Design and Applications of Two Low Frequency Guided Wave Electromagnetic Measurement Structures

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by

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DEDICATION

To my parents and all electromagnetic people.
I would like to express my hearty appreciations to Prof. Walter D. Burnside for his intuitive and instructive guidance from which both I and this research benefit a lot. I also want to thank Prof. Leon Peters Jr. for taking the time to review this thesis and giving valuable comments. Special thanks are delivered to the dear ESL friends who have been involved in system setup, measurements and discussions during the course of this research.
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CHAPTER I
Introduction

Most of backscatter measurements can now be done in a compact range where a approximate plane wave is generated to illuminate the target and very accurate backscatter signatures of various targets can be obtained [1, 2]. Usually, the frequency coverage of these ranges is from 1 GHz to 100 GHz and the performance tends to degrade at lower frequencies. Furthermore, the cost to construct and equip a low frequency anechoic chamber is high and special maintenance is required. Thus, if low frequency and simple backscatter features such as the co-polarized backscattered field measurements are preferred, one may want to consider other types of ranges.

It is well known that parallel-plate waveguides made of two conductors can propagate transverse electromagnetic waves (TEM) which are locally equivalent to plane waves. This property was first utilized to measure the radar cross section (RCS) of targets put inside a big parallel-plate transmission line (PPTL) by Gans [3] with an attempt to overcome the requirement of a very sensitive receiving system for measuring backscattered field in the presence of a $1/r^2$ propagation attenuation. He also justified the theoretical basis for this new technique by applying the Lorentz reciprocity theorem [4] under some valid assumptions. Two PPTL ranges, “bus bar transmission line ranges”, were constructed by Gans in 1965. Since he was interested in the resonance and above resonance region, these ranges covered the frequency bands from 1 GHz to 4 GHz and from 7 GHz to 10 GHz, respectively.
Another five PPTL ranges, called as the “rail line ranges”, were constructed later in 1971 by Newreuther et al. [5] for resonance region scatter studies. Each of these ranges covered a different frequency band and the total of them provided a frequency coverage from 0.25 GHz to 11.0 GHz. Four types of errors such as standard range errors, multiple reflections, fringe effects and image coupling effects, were identified and also briefly discussed in [5].

In 1983, the Electroscience Laboratory at the Ohio State University constructed another PPTL range for the purpose of making two dimensional material measurements at frequencies below 2 GHz [6]. The low frequency measurement performance of this PPTL range was also compared with the Ohio State University Radio Telescope (OSURT) compact range in [6]. In this PPTL range, absorber has been used to seal the side openings of the range with an attempt to reduce the fringe effects and external noise. But this modification increases the field taper in the target zone and introduces additional internal reflected waves.

In previous work, the fringe effects have been studied in [5] by measuring the backscattered field of a small dipole moving between the two conducting plates; the cross range and down range field behavior in the target zone of the PPTL with modified absorber wall was given in [6]. However, no data of the overall internal and external field distribution has been collected yet. Thus, the wave mechanisms in a PPTL range have not been fully understood so that no significant improvements and applications of this type of range has been made so far.

Since there is increased interest and applications in the low frequency region, more present work needs to be done in this area. For example, a rigorous study of the classical PPTL range may provide a solution to this problem. That is the focus of this study.
In this thesis, the classical PPTL range will be extensively studied. The overall field behavior in this type of range will be studied from two approaches. The first approach is to probe its two dimensional field distribution directly, and the other approach is to calculate its fields using the uniform geometric theory of diffraction (UTD) [7, 8]. This study shows that there are some inherent problems and difficulties caused by the undesired fields diffracted from the classical PPTL range structures. A new type of range, the radial-plate (RP) range, will be proposed to provide a better measurement environment by shaping the plates in such a way that each side edge is a radial line going out from the feed point. A much simpler wave mechanism and more uniform cross range field distribution at any down range position have been obtained in a down scaled RP range and will be demonstrated in this thesis.

Another modified RP range, the height-compensated radial-plate (HCRP) range, will also be proposed to further modify the down range field amplitude in an RP range. The new HCRP range has a modified height variation to compensate the $1/r^2$ power density attenuation in the propagation direction. This result is much more uniform down range field which has been measured in a down scaled HCRP range.

Chapter II reviews the theoretical background of using a PPTL range to perform RCS measurements. The diffraction coefficients and diffraction terms used to calculate the field in the PPTL range are also defined. Chapter III studies the overall wave mechanisms in a down scaled PPTL model. A near field focusing (NFF) technique [9] is also used to locate spurious scatters from the probed data. The problems associated with a classical PPTL range are identified in this chapter. In Chapter IV, a down scaled model of the new RP range is tested. The overall field distribution is probed and presented. The RCS measurements of conducting spheres are used as examples to show the measurement performance of this new range. In
Chapter V, a down scaled model of the modified RP range, the HCRP range, is also tested. The theoretical background of the modification is provided. The overall field behavior, range characteristics and sphere measurements are also presented. Chapter VI provides some simple guidelines for designing and constructing a good RP or HCRP range. More measurement examples such as RCS, echo width, and dielectric constant are presented in Chapter VII. This chapter explores the wide variety of applications of these ranges. This document is concluded in Chapter VIII by presenting a table which compares the different ranges discussed in this document from several important aspects.
CHAPTER II
Theoretical Background

In this chapter, the theoretical basis for target RCS measurements in a semi-
infinite transmission line made of two perfectly conducting plates will be discussed. The first theoretical basis for this technique was given briefly by Gans [3] where he related the open circuit voltage measured in the transmission line to the RCS of the target under test by applying the Lorentz reciprocity theory [4]. A more detailed proof of the proportion between the open circuit voltage and the target RCS was also given by Gwynne [6]. It should be noted that this proportion was obtained with some important assumptions which turn out to be invalid for some cases and thus cause measurement errors. Although Gans and Gwynne both treated the parallel-plate transmission line (PPTL), a more general two-plate structure, as shown here, can give the same results.

2.1 Theory Review

Figure 1 shows a section view of a two conducting plate transmission line. The source is fed at the tip position through a coaxial line with its outer conductor connected to the bottom plate and the center conductor to the top one. A test target is then placed in between the plates, and the return signal is received by the same coaxial line. The closed surface $S$ is a source free region. Surfaces $S_c$, $S_a$, $S_t$ and $S_e$ are subsets of $S$ and taken just off of the conducting plates, the feed, the
Figure 1: A general two-conductor plate transmission line with test target in between.

target and the end of the transmission line, respectively. If \((\vec{E}^i, \vec{H}^i)\) denote the fields inside the transmission line with the target absent, and \((\vec{E}, \vec{H})\) denote the total fields with the target present, then the Lorentz reciprocity theorem tells us the following:

\[
\int_{S_a+S_e+S_c+S_t} (\vec{E} \times \vec{H}^i - \vec{E}^i \times \vec{H}) \cdot \hat{n} dS = 0. \tag{2.1}
\]

The integral over \(S_c\) vanishes since the tangential component of electric field must vanish on the perfect electric conductors. Let \((\vec{E}^s, \vec{H}^s)\) denote the backscattered fields of the target, then

\[
\vec{E}^s \equiv \vec{E} - \vec{E}^i \quad \text{and} \quad \vec{H}^s \equiv \vec{H} - \vec{H}^i. \tag{2.2}
\]

Substituting Equations (2.2) into Equation (2.1) and rearranging the equation, one obtains that

\[
\int_{S_a+S_e} (\vec{E}^s \times \vec{H}^i - \vec{E}^i \times \vec{H}^s) \cdot \hat{n} dS = -\int_{S_t} (\vec{E} \times \vec{H}^i - \vec{E}^i \times \vec{H}) \cdot \hat{n} dS. \tag{2.3}
\]

Assuming that only the dominant mode (TEM) can propagate, the integral over \(S_e\) will vanish at infinity since \(\vec{E}^s\) and \(\vec{E}^i\) are both related to \(\vec{H}^s\) and \(\vec{H}^i\) by the characteristic impedance. Now let's define several currents as follows:

\[
\vec{J}_{S_t} = \hat{n} \times \vec{H} \quad \text{on} \ S_t \tag{2.4}
\]
\( \vec{M}_{S_t} = \vec{E} \times \hat{n} \) on \( S_t \) \hspace{1cm} (2.5) \\
\( \vec{J}^s = \hat{n} \times \vec{H}^s \) on \( S_a \) \hspace{1cm} (2.6)

and

\( \vec{J}^i = \hat{n} \times \vec{H}^i \) on \( S_a \). \hspace{1cm} (2.7)

After some vector manipulation on Equation (2.3), and then substituting the above currents into it, one finds that

\[
\int_{S_a} (\vec{E}^s \cdot \vec{J}^i - \vec{E}^i \cdot \vec{J}^s) dS = \int_{S_t} (\vec{E}^i \cdot \vec{J}_{S_t} - \vec{M}_{S_t} \cdot \vec{H}^i) dS. \hspace{1cm} (2.8)
\]

Equation (2.8) can further be rewritten such that

\[
V^s I^i - V^i I^s = -\int_{S_t} (\vec{E}^i \cdot \vec{J}_{S_t} - \vec{M}_{S_t} \cdot \vec{H}^i) dS \hspace{1cm} (2.9)
\]

where the voltages \( V^s, V^i \) and currents \( I^s \) and \( I^i \), which appear at the antenna terminals, are related to the incident fields and scattered fields, respectively. The open circuit voltage of backscattered fields can then be found from Equation (2.9), such that

\[
V^{oc} = \frac{-1}{I^i} \int_{S_t} (\vec{E}^i \cdot \vec{J}_{S_t} - \vec{M}_{S_t} \cdot \vec{H}^i) dS. \hspace{1cm} (2.10)
\]

Now, if the same target is put in free space with a plane wave \( (\vec{E}_c^i = \hat{e} E_0 e^{-jkz}) \) illuminating it, the \( \hat{e} \) component of target’s backscattered field can be expressed as

\[
\hat{e} \cdot \vec{E}_c^s \approx \frac{-jk\eta}{4\pi R} \int_{S_t} (\vec{E}_c^i \cdot \vec{J}_{S_t}^c - \vec{M}_{S_t}^c \cdot \vec{H}_c^i) dS. \hspace{1cm} (2.11)
\]

Note that \( \eta = \sqrt{\frac{\varepsilon}{\mu}} \), and \( \vec{J}_{S_t}^c \) and \( \vec{M}_{S_t}^c \) are equivalent currents which are defined by

\( \vec{J}_{S_t}^c = \hat{n} \times \vec{H}^c \) on \( S_t \) \hspace{1cm} (2.12)

and

\( \vec{M}_{S_t}^c = \vec{E}^c \times \hat{n} \) on \( S_t \) \hspace{1cm} (2.13)
where notation “c” implies “correct”, and $\hat{n}$ is the unit vector normal to the surface $S_t$. In addition, $\vec{E}_c$ and $\vec{H}_c$ are the total fields with the target scattering in free space. By comparing Equation (2.10) with Equation (2.11), it is clear that both equations have the same integration forms. If one further make the two following assumptions:

1. the incident wave in the transmission line in the target zone is effectively the same as that in free space; i.e., uniform plane wave, and

2. the equivalent currents for both cases are also the same

then it is found that $V_{oc} \propto \hat{e} \cdot \vec{E}_c^2$. Using this result, it is easy to show that the measured power is proportional to the RCS of the same target in free space. The first assumption is usually accurate if only one TEM wave is propagating in the line, which can be done with a good design. However, the second assumption is in general not true due to the fact that our target is scattering in the presence of the top and bottom plates rather than in free space, and the differences arising from this fact will be considered as measurement errors for echo area measurements and be discussed in the next section.

2.2 Intrinsic Errors

In Section 2.1, two assumptions were made to obtain the proportion between the measured open circuit voltage and the backscattered fields. The violations to these assumptions are the intrinsic errors of this type of range. The first assumption requires plane wave incidence on the target. This may not be true for the following situations. First, the range contains higher propagation modes. Second, there is strong coupling between the target and the feed. Third, the finite range width
causes edge effects. Last, the finite range length causes coupling between the target and the back end. As will be shown in the following chapters, all these terms can be avoided by carefully designing the transmission line or applying proper data processing. Although two other terms may be added to the above potential sources; the phase and magnitude variations due to the fact that a short line source was used and thus a cylindrical wave was actually propagating in the transmission line. The errors arising from the amplitude and phase variations of incident waves have been well studied, and a possible correction to this problem has also been proposed [10].

The violations to the second assumption are more difficult to deal with. When the target is illuminated by a local plane wave in the transmission line, it will scatter in all directions, and the field strength in each direction is usually target dependent. With the presence of the top and bottom plates and all the edges associated with a parallel plate structure, fields scattered into other directions rather than in the backscatter direction may be reflected or diffracted back to the receiving antenna. As one might expect, these terms can become stronger if the target has a wide scattering pattern, and thus causing more errors. There are at least two difficulties in treating this kind of error. First, since this kind of range is usually operated at low frequencies, the common methods used to correct multi-path errors (using two antennas receiving at the same time at different positions) are not applicable. This method can be interpreted as putting the target in several down range positions and then taking the data. Because the distance between the two positions needs to be large enough to have a significant change in the measured data, it would be too large and become impractical. Second, this kind of error is target dependent such that it may even have a higher level than that of the desired signal. Although it is difficult to treat this problem, one possible solution will be given in later chapters by using a
special range structure which can cut off or reject the signals bouncing back between the top and bottom plates so that they will not be received.

