A Radial Line Slot Antenna for 12 GHz Satellite TV Reception

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Abstract—**The analysis and design of novel planar antennas which can radiate circularly polarized pencil beams in x-band are presented for the application of receiving direct broadcast from a satellite (DBS). This antenna belongs to a class of slotted waveguide antennas and high efficiency is expected in principle. A circular two-dimensional aperture is located on the top plate of a twofold radial waveguide and is excited by a radially inward traveling transverse electromagnetic (TEM) mode in the upper waveguide. Slots are arrayed on the aperture spirally in such a way that they can couple with the radial currents flowing over the aperture plate to produce a circularly polarized broadside beam. Promising performances of the antenna are predicted theoretically. Experiments are performed with respect to basic characteristics of the antenna and they show the validity of the design and analysis.**

I. INTRODUCTION

DIRECT BROADCAST from a satellite (DBS) using 12 GHz band has become in commercial use in Japan [1]. Subscriber antennas for receiving DBS should possess high gain characteristics of about 36 to 37 dBi. To satisfy this requirement, reflector antennas with the aperture diameters of more than 70 cm have been used. Antennas with curved reflectors, however, generally suffer from degradation of performances due to rain and snow falls on them. As one of the countermeasures, planar antennas using microstrips have also been studied and developed [2]–[5]. Unfortunately, for microstrip antennas with large aperture sizes, the conductor and the dielectric losses become notable; in designing a subscriber antenna stated before, its antenna efficiency is limited to 40–50 percent at most.

Goebels and Kelly [6] reported another approach of an x-band planar antenna using a radial waveguide with annular slots. Its feeding structure, however, is rather complicated since two modal pairs in the waveguide are necessary for circular polarization. Furthermore, no qualitative discussion with respect to its efficiency was carried out there.

This paper proposes a planar antenna of a new type and discusses its basic characteristics. It also belongs to a class of slotted waveguide antennas and high efficiency is expected in principle. Key features of this antenna consist in its slot arrangement and a twofold radial waveguide. The basic idea of this antenna named "radial line slot antenna" was first proposed by one of the authors in 1980 [7]. It has a circular aperture on a upper plate of a radial waveguide in which a rotationally symmetric transverse electromagnetic (TEM) traveling-wave mode is supported. Slots are arrayed on the aperture so that they can couple by a constant factor with the radial currents flowing over the aperture plate and produce circular polarization. Though the slot positions to realize a circularly polarized broadside beam are not symmetric, symmetry of the inner field to excite the aperture should not be appreciably disturbed. For the purpose of realizing a stable symmetric inner field, the radial waveguide is folded and is fed at the center of the lower waveguide. Then a radially inward traveling-wave mode is supported in the upper waveguide and couples to slots. For this structure, the following two advantages are also expected. 1) Coupling of the slots modifies a steeply tapered aperture amplitude distribution to a more uniform one and the directive gain is increased. 2) The power dissipated in a dummy load at the end point can be diminished and the antenna efficiency is improved. In designing a radial line slot antenna with high efficiency, a slow wave structure in the upper waveguide is also necessary to suppress grating lobes from the array.

By introducing a fictitious model for coupling between slots and an inner field, basic characteristics of the antenna such as radiation patterns, axial ratio, and antenna efficiency are calculated and its promising features are pointed out. Antenna efficiency of no less than 70 percent is predicted for example. Experiments are also performed with respect to the input voltage standing-wave ratio (VSWR), aperture distributions, and Fresnel radiation patterns. The results confirm the design and the analysis of this antenna.

II. ANTENNA DESIGN

A. Structure

Construction of a radial line slot antenna is presented in Fig. 1(a). Three plates equally spaced by d compose a twofold radial line waveguide, the top of which is an aperture with slots. In the following discussion, a cylindrical coordinate system (ρ, ψ, z) is used to denote the aperture, while a standard spherical coordinate system (r, θ, ϕ) denotes the point of observation. The power flow in this antenna is pictured in Fig. 1(b). Electromagnetic power is fed at the center ($\rho = 0$) of the lower waveguide through a coaxial cable and a radially outward traveling-wave mode is generated. At the outer edge of the waveguide ($\rho = \rho_M$), it is transferred into

