Nov. 28, 1967 J. A. ALGEO 3,355,738

MICROWAVE ANTENNA HAVING A CONTROLLED PHASE DISTRIBUTION

Filed Nov. 9, 1964 10 Sheets-Sheet 1

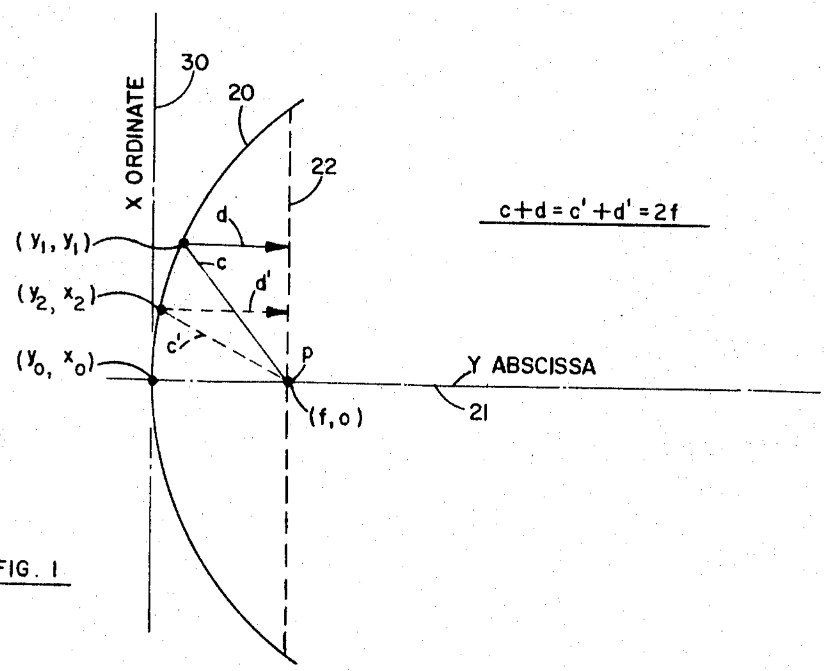

FIG. 1

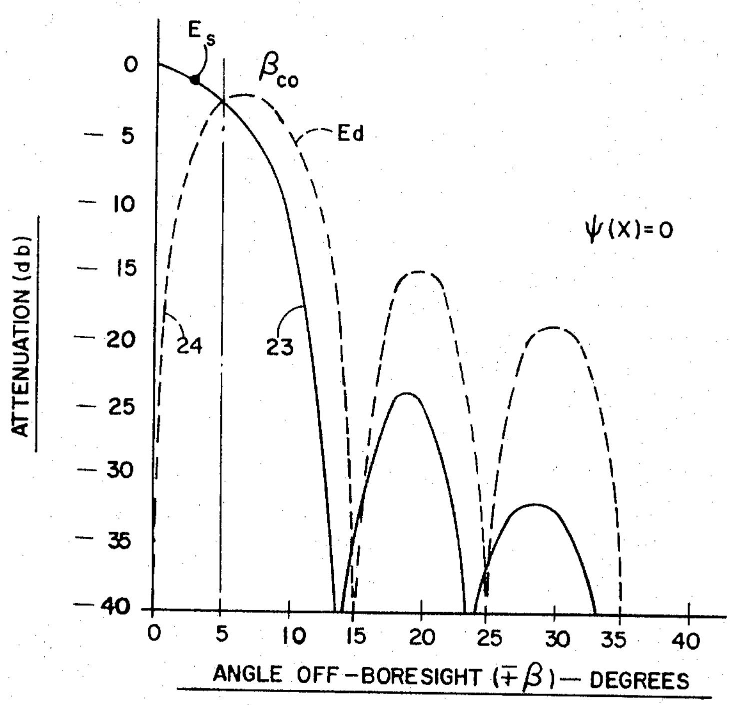

FIG. 2

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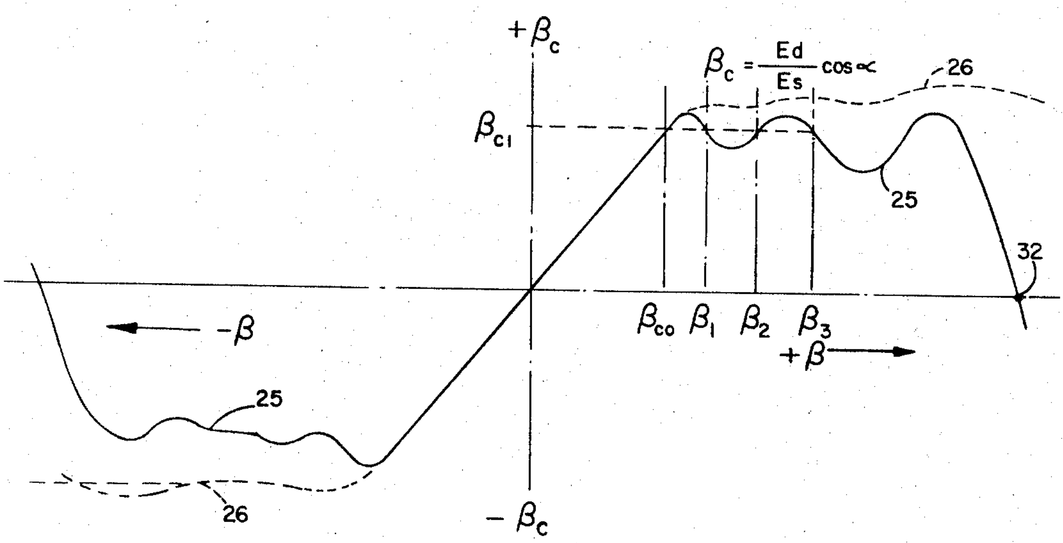

