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**MULTIPATH RESOLUTION
USING A ROTATABLE DUAL-POLARIZATION ANTENNA
FOR DIRECTION FINDING**

by
Lana D. Teany, B.Eng.

A thesis submitted to the
Faculty of Graduate Studies and Research
in partial fulfilment of the requirements
for the degree of

Master of Engineering

**Ottawa-Carleton Institute of Electrical Engineering
Department of Electronics
Carleton University
Ottawa, Canada
December 1990**

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submitted by Lana D. Teany, B.Eng.
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Prof. James S. Wight
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December 1990

ABSTRACT

The attainable accuracy of current direction finding (DF) systems is limited, primarily due to the wavefront distortion resulting from multipath (MP) propagation. A technique which exploits the polarization content of a wavefront is proposed in this thesis, as a means of obtaining more accurate measurements of direction-of-arrival (DOA) bearings. In this effort, a rotatable dual-polarization VHF direction finding antenna has been designed and fabricated, to measure the polarization diversity of a signal wavefront. Its operation is based upon a null-steering amplitude comparison technique. The antenna was tested in a controlled set of MP field measurements, and the experimental results were analyzed using a proposed DF correlation technique. Results indicate that this technique appears feasible for the estimation of transmitter bearings, in scenarios involving a single specular MP signal.

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LIST OF SYMBOLS AND ABBREVIATIONS

SYMBOLS

$\hat{x}$	unit vector in x direction
$\hat{y}$	unit vector in y direction
β	polarization angle or phase constant
β_2	secondary polarization angle
γ	grazing angle
$\Gamma_{\perp}$	perpendicular reflection coefficient
$\Gamma_{ }$	parallel reflection coefficient
δ	time-phase angle by which E_y leads E_x
Δ	angular separation between MP and DP signal
Δr	path length difference of reflected rays on irregular surface
ϵ	permittivity of medium
η	intrinsic impedance of medium
θ	bearing angle in azimuth plane
$\theta_{B }$	Brewster angle, or polarizing angle
θ_{DP}	transmitter bearing
θ_i	angle of incidence
θ_{MP}	MP signal bearing
θ_r	angle of reflection
θ_{refr}	angle of refraction
θ_t	angle of transmission
λ	signal wavelength
μ	permeability of medium
ρ	scattering coefficient
σ	conductivity of medium
τ	tilt angle of polarization ellipse
$\tau_{\perp}$	perpendicular transmission coefficient
$\tau_{ }$	parallel transmission coefficient
ϕ	signal phase
ω	signal angular frequency
a	loop element radius
A	multipath signal relative amplitude, general
A_{co}	MP signal co-polar relative amplitude
A_{HP}	horizontally polarized MP signal relative amplitude

$\mathbf{a}_n$	unit vector normal to surface
AR	axial ratio of polarization ellipse
A_{VP}	vertically polarized MP signal relative amplitude
A_x	MP signal cross-polar relative amplitude
A'	in-line MP attenuation
b	loop wire radius
c	speed of light or half-spacing separating loop turns
C	capacitance
d	dipole/loop element diameter or array element spacing
$\mathbf{E}$	electric field vector
$\mathbf{e}^+$	unit vector in direction of vertical (+) polarization
E^+	E-field component in direction of vertical (+) polarization
$\mathbf{e}^-$	unit vector in direction of horizontal (-) polarization
E^-	E-field component in direction of horizontal (-) polarization
E_{DP}	co-polar E-field of DP signal transmission
E_{EXT}	co-polar E-field of signals extraneous to MP model
E_{HORZ}	total received HP E-field
E_i	imaginary component of E-field
E_{MP}	co-polar E-field of MP signal transmission
E_r	real component of E-field
E_{VERT}	total received VP E-field
E_x	E-field component in x direction
E_{XDP}	cross-polar E-field of DP signal transmission
E_{XEXT}	cross-polar E-field of signals extraneous to MP model
E_{XMP}	cross-polar E-field of MP signal transmission
E_y	E-field component in y direction
F, f	frequency of signal
h	height of surface irregularity
$\mathbf{H}$	magnetic field vector
$\mathbf{k}$	propagation direction vector
l	dipole half-length
L	dipole length
$\mathbf{n}$	normal to local surface element
n	refractive index of medium
N	number of turns of loop element
OA	major axis of polarization ellipse
OB	minor axis of polarization ellipse

p	complex polarization factor
P	power
P_{HORZ}	horizontally polarized power reception
P_{VERT}	vertically polarized power reception
q	complex depolarization factor
R	receiving antenna
R^+, R^-	Fresnel reflection coefficients for vertical and horizontal planes
R_o	total ohmic loss resistance
R_m	Fresnel zone radius, in meters
R_r	radiation resistance
s	reflector wire spacing
t	time
T_1	transmitting antenna #1
T_2	transmitting antenna #2
u	propagation speed in medium
U	receiving antenna co-polar array pattern
U'	receiving antenna cross-polar array pattern
V	voltage
Z_{in}	input impedance
$\mathfrak{R}^+, \mathfrak{R}^-$	generalized reflection coefficients for vertical and horizontal planes

ABBREVIATIONS

AFC	automatic frequency control
AUT	antenna-under-test
balun	balance-to-unbalance transformer
BW	bandwidth
CW	continuous wave signal
dB	decibels
DF	direction finding
DFL	David Florida Laboratories
DOA	direction-of-arrival
DP	direct path propagation
DREO	Defence Research Establishment Ottawa
EMI	electromagnetic interference
EW	electronic warfare
HP	horizontally polarized signals

LOS	line-of-sight propagation
MHz	megahertz
MP	multipath propagation
op amp	operational amplifier
SNR	signal-to-noise ratio
SRI	Stanford Research Institute
SSE	sum of squared errors
VHF	very high frequencies (30-300 MHz)
VP	vertically polarized signals

CHAPTER 1

INTRODUCTION

1.1 INTRODUCTION

Direction Finding (DF) is a method of determining the direction-of-arrival (DOA) of a signal wavefront for the purpose of emitter location. Initial applications of DF systems were established in the early 1900's when the need for tactical emitter location arose, and through the years other uses have been found. Presently, DF systems are used for providing navigational aid to ships, for locating the origin of distress signals, and even for monitoring the migratory behaviour of wild animals; nonetheless, its uses in electronic warfare (EW) still prevail.

Advancement in DF research has led to the development of many different types of systems utilizing various different techniques. For the specific application of a landbased tactical VHF radio direction finding system, on which this thesis concentrates, strict requirements on system size, set-up time, and reliability have restricted the selection of such systems to one of three main types; these currently employed systems are the Watson-Watt, the pseudo-Doppler, and the phase interferometer [1, pp. 1-2], as will be discussed in Chapter 2. The maximum operational accuracy offered by current systems of these types can be expected to be in the range of 2-3 degrees [1], [2], [3]. Unfortunately, this accuracy is insufficient for current targeting purposes and thus research is continually being performed in an attempt to increase this value.

The accuracy of a DF system denotes the exactness with which a bearing determination can be made. This accuracy can be described in terms of two different measures, i) the equipment accuracy and ii) the operational accuracy. The

equipment accuracy is a generalized measurement of the entire system accuracy, and it is obtained while operating in an ideal environment¹. Resulting bearing errors may be due to poor system calibration, system orientation errors, receiver errors, etc. These errors are consistent between operating environments, and furthermore they can be precisely determined and designed to be within acceptable values. The operational accuracy, however, is a measure of the bearing accuracy obtained when operating in a realistic environment. Unfortunately this accuracy is directly dependent upon the selection of operating site, and thus it may vary strongly between system locations. Errors contributing to the degree of this accuracy are largely due to multipath (MP) propagation, in which signals arrive at the DF antenna along paths other than, and in addition to, the direct path (DP). Reception of these erroneous signals may result in large biasing errors of the determined DOA. Consequently, it is the objective of most current DF research projects to concentrate on increasing the operational accuracy by resolving the MP interference on DF systems.

1.2 BACKGROUND AND OBJECTIVES

The approach taken in this thesis is to utilize the polarization content of a signal, in an investigative study. A signal which is transmitted with a specific polarization should, assuming undisturbed free space propagation, maintain its polarization upon reception. However, once a signal has undergone a MP reflection, refraction, diffraction or reradiation it may be subject to varying degrees of signal depolarization. In other words, a measurement of the cross-polarization (cross-polar) response will indicate the receiving antenna response to the incoming cross-polar MP

¹ The term ideal environment refers to the fact that the bearing error measurement is not dependent on the test site selection. Instead, this measurement is an indication of only pure system errors.

signals only²; the co-polarization (co-polar) response however will measure the DP signal plus any existing co-polar MP signals. It is hypothesized then that the measurement and utilization of these two responses, for a given propagation environment, will enable the 'normalization' or effective removal of the MP disturbances. Bearing this in mind, in a 'null-steering' amplitude comparison DF technique the DOA accuracy will then be limited by the null definition of the antenna response along with any other system inaccuracies. Assuming these system inaccuracies to be small, then with the appropriate selection of antenna the nullwidth can be made quite narrow thus significantly increasing the DF accuracy.

It is the objective of this thesis to investigate this hypothesis and to evaluate its usefulness as a DF technique. To achieve this objective, a rotatable dual-polarization DF antenna is designed and built. This antenna is then used in a simple experimental evaluation of the proposed technique, and an analysis is performed of the measured data.

1.3 THESIS ORGANIZATION

Chapter 2 reviews some fundamental theory relevant to an understanding of the research topic. More specifically, two main topics are discussed: i) an introduction to current direction finding systems, and ii) a presentation of VHF propagation and multipath. The former topic will acquaint the reader with some current DF systems being used for landbased tactical VHF radio direction finding. Following this an examination will be made of some general effects of multipath propagation on signal polarization.

In Chapter 3, the specifics of the dual-polarization VHF direction finding antenna design are examined. Included in this chapter is a statement of the

² With this statement it is assumed that the receiving antennas exhibit good cross-polar isolation.

objectives and specifications of this design, along with a presentation of the measured antenna gain patterns.

A discussion of the purpose and procedure of the experimental measurements is given in Chapter 4, followed by a presentation of the measured experimental responses. Included in this chapter are relevant details useful in understanding and interpreting the obtained data.

Chapter 5 consists of an analysis and a theoretical evaluation of the experimental responses presented in the previous chapter. This analysis incorporates the use of a simulation routine as a comparison tool, in an attempt to correlate measured and theoretical data.

Finally, Chapter 6 summarizes the primary conclusions of the thesis and provides recommendations for possible future work.

CHAPTER 2

THEORY OF DIRECTION FINDING SYSTEMS AND MULTIPATH

This chapter introduces the basic theory relevant to the research topic. Emphasis is placed on acquainting the reader with the fundamentals of the direction finding technique, and with primary VHF propagation principles. Furthermore, the latter topic concentrates primarily on signal polarization.

2.1 INTRODUCTION TO DIRECTION FINDING

The subject of direction finding has become well established, with its origins dating back to the early 1900's. The basic intent of the DF technique is to measure a bearing corresponding to the bearing of a specific signal transmission, and this technique is discussed by many different authors [2], [5], [6], [7], and [8].

The measurement method of determining the DOA of a signal may generally be classified into three basic groups; these classifications are

- a) Amplitude Measurement DF,
- b) Phase Measurement DF, and
- c) Time Delay Measurement DF.

Actually, all of these methods initially utilize the phase content of the field vectors yet the sampled signals may then undergo an additional encoding process [8, p. 10-9], e.g., amplitude measurement DF is the transforming of phase relations into amplitude functions of the signal DOA, as is done in the Watson-Watt system.

2.1.1 Conventional DF Systems

Three conventional¹ DF systems for landbased tactical operation at VHF frequencies are the Watson-Watt, the pseudo-Doppler, and the phase interferometer [1, pp. 1-2]. The basic operation of these systems shall now be introduced.

i) Watson-Watt DF System:

The Watson-Watt is an example of an amplitude measurement DF system. Its operating principle is based on the simultaneous measurement of the magnitudes of two orthogonal components of the incident signal, with these components then being used as arguments in the arctangent function to determine the signal bearing. The system is realized by the use of two directional antennas located on orthogonal axes, with the requirement that the antennas have cosine gain patterns. At VHF frequencies, these antenna patterns are generally obtained by the use of two orthogonal two-element Adcock arrays.

Shown in Figure 2.1a [8, p. 10-71] is the antenna polar pattern illustrating the amplitude relationships of the signal components. The DOA of the signal θ is computed from the equation

$$\theta = \arctan\left(\frac{b}{a}\right) \quad (2.1)$$

with the orthogonal amplitude components a and b given by

$$\begin{aligned} a(\theta) &= K \sin(\theta + 90^\circ) = K \cos \theta, \\ b(\theta) &= K \sin \theta, \end{aligned} \quad \text{with } K \ll 2\pi \quad (2.2)$$

and with K given by $K = 2\pi d/\lambda$, where K is the spacing between antenna elements

¹ These systems have reached a mature state of their development.

in electrical degrees at the signal frequency, d is the element physical spacing and λ is the signal wavelength.

ii) Pseudo-Doppler DF System:

Illustrated in Figure 2.1b [8, p. 10-57] is the arrangement of a pseudo-doppler DF system: which consists essentially of a circular array of omni-directional antennas of constant spacing. The system name is derived from the fact that it uses the doppler frequency shifting principle for determining bearings.

The operational principles of this system consist essentially of i) the sequential sampling of each array element and ii) a notation of the resulting phase change between adjacent elements. An observed phase change of 2π rad/s is equivalent to a doppler frequency shift of 1 cycle; thus the effect of sequential sampling is to essentially phase modulate the received signal at a frequency equivalent to the rotation frequency ω_R .

If the rotation frequency is constant then the DOA of the signal may be determined by noting the position at which the doppler shift is maximal; at this point the DOA line is tangent to the circle and it is in the same direction of the sequential element sampling. Alternatively, a doppler frequency shift of zero may be used to indicate when the measuring antenna is tangential to the equiphase lines of the signal wavefront.

iii) Phase Interferometer DF System:

A simple phase interferometer DF system is composed of two antennas along a baseline, spaced a distance d apart, as shown in Figure 2.1c [4, p. 8]. With this system, the DOA of a signal propagating in the azimuth plane can be calculated by simultaneously measuring the phase of the signal incident at each element. The measured phase difference ϕ between the elements is expressed as

$$\phi = \frac{2\pi d \sin \theta}{\lambda} \quad (2.3)$$

where θ is the bearing angle of the incident wave and λ is the signal wavelength.

Rearranging this equation, the DOA can be determined from

$$\theta = \arcsin\left(\frac{\lambda \phi}{2\pi d}\right) \quad (2.4)$$

Unfortunately there is a 180° ambiguity in DOA measurements using this two-element array. To resolve this, a quadrature (square) array of two pairs of elements can be formed which share a common phase centre. With this configuration the phase difference for each element pair is given by

$$\phi_1 = \frac{2\pi d \sin \theta}{\lambda} \quad \text{and} \quad \phi_2 = \frac{2\pi d \cos \theta}{\lambda} \quad (2.5)$$

and the DOA is determined from

$$\theta = \arctan\left(\frac{\phi_1}{\phi_2}\right) \quad (2.6)$$

2.1.2 Rotating Cardioid DF System

The DF system employed in the experiments of this thesis is that of a rotating cardioid, as described by deRosa [8], and it is a type of amplitude measurement or 'null-steering' DF system. This system utilizes a directive antenna array by rotating the array in azimuth and thereby amplitude-modulating the received signal at the rotational rate. Thus, the transmitter bearing may be obtained by noting the orientation of the array corresponding to minimal signal reception.

The cardioid antenna pattern is obtained from the arrangement of two

a) Watson-Watt Antenna Polar Pattern

b) Pseudo-Doppler Element Arrangement

Figure 2.1 - Direction Finding Systems

c) Two-Element Phase Interferometer

d) Cardioid Antenna Pattern

Figure 2.1 - Direction Finding Systems

elements of omni-directional strength in the azimuth plane, separated by a spacing of $\lambda/4$ and with a quadrature phase relationship. This pattern is shown in Figure 2.1d [9, p. 125] and the complete design details of the dual-polarization VHF DF antenna, which is used in this thesis, are given in Chapter 3.

2.2 VHF PROPAGATION AND MULTIPATH

The modes of propagation of an electromagnetic wavefront are many and varied, with dependence on both the operating frequency and the path length. In addition, a propagation mode may be reclassified as unwanted multipath when it causes interference with the desired signal reception.

Section 2.2.2 discusses various radiowave propagation modes specific to the VHF band, i.e., 30-300 MHz, and it is followed by an introduction to multipath propagation in Section 2.2.3. In this latter section, specific reference is made to probable causes of multipath in the VHF band. In Section 2.2.4, an examination is made of a signal's polarization as it undergoes a multipath reflection. This discussion includes both smooth and rough surface evaluations, and a criterion for the classification of both. Prior to the aforementioned discussions, Section 2.2.1 deals with an important characteristic of an electromagnetic wave: its polarization.

2.2.1 Signal Polarization

Balanis [10, p. 48] defines the polarization of a propagated signal as "*that property of a radiated electromagnetic wave describing the time-varying direction and relative magnitude of the electric-field vector*". In its most general form, the orientation of this electric field vector $\vec{E}$ can be defined in terms of a polarization ellipse. Furthermore this ellipse can be expressed in terms of two linearly polarized components, E_x and E_y , such that

$$\vec{E} = E_x \hat{x} + E_y \hat{y} \quad (2.7)$$

where $\hat{x}$ and $\hat{y}$ are unit vectors in the x and y directions, respectively.

An illustration of such an ellipse is given in Figure 2.2a. It is characterized by an axial ratio AR expressed as

$$AR = OA/OB \quad (1 \leq AR \leq \infty) \quad (2.8)$$

which quantitatively expresses the ratio of the major to minor axes. In addition, the polarization ellipse may take on any orientation in which a tilt angle τ defines the angle separation between the major axis and the horizontal x-axis.

For a wave travelling in the positive z direction, out of the page, the two electric field components of this general ellipse as a function of time and position, referring to Figure 2.2a, are given by

$$E_x = E_1 \sin(\omega t - \beta z) \quad (2.9)$$

$$E_y = E_2 \sin(\omega t - \beta z + \delta) \quad (2.10)$$

where

E_x = amplitude of E-field component in x direction

E_y = amplitude of E-field component in y direction

E_1 = amplitude of wave linearly polarized in x direction

E_2 = amplitude of wave linearly polarized in y direction

δ = time-phase angle by which E_y leads E_x

ω = angular frequency

β = phase constant

There are two extreme cases of elliptical polarization: circular polarization and linear polarization. These two states are illustrated in Figures 2.2b and c, respectively, and are characterized by

$$\begin{aligned}
\text{circular polarization} - \quad & E_2 = E_1 \\
& \delta = \pm 90^\circ \\
& AR = 1 \\
\text{linear polarization} - \quad & OA = 1 \\
& OB = 0 \\
& AR = \infty
\end{aligned}
\tag{2.11}$$

The polarization type of primary concern in this thesis is linear, both vertical and horizontal. Shown in Figure 2.2c is a vertically polarized plane wave travelling in the positive z direction, out of the page. The electric field of this wave as a function of time and position is given solely by Equation (2.10) with $\delta = 0$, since $E_x = 0$. Similarly, horizontal polarization is expressed by the E-field of Equation (2.9) and it is represented by a vector orthogonal to that of Figure 2.2c.

The terms co-polarization (co-polar) and cross-polarization (cross-polar) are repeatedly used throughout this paper. Here co-polar refers to the polarization equivalent to a reference signal polarization, and cross-polar refers to the polarization orthogonal to the same reference signal polarization. The term depolarize is also used, and it signifies the process of a signal's initial polarization being altered.

