The Curtain Antenna Design

Cheng-Lee Chao

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THE CURTAIN ANTENNA DESIGN

BY

CHENG-LEE CHAO

A thesis submitted in partial fulfillment of the requirements for the degree Master of Science, Department of Electrical Engineering, South Dakota State College of Agriculture and Mechanic Arts

March, 1960

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Thesis Adviser

Head of the Major Department
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CHAPTER I
INTRODUCTION

This paper treats, in an elementary way, the antenna array problems, the effect of varying antenna array spacing and the relative power intensity of antenna arrays.

The radiation of a directive antenna might be projected from one to another with highest efficiency and a minimum of interference with adjacent stations. We have to improve the signal-to-noise ratio and discriminate against undesirable noise. The principles of radiation on the wave theory have been known for several centuries. The wave theory of light was first proposed by the English physicist Robert Hooke in 1665 and developed twenty years later by the Dutch scientist and mathematician Christian Huygens. When Gustav Hertz stated his celebrated experiments to verify Maxwell's theory, he was in full knowledge of these phenomena and their explanation, and invoked their use in proving the existence of electric wave.\(^1\) Augustin Fresnel, a French scientist, offered the wave theory of light in the early part of the last century. Although this viewpoint was held for many years it was later abandoned in favor of a new wave theory of light. By adopting the wave hypothesis a complete and adequate account of reflection, refraction, diffraction, interference, and polarization phenomena

was finally understood.

We consider each antenna as a spherical source of waves that radiates equal power in all directions. It is assumed that the current in each individual source is the same and is not materially affected in either magnitude or phase by its proximity to other sources.

The interference patterns resulting from a number of individual sources of waves, such as antennas, are dependent on both their special arrangement and the magnitudes and relative phases of their currents. We first discuss a very simple case of wave interference in the following paragraph.

We adopt the graphical method in a rough way to plot the interference patterns from two independent sources of spherical wave of the same amplitude. In Figure 1 an array of two antennas are separated by half wave length and phase difference in phase. At points where either two crests or two troughs arrive other crests and troughs arrive also, thereby neutralizing each other's effects. In the diagram we get a strong wave in northern and southern directions and two waves cancel out in eastern and western directions. In Figure 2 is the second case that two antennas are separated by half wave length and phase difference in 1/8 wave length. In Figure 3 is the third case which is separated by one wave length and phase difference in half period. Accompanying each figure is a directive diagram (Figures 1b, 2b and 3b) plotted in polar coordinates, which shows the effectiveness of the wave in each direction. The circle drawn outside each diagram indicates
the effect. If the radiation had proceeded from a single non-directional source similar to each of the above antennas, a field of semicircles would have resulted. The ratio between the areas of the circle and the inscribed diagram gives roughly the power improvement of such a device as manifested in the intensity of the radiated wave.

Most of the material of the field intensity pattern was given by Mr. R. M. Foster, who assembled it and published "Directive Diagrams of Antenna Arrays" in Bell Systems, Tech. Jour. 292, 1926.
Figure 1. An Array of Two Antennas Separate in Half Wavelength, Phase Difference in Phase
Figure 2. An Array of Two Antennas Separate in Half Wavelengths, Phase Difference in $1/8$ Wavelength
Figure 3. An Array of Two Antennas Separate in One Wavelength, Phase Difference in Half Period
CHAPTER II

CALCULATION OF LOGARITHMIC SPIRAL DIAGRAM

Most directive antenna systems now in general use for short waves may be regarded as special applications of the linear array. This type consists of two or more antennas having currents of equal amplitude, equispaced along the same straight line. The properties of such arrays have been treated very generally by Mr. Foster.

In Figure 4 is shown an array of many antennas separated by equal space. We consider a number of vertical elements arranged in line with the transmitting station an array. A transmission line is used to connect all vertical elements together. A coupling network is inserted between the antenna and the terminal end of transmission line to reduce standing waves. This impedance is not the characteristic impedance of the line, but rather a new impedance caused by the addition of the vertical elements. These elements will change the effective shunting impedance across the line and the line velocity.

\[ \text{Figure 4. An Array of Antennas Separate in Same Distance} \]
Considering first the array as being driven the current in each element is

\[ i_1 = I \]
\[ i_2 = I e^{2\alpha d} (\beta d + \theta) \]
\[ i_3 = I e^{2\alpha d} (\beta d + \theta) \]
\[ i_4 = I e^{2\alpha d} (\beta d + \theta) \]
\[ \ldots \]
\[ i_n = I e^{(n-1)\alpha d} ((n-1) (\beta d + \theta)) \]

Where \( \alpha = \) attenuation factor
\( \beta = \) phase constant = \( \frac{2\pi}{\lambda} \)
\( \theta = \frac{2\pi a}{\lambda} \)

The interference pattern of such an array can be obtained by adding vectorially the field intensities due to each of the elements. The field strength of each element is proportional to the magnitude of the current in this element. The summation of these fields is

\[ \vec{E} = \vec{E}_1 + \vec{E}_2 + \vec{E}_3 + \ldots + \vec{E}_n \]
\[ = K I \left[ 1 + e^{\alpha d+j(\beta d+\theta)} + e^{2(\alpha d+j(\beta d+\theta))} + \ldots + e^{(n-1)(\alpha d+j(\beta d+\theta))} \right] \]
\[ = K I \frac{1 - e^{n(\alpha d+j(\beta d+\theta))}}{1 - e^{(\alpha d+j(\beta d+\theta))}} \]

where \( K \) is a constant depending on element radiation efficiency.

