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Development of Planar Antennas

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SUMMARY

As a typical planar antenna in Japan, a microstrip antenna and radial line slot antenna are chosen and some original technologies are introduced for them. About the microstrip antenna, the analyzing method is described first and the method based on the theory of microstrip planar circuit [1],[2] born in Japan is introduced [3]. According to the formulas derived by this method, the design procedure considering the bandwidth is established [4]. In addition, it is shown clearly that a microstrip antenna can produce the circular polarizations at two kinds of frequencies with a single feed [5]. Furthermore, two kinds of broadband techniques [6],[7] born in Japan are picked up. About other unique microstrip antennas, they may be introduced in a suitable section each time. As for the RLSA, the history on invention is briefly presented. The radiation mechanisms depending on the slot-set arrangement and the excitation mode are discussed. The slot-coupling analysis to simulate the excitation of a two-dimensional uniformly-excited slot array is explained. The simple design based on the operation with traveling-wave propagation is also described. The technical progress to keep high efficiency in a wide gain range for satellite-TV reception is reviewed. Extensions of the RLSAs to millimeter-wave bands and plasma etching systems are finally summarized.

key words: *planar antenna, circular polarization, microstrip antenna, radial line slot antenna, parallel plate waveguide*

1. Introduction

An planar antenna is the general term of an antenna thin enough to wavelength, so that many types of antennas exist in it and the method of a classification is also various [8],[9]. Although microstrip antenna and slot antenna [10] can be given as a typical example, ones of the type shown in Fig. 1 [8] are known as another examples [11]–[15]. In Japan, research to microstrip antenna has been performed most briskly and widely in the planar antenna above and many interesting results have so far been reported for it. Also the radial line slot antenna is not only one with the highest efficiency of the examples shown in Fig. 1 but also the unique antenna born in Japan. From the above reason, the development of both antennas in Japan is mainly introduced in this paper.

The microstrip antenna (MSA) makes efficient radiator and is widely used in antenna applications. The basic geometry consists of a conducting radiator patch printed onto a grounded substrate. The shape of the patch, in principle, is arbitrary, though rectangular and circular designs are common in practice. Excitation can be achieved either with a coaxial probe, microstrip line, or some types of aperture electromagnetic coupling. Until today, MSA has been theoretically analyzed by various methods. In this paper, the analysis is described briefly, based on the variational method applied to the arbitrarily shaped microstrip planar circuit with multiterminals [1],[2], because planar circuit is the novel concept born in Japan. Furthermore, the modal-expansion based on the variational method [3] is the most useful in understanding the basic performance of the cavity type of antennas. That is, this method enables not only derivation of the equivalent circuit but also establishment of the design procedures. The conventional design procedure took the resonant frequency only into account, while this procedure is able to take the bandwidth into consideration [4]. In addition, according this equivalent circuit, an improved circular polarization technique can be derived and it is shown that the MSA can generally produce the circular polarizations at two kinds of frequencies with a single feed [5]. An MSA is inherently narrow band antenna, however, so that many broadband techniques have been also reported in Japan. In this paper, two kinds of broadband technique are introduced. One is use of parasitic element [6] and another is a sequential arrangement [7],[16]. The former is useful for making the bandwidth of antenna element itself broad and the latter is useful for making that of circularly polarized array antenna broad. Other interesting ideas may be introduced in a suitable section each time.

The radial line slot antenna (RLSA) is a slot array excited by a cylindrical TEM wave in a radial waveguide. It was proposed by Goto in 1980 [59]. The structure consisting of only a slotted radial waveguide with a coaxial feed at one point is so simple that it is suitable for mass production with low manufacturing cost for commercial uses. The RLSA has been developed for receiving direct broadcasting from a satellite (DBS). A radial waveguide is a closed and thicker feedline to have smaller transmission loss in comparison with planar feedlines like a microstrip line. Therefore the RLSA has higher efficiency than other flat antennas using pla-

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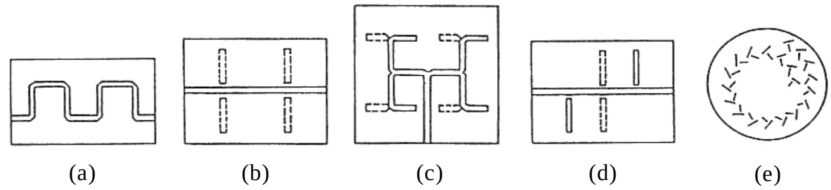

Fig. 1 Variety of typical planar antennas except microstrip antenna and slot antenna (a) Microstrip line antenna [11], (b) Printed slot antenna [12], (c) Printed dipole antenna [13], (d) Circular polarized printed antenna [14], (e) Radial line slot antenna [15].

nar feedlines. It keeps high efficiency more than 80% in a wide gain range from 25dBi to 37dBi at 12GHz band. It is more advantageous in high-frequencies using millimeter waves due to the low transmission loss. It is also suitable for high-power transmission such as plasma etching process because of its waveguide structure. A radial line is one of multimode waveguides, so that only the dominant TEM wave has to propagate in it and the excitation of higher modes due to the slot coupling has to be suppressed. Traveling-wave propagation with reflection-canceling slot sets is also required for the stable operation. In the latter part of this paper, the history on invention of the RLSA by introducing the simple structure is briefly presented at first. The radiation mechanisms depending on the slot-set arrangement and the excitation mode are discussed. The slot-coupling analysis using a rectangular waveguide with periodic boundary conditions on the narrow walls to simulate the excitation of a two-dimensional uniformly-excited slot array is explained. The simple design based on the operation with traveling-wave propagation is also described. The technical progress to keep high efficiency in a wide gain range for DBS reception is reviewed. Extensions of the RLSAs to millimeter-wave bands and plasma etching are finally summarized.

2. Microstrip Antenna (MSA)

2.1 Analysis for MSA

This section describes the analysis for MSA. Beginning about 1974, MSA has been theoretically analyzed using Transmission-line model [17], Cavity model [18], Cavity model considering the wall admittance [19], Modal expansion method [20], Method of moments [21], Theory of planar circuit with Method of moments [22], Unimoment Monte Carlo method [23], Finite-element method [24], Direct form of network analysis [25], Wire-grid method [26], Segmentation and desegmentation technique [27], Spectral Domain analysis [28]–[30], Modified Spectral Domain Moment Method [31], Bergeron's method [32], Boundary element method [33], Finite-Difference Time-Domain(FDTD) method[34]. FDTD method is one of developing the Maxwell' equations to difference equation and reducing directly it in a time- and space-domain. In Japan, FDTD method has also high popularity and many attractive simulation results have been reported using it [35]–[37]. Although FDTD method and the integral equation method, represented by the method of moments, is perhaps the most general, it requires considerable computational effort and yields few physical insights when designing. As opposed to this, the modal-expansion method using cavity model offers both simplicity and physical insight. The cavity model treats the region between a patch radiator and ground plane, as a cavity bounded by the electric walls and a magnetic wall along the periphery of the patch. Once the field distribution in the cavity is known, Huygens' principle may be applied to the magnetic wall. Hence, the radiation field is readily evaluated. According to this method, not only the design procedure considering the bandwidth can be established but also the equivalent circuit can be derived for an arbitrary shaped MSA with multi-terminals. This equivalent circuit is very useful to understand how MSA excites a circular polarization without any external Hybrid. So, the modal-expansion method based on the variational method is explained briefly below, because it is required of the explanation after following section.

The geometry of the analytical model and the coordinate system employed here are illustrated in Fig. 2, where C denotes the boundary line of the patch radiator, S is the physical area surrounded by the boundary line C, and $\hat{\boldsymbol{n}}$ is a unit vector, outer normal to the boundary. In many practical applications, the substrate is electrically thin, so the TM modes are superior in the cavity region. Assuming an $e^{j\omega t}$ time variation, the electric fields associated with the z-directed current source located at (x_m, y_m) is written as

$$E_z(x,y) = \frac{\sqrt{2S}}{t}\sum_{l=1}^{N}\frac{\varphi^{(l)}(x,y)}{j\omega C + \frac{1}{j\omega L^{(l)}} + g^{(l)}} \sum_{m=1}^{M} M_m^{(l)^*} I_m \quad \text{(in the cavity region).} \quad (1)$$

In the above equation, $\varphi^{(l)}(x,y)$ means the l-th eigen-

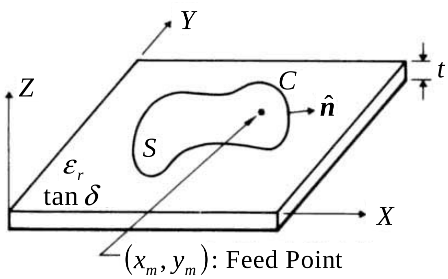

Fig. 2 Image of arbitrarily shaped MSA.

function which can be derived by variational method, and I_m is the effective value of current flowing into the m th terminal and the asterisk represents complex conjugation. The remaining parameters are defined in [3]. When the MSA has many terminals, the coupling between the terminals must be considered. In this case, the terminal currents must be derived from the following matrix equation [3]:

$$[\mathbf{I}] = [\mathbf{Z}']^{-1}\left[\mathbf{V}^{(in)}\right] \tag{2}$$

where

$$Z'_{n,m} = \begin{cases} Z_{n,m} + Z_{0m} & n = m \\ Z_{n,m} & n \neq m \end{cases}, \tag{3}$$

$$Z_{n,m} = \sum_{l=1}^{N} \frac{M_n^{(l)} M_m^{(l)^*}}{j\omega C + \frac{1}{j\omega L^{(l)}} + g^{(l)}}, \tag{4}$$

with Z_{0m} being the characteristic impedance of the m th terminal. According to Huygens' principle, the total radiation field can be calculated as a superposition of the contributions from all the modes considered and is expressed by

$$\boldsymbol{E} = \frac{\sqrt{2}S}{t}\sum_{l=1}^{N}\frac{\boldsymbol{E}_0^{(l)}}{j\omega C + \frac{1}{j\omega L^{(l)}} + g^{(l)}} \sum_{m=1}^{M}\varphi^{(l)^*}(x_m, y_m) I_m. \tag{5}$$

Here, it is important to know how an equivalent circuit is expressed for an arbitrary shaped MSA with multi-terminals. Such equivalent circuit can be easily obtained from Eq. (4) and expressed by Fig. 3.

