SITING CRITERIA FOR THE MICROWAVE LANDING SYSTEM (MLS):
MLS/ILS COLLOCATION AND RUNWAY HUMP SHADOWING

A Thesis Presented to the
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Master of Science

by
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FOREWORD

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INTRODUCTION

The concepts of the Microwave Landing System (MLS) date back to the early 1950's (1). From that time a number of implementations using microwave technology were proposed for use as precision approach and landing systems. However, only three types of MLS systems received extensive testing: a Doppler concept, a time reference scanning beam (TRSB) concept, and a DME based landing system (DLS). The scanning beam technique was accepted by the International Civil Aviation Organization (ICAO) in 1978 as the standard technique to be used worldwide (1). Improvements such as electronic scanning and solid state digital electronics have contributed to the development of the MLS.

MLS is designed to be an all-weather precision approach and landing system capable of meeting accuracies equivalent to ICAO Category III standards. MLS operates with an internationally standardized signal format which allows any aircraft equipped with a standard MLS receiver to make a precision approach to any MLS equipped runway. Also, MLS provides a continuous ground-to-air data link to the aircraft. The modular design of MLS makes it flexible and capable of meeting the needs of individual installations.
MLS offers a larger volume of coverage than the currently used Instrument Landing System (ILS). Specifically, the MLS format can provide proportional guidance over a maximum service volume of ±62 degrees in azimuth (typically ±40 degrees is implemented) and up to 30 degrees in elevation. In comparison, the ILS offers a maximum proportional guidance region of ±3 degrees in azimuth and ±0.7 degrees, about a fixed glide slope angle, in elevation. An MLS receiver coupled with a navigation computer will permit the full utilization of this increased coverage volume through executing approved segmented and curved approaches. This capability is very desirable for noise abatement, obstruction avoidance procedures, and the selection of approach profiles that best fit the performance capabilities of the aircraft.

Employing microwave frequencies allows MLS antennas to be electrically large while remaining relatively small physically. In addition, MLS antennas establish the guidance signal in space without the aid of a ground plane. These characteristics reduce the MLS's sensitivity to terrain irregularities and multipath interference due to objects commonly present in the airport environment.

Further information can be found in "Siting Criteria for Microwave Landing Systems (MLS)" developed by the Avionics Engineering Center, Ohio University (2).
purpose of this document is to provide the field engineer with siting criteria for the MLS. This document is constantly revised as new information becomes available on important siting topics related to the MLS. Two such topics are addressed in this thesis: attenuation of the azimuth signal due to a humped runway, and the collocated siting of MLS and ILS guidance systems at the same landing facility.
III PROBLEM STATEMENT AND DEFINITION

The MLS azimuth station is composed of the azimuth antenna, angle guidance equipment, distance measuring equipment (DME), and an AC power cabinet. This station is typically sited 1,000 ft. from runway stop end on the runway centerline extended. The azimuth guidance information in the runway region must be supplied down to a height of 2.44 m (8 ft.) above the runway surface. In cases where there exists a hump in the runway, either for drainage or to avoid costly excavation, line-of-sight between the azimuth antenna and a point 2.44 m (8 ft.) above the runway surface at threshold may be blocked.

Because of the high MLS source frequency, 5.03-5.09 GHz, the MLS is typically considered a line-of-sight system. In addition, the vertical pattern of the MLS azimuth antenna cuts off very sharply below the horizon to reduce the effect of ground reflections on the azimuth guidance signal. The consideration of these two factors results in concern that the azimuth signal may be appreciably attenuated due to signal blockage by a runway hump. Although the receiver would be relatively close to the azimuth antenna, the azimuth signal may be attenuated enough to cause an unacceptable signal-to-noise ratio in the shadowed region. This low signal-to-noise ratio could result in poor or unstable autopilot performance and general derogation of the azimuth guidance signal.
Therefore, the development of a computer model which accurately predicts the attenuation of the azimuth signal due to the runway hump is very desirable. This modeling capability would save many man-hours and the expense of performing field measurements at individual humped runway sites.

Presently there are two practical electromagnetic modeling methods that can be applied to account for diffraction effects: the Moment Method (MM) and the Geometrical Theory of Diffraction (GTD). Due to computational constraints MM is usually considered a low-frequency method, in comparison to GTD which is a high-frequency asymptotic method (3). Since the scattering object in this case is large in comparison to a wavelength, a high-frequency method is required for the runway hump problem. This fact makes GTD the only candidate.

The first goal of this thesis is the development of a GTD runway hump model. Achieving this goal requires implementing theory which is presented under ideal circumstances and modifying it to account for the characteristics of microwave propagation over real (non-ideal) earth.
The second goal involves analysis of MLS-ILS collocation. During the transition from ILS to MLS, many locations now providing ILS guidance will become host to MLS facilities also. However, before the transition can be implemented the effects of MLS structures on ILS signals must be determined. In addition the MLS must be sited so that the ILS service is preserved while permitting full utilization of the MLS.

Due to the present limitations of electromagnetic modeling, which will be discussed in further detail in section Vb, the collection of experimental data is the only practical method of obtaining the information required. Since the funding for this work is provided by an outside source, this project provided a unique opportunity to apply engineering skills to collect meaningful data in a cost-effective manner. The strategy is to determine which areas about the ILS antennas will be the most and the least sensitive to placement of the MLS equipment. This process requires a firm working knowledge of antenna theory and of the ILS itself. In view of the theory on how the ILS signals are established in space two hypotheses were developed. For collocating the azimuth antenna with the localizer, when it is required that the azimuth antenna be sited in front of the localizer, the hypothesis is that the minimum effect on the localizer course will occur when the azimuth antenna is sited symmetrically about the localizer
course centerline. For collocating the elevation antenna with the glide slope, when the elevation antenna is to be sited on the runway side of the glide slope, the hypothesis is that the minimum effect on the glide slope will occur when the elevation antenna is sited so that it does not penetrate the region through which the Fresnel zone migrates during an approach. To verify the two hypotheses and obtain quantitative information, flight measurements were performed.

Therefore, the second goal of this thesis is to validate the two hypotheses for the collocated siting of MLS and ILS equipment and to develop siting criteria for inclusion in reference 2. This process will require the development of test plans and data analysis.
IV GEOMETRICAL THEORY OF DIFFRACTION (GTD) MODEL FOR PREDICTING THE EFFECTS OF RUNWAY HUMP SHADOWING

a. Ground Plane and Edge Effects

In solving many antenna or radiation problems it becomes necessary to consider the effects of objects in close proximity to the radiating source. For example, the change in an antenna’s radiation pattern or impedance when placed at some finite distance above an infinite ground plane may be too important to ignore. Understanding the effects of an ideal ground plane on an antenna’s characteristics can be used to an advantage. An application of this knowledge, image theory, is the key to the development of the glide slope portion of the ILS. Although this is a good starting point, real ground planes are not infinite in size nor conductivity. There must be further investigation to predict the effects on the radiation pattern due to a finite ground plane. The truncation of the ground plane can range from an abrupt termination such as the roof of a building on which the antenna is mounted, or a gradual tapering off such as a sloping grade or hilly terrain.

In some problems objects need not be close to the source to merit consideration. The effects of an object, such as a mountain range, between a transmitting and receiving antenna certainly need to be considered. The
effect this geometry has on the field strength at the receiving antenna is an important consideration in a communications study. A diffraction theory is needed in order to predict the effects of edges or boundaries. Such a diffraction theory is particularly useful in determining a non-zero field where the incident field, line-of-sight, is blocked. Problems like these were the prime motivation in the development of the Geometrical Theory of Diffraction (GTD).

GTD is a high frequency technique used to determine the diffracted field given the incident field at the point of diffraction. Furthermore, at high frequencies, diffraction, like reflection and transmission, is a local phenomenon where the total diffracted field is proportional to the incident field (4). The coefficient relating the two fields is referred to as the diffraction coefficient. The characteristics of the diffraction coefficient depend on the properties of the incident field at the point of diffraction as well as the object responsible for the diffraction.

The diffraction coefficient, generally a dyadic, can be obtained by satisfying boundary conditions imposed by the diffracting surface. As stated earlier, the characteristics of the diffraction coefficient are dependent on the geometry of the problem. In most case it is desirable to solve the problem for a general geometry. This general problem is
referred to as a canonical problem. Two common canonical problems are wedge diffraction and diffraction from a smooth convex surface. There are other canonical problems, such as tip or vertex diffraction, but they are not as common.

There are several advantages to using GTD (3). It is a simple technique to use. Secondly, a complex problem can be broken up into simpler elements. Then the solutions of the individual elements can be combined through superposition to give the final result. By observing the characteristics of the diffracted field, GTD provides the user insight into the scattering mechanisms of the structure being investigated. Furthermore, GTD provides accurate results which compare well with measured data. Finally, it allows the user the flexibility to use this technique with other methods. This approach is commonly referred to as a hybrid technique.

In general, the diffraction coefficient can be represented in the form (3)(figure 4-1):

$$E^d(s) = E^i(Q) \cdot D \cdot A(s', s) e^{-jks}$$

(4-1)

where:

- $E^d(s)$ is the diffracted field at observation point p
- $E^i(Q)$ is the incident field at the point of diffraction
- $D$ is the diffraction coefficient
- $A(s', s)$ is the spatial spreading factor
OBLIQUE INCIDENCE

Figure 4-1 Oblique incidence on a two-dimensional conducting wedge.
A detailed discussion of this equation for the case of normal incidence on an infinitely conducting wedge will be taken up in the following section.

b. Wedge Diffraction Coefficients

In addition to providing a diffracted field there is another reason for selecting the GTD wedge for investigating the effect of a runway hump on the azimuth signal. Most runway profiles can be represented very well by a few straight line segments. Particularly, a humped runway profile can be represented as a diffracting wedge with a large wedge angle. This fact in conjunction with the advantages of GTD discussed previously make the GTD diffracting wedge a prime candidate for this problem.

Before discussing normal incidence edge diffraction, the general problem of oblique incidence edge diffraction will be presented. This overview will provide the reader an important benefit. This benefit is to see how the case of normal edge diffraction is derived from the more general oblique incidence case. Furthermore, once an understanding of normal incidence edge diffraction is reached the material presented in the following section could be used to solve a problem involving oblique incidence edge diffraction.
1. Oblique Incidence Edge Diffraction

Figure 4-1 illustrates the ray-fixed coordinate system for oblique plane wave incidence on a two-dimensional conducting wedge. The advantage of the ray-fixed coordinate system for oblique incidence is the dyadic diffraction coefficient is the sum of two dyads. In comparison, an edge fixed coordinate system would result in a dyadic diffraction coefficient composed of seven dyads. For the source at \((s', \beta_0', \phi')\) and the observation point \((s, \beta_0, \phi)\) the diffracted field has the general form (3):

\[
E^d(s) = E^i(Q) * D(\phi, \phi', n; \beta_0') * A(s, s') e^{-jks} \tag{4-2}
\]

The diffracted field can be represented in matrix form as follows:

\[
\begin{bmatrix}
E_{\beta 0}^i(s) \\
E_\phi^d(s)
\end{bmatrix} = \begin{bmatrix}
-D_s & 0 \\
0 & -D_h
\end{bmatrix} \begin{bmatrix}
E_{\beta 0'}^i(Q) \\
E_\phi'^i(Q)
\end{bmatrix} * A(s, s') e^{-jks} \tag{4-3}
\]

where

- \(E_{\beta 0'}^i(Q)\): is the incident E-field component parallel to the plane of diffraction.
- \(E_\phi'^i(Q)\): is the incident E-field component perpendicular to the plane of diffraction.
- \(D_s\): is the diffraction coefficient for horizontal (soft) polarization.
- \(D_h\): is the diffraction coefficient for vertical (hard) polarization.
The distance parameter $L$ and the spatial attenuation factor $A(s,s')$ are dependent on the incident field and are defined by:

$$L = \begin{cases} 
    \frac{s \sin^2 \beta_0'}{\rho \rho'/\rho + \rho'} & \text{plane wave incidence} \\
    \frac{s s' \sin^2 \beta_0'/s + s'}{s' / s(s + s')} & \text{cylindrical wave incidence} \\
    \frac{s' / s(s + s')}{1/s} & \text{spherical wave incidence}
\end{cases}$$

For far field observation ($s >> s'$, or $\rho >> \rho'$) the form for $L$ and $A(s,s')$ reduce to:

$$L = \begin{cases} 
    \frac{s \sin^2 \beta_0'}{\rho'} & \text{plane wave incidence} \\
    \frac{s' \sin^2 \beta_0'}{s' / s} & \text{spherical wave incidence}
\end{cases}$$

$$A(s) = \begin{cases} 
    \frac{1}{\sqrt{s}} & \text{plane, cylindrical, and conical wave incidence}
\end{cases}$$
2. Normal Incidence Edge Diffraction

For normal incidence edge diffraction, $\beta_0$ and $\beta_0'$ are equal to $\pi/2$ and the distance parameter $L$ takes the form

$$L=\begin{cases} 
  s & \text{plane wave incidence} \\
  \rho'\rho'/\rho + \rho' & \text{cylindrical wave incidence} \\
  ss'/s+s' & \text{spherical wave incidence}
\end{cases}$$

and for far field observations

$$L=\begin{cases} 
  s & \text{plane wave incidence} \\
  \rho' & \text{cylindrical wave incidence} \\
  s' & \text{spherical wave incidence}
\end{cases}$$

The geometry for normal incidence on a two-dimensional conducting wedge is shown in figure 4-2. The two boundaries, the incident shadow boundary and the reflection shadow boundary, divide the space about the wedge into three regions. The total field predicted by the vector sum of the direct and reflected rays will result in discontinuities at the incident shadow boundary and the reflection shadow boundary. These discontinuities represent a non-physical condition and were a strong indication that a diffraction term was missing from geometrical optics. In addition, the field in region III will be zero. The development of a wedge diffraction coefficient will allow the calculation of a non-zero field in region III and will provide a smooth
Figure 4-2 Normal incidence on a two-dimensional conducting wedge.
transition at the incident and the reflection shadow boundaries.

The diffraction coefficient for the two-dimensional wedge can be derived as follows (3):

1. Find the Green’s function solution in the form of an infinite series using modal techniques, and then approximate it for large values of kp (far-field observations).

2. Convert the infinite series Green’s function into integral form.

3. Perform on the integral form of the Green’s function a high-frequency asymptotic expansion using standard techniques (method of steepest descents).