For this general linear array of equally spaced elements the relative amplitude of the radiated field intensity is solved by a Logarithmic Spiral.
Figure 5 is Logarithmic Spiral diagram. We set
\[ r_0 = r_1 = r_2 = \ldots = r_n = e^{\alpha d} \]  
(3)
and
\[ |E_1| = |E_2| = |E_3| = \ldots = |E_{n-1}| = e^{\alpha d} \]
(4)

\[ |E_2| \quad |E_3| \quad |E_4| \quad \ldots \quad |E_n| \]

Considering \( \triangle OAB \) in Figure 5, we get
\[ |E_1|^2 = r_1^2 + r_0^2 - 2r_1r_0 \cos \delta \]  
(5)
\[ \frac{|E_{11}|^2}{r_0^2} = 1 + e^{-2\alpha d} - 2e^{-\alpha d} \cos \delta \]  
(6)

The summation of these field intensities can be obtained by the vector \( r_0 \) and \( r_{n+1} \):
\[ \Sigma |E|^2 = r_0^2 + r_{n+1}^2 + 2r_0r_{n+1} \cos n \delta \]  
(7)
\[ \left( \frac{\Sigma |E|}{r_0} \right)^2 = 1 + e^{-2(n+1)\alpha d} + 2e^{-\alpha(n+1)d} \cos n \delta \]  
(8)
Figure 5 is Logarithmic Spiral diagram. We set
\[ \frac{r_0}{r_1} = \frac{r_1}{r_2} = \frac{r_2}{r_3} = \cdots = \frac{r_n}{r_{n+1}} = e^{\delta d} \]  
and
\[ \frac{|E_1|}{|E_2|} = \frac{|E_2|}{|E_3|} = \cdots = \frac{|E_{n-1}|}{|E_n|} = e^{\delta d} \]

Figure 5. Logarithmic Spiral Diagram

Considering \( \triangle OAB \) in Figure 5, we get
\[ |E_1|^2 = r_1^2 + r_0^2 - 2r_1r_0 \cos \delta \]
\[ \frac{|E_1|^2}{r_0^2} = 1 + e^{-2\delta d} - 2e^{-\delta d} \cos \delta \]

The summation of these field intensities can be obtained by the vector \( r_0 \) and \( r_{n+1} \).
\[ \sum |E|^2 = r_0^2 + r_{n+1}^2 + 2r_0r_{n+1} \cos n \delta \]
\[ \left( \frac{\sum |E|^2}{r_0} \right)^2 = 1 + e^{-2(n+1)\delta d} + 2e^{-(n+1)\delta d} \cos n \delta \]
Combining equation (6) and (8)

\[
\left( \sum |E| \right)^2 = 1 + \frac{e^{-2(n+1)d} + 2e^{-2d} \cos \nu \cdot \cos \frac{\nu}{2}}{1 + e^{-2d} - 2e^{-d} \cos \delta} \tag{10}
\]

\[
\left| \frac{E}{E_1} \right| = \sqrt{\frac{1 + e^{-2(n+1)d} + 2e^{-2d} \cos \nu \cdot \cos \frac{\nu}{2}}{1 + e^{-2d} - 2e^{-d} \cos \delta}} \tag{11}
\]

If the attenuation factor can be neglected the general formula for the radiation pattern of the array is obtained

\[
\left| z \right| = E_1 \frac{\sin \frac{1}{2} \nu \cdot \cos \frac{\nu}{2}}{\sin \frac{1}{2} \delta} \tag{12}
\]

The gain, G, of an antenna in a given direction is defined as

\[4\pi \text{ times the ratio of the radiation intensity in that direction to the total power, considering the field-strength gain in the line of the array as compared to a single element.} \]

\[w_1 \text{ is delivered to the single element and } w_2 \text{ is delivered to the line of the array. } w_1 \text{ and } w_2 \text{ are} \]

\[w_1 = I_1^2 R_1 \tag{13}\]

\[w_2 = \frac{(I_2 Z_2)^2}{Z} \tag{14}\]

where

\[Z_2 = \text{base impedance of the line of the antenna array} \]

\[Z = \text{characteristic impedance in transmission line} \]

\[R_1 = \text{base impedance of the single element} \]

\[G = \frac{w_2}{w_1} = \frac{(I_2 Z_2)^2}{Z I_1^2 R_1} = \frac{R_2^2 Z_2^2}{Z R_1} \tag{15}\]
For example, we take a short vertical antenna to compare with the line of the array. Since the radiation resistance of a short vertical antenna above ground is

$$ R_1 = \frac{395}{\lambda_2} (\frac{h}{\lambda_2})^2 $$

(16)

where $h$ is the height of the single element, the power gain is

$$ G = \frac{1}{395} \frac{(\gamma)^2}{(h)^2} \frac{z_2^2}{z} R_k^2 $$

$$ = \frac{1}{395} \frac{(\gamma)^2}{(h)^2} \frac{z_2^2}{z} \left( \frac{1+e^{-2(n+1)\lambda_d} + 2e^{-(n+1)\lambda_d} \cos n \delta}{1+e^{-2\lambda_d} - 2e^{-2\lambda_d} \cos \delta} \right) $$

(17)

if the attenuation can be neglected. The general formula for the power gain is obtained

$$ G = \frac{1}{395} \frac{(\gamma)^2}{(h)^2} \frac{z_2^2}{z} \left( \frac{\sin \frac{1}{2} n \delta}{\sin \frac{\delta}{2}} \right)^2 $$

$$ = C \left( \frac{\sin \frac{1}{2} n \delta}{\sin \frac{\delta}{2}} \right)^2 $$

(18)

where $C = \text{constant}$.

