2-y_3)x+(x_3-x_2)y] \quad (2.6)$$

$$\alpha_2=\frac{1}{2A} [(x_3y_1-x_1y_3)+(y_3-y_1)x+(x_1-x_3)y] \quad (2.7)$$

$$\alpha_3 = \frac{1}{2A} [(x_1 y_2 - x_2 y_1) + (y_1 - y_2)x + (x_2 - x_1)y] \quad (2.8)$$

These equations are known as the element shape functions. Next, the elemental energy is given as

$$W_e = \frac{1}{2} \int \epsilon |\mathbf{E}_e|^2 dS = \frac{1}{2} \int \epsilon |\nabla V_e|^2 dS \quad (2.9)$$

Substituting equation (2.5) into equation (2.9) gives

$$W_e = \frac{1}{2} \sum_{i=1}^3 \sum_{j=1}^3 \epsilon V_{ei} \left[\int \nabla \alpha_i \cdot \nabla \alpha_j dS \right] V_{ej} \quad (2.10)$$

The bracket term is defined as

$$C_{ij}^{(e)} = \int \nabla \alpha_i \cdot \nabla \alpha_j dS \quad (2.11)$$

Writing equation (2.10) in matrix form gives

$$W_e = \frac{1}{2} \epsilon [V_e]^T [C^{(e)}] [V_e] \quad (2.12)$$

The 3x3 matrix, $[C^{(e)}]$, is called the element coefficient matrix and relates the coupling between nodes i and j . Now that a single element has been considered, the total energy of all elements is given as

$$W = \sum_{e=1}^N W_e = \frac{1}{2} \epsilon [V]^T [C] [V] \quad (2.13)$$

and

$$[V] = \begin{bmatrix} V_1 \\ V_2 \\ V_3 \\ \vdots \\ V_n \end{bmatrix} \quad (2.14)$$

N is the total number of elements; n is the total number of nodes; [C] is the global coefficient matrix. The global coefficient matrix from Figure 1 is a 4x4 matrix since it contains 4 nodes:

$$[C] = \begin{bmatrix} C_{11} & C_{12} & C_{13} & C_{14} \\ C_{21} & C_{22} & C_{23} & C_{24} \\ C_{31} & C_{32} & C_{33} & C_{34} \\ C_{41} & C_{42} & C_{43} & C_{44} \end{bmatrix} \quad (2.15)$$

Since the potential distribution across elements is continuous, the individual elements in the global coefficient matrix are found using a sum of all elements touching the specified node(s). For example,

$$C_{11} = C_{11}^{(1)} + C_{11}^{(2)}$$

since both elements (1) and (2) touch global node 1. Also,

$$C_{12} = C_{12}^{(1)}$$

since, only element (1) touches boundary between global nodes 1 and 2. The completed global coefficient matrix is

$$[C] = \begin{bmatrix} C_{11}^{(1)} + C_{11}^{(2)} & C_{12}^{(2)} & C_{12}^{(1)} + C_{13}^{(2)} & C_{13}^{(1)} \\ C_{21}^{(2)} & C_{22}^{(2)} & C_{23}^{(2)} & 0 \\ C_{21}^{(1)} + C_{31}^{(2)} & C_{32}^{(2)} & C_{22}^{(1)} + C_{33}^{(2)} & C_{23}^{(1)} \\ C_{31}^{(1)} & 0 & C_{32}^{(1)} & C_{33}^{(1)} \end{bmatrix} \quad (2.16)$$

The global coefficient matrix is symmetric, and for a large number of nodes, most of the elements in the matrix are zero. Next, it can be shown that Laplace's equation is solved when the energy in the solution region is minimized [1]. Thus, the partial derivatives of the energy W with respect to the nodal potential are set to zero:

$$\frac{\partial W}{\partial V_1} = \frac{\partial W}{\partial V_2} = \frac{\partial W}{\partial V_3} = \frac{\partial W}{\partial V_4} = 0 \quad (2.17)$$

Substituting equation (2.13) into equation (2.17) for global node 1 and using symmetry gives

$$\begin{aligned} 0 = \frac{\partial W}{\partial V_1} &= 2V_1 C_{11} + V_2 C_{21} + V_3 C_{31} + V_4 C_{41} + V_2 C_{12} + V_3 C_{13} + V_4 C_{14} \\ &= V_1 C_{11} + V_2 C_{21} + V_3 C_{31} + V_4 C_{41} \end{aligned} \quad (2.18)$$

In general,

$$\frac{\partial W}{\partial V_k} = 0 = \sum_{i=1}^n V_i C_{ik} \quad (2.19)$$

where n is the number of nodes. By taking the partial derivative of W with respect to the other node voltages, a set of simultaneous equations is obtained to solve for the unknown voltages.

The next step involves applying the boundary conditions. For this example, voltages are defined on at least two nodes, and the other node voltages are solved for using these voltages and the set of simultaneous equations. One way of solving for the unknown voltages is the iteration method. First, equation (2.18) is rearranged:

$$V_1 = -\frac{1}{C_{11}} (V_2 C_{21} + V_3 C_{31} + V_4 C_{41}) = -\frac{1}{C_{11}} \sum_{i=2}^4 V_i C_{1i} \quad (2.20)$$

and, in general

$$V_k = -\frac{1}{C_{kk}} \sum_{i=1, i \neq k}^n V_i C_{ki} \quad (2.21)$$

The iteration process begins by assigning the unknown voltages equal to zero and applying equation (2.21). This process continues with the other unknown voltages until the difference in subsequent iterations is reduced to the desired error.

This example solved Laplace's equation to find the unknown node voltages of the 2D model. The electric field can then be calculated from the voltage:

$$\mathbf{E} = -\nabla V$$

Also, the magnetic field can be found from the electric field:

$$\nabla \times \mathbf{E} = -\frac{\partial \mathbf{B}}{\partial t}$$

EMPro uses the same analysis as above but applied to the electromagnetic wave equations in the frequency domain and 3D space. The unknowns in this case are the electric fields of each smaller piece. Initial conditions and boundary conditions are then applied, and the resulting system of equations is solved. From the electric field, the magnetic field can then be found as well.

2.2 Finite Difference Time Domain

EMPro also includes a finite difference time domain (FDTD) simulator. The basic principles behind using the FDTD method involve dividing a structure into box-shaped cells and alternating between finding the electric and magnetic fields of the cells through time [1].

The simulation structure is divided into cells as shown in Figure 2, also known as Yee cells. The edges are assigned electric field components and magnetic field components for the faces. The location and direction of the vectors are centered on the edges and faces. They reflect the finite difference approximation, which is covered next.

Figure 2: FDTD Yee Cell with E-Field & H-Field Directions from [6]

To solve for the time-varying electric and magnetic fields of a structure, Maxwell's curl equations are used:

$$\nabla \times \mathbf{E} = -\frac{\partial}{\partial t} \mathbf{B}$$

$$\nabla \times \mathbf{B} = \mu_0 \mathbf{J} + \mu_0 \epsilon_0 \frac{\partial}{\partial t} \mathbf{E}$$

Expanding the curl operator gives the following scalar differential equations:

$$\begin{aligned}
\frac{\partial H_x}{\partial t} &= -\frac{1}{\mu} \left(\frac{\partial E_z}{\partial y} - \frac{\partial E_y}{\partial z} \right) & \frac{\partial E_x}{\partial t} &= \frac{1}{\varepsilon} \left(\frac{\partial H_z}{\partial y} - \frac{\partial H_y}{\partial z} - \sigma E_x \right) \\
\frac{\partial H_y}{\partial t} &= -\frac{1}{\mu} \left(\frac{\partial E_x}{\partial z} - \frac{\partial E_z}{\partial x} \right) & \frac{\partial E_y}{\partial t} &= \frac{1}{\varepsilon} \left(\frac{\partial H_x}{\partial z} - \frac{\partial H_z}{\partial x} - \sigma E_y \right) \\
\frac{\partial H_z}{\partial t} &= -\frac{1}{\mu} \left(\frac{\partial E_y}{\partial x} - \frac{\partial E_x}{\partial y} \right) & \frac{\partial E_z}{\partial t} &= \frac{1}{\varepsilon} \left(\frac{\partial H_y}{\partial x} - \frac{\partial H_x}{\partial y} - \sigma E_z \right)
\end{aligned}$$

To use these equations in software, a finite difference approximation is applied to each partial derivative. During one timestep, the electric and magnetic fields are calculated in succession. For example, the first equation is approximated as

$$\frac{H_x^{n+0.5} - H_x^{n-0.5}}{\Delta t} = -\frac{1}{\mu} \left(\frac{E_{z,(i,j+1,k+0.5)}^n - E_{z,(i,j,k+0.5)}^n}{\Delta y} - \frac{E_{y,(i,j+0.5,k+1)}^n - E_{y,(i,j+0.5,k)}^n}{\Delta z} \right)$$

The other two time-differential magnetic field equations are approximated in the same manner. The time-differential electric field equations are similarly approximated.

For example, the fourth equation from before:

$$\begin{aligned}
\frac{E_x^{n+1} - E_x^n}{\Delta t} &= \frac{1}{\varepsilon} \left(\frac{H_{z,(i+0.5,j+0.5,k)}^{n+0.5} - H_{z,(i+0.5,j-0.5,k)}^{n+0.5}}{\Delta y} - \frac{H_{y,(i+0.5,j,k+0.5)}^{n+0.5} - H_{y,(i+0.5,j,k-0.5)}^{n+0.5}}{\Delta z} \right. \\
&\quad \left. - \sigma \frac{E_x^{n+1} - E_x^n}{2} \right)
\end{aligned}$$

The other equations are approximated in the same manner. The current timestep is represented by n . The field component directions are labeled as x , y , or z . The field component locations are labeled as i , j , and k depending where they lie on the cell.

The initial electric and magnetic fields are determined by the source voltage and current, where the source is aligned to at least one of the edges of the initial cell. Since the source extends between two points, line integrals are used to compute the electric and magnetic fields. The electric field is related to the gradient of the voltage:

$$\mathbf{E} = -\nabla V_E$$

Also, Ampere's law relates the magnetic field to its source current:

$$\int \mathbf{B} \cdot d\mathbf{l} = \mu_0 I_{enc}$$

The entire structure is divided into these Yee cells, side by side. After the initial edges and walls are solved for, their adjacent edges and walls are solved in the next timestep. This looping of the finite difference equations continues until the simulation has converged to a certain extent.

For the simulation to converge, the simulation needs to reach steady state. In EMPro, convergence criterion is controlled by the user, and a decibel ratio controls when the simulation stops. For example, if a convergence criterion of -30 dB is entered, the simulation will stop when there is a 3% difference from steady state ($-30 \text{ dB} = 20 \log 0.03$).