It can be shown that the diffraction coefficient for this geometry (edge-fixed coordinate system) has the form (3):

\[ D = D_i + D_r \]  \hspace{1cm} (4-10)

The sign chosen depends on the incident field and the electrical characteristics of the diffracting surface. For a horizontally (soft) polarized electric line source (the electric field vector is parallel to the edge of the wedge)
above an infinitely conducting surface the diffraction coefficient has the form:

\[ D^s = D^i - D^r \]  \hspace{1cm} (4-11)

For a magnetic line source, vertical (hard) polarization, above an infinitely conducting surface the diffraction coefficient then is represented as:

\[ D^h = D^i + D^r \]  \hspace{1cm} (4-12)

Specifically, \( D^i \) and \( D^r \) take the forms:

\[ D^i = -\left( e^{-j\pi/4} \right) / 2n\sqrt{2\pi k^*} \]  \hspace{1cm} (4-13)
\[ \left[ C^+ (\beta^-, n) F(kp' g^+ (\beta^-)) + C^- (\beta^-, n) F(kp' g^- (\beta^-)) \right] \]

\[ D^r = -\left( e^{-j\pi/4} \right) / 2n\sqrt{2\pi k^*} \]  \hspace{1cm} (4-14)
\[ \left[ C^+ (\beta^+, n) F(kp' g^+ (\beta^+)) + C^- (\beta^+, n) F(kp' g^- (\beta^+)) \right] \]

where

\( \beta^+ = \phi + \phi' \),  \( \beta^- = \phi - \phi' \)  \hspace{1cm} (4-15)

\( C^+ (\beta, n) = \cot[ (\pi + \beta) / 2n] \)  \hspace{1cm} (4-16)

\( C^- (\beta, n) = \cot[ (\pi - \beta) / 2n] \)  \hspace{1cm} (4-17)

\[ F(kp' g (\beta)) = 2j\sqrt{kp' g(\beta)} \cdot e^{jkp' g(\beta)} * \]

\[ \int_{0}^{\infty} e^{-j\tau'} \frac{d\tau}{\sqrt{k\tau' g(\beta)}} \]  \hspace{1cm} (4-18)

\( (\tau' = \tau^2) \)
where the integers $N^+$ and $N^-$, can be, positive, negative, or zero to best satisfy the following:

\begin{align*}
2n\pi N^+ - (\phi'\phi) &= +\pi \quad (4-21) \\
2n\pi N^- - (\phi'\phi) &= -\pi \quad (4-22)
\end{align*}

The function $F[kp'g(\beta)]$ involves the Fresnel integral and is referred to as the transition function. This function provides the mechanism for continuity at the incident shadow boundary and the reflection shadow boundary. Away from these boundaries this function is nearly unity. $\beta^+$ and $\beta^-$ relate to the angular separation between the observation point and the incident shadow boundary and reflection shadow boundary, respectively.

Multiplying the incident diffraction and reflected diffraction coefficients by the magnitude of the incident field produces the incident diffracted and reflected diffracted fields at the wedge surface. $V_b$ is commonly referred to as the total diffracted field at the wedge surface and can be written as:
\[ \mathbf{V}_b(p', \beta^\pm, n) = \mathbf{V}_b^i(p', \beta^-, n) - \mathbf{V}_b^r(p', \beta^+, n) \]  
(4-23)

where

\[ \mathbf{V}_b^i = \mathbf{D}^i(p', \beta^-, n) e^{-jkp' / \sqrt{\rho'}} \]  
(incident diffracted field)

\[ \mathbf{V}_b^r = \mathbf{D}^r(p', \beta^+, n) e^{-jkp' / \sqrt{\rho'}} \]  
(reflected diffracted field)

C. GTD Two-Dimensional Electric Conducting Wedge

1. Computer Software Development for the GTD model

The MLS azimuth antenna provides lateral guidance for an aircraft making a precision MLS approach. This antenna is generally sited 1,000 feet back or more from the stop end of the runway on the runway centerline extended (figure 4-3). For many runway profiles line-of-sight between the azimuth antenna and threshold is blocked. To ensure that MLS requirements, particularly those for automatic landings, will be met it is important to develop a model which can accurately determine the power density in the shadow region.

The profile of these humped runways can be approximated very well with a few straight line segments. The geometry of this problem provides a direct application for the canonical GTD wedge. Even for azimuth antennas which are offset from the runway centerline the angle of incidence at the hump would be only a few degrees.
MLS RUNWAY HUMP MODELING

Figure 4-3 MLS azimuth antenna location and humped runway profile.
Therefore, the case of normal incident edge diffraction is sufficiently accurate for this problem.

Recall that the geometry for the two-dimensional conducting wedge consists of either an infinite electric or magnetic line source at a distance \( \rho' \) and angle \( \phi' \) above the wedge surface (figure 4-2). The observation point \( p \) is at a distance \( \rho \) and an angle \( \phi \). The total field is the sum of the direct, reflected, and diffracted fields.

The direct field at the observation point is easily calculated. Let \( \rho_i \) be the distance from the source to the point of observation. The direct field is then given by:

\[
E_d^i = \begin{cases} 
  e^{-jk\rho_i/\sqrt{\rho_i}} & \text{for } 0<\phi<\pi+\phi' \\
  0 & \text{elsewhere}
\end{cases}
\]  

(4-26)

By applying image theory the reflected field at the observation point can be determined. The image source is at a distance \( \rho' \) and angle \(-\phi'\). Defining the distance from the image source to the point of observation as \( \rho_r \) the reflected field is given as:

\[
E_r^i = \Gamma e^{-jk\rho_r/\sqrt{\rho_r}} \quad \text{for } 0<\phi<\pi-\phi
\]

(4-27)

where \( \Gamma \) is the complex reflection coefficient. For a horizontally polarized source above an infinitely
surface the reflected signal is out of phase with the direct signal and the reflection coefficient is -1.0. For a vertically polarized source above an infinitely conducting surface the reflected signal is in phase and the reflection coefficient is +1.0. These two cases are referred to as ideal imaging.

The incident diffracted field at the point of observation is calculated by multiplying $v_b^i$ by the appropriate spreading factor and the phase term. (Recall that the total diffracted field is $v_b = v_b^i + v_b^r$)

$$E^{id} = D^i e^{-jk'\rho} e^{-jk\rho}/(\sqrt{\rho'}*\sqrt{\rho}) \quad (4-28)$$

Similarly the reflected diffracted field at the point of observation is given by,

$$E^{rd} = D^r e^{-jk'\rho} e^{-jk\rho}/(\sqrt{\rho'}*\sqrt{\rho}) \quad (4-29)$$

Equations 4-26, 4-27, 4-28, and 4-29 were coded in FORTRAN. A computer subroutine from reference 3 was also coded. This subroutine calculates the CF product of the C and F function of equations 4-13 and 4-14. Inputs to the program are:
\( p \): observation distance in wavelengths
\( p' \): source distance in wavelengths
\( \phi' \): angle between source and wedge surface
\( \text{an} \): wedge angle where \((2-\text{an})\pi=\text{wedge angle (radians)}\)
\( \text{freq} \): frequency
\( \text{npol} \): polarization
\( \text{xstrt} \): starting angle (degrees)
\( \text{xend} \): ending angle (degrees)
\( \text{xinc} \): increment (degrees)

The program will produce the following outputs:

- geometrical optics field (direct and reflected)
- incident diffracted field
- reflected diffracted field
- total field
- total field in dB

A program listing for a double precision version of the program, GTDDP FORTRAN, can be found in Appendix A.

After successfully compiling the FORTRAN program, two test simulations were performed. The first test was to ensure that the image source was properly coded. For a horizontally polarized source at a distance \( h \) above the ground plane the number of lobes in the radiation pattern is given by the equation (4):

\[
\text{# of lobes} = \frac{2h}{\lambda} \quad \text{(where } h \text{ is in wavelengths)} \quad (4-30)
\]

Figures 4-4 through 4-12 show the lobing pattern of a horizontally polarized source \( 37\lambda \) above the ground plane. This source height was used since it is representative of the azimuth antenna’s phase center height at typical MLS installations. 74 lobes were produced, matching the number
Figure 4-4 Lobing pattern of a horizontally polarized source 3\(\lambda\) above the ground plane, 0-10 degrees.
Figure 4-5 Lobing pattern of a horizontally polarized source 37λ above the ground plane, 10-20 degrees.
Figure 4-6 Lobing pattern of a horizontally polarized source \(37\lambda\) above the ground plane, 20-30 degrees.
CYLINDRICAL SOURCE

Figure 4-7 Lobing pattern of a horizontally polarized source 37λ above the ground plane, 30-40 degrees.
Cylindrical Source

Figure 4-8: lobing pattern of a horizontally polarized source 37λ above the ground plane, 40°-50° degrees.
Figure 4-9 Lobing pattern of a horizontally polarized source 37\(\lambda\) above the ground plane, 50-60 degrees.
Figure 4-10  Lobing pattern of a horizontally polarized source 37λ above the ground plane, 60-70 degrees.
Figure 4-11  Lobing pattern of a horizontally polarized source 37λ above the ground plane, 70-80 degrees.
Figure 4-12 Lobing pattern of a horizontally polarized source 37λ above the ground plane, 80-90 degrees.
predicted by equation 4-30. (Due to the number of lobes, detailed illustration of the lobing pattern with a polar plot was unpractical.)

The second simulation consisted of an electric line source above an infinitely conducting half-plane. The source is one wavelength from the wedge edge ($\rho'$) and at an incident angle of 30 degrees ($\phi'$). The output of the program is shown in figure 4-13. The output for the same simulation is shown in figure 11.32 of reference 3 and is reproduced here in figure 4-14. The plots in these two figures are identical and indicate that no logic errors exist in the program. In addition, there was good agreement between hand calculations and the computer generated output for several points. This simulation was repeated for a magnetic line source (figure 4-15).

2. Microwave Landing System Runway Hump Model

In the previous section it was shown that there were no logic errors in the FORTRAN code and that the model results matched published results. However, the case tested was ideal in that the wedge surface was infinitely conducting. The next step is to investigate what modifications are required, if any, to the model to account for microwave propagation over real earth. One method for
GTD WEDGE PROBLEM
Horizontal (SOFT) Polarization

- total
- geometrical optics
- incident diffracted
- reflected diffracted

half-plane n=2
source distance = 1
incidence angle = 30

Figure 4-13 Horizontally polarized source above a half-plane.
Figure 4-14 Horizontally polarized source above a half-plane from reference 3.
GTD WEDGE PROBLEM

Vertical (HARD) Polarization

- total
- geometrical optics
- incident diffracted
- reflected diffracted

half-plane \( n=2 \)

source distance = 1 \( \lambda \)

incidence angle = 30

Figure 4-15 Vertically polarized source above a half-plane.
determining the model's accuracy or need for modification is to compare its results to measured data.

To obtain measured data it is required that a humped runway facility be identified which is being serviced by MLS also. One such runway is Ottawa runway 14, Ottawa, Quebec, Canada (5). At this facility MLS power density measurements were taken by a helicopter flying approximately 8 feet above the runway surface (see figure 4-16 for flight path profile). However, the power density measurements made at Ottawa were uncalibrated. Therefore, when comparing model results to the Ottawa power density measurements the computed results should be adjusted to best fit this measured data close to the antenna. The Ottawa data is compared to model results in the later part of the section. Before presenting the comparison, the additional modifications made to the model to apply it to this problem will be discussed.

The first addition to the FORTRAN program was code to simulate the helicopter's flight path. Secondly, since the MLS antennas are spherical radiators the spreading factors previously used (cylindrical spreading) were changed to account for spherical spreading. Code to perform the power density calculation was added to the FORTRAN program also. Furthermore, the MLS azimuth antenna has a cutoff characteristic at the horizon: the sidelobes below the
RUNWAY PROFILE - OTTAWA RUNWAY 14

Figure 4-16 Runway profile for Ottawa runway 14.
horizon are at least 14 dB below the pattern maximum (5). This pattern can be approximated by the expression (figure 4-17):

\[ E_{\text{dir}} = 1 + 0.8 \tanh(A \cdot \theta') \]  
\text{(direct signal)}

\[ E_{\text{ref}} = 1 - 0.8 \tanh(A \cdot \theta) \]  
\text{(reflected signal)}

A subroutine was added to the model to account for the vertical pattern of this MLS azimuth antenna. This subroutine calculates a pattern coefficient to scale properly the magnitudes of the direct and reflected signals in the main program. This enhancement permits the model to be modified easily to account for different vertical patterns.

The computed results, for a vertically polarized source above an infinitely conducting ground plane, versus Ottawa measured data are shown in figure 4-18. Figure 4-18 shows that there is poor agreement between the measured data from Ottawa and the computed results using this approach. Since these results were unsatisfactory, no effort was made to best fit the computed curve close to the antenna.

Further investigation was required to determine what characteristics the reflected and diffracted signals have at microwave frequencies above real earth. The first step was
MLS RUNWAY HUMP MODELING

Figure 4-17 Geometry for MLS antenna pattern equations.
Figure 4-18 Computed results (vertically polarized source above an infinitely conducting ground plane) versus Ottawa measured data.
to characterize the reflection coefficient as a function of incidence angle. The equation

$$\Gamma_v = \frac{(\varepsilon'_r \cos \theta - j \sqrt{\varepsilon'_r - \sin^2 \theta})}{(\varepsilon'_r \cos \theta + j \sqrt{\varepsilon'_r - \sin^2 \theta})}$$  \hspace{1cm} (4-33)

where:

$$\varepsilon'_r = \varepsilon_r - j(\sigma/\omega \varepsilon_0)$$  \hspace{1cm} (4-34)

$\theta$ is measured from normal to surface

was used to calculate the reflection coefficient for vertical polarization at MLS frequencies for a few common ground plane materials (4). The magnitude and phase plots for these materials are shown in figures 4-19 through 4-26. It is apparent that at incidence angles near 90 degrees the reflection coefficient for all realizable ground planes approaches -1. Therefore, at MLS frequencies -1 is a good approximation for the reflection coefficient when the receiver is at low altitudes over the runway since this geometry produces large incidence angles.