The length of array is determined by frequency and the characteristic of the transmission line. If the array becomes quite long the optimum velocity of propagation along the loaded line approaches that of free

---

\(^{2}\text{H. H. Skillings, Fundamentals of Electric Waves, Chapter XII, pp. 172-191, John Willey and Sons, Inc.: New York, 1948.}\)
space and tuning becomes difficult.
CHAPTER III

CALCULATION OF HIGH GAIN CURTAIN ANTENNA

The fundamental principles of the directive antenna were developed prior to the high frequency era. Gain was for many years a primary objective in engineering. For transoceanic communication with short waves, antennas with highly directional radiation characteristics are used. The high gain directive curtain antennas of Chinese radio stations set up in Formosa and directed toward Russia were to be determined experimentally. The high gain directive curtain antennas were built by A. D. Ring Company and consists of 16 dipoles in two vertical planes, each with 8 dipoles. One vertical plane is excited by radio frequency power, the other is used to reflect the radio wave forward. The following equations are the calculations of the horizontal characteristic and the vertical characteristic of the curtain antenna.

Figure 6. A Single Horizontal Dipole
In Figure 6 we consider the intensity of the electric wave at a point \( P \) a great distance from a dipole\(^3\)

\[
i = I_0 \cos \left( \frac{2\pi x}{\lambda} \right)
\]

(19)

according to Hertz equation\(^4\)

\[
d\varepsilon = \frac{2\pi C}{\lambda} i_x dx \cos \theta \sin (wt - \frac{2\pi x}{\lambda}) \cos \left( \frac{2\pi x}{\lambda} \sin \theta \right)
\]

(20)

where

\[
\frac{2\pi C}{\lambda} = \text{constant}
\]

\( i_x dx \) = current element

\( \cos \theta \) = direction of the radiation beam

\( \cos \left( \frac{2\pi x}{\lambda} \sin \theta \right) \) = phase difference

\[
\varepsilon = \frac{4\pi C}{\lambda} \cos \theta \sin (wt - \frac{2\pi x}{\lambda}) \int_{x=0}^{x=1} i_x \cos \left( \frac{2\pi x}{\lambda} \sin \theta \right) dx
\]

(21)

Substituting equation (19) in equation (21) and then integrating by parts we get

\[
\varepsilon = \frac{2\pi C I_0}{\lambda} \frac{\pi}{\lambda} \sin \left( \frac{2\pi}{\lambda} \sin \theta \right) \sin \left( wt - \frac{2\pi x}{\lambda} \right)
\]

(22)

The relationship between \( \varepsilon \) and \( H \) is\(^5\)

\[
\frac{\varepsilon}{H} = \sqrt{\frac{\mu}{c}}
\]

(23)

\(^3\)Simon Ramo and T. R. Whinnery, Fields and Waves in Modern Radio, Chapter 11, J. Wiley and Sons, Inc.: New York, 1944.


Therefore, the magnetic intensity $H$ at the point $P$ is

$$H = \frac{\mu_0 I}{r} \cos \left( \frac{\pi}{r} \right) \sin (\omega t - \frac{2\pi x}{r})$$  \hspace{1cm} (24)$$

The polting vector is

$$P = \frac{C}{4\pi} E \times H$$ \hspace{1cm} (25)$$

which is the radiation through the unit of surface perpendicular to the radius vector.

We now consider the horizontal characteristic; the horizontal group of four horizontal dipoles as shown in Figure 7.