2.2 Design Procedure with Consideration to Bandwidth

It is found from the previous analysis that the bandwidth of MSA is depend on the resonant frequency, the patch shape, and the dielectric constant and thickness of substrate. Now if $\sqrt{S}/t$ is introduced as a convenient parameter, the product of resonant frequency, denoted by f_0 and the square root of the physical patch area can be expressed as functions of $\sqrt{S}/t$ and the dielectric constant as follows [4]:

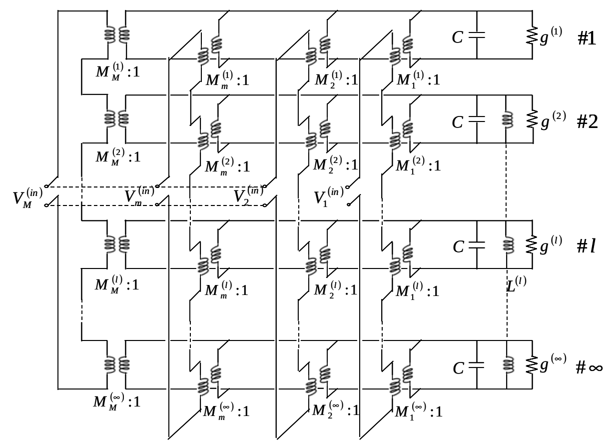

Fig. 3 Equivalent circuit for Arbitrarily shaped MSA with multiterminals.

$$\sqrt{S}f_0 \approx F_2\left(\frac{\sqrt{S}}{t}, \varepsilon_r\right). \tag{6}$$

Furthermore, if the radiation conductance component can be regarded as a dominant factor compared with other conductance components, the unloaded Q, denoted by Q_0, can also be approximated by the following equation as functions of $\sqrt{S}/t$ and ε_r [4]:

$$Q_0 \approx F_1\left(\frac{\sqrt{S}}{t}, \varepsilon_r\right). \tag{7}$$

These relationships can be arranged as a design charts by using computer simulation results for every given patch shape. For example, when the patch is circular, two design charts shown in Fig. 4(a) for $\sqrt{S}f_0$ and Fig. 4(b) for Q_0 are obtained [4]. In these figures, the dots indicate the results of experiments.

On the other hand, if the attentions are paid to only the dominant modes and other higher order modes are ignored, the following relationship is known to exist between Q_0 and relative bandwidth, denoted by B_r, in which the input VSWR is less than ρ [4]:

$$Q_0 B_r = \sqrt{(\beta\rho - 1)\left(1 - \frac{\beta}{\rho}\right)}, \tag{8}$$

where β is the coupling coefficient, defined by

$$\beta = \frac{G_0}{G'}, \tag{9}$$

with G_0 being the conductance of the transmission line and G' the conductance of the patch antenna. Note that Eq. (8) gives the maximum value and it can be obtained by

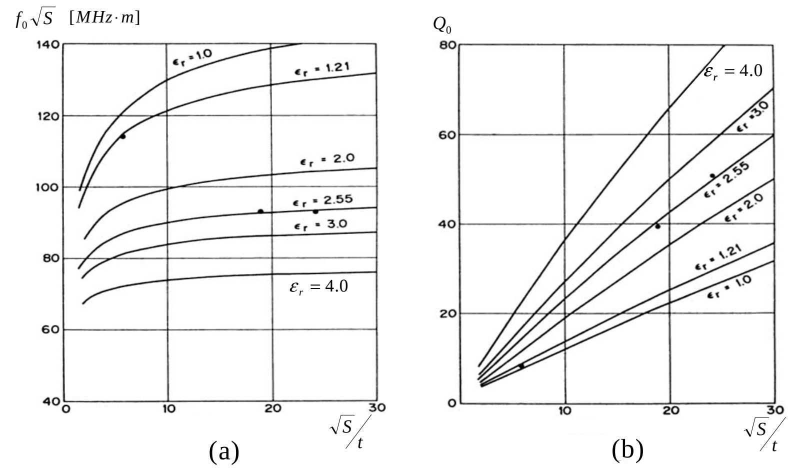

Fig. 4 Design charts for circular MSA against $\sqrt{S}/t$ with ε_r as a parameter, (a) design chart to determine resonant frequency, (b) design chart to determine unloaded Q value.

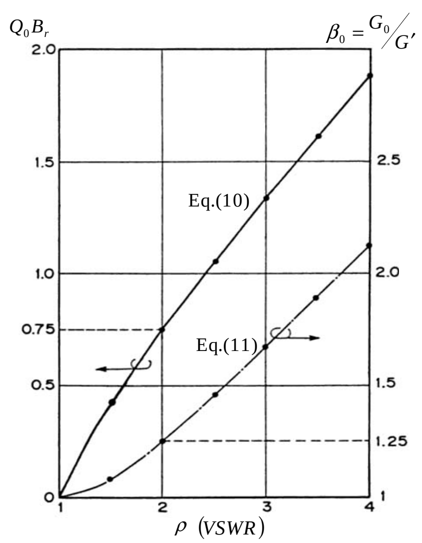

Fig. 5 Design chart to determine maximum value of Q_0, B_r, and coupling coefficient Equivalent (β_0) against desired VSWR.

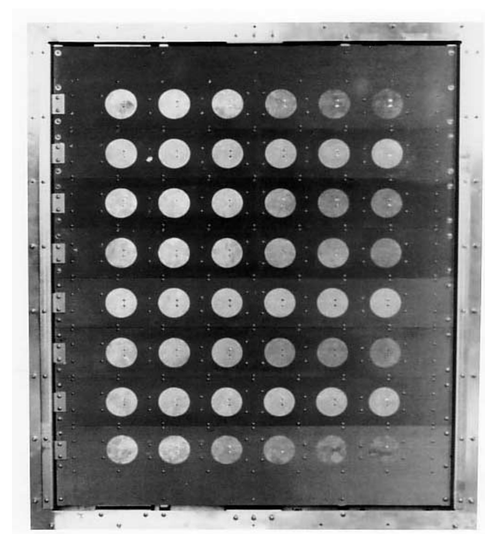

Fig. 6 Circularly polarized shaped beam array antenna with 8×6 dual feed elements.

$$Q_0 B_r = \frac{\rho^2 - 1}{2\rho}, \tag{10}$$

when the following condition is satisfied for the coupling coefficient:

$$\beta = \beta_0 = \frac{\rho^2 + 1}{2\rho}. \tag{11}$$

The above relationships are illustrated by the solid and chain-dotted lines, respectively, as a function of input VSWR ρ in Fig. 5 [4].

So, it is understand from the discussion above that a MSA can be designed using three kinds of figures as design charts. That is, from the solid line in Fig. 5, one can obtain Q_0 required for the antenna when the desired bandwidth and VSWR are specified. Also, from the chain-dotted line, one can determine the characteristic impedance necessary for the feed line when the position of the feed point is given. Next, the necessary value of $\sqrt{S}/t$ can be determined from Fig. 4(b) for the dielectric constant to be used. Finally, the physical patch area S can be derived by substituting the resonant frequency into the results found from Fig. 4(a). Once S obtained, the patch dimension and the substrate thickness can be determined immediately. Although only one example of a circular MSA is represented here, the design approach described above is applicable to any MSA with arbitrary shape.

2.3 Excitation of Circular Polarization

Many applications have been proposed for MSA. One of the most interesting concerns its use for transmitting and receiving circularly polarized waves (CP waves). A circularly polarized MSA is classified as being either

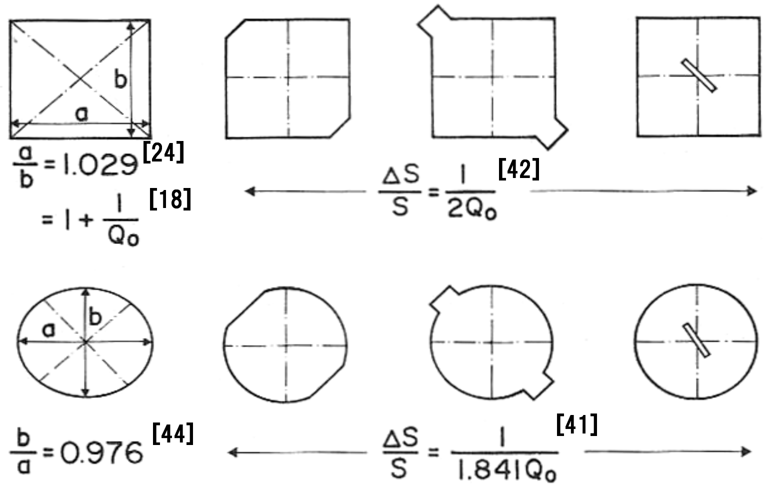

Fig. 7 Typical examples of singly fed circularly polarized MSA.

dual or single feed type, depending on the number of feed points necessary to excite the CP waves. The dual type is very popular, because it can conveniently radiate the CP waves from standard circular or square MSA using a 90° hybrid. Figure 6 shows the first shaped beam array antenna using such dual feed elements in Japan [38]. In this case, however, the influence on axial ratio of the higher-order modes may become a serious problem, together with an increase in bandwidth. Such degraded axial ratios can be improved by adopting a driving technique that will not excite the TM_{210} modes, which is the most troublesome unwanted mode [39].