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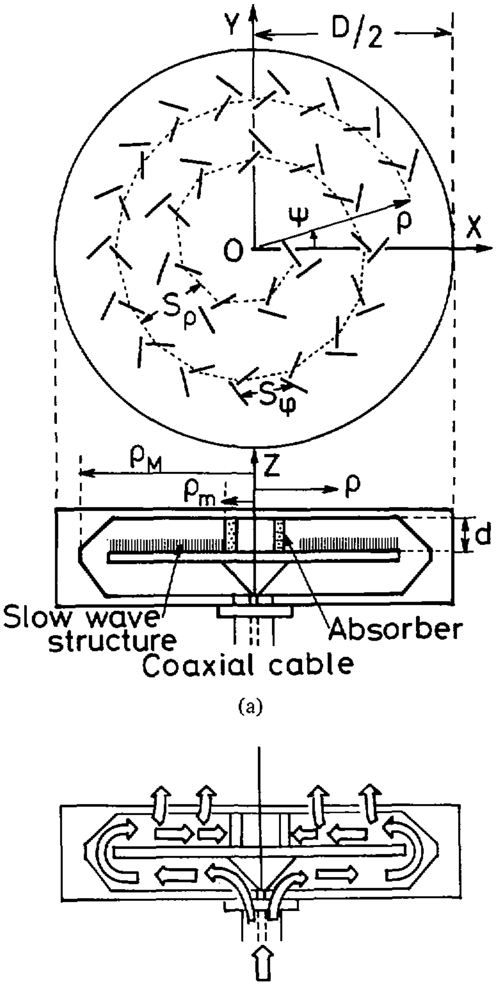

Fig. 1. A radial line slot antenna and its design parameters. (a) Structure. (b) Power flow in the waveguide.

a radially inward traveling-wave mode in the upper waveguide. Then some portions of the energy are radiated from the slots while the remainder reaches a dummy load (absorber) at $\rho = \rho_m$. In contrast with the standard radial waveguide, this folded one supports the radially inward (not outward) traveling-wave mode. This design is based on experiments which show that the symmetry of this inward traveling-wave mode is more stable than that of the outward traveling-wave mode in the presence of asymmetrically positioned slots. There are two more advantages for this structure viewed in terms of antenna efficiency as will be explained later.

The outer edge of the waveguide and the feeding point have 45° tapered structures so as to suppress reflection from them. Slots on the top plate consist of many slot pairs each one of which is a unit radiator of a circularly polarized wave. Slot pairs are arrayed in sequence along a design spiral. S_ρ and S_ϕ indicate the spacings between adjacent slot pairs along the ρ and ϕ direction, respectively. In order to suppress grating lobes from the array, both S_ρ and S_ϕ must be smaller than the free space wavelength λ_0. In our design, S_ρ must equal λ_g the wavelength in the upper waveguide, while S_ϕ is determined arbitrarily. To satisfy these conditions, a slow-wave structure may be adopted in the upper waveguide and $S_\rho (= \lambda_g)$ is diminished to $\zeta\lambda_0$, where $\zeta (< 1.0)$ is the slow wave factor.

B. Aperture Distribution

Slots are designed so as to couple with the currents on the top plate by a constant factor; so the magnetic field in the upper waveguide determines an excitation coefficient of a slot.

In the first place, we discuss the magnetic field neglecting the coupling to slots. For $d < \lambda_g/2$, the only possible symmetrical mode is TEM one whose magnetic field is represented as follows:

$$H_\phi = H_1^{(1)}(k\rho) - \Gamma \frac{H_1^{(1)}(k\rho_m)}{H_1^{(2)}(k\rho_m)} H_1^{(2)}(k\rho) \tag{1}$$

where $k (= 2\pi/\lambda_g)$ is the wavenumber in the waveguide and Γ is the complex reflection coefficient of the absorber at $\rho = \rho_m$. $H_1^{(1)(2)}(z)$ is the Hankel function of the first (second) kind of order one. For $k\rho \gg 1$, H_ϕ can be written in a more simple form as

$$H_\phi \simeq \sqrt{\frac{2}{\pi k\rho}}\, e^{J(k\rho - 3/4\pi)} - \Gamma \sqrt{\frac{2}{\pi k\rho}}\, e^{J\{k(2\rho_m - \rho) - 3/4\pi\}}. \tag{2}$$

The first term represents the radially inward traveling-wave mode while the second term represents the radially outward traveling-wave mode. The only existing aperture currents associated with this field are radial ones which are proportional to H_ϕ. Consequently, the excitation coefficient of the slot at ρ is given by (2). If the slot density is almost constant on the aperture, (2) also gives the aperture distribution.

Now suppose that $\Gamma = 0$ and only the radially inward traveling-wave mode (the first term) exists. Its amplitude has the tapered distribution of the form of $\rho^{-0.5}$ and is not favorable in terms of the directive gain of an antenna.