FIG. 3

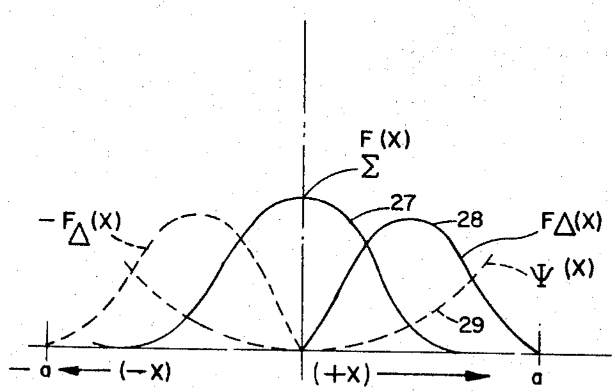

FIG. 4

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MICROWAVE ANTENNA HAVING A CONTROLLED PHASE DISTRIBUTION

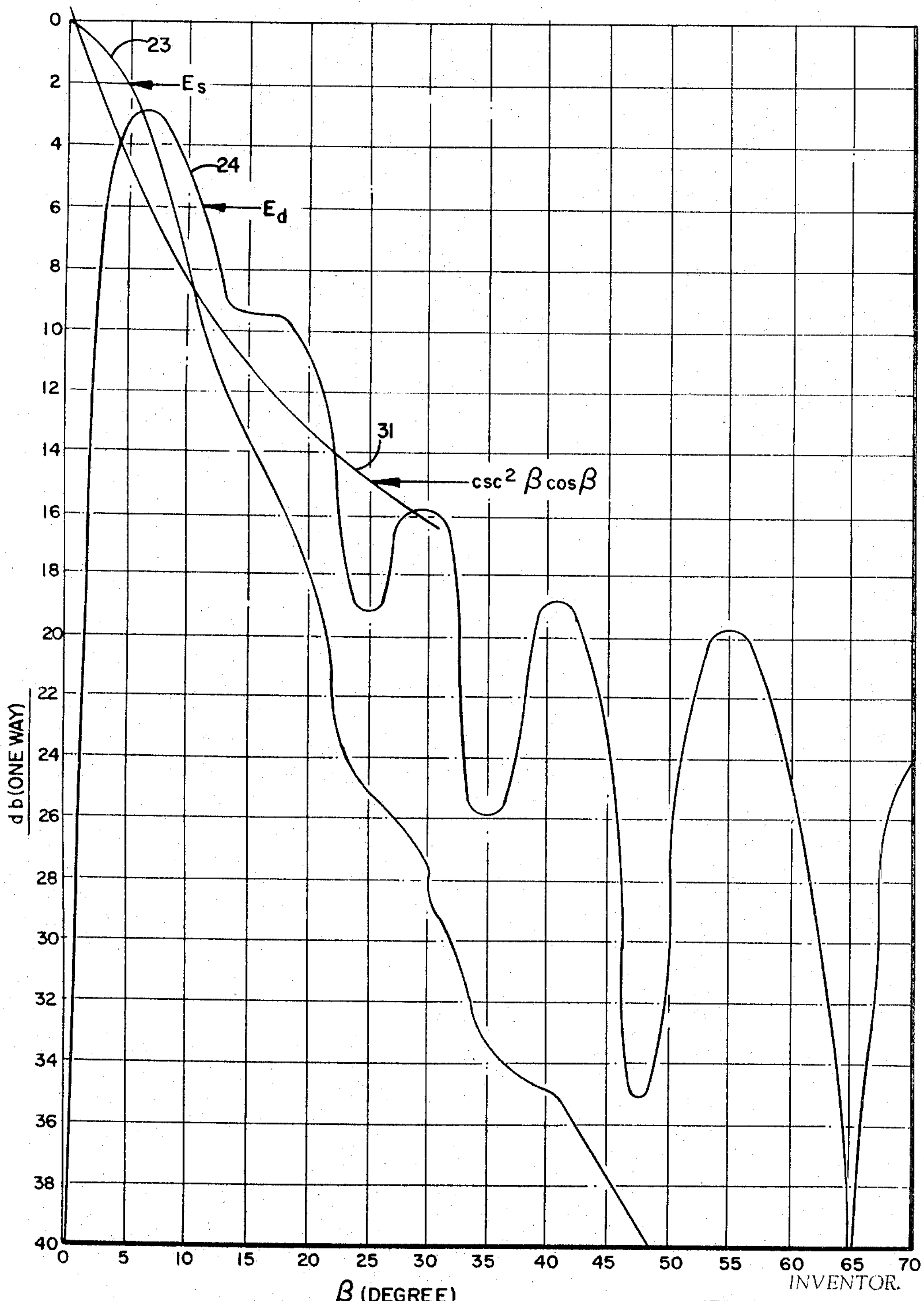

FIG. 5

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MICROWAVE ANTENNA HAVING A CONTROLLED PHASE DISTRIBUTION

Filed Nov. 9, 1964 10 Sheets-Sheet 4

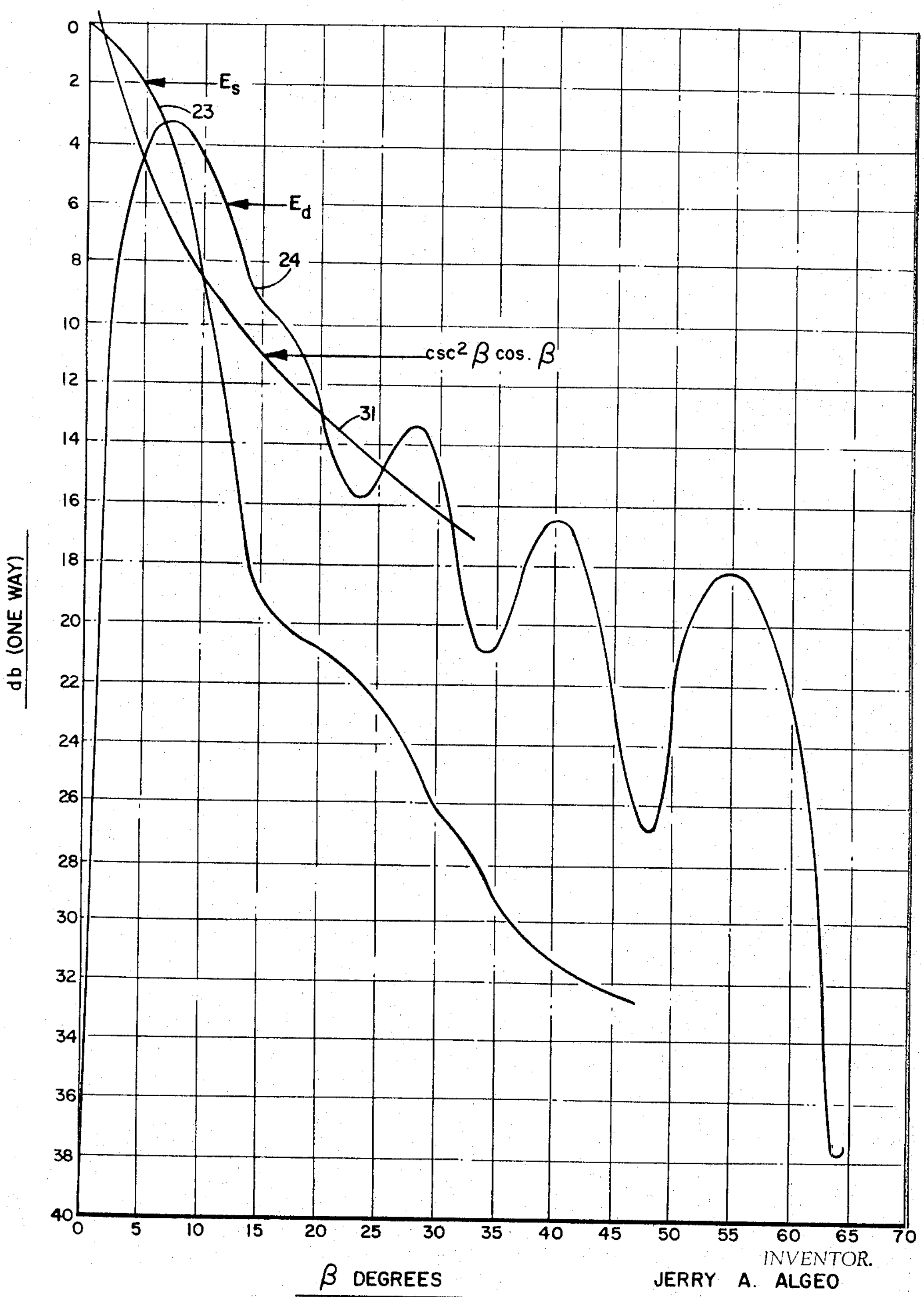