Since in this case an infinitely conducting surface is a poor approximation for real earth, the question arises as to the proper sign to be used in equation 4-10. Like reflection, diffraction is a local phenomenon dependent upon the characteristics of the diffracting surface. In view of the behavior of the reflection coefficient at grazing angles, it follows that the proper form should be that of equation 4-11. To verify the use of equation 4-11 to
REFLECTION COEFFICIENT AT 5.05 GHz
FOR VERTICAL POLARIZATION OVER SEAWATER

Figure 4-19 Magnitude of reflection coefficient (5.05 GHz) for vertical polarization over sea water.
REFLECTION COEFFICIENT AT 5.05 GHZ
FOR VERTICAL POLARIZATION OVER SEA WATER

Figure 4-20 Phase of reflection coefficient (5.05 GHz) for vertical polarization over sea water.
Figure 4-21  Magnitude of reflection coefficient (5.05 GHz) for vertical polarization over fresh water.
Figure 4-22 Phase of reflection coefficient (5.05 GHz) for vertical polarization over fresh water.
Figure 4-23 Magnitude of reflection coefficient (5.05 GHz) for vertical polarization over ground.
REFLECTION COEFFICIENT AT 5.05 GHz
FOR VERTICAL POLARIZATION OVER SANDY SOIL

Figure 4-25 Magnitude of reflection coefficient (5.05 GHz) for vertical polarization over sandy soil.
REFLECTION COEFFICIENT AT 5.05 GHz

FOR VERTICAL POLARIZATION OVER SANDY SOIL

Figure 4-26 Phase of reflection coefficient (5.05 GHz) for vertical polarization over sandy soil.
calculate the diffracted field a simulation was performed using equation 4-12. The plot in figure 4-27 shows the magnitude of the electric field along the runway. The two discontinuities further confirm the use of equation 4-11 to calculate the diffracted field. The model results (using -1.0 for the reflection coefficient and equation 4-11 to calculate the diffracted field) versus measured data are shown in figure 4-28. The computed results were adjusted 8.5 dB watts/m² for best fit close to the antenna. Figure 4-28 shows that the computed results have the same trend as the measured data. In performing the measurements at Ottawa strong multipath from a hangar was noticed in the threshold region (6). Depending on the gain of the receiving antenna, the presence of this multipath may account for the difference between the model results and measured data in that region. Also, there is some subjectivity in selecting the wedge angle to be used when modeling the humped runway. This factor will influence the results slightly. However, as further experience is gained in runway hump modeling the intent would be to develop a methodology for determining the wedge angle so to improve or optimize the model’s accuracy. The listing for this model, GTDRUN FORTRAN, can be found in Appendix A.
Figure 4-27 Magnitude of electric field versus receiver distance from azimuth antenna. Diffracted field calculated with equation 4-12.
Figure 4-28  Predicted power density versus Ottawa measured data.  
(Diffracted fields calculated with equation 4-11).
3. Investigation of Azimuth Signal Attenuation Due to Humped Runways.

To provide insight into the effect of various humped runway geometries on the azimuth signal power density, three humped runway profiles were investigated. For each of these profiles two simulations were performed. The first simulation was performed to determine the power density at the lower limit of coverage at threshold. The second simulation was performed to investigate the power density that would exist in the normal approach-to-landing flight profile.

Figures 4-29, 4-30, and 4-31 show power density plots for three runway profiles. The geometry for the runway profiles is shown in figure 4-32. Each plot contains power density curves for MLS azimuth antenna phase center heights (PCH) of 1.83 m. (6 ft.), 5.49 m. (18 ft.), and 12.2 m. (40 ft.) along with the minimum power density requirement for the MLS azimuth data signal. Throughout the runway region, the minimum power density requirement must be met down to a height of 2.44 m. (8 ft.) above the runway surface. The power density plots indicate for humped runways, even with a small slope, the azimuth antenna may need to be elevated to achieve the minimum power density requirement.
POWER DENSITY PLOTS

runway slope=.2 degrees

Figure 4-29 Power density plots for a runway with a 0.2 degree down slope.
Figure 4-30 Power density plots for a runway with a 0.4 degree down slope.
Figure 4-31. Power density plots for a runway with a 0.6 degree down slope.
RUNWAY PROFILE FOR POWER DENSITY PLOTS

Figure 4-32  Geometry of runway profiles for power density plots given in figures 4-29, 4-30, and 4-31.
To provide further insight into the effects of humped runways on the MLS azimuth signal the receiver path was changed to provide a threshold crossing height (TCH) of 15.2 m. (50 ft.) (figure 4-33). This path is more representative of an approach aircraft's position while crossing threshold. Figure 4-34 contains the power density plots for an azimuth antenna with a PCH of 1.83 m. (6 ft.) for the three runway profiles modeled. As expected, these plots indicate an improvement in power density for an aircraft crossing threshold at 15.2 m. (50 ft.) when compared to the results obtained at 2.44 m (8 ft.).

Further conclusions and recommendations concerning this runway hump analysis will be presented in section VI.
Figure 4-33 Geometry of runway profiles for low approaches to humped runway with a 15.2 m (50 ft.) threshold crossing height.
POWER DENSITY PLOTS: PCH=1.83m. -6 ft.

low approach to hump with a 50 ft. TCH

- x runway slope=.2 degree
- o runway slope=.4 degree
- ▼ runway slope=.6 degree
- ■ MINIMUM LEVEL

Figure 4-34  Power density plots for low approaches to humped runway with a 15.2 m (50 ft.) threshold crossing height.
a. Introduction to the Instrument Landing System.

The ILS has been in service since 1943 (9). It supplies the pilot with elevation and azimuth information permitting precision approach capabilities. ILS's are classified by the quality of guidance they provide. These classifications, or categories, are characterized by their decision heights. The decision height is a point on the ILS approach, of a specified altitude, where the pilot's reliance on the instrument approach service terminates. Once at the decision height the pilot must execute the approach visually or perform approved missed approach procedures. The system is composed of a glide slope antenna, localizer antenna, and an outer, middle, and inner marker beacon, as required. For a facility that provides a localizer back course for departure or missed approach guidance, a back course marker beacon may be included with the system equipment. The system configuration is shown in figure 5-1. The localizer and glide slope signals are radiated in different frequency bands on paired channels.

The localizer provides lateral guidance for an aircraft making an ILS approach. It radiates an AM signal in the VHF band between 108 and 112 MHz (10). The localizer course is defined as the locus of points which forms a vertical plane where the depth of modulation of the
Figure 5-1  ILS system configuration.
The glide slope provides vertical guidance for an aircraft making an ILS approach. The glide slope radiates an AM signal in the UHF band between 329 and 335 MHz (10). The glide slope course, or glide path, is the locus of points, forming an inverted cone, where the depth of modulation of the 90 Hz and 150 Hz sidebands is equal. This cone originates at the base of the glide slope antenna mast. The glide path angle is typically 3 degrees in elevation, although glide path angles between 2.5 and 3.5 degrees are used (10). Again, the guidance information is obtained by detecting the DDM of the 90 Hz and 150 Hz sidebands. This information is displayed to the pilot by the glide slope CDI needle.

The marker beacons are used to indicate particular points on the instrument approach path. They operate at 75 MHz and transmit a keyed audio tone. The marker beacon
information is provided to the pilot in the form of an indicator light and an aural signal. Each type of marker beacon has a unique audio tone, frequency, and indicator color, and each installation has a unique keying code (10).

Above and below the glide path is the proportional guidance region. Similarly, there is a proportional guidance region to the left and the right of the localizer course (figures 5-2, 5-3). In this region the angular displacement of the aircraft from either the localizer course and/or the glide path is proportional to the respective needle displacement on the aircraft’s CDI. Outside the proportional guidance region is the clearance section. It is in this region that the appropriate needle of the CDI cross pointer is deflected full scale. For the localizer this indicates to the pilot a hard fly left or fly right is needed to enter the localizer proportional guidance region. Similarly, for the glide slope this indicates a hard fly up or fly down. The specifics of how these regions are electromagnetically established in space will be addressed in further detail in sections e. and f.

Figure 5-4 shows a simplified block diagram of the deflection circuit used in the ILS receiver. The input to the AM detector is held constant (±1 dB). The detector output is then fed to two bandpass filters, with center frequencies of 90 Hz and 150 Hz. The rectified audio
Figure 5-2  Glide slope proportional guidance and clearance sectors.
Figure 5-3  Localizer proportional guidance and clearance sectors.
SIMPLE DEFLECTION CIRCUIT

Figure 5-4  Deflection circuit for CDI needles.
voltages developed across $E_1$ and $E_2$ are proportional to the 90 Hz and 150 Hz modulation depths (11). Any difference in these two voltages will cause a current to flow through the CDI meter. This current is proportional to the difference in the depth of modulation (DDM) of the two audio tones (11). In the receiver there are two deflection circuits like the one described above, one circuit for the localizer portion of the cross pointer and the second for the glide slope portion.

When the aircraft deviates off either the glide slope or localizer course, the magnitude of the detected audio tones is no longer equal. Thus, $E_1$ is not equal to $E_2$ causing current to pass through the CDI meter, driving the respective needle from the center location. The polarity is such that the needle displacement is in the direction the pilot must fly to correct the course deviation.

b. Present Limitations of Mathematical Models

At many airdromes where ILS service is desired the environment is usually electromagnetically clotteded. At these airdromes it would be advantageous to have the capability to accurately predict ILS performance. The previous statement applies to most navigational aids to some extent also. A useful tool for determining the effect
of re-radiating object on a navigation aid is a validated mathematical model. Such a model has been developed by the Avionics Engineering Center of Ohio University for the ILS.

The ILS model mentioned above has been successfully validated with an extensive amount of flight data (12, 13, 14). However, the cases studied consisted of electrically large objects, usually in the far-field. The theoretical basis for this model is physical optics (with a subsequent addition to include GTD techniques). Several simulations were performed with this model, by personnel outside of Ohio University, to predict the effects of the azimuth antenna structure on the ILS localizer signal (15). As expected, the modeled results did not compare well with the measured data and the reason for this is discussed in the following text.

Before writing an electromagnetic code or implementing an available mathematical model one should consider the nature and geometries of the problem that is to be investigated. Items such as the electrical size of the scattering object in comparison to the source wavelength will give direction to the appropriate types of models. Will the scatterer be in the near or far field? Even the point of observation or region of interest will dictate which methods can be used. Careful consideration at the beginning will save time and effort.
As stated earlier, GTD is an extension of ray optics which includes diffracted rays. Because of its ease of use it would be a first consideration. However, it is desired in the case of siting the MLS azimuth station to place the station as close as possible to the ILS localizer. Similarly, in the collocated siting of the MLS elevation station with the ILS glide slope it is desired to investigate the effects of siting the elevation station directly in front of the glide slope. This means that the scatterer for most locations will be in the near field. In this case, the incident field on the scatterer is not ray optic and GTD is therefore ruled out as a possible method.

As previously mentioned, a physical optics model was considered, but proved to be inadequate. This follows since a physical optics model will provide accurate results in a non-shadowed region as long as the assumed current of $2n \times H_{\text{inc}}$ across the illuminated side of the object is representative of the actual current induced on the scatterer. That is the current on the illuminated side of the object is set equal to twice the magnitude of the incident magnetic field, $2n \times H_{\text{inc}}$. Furthermore, the current on the non-illuminated sides of the object are assumed to be zero.
To investigate the validity of the two previous assumptions, an exact method for determining the induced surface current on an object is needed. One technique having this capability is the Method of Moments. This technique was implemented to investigate the accuracy of the physical optics assumptions for a simple scatterer having approximately the same electrical size as the MLS equipment at the ILS wavelengths.

The results of a Moment Method model for an infinitely long square cylinder are shown in figure 5-5. Two cases were modeled. These two cases bound the minimum and maximum width and depth dimensions of the MLS equipment at the respective ILS wavelength. Specifically, at the ILS localizer wavelength, \( k\alpha = 5.0 \) and \( k\alpha = 1.0 \) \((k\) is the wave number) corresponds to azimuth equipment dimensions of 2\(\alpha = 15\) feet and 2\(\alpha = 3\) feet, respectively. Similarly, at the ILS glide slope wavelength the two cases correspond to MLS elevation antenna dimensions of 1 and 5 feet. The reader may refer ahead to figures 5-19 and 5-27 for the MLS equipment dimensions.

For both cases a transverse-magnetic wave is normally incident one side of the box. Note that for both cases the current across the illuminated side of the box, predicted by the Moment Method model, is not the postulated \( 2\pi \alpha H_{inc} \). Observe as the wavelength of the source is increased.
Figure 5.5: Moment Method results: current induced on a square cylinder due to a normally incident transverse-magnetic wave.
(ka=1.0) the postulated physical optics current is a poor approximation of the actual current on the illuminated side. More important, however, is the fact the actual currents on the non-illuminated sides of the box are non-zero and of sufficient magnitude to warrant consideration. With a physical optics model the currents on the non-illuminated sides are assumed to be zero. This assumption results in an inability to predict a non-zero diffracted field making a physical optics model inappropriate for this problem.

This leaves the Moment Method as the only feasible modeling technique for this problem. This being the case, some discussion will be devoted to the Moment Method.

Many electromagnetic radiation problems can be represented by an integral equation of the general form (4):

$$\int I(z')K(z,z')dz'=-E^i(z)$$

(5-1)

$I(z')$ is the unknown current induced on the surface of the scatterer by the incident E-field $E^i$. $K(z,z')$ is the appropriate Green's function. Once the surface current on the scatterer is known the fields everywhere in space, external to the scatterer, can be calculated.
The unknown current may be represented by an expansion function $F_n$ having the form:

$$I(z') = \sum_{n=1}^{N} I_n F_n(z')$$  \hspace{1cm} (5-2)

where $I_n$'s are complex expansion coefficients and

$$F_n(z') = \begin{cases} 
1 & \text{for } z' \text{ in } \Delta z'_n \\
0 & \text{otherwise} 
\end{cases}$$  \hspace{1cm} (5-3)

Substituting (5-2) into (5-1) and using (5-3) gives:

$$\sum_{n=1}^{N} I_n \int_{\Delta z'} K(z_m, z') \, dz' = -E^i_z(z_m)$$  \hspace{1cm} (5-4)

for convenience let:

$$f(z_m, z'_n) = \int_{\Delta z'} K(z_m, z') \, dz'$$  \hspace{1cm} (5-5)

Enforcing the integral equation at $N$ discrete points, or segments, on the scatterer surface will result in $N$ simultaneous linear equations (4) which can be represented in matrix form as:
or in matrix notation:

$$\begin{bmatrix}
  f(z_1, z_1') & f(z_1, z_2') & \ldots & f(z_1, z_n') \\
f(z_2, z_1') & f(z_2, z_2') & \ldots & f(z_2, z_n') \\
  \vdots & \vdots & \ddots & \vdots \\
f(z_n, z_1') & f(z_n, z_2') & \ldots & f(z_n, z_n')
\end{bmatrix}
\begin{bmatrix}
  I_1 \\
  I_2 \\
  \vdots \\
  I_n
\end{bmatrix}
= 
\begin{bmatrix}
  -E_{iz}(z_1) \\
  -E_{iz}(z_2) \\
  \vdots \\
  -E_{iz}(z_n)
\end{bmatrix}
$$

(5-6)

The number of segments chosen must be large enough so that the discrete current calculated by the integral equation for one segment does not vary greatly with those surrounding it. Typically the scatterer should be broken into segments that are smaller than a tenth of a wavelength. Once the code is developed and executable the number of segments should be increased to check for convergence. Once the result converges there is no need to continue to increase N.