2.3 Analysis of the PPTL Range Using The Uniform Geometrical Theory of Diffraction (UTD)

Since the structure of a PPTL is simple and contains only straight edges and planar corners, the UTD method can be applied to study the field behavior if the following two conditions are satisfied:

1. All the dimensions of the basic structure must be at least a quarter of a wavelength. Note that the two plates can be separated by a small spacing such that higher order modes are not significant.

2. The higher order diffraction terms have been properly included.

In this section, the definition of the various parameters used in applying the UTD method will be given. As it will be shown in Chapter III, these diffraction terms can indeed predict the field behavior accurately and agree very well with the measured data.

2.3.1 Definitions

Figure 2 illustrates the geometry of the edges and corners considered in this problem. The various features are described below:

- Structure Simplification – As described earlier, a PPTL range consists of two sections – the transition section and the PPTL section. Each section is composed of two perfectly conducting planar plates. The top plates of these two sections are not coplanar, in that the transition expands linearly in width from
Figure 2: Definitions of edges and corners used in the UTD solution.
the throat to the parallel or test section. The bottom structure can be made of a single plate which makes it easier to construct and support from the bottom. The length of the transition section is usually long enough to obtain smaller cross range field variations. With these assumptions, the problem becomes finding the fields between the two parallel plates excited by a electric line source, as illustrated in Figure 2.

**Incident Field** – The height between the top and bottom plates is so small that only the transverse electromagnetic waves (TEM) can propagate in the transmission line. A short coaxial line with its inner conductor connected to the top plate and outer conductor to the bottom plate is used as the transmitter and receiver. It feeds the transmission line from the tip point of the transition section and then radiates into the range cylindrically. Since the electric field is always normal to the conductors, all edges and corners are effectively illuminated by this cylindrical wave at a grazing angle.

**Edge Diffracted Field** – The edges which will be included in calculating the diffractions are also shown in Figure 2, and are denoted as $E^{mn}$, where 'm' specifies the plate and 'n' the edge. It should be noted that the side edges in the transition section have not been included in calculating edge diffractions due to the fact that the wave is propagating along these edges, and thus no strong diffraction would be expected from them. Two classes of edge diffractions will be included, the first and second order edge diffractions, and they are defined in the following ways:
First order edge diffraction – The diffraction from the edges which are
directly illuminated by the source field and then diffract to the field obser-
vation point. For example, there are 6 possible first order edge diffraction
terms shown in Figure 2; i.e., \( E_{11}^{11}, E_{12}^{12}, E_{13}^{13}, E_{21}^{21}, E_{22}^{22} \) and \( E_{23}^{23} \).

Second order edge diffraction – The diffraction from the edges which are
illuminated by first order edge diffracted fields and then diffract to the
observation point. For example, there are 30 possible second order edge
diffractions shown in Figure 2. Every second order edge diffracted field is
associated with two of the following: \( E_{11}^{11}, E_{12}^{12}, E_{13}^{13}, E_{21}^{21}, E_{22}^{22} \) and \( E_{23}^{23} \).

Corner Diffracted Field – The corners which will be included in calcu-
lating the corner diffractions are also demonstrated in Figure 2, denoted as
\( C^{mn} \). The superscript ‘\( m \)’ specifies the plate and ‘\( n \)’ identifies the corner un-
der study. Each corner has two edges associated with it. Only the corner
diffractions which are directly illuminated by the source field and then diffract
for a second time to the observed field point will be considered here. There
are 8 corners shown in Figure 2, which create the corner diffracted fields:
\( C_{11}^{11}, C_{12}^{12}, C_{13}^{13}, C_{14}^{14}, C_{21}^{21}, C_{22}^{22}, C_{23}^{23} \) and \( C_{24}^{24} \).

2.3.2 The UTD Formulas

The general problem of a spherical wave illuminating a three dimensional curve
dge using the UTD method was solved by Kouyoumjian and Pathak[7]. Since a
half-plane edge is just a special case of a curve edge, its diffracted field is given by

\[
\begin{bmatrix}
    E_{\rho_o}^d \\
    E_{\phi}^d
\end{bmatrix} \sim \begin{bmatrix}
    -D_s & 0 \\
    0 & -D_h
\end{bmatrix}
\begin{bmatrix}
    E_{\rho_o}^{i'} \\
    E_{\phi}^{i'}
\end{bmatrix} \sqrt{\frac{s'}{s(s + s')}} \cdot e^{-jks}
\]

(2.14)
in which $D_s$ and $D_h$ are given by

$$D_{s,h}(L,\phi,\phi';\beta_o) = \frac{-e^{-j\pi/4}}{2\sqrt{2\pi}ksin\beta_o} \cdot \left\{ \frac{F[kLa(\beta^-)]}{\cos^{\beta^-}_2} + \frac{F[kLa(\beta^+)]}{\cos^{\beta^+}_2} \right\}$$  \hspace{1cm} (2.15)$$

where

$$k = \frac{2\pi}{\lambda}$$  \hspace{1cm} (2.16)

$$\beta^\pm = \phi \mp \phi'$$  \hspace{1cm} (2.17)

$$L = \frac{ss'}{s+s'}sin^2\beta_o$$  \hspace{1cm} (2.18)

$$a(X) = 2\cos^2\frac{X}{2}$$  \hspace{1cm} (2.19)

and

$$F(X) = 2j\sqrt{X}e^{iX} \int_{\sqrt{X}}^\infty e^{-jr^2} \, dr.$$  \hspace{1cm} (2.20)

For grazing incidence ($\phi' = 0$) with electric field normal to the edge (hard case), Equation (2.15) can further be simplified as

$$D_{s,h}(L,\phi,\phi';\beta_o) = \frac{-e^{-j\pi/4}}{\sqrt{2\pi}ksin\beta_o} \cdot \frac{F[kLa(\phi)]}{\cos^{\phi}_2}.$$  \hspace{1cm} (2.21)

The definitions of $\beta_o$, $\phi$ and $\phi'$ are illustrated in Figure 3, where $\beta_o$ is the acute angle between the incident ray and the edge. The polar angles $\phi$ and $\phi'$ are the angles between the lit side surface and the rays projected onto a plane normal to the half plane containing the edge.
Figure 3: Half-plane edge diffraction.
The diffracted field from a corner formed by two straight edges can be found by first calculating the diffracted field associated with the corner along one edge, and then the other. The heuristic formula for calculating the diffracted field associated with one corner and one straight edge, as shown in Figure 4, was proposed by Burnside and Pathak [8]. For all cases studied here, the corner diffracted field results are computed for grazing incidence with its electric field normal to the plane containing the corner. Thus, the corner diffracted field is given by

$$ E_\phi^c \approx MY_0 \frac{\sin \beta_c \sin \beta_{oc}}{(\cos \beta_{oc} - \cos \beta_c)} \cdot F[kLc(\pi + \beta_{oc} - \beta_c)] e^{-jks} $$

(2.22)
where

\[ M = -E^{i\phi}_h(Q_c)C_h(Q_E)Z_o \sqrt{\frac{8\pi}{k}} e^{-j\pi/4} \]  \hspace{1cm} (2.23)

and

\[ C_h(Q_E) = \frac{-e^{-j\pi/4}}{\sqrt{2\pi k \sin \beta_o \cos \frac{\phi}{2}}} \left| F[kLa(\phi)] F[kLc(\pi + \beta_{oc} - \beta_c)] \right|. \] \hspace{1cm} (2.24)

The functions \( F() \), \( a() \) and distance parameter \( L \) are the same as those defined in Equations (2.20), (2.19) and (2.18).

Using the above formulas, the first order edge, second order edge and corner diffracted fields can be calculated. It is noted that since the incident electric field is oriented normal to the plane containing the edges and corners, the slope diffractions contributions [11] have been ignored in the previous formulas.

2.3.3 Calculated Results

In this section, the calculated edge diffracted field and corner diffracted field derived in the previous section will be presented. The PPTL geometry to be calculated is shown in Figure 5, and the frequency is 2000 MHz.

The first order edge diffracted field distribution alone is shown in Figure 6. It is clear that these terms are dominated by the terms arising from the side edges (i.e., \( E^{11}, E^{13}, E^{21}, E^{23} \)). In the brighter region, only the diffracted fields from the back edges exist (i.e., \( E^{12} \) and \( E^{22} \)). The side edges of PPTL can also be clearly recognized from this figure (two bright lines), since the diffracted fields from the top plate and the bottom plate have opposite polarizations at the point right in between the two edges, and tend to cancel each other. Two darker regions extended from the back corners are due to the poor resolution of the *diffraction point searching procedure* in the computer codes when the diffraction point is close to the ends of the edges.

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Figure 5: The region of study in the PPTL range.
Figure 7 shows the distribution of the incident and first order edge diffracted fields. It is noted that fields exterior to the side edges are much weaker than the interior ones. This is because these regions are right at the lit sides of edge diffraction shadow boundary, and thus each edge diffraction tends to cancel one half of the incident field. Thus, the top and bottom edge diffracted fields tend to cancel the incident field. The remaining fields would be the back edge diffracted fields. Another thing to be noted is that there are several discontinuous lines, and one should be most concerned about those in the interior region where the target is going to be tested. This means that higher order diffraction mechanisms need to be included in our solution to make the interior total field as continuous as possible.

First, let us try to include the second order edge diffracted field whose distribution is shown in Figure 8. The maximum level of these terms is about 10dB less than that of the first order terms. Adding these terms to the previous total field, one obtains a new total field distribution with the first and second order edge diffracted fields included as shown in Figure 9. Note that there is not much change in the distribution and the discontinuities are still present. This means that the contributions from the second order edge diffracted field are insignificant at this frequency.

Next, the corner diffracted fields are added to the solution. The amplitude distribution of corner diffracted fields is shown in Figure 10, which has about the same maximum magnitude level as the second order edge diffracted field, and the new total field distribution is shown in Figure 11. It is interesting to find that the previous discontinuities have disappeared, which implies that they were caused by the lack of corner diffraction.

From the above analysis, one finds that the major diffraction mechanisms in a PPTL range are the first order edge diffracted field and the corner diffracted field.
This theoretical prediction will be further validated by showing that it agrees with the measured data in the next chapter.
Figure 6: First order edge diffracted field in the PPTL.
Figure 7: Field distribution of incident and first order edge diffracted fields.
Figure 8: Second order edge diffracted field of PPTL.
Figure 9: Amplitude distribution of first and second order edge diffracted fields and incident field.
Figure 10: Amplitude Distribution of corner diffracted fields.
Figure 11: Total field distribution in PPTL.
2.4 Summary

It has been shown that the open circuit voltage measured at the receiving antenna of a two-plate transmission line with a test target is proportional to the free space backscattering RCS of the same target if two assumptions are met. The violation to these two assumptions due to the presence of the corners, edges and conducting plates cause the main intrinsic errors in this type of range. The UTD method has been applied to find the first and second order edge diffracted fields and the corner diffracted field in a PPTL range. In the next chapter, this technique will be further used to find the theoretical solutions of an actual PPTL range model and then compare these results with the measured data. The validity of this RCS measurement technique will also be presented in the following chapters by comparing the measured results of different types of targets with their exact solutions.
CHAPTER III
Classical Type Parallel-Plate Transmission Line (PPTL) Range

In this chapter, some important characteristics of a classical type Parallel-Plate Transmission Line Range, as shown in Figure 12, will be studied by analyzing several measured results taken in the model shown in Figure 13, which will be called Model #1 hereafter.

A PPTL Range is made of two isolated conducting plates with one on the top and the other at the bottom. This kind of range can usually be divided into two sections – the transition section and the parallel plate section. The purpose of the transition section is to provide a smooth transition for a wave propagating from the feed antenna to the PPTL section without introducing multi-path interference [12] and thus reducing the excitation of TE or TM modes in the PPTL section.

The PPTL section is basically a finite parallel plate transmission line where the dominant propagating mode is TEM wave. The TM modes can be cut off by choosing the height to be less than a half wavelength corresponding to the highest frequency of operation. Since the use of the transition section has largely reduced the chance of the excitation of TE modes, the width of the PPTL can be larger than a half wavelength at the highest frequency of operation.

As one knows, a TEM wave is locally a uniform plane wave, and thus satisfies the first assumption made in Section 2.1. A PPTL can then be used as a range to measure the RCS of targets as if they were in free space. However, there are some
problems and difficulties in this type of range which will be discussed later in this chapter.

3.1 Model #1 Structure Description

The configuration of Model #1 is shown in Figure 13. Several design features of this model in comparison with a classical type one (see Figure 12) are listed below:

- Instead of using a symmetric structure of top and bottom plates, the bottom plate remains flat. A flat bottom plate has the following two advantages: (1) it is more convenient to support the range from the bottom; (2) different target height can be easily fitted into the range. Besides, the modification will not change any characteristics of the range because if one applies image theory with the bottom as its image plane, one can get exactly represent the classical type.
Figure 13: Structure of the Model #1, Parallel-Plate Transmission Line (PPTL) range.