2.2.2 Radiowave Propagation Modes

Wave propagation consists of mutually perpendicular electric and magnetic field lines propagating in an equi-phase wavefront. This wavefront travels along the direction perpendicular to both field lines in a direction in agreement with the right hand rule.

Propagation of a radiowave may occur via one of many types of modes. The type of signal propagation mode utilized in a system is dictated by the system, the system application, the operating frequency, and the transmission path length. A number of signal propagation modes exist and some of particular interest to the tactical communicator are illustrated in Figure 2.3 [11, p. 2-2].

a) Polarization Ellipse at Tilt Angle τ

b) Circular Polarization

c) Linear Polarization

Figure 2.2 - Illustration of Polarization States [9, pp. 71-72]

Figure 2.3 - Signal Propagation Modes

In this thesis, concentration is placed on the operation of a landbased, VHF (60 MHz), dual-polarization direction finding antenna over a short range (tens of km's). For this application the propagation emphasis is on ground wave signals, more specifically dealing with space waves.

The propagation of a ground wave refers to the broad category of waves which travel close to the earth, and which do not reach the receiver via a reflection or scatter by the ionosphere (upper atmosphere). This category is then further divided into two main subdivisions: space waves and surface waves.

A surface wave is one which travels in direct contact with the earth's surface, along the earth/troposphere (lower atmosphere) interface. It is generated by a transmitting antenna which induces current in the earth's surface. The efficient transmission of such a mode, however, is dependant on several variables. The conductivity σ and permittivity ϵ of the ground control the wave attenuation; in addition, so does the frequency of operation since attenuation increases with increasing frequency. The efficiency of transmission decreases also with increasing antenna height from ground, as well as with a deviation of the transmission polarization from vertical.

The second subdivision of wave modes is space waves. All received ground waves, apart from surface waves, fit into this classification. In general, these waves travel through the troposphere and thus are not subject to the same propagation inefficiencies, with poor earth conditions, as are surface waves.

Analysis of a received signal is generally simplified by assuming that the instantaneous reception of several different propagation modes may each be treated independently and that the resultant response is the vectorial addition of all these signals. Thus the reception of the ground wave consists of the dual reception of both surface and space waves. However, if the height of the receiving antenna is made high enough then it may be assumed that, at the frequency of interest, the reception of surface waves is negligible in comparison to space wave reception.

It now remains necessary to further classify space waves and to examine any polarization dependent characteristics within these. A space wave may arrive at a receiver via a line-of-sight (LOS) mode or a non-LOS mode, see Figure 2.3. Under the classification of LOS is direct wave propagation, and all of the other remaining space wave modes may generally be viewed as multipath propagation. The following section elaborates on the origin and effects of these multipath modes.

2.2.3 Multipath Propagation

Multipath is the phenomenon whereby electromagnetic waves arrive at the receiving antenna along paths other than, and generally in addition to, the direct path. When dealing with VHF propagation of a space wave, the phenomena of multipath may be realized by any of several different means. These signal propagation disturbances may be generalized into the categorization of four main groups: reflection, refraction, diffraction, and reradiation. Welch [12, pp. 54-96] discusses these topics, and each will now be introduced and briefly described.

i) Reflection:

Reflection of an electromagnetic wave, partial or whole, occurs when a signal becomes incident upon a second transmission medium along its propagation path. This signal, upon contact with the interface, returns either partially or completely to its initial medium rather than being fully transmitted through. The reflected signal properties (strength, phase, polarization, and direction of reflection) are functions of the reflecting surface and the signal angle of incidence.

The interface or plane between the two adjoining media may be defined as either 'smooth' or 'rough'. The general distinction between these is that a smooth plane gives rise to specular reflection in a single direction, whereas a rough plane will cause a diffuse scattering of the incident signal in many directions.

The laws governing smooth plane reflection are well-known and the reflected signal will depend on several parameters: the signal wavelength, the angle of incidence, and the electrical properties (permittivity ϵ , permeability μ , and conductivity σ) of the two adjoining media. Unfortunately though, the analysis of rough surface scatter is much more complex, and it involves the computation of many different variables. No general solution has yet been established, but progress has been made in the analysis of several specific cases. A more detailed examination of the effects of these two reflection types will be given in the next section.

A reflected signal may reach the receiving antenna by either of two ways: LOS propagation with reflection off of the ground, and/or with reflection off of any interfering structures on the earth's surface. In this case ground refers to both terrain and water, and the structures mentioned may include vegetation, trees, buildings, etc. Shown in Figure 2.4a is an illustration of MP reflection.

ii) Refraction:

Refraction of an electromagnetic signal refers to the bending of the signals propagation path upon its transmission i) from one medium into another of differing dielectric constant, or ii) through a medium of differing temperature and/or density. This bending results from a change in propagation speed of the wave upon entrance to the second medium. The resulting wave path (i.e., defined direction, degree of bending, and rate of bending) is dependent upon the characteristic relative speed of propagation of the wave in the second medium, and upon the continuity of change in the dielectric constant at the medium interface.

A measure of the degree of wave bending of an incident signal, relative to straight path propagation, is given by the refractive index of the medium. This is expressed as

$$n_2 = \frac{\sin \theta_i}{\sin \theta_{\text{refr}}} n_1 \quad (2.12)$$

where n_1 and n_2 are the refractive indices of the initial and final propagation mediums respectively, θ_i is the angle of incidence, and θ_{refr} is the angle of refraction. Note that for free space propagation of the incident signal, n_1 has an approximate value of unity. The refractive indices are given by the relations

$$n_1 = c \sqrt{\mu_1 \epsilon_1} = \frac{c}{u_1} \quad n_2 = c \sqrt{\mu_2 \epsilon_2} = \frac{c}{u_2} \quad (2.13)$$

where μ , ϵ and u are respectively the permeability, permittivity, and propagation speed quantities in their respective mediums, and c is the speed of light.

For VHF signal propagation, an example of refraction is ducting. Ducting is the phenomenon whereby a transmitted signal is refracted towards the earth at a certain height in the troposphere and then reflected back upon contact with the earth, thus confining the signal to propagate within a thin layer of the atmosphere near the earth. The initial signal refraction is caused by a gradual decrease in the index of refraction of the atmosphere with increasing height, this being sufficient to return the signal to earth. Note that ducts are primarily formed due to a variation of water vapour content, and thus they commonly exist over water but not land. Figure 2.4b demonstrates the operation of a duct.

iii) Diffraction:

Diffraction is the phenomenon whereby a radiowave bends to some extent as it passes over or around an obstacle, rather than being totally blocked by this obstacle. The region into which this bending occurs is termed the shadow zone or shadow region, which essentially is the geometrical shadow that one would envision by sight. The degree of bending is a function of both the propagation wavelength, and the edge thickness of the object. Furthermore, the diffraction effect is observed to increase with either an increase in wavelength or a decrease in edge thickness.

Diffraction effects on signals at VHF frequencies may occur with obstacles such as trees, buildings, ridges, hills, mountains, etc., basically for any obstacle which is comparable or larger in size to the signal wavelength. By considering the signal wavefront to consist of half-period zones of alternating phase with respect to the LOS, which are called Fresnel zones, then the partial signal blockage by an obstacle may be determined by a summation of the contributing zones, as suggested by Welch [12, pp. 90-93]. Using this analysis, the relative field strength for diffraction by a straight edge is illustrated in Figure 2.4c. Also, Figure 2.4d shows an example of radiowave diffraction by a ridge where the signal strength received at the different receiving antennas depends on the Fresnel zone distance from the free-space propagation path, as given in Figure 2.4c.

It is generally accepted that if there is no obstacle within 0.6 of the first Fresnel zone from the propagation path, then the diffraction effects on the transmitted signal may be assumed negligible. A measure of the first Fresnel zone radius R_m , in meters, is given by Freeman [13, p. 180] as

$$R_m = 17.3 \sqrt{\frac{d_1 d_2}{FD}} \quad (2.14)$$

where F = frequency of signal (in GHz), d_1 = distance from transmitter to path obstacle (in km), d_2 = distance from path obstacle to receiver (in km), $D = d_1 + d_2$, total path length (in km).

iv) Reradiation:

Reradiation is the concept of a propagating signal wavefront intercepting an antenna structure, thus causing a modified wavefront to emanate from it. This modified wavefront will possess characteristics of the reradiation source. Such qualities may include a directional signal strength defined by the secondary antenna radiation pattern, and a polarization defined by the reradiating elements. In addition, a change in signal phase may result. An example of signal reradiation is

a) Reflection

b) Refraction

Figure 2.4 - Categorization of Multipath Propagation

T - transmitting antenna
 R_a, R_b, R_c - receiving antennas

c) Relative Field Strength E for Diffraction by a Straight Edge

d) Diffraction by a Ridge, for Different Receiver Heights

e) Reradiation

Figure 2.4 - Categorization of Multipath Propagation

given in Figure 2.4e.

The above introduction has commented on some of the basic VHF propagation modes. It is important for one to realize that the reception of a VHF signal may include one or more of these unwanted multipath signals, in which case the total received signal is the summation of each individual component. Furthermore, each signal component may be defined by its own amplitude, phase and polarization, and thus it may possibly combine constructively or destructively with the other signal components. For this reason, it is important for one to be aware of all possible signal interference effects and to appropriately account for these in all ensuing analyses.

For reference, a paper by Saxton [14] presents measurements of attenuation caused by trees and buildings on VHF signals, for both vertical and horizontal polarization. In addition, a similar paper by Rice [15] expands on these terrain cover effects on VHF propagation.

Of specific interest to the experimental measurements of Chapter 4 is the analysis of reflections. Consequently, the next section will concentrate on the effects of smooth and rough plane reflection, with specific emphasis on signal polarization.

2.2.4 Effects of Reflection on Signal Polarization

As mentioned in Section 2.2.3, reflection can be categorized into two general groups: i) reflection from a smooth surface, and ii) reflection from a rough surface. The theory behind smooth surface specular reflection is well-known, including the effects on signal polarization. Unfortunately though, rough surface diffuse scattering is a much more complex topic and further research is required. In the following, the behaviour of a specular reflection will be briefly analyzed according to its governing laws. Also an approach to rough surface scattering, as given by Beckmann [16], will be presented as an introduction to the complexity of the problem at hand.

Before commencing with these analyses, it is primarily important that the reader be able to qualitatively classify a reflecting surface as being either smooth or rough. A derivation by Rayleigh [16, pp. 9-10] shall be used in which a relation is formulated involving three propagation parameters: i) signal wavelength, ii) surface roughness, and iii) signal grazing angle. This derivation is called the 'Rayleigh criterion' and it is illustrated in Figure 2.5 [16, p. 9].

In this figure, rays 1 and 2 are incident on an irregular surface and they are observed to have, upon reflection, a path length difference Δr given by

$$\Delta r = 2h \sin \gamma \quad (2.15)$$

and a corresponding phase difference $\Delta \phi$ of

$$\Delta \phi = \frac{2\pi}{\lambda} \Delta r = \frac{4\pi h}{\lambda} \sin \gamma \quad (2.16)$$

where h is the height of the surface irregularity, γ is the grazing angle, and λ is the signal wavelength.

If the phase difference indicated by Equation (2.16) is sufficiently small, i.e., $\Delta \phi \approx 0$, then the two rays will be approximately in phase and the surface can be classified as smooth. However, as the phase difference approaches π the two rays start to interfere destructively, thus cancelling the net energy in the reflected direction. This energy is then directed elsewhere, more specifically in diffusely scattered directions.

Rayleigh uses this fact to arbitrarily choose a mean value of $\Delta \phi = \pi/2$ as the phase criterion² distinguishing the two surface classifications. Substituting this into Equation (2.16) one obtains the 'Rayleigh criterion' which defines the maximum surface irregularity height for a surface to be considered smooth. This is given as

$$h < \frac{\lambda}{8 \sin \gamma} \quad (2.17)$$

² Although the selection of a mean value of phase seems to be a very crude criterion, it is widely accepted and used in engineering applications.

Figure 2.5 - Derivation of the Rayleigh Criterion for Surface Roughness

where h , λ , and γ are as previously defined.

i) Specular Reflection:

The mathematics involved in the derivation of the following equations is very lengthy but relatively straightforward. Consequently just the relevant results are presented with a short introduction; clarification of any point can be made by viewing Sadiku [17, pp. 461-478].

Figure 2.6a illustrates the oblique incidence of a plane wave to a smooth boundary, as dealt with by Sadiku [17]. Here $\mathbf{k}$ is the propagation vector defining the direction of wave propagation, with a magnitude of $k^2 = k_x^2 + k_y^2 + k_z^2 = \omega^2 \mu \epsilon$. Additionally, θ_i is the angle of incidence, θ_r is the angle of reflection, θ_t is the angle of transmission, and $\mathbf{a}_n$ is a unit vector normal to the boundary. The plane of incidence is the plane defined by $\mathbf{k}$ and $\mathbf{a}_n$, and the vectors $\mathbf{E}$, $\mathbf{H}$, and $\mathbf{k}$ are all mutually orthogonal.

The fact that the tangential component of $\mathbf{E}$ must be continuous at the boundary $z=0$ leads to the result $\mathbf{E}_i(z=0) + \mathbf{E}_r(z=0) = \mathbf{E}_t(z=0)$. Using this boundary condition and Maxwell's equations to define the waves, one obtains the results $k_i \sin \theta_i = k_r \sin \theta_r$ and $k_i \sin \theta_i = k_t \sin \theta_t$. Finally utilizing the condition for lossless media, $k_i = k_r = \omega(\mu_1 \epsilon_1)^{0.5}$, leads to the results

$$\theta_r = \theta_i \quad (2.18)$$

and

$$n_1 \sin \theta_i = n_2 \sin \theta_t \quad (2.19)$$

where $n_1 = c(\mu_1 \epsilon_1)^{0.5}$ and $n_2 = c(\mu_2 \epsilon_2)^{0.5}$ are the refractive indices of media 1 and 2, respectively. Equation (2.19) is known as Snell's law which defines the direction of specular reflection.

Specular reflection can be analyzed thoroughly by studying the reflection of a plane wave at oblique incidence to a smooth boundary for two special cases: one case in which the $\mathbf{E}$ -field is parallel to the plane of incidence, and the other in which

the E-field is perpendicular to it. A wave with any other polarization may be considered as a linear combination of these two cases.

a) Parallel Polarization:

For parallel polarization the E-field lies in the xz plane, the plane of incidence, as shown in Figure 2.6b. Once again Maxwell's equations are utilized to obtain the field components. Requiring that $\theta_r = \theta_i$ and that the tangential components of **E** and **H** be continuous at the boundary $z=0$, leads to the derivation of Fresnel's equations describing the parallel reflection and transmission coefficients, respectively, as

$$\Gamma_{||} = \frac{E_{ro}}{E_{io}} = \frac{\eta_2 \cos \theta_i - \eta_1 \cos \theta_i}{\eta_2 \cos \theta_i + \eta_1 \cos \theta_i} \quad (2.20)$$

and

$$\tau_{||} = \frac{E_{to}}{E_{io}} = \frac{2\eta_2 \cos \theta_i}{\eta_2 \cos \theta_i + \eta_1 \cos \theta_i} \quad (2.21)$$

where E_{ro} , E_{io} and E_{to} are respectively the reflected, incident and transmitted electric field magnitudes at $z=0$, and η_2 and η_1 are complex quantities representing the intrinsic impedances of media 2 and 1, respectively.

The reflection and transmission coefficients given by Equations (2.20) and (2.21) characterize the relative magnitude and phase of the reflected and transmitted components, respectively, relative to an incident wave on a specified boundary at a defined angle of incidence. Observing Equation (2.20) one notes that a condition exists where $\Gamma_{||} = 0$. This condition occurs for an incident signal at the Brewster angle $\theta_{B||}$, or the polarizing angle. At this point an arbitrarily polarized incident wave will be reflected with only the component of **E** perpendicular to the plane of incidence. Restated, an arbitrarily polarized wave incident on a boundary at an angle of incidence $\theta_i = \theta_{B||}$ will have only its cross-polar component reflected.

b) Perpendicular Polarization:

The case of perpendicular polarization is illustrated in Figure 2.6c, and it has the E-field perpendicular to the plane of incidence, the xz plane. A similar analysis to that performed for parallel polarization is carried out, and the following results are obtained for the perpendicular reflection and transmission coefficients:

$$\Gamma_{\perp} = \frac{E_{r\perp}}{E_{i\perp}} = \frac{\eta_2 \cos \theta_i - \eta_1 \cos \theta_t}{\eta_2 \cos \theta_i + \eta_1 \cos \theta_t} \quad (2.22)$$

and

$$\tau_{\perp} = \frac{E_{t\perp}}{E_{i\perp}} = \frac{2\eta_2 \cos \theta_i}{\eta_2 \cos \theta_i + \eta_1 \cos \theta_t} \quad (2.23)$$

where the parameters are as defined for the parallel case.

Rearrangement of Equation (2.22) to examine the instance where $\Gamma_{\perp} = 0$ reveals that a Brewster angle $\theta_{B\perp}$ does not normally exist.

The following concluding remarks can be made from the above analysis. Primarily, the reflection coefficient for either incident orthogonal polarization is dependent on the angle of incidence, the angle of transmission, and the intrinsic impedances of the media. In general, a specularly reflected linear field is not depolarized if the incident wave is of either pure parallel ($E \perp y$) or pure perpendicular ($E \parallel y$) polarization relative to the plane of incidence. Neither will there be depolarization if the reflection coefficients for parallel and perpendicular polarization are equivalent. However, the only important occurrence of this latter case arises near grazing incidence onto a plane of finite conductivity, at which point both reflection coefficients are approximately -1.

Any other incident wave will generally encounter some degree of depolarization due to unequal reflection of the orthogonal E-field components. In addition, for a perfectly conducting reflection plane an incident wave of a perpendicular (parallel) polarization will encounter a 180° (0°) phase change. For a non-perfect conducting interface the phase change is dependent on the phase terms

a) Oblique Incidence Reflection Geometry

b) E Parallel to Plane of Incidence

c) E Perpendicular to Plane of Incidence

Figure 2.6 - Illustration of Specular Reflection [17, pp. 472-476]

of the intrinsic impedances. Also note that for a rotational polarization the sense of rotation is reversed upon reflection.

In the exceptional case in which the angle of incidence is equivalent to the Brewster angle, an arbitrarily polarized incident wave will have only its component of perpendicular polarization being reflected. This means that at $\theta_i = \theta_{B||}$, an incident wave of a parallel polarization will ideally have no reflection but an incident wave of a perpendicular polarization will be totally reflected. Note that these ideal values are modified in reality to account for ground losses. Shown in Figure 2.7 are experimental measurements of the reflection coefficients of vertical and horizontal polarization³, both magnitude and phase, over perfectly plane earth.

ii) Diffuse Scattering:

In the following, a presentation of the depolarization results for rough surface diffuse scattering is given without a complete mathematical derivation. Reference should be made to Beckmann [16, pp. 152-177] for details of these implied calculations.

No general and rigorous solution has yet been derived to handle all instances of rough surface scatter; nevertheless a solution for several specific cases has been solved. In particular, depolarization equations have been formulated to define instances of:

- rough surface scatter in which the radii of curvature of the irregularities are large compared to the wavelength, i.e., "locally flat", and the surface is of arbitrary conductivity, and
- rough surface laterally scattered waves.