Unfortunately, prewritten Moment Method software with the ability to model readily the geometry imposed by the collocated siting of ILS and MLS equipment is not available. In addition, the derivation, coding, and model verification would take a considerable length of time, and funding. This level of effort would be beyond the scope of the task intended by the sponsor.
Furthermore, a computational consideration is the amount of storage space available to store the $N^2$ elements. For this particular application $N$ would have a minimum average value of 100. This number is achieved by realizing that each face of the MLS equipment must be divided into segment that are no larger than 1/10 of a wavelength long at the appropriate ILS source wavelength. For example, the MLS elevation system, (elevation antenna enclosure, electronics cabinet, and battery backup box: figure 5-27) would be modeled by 15 faces whose dimensions on the average are greater than one wavelength at the ILS glide slope frequency. Thus, a minimum of 150 segments would be required. Similar analysis of the azimuth system structure (figure 5-27) yields about 70 segments as a minimum value. Therefore, on the average a minimum of 100 segments would be required. For single precision real numbers this would require approximately 40K bytes of storage for the $N^2$ segment values. In addition, storage would also be required to invert the $N \times N$ matrix.

The discussion in the previous paragraph is limited to illumination of the MLS equipment structures by a single source. Most modern ILS glide slope antennas consist of six to nine individual dipoles (two or three sets of three) and the ILS localizer may consist of an eight to fourteen element array. Therefore, the storage requirement for illumination by a single source may need to be multiplied
six to fourteen times depending on the type of localizer or glide slope modeled. (This multiplication of six to fourteen can be avoided if the element factor and array factor of the antenna of interest is known and only far-field observation were of interest. However, this is not the case for the problem present by MLS-ILS collocation)

Recall, the number of segments, \( N \), must be increased until convergence of the results is obtained and that this process is required to ensure that \( N \) is large enough to provide accurate results. Increasing the number of segments to just two or three times the minimal required \( N \) (100) greatly increases the storage requirements. Furthermore, it is not certain that increasing \( N \) by two or three times its minimum value would be sufficient enough to show convergence of the results.

Finally, additional storage space is also required to perform the Sommerfeld integration to account for propagation over real earth and to handle the radiation integral which is required to construct the total field. Also, for propagation over a ground plane the six to fourteen antenna elements discussed earlier would have six to fourteen image antennas to account for ground reflections. Given the previous discussions, one can see that this problem quickly grows in complexity. However, still this problem could feasibly be handled with a large
mainframe or supercomputer, provided such a resource were available, and one was given sufficient time and funding.

Aside from the above considerations related to the Moment Method technique, the incident near fields of the various glide slope and localizer systems are not readily available. Before a Moment Method model could be developed and used this knowledge would have to be obtained. Unfortunately, this problem further reduces the ability to readily develop a Moment Method model. Since the results were needed by the sponsor in a very short time frame (four to six months) this approach would have been much too time consuming.

c. Advantages of Flight Measurements

Now that the possibility of readily using a mathematical model has been exhausted, one must consider performing flight measurements to determine and quantify the effects due to the physical presence of the MLS equipment on the ILS guidance signals.

The first advantage is that flight measurements have been performed to evaluate the performance of ILS systems for years. Much experience has been gained in performing these measurements by Avionics Engineering Center
personnel. In addition, the accuracies and limitations of this method are well understood, and the flight measurement process has been refined. Even if computer predictions were available, the flight measurement results obtained by a Federal Aviation Administration flight check would be the deciding factor in the recommissioning of an ILS facility after the MLS equipment was installed.

Given the considerations discussed in the previous section, for this application the information needed can be obtained in a shorter amount of time than would be possible if a computer modeling approach were used. Again, this is an important consideration since these results were needed in a timely manner to satisfy contractual obligations. Also, this information can be used to validate a computer model at a later date when the capability for a Moment Method solution is possible.

d. Flight Measurement Process

1. Flight Measurement System

The Ohio University flight measurement system consists of two groups of equipment, airborne and ground-based. Performing ILS flight measurements requires a three man crew composed of a pilot, panel operator, and
theodolite operator. The aircraft used to perform the flight measurements must be capable of carrying the pilot, panel operator, and airborne equipment. In addition, the aircraft must also be able to supply the additional power required for the airborne equipment. However, for cost and time effectiveness the aircraft should also be able to transport the theodolite operator and ground equipment to the flight measurement location. A Bonanza A-36 was used to perform the flight measurements for this investigation.

The aircraft must be equipped with an ILS receiver so the pilot can fly the ILS approach. It must also have an equipment pallet to restrain the airborne measurement equipment while in flight. The airborne equipment is composed of two 2-channel chart recorders, and a 15 channel ILS mini-laboratory, which was fabricated at the Avionics Engineering Center, Ohio University. In addition to providing the input to drive the chart recorders it is currently being modified to provide digital tape storage of the flight measurement data.

The ground-based measurement equipment is composed of a Warren Knight Theodolite and a Reaction Instruments, Inc. model 6050 UHF Theodolite Telemetric Transmitter, more commonly referred to as a radio telemetry transmitter (RTT). The theodolite axes are equipped with decoders which supply angular information to the RTT. The RTT,
being properly calibrated and referenced, transmits a signal whose DDM corresponds to the angular location the theodolite is "looking". The DDM signal from the RTT and the DDM received by the aircraft are inputs to a differential amplifier. It is the difference in the two DDM signals that represents the error in the ILS system (This error is actually recorded in microamperes(µAs) and not DDM). This error is recorded by a chart recorder and the resulting trace is referred to as the differential amplifier trace. Also, a tone, generated by a foot switch, can be transmitted to the aircraft to mark a specific event in the measurement process. In addition to the measurement equipment there are two communication transmitters included in the ground equipment. One transmitter is used for ground-to-air communication and the second one is a back-up which can be used to monitor tower-to-aircraft communications.

2. Aircraft Flight Check Patterns

Four distinct flight paths or patterns are flown by the aircraft to characterize the electromagnetic field radiated by the ILS antennas. The specific pattern flown depends on which antenna system is being investigated and what type of information is needed. The analysis of the
flight recordings produced from these flight check patterns will be addressed in section e.

A pattern "A" is flown to obtain information on the ILS course structure. This pattern provides a measure of the amount of roughness or noise present in the guidance signal, characterizing the flyability of the course. This pattern is performed for both localizer and glide slope structure measurements. This flight pattern starts at least 1 to 2 miles outside the outer marker beacon (figure 5-6). The aircraft generally makes a 3 degree approach on centerline, however, there are occasions when this pattern is flown to the right or left of the localizer course, or above or below the glide path. When the antenna system of interest is the glide slope this pattern terminates once the aircraft is past the glide slope antenna. However, for localizer measurements the pattern extends to the stop end of the runway.

A pattern "B" is flown for glide slope measurements. The information obtained is the width and symmetry of the ILS glide slope course. The flight path is flown at a constant altitude; this altitude is chosen so that the aircraft will pass though elevation angles between 1.9 and 4.1 degrees (figure 5-7).
Figure 5-6 Flight check pattern A.
Figure 5-7  Flight check pattern B.
A pattern "C" is usually performed for localizer width and symmetry measurements. The flight path is a perpendicular cut, at a constant altitude, across runway centerline at the outer marker as shown in figure 5-8. In some cases a pattern "C" is performed to investigate the three dimensional characteristic of the glide path, referred to as tilt.

A pattern "D" is performed if localizer clearance information is needed in addition to width and symmetry. This pattern is shown in figure 5-9.

3. Flight Measurement Process

Once the theodolite is sited at the facility it is carefully leveled and aligned. Then the total measurement system is fully calibrated by reference to a IFR 401-L signal generator, traceable to the National Bureau of Standards. Periodically, the flight measurement equipment is recalibrated to reduce system errors due to environmental changes. The pilot will then proceed to fly the pattern required to obtain the data of interest. The measurement system circuitry is very sensitive. The keying of a communications transmitter can induce noise in the recording system and if not noted this may be interpreted as a course perturbation. It is for this reason that every
Figure 5-8  Flight check pattern C.
Figure 5-9 Flight check pattern D.
attempt is made to minimize air-to-ground communication when a flight recording is in process. Currently a digital system is being developed, but it was not ready for actual field measurement work.

The telemetry unit is used in performing pattern "A" measurements. The telemetry unit continuously transmits angular information provided by the theodolite to the aircraft. This information is processed with the information provided by the ILS. The cross hairs of the theodolite must be kept on the aircraft's ILS antenna at all times. Since the ILS is an angular system, this is very important at close range (within a mile from threshold) since a small theodolite tracking error will produce significant errors in the recording. In addition, the ILS tolerances become more stringent close in for higher category facilities. For any particular structure measurement at least two pattern "A"s are performed. The chart recordings must repeat within ±2 or 3 μAs to confirm the measured data.

For pattern "B" measurements the theodolite is positioned so that it is at an elevation angle of 2 degrees. When the aircraft passes through the horizontal cross hair of the theodolite a double tone is generated by the theodolite operator to signify the start of the measurement and identify the elevation angle as 2 degrees.
This places two event marks, generated by the panel operator, on the chart recording. The theodolite is then incremented .1 degree in elevation. Again as the aircraft passes through the horizontal cross hair a single tone is generated signifying 2.1 degrees. This process is repeated and each time the aircraft passes through the horizontal cross hair a tone is generated. The 2.0, 3.0, and 4.0 degree elevations are denoted with double tones. This type of measurement requires the theodolite operator to exercise great skill since there is only one second between event marks.

The procedure for pattern "C" and "D" measurements is similar to that of the pattern "B" only the theodolite is incremented in the lateral plane.

**e. Instrument Landing System Tolerances**

Figure 5-10 illustrates the location of the various ILS points and zones in relation to the runway. Figure 5-11 shows the primary ILS glide slope tolerances as they would appear on the chart recording (produced by a pattern A). The tolerances shown are those germane to the analysis of the chart recordings produced for the MLS-ILS collocation work. Additional tolerances do exist and can be found in reference 16. For ease of illustration the
Figure 5-10 ILS points and zones.
Figure 5-11 Primary ILS glide slope tolerance, graphical representation on flight recording from a pattern A.
course (differential amplifier trace) is shown as ideal in that there is no noise.

The first step in the analysis of a flight recording is to draw in the ILS points A, B, C, and threshold. There are mile marks along the top of the chart recording, not shown here, to provide the scaling. Then the second step is to draw in and label the ILS zones. In Zone 2 a graphical straight line average of the course is drawn. Then, in Zone 3 a graphical average course is drawn. In zone 3 the average course is permitted to curve as long as it remains inside of the alignment brackets. In this ideal case the average course is a straight line at the 0 μA level. Finally the structure tolerances are drawn in. For a category I (CAT I) glide slope the structure tolerance is ±30 μA about the average course in all zones. Also note that zone three terminates at ILS point C and not threshold for CAT I glide slopes. This termination at point C is due to the decision height point for CAT I facilities. For CAT II and III facilities the tolerances are as shown.

In the presence of noise the average course may be biased from the 0 μA reference. This will also bias the average path and the structure tolerance limits, but not the alignment brackets.
For analysis of a localizer pattern "A" the procedure is the same but the localizer tolerances are applied. An example of an ideal case is shown in figure 5-12. Again the presence of noise may bias the average course and the structure tolerance limits.

For some positions of the MLS elevation mock-up an additional ILS tolerance had to be applied to the pattern "A" recordings. This tolerance is course reversal. The average course is allowed to bend or have a slope in it and there is no tolerance on the slope as long as the course remains within the alignment brackets. However, the rate at which the slope of the course changes is limited by the reversal tolerance. The limit for the reversal is 25 \( \mu \text{A} \). Due to its complex nature, the analysis of a course reversal is not addressed in this document but can be found along with the other tolerances discussed above in reference (16).

Patterns B, C, and D produce chart recordings that have a crossover as shown in figure 5-13. For localizer recordings, patterns C and D, the 0\( \mu \text{A} \) and the \( \pm 150 \mu \text{A} \) points are located. Then vertical lines are drawn from these points up to the event marks. The angles these points represent can be interpolated from the event marks. Once these three angles are known the width and symmetry of the localizer course can be determined. The procedure for
Figure 5-13 Width and symmetry analysis of a crossover pattern produced by a pattern B, C, or D.
the glide slope is identical except the 0 µA and the ±75 µA points are used.

f. Localizer Flight Measurement Results

1. Establishment of Localizer Course

In order to determine which regions in front of the localizer array will be least sensitive to the placement of the MLS azimuth station it is necessary to understand the radiation characteristics of the localizer. The radiation patterns for different types of localizer arrays vary in directivity and strength of the clearance signal. However, these radiation patterns have similar characteristics in the proportional guidance region.

The localizer radiates a carrier and sideband (CSB) signal and a sideband-only signal (SBO). The CSB radiation pattern (figure 5-14) has its maximum on centerline. The sidebands of the CSB signal are in phase with each other and of equal magnitude. These sidebands are also in phase with the carrier and the magnitude relationship of the sidebands to the carrier is fixed.

The radiation pattern of the SBO energy has a null on centerline (figure 5-15). On centerline the magnitude of
the 90 Hz modulation from the two lobes is equal but out of phase by 180 degrees. This is also characteristic of the 150 Hz modulation. It should be noted that the phase relationship between the 90 Hz and 150 Hz modulation in the same lobe is 180 degrees.