FIG. 6

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MICROWAVE ANTENNA HAVING A CONTROLLED PHASE DISTRIBUTION

Filed Nov. 5, 1964 10 Sheets-Sheet 5

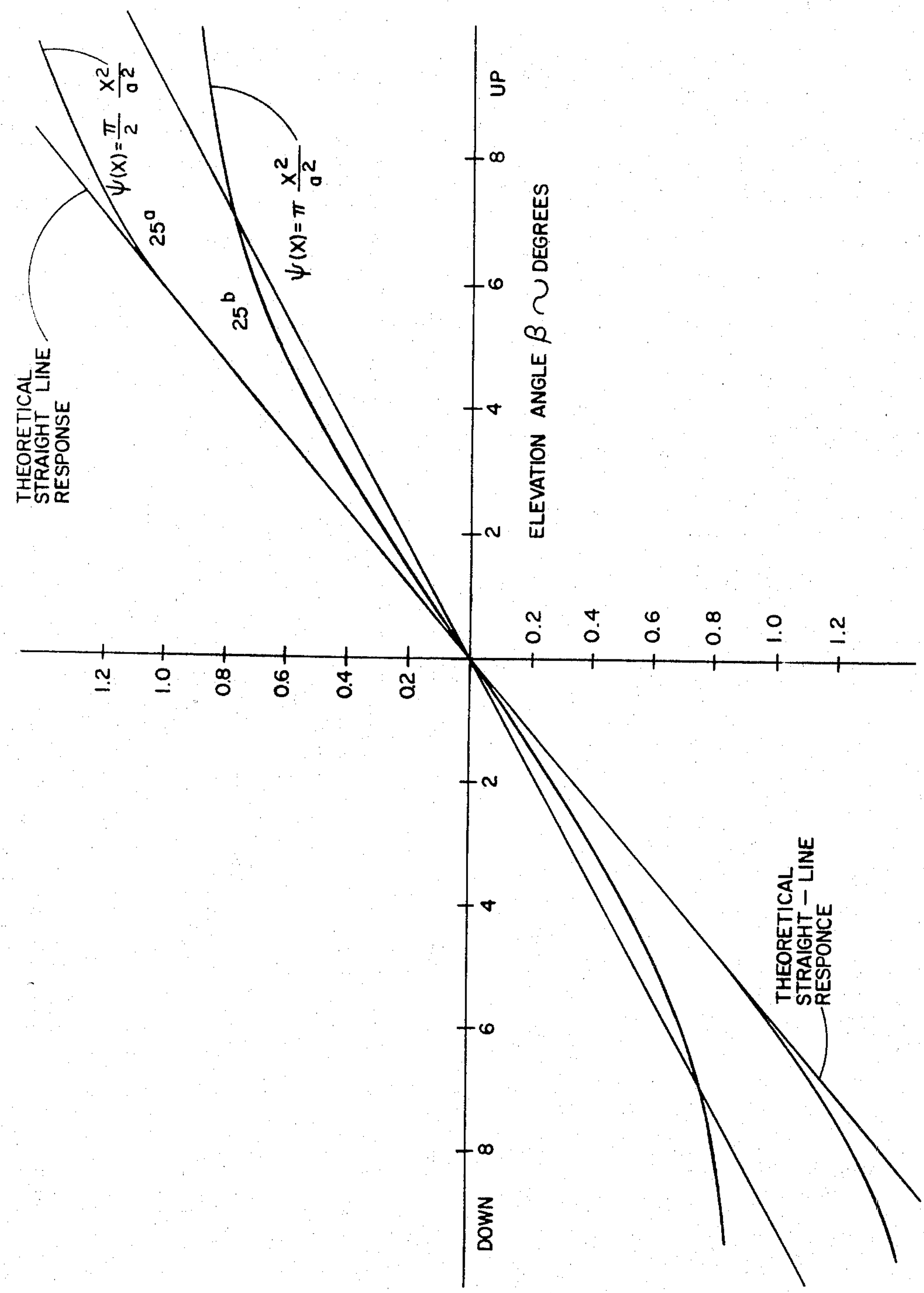

FIG. 7

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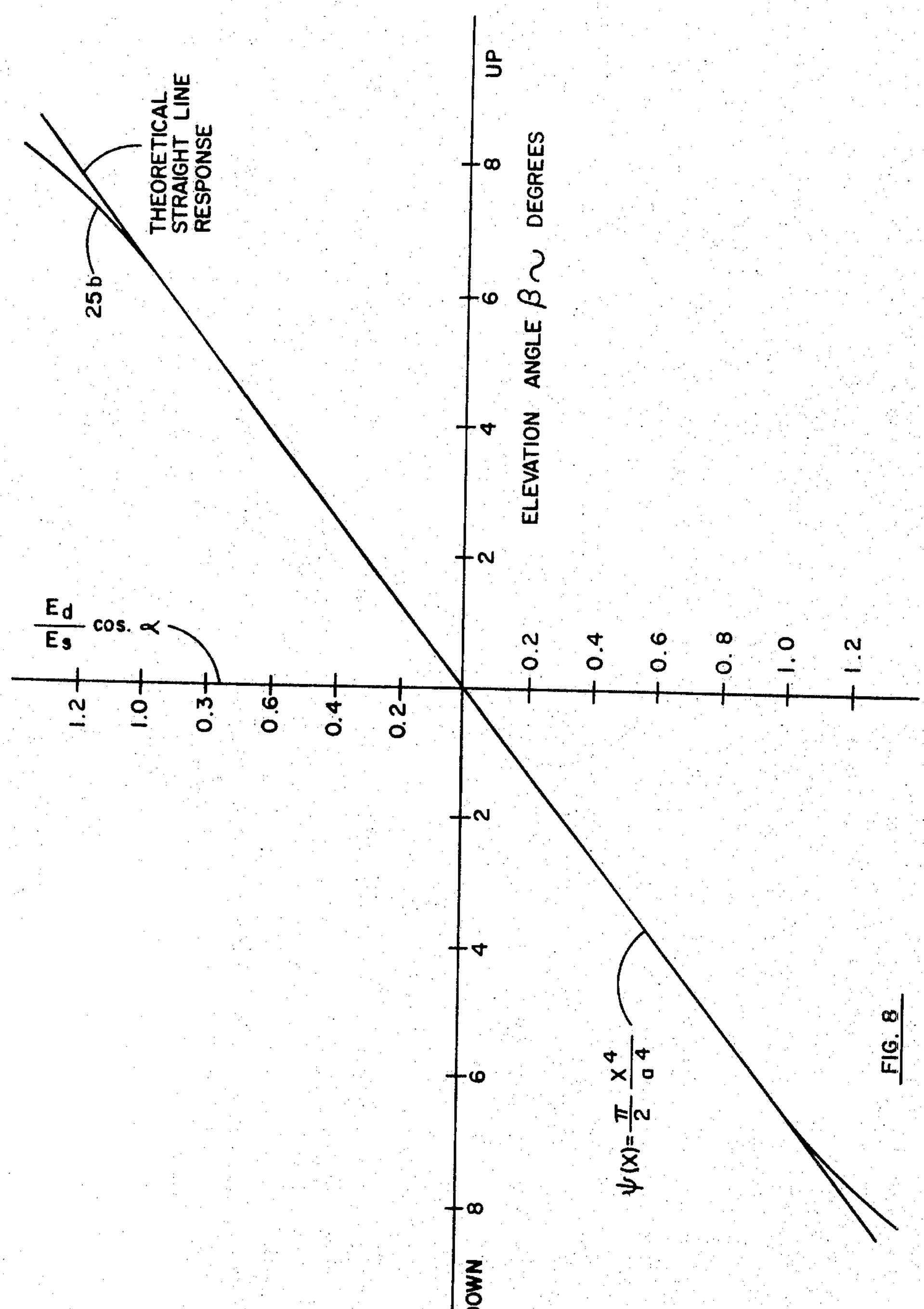

FIG. 8

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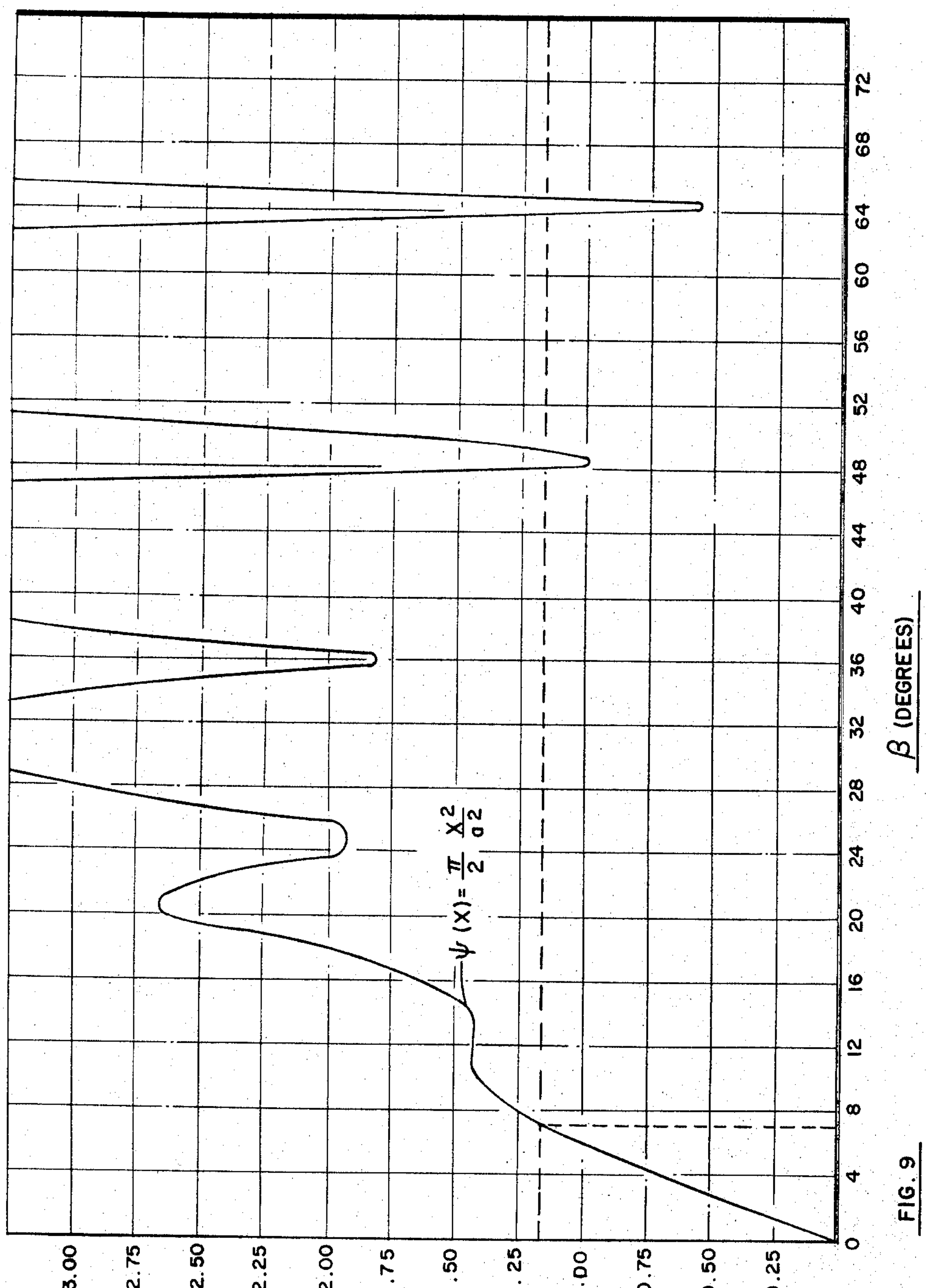