³ In this instance the polarization is in reference to a ground reflection; thus vertical is equivalent to parallel polarization, and horizontal is equivalent to perpendicular polarization.

a) Magnitude Response

b) Phase Response

Figure 2.7 - Measured Reflection Coefficients over Plane Earth [16, pp. 220-221]

A particular case which still requires more research is that of rough surface scatter in which the radii of curvature of the irregularities are small, the area covered by the irregularities is large, and the surface is of finite conductivity. This instance may describe, at VHF frequencies, the scatter from a surface formed by tree-tops of a forest.

In the following, the scattering geometry and the coefficient definitions shall first be defined, followed by a presentation of the results.

Figure 2.8 shows the scattering geometry which is used. This geometry indicates the plane of incidence defined by the vectors $\mathbf{k}_1$ and $\mathbf{z}_0$ and the plane of scattering defined by similar vectors $\mathbf{k}_2$ and $\mathbf{z}_0$. The symbolism used is defined as

$\mathbf{e}^+, \mathbf{e}^-$ - unit vectors in the direction of vertical (+) and horizontal (-) polarization, respectively

$\mathbf{k}$ - propagation vector defining direction of wave propagation

$\theta_1, \theta_2, \theta_3$ - angle of incidence, angle of scattering within plane of incidence and lateral scattering angle, respectively

Note that subscripts 1 and 2, respectively, are used for quantities associated with the incident wave and the scattered wave; also the vectors $\mathbf{e}_1^+, \mathbf{e}_1^-$, and $\mathbf{k}_1$ (similarly $\mathbf{e}_2^+, \mathbf{e}_2^-$ and $\mathbf{k}_2$) are mutually perpendicular. The orientation is such that $\mathbf{e}_1^+ \perp \mathbf{y}_0$ and $\mathbf{e}_1^- \parallel \mathbf{y}_0$ always, but $\mathbf{e}_2^+ \perp \mathbf{y}_0$ and $\mathbf{e}_2^- \parallel \mathbf{y}_0$ only for $\theta_3 = 0$.

A complex polarization factor p can be created to uniquely define any polarization of an electromagnetic wave. It is represented as

$$p = \frac{E^+}{E^-} \quad (2.24)$$

where E^+ and E^- are complex scalars of the E-field components in the direction of vertical (+) and horizontal (-) polarization, respectively, with each having the time factor $\exp(j\omega t)$ suppressed.

The tilt angle τ , as defined in Section 2.2.1, is related to p by the relation

The Plane of Incidence (I)
The Scattering Plane (S)

Vertical Polarization (+)
Horizontal Polarization (-)

Figure 2.8 - General Geometry for Diffuse Scattering [16, p. 154]

$$\tan 2\tau = \frac{1 - |p|^2}{2|p|^2} \cos(\arg p) \quad (2.25)$$

where $\arg p = \arctan (\text{Im } p / \text{Re } p)$. Also an angle α is defined by

$$\tan \alpha = \pm |p| \quad (2.26)$$

which is the angle between $\mathbf{e}^-$ and $\mathbf{E}$ assuming E^+ and E^- are in phase, i.e., $\arg p = 0$.

A complex depolarization factor q uniquely determines the depolarization of any incident wave and it is given by

$$q = \frac{p_2}{p_1} \quad (2.27)$$

in which p_2 and p_1 are the polarization factors for the scattered and the incident waves, respectively. This can be expressed using Equation (2.24) as

$$q = \frac{\mathfrak{R}^+}{\mathfrak{R}^-} \quad (2.28)$$

where

$$\mathfrak{R}^+ = \frac{E_2^+}{E_1^+} \quad \text{and} \quad \mathfrak{R}^- = \frac{E_2^-}{E_1^-} \quad (2.29)$$

are generalized reflection coefficients for any scattering plane. These coefficients reduce to what are termed Fresnel coefficients R^+ and R^- , when the scatterer is a properly oriented smooth plane.

One must also account for the possible occurrence of a wave of one polarization incident on a scatterer and giving rise to a cross-polarized field component. This depolarization is such as may take place when a plane element scatterer of a rough plane is not perpendicular to the plane of incidence. Thus the components E_{2c}^+ and E_{2c}^- are defined such that

$$E_2^+ = E_{2||}^+ + E_{2c}^+ \quad (2.30)$$

and

$$E_2^- = E_{2||}^- + E_{2c}^- \quad (2.31)$$

where E_{2c}^+ represents a vertically polarized scattered field caused by a horizontally polarized incident field, E_{2c}^- represents a horizontally polarized scattered field caused by a vertically polarized incident field, and $E_{2||}^+$ and $E_{2||}^-$ represent scattered fields due to incident fields of the same polarization.

Substituting Equations (2.30) and (2.31) into Equation (2.29) gives

$$\mathfrak{R}^+ = \mathfrak{R}_{||}^+ + \frac{1}{p_1} \mathfrak{R}_c^+ \quad (2.32)$$

$$\mathfrak{R}^- = \mathfrak{R}_{||}^- + p_1 \mathfrak{R}_c^-$$

where

$$\mathfrak{R}_c^+ = \frac{E_{2c}^+}{E_1^-} = \frac{p_1 E_{2c}^+}{E_1^+} \quad (2.33)$$

$$\mathfrak{R}_c^- = \frac{E_{2c}^-}{E_1^+} = \frac{E_{2c}^-}{p_1 E_1^-}$$

Thus from Equations (2.28) and (2.32), q can be restated as

$$q = \frac{p_1 \mathfrak{R}_{||}^+ + \mathfrak{R}_c^+}{p_1 \mathfrak{R}_{||}^- + p_1^2 \mathfrak{R}_c^-} \quad (2.34)$$

For longitudinal scattering, i.e., scattering in the plane of incidence, the condition $\mathfrak{R}_c^+ = \mathfrak{R}_c^- = 0$ holds and thus q can be expressed as

$$q = \frac{\mathfrak{R}_{||}^+}{\mathfrak{R}_{||}^-} = \frac{E_{2||}^+}{E_{2||}^-} = \frac{\rho^+}{\rho^-} \quad (2.35)$$

where $\rho = E_2/E_{20}$ is a scattering coefficient relating the scattered E-field to the field reflected in the direction of specular reflection, $\theta_2 = \theta_1$, by a smooth, perfectly conducting plane under the same conditions as the scattering plane, and for an incident wave of horizontal polarization used as the reference field. Note from Equation (2.35) that q depends on the properties of the scatterer and the scattering geometry θ_1 , θ_2 , and θ_3 , but it is independent of the incident wave polarization.

For lateral scattering, the condition $\mathfrak{R}_c^+ = \mathfrak{R}_c^- = 0$ is no longer valid and thus a basic depolarization formula which is valid for any rough surface or any other scatterer can best be expressed by multiplying Equation (2.34) by p_1 as suggested by Equation (2.27) to give

$$p_2 = \frac{p_1 \mathfrak{R}_{||}^+ + \mathfrak{R}_c^+}{\mathfrak{R}_{||}^- + p_1 \mathfrak{R}_c^-} \quad (2.36)$$

Note the following implications of the value for q :

- $q=1$ implies no depolarization
- $q=-1$ implies that the sense of rotational polarization is reversed and, for linear polarization, that the polarization plane is turned through an angle of $-2\alpha_1$
- $q=\{0,\infty\}$ implies a "filter" scattering in which only horizontal or vertical polarization, respectively, will be scattered regardless of the incident wave polarization

a) Longitudinal Scattering Depolarization:

Analysis of depolarization by longitudinal scattering, i.e., in the plane of incidence, reveals the general conclusion stated by Beckmann [16, p.163] that

"The field scattered by the rough surface into the direction (θ_2, θ_3) is depolarized in the same way as the field reflected by a plane of the same electrical properties inclined in such a way that it will reflect the incident field into the same direction (θ_2, θ_3) ."

Therefore for irregularities having radii of curvature large relative to λ , the problem reduces to solving the depolarization caused by inclined planes, for any given inclination of the plane, and for arbitrary electrical properties.

The resulting depolarization factor for $\theta_3=0$ is given as

$$q = \frac{R^+(\zeta)}{R^-(\zeta)} \quad (2.37)$$

where

$$\zeta = \frac{\theta_1 + \theta_2}{2} \quad (2.38)$$

Equation (2.37) implies that all incident waves will be depolarized, with the exception of a purely vertical or a purely horizontal incident polarization, or unless $R^+ = R^-$. In addition, Equations (2.37) and (2.38) hold regardless of whether the surface is random or periodic.

If the radii of curvature of the irregularities are small compared with λ then local diffraction effects occur and q must be determined from particular rough surface models such as the Twersky model [16, pp. 164-167].

b) Lateral Scattering Depolarization:

Shown in Figure 2.9a is the geometry for lateral scattering. This figure includes a "local" plane of incidence, equivalently called a "local" plane of scattering, which is determined by the plane containing the vectors $\mathbf{k}_1$, $\mathbf{k}_2$ and $\mathbf{n}$, where $\mathbf{n}$ is the normal to a local surface element.

Unlike the situation for longitudinal scattering, scattering of purely horizontal or purely vertical polarization off of a plane such as defined in Figure 2.9a will result in depolarization. An example is illustrated in Figure 2.9b.

Thus the condition $\mathcal{R}_c^+ = \mathcal{R}_c^- = 0$, as applicable for longitudinal scattering, no longer holds. The generalized reflection coefficients and the depolarization for an arbitrarily inclined plane of arbitrary incident polarization can be determined using the orientation indicated in Figure 2.10. Note that the geometry is uniquely defined by three independent angles:

- γ - local grazing angle
- β - polarization angle, or the angle between E_1 and the intersection of the wavefront with the reflecting plane
- β_2 - angle between E_2^- and the intersection of the wavefront with the reflecting plane (angle not illustrated)

a) Lateral Scattering Geometry

b) Depolarization by a Tilted Plane Element

Figure 2.9 - Lateral Scattering

Figure 2.10 - General Orientation of Planes for Local Surface Element

The obtained results are

$$\mathfrak{R}^+ = \cos 2\beta + \frac{1}{p_1} \sin 2\beta \quad (2.39)$$

$$\mathfrak{R}^- = -\cos 2\beta + p_1 \sin 2\beta$$

which represent the generalized reflection coefficients. The generalized depolarization equations are given by

$$p_2 = \frac{p_1(1 - \tan \beta \tan \beta_2) + \tan \beta + \tan \beta_2}{p_1(\tan \beta + \tan \beta_2) + \tan \beta \tan \beta_2} \quad (2.40)$$

for a perfectly conducting plane and by

$$p_2 = \frac{p_1(R^- \tan \beta \tan \beta_2 + R^+) - R^- \tan \beta + R^+ \tan \beta}{R^- + R^+ \tan \beta \tan \beta_2 - p_1(R^- \tan \beta_2 - R^+ \tan \beta)} \quad (2.41)$$

for a plane of arbitrary conductivity, where β is given by

$$\sin \beta = \frac{\sin \theta_3 \sin \theta_2}{\sqrt{[1 - (\cos \theta_1 \cos \theta_2 - \sin \theta_1 \sin \theta_2 \cos \theta_3)^2]}} \quad (2.42)$$

and β_2 is given by

$$\cos \beta_2 = \cos \beta \cos \theta_3 - \sin \beta \cos \theta_1 \sin \theta_3 \quad (2.43)$$

The results of Equations (2.40) and (2.41) for smooth planes apply equally to a rough surface provided that it is "locally flat" as previously defined.

An attempt has been made in this chapter to familiarize the reader with the fundamental theory supporting the thesis research topic. Particular emphasis has been placed on MP propagation, and on explaining the polarization dependency of an obstacle in behaving as a source of wavefront distortion.

In the following chapter, the design aspects of the dual-polarization VHF direction finding antenna will be presented.

CHAPTER 3

DESIGN OF DUAL-POLARIZATION VHF DIRECTION FINDING ANTENNA

At this point the reader has been acquainted with the interfering effects of multipath on DOA measurements. More specifically, Chapter 2 has examined the possibility of a depolarization of an incident signal upon encountering a MP disturbance; particular emphasis was placed on MP reflection. These introductory topics provide the justification for developing an antenna to independently measure the received power for two orthogonal polarizations, both vertical and horizontal; this chapter will begin with an introduction of such arguments. Following this, complete details of the design of the dual-polarization VHF direction finding antenna will be presented along with measured antenna radiation patterns.

3.1 OBJECTIVES AND SPECIFICATIONS OF DIRECTION FINDING ANTENNA DESIGN

Referring back to Section 2.2.3, one may recall how a signal can be depolarized upon encountering a MP reflection, refraction, diffraction or reradiation. Furthermore, the characteristics of these MP signals may provide invaluable information for the study of direction finding. A specific, hypothetical MP scenario will now be examined, enabling a clarification of the antenna design objectives. Following this, a summary of the principal specifications of the antenna design will be presented.

i) Objectives:

Consider the MP scenario illustrated in Figure 3.1a, where the $\theta = 0^\circ$ baseline is as shown. A transmitter is located at a bearing of $\theta = 335^\circ$, relative to the baseline, and it emits a vertically polarized, CW signal. In addition, a single prominent obstacle is located at a bearing of $\theta = 25^\circ$, giving an angular offset of $\Delta = 50^\circ$ from the DP signal. The reflective nature of this obstacle at the frequency of interest is such as to cause a depolarized specular reflection of the incident signal towards the receiving antenna. Also the distance of this reflecting object from the transmitter/receiver line-of-sight is sufficiently small, i.e., the distance travelled by the MP signal is no larger than double the distance travelled by the DP signal, so as to obtain a ratio of the MP signal strength to the DP signal strength of a reasonable value, i.e., in the range of about 0.5 to 1.

At the receiver site, a measurement of the vertically polarized signal strength includes not only the DP signal but also a component of the MP signal. Consequently a DOA determination, by application of a 'null-steering' amplitude comparison DF technique, will be contaminated by the presence of the MP signal and an erroneous bearing may result. This measured bearing will depend upon the manner of addition of the two signals; specifically, the degree of contamination depends upon the relative signal strengths and phases of the two signals, in addition to their angular separation, as will be examined in Chapter 5. At present, suffice it to say that the constructive addition of these two signals will result in a negligible disturbance in the null response, thus providing an accurate indication of the correct DOA. However, if these signals combine destructively, they may cause a significant shift in the null response from the correct bearing, or may even totally distort the null response.

Similarly, a measurement of the cross-polar response will be comprised of all received cross-polar components. However, in this instance the measurement includes only the cross-polar component which is present in the MP signal; the

assumptions made here are that no depolarization of the signal occurs along the DP and that sufficient cross-polar isolation of the receiving antenna elements is achieved.

It suffices to say that the application of a 'null-steering' technique with the co-polar responses alone is insufficient in determining the correct DOA. Nevertheless, there is useful information contained in the cross-polar responses.

The above-mentioned phenomena give rise to an interesting concept — it may be possible to concurrently use the measured co-polar and cross-polar responses to establish a new technique enabling the accurate determination of a transmitters bearing. As mentioned for the simplified scenario of Figure 3.1a, a measured response of the received cross-polar components will contain only a single null. In addition, this null should be centered on the absolute bearing of the MP, θ_{MP} , there being no other cross-polar components received to interfere with the 'null-steering' response. This case is illustrated in Figure 3.1b where one can notice the null of the cross-polar response situated at an angle of $\theta = \theta_{MP} = 25^\circ$. The co-polar response, as mentioned, will depend on the relative amplitudes and phases of the two signals. One possible configuration of these responses is shown in Figure 3.1b.

It is thus hypothesized that if a measured co-polar response can be used to determine the MP/DP angular offset variable Δ , and if a measured cross-polar response can be used to determine the absolute MP bearing θ_{MP} , then an accurate and correct transmitter bearing θ_{DP} can be determined from

$$\begin{aligned}\theta_{DP} &= \theta_{MP} - \Delta & , \theta_{MP} \geq \Delta \\ &= 360^\circ + \theta_{MP} - \Delta & , \theta_{MP} < \Delta\end{aligned}\tag{3.1}$$

To obtain these unknown values it is proposed that a 'DF Correlation Technique' be used in which the measured co-polar response is compared with a database of simulated responses, these responses being functions of the variables concerned. From this comparison a 'best fit' response may then be determined; one possibility is to compute an error term which is based on a summation of the squares

a) MP Scenario

Cross-Polar Response

Co-Polar Response

b) Ideal Responses, for Use in 'DF Correlation Technique'

Figure 3.1 - Illustration of Signal Depolarization Upon Reflection

of the difference between the measured and the simulated response for all values of θ . The selection of this 'best fit' would reveal the identity of all involved variables¹, including the MP/DP angular offset variable Δ . Also the variable θ_{MP} would be revealed from the null bearing in the cross-polar response.

Note that the accuracy of the DOA estimation depends chiefly on the definition of the null of the receiving antenna radiation pattern, assuming all other non-MP system inaccuracies to be negligible. This implies that the required DF accuracy can be attained from the appropriate selection of a receiving antenna radiation pattern.

The experimental analysis to be presented in this thesis concentrates on the presence of only one MP route. This is done as a simplification, however, and does not necessarily prevent the subsequent extension of this analysis to the more general case involving several MP routes, and more complex evaluation routines.

Summarizing, the primary objective of the dual-polarization DF antenna design is to enable separate measurement of the co-polar and the cross-polar signal strengths while employing a 'null-steering' amplitude comparison DF technique. This objective is incorporated with the objective of the overall thesis which, one may recall, is to investigate a new DF technique to enable increased accuracy in DOA determinations.

ii) Specifications:

In order to achieve the above-mentioned objectives, the following antenna specifications are required:

- two antennas, for separate co-polar and cross-polar reception, each possessing a single null radiation pattern in the azimuth plane to allow for

¹ The implication here is that the simulation is 'perfect', in that it takes into account all possible factors affecting the received power response.

null-steering and to eliminate bearing ambiguity due to multiple null patterns,

- a high degree of cross-polarization isolation of the receiving elements, ≥ 20 dB, to ensure polarization purity of the received signal measurement,
- a sufficient signal-to-noise ratio (SNR) of the receiving elements, $\text{SNR}_{\text{dB}} \geq 20$ dB, to avoid excessive null-filling of the antenna pattern resulting from noise signals,
- an antenna operating frequency of 60 MHz which is a frequency of interest in landbased tactical communications,
- a lightweight construction and a compact, portable design for ease of field measurements, and
- a manually rotatable structure and bearing measurement capability.

3.2 DESIGN AND FABRICATION OF DIRECTION FINDING ANTENNA

The dual-polarization VHF direction finding antenna was designed to meet the specification requirements outlined in the previous section. A photograph of this antenna is given in Figure 3.2.

The antenna consists of a pair of vertical dipoles and a pair of horizontal loops, to receive signals of vertical and horizontal polarization, respectively. The elements are supported by phenolic arms of adjustable length, situated on top of a phenolic mast and a metal tripod. The tubular mast rests on the tripod base and it is supported by the tripod centre column, thus enabling the entire element structure to be rotated. In addition, a protractor provides the capability of recording relative bearing readings.

In the sections to follow, the major design aspects of this antenna are examined. In particular, concentration is placed on the designs of the antenna array,

Figure 3.2 - Dual-Polarization VHF Direction Finding Antenna

the antenna element, the balun network, and other critical features.

3.2.1 Antenna Array Design

As mentioned previously in Section 3.1, the desired radiation pattern of each antenna must possess a single, deep null. From antenna theory [9, pp. 118-130], any desired radiation pattern can be obtained through the proper configuration of an array of individual elements. Specifically, by pattern multiplication² one can obtain the desired total field pattern through the multiplication of the individual source pattern by the array pattern, as illustrated in Figure 3.3a. In this example the source pattern is that produced by a single nonisotropic point source centered on the origin, and the array pattern is the result of two isotropic point sources separated by a distance d and located symmetrically with respect to the origin.

