IEEE TRANSACTIONS ON ANTENNAS AND PROPAGATION, VOL. 62, NO. 5, MAY 2014

A Two-Port Model for Antennas in a Reverberation Chamber

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Abstract—We present an improved model for characterizing fundamental antenna parameters based on the concept of a passive realizable two-port network. The proposed model is consistent with circuit theory-based approaches and allows for quantitative description and subsequent experimental determination of antenna efficiency, loss, and mismatch. A derivation of the model is given along with a discussion of the implications of the model. In addition, we discuss how the model behaves in a low-loss reverberation chamber, and give analytical and numerical methods for estimating the parameters of the model in a low-loss environment. Derivation and application of the model is carried out in the context of a reverberation chamber, but the results are potentially applicable to an arbitrary environment.

Index Terms—Antenna efficiency, antenna model, environment characterization, parameter estimation, radiation efficiency, reverberation chambers, two-port networks.

I. INTRODUCTION

In modern communication systems, the precise quantification of losses and mismatches of an antenna is a serious challenge. Even for classical antenna designs with accepted analytical models, empirical validation of basic parameters is difficult. With the advent of greater computing resources, numerical modeling of antennas has become more accessible, but even sophisticated numerical models inherently involve approximations and are prone to error when used for applications for which they were not originally intended. The situation of antenna characterization is even more cumbersome in the presence of manufacturing defects and elaborate designs in complex environments. Unknown antenna loss, efficiency, and impedance mismatch characteristics propagate into any data acquired with a real antenna and can be the source of significant error. We present an improved model for antennas that accounts for fundamental antenna characteristics including most imperfections. The model is robust, and can serve as a first order correction for many environments. It is composed of a passive, two-port network for which the parameters can be obtained or estimated from experimental data.

The modeling of an antenna as a two-port network is not new. In [1], Rogers, et al. show an antenna as a two-port network for use in calculations related to communications channels. The model shown here is of similar nature, but approached from a different perspective.

Similar work has also been done by She, where it is referred to as “loss evaluation” [2]. In She’s method, certain antenna parameters can be calculated by determining the individual losses of each antenna component. She’s model works well for antennas in which the individual loss components can be easily determined. A similar calculation approach can be used to determine the parameters for our model.

After the parameters of the model network are determined or estimated in one environment, the results should be able to serve as a first-order correction when applied to the same antenna in an arbitrary environment. The model is designed to be general, and apply to the full range of practical antennas. The model proposed here is consistent with other models that rely on circuit theory to account for the parameters [1], [3]–[5].

Following the basic derivation in Section II, we describe the characteristics of each of the network’s parameters. In some instances we will be able to determine or estimate the parameter, while in others we can only describe its characteristics and establish bounds for its value.

Sections III and IV outline the equations governing the two-port model and each of its parameters. For simplicity, the case where there is only one antenna in the environment will be discussed first. Section IV provides details on how the equations change if two antennas are present.

The first application of the model will be in Section V where we detail one possible way to estimate the model network parameters if the antenna can be placed in a very low-loss environment. From this estimation, a calculation of the lower bound of transmitting and receiving efficiency is possible. This assumption is the same as assuming the walls of the reverberation chamber have a small amount of loss.

Section VI shows a validation of this particular solution of the model network. Antenna efficiency will be calculated from measured data, and compared to theoretical calculations. These calculations and comparisons allow us to examine the capabilities of the two-port model and identify some of the weaknesses or inaccuracies that exist.

II. ANTENNA MODEL AND NOTATION

The two-port model developed here was first proposed by Ladbury and Hill in [6], where a partial derivation was given.
This approach simulates a real antenna [Fig. 1(a)] as an ideal antenna connected to an unknown two-port network [Fig. 1(b)]. This ideal antenna has three characteristics: it is lossless as both a transmitter and a receiver, perfectly matched when used as a transmitter in free space, and retains all pattern and scattering characteristics from the original antenna. The imperfections of the real antenna (mismatch and loss) are incorporated in the unknown network, while the scattering and pattern characteristics are not. As with any other two-port network, there are six parameters that need to be considered: the intrinsic impedance of each port, the complex reflection coefficient of each port, and the transmission/coupling coefficient between each port. Obtaining the value of all six parameters enables one to calculate several antenna parameters.

Though the model we outline is restricted to two ports, a multiport model would be more complete because it would allow us to account for the scattering and pattern characteristics. In a multiport model, one port would represent the connector on the antenna, and each remaining port would represent a possible mode or ray path inside the environment or reverberation chamber [7]. This approach would be extremely difficult to model mathematically, as the number of possible ray paths inside a highly reflective environment is very large. Instead of having one port for each ray path, this model is simplified by accounting for the overall statistical nature of a highly reflective environment. This simplification is similar to placing an ideal antenna in an ideal reverberation chamber [8]. This model also assumes that the ideal antenna maintains the same physical size, pattern, and scattering characteristics of the real antenna.

The network is attached to the ideal antenna via a virtual port, denoted in Fig. 1(b) as Port 1ß. For simplicity, we will assume that the intrinsic impedance \( Z_0 \) of Port 1ß is the same as Port 1, or \( Z_{01} = Z_{01ß} \). For most applications this is 50 \( \Omega \). We will also make the simplifying assumption that the real antenna, and therefore the two-port network, is reciprocal.

Fig. 2 depicts a signal flow graph representation of the two-port model. Here, \( T_1 \) represent the transmission from Port 1 into the environment, and \( R_1 \) represents the signal received from the environment by antenna 1. \( S_1 \) is the portion of the received signal from the environment that is reflected back into the environment, and \( \Gamma_1 \) is the input reflection looking into port 1. The 1 subscripts indicate the port or antenna number being referenced by the equation. \( S_{11}^E \) represents the amount of energy reflected back to the antenna from the environment. In this case, the superscript \( E \) is used to denote “environment.” The parameters of the environment (reverberation chamber or otherwise) will be discussed in greater detail in Section IV. Since Port 1ß is a virtual port, we are free to define the “location” of the reference port in a way that is mathematically convenient. In other words, we can attach a lossless length of matched transmission line so that we have control of the phase of either the transmission coefficient \( \Gamma_1 \) or the reflection coefficient \( S_1 \) (but not both simultaneously). We choose to set the phase of \( T_1 \) to zero. This, combined with the assumptions of reciprocity and equal port impedances give us \( T_1 = R_1 \), and both parameters are real.

