A Parabolic Cylinder Antenna with Very Low Sidelobes

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Abstract—A method for designing an offset-fed parabolic cylinder antenna which has sidelobes of −50 dB or less over a 15-percent frequency band is presented. The designed antenna was built and tested, and shown to have approximately −46-dB sidelobes, although over a slightly different band than designed for.

I. INTRODUCTION

A T PRESENT there is strong interest in the development of tactical radars which can maintain −50-dB azimuthal sidelobes over a 15-percent bandwidth. The principal limitations on developing a reflector antenna for this application are blockage (both in the near and Fresnel zones), reflector-surface tolerance errors, and spillover of the feed illumination beyond the edges of the reflector. In order to eliminate the adverse effects of blockage we have studied the radiation properties of an offset-fed parabolic cylinder. This structure was considered because it is simple to construct and may be readily scanned in the vertical plane. In this paper we will describe the design and test of this system and show that it meets the aforementioned sidelobe and bandwidth criteria.

II. ANALYTICAL PRELIMINARIES

We consider a section of a parabolic cylinder (satisfying the equation \( z = x^2/4F \)) with a line-source feed at its focus, as shown in Fig. 1. The electric field incident upon the reflector due to the line-source feed can be written as

\[
E_i = [\hat{x} F_x(\psi) + \hat{y} F_y(\psi)] e^{-ikr}, \tag{1}
\]

where \( k \) is the wavenumber, \( \hat{x} \) is a unit vector normal to the plane of the paper, \( \hat{y} = \hat{x} \cos \psi + \hat{z} \sin \psi \), \( \hat{x} \) and \( \hat{z} \) are unit vectors along \( x \) and \( z \), respectively, \( r \) is the distance from the center of the feed to an arbitrary point on the reflector, \( F_x(\psi) \) is the field pattern of the vertically polarized signal, and \( F_y(\psi) \) is the field pattern of the horizontally polarized signal. If this field is incident upon the reflector shown in Fig. 1 it is readily shown, by using the physical optics approximation, that the radiated magnetic field is given by [1, 2]

\[
H_z = A_0 \int_{\psi_0}^{\psi_1} d\psi \sec \left( \frac{\psi}{2} \right) \left[ \hat{y} F_x(\psi) - \hat{x} F_y(\psi) \right]
\cdot \left[ \cos \theta - \sin \theta \tan \frac{\psi}{2} \right]
\cdot \exp \left[ i2kF \tan \left( \frac{\psi}{2} \right) \right]
\cdot \left[ \sin \theta - \sin^2 \frac{\theta}{2} \tan \frac{\psi}{2} \right]. \tag{2}
\]

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Y_0 \) is the admittance of vacuum, \( R \) is the distance from the center of the reflector to the field point, \( \theta = x \cos \theta - \frac{z}{2} \sin \theta \) is a unit vector in the \( \theta \) direction, and \( \psi_0, \psi_1 \) are the angular coordinates of the two end points of the reflector, as shown in Fig. 1. In writing (2) we have tacitly assumed that the reflector is of infinite extent in the direction normal to the plane of the paper. We have shown that this approximates the azimuthal field pattern of a cylinder which is large (in comparison with the horizontal dimension) but finite in the vertical direction, provided there is sufficient vertical taper so that the illumination of the top and bottom edges is at least −20 dB (the top and bottom edges give rise to a cross-polarized field component).

By using (2) we have been able to obtain an excellent analytical approximation for the near-in sidelobe level (i.e., for \( \theta \ll 1 \)). If we assume that the feed is vertically polarized, has a symmetric radiation pattern, and is pointed at the center of the reflector, we find that the envelope of the relative sidelobe level of the radiation pattern at \( \theta \) is

\[
S(\theta) \equiv \frac{2}{\pi^2} \left( \frac{\lambda}{F} \right)^2 \left| \frac{F_y(\psi_1)}{F_x(\psi_0)} \right|^2 \left( \frac{1}{\sigma_\psi} \right)^2 \tag{3}
\]