Superimposing the SBO and CSB patterns produces a composite pattern as illustrated in figure 5-16. Note that there is maximum carrier power provided in the approach direction. In addition, there are two modulation lobes formed as shown. The signal strengths of the 90 Hz and 150 Hz modulations are equal on centerline. Thus the difference in the DDM on centerline is zero. Off centerline either the 90 Hz or the 150 Hz modulation dominates and the DDM is non-zero.

2. Symmetry Criteria

DDM is given by:

$$\text{DDM} = |M_{150}| - |M_{90}|$$

Furthermore $M_{150}$ and $M_{90}$ can be expressed as:

$$M_{150} = M_{150}/E_c \quad \text{and} \quad M_{90} = M_{90}/E_c$$

where

$E_c$ = field strength of the carrier

$M_{150}$ = total field strength of the 150 Hz sidebands

$M_{90}$ = total field strength of the 90 Hz sidebands
Since the total 90 Hz and 150 Hz sideband powers are vector sums of the sidebands from the CSB and SBO signals, it is important that the phase relationships between the field components remain unchanged by the scatterer. In addition, the DDM depends on the magnitude relationship of the carrier to the total sideband energy. Therefore, any attenuation caused by the scatterer must affect the carrier and sideband magnitudes similarly.

Placing a symmetrical scatterer symmetrically about the phase center of the localizer should produce the least effect on the localizer course. This assumes that the scatterer is not large enough to attenuate severely the localizer signal in the approach direction due to aperture blockage. For this geometry the incidence angle and source-to-scatterer distances for opposing localizer element pairs are equal (figure 5-17). Therefore, the diffracted field is symmetrical about the centerline, and the magnitude and phase relationships between opposing localizer pairs are preserved. The proper phase and magnitude characteristics of the composite field are preserved and the localizer course is theoretically unaffected.

For a symmetrical scatterer positioned offset from centerline the source-to-scatterer distances and angles of incidence for opposing localizer pairs are not equal. Thus
Figure 5-17: Geometry for symmetrical placement of scatterer.
the phase and magnitude relationships are altered. This alteration can have a significant effect on the localizer course. This effect will decrease as the scatterer is moved so far off centerline that it is not significantly illuminated by localizer energy.

At many airport facilities real estate in front of the localizer is limited. For these facilities it will be necessary to place the azimuth station very close to the localizer. Even if symmetry is maintained, near field effects due to the scatterer can significantly alter the localizer course. Furthermore, if the scatterer is close enough to the localizer it may couple with it changing the radiation characteristics of the array. Again the relationships required to establish the RF lobes will not exist.

The scatterer in this study is a mock-up of the contractor’s MLS azimuth station. The mock-up itself is not symmetrical. This type of structure will cause some effect on the course even when sited symmetrically with respect to the localizer. Flight measurements will be required to determine the magnitude of this effect for different distances ahead of the localizer. In addition, the sensitivity of the localizer signal to offset siting of the MLS azimuth station must be quantified. To obtain this
data some of the flight measurements must be made with the mock-up at offset positions.

3. Flight Measurement Results

Early work was performed at Airborne Express Airpark, Wilmington, Ohio, to verify the symmetry hypothesis. In performing these flight measurements only the antenna portion of the MLS mock-up was used (figure 5-18). The localizer was a 14-element log periodic dipole array. Three locations were chosen for placement of the antenna, two on centerline and one offset 21 feet. The centerline positions produce virtually no effect. The offset position had significant effect on the localizer course placing the system out-of-tolerance with a 22 uA course alignment shift [17]. A second set of flight measurements was later performed to further characterize this sensitivity to asymmetrical siting [18].

The antenna mock-up used for the initial measurements was constructed of 2x4's and metal sheeting requiring six to eight men to move. Since this manpower was not always available and the time required to make the flight measurements resulted in an inefficient use of this manpower, it was necessary to construct a lighter mock-up. To reduce the weight, a 3-dimensional mock-up of the MLS
METAL FACING ON PLYWOOD SURFACE

Figure 5-18 MLS azimuth antenna mock-up.
azimuth station (azimuth antenna, angle equipment, DME electronics) was constructed of PVC pipe and covered with 1 inch chicken wire (figure 5-19). This mock-up was used in subsequent tests. To reduce further the manpower required to move it the mock-up was placed on a sled. This modification permitted the mock-up to be pulled by a van for long distances (more than 25 feet) or moved by two men for shorter distances.

A test grid (figure 5-20) was developed to measure the effects of the mock-up placed on centerline and for offsets up to 50 ft. This grid was used in collecting flight data at Dayton International Airport, Dayton, Ohio, and Tamiami Airport, Miami, Florida [19,20]. At these two facilities the effects of the azimuth mock-up on the GRN-27 and the eight element V-ring localizers, respectively, were measured. At each facility the increment for the individual rows was determined by the rate of change in the flight data. Since the azimuth mock-up is non-symmetrical, flight measurements were performed at Tamiami Airport for offsets to either side of the localizer course centerline.

Flight measurements show that placement of the mock-up on the localizer centerline produced the least perturbation. Furthermore, the characteristics of the effects caused by the MLS mock-up placed in front of the localizer correlate well with the theory. Table 5-1
Figure 5-19 1 degree MLS azimuth mock-up.
summarizes the data collected for the three types of localizers. This table is composed of two headings. The data under the centerline heading shows the effect on the localizer course for the MLS azimuth station mock-up located on the localizer course centerline for placements of 50 and 100 feet ahead of the localizer array. The offset heading contains the data for the offset positions which produced the maximum disturbances on the localizer course for distances ahead of the localizer of 50 and 100 ft. The numbers shown in parentheses are the offsets from the localizer course centerline where the MLS mock-up produced the effect.

In closing it should be stated that the localizer/azimuth collocation work presented was limited to siting the azimuth system ahead of the localizer array. Furthermore, this work was limited to investigating the effects due to the physical presence of the azimuth equipment on the localizer front course. Additional work will be required to complete the localizer/azimuth collocation criteria. Further discussion of this work is presented in section VI.
<table>
<thead>
<tr>
<th>SYSTEM</th>
<th>CENTERLINE</th>
<th>OFFSET Pilot's Left/Right on Approach</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>@50'</td>
<td>@100'</td>
</tr>
<tr>
<td>GRN-27</td>
<td>10 µA</td>
<td>1 µA</td>
</tr>
<tr>
<td>V-RING (8)</td>
<td>9 µA</td>
<td>5 µA</td>
</tr>
<tr>
<td>LPD (14)*</td>
<td>1 µA</td>
<td>1 µA</td>
</tr>
</tbody>
</table>

* Effects of MLS azimuth antenna only.

Table 5-1  Localizer results: Course shift in presence of MLS azimuth mock-up.
g. Glide Slope Flight Measurement Results

1. Establishment of the Glide Slope Course

As with the localizer, the glide slope course is established through RF lobe formation (21). In the case of the glide slope the lobes are vertical. The main difference is that the glide slope system uses the ground in front of the glide slope antenna to establish the signal in space. The glide slope can be analyzed by removing the ground plane and supplying the proper image antennas.

The CSB and SBO patterns for a glide slope system are shown in figures 5-21 and 5-22. The CSB and SBO patterns vary for the three types of systems, but for purposes of discussion generalized patterns will be used. The superposition of the CSB and SBO patterns will form two lobes which are predominantly AM modulated either by 90 Hz or 150 Hz (figure 5-23). For a detailed description of the individual glide slope types the reader may refer to reference (21).
Figure 5-22  Idealized glide slope upper antenna pattern, sideband-only.
Figure 5-23 Idealized glide slope composite pattern of lower and upper antenna.
2. Fresnel Zone Migration

As previously stated, the glide slope depends on the ground plane in front of it to establish the path in space. The Fresnel zone represents the portion of the ground plane which is significant in forming the reflected (image) signal required to establish the glide path. The Fresnel zone is elliptical in shape and lies on the line between the glide slope antenna and the airborne receiver. As the receiver moves along the approach path, the Fresnel zone will migrate so that it remains on the line between the receiver and the glide slope antenna mast (figure 5-24).

If the effects of the scatterer in the ground plane region are to be minimized the scatterer must avoid penetrating the region though which the Fresnel zone migrates. In general this indicates that the MLS elevation station should be sited on the runway side of the diagonal between the glide slope antenna mast and the threshold centerline. Although this siting configuration will produce the least disturbance of the glide slope course, flight measurements are needed to determine the magnitude of the affect.

In addition, some glide slope facilities are offset only 250 feet from centerline. Clearance obstruction criteria prohibits the siting of the MLS elevation station
Figure 5-24  Fresnel zone migration.
any closer than 255 feet from runway centerline. In this case it would be desirable to know if the MLS elevation station could be sited directly in front of the glide slope antenna. Although this would cause obstruction of the ground plane and blockage of the antenna aperture, flight measurements were performed to determine if the effects due to this siting configuration are severe enough to prohibit its consideration.

3. Flight Measurement Results

Preliminary flight measurements were performed at Airborne Express Airpark, Wilmington, Ohio. This set of measurements was performed at the same time as the initial localizer work at Wilmington. In performing these initial measurements only the antenna portion of the MLS elevation station was used (figure 5-25). Three positions were chosen for placement of the MLS elevation antenna. The location of these positions relative to the capture effect glide slope (Category II) is shown in figure 5-26. Although none of the positions resulted in a glide slope out-of-tolerance condition, the most favorable placement of the MLS elevation antenna is in the region about position one.
Figure 5-25  MLS elevation antenna mock-up.
Figure 5-26  MLS elevation antenna positions.
To simulate the MLS elevation station (antenna, electronics cabinets, AC power panel) a three dimensional mock-up was constructed of PVC pipe and 1 inch chicken wire (figure 5-27). This mock-up was used in flight measurements made at Tamiami International Airport (TMB), Miami, Florida. At this facility flight measurements where performed on three types of glide slope systems: the null reference glide slope, the capture effect glide slope, and the sideband reference glide slope [15].

In performing the flight measurements at Tamiami the effects of penetrating the Fresnel zone with the MLS elevation station mock-up became more pronounced. Table 5-2 contains the flight data for the sideband reference glide slope. The data collected for this system show it to be the most sensitive to placement of the MLS elevation mock-up. Figure 5-28 shows the location of each mock-up position relative to the glide slope antenna mast. The diagonal between threshold and the glide slope antenna mast intersects the 1 and 1 1/2 degree MLS elevation station rows at offsets of 332 and 348 feet, respectively. The flight data show that for offsets greater than 350 feet the effects of the elevation mock-up are severe. Offsets of less than 350 feet produced a lesser effect, exhibiting a decrescendo as the offset approaches the minimal 255 feet offset. These results correlate well with the Fresnel zone analysis.
Figure 5-27 MLS elevation mock-up.
Table 5-2 Structure data, sideband reference glide slope.

<table>
<thead>
<tr>
<th>POSITION</th>
<th>ZONE 2</th>
<th>ZONE 3</th>
<th>CAT</th>
<th>COMMENTS</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 1/2°</td>
<td>µA</td>
<td>% Tol.</td>
<td>µA</td>
<td>% Tol.</td>
</tr>
<tr>
<td>1 (450°)</td>
<td>&gt;20</td>
<td>&gt;100</td>
<td>&gt;30</td>
<td>&gt;100</td>
</tr>
<tr>
<td>2 (425°)</td>
<td>44</td>
<td>220</td>
<td>&gt;30</td>
<td>&gt;100</td>
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<td>3 (400°)</td>
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<td>29</td>
<td>145</td>
</tr>
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<td>5 (350°)</td>
<td>8</td>
<td>40</td>
<td>12</td>
<td>60</td>
</tr>
<tr>
<td>6 (255°)</td>
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<td>50</td>
<td>8</td>
<td>40</td>
</tr>
<tr>
<td>1°</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>1 (450°)</td>
<td>31</td>
<td>155</td>
<td>26</td>
<td>130</td>
</tr>
<tr>
<td>2 (400°)</td>
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<td>145</td>
<td>28</td>
<td>140</td>
</tr>
<tr>
<td>3 (375°)</td>
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<td>---</td>
<td>---</td>
<td>---</td>
</tr>
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<td>13</td>
<td>65</td>
</tr>
<tr>
<td>5 (325°)</td>
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<td>---</td>
</tr>
<tr>
<td>6 (255°)</td>
<td>6</td>
<td>30</td>
<td>8</td>
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<td>60</td>
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<td>50</td>
</tr>
</tbody>
</table>
Figure 5-28 MLS elevation mock-up positions.
In closing it should be pointed out that the glide slope/elevation antenna collocation work presented was limited to developing siting criteria for siting the MLS elevation antenna between the glide slope antenna and the runway. Consideration has been given to siting the elevation antenna outboard of the glide slope antenna and this work is scheduled for the future.
VI CONCLUSIONS AND RECOMMENDATIONS

a. MLS Runway Hump Model

To aid in siting the Microwave Landing System (MLS) azimuth antenna at humped runway facilities it is desirable to have a mathematical model which can provide an accurate estimate of the power density in the shadowed region of the runway. Such a model could be used in determining the azimuth antenna phase center height required to satisfy the MLS power density requirement in this shadowed region.

It was demonstrated that the Geometrical Theory of Diffraction (GTD) could be used to handle the runway hump problem. Specifically, it was shown that the GTD diffracting wedge model could be implemented readily to the runway hump problem after a few modifications. The modifications were required to account for the vertical pattern of the MLS azimuth antenna and the characteristics of microwave propagation over real (non-ideal) earth.

Through comparison to measured data the GTD runway hump model presented was shown to predict, with reasonable accuracy, the power density trends for the humped runway scenario. Since the measured data available were uncalibrated, the ability of this model to predict absolute power densities could not be validated. In addition, since
the measured data were collected by a helicopter there are position uncertainties which influence the precision of the measured data, due to the very short source wavelength.

It is recommended that this model be validated against calibrated measurements. In addition, the measured data should be collected at surveyed points in order to determine the model's accuracy with reasonable precision. However, in view of the accuracies achieved in other applications using these same GTD wedge diffraction coefficients (e.g. the power pattern for a \( \lambda/4 \) monopole above a finite square ground plane [3]) the power density predictions should have an accuracy of at least \( \pm 3 \) dB.

In addition, it is recommended that consideration be given to developing a methodology for determining the wedge angle to be used in modeling a humped runway. The goal of this methodology would be to provide guidance in determining a wedge angle for a given runway profile so to improve the accuracy of the simulation results.