- The transition section is filled with styrofoam in between the top and bottom plates. This keeps the shape from deforming due to the weight of the top plate. Since the material properties are approximately the same as those of free space, its effects on the fields can be neglected.

- The feed point is offset by a very small amount from the tip which provides better shielding from the backward radiation of the feed such that most of the energy is delivered to the range. In other word, the amount of energy received from exterior scatterers by the feed is also reduced. This is illustrated in Figure 14.
3.2 Characteristics of Model

3.2.1 Field Probing Using Backscatter Method

The method of using the backscatter measurement of a sphere to do the field probe will be discussed in this section.

First, let us consider a scattering problem as shown in Figure 15(a), where an isotropic point source radiates in free space with a scatterer present, Scatterer 1. Suppose that the fields at point $F$ are to be determined. The most direct way is to use a small ideal “isotropic” receiving antenna to detect the field as shown in Figure 15(b). But the cable connected to the receiving antenna may also cause some measurement errors. So, let us put a small perfectly conducting sphere (Scatter 0) at point $F$ and measure the fields scattered by the sphere at the source point, as shown in Figure 15(c). Assuming that the scattered fields of the scatterer and the sphere can be express as

$$\vec{E}_1^s = S_1(\bar{r}, \theta) \vec{E}_1^i$$  \hspace{1cm} (3.1)

and

$$\vec{E}_0^s = S_0(\bar{r}, \theta) \vec{E}_0^i$$  \hspace{1cm} (3.2)
where

\[ \tilde{E}_1 = \text{field incidence on the scatterer (Scatterer 1)} \]

\[ \tilde{E}_0 = \text{field incidence on the sphere (Scatterer 0)} \]

\[ \tilde{E}^s_1 = \text{field scattered from Scatterer 1} \]

\[ \tilde{E}^s_0 = \text{field scattered from Scatterer 0} \]

\[ \text{S}_1 = \text{scattering matrix (dyadic) of Scatterer 1} \]

\[ \text{S}_0 = \text{scattering matrix (dyadic) of Scatterer 0} \]

\[ \vec{p} = \text{distance vector from the scatterer to the field point, and} \]

\[ \theta = \text{bistatic scattering angle with respect to incident direction.} \]

Then the field at point F in the absence of the sphere is given by

\[ \tilde{E}^c = \tilde{E}_0^i + S_1 \tilde{E}_1^i. \] (3.3)

The backscattered fields in the presence of the sphere consists three paths (see Figure 15), which are denoted as paths 1, 2 and 3. Each of them is defined as follows,

**Path 1** - direct backscatter

\[ \tilde{E}^s_{p1} = S_0(\vec{r}_0, 0) \tilde{E}_0^i \] (3.4)

**Path 2** - indirect backscatter

\[ \tilde{E}^s_{p2} = S_1 S_0(\vec{r}_1, 0) S_1 \tilde{E}_1^i \] (3.5)

**Path 3** - bistatic scattering

\[ \tilde{E}^s_{p3} = S_0(\vec{r}_0, \alpha) S_1 \tilde{E}_1^i. \] (3.6)
Figure 15: Field probing using a backscatter measurement.
Using Equations (3.4), (3.5) and (3.6), the total scattered fields can then be expressed as

$$\vec{E}^s = \vec{E}^s_{p1} + \vec{E}^s_{p2} + \vec{E}^s_{p3}. \quad (3.7)$$

Note that the second term in Equation (3.7) involves the third order diffraction mechanism, which is usually small in comparison with the other two terms and can be neglected. Thus, the total scattered fields can be approximated by

$$\vec{E}^s \approx S_0(\vec{r}_0, 0) \vec{E}_0^i + S_0(\vec{r}_0, \alpha) S_1 \vec{E}_1^i \quad (3.8)$$

or

$$\vec{E}^s \approx S_0(\vec{r}_0, 0) \left[ \vec{E}_0^i + \xi(\alpha) S_1 \vec{E}_1^i \right] \quad (3.9)$$

where $\xi$ is defined as

$$\xi(\alpha) = \frac{S_0(\vec{r}_0, \alpha)}{S_0(\vec{r}_0, 0)} \quad (3.10)$$

which is the ratio of the bistatic scattered field to the backscattered field; i.e., it's a bistatic scattering pattern with its value normalized to the value in the backscatter direction. Figure 16 displays the $\xi$ function of a 1" perfect conducting sphere at the 300 MHz, 2 GHz, 10 GHz and 20 GHz, respectively. It is seen that $\xi(\alpha) \approx 1$ near the backscatter direction. Under this circumstance, Equation (3.9) becomes approximately

$$\vec{E}^s \approx S_0(\vec{r}_0, 0) \left[ \vec{E}_0^i + S_1 \vec{E}_1^i \right]. \quad (3.11)$$

Comparing Equation (3.11) with Equation (3.3), it is found that

$$\vec{E}^s \approx S_0(\vec{r}_0, 0) \vec{E}_c. \quad (3.12)$$

The above equation shows that the direct probe fields are related to the backscatter measurement of a perfect conducting sphere by the backscatter dyadic $S_0$ of the sphere at point F (or $\vec{r}_0$). For uniform plane wave incidence, $S(\vec{r}, 0)$ of a perfect
conducting sphere can be found as

$$S(\vec{r}, 0) \approx C \frac{e^{-jkr}}{r}$$  \hspace{1cm} (3.13)

where $C$ is a constant dyadic. Using Equation (3.13) in Equation (3.12), one can approximately find the direct probe field by correcting the backscatter probe field, such that

$$\vec{E}^c \approx \vec{E}^s \cdot C^{-1}re^{jkr}$$  \hspace{1cm} (3.14)

where $r$ is the distance between the probing sphere and the receiving antenna. Let the field $\vec{E}_0$ at point $r_0$ be the reference point and $\vec{r}$ be an arbitrary field point, then one has

$$\vec{E}_0^c \approx \vec{E}_0^s \cdot C^{-1}r_0e^{jkr_0}$$  \hspace{1cm} (3.15)

and

$$\vec{E}^c \approx \vec{E}^s \cdot C^{-1}re^{jkr}.$$  \hspace{1cm} (3.16)

The relative amplitude and phase with respect to the reference point can then be defined as

$$\text{Corrected Relative Field (CRF) Amplitude} \equiv \frac{\vec{E}^c}{\vec{E}_0^c} = \frac{\vec{E}^s}{\vec{E}_0^s} \frac{r}{r_0}$$  \hspace{1cm} (3.17)

$$\text{Corrected Relative Field (CRF) Phase} \equiv \angle \frac{\vec{E}^c}{\vec{E}_0^c} = \angle \frac{\vec{E}^s}{\vec{E}_0^s} + k(r - r_0)$$  \hspace{1cm} (3.18)

where $k$ is the wave number. Equations (3.17) and (3.18) are exactly the relationships which will be used to find the true field distribution from the backscatter probe data in this thesis.

In this section, it has been shown that the field probing can be accomplished by using a moving perfectly conducting sphere and then measuring its backscattered field. This is much easier than moving a receiving antenna with a cable connected to it. Equation (3.12) clearly shows that the two measurements are related by
Figure 16: $\xi$ function of 1" perfectly conducting sphere as a function of the bistatic scattering angle $\alpha$. 
the backscatter property of the sphere. It should be noted that the following two assumptions have been made to obtain the above results:

1. The third order diffraction mechanism in Equation (3.5) should be small enough to be neglected. This assumption is normally a valid one.

2. The sphere should have a wide bistatic scattering pattern so that $\xi$ in Equation (3.10) approximately equals to unity. From Figure 16, it was found that a larger sphere could provide more accurate results. However, a large sphere will not sense the field at a point. Thus, there is a trade off in choosing a proper size for the sphere to do the field probing. Another important property of the sphere which has been implicitly utilized is that it gives the same bistatic scattering pattern with respect to the incident direction for any angle of incidence.

In next section, this technique will be used to probe the field of the Model §1 PPTL range, and, as one will see, the results agree very well with the theoretical ones.

### 3.2.2 Probe Field Results

In this section, the field distribution in Model §1 will be studied by doing both down range and cross range field probing. The configuration of this probing system is illustrated in Figure 17. A 1" conducting sphere is used as a scatterer and is moved in the cross range direction at each down range position. Then both the amplitude and phase of the backscattered fields are received by the same feed antenna and recorded. Figure 18 is a three dimensional plot of the measured amplitude distribution in Model §1. The measured amplitude and phase distribution of the CRF are also shown in Figures 19 and 20. It can easily be seen that the wave inside the range is not really
Figure 17: Configuration of 2D field probing system.
a simple plane wave. Instead of having just one wave (TEM) propagating in the range, as was expected, the interference patterns show that there are other waves existing in the range as well. In order to analyze the sources of these extra signals, the UTD method was used to theoretically find the possible first and second order edge diffractions and first order corner diffractions, as discussed in Section 2.3. The UTD amplitude results are shown in Figure 21.
It is interesting to find that if one compares this calculated result with the measured amplitude data shown in Figure 19, both plots have almost the same interference pattern. The similarity in the field interference pattern between the measured data and the UTD solution implies that the main wave mechanisms inside a classical type Parallel-Plate Range are the same as what one would expect using UTD analysis, and they are summarized as follows:

1. A cylindrical wave radiates from the feed antenna with its electrical field normal to the top and bottom plates.

2. Edge diffractions occur when the edges on each plate in the PPTL section are illuminated by the grazing cylindrical wave.

3. There are two right angle corners at the back of each plate, and two obtuse angle corners formed by the connection between the transition section and the PPTL section on each plate. That is, there are a total of eight corner diffraction terms. All of these corners are also effectively illuminated by a grazing cylindrical wave. From the UTD analysis, it was found that diffractions from the front corners \( C^{11}, C^{14}, C^{21} \) and \( C^{24} \) contaminate the field at the target zone more than the others.
Figure 19: CRF amplitude distribution in PPTL range at 2000 MHz.
Figure 20: CRF phase distribution in PPTL range at 2000 MHz.
Figure 21: UTD calculated field distribution in PPTL range at 2000 MHz.
In next section, a near field focusing technique [9] is used to analyze the probed data so that the various sources of stray signals can be located. It will be interesting to see if the significant image terms are very close to the front corners ($C^{11}, C^{21}, C^{14}, C^{24}$) which has been shown in this section.

### 3.3 Locations of Spurious Scatterers

The near field focusing (NFF) method has been used to locate the spurious sources in a compact range by using the field probe data [9]. In this method, the fields are probed on a plane at a fixed down range position, and then a searching algorithm is used to estimate the locations of the spurious signals might occur. This searching procedure is basically a coherent phase matching procedure which will be briefly discussed latter. In this study, the fields are probed at many down range positions. Instead of just matching the phases to a cross range probe data, the same procedure was repeated for each down range position. That is, for each searching point, its phase is matched to both cross range and down range probe data, and thus provides more accurate position of spurious scatters than those obtained from doing phase matching for just one cross range probe data set.

#### 3.3.1 Phase Matching Method

In Section 3.2.1, it was shown that the major signal returns from the target are the direct backscatter (Path 1) and the bistatically scattered term (Path 2) when a conducting sphere was used to do the field probing. If the fields propagating through the direct path between the primary source and the sphere are removed, then the problem of finding the position of a spurious signal source is illustrated in Figure 22.
Figure 22: Locating the spurious scatterer using the near field focusing method.
The probed field, $E_p(i,j)$ at $(i,j)$ position, can usually be written as

$$E_p(i,j) = A(i,j)e^{j\phi}e^{-jk[R_0(i,j)+R_1(i,j)+R_2(i,j)]}$$  \hspace{1cm} (3.19)

where

$k = \text{wave number}$

$A(i,j) = \text{the amplitude of probed field when sphere is at (i,j) position}$

$\phi = \text{a constant phase term}$;

$R_0 = \text{the distance between the primary source and real spurious scatterer}$

$R_1 = \text{the distance between the real spurious scatterer and the probing sphere}$

$R_2 = \text{the distances between the primary source to the probing sphere}$

$(i,j) = \text{the (i,j)th position of the sphere, and}$

If the phase is multiplied by its conjugate $\exp[jk(R_0(i,j)+R_1(i,j)+R_2(i,j))]$ at each sphere position, the probed fields will then be in phase for all positions and can be summed up coherently. On the other hand, if another conjugate phase term is used that is related to another searching point, then the phase of the probed fields will not be coherent for every probing position and thus will be destructively summed to a very small value. With this approach, one can locate the position of the spurious scatters by observing the searching results. The larger magnitude means a better phase matching, and thus indicates a larger possibility of a stray signal source.

