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3,005,985

PRE-PROGRAMMED SCANNING ANTENNA

Filed Sept. 19, 1957

2 Sheets-Sheet 1

FIG. 1

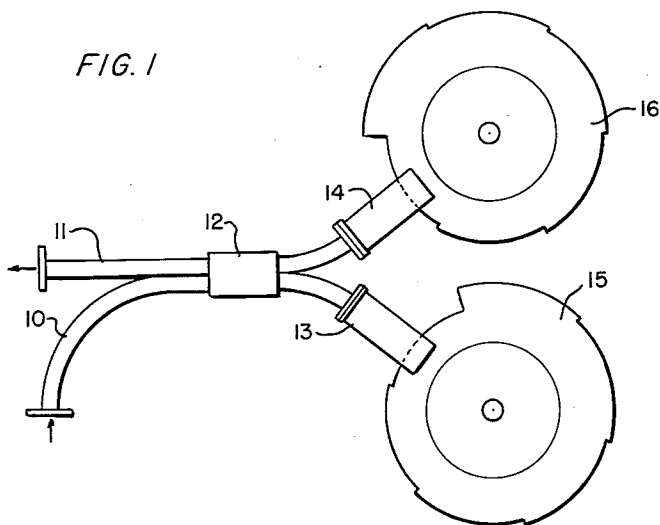


FIG. 2

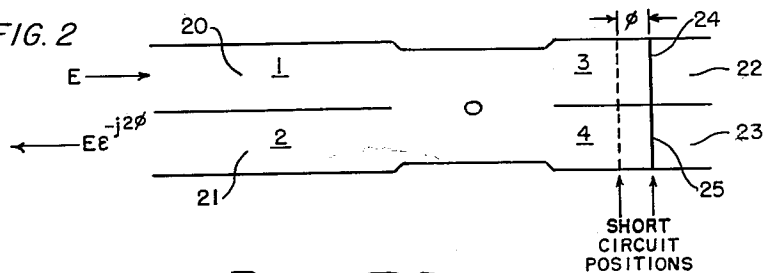
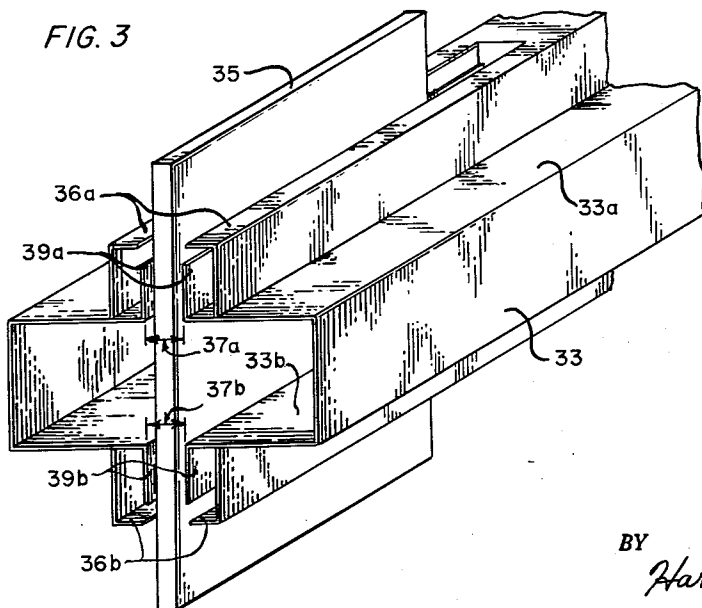