To provide insight on the effects of a humped runway on the MLS azimuth signal three runway profiles were modeled. These simulations were performed to investigate the ability to meet the minimum MLS power density requirement. Power density predictions from the GTD model indicate that the MLS azimuth antenna may need to be
elevated, from the minimum phase center height, to meet the minimum power density requirement down to a height of 2.44 m. (8 ft.) at threshold, when sited at a facility with a humped runway.

Additional simulations were performed to investigate the effects of a humped runway on the azimuth signal during normal aircraft landing operations. In these simulation the same runway profiles were modeled but the receiver path was changed to reflect an aircraft crossing threshold at 15.24 m (50 ft.) The power density predictions for a 15.24 m. (50 ft.) TCH show a significant increase in power density compared to that available for the same runway profile at 2.44 m. (8 ft.) above the runway at threshold, as expected. Therefore, it is expected that at most humped runway airdromes, even if the power density is not satisfied at 2.44 m, that there should be little effect on landing operations for Category I and II facilities. Further analysis of this problem will be required to determine the impact of a humped runway on Category III facilities and facilities requiring the use of the (approach) azimuth for departure guidance. (Typically the back azimuth is used for departure guidance.)
b. MLS-ILS Collocation

In the transition period from the Instrument Landing System (ILS) to the Microwave Landing System (MLS) both landing systems will service the same landing facility. This requires that siting criteria be developed which permit the collocated siting of MLS and ILS equipment. Two hypotheses were developed based on the theory of how the ILS signals are established in space. When collocating the MLS azimuth antenna ahead (approach side) of the ILS localizer antenna, the hypothesis is that the least effect on the localizer course will occur when the azimuth antenna is sited symmetrically about the localizer course. When collocating the MLS elevation antenna to the runway side of the ILS glide slope, the hypothesis is that the minimum effect on the glide slope signal will occur if the elevation antenna is sited outside the region which the Fresnel zone migrates during an approach. Based on these two hypotheses, flight measurement test plans were developed for validation of the hypotheses and to collect quantitative information.

The flight data show that the least effect on the localizer course occurs when the MLS azimuth station is sited on the localizer course centerline. It is also necessary that the MLS azimuth station be ahead of the localizer at least 100 feet. In the presence of
appreciable localizer course roughness or when siting ahead of an eight element localizer, increasing the distance between the localizer and the MLS azimuth station should be considered. If this is not possible or if it is questionable that the system will remain in tolerance, flight measurements using an MLS azimuth mock-up should be performed.

The glide slope data collected supports the hypothesis that the worst disturbances will occur when the scatterer penetrates the Fresnel zone or is positioned such that it blocks a portion of the antenna aperture. For a glide slope offset 400 to 450 feet from centerline, siting the MLS elevation station with an offset of 255 feet from runway centerline should have little effect on the glide path.

At the present time there is no information on the effects caused by siting a MLS azimuth station ahead of the localizer on the localizer back course. In addition, the case of siting the MLS station behind (departure side) the localizer and the effects this would have on both the localizer front and back courses have not yet been investigated. It is recommended that this type of work be performed.
As the glide slope offset is varied, an MLS elevation antenna positioned at an offset of 255 feet from runway centerline moves deeper laterally into the glide slope radiation pattern. In addition to avoiding penetration of the Fresnel zone, multipath effects may need to be considered when siting the MLS elevation station with a glide slope that is offset from centerline less than 400 feet. Furthermore, information on how much lateral pattern penetration, by the elevation antenna, is permitted will be useful in siting the elevation antenna at offsets from the runway which are greater than the glide slope offset. It is recommended that additional work be performed to develop a family of curves for different glide slope offsets using the sideband reference glide slope.

These data have been collected to provide information on the effects of siting MLS equipment at facilities currently served by ILS guidance. To increase confidence in the criteria established for siting the MLS equipment at an ILS facility it is recommended that additional field measurements be performed at future MLS-ILS sites. This type of work will also give insight to any unique site dependent characteristics, if they exist.

The MLS-ILS collocation work presented in this thesis represents one side of the collocation problem: the effects of MLS equipment on the ILS signals. To provide a complete
set of collocation criteria the effects of ILS structures on MLS signals must also be considered. Since the elevation antenna phase center height generally places the elevation antenna ahead of the glide slope antenna, a study to determine the effect of the glide slope antenna structure on the MLS elevation signal would not be necessary. However, there will be instances where the azimuth antenna will be sited behind (departure side) the localizer. Since real estate is limited, the localizer may be in the near field of the azimuth antenna. Therefore, the effects of the localizer antenna array’s physical structure on the azimuth signal should be determined. It is recommended that future work tackle this problem.
REFERENCES


[10] DOT-FAA order 6750.16A.


APPENDIX I

1. GTDDP FORTRAN Listing
2. GTDRUN FORTRAN Listing
3. GTDPLT FORTRAN Listing
4. MIKEPL FORTRAN Listing
1. GTDDP FORTRAN listing
THIS PROGRAM CALCULATES THE FAR FIELD FOR THE GTD WEDGE PROBLEM. THIS IS A DOUBLE PRECISION VERSION. DATA OUTPUT FOR THE GEOMETRICAL FIELD, THE INCIDENT DIFFRACTED FIELD, REFLECTED DIFFRACTED FIELD, TOTAL FIELD, AND TOTAL FIELD IN DB ARE AVAILABLE.

FILEDEFS
14 GEOMETRICAL FIELD (DIRECT AND REFLECTED COMPONENTS)
15 INCIDENT DIFFRACTED FIELD
16 REFLECTED DIFFRACTED FIELD
18 TOTAL FIELD
19 TOTAL FIELD IN DB

INPUTS
ROWL: OBSERVATION DISTANCE IN WAVELENGTHS
ROWPL: SOURCE DISTANCE IN WAVELENGTHS
AN: WEDGE ANGLE PARAMETER WHERE (2-AN) PI = WEDGE ANGLE IN RADIANS
PHIP: ANGLE BETWEEN SOURCE AND WEDGE IN DEGREES
FREQ: FREQUENCY
XSTRT: STARTING ANGLE
XEND: ENDING ANGLE
XINC: ANGLE INCREMENT

IMPLICIT COMPLEX*16 (C)
IMPLICIT REAL*8 (A,B,H,O-Z)

INPUT AREA

XSTRT=0.0D0
XEND=359.0D0
XINC=1.0D0
ROWL=100.0D0
ROWPL=1.0D0
ROWCF=ROWL*ROWPL/(ROWL+ROWPL)
ETMAX=0.0D0

POLARIZATION 0=HARD; 1=SOFT
NPOL=1
AN=2.0D0
PHIP=30.0D0
FREQ=5.05D09

END INPUT AREA

WL=3.0D08/FREQ
PI=DARCOS(-1.D0)
PHIPRD=PHIP*PI/180.D0
AK=2.0D0*PI/WL
ROW=ROWL*WL
ROWP=ROWPL*WL
XP = ROWP * DCOS (PHIPRD)
YP = ROWP * DSIN (PHIPRD)
XPP = XP
YP = -1.0 * YP

C CALCULATE BOUNDARIES
REG1 = 180.0 - PHIP
REG2 = 180.0 + PHIP

C CALCULATE DO LOOP PARAMETERS
ILOOP = INT (SNGL ((XEND - XSTRT) / XINC)) + 2
WRITE (6, 1220) ILOOP

1220 FORMAT (1X, I4)
DO 10 I = 1, ILOOP

C PHI = DFLOAT (I - 1) * XINC + XSTRT
PHIRR (I) = PHI
PHIRAD = PHI * PI / 180.0
X = ROW * DCOS (PHIRAD)
Y = ROW * DSIN (PHIRAD)
BMIN = PHI - PHIP
BPLUS = PHI + PHIP
DDR = DSQRT ((X - XP) ** 2 + (Y - YP) ** 2)
DR = DSQRT ((X - XPP) ** 2 + (Y - YPP) ** 2)

C CALCULATE DIRECT FIELD
AKDDR = AK * DDR
CAKDDR = DCMPLX (0.0, - AKDDR)
CEDIR = CDEXP (CAKDDR) / DSQRT (DDR)
IF (PHI .GT. REG2) CEDIR = DCMPLX (0.0, 0.1D-20)

C CALCULATE REFLECTED FIELD
AKDR = AK * DR
CAKDR = DCMPLX (0.0, - AKDR)
CEREF = CDEXP (CAKDR) / DSQRT (DR)
IF (NPOL .EQ. 1) CEREF = -1.0 * CEREF
IF (PHI .GT. REG1) CEREF = DCMPLX (0.0, 0.1D-20)
CEGO = CEDIR + CEREF
EGO = CDABS (CEGO)
WRITE (14, 30) PHI, EGO

30 FORMAT (1X, F7.2, 3X, D14.7)

C CALCULATE INCIDENT DIFFRACTED FIELD
AKROWP = AK * ROWP
CKROWP = DCMPLX (0.0, - AKROWP)
AKROW = AK * ROW
CKROW = DCMPLX (0.0, - AKROW)
CALL CF (RCF, UCF, ROWCF, BMIN, AN)
WRITE (20, 2001) PHI, N1, N2, IDD1, A1, A2, BOLT1, BOLT2
CDICF = DCMPLX (RCF, UCF)
P14 = PI / 4.0
CPI4 = DCMPLX (0.0, - PI4)
AKPI2 = 2.0 * PI * AK
CCD = -1.0 * CDEXP (CPI4) / (2.0 * AN * DSQRT (AKPI2))
CDI=CDICF*CCD
DI=CDABS(CDI)/(DSQRT(ROW)*DSQRT(ROWP))
WRITE(15,30) PHI,DI

CALCULATE REFLECTED DIFFRACTED FIELD
CALL CF(RCF,UCF,ROWCF,BPLUS,AN)
CDRCF=DCMPLX(RCF,UCF)
CDR=CDRCF*CCD
IF(NP0L.EQ.1) CDR=-1.0*CDR
DR=CDABS(CDR)/(DSQRT(ROW)*DSQRT(ROWP))
WRITE(16,30) PHI,DR
CEDIF=CDEXP(CKROWP)*CDEXP(CKROW)*(CDI+CDR)/(DSQRT(ROW)*DSQRT(ROWP))

CET=CEDIF+CEREF+CEDIR
ET=CDABS(CET)
WRITE(18,30) PHI,ET
IF(ET.GT.ETMAX) ETMAX=ET
ETR(I)=ET
10 CONTINUE

CALCULATE FAR FIELD IN DB
DO 20 II=1,ILOOP
IF(ETR(II).LE.0.0D0) ETR(II)=1.0D-40
ETDB(II)=20.0D0*DLOG10(ETR(II)/ETMAX)
IF(ETDB(II).LT.-90.0D0) ETDB(II)=-90.0D0
WRITE(19,30) PHIRR(II),ETDB(II)
20 CONTINUE
WRITE(6,70) ETMAX
70 FORMAT(1X,f20.10 MAXIMUM VALUE OF E=,D14.7)
STOP
END

SUBROUTINE CF(RCF,UCF,R,ANG,FN)
IMPLICIT REAL*8 (A-H,O-Z)
COMPLEX*16 DCMPLX,CDEXP
COMPLEX*16 Top,CTEXP,UPPI,UNPI
PI=PIRADS(-1.0D0)
TPI=2.0D0*PI
ANG=ANG*PI/180.0D0
Top=DCMPLX(0.0D0,2.0D0*DSQRT(TPI*R))
N=IFIX(SNGL((PI+ANG)/(2.0D0*FN*PI)+0.5D0))
DN=DFLOAT(N)
A=1.0+DCOS(ANG-2.0D0*FN*PI*DN)
Botl=DSQRT(TPI*R*A)
CTEXP=CDEXP(DCMPLX(0.0D0,TPI*R*A))
CALL FRNELS(AC,S,BOTL)
FORMAT(1X,2(D14.7))
AC=DSQRT(PI/2.0D0)*(0.5D0-AC)
S=DSQRT(PI/2.0D0)*(S-0.5D0)
RAG=(PI+ANG)/(2.0D0*FN)
TSIN=DSIN(RAG)
TS = DABS(TSIN)
X = 10.0
Y = 1.0 / X ** 5
IF (TS .GT. Y) GO TO 442
ACOMP = -DSQRT(2.0 * FN * DSIN(ANG / 2.0 - FN * PI * DN))
IF (DCOS(ANG / 2.0 - FN * PI * DN) .LT. 0.0) ACOMP = -ACOMP
GO TO 443

442 DP = DSQRT(A) * DCOS(RAG) / TSIN
ACOMP = DP

443 UPP = TOP * CTEXP * ACOMP * DCMLX(AC, S)
N = IFIX(SNGL((-PI + ANG) / (2.0 * FN * PI) + 0.5D0))
DN = DFLOAT(N)
A = 1.0 + DCOS(ANG - 2.0 * FN * PI * DN)
BOTL = DSQRT(TPI * R * A)
CTEXP = CDEXP(DCMLX(0.0, TPI * R * A))
CALL FRNELS(AC, S, BOTL)
AC = DSQRT(Pi / 2.0) * (0.5D0 - AC)
S = DSQRT(Pi / 2.0) * (S - 0.5D0)
RAG = (PI - ANG) / (2.0 * FN)
TSIN = DSIN(RAG)
TS = DABS(TSIN)
IF (TS .GT. Y) GO TO 542
ACOMP = DSQRT(2.0) * FN * DSIN(ANG / 2.0 - FN * PI * DN)
IF (DCOS(ANG / 2.0 - FN * PI * DN) .LE. 0.0) ACOMP = -ACOMP
GO TO 123

542 DP = DSQRT(A) * DCOS(RAG) / TSIN
ACOMP = DP

123 UPI = TOP * CTEXP * ACOMP * DCMLX(AC, S)
ANG = ANG * 180.0 / PI
RCF = DREAL(UPP + UPI)
UCF = DIMAG(UPP + UPI)
RETURN
END