After analyzing many configurations, a simple, two element 'end-fire' linear array is chosen. The field pattern is that of a cardioid, as shown in Figure 3.3b, and it is defined by the equation

$$E = \cos\left(\frac{\pi}{4} + \frac{\pi}{4}\cos\phi\right) \quad (3.2)$$

where ϕ is as indicated in the figure. Note that the space pattern is a figure-of-revolution of this pattern around the x-axis. Also, for reference, the derivation of Equation (3.2) is given in Appendix A.

To satisfy the total array pattern, as explained in Appendix A, the two elements must be i) in phase quadrature, ii) physically separated a distance of $\lambda/4$, and iii) producing signals of equal amplitude. In addition, the individual elements

² It is assumed that the reader has sufficient knowledge of the concept of pattern multiplication. This concept is a fundamental topic dealt with by most books which cover general antenna theory, and any further details may be obtained from such references as [9], [10], [18], [19], [20], and/or [21].

a) Example of Pattern Multiplication [9, p. 130]

b) Cardioid Field Pattern [9, p. 125]

Figure 3.3 - Antenna Radiation Patterns

must be isotropic sources in the plane of interest, the horizontal plane. Thus, two vertical dipoles are used for the reception of vertical polarization, and two horizontal loops are used for the reception of horizontal polarization. Figures 3.4a and b illustrate the desired polarization reception of these elements.

3.2.2 Antenna Element and Balun Design

After selection of the array design, one has to calculate the element dimensions and design a balance-to-unbalance transformer (balun). The element dimensions are chosen in an iterative process to obtain a reasonable radiation resistance thus ensuring sufficient power reception, while limiting the antenna size to avoid arm sagging due to excessive weight. The balun is required to join the balanced element terminals to the unbalanced coaxial transmission line.

i) Element Design:

The design of the element pairs, both dipoles and loops, follow the guidelines for a 'small antenna'. A 'small antenna' is defined as one in which all element dimensions are small in relation to the wavelength of operation λ ; it is used in this situation due to the physical size restrictions of the structure which have been imposed. Detailed calculations of the element dimensions are given in Appendix B. To follow, the particulars of the construction of these elements are discussed.

The dipole is constructed out of a lightweight aluminum rod with the dimensions given in Figure 3.5a. Semi-rigid coaxial feed lines with SMA connectors are used to provide shielding of the signal and to reduce unwanted cross-polar component pickup. In addition, a phenolic block separates the two monopoles from short circuiting, and two plastic screws are provided to securely hold the elements in place.

The dimensions for the loop element are shown in Figure 3.5b. Its design

a) Vertical Polarization Receiving Element - Dipole

b) Horizontal Polarization Receiving Element - Loop

Figure 3.4 - Polarization Reception of Elements

is that of a standard shielded-loop antenna, as given by Jasik [22, pp. 5-19 - 5-21] and by other authors [23], [24]. The purpose of this design is to provide geometrical symmetry for the loop, about a plane perpendicular to the element. It is constructed by completely shielding the loop conductor by an external conductor, with only a small gap in the shield geometrically opposite the element terminals. With this arrangement, the ends of the gap form the new effective terminals of the loop.

To realize this standard loop design, a 50 Ω RG174 coaxial transmission line is used for the loop with a small gap cut in the outer shield to prevent current flow. This transmission line is wrapped twice about a plexiglass disk, in accordance with the two turn design in Appendix B. In addition, BNC connectors are used to form a reliable link between the element and the balun terminals.

ii) Balun Design:

The balun must function, as mentioned, to connect the balanced element terminals to the unbalanced coaxial transmission line. In addition it must also contend with the large impedance mismatch between these components; the elements have a highly reactive input impedance while the coaxial transmission line has a characteristic impedance of 50 Ω .

To perform as such, an operational amplifier (op amp) is used in a differential input configuration. In this form, the amplifier will function to sense the voltage difference across the element input terminals and to amplify this value by a set gain, as determined by the op amp biasing circuit. Unfortunately, this gain is forced low as a result of both the internal feedback resistance of the selected op amp and the large element input impedance.

In order to increase the SNR at the signal output, which has been listed as a design specification in Section 3.1, a second amplifier is cascaded in series with the first to provide input signal amplification over noise levels. With this amplifier, a gain of 20 dB is designed for each element channel. Shown in Figure 3.6 is a

a) Dipole Dimensions

b) Loop Dimensions

Figure 3.5 - Dimensions of Receiving Elements

Figure 3.6 - Schematic of Balun Network

schematic of the balun network. Also, Appendix C gives details of the circuit calculations, the circuit diagrams and the biasing schematics that are used.

After the assembly of the four baluns, it is required that the amplification of each be measured, and that any necessary compensation be made so as to obtain an amplitude match within each element pair. Without this tuning, the signal strengths produced by both elements in each pair would differ, thus altering the antenna response designed for in Section 3.2.1. Accordingly, the balancing of these amplification units was performed in the lab prior to their use for field measurements; an HP 3335A signal generator and an HP 8568B spectrum analyzer were used for this purpose. The measured levels of gain were determined to be 31 dB and 25 dB for the dipole elements and for the loop elements, respectively.

3.2.3 Critical Features

Apart from the antenna array, the element, and the balun network designs, there are other important design aspects to consider. In the following, the particulars of these remaining items are presented.

One of the main specifications of the antenna design, as listed in Section 3.1, is to ensure a high degree of cross-polar isolation of the receiving elements. This is considered in all the aforementioned designs, where applicable. In addition, several specific precautions that are taken are mentioned below:

- both cables and dc power lines are tied securely to the antenna supporting mast and arms; this is done in an attempt to reduce cable twisting with antenna rotation, thereby reducing fluctuations in received signal level,
- the structural support of the antenna is made out of non-metallic components, where at all possible, to reduce cross-polar reflections and reradiations,
- dc power lines are enclosed in a shielded cable to reduce unwanted signal

pickup,

- all circuit grounds are connected and examined according to recommendations by Ott [25], and
- the receiving elements are designed sufficiently small in the cross-polar direction to increase cross-polar isolation.

Another design consideration is the height of the antenna above ground. There are no specific details given in the literature as to what is the optimal antenna height for a given application; however, as mentioned in Chapter 2, it is desired to have signal reception of only the space wave propagation mode but not of any existing surface waves. This requirement is a function of both the antenna height and the frequency, and thus it is used as the criterion for the height selection. Palmer [11, p. 2-3] cites an example of expected signal strengths for these propagation modes at VHF frequencies. For a vertically polarized antenna 2 m off the ground and operating at 300 MHz, the space wave can typically be expected to have a signal strength 30 dB greater than the surface wave. This ratio is even greater for horizontal polarization, and a difference of about 70 dB may be expected [11]. With these values in mind, an antenna height of 2 m was selected. Note that one does not desire the antenna to be too high, as this would reduce its compact portable design.

3.3 ANTENNA RESPONSE MEASUREMENTS

Before commencing with the analysis stage of the DF measurements, it is required that the antenna pairs be characterized so as to determine the deviation between ideal and measured antenna radiation patterns, the level of cross-polar isolation attained, and the measured null sensitivity. A complete characterization of the DF antenna includes the independent measurement of the co-polar and the cross-

polar radiation patterns, for both the dipole and the loop pairs. These patterns may then be used in the experimental response analysis, to denote the departure of the measured results from the ideal expectations.

The David Florida Laboratory (DFL) roof top range facility, located at Shirley's Bay, was used to record the antenna radiation patterns³. This facility operates by elevating the antenna-under-test (AUT) high above the roof in an attempt to minimize all multipath reflections. A sketch of the test set-up is shown in Figure 3.7.

For the test set-up, note that the AUT is elevated by a 3 m aluminum supporting post and it is accurately oriented with an Orbit model AL-4306-1 Azimuth-over-Elevation Positioner. The positioner, in turn, is located on a galvanized steel platform on top of DFL's High Bay 3 at a height of 20.73 m above ground level. The source antenna is located on top of the roof's mobile source tower at a short range of 19 m from the AUT, which is within the AUT far-field, and it is at an equivalent height. This source is a biconical antenna designed for operation in the frequency range 20 - 200 MHz, and it is appropriately oriented to transmit the desired polarization. The antenna measurement system consists of the Scientific-Atlanta model 1750 receiver and model 1520 rectangular recorder system.

Both the co-polar and the cross-polar responses for each pair of elements were measured independently by rotating the AUT through a full 360° in azimuth, and obtaining a continuous reading. The set of responses for the dipole elements are shown in Figure 3.8 while those for the loop elements are shown in Figure 3.9.

In obtaining the co-polar responses, it was initially observed that the antenna was peaking off of its center frequency of 60 MHz. To compensate for this, the required arm length adjustments to retune the elements were calculated and the

³ The DFL anechoic chamber facility could not be used for these measurements due to its low cutoff frequency of 250 MHz.

Figure 3.7 - DFL Roof Top Range Test Set-Up

Scale:

5 dB per main division

Figure 3.8 - Measured Antenna Patterns for Dipole Array

Figure 3.9 - Measured Antenna Patterns For Loop Array

necessary structural alterations were made. Additionally, it was observed that the cross-polar power levels relative to corresponding co-polar responses were unrealistically high. Since extensive precautions had been taken in the design of the antenna to eliminate and/or reduce the cross-polar pickup, i.e., ensuring proper cable shielding, etc., it was assumed instead that the large power levels were due to multipath reflections from the range⁴. Consequently, the cross-polar responses illustrated in Figures 3.8 and 3.9 were obtained from supplementary measurements in which a test field on the DREO site was used.

From an observation of the dipole and the loop antenna patterns, some general comments may be made about the response shapes. Primarily, the measured co-polar responses for both the dipole and the loop arrays demonstrate a departure from the ideal cardioid pattern. Specifically, the patterns exhibit a null-filling⁵ and a null-shifting characteristic. From Figure 3.8, the dipole array pattern indicates a shift in null bearing of 25° and a null depth, relative to peak power, of 35 dB. Similarly, the loop responses of Figure 3.9 reveal a 15° null bearing shift and a corresponding null depth of 25 dB. The level of cross-polar isolation⁶ attained is approximately 21 dB, which applies equally for both pairs of elements. Furthermore, the measured null sensitivity⁷ is approximately 3° and 6°, respectively, for the dipole array and the loop array responses.

⁴ The roof top range is 'useable' at 60 MHz, but for such a low frequency it is non-optimal and thus it is not guaranteed to be reflection-free. Additionally, the mentioned source antenna has a very broad beamwidth thereby increasing the possibility for signal reflections.

⁵ The co-polar radiation pattern null-filling may partially be due to co-channel interference resulting from the signal reception of other broadcasting stations which operate at 60 MHz. This explanation is quite probable, as repetitions of the measured responses showed differing null depths. Therefore to obtain the polar plots shown, an averaging of the recorded data was performed.

⁶ The cross-polar isolation is defined as the power level difference between the co-polar peak power and the cross-polar peak power.

⁷ Null sensitivity is defined as the null width, in degrees, between two specified equal amplitude points, one to either side of the null minimum; its value is an indication of the antenna array bearing measurement capability. Arbitrarily these points were chosen to be +1 dB above null bottom.

In conclusion, the measured antenna patterns differ from the ideal cardioid response in two important characteristics: i) a more shallow null depth, and ii) a shift in null bearing. Furthermore, these differing pattern attributes will have an affect on all ensuing experimental responses, to an extent as yet undefined. The attained levels of cross-polar isolation, approximately 21 dB for both element pairs, are sufficient to satisfy the initial antenna design specification, requiring an isolation ≥ 20 dB. Finally, the obtained values of null sensitivity give an indication of the null-steering technique precision. The values of 3° and 6° are satisfactory for investigative purposes; however, an antenna response with a more well-defined null should be incorporated into an actual DF system design.

Having introduced several aspects of the DF antenna design and illustrated the measured antenna radiation patterns, the following chapter will concentrate on presenting the experimental procedure and results.

CHAPTER 4

EXPERIMENTAL PROCEDURE AND RESULTS

This chapter focuses on the experimental aspects of the thesis work. More specifically, a set of experimental measurements is performed with two primary intentions: i) to demonstrate the diversity of received power responses caused by the interference of a single MP signal with the DP signal, and ii) to observe if a correlation may be obtained of the measured responses with the theoretically predicted results, as will be examined in Chapter 5. Obtaining such a correlation will establish a proof-of-concept for the multipath resolution technique using orthogonal polarization components, and it will be sufficient to encourage future investigations.

An introduction to the experimental set-up that is used for performing the field measurements, and an explanation as to its realization are given in Section 4.1. Also included is a statement of all assumptions made while outlining the experiment. Section 4.2 examines various topics relevant to performing the field measurements, and Section 4.3 concludes with a presentation and discussion of the measured responses.

4.1 DESIGN AND REALIZATION OF EXPERIMENTAL SET-UP

Considering the complexity involved in the study of multipath, it was decided that the best approach to the polarization diversity problem would be to create and utilize a simple, artificial MP environment for the experimental measurements rather than concentrating on naturally occurring signal disturbances. In this capacity, one can study the MP effects of a known and well-defined environment while maintaining reasonable control over the introduction of unknown system parameters.

The key concept in the design of the experimental configuration is to create a simple, artificial environment consisting of a polarization-pure DP signal, and one single dominant MP signal comprised of both co-polar and cross-polar components. The creation of this environment is based on several underlying assumptions, as follows:

- all received signals are propagating in the space wave mode, i.e., there is no reception of surface waves, due to the sufficient height elevation of the receiving antenna above the ground surface,
- the DP signal does not experience any depolarization along its propagation path,
- the MP signal propagates with equal components of vertical and horizontal polarization,
- the receiving antenna is in the far-field of both the DP and the MP transmitting antennas thus ensuring the measurement of a plane wave, and
- the propagating DP and MP signals experience no attenuation other than the calculated free-space propagation loss.

With these assumptions made, it is now necessary to define each received signal in terms of both its signal strength and its phase. For convenience, the DP signal can be normalized such that it has a magnitude of one and a phase of 0° . In turn, the MP signal can be represented as having a co-polar amplitude, a cross-polar amplitude, and a phase, relative to those of the DP signal, given by the three variables A_{co} , A_x , and ϕ , respectively. A fourth variable Δ must also be specified to indicate the angular separation in azimuth between the two propagation paths of the DP and the MP signals. This devised signal reception, as just described, is illustrated in Figure 4.1a.

In order to realize this arrangement, it is of primary importance to choose an appropriate field site and of secondary importance to select the components of equipment which make up the system. The selection of these two items is illustrated

in Figure 4.1b, and a discussion follows.

The important factors considered in the selection of a field site are its location, its accessibility, and its suitability to the desired measurements. Due to equipment transportation problems and associated set-up time constraints, it was decided to use a field on the DREO grounds at the Shirley's Bay site. The selected site is a large field composed of relatively flat¹ grass-covered terrain, enclosed on three sides by a row of distant trees (distance $> 20\lambda$) and on the fourth side by a gravel access road. The intention in using this field is to be able to set up the complete DF system in the field centre with confidence that only negligible interference effects (reflections, etc.) from surrounding obstacles will reach the receiving antenna, thus preventing undesired pattern perturbations. A few antenna towers are situated around the site but these are assumed to be far enough away (distance $> 40\lambda$) such that reradiation and reflection effects are negligible. Similarly, the row of trees is sufficiently far such that reflected power is negligible compared to the simulated MP power. Note that the Stanford Research Institute (SRI) has developed a technique, as described by Barker [26, pp. 538-544], specifically for measuring the radiation patterns of antennas at VHF frequencies to ensure their proper operation when located in an operational environment. This technique can be employed as a verification of the above assumptions, but it is not considered feasible due to the degree of involvement and equipment necessary. Rather it is mentioned here as a matter of interest.

The system equipment is comprised of three main components:

- the dual-polarization DF antenna for signal reception,
- a transmitter for signal generation, and
- a MP generator.

The receiving antenna has been discussed in Chapter 3 and the transmitter selection

¹ The term 'relatively flat' implies that the surface is 'smooth' by the definition of Chapter 2.

Co-Polar Reception

Cross-Polar Reception

a) Reception of Signal Components

Figure 4.1 - Experimental Set-Up

b) Experimental Field Realization

Figure 4.1 - Experimental Set-Up

is mentioned later. The MP generator, however, requires a little more consideration.

The concept of a MP generator involves the generation of a MP signal from a single, specular reflection. This MP signal will obey Snell's Law whereby the angle of incidence θ_i is equal to the angle of reflection: θ_r , such that maximum received power is attained.

Some experimentation was initially performed so as to determine an appropriate source of MP specular reflection, and altogether three different sources were tested. The first two sources tested are reflectors: one reflector is capable of providing reflections of varying strengths and polarization, for a given incident power; the second reflector is capable of providing only a single-strength, single-polarization reflection. Unfortunately, neither of these two sources provided the desired reflection². The third source is utilized for the set of field measurements which follow, and it consists essentially of a second transmitter simulating the propagation of a reflected signal. The design and operation of these three sources are briefly described below.

4.1.1 Grid Reflector as Multipath Source

A grid reflector, see Figure 4.2a, is designed and built to enable a specular reflection of the transmitted signal while controlling the ratio of co-polar to cross-polar reflection components.

In the design, the reflector diameter and the wire spacing have to be selected. The dimension of the reflector diameter is chosen after consideration of the diffraction effects of an obstacle of a given radius and for a given signal wavelength. Kerr [27, pp. 445-465] has performed such an analysis in which the

² The received power responses with the reflector present were essentially equivalent to those responses obtained with no reflector. Thus the strength of the MP reflection was negligible.

back-scattering of an illuminated sphere has been calculated; note that this analysis applies sufficiently well to an obstacle of circular geometry, provided that an additional multiplicative constant is used. Using the dimensions corresponding to the occurrence of peak power back-scatter, a reflector diameter of 2.4 m is chosen. This size is a good compromise between being large enough to function as a reflecting surface (not transparent) at the wavelength of interest, and being small enough to be of practical size (easy handling) for field measurements.

The reflector is constructed out of pine wood timbers (1" X 1") so as to keep the structure light, with two cross-bars for structure support. Parallel lengths of wire strung across the structure in a grid, and stapled taut at the ends, are used to provide the desired reflection. A sufficient wire spacing is needed so as to eliminate coupling between lines yet provide adequate signal reflection. Following the design given by Kraus [9, pp. 513-519] for a similar reflector at the same frequency, a wire spacing of $s = 10$ cm is used.

The wires function to reflect the transverse component of the E-field, in relation to the wire, while allowing for unaltered transmission of the normal component, as shown in Figure 4.2b. Thus as the grid of wires is rotated, various ratios of co-polar to cross-polar reflection components are made possible.

The shape of the reflector is designed for such a rotation. The octagonal edges allow for four unique co-polar to cross-polar reflection ratios while a plastic securing screw through the base support maintains the desired orientation.

As mentioned, field trials were performed with this reflector but the results proved unsuccessful. It appears that too little power from the MP route is actually being captured by the receiving antenna. Essentially the result is that the received responses are independent of the presence of the reflector. In other words, the reflector is not behaving as a strong source of MP reflection.

a) Reflector Structure

E_{vert} - total incident
E-field

E_{trans} - transverse reflected
E-field component

E_n - normal transmitted
E-field component

s - wire spacing

b) Reflector Operation

Figure 4.2 - Grid Reflector

4.1.2 Plane Sheet Reflector as Multipath Source

Having already built the grid reflector described in Section 4.1.1, it was decided to try utilizing this structure in another configuration. In this application, a sheet of reflective foil is secured to the frame in lieu of the grid of wires; the idea here is to increase the power reflected towards the receiving antenna by eliminating the transmitted component. Figures 4.3a and b indicate the reflector and its operation, respectively.