To begin the analysis of the model network, we can examine how such a network would represent an antenna in free space. If the ideal antenna from Fig. 1(b) is placed in free space, then there will be no reflection from the environment, and therefore the \( S_{11}^E \) will be zero. This would be equivalent to replacing the ideal antenna in Fig. 1(b) with a non-reflecting load. This gives a reflection coefficient at port 1 of \( \Gamma_1 \), or the reflection coefficient of the network when terminated by a non-reflecting termination. For an incident power of \( P_{inc} \), the reflected power will be \( P_{inc} |\Gamma_1|^2 \), the net power into the antenna will be \( P_{inc} (1 - |\Gamma_1|^2) = P_{inc} m_T \) where \( m_T = 1 - |\Gamma_1|^2 \) is the transmission mismatch factor. The power into the ideal antenna, which is the transmitted power, will be \( P_{inc} |T_1|^2 \).

From these basic definitions, equations for the efficiency of the real antenna can be developed. According to IEEE 145-1993, radiation efficiency is defined as: “The ratio of the total power radiated by an antenna to the net power accepted by the antenna from the connected transmitter” [9]. We now introduce two terms. The transmission efficiency of an antenna is simply the radiation efficiency for an antenna in free space [9]. Similarly, we define the transmission efficiency of a two-port network depicted in Fig. 2 as the ratio of the power dissipated by a
reflection-less load attached to the virtual port of the network to power accepted by the network at port 1, which is analogous to the radiation efficiency outlined in [9]. This efficiency (denoted as $\eta_{T1}$) is equivalent to the radiation efficiency of the real antenna, or

$$\eta_{T1} = \frac{T_1^2}{1 - |\Gamma_1|}. \quad (1)$$

An optional subscript has been added to $\eta_{T}$ to denote the proper antenna of interest. In most cases, we will omit this subscript.

When used as a receiving antenna, we will assume that the receiving instrument attached to Port 1 also represents a non-reflecting load. This assumption simplifies the description, but imperfect instruments can be included with a few modifications to the equations. Receiving characteristics of an antenna are more complicated than the corresponding transmitting characteristics. In the receiving case, every incident plane wave will result in scattering in all directions. The portion of the incident energy that is received by the antenna and reflected by the two-port network will also be scattered (retransmitted) in all directions. As a result, discussing the receiving reflection coefficient of the antenna is somewhat ambiguous, and phase angles have very little meaning. However, referring to the model depicted in Figs. 1 and 2, we can define $P_{\text{rec}}$ as the power received by an ideal antenna for a given plane-wave excitation, the power reflected by the two-port network and the antenna is $P_{\text{rec}} |S_1|^2$, net power received by the antenna will be $P_{\text{rec}}(1 - |S_1|^2) = P_{\text{rec}} m_{R1}$ where $m_{R1} = 1 - |S_1|^2$ is the receiving mismatch factor. The power into the measurement instrument, which is the measured power, will be $P_{\text{rec}} |R_1|^2$.

Conceptually, modeling the receiving antenna inside an ideal reverberation chamber is equivalent to connecting Port 1/3 of the two-port network to a random source that has the same statistical properties as the reverberation chamber. In this case, the signal measured at Port 1 will have the same statistical properties, but will be scaled by $R_1$. This result is consistent with [8] and [10].

These definitions lead to an equation for the receiving efficiency. Unfortunately, IEEE 145-1993 does not define “receiving efficiency” [9]. We define the receiving efficiency of the two-port network to be the ratio of the power dissipated by a reflection-less load attached to Port 1 of the network [refer to Fig. 1(b)] to the amount of power accepted by the network from a source (of impedance $Z_0$) connected to port 1/3. This can be expressed algebraically as

$$\eta_{R1} = \frac{|R_1|^2}{1 - |S_1|^2}. \quad (2)$$

The receiving efficiency of the antenna being modeled can be expressed as a ratio of the total amount of power available to the antenna (over $4\pi$ steradians) to the power accepted by the antenna. The efficiency expressed in (2) is equivalent to the efficiency of the real antenna.

Because the model network is bound by the strict realizability constraints of a passive two-port network, the transmission efficiency, receiving efficiency, and the transmission and receiving mismatch factors must all fall between 0 and 1 (inclusive).

The definition of transmission efficiency (and thus radiation efficiency) used here is consistent with the IEEE 145-1993 standard definition of radiation efficiency. This definition does not consider power reflected from the antenna due to an impedance mismatch in the efficiency calculation. Therefore, if 1 W is incident on the input port of an antenna, 0.5 W is reflected, and the remaining 0.5 W is radiated, the antenna is considered to be 100% efficient.

From the above analysis, a reasonable first step to model a real antenna is to assume that $\Gamma_1$ for our two-port network is the same as the free-space reflection coefficient $\Gamma_{1,FS}$ for a real antenna in free space. If the efficiency is known, we can use the efficiency relationship in (1) or (2) and the antenna’s reflection coefficient to determine $|T_1|$, but efficiency is not generally known or provided by antenna manufacturers, and can be difficult to measure. One possibility is to estimate the efficiency with a reverberation chamber [8], [10].

For a reciprocal network, we know that $T_1 = R_1$. However, we cannot assume that $\Gamma_1 = S_1$. We will show in Section III that for small losses, $\Gamma_1$ will be similar to $|S_1|$, but they may be significantly different in higher loss networks. In the case where the network is lossless, $|\Gamma_1| = |S_1|$.

With the model defined, the challenge of estimating each network parameter in the two-port network is apparent. Using the limits of physically realizable networks to bound the problem, we can analytically determine the allowable ranges for the parameters.

### III. SINGLE ANTENNA EQUATIONS

With the basic definitions outlined in Section II, a series of equations can be developed that will allow bounds to be established for the individual parameters.

For simplicity, these equations will be shown in the context of a single antenna inside a reverberation chamber. Therefore, $S_{11}^{RC}$ and network parameters with a “1” subscript will be used. However, we allow for the possibility of different or multiple antennas which would require changing the subscript from “1” to “1...n.” The details of modeling multiple antennas is discussed in Section IV.

In an ideal reverberation chamber, the electric field is statistically uniform throughout the working volume of the chamber. This uniformity is achieved by changing the boundary conditions using a rotating metal paddle. In a fundamental sense, stirring creates a new electromagnetic environment that contains a distribution of energy physically different than the previous environment. A single chamber can be stirred to create hundreds of unique but statistically equivalent electromagnetic environments. Models indicate that the ensemble of such environments yields a statistically uniform environment wherein electromagnetic parameters are independent of location and/or orientation in the working volume of the chamber [8].