C**********************************************************
SUBROUTINE FRNELS(C, S, XS)
IMPLICIT REAL*8 (A-H, O-Z)
DIMENSION A(12), B(12), CC(12), D(12)
A(1) = 1.595769140D0
A(2) = -0.000001702D0
A(3) = -6.808568854D0
A(4) = -0.000576361D0
A(5) = 6.920691902D0
A(6) = -0.016898657D0
A(7) = -3.050485660D0
A(8) = -0.075752419D0
A(9) = 0.850663781D0
A(10) = -0.025639041D0
A(11) = -0.150230960D0
A(12) = -0.034404779D0
B(1) = -0.000000033D0
B(2) = 4.255387524D0
B(3) = -0.000092810D0
\begin{verbatim}
B(4) = -7.7800204000D0
B(5) = -0.009520865D0
B(6) = 5.075161298D0
B(7) = -0.138341947D0
B(8) = -1.363729124D0
B(9) = -0.403349276D0
B(10) = 0.70222016D0
B(11) = -0.216195929D0
B(12) = 0.019547031D0
CC(1) = 0.0D0
CC(2) = -0.024933975D0
CC(3) = 0.000003936D0
CC(4) = 0.005770956D0
CC(5) = 0.000689892D0
CC(6) = -0.009497136D0
CC(7) = 0.011948809D0
CC(8) = -0.006748873D0
CC(9) = 0.000246420D0
CC(10) = 0.002102967D0
CC(11) = -0.001217930D0
CC(12) = 0.000233939D0
D(1) = 0.199471140D0
D(2) = 0.000000023D0
D(3) = -0.009351341D0
D(4) = 0.00023006D0
D(5) = 0.0048514661D0
D(6) = 0.001903218D0
D(7) = -0.017122914D0
D(8) = 0.029064067D0
D(9) = -0.027928955D0
D(10) = 0.016497308D0
D(11) = -0.005598515D0
D(12) = 0.000838386D0

IF(XS.LE.0.0D0) GO TO 414
X=XS
X=X*X
FR=0.0D0
FI=0.0D0
K=13
IF(X-4.0D0) 10,40,40
10 Y=X/4.0D0
20 K=K-1
FR=(FR+A(K))*Y
FI=(FI+B(K))*Y
IF(K-2) 30,30,20
30 FR=FR+A(1)
FI=FI+B(1)
C=(FR*DCOS(X)+FI*DSIN(X))*DSQRT(Y)
S=(FR*DSIN(X)-FI*DCOS(X))*DSQRT(Y)
RETURN
40 Y=4.0D0/X
50 K=K-1
FR=(FR+CC(K))*Y
FI=(FI+D(K))*Y
\end{verbatim}
IF (K-2) 60, 60, 50
60
FR=FR+CC(1)
FI=FI+D(1)
C=0.5D0+(FR*DCOS(X)+FI*DSIN(X)) * DSQRT(Y)
S=0.5D0+(FR*DSIN(X)-FI*DCOS(X)) * DSQRT(Y)
RETURN
414
C=-0. D0
S=-0. D0
RETURN
END
2. GTDRUN FORTRAN listing
THIS PROGRAM CALCULATES THE POWER DENSITY FOR THE MLS AZIMUTH SIGNAL WHEN THE AZIMUTH ANTENNA IS SITED AT A HUMPED RUNWAY FACILITY. THE GEOMETRICAL THEORY OF DIFFRACTION (GTD) TWO-DIMENSIONAL CONDUCTING WEDGE IS USED TO PREDICT THE AZIMUTH SIGNAL STRENGTH. ALTHOUGH THE AZIMUTH ANTENNA RADIATES A VERTICALLY POLARIZED SIGNAL, INVESTIGATION OF THE REFLECTION COEFFICIENT FOR PROPAGATION OVER REAL EARTH AT MICROWAVE FREQUENCIES INDICATE THAT AT GRAZING ANGLES (AND UP TO 15 DEGREES) THE APPROPRIATE REFLECTION COEFFICIENT IS CLOSELY APPROXIMATED BY -1.0. THEREFORE, TO AVOID CONFUSION ONE MUST REALIZE THAT THE GTD WEDGE MODEL HAS BEEN ADJUSTED FROM ITS IDEAL PRESENTATION IN TEXTBOOKS TO ACCOUNT FOR MICROWAVE PROPAGATION OF REAL EARTH.

FILEDEFS

11 DIRECT ELECTRIC FIELD
12 REFLECTED ELECTRIC FIELD
14 GEOMETRICAL FIELD (DIRECT AND REFLECTED COMPONENTS)
15 INCIDENT DIFFRACTED FIELD
16 REFLECTED DIFFRACTED FIELD
18 TOTAL FIELD
19 TOTAL FIELD IN DB
20 POWER DENSITY

INPUTS

ROWP: SOURCE DISTANCE IN METERS
AN: WEDGE ANGLE PARAMETER FOR RUNWAY HUMP WHERE (2-AN)*PI=WEDGE ANGLE IN RADIANS
PHIP: ANGLE BETWEEN SOURCE AND WEDGE SURFACE IN DEGREES
FREQ: SOURCE FREQUENCY
XSTRT: STARTING POSITION = 0 AT ANTENNA (METERS)
XEND: ENDING POSITION (METERS)
XINC: INCREMENT (METERS)
SLP: SLOPE OF RECEIVER PATH IN DEGREES WHICH IS USED WHEN THE RECEIVER IS ON THE FAR SIDE (SHADOWED AREA) OF THE RUNWAY HUMP. THIS ANGLE IS SET TO 360-WEDGE ANGLE TO HAVE THE RECEIVER PATH AT A CONSTANT HEIGHT (RCVHGT) ABOVE THE RUNWAY SURFACE.
RCVHGT: RECEIVER HEIGHT ABOVE WEDGE SURFACE (METERS)
COFIM: REFLECTION COEFFICIENT FOR REFLECTED FIELD FOR THE RUNWAY HUMP PROBLEM THIS FACTOR IS -1.
PTRAN: POWER TRANSMITTED IN WATTS
DTRAN: GAIN OF TRANSMITTING ANTENNA IN DB
NPOL: SIGN FOR REFLECTED DIFFRACTED FIELD. THIS
FACTOR IS SET TO 0 OR 1: 1 FOR THE RUNWAY HUMP PROBLEM.

IMPLICIT COMPLEX*16 (C)
IMPLICIT REAL*8 (A,B,D-H,O-Z)
DIMENSION ETR(6000),ETDB(6000),XRAY(6000),ETPD(6000)

INPUT AREA

PTTRAN=20.0D0
DTRAN=6.310D0
XSTRT=200.0D0
XEND=3350.00D0
XEND=400.0D0
XINC=5.0D0
RCVHGT=2.440D0
COFIM=-1.0D0
SLP=180.3D0
ROWP=2000.0D0
ETMAX=0.0D0
NPOL=1
AN=1.0017D0
PHIP=.349D0
FREQ=5.05D09

END INPUT AREA

DEFINE CONSTANTS AND CONVERT DEGREES TO RADIANS ET CETERA
WL=3.0D08/FREQ
PI=DARCOS(-1.0D0)
PHIPRD=PHIP*PI/180.0D0
SLPRAD=SLP*PI/180.0D0
AK=2.0D0*PI/WL
ROWPL=ROWP/WL

CALCULATE SOURCE COORDINATES
XP=ROWP*DCOS(PHIPRD)
YP=ROWP*DSIN(PHIPRD)

CALCULATE IMAGE COORDINATES
XPP=XP
YPP=-1.0D0*YP

CALCULATE SHADOW BOUNDARIES
REG1=180.0D0-PHIP
REG2=180.0D0+PHIP

CALCULATE DO LOOP PARAMETERS
ILOOP=INT(SNGL((XEND-XSTRT)/XINC))+2

LOOP INCREMENTS RECEIVER ALONG RUN WAY. FOR EACH POINT ROW,PHI, X, Y, ROWL, AND ROWCF MUST BE RECALCULATED
DO 10 I=1,ILoop
XT=DFLOAT(I-1)*XINC+XStRT
IF (XT .GT. ROWP) GO TO 120
Y=RCVHGT
IF(ROWP .EQ. XT) ROWP=ROWP+1.0D-5
PHIRAD=DATAN(Y/ (ROWP-XT))
ROW=DSQRT((ROWP-XT)**2+y**2)
ROWL=ROW/WL
PHI=PHIRAD*180.0D0/PI
GO TO 130
120
Y=(ROWP-XT)*DTAN(SLPRAD)+RCVHGT
PHIRAD=PI+DATAN(Y/ (ROWP-XT))
PHI=PHIRAD*180.0D0/PI
ROW=DSQRT((ROWP-XT)**2+y**2)
ROWL=ROW/WL
130
ROWCF=ROWL*ROWL/ (ROWL+ROWL)
X=ROWP-XT
XRAY(I)=XT
BMIN=PHI-PHIP
BPLUS=PHI+PHIP
C
C SOURCE-OBSERVATION, AND IMAGE-OBSERVATION DISTANCES
DDR=DSQRT((X-XP)**2+(Y-YP)**2)
DR=DSQRT((X-XPP)**2+(Y-YPP)**2)
C
C CALCULATE DIRECT FIELD
AKDDR=AK*DDR
CAKDDR=DCMPLX(0.0D,0,-AKDDR)
CALL ANTPAT(DIRCOF,DDR,ROW,ROWP,0)
C
C CHOSE SPREADING FACTOR FOR DIRECT FIELD
DIRSF=1.0D0/DDR
C
C DIRSF=1.0D0/DSQRT(DDR)
CEDIR=DIRCOF*CDEXP(CAKDDR)*DIRSF
EDIR=CDABS(CEDIR)
IF(PHI .GT. REG2) CEDIR=DCMPLX(0.0D0,0.1D-50)
EDIR=CDABS(CEDIR)
WRITE(11,30) XT,EDIR
C
C CALCULATE REFLECTED FIELD
AKDR=AK*DR
CAKDR=DCMPLX(0.0D,0,-AKDR)
CALL ANTPAT(REFCOF,DR,ROW,ROWP,1)
C
C CHOSE SPREADING FACTOR FOR REFLECTED FIELD
REFSF=1.0D0/DR
C
C REFSF=1.0D0/DSQRT(DR)
CEREF=COFIM*REFCOF*CDEXP(CAKDR)*REFSF
IF(PHI .GT. REG1) CEREF=DCMPLX(0.0D0,0.1D-20)
EREF=CDABS(CEREF)
WRITE(12,30) XT,EREF
CEGO=CEDIR+CEREF
EGO=CDABS(CEGO)
WRITE(14,30) XT,EGO
CHOOSE SPREADING FACTOR FOR DIFFRACTED FIELDS
DIFS=DSQRT(ROWP/(ROW*(ROW+ROWP)))/ROWP

CALCULATE INCIDENT DIFFRACTION COEFFICIENT (CDI)
AKROWP=AK*ROWP
CKROWP=DCMPLX(0.D0,-AKROWP)
AKROW=AK*ROW
CKROW=DCMPLX(0.D0,-AKROW)
CALL CF(RCF,UCF,ROWCF,BMIN,AN)
CDICF=DCMPLX(RCF,UCF)
PI4=PI/4.D0
CPI4=DCMPLX(0.D0,-PI4)
AKPI2=2.D0*PI*AK
CCD=-1.D0*CDEXP(CPI4)/(2.D0*AN*DSQRT(AKPI2))
CDI=CDICF*CCD

CALCULATE MAGNITUDE OF INCIDENT DIFFRACTED FIELD
DI=CDABS(CDI)*DIFS
WRITE(15,30) XT,DI

CALCULATE REFLECTED DIFFRACTION COEFFICIENT (CDR)
CALL CF(RCF,UCF,ROWCF,BPLUS,AN)
CDRCF=DCMPLX(RCF,UCF)
CDR=CDRCF*CCD
IF(NPOL.EQ.1) CDR=-1.D0*CDR

CALCULATE MAGNITUDE OF REFLECTED DIFFRACTED FIELD
DR=CDABS(CDR)*DIFS
WRITE(16,30) XT,DR

CALCULATE TOTAL DIFFRACTED FIELD
CEDIF=CDEXP(CKROWP)*CDEXP(CKROW)*(CDI+CDR)*DIFS

CALCULATE TOTAL FIELD
CET=CEDIF+CEREF+CEDIR
ET=CDABS(CET)
WRITE(18,30) XT,ET
IF(ET.GT.ETMAX) ETMAX=ET
ETR(I)=ET
CONTINUE

CALCULATE FAR FIELD IN DB

DO 20 II=1,ILOOP
IF(ETR(I).LE.0.0D0) ETR(I)=1.0D-40
ETDB(I)=20.D0*DLOG10(ETR(I)/ETMAX)
IF(ETDB(I).LT.-100.0D0) ETDB(I)=-100.0D0
WRITE(19,30) XRAY(I),ETDB(I)

CALCULATE POWER DENSITY
PDENS=PTRAN*DTRAN*ETR(I)*ETR(I)
ETPD(I)=10.0D0*DLOG10(PDENS/377.0D0)
WRITE(20,30) XRAY(I),ETPD(I)
SUBROUTINE CF(RCF, UCF, R, ANG, FN)
IMALPICATION REAL*8 (A-H, O-Z)
COMMPLEX*16 DCMLPX, CDEXP
COMMLEX*16 TOP, CTEXP, UPPI, UNPI
PI=DARCOS(-1.0D0)
TPI=2.0D0*PI
ANG=ANG*PI/180.0D0
TOP=DCMPLEX(0.0D0, 2.0D0*DSQRT(TPI*R))
N=IFIX(SNGL((PI+ANG)/(2.0D0*FN*PI)+0.5D0))
DN=DFLOAT(N)
A=1.0D0+DCOS(ANG-2.0D0*FN*PI*DN)
BOTL=DSQRT(TPI*R*A)
CTEXP=CDEXP(DCMPLX(0.0D0, TPI*R*A))
CALL FRNELS(AC, S, BOTL)
FORMAT(1X, 2(D14.7))
AC=DSQRT(PI/2.0D0)*(0.5D0-AC)
S=DSQRT(PI/2.0D0)*(S-0.5D0)
RAG=(PI+ANG)/(2.0D0*FN)
TSIN=DSIN(RAG)
TS=DABS(TSIN)
X=10.0D0
Y=1.0D0/X**5
IF(TS.GT.Y) GO TO 442
ACOMP=-DSQRT(2.0D0)*FN*DSIN(ANG/2.0D0-FN*PI*DN)
IF(DCOS(ANG/2.0D0-FN*PI*DN).LT.0.0D0) ACOMP=-ACOMP
GO TO 443
DP=DSQRT(A)*DCOS(RAG)/TSIN
ACOMP=DP
443 UPPI=TOP*CTEXP*ACOMP*DCMPLX(AC, S)
N=IFIX(SNGL((-PI+ANG)/(2.0D0*FN*PI)+0.5D0))
DN=DFLOAT(N)
A=1.0D0+DCOS(ANG-2.0D0*FN*PI*DN)
BOTL=DSQRT(TPI*R*A)
CTEXP=CDEXP(DCMPLX(0.0D0, TPI*R*A))
CALL FRNELS(AC, S, BOTL)
AC=DSQRT(PI/2.0D0)*(0.5D0-AC)
S=DSQRT(PI/2.0D0)*(S-0.5D0)
RAG=(PI-ANG)/(2.0D0*FN)
TSIN=DSIN(RAG)
TS=DABS(TSIN)
IF(TS.GT.Y) GO TO 542
ACOMP=DSQRT(2.0D0)*FN*DSIN(ANG/2.0D0-FN*PI*DN)
IF(DCOS(ANG/2.0D0-FN*PI*DN).LE.0.0D0) ACOMP=-ACOMP
GO TO 123
542 DP=DSQRT(A)*DCOS(RAG)/TSIN
ACOMP=DP
123 UNPI=TOP*CTEXP*ACOMP*DCMPLX(AC, S)
ANG = ANG * 180.D0 / PI
RCF = DREAL (UPPI + UNPI)
UCF = DIMAG (UPPI + UNPI)
RETURN
END