In the discussion stated above, the coupling between the field and the slots is neglected. If the assumption holds that at every point on the aperture the power radiated from slots per area is proportional to that in the waveguide and the rotational symmetry of the field is not appreciably disturbed, the coupling can be taken into account. Using the parameter $\alpha(m^{-1})$ for the coupling proportionality, we get the aperture distribution $F(\rho)$ as follows:

$$F(\rho) = \sqrt{\frac{\alpha}{\rho}}\, e^{(jk+\alpha)\rho} - \Gamma \sqrt{\frac{\alpha}{\rho}}\, e^{(jk+\alpha)(2\rho_m - \rho)}. \tag{3}$$

The excitation coefficient of the slot located at ρ with equal density is also given by this equation.

For the ideal case of $\Gamma = 0$, the initial amplitude taper of $\rho^{-0.5}$ is modified by the factor of $e^{\alpha\rho}$ and for some $\alpha (> 0)$, the aperture distribution is improved to be more flat. This is the second reason for the selection of the twofold radial waveguide, irrespective of its complexity. In addition, once an almost uniform aperture excitation is realized, the portion of the power wasted in the absorber is only ρ_m/ρ_M of the total power. This is the third advantage of this structure.

Aperture distributions for different α are calculated for $\Gamma =$

0 and presented in Fig. 2. Results for α of about 1–5 become rather flat and are favorable in terms of the directive gain of an antenna.

As is explained in the next section, the radially outward traveling-wave mode (the second term) contributes only to the cross-polarized radiation and does not have any effects on the co-polarization. Above discussion for $\Gamma = 0$, therefore, also applies in the case $\Gamma \neq 0$.

C. Slot Arrangement

A design of the slot arrangement consists of two steps. The first one is the design of one slot pair while the second one is the arrangement of sequential slot pairs. In the design, only a radially inward traveling-wave mode is considered. In addition, its amplitude variation is neglected while the phase delay associated with $H_1^{(1)}(k\rho)$ is taken into account. The slots are numbered ($n = 1$ to N) in the increase order of the coordinate as are shown in Fig. 3.

ρ_n	the distance from origin O to the center P_n of the slot #n
β_{2n-1}	the angle $P_{2n-1}OP_{2n}$
Θ_{2n-1}	slot inclination angle from radial direction
$2L$	slot length
δ	the distance between two slots.

The parameters δ and $2L$ are constant for every slot pair and are given in advance.

Fig. 3(a) shows the parameters for designing the first slot pair ($n = 1, 2$). Two slots #1 and #2 make the right angle with each other. Furthermore, both slots are inclined by the equal angle Θ_1 from the radial direction and they are excited with equal amplitude. So if they are excited with a relative phase shift of $\pi/2$, they work as an unit radiator of the right-hand circular polarization (RHCP). The condition is written as follows:

$$\arg\ (H_1^{(1)}(k\rho_2)) - \arg\ (H_1^{(1)}(k\rho_1)) = \frac{\pi}{2} \tag{4}$$

where arg(x) indicates the phase of x. Furthermore, the geometry requires the next relation:

$$\rho_2 \sin \Theta_1 - \rho_1 \cos \Theta_1 = L + \delta \tag{5}$$

$$\beta_1 = 2\Theta_1 - \frac{\pi}{2}. \tag{6}$$

Solving these transcendental equations (4)–(6) for the specified initial value of ρ_1, we obtain ρ_2, β_1, and Θ_1. According to the design in which β_1 and Θ_1 are parameters to be determined, relatively longer slots can be accommodated compared with the design in which $\beta_1 = 0$ and $\Theta_1 = \pi/4$ is prescribed.

In the next place, the position of the next slot pair is derived. Fig. 3(b) shows the design parameters, where ρ_3 and Φ_3 are to be determined for given ρ_1 and S_ϕ. Design policy to produce a circularly polarized broadside beam can be summarized as that slot pairs geometrically rotated by Φ_3 should obtain the same amount of electrical phase advance.