For the grid reflector, there is a loss of power upon reflection due to the normal field component transmitted through the wire grid. However in this case, neglecting reflector losses, the entire incident electric field is reflected due to the lack of polarization sensitivity of the reflecting surface. By the same token however, this reflector will only be useful in examining the DF antenna reception of a co-polarized reflected signal, or rather the co-polar signal response disturbance effects.

Unfortunately, no measurements are possible with this configuration since the wind resistance of the structure is too large. During the field trials, even a light wind was causing a billowing effect of the foil and the reflecting surface itself was twisting, thus making for very unreliable, inconsistent reflections. As a result it was decided to try a new approach.

4.1.3 Second Transmitter as Multipath Source

Having been unsuccessful in generating an actual signal reflection of comparable magnitude to the DP signal, it was decided instead to simulate this MP signal. In doing so, the signal is produced by a second transmitting antenna (the MP antenna) placed at the position of the desired reflection, and referenced in amplitude and phase to the DP transmitting antenna. This referencing is required to enable a useful analysis of the obtained responses. Figures 4.4a and b illustrate

a) Reflector Structure

Polarization
Maintained upon
Reflection

θ_i - angle of incidence

θ_r - angle of reflection

b) Reflector Operation

Figure 4.3 - Plane Sheet Reflector

the set-up and details, respectively, of this arrangement.

The MP antenna is oriented at a 45° angle, with respect to the ground surface, so as to transmit equal strengths of so-called co-polar and cross-polar reflective components. The desired orientation of this antenna is enabled by the use of the wooden structure of Section 4.1.1/4.1.2 (without the wires) as support.

As mentioned, Figure 4.4b shows the details in referencing the two antennas. A single transmitter is used as the signal source with a power splitter at its output. This transmitted signal is in turn split, with a 0° phase difference, into two paths each terminated with a whip antenna. The desired relative amplitude and phase of the MP antenna transmission is attained through the use of in-line attenuators and cable length adjustments, respectively. In addition, the relative amplitude calculation takes into account the differing propagation losses along the two paths (the DP and the MP routes).

The aforementioned field arrangement is used to realize the experimental set-up design, as described earlier. Using this set-up, several sets of field measurements are obtained and any reference to these measurement sets henceforth refers to the use of the MP antenna as described. These measurement sets are defined in the following section.

4.2 FIELD MEASUREMENTS

It can be very difficult at times to examine and thoroughly understand experimental data without a priori knowledge of the measurement data and the experimental methodology. This section on field measurements is included to give the reader an appreciation of the details of the field trials, and to aid in understanding the measured responses presented in the next section.

This section is divided into four subsections dealing with the measurement set description, the system set-up procedure, the measurement procedure, and

a) Field Set-Up

b) Details of Antenna Referencing

Figure 4.4 - Simulated Multipath Source Antenna

field/system observations and comments.

4.2.1 Measurement Set Description

As observed in Section 4.1, the multipath reflection is modelled and characterized by the provision of four variables:

A_{co} - MP co-polar relative amplitude,

A_x - MP cross-polar relative amplitude,

ϕ - MP relative phase, and

Δ - angular separation between DP and MP signal.

Henceforth the variable A will be used interchangeably to denote either A_{co} or A_x , as the situation warrants.

Preliminary simulation analyses³ have shown that changes in the variables Δ and ϕ for constant A , result in response trends comparable to those observed with changes to both A and ϕ for constant Δ . With this in mind, inferences can be made from the observation of only one set of these trends and then applied with equal success to the second set. The proposed measurement sets do precisely that; responses are observed for changing A and ϕ , while Δ is fixed.

More specifically four measurement sets are obtained, each consisting of both the co-polar and the cross-polar responses. Each of these sets is measured with a different ϕ setting, at a constant separation $\Delta = 30^\circ$. In selecting the ϕ settings, the aforementioned simulation analyses were utilized to reveal that the most severe response perturbations occur at a MP relative phase of $\phi = 180^\circ$, where the two signals destructively interfere. With this in mind, three phase settings are chosen at and around $\phi = 180^\circ$, and one phase setting is selected at $\phi = 0^\circ$ to observe constructive interference.

³ These simulation analyses will be further discussed in Chapter 5.

Within each set three signal amplitude adjustments are made (0 dB, 3 dB, and 6 dB in-line attenuation added), plus one normalizing response. The normalizing response is a measurement of the signal reception with only the DP signal transmission, the MP port having been terminated. A summary of these measurements is given in Table 4.1 for reference.

The variable A in Table 4.1 is obtained by normalizing the transmitted MP signal to the transmitted DP signal, calculating the difference in propagation losses between the DP and the MP signal paths, adding this difference to the additional in-line MP attenuation A' , and dividing this amount equally between the co-polar and cross-polar signal strengths (since the antenna is at a 45° angle). A sample calculation is given in Appendix D.

Note that as a result of the aforementioned MP antenna orientation, the value of A is equivalent for both the co-polar and the cross-polar reflected signals. In this manner, Table 4.1 applies equally to either orthogonal polarization. In addition the propagation loss difference, between the DP and the MP signal paths, changes minimally with a change in ϕ . Thus the value of A is equivalent for all sets at a given A' .

4.2.2 Direction Finder System Set-Up

The set-up procedure of the DF system consists of the assembly of three main components:

- the two transmitting antennas (one for DP signal and one for MP signal),
- the receiving dual-polarization DF antenna, and
- the receiver unit.

In the following, these are described in more detail.

Table 4.1 - Summary of Field Measurement Sets

Set # *	Angle Between DP and MP, Δ (deg.)	MP Relative Phase, ϕ (deg.)	In-line MP Atten., A' (dB)	MP Relative Amplitude. A
1A	30	180	0	0.79
1B	"	"	3	0.56
1C	"	"	6	0.40
1D**	"	"	N/A	N/A
2A	30	0	0	0.79
2B	"	"	3	0.56
2C	"	"	6	0.40
2D**	"	"	N/A	N/A
3A	30	160	0	0.79
3B	"	"	3	0.56
3C	"	"	6	0.40
3D**	"	"	N/A	N/A
4A	30	200	0	0.79
4B	"	"	3	0.56
4C	"	"	6	0.40
4D**	"	"	N/A	N/A

* corresponds to both co-polar and cross-polar responses

** corresponds to the 'normalizing' response in which case the power splitter is terminated with a 50 Ω load in the MP signal path

i) Transmitting Antennas:

For transmission purposes a RACAL Jaguar manpack radio, model BCC 66HB, is used. This functions as the signal source for the two transmitting antennas, and it is primarily set to a low transmission power of 10 mW, unless otherwise specified. The transmission is a CW signal. Also, to enable the tagging of the transmitted signal a tape recorder with a prerecording is played into the transmitting handset.

The nature of the field measurements require that the two transmitting antennas be referenced to one another, in both signal strength and phase. To allow this a 0° phase power splitter is connected to the output signal of the Jaguar, and both of the split signals are in turn connected, by a predetermined length of cable, to the appropriate antenna. Any required phase difference between the two transmissions is achieved by using various length cables in the MP source antenna path. Similarly, any required signal amplitude difference is attained by the use of in-line attenuators in the same path.

The antennas themselves are 2 m long, vertical whip elements with attached 50 Ω matching units, model BCC 588. As a consequence of their long length, each antenna requires some manner of support. The DP antenna is set to transmit a vertically polarized signal, and thus it is erected vertically. A portion of particle board underneath the antenna is used to steady the antenna and to ensure a polarization-pure transmission.

The MP antenna is set to transmit a signal of equal co-polar and cross-polar components. As such, the wooden reflector frame of Section 4.1.1 is used for support and to attain a 45° angle slope of the antenna with respect to the ground surface.

ii) Receiving DF Antenna:

One may recall, from the description of the dual-polarization receiving antenna described in Chapter 3, how the antenna mast is supported by a metal tripod

with adjustable legs. These leg adjustments are used to level the entire antenna, thus ensuring the correct vertical orientation of the structure. In addition to the mast orientation, each of the elements in use must also be correctly oriented for proper polarization reception. Note that only one pair of elements is in use at a time as a precaution against coupling interference.

Since the elements are removable and the element arm lengths are adjustable, a check must be made to ensure that the correct element separation for proper pattern formation is attained. Recall from Chapter 3 that manual response tuning necessitated a change in element separation.

Relative bearing readings are made possible by the use of a precision protractor mounted on the tripod. This protractor must be correctly positioned such that the two-element array, the DP transmitting antenna, the bearing indicator, and the 0° protractor bearing are all aligned. At this point the protractor is then securely attached to the tripod, while the bearing indicator rotates with both the mast and the array of elements.

Finally the element matching units must be powered up, by use of the $\pm 12V$ battery located under the tripod. This power line and the transmission cable must both be securely attached to the antenna support.

iii) Receiver Unit:

The receiver is a Singer Electromagnetic Interference (EMI) field intensity meter, model NM-37/57. It is positioned a sufficient distance from the receiving antenna, approximately 5 m, in order to reduce interference, and it is connected by an RG-58C cable of constant length.

The correct signal reception is verified by utilizing a set of headphones, and signal lock is attained with the receiver unit automatic frequency control (AFC) and a signal BW of 5 kHz.

4.2.3 Measurement Procedure

The responses may be measured after the system set-up of Section 4.2.2 is complete and all units have been powered up. This measurement procedure is very repetitive in nature and it consists of the following basic steps:

- a) aligning bearing indicator to desired azimuth angle, and recording received power level,
- b) repeating step a) for desired step interval until a full 360° revolution has been completed,
- c) altering the variables A and/or ϕ , as required, by adjusting attenuation and/or cable lengths in the MP source path,
- d) repeating steps a) through to c) until all measurement sets have been completed for one pair of elements,
- e) replacing the pair of measuring elements with the second set, and then repeating steps a) through to d) until complete.

The above steps have been over-simplified; there are in fact several complicating factors to contend with such as adverse weather conditions, equipment malfunctions, etc. Also, the measurements themselves are very time consuming.

4.2.4 Field/System Observations and Comments

During the course of the field measurements, several observations were noted. The following comments deal with the importance of these observations:

- a) Preliminary field trials⁴ had revealed a potential difficulty in taking measurements in tall (≥ 25 cm.) grass. Primarily, the received power levels

⁴ Various different field trials were initially performed to acquire an understanding of the operation and of the characteristic problems of the DF receiving antenna, while in an operational environment.

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of/de

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MICROCOPY RESOLUTION TEST CHART
NATIONAL BUREAU OF STANDARDS
STANDARD REFERENCE MATERIAL 1010a
(ANSI and ISO TEST CHART No. 2)

seemed to fluctuate significantly for a given bearing, most probably due to variations in ground reflections. To diminish these problems and to obtain approximately constant results, the completed measurement sets were performed in a field of cut grass.

- b) Preliminary field trials, again, showed that loose cables and/or power supply lines, windy conditions, and extraneous co-channel interference, all altered the received power levels. In fact, fluctuations and twisting of the mast cable position were observed to alter the received power readings in a range of ± 4 dB. As a preventative measure, the cables were securely attached to the antenna mast, arms and tripod base, leaving only a small loop of cable loose to allow for rotation. This resulted in a dramatic reduction of the signal fluctuations giving a new range in power readings of ± 0.5 dB. In addition, responses taken in windy conditions were repeated for verification of the results, and the reception of unwanted co-channel interference was partially alleviated by utilizing as narrow a receiver BW as possible. Nevertheless, it should be mentioned that some extraneous signals were received, as is evident from the occurrence of spurious dips in the measured responses.
- c) As mentioned in Section 4.2.3, the measurement procedure is very time consuming due to the manual method of recording data. The introduction of an automatic data acquisition system would enable more thorough experimentation by cutting down on the required field time.

4.3 MEASURED RESPONSES

As mentioned in Section 4.2.1, four different sets of measurements were performed and these measurements correspond to the various settings indicated in Table 4.1. The measured co-polar and cross-polar response curves for each of these

sets, 1 through 4, are shown in Figures 4.5-4.8, respectively. Recall from Table 4.1 that each set includes four different trials; each of these trials is represented by a different curve on the plots as indicated by the given legend. In addition, note that the absolute power levels vary for each graph and thus the response sets are plotted over different power ranges. Nevertheless, this power range is kept at a constant value of 30 dB so as to keep all graphs on the same scale for comparison purposes.

From an overall observation of the measured responses, some general comments can be made. Specifically, the significance of the shape of response D for all of the measurement sets may be noted. Recall that this measurement is of the normalizing response in which case the MP antenna has been terminated. Hence the co-polar response to the DP signal with no MP interference is effectively an indication of the dipole antenna co-polar radiation pattern, while in the defined operational environment. Similarly the corresponding cross-polar response is an indication of the level of cross-polar isolation of the loop elements, and/or an indication of the presence of additional signal components which are extraneous to the assumed experimental environment. With this in mind, note that the general shape of all responses is that of a single null; however, this null is quite broad and shallow which may mean that high noise levels are causing null-filling. In addition, the difference in power levels between the cross-polar and the co-polar responses for this specific measurement is quite small, perhaps indicating the presence of signal interference in the former responses.

It is advantageous to compare the responses of Figures 4.7 and 4.8, in which the MP relative phase is 160° and 200° , respectively. These two sets of experimental responses should ideally be identical, since there exists an antenna response symmetry about a phase of 180° . In general, the obtained results are quite similar with regards to response shape and absolute power level. However, the normalizing co-polar response of Set 3D exhibits additional spurious signal dips over the corresponding response of Set 4D. These dips may be a result of co-channel

a) Vertical Polarization Response (Co-Polar)

b) Horizontal Polarization Response (Cross-Polar)

Figure 4.5 - Experimental Responses for Measurement Set 1

a) Vertical Polarization Response (Co-Polar)

b) Horizontal Polarization Response (Cross-Polar)

Figure 4.6 - Experimental Responses for Measurement Set 2

a) Vertical Polarization Response (Co-Polar)

b) Horizontal Polarization Response (Cross-Polar)

Figure 4.7 - Experimental Responses for Measurement Set 3

a) Vertical Polarization Response (Co-Polar)

b) Horizontal Polarization Response (Cross-Polar)

Figure 4.8 - Experimental Responses for Measurement Set 4

interference, in which the receiving antenna is picking up extraneous signals operating at the same frequency.

As an additional observation, one may note the relation of power levels between the co-polar and the cross-polar responses for the different measurement sets. All responses have corresponding power levels which appear to be dependent upon both the variables A and ϕ . Furthermore, the minimum and maximum divergence of power levels, between co-polar and cross-polar responses belonging to one measurement set, occurs for the cases of constructive and destructive signal interference.

In conclusion, a demonstration of the diversity of received responses corresponding to the wavefront distortion resulting from multipath interference has been given. Even in such a simplified study, it is apparent that multipath propagation has a significant affect on the signal wavefront and thus it can not be ignored nor considered lightly.

One may recall from Chapter 3 the hypothesis of expected co-polar and cross-polar responses, and these may be related to the experimental data. It was suggested that the form of the co-polar response would vary, depending upon the settings of the MP variables A , ϕ and Δ , but that the cross-polar response would be that of a single null centered on the MP bearing. From an observation of the experimental cross-polar responses, it appears that there is no specific null to indicate the value for θ_{MP} , contrary to the initial hypothesis. This phenomenon may be an indication of having over-simplified the initial assumptions of signal addition, i.e., the assumptions of no depolarization of the DP signal, no extraneous signal interference, etc. Thus incorporated into the experimental data may be the measurement of multiple propagating cross-polar MP signals, as opposed to the assumption of a single MP signal. Further comments will be made in the following chapter concerning this issue.

From an observation of the variety of experimental co-polar responses

obtained, it is apparent that the precise values of signal strength and phase for the MP signal contribute significantly to the formation of the measured responses. Furthermore, the indication of the co-polar DP null has been altered by the presence of the MP signal, as expected, thereby rendering ineffective the technique of DOA determination by null-steering alone, and supporting instead an investigation of the correlation technique for direction finding applications. Nevertheless, it is not evident whether the experimental co-polar response also suffer from the introduction of additional co-polar MP signals. In Chapter 5, these measured responses will be correlated with the theoretical predictions to see if the variable Δ may be extracted.

CHAPTER 5

ANALYSIS OF DATA FOR DIRECTION FINDING

In this chapter, the measured experimental responses of Chapter 4 are analyzed to investigate the possibility of achieving an accurate DF technique through polarization diversity of the received signals.

As discussed in Chapter 3, the analysis technique of concern involves i) correlating the co-polar responses to theoretical data to determine the MP/DP signal bearing separation, while ii) using the location of the null of each cross-polar response to indicate the angle of arrival of the MP signal. With these two pieces of information, it was proposed that the direct path signal bearing could be accurately determined. Accordingly, a correlation is performed of the experimental co-polar data with a theoretical database, and the predicted values for the MP/DP signal bearing separation angle are presented, for four representative data responses. In addition, the cross-polar data is examined in an attempt to extract the value for the MP signal bearing.

Unfortunately, the cross-polar experimental responses appear to exhibit indications of more than one MP signal, thereby preventing the justification of the DF Correlation Technique. Nevertheless, it is possible to illustrate the theoretical effects of having introduced additional multipath signals into the MP model, as an aid to further understanding the measured data. Therefore, the effects of multiple MP signals on the DF correlator are next examined.

Section 5.1 introduces a simulation model of the MP experiment, from which the theoretical database for the correlation process is produced. In addition, graphical illustrations of the simulation results are presented, to provide insight to the theoretical signal behaviour. In Section 5.2, the results of the correlation

technique are presented along with the effects of multiple MP signals. The analysis is concluded with a discussion, in Section 5.3, of the main findings.

5.1 SIMULATION MODEL OF MULTIPATH EXPERIMENT

This section functions as a pre-analysis preparation, by introducing various topics of concern for the data evaluation. Primarily, two routines are created to simulate the independent reception, at the dual-polarization receiving antenna, of the vertical and the horizontal components of one DP and one MP signal. These signals are of defined strengths, phases, and bearing separation. As mentioned, the basic intent for the development of these routines is to create a database of simulated responses for the purpose of correlation with experimental data. In addition to the programs, a presentation of the simulation results is provided, for insight to the theoretical signal behaviour.

The development of the simulation routines is outlined in Section 5.1.1 along with a description of the MP model used. In Section 5.1.2, the simulation results are graphically presented and examined, corresponding to the use in these routines of both the ideal cardioid antenna pattern and the defined MP model variables.

5.1.1 Development of Software Programs

The simulation routines are developed by modelling the signal propagation for the experimental set-up of Section 4.1, as illustrated in Figure 5.1a. There are two basic routines which perform the function of adding all similarly polarized signals incident on the receiving antenna. One routine is devoted to the summation of all vertically polarized (VP) signals, and the other routine is for all horizontally polarized (HP) signals.

In the MP model there are two signals of primary concern, one representing

the DP and one representing the MP signal, as illustrated in Figure 5.1a. It is assumed that all other signals incident at the receiver are of negligible signal strength. In addition, the DP signal is assumed to maintain its initial polarization during propagation, which is vertical, thus introducing no HP components along its propagation path. The two signals are characterized by their relative strengths, phases and bearing separation. More specifically, the DP signal consists solely of a pure VP component, with a normalized amplitude and phase expressed as $1e^0$. The solitary MP signal consists of equal VP and HP components, each with a specified amplitude and phase relative to the DP signal given by $A_{VP}e^{j\phi}$ and $A_{HP}e^{j\phi}$, respectively, where $A_{VP} = A_{HP} = A$. Also note that the MP signal is offset from the DP signal by a bearing separation of Δ° , in a clockwise direction.