In general, CP waves can always be produced by exciting orthogonally polarized fields of equal amplitude but 90° out of phase. Such radiation fields can also be excited from a singly fed MSA by modifying the patch shape to a suitable one. So the singly fed circularly polarized MSA requires no external polarizer. Typical examples of such MSA are shown in Fig. 7 [18],[24],[40]–[44]. These MSAs radiate CP waves when fed at a point on the chain-dotted lines in Fig. 7. In this figure, the conditions imposed on the patch geometry to obtain the CP waves are also given, where ΔS is the variation in area from the original patch geometry and Q_0 is the unloaded Q value for the original patch antenna. In these examples, the mechanism by which the CP waves are excited can be broadly understanding by separating the degenerative modes in the cavity. But MSA can also radiate CP waves with single feed by inserting any short-circuited or impedance-loaded post between patch radiator and ground plane [45]. Although a single feed type is very useful, it is not easy to analyze such nonstandard geometry, because in many cases its eigenvalues and eigenfunction cannot be solved by ordinary separation of variables. According to the previous

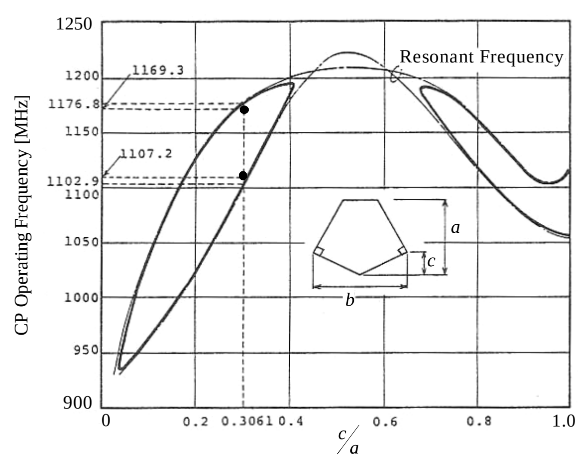

Fig. 8 CP operating frequency verses aspect ratio (c/a) for a pentagonal microstrip antenna with $b/a = 1.0603, a = 100mm, t = 3.2mm, \varepsilon_r = 2.55, tan\delta = 0.0018$, solid line: theoretical, dot dash line: experimental.

formulation based on modal-expansion, however, the radiation field is given by Eq. (5), even when the antenna has arbitrary shape but also multiterminals. For the MSA to radiate the CP waves with a single feed, not only must a pair of orthogonally polarized modes be excited in the resonator, but the ratio of both radiation fields associated with them must be equal to $\pm j$. By solving these relation with respect to frequency, the CP operating frequencies, at which the ideal CP waves are excited, can be derived together with the conditions concerning the corresponding feed point [5]. As a result, two kinds of CP operating frequencies can be obtained and all of the feed location loci, consisting of the optimum feed point (x_0, y_0) are determined numerically. Figure 8 shows as an example, the relationship between the CP operating frequency and aspect ratio

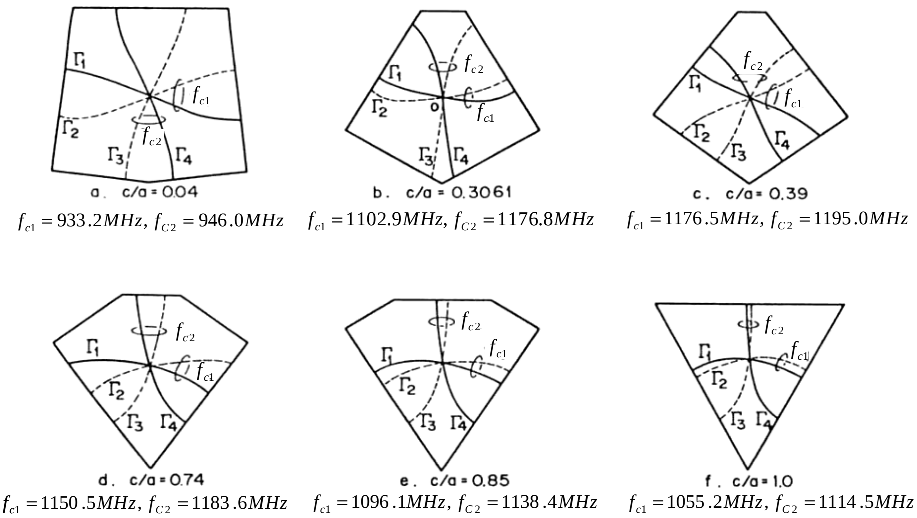

Fig. 9 Variation of feed location loci with respect to aspect ratio (c/a) for a pentagonal microstrip antenna with $b/a = 1.0603, a = 100mm, t = 3.2mm, \varepsilon_r = 2.55, tan\delta = 0.0018$, dot line: RHCP, solid line: LHCP.

c/a of pentagonal shaped MSA [5]. In this figure, the solid lines represent the calculated relationships and the two dots are measured results for the pentagon proposed by Weinschel [46]. The two chain-dotted lines denote the resonant frequencies of the two orthogonal modes contributing to circular polarization. Figure 9 shows the numerical feed location loci for several aspect ratios [16]. In these figures, Γ_1 and Γ_2 show the loci when the corresponding CP operating frequency is f_{c_1}, and Γ_3 and Γ_4 show the loci when it is f_{c_2}. It is found from these results that even triangular MSA can radiate the CP waves at two kinds of frequencies [47].

2.4 Wideband Technique

Microstrip antenna has found various applications because of their low profile and conformal nature. Despite their many attractive features, however, their narrowband characteristics have remained a major disadvantage. The bandwidth that can be obtained with the standard microstrip antenna is typically on the order of 1 to 2 % for a VSWR of less than 2. For the sake of practical use, wider bandwidth would be better. So, some useful techniques are already proposed for increasing the bandwidth. They are categorized into the following methods: (1) using a substrate with low unloaded Q [48],[49]; (2) Employing a wideband impedance-matching network [50]; (3) Adding some parasitic element [6],[51],[52]. In the above, the most interesting technique is (3). The typical antenna based on the technique of (3) is the stacked MSA, which consists of feeding and parasitic patches, where the upper patch is parasitic element. Each patch is fabricated on electrically thin substrate and separated by a region of air or foam with $\varepsilon_r \approx 1$. When the air region is small, two resonances are expected. In this case, broadband characteristics can be obtained by designing the diameter of the parasitic element to be larger than the feeding element. Although it is said that this type of MSA appeared to be first discussed by Sabban [53], the same report was made also in Japan [6],[51]. The spectral domain method is suitable for the analysis of this type of MSA and when the shape of patch is circular, the analysis result was reported using the Hankel transformation [52].

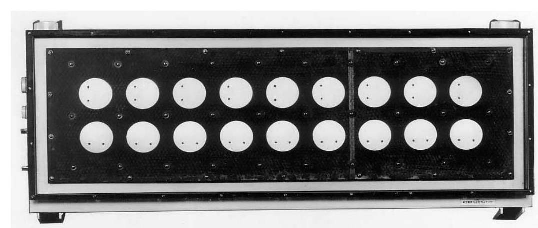

Fig. 10 Circularly polarized wideband phased array antenna employing paired-element configuration, which consists of circularly polarized elements with dual feed.

By the way, if a circularly polarized radiator is used as the element in an array antenna, it is possible to make the bandwidth of such array antenna broad by canceling out the reverse-polarized component. A sequential rotation arrangement is the most useful technique [7] and the simplest example is a paired arrangement, as shown in Fig. 10 [54]. This is the phased array antenna first in Japan, which used MSA. Another example adopting the paired arrangement technique is shown in Fig. 11 [16],[55],[56], where each element is a singly fed circularly polarized MSA with small notches.

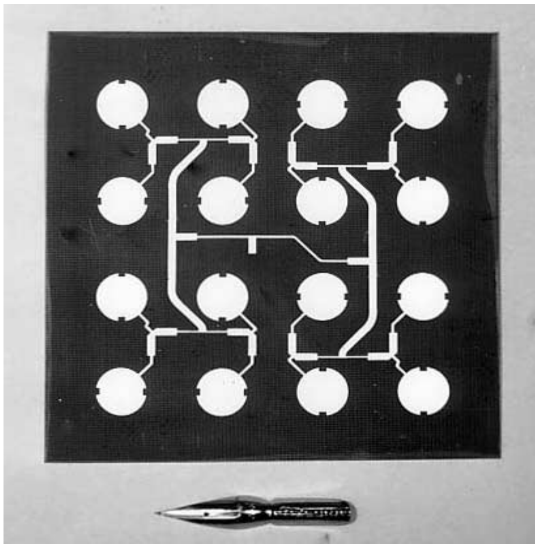

Fig. 11 Circularly polarized array antenna employing paired-element configuration, which consists of singly fed circularly polarized elements.

These paired elements are rotated orthogonally on the coplanar plane according to the following sequential rule [7]:

$$\phi_n = (n-1)\frac{\pi}{N} \quad (n = 1, 2, \cdots\cdots, N) \tag{12}$$

where ϕ_n denotes the mechanical rotation angle to be given to the n-th element. N is the number of elements contributing to the sequential arrangement, but not the total element number in the array. That is, $N = 2$ in the case of array antenna consisting of paired elements, as shown in Figs. 10 and 11. In such a sequential arrangement, of course, each element must be excited to compensate for the additional phase rotation suffered due to the mechanical rotation. In practice, each element in the sequential arrangement is excited through a corresponding differential path length in the feed line so that reflected waves from each element have a different phase shift. As a result, the reflected waves from all elements cancel each other at the input terminal. A sequential array antenna that satisfies the conditions for perfect circular polarization naturally has no reflections at the input terminal [7].

As other attractive development to wideband techniques, that to dual frequency operation can be picked up. Although it cannot be described in detail here because shortage of space, the unique examples of dual frequency operation are shown in reference [57],[58]. Since a MSA has inherently high quality factor, it can be said that dual frequency operation of MSA is suitable for the use as a self-diplexing antenna.

3. Radial Line Slot Antenna

3.1 Invention of the Radial Line Slot Antenna

The RLSA was proposed by Goto in 1980 [59]. Around 1980, planar antennas using a microstrip line or a triplate lines were mainly investigated for DBS reception. These planar antennas have a disadvantage because of large transmission loss. The transmission loss of a microstrip line is 4–6dB/m and that of a triplate line is 2.7–5.6dB/m at 12 GHz, which are much higher than the loss of a hollow rectangular waveguide of 0.2dB/m [60]. A rectangular waveguide array has conventionally another disadvantage in terms of complicated structures, which was used mainly for special uses such as in military or satellite communications. The RLSA with a single-layer structure as shown in Fig. 12 was invented as a circularly-polarized slot-pair array excited by a coaxial line in a radial waveguide with small transmission loss and a simple structure. It does not have a complicated feeding structure such as a microstrip patch array where each element is fed by a microstrip line. The RLSA uses non-resonant slots to control the coupling by their length to get desired aperture illumination. Rectangular waveguide arrays using non-resonant slots had rarely been in practical use before because resonant slots were typically used and the offset or the tilted installation was usually adopted for the power control. Radiation with circular polarization in the RLSA was confirmed in the first model consisting of eight pairs on a hollow radial waveguide [59]. However the antenna efficiency had been low before finding that a conical grating lobe in a circular array gave serious gain degradation [61], which is different from pencil grating lobes in a rectangular array. A slow-wave circuit of corrugation or dielectric was begun to be installed in radial waveguide to suppress the grating lobe [62]. A double-layer structure as shown in Fig. 12 was also introduced to compensate the decay of the internal field by the slot radiation with accumulation of the inward incident wave to enhance the uniformity [63]. In the double-layered RLSA, slots with constant length could be applied to get uniform excitation. Desired stable operation was confirmed at first by experiments that 35.0dBi gain with 57% efficiency was obtained in a 12GHz model with a 60cm-diameter aperture [64].