3. COMPARISON OF SIMULATION METHODS

EMPro includes two separate 3D EM simulators: one base on FEM and one base on FDTD. FEM simulates in the frequency domain and FDTD simulates in the time domain. The methods outlining their basic operation are described in the previous sections, and their advantages are discussed below [2].

3.1 Meshing

One of the main advantages of using EMPro's FEM simulator over its FDTD simulator involves its ability to generate more accurate meshes over curved structures. In FEM, tetrahedra are used to mesh the model; whereas, in FDTD, boxes are used. The differences in meshing are shown below in Figure 3 for a sphere. The FEM mesh provides a more accurate depiction of the sphere, as the FDTD version relies on cubes.

Figure 3: FDTD Gridding (Left) Vs. FEM Meshing (Right) from [7]

In 2011, EMPro released a conformal meshing option for its FDTD simulator. This option allows for curved structures to be modeled more accurately. Because of this, there is no distinguishable meshing/gridding advantage of one method over the other.

3.2 Coding

Programming a complete FEM simulator from scratch is much more difficult than programming an FDTD simulator. There are much more steps involved, as various matrices are generated; also, the entire simulation space is taken into account before the solution is found, which is evident from finding the global coefficient matrix before solving.

FDTD is easier to program from scratch, as the solution takes a more natural time-stepped progression throughout the simulation space. Simple difference equations are solved instead of more calculation-intensive matrix equations.

3.3 Solution Method

FEM is based on solving matrix equations. As the number of nodes increases for an FEM model, the number of unknowns increases. This generates very large matrices to be solved for.

FDTD is based on iterative time stepping. Since the unknowns are solved for directly from the difference equation form of Maxwell's equations, simulations are less memory intensive than FEM.

3.4 Frequency Domain vs. Time Domain

EMPro's FEM simulator solves Maxwell's equations at discrete frequencies, and frequency sweep plots are obtained through curve fitting; time-domain results can be obtained by using the Fourier transform.

FDTD uses a time-stepping approach. This allows for broadband results using a short time pulse as the source and applying a discrete Fourier transform; single frequency results can also be obtained by applying a sinusoidal source waveform.

3.5 Choosing FDTD

The decision to use EMPro's FDTD simulator over its FEM came down to FDTD's ability to produce broadband results when exact frequencies are not known; this was very useful when performing filter tuning. Also, since FDTD computes the E-field and H-fields directly through time, animated views of these fields are easily obtained throughout the model. Table 1 below summarizes the advantages and disadvantages of both methods.

	FEM	FDTD
Volumetric Sampling	<ul style="list-style-type: none"> ▪ Tetrahedron-based meshing. <i>Pros: Great for rounded objects.</i> <i>Cons: Memory intensive for very large objects (cars, planes, etc.)</i> 	<ul style="list-style-type: none"> ▪ Box-shaped gridding. <i>Pros: Great for very large objects (cars, planes, etc.).</i> <i>Cons: Not as accurate for rounded objects.</i>
Solution Method	<ul style="list-style-type: none"> ▪ Based on solving matrix equations. <i>Pros: Fields at any point in model can be obtained from interpolation.</i> <i>Cons: More memory intensive due to solving matrices.</i> 	<ul style="list-style-type: none"> ▪ Based on solving difference equations. <i>Pros: Less memory intensive due to solving simple algebra equations.</i> <i>Cons: Fields are only available at cell edges.</i>
Simulation Results	<ul style="list-style-type: none"> ▪ Adaptive frequency solution method. <i>Pros: Smoother S-parameter results as points are curve-fitted using adaptive sweep.</i> <i>Cons: Broadband results are very memory intensive.</i> 	<ul style="list-style-type: none"> ▪ Iterative time-stepping solution method. <i>Pros: Broadband results can be obtained due to time-based approach and FFT.</i> <i>Cons: S-parameter results are not curve-fitted.</i>

Table 1: Summary of FEM and FDTD Methods

In general, any structure can be simulated using either method. For high-Q RF circuits, FEM is recommended for more accurate results. It is more advantageous to use FDTD when simulating objects such as antennas, planes, and cars, as simulation times are decreased compared to FEM.

4. FILTER THEORY

All of the filters designed in this thesis are based on coaxial resonator theory, either half-wave or quarter-wave. Coaxial resonators were chosen for this project, as they are an extension of transmission line theory and easier to tune. To excite these resonators, two kinds of field coupling probes were used: magnetic and electric. As mentioned in the introduction, the filters were mainly chosen based on simplicity and cost. The two TEM filters include a trough shape, which is shown later in Figure 9 and Figure 10.

4.1 Half-Wave Resonator

The first filter investigated includes a half-wave resonator. Both sides of the resonator are shorted allowing standing waves to be supported [3]; these standing waves have a wavelength equal to twice the length of the resonator. The electric field amplitude is maximum at the center of the resonator and minimum at the shorted ends, while the magnetic field is the inverse. See Figure 4 below.

Figure 4: Half-Wave Resonator E-Field & H-Field Amplitude

These amplitudes are better described through lossless frequency-domain, plane wave solutions of the wave equations:

$$E_\rho(z) = E_\rho^+ e^{-j\beta z} + E_\rho^- e^{+j\beta z} = E_\rho^+ [e^{-j\beta z} + \Gamma e^{+j\beta z}]$$

Γ is the reflection coefficient of the short circuit walls. The equations below describe the distributions in Figure 4 and show how the electric field changes versus distance along the line:

$$|E_\rho(0)| = E_\rho^+ [e^{-j\beta 0} + (-1)e^{+j\beta 0}] = 0$$

$$\left| E_\rho \left(\frac{\lambda}{4} \right) \right| = E_\rho^+ [e^{-j\frac{\pi}{2}} + (-1)e^{+j\frac{\pi}{2}}] = 2E_\rho^+$$

$$\left| E_\rho \left(\frac{\lambda}{2} \right) \right| = E_\rho^+ [e^{-j\pi} + (-1)e^{+j\pi}] = 0$$

Also from Faraday's law,

$$H = \frac{\nabla \times E}{j\omega\mu}$$

and

$$H_\phi(z) = \frac{1}{\eta} (E_\rho^+ e^{-j\beta z} - E_\rho^- e^{+j\beta z}) = \frac{1}{\eta} E_\rho^+ [e^{-j\beta z} - \Gamma e^{+j\beta z}]$$

Also, for a coaxial resonator with outer radius 'b', and inner radius 'a',

$$E_\rho = \frac{j2E_0}{\rho \ln \frac{b}{a}} \sin \beta z$$

and

$$H_\phi = \frac{2E_0}{\eta \rho \ln \frac{b}{a}} \cos \beta z$$

From this relation, the magnitude of the magnetic field is minimum in the middle of the conductor and maximum at the shorted ends.

To relate voltage to the electric field and current to the magnetic field, a Gaussian cylinder surface is placed between the center and outer conductors. Gauss's law can then be applied at a cross section of the resonator to find the direction of the electric and magnetic fields. Gauss's law for the electric and magnetic fields can then be described as

$$\oint \epsilon_0 \vec{E} \cdot d\vec{S} = Q_{enclosed}$$

and

$$\oint \vec{H} \cdot d\vec{l} = I_{enclosed}$$

After solving for the coaxial case where the direction of propagation is along the transmission line, the electric field only points in the radial direction, normal to the center conductor and the outer conductor:

$$\vec{E} = \frac{\rho_l}{2\pi\epsilon_0\rho} \vec{a}_\rho$$

ρ_l is the charge per unit length, and ρ is the radius of the Gaussian surface. Similarly, the magnetic field circles the center conductor normal to the electric field lines:

$$\vec{H} = \frac{I_{enclosed}}{2\pi\rho} \vec{a}_\phi$$

See Figure 5 below.

Figure 5: E-Field & H-Field Lines in a Coaxial Transmission Line

4.2 Quarter-Wave Resonator

The quarter-wave resonator allows standing waves whose wavelength is equal to four times the length of the resonator. This is accomplished by shorting one end and opening the other [3]. By doing this, certain standing waves are supported whose electric and magnetic fields differ in magnitude at both ends. See Figure 6 below. For the field directions, see Figure 5 above.

Figure 6: Quarter-Wave Resonator E-Field & H-Field Amplitude Distribution in the Z-direction

Transmission line theory can then be used to verify the magnitude of the electric and magnetic fields along the resonator. This time, the reflection coefficient is equal to 1 since the right end of the resonator is an open circuit.

$$E_{\rho}(0) = \frac{j2E_0}{\rho \ln \frac{b}{a}} \sin \beta(0) = 0 = \min. \quad E_{\rho}\left(\frac{\lambda}{4}\right) = \frac{j2E_0}{\rho \ln \frac{b}{a}} \sin \frac{2\pi}{\lambda} \frac{\lambda}{4} = \frac{j2E_0}{\rho \ln \frac{b}{a}} = \max.$$

From these equations, the resonator resembles a quarter-wave transformer. Similarly, as from before, the current is inversely proportional to the voltage. The connection between voltage and the electric field can then be made; as well as current and the magnetic field.

4.3 Coaxial Resonator Q

The quality factor (Q) describes how narrow the bandwidth of a resonator is compared to its center frequency. For a half-wave coaxial resonator, it is described as

$$Q = \frac{2\pi \times (\text{Energy Stored})}{\text{Energy Dissipated Per Cycle}}$$

$$= \frac{\omega \epsilon \int_V |\mathbf{E}|^2 dV}{R_s \int_{center} |\mathbf{H}|^2 dS + R_s \int_{outer} |\mathbf{H}|^2 dS + R_s \int_{ends} |\mathbf{H}|^2 dS}$$

where only the conductor loss is taken into account, as air fills the space between the conductors. This can be simplified as

$$Q_{unloaded, \lambda/2} = \frac{\pi \eta \ln \frac{b}{a}}{R_s \left[l \left(\frac{1}{a} + \frac{1}{b} \right) + 4 \ln \frac{b}{a} \right]}$$

where R_s is the surface resistance:

$$R_s = \sqrt{\frac{\omega \mu_0}{2\sigma}}$$

‘b’ is the outer radius, ‘a’ is the inner radius, and η is the electromagnetic impedance taking into account the conductivity of the conductors. Similarly, for the quarter-wave coaxial resonator,

$$Q_{unloaded, \lambda/4} = \frac{2\sqrt{\pi f \mu \sigma} \ln \frac{b}{a}}{\frac{1}{a} + \frac{1}{b}}$$

Also, when an external circuit is coupled to the resonator, the power lost through this external circuit is taken into account. This is known as loaded Q and can be modeled as

$$Q_{loaded} = \frac{\omega_0 W_T}{P_C + P_D + P_R + P_{Ext}}$$

To find the coupling coefficient, first define

$$k = \frac{P_{Ext}}{P_C + P_D + P_R}$$

then

$$Q_{loaded} = \frac{Q_{unloaded}}{1 + k}$$

and

$$1 + k = \frac{Q_{unloaded}}{Q_{loaded}}$$

where k is the coupling coefficient. It is a measure of the power dissipated in the resonator to the power dissipated from the external circuit. Three types of coupling are obtained from this, where Q_{ext} pertains to the external circuit:

$$\textit{Critical Coupling} \rightarrow Q_{unloaded} = Q_{ext}$$

$$\textit{Over Coupling} \rightarrow Q_{unloaded} > Q_{ext}$$

$$\textit{Under Coupling} \rightarrow Q_{unloaded} < Q_{ext}$$

Visually, on a Smith chart, S11 measurements show if the coupling is one of the three mentioned above. If the center frequency passes through the center of the chart, the circuit is critically coupled; if it passes through a smaller resistance on the Smith chart, it's over-coupled; and if it passes through a larger resistance than the center, under-coupled. See Figure 7 on the following page.