The simulation routines calculate responses using the bearing convention illustrated in Figure 5.1b. As shown, the antenna rotates in a clockwise direction and steers the array pattern null in a similar fashion. The azimuth bearing angle θ is a measure of the counterclockwise angle separation of the DP nominal bearing from the peak of the antenna radiation pattern¹. Thus an initial bearing angle of $\theta=0^\circ$ aligns these two reference planes. This MP model description is the foundation for the following simulation routine development.

In general, the total received E-field E_{tot} for a receiving antenna of a specified polarization can be expressed as

$$E_{tot} = E_{DP} + E_{MP} \dots \{ + E_{EXT} + E_{XDP} + E_{XMP} + E_{XEXT} \} \quad (5.1)$$

where

¹ This definition is clear for the use of an ideal cardioid response as the antenna radiation pattern, but it is somewhat vague for the application of a radiation pattern involving a shifted null. In this latter instance, the peak of the antenna radiation pattern is taken as the co-linear direction of the two array elements in which they are to constructively add, regardless of whether this direction exhibits maximum gain.

a) Addition of Signal Components

b) Bearing Convention for Antenna Rotation

Figure 5.1 - Simulation Model

E_{DP} = co-polar² E-field of DP signal transmission,
 E_{MP} = co-polar E-field of MP signal transmission,
 E_{EXT} = co-polar E-field of signals extraneous to MP model,
 E_{XDP} = cross-polar² E-field of DP signal transmission,
 E_{XMP} = cross-polar E-field of MP signal transmission, and
 E_{XEXT} = cross-polar E-field of signals extraneous to MP model.

The bracketed terms in this expression are included to account for possible sources of error in the MP model description. The last three terms indicate cross-polar E-field components, relative to the receiving antenna, and thus they are only present in significant quantities when the receiving antenna elements exhibit poor cross-polar isolation. Also the terms E_{EXT} and E_{XEXT} represent signals extraneous to the MP model, and as such they include possible ground reflections, reflections off of other obstacles, co-channel interference, and any other received signal not accounted for in the MP model.

The equation for signal strength E may be expressed in a more useful form as a power response P using the formula

$$P \text{ (dB)} = 10 \log \left\{ [E_r]^2 + [E_i]^2 \right\} \quad (5.2)$$

where E_r and E_i are the real and imaginary E-field terms, respectively.

As mentioned, two separate routines are developed for the simulation of the antenna reception of orthogonally polarized signals. One routine, entitled CMPFILES, is concerned with the summation of all incident VP signals. The second routine is named XMPFILES, and it operates on all incident HP signals. These two routines will now be further discussed.

² The terms co-polar and cross-polar, as used here, are in reference to the polarization of the receiving antenna elements.

i) VP Signal Summation Routine:

In calculating the total received VP E-field, E_{VERT} , Equation (5.1) can be reduced by assuming that the receiving antenna elements possess good cross-polar isolation, that no significant extraneous signals are received, and that the DP signal maintains its initial polarization. With these assumptions E_{VERT} can be expressed as

$$\begin{aligned} E_{\text{VERT}} &= E_{\text{DP}} + E_{\text{MP}} \\ &= e^{j0} U(\theta) + A e^{j\phi} U(\theta - \Delta) \end{aligned} \quad (5.3)$$

where U represents the receiving antenna co-polar array pattern as a function of bearing. Note that this expression is simply the sum of the strengths of all incident signals, each multiplied by the radiation pattern strength at the appropriate bearing.

The term $e^{j\phi}$ can be written as

$$e^{j\phi} = \cos\phi + j\sin\phi \quad (5.4)$$

so that Equation (5.3) becomes

$$E_{\text{VERT}} = U(\theta) + A \cos\phi U(\theta - \Delta) + j A \sin\phi U(\theta - \Delta) \quad (5.5)$$

This final equation is the general expression for VP E-field reception, for use with any arbitrary radiation pattern.

In the ideal case, the antenna pattern corresponds to that of a cardioid response. Recall from Chapter 3 that the basic cardioid function³ is expressed as $U(\theta) = \cos[\pi/4 + \pi/4\cos(\theta + \pi)]$. Substituting this into Equation (5.5) gives

³ In this instance the bearing angle θ has been replaced with the angle $\theta + \pi$; this serves the purpose of rotating the array reference plane so as to begin rotation on the array peak rather than on the array null.

$$E_{\text{VERT}} = \cos\left[\frac{\pi}{4} + \frac{\pi}{4} \cos(\theta + \pi)\right] + A \cos \phi \cos\left[\frac{\pi}{4} + \frac{\pi}{4} \cos(\theta + \pi - \Delta)\right] + jA \sin \phi \cos\left[\frac{\pi}{4} + \frac{\pi}{4} \cos(\theta + \pi - \Delta)\right] \quad (5.6)$$

which is of specific use in calculating ideal responses.

Equation (5.5) may be expressed in terms of power, using Equation (5.2), to obtain

$$P_{\text{VERT}}(\text{dB}) = 10 \log \left\{ [U(\theta) + A \cos \phi U(\theta - \Delta)]^2 + [A \sin \phi U(\theta - \Delta)]^2 \right\} \quad (5.7)$$

which is the general VP power reception expression used in the CMPFILES routine to calculate the received responses. This simulation routine is listed in Appendix E. It includes the values of A , ϕ , and Δ which are entered as constants. In addition, the user can specify and alter the input data file of the co-polar antenna response. This is important since it enables the possible evaluation of an infinite series of responses. In particular, the cardioid response can be used as input to perform a theoretical evaluation of the ideal expected responses, and to observe trends in these responses. Also, the measured antenna response data file may be used as input for a comparative evaluation of the experimental field measurement data.

ii) HP Signal Summation Routine:

Similar to the previous routine, the assumptions that no signals extraneous to the MP model are being introduced, and that no depolarization of the DP signal has occurred, will still be maintained. However, to demonstrate how a deficiency in the cross-polar isolation of the receiving elements is accounted for, the terms E_{XDP} and E_{XMP} will be included in the simulation.

Therefore from Equation (5.1), the total received HP E-field, E_{HORIZ} is given by

$$\begin{aligned}
E_{\text{HORZ}} &= E_{\text{MP}} + E_{\text{XDP}} + E_{\text{XMP}} \\
&= A e^{j\phi} U(\theta - \Delta) + e^{j0} U'(\theta) + A e^{j\phi} U'(\theta - \Delta)
\end{aligned} \tag{5.8}$$

where U and U' represent the receiving antenna co-polar and cross-polar array patterns, respectively, as a function of bearing, and A and ϕ are the relative amplitude and phase, respectively, of the MP signal to the DP signal.

Expanding this expression with the use of Equation (5.4) gives

$$\begin{aligned}
E_{\text{HORZ}} &= A \cos \phi U(\theta - \Delta) + jA \sin \phi U(\theta - \Delta) + U'(\theta) \\
&\quad + A \cos \phi U'(\theta - \Delta) + jA \sin \phi U'(\theta - \Delta)
\end{aligned} \tag{5.9}$$

which in terms of power from Equation (5.2) is

$$\begin{aligned}
P_{\text{HORZ}} (dB) &= 10 \log \left\{ \left[A \cos \phi (U(\theta - \Delta) + U'(\theta - \Delta)) + U'(\theta) \right]^2 \right. \\
&\quad \left. + \left[A \sin \phi (U(\theta - \Delta) + U'(\theta - \Delta)) \right]^2 \right\}
\end{aligned} \tag{5.10}$$

This is the general expression for HP power reception, and it is used in the XMPFILES routine to calculate such responses. This routine is listed in Appendix E along with CMPFILES and it allows the values of A , ϕ , and Δ to be entered as constants. Similarly, the input data files of the co-polar and the cross-polar antenna responses may be specified.

The simulation routines in Appendix E have been developed to simulate the addition of specific signal components. Nevertheless, for additional evaluation purposes they may be expanded to be of a more general form by using all of the terms given in Equation (5.1). Also shown in Appendix E is CARDIROID which is a short program to create the data file of an ideal cardioid response; this data file is then used as input in both CMPFILES and XMPFILES. Additional data files are manually created from the measured antenna responses of Section 3.3.

5.1.2 Simulation Results

It is the intent of this section to demonstrate the theoretical response behaviour corresponding to the summation of various received signals, for application to the analysis of the measured responses. More specifically, a study of the ideal addition of two signals will be presented. This study will concentrate on evaluating the simulation model of Figure 5.1, and it will make use of the cardioid radiation pattern, and the CMPFILES and XMPFILES programs. In addition, the settings of the MP model variables A , ϕ , and Δ will approximate those of the field measurement sets in Table 4.1.

To begin with, the effects of a variation in MP relative amplitude A on the measured VP and HP responses will be analyzed. This will be followed by an observation of the effects of a variation in signal bearing separation Δ on the measured VP responses. In all of these studies, consideration will be given to the corresponding effects of phase ϕ on such data. In addition, a multiple number of response plots are included for visual aid.

i) Variation in MP Relative Amplitude, A :

A series of three dimensional plots are shown in Figures 5.2a-d to illustrate the changes in received VP power responses with a progressive increment in the co-polar MP signal relative strength. Recall that the three MP model variables are MP relative amplitude A , MP relative phase ϕ , and signal bearing separation Δ . In this instance, the DP signal is normalized to a unity strength and the variable A assumes the values of 0.1, 0.3, 0.6 and 0.9, progressively, for each separate figure. In addition, all plots correspond to a constant bearing separation of $\Delta = 30^\circ$ and the phase term ϕ is variable.

The three axes of these plots are relative power, azimuth bearing, and MP relative phase. As such, a lateral cut of any of these plots will provide the ideal

power response for a unique setting of any of the three MP variables indicated. Recall the settings of the experimental MP variables from Table 4.1, and note that the corresponding responses can be interpolated for comparison from the figures shown.

Several observations can be made by studying these plots, and some important features will now be discussed.

From a general observation of the figures, it is apparent that the greatest degree of disturbance to the original cardioid response occurs at a MP relative phase of 180° . At this phase the two signals destructively interfere with each other. Furthermore, the least response disturbance occurs at $\phi=0^\circ$ where the two signals constructively add upon reception. In order to observe the activity at these two phases more clearly, and in more detail, a linear response plot is recorded of each. These are shown in Figures 5.3a and b, with each plot including the detail of the four different MP amplitudes mentioned.

From the curves in Figure 5.3a, corresponding to a phase of 0° , note how the general shape of the responses seem to have been unaltered in comparison to the original cardioid response. In particular, all four curves exhibit only one null. For $A=0.1$, this null appears to be quite deep and it is approximately centered at $\theta=180^\circ$, i.e., at the DP signal bearing. However there is a tendency, with increasing MP amplitude, for the null to get filled in and to shift in bearing. In fact, at $A=0.9$ the null has shifted approximately 15° from its proper bearing to a location approximately mid-way between the DP and the MP signal bearing, and the null depth⁴ has decreased by about 8 dB compared with the depth for $A=0.1$. Also note that even for small values of A , e.g., $A=0.1$, this null-shifting characteristic is already apparent.

In Figure 5.3b the responses for a phase of 180° are shown. Evident in all

⁴ The null depth is defined as the absolute power difference between the null power and the peak power.

a) MP Relative Amplitude, $A = 0.1$

b) MP Relative Amplitude, $A = 0.3$

Figure 5.2 - Ideal Responses of VP Signals, Indicating a Variation of A

c) MP Relative Amplitude, $A = 0.6$

d) MP Relative Amplitude, $A = 0.9$

Figure 5.2 - Ideal Responses of VP Signals, Indicating a Variation of A

a) Constructive Interference, $\phi = 0^\circ$

b) Destructive Interference, $\phi = 180^\circ$

Figure 5.3 - Ideal Responses of VP Signals, Indicating a Variation of A

four curves is the distinct alteration of the original response into two deep nulls. Note that as A is increased, the bearing separation between the two nulls also increases. Furthermore, the location of these nulls is such that a primary null is centered approximately on the true DP signal bearing, and a secondary null is off to one side. Note that the primary null is so-named since it is present at all phases, whereas the secondary null exists only for phases at and around 180° . Similar to the responses in Figure 5.3a, the primary null starts to drift in bearing away from $\theta = 180^\circ$ as A is increased. In addition, the relative null depths of the two nulls vary with A yet they do not seem to follow any observable trend.

In a similar analysis, the received HP power responses may also be presented. These responses are shown in Figure 5.4 and they correspond once again to the four mentioned MP amplitudes. They are obtained with the aid of the XMPFILES program.

In Figure 5.4, all four curves depict the cardioid response shape. This fact is intuitive since there is only one incident signal. In addition, note that the null is consistently centered on a bearing of 210° which corresponds to θ_{MP} ; thus the cross-polar response is useful in determining this bearing. It is cautioned, however, that the bearing determination is only as accurate as the nullwidth, and for $A = 0.9$ the nullwidth is still a few degrees at a corresponding null depth of 70 dB. This implies that a receiver with a good SNR is required, i.e., in this instance $SNR_{dB} \geq 70$ dB, such that the signal bearing may accurately be measured even at low input power levels.

ii) Variation in Signal Bearing Separation, Δ :

The plots indicating the effect of a variation in bearing separation Δ on received VP responses have been calculated for Δ values of 5° , 10° , 20° and 30° , corresponding to a MP signal relative amplitude of 0.7, and they are illustrated in Figure 5.5. These responses portray trends very similar to those of Figure 5.3, and

Figure 5.4 - Ideal Responses of HP Signals, Indicating a Variation of A

a) Constructive Interference, $\phi = 0^\circ$

b) Destructive Interference, $\phi = 180^\circ$

Figure 5.5 - Ideal Responses of VP Signals, Indicating a Variation of Δ

to follow the principal traits will be briefly discussed.

Once again a distinction can be made between the responses for phases of 0° and 180° . For a phase of 0° , as in Figure 5.5a, all responses exhibit the general cardioid shape which is one single null. In addition, they experience a null-filling and a null-shifting, with an increase in Δ . For the responses with a phase of 180° , see Figure 5.5b, the general shape illustrates two nulls. For a small bearing separation, the primary null is centered on the true DP bearing and the secondary null is off to one side. As Δ increases however, the secondary null rapidly moves away while the primary null slowly drifts off centre. As an example, at $\Delta = 30^\circ$ the primary null and the secondary null are located at approximate bearings of 195° and 100° , respectively, as compared to the DP bearing of 180° . Furthermore, a primary null-filling is observed with an increase in Δ , similar to that occurring for the 0° phase responses.

The above mentioned trends for the specific MP model indicated should be of use in evaluating the measured responses of Chapter 4. In the following section, such an analysis is made.

5.2 EXPERIMENTAL DATA APPLIED TO DF CORRELATOR

In the previous section, a simulation model has been developed for the chief purpose of creating a database of theoretical responses for correlation purposes. Applying this theoretical database to the DF correlation technique, the experimental responses are then analyzed and an evaluation of the correlation process is made. After drawing the conclusion that more than one MP signal was present during the field tests, even though these tests took place in a controlled experimental environment, it was decided to evaluate the effects of multiple MP signals on the described MP model. This secondary analysis is done in an attempt to better understand the experimental data.

Section 5.2.1 presents the results of the correlation technique for the

experimental data. Included in this presentation are comments on its effectiveness. In Section 5.2.2, the secondary analysis concerning the effects of multiple MP signals on received responses, is presented.

5.2.1 Results of Correlation Technique

Recall from Section 3.1 that the correlation technique for determining a transmitters bearing, involves the twofold process of i) correlating the measured co-polar responses to theoretical data to determine the MP/DP signal bearing separation angle, while ii) using the location of the null of the experimental cross-polar response to indicate the angle of arrival of the MP signal. With this procedure in mind, the experimental data of Chapter 4 is analyzed.

Refer back to Figures 4.5b-4.8b, and note the shape of the experimental cross-polar responses. Instead of the data indicating a well-defined single null, as expected, the responses appear distorted. More specifically, the data indicates two prominent nulls and the bearing location of these nulls appears to be dependent upon the MP signal phase. Furthermore, upon comparison of these responses with the simulation results of Section 5.1.2, it is feasible to suggest that the observed distortion is due to the introduction of one or more additional signal components. For example, compare response 1A of Figure 4.5b with the responses of Figure 5.3b, and observe the similarities. The latter set corresponds to the addition of two out-of-phase signals, separated by an angle of 30° , and thus the shape of the measured responses may suggest the addition of similar signals.

Hence, it is concluded that the experimental measurement of the cross-polar responses has been contaminated by the presence of multiple cross-polar MP signals rather than just the anticipated single source. Consequently, it is futile to use the cross-polar null-indicated bearing of these measurements as an estimate of the MP signal bearing. Nevertheless, the presence of additional MP signals in the cross-polar

responses does not necessarily imply their presence in the co-polar responses, and thus the co-polar correlation process is next presented.

The co-polar response correlation was performed for the experimental data sets 1A, 2A, 3A and 4A, against a database of simulated responses corresponding to the following variable range settings: $\Delta = 1$ to 360° , $\phi = 1$ to 180° , and $A = \{0.7, 0.8, \text{ and } 0.9\}$. Note that the phase range for the correlation is only from 1 to 180° , due to the fact that the ideal responses are symmetrical about this latter phase. Also, the range for the variable A is selected from 0.7 to 0.9 , so as to consider only the amplitudes of the MP signals which have the greatest distortion effect on the obtained responses.

The resulting predictions for the value of Δ , as obtained from these correlations, are tabulated in Table 5.1. As is given in this table, the sum of squared errors term (SSE) is used as an indicator as to which simulation response is the 'best-fit'. Specifically, the correlation which yielded the smallest value for this error term, was selected.

Table 5.1 - Co-Polar Correlation Predictions of Δ

Data Set #	SSE	Predicted Δ	Actual Δ
1A	1.238	11°	30°
2A	153.9	37°	30°
3A	1.215	13°	30°
4A	1.138	13°	30°

The predicted values for Δ vary between 11° and 37° , as illustrated in Table 5.1. The true separation angle is 30° , and thus the given results exhibit a

definite offset. However, this offset is suspected to be due to the presence of additional signals which are further distorting the measured responses, and it is not to be interpreted as an evaluation of the technique. Note from the table that all four 'best-fit' correlations have corresponding MP variable values of approximately $A = 0.9$ and $\phi = 160^\circ$ (the correct MP phase), except for data set 2A in which the phase term is 0° .

From the obtained results it was concluded that, similar to the cross-polar responses, the introduction of multiple MP signals rather than just the expected one, resulted in a further distortion of the co-polar responses. Thus, an accurate determination of Δ can not be obtained from the existing data. Nevertheless, the correlation process has examined the usefulness of this DF technique and it has demonstrated the possible extraction of Δ values.

It is apparent that although the DF correlation technique is feasible, it is also very susceptible to the effects of multiple scatterings. If such multiple scatterings exist, then the correlator may not be able to distinguish between the measured response and a similar simulation response involving only two signals. This would result in a false indication of Δ . Indeed, this appears to have occurred with the experimental measurements. As has been demonstrated, the performed correlations yielded relatively small SSE terms, i.e., $SSE \approx 1.2$, but these resulted in bearing errors between 7° and 17° .

5.2.2 Effects of Multiple MP Signals

As just concluded in Section 5.2.1, both the experimental co-polar and cross-polar responses have been affected by the introduction of additional MP signals. Thus, no accurate indication of the DP transmitter bearing could be determined from these measurements. It is the objective in this section to evaluate the effects of multiple MP signals on the received responses, in an attempt to better comprehend

the experimental data.

The approach taken in this section is to illustrate a possible transformation of the responses, from an initial theoretical shape to a final form in which they resemble the experimental data of Section 4.3. The simulation routines CMPFILES and XMPFILES are used, and progressive modifications to the simulation model are made, as necessary.