**Figure 7. The Horizontal Group of Four Horizontal Dipoles**

The spacing between the dipoles is represented by $d$. The angle between the normal line $Y$-axis and the direction of the radiation beam is indicated by $\theta$. According to equation (22) we get
\[
E = \frac{2Cl_0 \cos \left( \frac{\pi}{\lambda} \right)}{r} \left[ \sin \left( \omega t + \frac{2\pi}{\lambda} \cdot \frac{d \sin \theta}{2} \right) \right.
+ \sin \left( \omega t + \frac{2\pi}{\lambda} \cdot \frac{3d \sin \theta}{2} \right)
+ \sin \left( \omega t - \frac{2\pi}{\lambda} \cdot \frac{3d \sin \theta}{2} \right)
+ \sin \left( \omega t - \frac{2\pi}{\lambda} \cdot \frac{d \sin \theta}{2} \right) \left. \right]
\]
\[
= \frac{2Cl_0 \cos \left( \frac{\pi}{\lambda} \right)}{r} \left[ \sin \left( \omega t + \frac{\pi}{\lambda} \cdot \frac{d \sin \theta}{2} \right) \right.
+ \sin \left( \omega t + \frac{3\pi}{\lambda} \cdot \frac{d \sin \theta}{2} \right)
+ \sin \left( \omega t - \frac{\pi}{\lambda} \cdot \frac{3d \sin \theta}{2} \right)
+ \sin \left( \omega t - \frac{\pi}{\lambda} \cdot \frac{d \sin \theta}{2} \right) \left. \right]
\]
\[
= \frac{2Cl_0 \cos \left( \frac{\pi}{\lambda} \right)}{r} \left[ 2 \sin \omega t \right] \left[ \cos \left( \frac{\pi}{\lambda} \cdot \frac{d \sin \theta}{2} \right) + \right.
\cos \left( \frac{3\pi}{\lambda} \cdot \frac{d \sin \theta}{2} \right) \left. \right]
\]
\[
= \frac{8Cl_0 \cos \left( \frac{\pi}{\lambda} \right)}{r} \sin \omega t \left[ \cos \left( \frac{\pi}{\lambda} \cdot \frac{d \sin \theta}{2} \right) \right.
+ \cos \left( \frac{3\pi}{\lambda} \cdot \frac{d \sin \theta}{2} \right) \left. \right]
\]

For \( d = \frac{\lambda}{2} \) we get
\[
E = \frac{8Cl_0 \cos \left( \frac{\pi}{\lambda} \right)}{r} \sin \omega t \cos \left( \frac{\pi}{\lambda} \sin \theta \right) \quad (26)
\]

The relative horizontal characteristic of four dipoles radiating in phase is obtained as follows:
\[
k_{4h} = \frac{1}{\cos \theta} \cos^2 \left( \frac{\pi}{2} \sin \theta \right) \cos \left( \frac{\pi}{\lambda} \sin \theta \right) \quad (27)
\]

For eight dipoles and \( d = \frac{\lambda}{2} \) we get the relative horizontal characteristic as follows:
\[ K_{8h} = \frac{1}{\cos \theta} \cos^2(2 \sin \theta) \cos(\pi \sin \theta) \cos(2\pi \sin \theta) \cos(3\pi \sin \theta) \]  
(28)

For twelve dipoles and \( d = \lambda/2 \) we get the relative horizontal characteristic

\[ K_{12h} = \frac{1}{\cos \theta} \cos^2(2 \sin \theta) \cos(\pi \sin \theta) \cos(2\pi \sin \theta) \cos(3\pi \sin \theta) \]  
(29)

Therefore, we get a general formula

\[ K_{nh} = \frac{1}{\cos \theta} \cos^2(2 \sin \theta) \cos(\pi \sin \theta) \cos(2\pi \sin \theta) \cos(3\pi \sin \theta) \cos(n\pi \sin \theta) \]  
(30)

where \( n = 4, 8, 12, 16 \).

The polar plot of the relative horizontal power intensity radiation of our horizontal half-wave dipoles is shown in Figure 8. The range of \( K \) is a function of \( \theta \) for the four horizontal dipoles. We get the following table and field pattern in Figure 8.

<table>
<thead>
<tr>
<th>( \theta )</th>
<th>( K_4 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>100</td>
<td>0.805</td>
</tr>
<tr>
<td>150</td>
<td>0.6</td>
</tr>
<tr>
<td>300</td>
<td>0</td>
</tr>
<tr>
<td>450</td>
<td>-0.17</td>
</tr>
<tr>
<td>600</td>
<td>-0.08</td>
</tr>
<tr>
<td>750</td>
<td>-0.095</td>
</tr>
<tr>
<td>900</td>
<td>0</td>
</tr>
</tbody>
</table>
Considering the polar plot of relative horizontal power intensity of eight horizontal half-wave dipoles fed by currents equal in magnitude, we get the following table and field pattern in Figure 9.

<table>
<thead>
<tr>
<th>Angle (°)</th>
<th>K</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>10°</td>
<td>0.372</td>
</tr>
<tr>
<td>15°</td>
<td>-0.036</td>
</tr>
<tr>
<td>30°</td>
<td>0</td>
</tr>
<tr>
<td>45°</td>
<td>0.047</td>
</tr>
<tr>
<td>60°</td>
<td>-0.0515</td>
</tr>
<tr>
<td>75°</td>
<td>-0.092</td>
</tr>
<tr>
<td>90°</td>
<td>0</td>
</tr>
</tbody>
</table>

We compare these two field patterns in Figure 8 and Figure 9. We find that the field pattern of eight horizontal dipoles is much sharper than that of four dipoles.