If $S_{11}^{RC}$ is measured inside an ideal reverberation chamber, it will have a uniformly distributed phase ($0$ to $2\pi$), its magnitude will be independent of the phase, and it will have an unknown distribution bounded between 0 and 1. If the losses are high, such that the average reflection coefficient is much less than 1, a Rayleigh distribution can be used to approximate the distribution of the magnitude. When expressed as a complex quantity,
the real and imaginary components are independent and normally distributed with zero mean and identical variances [10].

This concept of placing an antenna into a reverberation chamber which is described by the random chamber reflection coefficient \(S_{11}^{RC}\) is analogous to terminating the unknown two-port network with a load that has a random reflection coefficient \(S_{11}^{RC}\), which will have the statistical properties of the reverberation chamber. If the reflection coefficient of Port 1, \(S_{11}\), is examined while the random load is still attached to port 1/2, the result will also be random.

With an idea of how the model performs in an ideal reverberation chamber, a series of equations can be developed that express these characteristics. These equations are rooted in waveguide junction theory [11], and will be centered around the reflection coefficients \(S_{11}^{RC}\) and \(S_{11}\).

Waveguide theory [11] already provides an expression for \(S_{11}\), given \(S_{11}^{RC}\):

\[
S_{11} = \Gamma_1 + \frac{T_1 R_1 S_{11}^{RC}}{1 - S_{11} S_{11}^{RC}} = \Gamma_1 + \frac{T_1 R_1}{S_{11}^{RC} - S_{11}}. \tag{3}
\]

If we write \(S_{11}^{RC}\) in polar form as \(S_{11}^{RC} = re^{i\phi}\) we can rewrite (3) as

\[
S_{11}(r, \phi) = \Gamma_1 + \frac{T_1 R_1 re^{i\phi}}{1 - S_1 r e^{i\phi}} = \Gamma_1 + \frac{T_1 R_1}{r^{-1} e^{-j\phi} - S_1}. \tag{4}
\]

Utilizing the characteristics of an ideal reverberation chamber, we can assume that the magnitude and phase of \(S_{11}^{RC}\) are independent of each other. Given this, it is useful to examine the effects of changing the magnitude and phase of \(S_{11}^{RC}\) separately. If the magnitude is held constant while the phase is varied, this would be conceptually equivalent to attaching a sliding mismatch to the two-port network. This problem has been well characterized in [11]. By use of this sliding mismatch concept and (3), a \(S_{11}^{RC}\) circle (of radius \(r\), centered at the origin) is transformed to a new circle with the following radius \(r_1\) and center \(C_1\):

\[
r_1 = \frac{|T_1 R_1 S_{11}^{RC}|}{1 - |S_{11}^{RC}|^2} \tag{5}
\]

and

\[
C_1 = \Gamma_1 + r_1 |S_{11}^{RC}| e^{j(\phi_{31} + \phi_{11} - \phi_{11}^{RC})} \tag{6}
\]

where \(\phi_{31}\) represents the phase component of the indicated parameter. Applying our simplifying assumptions of reciprocity, matched port impedance, and zero phase for \(T_1\), (6) simplifies to

\[
C_1 = \Gamma_1 + r_1 S_{11}^{RC} \tag{7}
\]

where the bar over \(S_1\) denotes the complex conjugate. If \(C_1\), \(\Gamma_1\), \(r_1\), and \(r\) can be measured or estimated, then it is possible to determine \(S_1\) (both magnitude and phase) as

\[
S_1 = \frac{(C_1 - \Gamma_1)}{r_1 r}. \tag{8}
\]

We will make use of this in Section III-C. For the very special case of complete reflection (\(|S_{11}^{RC}| = r = 1\), these equations can be further simplified. From (5) we get

\[
r_1 = \frac{|T_1 R_1 S_{11}^{RC}|}{1 - |S_{11}^{RC}|^2} = \frac{|T_1 R_1|}{1 - |S_{11}|^2} = \frac{R_1}{r - S_1}. \tag{9}
\]

and (8) simplifies to

\[
S_1 = \frac{(C_1 - \Gamma_1)}{\eta r}. \tag{10}
\]

These simplifications give some hope that we may be able to estimate \(S_1\) from measured data in a realistic environment.

Fig. 3 shows the transformation of several circles from the \(S_{11}^{RC}\) plane (a) to the \(S_{11}\) plane (b). In these examples, the magnitude of \(S_{11}^{RC}\) was stepped from 0.0 to 1.0 in 0.1 increments. The network parameters are set to \(|T_1| = 0.8\) at a phase angle of 45°, \(T_1 = R_1 = 0.404\) with a 0° phase angle, and \(|S_1| = 0.8\) at a phase angle of 315°. These parameters result in both the transmitting and receiving efficiencies being 1.

A. Estimating \(\Gamma\)

Of the parameters in the two-port antenna model, \(\Gamma\) is the easiest to estimate. It is also the parameter we can determine with the lowest uncertainty.

We proceed by evaluating the expected value of \(S_{11}\). For this analysis, we assume the phase \(\phi_{11}^{RC}\) of \(S_{11}^{RC}\) is uniformly distributed. To keep this analysis general, we assume that \(S_{11}^{RC}\) has an arbitrary distribution of \(f_r(r)\), where \(0 \leq r \leq 1\). For simplicity, we will let \(|S_{11}^{RC}| = r\) and let \(\phi_{11}^{RC} = \phi\). Because \(\phi_{11}^{RC}\) and \(|S_{11}^{RC}|\) are independent, the joint distribution can be written as

\[
f_{r,\phi}(r, \phi) = f_r(r) f_\phi(\phi) = \frac{1}{2\pi} f_r(r), \quad 0 \leq \phi < 2\pi, \quad 0 \leq r \leq 1. \tag{11}
\]

The expected value of the measured value of \(S_{11}\), which is a function of the complex random variable \(S_{11}^{RC}\) [as given in (4)] and the joint distribution given in (7), can be obtained by integration [12]:

\[
E(S_{11}) = \int_0^1 \int_0^{2\pi} S_{11}(r, \phi) f_{r,\phi}(r, \phi) r \, d\phi \, dr. \tag{12}
\]

This integral can be simplified:

\[
E(S_{11}) = \frac{1}{2\pi} \int_0^1 \int_0^{2\pi} f_r(r) S_{11}(r, \phi) r \, d\phi \, dr
\]

\[
= \Gamma_1 + \frac{T_1 R_1}{2\pi} \int_0^1 \int_0^{2\pi} f_r(r) \left[ \int_0^{2\pi} e^{-i\phi} S_{11} \, d\phi \right] r \, d\phi dr.
\]

(13)

The bracketed term in (13) integrates to 0 for \(0 \leq |r| \leq 1\) and \(S_{11} < 1\), which implies that

\[
E(S_{11}) = \Gamma_1. \tag{14}
\]
This validates the long held assumption that the free-space reflection coefficient of an antenna is equivalent to the ensemble average of $S_{11}$ observed in a reverberation chamber [10]. This validation is a powerful result that will allow measured $S_{11}$ data (assuming a Raleigh distributed magnitude and uniformly distributed phase) to be used in the calculation of the reflection coefficient ($\Gamma_1$) of the unknown two-port network. With a reverberation chamber, $S_{11}$ data can be measured at each stirrer position (e.g., 100 stirrer positions), then averaged to obtain an estimate of $\Gamma_1$.