Both the measured co-polar responses, i.e., responses of all VP signals, and the measured cross-polar responses, i.e., responses of all HP signals, are analyzed. However to avoid overwhelming the reader with data, not all of the experimental data sets are illustrated. Instead, one representative measurement set out of the total four, see Table 4.1, is chosen to demonstrate the transitions. Nevertheless, the given analysis and all observed trends are applicable to all four sets.

i) Measured Co-Polar Response Comparison:

The representative measurement set chosen for this analysis is Set 4 from Table 4.1, in which $\Delta=30^\circ$, $\phi=200^\circ$, and $A=(0.79, 0.56, \text{ and } 0.4)$. A series of corresponding plots are shown in Figures 5.6a-d to illustrate the possible response transformation, and each of these graphs will be successively discussed. Note that each plot has three curves, representing all mentioned values of A .

In Figure 5.6a, the theoretical responses are illustrated corresponding to the use of both the cardioid radiation pattern and the indicated variable set, in the CMPFILES simulation routine. These obtained curves indicate the ideally 'expected' results for the given scenario. Note the general characteristics of the observed results. The DP is located at a bearing of 180° , yet the primary null of the responses appears to be offset from this location. In addition, there is the formation of a secondary null off to one side of the DP. The presence of this second null is more noticeable, and the bearing offset from 180° is less, for weaker values of A . Consequently, the use of a null-steering technique to determine the signal DOA for

this scenario will give erroneous results.

A natural progression in this analysis is to substitute, in the simulation, the ideal cardioid pattern with the actual range-measured co-polar antenna pattern of the dipole array, as given in Figure 3.8. This substitution is illustrated by the responses of Figure 5.6b. Note that the additional use of the cross-polar power response has a negligible effect on the obtained results, and thus its use is neglected in the simulation.

It is apparent, upon comparison of this figure with Figure 5.6a, that a slight alteration in the co-polar antenna pattern results in significant changes to the obtained responses. The responses in Figure 5.6b are no longer smooth. Instead fluctuations in received power are apparent, corresponding to similar fluctuations or irregularities in the antenna pattern. More specifically, note the formation of new nulls whose null depth is approximately equivalent to the aforementioned secondary null, at least for the case of $A=0.79$. As A decreases these new nulls are less apparent. Also note that the location of the primary null has drifted off center. This is due to the fact that the co-polar antenna pattern, itself, has an off-center null.

A comment may also be made on the effect of this pattern alteration on the other three measurement sets of Table 4.1, which have not been illustrated. It is observed that at a phase of 0° , the responses have not been altered too much except for a decrease in null depth. At a phase of 180° however, the largest response disturbance is encountered. In fact, as an extreme example of the response shown, deep spurious nulls exist in the power responses for $\phi=180^\circ$, and these nulls correspond to the location of each of the antenna pattern irregularities.

The responses of Figure 5.6b must be further modified in order to increase their resemblance to the measured data. Recalling from Section 5.2.1, it was concluded that there possibly was an additional, significant MP signal present during the experimental response measurement. Consequently, to model this scenario an additional MP signal was employed in the simulation. The resulting responses thus

a) Simulated Response using Ideal Cardioid Pattern

b) Simulated Response using Measured Dipole Array Antenna Pattern

Figure 5.6 - Responses Indicating Co-Polar Data Transition

c) Simulated Response with Additional Multipath Signal

d) Measured Field Response

Figure 5.6 - Responses Indicating Co-Polar Data Transition

suggest one possible outcome, for such an addition. With this modification made, the simulation results of Figure 5.6c are obtained.

Comparing Figure 5.6c to Figure 5.6b, one can observe how the primary null has been filled in. Now there is instead a wide, shallow null in this general bearing area. Shown in Figure 5.6d, for comparison, are the measured responses repeated from Chapter 4. As is observed, these latter two responses approximately resemble one another. The null depth of the curves in Figure 5.6c is about 15 dB compared with a depth of about 12.5 dB in Figure 5.6d. In addition, the general response shapes are similar. There is one general broad null, and a gradual rising to either side.

The agreement between the latter two sets of curves is not exact, but further modifications to the simulation can be made in order to attain a better correlation. Recall from Section 5.1.2 that slight changes to the magnitudes of the variables A and/or ϕ resulted in significant alterations to the simulated responses. Consequently, the exact match between measured and simulated responses will then be obtained with the correct setting of all possible modification variables. In conclusion, what is important to note here is that an explanation can be provided to suggest the possible transition of the theoretical data to the measured results.

ii) Measured Cross-Polar Response Comparison:

A similar illustration to the data transition shown for the co-polar responses, may also be given for the cross-polar responses. In this instance the same measurement set will be used for the analysis, which is Set 4 from Table 4.1. Shown in Figures 5.7a-d are graphs representing the possible transformation of the responses from ideal to actual. To follow, these graphs shall be successively discussed.

In Figure 5.7a, the theoretical response set is given for the three indicated strengths of A . These responses are essentially cardioid patterns with their nulls centered on a bearing of 195° , indicating a MP bearing of 15° . Therefore, if

responses such as these could be reproduced during field trials, then a correct MP bearing determination could be made. Unfortunately, these responses do not resemble the measured data of Section 4.3 and thus a modification to the ideal simulation model variables is necessary.

In Figure 5.7b, the simulation results are shown which represent a substitution of the ideal cardioid antenna pattern with both the range-measured co-polar and cross-polar antenna patterns of the loop array, as given from Figure 3.9. The most significant observation of this change is a shifting of the response null back towards a bearing of 180° . This occurs as a direct result of the null offset existing in the loop array co-polar antenna pattern, itself. In addition, the null is filled in such that a new null depth of approximately 27 dB is attained. Apart from these alterations and the observed small decrease in peak power levels, these responses have been only slightly affected by the change in antenna patterns.

These results indicate that further alterations to the simulation are required. Recall from Figure 4.8b that the relative power level of the experimental response of measurement D is comparable to the remaining responses in that set. However, measurement D in this instance is essentially an indication of the cross-polar power level. Consequently, the relatively large power level indicates the introduction of a significant⁵ cross-polar signal component and/or suggests a low cross-polar isolation of the receiving antenna element. To account for this in the simulation, an additional MP signal is modelled. With this adjustment made, the responses of Figure 5.7c are obtained. Note from a comparison of these responses with those of Figure 5.7b how the deep sharp null has been transformed into a broad shallow null.

Shown in Figure 5.7d are the experimental responses repeated from Section 4.3. Comparing the latter two graphs, it is possible to see similarities

⁵ The plots of Figure 5.4 give an indication as to what values of MP signal relative amplitude result in cross-polar power responses of a sufficient magnitude to distort the measured co-polar responses.

a) Simulated Response using Ideal Cardioid Pattern

b) Simulated Response using Measured Loop Array Antenna Pattern

Figure 5.7 - Responses Indicating Cross-Polar Data Transition

c) Simulated Response with Additional Multipath Signal

d) Measured Field Response

Figure 5.7 - Responses Indicating Cross-Polar Data Transition

between these. The null depth of the responses in Figure 5.7c is approximately 14 dB, compared with a depth of about 10 dB for those of Figure 5.7d. In addition, the general response shapes are similar in that neither has a dominant null.

Although the latter two graphs are not exact replicas of one another, it is believed that with further adjustment of the appropriate simulation variables that comparable results may be obtained. Similar to the comment made for the co-polar response comparison, the exact characteristics of the MP signal significantly affect the obtained responses; recall the simulation results of Section 5.1.2. In addition, the simulation of an additional MP signal is hard to do, since its defining characteristics, i.e., signal relative strength, phase, and bearing, are unknown. Nevertheless, further such modifications may lead to a closer resemblance of responses.

5.3 DISCUSSION

In this chapter, a theoretical illustration has been given of a technique by which a transmitter bearing may be determined, for the instance involving only one MP signal. The original hypothesis of Section 1.2 suggested that the measurement and utilization of both the co-polar and the cross-polar responses, for a given propagation environment, would enable the 'normalization' and the effective removal of all MP interference effects. Furthermore, this 'normalization' would be allowed due to the differing signal reception responses, provided from these two separate measurements. The co-polar and the cross-polar responses of Figures 5.6a and 5.7a, respectively, illustrated this concept. These responses correspond to the ideal simulation results, of the signal addition of both one DP and one MP signal. From these responses it was observed that an erroneous measure of the signal direction-of-arrival would be obtained from the sole use of a 'null-steering' DF technique operating on the measured co-polar response. However, the use of the corresponding cross-polar response would give an accurate indication of the MP

signal bearing, and this combined information could then be incorporated into an evaluation routine to extract the required DP signal bearing information.

The DF correlation technique was applied to the experimental data, but unfortunately a complete evaluation could not be performed to determine the achieved accuracy of this technique. It was assumed i) that one or more additional MP signals were present during the experimental response measurement, and ii) that these additional signals caused a noticeable distortion in the measured data. This distortion was such that i) the cross-polar response did not indicate a single null as expected, and ii) a correlation of the co-polar response yielded relatively appealing SSE terms which unfortunately resulted in significant bearing errors.

Next an evaluation of the effects of multiple MP signals on the received responses was performed, in an attempt to understand the measured data. This evaluation entailed an illustration of a possible data transformation, for both the co-polar and the cross-polar responses, from an initial theoretical form to a final form in which a resemblance to experimental data was obtained. Hence by simulating the propagation of multiple MP signals, the experimental responses appeared to be attributable to MP effects. This evaluation has essentially been a theoretical illustration of the response degradation caused by an additional number of MP signals, and it has thus established the limitations of this technique at the present time.

The thesis objective has been to ascertain whether or not the polarization content of a received signal may be utilized in resolving the wavefront interference effects of MP propagation, thereby permitting an increase in direction finding accuracy. From the theoretical analysis, one may conclude that the suggested MP resolution technique looks promising in terms of achieving this goal, for an application involving only one dominant MP signal. However, the use of this technique for a general evaluation purpose involving more than one multipath signal, is restricted due to the ambiguity of the response distortions. A summary of these

conclusions is given in Table 5.2.

Table 5.2 - Application of Dual-Polarization Null-Steering Technique, for DF Measurements

No. of Signal Paths		
1	2	3,4,5...n
X	X	N/A

CHAPTER 6

CONCLUSIONS AND RECOMMENDATIONS

6.1 Summary

A rotatable dual-polarization VHF direction finding antenna has been designed, built and characterized, so as to resolve the effects of MP interference for achieving signal direction-of-arrival measurements. The principal operation of this antenna is based on performing independent measurements of orthogonally polarized signals, through a 'null-steering' amplitude comparison DF technique. In the design process, several constraints had to be met to allow for proper operation. These constraints included such concerns as obtaining i) an adequate element cross-polarization isolation to ensure polarization purity of the received signals, ii) a sufficient antenna balun network signal-to-noise ratio to avoid excessive pattern null-filling, and iii) an appropriate antenna size for both portability and for the reduction of element sagging which introduces cross-polar errors. Specifically, a cross-polar isolation of 21 dB was obtained for both pairs of elements, and SNR's of 31 dB and 25 dB were achieved for the dipole and the loop element balun networks, respectively. Furthermore, the antennas were designed according to the guidelines for a 'small antenna' in which the largest element dimension is less than $\lambda/2$.

Characterizations of the antenna array radiation patterns were recorded at both the DFL range facility and at a DREO site, thus providing the necessary information for subsequent experimental data analysis. The measured patterns possessed attributes of null-filling and null-shifting, relative to the theoretical cardioid response. Specifically, the measured dipole array response indicated a null bearing shift of 25° and a null depth of 35 dB. Similarly, the loop array response exhibited

a corresponding null bearing shift and a null depth of 15° and 25 dB, respectively. Additionally, measured null sensitivities of approximately 3° and 6° were extracted from the dipole array and the loop array responses, respectively, providing an indication of the available null-steering technique precision.

Using the designed DF antenna, a set of experimental measurements was performed to study the effects of a well characterized, single MP signal combined with the direct signal at the receiver. The intent in recording these measurements was i) to demonstrate the diversity of received power responses caused by the interference of these two signals, and ii) to observe if experimental results comparable with theoretical data could be obtained for the given propagation environment. The experimental set-up incorporated the use of a generated DP signal and an artificially generated MP signal. In an attempt to implement the latter, three different types of MP source were investigated. The first two sources consisted of electromagnetic reflectors, with subtle differences, in which a MP specular reflection was generated. Unfortunately, the reflected signal strength in both instances was too weak to have an effect on the direct path signal. The third type of MP source that was utilized, consisted of a transmitting whip antenna to simulate a MP signal. This antenna was referenced in amplitude and phase to the DP signal antenna, and it was oriented at a 45° angle with respect to the ground, so as to transmit equal strengths of co-polar and cross-polar reflective components. The settings of the MP model variables for the experimental measurement sets were selected so as to observe the diversity of received responses, corresponding to a constant MP/DP signal bearing separation, and to a specific range of MP signal relative amplitudes and relative phases.

Upon observation of the experimental data it was confirmed that, as expected, the co-polar responses exhibited interference effects which distorted the single null response, thereby exposing the limitations of the null-steering technique for DOA measurements. However, the cross-polar responses also exhibited

distortion, contrary to the hypothesis of Chapter 1. It appeared that the initial assumptions about the experimental incident signal components were oversimplified, and that there were in fact additional signal contributions not accounted for, which caused the observed response distortions.

Following the presentation of the experimental responses, the data was analyzed using the 'DF correlation technique'. Specifically, this technique involves the correlation of the co-polar responses to theoretical data, incorporated with an observation of the cross-polar response null to indicate the MP signal bearing. Co-polar correlation predictions of Δ , corresponding to four representative data sets, gave predicted values for Δ of 11° , 37° , 13° , and 13° . Unfortunately, these values were offset from the true bearing, and they were assumed to be attributable to the presence of an additional MP signal. The cross-polar responses indicated similar results. However, due to the nature of direction finding at VHF frequencies, the observed response distortions could also be attributable to high noise levels, co-channel interference, etc.

The analysis proceeded with a presentation of the effects of multiple MP signals on the DF correlator. A better understanding of the experimental data was gained, by illustrating the response degradation caused by multiple MP signals. In conclusion, the DF correlation technique appears to be viable for the estimation of transmitter bearings, in scenarios involving only one dominant MP signal. However, for general evaluation purposes involving multiple MP signals, the technique is limited due to the ambiguity of the response distortions.

6.2 Future Considerations

The main emphasis of this thesis has been to investigate the use of orthogonally polarized signal components for achieving direction finding measurements. In this effort, a simple artificial scenario, which combined the

propagation of one direct path and one multipath signal, was utilized for experimental measurements, and a technique for 'normalizing' the received signal wavefront and extracting the direct path signal bearing was proposed. A possible source of a single specular multipath signal, in which this type of technique may have application, is the reflection from a distant mountain, where the surrounding terrain is relatively smooth. Nevertheless, in a more general propagation environment the DF system would have to contend with any number of multipath signals, which are of various magnitudes and phases. The present correlation technique is limited due to the ambiguity of the response distortions. Consequently, the natural progression of the performed analysis would be to investigate the feasibility of expanding the data evaluation to satisfy the general case. This could possibly involve the creation of an elaborate simulation routine to account for all possible signal combinations, and/or it may necessitate the use of a terrain analysis database, specific to the propagation environment.

An improvement to the developed DF system would involve the use of an automatic data acquisition system, to allow for the concurrent measurement of both the received co-polar and cross-polar signals. In this thesis, experimental measurements of such signals were performed consecutively in an environment considered to be essentially time independent. However, in a realistic scenario this is not the case, and the presence of time dependent MP propagation would necessitate the use of concurrent measurements. Furthermore, the addition of such an acquisition system would include the definite advantage of increased speed in data recording over that offered by the existing manual method.

In performing the null-steering DF technique, a cardioid antenna pattern was used for simplicity in order to avoid the 180° bearing ambiguity associated with the dipole antenna response and other such responses. Although this radiation pattern has only a single null, the associated nullwidth is quite broad. This is insignificant in the described experiments since the antenna was used for the sole purpose of

investigating the DF technique, and not for determining operational accuracies. Nevertheless, the operation of this technique in an actual system would require the use of an antenna with a narrow nullwidth response, and thus such an antenna structure should be employed.

The idea to utilize an antenna null rather than an antenna peak, in a steerable DF technique, was considered due to the increased definition of antenna nullwidths over beamwidths. However, the presented analysis has illustrated the ambiguity of the given technique for more than one multipath signal, which is due to the nature of the null-steering technique. Therefore, it may be beneficial to return to the original concept of antenna beam-steering, and to investigate the possibility of designing an antenna with a sufficiently narrow beamwidth and low sidelobe/backlobe levels.

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APPENDICES

APPENDIX A

DERIVATION OF CARDIOD FIELD PATTERN

The approach taken by Kraus [9, pp. 118-125] in the derivation of the cardioid radiation field pattern is to regard the pattern as being produced by an array of point sources. More specifically, the array consists of two point sources which are both identical and isotropic elements.

Referring to Figure A.1, an illustration is given of the wide variety of field patterns available from the use of two vertical elements with various element phase difference δ and element separation distance d .

The field pattern of interest is the cardioid shape shown, with $d = \lambda/4$ and $\delta = 90^\circ$. This can be analyzed by considering the configuration of two point sources of the same amplitude and in phase quadrature. A physical separation distance d is maintained between the two elements and they are oriented as shown on the coordinate system in Figure A.2. Expressing d in radians, one obtains $d_r = (2\pi d)/\lambda$.

The origin of the coordinates is taken as the reference for phase, and thus source 1 is retarded by 45° and source 2 is advanced by 45° . Furthermore, observing the element arrangement in Figure A.2, it is noted that at a far distance in the direction ϕ , the field from source 1 is retarded by an additional $\frac{1}{2}d_r \cos\phi$ while the field from source 2 is advanced by the same amount. Combining these to form the total field at a far distance r in the direction ϕ , one obtains the expression

$$E = E_0 e^{-j\left(\frac{\psi}{2} + \frac{\pi}{4}\right)} + E_0 e^{-j\left(\frac{\psi}{2} - \frac{\pi}{4}\right)} \quad (\text{A.1})$$

where $\psi = d_r \cos\phi$ and E_0 represents the amplitude of the field components at a distance r .

Equation (A.1) may be rewritten as

Figure A.1 - Horizontal-Plane Field Patterns of 2 Vertical Elements as a Function of the Phase Difference δ and the Spacing d . Both elements are the same length and they have current of equal magnitude. The circles indicate the field intensity of a single reference element of the same length with the same power input [9, p. 452].

Figure A.2 - Relation to Coordinate System of 2 Isotropic Point Sources Separated by a Distance d [9, p. 119].

$$E = 2E_o \frac{e^{+j\left(\frac{\psi}{2} + \frac{\pi}{4}\right)} + e^{-j\left(\frac{\psi}{2} + \frac{\pi}{4}\right)}}{2} \quad (\text{A.2})$$

which by application of a trigonometric identity becomes

$$E = 2E_o \cos\left(\frac{\psi}{2} + \frac{\pi}{4}\right) = 2E_o \cos\left(\frac{d_r}{2} \cos\phi + \frac{\pi}{4}\right) \quad (\text{A.3})$$

Normalizing by letting $2E_o = 1$ and by setting the separation distance to the desired amount, $d = \lambda/4$ or $d_r = \pi/2$, Equation (A.3) becomes

$$E = \cos\left(\frac{\pi}{4} + \frac{\pi}{4} \cos\phi\right) \quad (\text{A.4})$$

This final equation gives the desired cardioid-shaped field pattern.