We now calculate the vertical characteristic. The separate distances between the dipoles is denoted by d and the angle of elevation between the radiation beam and the Y-axis is indicated by Φ. In Figure 10 we suppose four vertical dipoles are parallel to Z-axis and it will not be reflected by the surface of the earth.
Figure 8. Horizontal Characteristic in Rectangular Coordinates
Figure 9. Horizontal Characteristic in Rectangular Coordinates
Figure 10. Vertical Diagram

According to equation 21 we get

\[
Ey = \frac{2CI_0}{r} \frac{\pi}{\cos \theta} \left[ \sin \left( \omega t - \frac{2\pi}{\lambda} \cdot \frac{d \sin \phi}{2} \right) + \sin \left( \omega t + \frac{2\pi}{\lambda} \cdot \frac{d \sin \phi}{2} \right) + \sin \left( \omega t - \frac{2\pi}{\lambda} \cdot \frac{3d \sin \phi}{2} \right) \right]
\]

\[
= \frac{2CI_0}{r} \frac{\pi}{\cos \theta} \left[ 2 \sin \omega t \cos \left( \frac{\pi d \sin \phi}{\lambda} \right) + 2 \sin \omega t \cos \left( 3\frac{\pi d \sin \phi}{\lambda} \right) \right]
\]

\[
= 2CI_0 \frac{\pi}{r} \frac{\cos \left( \frac{\pi \sin \theta}{\cos \theta} \right)}{\cos \theta} \cdot 2 \sin \omega t \left[ \cos \left( \frac{\pi d \sin \phi}{\lambda} \right) + \cos \left( 3\frac{\pi d \sin \phi}{\lambda} \right) \right]
\]
\[
\frac{E_y}{r} = 8\pi I_0 \cos(\omega t) \cos(\frac{2\pi d \sin \phi}{\lambda}) \cos(\frac{\pi d \sin \phi}{\lambda}) x \sin \omega t
\]

When \( \theta = 0 \) we get

\[
E_y = \frac{8\pi I_0 \sin \omega t \cos(\frac{2\pi d \sin \phi}{\lambda}) \cos(\frac{\pi d \sin \phi}{\lambda})}{r}
\]

The vertical field intensity is proportional to \( \cos(\frac{2\pi d \sin \phi}{\lambda}) \times \cos(\frac{\pi d \sin \phi}{\lambda}) \) if the distance between the dipoles is equal to \( \frac{\lambda}{2} \).

The relative horizontal field intensity is

\[
K = \cos(\pi \sin \phi) \cos(\frac{\pi \sin \phi}{2})
\]

if the earth near an antenna must be taken in account.\(^6\) The earth acts as a mirror which reflects the radiated wave upward but two difficult problems can result.

1. **Effect of earth conductivity**
2. **Effect of earth curvature**

Normally, we assume that the earth is plane and perfectly conducting. The perfectly reflecting surface of the earth is replaced by a mirror image of the overhead antenna arrangement reflected at the surface.

The radiated beam passes through the center of symmetry as in Figure 11 and the angle between the radiated beam and \( z \)-axis is designated by \( \phi \).

---

By using the same method of calculation as in equation (31) we obtain

\[ Ey = \frac{2C_0}{r} \cos \left( \frac{2}{r} \sin \phi \right) \left[ \sin(\omega t - \frac{2\pi}{\lambda} \sin \phi) ight. \\
\left. + \sin(\omega t - \frac{2\pi}{\lambda} \cdot 2d \sin \phi) + \sin(\omega t - \frac{2\pi}{\lambda} \cdot 3d \sin \phi) \\
+ \sin(\omega t - \frac{2\pi}{\lambda} \cdot 4d \sin \phi) - \sin(\omega t + \frac{2\pi}{\lambda} \sin \phi) \\
- \sin(\omega t + \frac{2\pi}{\lambda} \cdot 2d \sin \phi) - \sin(\omega t + \frac{2\pi}{\lambda} \cdot 3d \sin \phi) \\
- \sin(\omega t + \frac{2\pi}{\lambda} \cdot 4d \sin \phi) \right] \]

Let \( \Phi = 0 \) and \( d = \frac{\lambda}{2} \)

\[ Ey = -\frac{2C_0}{r} (2 \cos \omega t) \left[ \sin(\pi \sin \phi) + \sin(3\pi \sin \phi) \\
+ \sin(2\pi \sin \phi) + \sin(4\pi \sin \phi) \right] \]

\[ Ey = \frac{8C_0}{r} \cos \omega t \cos(\frac{\pi}{2} \sin \phi) \left[ \sin(\frac{3\pi}{2} \sin \phi) \\
+ \sin(\frac{7\pi}{2} \sin \phi) \right] \]

\[ = \frac{16C_0}{r} \cos \omega t \cos(\frac{\pi}{2} \sin \phi) \cos(\pi \sin \phi) \sin(\frac{5\pi}{2} \sin \phi) \]

(33)

The relative vertical characteristic of four dipoles radiating in phase is obtained

\[ Ky = \cos(\frac{\pi}{2} \sin \phi) \cos(\pi \sin \phi) \sin(\frac{5\pi}{2} \sin \phi) \]

(34)
The earth reflects perfectly, and therefore acts like a perfect conductor (infinite conductivity) or like a perfect insulator with an infinitely high dielectric constant. In this case the distance of the dipole from the ground is very important. The perfectly reflecting earth surface can be replaced, from the viewpoint of radiation, by the mirror image of the overhead antenna arrangement, reflected at the surface.
From equation 34, we get

<table>
<thead>
<tr>
<th>$\phi$</th>
<th>$\cos\left(\frac{\pi}{2} \sin \phi\right)$</th>
<th>$\cos\left(\frac{\pi}{2} \sin \phi\right)$</th>
<th>$\sin\left(\frac{5\pi}{2} \sin \phi\right)$</th>
<th>$\text{K}_y$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>5</td>
<td>0.99</td>
<td>0.96</td>
<td>0.63</td>
<td>0.6</td>
</tr>
<tr>
<td>7.5</td>
<td>0.98</td>
<td>0.92</td>
<td>0.845</td>
<td>0.74</td>
</tr>
<tr>
<td>10</td>
<td>0.96</td>
<td>0.86</td>
<td>0.981</td>
<td>0.81</td>
</tr>
<tr>
<td>15</td>
<td>0.92</td>
<td>0.69</td>
<td>0.892</td>
<td>0.57</td>
</tr>
<tr>
<td>20</td>
<td>0.86</td>
<td>0.47</td>
<td>0.438</td>
<td>0.177</td>
</tr>
<tr>
<td>30</td>
<td>0.71</td>
<td>0</td>
<td>-0.707</td>
<td>0</td>
</tr>
<tr>
<td>40</td>
<td>0.53</td>
<td>-0.44</td>
<td>-0.944</td>
<td>0.22</td>
</tr>
<tr>
<td>50</td>
<td>0.36</td>
<td>-0.74</td>
<td>-0.277</td>
<td>0.074</td>
</tr>
<tr>
<td>60</td>
<td>0.21</td>
<td>-0.91</td>
<td>0.5</td>
<td>0.095</td>
</tr>
<tr>
<td>70</td>
<td>0.1</td>
<td>-0.98</td>
<td>0.89</td>
<td>0.087</td>
</tr>
<tr>
<td>80</td>
<td>0.03</td>
<td>-1</td>
<td>0.996</td>
<td>0.03</td>
</tr>
<tr>
<td>90</td>
<td>0</td>
<td>-1</td>
<td>1</td>
<td>0</td>
</tr>
</tbody>
</table>

Figure 12 shows the vertical characteristic in rectangular coordinates, and in values which are linearly proportional to the field strength.

The angle of elevation for the main maximum was approximately 10 degrees when calculated.

The radiation pattern for a curtain antenna is chosen for the best frequency, orientation, and vertical angles of radiation. These factors determine the height, length, and orientation of the antenna.

The vertical angles 10 degrees as functions of layer height and
Figure 12. Vertical Characteristic on Rectangular Coordinates
HOP LENGTH ARE SHOWN AS FOLLOWS.\footnote{Ionospheric Radio Propagation, National Bureau Standards (U.S.) Circ. 462. An essential reference for high frequency propagation engineering.}

<table>
<thead>
<tr>
<th>Layer Height</th>
<th>Length for a Hop</th>
</tr>
</thead>
<tbody>
<tr>
<td>E layer (100 Km)</td>
<td>900 Km</td>
</tr>
<tr>
<td>F layer (250 Km)</td>
<td>1900 Km</td>
</tr>
<tr>
<td>F layer (300 Km)</td>
<td>2150 Km</td>
</tr>
<tr>
<td>F layer (350 Km)</td>
<td>2400 Km</td>
</tr>
<tr>
<td>F layer (400 Km)</td>
<td>2700 Km</td>
</tr>
</tbody>
</table>

We design a high gain curtain antenna to give a high gain in the forward direction and considerable suppression of backward radiation. A reflector is valuable for increasing the sensitivity of a dipole\footnote{Reference. For further information on dipole theory see: P. S. Carter, \textit{Proc. I. R. E.}, p. 1017, June, 1932. G. H. Brown, \textit{Proc. I. R. E.}, p. 86, January, 1937. Cheng-lee Chao, "High-gain Microwave Antenna Design," \textit{Chinese Radio Association}, June, 1937.} in a weak signal area, and the unidirectionality which arises with its use is particularly helpful in diminishing interference from sources in the rear. In order to explain the action of a reflector on a curtain antenna, suppose that a current $i_k$, flowing in the reflector, induces a voltage $e_1$ in the dipole and that a current is flowing in the dipole induces a voltage $e_2$ in the reflector. Then it can be proved by the Reciprocity Theorem that
\[
\frac{e_2}{i_d} = \frac{e_1}{i_r} = Z_m
\]  

\(Z_m\) is called the mutual-impedance between the dipole and reflector.

Consider now the array in Figure 13. If an incoming wave travelling in the direction AB generates a voltage \(e_h \cos \omega t\) in the dipole, then due to the time taken for the wave to travel from the dipole to the reflector, the wave generates a voltage \(e_r \cos(\omega t - n \lambda)\) in the reflector. Thus the current in the dipole and reflector are determined by the equations

\[
\begin{align*}
\quad \quad h_d e \cos \omega t &= i_d \left( Z_d + Z_1 \right) + i_r Z_m \\
\quad \quad h_r e \cos(\omega t - n \lambda) &= i_r Z_r + i_d Z_m
\end{align*}
\] (36)

Figure 13. The addition of a Reflector Increases the Sensitivity in the Direction AB and decreases it in the Direction NM
where \( i_d \) and \( i_r \) are the dipole and reflector currents, \( Z_d \) and \( Z_r \) are the respective impedances, and \( h_d \) and \( h_r \) are the effective coefficients, \( Z_l \) is the impedance of the transmission line. If the spacing between the dipole and reflector is \( n \lambda \),

\[
  i_d = h_d e \cos wt - \frac{Z_m}{Z_r} h_r e \cos \left( wt - n \lambda \right)
\]

equation (37) shows that a parasitic reflector has two general effects on a dipole.