where \( S(\theta) \equiv |H_z(\theta)/H_z(0)|^2 \), \( \lambda \) is the signal wavelength, \( \psi_B \) is the total angle subtended at the feed by the reflector, and \( \psi_\sigma = (\psi_0 + \psi_1)/2 \). Equation (3) is valid provided \( \theta < 1 \) but \( 2kF \sin \theta \gg 1 \). As an example of the utility of (3) let us suppose that we have a parabolic cylinder with \( \psi_0 = 0 \), \( \psi_1 = \pi/2 \), \( kF = 188.5 \), and an edge taper of −36 dB. The sidelobe level predicted at \( \theta = 5.7^\circ \) by (3) is −56.6 dB, whereas the exact computer evaluation of (2) gives −55.7 dB.
Equation (3) is the desired result and relates the envelope of the sidelobes at $\theta$ to the edge illumination, assuming that the reflector surface exactly satisfies the equation $z = x^2/4F$. Unfortunately, because of random errors, the reflector surface will never be exactly a parabolic cylinder, and the phase errors produced by manufacturing imperfections will also contribute to the sidelobe level. By extending the analysis of Ruze [4] we have shown [5] that the sidelobe level produced at $\theta$ by the random surface imperfections will be highest (i.e., worst case) if the lateral correlation length $a$ of the surface bumps is given by $a = 2k \sin \theta$. When $a$ has this value the average sidelobe level produced at $\theta$ by the reflector-surface imperfections is

$$\langle S_R \rangle = \frac{18.49 \delta^2 \cos^4 \left(\frac{\theta}{2}\right)}{\gamma A \sin^2 \theta},$$

where $A$ is the area of the reflector (projected onto the $z = 0$ plane), $\gamma$ is the illumination efficiency, and $\delta$ is the rms surface bump height. Because the probability density satisfied by the relative sidelobe level at $\theta$ is $p(S_R) = \langle S_R \rangle \exp (-\langle S_R \rangle/S_R)$ it can be shown that the probability that the sidelobe level due to surface bumps will be less than some specified value $S_0$, for all $\theta \geq \theta_0$, and is

$$P(S_R < S_0) = 1 - \exp \left[ \frac{-S_0 A \gamma \sin^2 \theta_0}{18.49 \delta^2 \cos^4 \left(\frac{\theta_0}{2}\right)} \right].$$

By specifying $P(S < S_0)$, $S_0$, $A$, etc., we can use (5) to calculate the required surface tolerances. For example, suppose that the aperture area is 225 ft$^2$, its efficiency is 0.42, the operating frequency is 3.34 GHz, and we desire that there be a 99-percent chance that any sidelobe beyond $\theta = \theta_0$ will be below $S_0 = 10^{-5}$ ($50$ dB). Then from (5) we find that the required surface tolerance is $\delta \leq 4.87 \times 10^{-3}$ in.

III. FEED CONSIDERATIONS

Once we have determined the tolerance requirements from (5) and the edge illumination requirements from (3), the problem is to design a feed which can simultaneously 1) illuminate the reflector with a relatively high efficiency, 2) provide the required edge illumination, and 3) limit the spillover of the feed radiation beyond the reflector edges. This last constraint is a somewhat difficult one because the gain of the parabolic cylinder is rather low. Therefore, if we require that all of the reflector sidelobes be more than 50 dB below the main beam level, and the gain of the reflector is 14 dB, then the sidelobes of the feed antenna must be more than 36 dB below the peak of the feed pattern. As an example, suppose that the parabolic cylinder occupies the angular region from $\psi = 0$ to $\psi = \pi/2$, and we point the center of the feed main beam at $\psi = \pi/4$. If the feed-antenna sidelobes were no problem we could easily achieve the required edge illumination with a conventional (parallel plate) waveguide feed with the aperture distribution $\cos (n \xi/b)$, where $b$ is the waveguide width and $\xi$ is the coordinate measured from the aperture center. However, as we clearly see from Fig. 2 (in Fig. 2, $\psi = \psi - (\pi/4)$) this leads to sidelobe levels which are too high. One way to achieve lower feed sidelobes is to allow higher order modes in the feed-aperture distribution. When the feed-aperture distribution is $\cos (n \xi/b) + 0.1666 \cos (3n \xi/b)$ it is evident from Fig. 2 that