$$\Phi_3 = \arg\ (H_1^{(1)}(k\rho_3)) - \arg\ (H_1^{(1)}(k\rho_1)). \tag{7}$$

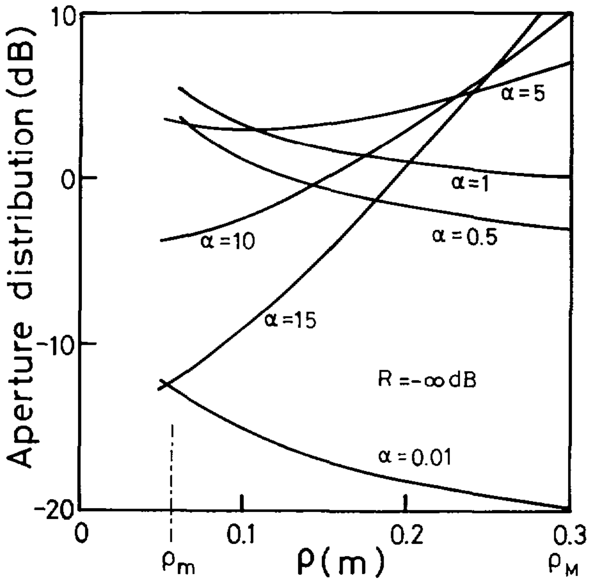

Fig. 2. Aperture distributions for different coupling factor α.

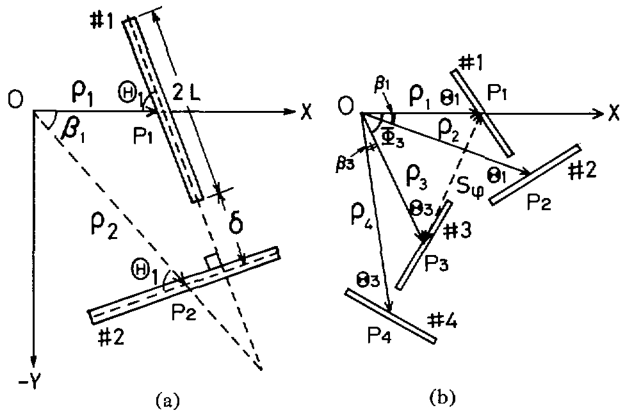

Fig. 3. Coordinates and design parameters for the slot arrangement on the aperture. (a) The first slot pair. (b) The first two slot pairs.

Basic geometries together with the result in (7) require the following relation:

$$S_\phi^2 = \rho_1^2 + \rho_3^2 - 2\rho_1\rho_3 \cos \{\arg\ (H_1^{(1)}(k\rho_3)) - \arg\ (H_1^{(1)}(k\rho_1))\}. \tag{8}$$

Consequently, (8) determines ρ_3 and then Φ_3 is calculated by (7). Arrangements for successive slots can also be determined by making iterative use of (4)–(8). In more detail, β_3 and Θ_3 are derived from ρ_3 in (4)–(6), while ρ_5 and Φ_5 are determined from ρ_3 by (7) and (8). In this design, the change in Θ_n for different slots over the aperture is considerably small. Therefore, every slot couples with the radial currents by almost a constant factor. If the relation $k\rho \gg 1$ holds, (7) is simplified to give

$$\rho_3(\psi) = \psi \cdot \lambda g/2\pi + \rho_1. \tag{9}$$

This result indicates that the trace on which slot pairs are located is a kind of spiral. Since $\rho(\psi + 2\pi) - \rho(\psi) = \lambda_g$, S_ρ should become approximately λ_g. Consequently, if we set S_ϕ constant, all the slot pairs are arrayed with almost equal

TABLE I
DESIGN PARAMETERS OF ANTENNAS (*A*) AND (*B*)

Type	D (m)	ρ_M (m)	ρ_m (m)	d (m)	δ (m)	2L (m)	S_ϕ/λ_0	S_ρ/λ_0	N
(A)	0.6	0.3	0.06	0.0075	0.001	0.0125	0.7	1.0	1238
(B)	0.6	0.3	0.06	0.0075	0.001	0.0125	0.5	0.8	2184

density over the aperture, that is, one pair for each area of $S_\rho \times S_\phi$.

D. Fabrication

For the further analyses and experiments, different types of antennas with diameters of 0.6 m are fabricated. The detailed design parameters are listed in Table I. Antenna (*A*) has no slow-wave structure. Antenna (*B*), on the other hand, has a dielectric sheet ($\epsilon_r = 2.5$) with the thickness of 3 mm in the bottom of the upper waveguide and λ_g is diminished to 0.8 λ_0 for the grating lobe suppression. It is supposed that the dielectric sheet affects not only the wavelength λ_g but also the coupling factor α. Smaller S_ϕ is chosen for antenna (*B*) so as to obtain a coupling factor α comparable with antenna (*A*).