There are some subtle implications of the results here that require some additional comments. Since the bracketed term in (13) integrates to 0 for all $\tau$ between 0 and 1, the mean of each circle in Fig. 3(b), (or any other $S_{11}$ circle generated using a sliding mismatch with uniformly spaced phase samples), is $\Gamma_1$, although (6) clearly indicates that the center of each circle is offset from $\Gamma_1$ unless $r$, $T$, and/or $S_1$ are zero. For example, in the case of Fig. 3(b), the circle generated for $\tau = 1$ results in an $S_{11}$ circle centered at the origin, but with a mean of $\Gamma_1$. This seems non-intuitive, but can be explained if we take $N$ uniform phase samples in the $S_{11}^{RC}$ plane, and transform those samples into the $S_{11}$ plane, as shown in Fig. 4. The higher density of points near $\Gamma_1$ is sufficient to shift the mean of each circle away from its center. In fact, the offset between $C_1$ and $\Gamma_1$ provides information about the magnitude and phase of $S_{11}$, as given by (10).

![Fig. 3](image1)

(a) shows a series of circles representing $S_{11}^{RC}$ data. (b) shows the same data transformed to represent $S_{11}$.

![Fig. 4](image2)

Data from Fig. 3(b) are re-graphed as individual points to show their distribution.

### B. Characteristics of $T$ and $R$

In Section II, we stated that the intrinsic impedance of both ports of the model network are assumed to be equal. This assumption coupled with the reciprocity of the network leads to $T_n = R_n$. The identical port impedances and reciprocal network do not mean the functions of the model network (e.g., transmitting and receiving efficiency) are equal. Differences in the $\Gamma_n$ and $S_n$ terms can cause the transmitting and receiving characteristics of the model network to be different.

In Section V we will show one example of how the $T$ and $R$ parameters can be determined. If the efficiency of the antenna being modeled is known and $\Gamma_1$ can be determined using (14), then $|T_1|$ can be determined using a version of (1):

$$|T_1| = \sqrt{|r_1| (1 - |\Gamma_1|^2)}.$$  \hfill (15)

### C. Characteristics of $S$

The parameter $S$ is the most difficult to quantify because it represents a process that is not readily accessible theoretically or experimentally. Instead of solving for a unique $S_n$ value, a range of possible values can be obtained. Utilizing information about the other three network parameters, possible $S_n$ values can be bounded. Since the random and statistical nature of reverberation chambers prevents us from determining an exact value, we will show a series of constraints that can be placed on the value of $S_n$. 

![Diagram](image3)
One such set of constraints are the strict realizability constraints. Any passive, linear, time-invariant network must satisfy all of these conditions. From waveguide theory [11], these constraints are simplified to account for identical intrinsic impedances. The three stipulations for a realizable network are

\[
0 < \frac{|T_1|^2}{1 - |\Gamma_1|^2} = \eta_{T1} < 1 \quad (16a)
\]

\[
0 < \frac{|R_1|^2}{1 - |S_1|^2} = \eta_{R1} < 1, \quad \text{and} \quad (16b)
\]

\[
\left(1 - |\Gamma_1|^2 - |T_1|^2\right)\left(1 - |S_1|^2 - |R_1|^2\right) - \Gamma_1 R_1 + T_1 S_1 > 0. \quad (16c)
\]

Note that the first two constraints are based on the transmitting and receiving efficiencies of the model network.

The realizability constraints combined with the equations for the other three parameters can considerably limit the possible values of \(S_1\). To further describe the value of \(S_1\), we can take advantage of reciprocity and obtain a range of possible \(S_1\) values through algebraic manipulation. The algebraic result will be a locus of solutions for \(S_1\), all located within a circle, and subject to the following constraints:

1. The distance from the center of the \(S_1\) circle to the origin of the complex plane is

\[
\frac{|\Gamma_1^1 T_1|^2}{1 - |\Gamma_1|^2} = |\eta_{T1}|. \quad (17a)
\]

2. The radius of the \(S_1\) circle will be

\[
1 - \frac{|T_1|^2}{1 - |\Gamma_1|^2} = 1 - \eta_{T1}. \quad (17b)
\]

3. The phase angle \(\phi_{S1}\) to the center of the \(S_1\) circle will be

\[
2\phi_{T1} - \phi_{S1} + \pi. \quad (17c)
\]

The \(2\phi_{T1}\) term in the (17c) becomes zero as a result of our assumption that the phase of \(T\) is zero.

Given these three sets of constraints, \(S_1\) can only be definitively determined in a few special cases. From (17b), the possible solutions for \(S_1\) can be reduced to a small circle if the efficiency is high. In cases where the efficiencies are low, there will be a large range of possible \(S_1\) values. In Section V-C, we will detail an iterative method whereby the \(S_1\) parameter can be approximated when the antenna is in a low-loss/highly reflective environment.

If the model network is lossless (or assumed to be), then the efficiencies shown in (17a) and (17b) will be equal to 1, and the third constraint, (17c), becomes \(\Gamma_1^1 R_1 + T_1 S_1 = 0\), or \(S_1 T_1 = -\Gamma_1^1 R_1\), and since \(T_1 = T_1\), we end up with \(S_1 = -\Gamma_1^1\). When the set of three bounds are applied with \(\eta_{T1}\) and \(\eta_{R1}\) equal to 1, the range of possible \(S_1\) values shrinks to a single point. Assuming the network is reciprocal and the phase of the \(T\) and \(R\) parameters is zero, this becomes one of the few cases where \(S_1\) can be uniquely determined.

In cases where the model network is nearly lossless, the efficiencies will be high, causing the radius of possible \(S_1\) values to be very small. For these cases, it may be reasonable to assume a value for \(S_1 \approx -\Gamma_1\). If the network has more loss, the radius of solutions will grow. Assuming a value for \(S_1\) in these cases could lead to large errors in subsequent applications.

IV. TWO-ANTENNA CASE

Up to this point, discussion of the two-antenna model has focused on a single antenna inside an arbitrary environment. However, many common reverberation chamber applications require two antennas be used. If we wanted to model the case where there are two antennas present in a reverberation chamber, the previous discussions could be expanded. Now, instead of a single two-port network representing the imperfections of a single antenna, two additional networks are required: one for the reverberation chamber and one for the second antenna. This set of networks can be thought of as three cascaded two-port networks, as shown in Fig. 5. Though we specifically speak about using a reverberation chamber (denoted as \(S_{\text{RCH}}\)), this method could be used with any environment (denoted as \(S_{\text{env}}\)). The two antennas present in the environment need not be similar because each will be represented by its own network.

The notation for two antennas in an arbitrary environment is shown in Fig. 5. The notation remains similar to the single antenna case, but the subscript of each parameter is adjusted to indicate the appropriate port.

Equations for the two-antenna case become significantly more complicated because three networks are being cascaded: antenna-environment-antenna. In this configuration, the goal of the two-port model is the same: to remove the effects of the antennas, allowing a better characterization of the environment. This becomes more difficult when two antennas are present in the chamber, as the effects of both antennas must be removed.

To describe \(S_{11}\) of a series of cascaded networks, a modified version of (3) can be used. From (3), additional terms are included to account for the environment, the scattering due to the antennas, allowing a better characterization of the environment. However, many common reverberation chamber applications require two antennas be used. If we wanted to model the case where there are two antennas present in a reverberation chamber, the previous discussions could be expanded. Now, instead of a single two-port network representing the imperfections of a single antenna, two additional networks are required: one for the reverberation chamber and one for the second antenna. This set of networks can be thought of as three cascaded two-port networks, as shown in Fig. 5. Though we specifically speak about using a reverberation chamber (denoted as \(S_{\text{RCH}}\)), this method could be used with any environment (denoted as \(S_{\text{env}}\)). The two antennas present in the environment need not be similar because each will be represented by its own network.

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To describe \(S_{11}\) of a series of cascaded networks, a modified version of (3) can be used. From (3), additional terms are included to account for the environment, the scattering due to the second antenna, and the interactions between the two antennas. The full expression for \(S_{11}\) is shown as

\[
S_{11} = \Gamma_1 + \frac{T_1 \left( S_{11}^{\text{RCH}} + S_{12}^{\text{RCH}} S_{12}^{\text{env}} \right) R_1}{1 - S_1 \left( S_{11}^{\text{RCH}} + S_{12}^{\text{RCH}} S_{12}^{\text{env}} \right)} \quad (18)
\]

\[
S_{21} = \frac{T_1 S_{12}^{\text{RCH}} R_2}{1 - S_1 S_{12}^{\text{RCH}} \left( 1 - S_2 S_{22}^{\text{env}} \right)} \quad (19)
\]

and is the result of algebraically solving for \(S_{11}\) from the set of three cascaded network matrices [11]. An examination of (3) shows that it is a special case of (18) where \(S_2 = 0\).
An equation for the signal that propagates through the environment can also be developed. Similar to (18), it is possible to compute $S_{21}$ (note the lack of superscripts) by solving for it from the set of cascaded network matrices. The result of this is (19), which accounts for the imperfections and scattering of both antennas, and represents the signal propagating from test port to port.

These two equations can also be used to provide expressions for $S_{12}$ and $S_{22}$, by simply changing the terms to apply to the opposite ports. The assumptions outlined in Section III for each individual two-port network still apply when they are cascaded.

From (18) and (19), we can see that the only four directly-measurable parameters are $S_{11}$, $S_{21}$, $S_{12}$, and $S_{22}$. From these, the only parameters we can directly estimate from the measured data are $\Gamma_1$ and $\Gamma_2$. Depending on the type of environment between the two antennas, we may be able to estimate the characteristics of the environment ($S_{11}^{RC}$, $S_{21}^{RC}$, $S_{12}^{RC}$, and $S_{22}^{RC}$).

If the two antennas are separated, such that they have no significant scattering or multiple interactions between them, the parameters of the two-port network would be bounded by the equations in Section III. With each antenna properly characterized by the two-port network, the characteristics of the environment can be obtained.

However, if the two antennas are strongly interacting with each other, the single antenna solutions cannot be applied. Antenna-to-antenna interactions can alter their respective radiating and receiving properties. If some assumptions are made about the environment (e.g., high- or low-loss), other methods could be used to find some of the antenna parameters. For example, if antenna efficiency is obtained using another method [13], [14], it could be used to help identify some of the parameters of the model network, and of the cascaded networks. This would then lead to approximations about the parameters of the environment.

V. CALCULATING A LOWER BOUND FOR ANTENNA EFFICIENCY

One method for estimating the four parameters of the model network can be derived if we assume a very low-loss environment. This derivation begins with the efficiency equations. By manipulating these equations, the four network parameters can be estimated. In Section III, equations were developed assuming the reverberation chamber was ideal. If we want to relax those assumptions, and estimate the network in a low-loss environment, we need to understand how the change in environment will affect the results of the two-port model.

We define a low-loss environment to be one where the power absorbed by the walls is approximately the same or less than the power absorbed by the antennas. A reverberation chamber is well suited for this application because it can be operated at frequencies where the losses are low and the amount of reflected energy [due to the walls and stirrer(s)] is high. The degree to which this assumption is true in practice is what makes the following calculation a lower bound of antenna efficiency.

In cases where the environment has sufficient loss (<70%), the statistical distribution of reverberation chamber has been studied in detail [8], [10]. These showed that, under some general conditions (well-matched antennas, over-moded chamber, and sufficient loss), that the phase of any measured $S$-parameter will be uniformly distributed over all possible phases (0 to $2\pi$) and the magnitude will have an approximately Rayleigh distribution. The magnitude cannot have a true Rayleigh distribution, since a Rayleigh distribution has a nonzero probability that a magnitude sample can exceed the unit circle. Even so, the Rayleigh model proved to be useful even for predicting extrema, as shown by close agreement between theoretical and measured peak-to-average ratios. This appears to be consistent as long as the power absorbed by the chamber is much greater than the power absorbed by antennas (or anything connected to the antenna) in the chamber, because under high-loss conditions, measured values near the unit circle are very unlikely.

As chamber losses decrease (which generally occurs as frequency decreases), the Rayleigh model becomes less and less appropriate. This is illustrated by a much lower peak-to-average ratio. So, even though average coupling continues to increase as loss decreases, the peak is more constrained. However, the phase still appears to be uniformly distributed. In the most extreme case of a lossless chamber containing a single transmitting antenna, the reflection coefficient must be 1 (all power radiated by the antenna must also be received by the antenna in a lossless chamber). In this case, we also assume that the phase is uniformly distributed.

Based on these observations, we will make some general predictions regarding the behavior of any of the reverberation chamber $S$-parameters ($S_{11}^{RC}$, $S_{21}^{RC}$, $S_{12}^{RC}$, and $S_{22}^{RC}$). For all conditions we will assume a uniform phase distribution. For higher chamber loss conditions, we will assume that the magnitude of the chamber $S$-parameters has a truncated Rayleigh distribution, constrained between 0 and 1. As loss decreases, the probability density near 1 increases, which means that the probability of observing a random sample near 1 increases. As loss decreases further, the probability density shifts even more towards 1. Regardless of the nature of the true distributions, we will simplify the assumptions to the following for the low-loss case:

1) The phase is uniformly distributed.
2) There is a good chance that, given several hundred samples of the magnitude, that at least three will be “close” to 1.

Unfortunately, we cannot directly measure the reverberation chamber $S$-parameters, which means the distributions cannot be directly observed. However, we can take measurements at a few frequencies where the antennas appear to be well matched. Fig. 6(a) shows an example of 1000 samples taken at 265.0 MHz with a dual-ridged horn antenna. Fig. 6(b) shows a histogram of the magnitude from Fig. 6(a).

If we reexamine the case of a lossless chamber containing a single transmitting antenna, we know that $|S_{11}^{RC}| = 1$, and assume a uniformly distributed phase, so all samples will fall on the unit circle, and the analysis above in Section III is appropriate. We can now relax the assumptions of losslessness and assume instead that a few (at least 3) samples have a magnitude near one. Now, only a few of the samples fall near the unit circle, and the analysis from Section III is more complicated. However, if we take the smallest circle that contains all of the samples, this should be very close to the unit circle. If we transform all of the points and the enclosing circle to $S_{11}$ using (3), then the transformed circle will still contain all of the transformed points, and
There currently is no way to experimentally verify the receiving efficiency of an antenna. Consider a plane wave incident on an antenna under test (AUT). As the plane wave reaches the antenna, part of the signal will be received by the antenna, and part of it will be scattered by the antenna. We can measure the portion of the signal that is received by the antenna, but cannot yet measure the portion of the signal that is scattered by the antenna.

### B. Transmission Efficiency

Once the receiving efficiency is known, calculating the transmission efficiency requires a few more steps.

Mathematically, transmission efficiency is defined by (1). As with the receiving efficiency, not all of the required parameters are known. However, it is possible to write the transmission efficiency in terms of receiving efficiency. Beginning with (1), the denominator can be replaced with $R_1$ (due to reciprocity). The $R_1$ denominator can then be expressed in a version of (2), solved for $R_1$

$$\eta_T = \frac{\eta_R \left(1 - |S_{11}|^2\right)}{1 - |\Gamma_1|^2}.$$  \hspace{1cm} (20)

In previous sections, methods for estimating each of the parameters in (20) have been described.

Recall that in Section III, we discussed that the true value of $\Gamma_1$ is very difficult to calculate. Usually, $\Gamma_1$ can be estimated because it is dependent on the radius of the smallest circle that bounds all of the measured $S_{11}$ data.

### C. Calculating Efficiency—An Iterative Approach

From (2) and (20), we see that in order to calculate the transmission efficiency, we first need to find the receiving efficiency. Receiving efficiency is determined by $\Gamma_1$, which is the radius of a circle that bounds all data points in the $S_{11}$ plane. In the case of a reverberation chamber, each data point could represent the complex reflection coefficient measured at a different paddle position.

When the chamber is lossless ($S_{11}^{RC} = 1$), measured data values will be within the unit circle, and calculated $S_{11}^{RC}$ values will be on the unit circle. Thus, the radius of the minimum circle will be 1. If the reverberation chamber is not lossless, and $S_{11}^{RC} < 1$, the data points will not extend to the unit circle. The more loss present in the chamber, the further from the unit circle the data may be. Similarly, as the loss increases, the radius of the minimum circle will decrease. This decrease in the radius will result in a lower estimate of receiving and transmission efficiency simply due to a change in the environment.

The other potential issue with the use of a minimum radius bounding circle arises when the AUT has a significant impedance mismatch. In this case, $S_{11}^{RC}$ may be appropriately distributed, but the impedance mismatch will cause the $S_{11}$ points to cluster close to the edge of the unit circle, as in Fig. 7. When a minimum radius circle is applied, the circle could cross the unit circle, resulting in a set of equations that cannot be solved.

One possible approach to circumvent this problem would be to determine the smallest bounding circle that encompasses all points and uses the unit circle as one of the three points in determining the minimum bounding circle. This is what we hope
to accomplish using an iterative method outlined below. This approach begins with a set of measured $S_{11}$ data. The easiest way to describe this iterative method is to show a sample calculation. The same set of data shown in Fig. 7 will be used for the sample calculation of efficiency. This data came from the measurement of a single dual ridged horn antenna inside a reverberation chamber. For Fig. 7 and the efficiency calculations, there are 1000 stirrer positions at a frequency of 468.8 MHz.

The iterative method begins by assuming the antenna is 100% efficient. This assumption, albeit incorrect, allows us to calculate an initial estimate of all four of the parameters in the two-port network. These four parameters will be used as a starting point for the iterative process. As the iterations progress, these four parameters will change to more reasonable values. Derived from the strict realizability constraints, the following three equations can be used to calculate the initial network parameters

\begin{align}
\Gamma_1 &= \langle S_{11} \rangle, \\
S_1 &= -\Gamma_1, \\
T_1 - R_1 &= \sqrt{1 - \Gamma_1^2}. 
\end{align}

Once the four network parameters are calculated, assuming the antenna is 100% efficient, the data are transformed to the $S_{01c}$ plane by use of (3), solved for $S_{11c}^{RC}$:

\begin{equation}
S_{11c}^{RC} = \frac{S_{11} - \Gamma_1}{T_1 R_1 + (S_{11} - \Gamma_1) S_1}.
\end{equation}

The goal with this transformation is to redistribute the points to have a more uniform phase. After this transformation is complete, the minimum radius bounding circle is found. Fig. 8 shows the original $S_{11}$ data transformed to the $S_{11c}^{RC}$ plane (notice the more uniform phase distribution). The solid red circle around the data is the minimum radius bounding circle.

Now that the minimum radius circle bounding all of the $S_{11c}^{RC}$ points has been found, the circle (not the data) must be transformed back to $S_{11}$ with (3).

Note that (22) contains several operations: reciprocal, scale, and offset. Each of these operations must be done on the circle as a whole. By use of a bilinear transformation technique, (3) can be computed using only the radius and center of the circle [11]. The key to this transformation is that if the circle is transformed (as well as all of the points), the transformed points will still all be within the transformed circle.

Fig. 9 shows the circle transformed back into the $S_{11}$ plane. From this point, the center and radius of the circle in the $S_{11}$ plane can be used to calculate a new $S_1$

\begin{equation}
S_{1,new} = \frac{1}{S_{11,\text{radius}}} (S_{11,\text{center}} - \Gamma_1)
\end{equation}
where $S_{1,\text{new}}$ is the new $S_1$ parameter calculated from the points of the circle, $S_{11,\text{radius}}$ and $S_{11,\text{center}}$ are the radius and center of the circle used to calculate the new $S_1$, respectively.

Unless the antenna being used is truly 100% efficient, this process may need to be repeated at least once more. This means calculating a set of new network parameters and repeating the $S_{1}$ and $S_{11}^{RC}$ transformations again, as described above. To repeat the process, a new $T_1$ must be calculated:

$$T_{1,\text{new}} = \sqrt{S_{11,\text{radius}}(1 - |S_{\text{new}}|^2)}.$$

As the process is repeated a second time, Fig. 10 shows the original measured $S_{11}$ data transformed into the $S_{11}^{RC}$ plane. To compare the results from iteration to iteration, compare Figs. 8–10.

A careful examination of Fig. 10 reveals that the minimum radius bounding circle has points that fall outside of the unit circle. In this example, the difference is very difficult to see, but it can be more pronounced if the antenna’s mismatch is more severe. While an exact mathematical explanation for this has not yet been determined, this appears to be an artifact of the processing that occurs only in intermediate iterations, and only for a small fraction of the data points. A probable cause for this artifact is that we are attempting to find parameters that would map points directly to the unit circle, even though all network parameters are not yet estimated. Further investigation is required to determine the cause of this anomaly. We would be more concerned if any of the redistributed points or center of the circle exceed the unit circle.

Similar to Fig. 9, Fig. 11 shows the second iteration of data after the minimum radius bounding circle has been transformed to the $S_{11}$ plane. The calculation ends when the lower bound of radiation efficiency stops significantly changing from iteration-to-iteration. In practice, this is usually after three iterations.

The iterative method remains a lower bound calculation because the reverberation chamber and paddle are not ideal, and do contain some loss. As the amount of loss increases, the lower the lower-bound calculation of efficiency will be. This method yields results that are consistent with those in [15].

VI. ANALYTICAL EFFICIENCY CALCULATIONS AND MEASUREMENTS

As a first step toward validating the two-port antenna model, we will examine its specific application to antenna efficiency. If the model and calculations are shown to be consistent with analytical calculations, we can have some confidence that the model is functioning properly.

In order to provide a basic analytical validation, we consider a monopole antenna with finite conductivity over an infinite ground plane for which the transmission efficiency is analytically tractable under standard antenna theory assumptions. We define a dipole of radius $a$, length $l$ and conductivity $\sigma$. By definition, the radiation efficiency is

$$\eta_R = \frac{P_R}{P_R + P_L},$$

where $P_R$ is the radiated power and $P_L$ is the Ohmic power loss [16]. By use of image theory, the current distribution on the antenna (along the $z$ axis) is

$$I(z) = I_0 \sin [k(l - z)],$$

where $k$ is the wavenumber and $I_0$ is the maximum current. With the given current distribution, the radiated power is given by an integral over the space above the conducting ground plane:

$$P_R = \frac{\eta_R I_0^2}{4\pi} \int_0^\infty \frac{\cos(kl \cos \theta) - \cos[kl]^2}{\sin(\theta)} d\theta$$

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where $\eta_0$ is the impedance of free space. Equation (27) can be readily evaluated numerically. To evaluate ohmic losses, we begin with an expression for total resistance $R$ for any conductor of uniform conductivity:

$$R = \frac{1}{\sigma} \frac{\ell}{A}$$

(28)

where $\ell$ is the conductor length and $A$ is the cross-sectional area of current flow. In the antenna, the current will flow primarily on the surface, penetrating the conductor to a depth of the skin depth $\delta = \sqrt{2/\omega \mu_0 \sigma}$. Under the conditions of $\ell \ll a$ the cross sectional area of current flow in the monopole is $A = 2\pi a\delta$, allowing us to write the Ohmic loss as

$$P_L = \int_0^{\ell} \frac{I(z)^2}{2\pi \sigma a^2} dz.$$  

(29)

The integral in (29) can be evaluated as

$$P_L = \frac{I_0^2}{2\pi a} \sqrt{\frac{\pi \mu_0}{\sigma}} \left[ \frac{1}{2} - \frac{1}{4k\delta} \sin(2k\ell) \right].$$

(30)

The above analysis is based on a perfectly conducting infinite ground plane. If we assume that the ground plane has a finite conductivity, it is not unreasonable to also assume that losses in the ground plane are equal to that in the antenna. The expression for efficiency then becomes

$$\eta_R = \frac{P_R}{P_R + 2P_L}.$$  

(31)

We can estimate the efficiency of a given monopole using this set of equations. This calculation can then be compared to measurements made with a precision constructed monopole inside a reverberation chamber. For this comparison, three different monopoles were constructed. Each monopole was designed to have a different resonant frequency: 750 MHz, 1 GHz, and 1.5 GHz, respectively. Fig. 12 shows one of the monopoles constructed for this measurement.

The diameter of each of the three monopoles is 6.4 mm. The lengths are 100.4 mm, 62.7 mm, and 41.4 mm, respectively. The ground plane of each monopole has a diameter of 298.4 mm. The monopoles are constructed from Copper 101, which has a nominal conductivity of $5.8 \times 10^7$ S/m. Given these dimensions, the efficiency we expect for each monopole is shown as a dashed line in Fig. 13.

The calculation indicates that the transmission efficiency of each monopole is nearly flat (within 10 %) across the frequency range 500 MHz to 1.5 GHz. This is the result of the definition of efficiency we are using; energy reflected by the monopole is not considered to be a loss.

Also shown in Fig. 13 is the lower bound of transmission efficiency as estimated by the iterative method shown in Section V. In the case of the 750 MHz monopole, the measured data are within 5 % of the analytical model at the lower frequencies. As the frequency increases, the lower bound diverges from the analytical calculation.

This divergence has two main contributors. First, the assumption that the chamber is lossless for at least a few paddle positions becomes weaker as frequency increases. At lower frequencies, this assumption is close to being correct. As the frequency increases, losses in the stirrer and chamber begin to increase. This increase in loss shrinks the radius of the measured data, and thus results in a reduced lower-bound efficiency calculation.

Second, the analytical calculation assumes that the ground plane is infinite in size. This allows for simple assumptions about the current structure on the monopole. On an infinite ground plane, the monopole will appear to be a perfect dipole due to imaging currents [5]. If the ground plane is not infinite, the image currents are not perfect, and the current structure is less favorable. This leads us to underestimate the losses in the monopole and ground plane, causing a slightly higher theoretical efficiency.

The data for the 1 GHz and 1.5 GHz monopoles follow the same fall off as for the 750 MHz monopole. This gives a very clear indication that the loss of the reverberation chamber and stirrers is dominating the calculation. For comparison, a broadband (200 MHz–2 GHz) dual ridged horn was measured in the same reverberation chamber at the same frequencies as the three monopole antennas. These antennas are designed to
be reasonably well matched, low-loss, and efficient at all frequencies in this band. The similar fall off between the horn and three monopole antennas is performing equally poorly for all antennas in cases where the antenna mismatch may be severe. Each of the three monopole antennas has a severe mismatch when it is operating outside of its resonant frequency.

Despite this drop-off at higher frequencies, this comparison provides some support for the two-port model being valid. If the lower bound of efficiency were higher than the analytical solution, it would indicate that the radius of measured $S_{11}$ was artificially large.

Future research may indicate that this large difference can be corrected. If the loss of the chamber is known from other means (e.g., a measurement of insertion loss), it may be possible to remove the effects of chamber loss on the lower bound efficiency measurement.

The next step towards validating this antenna model would be to perform a finite element analysis. This analysis would require a detailed model of a monopole be created and simulated to calculate the transmission efficiency.

A. Measurement Setup and Uncertainties

For the measured data in Fig. 13, the antenna under test was the only antenna inside the reverberation chamber during the measurement. Complex $S_{11}$ was measured by use of a vector network analyzer (VNA). Data were acquired at 31.25-Hz intervals.

The reverberation chamber measures 4.2 m in length, 3.6 m width, and 2.9 m in height. There are two stirrers inside the chamber. One runs side to side, and the other from floor to ceiling. Each paddle was stepped 23 times to create 529 different environments. These data were then used to calculate the lower bound of antenna efficiency. Data for the horn and 750 MHz monopole antennas began at 500 MHz, and ended at 2 GHz. Data for the 1 GHz and 1.5 GHz monopole antennas began at 750 MHz and 1 GHz, respectively.

Uncertainties associated with the measurement of $S_{11}$ include uncertainties of the VNA (e.g., receiver nonlinearity), effects of impedance mismatch, and connector repeatability. These uncertainties total 3.3% (coverage factor $k = 2$). In addition to these uncertainties, additional consideration is given to how the measurement uncertainty propagates through the model.

To assess these uncertainties, a bootstrap analysis was performed [17]. For this analysis, 100 random data sets were taken from each set of 529 stirrer positions. For each random data set, transmission efficiency was calculated. The standard deviation of the results are used as an indication of the uncertainty of that parameter. The VNA uncertainty and standard deviation are combined (via the root of the sumsquared of the components) and a coverage factor of $k = 2$ is applied. Fig. 14 shows an envelope of the total expanded uncertainty of the lower bound of radiation efficiency.

Fig. 14 indicates the uncertainty of the lower bound estimate. Although the uncertainty of the lower bound is low, this does not imply that the lower bound is actually a good estimate of the efficiency of the antenna. In some cases, the true efficiency may be significantly higher.

VII. Applications and Future Work

The proposed antenna model has the potential to isolate and remove the effects that an antenna has on the measurement of an environment. This would have a significant impact on our ability to measure an environment utilizing an antenna. If all four parameters of the model could be determined, and an antenna’s effects removed, the uncertainty of a variety of measurements would decrease significantly.

Here, we have shown one way in which the model network parameters can be estimated for a special set of conditions: when the reverberation chamber is lossless, or has very little loss. For one of the parameters, $S_{11}$, the estimation is weak. Additional work is required to develop a deterministic solution for $S_{11}$.

The lower bound on antenna efficiency discussed in Section V is only one example of an application. Once estimates for the model’s parameters have been determined, there are sure to be a wide variety of other applications. This will be the focus of future work: what other characteristics can be calculated from the two-port model? Can we gain a better understanding of the antenna’s mismatch or loss characteristics?

Revisions or supplements to the model may also be possible. By supplying supplemental information about the environment (e.g., loss characteristics) it may be possible to revise the estimations of the network parameters to include this information.

Developing better estimations of the two-port model parameters when there is only a single antenna in an arbitrary environment would result in significant progress toward solving the case where two antennas are in an arbitrary environment (Section IV). If the two-antenna case was solved, we could thoroughly describe how signals propagate through an environment, while utilizing the two-port model to remove the effects of both antennas. This has the potential to significantly improve measurements done to characterize wireless channels [18].

Another important future research topic would be to focus on the difference between the measured data and the analytic calculation. This difference may yield some information about the environment the antenna is operating in. Consider the case where an antenna is characterized in a lossless reverberation.
chamber. In this environment the two-port model parameters are estimated or determined, and the parameters of the antenna become known. If the same antenna is then put in an unknown environment, any difference between the expected efficiency and the lower bound could be considered an aspect of the environment. This could provide a useful method for characterizing an unknown environment.

VIII. CONCLUSION

An improved model for antennas in a reverberation chamber has been presented. The proposed model works by accounting for antenna imperfections (e.g., impedance mismatch and loss) with an unknown two-port network. One method for estimating the parameters of the two-port network under special conditions was detailed. In this case, the network was estimated by inserting the antenna into a reverberation chamber, $S_{11}$ data were measured, and a lower bound of antenna efficiency was calculated. As a preliminary validation of the model, three nearly ideal monopole antennas were modeled and constructed. The efficiency of each monopole was calculated from measured data and compared to the numerical efficiency of the model. The overall model performance and applications were discussed. Areas for future work were identified, including a brief examination of the case where two antennas are present in an arbitrary environment.

ACKNOWLEDGMENT

The authors would like to thank Dr. Ryan J. Pirkl for his helpful input and ideas concerning the revised network notation. They would also like to thank Masih Hajihabib for his efforts in constructing the monopole antennas.

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