![Figure 2. Radiation patterns of planar feed apertures for different mode distributions.](image-url)

the feed-antenna pattern is satisfactory. We have not extended our plots beyond $\psi = \pm 90^\circ$ because the expression

$$F_\psi(\varphi) = C_1 \cos \left( \frac{n \varphi}{\lambda} \sin \varphi \right) \left[ \frac{1}{(2b/\lambda)^2} - 1 \right]$$

$$- \left[ \frac{3a}{(2b/\lambda)^2} - 9 \right]$$

which we have used for $F_\psi(\varphi)$, is not valid beyond these values of $\varphi$. In (6) $C_1$ is a constant, and the feed-aperture distribution is assumed to be $\cos (n \xi/b) + \alpha \cos (3n \xi/b)$. A computer program has been written which utilizes (2) to calculate the radiation pattern of the reflector. When $F_\psi$ is given by (6) with $b = 2.7 \lambda$ and $\alpha = 0.1666$ it is found that when the feed is pointed at the center of a parabolic cylinder with $kF = 188.5$ and occupying the angular region $\psi = 0$ to $\psi = \pi/2$, the secondary radiation pattern is as shown in Fig. 3. We observe that all sidelobes are less than 50 dB below the main beam.

We would now like to put all our previous ideas together to design a parabolic cylinder antenna, with a $2^\circ$ (3-dB) beamwidth and $-50$-dB sidelobes, which can operate from 3.1-3.6 Hz. We choose a reflector which satisfies $z = x^2/4F$ and pick $\psi_0 = 5^\circ$, $\psi_1 = 80^\circ$, and $F = 8.802$ ft. This leads to a reflector diameter $D$ (see Fig. 1) of 14 ft. This reflector has a gain of 14 dB, so that in order to get sidelobes smaller than $-50$ dB we require a feed antenna such that 1) the edge illumination on the reflector is $-36$ dB relative to the illumination at the center, and 2) the spillover of the feed pattern beyond the edges of the reflector must be smaller than $-36$ dB. This can be achieved by a feed (located at the focus$^1$ of the reflector)

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$^1$ We have assumed that the center of the feed lies at the focal point of the reflector. Longitudinal and transverse feed displacements lead to pattern broadening and coma, respectively. In order to avoid these effects we require that longitudinal and transverse feed placement errors be less than 1.05 in.
PARABOLIC CYLINDER ANTENNA

Fig. 3. Radiation pattern of a parabolic cylinder with $\psi_0 = 0, \psi_1 = \pi/2, kF = 188.5$ illuminated by a feed aperture with $b = 2.7 \lambda$ and distribution $\cos(n\xi/b) + 0.1666 \cos(3n\xi/b)$.

with a planar aperture of width $b = 10.895$ in and with the aperture distribution $\cos(n\xi/b) + \alpha \cos(3n\xi/b)$. It can be shown [6] that the far-field sidelobe level produced by the scattering of the secondary radiation by this feed is at most $-61.57$ dB, so that blockage is clearly of no concern.

Typical feed radiation patterns $F_\alpha(\psi)$, where $\psi$ is measured relative to the angular coordinate of the center of the reflector, are shown in Fig. 4 for different values of $\alpha$, assuming the frequency is chosen at the center (3.35 GHz) of the required operating band. By making similar plots of the patterns for 3.1 and 3.6 GHz we have found that $\alpha = 0.14$ is a good compromise which leads to edge illuminations below $-36$ dB and spillover below $-36$ dB over the entire band of operation (3.1-3.6 GHz). If we choose $\alpha = 0.14$ and assume that the feed-aperture distribution remains equal to $\cos(n\xi/b) + 0.14 \cos(3n\xi/b)$ over the entire frequency band, we obtain the theoretical radiation pattern shown in Fig. 5 for $f = 3.35$ GHz. Observe that the 3-dB beamwidth is approximately 1.8° and the sidelobes are generally well below $-50$ dB over the entire band. In the next section we will discuss a method for achieving the required feed distribution.