FIG. 9

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MICROWAVE ANTENNA HAVING A CONTROLLED PHASE DISTRIBUTION

Filed Nov. 3, 1964 10 Sheets-Sheet 8

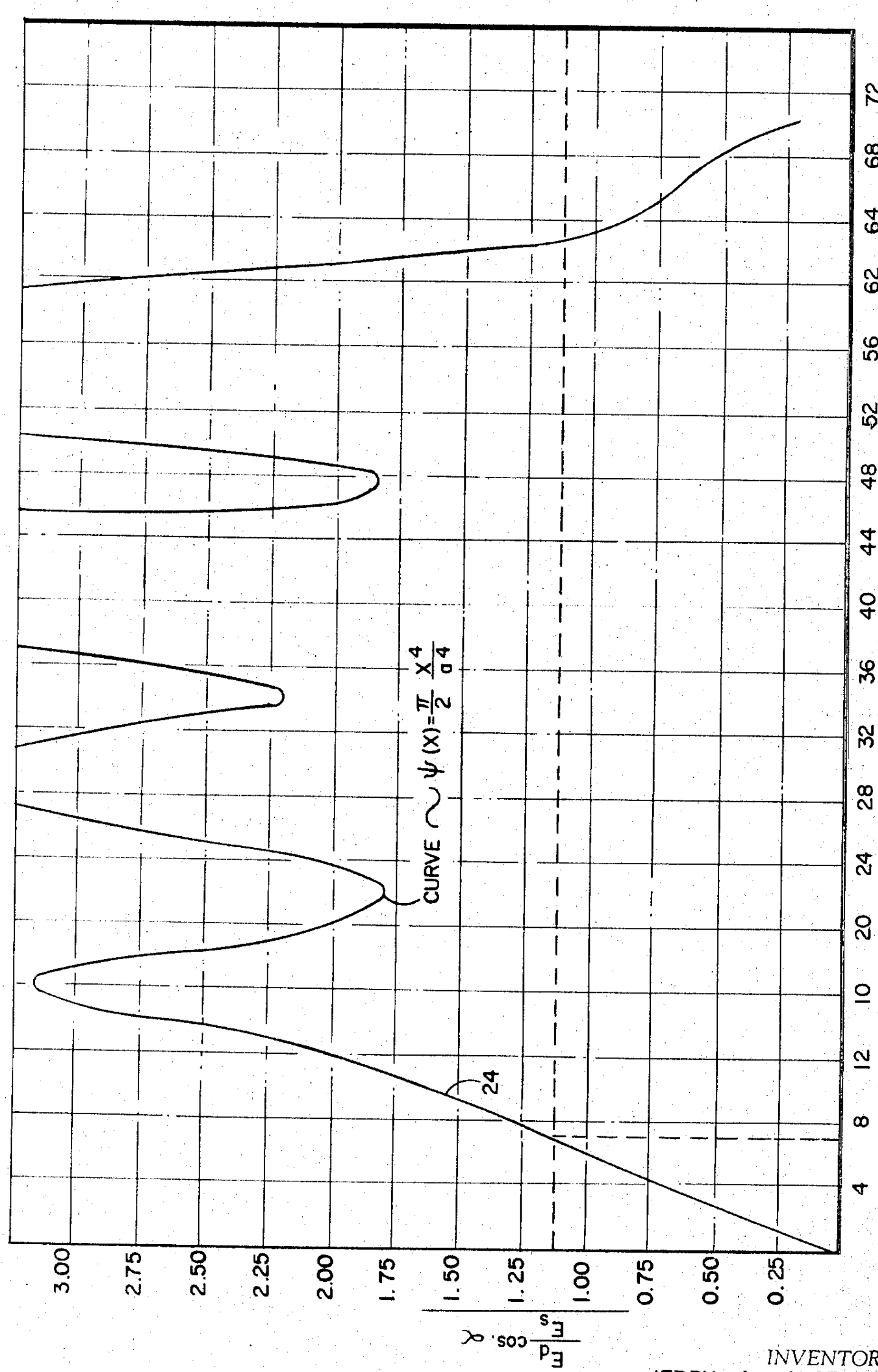

FIG. 10

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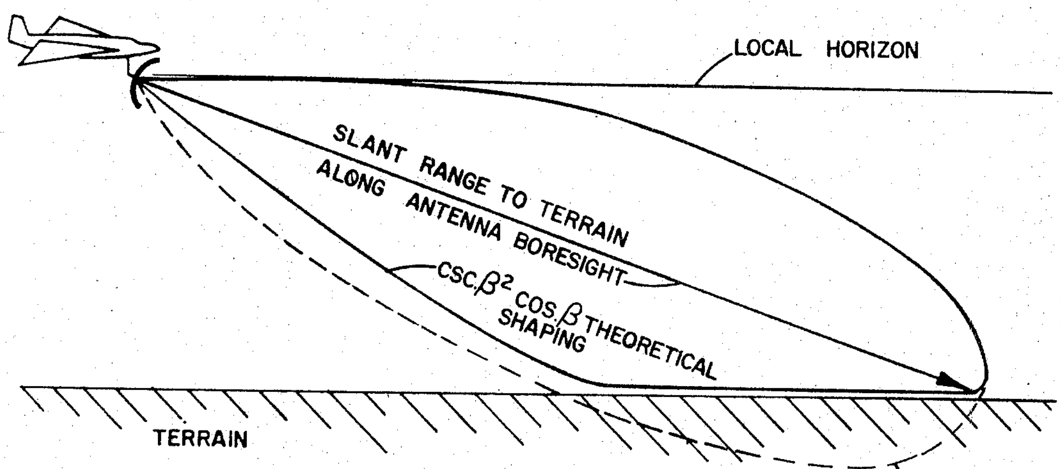

FIG. 11

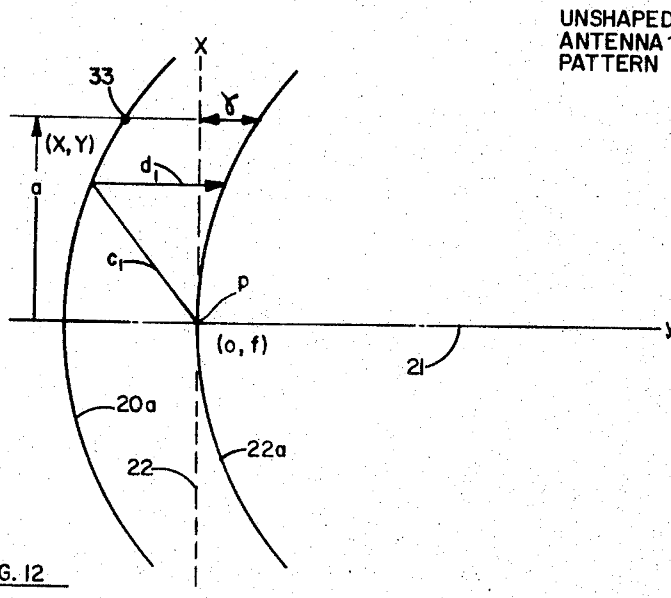

FIG. 12

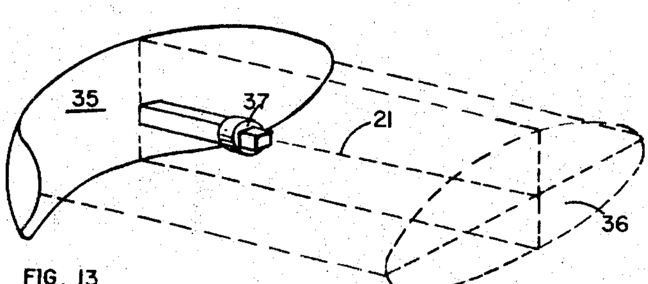

FIG. 13

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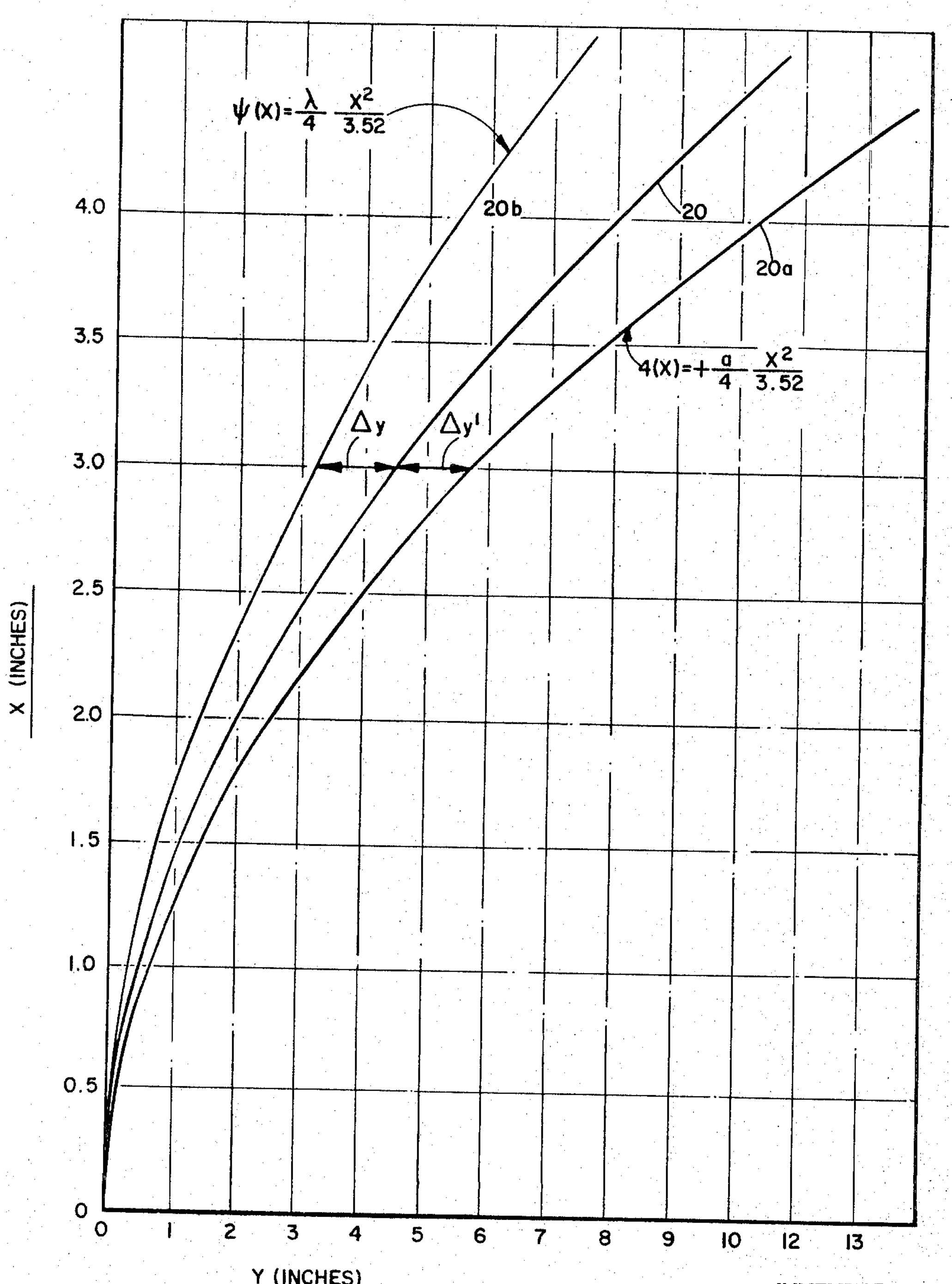

FIG. 14

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3,355,738

MICROWAVE ANTENNA HAVING A CONTROLLED PHASE DISTRIBUTION

Jerry A. Algeo, Buena Park, Calif., assignor to North American Aviation, Inc.

Filed Nov. 9, 1964, Ser. No. 409,912
7 Claims. (Cl. 343—779)

ABSTRACT OF THE DISCLOSURE

A reflector type microwave antenna for monopulse applications, in which the curvature of the reflector is varied from a true parabola to provide a controlled phase distribution corresponding to a continuous even-valued function of a power greater than unity of the aperture dimension. Such controlled phase distribution reduces the beta-anomalies occurring in the monopulse data processor with which such monopulse antenna is intended to cooperate.

Background of the invention

Monopulse systems for measuring the target angle or angle-off-boresight of a detected target (situated within the antenna beamwidth) in a given plane containing the antenna boresight axis or a radiation axis of symmetry, employ an antenna having at least two feedhorns to provide two received signals. Also employed are conventional sum-and difference monopulse receivers responsive to the sum of and the difference between the two received signals to provide a target angle signal indicative of the angle of a detected target off the antenna boresight axis. The sum signal itself is ordinarily used for target display purposes.