FIG. 3



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FIG. 4A

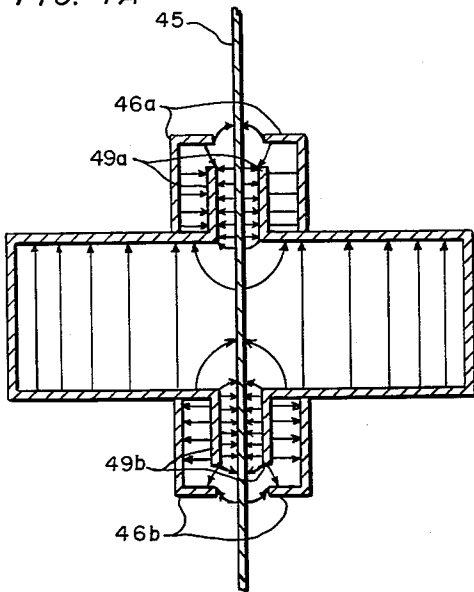


FIG. 4B

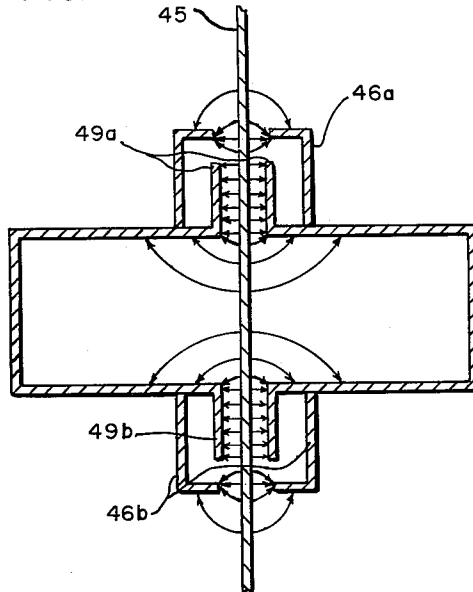
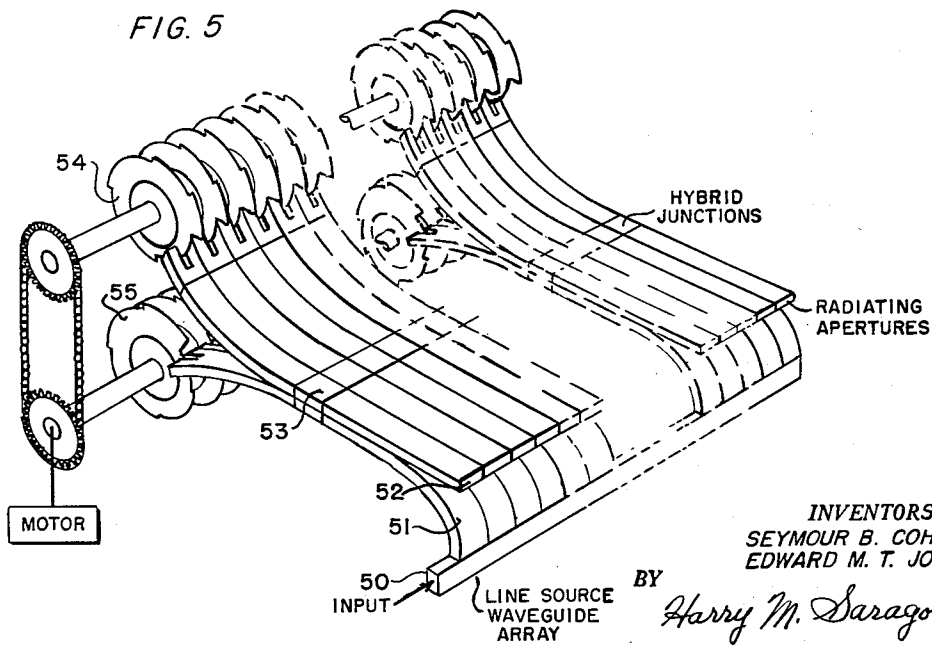


FIG. 5



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PRE-PROGRAMMED SCANNING ANTENNA
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4 Claims. (Cl. 343-777)

The invention described herein may be manufactured and used by or for the Government for governmental purposes, without the payment of any royalty thereon.