1. It changes the apparent impedance of the dipole from its original value, \( Z_d \), to a new value

\[
  \left( Z_d - \frac{Z_m^2}{Z_d} \right)
\]

2. It causes an additional voltage of amount

\[
  -\frac{Z_m}{Z_r} h_r e \cos \left( wt - n \lambda \right)
\]

to be effectively generated in the dipole. In the latter expression, \( n \lambda \) represents the difference in phase between the incident waves reaching the dipole and the reflector.

The first of the two above listed effects causes a change in the apparent dipole impedance, which effect varies with frequency.

Examining the expression

\[
  -\frac{Z_m}{Z_r} h_r e \cos \left( wt - n \lambda \right)
\]

it is seen that the angle, \( n \lambda \), has opposite signs for waves coming from the directions AB and MN in Figure 13. Consequently, if the spacing
between the dipole and reflector is \( \gamma/4 \), the voltage due to the reflector will be a subtracting voltage for a wave travelling in the direction MN and an adding voltage for a wave travelling in the direction AB. This is an explanation of the unidirectionality of a dipole plus reflector array.

Reference to pp. 258-261, *Radio Antenna Engineering* by Dr. Laport, shows that the phase angle of \( Z_r \) changes rapidly with frequency. Furthermore, the change in phase of \( Z_m \) with frequency is such as to accentuate this effect. Consequently, the voltage due to the reflector

\[
- \frac{Z_m}{Z_r} h \cdot e \cos (\omega t - n\lambda)
\]

changes from an adding voltage to a subtracting voltage if the frequency is decreased by a few megacycles only.

If the induced current in the reflector is \( KI_0 \), the horizontal characteristic pattern is

\[
E_x = \frac{20I_0}{\pi x} \cos \left( \frac{\pi}{\cos \theta} \right) \left[ \sin (\omega t - \frac{2\pi}{\lambda} x \cdot \frac{n\lambda}{2} \sin \theta) \right.
\]

\[
+ K \sin (\omega t + \frac{2\pi}{\lambda} x \cdot \frac{n\lambda}{2} \sin \theta) \right]
\]

(38)

If we adopt the graphical method in a rough way to plot the interference pattern from the dipole and reflector which is the same as page 2.

Before we follow the former formulas to design a high-gain curtain antenna, as to my experience we have to consider several
factors as follows:

(1) The field intensity gain.

(2) The distance from the radio station to the receiving station in Km.

(3) The variation of the height of the ionosphere characteristic from time to time. (The information is offered from the Central Radio Propagation Laboratory, Washington, U.S.A.)

(4) The antenna field needs a cleared area for good wave reflection for a desired radiation angle.
CHAPTER IV

CONSTRUCTION OF THE CURTAIN ANTENNA AND IMPEDANCE-MATCHING TECHNIQUES

Figure 14. Directive High-gain Curtain Antenna (Shown Schematically)
In Figure 14, the directive high-gain curtain antenna set up in Formosa, China and directed toward Russia was to be determined experimentally. The directive radio system consists of 32 dipoles in two vertical planes each with 16 dipoles. There are two vertical planes, one which is excited by the transmitter and the other which is radiation coupled and serves as a reflector. The dipoles are connected to the vertical feed lines, and are excited in the same phase. Figure 14 shows the reflector dipoles which are not fed, and the tuning stub used to change the phase and magnitude of reflector currents.  

The antenna circuit from c to f in Figure 14 can be analyzed conveniently, using the corresponding equivalent circuit as follows.

---

A 600 ohms feeder and 495 ohms connecting bar is used in the antenna circuit, \( C_1, C_2 \) and \( C_3 \) are the insulator capacity instance. We assume for tuning frequency = 17.6 mc

\[
\begin{align*}
1_{bc} &= 324^\circ \\
1_{cd} &= 723^\circ \\
1_{de} &= 16^\circ \\
1_{ef} &= 16^\circ \\
X_{C_1} &= -j.38 \\
X_{C_2} &= -j.15 \\
X_{C_3} &= -j.316
\end{align*}
\]

The problem is to find a driving point b on the line where the conductance \( g/g_0 \) and the suscep-
tance \( B/g_0 \) is equal to \( .559+j.11 \).

(We usually use a V.H.F. admittance bridge, Wyne Kur Ltd. Type 801)

---


Figure 16. The Smith Chart
serial No. 321 and a bridge oscillator to measure the conductance and
susceptance.) Considering a lossless transmission line is applied on
the circuit. In Figure 16, we get the admittance of the point f

<table>
<thead>
<tr>
<th></th>
<th>I</th>
<th>II</th>
<th>III</th>
<th>IV</th>
</tr>
</thead>
<tbody>
<tr>
<td>((Y/Y_o)) d.p</td>
<td>(\frac{1}{2}(Y/Y_o)) d.p</td>
<td>(l_{bc})</td>
<td>(X_{c1})</td>
<td>(l_{cd})</td>
</tr>
<tr>
<td>(.599 + j 1.11)</td>
<td>(.299 + j .555)</td>
<td>(324^\circ)</td>
<td>(-j .38)</td>
<td>(723^\circ)</td>
</tr>
</tbody>
</table>

\[X_{c2} = (1.0 + j 1.11) \times \frac{1.67}{2.02}\]
\[-j .15 + j .92\]

\[Y_f/2.20 = 0.49 + j 0.08\]

Now change \(Y_f/2.02\) into \(Y_f/1.67\)

\[Y_f/Y_o = Y_f/1.67 = Y_f/2.02 \times 2.02/1.67 = .6 + j 0.0968\]

\(\frac{1}{2} Y_f/Y_o\) is a single branch admittance in the point f.

\[\frac{1}{2} Y_f/Y_o = 0.3 + j 0.0484\]

In Figure 18, we match the antenna circuit from load to generator.