APPENDIX B

ELEMENT SIZE CALCULATIONS

To determine the element dimensions, the input impedance is calculated for various element sizes; the optimal dimensions are then chosen which satisfy the constraints on the radiation resistance and the element weight. Shown are the calculations of both the dipole and the loop dimensions.

i) Dipole Dimensions:

Details of the dipole dimensions are shown in Figure 3.5a and they are reiterated in Figure B.1 for convenience.

The optimum length of a dipole in terms of radiation resistance is $\lambda/2$. However, in this instance $\lambda = 5$ m and thus a dipole length of $\lambda/2$ will result in an antenna which is too large and heavy; this in turn would cause arm sagging which disturbs the polarization purity of reception.

Instead, the element must be reduced in size from this optimum length. The input impedance Z_{in} for a short antenna, defined as one which satisfies the constraints $L < \lambda/2$ and $L \ll \lambda$ where L is the largest dimension of the element, is given by Jasik [22, p. 4-5] as

$$Z_{in} = 20(kl)^2 - j \frac{120}{kl} \left[\ln \left(\frac{l}{a} \right) - 1 \right] \quad (B.1)$$

where $k = 2\pi/\lambda$, $l = L/2$, and a is the dipole radius.

Thus the dipole length is chosen to satisfy the constraints imposed on Equation (B.1). Note from this equation that the radiation resistance for a given frequency is dependant, in direct proportion, upon only the half length dimension l . Consequently, the length is chosen as large as possible to obtain the maximum

radiation resistance while still satisfying the constraints of Equation (B.1). Specifically, a dipole length of $L = 1/8\lambda$ or 62 cm $\ll \lambda$ is selected. See Table B.1 for the effect of increasing element half length on radiation resistance.

For the diameter of the dipole d , it is required that $d \ll L$. If this relation is not satisfied, then the possibility of the reception of significant cross-polar levels increases. Table B.1 shows the effect of different element radius on the radiation reactance. It is seen that the reactance is inversely proportional to the element radius. One requires the radius to be large enough so as to make the dipole rigid and thus prevent bending, while in turn producing a relatively small reactive component. However, the radius must not be too large so as to pick up the cross-polar field component of the signal across its diameter. Accordingly, a radius of $a = 0.476$ cm is chosen.

The input impedance of the dipole, from Equation (B.1), is then $Z_{in} = (3.035 - j978.44) \Omega$ and the dipole dimensions are as specified: dipole diameter $d = 0.952$ cm and dipole length $L = 62$ cm.

ii) Loop Dimensions:

Details of the loop dimensions are shown in Figure 3.5b and they are reiterated in Figure B.2 for convenience.

Similar to the dipole design, the loop dimensions must be reduced in size from the optimum $\lambda/2$ diameter which is too large. Again, the antenna input impedance will first be calculated and then used to select the appropriate element dimensions.

For the loop, the input impedance Z_{in} can be defined by

Figure B.1 - Dipole Dimensions

Table B.1 - Impedance Values Z_m for Various Dipole Dimensions

		Element radius, a (cm)		
		0.794	0.476	0.318
Element half-length, l (cm)	20.8	1.366-j1040.15 Ω	1.366-j1275.06 Ω	1.366-j1460.24 Ω
	25.0	1.974-j935.65 Ω	1.974-j1131.10 Ω	1.974-j1285.17 Ω
	31.0	3.035-j820.83 Ω	3.035-j978.44 Ω	3.035-j1102.69 Ω

$$\begin{aligned} Z_{in} &= R + jX \\ &= (R_r + R_o) + j\omega L \end{aligned} \quad (B.2)$$

where R_r is the radiation resistance, R_o is the total ohmic loss resistance, and $j\omega L$ is the loop reactance.

In Equation (B.2), the radiation resistance for a small circular loop is given by Balanis [10, p. 170] as

$$R_r = \eta \left(\frac{8}{3} \pi \right) \left(\frac{a}{\lambda} \pi \right)^4 N^2 \quad (B.3)$$

where $\eta = (\mu/\epsilon)^{1/2}$ is the intrinsic impedance of the medium, and μ and ϵ are the corresponding permeability and permittivity, respectively, a is the loop radius, and N is the number of turns of the loop wire.

Equation (B.3) can be simplified for the case of an air core loop where $\eta = 120\pi$, thus giving a radiation resistance of

$$R_r = 320\pi^6 \left(\frac{a}{\lambda} \right)^4 N^2 \quad (B.4)$$

As a direct result of the large wavelength in Equation (B.4), $\lambda = 5$ m, the radiation resistance is limited in value. Thus for a practical loop diameter and number of turns, R_r is chosen to be 0.5Ω . Using this value in Equation (B.4), one obtains a selection for the loop radius and number of turns, as shown in Table B.2. From this table, the two turn loop of radius 12.6 cm is then chosen.

The loss resistance for the same small antenna is given by Balanis [10, pp. 171-172] as

$$R_o = \frac{Na}{b} R_s \left(\frac{R_p}{R_o} + 1 \right) \quad (B.5)$$

Figure B.2 - Loop Dimensions

Table B.2 - Number of Turns N and Loop Radius a , corresponding to $R_r = 0.5$

Number of Turns, N	Loop Radius, a (cm)
12	5.2
10	5.6
8	6.3
6	7.3
4	8.9
3	10.3
2	12.6

where b is the wire radius, R_s is the conductor surface resistance due to skin effect, R_p is the ohmic resistance due to proximity effect, and R_o is the ohmic skin effect resistance per unit length.

In Equation (B.5), R_s is calculated from $(\omega\mu_o/2\sigma)^{1/2}$ where ω is the radian frequency, μ_o is the permeability of free space and σ is the wire loop conductivity. At a frequency of 60 MHz, $R_s = 2 \text{ m}\Omega$. The ratio R_p/R_o is read off of a plot [10, p. 172] as a function of the ratio c/b , where $2c$ is the loop separation.

A ratio of wire separation to wire diameter c/b of 5 is chosen, thus giving a ratio R_p/R_o of 0.05. The ohmic resistance from Equation (B.5) is now calculated to be 1.32Ω . Thus the loss resistance R_o is larger than the radiation resistance R_r , resulting in a poor efficiency. However, this will not be of too much concern, as absolute power levels are not important.

The loop reactance is calculated from the equation [22, p. 5-4]

$$L = a\mu \left[\ln \left(\frac{16a}{2b} \right) - 2 \right] \quad (\text{B.6})$$

from which jX is determined to be $j347.2 \Omega$. Thus the total input impedance of the loop is $Z_{in} = (1.82 + j347.2) \Omega$. Summarizing, the loop dimensions are as follows: loop diameter $d = 25.2 \text{ cm}$, wire separation $2c = 4 \text{ mm}$ and wire diameter $2b = 0.8 \text{ mm}$.

APPENDIX C

BALUN DESIGN

The balun network consists of two op amp stages cascaded together. Stage 1 is designed to act as a voltage and an impedance transformer, while providing little gain. Stage 2 is designed chiefly to provide the necessary signal amplification. Details of these designs follow.

i) Stage 1:

This stage consists of a CLC103AI operational amplifier (op amp) in a differential input configuration, as shown in Figure C.1a. The resistance R_2 in the negative feedback loop is a built-in resistor (1500Ω) of the CLC103 chip. The gain G is the ratio of R_2/R_1' where $R_1' = R_1 + |Z_{in}|/2$. Here the element input impedance $|Z_{in}|$ is divided equally between both op amp input terminals, and R_1 is a precision resistor added in series to obtain the desired gain.

In selecting R_1' , consideration is given to both the desired gain and bandwidth. From the op amp specifications, it appears that a flat power response can be obtained at the frequency of interest, 60 MHz, with a closed loop voltage gain of 2. Thus $R_1' = 750 \Omega$, for both element pairs.

The element input impedances, as calculated in Appendix B, are

$$\begin{aligned} Z_{in} \text{ (dipoles)} &= (3.035 - j978.46) \Omega \\ Z_{in} \text{ (loops)} &= (1.82 + j347.2) \Omega \end{aligned} \tag{C.1}$$

From these values the required precision resistors are calculated, as shown:

Dipoles:

$$|Z_{in}| = 978.5 \, \Omega$$

$$|Z_{in}|/2 = 489.25 \, \Omega$$

$$\begin{aligned} R_1 &= R_1' - |Z_{in}|/2 \\ &= (750 - 489.25) \, \Omega \\ &= \underline{260.8 \, \Omega} \end{aligned}$$

Loops:

$$|Z_{in}| = 347.2 \, \Omega$$

$$|Z_{in}|/2 = 173.6 \, \Omega$$

$$\begin{aligned} R_1 &= R_1' - |Z_{in}|/2 \\ &= (750 - 173.6) \, \Omega \\ &= \underline{576.4 \, \Omega} \end{aligned}$$

After calculating the desired resistor values, a biasing circuit is configured. This is shown in Figure C.1b. As is characteristic of an op amp, the output terminal has a low impedance $R_{out} = 1.3 \, \Omega$. However, we wish to feed a $50 \, \Omega$ impedance coaxial cable and thus a resistor $R_4 = 48.7 \, \Omega$ is placed in series to achieve this. Note that this has the effect of halving the output voltage, and thus one is left with a unity voltage gain.

ii) Stage 2:

This stage is provided to amplify the received signal level over any level of noise picked up or generated down the transmission line. A Mini-Circuits MAR-6 cascable amplifier is used because of its compact size and its desirable gain at 60 MHz, a typical power gain of about 20 dB.

To design the amplifier section, a microstrip board has to be constructed and the biasing elements calculated¹. A small square of duroid is used and the appropriate $50 \, \Omega$ microstrip lines are etched on. The amplifier circuit configuration is shown in Figure C.2.

A bias stabilization resistor is used in the power supply path to provide bias stability with temperature fluctuations. It is calculated from $R_c = (V_{cc} - V_d)/I_d [\Omega]$

¹ The guidelines supplied by Mini-Circuits entitled "A Handy 'how-to-use' Guide for MAR monolithic drop-in amplifiers", pages 2-7, are used.

a) Schematic

b) Biasing Circuit

Figure C.1 - Stage 1 Design

where V_{cc} is the power supply voltage [V], V_d is the voltage level at the DC input terminal of the MAR-6 [V], and I_d is the quiescent bias current drawn by the MAR 5 [A].

Using a power supply of +12 V, a resistor value of $R_c = 531 \Omega$ is calculated. In series with R_c is an RF choke, of a sufficient value to keep this resistor from appearing in parallel with the load circuit and thus degrading the output match. Also, DC blocking capacitors, $C = 1000$ pF, are used in both the RF input and output lines to provide isolation.

The two stages are cascaded together with 50 Ω RG174 cable for the signal transmission, and an illustration of the circuit hardware for these two stages is shown in Figure C.3. Furthermore, the entire balun network is enclosed in a shielded box, which is used as the power ground for the circuits. These circuits are fed with a power supply line of ± 12 V.

Figure C.2 - Biasing Circuit for Stage 2 Design

Figure C.3 - Circuit Hardware of Cascaded Stages

APPENDIX D

SAMPLE CALCULATION OF MP SIGNAL RELATIVE AMPLITUDE, A

This appendix gives details on how the MP signal relative amplitude values A , found in Table 4.1, are calculated. A sample calculation is demonstrated corresponding to a MP relative phase of 180° ; however, equivalent calculations were performed for $\phi = 0^\circ$, 160° and 200° .

Recall the equipment arrangement for the field measurement sets of concern, as shown in Figure D.1. The signal paths T_1 -R and T_2 -R utilize free-space as a propagation medium, whereas the signal path T_1 - T_2 uses an RG58c co-axial transmission line. Initially it is necessary to calculate the required cable length for path T_1 - T_2 . This is achieved by calculating a free-space path length (greater than the physical path length) to permit the proper MP relative phase, and then by multiplying this value by the relative propagation speed in the cable. The obtained results are summarized in Table D.1.

After the required cable length has been calculated, the propagation loss values need to be determined. For the two signal transmission media, the propagation losses are defined as

$$\text{RG58c Cable Loss (@ 60MHz)} = 11.86 \times 10^{-2} \text{ dB/m} \quad (\text{D.1})$$

$$\text{Free-Space Propagation Loss (dB)} = 20\log D_{\text{km}} + 20\log f_{\text{MHz}} + 32.44 + L_o$$

where D_{km} is the propagation distance in km, f_{MHz} is the operating frequency in MHz, and L_o accounts for any additional losses. As a simplification, it is assumed that $L_o = 0$ in the calculations. Table D.2 summarizes the obtained propagation losses, utilizing these equations.

From Table D.2, the total DP propagation loss L_{DP} is 40.04 dB while the loss for the MP path L_{MP} is 39.04 dB + A' , where A' is the added in-line MP signal attenuation in dB. Ascertaining the difference P in propagation losses between both signal paths, i.e., $L_{DP} - L_{MP}$, enables a determination of A by using the equation

$$A = V_1 = 10^{P/20} * V_o * \sin(45^\circ) \quad (D.2)$$

where V_o = DP reference voltage,

$A = V_1$ = MP Signal Relative Amplitude,

P = power difference between signal paths.

Note that the $\sin(45^\circ)$ term is used to obtain the equivalent co-polar and cross-polar components of the MP signal. A summary of the latter calculations, and a presentation of the determined values for A , are given in Table D.3. Also note that the change in cable lengths to obtain the different phases is sufficiently small such that the illustrated values of A apply equally for all phases.

Figure D.1 - Arrangement of Equipment for Field Trials

Table D.1 - Required Cable Length for $\phi = 180^\circ$

Blk #	1	2	3	4	5	6
ϕ	T_1 -R (m)	T_2 -R (m)	$(1+4-2+n\lambda)$ free-space path length T_1 - T_2 (m)	Add.MP Path Length (m)	Relative Prop. Speed in Cable	(3×5) T_1 - T_2 Required Cable Length (m)
180°	40	25.52	$21.98 + 3\lambda$	7.5	0.66	24.41

Table D.2 - Propagation Loss Values for $\phi = 180^\circ$

Path	Path Length (m)	Free-Space Loss (dB)	Cable Loss (dB)
T ₁ -R	40	40.04	N/A
T ₁ -T ₂	24.41	N/A	2.90
T ₂ -R	25.52	36.14	N/A

Table D.3 - Summary of Calculated A Values

In-line MP Atten. A' (dB)	Δ Prop. Loss $L_{DP} - L_{MP}$ (dB)	A
0	1	0.79
3	-2	0.56
6	-5	0.40

APPENDIX E

SIMULATION PROGRAMS

This appendix contains two Pascal simulation routines developed for the analysis purposes of Chapter 5. The CMPFILES routine calculates the received vertically polarized response to the addition of one DP signal and one MP signal. Similarly, the XMPFILES routine calculates the received response to the addition of these same signals, but focuses on the horizontally polarized signal components rather than on the vertical.

The values of A , ϕ - phi, and Δ - del may be chosen by the user. Furthermore, the input antenna radiation pattern file may be set accordingly. XMPFILES allows for the input of both the co-polar and the cross-polar patterns while CMPFILES utilizes only the co-polar response. In addition to these two routines, the CARDIOID program is included which creates the data file of the cardioid antenna response. This file is then used as input to the two routines.

i) CMPFILES:

```
program cmpfiles (patt1, cmpout, input, output);
    {accepts an antenna response data file as input, and
    from this calculates the received co-polar response for
    the addition of one DP signal and one MP signal.
    patt1 - ideal cardioid response
    patt2 - actual antenna response                                }

type
    ray = array[1..360] of real;
var
    U                : ray;
    E1, E2, Power    : real;
    patt, cmpout      : text;
    I, J              : integer;
```

```

const
    pi      = 3.141592654;
    loge    = 0.4342944;
    A       = 0.8;          {user defined}
    Phi     = 180/180*pi;   {user defined}
    Del     = 30;           {user defined}

begin
    assign (patt, 'c:\language\tp\dfiles\patt1');
                                {assigns input antenna pattern file to file variable patt}
    reset (patt);
                                {moves pointer to beginning of external file 'patt'}
    assign (cmpout, 'c:\language\tp\dfiles\cmpout');
                                {assigns output data file to file variable cmpout}
    rewrite (cmpout);
                                {creates an empty external file for simulated vertically
                                polarized MP response values}
    for I:= 0 to 359
        do readln (patt, u[i]);
    for J:= 0 to 359
        do begin
            if del>j
                then begin
                    E1:= U[j] + A*cos(phi)*U[j+360-del];
                    E2:= A*sin(phi)*U[j+360-del]
                end
            else
                begin
                    E1:= U[j] + A*cos(phi)*U[j-del];
                    E2:= A*sin(phi)*U[j-del]
                end;
            Power:= 10*loge*ln( sqr(E1) + sqr(E2) );
            writeln (cmpout, J:3, ' ', Power:8:3);
        end;
    close (cmpout)
end.

```

ii) XMPFILES:

```

program xmpfiles (patt, pat, xmpout, input, output)
    {reads in values of co-polar and cross-polar antenna
    radiation patterns, and from this calculates the received
    cross-polar response to the addition of one DP signal}

```

```

        and one MP signal.
        patt1 - ideal cardioid response
        patt2 - actual co-polar antenna response
        pat2 - actual cross-polar antenna response    }

type
    ray = array[1..360] of real;
var
    UC, UX          : ray;
    E1, E2, Power    : real;
    patt, pat, xmpout : text;
    I, J             : integer;
const
    pi      = 3.141592654;
    loge     = 0.4342944;
    A        = 0.8;        {user defined}
    phi      = 200/180*pi; {user defined}
    del      = 30;         {user defined}

begin
    assign (patt, 'c:\language\tp\dfiles\patt2');
        {assigns input antenna co-polar data file to file variable
        patt}
    assign (pat, 'c:\language\tp\dfiles\pat2');
        {assigns input antenna cross-polar data file to file
        variable pat}
    reset (patt);
        {moves pointer to beginning of external file 'patt'}
    reset (pat);
        {moves pointer to beginning of external file 'pat'}
    assign (xmpout, 'c:\language\tp\dfiles\xout');
        {assigns output data file to file variable xmpout}
    rewrite (xmpout);
        {creates an empty external file for simulated cross-polar
        MP response values}
    for I:= 0 to 359
        do begin
            readln (patt, UC[i]);
            readln (pat, UX[i]);
        end;
    for J:= 0 to 359
        do begin
            if del > j
                then begin

```

```

        E1:= (A*cos(phi)*UC[j+360-del])
            + (UX[j] + A*cos(phi)*UX[j+360-del]);
        E2:= (A*sin(phi)*UC[j+360-del])
            + (A*sin(phi)*UX[j+360-del]);
    end
else
    begin
        E1:= (A*cos(phi)*UC[j-del])
            + (UX[j] + A*cos(phi)*UX[j-del]);
        E2:= (A*sin(phi)*UC[j-del])
            + (A*sin(phi)*UX[j-del]);
    end;
    Power:= 10*log10( sqrt(E1) + sqrt(E2) );
    writeln (xmpout, J:3, ' ', Power:8:3);
end;
close (xmpout)
end.

```

iii) CARDIROID:

```

program cardioid (patt1, output);
    {creates external 'patt1' file containing cardioid response
    for complete azimuth bearing rotation          }

type
    ray = array[1..360] of real;
var
    U      : ray;
    theta  : real;
    i      : integer;
    patt    : text;
const
    pi     = 3.141592654;

begin
    assign (patt, 'c:\language\tp\dfiles\patt1');
    {assigns input data file to file variable 'patt'}
    rewrite (patt);
    for I:= 0 to 359
        do begin
            theta:= I/180*pi;
            U[i]:= cos(pi/4 + pi/4*cos(theta+pi) );
            writeln (patt, u[i]:9:4, ' ', I:3);
        end
    end
end.

```



```
    end;  
close(patt);  
end.
```

END

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FIN