In the design of such monopulse receiver systems, the aperture of a prior art conventional antenna may have a rectangular shape which provides a uniform field distribution across the aperture. Such rectangular aperture normally has a substantial associated side-lobe pattern or response. Antennas having such antenna side-lobe response or radiation pattern provide illumination of targets lying within such side-lobes; and are, therefore, sensitve to energy reflected from such illuminated target. Further, such a rectangular aperture normally provides more than a single null in the response of the difference signal as a function of target angle-off-boresight which produces certain anomalies in the determination of the target angle-off-boresight from such signal. For example, the detection of a target lying within the side-lobe response of an antenna may result in the generation of target angle signals falsely indicating a target angle-off-boresight lying within the angular width of the antenna main lobe response.

A discussion of such ambiguities, together with one means of attempting to reduce such ambiguities in the monopulse technique for measuring target angle-off-boresight, is described in copending U.S. Patent Application Serial No. 258,183 filed February 13, 1963, by R. E. Hovda et al., now Patent No. 3,283,322 and assigned to North American Aviation, Inc., assignee of the subject invention. Such means comprises shaping the amplitude distribution or combined aperture field distributon from a conventional rectangular shape to achieve a gabled amplitude distribution. Such gabled amplitude aperture distribution is achieved by physically shaping the frontal area of an antenna reflector or physically shaping of the antenna feedhorn apertures or correcting the dipole elements of a flat plate monopulse antenna. Such shaping of the amplitude distribution is to be distinguished from shaping the phase distributon across the antenna aperture. Hence, the teachng of the above cited U.S. application Serial No. 258,183, filed February 13, 1963, relates to the use of an antenna having a conventional planar phase front across the aperture thereof, and for which the amplitude distribution is changed from a rectangular shape to a gabled shaped. Such amplitude shaping technique, for reducing ambiguities in the monopulse receiver difference channel output, is extremely difficult and expensive to employ, requiring a great deal of empirical effort to realize a satisfactory configuration for a specific design. Further, such reflector area shaping, and like means for effecting amplitude distribution shaping of the effective antenna aperture, has little practical effect upon phase-distribution phenomenon which contribute to the monopulse anomalies.

By means of the concept of the subject invention, there is provided efficient means for achieving a controlled phase distribution across the antenna aperture, whereby the monopulse anomalies are substantially and effectively reduced.

In a preferred embodiment of the subject invention, there is provided a monopulse receiving antenna for producing a first and second microwave output in response to received microwave signals, and including means for providing a selected aperture phase distribution in the antenna. Such selected aperture phase distribution means comprises a curved reflecting surface defining a central axis originating at a point in such curved surface and oriented perpendicularly thereto at the point of origin, a feed point being located on the central axis and spaced apart from the curved surface. There is further provided an array of at least two feedhorns mounted at the feed point for illuminating the curved surface. The selected curvature of the curved surface cooperates with the feedhorns to provide an aperture phase distribution which varies from a planar phase front as an even valued function of a power or exponent greater than unity of the aperture dimension. The effect of such selected phase distribution is to control both the far field amplitude pattern of the sum and difference of the two received signals, together with the time-phase angle between the sum and difference, so as to substantially reduce anomalies occurring in the monopulse measurement of target angle-off-boresight. Hence, improved performance in reducing monopulse signal anomalies, is achieved. Further, the required curvature of the curved reflector for effecting such result, for a specific application, can be accurately specified and efficiently determined by convenient analytical techniques rather than requiring expensive and laborious empirical procedures. Accordingly, it is an object of the subject invention to provide an improved monopulse antenna.

It is also an object of the invention to provide a monopulse antenna having an improved aperture phase distribution.

It is another object of the subject invention to provide a monopulse antenna having a selected phase distribution for reducing monopulse anomalies.

It is still another object of the invention to provide a monopulse antenna providing reduced monopulse anomalies that is inexpensive and efficient to manufacture.

It is a further object of the invention to provide an antenna having a selected aperture phase distribution.

It is a still further object of the invention to provide an antenna having an aperture phase distribution which deviates in a selected manner from a planar phase front.

These and other objects of the invention will become apparent from the following description, taken together with the following drawings in which:

FIG. 1 is a diagram of the geometry of a meridional plane, including the axis of symmetry, of a parabolic reflector, and illustrating the phenomenon of a planar phase front.

FIG. 2 is a diagram of the equivalent far field sum and different monopulse patterns for an exemplary monopulse antenna designed to provide a planar phase front.

FIG. 3 is a curve of a normalized monopulse difference signal

$$\beta_c=\frac{\Delta}{\Sigma}\cos\alpha \text{ versus } \beta$$

for an exemplary monopulse antenna designed to provide a planar phase front, and further illustrating a desired response for β_c versus β.

FIG. 4 is a diagram of exemplary near-field sum and difference monopulse patterns for a monopulse antenna, and illustrating an associated polynomial phase deviation in the aperture phase distribution.

FIGS. 5 and 6 are exemplary monopulse far-field patterns for antennas having a respective exemplary quadratic and quartic phase deviation from a planar phase front.

FIGS. 7 and 8 are curves of only the quasi-linear portions of the corresponding normalized difference function resulting from the far-field patterns of FIGS. 5 and 6, illustrating the linearity thereof.

FIGS. 9 and 10 are curves of the corresponding normalized difference function resulting from the far-field patterns of FIGS. 5 and 6, illustrating the reduction in the severety of the monopouse anomalies obtained.

FIG. 11 is an elevation view of an airborne antenna, the boresight of which is depressed relative to the horizontal and the pattern shape of which is adapted for illumination of a selected portion of the terrain with an even intensity.

FIG. 12 is a diagram of the geometry of a central meridianal plane of an exemplary antenna employing the concept of the invention, and illustrating a phase front deviating from a planar phase front.

FIG. 13 is an illustration of an exemplary arrangement of an antenna embodying the concept of the invention.

FIG. 14 is a family of curves for an exemplary spoiled parabola antenna design, showing that the "spoiled" curve may be generated by bending the antenna reflector surface either frontwardly or backwardly of the nomial true parabola by a like amount.

In the figures, like reference characters refer to like parts.

In a conventional monopulse radar antenna, there is provided a parabolic reflecting surface in cooperation with at least two microwave feedhorns located proximate the focal point P of the parabolic reflecting surface **20**, and oriented to illuminate the reflecting surface, as shown schematically in FIG. 1. The two feedhorns may be fed or excited in time phase to provide a combined transmitted beam of microwave energy, comprised of the mutual inphase radiations of the separate horns, as reflected from the parabolic reflector. Echoes of the transmitted beam of microwave energy, received by the concave reflecting surface are reflected therefrom and concentrated at, or focussed upon, the focal point P where the feedhorns are located, whereby a first and second received signal is produced. These two received signals may then be combined by hybrid microwave mixing means to produce a monopulse sum signal indicative of the sum of the two received signals. and a monopulse difference signal indicative of the difference between the two received signals, as is well understood in the monopulse microwave art.

The reason for using a parabolic shape for the reflecting surface **20** and locating the microwave feedhorns at the focal point of the parabolic shape, is to provide a high degree of directivity, as is well known. Transmitted energy reflected by the dish will be directed substantially in a single direction (i.e., parallel to the center axis **21** of the parabola upon which the focal point P lies). In other words, energy radiating from substantially the focal point P and directed toward the reflector is reflected in a series of rays all substantially mutually parallel to the central axis **21** thereof. Similarly, rays of microwave echoes, received from an illuminated target lying within the far-field pattern of the antenna, and upon the central axis thereof, are substantially parallel as to be focussed upon, or concentrated at the feedhorns, thereby providing an efficient, directive antenna. That is to say, the ray described by line segment d in FIG. 1 is not only parallel to central axis **21**, but is also mutually parallel with any other ray d' radiating from focal point P and reflected from parabolic surface **21**.

Another characteristic of a prior art parabolic antenna having feedhorns located at the focal point thereof is that a substantially planar phase front is provided. In other words, energy radiated from focal point P and reflected from any point (say, y_1, x_1) on parabolic surface **20** takes the same time to reach a plane **22** normal to the central axis as energy originating at point P and reflected from any other point (say, y_2, x_2) on surface **20**. Such property of a parabolic antenna is well known, resulting from both the collimating feature of a parabola and the definition thereof as a curve generated by locus of points the sum of the radial distance to which from a focal point P and the axial distance of which to normal plane **22** describing a fixed distance, the revolution of the curve **20** about the central axis **21** generating the paraboloid of the antenna reflecting surface. The planar phase front property is also referred to as a constant phase distribution across the antenna aperture.

Referring to the control axis **21** as the y abscissa and a perpendicular **30** thereto as the x ordinate in FIG. 1, the familiar equation for the parabola demonstrates the near field planar phase front phenomenon. Referring to FIG. 1:

$$c+d=c^1+d^1=K=2f \tag{1}$$

Where:

$$c=[(y-f)^2+x^2]^{1/2}$$

and

$$d=f-y$$

Rearranging Equation 1:

$$c=2f-d \tag{2}$$

Substituting for c and d:

$$[(y-f)^2+x^2]^{1/2}=2f-(f-y) \tag{3}$$

Squaring both sides:

$$(y-f)^2+x^2=(f+y)^2 \tag{4}$$

$$y^2-2fy+f^2+x^2=f^2+2fy+y^2 \tag{5}$$

$$-2fy+x^2=2fy \tag{6}$$

$$x^2=4fy \tag{7}$$

The development of such geometry is set forth here in order to allow an appreciation of changes thereto, to be described more fully hereinafter.

An exemplary arrangement of such prior art monopulse antenna comprising a parabolic reflector, feedhorns and microwave feedbridge is shown, for example, in FIG. 3 of U.S. Patent No. 3,071,769 issued January 1, 1963, to G. M. Randall et al. for a Four Horn Feed Bridge. Typical Fraunhaufer sum and difference patterns (or variation of the far-field sum and difference signals corresponding to a variation in the angle-off-boresight or angular displacement, β, of a target from the central, or boresight, axis) of such prior art antenna are shown in FIG. 2 hereof.