III. Theoretical Analysis

A. Radiation Pattern

Using the slot excitation coefficients derived from (3) together with their positions determined by (4)–(8), we can obtain radiation patterns through cumbersome but straightforward calculation. Fig. 4 presents a wide-angle radiation patterns of antenna (*B*) where $\alpha = 1$ is used for calculation. The condition $S\rho < \lambda_0$ is satisfied in this antenna and the grating lobes do not appear. Since the edge illumination level is rather low for $\alpha = 1$, the sidelobe levels are very low and beamwidth (BW) is as wide as 2.7°. The discrepancies between the patterns in $x - z$ plane ($\phi = 0°, 180°$) and $y - z$ plane ($\phi = 90°, 270°$) are sufficiently small and almost rotationally symmetric radiation patterns are obtained. Furthermore, E_θ and E_ϕ have almost the equal amplitude and the relative phase shift of $\pi/2$ in the broadside. So, the pure circular polarization is expected.

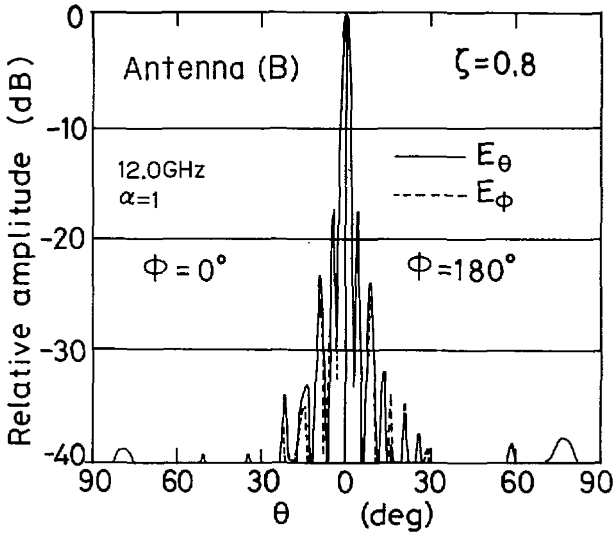

Fig. 4. Calculated far-field radiation patterns of E_θ and E_ϕ. Antenna (*B*) where $\alpha = 1$ is assumed.

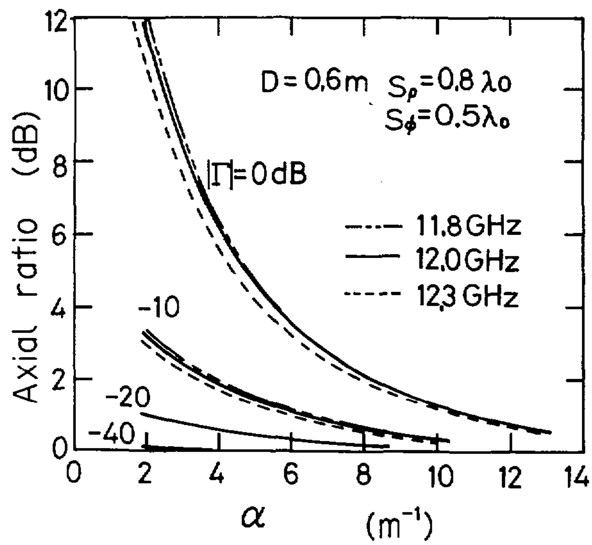

Fig. 5. Calculated axial ratio of antenna (*B*) versus coupling factor α for different reflection coefficients of the dummy load.

B. Axial Ratio

In a radial line slot antenna, the radially inward traveling-wave mode radiates the co-polarized wave while the radially outward traveling-wave mode radiates the cross-polarized wave. Accordingly, the reflection coefficient Γ of the absorber should be small for high polarization purity of antennas. In addition, if the coupling factor α is large and the wave decays rapidly, the power reaching the absorber becomes small and cross-polarized wave is also suppressed. Fig. 5 shows the axial ratio of antenna (*B*) for different coupling factors and the reflection coefficient of the absorber. Consider $|\Gamma| = -15$ dB. The axial ratio of about 1 dB is realized for α more than five. In addition, for α more than 10, the axial ratio approaches to unity and even a metal short ($|\Gamma| = 1$) can be used in place of the absorber. Furthermore, the frequency dependence of the axial ratio is very small.

These results are not peculiar for antenna (*B*) but general for all the radial line slot antennas.