IV. FEED DESIGN

We now discuss a feed design which can give the low spillover we require. In particular, in order to achieve the desired feed beamwidth we require that the feed aperture be $3.09 \lambda$ wide at 3.35 GHz, with an aperture distribution of $\cos(n\xi/b) + 0.14 \cos(3n\xi/b)$. In addition, we require that the feed be compact (less than 1 ft deep) and not contain any lenses. These requirements force us to exclude horns because they would have to be too deep in order to have small quadratic phase errors across the aperture. The design we actually chose was suggested by W. Rotman [7] and is shown in Fig. 6. Because $b = 3.09 \lambda$ it is clear that all modes up to the TE_{60} can propagate, whereas we desire to have only the TE_{10} plus TE_{30}. If the feed probes are symmetrically located about the center of the guide then the TE_{20}, TE_{40}, and TE_{60} will not be excited. Moreover, if the probes are located at the zeros of the TE_{60} mode, as shown, then this mode will not be excited. Therefore only the TE_{10} and TE_{30} will exist. We will now calculate the probe currents required to produce the desired fractions of TE_{10} and TE_{30} modes.
In the region \( z < 0 \) in Fig. 6 we can write the electric field as

\[
E = \hat{y} \sum_{n=1}^{\infty} C_n \sin \left( \frac{n \pi x}{b} \right) \sinh \Gamma_n (z + l),
\]

where \( \hat{y} \) is a unit vector, \( \Gamma_n^2 = (n \pi/b)^2 - (2 \pi/\lambda)^2 \), and \( C_n \), \( D_n \), \( R_n \) are unknown coefficients. Because the aperture is more than \( 3 \lambda \) wide it can be shown that the reflection coefficient \( R_n \) will be small for both the TE\(_{10}^1 \) and TE\(_{30}^1 \) modes. If we ignore \( R_n \) and require that the electric field be continuous across the \( z = 0 \) plane we get

\[
C_n \sinh \Gamma_n l = D_n.
\]

If we assume that the feed probes are infinitesimally thin we can write their total current density as

\[
J = \hat{y} \sum_{p=1}^{4} \beta_p \delta(x) \delta(x - x_p)
\]

where \( \beta_p \) is to be calculated and \( x_p = pb/5 \). If we substitute (10) into Maxwell's wave equation, multiply by \( \sin(n \pi x/b) \), and integrate we get

\[
\frac{d^2 e_n}{dz^2} + \left( k^2 - \frac{n^2 \pi^2}{b^2} \right) e_n = i \omega \mu_0 \delta(x) \sum_{p=1}^{4} \beta_p \sin \left( \frac{n \pi x_p}{b} \right),
\]

where

\[
e_n = \int_{0}^{b} dx E(x, z) \sin \left( \frac{n \pi x}{b} \right).
\]

\( \omega \) is the radian frequency, and \( \mu_0 \) is the permeability. Upon integrating (11) from \( z = 0^- \) to \( z = 0^+ \) we find

\[
\left( \frac{d e_n}{dz} \right)_{0^+} = i \omega \mu_0 \sum_{p=1}^{4} \beta_p \sin \left( \frac{n \pi x_p}{b} \right).
\]

If we use (7) and (8) in (12) we get

\[
\Gamma_n D_n + C_n \Gamma_n \cosh \Gamma_n l = -i \omega \mu_0 \sum_{p=1}^{4} \beta_p \sin \left( \frac{n \pi x_p}{b} \right).
\]

Finally, upon combining (9) and (13) we obtain

\[
D_n = \frac{-i \omega \mu_0}{\Gamma_n (1 + \cosh \Gamma_n l)} \sum_{p=1}^{4} \beta_p \sin \left( \frac{n \pi x_p}{b} \right).
\]

When there is perfect symmetry, and \( x_p = bp/5 \), we find that only \( D_1 \) (TE\(_{10}^1 \) mode) and \( D_3 \) (TE\(_{30}^1 \) mode) are nonzero. From (14) we can solve for \( \beta_1 = \beta_4 \) and \( \beta_2 = \beta_3 \) in terms of \( D_1 \) and \( D_3 \). If we require that the aperture distribution consist of a TE\(_{10}^1 \) plus some fraction \( \alpha \) of a TE\(_{30}^1 \) mode then

\[
D_3 = \alpha \exp \left( (\Gamma_3 - \Gamma_1) l \right) D_1.
\]