C**********************************************************
SUBROUTINE FRNELS (C, S, XS)
IMPLICIT REAL*8 (A-H, O-Z)
DIMENSION A(12), B(12), CC(12), D(12)
A(1) = 1.595769140D0
A(2) = -0.000001702D0
A(3) = -6.808568854D0
A(4) = -0.000576361D0
A(5) = 6.920691902D0
A(6) = -0.016898657D0
A(7) = -3.050485660D0
A(8) = -0.075752419D0
A(9) = 0.850663781D0
A(10) = -0.025639041D0
A(11) = -0.150230960D0
A(12) = 0.034404779D0
B(1) = -0.000000033D0
B(2) = 4.255387524D0
B(3) = -0.000092810D0
B(4) = -7.780020400D0
B(5) = -0.009520865D0
B(6) = 5.075161298D0
B(7) = -0.138341947D0
B(8) = -1.363729124D0
B(9) = -0.403349276D0
B(10) = 0.702222016D0
B(11) = -0.216195929D0
B(12) = 0.019547031D0
CC(1) = 0.0D0
CC(2) = -0.024933975D0
CC(3) = 0.000003936D0
CC(4) = -0.024933975D0
CC(5) = 0.000003936D0
CC(6) = 0.005770956D0
CC(7) = -0.009497136D0
CC(8) = -0.011948809D0
CC(9) = 0.000024642D0
CC(10) = 0.0002102967D0
CC(11) = -0.001217930D0
CC(12) = 0.000233939D0
D(1) = 0.199471140D0
D(2) = 0.000000023D0
D(3) = -0.009351341D0
D(4) = 0.000023006D0
D(5) = 0.0048514661D0
D(6) = 0.001903218D0
D(7) = -0.017122914D0
D(8)=0.029064067D0
D(9)=-0.027928955D0
D(10)=0.016497308D0
D(11)=-0.005598515D0
D(12)=0.000838386D0
IF(XS.LE.0.0D0) GO TO 414
X=XS
X=X*X
FR=0.D0
FI=0.D0
K=13
IF(X-4.D0) 10,40,40
10 Y=X/4.D0
20 K=K-1
FR=(FR+A(K))*Y
FI=(FI+B(K))*Y
IF(K-2) 30,30,20
30 FR=FR+A(1)
FI=FI+B(1)
C=(FR*DCOS(X)+FI*DSIN(X))*DSQRT(Y)
S=(FR*DSIN(X)-FI*DCOS(X))*DSQRT(Y)
RETURN
40 Y=4.D0/X
50 K=K-1
FR=(FR+CC(K))*Y
FI=(FI+D(K))*Y
IF(K-2) 60,60,50
60 FR=FR+CC(1)
FI=FI+D(1)
C=0.5D0+(FR*DCOS(X)+FI*DSIN(X))*DSQRT(Y)
S=0.5D0+(FR*DSIN(X)-FI*DCOS(X))*DSQRT(Y)
RETURN
414 C=-0.D0
S=-0.D0
RETURN
END

SUBROUTINE ANTPAT(PATCOF,R,ROW,ROWP,ICHECK)

THIS SUBROUTINE CALCULATES THE ANTENNA GAIN IN THE VERTICAL PLAN FOR THE AZIMUTH ANTENNA BASED ON THE INFORMATION OBTAINED FROM THE HAZELTINE CORPORATION. SINCE THE AZIMUTH ANTENNA HAS A SHARP CUTOFF AT THE HORIZON THE POWER INCIDENT AT THE "VERTEX" OF THE RUNWAY HUMP CAN VERY GREATLY DEPENDING ON THE ANTENNA PHASE CENTER HEIGHT ABOVE THE RUNWAY.

IMPLICIT REAL*8 (A-H,O-Z)
A=.1215D1
PI=DARCOS(-1.0D0)
A1=(ROW**2-(R**2+ROWP**2))/(-2.0D0*R*ROWP)
THETA=DARCOS(A1)
ARAD=1.215D0*PI/180.0D0
A2 = THETA * ARAD
IF (ICHECK .EQ. 1) GO TO 40
PATCOF = 1.0D0 + .8D0 * DTANH(A2)
GO TO 100
40  PATCOF = 1.0D0 - .8D0 * DTANH(A2)
100  RETURN
END
3. GTDPLT FORTRAN listing
THIS PROGRAM USES THE DISSPLA PLOTTING PACKAGE AT OHIO UNIVERSITY TO PRODUCE A POLAR PLOT. THE COMMAND STRUCTURE IS STRAIGHT FORWARD, BUT ONE SHOULD READ THE FIRST TWO SECTIONS OF THE DISSPLA MANUAL BEFORE USING THIS PROGRAM.

```
DIMENSION PHI(721), DBR(721)
CALL HP7470
CALL XNAME ('RELATIVE POWER IN DB$', 100)
CALL YNAME ('$', 100)
CALL PAGE (11., 8.5)
CALL AREA2D (6.0, 6.0)
CALL TRIPLX
CALL BASALF ('L/CSTDr)
CALL MIXALF ('STANDARDt)
CALL HEADIN ('(POWER PATTERN-GTD WEDGE
* PROBLEM)$', 100, 1.2, 2)
CALL HEADIN ('(H)ORIZONTAL ((SOFT)) (P)OLARIZATION
* $', 100, .8, 2)
CALL POLORG (-90.)
CALL POLAR (3.14159/180., 30., 3., 3.)
CALL DASH
CALL GRID (1, 3)
CALL RESET ('DASH')
DO 10 I=1, 721
   II=I-1
   PHI(I)=FLOAT(II)/2.
   READ (20, 30, END=40) DB
   DBR(I)=DB
10   CONTINUE
30   FORMAT (1X, E14.7)
40   CALL CURVE (PHI, DBR, 721, 0)
CALL ENDPIL (0)
CALL DONEPL
STOP
END
```
4. MIKEPL FORTRAN listing
THIS PROGRAM USES THE DISSPLA PLOTTING PACKAGE. IT IS RECOMMENDED THAT YOU READ THE FIRST TWO SECTIONS OF THE DISSPLA MANUAL BEFORE USING.

DIMENSION X(720), Y(720), XGOR(720), YGOR(720), XDIR(720), YDIR(720),
* X1(1), Y1(1), X2(1), Y2(1), X3(1), Y3(1), X4(1), Y4(1)
CALL PTEKAL
CALL XNAME ('OBSERVATION ANGLE IN DEGREES$', 100)
CALL YNAME ('NORMALIZED FAR-FIELD MAGNITUDE$', 100)
CALL XNAME ('DISTANCE FROM AZIMUTH ANTENNA$', 100)
CALL YNAME ('E-FIELD MAGNITUDE$', 100)
CALL PAGE (11.0, 8.0)
CALL AREA2D (8.0, 5.5)
CALL TRIPLX
CALL BASALF ('\text{L/CSTD}'$
CALL MIXALF ('\text{STANDARD}'$
CALL HEADIN ('\text{GTD WEDGE PROBLEM}$', 100, 1.2, 2)
CALL HEADIN ('\text{(H)ORIZONTAL ((SOFT)) (P)OLARIZATION}$', 100, 8.2)
CALL RESET ('\text{TRIPLX}')$
CALL GRAF (0.0, 1000.0, 4000.0, 0.0, .5, 1.0)
CALL RLMESS ('\text{TOTAL}$', 100, 240., 2.40)
CALL RLMESS ('\text{GEOMETRICALOPTICS}$', 100, 240., 2.22)
CALL RLMESS ('\text{INCIDENT DIFFRACTED}$', 100, 240., 2.04)
CALL RLMESS ('\text{REFLECTED DIFFRACTED}$', 100, 240., 1.86)
CALL RLMESS ('\text{HALF-PLANE N=2}$', 100, 240., 1.68)
CALL RLMESS ('\text{SOURCE DISTANCE= 1}$', 100, 240., 1.50)
CALL RLMESS ('\text{INCIDENCE ANGLE= 30}$', 100, 240., 1.32)
CALL RESET ('\text{MIXALF}')$
CALL RESET ('\text{BASALF}')$
CALL BASALF ('\text{L/CGREEK}')$
CALL MIXALF ('\text{GREEK}')$
CALL RLMESS ('\text{L}$', 100, 340., 1.53)
X1(1)=230.0
Y1(1)=2.43
X2(1)=230.
Y2(1)=2.25
X3(1)=230.
Y3(1)=2.07
X4(1)=230.
Y4(1)=1.89
CN=.1832513
I PLOT=-2
DO 10 I=1,720
READ (8, 30, END=40) XXTOL, YYTOL
FORMAT (1X, F7.2, 3X, E14.7)
X(I)=XXTOL
Y(I)=YYTOL/CN
READ (20, 30, END=40) XXGO, YYGO
XGOR(I)=XXGO
YGOR(I)=YYGO/CN
READ(21,30,END=40) XXDI, YYDI
XDIR(I)=XXDI
YDIR(I)=YYDI/CN
READ(22,30,END=40) XXDR, YYDR
XDRR(I)=XXDR
YDRR(I)=YYDR/CN
IPLT=IPLT+1
10 CONTINUE
40 CALL MARKER (4)
CALL CURVE (X1,Y1,1,-1)
CALL CURVE (X,Y,IPLT,25)
CALL MARKER (5)
CALL CURVE (X2,Y2,1,-1)
CALL CURVE (XGOR,YGOR,IPLT,25)
CALL MARKER (6)
CALL CURVE (X3,Y3,1,-1)
CALL CURVE (XDIR,YDIR,IPLT,25)
CALL MARKER (7)
CALL CURVE (X4,Y4,1,-1)
CALL CURVE (XDRR,YDRR,IPLT,25)
CALL ENDPL (0)
CALL DONEPL
STOP
END
APPENDIX II

1. Samples of Flight Recordings
1. Samples of Flight Recordings

The six flight recordings presented in this appendix form a representative sample of the approximately 500 flight recordings generated and analyzed to develop the MLS-ILS collocation criteria presented in this document. Since the actual flight recordings are required to remain on file at Ohio University, only xerox copies could be presented in this document. In addition, the average chart recording is about four feet in length, therefore, only a portion of the chart recording is presented herein. The portion chosen represents the part of the recording from threshold to about 2 nautical miles from threshold (in the approach direction with threshold to the readers left). However, this portion of the flight recording is sufficient enough to show examples of the effects measured during the MLS-ILS collocation study. These flight recordings are presented with a brief narrative.

Figures A1, A2, and A3 are copies of the localizer flight recordings generated at Airborne Express Airpark, Wilmington, Ohio. Figure A4 shows the locations of the azimuth antenna system mockup for the flight measurement tests performed at Airborne. Figure A1 represents a baseline or reference measurement of the localizer course.
SEE BOOK POCKET FOR FLIGHT RECORDINGS
SEE BOOK POCKET FOR FLIGHT RECORDINGS
SEE BOOK POCKET FOR FLIGHT RECORDINGS
Figure A4  Locations of MLS azimuth system mockup relative to the ILS localizer.
It is evident from this recording that the localizer course is one of high quality showing only 2 to 3 microamperes (μAs) of structure roughness and an alignment value of -2 μAs.

Figure A2 represents a flight recording generated with the MLS azimuth mockup sited on the localizer course centerline, 94.0 feet in front of the localizer array. Note that this recording is almost identical to the one presented in figure A1. The structure roughness is again about 2 to 3 μAs with an alignment value of -1 μA, which is within the ±2 μA measurement accuracy.

Figure A3 represents a flight recording with the MLS azimuth mockup site at 21.8 feet offset from runway centerline, 94 feet ahead of the localizer array. Since this azimuth system location will block a significant portion of the sideband only energy, a noticeable effect was expected on the localizer course. Figure A3 shows that the structure roughness remained about 2 to 3 μAs, but that an alignment value of -22 μA was recorded. This alignment value exceeds tolerance and the localizer monitor system did shut the localizer system down. The localizer monitor system had to be switched into its bypass mode to permit this flight measurement to be made.
Figures A5, A6, and A7 present flight data measured at Ohio University's test facility in Tamiami, Florida. The sideband reference glide slope system was used in making these flight recordings. Figure A8 shows the locations of the MLS elevation system mock-up relative to the glide slope for the flight recording presented. Figure A5 contains the baseline measurement for the sideband reference glide slope system. The baseline measurement shows the system to have a 2.96 degree path angle and a structure roughness of 6 to 8 μAs.

Figure A6 shows the flight data measured with the MLS elevation system mock-up site 255 feet from the runway centerline and setback 810 feet from threshold. This elevation antenna location is outside of the glide slope's Fresnel zone migration region. The flight measurement shows the system to have a 2.97 degree path angle and a structure roughness of 6 to 8 μAs.

Figure A7 show the flight data recorded for the MLS elevation system mock-up site 450 feet from runway centerline (setback 810 feet from threshold), directly in front of the glide slope. The record shows that the system has a 2.82 degree path angle and a 45 μA course reversal. Both of these parameters are out-of-tolerance.
SEE BOOK POCKET FOR FLIGHT RECORDINGS
SEE BOOK POCKET FOR FLIGHT RECORDINGS
SEE BOOK POCKET FOR FLIGHT RECORDINGS
Figure A8  Locations of MLS elevation system mock-up relative to the ILS glide slope