C. Efficiency

With slot excitation coefficients given by (3), antenna gain is written in a form of

$$G = \frac{4\pi r^2 \mathbf{E}(0, 0) \cdot \mathbf{E}^*(0, 0)}{\iint \mathbf{E}(\theta, \phi) \cdot \mathbf{E}^*(\theta, \phi) r^2 \sin\theta \, d\theta \, d\phi} \cdot (1 - e^{2\alpha(\rho_m - \rho_M)}). \quad (10)$$

The fractional part represents the directive gain while the additional factor accounts for the power loss dissipated in the dummy load. The latter part approaches to unity when the diameter ρ_M becomes large and (or) the coupling factor α increases. Antenna efficiency is calculated from the following relation.

$$\eta = \left(\frac{\lambda_0}{\pi D}\right)^2 \cdot G. \tag{11}$$

The dependence of the antenna efficiency upon the coupling factor and the spacings S_ρ and S_ϕ is discussed first. Fig. 6 shows the efficiency of antennas with a diameter (D) of 0.6 m at 12 GHz. It is found that the antenna efficiency takes its maximum at $\alpha \simeq 5$. Decrease of the efficiency for $\alpha < 5$ stems from the power loss in the dummy load while that for $\alpha > 5$ stems from the nonuniform aperture distribution where the amplitude decreases too rapidly toward its center. With respect to the spacings, the following are understood. When S_ρ and (or) S_ϕ increase and approach λ_0, the efficiency seriously decreases due to the appearance of the grating lobes. In addition, the smaller spacings generally increase the efficiency, while it is saturated at about $S_\rho = 0.6\ \lambda_0$ and $S_\phi = 0.7\ \lambda_0$ and no drastic improvement is found for S_ρ and S_ϕ smaller than these. Suppose that the coupling factor α is controlled to be about five for $S_\rho = S_\phi = 0.8\ \lambda_0$ for example, sufficiently high efficiency of more than 70 percent is realized in the frequency range of 11.7–12.3 GHz.

The efficiency of antennas with different diameters is also compared. Almost the similar results are obtained except that the coupling factor α at which the efficiency takes its maximum is in inverse proportion to the diameter D.

IV. Experiments

A. Input VSWR

A radial line slot antenna is fed by a coaxial cable through a simple N-type connector. The input VSWR is first examined since certain amount of reflection is expected for the twofold radial line waveguide. Experimental results of return loss for antennas (A) and (B) are presented in Fig. 7. The wavy lines indicate calibration levels. In the frequency range of 11.8–12.5 GHz, antenna (A) with the absorber at its center has the maximum reflection of about -12 dB. This reflection consists of the contributions from the connector, the edge of the radial waveguide, and the slots. The reflection level increases to about -8 dB for antenna without absorber. It must be added that the reflection for antennas without 45° tapers at the center and the edge of the waveguide is about -1 dB and intolerably high. Therefore the roles of 45° tapers are important.

B. Aperture Distribution

Fig. 8 shows the envelopes of the field strength measured at $z = 6$ cm over the aperture of antenna (B). If the coupling between slots and the field is negligibly small, the distribution should have the form of $\rho^{-0.5}$. In Fig. 8, on the contrary, the amplitude decays toward the center of the aperture and considerable amount of coupling is estimated. Small increase of the amplitude at the vicinity of ρ_m is due to the reflected waves from the imperfect absorber. Comparison of the distributions in Fig. 8 with the theoretical ones in Fig. 2 clarifies the validity of the coupling model in the analysis. The coupling factor measured for antenna (B) at 12.2 GHz is about five. The change due to frequency is not serious though α is a bit larger for 12.0 GHz. The value α experimentally obtained here is not so different from the desired one which maximizes the aperture efficiency in theory. Besides, the coupling factor α for antenna (A) without the slow-wave structure is also measured to be about 10 at 12.0 GHz.

The coupling factor α depends on many parameters, that is, the slot length, the slot spacings and the thickness of the radial waveguide etc. A detailed design of this factor is left as future problems. Nevertheless, the possibility of a high efficiency antenna has partly been verified by these experiments.

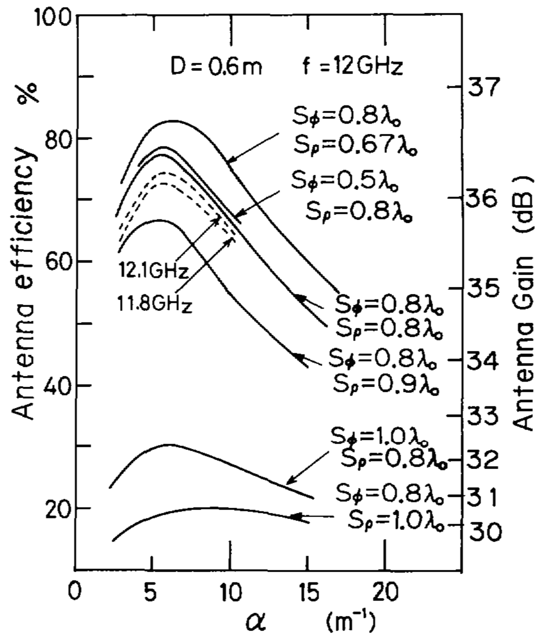

Fig. 6. Calculated antenna efficiency for different slot spacing S_ρ and S_ϕ. $D = 0.6$ m, $f = 12$ GHz. $\rho_M = 0.3$ m, $\rho_m = 0.06$ m.