We then find from (14) and (15) that

\[
\beta_1 = 0.47022 + 0.76084 \Lambda
\]

\[
\beta_2 = 0.76084 - 0.47022 \Lambda
\]

where

\[
\Lambda = \left( \frac{\Gamma_3}{\Gamma_1} \right) \left( \frac{1 + \coth \Gamma_3 l}{1 + \coth \Gamma_1 l} \right) \alpha \exp \left( (\Gamma_3 - \Gamma_1) l \right).
\]

For a feed operating at 3.35 GHz with \( L = 8.12 \) in, \( l = 0.88 \) in, and \( b = 10.9 \) in we find that \( \beta_1 = \beta_4 = 0.588909 \exp(-i 16.07193^\circ) \) and \( \beta_2 = \beta_3 = 1 \). Now that the feed parameters have been specified we must estimate the feed bandwidth and assess the effect of errors in the probe location and excitation on the feed-radiation pattern. If we assume that the probe excitations are relatively insensitive to frequency changes it is clear that the primary factor which influences the feed bandwidth is the relative phase between the TE\(_{10}^1 \) and TE\(_{30}^1 \) modes at the feed aperture. From (15) it is clear that at the center frequency this phase difference is

\[
\Delta \Phi_0 = \frac{2 \pi L}{\lambda_0} \left[ \left( 1 - \frac{\lambda_0^2}{4 b^2} \right)^{1/2} - \left( 1 - \frac{9 \lambda_0^2}{4 b^2} \right)^{1/2} \right]
\]

where \( \lambda_0 \) is the wavelength corresponding to the center frequency. If we now let \( \lambda = \lambda_0 + \delta \lambda \), and also use the fact that \( \lambda_0 \ll 2b, 3 \lambda_0 \ll 2b \), we find for the phase error (i.e., the actual phase difference minus the desired phase difference) between TE\(_{10}^1 \) and TE\(_{30}^1 \) modes at the aperture

\[
\delta \phi \equiv 2 \pi \left( \frac{L}{b} \right) \left( \frac{\lambda_0}{b} \right) \left( \frac{\delta \lambda}{\lambda_0} \right). \tag{20}
\]

A computer study of the feed-radiation pattern indicates that the pattern is nearly insensitive to the phase change \( \delta \phi \) so long as \( \delta \phi < \pm 8^\circ (\pm 0.14 \text{ rad}) \). If we use this result in (20) we find that \( \delta \lambda/\lambda_0 = \pm 0.093 \). Consequently, the total bandwidth of the feed is at least 18.6 percent.

We have calculated the radiation pattern and have used Monte Carlo methods to study the effect of all the errors on this pattern. We have found that the following tolerances are sufficient to achieve the edge illumination and spillover requirements:

\[
\frac{\epsilon_{p/b}}{\epsilon_0} = \pm 0.0015 \tag{21}
\]

\[
\delta \varphi_p = \pm 2.5^\circ \tag{22}
\]

\[
| r_p | = 1 \pm 0.035. \tag{23}
\]

2 That is, at \( \lambda = \lambda_0 + \delta \lambda \) the aperture distribution is \( \cos \left( \pi x/b \right) + \alpha \cos \left( 3 \pi x/b \right) \exp(i \delta \phi) \) instead of \( \cos \left( \pi x/b \right) + \alpha \cos \left( 3 \pi x/b \right) \), as it is at \( \lambda = \lambda_0 \).
In particular, when we Monte-Carloed from a distribution in which $-0.0015 \leq \epsilon_p/b \leq 0.0015$, $-2.5^\circ \leq \delta \phi_p \leq 2.5^\circ$, and $0.965 \leq |\tau_p| \leq 1.035$ the spillover was greater than $-35$ dB in only about eight percent of the trials.

V. MEASURED DATA

Although we have designed the antenna to operate over the frequency band from 3.1–3.6 GHz, it is convenient, in order to limit the size of the structure, to scale up the frequency by a factor of 3 when taking measurements. Also, in order to simulate the two-dimensional nature of the parabolic cylinder we have built the pillbox structure shown in Fig. 7. The purpose of the flanges is to provide a smooth transition from the parallel-plate structure into free space.