Referring to FIG. 2 there is illustrated the typical far-field sum and difference patterns resulting in a monopulse receiving antenna employing a near-field planar phase distribution. The sum pattern, indicated by curve **23**, is seen to be comprised of a main lobe having an axis of symmetry corresponding to the boresight axis, and a series of lesser (side) lobes displaced symmetrically relative to the boresight axis, the antenna beamwidth being conventionally described by the angle at which the energy pattern at a given range is attentuated to −3 db relative to the on-boresight energy level. Within the beamwidth of the representative sum pattern of FIG. 2, the difference pattern amplitude **24** is seen to vary somewhat linearly with the angle, β. The sharp nulls in each of the sum and difference patterns indicate rapid phase reversals between the several lobes, as well as amplitude attenuations, and account for the anomalies to be observed in the determination of the angle-off boresight of a detected target.

One monopulse receiver technique commonly employed in the determination of the angle-off-boresight β of a detected target is the measurement of the function

$$\frac{E_d}{E_s} \cos \alpha$$

where E_d and E_s are a monopulse receiver sum and difference signals, respectively, and α is the phase angle difference between them. The ratio E_d/E_s is conventionally achieved by AGC amplifiers in the sum and difference channels, commonly employing the sum channel output as the AGC control signal; while the factor $\cos \alpha$ is obtained by the phase detection of the normalized difference signal (E_d/E_d), employing the AGC processed sum signal as a phase reference. Such arrangement is well understood in the art, being illustrated, for example, in FIG. 3.12 on page 57 of "Introduction to Monopulse" by Rhodes, published by McGraw-Hill (1959).

Other functions varying somewhat linearly with angle-off-boresight are also employed as measures of the angle-off-boresight of a detected target, and are also shown in the above cited "Introduction to Monopulse." However, all such so-called linear functions of β are sufficiently linear over only a restricted region (within the main lobe) about $\beta=0$. Further, the monopulse difference signal normalizing means for generating such linear function signal is inherently responsive to as much of the side lobe pattern as is above the receiver threshold response. Accordingly, where the planar front, of the monopulse antenna aperture phase distribution, produces sharp nulls in the difference amplitude far-field pattern or several crossovers of the sum and difference far-field amplitude patterns, then anomalies may occur in the above described linear monopulse functions. In other words, a selected function of the monopulse sum and difference signals may exhibit the same amplitude and sense for more than one combination of sense and amplitude of the angle β. Hence, such condition provides an ambiguity in the indication β_c of an actual angle-off-boresight β of a detected target. Such ambiguities are herein referred to as monopulse anomalies. The anomalies occurring in the function,

$$\beta_c = \frac{E_d}{E_s} \cos \alpha$$

for the typical antenna response of FIG. 2 are shown in FIG. 3, although it is to be understood that anomalies would similarly exist in the function,

$$\beta_c = \log \left| \frac{1+K\beta_c}{1-K\beta_c} \right|$$

Referring to FIG. 3, there is plotted the function,

$$\beta_c = \frac{E_d}{E_s} \cos \alpha$$

versus β, for the planar aperture phase distribution indicated by the typical antenna pattern of FIG. 2. Such function is illustrated by curve 25, showing a plurality of anomalies or alternate values of β, for which a given amplitude and sense combination of the function

$$\frac{E_s}{E_d} \cos \alpha$$

are repeated. For example, the value β_{c1} occurs at least at β_{co}, β_1, β_2 and β_3 in FIG. 3. A most serious anomaly occurs in the region of point 32, corresponding to the on-boresight region or origin of FIG. 3.

A preferred characteristic of the function

$$\frac{E_d}{E_s} \cos \alpha$$

is shown in FIG. 3 by the exemplary dotted curve 26, which illustrates no ambiguities. Such freedom from ambiguity has been sought in the prior art by adjustment of the near-field amplitude distribution across the antenna aperture, while retaining a constant phase distribution thereacross. However, such technique is extremely difficult to implement in practice, and has produced results of only limited success. It has been discovered to be more effective and efficient to remove such ambiguities by varying the phase distribution across the antenna aperture from the prior near-field planar phase front. Such deviation from a planar phase front has been referred to in the prior art as a phase error because such deviations in the construction and performance of prior art (planar phase front) antennas were deemed errors. However, such term as employed herein, refers to a selected deviation from a planar phase front or constant aperture phase distribution which is deemed desired or preferred for providing monopulse antenna pattern shaping.

The amplitude and phase distribution across the aperture of an antenna are so-called near-field effects, the aperture (a) corresponding to the maximum dimension of the antenna reflector in the coordinate or dimension (x) of interest.

Where the phase distribution or shape of the phase front across the antenna aperture is made to vary relative to the theoretical time phase occurrence of a ray originating at the focus P (of FIG. 1) and reflected from point (y_0, x_0) of surface 20 along central axis 21 in FIG. 1, the sharp nulls in the far-field patterns of FIG. 2 are observed to be somewhat avoided as shown in FIG. 5 by means of the near-field pattern characteristics shown in FIG. 4.

Referring to FIG. 4, there is illustrated a typical set of near-field or aperture patterns provided by a pair of microwave feedhorns used to illuminate a reflector. Curve 27 represents a typical sum pattern aperture amplitude distribution, curve 28 represents the corresponding difference pattern aperture amplitude distribution, and curve 29 represents a phase distribution across the antenna aperture and having a selected deviation from a planar phase front. Curve 29 indicates, in other words, that the wave front of the energy at the extremities of the aperture does not occur in phase with that portion of the wave front at the center of the aperture. Because, for a symmetrical aperture, the sense of the time phase effects are similar for corresponding portions of the wave front on opposite sides of the center of the aperture, the phase deviation may be described as an even valued function. Where the phase deviation varies linearly across the aperture from the center thereof outwardly to the extremity thereof, the phase deviation may be referred to as a linear phase deviation. Where it increases as a function of some other power or exponent greater than unity, the phase deviation may be referred to generally as a polynominal phase deviation.

Now, the far-field sum and difference patterns of a horn-fed reflector antenna with phase deviation from a planar phase front (e.g., having "phase error") can be approximated by the following expression:

$$G(\beta) = \int_{-a}^{+a} F(x) e^{jk\left[\frac{x}{a} \sin \beta - \psi(x)\right]} dx \quad (8)$$

Where:

$G(\beta)$=Far-field voltage pattern
$F(x)$=Amplitude distribution function of aperture
$\psi(x)$=Phase deviation distribution
β=Pattern angle variable
j=Phasor or imaginary notation
x=Aperture coordinate or dimension variable
$k=2\pi/\lambda$
$2a$=Aperture size (measured on the x coordinate), and
λ=Free space wavelength

The sum and difference amplitude distribution function, $F(x)$, can be approximated by cosine and sine functions respectively.

Now, because the (sum or difference) far-field voltage pattern $G(\beta)$ is an integral function of the near-field phase deviation and associated one of the (sum or difference) amplitude near-field or aperture distribution functions

$F(x)$, it is to be appreciated, from FIG. 4, that the phase deviation (Curve **29**) will have less effect upon the resulting far-field sum pattern E_s than it will upon the far-field difference pattern E_d. For example, it can be seen from FIG. 4 that as the phase deviation increases (Curve **29**), the microwave sum signal amplitude distribution is decreasing while the corresponding difference signal amplitude distribution is increasing, the maximum of the sum pattern amplitude distribution occurring at the center of the aperture and the maxima of the difference pattern amplitude distribution occurring at a displacement on either side of the center of the aperture. Hence, a polynomial phase deviation, or polynomial phase error, of a power greater than unity in the phase distribution across the near-field or antenna aperture, would produce a greater effect on the integral expression of Equation 8 as applied to a difference pattern, than to a sum pattern. In other words, the area described by phase deviation Curve **29** under the difference pattern Curve **28** is larger than that described by such curve under the sum pattern Curve **27** in FIG. 4; hence, the associated effect upon the corresponding integral function of Equation 8 will be correspondingly greater.

The effect of such near-field phase deviation from a planar phase front, is to "fill in" or reduce the nulls in the far-field response (of FIG. 2, for example). Further, because such effect is more pronounced for the far-field difference pattern than for the far-field sum pattern, the general level of the difference pattern side-lobe response is increased above that of the sum pattern side lobe response, whereby the monopulse anomalies in the selected functions of the sum and difference signal such as, for example,

$$\frac{E_d}{E_s}\cos\alpha$$

are reduced.

Such near-field phase deviation from a planar phase front may be achieved in practice by varying (increasing or decreasing) the curvature of the antenna reflector from a true parabola. Although such curvature variation slightly disturbs the collimating effect provided by a true parabolic reflector, such disturbance is relatively minor in nature, the major effect is the time-phase effect produced by the space-phasing or curvature deviation upon the microwaves processed thereby.

In the calculation of the far-field sum and difference patterns from the near-field patterns or aperture distributions by means of Equation 8 the following amplitude distribution functions $F(x)$ for sum (Σ) and difference (Δ) patterns are representative:

$$F_\Sigma(x)=\cos\frac{1.443x}{a} \tag{9}$$

$$F_\Delta(x)=.7943\sin\frac{2.042x}{a} \tag{10}$$

The phase error distribution function $\psi(x)$ in Equation 8 (for a "spoiled" parabola or reflector deviating from a true parabola) is expressed as the magnitude, γ, of the phase error existing at the edges of the aperture (relative to the center thereof), multiplied by a selected power of the ratio of the aperture ordinate x to the aperture limit a. For example, a linear, or first power, phase error may be analytically described as

$$\psi(x)=\frac{\gamma x}{a}$$

a quadratic, or second power, phase error described as

$$\psi(x)=\frac{\gamma x^2}{a^2}$$

and a quartic or fourth power phase error as

$$\psi(x)=\frac{\gamma x^4}{a^4}$$

where γ is defined as the magnitude of phase error at the edges of the aperture. More generally,

$$\psi(x)=\sum_{n=1}^{\infty}\frac{\gamma_n|x|^n}{a^n} \tag{11}$$

Exemplary monopulse sum and difference far-field patterns and the associated functions,

$$\beta_c(\beta)=\frac{E_d}{E_s}\cos\alpha$$

to be obtained with selected polynomial phase distributions are shown in FIGS. 5, 6, 7 and 8. These figures represent the improved responses obtainable from a horn fed monopulse antenna having an aperture or maximum reflector dimension ($2a$) of seven inches and a focal length of 5 inches.