3.2 Radiation Mechanisms

Figure 12 shows the structure of RLSAs. Two conductor disks form a single-layer radial waveguide. Power is fed at the center and is transferred into an outward cylindrical TEM traveling wave. It is gradually radiated from slot sets arranged on the top plate. The shape of the main beam depends on a mode propagating in the waveguide and arrangement of the slot sets.

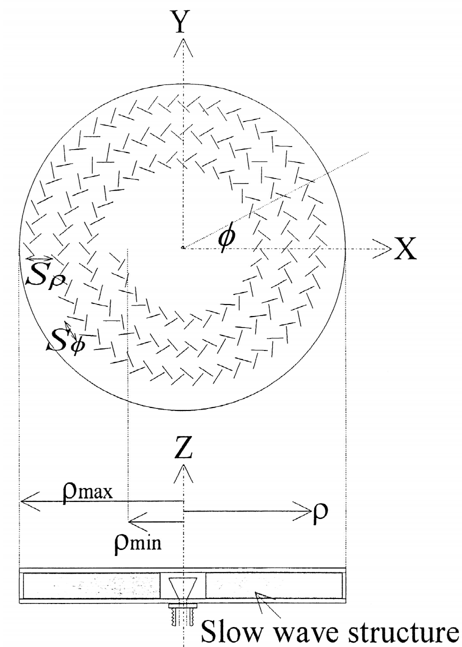

Fig. 12 Structure of RLSAs.

The polarization is determined by the configuration of the slot set. Before the RLSA, another slot array on a radial waveguide was proposed as an annular slot planar antenna by Goebels and Kelly [65]. However, the RLSA is different from their antenna in terms of the excitation mode for slots. Their antennas require the longitudinal (radial) component (H_ρ) of magnetic field as well as the transverse (circumferential) one (H_ϕ) for the slot excitation, however the RLSA do not need the longitudinal one at all.

The original RLSA [59] had spiral arrangement of slot pairs excited by a coaxial TEM traveling mode to get a main beam with circular polarization in the boresight as shown in Fig. 12. The slot set on the radial waveguide consists of two slots as a unit radiator of circular polarization. The two slots are perpendicular to each other and are excited with equal amplitude and phase difference of 90 degrees. They are separated by a quarter of the guided wavelength in the radial direction with a coupling angle of ±45 degrees. The slot pairs are arranged spirally with a pitch S_ϕ of about one guided wavelength to have the same instantaneous orientation of the circular polarization vector, which can be controlled by the phase. The pair spacing S_ϕ in the circumferential direction should be small not to excite higher modes. The direction of incident TEM wave propagation for a slot pair in Fig. 12 determines the sense of circular polarization. The inward propagation gives the right-handed circular polarization while the outward propagation gives the left-handed one. The quarter-wavelength slot spacing in a pair for circular polarization simultaneously satisfies the reflection cancellation because the reflected waves from the two slots

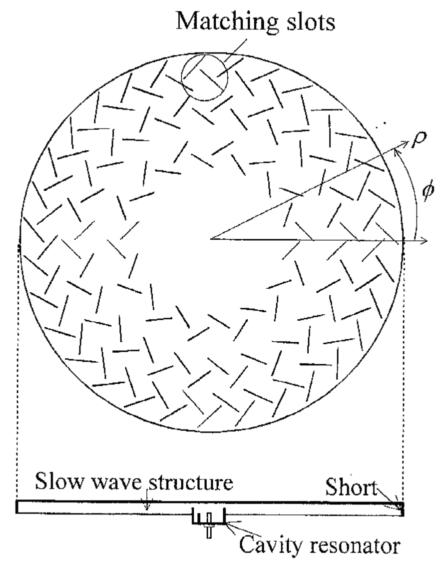

Fig. 13 Concentric-array RLSA.

have phase difference of 180 degrees.

Another technique to get a boresight beam with the circular polarization is the combination of concentric arrangement of the slot pairs and excitation by the rotating mode, which was named a concentric-array RLSA as shown in Fig. 13. The rotating mode is defined to be uniform in amplitude and to have a linear taper of 2π in phase in the ρ-direction. The same instantaneous orientation of circular polarization vectors in all the slot pairs is obtained by the incident wave in this case while it was realized by the pair arrangement in the previous case. There are various ways to excite the rotating mode in a radial waveguide. The feeding structure is installed at the center of a radial waveguide. A cum-shaped dielectric substrate to have a different path length in the ρ-direction can be adopted [66]. A magnetic or electric-wall circular cavity resonator with perturbation elements to separate the degenerated modes is also applicable [67]–[69]. The use of a ring slot with perturbation elements or a crossed slot cut on the bottom plate is another method [70],[71].

A conical beam can be achieved in the concentric pair arrangement with excitation of the coaxial mode [72]. The instantaneous orientation of circular polarization vector has the same angle to the radial direction in all the slot pairs and the total radiation in the boresight is cancelled out. When all the pairs are excited in phase on a planar radial waveguide, the conical beam can not cover a wide range in the elevation plane. In order to do so, the pairs should be arrayed on a spherical-shape radial waveguide [73] or the excitation should be controlled both in amplitude and phase on a planar radial waveguide [74].

Linear polarization can be obtained by different configuration of a slot pair as shown in Fig. 14 [75]. In

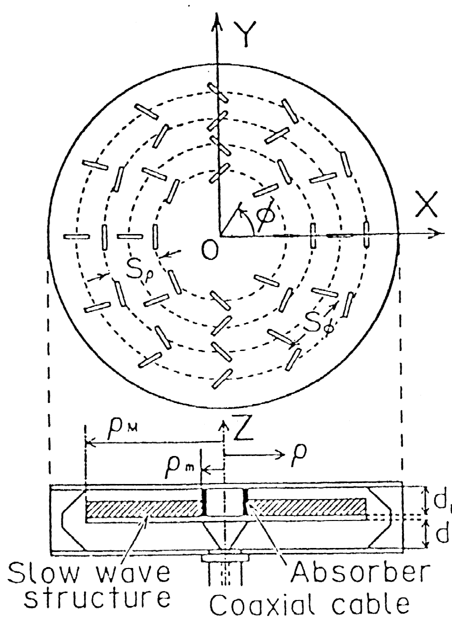

Fig. 14 Linearly-polarized RLSA.

a pair, two slots have the same length, are spaced by a half of the guided wavelength and make a right angle to each other, assuming that the slot excitation is proportional to $\sin\theta$ where θ is an angle taking from the radial direction. A pair at ϕ in Fig. 14 should be rotated by an angle of $\phi/2$ for the x-directed linear polarization. Design of slot arrays with various radial directions should be changed for linear polarization, which is different from that for circular polarization. In this original configuration of linear polarization, reflections of the two slots in a pair are summed up in phase because their path-length difference is one wavelength. Reflection-canceling configuration [76] of a slot set spaced by a quarter of guided wavelength or beam-tilting [77] spaced by other than one guided wavelength between adjacent slot pairs was proposed to suppress the overall reflection of the antenna [78].

Monopulse operation [79],[80] or dual-beam radiation [81] were realized by feed from multiple input ports.

3.3 Analysis and Design

3.3.1 Analysis of Slot Coupling using a Rectangular Waveguide with Periodic Boundary Walls

In a design of a RLSA, control of the slot coupling is a key technology. Previously the slot coupling was estimated and controlled on the basis of measured data; only the slot length was utilized [82]. Many parameters, such as the arrangement and the configuration of slot sets, the waveguide height and the permittivity of dielectric filling it, should be optimized by an analysis for an advanced design of the antenna. More than a thousand slot pairs are cut spirally in a circular plate

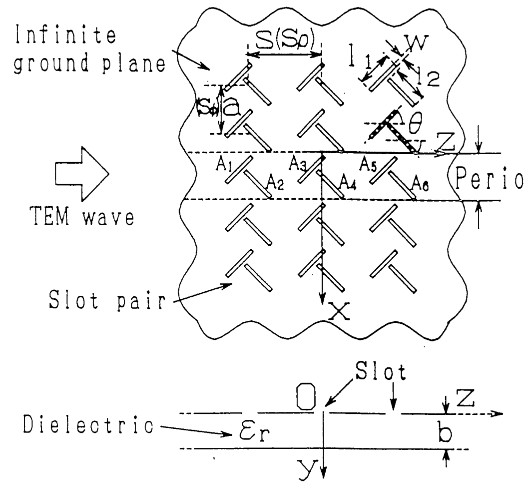

Fig. 15 Analysis model.

of a 12GHz RLSA with 60cm diameter; analyzing the actual structure including all the slots is difficult.

Derivation of an analysis model to simulate the normal operation of a RLSA with rotational symmetry by a TEM wave excitation is important. A simple linear model for a uniformly excited two-dimensional slot array on a parallel plate waveguide was proposed [83]. Figure 15 shows the analysis model. Two infinite conductor plates comprise a parallel plate waveguide. Slot pairs are two-dimensionally as arrayed in a lattice. The spacing between adjacent pairs is a $(= S_\phi)$ in the x direction, and s $(= S_\rho)$ in the z direction. The waveguide is filled with a dielectric with the relative permittivity ε_r. A plane TEM wave is incident in the $+z$ direction. All the slot pairs in each row along the x axis are assumed to be excited with equal amplitude. The field has a period of a in the x direction. Once the excitation coefficients of the slot pairs are analyzed by the method of moments [83], the coupling factor a and the slow wave factor ζ [84] are determined. To express the periodicity of the internal field in the x direction, the parallel plate waveguide is replaced by a rectangular waveguide with periodic boundaries in the narrow walls. The infinite array in the x direction in the external region is also truncated. The truncation size is determined so that the coupling and slow wave factors may converge. This model can be applied to other two-dimensional slot arrays on a rectangular parallel plate waveguide [85],[86].

The effects of various parameters on the slot coupling can be discussed by using this model. The frequency characteristics of the coupling and the slow wave factors are predicted and compared with experiments. Figure 16 shows the slow wave factor for different permittivity of the dielectric in the waveguide, for example. The slow wave factors in measurements are predicted accurately by the analysis as well as the coupling factor. The analysis simulates well a rapid change at the slot

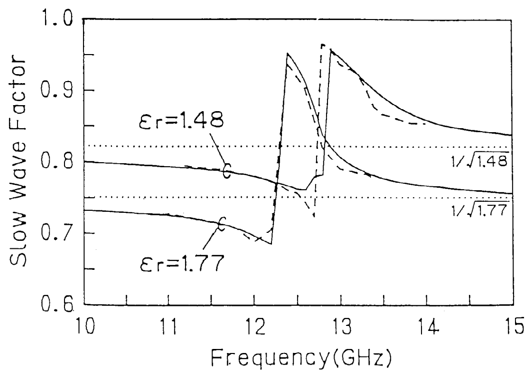

Fig. 16 Slow wave factor for different permittivity of dielectric (Solid line: calculated, dotted line: measured).

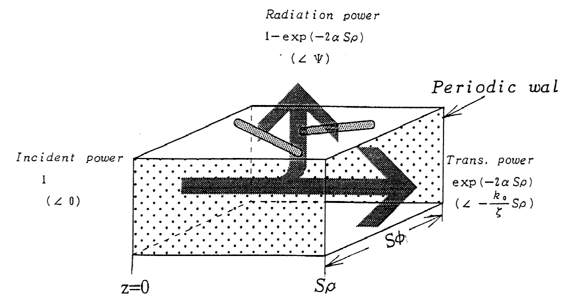

Fig. 17 Coupling of a unit slot pair.

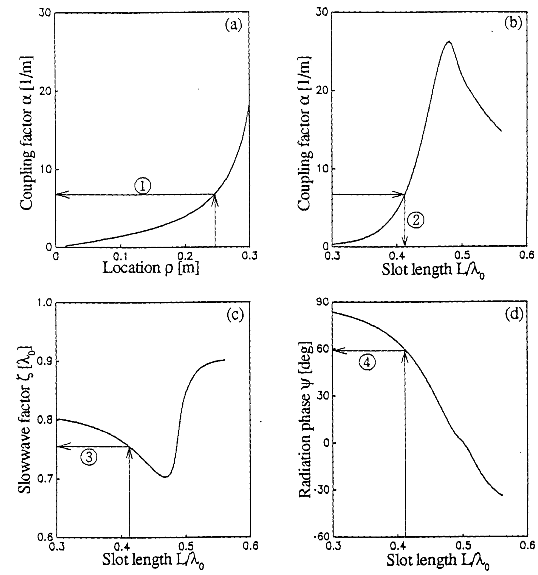

Fig. 18 Coupling parameters.

resonance. As the permittivity becomes large, the coupling shifts to the lower frequency. The slow wave factor is greatly perturbed by the slot coupling, but it approaches the unperturbed value $(1/\sqrt{\varepsilon_r})$ at frequencies far from the resonance.

3.3.2 Design for Uniform Aperture Illumination in Single-layer RLSAs

The design is based on the traveling-wave excitation of reflection-canceling slot sets. The derivation of a desired amplitude distribution is according to the power conservation. The parameters of slot pairs are determined from the inner portion to the outer one of an aperture in turn to get stronger coupling gradually. The amplitude of each slot pair is controlled by the length. Almost of the slots are non-resonant. Nonresonant slots give phase shift in coupling. Phase control is also required to get uniform phase distribution. Phase adjustment between adjacent slot pairs can easily be obtained by the spacing because the slot pairs are excited by a traveling wave. A straightforward theoretical method of the slot design for uniform aperture illumination was proposed [87].

Before the design, a unit slot pair is analyzed by the method described in the previous sub-section. Figure 17 depicts the perturbation for a unit slot pair in a period S_ρ in terms of the coupling and the slow wave factors. An incident unit power is split into the slot radiation $1-\exp(-2\alpha S_\rho)$ and the transmission $\exp(-2\alpha S_\rho)$. The phase of the radiated field is ψ while that of the transmission field is $-k_0 S_\rho/\zeta$. The coupling parameters, such as α, ζ and ψ, are summarized in Fig. 18 [87], as functions of the slot length L normalized by the free-space wavelength λ_0.

The design procedure is presented. First, the coupling factor $\alpha(\rho)$ required for desired aperture illumination is derived as a function of ρ as shown in Fig. 18(a). As unique difficulty of the RLSA, slot coupling can not be increased arbitrarily so as not to disturb the rotational symmetry of the field excessively. Secondly, the slot length and the position are determined over the aperture. This is a kind of non-uniform array designs. Assuming the variation of the slot length and spacing is small, we determine them by making use of the coupling parameters. For some point at ρ, the slot length L is determined from Fig. 18(b) to realize $\alpha(\rho)$. Then the slow wave factor $\zeta(\rho)$ and the phase $\psi(\rho)$ of the radiated field are obtained in Fig. 18(c) and (d). The pair radial spacing S_ϕ is set to be the local guide wavelength $\lambda_g(=\zeta(\rho)\lambda_0)$. This procedure is repeated to cover whole the aperture. The slot design ends after shifting all the pairs by the amount of $\psi(\rho)/k_0$ to assure the aperture phase uniformly. All the design procedures are carried out theoretically. This design can be applied to other distributions of coupling factor for non-uniform illumination, such as a Taylor distribution to suppress the sidelobes [88],[89].

3.4 Applications

3.4.1 Technical Progress for DBS Reception

Figure 19 shows the measured efficiency of various types

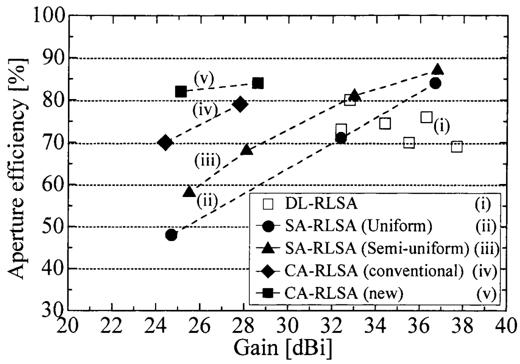

Fig. 19 Measured efficiency as functions of the gain at 12GHz band.

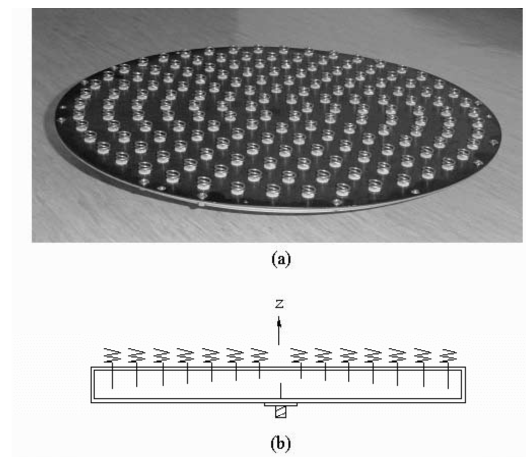

Fig. 20 Low-profile helical array antenna fed from a radial waveguide (a) Picture, (b) Structure.

of RLSAs as functions of the gain at 12GHz band [90]. The results of the double-layered RLSA are marked by white squares in the figure. In the double-layered RLSA, an E-bend to transfer a cylindrical TEM wave from the lower layer to the upper one as well as a coaxial-to-radial line transformer was designed by the two-dimensional finite element method including the rotational symmetry and the overall reflection was suppressed below −15dB [91]. The slot length was designed by introducing an impedance boundary experimentally at the place of the slot aperture to express the radiation [82] and 75% efficiency (gain: 36.3dBi) was achieved in a 60cm-diameter antenna at 12GHz band [92].

Uniform aperture illumination in single-layered RLSAs was designed by the method described in the previous section. The results are denoted by black circles in Fig. 19. In single-layered RLSA, no absorbers are used at the end of the radial waveguide and the residual power is reflected by a conducting wall. 84% efficiency was obtained at a 60cm-aperture antenna [93]. When the aperture diameter decreases, the efficiency is degraded because of notable termination loss. One way to improve the efficiency is to install a matching spiral slit at the end of the radial waveguide to radiate all the residual power [94]. Another way is to apply for semi-uniform illumination where the field is uniform inside and tapered outside of the inflection point on an aperture to reducing the termination loss and to maximize the efficiency [95]. The efficiency enhancement of 9% was realized in a 40cm-aperture antenna.

The spiral arrangement of slot pairs with the coaxial-mode feed on a small aperture degrades the efficiency because the innermost pair with strong coupling deviates the rotational symmetry of the internal field in the radial waveguide. The concentric arrangement with the rotating-mode feed [68] does not disturb the rotational symmetry even in the small aperture [96]. By installing matching slot pairs to radiating the residual power with circular polarization at the end, the efficiency of 79% and 70% were obtained in a 23cm and 16cm-diameter antenna, respectively, as presented by black diamond marks in Fig. 19 [97]. These were about 10% higher than that of spiral arrangement with the same size. Finally, the parameter modification of slot pairs including the mutual coupling gave the efficiency enhancement up to 84% and 82% in a 24cm and 16cm-diameter antenna, respectively, as shown by black squares in Fig. 19 [90].

Dual-polarization operation is achieved by propagating an inward and outward traveling wave simultaneously in a radial waveguide [98]. The combination of the double layer in the inner part and the single layer in the outer part by installing a power divider at the middle of a radial waveguide enhanced the gain bandwidth because the long line effect is reduced [99]. A RLSA without filling dielectric in a radial waveguide is attractive in terms of cost reduction. Its operation was confirmed by introducing large slow-wave effect of strongly-coupled non-resonant slots in a thin hollow waveguide [100].

A helical antenna [101] and a microstrip antenna [102] are also used as an element instead of a slot pair as shown in Fig. 20 and Fig. 21, respectively. Both types of the elements are designed to radiate a circularly polarized wave in the boresight. Each element in an array is placed on a radial waveguide and is excited by a probe inserted into it. The elements are arrayed annularly. The amplitude of the excitation is determined by the probe length. The phase of the circularly-polarized wave is controlled by the rotation angle of each element without changing the spacing in the radial direction between adjacent elements. The microstrip array antenna on a radial waveguide is applied to practical use of mobile DBS reception with mechanical steering only in the azimuth plane [103].

3.4.2 Plasma Etching Systems

RLSAs have many unique features suitable for high-density plasma sources such as low profile and high power handling capability and aperture field unifor-

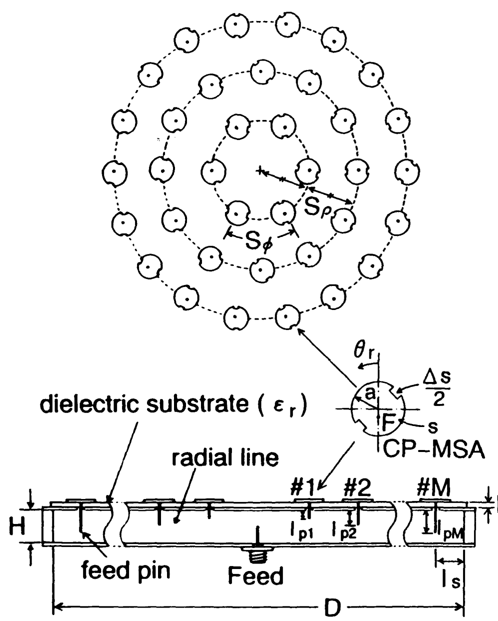

Fig. 21 Microstrip array antennas fed by radial line.

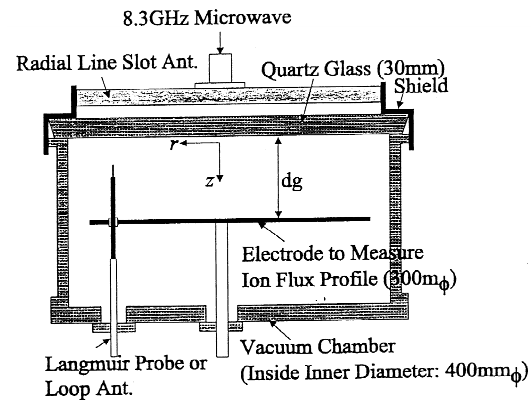

Fig. 22 RLSA in a cylindrical vacuum chamber.

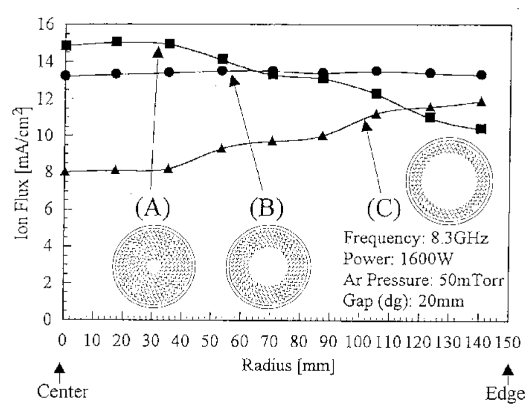

Fig. 23 Distribution of ion flux.

mity. It was applied to the electron cyclotron resonance (ECR) plasma system [104]. However this system requires a uniform and strong magnetic field supported by a coil outside the vacuum chamber; the system is bulky and complex. As another difficulty, no diffusion of electrons transverse to the magnetic field occurs in ECR and the uniformity of the plasma on a wafer directly reflects the electromagnetic distribution on the antenna aperture. A RLSA was adopted to another plasma process system which does not need the external magnetic field. Figure 22 shows the RLSA in a cylindrical vacuum chamber used in the plasma process [105]. The system is simple and compact. Microwave energy is radiated into the vacuum chamber through a quartz glass and is transferred to electrons. High electron density of more than $1 \times 10^{12}\mathrm{cm}^{-3}$ is measured.

The electron density as well as the magnetic field decays rapidly near the quartz glass. It is suggested that plasma is excited by the surface of the glass plate and diffuses towards to the electrode. The electron density below the quartz glass surface is far beyond the cutoff density. Microwaves incident on the plasma are completely dissipated by high-density plasma near the quartz plate. Figure 23 shows the ion flux distribution for three kinds of aperture blocking. The ion flux can be controlled by the blocking or the slot design. Good uniformity of $\pm 2\%$ is realized for the blocking size (B), where the inner three rounds of the slot pairs are omitted. A RLSA was introduced also to the surface-wave-coupled plasma generation [106], where parallel slots with linear polarization are arrayed annularly with a narrow spacing of 0.1 wavelengths in the radial direction to generate a surface wave decaying exponentially from the aperture. Uniform high-density plasma beyond the cut-off was realized over a large area in the system [107].

3.4.3 Millimeter-wave Applications

The design of RLSAs is extended to higher frequencies up to 60GHz [108]. A slotted waveguide array is the most attractive candidate for high-gain planar antennas in the millimeter wave band, having smaller transmission loss in comparison with a microstrip and triplate line. Parallel plates used in a RLSA have no side walls, which dispenses with electrical contact between them. The RLSA is the most promising candidate of low-cost and mass producible planar antenna. The design procedure and the analysis model for 12GHz-band RLSAs are used. Two types for the 60GHz band are manufactured with different structures. The first one consists of three separate parts (a slotted plate, a dielectric sheet and a bottom plate) to be piled up. The transmission loss remains small. However, the misalignment of the two plates has to be carefully minimized as the frequency becomes higher. As an alternative structure to

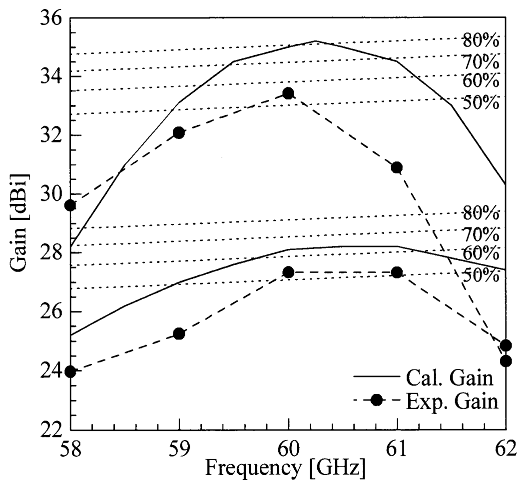

Fig. 24 Gain of 60GHz RLSA.

cope with this problem, the second type utilizes an existing PTFE substrate with copper foils on both sides which assures structural accuracy as well as productivity. Transmission loss of the substrate is estimated to be four times larger than the previous type due to the dielectric loss of PTFE. Figure 24 shows the gain of antennas using the dielectric substrates with different diameters (50mm and 100mm). The peak gain in the measurements is 27.3dBi in the 50mm antenna and 33.4dBi in the 100mm antenna. The efficiency is around 55% in the both. The gain is lower than the prediction by about 1dB in the 50mm antenna and about 2dB in the 100mm antenna. The possible reason for the gain reduction seems to be the positioning error of the feed point. This error is estimated to be 0.5mm in the 100mm antenna.

4. Conclusion

Development of the microstrip antenna and radial line slot antenna in Japan was described. About the former, a useful and accurate method was derived to analyze the behavior of an arbitrarily shaped MSA with multiterminals and its equivalent circuit was shown. Next, a design procedure able to take the bandwidth into consideration was established. Although two of three design charts used here were constructed from numerical results, the reader could prepare similar charts on the basis of accumulated experimental results. Furthermore, as an interesting feature of a MSA, it was made clear that it has two inherent CP operating frequencies at which perfect circular polarization is achieved with a single feed. In Sect 2.4, useful wideband technique, called sequential arrangement technique was introduced.

As for the RLSA, the low transmission loss of waveguide gives high-efficiency characteristics, which is advantageous especially in high-gain and/or high-frequency range in comparison with other types of planar antennas. Traveling wave excitation of slots results in the stable operation with rotational symmetry in a parallel plate waveguide and the simple design procedures based on the power conservation. The introduction of the periodic boundary condition in the narrow walls in a rectangular waveguide predicts well the coupling in a uniformly-excited two-dimensional slot array. The extensions to plasma etching system and millimeter-waves are promising for future applications of RLSAs.

References

[1] T. Ohkoshi and T. Miyoshi, "The planar circuits - An approach to microwave integrated circuitry," IEEE Trans. Microwave Theory & Tech., vol.MTT-20, pp.245–252, April 1972.

[2] H.J. Pang and T. Anada, "Computer analysis of microstrip planar circuits," IEICE Technical Report, MW73-7, April 1973. (in Japanese)

[3] Y. Suzuki and T. Chiba, "Computer Analysis Method for Arbitrarily Shaped Microstrip Antenna with Multi-terminals," IEEE Trans. Antennas & Propag., vol.AP-32, pp.585–590, June 1984.

[4] Y. Suzuki and T. Chiba, "Designing Method of Microstrip Antenna Considering the Bandwidth," IECE Trans., vol.E67, pp.488–493, Sept. 1984.

[5] Y. Suzuki and T. Chiba, "Improved Theory for a Singly-Fed Circularly Polarized Microstrip Antenna," IECE Trans., vol.E68, pp.76–82, Feb. 1985.

[6] T. Taga, H. Mishima, and T. Kanehori, "Broadband microstrip antenna for UHF-band," Nati. Conv. Rec., IECE Japan, S6-6, March 1979. (in Japanese)

[7] T. Teshirogi, M. Tanaka, and W. Chujo, "Wideband Circularly Polarized Array Antennas with Sequential Rotations and Phase Shift of Elements," Proc., ISAP, pp.117–120, Aug. 1985.

[8] M. Haneishi, "Planar Antenna," Journal of ITE, vol.38, no.11, pp.976–984, 1984. (in Japanese)

[9] K. Itoh, and T. Teshirogi, "Thin Antenna Technology," IECE Trans., vol.J71-B, no.11, pp.1217–1227, Nov. 1988. (in Japanese)

[10] Y. Yoshimura, "A microstrip line slot antenna," IEEE Trans. Microwave Theory & Tech., vol.MTT-20, pp.760–762, 1972.

[11] C. Wood, P.S. Hall, and J.R. James, "Design of wideband circularly polarized microstrip antennas and array," IEE Conf. Antennas & Propag., IEE Conf. Publ. 169, Pt1, pp.312–316, 1978.

[12] K. Nakaoka K. Itoh, and T. Matsumoto, "Microstripline Slot Array Antenna," IECE Trans., vol.61-B, 11, pp.943–950, Nov. 1978. (in Japanese)

[13] I.E. Rana, and N.G. Alexopoulous, "Current Distribution and Input Impedance of Printed Dipoles," IEEE Trans., Antennas & Propag. AP-29, no.1, pp.99–105, Jan. 1981.

[14] K. Ito, N. Aizawa, and N. Goto, et al., "Circularly Polarized Printed Array Antennas Composed of Strips and Slots," Electron. Lett., no.15, 25, pp.811–812, Dec. 1979.

[15] M. Ando, T. Numata, J. Takada and N. Goto, "A Linearly Polarized Radial Line Slot Antenna," IEEE Trans. Antennas & Propag., vol.AP-36, no.12, pp.1675–1680, Dec. 1988.

[16] M. Haneishi, S. Yoshida, and N. Goto, "Patch Antenna and its Pair," IEICE Technical Report, AP81-102, pp.39–42, Nov. 1981. (in Japanese)

[17] R.E. Munson, "Conformal Microstrip Antennas and Mi-

crostrip Phased Arrays," IEEE Trans. Antennas & Propag., vol.AP-22, pp.74–78, Jan. 1974.
[18] W.F. Richards, Y.T. Lo, and D.D. Harrison, "An Improved Theory for Microstrip Antennas and Applications," IEEE Trans. Antennas & Propag., vol.AP-29, pp.38–46, Jan. 1981.
[19] T. Fujimoto, K. Tanaka, and M. Taguchi, "Analysis of Elliptical Microstrip Antennas with and without a Circular Slot," IEICE Trans. Commun., vol.E83-B, no.2, pp.386–393, Feb. 2000.
[20] K.R. Carver, "A modal Expansion Theory of the Microstrip Antenna," AP-S Int. Symp. Digest, pp.101–104, 1979.
[21] J.H. Newman, and P. Tulyathan, "Analysis of Microstrip Antennas Using Moment Methods," IEEE Trans. Antennas & Propag., vol.AP-29, pp.47–53, Jan. 1981.
[22] N. Inagaki and K. Ariyoshi, "Numerical Analysis of Microstrip Antenna," AP-S Int Symp. Digest, pp.609–612, 1980.
[23] E.L. Coffey, and K.R. Carver, "Towards the Theory of Microstrip Antenna Patterns," Proc. Ant. Appl. Symp., Univ. of Illinois, Urbana, IL, April 1977.
[24] K.R. Carver, and E.L. Coffey, "Theoretical Investigation of the Microstrip Antenna," Technical Report, PT-00929, Physical Science Laboratory, New Mexico State Univ., Las Cruces, NM, Jan. 1979.
[25] E.L. Coffey, "DFNA Analysis of Microstrip Antenna," AP-S Int Symp. Digest, pp.613–616, 1980.
[26] P.K. Agrawal, and M.C. Bailey, "An Analysis Technique for Microstrip Antennas," IEEE Trans. Antennas & Propag., vol.AP-25, pp.756–759, Nov. 1977.
[27] K.C. Gupta, and P.C. Sharma, "Segmentation and Desegmentation Techniques for an Analysis of Planar Microstrip Antennas," AP-S Int. Symp. Digest, pp.19–22, 1981.
[28] K. Araki and T. Itoh, "Hankel Transform Domain Analysis of Open Circular Microstrip Radiating Structures," IEEE Trans. Antennas & Propag., vol.AP-29, pp.84–89, Jan. 1981.
[29] T. Itoh and W. Menzel, "A Full-Wave Analysis Method for Open Microstrip Structures," IEEE Trans. Antennas & Propag., vol.AP-29, no.1, pp.63–68, Jan. 1981.
[30] D.M. Pozar, "Input Impedance and Mutual Coupling of Rectangular Microstrip Antennas," IEEE Trans. Antennas & Propag., vol.AP-30, no.6, pp.1191–1196, Nov. 1982.
[31] N. Ishii, T. Fukasawa, and K. Itoh, "Analysis of High-Tc Superconducting Microstrip Antenna Using Modified Spectral Domain Moment Method," IEICE Trans. Electron., vol.E77-C, no.8, pp.1242–1248, Aug. 1994.
[32] T. Kashiwa, S. Koike, N. Yoshida, and I. Fukai, "Three Dimensional Analysis of Patch Antenna by Bergeron's Method," IECE Trans., vol.J71-B, pp.576–584, April 1988. (in Japanese)
[33] H. Ohmine, H. Mizutamari, and Y. Sunahara, "A New High Gain Circularly Polarized Microstrip Antenna with Diagonal Short," IEICE Trans. Commun., vol.E80-B, no.7, pp.1090–1097, July 1997.
[34] A. Reineix, and B. Jecko, "Analysis of Microstrip Patch Antennas Using Finite Difference Time Domain Method," IEEE Trans. Antennas & Propag., vol.AP-37, pp.1361–1369, Nov. 1989.
[35] T. Ohnishi, T. Kashiwa, and I, Fukai, "Analysis of Microstrip Antenna on a Curved Surface Using the Conformal Grids FD-TD Method," IEICE Trans., vol.J75-B-II, no.12, pp.957–963, Dec. 1992. (in Japanese)
[36] E. Nishiyama and S. Egashira, "The Analysis of Staked Microstrip Antenna with high gain using the FDTD method," ITE Technical Report, vol.22,no.7, pp.19–24, Jan. 1988. (in Japanese)
[37] T. Arima, T. Uno, and Y. Takahashi, "Improvement of FDTD Calculation Accuracy for Analyzing Rectangular Patch Antenna by Using Quasi-static Approximation," IEICE Trans., vol.J85-B, no.6, pp.1001–1004, June 2002. (in Japanese)
[38] S. Ohmori, U. Morikawa, N. Miyano, Y. Suzuki, and T. Chiba, "Circularly polarized shaped beam antenna for maritime communication," Nati. Conv. Rec., IECE Japan, S1-4, pp.403–404, 1980. (in Japanese)
[39] T. Chiba, Y. Suzuki, and N. Miyano, "Suppression of Higher Modes and Cross Polarized Component of Microstrip Antenna," AP-S Int. Symp. Dugest, pp.285–288, 1982.
[40] J.L. Kerr, "Microstrip polarization techniques," Proc. 1978 Antenna Appl. Symp., Allerton Park, IL, 1978.
[41] M. Haneishi, S. Yoshida, and N. Oka, "Back-Feed Type Circularly Polarized Microstrip Disk Antenna by One-Point Feed," IECE Trans., vol.J63-B, pp.559–565, June 1980. (in Japanese)
[42] M. Haneishi, S. Yoshida, "A Design Method of Circularly Polarized Microstrip Antenna by One-Point Feed," IECE Trans., vol.J64-B, pp.225–231, April 1981. (in Japanese)
[43] L.M. Kakoi, "Corner Fed Electric Microstrip Dipole,2" Naval Missile Center, Point Mugu, CA, March 1978.
[44] S.A. Long, L.C. Shen, D.H. Schaubert, and F.G. Farrar, "An Experimental Study of the Circular Polarized Elliptical Printed-Circuit Antennas," IEEE Trans. Antennas & Propag., vol.AP-29, pp.95–99, Jan. 1981.
[45] S. Tokumaru and S. Fukui, "Microstrip Antennas Having Posts for Circular Polarization," IECE Trans., vol.J67-B, no.5, pp.529–536, May 1984. (in Japanese)
[46] H.D. Weinschel, "A cylindrical array of circularly polarized microstrip antenna," AP-S Int. Symp. Dugest, pp.177–180, 1975.
[47] Y. Suzuki, N. Miyano, and T. Chiba, "Circularly Polarized Radiation from Singly-Fed Equilateral-Triangular Microstrip Antenna," Proc. IEE, Pt.H, vol.134, pp.194–198, April 1987.
[48] L. Murphy, "SESAT and SIR-A Microstrip Antennas," Proc. Workshop on Printed Circuit Antenna Technology, New Mexico State University, Las Cruces, NM, pp.18.1–18.20, Oct. 1979.
[49] Y. Suzuki, N. Miyano, and T. Chiba, "Expanding the Frequency Bandwidth of a Microstrip Antenna," AP-S Int. Symp. Digest, pp.617–620, 1980.
[50] H.F. Pues and A.R. Van De Capelle, "An Impedance-Matching Technique for Increasing the Bandwidth of Microstrip Antennas," IEEE Trans. Antennas & Propag., vol.AP-37, pp.11345–1354, Nov. 1989.
[51] H. Itami and T. Hori, "Broad Band Circular Polarized Microstrip Antenna," Nati. Conv. Rec., IECE Japan, p.642, 1982. (in Japanese)
[52] K. Araki, H. Ueda, and M. Takahashi, "Numerical Analysis of Circular Disk Microstrip Antenna with Parasitic Elements," IEEE Trans. Antennas & Propag., vol.AP-34, pp.1390–1394, Dec. 1986.
[53] A. Sabban, "A new broadband stacked two-layer microstrip antenna," AP-S Int. Sym. Digest, pp.63–66, 1983.
[54] T. Chiba, Y. Suzuki, N. Miyano, S. Miura, and S. Ohmori, "A phased array antenna using microstrip patch antenna," Proc. 12th Eur. Microwave Conf., pp.472–477, 1982.
[55] M. Haneishi, S. Yoshida, and N. Goto, "A Broadband Microstrip Array Composed of Single-Feed Type Circularly Polarized Microstrip Antenna," AP-S Int. Symp. Digest, pp.160–163, 1982.
[56] M. Haneishi, S. Saito, S. Yoshida, and N. Goto, "A Circularly Polarized Planar Array Composed of the Microstrip

Pairs Element," IEICE Technical Report, AP83-64, 1983. (in Japanese)

[57] N. Goto and K. Kaneta, "Ring patch antennas for dual frequency us," AP-S Int. Sym. Digest, pp.944–947, June 1987.

[58] H. Iwasaki, T. Nakajima, and Y. Suzuki, "Circularly Polarized Self-Diplexing Array Antenna with Higher Gain Operation," IEEE Trans. Antennas & Propag., vol.43, no.10, 1156–1159, Oct. 1995.

[59] N. Goto and M. Yamamoto, "Circularly polarized radial-line slot antennas," IECE Tech. Rept., AP80-57, Aug. 1980. (in Japanese)

[60] S. Tsurumaru, "The developement and the technological trend of circularly-polarized planar antennas," Techno System Seminar Rec., Jan. 1988. (in Japanese)

[61] N. Goto, N. Miyahara and K. Arimura, "Radiation pattern of a radial line slot antenna," IECE Tech. Rept., AP81-90, Oct. 1981. (in Japanese)

[62] S. Ito, M. Ando and N. Goto, "An analysis of radial line slow wave circuit," IECE Tech. Rept., AP84-70, Nov. 1984. (in Japanese)

[63] M. Ando, K. Sakurai, N. Goto, K. Arimura and Y. Ito, "A radial line slot antenna for 12 GHz satellite TV reception," IEEE Trans. Antennas Propagat., vol.33, no.12, pp.1347–1353, Dec. 1985.

[64] M. Ando, K. Sakurai and N. Goto, "Characteristics of a radial line slot antenna for 12 GHz band satellite TV reception," IEEE Trans. Antennas Propagat., vol.34, no.10, pp.1269–1272, Oct. 1986.

[65] F.J. Goebels Jr., K.C. Kelly; "Arbitrarily polarized planar antennas," 1959 IRE Natl. Conv. Record, pp.119–127, March 1959.

[66] H. Sumiyoshi, H. Arai and N. Goto, "The feed circuit for exciting a rotating electric field in a radial waveguide," IEICE Spring Natl. Conf., B-51, March 1993. (in Japanese)

[67] H. Sumiyoshi, H. Arai and N. Goto, "Feed circuit by cavity resonator for radial-line waveguide," IEICE Tech. Rept., AP93-33, May 1993. (in Japanese)

[68] H. Arai and N. Goto, "Novel feed circuit for radial-line waveguide," Proc. of Ninth Intl. Conf. Antennas Propagat., vol.1, pp.231–234, 1995.

[69] M.E. Bialkowski, P.W. Davis, and R.S. Varnes, "Analysis of a recessed cavity radiating into a radial waveguide," Proc. of Asia-Pacific Microwave Conf., pp. 800–803, Oct. 1995.

[70] Y. Kigure, A. Akiyama, J. Hirokawa and M. Ando, "A radial line slot antenna fed by a ring slot coupled planar circuit," Korea Japan Joint Conf. '98, Sept. 1998.

[71] K. Sudo, T. Hirano, J. Hirokawa and M. Ando, "An analysis and a design for excitation of a rotating mode in a radial waveguide by a cross slot-coupled rectangular waveguide," IEICE Tech. Rept., AP2001-206, Feb. 2002. (in Japanese)

[72] J. Takada, A. Tanisho, K. Ito and M. Ando, "Circularly polarized conical beam radial line slot antenna," Electronics Letters, vol.30, no.21, pp.1729–1730, Oct. 1994.

[73] C. Phongcharoenpanich, M. Krairiksh and J. Takada, "Investigations of radiation characteristics of a circularly polarized conical beam spherical slot array antenna," IEICE Trans. Electron., vol.82, no.7, pp.1242–1247, July 1999.

[74] A. Akiyama, J. Hirokawa, M. Ando, E. Takeda and Y. Arai, "60GHz band small aperture conical beam radial line slot antennas," IEICE Trans. on Electron., vol.82, no.7, pp.1229–1235, July 1999.

[75] M. Ando, T. Numata, J. Takada and N. Goto, "A linearly-polarized radial line slot antenna," IEEE Trans. Antennas Propagat., vol.36, no.12, pp.1675–1680, Dec. 1988.

[76] J. Takada, M. Ando and N. Goto, "A reflection canceling slot set in a linearly polarized radial line slot antenna," IEEE Trans. Antennas Propagat., vol.40, no.4, pp.433–438, April 1992.

[77] J. Takada, T. Numata, M. Ando and N. Goto, "A beam-tilted linearly polarized radial line slot antenna," IEICE Spring Natl. Conf., B-107, March 1988. (in Japanese)

[78] P.W. Davis and M.E. Bialkowski, "Comparing beam squinting and reflection cancelling slot methods for return loss improvement in RLSA antennas," IEEE AP-S Intl. Symp. Digest, vol.3, pp. 1938–1941, 1997.

[79] H. Miyashita and T. Katagi, "Four-point fed radial line planar antenna array for monopulse tracking operation," IEEE AP-S Intl. Symp. Digest, vol.1, pp.136–139, 1993.

[80] M.S. Castafier, M.S. Perez and M.V. Isasa, "Design of monopulse radial line slot antennas," IEEE AP-S Intl. Symp. Digest, vol.4, pp. 2774–2777, 1999.

[81] J. Takada, T. Tanahashi, K. Ito and M. Ando, "A dual beam linearly-polarized radial line slot antenna," IEEE AP-S Intl. Symp. Digest, vol.3, pp.1624–1627, 1993.

[82] H. Sasazawa, Y. Oshima, M. Ando and N. Goto, "Slot coupling in a radial line slot antenna for 12 GHz band satellite TV reception," IEEE Trans. Antennas Propagat., vol.39, no.9, pp.1221–1226, Sept. 1988.

[83] J. Hirokawa, M. Ando and N. Goto, "An analysis of slot coupling in a radial line slot antenna for DBS reception," IEE Proc., pt.H, vol.137, no.5, pp.249–254, Oct. 1990.

[84] M. Ando, K. Nishimura, S. Ito, K. Sakurai and N. Goto, "A radial line slot antenna with a slow-wave corrugation for 12GHz band satellite TV reception," IEEE AP-S Intl. Symp., vol.2, pp.919–922, June 1986.

[85] J. Hirokawa, M. Ando and N. Goto, "Waveguide-Fed Parallel Plate Slot Array Antenna," IEEE Trans. Antennas Propagat., vol.40, no.2, pp.218–223, Feb. 1992.

[86] J. Hirokawa and M. Ando, "Single-Layer Feed Waveguide consisting of Posts for Plane TEM Wave Excitation in Parallel Plates," IEEE Trans. Antennas & Propagat., vol.46, no.5, pp.625–630, May 1998.

[87] M. Takahashi, J. Takada, M. Ando and N. Goto, "A slot design for uniform aperture field distribution in single-layered radial line slot antennas," IEEE Trans. Antennas Propagat., vol.39, no.7, pp.954–959, July 1991.

[88] M. Takahashi, Y. Nakagawa and M. Abe, "Low sidelobes for radial line slot antennas," Electron. Commun. in Japan, Wiley InterSci., pt.1, vol.80, no7, pp.70–76, July 1997.

[89] J. Hirokawa and M. Ando, "Sidelobe suppression in 76 GHz post-wall waveguide-fed parallel plate slot arrays," IEEE Trans. Antennas Propagat., vol.48, no.11, pp.1727–1732, Nov. 2000.

[90] A. Akiyama, T. Yamamoto, M. Ando and N. Goto, "Numerical optimisation of slot parameters for a concentric array radial line slot antenna," IEE Proc. Microw. Antennas Propag., vol.145, no.2, pp.141–145, April 1998.

[91] M. Ando, S. Ito, K. Sakurai and N. Goto, "Suppression of reflection in a radial line slot antenna for 12GHz band satellite TV reception," IEEE AP-S Intl. Symp., vol.2, pp.898–901, 1987.

[92] M. Ando, H. Sasazawa, S. Nishikata and N. Goto, "A slot design of radial line slot antennas," IEICE Trans., vol.71-B, no.11, pp.1345–1351, Nov. 1988. (in Japanese)

[93] M. Takahashi, J. Takada, M. Ando and N. Goto, "Characteristics of small-aperture, single-layered, radial line slot antennas," IEE Proc., pt.H, vol.139, no.1, Feb. 1992.

[94] M. Ando, M. Natori, T. Ikeda and N. Goto, "A matching spiral for a single-layered radial line slot antenna," IEICE Trans., vol.73-E, no.8, pp.1322–1325, Aug. 1990.

[95] M. Takahashi, J. Takada, M. Ando and N. Goto, "Aperture illumination control in radial line slot antennas," IEICE Trans. Commun., vol.76, no.7, pp.777–783, July 1993.

[96] M. Ueno, M. Takahashi, J. Hirokawa, M. Ando, N. Goto and H. Arai, "A rotating mode radial line slot antenna -concentric array-," IEICE Tech. Rept., AP93-43, June 1993. (in Japanese)
[97] S. Hosono, J. Hirokawa, M. Ando, N. Goto and H. Arai, "A rotating mode radial line slot antenna fed by a cavity resonator," IEICE Trans. Commun., vol.78, no.3, pp.407–413, March 1995.
[98] M. Takahashi, M. Ando, N. Goto, Y. Numano, M. Suzuki, Y. Okazaki and T. Yoshimoto, "Dual circularly polarized radial line slot antenna," IEEE Trans. Antenna Propagat., vol.43, no.8, pp.874–876, Aug. 1995.
[99] T. Yamamoto, M. Takahashi, M. Ando and N. Goto, "Enhancement of band-edge gain in radial line slot antennas using the power divider -a wide-band radial line slot antenna-," IEICE Trans. Commun., vol.78, no.3, pp.398–406, March 1995.
[100] K. Ichikawa, J. Takada, M. Ando and N. Goto, "A radial line slot antenna without a slow wave structure," IEICE Trans., vol.75-B-II, no.6, pp.363–369, June 1992. (in Japanese)
[101] H. Nakano, H. Takeda, Y. Kitamura, H. Mimaki and J. Yamauchi; "Low-profile helical array antenna fed from a radial waveguide," IEEE Trans. Antennas Propagat., vol.40, pp. 279–284, March 1992.
[102] O. Shibata, S. Saito and M. Haneishi, "Radiation properties of microstrip array antennas fed by radial line," Electron. