ABSTRACT

Cylindrical surface waveguides are extremely useful for transporting a surface wave in one dimension with low attenuation when the medium surrounding the surface waveguide has low attenuation. Additionally, surface waveguides guide a pure low-order mode with no cutoff frequency and low distortion, while higher-order modes are quickly attenuated. In this work, we develop simulations and conduct experiments to design a large-radius surface waveguide which can be measured with existing lab equipment. Additionally, we develop and examine the effects of small, simply-shaped couplers, which are highly inefficient at launching surface waves. We also measure the effects of bends on a large-radius surface waveguide (approximately 0.24\(\lambda\)). We study the propagation of the surface wave with a surrounding medium of air and of sand, which is a lossy medium.
To my parents, for their love and support
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1.1 Motivation

The existence of cylindrical surface waves were predicted by a German physicist, Arnold Sommerfeld in 1909 [1]. However, the existence of such waves was debated widely until the 1950s when Georg Goubau made convincing measurements of the surface waves [2, 3] (see Collin’s discussion of the argument [4]).

Surface waveguides are extremely useful in that they have low frequency distortion, no cutoff frequency of the fundamental mode, and low signal attenuation. Additionally, non-fundamental modes have large attenuation, which means that it is trivial to have a low-distortion transmission, assuming the fundamental mode is adequately excited. Surface waveguides are particularly interesting for long-distance, low-attenuation power transfer. However, surface waveguides have major problems with radiation when the waveguide is bent and usually require large wave-launchers for efficient launching of the waves.

Surface waves have several useful applications. Surface waveguides are well-suited for long distance transmission [5] due to low attenuation on the line and purity of the fundamental mode. Surface waveguides have been shown to be especially useful in terahertz transmission because traditional microwave waveguides have impractical conductive loss and fiber-optic waveguides have impractical dielectric loss [6, 7, 8, 9]. Since surface waves are primarily transported in the medium external to the conductor (usually air), the losses are much lower than those experienced in traditional waveguides. Additionally, since surface waveguides have such low loss, it is well-suited to applications where loss is a limiting factor although the maximum power throughput can be limited by different surface modifications such as dielec-
tric loss in a dielectric covered line [10]. The large field extension of surface waveguides can be used as an advantage in sensing applications. For example, Salman used a surface waveguide to feed antennas in a security fence which detects intruders [11]. Others have used Goubau lines (a type of surface waveguide with a dielectric surface modification) to measure the concentration of high loss gases [12]. Surface waves have also been applied to the biological sciences to explain how nerves work [13].

1.2 Structure

In this work, we will show that we have developed a meaningful way to simulate surface waveguides. We will design and build a large-radius waveguide, which is similar enough to existing theory to create meaningful comparisons but also has a large enough radius to demonstrate the effect of lower attenuation on larger radius surface waveguides. We will show modifications that allow for more practical waveguide property measurement with existing equipment, such as tapering the surface waveguide within the flared launching horns that we design. We will test the effect of bends on our larger-radius structure, which has been shown to create a lot of loss for surface waves excited on Goubau surface waveguides, which have a dielectric coating to reduce field extension [14]. We will design simply-shaped, small surface wave launchers, which we expect to be much more lossy than the traditional launchers because efficiency is generally related to the launcher size; a larger launcher generally yields a smaller impedance discontinuity. These new launchers will allow hollow surface waveguides to be multi-purpose and transport a material on the inside of the line, like gas or water. We will also test the effect of a lossy media, in this case sand, around and near the surface waveguide.

In Chapter 2, we will discuss the theory of surface waveguides. In Chapter 3, we will describe the design of our practical surface waveguide in simulation with FEKO, a method of moments based, full-wave solver. We will investigate the effect of radius size, tapering of the surface waveguide within flared coaxial launchers, and the effects of discontinuities necessary to create a practical and cost-effective surface waveguide. Additionally, we will investigate geometric concessions necessary in simulation, such as cylindrical approximation by a cuboid and finite mesh density, to fit a large simulation.
 (> 100\(\lambda\)) with small details into 8 GB of RAM. It is necessary to keep the simulations stored in RAM rather than in virtual RAM in the hard drive because storing simulation details increases the simulation time by at least an order of magnitude, and simulations for these large complex structure are already on the order of days. In Chapter 4, we will verify the functionality of the complete surface waveguide developed in simulation and compare experimental and simulated measurement of new surface waveguide developments such as the effects of bends on a large-radius waveguide, the effects of small, simple surface wave couplers, and the effects of lossy media surrounding and near the surface waveguide. In Chapter 5, we will discuss the implications of these results.
2.1 Field Equations and Modes

The field equations for an infinitely long surface waveguide are well-known [15, 2]. Very quickly after Sommerfeld derived the surface wave equations, Hondros showed that higher-order modes that are circularly-asymmetric are very quickly attenuated and are not worth considering as propagating modes [15]. Stratton provides an argument for the damping of non-fundamental modes (in English) [16], which simply points out that the solution of the eigenvalue equation when there is an angularly-varying mode yields a large imaginary part to the propagation constant along the surface waveguide, $k_z$. This effect can be mitigated by a large conductor conductivity, but this conductivity needs to be so large that even copper is not conductive enough [16].

For an infinitely long surface waveguide of radius $a$ oriented so the center of the cylinder is along the z-axis as shown in Figure 2.1, the fundamental
mode is of the form (assuming time harmonic waves) [2]:

\[ E_z = AJ_0(k_{p,c}\rho)e^{-jk_zz} \]  

(2.1)

\[ E_\rho = AJ_kz^2 J_1(k_{p,c}\rho)e^{-jk_zz} \]  

(2.2)

\[ H_\phi = AJ_\omega\epsilon_kz J_1(k_{p,c}\rho)e^{-jk_zz} \]  

(2.3)

and \( H_z = H_\rho = E_\phi = 0 \), for \( r < a \) and, where \( E_u \) and \( H_u \) represent the electric and magnetic fields respectively in the \( u \) direction, and \( k_c^2 = k_z^2 + k_{p,c}^2 \), where \( k_c = e^{-j\pi/4} \sqrt{\omega\mu_c\sigma_c} \), which is the conductor wavenumber assuming the conductivity of the material is large [16]. Similarly, the fields outside the waveguide \( r > a \) are given by:

\[ E_z = AH_0^{(1)}(k_{p}\rho)e^{-jk_zz} \]  

(2.4)

\[ E_\rho = AH_kz H_1^{(1)}(k_{p}\rho)e^{-jk_zz} \]  

(2.5)

\[ H_\phi = AH_\omega H_1^{(1)}(k_{p}\rho)e^{-jk_zz} \]  

(2.6)

where \( k^2 = k_z^2 + k_\rho^2 \), where \( k = \omega \sqrt{\epsilon \mu} \), and where all other field components are zero. The magnitudes of these fields are shown in Figure 2.2. It is clear that the waves decay slowly outside the waveguide and appear to propagate in the expected sinusoidal distribution.

The eigenvalue equation found from the boundary conditions at \( a \) is

\[ \frac{\mu k_\rho H_0^{(1)}(k_{p}\rho)}{k^2 H_1^{(1)}(k_{p}\rho)} = \frac{\mu_c k_{p,c} J_0(k_{p,c}\rho)}{k_c^2 J_1(k_{p,c}\rho)} \]  

(2.7)

With equation 2.7, the dispersion equations, and the definitions of \( k \) and \( k_c \), it is possible to solve for \( k_{p,c}, k_\rho, \) and \( k_z \).

For these fields to exist, one must have a surface modification such as a lossy conductor, a dielectric coating, or corrugated or twisted waveguide [2]. For large conductivities, the loss is typically in the derivation, other than being a necessary condition for the surface wave to exist. Additionally, surface modifications must be chosen carefully; Sharp and Goubau found that rain and ice on a surface waveguide increase transmission loss drastically [17].
2.2 Surface Impedance

Surface impedance is a simple way to gain physical insight from surface waveguides. It is especially useful because surface waveguides must have surface modifications which sometimes yields complex structures such as finite conductivity, dielectric coatings, and corrugation [18]. For a cylindrical surface waveguide along the z-axis, such as in Figure 2.1, the surface impedance is given by:

\[ Z_s = \frac{E_z}{H_\phi} \]  

(2.8)

where \( E_z \) and \( H_\phi \) are the fields found in the previous section. Roberts [18] found that the surface impedance definition reinforces the surface wave mode since the equation for the surface impedance only yields one physical solution for a surface waveguide with a very small loss. The surface impedance for the surface waveguide is inductive; there is no solution to the capacitive case [18]. The natural modes of the surface waveguide, i.e., the modes without a specific forcing function, can be found by matching the surface impedances at the border of the conductor of the surface waveguide and the outside medium [16].

2.3 Effect of Increasing Radius

Increasing the surface waveguide radius has several effects. Most importantly, the field confinement of a surface waveguide decreases with increasing radius. This becomes clear from the solutions to the eigenvalue equation.
Goubau walks through a simple case for a radius much larger than the skin depth $k_{\rho,c}a \gg 1$ and not too large $k_{\rho}a \ll 1$, from which we can derive understanding without the use of a numerical solver [2]. Through using asymptotic approximations for the Hankel and Bessel functions, Goubau is able to separate the magnitude and the phase of the complex eigenvalue equation, which then can be solved sequentially. As the field confinement decreases, loss in the surface waveguide also decreases since the surface wave is bound to the surface waveguide through loss. Thus, increasing the surface waveguide radius decreases loss in the system. One can also find the phase velocity from the solution to the eigenvalue equation, which is related to the real part of $k_z$. As the radius of the waveguide increases, the phase velocity approaches the speed of light (from below since a surface wave is a slow wave). Decreasing field confinement means decreasing loss in the surface waveguide, which agrees with our finding the phase velocity is increasing.

The effect of increasing the surface waveguide radius can be negated by changing the material parameters to have a lower conductivity or to have a higher permittivity, which Berceli showed has a bigger effect on surface waveguide loss than permittivity with the same loss tangent [19]. However, this also means that if it is necessary to use a lossy or magnetic metal (or a metal that is both lossy and magnetic) it is possible to mitigate the effects by increasing the radius of the waveguide.

Additionally, increasing the waveguide radius has structural benefits, in that it is possible to build a sturdier waveguide with less sagging. As we will discuss later, bends in the surface waveguide lead to radiation loss, so less sagging in the surface waveguide will increase the transmission parameter through the line.

2.4 Cutoff Frequency

There is no cutoff frequency for the fundamental mode, which is a big advantage of surface waveguides over other types of waveguides. Other authors have shown that the $k_z$ propagation constant never vanishes in Equation 2.7 for metallic waveguides, which have a high conductivity and a finite loss [16, 20]. Unfortunately, this equation is only solvable numerically, so it is a bit more difficult to gain an intuitive sense of the behavior. Since there is no
2.5 Bends in Surface Waveguides

Bends have always been a huge problem for surface waveguides; even sags in the line cause significant loss. To develop an intuitive understanding of radiation around a bend it is helpful to think of phase fronts along the surface waveguide. When the waveguide is straight, the phase fronts are all perpendicular to the axis of the waveguide, with a phase velocity slightly less than the speed of light. When the surface waveguide bends, the phase front on the outside of the bend would need to travel much faster to stay perpendicular to the surface waveguide. When the wave is forced to travel faster than the speed of light, it radiates instead. Goubau and Sharp [10] and later Chiba [14] published experimental studies of Goubau lines, which are metallic lines covered with a dielectric coating that are another type of surface waveguide lines. Since the dielectric coating serves to decrease the radial field extension, and more closely bind the surface wave to the conductor, we expect these experimental results to have less loss than a bare copper surface waveguide. These experimental studies found that losses are non-linear with respect to bend angle. In fact, it is desirable to divide a large bend angle into several smaller bend angles. Additionally, the authors found that the amount of radiation is far more strongly dependent on the total bend angle of the surface waveguide rather than the radius of curvature of the bend.

Nakamura et al. [21] were the first to develop a theoretical method which could be used to accurately predict the radiative bend loss with the trav-
eling wave method. In this method, the electric fields were used to find a transmission-line equivalent of the surface waveguide. The bends were approximated by different characteristic impedances of the line. Then from the characteristic impedances, Nakamura found the reflections and transmission coefficients of the line. By assuming the bend is a discontinuity that causes radiation, he applied a pattern multiplication method, using the transmission and reflection as coefficients to find the current around the bend, which is then used to find the far field radiation pattern.

2.6 Launchers

Although Goubau mentions flared coaxial launchers as early as 1950 [2], he does not discuss how these devices work besides an intuitive understanding. He explains that the impedance of the coaxial cable is proportional to \( \log(b/a) \) so, by flaring the outer cable, the characteristic impedance of the coaxial line increases gradually. Since this is a log type relationship, eventually there is a point where increasing the size of the outer conductor makes little different to the characteristic impedance. At this point, he simply removes the outer conductor, which can be thought of as an outer conductor at infinity. While this makes intuitive sense, it does not completely describe what is happening in the coaxial launcher. This becomes more clear when thinking of the modes of a coaxial line and the modes of a surface waveguide line. The coaxial flare is a mode transducer, transforming the \( TEM \) wave of a coaxial cable to the \( TM_{01} \) wave of the surface waveguide.

Gunn [22] made a detailed experimental study about the effect of surface waveguide flared coaxial launchers. He developed several empirical formulas to design a surface waveguide for a surface waveguide with a thin radius:

\[
\frac{N_\rho}{N} = -\frac{2 \ln \frac{\xi}{a}}{\ln 2.2|\xi|} \tag{2.9}
\]

where \( N_\rho \) is the power within a certain radius \( \rho \), \( N \) is the total transmitted power, \( a \) is the radius of the waveguide, and \( \xi \) is a material parameter found by \( \xi = (-j0.89k_\rho a)^2 \). Additionally, to have a small phase curvature at the
Figure 2.3: An illustration showing the half-angle opening of a coaxial flare.

mouth of the launcher, the half-angle opening should follow the equation:

$$\tan \frac{\theta}{2} \leq \frac{\lambda}{4}$$

(2.10)

where $\theta$ is the half-angle opening and $\lambda$ is the free space wavelength of the desired signal as in Figure 2.3. There are other attempts to theoretically explain and model the field transition inside the coaxial flare launcher [23, 24]; however, the experimental results are roughly as accurate as Gunn’s results but with much more complex derivations. Since we are testing the launchers in simulation first, and mostly studying the surface wave on the line, rather than the efficiency of the launcher, Gunn’s equations are good enough.

Other more compact launchers have also been designed [25, 26], but they have not been as efficient or as simple to design as a coaxial flare launcher. Brown designed an annular slot launcher and compared it with a flared coaxial launcher with a radius of $1.5\lambda$ [25]. The optimal launcher that he measured had the slot $0.37\lambda$ above the surface of the waveguide. The finite ground plane for the annular slot extended beyond the slot, but the total structure appears to be smaller than the flared coaxial launcher. Brown carefully varied several design parameters, including slot width, distance from slot to waveguide, and the thickness of the slot material (so as to behave more like an annular parallel plate waveguide for thicker launchers). Although the annular slot theoretically performs much better than the coaxial launcher, measured results show that the flared coaxial launcher has a minimum efficiency of 63.8% and the annular slot has a maximum measured efficiency of 63%. Beal measured the effect of an array of slots, arranged circumferentially around a surface waveguide, rather than a single slot [26]. This launcher had a measured efficiency of 25% with an outer radius of $0.127\lambda$ for the entire structure.
CHAPTER 3

DEVELOPMENT OF SURFACE WAVEGUIDE

Simulation is helpful in designing waveguide structures from several perspectives. From the experimental side, it is much cheaper and much less time-consuming to use a software package to see the effect of adjustments to the structure. Custom-machining copper parts to test on the surface waveguide can be very expensive and can have a large lead-time. Full-wave simulations are helpful from the theoretical side when the geometries become too complex to be solved or it becomes very difficult to make meaningful approximations in the theory. We can see if approximations in the theory tend to hold true before manufacturing a structure and measuring it in the lab. However, it is important to note that, while simulations are a useful modeling tool, they can still have errors and effects which can only be shown through measurements in the lab. Simulations are helpful but not sufficient to test the validity of the final design, which should be measured in a laboratory setting.

We used FEKO, a method of moments-based solver, to develop a surface waveguide model and to simulate changes to the structure before building a structure. We chose a method of moments-based solver because the algorithm is very efficient for our simulation structure, which is generally in the form of a large metallic cylinder with metallic launchers on either end (no dielectric in the simulation). We made several geometrical approximations since we wanted a very large structure, on the order of $100\lambda$, to compare with experiments in the literature. In this chapter, we will examine each of these approximations in turn. The most prominent approximation is replacing the cylinder with an octagonal structure, which simplifies the meshing of the structure, allowing the simulation to fit in 8GB of RAM, which is the maximum available for the simulation computer. It is highly desirable to keep the simulation in RAM if possible because if more RAM is necessary, FEKO allocates virtual RAM on the hard drive; unfortunately, this increases the simulation time by at least an order of magnitude. Additionally, we had to
choose a slightly coarse mesh, which especially affects the simulated matching at the ports. To keep the simulations in a reasonable simulation time (a couple of days per simulation at the longest), we simulated at only a few discrete frequencies in our frequency band of interest.

Several other approximations were necessary from an experimental standpoint to create an effective yet low-cost experiment. First, we adapted the launching structure so that the large-radius surface waveguide could still mate with existing measurement equipment. Next, we used two separate pieces of commercially-available copper pipe, which required a coupler to join the two pieces of the structure.

Even after all the approximations, several simulations did not fit in memory for structure with more complex geometries. For these simulations, we compared a 10\(\lambda\) simulated waveguide with the 100\(\lambda\) waveguide built in the lab. With the simulated waveguide, we were able to measure trends. Even though the 10\(\lambda\) structure will incur more undesired coupling between the launching structures and more undesired radiation, both of which appear as throughput of the line, we can generally have a good idea of how the structure will change with frequency or with varying geometric parameters. For an unsuccessful comparison, see Section 4.2 which describes a compact surface waveguide launcher. This simulation was not successful because it was not large enough to allow effects, such as radiation from the launcher, space to attenuate, which could be fixed by having a larger simulation structure. For a successful comparison, see Section 4.3 which discusses the effect of bends.

3.1 Cylindrical Approximation

One of the most apparent problems about simulating a surface waveguide in FEKO is that a cylinder is very difficult to mesh efficiently. Thus, we substituted an octagon which was much simpler to mesh accurately. We created a small simulation with which to compare the two configurations. The surface waveguide line has a small radius relative to frequency 0.025\(\lambda\), which is not one of the goals of the research. Additionally the waveguide is roughly 10\(\lambda\) long rather than 100\(\lambda\) which means that the radiation from the launching horns will compromise the results for the surface wave. However, this is a good figure of merit to compare the results of the two types of lines.
Figure 3.1: The simulation structure for both the cuboid and the cylindrical lines.

If the results are similar, then it is likely that these two lines radiate and launch surface waves in the same way.

Figure 3.1 shows the simulation structure for comparing the cylindrical and cuboid surface waveguides. The structure is first simulated with the cylindrical line; then the structure is simulated with a cuboid line and the results of the two simulations are compared. The surface wave is launched using coaxial horn launchers, as discussed previously in section 2.6. Figure 3.2 shows a close-up of the two types of surface waveguides lines. The ideal cylindrical line is on the left. The blue line down the center of the structure is a symmetry plane, which is simply a simulation marker rather than a physical part of the line. The line on the right is the cuboid line. The cuboid line is made from intersecting two cuboids with square cross sections that each have a length of twice the desired surface waveguide radius. Thus, the midpoint of the faces of the line is at the correct radius; however, the points of the octagon are farther away by a factor of $1/\cos(22.5^\circ)$ which is an 8% maximum deviation. Figures 3.3 and 3.4 show the S-parameters of the cylindrical surface waveguide compared to the cuboid surface waveguide. Note that the reflection parameters are very similar, with slight deviations near the nulls. The transmission parameters show a slight loss with the cuboid approximation. Since the match is relatively unchanged, this implies that there is slightly more radiation that is not being received on the other end of the surface waveguide or that there are higher modes excited that are attenuated before reaching the other side of the line. The near field of the structure is shown in Figure 3.5. The fields are so similar that it is only
necessary to show one of the fields. This clearly shows that a surface wave is being transmitted between the two launchers. It is reassuring to the see the surface wave electric field dominates the radiated electric field, which decreases as $\frac{1}{r^2}$ and is non-negligible for a waveguide of length $10\lambda$.

3.2 Large Radius

Next we verify that our model has the effect we expect with large radius surface waveguides. To test this, we create a larger radius simulation to compare with a simulation with the same dimensions as the previous cuboid simulation, except with a length of $100\lambda$. For this simulation, we use the same conical launchers as previously. We simply adjust the dimensions on the coaxial side of the launcher to have an impedance of $50\Omega$ with the thicker inner conductor. The length of the surface waveguide is approximately $100\lambda$. The larger radius waveguide had a radius of $0.127\lambda$ while the smaller radius waveguide had a radius of $0.025\lambda$. The near field of a section of the larger radius waveguide is shown in Figure 3.6. The field distribution still shows that the surface waveguide is propagating in the expected fashion along the surface waveguide. Figure 3.7 shows the reflection parameters for the sur-
Figure 3.3: Comparing cuboid approximation surface waveguide reflection characteristics to those of a cylindrical surface waveguide.

Figure 3.4: Comparing cuboid approximation surface waveguide transmission characteristics to those of a cylindrical surface waveguide.
face waveguides. The large radius case has a much better match than the smaller radius. While the match is slightly better because of the large radius impedance characteristic, it is also better because there is more radiative loss since the launcher size was not increased with increased radius size. The large radius has between 42% and 51% power loss while the small radius has between 24% and 39% power loss. Figure 3.8 shows the throughput parameter for this configuration. The radiated field intensity along the surface waveguide decreases much more quickly than the surface wave field intensity, so the throughput increase is not a function of the increased radiation of the large radius waveguide. As shown in theory, large radius waveguides have lower loss than smaller radius waveguides.

3.3 Surface Waveguide Taper

We found it necessary to taper the surface waveguide for practical construction reasons, namely that we wanted to measure a large-radius surface waveguide with our existing measurement equipment, which does not have a large coaxial adapter. We decided to taper the surface waveguide within the launcher rather than on the surface waveguide line. This decreased building complexity and cost, although a more favorable solution could possibly be created from tapering the surface waveguide line in a less sensitive location.
Figure 3.7: Radius size effect on reflection parameters.

Figure 3.8: Radius size effect on transmission parameters.
Figure 3.9: Two views of the untapered waveguide structure. The view on the left is looking down into the conical launcher, where the waveguide transitions from a SWG to coaxial without changing dimension of the SWG. The view on the right shows the back of the launcher and the coaxial size.

However, by tapering the surface waveguide line within the launcher, we can lump in the tapering loss with the launcher effects and have a simple surface waveguide transmission line that does not change dimensions.

Figures 3.9 and 3.10 show the untapered and tapered structures, respectively. Note that the coaxial input to the launcher is much smaller on the tapered structure than on the untapered structure. The best length of taper was found by optimizing for maximum $S_{21}$ within FEKO using the Nelder-Mead algorithm which finds a local maximum. This requires a reasonable guess as to the correct length so that the system will find an absolute maximum or close to the absolute maximum. As our first guess, we choose that the taper is the length of the launcher, which will cause the impedance of the launcher to vary as slowly as possible. However, the optimizer shortened the length of the taper to roughly 55% of the coaxial launching horn length or roughly $0.64\lambda$. We found that having the taper the entire length of the flare decreases the system transmission parameters because the SWG discontinuity at the launcher opening causes more radiation. Also, the taper is not as short as possible because that would be a drastic impedance change, which would decrease the transmission parameter because more of the energy would be reflected from a poor match.

Figures 3.11 and 3.12 show the reflection and transmission parameters,
Figure 3.10: Two views of the tapered waveguide structure. The view on the left is looking down into the conical launcher, where the waveguide transitions from a SWG to coaxial with narrowing of the SWG radius. The view on the right shows the back of the launcher and the coaxial size.

respectively for the optimum taper length. The taper generally decreases the quality of the match, by as much as 18.4 dB at 4.8 GHz. Additionally, the taper decreases the transmission through the waveguide by as much as 1.5 dB over our frequency band of interest. This seems like a reasonable trade-off of surface waveguide complexity and loss through the waveguide for the ability to inexpensively measure the structure with existing equipment.

3.4 Mesh Density

For large structures with fine details, such as the taper and the small coaxial connection, there is a limit on how fine we can make the mesh and have the simulation fit into computer memory. To test this effect, we compared a 10\(\lambda\) tapered waveguide shown in Figure 3.13 with the mesh that will fit into simulation for the 100\(\lambda\) case and a denser mesh which should yield more accuracy.

The effects of the simulation are shown in Figures 3.14 and 3.15 for the reflection and transmission parameters, respectively. In Figure 3.14, we can clearly see that increasing the density of the surface waveguide mesh has a big effect on the reflection parameter, up to 7.6 dB or a 58.5% difference.
Figure 3.11: Comparing the reflection parameters of the tapered and untapered structures.

Figure 3.12: Comparing the transmission parameters of the tapered and untapered structures.
Figure 3.13: Structure used to test the effect of simulation mesh density.

Note that this is not necessarily indicative of how the $100\lambda$ simulation should perform because the coupling between the two launchers cannot be ignored since the waveguide is too short. Although it seems from this graph that the simulation should become more accurate at higher frequencies, the error introduced by finite mesh size is not that predictable so this is not the case. In Figure 3.15 shows the transmission coefficient for both meshes. In this case, the transmission coefficients do not differ by more than 0.06 dB or 0.75% in magnitude or $1^\circ$ in phase. Thus, it seems that the simulation mesh density used in the larger $100\lambda$ simulations will be useful for predicting the transmission coefficient but could have some significant errors when predicting the reflection coefficient.

3.5 Surface Waveguide Discontinuity

To inexpensively construct and transport the surface waveguide and materials, it was necessary to buy the materials in two sections and join the two sections with a pipe coupler. This pipe coupler introduced a discontinuity in the surface waveguide, so we wanted to see the effect of a discontinuity on a surface waveguide. Figure 3.16 shows the structure we used in simulation to test the effect of this discontinuity. The simulation used a $10\lambda$ surface waveguide with a large radius and coaxial output (i.e., no taper of the sur-
Figure 3.14: Reflection parameter of a simulation with varying mesh density.

Figure 3.15: Transmission parameter of a simulation with varying mesh density.
face waveguide in the coaxial launcher), to keep the structure simple and to enable quick simulation of the structure.

3.5.1 Discontinuity Radial Variation

The first parameter variation for the surface waveguide discontinuity was the discontinuity radius. Since the actual discontinuity radius was only 0.02λ larger than the surface waveguide radius, it is sufficient to test the effect of radial discontinuities for small radial variations.

Figures 3.17 and 3.18 show the reflection and transmission parameters, respectively. For small radial discontinuities, the changes in radial variation have no effect on the S-parameters, making the 0.09λ variation indistinguishable from the 0.18λ variation. Larger radial variations change the matching parameters to look like there is a slight radiation from the discontinuity, which presents as periodic fluctuations in the matching parameter and slight loss in the transmission parameter. However, this effect is still barely noticeable (i.e., less than 0.51 dB loss on all simulated frequencies) in the transmission parameter for radii that are 0.72λ larger than the surface waveguide radius! Thus, a very small discontinuity should present little problems for our actual measured structure.
Figure 3.17: Reflection parameters for various radial discontinuities.

Figure 3.18: Transmission parameters for various radial discontinuities.
3.5.2 Discontinuity Length Variation

We also examined the effect of the length of the surface waveguide discontinuity on the structure. As the actual surface waveguide discontinuity was approximately $0.85\lambda$, we investigated values close to that length and also chose discontinuities that we thought would have a maximum effect on the S-parameters such as lengths of $\lambda/2$ and $\lambda/4$.

Figures 3.19 and 3.20 show the reflections and transmission parameters, respectively. The match changes significantly when the discontinuity length changes; however, this is typical of transmission line sections with differing impedances. While the match does change with the differing discontinuity lengths, the transmission parameter changes less than 0.66 dB for the frequencies that we simulated. Thus, the length of the discontinuity will not affect our measurements appreciably.
Figure 3.20: Transmission parameters for various discontinuity lengths.
CHAPTER 4
EXPERIMENTAL DESIGN AND VERIFICATION

For large, complex structures, simulations are not always able to accurately predict waveguide behavior. While simulations are useful design tools, they must be used in conjunction with theory to design an experiment that can be used to verify the simulation and theoretical predictions. Additionally, we can experimentally measure effects that we are not able to simulate.

In this chapter, we will first measure the complete surface waveguide designed in simulation, using non-ideal features described in the previous chapter. We will then discuss a compact surface waveguide launcher, and the effects of a small surface wave launcher. Next, we will discuss the effect of a 45° and a 90° bend in the surface waveguide. We will conclude this discussion with a study of the effects of lossy media surrounding the surface waveguide and near the surface waveguide.

4.1 Surface Wave Transmission Line

The first structure to experimentally verify is the surface waveguide most similar to the waveguides in the literature, except with a larger radius. This provides a sanity-check to reassure us that our results are reasonable, and it is the simplest case to allow for troubleshooting of the instrument setup.

The simulation file used for reference contains a surface waveguide with the same dimensions as the actual structure: 90.5\(\lambda\) long with a 0.64\(\lambda\) long taper, and a 0.24\(\lambda\) radius. The launcher has a 1.17\(\lambda\) radius at the opening and has a half angle of 45°. Additionally, the model has a pipe coupler with radius 0.26\(\lambda\) and a length of 0.85\(\lambda\) which joins the two sections of surface waveguide. The results comparing the measurements and this simulation are included here.

The experiment test structure shown in Figure 4.1 was constructed with
Figure 4.1: The long, straight SWG in the anechoic chamber, with coaxial launchers on each end of the waveguide.

Figure 4.2: The coaxial launcher and foam supports on the end of the surface waveguide.
two pieces of copper plumbing soldered with a commercially-available coupler to yield a structure that was approximately $100\lambda$ at 5 GHz. The surface wave launchers shown in Figure 4.2 were constructed out of a plate of copper and copper sheet metal molded into two cones. The inner surface waveguide flares were created from a solid block of copper and were each connected to the center pin of an N-type connector which could simply be pushed into the outer part of the N-type connector which was attached to the plates of copper on the outer launching cones when it was time to take measurements. The structure could also be taken apart for ease of transport or for storage, which would prevent any undue stresses on the surface waveguide which might cause it to sag (which would add loss). The surface waveguide launchers were supported by styrofoam stands on top of PVC cubes. Additional styrofoam was inside the launchers to help support the weight of the surface waveguide and to prevent the weight of the structure from bending the inner pin of the N-type connectors on either end of the surface waveguide. Additionally, the surface waveguide was supported in the center of the waveguide with more foam on top of a PVC cube to prevent undue sagging of the line, which would present as extra insertion loss.
Figure 4.4: Phase of reflection parameter for a straight waveguide.

Figure 4.5: Magnitude of transmission parameter for a straight waveguide.
Figure 4.6: Phase of transmission parameter for a straight waveguide.

Figure 4.7: Closer look at phase of transmission parameter for a straight waveguide.
Figures 4.3 and 4.4 show the magnitude and phase of the reflection coefficients respectively. The simulated match is much better than the actual match. This discrepancy in the simulation is likely due to the finite mesh gridding, which especially causes errors in the reflection coefficient as discussed in Section 3.4. The mesh density is very important in simulation accuracy, especially around the surface waveguide taper within the launching structure, where the slight discontinuities can cause radiation. With the meshing structure more coarse than optimal, the surface waveguide has small, non-physical discontinuities which cause radiation in the simulation. Thus, the radiated wave in the simulation is not reflected, which decreases the simulated reflection coefficient magnitude. The reflection coefficient phase is off by a constant, which implies that there is an unaccounted-for phase shift somewhere in the structure. Figure 4.5 shows the magnitude of the transmission coefficient. While there is more loss in the measurement than the simulation, the trends between the two basically match. In the simulation, the transmission increases in performance slightly near 5 GHz, which flattens out in the measured results. As predicted in theory, the loss in the line increases as frequency increases because the line becomes electrically longer. This appears to be somewhat mitigated in the mid-region, possibly because the launchers become more efficient as the frequency goes up because they also become electrically larger. Figures 4.6 and 4.7 show the phase of the transmission coefficient. Since the surface waveguide is very long, the phase of this parameter is quickly varying. The slight offset of the phase shows that the simulation dimensions are very close but not exactly the same.

4.2 Compact Surface Waveguide Launchers

As shown in the previous sections, surface waveguide launching efficiency is directly correlated to size. However, we wanted to create a surface waveguide launcher that is more compact than the coaxial launchers and allows the copper piping to be dual-purpose, with water flowing on the inside of the pipes which is not possible with the flared coaxial launchers. We will refer to this compact structure as a surface wave coupler to differentiate it with the flared coaxial launchers.

We designed the surface wave coupler through optimization in FEKO.
Figure 4.8: The simulation structure used to find the optimum dimensions of the new surface wave coupler.

Figure 4.9: Front view of the compact launcher for surface waveguide. The conductor in the middle is pressure fit to the surface waveguide when the launcher is on the surface waveguide.
Figure 4.10: Side view of the compact launcher for surface waveguide. The outer conductor is made of copper tape.

Figure 4.11: Compact launcher installed on the surface waveguide. It is visible in the middle of the center, styrofoam support of the surface waveguide.
Since the simulation required too much memory to simulate the 100\(\lambda\) case for the surface waveguide, we simulated a surface waveguide of 10\(\lambda\), as shown in Figure 4.8. We used a Nelder-Mead method for optimization which finds a local maximum. Thus, this optimizer requires a good “first guess” to have a reasonable solution, since a local maximum is not necessarily a global maximum. In this case, we optimized the transmission parameter from a port on the surface wave coupler to a coaxial flare with the unused coaxial flare terminated with a system impedance of 50\(\Omega\). We optimized both the surface wave coupler radius and the surface wave coupler length to get the current solution.

To create the couplers we used copper tape as an outer coaxial shell for the surface waveguide and we used semi-rigid coaxial cable to excite the surface wave mode, as shown in Figures 4.9 and 4.10. The radius of the surface wave coupler is approximately 0.625\(\lambda\) at 5 GHz (3.75 cm), where \(\lambda\) is the free space wavelength, and the interior of the surface wave coupler support is tight around the surface waveguide. The length of the surface wave coupler is approximately 0.42\(\lambda\) at 5 GHz (2.52 cm). The interior surface waveguide contact of the surface wave coupler is flush against the surface waveguide when the surface wave coupler is mounted on the surface waveguide. A semi-rigid coaxial cable is connected to the surface wave coupler to excite the surface wave, as seen in Figure 4.9. The length of the semi-rigid coaxial cable is on the order of 2.5\(\lambda\) at 5 GHz, for convenience in the measurement since any current on the outside of the semi-rigid coax will also flow on the outside of the cables attached between the semi-rigid coaxial cable and the measurement device, in this case the vector network analyzer. We added the surface wave couplers to the current structure with the coaxial horns for measurement with the unmeasured ports terminated with 50\(\Omega\) loads, as shown in Figure 4.11. We then measured the performance of the new compact surface wave couplers.

4.2.1 Single Compact Launcher

In this case, we measure the effect of a signal surface wave coupler. The surface wave coupler is placed in the middle of the two coaxial launchers, very near the center of the waveguide. The surface wave coupler is not
Figure 4.12: Compact launcher installed on the surface waveguide. It is visible in the middle of the styrofoam support of the surface waveguide. The measurement is taken between the surface waveguide coupler in the center and the flared coaxial launcher on the left. The remaining coaxial launcher on the right (not shown) is terminated with the 50 Ω system impedance.
precisely in the center of the waveguide because it is built to fit on the surface waveguide rather than the discontinuity that joins the two pieces of the surface waveguide. Thus, the measurements in this case are between the coupler and the coaxial launcher which is further away from the wave coupler, as shown in Figure 4.12. The other coaxial launcher is terminated with the system impedance of 50Ω.

Figures 4.13 and 4.14 show the matching parameter magnitude and phase for the surface wave coupler and a coaxial launcher. The surface wave coupler is typically not as well matched as the coaxial horn, which is to be expected since the surface wave coupler is a much more abrupt transition from TEM mode to $\text{TM}_{01}$ mode, whereas the coaxial flared launcher is a very gradual change in impedance. It also appears that the surface wave coupler excites a radiative mode around 5.6 GHz which increases the match but has little effect on the system loss. Our original assumption for this design was that styrofoam had a permittivity very close to that of air. We would expect a radiating mode at a multiple of $0.5\lambda$. The resonance suggests that the permittivity of styrofoam is roughly 2, which would make the radial distance
Figure 4.14: Phase of reflection parameter for the new surface waveguide coupler.

Figure 4.15: Magnitude of transmission parameter for the new surface waveguide coupler.
Figure 4.16: Phase of transmission parameter for the new surface waveguide coupler.

Figure 4.17: Closer look at phase of transmission parameter for the new surface waveguide coupler.
from the surface waveguide to the radius of the surface wave coupler approximately $0.5\lambda$ at 5.6 GHz. Since the permittivity of styrofoam is lower than our predicted value, we also think that fringing capacitance between the surface wave coupler and the surface waveguide could have lowered the radiative resonance frequency. The phase difference is due to the fact that the surface wave coupler has a significant transmission line attached to the outer ring, which could be shorter if necessary but was kept long for ease of construction.

The first thing that becomes apparent about this surface wave coupler is that the simulation is not adequate for predicting the performance. This is most apparent in Figure 4.15. In this simulation, the transmission coefficient for the coupler and the coaxial launcher differ at most by 0.7 dB, meaning that there is very little difference in performance between the coaxial launcher and the surface wave coupler, which is counterintuitive because launcher efficiency of surface waveguide is usually dependent on the size of the launcher. The measured results confirm that this is not the case; the compact coupler does have between 17 dB and 23 dB of transmission loss for all the frequencies that we measured. Thus, the simulation has far too much radiative coupling between the coaxial launcher and compact coupler and can be neglected for the purposes of analysis for the performance of this system. Figures 4.15 - 4.17 show the transmission coefficient for the coaxial coupler and the coaxial flares. Note the the surface wave coupler is near the center of the waveguide where the coaxial flares are at the ends of the waveguide; thus the couplers should have less waveguide loss since the wave only has to propagate half of the distance that the wave launched from the coaxial launchers has to propagate. From the graph in Goubau [2], note that this surface waveguide is predicted to have less than 0.15 dB loss per 100 ft, which means that our loss is dominated by the coupler losses. Since the surface waveguide loss is negligible, to find the loss from the couplers we can simply subtract the coupler throughput from the coaxial throughput. We find that the coupler adds between 17 dB and 23 dB of loss relative to the flared coaxial launcher over our frequencies of interest and between 22 dB and 29 dB of loss per coupler.
4.2.2 Compact Launcher and Receiver

After characterizing a single surface wave coupler, we wanted to see if it was possible to measure the loss associated with using surface wave couplers for both transmitters and receivers. Since the surface wave couplers are not machined, there is a bit of variation between the two couplers. For this measurement, we placed the surface wave couplers approximately $3\lambda$ from the coaxial flared launchers on each side of the surface waveguide. We terminated both of the coaxial flared loads with the system impedance of $50\Omega$.

Figures 4.18 and 4.19 show the matching parameters of both surface wave couplers. The second surface wave coupler that we constructed seems to have a sharper resonance around 5.6 GHz, where the coupler is well-matched. This is likely due to a slight variation in construction, where the second surface waveguide coupler has a more uniform radial distance from the surface waveguide along the length of the coupler, whereas the first coupler had a slight variation in radial distance from the surface waveguide along its length. The resonance at 5.6 GHz could increase the throughput at this frequency through
Figure 4.19: Phase of reflection parameter for compact launcher and receiver.

Figure 4.20: Magnitude of transmission parameter for compact launcher and receiver.
radiation since the surface wave magnitude is low and might not necessarily dominate. However, the surface wave will still not attenuate as quickly as the radiated wave and the losses come primarily through mismatch losses. We can see an increase in the throughput around 5.6 GHz in Figure 4.20, with approximately 5 dB better signal throughput at this frequency where the couplers are better matched than at other frequencies. As we discussed with the single surface wave coupler, since mismatch losses dominate, a radiative mode can be excited at 5.6 GHz where the distance from the surface waveguide to the surface wave coupler is approximately $\lambda_g/2$, which can allow more throughput if it is well-matched. Figure 4.21 shows the phase of the transmission parameter, which behaves like we would expect with a linear progression of phase along the transmission line.
4.3 Bend in the Surface Waveguide

Bending the surface waveguide has a major effect on transmission efficiency. As we discussed previously, bending the surface waveguide increases radiation dramatically, which presents as loss in the surface waveguide. Since the surface waveguide is created from plumbing pipe, it is simple to modify the existing structure to see the effect of bends. We simply replace the unbent plumbing coupler that previously joined the two halves of the surface waveguide with a plumbing coupler with bends, in this case, 45° and 90°, respectively.

4.3.1 45° Bend

The surface waveguide structure with a 45° bend was simulated in FEKO to see the effect of the bend on throughput and matching of the structure. Figure 4.22 shows the simulation structure in FEKO. The simulated surface waveguide structure is only 10λ because the 100λ simulation structure would not fit in the 8GB of RAM available; however, this structure will be useful to predict trends. This simulation should actually be more useful than that of a 10λ surface waveguide without a bend because we expect less radiative coupling since the radiation from the launching horns is strongest near broadside. Additionally, this simulation has a bend with a non-zero radius.

Figure 4.22: Surface waveguide simulation structure with a 45° bend.
of curvature (i.e., the bend is not a corner on the inside of the bend), which reduces the amount of radiation from the bend and more closely matches the plumbing coupler geometry.

Figures 4.23 - 4.28 show the simulated and measured results. Figures 4.23 and 4.24 show the matching parameter of the surface waveguide without a bend compared to a surface waveguide with a bend. As discussed previously, errors in the reflection coefficients between simulation and measurements are a function of the finite mesh gridding and can be made more accurate with increasing the mesh density. Our accuracy in this case is limited by the computational resources. In the measured cases, it is clear that the 45° bend does not have a big difference on the reflection coefficient magnitude or phase. Figures 4.25 - 4.27 show the simulated transmission coefficient vs. measured transmission coefficient. In this case, the simulation clearly predicts the measurement trends, including the resonance around 5.5 GHz, although the simulation predicts the resonance around 70 MHz lower than the measured resonance at 5.57 GHz. Figure 4.28 shows the difference between the straight surface waveguide and the surface waveguide with a 45° bend for both the simulated and measured cases.

4.3.2 90° Bend

Figure 4.29 shows the simulation structure of the surface waveguide with a 90° bend. Again, we could only use a surface waveguide with a total length of 10λ because the mesh in the bend requires too much memory to simulate a surface waveguide with a 100λ length. Also this structure has a non-zero radius of curvature, i.e., the inside of the bend is not a sharp corner, which decreases radiation in the simulation and more closely matches the geometry of the plumbing coupler used to join the two halves of the surface waveguide. Figure 4.30 shows the surface waveguide with a 90° bend. The straight plumbing coupler on the surface waveguide was simply replaced by a plumbing coupler with a 90° bend. The center of the waveguide still rests on the foam support in the chamber to prevent sagging. Additionally, since we expected a large amount of radiation from the bend, we made sure to point the maximum expected radiation away from the measurement equipment, which is largely conductive, to reduce error in the results caused by reflections.
Figure 4.23: Magnitude of reflection parameter for a surface waveguide with 45° bend.

Figure 4.24: Phase of reflection parameter for a surface waveguide with 45° bend.
Figure 4.25: Magnitude of transmission parameter for a surface waveguide with 45° bend.

Figure 4.26: Phase of transmission parameter for a surface waveguide with 45° bend.
Figure 4.27: Closer look at phase of transmission parameter for a surface waveguide with 45° bend.

Figure 4.28: Loss between straight waveguide and a waveguide with a 45° bend.
Figure 4.29: Simulation structure of surface waveguide with a 90° bend.

Figure 4.30: Surface waveguide with a 90° bend.
Figures 4.31 and 4.32 show the magnitude and phase of the reflection coefficients, respectively. The simulated reflection coefficients again suffer from inaccuracies as a result of a relatively coarse mesh density limited by computational resources. In the measured cases, as in the 45° bend case, there is little difference in the reflection coefficient magnitude and phase between the straight surface waveguide and the surface waveguide with a 90° bend. The phase of the reflection coefficient has a slight offset, which is likely a slight calibration error. Figures 4.33 - 4.36 show the transmission coefficient and the bend loss created by inserting the 90° bend. Figure 4.33 shows the magnitude of the transmission coefficient. The simulations and measurements appear to be highly correlated. To better see the accuracy, we compare this with Figure 4.36, which shows the difference between the straight surface waveguide and the bent surface waveguide. The simulated and measured bend loss are relatively close, but the simulated bend loss fluctuates much more than the measured case. Figures 4.34 and 4.35 show that the phases of the simulated and measured transmission coefficients are very close. The slight change in offset seen in Figure 4.34 indicates that the bent plumbing coupler is slightly longer than the straight plumbing coupler.

4.4 Surface Waveguide in Lossy Media

In this section, we will discuss the effect of lossy media on the surface waveguide. We were unable to simulate this section because the MOM-solver is especially memory-inefficient at simulating large dielectric volumes and requires large amounts of RAM in the simulating computer to simulate such a structure. Additionally, it was unknown how materials that were much more lossy than the surface waveguide would affect the wave because, in addition to causing attenuation expected for a wave propagating through lossy media, it could possibly cause the surface wave to radiate. We extrapolate the measurements taken by Fano in [27] to find the sand to have a permittivity of approximately 3 for the purposes of calculating the guided wavelength and a loss tangent of approximately 0.1.

We tested the effect of sand on surface wave propagation by creating a box of dimensions 24 cm by 24 cm by 30.5 cm, which is $\geq 3\lambda_g$ where $\lambda_g$ is
Figure 4.31: Magnitude of reflection parameter for a surface waveguide with 90° bend.

Figure 4.32: Phase of reflection parameter for a surface waveguide with 90° bend.
Figure 4.33: Magnitude of transmission parameter for a surface waveguide with 90° bend.

Figure 4.34: Phase of transmission parameter for a surface waveguide with 90° bend.
Figure 4.35: Closer look at phase of transmission parameter for a surface waveguide with 90° bend.

Figure 4.36: Loss between straight waveguide and a waveguide with a 90° bend.
Figure 4.37: Measurement setup when adding lossy media around the surface waveguide. The plastic box has two holes in either end where the surface waveguide goes through and the box is filled with sand, which requires its own PVC support.

the wavelength in sand in each cross sectional dimension beyond the surface waveguide. We chose these dimensions so that the box of sand would encompass roughly 50% of the surface wave power [2]. Figure 4.37 shows the completed box and surface waveguide setup. The box was approximately $5.3\lambda_0$ in length along the surface waveguide because that was a practical length for the volume and weight of sand required to fill the box. We bought a plastic box off the shelf and drilled holes in either end and mounted the box on the surface waveguide. As shown in the figure, the box is not mounted symmetrically on the surface waveguide. This affects the match on either end since the structure is greatly asymmetric, but should not greatly affect the transmission coefficient except for the expected case of a different surface impedance along the surface waveguide. Although the box was fairly flush against the surface waveguide surface, we used packing tape to form a seal around each hole so that sand would not leak into the chamber. We then filled the box with sand to take the measurement.

Figures 4.38 - 4.42 show the measurement results. In this case, we are comparing the measurement results of the empty box on the surface waveguide with the results for the box full of sand. The sand does not appear to
change the reflection coefficient drastically, as shown in Figures 4.38 and 4.39, although the sand increases the reflection slightly, which intuitively makes sense since there is a sudden impedance change. The sand introduces more noise into the measurement, although the noise is not as bad as it appears from Figure 4.39. The drastic variations below 4.5 GHz are caused by the wrapping of the phase from $-180^\circ$ to $180^\circ$ so slight variations in that region show drastic changes on the graph. Figures 4.40 and 4.41 show the transmission coefficient magnitude and phase, respectively. While the phase only has a slight offset while going through the sand, the magnitude has a larger change. The loss introduced is non-linear with frequency; there is a peak in transmission around 4.8 GHz rather than a near-linear decrease in transmission in the case with no sand. We do not think that this is a function of the shape of the box of sand, which is $6.65\lambda$ by $6.65\lambda$ by $8.46\lambda$ at 4.8 GHz, but rather a function of the dielectric properties of sand, especially the sand loss. Figure 4.42 shows the insertion loss of the sand, which varies between 2 dB and 7 dB. This measurement indicates that long distance transmission in sand is likely not feasible since there is much more loss than there is in air (note that most of the previous loss was a function of launching inefficiencies). This intuitively makes sense because the benefit of using a surface waveguide in air is that the wave travels mostly in air, which has less loss than copper. When the surrounding material is replaced by air, waves that travel mostly inside the waveguide have the advantage.

4.5 Surface Waveguide Near Lossy Media

In addition to knowing how a surface wave will behave when propagating through a lossy medium, it is also useful to know how the surface will behave when propagating near a lossy medium. We expect the effect of the sand to be drastic very near the surface waveguide because, not only does it introduce loss directly for the surface wave propagating through the sand, it also introduces severe asymmetries which should excite higher order modes which are then quickly attenuated. We were unable to create a simulation in FEKO, which is a method of moments solver and highly inefficient at simulating dielectrics. Figure 4.43 shows the measurement setup in the lab. We used the same box to hold the sand as the box that held the sand while
Figure 4.38: The effect of sand surrounding the surface waveguide on the reflection coefficient magnitude.

Figure 4.39: The effect of sand surrounding the surface waveguide on the reflection coefficient phase.
Figure 4.40: The effect of sand surrounding the surface waveguide on the transmission coefficient magnitude.

Figure 4.41: The effect of sand surrounding the surface waveguide on the transmission coefficient phase.
measuring the wave propagation through the sand. The sand is placed closer to one coaxial launcher than other for ease of measurement and to enable easily changing the distance from the box of sand to the waveguide. We first took a measurement where the box of sand is touching the surface waveguide. Then, we took measurements for increased distances between the edge of the surface waveguide and the box of sand. Figures 4.44 and 4.45 show the magnitude and phase of the reflection coefficient for the box of sand at varying distances from the waveguide. The no sand measurement is a measure of the unmodified straight surface waveguide and is intended to be used as a reference. The 0λ distance measurement is where the box of sand is touching the surface waveguide. Since the sides of the box of sand are slightly curved, the box is only touching the surface waveguide at one point. From these two graphs, it is evident that the box of sand does not have a significant effect on the reflection parameter. In the phase plot, slight variations in the phase around −180° cause the phase to be wrapped around to 180°, which is not a large discontinuity as it appears on the graph.

Figures 4.46 - 4.48 show the transmission coefficient magnitude and phase
Figure 4.43: Measurement setup for testing the effect of sand near the surface waveguide.
Figure 4.44: The effect of sand at varying distances from the waveguide on the reflection coefficient magnitude. Note that $\lambda$ is the free space wavelength.
Figure 4.45: The effect of sand at varying distances from the waveguide on the reflection coefficient phase. Note that $\lambda$ is the free space wavelength.

and the magnitude of the insertion loss caused by the box of sand. Notice that the trend is different in this case from the trend of the propagation through sand. This trend more closely resembles the propagation of the surface waveguide. Also notice that the surface wave has much more loss when the sand is beside the surface waveguide rather than surrounding it. We think that the extra loss comes from severe asymmetry around the surface waveguide exciting higher-order modes, which would be quickly attenuated. Also, note that by adding more loss to the line, we were able to achieve the trend predicted in simulation for the plain surface wave transmission line, where the transmission coefficient through the line increases to a maximum around 5.3 GHz.
Figure 4.46: The effect of sand at varying distances from the waveguide on the transmission coefficient magnitude. Note that $\lambda$ is the free space wavelength.
Figure 4.47: The effect of sand at varying distances from the waveguide on the transmission coefficient phase. Note that $\lambda$ is the free space wavelength.

Figure 4.48: Transmission loss of the sand at varying distances from the waveguide. Note that $\lambda$ is the free space wavelength.
CHAPTER 5
CONCLUSIONS

5.1 Conclusions

In this thesis, we have shown that we can create meaningful simulations for surface waveguides and simple metallic structures using FEKO, a method of moments solver. For simple geometries, we can create complete simulations of the structure for structures around $100\lambda$ in length. For more complicated geometries, we can create simulations with shorter surface waveguides to see the trend as a parameter is varied. It is important to have important features, especially radiating features, $10\lambda$ or more apart to have meaningful results. Using the simulation techniques, we have developed a practical surface waveguide. We have demonstrated that larger radius surface waveguides have a higher transmission coefficients than smaller radius surface waveguides through simulation. We found that finite mesh density tends to affect the reflection coefficient accuracy more than the transmission coefficient accuracy. We found that tapering the surface waveguide from a large radius, which has advantageous transmission properties, to a smaller radius waveguide, which can be measured with existing equipment does not significantly affect the transmission coefficient but does require a larger mesh density than similar untapered structures.

Additionally, we have made experimental verifications and discoveries. We showed the effectiveness of larger coaxial launchers compared to the inefficient, smaller, simpler launchers. We have shown that bends have a significant detrimental impact on the transmission coefficient including the ability to excite a radiating mode in the $45^\circ$ case. Additionally, we have shown that transmitting surface waves in lossy media such as sand is a lot less practical, in most cases, than traditional waveguides (where the wave travels on the interior of the waveguide), because the loss in sand is very high.
5.2 Summary of Contributions

In addition to verifying existing theory, we have made several contributions that help design surface waveguides. First, we have developed flexible simulations that can be used to design surface waveguides. Additionally, we have discussed alternative methods to predict the behavior of surface waveguides through simulations, such as looking at a surface waveguide model with a shorter surface waveguide. We have created a surface waveguide that is practical to measure by evaluating the effects of a surface waveguide taper and a relatively small (\(~1\lambda\) outer radius) flared coaxial launcher which has the relatively steep opening half-angle of 45°. We have designed and characterized simple compact launchers which can be used to create a multi-purpose structure with water or gas flowing inside the structure and a surface wave on the outside of the structure. Additionally, we have examined the effect on lossy materials on surface waveguides and have found lossy materials to be undesirable.

5.3 Future Research

There are several directions to take future research. It could be useful to derive several cases for a surface waveguide in lossy media such as sand, if possible. This could be expanded to layered media such as switching from a sand to air surface waveguide boundary. Simulations need to be developed for large structures with fine features including small, possibly lossy, dielectric regions. Additionally, it could be useful to optimize small launchers for efficiency rather than simplicity, such as we did in our investigations.
REFERENCES