FIGS. 5 and 6 represent the far-field sum and difference patterns resulting from a quadratic and quartic phase error, respectively. In each case, γ (the magnitude of phase error at the edges of the aperture) was selected as $\pi/2$ radians. The response curves in each of FIGS. 5 and 6 demonstrate generally an improved side lobe response in which no sharp nulls occur and wherein no additional amplitude crossovers of the sum and difference patterns occur above -35 db. The anomaly region of FIG. 6 of 2 degrees width, occurring near $\beta=65°$, being below -35 db, is of no practical significance for the reason that such attenuated region would normally be below the threshold response of the receiver or radar data processor.

FIGS. 7 and 8 represent the quasi-linear response region of the function

$$\beta_c=\frac{E_d}{E_s}\cos\alpha$$

associated with the far field patterns of FIGS. 5 and 6 respectively. FIG. 7, the quadratic error case, in addition to response Curve **25***a* for

$$\psi(x)=\frac{\gamma x^2}{a^2}$$

where $\gamma=\pi/2$; also includes a Curve **25***b* showing the effect of ($\gamma=\pi$) in the expression for the phase error function, $\psi(x)$. It is to be thus observed that the value $\pi/2$ (Curve **25***a*) is to be preferred over the value π for γ (Curve **25***b*) in the quadratic case (FIG. 7); and that for the value, $\gamma=\pi/2$, the quartic case provides a more linear β_c response than the quadratic case (FIG. 8).

FIGS. 9 and 10 represent the side-lobe region, as well as quasi-linear response region, of the function,

$$\beta_c=\frac{E_d}{E_s}\cos\alpha$$

associated with the far-field sum and difference pattern of FIGS. 5 and 6, respectively; and illustrate the relative freedom from ambiguities in such function for values thereof corresponding to the linear regions of FIGS. 7 and 8. The quartic case of FIG. 10

$$\left[\beta_c(\beta)\text{ for }\psi(x)=\frac{\pi}{2}\frac{x^4}{a^4}\right]$$

is seen to be superior to the quadratic case of FIG. 12

$$\left[\beta_c(\beta)\text{ for }\psi(x)=\frac{\pi}{2}\frac{x^2}{a^2}\right]$$

for the reason that a greater range of values of β_c exist for which no ambiguity exists. For example, the values of

$$\beta_c=\frac{E_d}{E_s}\cos\alpha$$

for angles-off-boresight up to 10° do not occur at other angles-off-boresight, except those above 62°, for which the side lobe pattern is so attenuated as to be below the receiver threshold. In either case however, a limited response region, free of side-lobe response, may be obtained by suitable gating of β_c signals above a selected amplitude.

In the application of an airborne monopulse radar data processor to ground mapping, it is desired to achieve a far-field antenna pattern which evenly illuminates the terrain or area under survillance by the radar, as shown in FIG. 11. Accordingly, the beam pattern is preferably further shaped theoretically in the elevation plane in accordance with the $csc^2\beta \cos \beta$ function. In such application, the antenna of the airborne radar is stabilized relative to the local horizon (and declined relative thereto by a selected angle) in order to illuminate the terrain, terrain features being located within the beamwidth in terms of the angle-off-boresight as well as in terms of radial, or slant, range. The further shaping of the antenna pattern as a function of $csc^2\beta$ provides an even intensity pattern below the airborne antenna and forward thereof (out to the normal limits of the 3 db envelope), while the $\cos \beta$ function serves to attenuate any early ground return occurring directly under the airborne system (e.g., $\cos \beta = \cos 90° = 0$).

Such ground mapping criterion or $csc^2\beta \cos \beta$ function, optimized for an altitude of 10,000 feet and a ground range of 20 nautical miles, is shown superposed as Curve 31 in both of FIGS. 5 and 6. From a comparison of such curve with Curves 23 of each of FIGS. 5 and 6, it is clear that the quadratic phase error case (FIG. 5) will provide the best ground mapping operation while still allowing the generation of a ratio curve free of major monopulse anomalies.

Hence, the selection of a particular polynomial phase error $\psi(x)$ to be employed is determined by a particular application for which the monopulse antenna is intended. A preference for a particular polynomial phase error having been determined, the reflector contour providing such phase error must be determined. Because the phase deviation γ for the edge or maximum aperture dimension varies as a function of the wavelenth λ of the microwave energy employed, γ is preferably expressed in terms of wavelengths. For example, a selected value, $\pi/2$ (radians), for γ corresponds to $\lambda/4$ (wavelengths).

Where a selected value for γ (the magnitude of phase error at the edges of the aperture) and selected polynomial order for the phase error function $\psi(x)$ are preferred (say:

$$\gamma = \frac{\pi}{2}, \text{ and } \psi(x)\frac{\gamma x^2}{a^2}$$

for example, corresponding to

$$\psi(x) = \sum_{n=1}^{\infty} \frac{\gamma_n |x|^n}{a^n}$$

where all $\gamma_n = 0$, except γ_2) for a given antenna application, the associated curvature of the antenna reflector needs be determined. Such curvature may be referred to as a "spoiled parabola" for the reason that the curvature is essentially that of a paraboloid, deviating therefrom by an amount corresponding to the preferred polynomial phase distribution function. The derivation of an analytical expression approximating such shape may be more easily appreciated from a consideration of FIG. 12.

Referring to FIG. 12, there is illustrated the geometry of a meridianal plane, including the axis of symmetry, of a spoiled parabolic reflector, and illustrating the phenomenon of a polynomial phase deviation from a planar phase front. The general arrangement of FIG. 12 corresponds to the arrangement of FIG. 1, but for the difference in shape of meridianal reflector curve **20**a, resulting in the curved phase front **22**a, deviating from a planar phase front **22** by an amount which has been exaggerated in FIG. 12 for ease of exposition, and corresponding to the phase distribution Curve **29** of FIG. 4. Point **33** represents the edge of the reflector, corresponding to one-half the aperture dimension, $2a$. The symbol γ represents a selected phase-deviation from a planar phase front occurring at edge **33**, such as for example,

$$\gamma = \frac{\pi}{2} = \frac{\lambda}{4}$$

The curvature of the phase front Curve **22**a is selected to represent a preferred function $\psi(x)$, such as for example,

$$\psi(x) = \sum_{n=1}^{\infty} \frac{\gamma_n |x|^n}{a^n} = \frac{\pi x^2}{2a^2}$$

In determining the curvature of Curve **20**a, the following relationship must be satisfied:

$$2f = c + d \tag{12}$$

or

$$c = 2f - d \tag{13}$$

Now:

$$c = \sqrt{(y-f)^2 + x^2} \tag{14}$$

and

$$d = f - y + \frac{\gamma x^2}{a^2} \tag{15}$$

Substituting Equation 14 in Equation 13:

$$\sqrt{(y-f)^2 + x^2} = 2f - d \tag{16}$$

Then, substituting Equation 15 in the right hand member of Equation 16:

$$\sqrt{(y-f)^2 + x^2} = f + y - \frac{\gamma x^2}{a^2} \tag{17}$$

Squaring both sides of Equation 17:

$$y^2 - 2fy + f^2 + x^2 = f^2 + 2fy + y^2 - 2(f+y)\frac{\gamma x^2}{a^2} + \frac{\gamma^2 x^4}{a^4} \tag{18}$$

Rearranging:

$$4fy = x^2 - \frac{\gamma^2 x^4}{a^4} - 2(f+y)\frac{\gamma x^2}{a^2} \tag{19}$$

Combining terms:

$$4fy = \left[1 + \frac{2\gamma y}{a^2} + \frac{2\gamma f}{a^2}\right]x^2 - \frac{\gamma^2 x^4}{a^4} \tag{20}$$

In other words, Equation 20, deviates from a true parabola to the extent of the bracketed coefficient of x^2 and the added term,

$$-\frac{\gamma^2 x^4}{a^4}$$

More generally,

$$4fy = x^2 + \frac{2(y+f)\Sigma\gamma_n |x|^n}{a^n} - \frac{\Sigma\gamma_n^2 |x|^{2n}}{a^{2n}} \tag{21}$$

where $n > 1$. In other words:

$$4fy = x^2 + 2(y+f)\psi(x) - \psi^2(x) \tag{22}$$

The above derived analytical expressions assume that line segment d, as a reflected ray, is parallel to the axis of symmetry. However, according to wave propogation theory, line segment d, as a reflected ray, must be normal to the wave front. That is, it should be normal to a line which is tangent to Curve **22**a (in FIG. 12) at the point of intersection of segment d with Curve **22**a. In actual practice with exemplary microwave designs, the maximum error envolved has been shown to be less than $\frac{1}{60}$ of a wavelength. Accordingly, the rays, d, of the spoiled parabola of FIG. 12 may be assumed to be mutually parallel with the axis of symmetry **21**. In other words, the effect of so spoiling the parabola is more pronounced upon the phase front than upon the collimation of the reflected rays therefrom.