Figure 7: S11 Smith Chart Coupling Cases

4.4 Circular Cylindrical Resonator Modes and Q

The theory presented in this section relates to the filter in chapter 6, as the hollow cap will produce various TE and TM modes. With the tuning screw, the behavior is more complicated and will be explained later.

The TM_{01} mode is considered, as electric field probes are used for the copper end cap filter; they are inserted into the shorted ends of the cylinder in the direction of propagation. The TM_{01} field directions are shown below in Figure 8.

Figure 8: E-Field & H-Field inside TM_{01} Mode Circular Cylindrical Resonator from [8]

The electric field lines begin and terminate on the shorting plates at the ends of the cylinder, and the current travels along the cylinder walls. The magnetic field lines circle around the inside of the cavity. The TM01 resonant frequency is obtained from [8]

$$f = \frac{c \times 2.405}{2\pi a}$$

where 2.405 is the TM01 coefficient determined through Bessel functions, 'a' is the cylinder radius, and 'c' is the speed of light.

The unloaded Q can be calculated as [8]

$$Q = \frac{\eta}{R_s} \frac{2.405}{2(\frac{a}{d} + 1)}$$

where R_s is the surface resistance, η is the electromagnetic impedance, 2.405 is the coefficient for the TM01 mode, 'a' is the radius, and 'd' is the depth of the cavity.

4.5 Magnetic Coupling

One way to excite the resonator involves coupling magnetic fields into it. To do this, a probe forms a loop between the center conductor and the outer conductor of the resonator, then shorting against the outer conductor. See Figure 9 below.

Figure 9: Magnetic Coupling Loop Field Lines & Resonator Field Lines

By shorting the coupling loop at the same location as the resonator short, maximum magnetic coupling occurs. See the H-field magnitude distribution from Figure 4.

From Ampere's law, as current changes along the coupling wire, it generates a magnetic field around the wire. This, in turn, induces a field around the center conductor that induces a current. This relation is better described through Faraday's law:

$$V = \int \mathbf{E} \cdot d\mathbf{l} = - \int \frac{\partial \mathbf{B}}{\partial t} \cdot d\mathbf{S}$$

A current is induced along the center conductor as the magnetic field curls around it.

4.6 Electric Field Coupling

Electric coupling involves inserting a probe parallel to the electric field of the resonator. The probe is located at the resonator location having a maximum electric field

distribution. For the case of the half-wave resonator, this is at the halfway point. See Figure 9 and refer to the E-field resonator distribution from the previous sections. Similarly, energy can be coupled out of the resonator using another E-field probe or an H-field probe. Locations of both depend on the location of respective field maximums [3].

Figure 10: Electric Coupling Field Lines & Resonator Field Lines

The current in the center conductor is induced, as the electric field lines of the probe and the resonator align. The strength of the E-field decreases along the line from the center, and this change in potential produces current.

5. HALF-WAVE TROUGH RESONATOR FILTER

This experiment builds on an existing experiment involving the characterization of a 1 GHz half-wave coaxial resonator filter. The existing project involves building and measuring the filter based on approximations. The changes presented here include adding a simulation aspect to the project using EMPro. The filter's S-parameters will be characterized versus tuning changes. This allows for insertion loss tuning, bandwidth tuning, and field visualization.

The filter was simulated using EMPro, and the bandwidth and insertion loss were optimized. After the desired response was obtained in simulation, the filter model's dimensions were used to build the filter and measure it using a network analyzer. Comparisons were made between measured and simulated results.

The filter is based on 1010 MHz oscillator design [4]. The oscillator circuit included three chambers: one for the resonator, one for the oscillator circuit, and one for the amplifier circuit. The resonator portion of the circuit simply involved a 6 inch piece of tubing being supported by two walls; one at either end. This is shown in Figure 11, where the 3/8" brass tubing is the resonator.

Figure 11: Half-Wave Resonator Motivation inside 1010 MHz Oscillator[4]

Since only the resonator portion of the oscillator was used in this experiment, some modifications to the structure can be made. The oscillator and amplifier chambers were not necessary for the experiment, as the resonator was the main concern. Also one of the probes was moved to the opposite side of the resonator to provide for better isolation.

5.1 Modeling

The structure was modeled in EMPro after an existing filter made by Dr. Derickson. This was done to verify measurement results of the existing filter with EMPro simulation results. This filter is shown below in Figure 12.

The dimensions of the box were measured and estimated as follows: length = 126 mm, width = 25 mm, and height = 43 mm. The center, brass conductor has a diameter of 3/8 inches and is centered at a height of 15 mm. The walls of the filter were modeled as copper ($\sigma = 5.8\text{e}+07$ S/m), the tubing as 70-30 brass ($\sigma = 1.57\text{e}+07$ S/m), and the coupling loop wires as silver ($\sigma = 6.17\text{e}+07$ S/m). The dimensions of the coupling loop were estimated to be 15 mm x 5 mm, and they were centered in height with the brass tubing. 0.085" diameter semi-rigid coaxial cable was used for the ports.

Figure 12: Existing 1.2 GHz Half-Wave Coaxial Filter Built by Dr. Derickson

The modeled filter is shown in Figure 13 below. The unnecessary lengths of the base PCB were eliminated, and the extra lengths of semi-rigid coax were shortened to the box walls. Making the structure smaller aids in decreasing the simulation workload.

Figure 13: Half-Wave 1 GHz Filter Modeled in EMPro

After the filter was modeled, a near field sensor was added to the simulation space to provide E-field and H-field visualization. A solid box sensor was chosen for the entire simulation space. From the sensor definition, E-fields and H-fields were visualized versus frequency.

Initially, a broadband time pulse was used to excite port 1 as shown in Figure 14 on the following page. This was chosen to obtain a wide frequency response from the simulation. After the center frequency of the filter was found, a sinusoidal waveform was then substituted and applied at that frequency. By completing a steady state simulation, field visualization can be obtained at the center frequency.

Figure 14: Half-Wave Filter Source Waveform (Broadband Pulse)

The gridding pattern used for the filter included a minimum cell size of 0.5 mm and maximum of 14 mm. Once an accurate depiction of the filter was gridded, conformal meshing was applied for increased simulation accuracy. The final gridded structure is shown in Figure 15 below.

Figure 15: Cell Edges of Half-Wave Filter Gridded in EMPro

A more detailed screenshot of the filter grid is shown in Figure 16. Depending on the size of the feature, smaller or larger cell sizes were used to increase simulation accuracy and decrease simulation time.

Figure 16: Half-Wave Filter Grid Details

5.2 Simulation

With the filter modeled, simulation setup parameters were defined. S-parameters at port 1 were selected. To obtain accurate simulation results, a convergence termination criterion was chosen. This number establishes when the simulation would end based on the error between recent simulation calculations and overall simulation calculations. For example, a -20 dB termination criterion is equivalent to a 10% error:

$$20 \log_{10} 0.1 = -20 \text{ dB}.$$

The default -30 dB convergence termination criteria was chosen, as it translates to a 3% error in simulation results; lower termination criteria would significantly increase simulation time but decrease error only slightly. The S21 results are shown in Figure 17 below. The center frequency was 1.189 GHz with a half-power bandwidth of 23 MHz. Insertion loss was 2.492 dB.

Figure 17: Half-Wave Filter Simulation Results using Approximate Dimensions from Filter in Figure 11. $f_c = 1.19$ GHz, BW = 23.2 MHz, Q = 51, Loss = 2.5 dB

To achieve less insertion loss and narrower bandwidth, the location and size of the coupling loops were varied in EMPro and simulated. As the coupling loop was moved closer to the center conductor, less insertion loss was observed. This is shown in Figure 18 below. The loop was simulated at distances of 0.5 mm (green trace) and 3.5 mm (blue trace) from the center conductor.

Figure 18: Insertion Loss Increase as Coupling Loop Distance from Center Conductor Increases; A = 0.5 mm, 3 dB Loss; B = 3.5 mm, 9.7 dB Loss

By adjusting the length of the coupling loop parallel to the center conductor, the filter's bandwidth changed. This is shown in Figure 19 below and Table 2. As the coupling loop length was increased, the bandwidth widened, and the insertion loss decreased. As the length was decreased, the bandwidth narrowed, but the insertion loss increased.

Figure 19: Bandwidth vs. Varying Coupling Loop Lengths

	Coupling Length	Center Frequency	Bandwidth	Insertion Loss
Light Blue (a)	5 mm	1.176 GHz	13.3 MHz	3.00 dB
Red (b)	10 mm	1.199 GHz	32.1 MHz	1.99 dB
Green (c)	15 mm	1.210 GHz	42.1 MHz	1.89 dB
Dark Blue (d)	20 mm	1.233 GHz	46.6 MHz	2.31 dB

Table 2: Bandwidth vs. Varying Coupling Loop Lengths

Also, to verify the field distribution along the resonator, plots were obtained from the near field sensor. H-field and E-field plots are shown in Figure 20.

(a) E-Field Magnitude along Resonator; Red indicates largest magnitude at the center of the resonator, while purple represents very little field intensity at the short circuit end. See Section 4.1 for Theory.

(b) H-Field Magnitude along Resonator; Red indicates largest magnitude at the ends of the resonator, while purple represents very little field intensity at the center. See Section 4.1 for Theory.

Figure 20: Half-Wave Filter Field Distributions

S11 simulations were also completed for the filter using the dimensions of the existing one in Figure 12. This is shown below in Figure 21. Due to the discrete number of frequency points and lack of curve-fitting, the exact shape of the impedance is hard to see.

Figure 21: Smith Chart Impedance vs. Frequency of the Existing Filter from Figure 11; Under-Coupled Case

5.3 Measurements

S-parameter measurements for the existing filter from Figure 12 were captured using a network analyzer. S21 and S11 results are shown below in Figure 22.

Figure 22: Network Analyzer Measurements of S21 & S11 for existing Half-Wave Filter from Figure 12

The center frequency was 1.213 GHz with a half-power bandwidth of 21 MHz. The insertion loss was 0.943 dB.

Table 3 below summarizes the differences between the measured and simulated data. The measured center frequency and bandwidth agreed with simulation, but the insertion loss differed by 1.5 dB. The difference in insertion loss is due to the lack of data points available in simulation as seen in Figure 17. There was no curve fitting available, therefore a line connects successive frequency points.

	Center Frequency	Half-power Bandwidth	Insertion Loss
Simulated	1.189 GHz	23 MHz	2.492 dB
Measured	1.213 GHz	21 MHz	0.943 dB

Table 3: Simulated vs. Measured Filter Data

S11 was also measured for the existing filter and is shown below in impedance form on a Smith chart. See Figure 23. Time domain settings were applied to set the reference plane to the walls of the resonator. There is a large discrepancy between the measured and simulated impedance from Figure 21. This is due to the discrete number of points available for FDTD simulations.

Figure 23: Network Analyzer Impedance Measurement for Half-Wave Filter from Figure 12; Over-Coupled Case

6. COPPER END CAP FILTER

The second filter investigated included the design and measurement of a 2 GHz copper end cap filter. The design was based on an article by Kent Britain that outlines several 4 GHz - 10 GHz cap filters [5]. They were built using 0.5" – 1" diameter copper end caps. Figure 24 below shows his 3456 MHz filter.

Figure 24: 1 Inch Copper Cap Filter with Tuning Screw from [5]

The center screw provides the tuning, and electric field coupling is used from the protruding probes. An empty cap cavity, without the tuning screw, would allow true waveguide modes to propagate. Since the electric field is maximum at the open end of the screw, coupling occurs between the fringing fields of the probes and the screw. The filter behaves as a circular waveguide resonator without the tuning screw and looks like a coaxial quarter wave resonator as the screw depth is increased. The lowest frequency of the filter is around 1.5 GHz with the screw about a millimeter from the base.

The center frequency is further reduced by the capacitance created by the probes and the outer conductor. Because of this, by varying the height and spacing of the probes, the center frequency is tuned much lower than that of the waveguide cavity.

6.1 Modeling

To model the filter, the dimensions of a 2 inch diameter NIBCO-brand copper cap were used: depth = 1.46", height = 2.26", and width = 2.26". The thickness of the copper cap walls was 1/8". For the tuning screw, a brass rod was used. The probes were modeled after 0.085" diameter semi-rigid coax, and the probe diameter was 0.5 mm. A 60 mm² PCB was used for the base. The initial tuning screw depth was set to 20 mm and the port probes were set to 10 mm. These values would be tuned in simulation. All objects were defined as copper, except the tuning screw was defined as 70-30 brass, and the probes as silver. The modeled filter is shown in Figure 25. All objects were made translucent to show the tuning screw and port probes.

Figure 25: Copper Cap Filter Model; Model Objects Were Made Translucent to See Tuning Screw and Port Probes

To grid the filter, the minimum cell size used was 0.16 mm, and the maximum was 4.5 mm. These cell sizes were used to account for the details of the smallest and largest details of the filter. The filter grid is shown in Figure 26.

Figure 26: 2 GHz Copper Cap Filter Grid; Minimum Cell Size = 0.16 mm, Maximum = 4.5 mm

6.2 Simulation

For the 2" cap, the center frequency of the hollow cavity was calculated to be

$$f = \frac{c \times 2.405}{2\pi a} = 4.5 \text{ GHz}$$

The simulated hollow cap gave a resonance of 4 GHz as shown in Figure 27 on the following page. Differences in the calculated and simulated frequencies include estimations on the dimensions of the cap.

Figure 27: Simulated Center Frequency of Hollow Copper Cap

Different tuning screw lengths were simulated to observe the behavior of the filter (see Figure 28): 20 mm (light blue, e), 25 mm (red, d), 27 mm (green, c), 28 mm (pink, b), and 30 mm (blue, a). The probes were centered and spaced 20 mm. The various center frequencies are outlined in Table 4.

Figure 28: Simulated Copper Cap Center Frequency vs. Various Tuning Depths

	Tune Depth	Center Frequency
Dark Blue (a)	30 mm	1.76 GHz
Pink (b)	28 mm	1.92 GHz
Green (c)	27 mm	2.05 GHz
Red (d)	25 mm	2.12 GHz
Light Blue (e)	20 mm	2.51 GHz

Table 4: Copper Cap Center Frequency vs. Various Tuning Depths

Since the 27 mm tune depth provided the most accurate 2 GHz response, further simulations were performed with this length fixed at 27.2 mm. First, the probe spacing was varied among 10 mm, 20 mm, 30 mm, and 40 mm. The results are shown in Figure 29 and Table 5. Loss versus bandwidth tradeoffs are shown. Also, the frequency varies greatly; this suggests that changing the probe spacing also changes the center frequency of the filter.

Figure 29: Simulated Copper Cap S21 vs. Varying Probe Spacing

	Probe Spacing	Center Frequency	Half-Power Bandwidth	Insertion Loss
Light Blue (a)	10 mm	1.76 GHz	53.6 MHz	1.15 dB
Red (b)	20 mm	1.93 GHz	22.2 MHz	3.37 dB
Green (c)	30 mm	1.97 GHz	11.6 MHz	8.00 dB
Dark Blue (d)	40 mm	1.98 GHz	5.9 MHz	16.72 dB

Table 5: Simulated Copper Cap S21 vs. Varying Probe Spacing

Next, the length of the probes was varied, while the spacing was fixed to 30 mm. This is shown in Figure 30 and Table 6. Again, there is a tradeoff between loss and bandwidth. Also, the center frequency changes.

Figure 30: Simulated Copper Cap S21 vs. Varying Probe Height

	Probe Height	Center Frequency	Half-Power Bandwidth	Loss
Dark Blue (a)	20 mm	1.91 GHz	55.2 MHz	1.19 dB
Green (b)	15 mm	1.95 GHz	26.6 MHz	3.17 dB
Red (c)	10 mm	2.09 GHz	11.8 MHz	7.14 dB
Light Blue (d)	5 mm	1.98 GHz	9 MHz	22.6 dB

Table 6: Simulated Copper Cap S21 vs. Varying Probe Height

From the previous simulations, the following design was chosen: tuning screw length = 26 mm, probe spacing = 20 mm, and probe height = 10 mm. S21 of this design is shown below in Figure 31. These dimensions were chosen based on minimizing loss and narrowing the bandwidth.

**Figure 31: Simulated Copper Cap S21 of Final Filter Design;
Center Frequency = 1.96 GHz; BW = 18.8 MHz; Q = 104; Loss = 1.4 dB**

To verify the quarter-wave resonator response, the electric and magnetic fields were visualized in EMPro as shown in Figure 32 on the following page. The electric field is shown at a phase of 0 degrees, while the magnetic field is shown at 90 degrees. This shows their maximum magnitudes. The expected quarter-wave resonator response is seen at the open circuit end of the tuning screw, maximum E-field magnitude.

- (a) E-Field Magnitude along Resonator; Red indicates largest magnitude at the open end of the quarter-wave tuning screw, while purple represents very little field intensity on the conductor walls. The resonator behaves more as a coaxial resonator than a waveguide cavity, even though the port probes are oriented in the wrong direction.

- (b) H-Field Magnitude along Resonator; Red indicates largest magnitude at the top of the resonator, while purple represents minimum magnitude at the conductor walls.

Figure 32: Copper Cap Field Visualization

S11 simulations were completed for the copper cap filter. Shown in Figure 33 below are both FDTD and FEM simulations of the impedance on a Smith chart. This was the only filter where an FEM simulation was able to finish. The FDTD simulation finished in 20 minutes, while the FEM simulation finished in 40 minutes. The response was slightly under-coupled, and discrepancies are due to the adaptive sweep feature of FEM.

Figure 33: Simulated Copper Cap Smith Chart Impedance using both FDTD (top) and FEM (bottom) Methods

6.3 Measurements

The final filter design was built and measured. The finished filter is shown below in Figure 34, and its S21 measurements are shown in Figure 35.

Figure 34: Fabricated Copper End Cap Filter

Figure 35: Network Analyzer Measurement of S21 & S11 of Copper Cap Filter

Table 7 compares the responses of the simulated filter and the measured filter. The center frequency and bandwidth of the simulated design agreed with the measured,

but there was about 1.4 dB more loss. Discrepancies in loss are due to the discrete number of frequencies in simulation. The increase in loss is also due to the amount of lossy solder used. There also may be cold solder joints surrounding the cap. The rim of the cap may also not be completely flat against the PCB, allowing currents to flow underneath the rim.

	Center Frequency	Half-Power Bandwidth	Insertion Loss
Simulated	1.96 GHz	18 MHz	1.402 dB
Measured	2.00 GHz	11 MHz	2.853 dB

Table 7: Copper Cap Comparison of Simulated and Measured Results

Additional measurements were captured with the tuning depth varied. The center frequency decreases the further the screw is inserted into the cap, but the insertion loss increases. As the tuning screw approaches the floor of the filter, it begins to behave as a coaxial quarter-wave resonator with one end shorted and one end opened. Around a tuning depth of 32 mm, the insertion loss increases dramatically, as the direction of the electric field does not align with the electric field probes. This is shown in Figure 36 on the following page.

Figure 36: Copper Cap Center Frequency & Insertion Loss vs. Tuning Depth

Smith chart impedance was measured on the network analyzer as shown below in Figure 37. This agreed with the simulated results from Figure 33.

Figure 37: Network Analyzer Impedance Measurement for Copper Cap Filter; Slightly Over-Coupled

7. QUARTER-WAVE TROUGH RESONATOR FILTER

The third filter investigated involved the design of a 1 GHz quarter-wave coaxial resonator filter. The quarter-wave design was based on the article for the 1010 MHz oscillator and is a modification of the half-wave design outlined in Chapter 2.

The design was very similar to that of the half-wave filter. The built filter is shown below in Figure 38. Since the H-field is strongest at the shorted ends of the resonator, both coupling loops were placed at that end, but on opposite sides of the center conductor to provide for better isolation. To achieve a tunable response, the center conductor telescopes along the box cavity. Further details are explained in the next section.

Figure 38: Built Quarter Wave Filter

7.1 Modeling

The modeled quarter wave resonator filter is shown below in Figure 39. The box's dimensions are as follows: length = 90 mm, width = 40 mm, and height = 40 mm. The length dimension was chosen to allow for ample space for the brass tubing to be tuned. The width and height dimensions were chosen to allow for easier access in soldering the coupling loop wires. The initial loop size was set to 20 mm length-wise and 13.5 mm from the side walls. The same materials properties were used from the half-wave filter.

Figure 39: Quarter Wave Resonator Filter Model

After the structure was modeled, gridding was applied to the filter. The smallest cell size was 0.5 mm, and the largest was 14 mm. The cell sizes were varied based on the

size of the model feature. This also reduces the simulation time. After the initial grid was defined, conformal meshing was added to the structure. This provides more accurate results during simulation. The filter grid is shown below in Figure 40.

Figure 40: Quarter Wave Resonator Filter Grid

A broadband time pulse was used to excite the filter at port 1. This allows for a wide range of frequencies to be simulated in a single simulation. The pulse waveform is shown below in Figure 41.

Figure 41: Quarter Wave Filter Excitation Waveform

7.2 Simulation

After the filter was modeled, simulations were performed. The convergence criterion for all simulations was -30 dB. By changing the location and size of the coupling loops, different responses were obtained. Insertion loss was minimized by keeping the coupling wire as close as possible to the center conductor. Next, the coupling distance was held constant while varying the length of the coupling wire. This is shown in Figure 42 below and Table 8.

Figure 42: Quarter-Wave Filter S21 vs. Parallel Coupling Length

	Coupling Length	Center Frequency	Half-Power Bandwidth	Insertion Loss
Dashed Blue (a)	5 mm	0.88 GHz	26.55 MHz	7.2 dB
Pink (b)	10 mm	0.89 GHz	28.5 MHz	2.9 dB
Light Blue (c)	15 mm	0.91 GHz	33.7 MHz	2.2 dB
Dark Blue (d)	20 mm	0.93 GHz	40.6 MHz	2.1 dB
Red (e)	25 mm	0.94 GHz	42.2 MHz	2.1 dB
Green (f)	30 mm	0.95 GHz	44.6 MHz	2.6 dB

Table 8: Quarter-Wave Filter S21 vs. Parallel Coupling Length

From the previous coupling length simulation, 20 mm was chosen based on insertion loss and bandwidth. The simulation results are shown below in Figure 43. The center tune length was chosen after measuring the filter on the network analyzer. EMPro was used more as a verification method since the optimization process involves many individual simulations.

Figure 43: Simulated S21 & S11 of Final Quarter-Wave Filter Design; Center Frequency = 0.928 GHz; BW = 40 MHz; Q = 23; Loss = 2.1 dB

Again, E-field and H-field visualization is shown for the quarter-wave resonator in Figure 44. The expected E-field maximum at the open circuit end is seen. Also the expected H-field minimum at the open circuit end is seen.

- a) **E-Field Magnitude along Resonator; Red indicates largest magnitude at the open end of the quarter-wave resonator, while purple represents very little field intensity at the short circuit end.**

- a) **H-Field Magnitude along Resonator; Red indicates largest magnitude at the short circuit end of the quarter-wave resonator, while purple represents very little field intensity at the open circuit end.**

Figure 44: Quarter-Wave Filter Field Visualization

Smith chart impedance was simulated as shown in Figure 45 below. Again, the number of discrete frequencies around resonance is scarce. It is shown that the filter is under-coupled.

Figure 45: Simulated Smith Chart Impedance of Quarter-Wave Coaxial Filter

7.3 Measurements

A quarter wave resonator filter was built using fairly large coupling loops. The coupling distance was 13.5 mm, and the coupling length was 20 mm. These numbers were chosen based on insertion loss and bandwidth. From simulation, the insertion loss at coupling lengths less than 20 mm was lossy. Measurement results are shown below in Figure 46. The center frequency was tuned to 1 GHz, and the bandwidth was 50 MHz. Insertion loss was 0.759 dB. This tuned length was measured and simulated in EMPro.

Figure 46: Measured S21 for Quarter-Wave Filter; Tuning Length = 76 mm, Probe Coupling Length = 20 mm, Probe Coupling Spacing = 1 mm

Additional measurements were captured at various tuning distances. This is shown below in Figure 47 and Table 9. When the tuning rod makes contact with the other wall of the trough, the filter behaves as a half-wave resonator.

Figure 47: Measured Quarter-Wave Filter Center Frequency vs. Tuning Distance (at 90 mm, tuning rod contacts and shorts at the other wall of the filter); the tuning distance is defined as the length of the brass tubing inside the copper PC board cavity

Tuning Distance (mm)	Frequency (MHz)	Wavelength (mm)
40	1950	153.739722
50	1514	198.013513
60	1275	235.13134
70	1090	275.038952
80	944	317.576756
85	878	341.449269
87	808	371.03027
89	750	399.723277
89.5	709	422.838446
89.6	655	457.698409
89.7	615	487.467411
89.8	531	564.5809
89.9	434	690.766032
90	1787	167.762987

Table 9: Measured Quarter-Wave Filter Center Frequency vs. Tuning Distance

Next, the dimensions of the coupling loop were changed. Instead of the 20 mm coupling length of the loop, the length was shortened to 13 mm, creating a smaller loop and keeping the distance from the center conductor constant. The S21 response is shown below in Figure 48. With the shorter coupling distance, the bandwidth of the filter expectedly narrows from 53 MHz (at 20 mm) to 38 MHz (at 13 mm). As the simulation from Figure 42 showed, increasing the loop size increases the bandwidth. The bandwidth ratio in simulation from the 13 mm tuning length to 20 mm was $31.5 \text{ MHz}/40.6 \text{ MHz}=0.77$; the measured ratio was $38 \text{ MHz}/53 \text{ MHz}=0.72$, showing reasonable agreement.

Figure 48: Measured S21 for Quarter-Wave Filter; Tuning Length = 76 mm, Probe Coupling Length = 13 mm, Probe Coupling Spacing = 1 mm

Measured Smith chart impedance from the network analyzer is shown below in Figure 49. For this case, the filter is slightly under-coupled. These impedance measurements agreed with the simulated results from Figure 44.

Figure 49: Measured Smith Chart Impedance of Quarter-Wave Coaxial Filter; Slightly Under-Coupled Case

7.4 Implementation into EE 455

This quarter-wave coaxial filter experiment was implemented into Cal Poly's EE 455 course during spring 2012. The span of the experiment was intended for 3 weekly meetings, but due to a holiday during the last week, the experiment was cut short to 2 meetings. Because of this, one meeting was focused on using the software, while the other meeting was focused on building the filters.

For the first meeting, a brief EMPro overview was provided, and the students completed a tutorial that had them simulate the quarter-wave filter. The tutorial can be found in the appendix and includes modeling one of the walls of the filter, adding input and output port definitions, assigning materials to the filter, gridding the filter, and running an FDTD simulation to obtain the S-parameters. Also, a separate simulation was included that involved simulating the electric and magnetic near field at the filter's center frequency. This allows for viewing of the E-field or H-field magnitude distributions along the resonator.

For the second meeting, the students built and measured the filters. Because of time constraints, the design aspect of the experiment was replaced with a verification procedure. The students varied the location and size of their coupling loops, measured the filters, and verified their designs in EMPro. Through this, they were able to gain experience using the software, although not in a design sense.

It is recommended to have one meeting focus on how to use the software, a second on the actual filter design, and the third on building and measuring the filter.

7.5 Feedback from EE 455 during Spring 2012

Many students mentioned that a more basic tutorial be used, one that includes basic project creation on EMPro startup. It is recommended that an existing Agilent tutorial be used for this purpose. A microstrip low-pass filter tutorial can be found online at <http://edownload.soco.agilent.com/eedl/empro/2010/pdf/emprosim.pdf> that includes instructions from project creation through simulation.

Also, many students mentioned that the tutorial used during spring 2012 needs to include more commentary on why they were completing certain aspects of the tutorial. A revised version of the tutorial with more commentary is included in the appendices.

There was some positive feedback as well; one student used the 1 GHz filter in his senior project. He was trying to build the filter on some PCB with inductors and capacitors, much like how the rest of EE 455 is taught; but then remembered the tunable filter he built for EE 455 and tuned it to the different needed frequencies. They were working on a project related to the jamming of spread spectrum communications.

8. CONCLUSION & FUTURE WORK

Three filters have been investigated as a starting point to introducing 3D EM simulation software into Cal Poly's curriculum. The designs of these filters were meant to be more educational rather than refined, and the quarter-wave coaxial filter was implemented into EE 455 during spring 2012. Feedback from the course is discussed in section 7.5.

From the three filters that were investigated, it is recommended to use the quarter-wave coaxial resonator filter for future EE 455 courses. Implementation of the quarter-wave filter includes several advantages over the other two designs that were discussed. In comparison to the half-wave filter, the quarter-wave filter is essentially half the size. This leads to cheaper construction costs and faster simulation times. In comparison with the copper cap filter, again the quarter-wave filter is cheaper to build, as the price of the cap was rather expensive.

It is recommended to spread this lab over a span of 3 weeks, which includes 3 weekly lab meetings. The first lab meeting would include a brief overview of what EMPro does and its capabilities. Next, a microstrip EMPro tutorial, as mentioned in section 7.5, would be completed.

For the second lab meeting, the quarter-wave coaxial filter tutorial, found in Appendix A, would be completed. The students would use the completed tutorial and change the dimension parameters to find the critically coupled response of the filter. S11 Smith chart plots would be completed for verification.

The third lab meeting would involve the actual building of the filter and measuring it using the network analyzers. The critically coupled simulation S11 response would be compared to the measured one.

8.1 Future Work

There are many additions that could be made to the work presented in this thesis. Instead of using one type of coupling into the resonators, both electric and magnetic coupling could be used. Comparisons could then be made about the effects of both coupling methods.

Also, since the filters discussed in this thesis used TEM modes, it is suggested that waveguide filters be implemented. Simple rectangular cavities could be constructed from the same PCB used in this thesis. Electric or magnetic coupling could then be used to excite the various modes of the cavity. EMPro would be used to visualize the E-field and the H-field inside the cavity. By viewing these fields, the TE and TM modes can be verified. Also, tuning posts and walls could be added to the filter to improve the response.

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APPENDIX A: EMPRO TUTORIAL FOR EE 455 SPRING 2012

The tutorial below was used in EE 455 during spring 2012.

EE 455 EMPro Tutorial: 1 GHz Bandpass Filter

Bobby Pheng
Cal Poly, Spring 2012

Opening the EMPro tutorial

1. Startup EMPro, either from the desktop or the start menus; if there is a licensing error, specify the following licensing server: 27010@129.65.26.217
2. Download the project from “L:\EE-455\ee_455_tutorial.zip”
3. Unzip the “L:\EE-455\ee_455_tutorial.ep” project folder to the desktop, and open it in EMPro.

Creating the Missing Geometry

In EMPro, models can be built using the included modeling environment, or they can be imported from other programs: SolidWorks, AutoCAD, etc. In this section, most of the filter has already been built, and only one wall will be modeled using pre-loaded parameters.

1. In the geometry workspace window, select *Create > Geometry > Extrude*. See Figure 1.

Figure 1: Creating an Extrude Geometry

- Under the “Specify Orientation” tab, select *Presets > YZ Plane*. For the X-dimension of the origin, input 40 mm. See Figure 2.

Figure 2: Specify Geometry Orientation

- Under the “Edit Cross Section” tab, select the “Rectangle” tool. The outline of the wall can be drawn freely, or coordinates can be defined. To define coordinates, right-click anywhere on the workspace (this selects the workspace as the active region and not the Edit Cross Section Tab). Press the “Tab” key, and enter “0 mm” for both dimensions. See Figure 3.

Figure 3: Entering Starting Coordinates

- Press the “Tab” key again, and enter “box_length” and “box_height” for the coordinates. See Figure 4. As mentioned before, these values are defined in the “Parameters” tab, which is located to the far right of the program window.

Figure 4: Entering Ending Coordinates

5. Under the “Edit Cross Section” tab, select the “Circle Center, Radius” tool. Press the “Tab” key, and enter “probe_spacing_y” and “probe_spacing_z” for the coordinates. See Figure 5a. Similarly, press “Tab,” and enter “probe_outer_radius” for the radius. See Figure 5b.

(a)

(b)

Figure 5: Circle Coordinates

6. Under the “Extrude” tab, enter “pcb_thickness” for the Extrude Distance. Also make sure the extrude direction coordinates are 0, 0, and 1 respectively. See Figure 6. Press the “Done” button.

Figure 6: Extrude Options

7. “Model” appears in the project parts tree, which is located to the far left of the program window. Right-click it and rename it as “front_wall”.

Adding Ports

Now that the entire model of the filter has been constructed, the input and output ports will be defined. These will be used for S-parameter simulation later. A simple 1V, 50 ohm feed will be used to excite both ports.

1. In the project parts tree, right-click “Circuit Components/Ports,” and choose “New Circuit Component with” > “New Feed Definition.”

2. Zoom into the front face of the probe. See Figure 7. Using the arrow tool under “Endpoint 1,” hover over the face of the probe, and press the “c” key to center the port, then left-click. Enter the following coordinates for endpoint 2:

X: box_width + pcb_thickness

Y: probe_spacing_y

Z: probe_spacing_z - probe_outer_radius

See Figure 7 on the following page . Endpoint 1 should read: Center Of Face, and Endpoint 2: Parameterized Position.

Figure 7: Port 1 Coordinates

3. Under the “Properties” tab, name the component “port1”, and press done.
4. Similarly, create another port on the other side of the filter; create the component as in step 1, but select “No Definition” instead of “New Feed Definition.” Using the same method in step 2, select the Center of Face for Endpoint 1. Enter the Endpoint 2 coordinates (See Figure 8):

X: -pcb_thickness
Y: probe_spacing_y
Z: probe_spacing_z - probe_outer_radius

Figure 8: Port 2 Coordinates

5. Under the “Properties” tab, name the component, “port2”. For the Component Definition, select “50 ohm Voltage Source,” and press done.

Adding Definitions

Now that the ports have been defined, some project definitions will be added. Most importantly, the different materials used in the project are defined. Next, the source waveforms are defined. A simple time pulse will provide a broad frequency response.

1. Right-click “Materials” under the “Definitions” branch of the project tree. Choose “Select from Default Material Library.” Choose “Copper” and press “Add.” Also add 70-30 brass and silver.
2. Right-click on “Parts” in the project tree, and select “View Parts List (All Parts).” Select the 11 extrude objects and assign copper to them. See Figure 9. Assign silver to the “normal” and “parallel” objects. For the center and outer conductor, assign 70-30 brass.

Figure 9: Assigning Material

- Under the “Definitions” branch of the project tree, expand “Waveforms” and right-click “Automatic.” Select “Properties.” Under type, select “Broadband.” Press done.

Running a Simulation

To setup an FDTD simulation, first the model needs to be gridded. Small cell sizes are used for the smaller features of the model, while larger ones are used for larger features. Also, conformal FDTD meshing is applied to the entire model to provide more accurate results for rounded objects.

- Under “Simulation Domain” of the project tree, right-click “FDTD Grid,” and select “Properties.” Under the “Size” tab, choose “Advanced,” and edit the fields as shown in Figure 10. The base cell sizes should be “0.1*wavelength(maxFreq).” Press done.

Figure 10: FDTD Gridding

- Open the parts list, and right-click “base.” Open up the “FDTD Gridding Properties.” See Figure 11. Under “Fixed Points,” make sure that the “Use Automatic Fixed Points” checkbox is checked. Press “Apply,” then press “Copy to Clipboard.” Press done.

Figure 11: Setting FDTD Gridding Properties

- Select all 11 items in the parts list. Right-click them, and select “Edit > Paste”.

- Edit the FDTD Gridding properties of “parallel_1” as shown in Figure 12 below. The target cell sizes should be “0.01*wavelength(maxFreq)” and the “Y” checkbox should be unchecked. Click “Apply” then “Copy to clipboard” and “Done.” Back in the parts list, right-click “parallel_2” and select edit > paste to copy the same settings from “parallel_1.”

Figure 12: Automatic Grid Regions for Parallel Probe Components

5. Edit the FDTD gridding properties of “normal_1” and “normal_2” in the same fashion. See Figure 13. This time, the “X” box is unchecked. Apply the settings, copy them to the clipboard, and paste them for “normal_2” as in step 4.

Figure 13: Automatic Grid Regions for Normal Probe Components

6. Right-click all 11 objects, and enable conformal meshing. See Figure 14.

Figure 14: Conformal Meshing

7. Click the wrench icon near the top of the window. Make sure that “Port 1” is selected under the “Setup S-Parameters” tab. Under the “Specify Termination Criteria” tab edit the “Maximum Simulation Time” to “200000*timestep.” Press “Create & Queue Simulation.” See Figure 15.

This maximum simulation time is chosen to provide ample simulation time to reach the -30 dB convergence criterion.

Figure 15: FDTD Simulation Settings

Viewing the Results & Changing the Design

1. Click the “Results” tab near the far right of the program window. To view the S-parameters, select the desired simulation, sensor, domain, and result type. Double-clicking on a result will bring up the default view, while right-clicking will provide other viewing options. See Figure 16.

Figure 16: S21 Magnitude Plot

2. This simulation was completed for a specific coupling loop size. Various parameters of the project can be changed from the “Parameters” tab. Parameters of interest include: center_tune_length, probe_spacing_x, probe_spacing_y, and probe_spacing_z. These parameters control the length of the center tune, which controls the filter’s center frequency, and the probe spacing, which controls the size and location of the coupling loops.

After the parameters have been changed, separate simulations can be completed. Alternatively, there is also a parameter sweep tool in the simulation setup window.

Field Visualization

1. In the project tree, under “Sensors,” right-click “Near Field Sensors.” Select “New Solid Box Sensor.” Under the properties tab, make sure “New Solid Sensor Definition” is selected for Sensor Definition. Press Done.
2. In the project tree, under “Definitions > Sensor Data Definitions,” right-click “New Solid Sensor Definition.” Name it “Sensor1.” Check the Steady E and Steady H boxes. Press done. See Figure 17.

Figure 17: Adding a Sensor Definition

- Right-click on “Waveforms” under “Definitions” in the project tree, and add a new definition. Name it “field.” Choose “Sinusoid” for type. Enter “1 GHz” for the frequency. Press done. See Figure 18.

Figure 18: Adding a Waveform

- Edit the properties of “50 ohm Voltage Source” in the Definitions tree. For “Waveform,” choose “field” instead of “Automatic.” Press done.
- Click the simulation wrench icon. Under “Frequencies of Interest,” check the “Collect Steady-State Data” box and “Use Waveform Frequency.” Press “Create & Queue Simulation.”
- Under the “Results” tab, select the latest simulation under “Project: Simulation.” Choose “H-Field” under “Result Type,” then view the plot.

- Adjust the parameters to view specific planes. XZ-plane H-field is shown in Figure 19.

Figure 19: Field Visualization

APPENDIX B: LECTURE MATERIALS

3D Electromagnetic Simulation

Classical 3D EM Analysis

- Example:
Electric Field
around infinitely
long charged wire

- Solution:
Create Gaussian
Surface & apply
Gauss's law to find E

Classical 3D EM Analysis

- Example:
Dipole Antenna
Radiation Pattern

- Solution: Approximate
arm dimensions of
antenna & apply
Maxwell's equations

Classical 3D EM Analysis

- Problems involve simple, symmetric shapes.
- Assumptions are made to simplify analysis.
- More complex problems are subdivided into shapes whose solutions are known, and boundary conditions are applied where the subdivided shapes meet.
- Electric and/or magnetic fields can then be found by applying Maxwell's equations.

What if the problem looked like this?

- How would you find the electric/magnetic fields reflected from the aircraft?
- Solution: subdivide problem into solvable shapes and apply boundary conditions.
- Impractical to do by hand.
- Practical with the help of computers.

Computational Electromagnetics (CEM)

$$\begin{aligned} \nabla \cdot \vec{E} &= \frac{\rho}{\epsilon_0} & \oint \vec{E} \cdot d\vec{l} &= \frac{q}{\epsilon_0} \\ \nabla \times \vec{E} &= -\frac{\partial \vec{B}}{\partial t} & \oint \vec{B} \cdot d\vec{l} &= 0 \\ \nabla \times \vec{B} &= \mu_0 \vec{J} + \frac{1}{c^2} \frac{\partial \vec{E}}{\partial t} & \oint \vec{E} \cdot d\vec{l} &= -\mu_0 \frac{\partial q}{\partial t} \\ \nabla \cdot \vec{B} &= 0 & \oint \vec{B} \cdot d\vec{l} &= \mu_0 \int_V \vec{J} \cdot d\vec{l} \end{aligned}$$

- Subdivide problem into suitable shapes (volumetric Sampling).
- Approximate the fields on each shape using Maxwell's equations.
 - Integral or differential solvers.
- Combine solutions of each shape for a complete solution.

Classification of CEM Methods

- CEM methods are labeled based on how they solve Maxwell's equations.
- Integral Solvers:
 - Method of Moments (Agilent ADS)
- Differential Solvers:
 - Finite Element Method (Agilent ADS & EMPro)
 - Finite Difference Time Domain (Agilent EMPro)

Method of Moments (MoM)

- 2.5D EM simulator used for planar structures such as microstrip.
- Substrate and ground plane are assumed to be infinite.
- Integral-based solver.

Finite Element Method (FEM)

$$\begin{bmatrix}
 P_{11} & P_{12} & \dots & P_{1N} \\
 P_{21} & P_{22} & \dots & P_{2N} \\
 \vdots & \vdots & \ddots & \vdots \\
 P_{N1} & P_{N2} & \dots & P_{NN}
 \end{bmatrix}
 \begin{bmatrix}
 V_1 \\
 V_2 \\
 \vdots \\
 V_N
 \end{bmatrix}
 =
 \begin{bmatrix}
 Q_1 \\
 Q_2 \\
 \vdots \\
 Q_N
 \end{bmatrix}$$

- Full wave 3D EM solver for arbitrary 3D structures.
- Model is subdivided into polyhedrons; most often, tetrahedrons.
- Matrices are created relating the polyhedrons to each other.
- These matrices are then solved to find the electric field at each polyhedron, then combined for the whole model.
- Solutions are found at discrete frequencies.

FEM Electrostatic Example

- Find voltages along the quadrilateral.
- Start with Laplace's equation for voltage: $\nabla^2 V = 0$
- One solution to this equation involves minimizing the energy in the quadrilateral.
- To find the energy, voltages need to be defined.

FEM Electrostatic Example

- Divide the quadrilateral into triangles.
- Label global nodes (red), local nodes (blue), and elements (black)

FEM Electrostatic Example

- The voltage in each element can be determined
- $V_{\text{element}} = \frac{1}{4} (V_1 + V_2 + V_3 + V_4)$
- Build stiffness matrix defined as $K_{ij} = \int_{\Omega} \nabla \phi_i \cdot \nabla \phi_j d\Omega$
- $\begin{bmatrix} K_{11} & K_{12} \\ K_{21} & K_{22} \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} Q_1 \\ Q_2 \end{bmatrix}$
- Solving for $\begin{bmatrix} V_1 \\ V_2 \end{bmatrix}$ and substituting back into equation gives
- $V_{\text{element}} = \frac{1}{4} \begin{bmatrix} \phi_1 & \phi_2 & \phi_3 & \phi_4 \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \\ V_3 \\ V_4 \end{bmatrix}$
- The negative values (charge) voltages in the model voltage and it is the basis for solving Laplace's equation using finite elements.

FEM Electrostatic Example

- $\psi_{\Omega}(x,y) = \frac{1}{4} \begin{bmatrix} x_1^2 & x_1 y_1 & x_1^2 y_1 & x_1^2 y_1^2 \\ x_2^2 & x_2 y_2 & x_2^2 y_2 & x_2^2 y_2^2 \\ x_3^2 & x_3 y_3 & x_3^2 y_3 & x_3^2 y_3^2 \\ x_4^2 & x_4 y_4 & x_4^2 y_4 & x_4^2 y_4^2 \end{bmatrix} \begin{bmatrix} \psi_1 \\ \psi_2 \\ \psi_3 \\ \psi_4 \end{bmatrix}$
- Next, define:

$$a_{ij} = \frac{1}{4} \int_{\Omega} (\nabla \psi_i \cdot \nabla \psi_j) = \frac{1}{4} \int_{\Omega} (\nabla \psi_i \cdot \nabla \psi_j) dx dy$$

$$a_{ij} = \frac{1}{4} \int_{\Omega} (\nabla \psi_i \cdot \nabla \psi_j) = \frac{1}{4} \int_{\Omega} (\nabla \psi_i \cdot \nabla \psi_j) dx dy$$
- There are three degrees of freedom. They represent the three nodal values.
- Now, $\psi_{\Omega}(x,y) = \frac{1}{4} \begin{bmatrix} x_1^2 & x_1 y_1 & x_1^2 y_1 & x_1^2 y_1^2 \\ x_2^2 & x_2 y_2 & x_2^2 y_2 & x_2^2 y_2^2 \\ x_3^2 & x_3 y_3 & x_3^2 y_3 & x_3^2 y_3^2 \\ x_4^2 & x_4 y_4 & x_4^2 y_4 & x_4^2 y_4^2 \end{bmatrix} \begin{bmatrix} \psi_1 \\ \psi_2 \\ \psi_3 \\ \psi_4 \end{bmatrix}$

FEM Electrostatic Example

- Next, the element energy functional can be defined as:

$$W_{\Omega} = \frac{1}{2} \int_{\Omega} \epsilon \nabla \psi \cdot \nabla \psi dx dy = \frac{1}{2} \int_{\Omega} \epsilon \nabla \psi \cdot \nabla \psi dx dy$$
- Substituting the expression for element voltage into this gives:

$$W_{\Omega} = \frac{1}{2} \int_{\Omega} \epsilon \nabla \left(\frac{1}{4} \begin{bmatrix} x_1^2 & x_1 y_1 & x_1^2 y_1 & x_1^2 y_1^2 \\ x_2^2 & x_2 y_2 & x_2^2 y_2 & x_2^2 y_2^2 \\ x_3^2 & x_3 y_3 & x_3^2 y_3 & x_3^2 y_3^2 \\ x_4^2 & x_4 y_4 & x_4^2 y_4 & x_4^2 y_4^2 \end{bmatrix} \begin{bmatrix} \psi_1 \\ \psi_2 \\ \psi_3 \\ \psi_4 \end{bmatrix} \right) \cdot \nabla \left(\frac{1}{4} \begin{bmatrix} x_1^2 & x_1 y_1 & x_1^2 y_1 & x_1^2 y_1^2 \\ x_2^2 & x_2 y_2 & x_2^2 y_2 & x_2^2 y_2^2 \\ x_3^2 & x_3 y_3 & x_3^2 y_3 & x_3^2 y_3^2 \\ x_4^2 & x_4 y_4 & x_4^2 y_4 & x_4^2 y_4^2 \end{bmatrix} \begin{bmatrix} \psi_1 \\ \psi_2 \\ \psi_3 \\ \psi_4 \end{bmatrix} \right) dx dy$$
- And $\nabla \psi = \frac{1}{4} \begin{bmatrix} 2x_1 & y_1 & 2x_1 y_1 & 2x_1^2 y_1 \\ 2x_2 & y_2 & 2x_2 y_2 & 2x_2^2 y_2 \\ 2x_3 & y_3 & 2x_3 y_3 & 2x_3^2 y_3 \\ 2x_4 & y_4 & 2x_4 y_4 & 2x_4^2 y_4 \end{bmatrix} \begin{bmatrix} \psi_1 \\ \psi_2 \\ \psi_3 \\ \psi_4 \end{bmatrix}$
- $\nabla \psi$ is known at the element level, so we can write it as:

$$\nabla \psi = \frac{1}{4} \begin{bmatrix} 2x_1 & y_1 & 2x_1 y_1 & 2x_1^2 y_1 \\ 2x_2 & y_2 & 2x_2 y_2 & 2x_2^2 y_2 \\ 2x_3 & y_3 & 2x_3 y_3 & 2x_3^2 y_3 \\ 2x_4 & y_4 & 2x_4 y_4 & 2x_4^2 y_4 \end{bmatrix} \begin{bmatrix} \psi_1 \\ \psi_2 \\ \psi_3 \\ \psi_4 \end{bmatrix}$$
- Rewriting gives:

$$W_{\Omega} = \frac{1}{2} \begin{bmatrix} \psi_1 & \psi_2 & \psi_3 & \psi_4 \end{bmatrix} \begin{bmatrix} a_{11} & a_{12} & a_{13} & a_{14} \\ a_{21} & a_{22} & a_{23} & a_{24} \\ a_{31} & a_{32} & a_{33} & a_{34} \\ a_{41} & a_{42} & a_{43} & a_{44} \end{bmatrix} \begin{bmatrix} \psi_1 \\ \psi_2 \\ \psi_3 \\ \psi_4 \end{bmatrix}$$

FEM Electrostatic Example

- After a single element has been considered, all elements are assembled:

$$W = \sum_{\Omega} W_{\Omega} = \frac{1}{2} \begin{bmatrix} \psi_1 & \psi_2 & \psi_3 & \psi_4 \end{bmatrix} \begin{bmatrix} a_{11} & a_{12} & a_{13} & a_{14} \\ a_{21} & a_{22} & a_{23} & a_{24} \\ a_{31} & a_{32} & a_{33} & a_{34} \\ a_{41} & a_{42} & a_{43} & a_{44} \end{bmatrix} \begin{bmatrix} \psi_1 \\ \psi_2 \\ \psi_3 \\ \psi_4 \end{bmatrix}$$
- where $\begin{bmatrix} \psi_1 & \psi_2 & \psi_3 & \psi_4 \end{bmatrix}$ is the global stiffness matrix.
- and $\begin{bmatrix} a_{11} & a_{12} & a_{13} & a_{14} \\ a_{21} & a_{22} & a_{23} & a_{24} \\ a_{31} & a_{32} & a_{33} & a_{34} \\ a_{41} & a_{42} & a_{43} & a_{44} \end{bmatrix}$ is the global stiffness matrix.

FEM Electrostatic Example

- For this example,

$$\begin{bmatrix} a_{11} & a_{12} & a_{13} & a_{14} \\ a_{21} & a_{22} & a_{23} & a_{24} \\ a_{31} & a_{32} & a_{33} & a_{34} \\ a_{41} & a_{42} & a_{43} & a_{44} \end{bmatrix} = \begin{bmatrix} \frac{1}{12} & \frac{1}{24} & \frac{1}{24} & \frac{1}{12} \\ \frac{1}{24} & \frac{1}{12} & \frac{1}{24} & \frac{1}{24} \\ \frac{1}{24} & \frac{1}{24} & \frac{1}{12} & \frac{1}{24} \\ \frac{1}{12} & \frac{1}{24} & \frac{1}{24} & \frac{1}{12} \end{bmatrix}$$
- The eight elements share the global matrix in each other. They are determined by the local stiffness matrix of the element (which is specified below):

$$\begin{bmatrix} a_{11} & a_{12} & a_{13} & a_{14} \\ a_{21} & a_{22} & a_{23} & a_{24} \\ a_{31} & a_{32} & a_{33} & a_{34} \\ a_{41} & a_{42} & a_{43} & a_{44} \end{bmatrix} = \begin{bmatrix} \frac{1}{12} & \frac{1}{24} & \frac{1}{24} & \frac{1}{12} \\ \frac{1}{24} & \frac{1}{12} & \frac{1}{24} & \frac{1}{24} \\ \frac{1}{24} & \frac{1}{24} & \frac{1}{12} & \frac{1}{24} \\ \frac{1}{12} & \frac{1}{24} & \frac{1}{24} & \frac{1}{12} \end{bmatrix}$$
- Note: for larger problems with many elements, there will be many terms in the global stiffness matrix.

FEM Electrostatic Example

- With the global stiffness matrix defined, node voltages can be determined:

$$\begin{bmatrix} \psi_1 & \psi_2 & \psi_3 & \psi_4 \end{bmatrix} \begin{bmatrix} a_{11} & a_{12} & a_{13} & a_{14} \\ a_{21} & a_{22} & a_{23} & a_{24} \\ a_{31} & a_{32} & a_{33} & a_{34} \\ a_{41} & a_{42} & a_{43} & a_{44} \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \end{bmatrix}$$
- For the global matrix,

$$\begin{bmatrix} a_{11} & a_{12} & a_{13} & a_{14} \\ a_{21} & a_{22} & a_{23} & a_{24} \\ a_{31} & a_{32} & a_{33} & a_{34} \\ a_{41} & a_{42} & a_{43} & a_{44} \end{bmatrix} = \begin{bmatrix} \frac{1}{12} & \frac{1}{24} & \frac{1}{24} & \frac{1}{12} \\ \frac{1}{24} & \frac{1}{12} & \frac{1}{24} & \frac{1}{24} \\ \frac{1}{24} & \frac{1}{24} & \frac{1}{12} & \frac{1}{24} \\ \frac{1}{12} & \frac{1}{24} & \frac{1}{24} & \frac{1}{12} \end{bmatrix}$$
- Inverting the matrix gives:

$$\begin{bmatrix} \psi_1 & \psi_2 & \psi_3 & \psi_4 \end{bmatrix} = \begin{bmatrix} \frac{1}{12} & \frac{1}{24} & \frac{1}{24} & \frac{1}{12} \\ \frac{1}{24} & \frac{1}{12} & \frac{1}{24} & \frac{1}{24} \\ \frac{1}{24} & \frac{1}{24} & \frac{1}{12} & \frac{1}{24} \\ \frac{1}{12} & \frac{1}{24} & \frac{1}{24} & \frac{1}{12} \end{bmatrix}^{-1} \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \end{bmatrix}$$
- $\psi_1 = \frac{1}{12} \begin{bmatrix} \psi_1 & \psi_2 & \psi_3 & \psi_4 \end{bmatrix}$

FEM Electrostatic Example

- This equation is used to solve for unknown node voltages, given the voltages of other nodes.
- One method involves an iteration procedure.
- All unknown node voltages are set to zero, and the above equation is solved for until the change between iterations is negligible.
- Once node voltages are found, voltages along the element are found by using the shape functions.

EMPro FEM

- Model is meshed into tetrahedrons.
- More nodes are included for matrix calculations as shown in the tetrahedron to the right.
- The electric field is solved for each tetrahedron, and the field at any specific point in the tetrahedron is defined by the matrices calculated earlier.
- The magnetic fields are found from the electric fields.
- This is completed for each frequency in a sweep.

Finite Difference Time Domain (FDTD)

- Full wave 3D EM solver for arbitrary 3D structures.
- Model is subdivided into interlocking cubes, known as Yee cells.
- The electric and magnetic fields are found successively through time-stepping.
- Solutions are found in time.

EMPro FDTD

- Model is gridded into cubes (Yee cells).
- E-field & H-field directions are defined as shown to the right.
- These directions are important for computing finite differences.

EMPro FDTD

- Expanding Maxwell's curl equations gives:

$$\begin{aligned} \frac{\partial E_x}{\partial t} &= -\frac{1}{\mu} \left(\frac{\partial B_y}{\partial z} - \frac{\partial B_z}{\partial y} \right) & \frac{\partial B_x}{\partial t} &= \frac{1}{\epsilon} \left(\frac{\partial D_y}{\partial z} - \frac{\partial D_z}{\partial y} - \rho_v \right) \\ \frac{\partial E_y}{\partial t} &= -\frac{1}{\mu} \left(\frac{\partial B_z}{\partial x} - \frac{\partial B_x}{\partial z} \right) & \frac{\partial B_y}{\partial t} &= \frac{1}{\epsilon} \left(\frac{\partial D_z}{\partial x} - \frac{\partial D_x}{\partial z} - \rho_v \right) \\ \frac{\partial E_z}{\partial t} &= -\frac{1}{\mu} \left(\frac{\partial B_x}{\partial y} - \frac{\partial B_y}{\partial x} \right) & \frac{\partial B_z}{\partial t} &= \frac{1}{\epsilon} \left(\frac{\partial D_x}{\partial y} - \frac{\partial D_y}{\partial x} - \rho_v \right) \end{aligned}$$

EMPro FDTD

- Finite differences are applied to the previous equations. 'n' refers to the current timestep.

$$\begin{aligned} \frac{\partial E_x}{\partial t} &= -\frac{1}{\mu} \left(\frac{\partial B_y}{\partial z} - \frac{\partial B_z}{\partial y} \right) \\ \frac{\partial^2 E_x}{\partial t^2} &= \frac{1}{\mu \epsilon} \left(\frac{\partial^2 D_y}{\partial z^2} - \frac{\partial^2 D_z}{\partial y^2} - \frac{\partial^2 \rho_v}{\partial x^2} \right) \\ \frac{\partial^2 E_x}{\partial t^2} &= \frac{1}{\mu \epsilon} \left(\frac{\partial^2 D_y}{\partial z^2} - \frac{\partial^2 D_z}{\partial y^2} - \frac{\partial^2 \rho_v}{\partial x^2} \right) \end{aligned}$$

- From this, the current value of the E-field is found from previous values. The same method is also applied to the H-field components.

Comparison of FEM & FDTD

FEM	FDTD
<ul style="list-style-type: none"> Tetrahedron meshing is more accurate for rounded objects, as FDTD uses boxes. Adaptive frequency sweep is more accurate for high-Q devices. Better for RF/microwave components. 	<ul style="list-style-type: none"> Easier to understand and program. Produces broadband results from a single simulation. Much less memory intensive than FEM, as solutions involve simple difference equations compared to solving matrix equations. Better for antennas and large objects.

Popular 3D EM Simulators

-
 - Ansys HFSS
 - Most popular.
 - FEM
-
 - Agilent EMPro
 - FEM & FDTD
 - What Cal Poly uses.
-
 - CST Microwave Studio
 - FEM & FDTD

Agilent EMPro

- Powerful modeling environment.
- Choice of two simulators:
 - FEM
 - FDTD
- Integration with Agilent ADS

Conclusion

- Numerical solution methods are needed for complex structures where classical analysis becomes impractical.
- Numerical methods involve volumetric sampling and solving Maxwell's equations on the smaller pieces.
- Choosing a method depends on dimensions of the problem: planar → MoM, full 3D → FEM or FDTD.
- EMPro includes a powerful modeling environment coupled with either FEM or FDTD simulation depending on the model.

APPENDIX C: LAB MATERIALS

FDTD Simulations in Agilent EMPro

What is EMPro?

- EMPro is a 3D EM simulator.
- It solves problems where using classical analysis would be impractical due to problem complexity.
- EMPro includes a frequency-domain solver (finite element method) and a time-domain solver (finite difference time domain).

EMPro Simulation Procedure

- All project files are listed in the project tree.
- The process of completing a simulation involves moving down the project tree.

EMPro Simulation Procedure

- Parts (modeling)
- Objects can be created using EMPro's modeling environment or imported from Solidworks, AutoCAD, etc.
- Creating rectangular or cylindrical objects in EMPro involves drawing 2D faces and extruding them to create 3D shapes.
- For more complex objects, EMPro includes many other tools to create arbitrary shapes.

EMPro Simulation Procedure

- Example: Creating a Cylinder

EMPro Simulation Procedure

- After the entire design has been modeled, add input and output ports.
- Draw feed endpoints, or define them.
- Adding ports creates sources that can be edited later, under definitions.

EMPro Simulation Procedure

- After adding ports, add any desired sensors to view near or far fields.
- Near field sensors are useful for verify field configurations in resonators.
- Far field sensors are useful for looking at radiation patterns of antennas.

EMPro Simulation Procedure

- After adding sensors, project definitions are added.
- First define all relevant object materials (copper, teflon, brass, etc.)
- These can be added from EMPro's library or defined if the electromagnetic properties of the material are known.

EMPro Simulation Procedure

- Under circuit component definitions, the sources used for the ports can be found.
- 1 V, 50 ohm sources are default, but these values can be edited.
- Their source waveform can also be chosen from a dropdown menu. These are defined next.

EMPro Simulation Procedure

- The source waveforms are defined next.
- For broadband results, short time pulses can be used.
- For single frequency results, ramped sinusoids are assigned.

EMPro Simulation Procedure

- Next, if any sensors were added from before, their definitions can be edited.
- Various fields can be obtained by checking the appropriate boxes (E-field, H-field, etc.)

EMPro Simulation Procedure

- Next, boundary conditions can be applied. The default is absorbing for all 6 directions.
- Gridding or meshing can then be defined, depending on if FDTD or FEM simulation was chosen.
- Toggling between FDTD and FEM is located near the wrench icon at the top of the window.

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EMPro Simulation Procedure

- For FDTD gridding, smaller features require smaller cell sizes.
- When gridding, it is beneficial to visually inspect the grid after collating changes.
- This helps locate any unwanted short or open circuits.
- If so, smaller cell sizes can be applied and inspected.
- This process continues until the gridding is sized to the desired accuracy.

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EMPro Simulation Procedure

- Next, clicking the wrench icon opens the simulation setup window.
- FDTD setup is shown to the right.
- Options include parameter sweep, S-parameters, discrete frequency data, and termination criteria.
- For the most accurate results, leave enough maximum simulation time to reach the threshold.
- Simulations should end by reaching the threshold, rather than reaching the maximum time.

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Viewing the Results

- Clicking the results tab brings up all simulation data. Default data plots and others (Smith chart) are available for viewing.

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Conclusion

- Simulation in EMPro involves moving down the project tree.
- Depending on the project, either FEM or FDTD simulations can be completed.
- Specifically for FDTD simulations, gridding the structure accurately and efficiently is the most important aspect.