The present invention relates to a pre-programmed scanner used with an antenna for radiating electromagnetic energy in a highly directive pattern and for scanning through space in a predetermined manner. This type of scanner may find particular utility for directional antenna systems operating on wave-lengths from about three centimeters down to about four millimeters or less.

In general, known antenna systems of the type referred to above achieve a scanning action by introducing large and continuously variable phase shifts between the primary radio frequency feed source and the energy radiating means. It has heretofore been proposed to provide a pre-programmed scanner utilizing a linear array of radiating horns fed with constant-amplitude radio frequency signals whose phases are varied in a discrete fashion. However, in practice, this type of scanner design has been impeded by the lack of a suitable rapid-acting phase shifter.

A principal object of the present invention is to provide an improved directional scanning antenna for this purpose which is electrically and mechanically simple due to the use therein of a rapid-acting phase shifter.

Accordingly a further object of the present invention is to provide a novel phase shifter which is capable of being varied in a pre-programmed manner by means of a rapid, continuous, mechanical rotation.

A major advantage of the scanner of the present invention is that it may be programmed in a predetermined manner to direct the antenna radiation at various angles in any arbitrary periodic manner, rather than in the sinusoidal or sawtooth fashion of the prior devices. Thus, for example, by arranging time consecutive beam positions to be non-consecutive in angular position, the possibility of false echoes from targets beyond the nominal range of the radar may be minimized.

Another advantage of the pre-programmed scanner is that the radiated beam remains stationary at discrete angular positions during the length of time that it takes energy to make a round trip between the antenna and the target. Thus there is no scanning loss due to the radiated beam pointing in slightly different directions at the time of transmission and reception of the radar pulse. This advantage is offset to some extent by the fact that only a discrete number of targets within the limits of scan of the radar receive maximum illumination, while those in between adjacent beam positions receive somewhat less energy. Because the pre-programmed scanner phase shifters need only introduce phase delays between 0 and 360 degrees their losses are relatively independent of beam position, in contrast to the losses in scanners which employ phase shifters that introduce many times this phase shift at wide scan angles.

The programming possibilities are of particular importance because the beam position may be changed in steps in any desired sequence. In addition to the usual consecutive beam position, the beam positions may be interlaced and their durations may be unequal. Many new system possibilities are made feasible by this feature. Previously, the frequency-scanning antenna was the only type that could function in this way, but with the disad-

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vantage of requiring fixed, highly-stable transmitter and local oscillator sources, with consequent great complications in circuitry. On the other hand, the pre-programmed scanner can operate with ordinary magnetron and klystron sources operating anywhere in a wide frequency band. Hence, the pre-programmed scanner disclosed here is far less susceptible to jamming than the frequency scanner.

The aforementioned and additional advantages of the novel scanning antenna of the present invention may thus be summarized as follows: the performance is good over a wide frequency band; the scan cycle may be programmed in any desired predetermined manner; more than one scan cycle may be obtained per rotation of the drive shaft, if desired; and the construction tolerances are simple to obtain in manufacture.

As indicated generally above, a directive antenna array embodying the concept of the present invention requires a primary feed source of radio frequency energy, an ultimate electromagnetic energy radiating means, and a phase shifting device interposed between the feed source and the radiating means. According to the invention the rapid-acting phase shifter may utilize a known hybrid junction such as a three-db directional coupler and a pair of novel variable short circuits in a waveguide system. A variable short circuit which effects the phase shift comprises a length of waveguide having a conductive reflective member passing through or inserted into a slot cut in the opposing walls of the waveguide. The reflective member should not physically contact the slot edges in the waveguide wall, and chokes are added to the outside of the wall to prevent energy from leaking out of the structure between the edges of the slot and the reflective member.

Further objects and advantages of the invention will become apparent as the detailed description of a preferred embodiment of scanning antenna dimensioned to operate in the so-called "centimeter" frequency region at around 9500 megacycles proceeds, reference being made to the accompanying drawings wherein:

FIG. 1 shows one rapid acting phase shifter;

FIG. 2 is a schematic illustration useful in describing the operating principles of the invention;

FIG. 3 is an isometric line drawing showing the mechanical configuration of one form of a non-contacting short circuit member for effecting phase shift in a waveguide;

FIG. 4 shows the electric field lines of energy distributions which may be established within the waveguide structure of FIG. 3; and

FIG. 5 is an isometric view of a pre-programmed scanning antenna array made up of a plurality of elements or sections of the type shown in FIG. 1.

Rapidly scanning millimeter wavelength antennas used with ground radar sets will, in general, have a rapid scanning motion in the horizontal plane and considerably slower scanning motion in the vertical plane. One example of a scanner with these characteristics is the two-beam mortar locator. Another is the television type scanner which presents azimuth and elevation information to the operator. Since the vertical scan rate is relatively slow, scanning in the vertical plane can be accomplished by purely mechanical means. However, rapid scanning in the horizontal plane is more difficult. Optical type line-source scanners using lenses to collimate energy from a rapidly moving feed-source have produced the desired rapid scanning. Such lenses are not entirely satisfactory since they do not transmit all the energy incident upon them. The line-source scanner of our invention consists of an array of radiating horns fed with arbitrary phases such that a beam will radiate in any pre-pro-

grammed direction. As discussed, this type of scanner requires a novel rapid-acting phase shifter.

FIG. 1 shows a single section of the pre-programmed scanner having a phase shifter which is variable in a pre-programmed sequence by a rapid continuous mechanical rotation. The entire scanner assembly consists of a plurality of these sections in a line array as shown in FIG. 5. In FIG. 1, input waveguide 10 is coupled to the input of hybrid junction 12. The output waveguide section 11 lies parallel to section 10 and, in practice, the two waveguides may share a common intermediate broad or narrow boundary wall. Each waveguide terminates in slotted stub 13 and 14 respectively. Variable radii discs 15 and 16 are rotated within the respective slotted openings. Energy fed into the phase shifter at input 10 emerges from output 11 with a phase determined by the position of the variable short circuit discs 15 and 16 in the arms 13 and 14.

FIG. 2 schematically shows the theory of operation of the phase shifter. Movable short circuits 24 and 25, equivalent to stepped discs 15 and 16, are positioned in arms 22 and 23 of the hybrid coupler. Energy is fed into arm 20 of the coupler and emerges from arm 21. Moving the shorting plungers an electrical distance ϕ changes the phase of the output signal by 2ϕ .

FIG. 3 shows the details of the non-contacting slot arrangement used in the terminating stubs 13 and 14 of the phase shifter. The metallic disc portion 35, a portion of discs 15 and 16, fits in the slots 37a and 37b cut in the top and bottom walls 33a and 33b of the rectangular waveguide 33. Chokes 36a and 36b are added to the outside of the waveguide 33 to prevent energy from leaking out of the structure between the serrated edges 39a and 39b of the slots 37 and the metal fin 35.

The electric field lines of the modes that can exist in the non-contacting short, previously described, are illustrated in FIGS. 4A and 4B. The mode shown in 4A is the ordinary TE_{10} mode which is below cut-off. As disclosed, the half wavelength chokes 46a and 46b on the outside of the guide prevent this mode from radiating. The second possible mode, illustrated in FIG. 4B is the ordinary TEM transmission line mode. This mode exists if the metallic fin 45 is asymmetrically positioned. This mode can be prevented from propagating by the known expedient of slotting the choke at 49a and 49b in a well known manner. Under the conditions that these serrations are effectively a quarter-wavelength deep, the attenuation suffered by the TEM mode is very great. Measurements on this type of non-contacting short circuit element indicate that at about 9500 mc., the standing wave ratios of greater than 40 db are easily obtained. This represents less than a 4% power loss.

The rapid phase variation and pre-programmed scanning is accomplished by a line array of phase shifters of the type shown in FIG. 1. Rotating the serrated discs in the slots of the terminating stubs produces the particular scan. The shorting vanes are in the form of large diameter thin discs and having each set attached to a common shaft. The phase shifters can be arranged in various manners to produce a particular pre-programmed scanner. In the scanner, energy of constant phase is fed into the input waveguides 51 of the scanner from a line source 50 and emerges from the radiating apertures 52 with phases determined by the positions of the toothed wheels 54 and 55 in the non-conducting short circuit stubs, as shown in FIG. 5.

Quantitative expressions for the wheel radii and rotation speeds necessary for a scanner of the type shown in FIG. 5 in terms of the pertinent system parameters are:

$$R = \frac{nL}{2\pi F}$$

$$\text{r.p.m.} = \frac{\text{prf}}{nm} \times 60$$

(1)

where

R = wheel radius

r.p.m. = revolutions per minute of the wheel

n = number of beam positions to be scanned per wheel revolution

m = number of pulses per beam position

prf = pulse repetition frequency of the radar

L = average length of wheel circumference required for switching (i.e., a distance equal to or less than the height of the non-contacting short circuit)

F = fraction of a scanning period that it takes to switch from one beam position to another (i.e., the fractional dead time)

Table I lists some typical radii and wheel rotation speeds that are required for a millimeter scanner, operating at 8.6 mm. = 0.339 in., assuming a value of $L = 0.17$ in., equal to one-half the height of the non-contacting short circuit.

TABLE I

Size and rotation speed of serrated discs for use in a millimeter ($\lambda = 8.6$ mm.) scanner

prf	n	m	L (in.)	F	r.p.m.	R (in.)
6,000	42	4	0.17	0.25	4,290	4.49
6,000	42	4	0.17	0.10	4,290	11.36
6,000	42	3	0.17	0.25	5,720	4.49
6,000	42	3	0.17	0.10	5,720	11.36
4,000	42	4	0.17	0.25	2,860	4.49
4,000	42	4	0.17	0.10	2,860	11.36
4,000	42	3	0.17	0.25	3,810	4.49
4,000	42	3	0.17	0.10	3,810	11.36
6,000	22	4	0.17	0.25	8,190	2.38
6,000	22	4	0.17	0.10	8,190	5.95
6,000	22	3	0.17	0.25	10,900	2.38
6,000	22	3	0.17	0.10	10,900	5.95
4,000	22	4	0.17	0.25	5,450	2.38
4,000	22	4	0.17	0.10	5,450	5.95
4,000	22	3	0.17	0.25	7,280	2.38
4,000	22	3	0.17	0.10	7,280	5.95

The periodic phase perturbations along the radiating aperture of the scanner, which is effectively a line source composed of a large number of separate radiators cause a deterioration of the radiation pattern. The following shows the magnitude of this effect.

A uniform line source of length L will radiate a beam in a direction measured with respect to the normal, if the total phase delay ψ along the array is

$$\psi = \frac{2\pi L}{\lambda} \sin \theta_0$$

If an array is composed of N separate radiators of width L/N , it is necessary to adjust the phase difference ψ' between individual radiators so that $\psi' = \psi/N$, in order to obtain a beam at an angle θ_0 . When the array is composed of horns, each one of which is assumed to have a constant phase over its aperture, it is seen that the resultant phase distribution of the array is a sawtooth approximation to the desired linear phase variation. This periodicity in the phase front will result in reduced gain and spurious side lobes.

The radiation from an array of length L , neglecting the obliquity factor, is

$$g(\sin \theta) = \int_{-\frac{L}{2}}^{+\frac{L}{2}} f(x') e^{j \frac{2\pi}{\lambda} x' \sin \theta} dx' \quad (3)$$

where

$$f(x') = A(x') e^{-j \left[\frac{\psi x'}{L} + \sum_{n=1}^{\infty} \alpha_n \sin \frac{2\pi n N x'}{L} \right]}$$

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If the substitutions

$$x' = \frac{Ly}{2} \text{ and } u = \frac{\pi L}{\lambda} \sin \theta$$

are made in Equation 3, it is found that

$$g(u) = \frac{L}{2} \int_{-1}^{+1} A(y) \epsilon^{jy(u-\frac{\psi}{2}) - \sum_{n=1}^{\infty} \alpha_n \sin \pi n Ny} dy \quad (4)$$

The Fourier coefficients α_n can be evaluated in the usual fashion. For example, for a sawtooth phase variation it is found that

$$\alpha_n = \frac{\psi}{\pi n N}$$

In order to integrate Equation 4 it is necessary to make use of the following identities:

$$\epsilon^{-j\alpha_n \sin \pi n Ny} = \sum_{\xi=-\infty}^{\xi=\infty} j_{\xi}(\alpha_n) \epsilon^{-j\xi \pi n Ny}$$

and

$$J_{-\xi}(\alpha_n) = (-1)^{\xi} J_{\xi}(\alpha_n)$$

Equation 4 becomes

$$g(U) = \frac{L}{2} \int_{-1}^{+1} A(y) \epsilon^{jy(u-\frac{\psi}{2})} \sum_{n=1}^{\infty} \sum_{\xi=-\infty}^{\infty} J_{\xi}(\alpha_n) \epsilon^{-j\xi \pi n Ny} dy \quad (5)$$

If the summation in the integrand of Equation 5 is written explicitly,

$$g(U) = \frac{L}{2} \int_{-1}^{+1} A(y) \epsilon^{jy(u-\frac{\psi}{2})} [\alpha_0 + \alpha_{-1} \epsilon^{-j\pi Ny} + \alpha_1 \epsilon^{j\pi Ny} + \alpha_{-2} \epsilon^{-j2\pi Ny} + \alpha_2 \epsilon^{j2\pi Ny} + \dots] dy \quad (6)$$

is obtained. Now let the pattern of the aperture with no periodic phase variations be

$$G(u - \frac{\psi}{2}) = \frac{L}{2} \int_{-1}^{+1} A(y) \psi(u - \frac{\psi}{2}) dy \quad (7)$$

Then, using the complex translation theorem, it is seen that

$$g(u) = \alpha_0 G(u - \frac{\psi}{2}) + \alpha_{-1} G(u - \frac{\psi}{2} - N\pi) + \alpha_1 G(u - \frac{\psi}{2} + N\pi) + \alpha_{-2} G(u - \frac{\psi}{2} - 2N\pi) + \dots \quad (8)$$

Thus the periodic phase disturbances give rise to a diffraction pattern having the same shape as the pattern of the aperture with no disturbances, as well as a series of auxiliary patterns of the same shape but smaller amplitude, split off at equal angles from the main beam. These auxiliary patterns will all have their peaks at imaginary angles θ , if

$$\frac{L}{N} \leq \frac{\lambda}{1 + |\sin \theta_0|} \quad (9)$$

otherwise, they will appear at real angles.

In general, the coefficients $\alpha_0, \alpha_{-1}, \alpha_{+1}, \dots$ must be evaluated by expanding the integrand of Equation 5. However, for the case of sawtooth phase variations, it is possible to obtain another expression from simple array theory for $g(U)$ when $A(y)$ is constant.

The expression is, for N even,

$$g(u) = \frac{1}{N} \frac{\sin(\frac{u}{N})}{\frac{u}{N}} \frac{\sin[N(\frac{u}{N} - \frac{\psi}{2N})]}{\sin(\frac{u}{N} - \frac{\psi}{2N})} \epsilon^{-j(N-1)(\frac{u}{N} - \frac{\psi}{2N})} \quad (10)$$

Thus it is seen, for example, that the coefficient α_0 is obtained by substituting $U = \psi/2$ in Equation 10, α_{-1} is obtained by letting $U = 4/2 + N\pi$. Although these coefficients $\alpha_0, \alpha_{-1}, \alpha_1, \dots$ are evaluated for the specific case of uniform amplitude distribution along the radiating horns, it is seen from the general derivation that they will have the same value no matter what amplitude distribution is used. The values of some of these coefficients are tabulated in Table II below, as a function of the phase difference ψ' between adjacent horns.

TABLE II

Normalized amplitude coefficients of the principal and auxiliary patterns of an aperture with arbitrary amplitude distribution but sawtooth phase variations

ψ'	α_0	α_{-1}	α_1	α_{-2}	α_2	α_{-3}	α_3
$\pi/8$	0.994	0.058	-0.066	0.030	-0.032	0.021	-0.021
$\pi/4$	0.974	0.108	-0.139	0.057	-0.065	0.039	-0.042
$\pi/2$	0.900	0.180	-0.300	0.100	-0.129	0.069	-0.082

The ratio of the coefficients α_1/α_0 can be interpreted only as the ratio of the amplitude of the first auxiliary lobe to that of the main radiated beam. It is not to be construed as the ratio of the first side lobe in the diffraction pattern to that of the maximum of the radiated beam. This latter quantity can be determined only by computing in detail the diffraction pattern for each assumed aperture amplitude distribution. If the inequality of Equation 9 is not satisfied and the first auxiliary lobe maximas appear at real angles, the ratio of α_1/α_0 will probably set a lower bound on attainable side-lobe level. If, on the other hand, the inequality of Equation 9 is satisfied and no auxiliary lobe maxima appear at real angles the side-lobe level of the diffraction pattern will be nearly the same as that of the diffraction pattern of the particular antenna which has no phase variations in its aperture.

It is not possible to consider α_0^2 as the relative gain of the antenna because the main and auxiliary diffraction patterns are not orthogonal, powerwise. When the individual radiating horns are far enough apart so that the inequality of Equation 9 is not satisfied, α_0^2 gives a very qualitative estimate of the relative gain. It might be higher or lower than the true value, depending on whether the auxiliary lobes are displaced by large or small angles from the main lobe. When the individual horns are close enough that the inequality of Equation 9 is satisfied the relative gain would probably be much nearer unity than indicated by α_0^2 .

The following is an example of the dimensions of the phase shifter of FIG. 1 operating at approximately 9500 mc. The radii of serrated discs 15 and 16 is 3.87, 3.71, 3.55, 3.39, 3.23 and 3.07 inches. The discs were made of No. 14 gage brass and have their center of rotation spaced 8 1/2 inches apart.

While there has been described what is at present considered the preferred embodiments of this invention, it will be obvious to those skilled in the art that various changes and modifications may be made herein without departing from the invention.

What is claimed is:

1. A pre-programmed scanning antenna comprising a line array of input waveguides, a feed waveguide for coupling radio frequency energy to one end of said input waveguides, a line array of output waveguides, a radiating aperture at one end of each output waveguide, a plurality of directional couplers interconnecting each input waveguide with a respective output waveguide, a plurality of variable phase shifters terminating the other end of the input waveguides and the output waveguides.

2. A pre-programmed scanning antenna according to claim 1 wherein each variable phase shifter comprises a slotted terminating stub and a serrated disc rotated within the slot to vary the electrical length of said input and output waveguides.

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3. A line source scanning antenna system for directing radiation at various angles in a predetermined sequence comprising a line array of input and output waveguides, a separate directional coupler for coupling radio frequency energy from each input waveguide to an output waveguide, a slotted stub terminating one end of each waveguide, a variable radius disc rotated in each slot for varying the electrical length of each waveguide, means for applying radio frequency energy to the other end of said input waveguides, radiating means attached to the other end of said output waveguides.

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4. A line source scanning antenna system according to claim 1 wherein the slotted stub includes radio frequency chokes for preventing extraneous radiation at the slot.

References Cited in the file of this patent**UNITED STATES PATENTS**

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