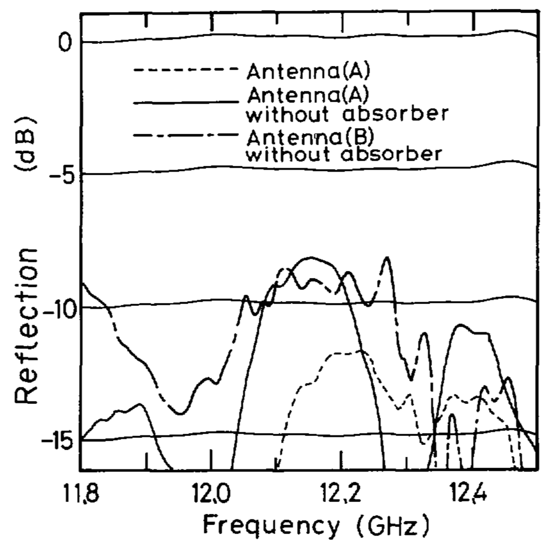

Fig. 7. Measured input VSWR (reflection) for antennas (A) and (B).

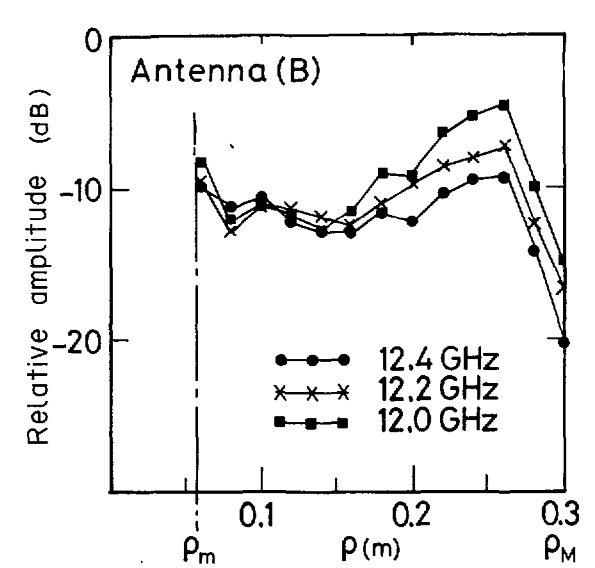

Fig. 8. Aperture distributions measured at 0.06 cm from the aperture of antenna (*B*). ρ_M = 0.29 m, ρ_m = 0.06 m, ζ = 0.81, f_0 = 12.2 GHz.

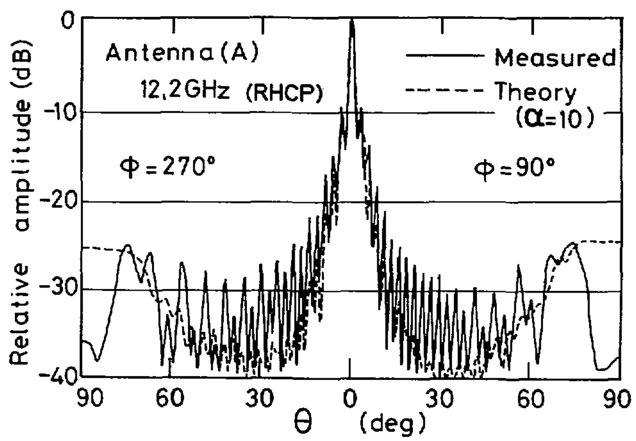

Fig. 9. Fresnel radiation patterns of antenna (*A*) in $y - z$ plane at 12.2 GHz. α = 10 is used for the calculation. r = 5 m, right-hand circular polarization.

C. Radiation Pattern

The basic operations of the radial line slot antenna such as the rotational symmetry of the inner fields and the grating lobe suppression are discussed via radiation patterns measured in Fresnel region (r = 5 m). The wide-angle patterns of antenna (*A*) are presented in Fig. 9. The first sidelobe levels are about -10 dB and a relatively high edge illumination level of the aperture is suggested. The grating lobes of -23 dB appear at $\theta = 75°$. Only the pattern in $Y - Z$ plane is presented, since it is almost rotationally symmetric. A pattern calculated for the coupling factor $\alpha = 10$ is also presented for comparison. They are showing qualitative agreements, though slight discrepancies are observed at $\theta = 30°$. Fig. 10 shows the results for antenna (*B*), where $\alpha = 5$ is used for calculation. Almost the similar results with antenna (*A*) are derived except that the grating lobes are perfectly suppressed by the slow-wave structure.

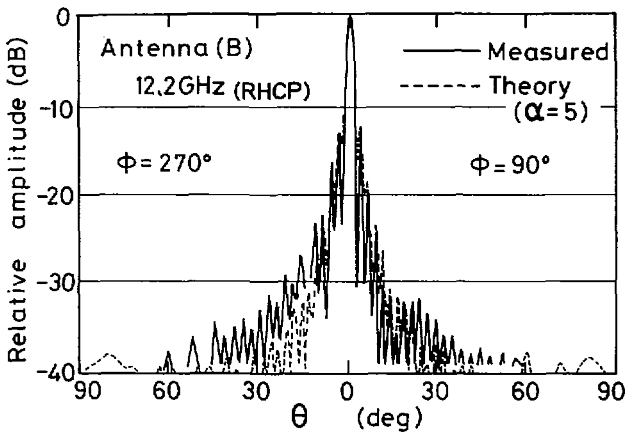

Fig. 10. Fresnel radiation patterns of antenna (*B*) in $y - z$ plane at 12.2 GHz. α = 5 is used for the calculation. r = 5 m, right-hand circular polarization.

The agreement of the experiments with the theory indicates the validity of the design and analysis of the radial line slot antenna.

Conclusion

A planar type slotted waveguide antenna is proposed and its design method is summarized. A radially inward traveling TEM mode is adopted so as to increase the antenna efficiency. Antenna characteristics are analyzed and the promising performances are pointed out. Experimental results are also presented to examine the basic antenna performances. All these results indicate the validity of the design and the analysis of this antenna.

Optimization of the coupling of slots together with measurements of the antenna efficiency in the far field is open for further investigation.

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References

[1] Y. Ueda, K. Ishibashi, and Y. Ichikawa, "Operational broadcasting satellite program in Japan," IAF, Rome, Italy, 1981.

[2] E. Rammos, "New wideband high-gain stripline planar array for 12 GHz satellite TV," *Electron. Lett.*, vol. 18, no. 6, pp. 252-253, Mar. 1982.

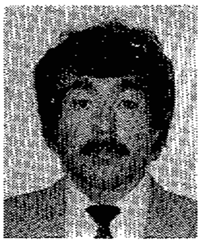
[3] J. C. Williams, "A 36 GHz printed planar array," *Electron. Lett.*, vol. 14, no. 5, pp. 136-137, Mar. 1978.

[4] M. Haneishi, S. Saito, S. Yoshida, and N. Goto, "A circularly polarized planar arrays composed of the microstrip pairs element," IECE Japan, Tech. Rep., AP83-64, Sept. 1983.

[5] S. Nishimura, Y. Sugio and T. Makimoto, "A crank-type microstrip line planar antenna for circular polarization," in *Conf. Rec., Nat. Conv. IECE Japan*, vol. 3, S7-16, 1983, pp. 299-300.

[6] F. J. Goebels and K. C. Kelly, "Arbitrary polarization from annular slot planar antennas," *IRE Trans. Antennas Propagat.*, vol. AP-9, no. 4, pp. 342-349, July 1961.

[7] N. Goto and M. Yamamoto, "Circularly polarized radial-line slot antennas," IECE Japan, Tech. Rep., AP80-57, Aug. 1980.



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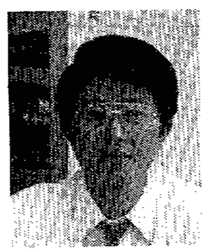
After working at the Furukawa Electric Corp. from 1963 to 1968, he was a Consulting Engineer in the U.S. concentrating on CATV and the security field, from 1968 to 1970. On returning to Japan, he expanded his business and established Arimura Institute of Technology in 1974. The company's activities are the distribution and research and development of Tx and Rx antennas in TV and FM field and IC card-Smart card. He is first inventor and the patent holder of IC card. He has worked on the development of such IC card systems and DBS receiving antennas for future use.



Yoshiharu Ito was born in Hita City, Ohita, Japan, on February 1, 1951. He received the B.S. degree in electrical engineering from Kumamoto University, Kumamoto, Japan, in 1973.

He joined Japan Broadcasting Corporation (NHK) in 1973, and has engaged in design and development of various antennas for terrestrial broadcasting systems at NHK Engineering Headquarters, Tokyo.

Mr. Ito is a member of the Institute of Television Engineers of Japan.