![Figure 17. Equivalent Circuit](image-url)
Figure 18. The Smith Chart
On the point $f$, the $Y/Y_0$ of point $f$ is equal to $2(0.5 + j0) = 1$. In Figure 19, we match the feeder line from $f$ to the driving point $b$.

The East Bay and West Bay in the curtain antenna are designed in the same impedance, hence, the admittance in the driving point is equal to $Y/Y_0 = 2(0.8 + j1.18)$. In Figure 20 shows the total admittance matching of entire antenna.
Figure 19, The Smith Chart
Figure 20. The Smith Chart
The susceptance of the open stub is made equal but of opposite sign to the susceptance of the feeder at the point a, and then, the standing wave ratio on the transmission line is unity. If any impedance mismatch loss between the antenna and transmission line will subtract from its gain, maximum gain will be realized only when the mismatch is eliminated.

**Impedance - Measuring Equipment.**

A V.H.F. Admittance Bridge, Wyne Kur Lab. Ltd. Type 801 Serial No. 321 was used as the main equipment in this test. A General Radio Type
1330A, Serial number 580, Bridge Oscillator having an output up to 5 volts was used as the generator voltage for the bridge. A Hallicrafter model S40A receiver with seven-frequency bands was used as the bridge detector. In making the measurements over a range of frequencies from 2 MC to 50 MC, care was taken to get accurate values of the resistance component. Corrections were made to take into account the loss in the capacitor placed in series with the unknown. The loss of this capacity varied considerably over the frequency range. The generator signal was unmodulated and the null point was determined by the dip of the "R" meter in the Hallicrafter receiver.
CHAPTER V

DISCUSSION OF RESULTS

We can use the former formulas to design a high-gain curtain antenna, but according to my experience we found out several points to which we have to pay special attention.

The antenna system is in this discussion to include the coupling network between the transmitter and antenna. It is important to include this network since its losses may be an appreciable factor in determining the overall efficiency of the antenna system. In both the theoretical and practical case, as the antenna is made shorter, the radiation resistance decreases and the capacitive reactance increases. To transform this antenna impedance to a value that will properly load the transmitter, it is common practice to insert a coil in series with the antenna that will neutralize the capacitive reactance. It is desirable to take measures to increase the antenna terminal resistance and lower the capacitive-reactance component. Both of these conditions are improved by proper top loading of the antenna.

In actual practice the theoretical values of the field intensity can not be realized because of loss resistance in the conductors of the curtain antenna and coupling network, finite conductivity of the ground system, and dielectric losses in the insulator. The ratio of power radiated to power input to the antenna system can be taken as the criterion of over-all performance of the antenna system.

antenna system efficiency = \( \frac{P_T}{P_{in}} \times 100\% \)
The power radiated from the antenna can be determined by measuring the unattenuated root-mean-square field intensity at one mile and comparing it with the theoretical unattenuated field intensity. The antenna-system input power can also be considered as the transmitter output power since power supplies the losses in the antenna-system coupling network between the transmitter terminals and the antennas. The total power lost in the antenna system can be expressed as follows:

\[
\text{total power lost in the antenna system measured in watts} = \frac{\text{antenna-system power lost measured in watts}}{\text{coupling-network power lost measured in watts}} + \frac{\text{antenna-resistance power lost measured in watts}}{\text{insulator dielectric power lost measured in watts}} + \frac{\text{ground-system power lost measured in watts}}{\text{power lost measured in watts}}
\]

The band width is considered to be the frequency band within which the power is equal to or greater than one-half the power at resonance. One often hears the question whether one type of directive antenna is better than some other type. The answer usually depends on an economic condition. So a high-gain broad band curtain antenna was designed which was more convenient than the single frequency curtain antenna.

An experimental investigation was made of the radiation of a short-
wave directive antenna at Panchiao, Formosa, China. Measurements were made on the ground and also in the air, using a balloon. The results were compared with calculated radiation characteristics. In the case of the horizontal radiation characteristic the agreement between measurement and calculation is comparatively good.

For the vertical radiation characteristic curves calculations were made for two limiting cases:

a. Nonreflecting earth surface.

b. Perfectly reflecting earth surface.

The measured vertical characteristic agrees well with case b, from which it follows that the surface of the earth, under the conditions prevailing at Panchiao, shows a very high reflecting power. This is also confirmed by the fact that the field strength increased very rapidly from 0 at the surface of the earth as the height increases. It is also found that the radiated energy is highly concentrated in the desired direction in the horizontal as well as in the vertical plane.
LITERATURE CITED