The feed structure was constructed in two steps: first the probe-feed waveguide shown in Fig. 6 was constructed (with $l = 0.543$ in, $L = 2.70$ in, and $b = 3.63$ in), and then the feed network was built. The fixed-value power divider in the feed network consisted of a hybrid, a two-way power divider, and a cable length (for differential phase shift). The two higher power ports fed the two inner probes of the waveguide feed, whereas the remaining ports had lengths of cable designed to provide a 16.7° phase shift and fed the two other probes in the waveguide in Fig. 6. The phase and amplitude response of a number of hybrids and power dividers was carefully measured and the best components selected; a similar procedure was followed with the line lengths used for the 16.7° phase shift. As a result of this care the measured amplitude and phase values of the network were within 0.2 dB and 1.5°, respectively, of the design values. However, it is important to note that the aforementioned values are at the input to the feed probes and the current amplitudes and phases may be different from these, because of mutual coupling between the probes and geometrical effects (i.e., the outer probes are nearer to the side walls than the inner ones). In order to assess the effect of the different environment seen by the inner and outer probes we have measured the self impedance of each probe, with all the other probes terminated in a matched load. The two inner probes (2 and 3) had nearly an identical impedance plot as a function of frequency. Also, the two outer probes had nearly the same impedance characteristics, but these differed somewhat from those of the two inner probes.

![Fig. 7. Pillbox structure used to test parabolic cylinder.](image)

![Fig. 8. Feed radiation pattern at 10.7 GHz.](image)

Because the frequency characteristics of all the probes are not identical, we expect that the radiation pattern of the feed waveguide will change somewhat as the frequency is varied. Consequently, the actual feed bandwidth may not be as large as predicted by using (20), because in deriving (20) we assumed that the probe characteristics were frequency independent.

We next measured the radiation pattern of the feed antenna as the frequency was varied from 9.1 to 11.3 GHz in increments of 0.1 GHz. A typical pattern is shown in Fig. 8. In general, the feed response fails to meet the criteria (e.g., sidelobes below $-36$ dB) determined in Section III at the lower edge of the band but is excellent at the middle and upper portion of the band.

Upon completing the test of the feed antenna we next proceeded to measure the far-field radiation pattern of the parabolic cylinder antenna. However, before making these measurements we placed absorber along the side walls of the pillbox in Fig. 7, so as to substantially reduce their effect. It would be desirable to completely eliminate the side walls, but of course they were necessary for structural support. The measurements were made by employing the “round-house” antenna positioner at our Ipswich, MA, antenna range, with the parabolic-cylinder pillbox used as a receiving antenna. The antenna was mounted on the positioner and then rotated through 90° as shown in Fig. 9, so that the azimuth plane of the parabolic cylinder is actually in the vertical direction. The main beam of the parabolic cylinder was first pointed at the transmitter (located ½ mi away) and then rotated upwards to obtain the sidetone structure for $\theta = 0$ to 90°. This procedure was then repeated to obtain results for $\theta = 0$ to $-90^\circ$. Because the main beam does not hit the ground there is no power scattered into the sidetones by main beam ground clutter.

Some measured radiation patterns of the parabolic cylinder antenna are shown in Figs. 10–12. Observe that the sidetones of the radiation pattern are approximately $-50$ dB or below at the higher edge of the band but above $-50$ dB at the lower edge.

4 In addition, chokes were placed along the flared edges of the pillbox in order to reduce the backscatter from the mount and supporting struts.

3 The allowed tolerances were 0.3 dB and 2.5°.
VI. CONCLUSION

We have demonstrated that it is possible to design and build an offset-fed parabolic cylinder which has azimuthal sidelobes which are 46 dB or lower over a frequency band of approximately 15 percent. However, because the approximations we made in analytically modelling the feed are not strictly valid, and also because of tolerance errors in actually building the feed, the actual band over which all sidelobes were nearly -50 dB (actually -46) or below was from 9.8–11.3 GHz, rather than the band from 9.3–10.8 GHz for which we designed.

REFERENCES

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