In the previous discussion, it was assumed that the direct path term had been removed before searching. This is not a necessary step and the searching results will show the presence of the primary source. However, since the magnitude of the direct
path (Path 1) signal is usually much greater than the indirect path (Path 2), the positions of spurious scatters may be obscured in the results leaving the primary source in the image. Thus, it is usually desirable to remove the primary source before applying the NFF algorithm. There are many ways to do so. A smoothing method, for example, has been suggested by Gupta [13]. If the primary source radiates a cylindrical or spherical wave, then it is first changed into uniform signal by multiplying \( \exp(j2kR_2) \). Next the direct signal can be obtained by smoothing the field data since other indirect path signals will be removed after smoothing. Then the direct path signal can easily be subtracted out from the total field to obtain the indirect path field.

3.3.2 Locations of Spurious Scatterers

The field probe data of Model \#1 at 2000 MHz was processed using the near field focusing technique discussed previously. The results are shown in Figures 23 and 24. In Figure 23, the direct backscatter signals were included in the processing, and thus the result shows a strong signal coming from the primary source position. As discussed earlier, it is very hard to identify other source positions from this figure. Figures 24 shows the results of NFF by using 11 point smoothing to remove the primary source. Now, two possible spurious source locations are clearly identified. It is also interesting to find that these positions are very close to the front corner positions of Model \#1. These results prove again that the spurious signals in Model \#1 are arising from the corner and edge diffractions as was pointed out by the UTD analysis.
Figure 23: Positions of spurious scatterer including the direct backscatter signals, Kaiser-Bessel weights.
Figure 24: Positions of spurious scatterer using 11 point smoothing to remove the direct backscatter signals, Kaiser-Bessel weights.
3.4 Summary

In this chapter, the field behavior of a classical type Parallel-Plate Range was studied by using 2D field probing. The theoretical field distribution has also been calculated by using a UTD method. It has been shown that these two results have similar field interference patterns. It was then shown that the edge and corner diffractions in a classical type PPTL range are significant in the low frequency band, and need to be taken into consideration when designing the range. Another method, the near field focusing method, was used to locate the stray signal source positions from the probe data. These results have proved again that the corners of Model 1 are the major sources of the stray signal in a PPTL range.

One might also think about using absorber to treat the edges and corners, as was used in [6]. But there are at least three disadvantages with this method. First, it increases the difficulty in placing or removing targets without changing the background since the range is enclosed by absorber and one needs to remove some to add or remove the target. Second, since the use of this kind of range is basically for low frequency measurements, it would be difficult to find a good absorber which gives little reflection. Thus, by doing so, one just substitutes the diffraction problems with reflection problems, which might become more serious. Third, the absorber will cause a cross range taper in field distribution and thus reduce the size of available target zone.

Realizing the problems of a classical parallel-plate range makes it possible to do a better design, and leads to a new kind of range which will be discussed in the next chapter. It will also be shown that this new design can indeed solve the diffraction problems suffered by a classical type PPTL range and provide better target zone fields.
CHAPTER IV
The Radial-Plate (RP) Range

In the previous chapter, it was shown that the classical Parallel-Plate range has edge and corner diffracted field problems. To avoid these problems, it is desirable to design a range such that all the edges are aligned to the propagation direction of the source field. This requires that all the edges lie along radial lines going out from the source point. Thus, this new kind of range is called as a Radial-Plate RP range. A model of this new type of range is shown in Figure 25, and will be further described in the next section. It will be shown that such a structure basically eliminates the diffractions associated with a classical PPTL range.

In this chapter, the general field characteristics, such as field distribution, noise level, system response and sphere RCS measurements of the RP range will be studied. All of these properties will be studied using the model shown in Figure 25, which will be referred to as Model §2 hereafter. The advantages and disadvantages of this RP range will also be discussed at the end of this chapter.

4.1 Structure Description of an RP range – Model §2

The body of an RP range, as shown in Figure 25, is composed of two isolated triangular conducting plates. The bottom plate remains horizontal for the same reason discussed in Model §1; while, the height of the top plate increases linearly in the down range direction. The range is fed in the same way as the classical structure
Figure 25: Geometry of the RP range.
shown in Figure 14. To maintain the shape, the front section is again filled with styrofoam. At the back of the range, absorber is used to reduce the diffracted field from the back truncation. It also prevents the field from radiating into the room containing the range.

4.2 General Field Distribution

In order to understand the field behavior in this new type of range, the same probing technique discussed for Model $\text{UL}$ is used again to find the field distribution over the region of interest. The measured amplitude distribution of the backscattered field at 1000 MHz are shown in Figures 26 and 27. The amplitude distribution shows a more uniform cross range field and a smoother decaying down range field than found in Model $\text{UL}$. The small ripple appearing in the amplitude plot is caused by the interference of the signals returning from the range's back end where the absorber have been placed to reduce a large part of this returning signal. The physical edges of the range are also shown by the two dash lines. One should also notice that the fields are well confined in the range at this frequency. That is, the fields attenuate very fast as they leave the range. This is a very desirable effect because it assures better efficiency, better external clutter isolation and less external radiation.

The measured phase of the backscattered at 1000 MHz is shown in Figure 28. One should notice that the field propagating inside the range has a spherical phase and there is large phase variation near the range's side edges. More detail field characteristics will be studied in the following sections by examining its cross range and down range field behavior at various frequencies.
Figure 26: The backscattered field amplitude distribution in an RP range at 1000 MHz.
Figure 27: Backscattered field amplitude distribution in an RP range at 1000 MHz.
Figure 28: Backscattered field phase distribution in an RP range at 1000 MHz.
4.2.1 Cross Range Field Behavior

In this section, the cross range field behavior at various frequencies is studied in more detail by examining the cross range field probed at the 50'' down range position. In order to isolate the contributions from the back end and the mismatch at the feeding point, the frequency domain data was processed using data smoothing. The cross range CRF for a lower frequency band (200 MHz - 1000 MHz) and a higher band (1200 MHz - 2000MHz) are shown in Figures 29 and 30, respectively. One interesting feature of these results is that the field amplitude drops more than 25 dB outside the range. One should not be too surprised at this result since the structure is designed such that it guides the wave "naturally" inside the range, and thus most of the energy will remain inside the range. The field behavior outside the range is dominated by the edge waves which propagate along the waveguide–free space interfaces along the two sides of the range with a free space velocity. These edge waves are launched by the feed antenna close to the edges. Another important feature is that the fields are quite uniform in the middle region where the target zone of the range can be specified. For example, if the target zone requires less than 10 degrees phase variation and 1dB amplitude variation, then the target zone size for Model §2 is about 6'' for frequencies less than 2000 MHz (see Figures 31 and 32). This is about 30% of the range width. It is also noted that the size of target zone can further be increased for lower frequency measurements.
Figure 29: Lower frequency band cross range CRF distribution at 50\degree down range position.

Figure 30: Higher frequency band cross range CRF phase distribution at 50\degree down range position.
Figure 31: Lower frequency band target zone cross range CRF distribution at 50" down range position.

Figure 32: Higher frequency band target zone cross range CRF distribution at 50" down range position.
High frequency behavior of edge waves have been studied in [14] which also provides some useful high frequency asymptotic formulas for half-plane edge and wedge problems. However, the behavior of the edge wave at low frequencies and the coupling of two edge waves are still not well understood and need further study.

The cross range field of Model $\mathcal{H}2$ can also be evaluated numerically by using the moment method at the frequencies of 100 MHz and 300 MHz. Figures 33 and 34 compare the measured data with the calculated results using the moment method. It can be seen that they show very good agreement. Note also that these results are evaluated at a fixed distance ($50''$) from the tip point due to the requirements of moment method codes used by the author. Thus, the amplitude and the phase of the linear probed data were also adjusted to the circular arc trace where the values were calculated using the moment method. The adjustment is based on the assumption of a $e^{-jkr}$ field behavior which is not true when the field is close to the edges. This might explain the small discrepancies around the edges.
Figure 33: Comparison between the measured and calculated CRF at 100 MHz.

Figure 34: Comparison between the measured and calculated CRF at 300 MHz.
4.2.2 Down Range Field Behavior

The down range field behavior was also probed and is shown in Figure 35. This sampled data has been smoothed using a 151 point moving window, and the theoretical curves of the backscattered data are also shown for comparison. As one can see from these figures, the measured curves agree very well with the theoretical ones, which proves that the incident field is basically a spherical wave. One can also see from these results that the CRF field strength varies as $1/R^2$ for the RP range. This might introduce a measurement error for large down range targets.

![Figure 35: The backscattered field variations in the down range direction of RP range.](image-url)
4.3 Reflection Coefficient Frequency Response at Feed Point

The frequency response of the reflection coefficient at the feed point is shown in the Figure 36. Note that the frequency range is from 47.3 MHz to 2 GHz. The higher frequency part is plotted for reference. It is quite interesting to find that the reflection coefficient at the feed point using the feed structure shown in Figure 14 is very constant ($\approx -15$ dB for Model \#2). This reflection coefficient shows that there is still an impedance mismatch between the range's input impedance and the source impedance. Let $\Gamma$ denote the reflection coefficient and $S_{11}^t$ be the actual reflection coefficient of the test target. If there are no other mechanism which will cause errors in the reflection measurement, then the measured reflection coefficient, $S_{11}^m$, can be expressed as, [15],

$$S_{11}^m = \frac{S_{11}^t}{1 - \Gamma S_{11}^t}.$$  

(4.1)

From the above equation, it is clear that the impedance mismatch at the feed point can cause errors in RCS measurements. Although this error may be partially or even completely corrected by using the proper frequency domain or time domain calibration techniques [15], it is desirable to improve the matching condition and thus reduce the error before doing any correction. One possible method is to adjust the input impedance is to vary the height between the top and bottom plates at the feed point (see Figure 14) so that a better impedance condition could be obtained.
Figure 36: The reflection coefficient frequency response of Model #2.
4.4 Clutter and Noise Level

In order to know the ability of the RCS measurement in a range, it is usually important to determine the calibrated noise and clutter level. This can be obtained as follows: (1) take the target-free RCS measurement twice and record each data set, (2) subtract one from the other, and (3) calibrate the subtracted data with the RCS measurement of a reference sphere located at the target zone center. That is

\[
\text{Noise Level} = \frac{BKG - BKG}{C(f)} \tag{4.2}
\]

where BKG denotes the measurement with no target present and \( C(f) \) is the calibration function defined as

\[
C(f) = \frac{\text{measured RCS of reference sphere response-BKG}}{\text{exact RCS of reference sphere}} \tag{4.3}
\]

Since the down range field is not uniform in Model §2, the calibration function \( C(f) \) will be a function of the down range position. Thus, the noise level will be a function of down range position.

The measured noise and clutter RCS in Model §2 at the 50" down range position is shown in Figure 37. A 1" diameter sphere was used as the reference target. Both impulse response and frequency response are represented. The noise appears as random variations modulating on the clutter. The largest clutter is from the subtraction residue of mismatch reflection at the feed point, as shown in the impulse response. This clutter also cause the rising of the measured data. Since the target zone is of most interest, one can obtain the clutter in this region by smoothing the raw data. The smoothed data is also shown by the bold line in Figure 37. It is noted that the noise will be removed after smoothing and only the clutters remain. The smoothed result shows a clutter level of about -90 dBSM.

A very convenient and useful figure which gives the relative error in RCS measurements for different RCS clutter levels was introduced by Blacksmith and et al.
Figure 37: The calibrated noise and clutter RCS in Model §2.
Figure 38: The RCS error caused by the range's RCS clutter with respect to the RCS of true target.

[16]. Note that the RCS clutter level defined here is called as "net extraneous signal" in their paper. This figure is plotted in Figure 38, where $\sigma_e$ is the relative RCS error, and $S_e$ is the ratio of clutter to the true target RCS level. Thus, $S_e$ in dB is the required clutter level below the level of true target's RCS so that the relative RCS error is less than the desired value. For example, if one wants the relative RCS error to be less than 10 percent (i.e. $\sigma_e = -1.0$ dB) then, from the figure, it reads that $S_e \approx -25$ dB, which means the RCS noise level must be at least 25 dB below the true target RCS. That is, if a target with RCS of -65 dBSM is measured in Model #2 where the RCS noise level is $\approx -90$ dBSM, then the measurement error will be less than 1 dB!
4.5 Sphere Measurement

Perfectly conducting spheres measurements in Model #2 are presented in this section to examine the measurement performance of this kind of range. Both the time domain and frequency domain results are studied. Two spheres, one with a diameter of the 0.5" and the other with a diameter of 0.25" are used as the test targets, respectively. Another 1" sphere will be used as the calibration target.

Figure 39 shows the time domain response of 0.5" sphere. Figure 39(a) is the raw data before doing the background subtraction and 39(b) is the response after the background subtraction. There are five main signals in the raw data which identified as follows:

1. source internal reflections
2. mismatch at the source output port
3. mismatch at the range feed point
4. signal return from the test sphere, and
5. background return from the back of the range.

Signal (1) results from reflections occurring inside the network analyzer (HP8753C), which are caused by its internal circuit mismatch. Signal (2) is caused by the impedance mismatch between the signal line and the source’s output port. (3) is the reflection from the feed antenna at the range’s input. (4) is the desired signal returning from the conducting sphere. (5) is the signal returning form the range’s back edges. As one can see, the background signals, (1), (2), (3) and (5), are so large that the desired signal can not be clearly identified. This is expected result because they have not yet been specially treated and improved for this prototype.
model. However, since the whole system is stable enough, these background signals can be greatly reduced by background subtraction and only a small portion of clutter signals remain after subtraction, as one can see in Figure 39.

It is noted that, theoretically, one can reduce the (1) and (3) by improving the matching condition and treating the back structure; e.g., modifying the feed structure and using better absorber material; however, this becomes much more difficult at low frequency. Alternatively, they can be further reduced by using data processing techniques. The two most common processing methods are time domain gating and frequency domain smoothing, and the latter is adopted throughout this thesis. It should be noted that the number of points for smoothing must be limited so that no distortion is introduced and no target information is lost after smoothing. This will become clearer in Chapter VI.

Figure 40 shows the measured and calculated RCS of a 0.5" diameter sphere. Basically, the measured data follows the calculated curve except for the random noise and small clutter errors at the low end of the frequency band. These errors can be further removed by smoothing the data as shown in Figure 41. One should note that, after smoothing, the data shows about a 2 dB error which is caused by the coupling between the sphere and the range plates and will be discussed in Chapter VII. This error was introduced by the 1" reference sphere through the calibration process. A similar result was obtained for the 0.25" diameter sphere measurement as shown in Figures 42 and 43 except that the sphere RCS level in low frequency band is now lower than the noise and clutter level.
Figure 39: The measured impulse response of a 1" sphere: (a) before background subtraction (b) after background subtraction.
Figure 40: The RCS of a $\frac{1}{2}''$ sphere measured in the RP range without smoothing.
Figure 41: The RCS of a $\frac{1}{2}''$ sphere measured in the RP range using 101 points smoothing.
Figure 42: The RCS of a $\frac{1}{4}''$ sphere measured in the RP range without smoothing.
Figure 43: The RCS of a $\frac{1}{4}''$ sphere measured in the RP range using 101 points smoothing.
4.6 Summary

In this chapter, a new type of range for low frequency scattering measurements has been introduced. It processes the advantages of the classical type PPTL range for low frequency measurements but has much better field properties and better measurement performance. Its field behavior has shown that the incident wave illuminating the target zone is simply a spherical wave and that two edge waves exist in the vicinity of the side edges. This new range also has very low RCS Noise Level (-90 dB SM) which allows it to measure very small targets. Examples for sphere RCS measurements have been presented to illustrate this ability. It should be pointed out that the construction and the maintenance is very easy and very low-cost. Two possible future improvements of the range performance have also been pointed out in the chapter, which are to improve the matching condition at the feed point and the reduction of the signals returning from the range truncation edges.

In next chapter, a modified RP range will be introduced. Most of the merits contained in the RP range can also be found in this revised range. Furthermore instead of having a \( \frac{1}{r} \) decay in the down range field distribution, this next range can be designed with a variable down range amplitude performance.
CHAPTER V

Height-Compensated Radial-Plate (HCRP) Range

In the previous chapter, it was shown that the Radial-Plate Range has a uniform cross range field distribution; whereas, the down range field has a $\frac{1}{r}$ attenuation, due to the fact that the incident field is simply a spherical wave. This rapid decay in the down range field might cause measurement errors when the target's length is large or when measuring layered material characteristics.

A new height-compensated radial-plate (HCRP) range is introduced in this chapter. This range is significant because one can adjust the down range field amplitude distribution. The idea for this modification is that by changing the height between the top and bottom plates of the range, one can compensate for the $1/r$ field attenuation in the RP range. Consequently, a more uniform down range field can be obtained. Thus, the "Height-Compensated Radial-Plate (HCRP)" may be a proper name for this new type of range. It should be noted that a "uniform down range field" is not always the desired case for every application. In general, any down range field distribution can be obtained using this concept as long as the desired field amplitude varies smoothly and continuously. The configurations, field properties, measurement capabilities and limitations of this new range are studied in the following sections. It will be shown that a height-compensated radial-plate range can provide a very good and convenient environment for making many low frequency measurements.
5.1 Down Range Field Amplitude Adjustment in a HCRP Range

From the previous chapter, it is found that most of the energy delivered to a RP range is guided in the range and is propagating as a spherical wave. Thus, it can be assumed that all the input power, \( P_i \), delivered by the feed antenna, propagates spherically in the range and that the power density at a given distance can be denoted by \( I(r) \), which is given by

\[
I(r) = \frac{P_i}{S(r)}
\]  

(5.1)

where \( S(r) \) is the area of constant phase front at the distance \( r \). Figure 44 shows how the power is distributed in an RP range, where the angles \( \theta \) and \( \phi \) are, respectively, the flare angles in the horizontal and vertical planes of the range. Since \( S(r) = r^2 \theta \phi \), Equation (5.1) becomes

\[
I(r) = \frac{P_i}{\theta \phi} \cdot \frac{1}{r^2}
\]  

(5.2)

Usually, the angle \( \phi \) is so small that the curve length \( r \phi \) can be approximated by the height, \( H(r) \). Then Equation (5.2) can also be written as

\[
I(r) = \frac{P_i}{\theta} \cdot \frac{1}{rH(r)}
\]  

(5.3)

Since \( P_i \) and \( \theta \) are both constants, the power density at a distance \( r \) is completely determined by \( rH(r) \). Thus, by varying the height, \( H(r) \), it is possible to obtain a desired power density distribution, \( I(r) \). For example, if one wants

\[
H(r) = \frac{A}{r}
\]  

(5.4)

where \( A \) is a constant, then he can obtain a uniform down range field distribution. This is exactly the motivation for building a height-compensated radial-plate (HCRP) range.
Figure 44: The power density of a spherical wave in an RP range.

5.2 Structure Description of an HCRP range – Model 3

The geometry of an HCRP range (Model 3) is shown in Figure 45. Basically, Model 3 has the same geometry as Model 2 discussed in previous chapter except that the heights in regions B and C are varying in a different way. Region B is basically a parallel-plate region which serves as a transition region to avoid a sudden large curvature change from region A to region C. Region C is the height-compensated region where the height is varying according to Equation (5.4). The parameter A in Equation (5.4) is used to adjust the height of the top plate to a desired level. For example, in Model 3, the constant A is chosen such that the height in region C is varying from 3" to 2". It is important to keep the angle of the wedge at the junction of two adjacent regions as small as possible such that the wedge diffracted fields will not significantly affect the range field distribution. Also, the rate of change of the height in Region C can not be too large and that curve surface diffraction will not corrupt the range field quality. Consequently, each region should be built long enough to avoid the above two potential problems.
Figure 45: Geometry of Model #3, the HCRP range.
5.3 General Field Distribution

The field distribution in the HCRP range, Model \#3, at 1000 MHz was measured using the same sphere probe technique described earlier. The probed amplitude distribution is presented in both a 3D plot and a gray-scaled plot as shown in Figures 46 and 47, respectively. As can be seen in these figures, the field distribution is similar to that of Model \#2 (Figure 26) except that the measured down range field no longer has a $1/r^2$ (backscatter) attenuation.

The CRF phase distribution plotted in Figure 48 shows that the field inside the range also has a spherical phase distribution. Since the phase angle domain is chosen from -180 to 180 degrees, there is jump in the phase plot when the phase is less than -180 degrees. As was found in Model \#2, the field amplitude outside this range drops very fast to a small value. The physical positions of the edges are also shown by the dash lines. More detail field behavior will be presented in the following sections by studying its cross range and down range field characteristics.
Figure 46: 3D plot of field amplitude distribution in a Height-Compensated Radial-Plate range at 1000 MHz using sphere probing.
Backscattered Field Amplitude Distribution in HCRP Range

\[ f = 1000 \text{ MHz} \]

\[ -41. \quad -36. \quad -31. \quad -26. \quad -21. \quad -16. \quad -11. \]

MAGNITUDE (dB)

-\( -16 \) -\( -14 \) -\( -12 \) -\( -10 \) -\( -8 \) -\( -6 \) -\( -4 \) -\( -2 \) \( 0 \) \( 2 \) \( 4 \) \( 6 \) \( 8 \) \( 10 \) \( 12 \) \( 14 \) \( 16 \)

Cross Range Position (inch)

Down Range Position (inch)

Figure 47: Backscattered field amplitude distribution in a Height-Compensated Radial-Plate range at 1000 MHz using sphere probing.
Backscattered Field Phase Distribution in HCRP Range

\( f = 1000 \text{ MHz} \)

Figure 48: Backscattered field phase distribution in a HCRP range at 1000 MHz.
5.3.1 Cross Range Field Property

The cross range CRF amplitude and phase distributions at several frequencies are shown in Figures 49 and 50. The same large attenuation (25 dB) outside the range can also be observed in Model II 3. Near the edges, the incident field is also affected by another wave mechanism, the edge wave, which propagates along the range's side edges and is excited by the near-edge source located at the tip of the range. If the phase and amplitude variations in the target zone are limited to 10 degrees and 1 dB, the available target zone size will be about 5" (see Figures 51 and 52), which is about 30% of the range width (15"), the same percentage as that in a Radial-Plate Range. This number will vary with the down range position where the cross range field data is measured.
Figure 50: High frequency band cross range CRF distribution at 50'' down range position.

Figure 51: Lower frequency band target zone cross range CRF distribution at 50'' down range position.
5.3.2 Down Range Field Characteristics

The down range field variation in Model 3 was measured using the sphere probe technique, and the results are shown in Figure 53. Note that Figure 53(a) compares the amplitude variations at several frequencies with a theoretical spherical wave, $|E^b| \propto \frac{e^{-j2kr}}{r^2}$, and Figure 53(b) compares the measured phase variation with an ideal spherical wave at 1000 MHz.

The first thing one can find is that the measured backscattered field has the same phase variation as that of a backscattered spherical wave. This agreement also occurs at other frequencies but are not plotted here for clarity. Second, instead of having a $1/r^2$ attenuation, the amplitude of the measured backscattered field in the down range direction has a much more uniform distribution.
The above results prove that the field compensation theory derived in Section 5.1. With this improvement, more accurate down range measured data can be obtained. Nevertheless, one should note that the down range fields at higher frequencies still have some attenuation. Since the higher the frequency the more power is lost in the range due to radiation and finite plate conductivity. This power loss causes extra attenuation which has not been taken into account when finding the height to compensate the spherical attenuation.
5.4 Reflection Coefficient Frequency Response at Feed Point

The frequency response of the reflection coefficient ($S_{11}$) of this HCRP range model in the absence of any target is shown in Figure 54. The reflection coefficient frequency response behavior for this HCRP range is similar to that obtained for the RP range. At the low frequency band ($< 2 \text{ GHz}$) the reflection coefficient is on the average about -12 dB and has less than 2.2 dB variation. As discussed for Model #2, any reflection from the feed antenna itself will cause a RCS measurement error, and one should avoid this situation by improving the impedance matching condition between the source and the feed antenna.
Figure 54: The frequency response of the reflection coefficient of Model #3.
5.5 Clutter and Noise Level

The clutter and noise level in region C of the Model 3 range at \( Z = 50'' \) down range position is shown in Figure 55. The reference target was a 1'' diameter sphere. Both impulse response and frequency response with/without smoothing applied are presented. Compare this result with that measured in the RP range model, Figure 37, one finds a similar behavior, except that there appears a little more clutter based on the sinusoidal variations. The target zone clutter level was obtained using 101 point smoothing and is shown by the bold line. The clutter RCS is less than -90 dBSM, which is approximately the same as that found in the RP range. It is also noted that the impulse response shows more and higher clutter levels than that in the RP range and the feed mismatch still causes the most significant clutter term. As discussed in the RP range, this results provides a knowledge of the relative RCS measurement error when used with Figure 38. Since the down range in the compensated region has quite uniform field distribution, the noise level will be approximately constant at different down range positions.
Figure 55: The calibrated clutter and noise RCS in a HCRP range.
5.6 Sphere Measurement

Perfectly conducting spheres are measured next in the model §3 range to study its measurement performance. The impulse response of a 1" sphere at 43" down range position is shown in Figure 56. Note that Figure 56(a) is the raw data without using background subtraction; whereas, Figure 56(b) is the data with the background subtracted out. In Figure 56(a), one should observe five main signals whose source are identified as

1. mismatch at the source output port
2. mismatch at the range feed point
3. signal return from the test sphere, and
4. background return form the back end of the range.

In this measurement, an HP8753 was used as the source. (1) is caused by the impedance mismatch of the power divider and signal lines connected to the source output port. (2) is caused by the impedance mismatch between the signal line and the range input impedance. Reducing this reflection term requires a change in range input impedance and this needs careful and ingenious modifications of the range structure near the tip point without introducing other wave mechanisms. Note that the test target is usually positioned at a far distance such that the coupling between the target and the source, or between the target and the feed point is negligible. Thus, (2) is usually independent of the target and can be sufficiently removed after background subtraction. However, the reduction of these reflections can improve the system efficiency and is still an important issue in future studies.

The sphere return, signal (3), is quite "clean" and thus indicates little coupling with the range. (4) is another large unwanted background signal which is the field
scattered by the back end of the range. Although absorber material has been put at the back to reduce the scattering, it is not possible to achieve the needed attenuation at low frequencies. That is why one still has this large signal. It is possible to incorporate the absorber with the modification of the range back structure to reduce this backscattering and is another important issue for future study. Note that background subtraction can also remove most of (4) such that only a small amount of residue is left.

Although background subtraction can remove most of the unwanted background signals, there are still some residues left. Data processing technique is usually required to further remove these residues and improve the purity of the target signals.

The RCS of a 0.5" sphere measured in model $U_3$ is shown in Figures 57 and 58 with 101 point smoothing applied to the latter. A 1" sphere was used as the reference target for this measurement. As one can see, very good agreement with the exact solution has been obtained down to -90 dBSM. Below -95 dBSM, large errors occur because of the range's noise level (see Figure 55). A smaller sphere was also measured in this range, and the results are shown in Figures 59 and 60, with the latter being smoothed. This time, although the exact solution does indeed pass through the middle of the measured data above -90 dB, basically there is about a 5 dB error variation in these results. Below -90 dBSM, one has noise. The reason for having larger errors in measuring the RCS of this smaller sphere is simply because the signal level is now very close to the range's noise level. Based on these results, it appears that the RP's range noise level is about 10 dB lower that of the HCRP range.
Figure 56: The impulse response of a 1" sphere, (a) before background subtraction (b) after background subtraction.
Figure 57: The RCS of a $\frac{1}{2}$" sphere without smoothing.
Figure 58: The RCS of a $\frac{1}{2}$" sphere using 101 point smoothing.
Figure 59: The RCS of a $\frac{1}{4}$" sphere without smoothing.
Figure 60: The RCS of a $\frac{1}{4}$" sphere using 101 point smoothing.
5.7 Summary

In this chapter, a new type of range, the HCRP range, has been introduced, which gives an adjustable down range field distribution. The motivation and theory associated with this design have also been discussed. A range which creates a uniform down range field has been designed and proved through Model §3 and its measured results. It has also been shown that, as the frequency decreases, the range field becomes more uniform, which allows this range to do better low frequency measurements for longer down range targets. Both cross range and down range field distributions have been examined separately to verify the uniformity and purity of the field. In the last section, conducting sphere measurements were used to illustrate the good measurement performance and accuracy. From the sphere impulse response, it was determined that more work should be done to better match the input impedance at the feed point and to reduce the signals returning from the back end. This would reduce the background signals and improve the range’s sensitivity.

It is interesting to note that each region A, B and C in model §3 has a different down range field behavior. Region A has a down range field behavior similar to Model §2 discussed in the previous chapter, whose range field been shown to be spherical. That is, the range field has a $1/r$ attenuation. Region B is basically a classical PPTL with modified side edges to avoid corner and edge diffractions discussed in Chapter III. The range field inside this region is basically a cylindrical wave and has a $1/\sqrt{r}$ attenuation. Region C, as found in this chapter, has a very uniform down range variation. In addition, the phase of the range field in all three regions has the same $-jkr$ variation. This unique property of an HCRP range means that it can be used for a wider variety of applications than the other two studied previously. More measurement examples will also be presented in Chapter VII.
Like a classical PPTL range or a RP range, building and maintaining a HCRP range is quite easy and low-cost, such that measurements in this range can also be done very efficiently.
CHAPTER VI
RP and HCRP Range Design Considerations

In the previous chapters, the RP and HCRP ranges were shown to be very useful for low frequency RCS measurements. It has also been shown that the incident field is basically a spherical wave. This is the direct result of forcing the side edges of these range to be aligned to the radial lines going out from the feed antenna. But there are the two degrees of freedom in choosing the vertical and the horizontal flare angles (see Figure 44) that have not yet been specified. The proper down range length needs also to be determined and is another degree of freedom. Thus, there are three important undetermined parameters in designing a good RP or HCRP range. The determination of these parameters, however, is not arbitrary and is closely related to the target zone specifications and desired signal properties. The purpose of this chapter is to find these relationships and, hopefully, provide some guide lines to determine these parameters and to ensure a good target zone field and quality measured data. Since an HCRP range has almost the same design requirements as an RP range, most discussions in this chapter will focus on an RP range unless otherwise specified.
6.1 The Proper Target Zone-to-Edge Distance

Consider that a perfectly conducting cylinder with the same size as the circular target zone (radius \( R_t \)) is positioned at the center of the target zone (see Figure 61). Let us assume that a background scatterer is located at the position shown in the same figure and has some coupling effect with the target. This undesired coupling will arrive in the received data along with the desired target signals. Then assume that the frequency is scanned from the lowest frequency \( f_1 \) to the highest frequency \( f_2 \) and \( N \) sampled data points are collected at the receiver with a step of \( \Delta f \). After collecting this data, a certain type of smoothing window (\( N_s \) points) is applied to the sampled data to remove the undesired background signals with little loss and distortion in the target information. It is also assumed that one wants to keep the target scattering mechanisms up to the first-order creeping wave term. Then there are two questions to be answered:

“(1) What should be the minimum distance between the background scatterer and the target such that one can remove the background signal using frequency domain smoothing? (2) What is the maximum number of points in smoothing without causing serious distortion?”
Figure 61: Time domain responses of the target and range scatterer.

Before answering these questions, several important parameters appearing in the diagram shown in Figure 61 are defined as follows:

\[ BW \equiv \text{the measured frequency band width, } (f_2 - f_1) = (N - 1)\Delta f. \]

\[ T_c \equiv \text{the time delay for signals traveling from the time reference plane to the target center and then return to the reference plane; i.e., } S_c/c (\text{sec}), \text{ where } c = \text{free space velocity of light.} \]

\[ \Delta T_f \equiv \text{the time difference between the signals returning from the target front point and the target center; i.e., } 2R_t/c (\text{sec}). \]
\( \Delta T_{cp} \equiv \) the time difference between the signals returning from the target center and the first order creeping wave term; i.e., \( \pi R_t / c \) (sec).

\( \Delta T_b \equiv \) the time difference between the signals returning from the target center and the range scatterer; i.e., \( 2R_b / c \).

\( W_p \equiv \) the half pulse width (first null position) of the time domain signal due to the finite number of frequency domain samples.

\( W_s \equiv \) the time duration from the peak to the first null of the time gate which is the result of convolving the sampled data with a smoothing window.

\( N_s \equiv \) number of points in the smoothing window.

The pulse width of the signal, is caused by the spatial leakage which is the result of using a finite width spectral domain data set. The value of \( W_p \) is closely related to the weighting function used upon the sampled points to reduce the leakage (sidelobe level) before applying the Inverse Fourier Transform or Inverse Discrete Fourier Transform [17, 18]. The first null position for several commonly used windows can easily be found and are listed below, along with their peak sidelobe levels (SLL) and the equivalent Kaiser window parameter \( \alpha \) found in [18].

Let us define a pulse width parameter, \( \beta \), such that

\[
W_p = \frac{\beta}{BW}. \tag{6.1}
\]

Then from the above table, one can find that \( \beta = 1 \) for a Rectangular window, \( \beta = 1.5 \) for Hanning window, \( \ldots \), etc.

The half gate width, \( W_s \), is also related to the type and number of points used in the smoothing window. Using the previous comments, one can find the common
Table 1: The half pulse widths and sidelobe levels of common windows

<table>
<thead>
<tr>
<th>Window Type</th>
<th>$W_p$</th>
<th>SLL</th>
<th>Equivalent Kaiser Window Parameter, $\alpha$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rectangular Window</td>
<td>$\frac{1}{BW}$</td>
<td>$\approx -13$ dB</td>
<td>0</td>
</tr>
<tr>
<td>Hanning Window</td>
<td>$\approx 1.5 \cdot \frac{1}{BW}$</td>
<td>$\approx -31$ dB</td>
<td>$\approx 0.4233$</td>
</tr>
<tr>
<td>Hamming Window</td>
<td>$\approx 2.0 \cdot \frac{1}{BW}$</td>
<td>$\approx -41$ dB</td>
<td>$\approx 1.5470$</td>
</tr>
<tr>
<td>Kaiser Window</td>
<td>$\approx 2.9 \cdot \frac{1}{BW}$</td>
<td>$\approx -47$ dB</td>
<td>2</td>
</tr>
<tr>
<td>Blackman Window (K=3)</td>
<td>$\approx 3.5 \cdot \frac{1}{BW}$</td>
<td>$\approx -57$ dB</td>
<td>$\approx 2.2409$</td>
</tr>
</tbody>
</table>

range for $W_s$ (with $\alpha$ from 0 to 2), such that

$$\frac{1}{(N_s - 1)\Delta f} < W_s < 3 \cdot \frac{1}{(N_s - 1)\Delta f}. \quad (6.2)$$

The relative time positions of each signal as well as target range gate are also illustrated in Figure 61. It is noted that $\Delta T_f < \Delta T_{cp}$. To avoid loosing the target information, one requires that

$$W_s > \Delta T_{cp} + W_p. \quad (6.3)$$

Combining the conditions given in (6.2) and (6.3), one obtains the following condition:

$$\frac{1}{(N_s - 1)\Delta f} > \Delta T_{cp} + W_p \quad (6.4)$$

Substituting the definitions of $\Delta f, \Delta T_{cp}$ and Equation (6.1) into the above inequality and rearrange the equation, one finds that

$$N_s - 1 < \frac{N - 1}{\beta + \pi R_t \cdot \frac{BW}{c}} \quad (6.5)$$
For simplicity, a new parameter called the "band wavelength" is introduced and defined as

$$\lambda_b \equiv \frac{c}{BW}. \quad (6.6)$$

Using this definition in inequality (6.5) and assuming that \( N \gg 1 \) and \( N_s \gg 1 \), one finds the approximate maximum number of points for smoothing without losing the target information, such that

$$N_s < \frac{N}{\rho + \beta} \quad (6.7)$$

where \( \rho \) is also a new parameter called the "extension factor" and is defined as \( \rho \equiv \frac{\pi R_t}{\lambda_b} \). This inequality is a very important and a useful relationship to find the maximum number of smoothing points and is valid for any kind of smoothing window as long as the associated \( \beta \) value is properly chosen. If one wants to include other signals besides the creeping terms, \( \pi R_t \) should be replaced by the maximum path difference among all the target signals to find the new extension factor \( \rho \), and then use (6.7) to determine the approximate maximum number of points for smoothing.

The relative error for the various signals other than the main signal (located at the gate center) can also be found by taking the gate window weighting function at different time positions into account, which is beyond the scope of this thesis.

The minimum number of points required in doing the smoothing to remove background signals can be found from the following relationship:

$$W_s < \Delta T_b - W_p. \quad (6.8)$$

From (6.2), it is equivalent to require that

$$3 \cdot \frac{1}{(N_s - 1)\Delta f} < \Delta T_b - W_p. \quad (6.9)$$
Using the definition of $\Delta T_b$ and Equation (6.1) to the above equation, one obtains that

$$N_s - 1 > \frac{3(N - 1)}{\rho_b - \beta}$$

(6.10)

where $\rho_b$ is the extension factor for the backscatter signal, and is defined as

$$\rho_b \equiv \frac{2R_b}{\lambda_b}$$

(6.11)

Since $N_s \gg 1$ and $N \gg 1$, it follows that

$$N_s > \frac{3N}{\rho_b - \beta}$$

(6.12)

From Equation (6.7) and (6.12), it is found that in order to avoid aliasing in the target signals due to the contribution from the background signals when doing smoothing, the following relationship must be satisfied:

$$\frac{3N}{\rho_b - \beta} < \frac{N}{\rho + \beta}$$

(6.13)

Simplifying the above inequality, one obtains the minimum distance required for the background scatterer, which is given by

$$\rho_b > 3\rho_t + 4\beta$$

(6.14)

where $\rho_t$ is the maximum target extension factor. In the previous special case, the maximum target extension factor is determined by the first order creeping term and thus $\rho_t = \rho$. The required distances for a range scatterer according to condition (6.14) with $\beta = 2$ for three different sizes of target zone at different angles are illustrated in Figure 62. The condition set in (6.14) provides an important guide in range design when the range clutter signals need to be removed by smoothing the measured data. This condition will further be used in the following sections to decide the proper dimensions of the RP and HCRP ranges.
Figure 62: Minimum distance for a range scatterer to be removed using proper smoothing.

6.2 The Shortest Down Range Position for the Target Zone

One important parameter in designing an RP range or an HCRP range is the down range size which is the total length from the feed antenna to the back end of the range. A proper down range depth should be able to provide the desired field illumination in the target zone and also be able to isolate the target signals from the undesired signals from the back end. It is obvious that this depth is strongly dependent on the target zone specifications. It is also preferred to have a smallest down range depth to cut the construction cost and to minimize the space it occupies.

Suppose that one wants a circular target zone whose diameter is $D$ and the illumination field's amplitude and phase variation are required to be less than $\delta$ dB and $\epsilon$ radians, respectively. Since the signals from the back end can not be completely removed using background subtraction, one needs to do frequency domain smoothing
to remove them. Using this information, the minimum down range depth can be found.

From the field study of the RP and HCRP range models, it is found that the amplitude requirement is easier to satisfy than the phase one. Thus, one just needs to meet the phase variation requirement, and the amplitude variation is usually achieved at the same time. Since the incident field is a spherical wave, the maximum cross range phase variation, \( \phi \), in the target zone with its center located at a distance \( R \) for the feed antenna can be easily found to be

\[
\phi \approx \frac{2\pi}{\lambda_{\text{min}}} \frac{D^2}{8R}
\]

(6.15)

where \( \lambda_{\text{min}} \) is the shortest wavelength and \( (D/2R)^2 \ll 1 \) has been assumed. Since one requires that \( \phi < \epsilon \), the following condition for the distance \( R \) can be obtained, such that

\[
R > \frac{2\pi}{\lambda_{\text{min}}} \frac{D^2}{8\epsilon}.
\]

(6.16)

The condition given in Equation (6.16) then defines the required distance from the feed antenna to the target zone center for a given allowed amplitude and phase variation in the target zone. If one wants \( \epsilon < \frac{\pi}{2} \), for example, then \( R > \frac{2D^2}{\lambda_{\text{min}}} \), which is exactly the commonly used far field condition for antenna pattern measurements.

6.3 The Minimum Horizontal Flare Angle

In this section, the horizontal flare angle of an RP or an HCRP range will be determined. The choice of this angle is based on two considerations, the cross range size of the target zone and the possible excitation of higher order propagation modes. Let the horizontal flare angle be \( \psi_H \) and the target zone size be \( D \) at a distance \( R \), as shown in Figure 63. The minimum value of \( \psi_H \) can be decided by using the relationship derived in (6.14). First, one needs to plot the \( \rho_b = 4\beta + 3\rho \) curve, as
done in Figure 62, corresponding to the given $D, R$ and the frequency bandwidth of the operation. As discussed earlier, this curve gives the minimum distance for range clutter signals other than those of the target to be isolated and removed. Thus, if there is coupling between the side edges of the range and the target, the edges should be kept outside the $\rho_b = 4\beta + 3\rho$ curve. From the tangent points $P$ and $Q$ shown in Figure 63, one can easily find the required minimum angle $\psi_H$.

![Figure 63: Determination of the horizontal flare angle in RP and HCRP ranges.](image)

When the width of the range is greater than one half wavelength, higher order propagating modes may appear and cause large amplitude and phase variation in the target zone. In practice, the cross range dimension in a RP range or an HCRP range is usually greater than one half wavelength at the highest frequency. However, since the range is fed near the tip region where the dimension is just a small fraction
a wavelength, higher order propagating modes will be cut off in this region. Furthermore, both RP and HCRP ranges have very smooth edges which extend from the tip point to the target zone; thus, the chance of higher order mode excitation has also been significantly reduced. As a consequence, although the width of the cross range at the target zone is normally several wavelengths, higher modes have not been observed in experiments for both ranges when the feed antenna is located close to the tip and is positioned at the center between two edges. However, higher order modes have been observed when the feed antenna is too far away from the range tip. Another good reason for keeping the horizontal flare angle as small as possible is to decrease the dimension of the range, which minimizes the construction cost and its occupancy in the room. However, the edge waves which has been observed in the cross range field distribution reduce the available target zone size in the cross range direction. For example, from the previous findings, only 30% of the range width can provide less than 1 dB amplitude and 10 degrees phase variations.
6.4 The Maximum Vertical Flare Angle

Similar to the horizontal dimension, the vertical height of the range is normally kept less than half a wavelength at the highest frequency to avoid the problem of higher order propagating modes. However, since the range is fed in the tip region, as discussed earlier, and is smoothly extended from that smaller region to the larger target region, this half a wavelength limitation may be extended. If the height at the target zone is chosen to be less than a half wave length ($\frac{\lambda_{min}}{2}$) at the highest frequency ($f_{max}$) in an RP range with the shortest down range depth found in Section 6.2, as shown in Figure 64, then the maximum vertical flare angle is given by

$$\psi_V < \tan^{-1} \left[ \frac{2\epsilon/\pi}{(D/\lambda_{min})^2} \right]$$  (6.17)

![Diagram](https://via.placeholder.com/150)

Figure 64: Determination of the vertical flare angle in RP and HCRP ranges.
6.5 Summary

In this chapter, the methods to determine the values of three important range design parameters – minimum down range depth, minimum horizontal flare angle and the maximum vertical flare angle have been derived. These have been developed based on the satisfactory isolation from range error signals and the permissible amplitude and phase variation of the illuminating field in the target zone. Using these guidelines, one can construct an RP or HCRP range with good range performance and, at the same time, minimize the cost and room space for these ranges. Other factors, such as coupling between the target and the plates, multi-path and edge waves, which may also affect the above results are still under study.
CHAPTER VII
Applications and Measurement Examples

In this chapter, various measurements are performed using the previously described ranges, models §2 and §3. These examples are divided into the following three groups:

1. 3D Measurements - measuring the echo area of three dimensional (3D) targets such as conducting spheres and conducting plates.

2. 2D Measurements - measuring the echo width of two dimensional (2D) targets such as conducting cylinders and strips.

3. 1D Measurements - measuring the material dielectric constant of one dimensional (1D) layer targets such as glass, plexiglas, styrofoam, absorber and microwave laminate.

The various types of difficulties and errors associated with these types of measurements will be discussed along with each example. From these examples, valuable information for future improvement of these prototype models can be obtained. The good measurement performance and versatility of the proposed ranges will be shown through these examples.

It should be noted that since Models §2 and §3 are prototype models built for evaluation purposes, the design criteria discussed in Chapter VI have not been taken into account when they were built; as a result, some errors appearing in these
examples could be avoided if the proper design criteria had been used. These kinds of errors will be further discussed when they appear in the examples.

7.1 Three Dimensional (3D) Target RCS Measurement - Square Conducting Plate

The RCS measurement results for 0.5" and 0.25" conducting spheres have been shown in the Sections 4.5 and 5.6 where a 1" sphere was used as the calibration target. Those results have also been shown to agree well with the theoretical solutions. In this section, several conducting square plates will be measured as the other 3D example. The plates will be mounted broadside and the side length varied as listed below:

1. 1cm×1cm plate
2. 2cm×2cm plate
3. 3cm×3cm plate
4. 4cm×4cm plate
5. 5cm×5cm plate

It is noted that these plates were measured in both Models 2 and 3, where the height between the floor and ceiling of each range is about 5.7 cm. Thus, for plates (4) and (5), the top and bottom edges will be very close to the floor and ceiling of the range. The measurement results calibrated using a 1" sphere are shown in Figures 65–69. In each figure, the theoretical solution obtained from the moment method is also shown by the dash line, for comparison. One can clearly see that there is about a 4dB which occurs in each of these measurements. In addition, the differences between the calibrated data and the exact solutions are about the same.
for all cases. After carefully studying the data, it WAS found that this error was introduced by the calibration target – the 1" sphere.

As discussed before, there is an intrinsic error arising from the coupling between the target and the range's floor and ceiling conducting plates. Especially if a large amount of the power is scattered toward the floor and the ceiling, this error will become more significant. Due to the bistatic scattering nature of a sphere, which has a wide bistatic scattering pattern such as the one shown in Figure 16 for a 1" sphere, additional signals besides the desired direct backscattered signal will be received. These additional signals result from the interaction between the sphere and the range's plates. Some of these bouncing terms may be smoothed out if they have longer time delays with respect to the direct backscatter signal. Others will appear in the received data as measurement errors.

Consequently, the sphere is not an appropriate calibration target for these measurements. In order to obtain a more accurate calibration, one needs a target which has high directivity in the backscatter direction and whose theoretical solution can be found easily. A convenient choice is a dihedral corner reflector.

A dihedral corner reflector has been proposed to be used as an calibration target for RCS measurements due to its polarization property and different backscatter directivities in different scanning plane [19]. The theoretical scattered fields of a large dihedral have been calculated by using the PO and PTD method [20] or the UTD method [21]. The moment method is also a convenient way to calculate the scattered field for a small dihedral corner reflector.

The geometry of a dihedral corner reflector which is used as a calibration target in several measurements is shown in Figure 70. Its theoretical solution has been calculated using the moment method. Figures 71 shows the calculated E-plane bistatic scattering patterns of this dihedral corner reflector at 2 GHz. This pattern
Figure 65: The radar cross section (RCS) of the broadside of a 1cm×1cm square conducting plate. The data was calibrated with respect to a 1" conducting sphere.
Figure 66: The radar cross section (RCS) of the broadside of an 2cm×2cm square conducting plate. The data was calibrated with respect to a 1" conducting sphere.
Figure 67: The radar cross section (RCS) of the broadside of an 3cm×3cm square conducting plate. The data was calibrated with respect to a 1" conducting sphere.
Figure 68: The radar cross section (RCS) of the broadside of an 4cm×4cm square conducting plate. The data was calibrated with respect to a 1” conducting sphere.
Figure 69: The radar cross section (RCS) of the broadside of an 5cm x 5cm square conducting plate. The data was calibrated with respect to a 1" conducting sphere.
shows a much weaker scattering at wide angles. Since the range’s floor and ceiling are at the wide angles with respect to its backscatter direction, the coupling between them and the dihedral will be reduced. In this sense, a dihedral corner reflector is a better calibration target for 3D echo area measurements than a sphere.

The measured data of the previous plates calibrated with respect to the dihedral are shown in Figures 72 – 76. One should notice the good agreement between the measured and calculated data, except for the 5cm×5cm plate in the high frequency region. This plate, as noted earlier, has its top and bottom edges very close to the ceiling and floor of the ranges, and will suffer more significant coupling errors.

An interesting measured result is shown in Figure 77 where the measured RCS of a 1" sphere calibrated with respect to the reference dihedral corner reflector. Comparing the measured data with the exact solution (dot line), one can clearly see error caused by the coupling effect and why Figures 65 – 69 had these discrepancies. This result suggests that one can evaluate the errors due to the ranges’ plates by doing the sphere RCS measurement by calibrating with respect to a dihedral corner reflector. This knowledge is important for improving the range’s measurement performance.
Figure 70: The geometry of the reference dihedral corner reflector used in the measurement.
POLARIZATION: $\Theta$-IN $\Theta$-OUT
ELEV. PLANE: $\phi = 90^\circ$ $R = \infty$
MAXIMUM = $-26.4$ DB/M$^2$ 10 DB/DIV.

Figure 71: Vertical polarized bistatic scattering pattern of the reference dihedral corner reflector.
Figure 72: The radar cross section (RCS) of the broadside of an 1cm×1cm square conducting plate. The data was calibrated with respect to a dihedral corner reflector.
Figure 73: The radar cross section (RCS) of the broadside of an 2cm×2cm square conducting plate. The data was calibrated with respect to a dihedral corner reflector.
Figure 74: The radar cross section (RCS) of the broadside of an 3cm×3cm square conducting plate. The data was calibrated with respect to a dihedral corner reflector.
Figure 75: The radar cross section (RCS) of the broadside of an 4cm×4cm square conducting plate. The data was calibrated with respect to a dihedral corner reflector.
Figure 76: The radar cross section (RCS) of the broadside of an 5cm×5cm square conducting plate. The data was calibrated with respect to a dihedral corner reflector.
Figure 77: The radar cross section (RCS) of a 1" conducting sphere calibrated with respect to the reference dihedral corner reflector.
7.1.1 Conclusion

It has been shown in this section that accurate echo area measurements can be achieved in the new RP or HCRP range if the target is not strongly coupled with the ranges' floor and ceiling plates. It has also been shown that the dihedral corner reflector is much more suitable as the calibration target.

7.2 Two Dimensional (2D) Echo Width Measurement

Another feature of the classical PPTL range or the new RP and HCRP ranges is that they can measure the echo width of two dimensional targets; i.e., infinitely long cylindrical targets with arbitrary cross section. This is obtained by assuming that the floor and ceiling conducting plates in these range are infinitely large so that one can apply image theory [4] as illustrated in Figure 78. Another assumption has also been implied here, that is, these two plates are parallel to each other so that all the images lie along the same straight axis. Of course, this is true for a classical PPTL range. For a RP or HCRP range, the ceiling is not actually parallel to the floor. However, as long as the height is small, which has been required in the range design to avoid diffractions and higher order waveguide modes excitation, this error is negligible. The examples for echo width measurements of several conducting cylinders and strips are given below to prove this concept.
Figure 78: Applying image theory in the RP and HCRP range to measure the echo width of infinitely long cylinders with an arbitrary cross section.
7.2.1 Echo Width of 2D Strip

The first 2D measurement example is to determine the echo width of various strips with normal incidence (TM case). An measurement example of a 1" strip with edge-on incidence will also be presented. The width of the strips to be measured in the RP range and the HCRP range are list below:

- **RP** - 1", 2", 3", 4", 5", 6", 8", 14"
- **HCRP** - 1", 2", 3", 4", 5", 6", 8", 12"

The above strips were U-shaped such that they were in good contact with the floor and the ceiling as shown in Figure 79. A 0.5" cylinder was used as the calibration target whose theoretical solution has been found by including thirty terms in its eigenfunction expansion [4]. The calibrated strip echo width was then plotted along with their theoretical solutions obtained via the moment method. The results measured in the RP and HCRP ranges are shown in Figures 80 - 88.

Basically the measured data agrees well with the theoretical solutions, especially for the narrower strips (1" and 2"). For wider strip lines, one can clearly see some variations in the data which oscillate around the theoretical solutions. This variation was found to be caused by the fact that the separation between the targets and the back of the range was too short (≈ 1.2 ns) to be completely removed by smoothing the data. There are two main signals which are coming from the back. The first one happens when the back end is illuminated by the incident wave, it creates some reflected and diffracted fields which are later received by the receiver. Another one is the background change when part of the illumination is shadowed by the target put into the range. This change generates a pseudo signal returning from the back whenever background subtraction is used. This shadowing effect can also be thought of as the coupling between the target and the back end of the range. It is obvious
that the larger the target width the greater the shadowing. This can explain the increased errors associated with the wider strips. One should note that, in a real range, the distance between the target zone and the back end should be dictated by the criteria given in Chapter VI and any signals coming from the back can be sufficiently removed by smoothing. It is noted that these errors also exist in the 3D RCS measurements, however, the blockage by the targets in those cases was usually small so that those errors were not as significant as they are in the 2D echo width measurements.

It is also noted that the total width of the range where the strips were measured is about 18". So when the 12" strip was measured in the range its sides edges were very close to the side edges of the range and some errors will be contributed by the edge waves and the coupling between these edges and the strip. On the other hand, since wider strips have lower scattering sidelobes, this coupling with the side edge should be reduced somewhat.

Figure 89 also shows the echo width of a 1" strip with edge-on incidence (TM case). The data was obtained from both RP and HCRP ranges and was calibrated with respect to a 0.5" diameter conducting cylinder. A theoretical curve calculated via the moment method is also plotted for comparison. One can clearly see the very good agreement between the calculated and measured results.
Figure 79: Measuring the echo width of an infinitely long strip in the RP or HCRP range.
Figure 80: The echo width of a 1" strip measured in the RP and HCRP range models.
Figure 81: The echo width of an 2" strip measured in the RP and HCRP range models.
Figure 82: The echo width of an 3" strip measured in the RP and HCRP range models.
Figure 83: The echo width of an 4" strip measured in the RP and HCRP range models.
Figure 84: The echo width of an 5" strip measured in the RP and HCRP range models.
Figure 85: The echo width of an 6" strip measured in the RP and HCRP range models.
Figure 86: The echo width of an 8" strip measured in the RP and HCRP range models.
Figure 87: The echo width of a 12" strip measured in HCRP the RP range model.
Figure 88: The echo width of a 14" strip measured in the RP range model.
Figure 89: The measured echo width of a 1" strip with edge-on TM incidence.
7.2.2 Echo Width of 2D Circular Conducting Cylinders

The next 2D measurement example is to determine the echo width of circular conducting cylinders in the HCRP range. The diameters of the cylinders to be measured are $\frac{1}{4}''$, $\frac{1}{2}''$, $\frac{9}{8}''$ and $2''$, respectively, and the measured data is calibrated with respect to the $1''$ strip used in the previous section. These results are shown in Figures 90 - 93. The theoretical solution is calculated using the eigenfunction expansion method and is also plotted for comparison. Again, one can observe each measured data set tends to pass through the exact solutions except for a small oscillation error. These errors are mainly caused by the same mechanism as discussed

![Graph of echo width vs. frequency](image)

**Figure 90:** The echo width of an $\frac{1}{4}''$ circular conducting cylinder measured in a HCRP range model.
Figure 91: The echo width of an \( \frac{1}{2}'' \) circular conducting cylinder measured in a HCRP range model.

in the strip measurements; i.e., the undesired signals returning from the back of the range. One can also note that the period of this oscillation is about the same (\( \approx 0.7 \) GHz) for each case as well as the previous strip measurements. This is because both targets were placed at the same position. For the \( \frac{1}{4}'' \), \( \frac{1}{2}'' \), and \( \frac{9}{8}'' \) cylinders, the measured data passes through the exact solution correctly, except for the oscillation errors discussed earlier. As the dimension of the cylinder increases, it is found that the magnitude of the oscillation also increases similar to the strip measurements. In the \( 2'' \) cylinder case, the error is so large that the signal behavior has been obscured. This increase in the error magnitude is due the larger coupling
effect between the larger cylinder and the back end of the range. This can be seen from its time domain response as shown in Figure 94 (solid line), in which several bouncing terms between the cylinder and the back are observed. It also can be understood from the calculated bistatic echo width pattern of the conducting cylinder at different frequencies as shown in Figure 95. This figure also points out another problem with using a cylinder as a calibration target. That is, a cylinder has a very isotropic scattering pattern (i.e. -90 degree – 90 degree with incidence at 0 degree). Similar to a sphere in the 3D case, this nearly isotropic scattering may create some undesired signals bouncing between the two sides of the range, causing
calibration errors. When the dimension of the cylinder in terms of the wavelength increases, stronger scattering occurs in the backscatter direction and thus increasing the coupling between the cylinder and its background. As mentioned earlier, since the separation between the target and the back is not far enough in the HPRC range model, this coupling can not be completely removed and thus causes a larger error.

It is interesting to look at the third curve shown in Figure 93, which was measured by putting the same cylinder further away from the back of the range. One should note that the measured result has greatly improved and is very close to the exact solution. Its time domain response is shown in Figure 94 (dot line), which illustrates...
Figure 94: The time domain response of the 2" diameter conducting cylinder measured at two different positions.

that the coupling between the cylinder and the back has been largely reduced. The remaining error is caused by the coupling effects from the side edges which are now about 6" from the cylinder.
Figure 95: The calculated bistatic scattering pattern of a 2'' conducting cylinder.
7.2.3 Echo Width of Linearly Periodic Array

Next interesting example is to show how the echo width of an infinitely long linearly periodic structure can be measured in the RP and HCRP ranges. The example used here is an infinitely long array of bent wires, as shown in Figure 96. This was done by putting a thin wire section in the range so that it and its images form the desired array. This is also illustrated in Figure 97. Since, in the previous sections, the examples given for 2D measurements contain no mutual coupling between the two image elements, this example should provide more information concerning general 2D echo width measurements. It should be noted that the presence of the images will change the equivalent current distribution on the target, and thus causing a measurement error due to the violation of the second basic assumption made in Chapter II. The images will affect the measurements in such a way that they generate the signals bouncing between the range plates or between the target and the range plate. Some of these terms may bounce back toward the receiver. If they eventually reaches to the receiver then a measurement error will occurs. However, as discussed in Chapter VI, the geometry in the transition region tends to suppress these bouncing terms and reduces this error. Thus, another purpose of this measurement is to verify the above assumption.

Figure 98 shows the comparison between the measured and calculated data. The periodic moment method with sinusoidal basis and testing functions was used to calculate these results in 100 MHz step. The measured data was smoothed using 101 points. The length of the measured wire section was 8.2 cm and the wire length in each periodic cell is then 16.4 cm which is the wavelength at about 1.83 GHz. This agrees with the resonance frequency observed in Figure 98. The measured data tends to follow the calculated one very well. The bouncing terms can easily
Figure 96: The geometry of an infinitely long bent wire array.

Figure 97: Measuring the bent wire array in RP and HCRP ranges.

be excited by this 45 degree wire; however, no additional error was observed in this measurement.
Figure 98: The echo width of an infinitely long bent wire array.
7.2.4 Conclusion

In this section, the echo width of different sizes of conducting strips and cylinders have been measured in the HCRP range. This equivalence is obtained by applying image theory to the floor and ceiling plates which are uniquely used in PPTL, RP and HCRP ranges. Although, in earlier discussions, it was found that these plates cause a measurement error due to the bouncing terms in 3D target RCS measurements. The measurement results in this section clearly show that the presences of these plates make the accurate 2D echo width measurement become a unique feature of these ranges. It was also found that the strip is more suitable as a calibration target than a cylinder due to its lower bistatic scattering levels.
7.3 Material Characteristic Measurement with 1D Layer

From the previous studying, it was found that both the RP and HCRP ranges guide waves well at low frequencies. This fact makes it possible to do an accurate $S_{11}$ measurement. This important parameter can be used to find the material characteristics, such as permittivity and permeability of a homogeneous material layer. If both the permittivity and permeability are to be determined, one also must measure the $S_{21}$ parameter which can also be obtained in the RP or HCRP range either by introducing a transmission port or extracting the signals returning from the slab's second surface along. However, only the reflection coefficient and the dielectric constant will be measured in the following example. The dielectric constant of a type of absorber which is commonly used in anechoic chambers to reduce reflections will be measured in both RP and HCRP ranges. As a comparison, this absorber will also be measured in an HP805A Slotted Line.

Figure 99 illustrates how a homogeneous material layer is measured in the target zone of an RP range. As in 2D measurements, image theory can be applied to obtain an infinite dimension in the vertical direction. Measurement error due to the truncated horizontal dimension could be minimized by measuring different size samples. The measured reflection coefficient was calibrated with respect to the reflections from a perfect conducting wall. This reflection coefficient was used to find the corresponding complex dielectric constant. Two absorber layers, 9.5 and 6 cm thick, were measured in the RP and HCRP range, respectively. The resultant dielectric constants are shown in Figure 100. The solid lines also show the dielectric constant obtained from the transmission measurement in a HP805A Slotted Line. The layer thickness of this case was 8 cm. One can see that all the measured data are in reasonable agreement. It is also noted that the data measured in the RP
range and HCRP range show a little distortion at the end bands caused by the smoothing process used to remove the range clutter sources. The data from HCRP range seems to deviate more than the RP relative to the slotted line in the 300 MHz to 800 MHz frequency range. This was introduced in the calibration procedure where a conducting plate was used as a short-end reference. When this plate is put in the range and does not make good contact with the top and bottom plates of the range, slits may appear. Experiments showed that the presence of these slits cause unstable measurements at some frequencies depending on the size and the position of these slits. It was also found that even a tiny slit can cause a large error in the determination of the dielectric constant.

Figure 99: Measuring the $S_{11}$ of a homogeneous material layer.
Figure 100: Real and imaginary parts of the dielectric constant of an absorber measured in RP range model and HP805A slotted line.
7.4 Summary

In this chapter, examples of the radar cross section, echo width and dielectric constant measurements of 3D, 2D and 1D targets have been presented. All results have been shown to agree with their appropriate calculated data, which means that the new RP and HCRP ranges can provide accurate measurement results. These results have illustrated the wide variety of measurements that can be successfully performed in these types of ranges.
CHAPTER VIII
Conclusion

In this document, three kinds of low frequency wave guided ranges have been discussed. The classical parallel-plate (PPTL) range has been shown to contain internal diffraction problems which contaminate the field uniformity. These diffractions tend to corrupt the measurements and also reduce the available target zone size. Two new related ranges, the radial-plate (RP) and height-compensated radial-plate (HCRP) ranges, have been designed to overcome some of the problems associated with the PPTL range and provide much better target zone fields. Measurement results have shown that very accurate data can be obtained in these new ranges in a very efficient way. A wide variety of applications have been explored through the examples given. The previous study shows that both ranges have about the same measurement performance quality. The various attributes of these ranges are listed in Table 2. As can be seen by this summary both structure provide similar performance with the HCRP having additional advantage of a more uniform down range field distribution. Although it is more complex to build. So one should decide which range to use based on whether he intends to measure large down range targets or not.

Future work will focus on the following issues:

- Improve the matching condition at the feed point to increase the measurement efficiency and reduce the clutter.
• Reduce the clutter from the back end of the range.

• Reduce the edge waves which propagates along the two side edges.

• Introducing a transmission port for material measurements.
Table 2: Comparison between RP and HCRP range.

<table>
<thead>
<tr>
<th></th>
<th>RP</th>
<th>HCRP</th>
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<tbody>
<tr>
<td>CROSS RANGE FIELD AMPLITUDE</td>
<td>spherical wave + edge waves</td>
<td>spherical wave + edge waves</td>
</tr>
<tr>
<td>CROSS RANGE FIELD PHASE</td>
<td>spherical</td>
<td>spherical</td>
</tr>
</tbody>
</table>
| DOWN RANGE FIELD AMPLITUDE     | $1/R$                  | region A: $1/R$
|                                |                       | region B: $1/\sqrt{R}$
|                                |                       | region C: uniform     |
| DOWN RANGE FIELD PHASE         | $-jkR$                 | $-jkR$                 |
| TOP PLATE ANGLE                | constant angle         | region A: constant angle
|                                |                       | region B: horizontal
|                                |                       | region C: slowly varying |
| CLUTTER LEVEL                  | -90 dBSM              | -85 dBSM              |
| REFLECTION COEFFICIENT         | -20 dB                | -15 dB                |
| GEOMETRY COMPLEXITY            | simple                 | height varying required |
BIBLIOGRAPHY


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