In a practical application, the spoiled parabolic curve segment **22**a (of FIG. 12), rotated about axis **21** would generate a spoiled paraboloid surface corresponding to the reflector shape desired, the monopulse microwave feedhorns being located at focal point P and oriented to illuminate the surface so generated. In a specific design for an airborne monopulse mapping system, the antenna would be of the azimuthal scanning type and have a fan-shaped (sum) pattern. By means of such pattern, a broad angular beamwidth is provided in elevation, and a very narrow beamwidth is provided in azimuth. The angular position of a target is determined in elevation by monopulse techniques for determining the angle-off-boresight of a detected target, while the azimuth position is determined by azimuthal scanning of the narrow azimuthal beamwidth, as is well understood in the art. Such fan-shaped (wide elevation and narrow azimuthal) beamwidths are achieved by employing a relatively small elevation aperture dimension relative to the azimuthal aperture dimension, whereby the frontal area of the spoiled parabola reflector resembles an ellipse, the minor axis of which lies in the elevation or vertical plane, as shown in FIG. 13.

In FIG. 13 is shown an exemplary design of an antenna employing the concept of the invention. The antenna comprises a reflector **35** representing a spoiled parabola in the elevation plane, and a conventional parabola in the azimuth plane, an elliptical frontal area as indicated by the dotted ellipse **36**. The antenna of FIG. 13 further comprises a multi-horn microwave feed structure **37**, the openings of which are clustered about the axis of symmetry **21** and located at the focal point of surface **35**.

In an exemplary arrangement of the device of FIG. 13 for x-band about (9.3 kilomegacycles per second) frequencies, the major axis was 24 inches, the minor axis 7 inches and the focal length of 5 inches. A reflector illumination having an edge taper of -18 db was employed in the elevation (minor axis) plane, and azimuth edge taper of -13 db selected. By means of such geometry, the sum and difference elevation patterns were optimized while reducing spill-over energy, and maximum gain and low side lobe levels achieved (to compensate for aperture blockage due to the center feed arrangement).

A half-contour of a meridianal elevation plane through the exemplary antenna, illustrating the shape of the reflector in the elevation plane, is shown in FIG. 14, relative to a true parabola.

Such elevation axis geometry is plotted in FIG. 14 for the following selected quadratic case, although it is to be understood that the scope of the invention is not limited to such case:

$$\gamma=\frac{\pi}{2}=\frac{\lambda}{4}$$

$$a=\frac{7''}{2}\text{ or 3.5 inches}$$

$$\psi(x)=\frac{\lambda x^2}{4(3.5)^2}$$

Because the amount $|\gamma|$ of the desired aperture phase deviation at the aperture edge may be achieved by bending the reflector curvature to either side of the parabola **20**, as shown by Curves **20**a and **20**b, it is to be understood that the invention is not limited to spoiling the parabola to just to one side thereof, but that spoiling may be achieved by deviation to either side thereof. Further, the curvature of the azimuthal plane may be a true parabola rather than the spoiled parabola computed for the elevation plane, without reducing the effect of the spoiled parabola in the elevation plane upon the elevation axis of the monopulse system utilizing the invention.

Hence, it is to be appreciated that an improved monopulse antenna structure has been disclosed for providing improved off-boresight data processing and minimizing the effects of monopulse ambiguities.

Although the invention has been described and illustrated in detail, it is to be clearly understood that the same is by way of illustration and example only and is not to be taken by way of limitation, the spirit and scope of this invention being limited only by the terms of the appended claims.

I claim:

1. In a monopulse receiving antenna for producing first and second monopulse microwave signals in response to received microwave signals of a selected carrier frequency, means for providing a selected aperture phase distribution in said monopulse receiving antenna comprising

a continuous curved reflecting surface defining a central axis originating at a point in said curved surface and oriented perpendicularly to said surface at said point of origin, a feed point on said central axis and spaced apart from said curved surface; and

an array of at least two feedhorns monuted at said feed point for illuminating said curved surface; and

the curvature of said curved surface cooperating with said feedhorns to provide an aperture phase distribution which varies from a planar phase front as a continuous even valued function of a power greater than unity of the aperture dimension.

2. In a monopulse receiving antenna for producing first and second monopulse microwave signals in response to received microwave signals of a selected carrier frequency, means for providing a selected aperture phase distribution deviating from a planar phase front in a monopulse receiving antenna comprising

a continuous curve reflecting surface defining a central axis originating at a point in said curved surface and oriented perpendicularly to said surface at said point of origin, a feed point on said central axis and spaced apart from said curved surface; and

an array of at least two feedhorns mounted at said feed point for illuminating said curved surface;

said curved surface deviating from a parabola as a continuous function of the aperture of said antenna, whereby anomalies are reduced in the amplitude ratio of the difference between said received signals to the sum thereof.

3. In a monopulse receiving antenna for producing sum and difference monopulse microwave signals in response to received microwave signals of a selected carrier frequency, means for providing a controlled aperture phase distribution in a monopulse receiving antenna comprising

a continuous curved reflecting surface defining a central axis originating at a point in said curved surface and oriented perpendicularly to said surface at said point of origin, a feed point on said central axis and spaced apart from said curved surface; and

an array of at least two feedhorns mounted at said feed point for illuminating said curved surface;

the curvature of said curved surface cooperating with said feedhorns to provide an aperture phase distribution which varies from a planar phase front as a continuous even valued function of a power greater than unity of the aperture dimension.

4. In a monopulse receiving antenna for producing sum and difference monopulse microwave signals in response to received microwave signals of a selected carrier frequency, means for providing a selected phase distribution in a monopulse receiving antenna comprising

a continuous concave curved reflecting surface defining a central axis originating at a point in said curved surface and oriented perpendicularly to said surface at said point of origin, said curved surface having a focal point spaced along said central axis on the concave side of said curved surface;

an array of at least two feedhorns mounted at said focal point for illuminating said curved surface;

said curved surface deviating from a parabola as a continuous function of the aperture of said antenna, whereby anomalies are reduced in the product of the ratio of said difference signal to said sum signal and the cosine of the phase angle between them.

5. In a monopulse receiver including monopulse difference signal normalizing means and an antenna having a hybrid microwave feed structure in cooperation with an array of at least two feedhorns defining an aperture for

producing separate microwave receiver signals indicative of the sum and difference respectively of the microwave energy received by said feedhorns, means for reducing the anomalies existing in a normalized monopulse difference signal which varies substantially only with the angle-off-boresight of a detected target, comprising

a curved reflecting surface defining a central axis originating at a point in said surface and oriented perpendicularly to said surface at said point of origin and having a feed point along said central axis spaced apart from said curved reflecting surface;

said array of monopulse feedhorns being mounted at said feed point and oriented for illuminating said curved surface, said array defining a plane containing said central axis;

the curvature of said curved surface being defined relative to said central axis and the plane of said array substantially in accordance with the following relationship;

$$4fy=x^2+2(f+y)\psi(x)-\psi(x)^2$$

where:

$$\psi(x)=\sum_{n=1}^{\infty}\frac{\gamma_n|x|_n^n}{a^n}$$

f=distance of said feed point from said point of origin of said curved surface.

y=a coordinate distance of a point on said curved surface measured along said central axis from said point of origin.

x=a corresponding coordinate distance of said point measured perpendicularly from said central axis in a direction parallel to the orientation plane of said array.

$\psi(x)$=the phase distribution across the aperture relative to the center thereof.

a=one-half the aperture size of said reflector in the plane of said array (e.g., x_{max}).

γ=magnitude of the phase error at the edges of the aperture.

6. In a monopulse receiver including monopulse difference signal normalizing means and an antenna having a hybrid microwave feed structure in cooperation with an array of at least two feedhorns defining an aperture for producing separate microwave receiver signals indicative of the sum and difference respectively of the microwave energy received by said feedhorns, means for reducing the anomalies existing in a normalized monopulse difference signal which varies substantially only with the angle-off-boresight of a detected target, comprising

a curved reflecting surface defining a central axis originating at a point in said surface and oriented perpendicularly to said surface at said point of origin and having a feed point along said central axis spaced apart from said curved reflecting surface;

said array of monopulse feedhorns being mounted at said feed point and oriented for illuminating said curved surface, said array defining a plane containing said central axis;

the curvature of said curved surface being defined relative to said central axis and the plane of said array substantially in accordance with the following relationship:

$$4fy=\left[1+\frac{2\gamma y}{a^2}+\frac{2\gamma f}{a^2}\right]x^2-\frac{\gamma^2}{a^2}x^4$$

where:

f=distance of said feed point from said point of origin of said curved surface.

y=a coordinate distance of a point on said curved surface measured along said central axis from said point of origin.

x=a corresponding coordinate distance of said point measured perpendicularly from said central axis in a direction parallel to the orientation plane of said array.

a=one-half the aperture size of said reflector in the plane of said array (e.g., x_{max}).

γ=magnitude of the phase error at the edges of the aperture.

7. In a monopulse antenna, having a hybrid microwave feed structure in cooperation with an array of at least two feedhorns defining an aperture for producing separate microwave receiver signals indicative of the sum and difference respectively of the microwave energy received by said feedhorns, means for reducing the anomalies existing in the product of the ratio of said difference to said sum and the cosine of the phase angle difference between them, comprising

a concave reflecting surface defining a central axis originating at a point in said surface and oriented perpendicularly to said surface at said point of origin and having a focal point spaced along said central axis on the concave side of said curved reflecting surface;

said array of monopulse feedhorns being mounted at said focal point and oriented for illuminating the concave side of said curved surface, said array being perpendicular to said central axis;

the curvature of said curved surface being defined relative to said central axis and the direction of said array substantially in accordance with the following relationship:

$$x^2+\left[2(f+y)\frac{\sum\gamma_n|x|^n}{a^n}-\frac{\sum\gamma^2_n|x|^{2n}}{a^{2n}}\right]$$

where:

f=distance of said focal point from said point of origin in said curved surface.

y=a coordinate distance of a point on said curved surface measured along said central axis from said point of origin.

x=a corresponding coordinate distance of said point measured from said central axis in a direction parallel to the orientation of said array.

a=one-half the aperture size of said curved surface in the x direction.

γ=magnitude of the phase error at the edges of the aperture relative to the center thereof.

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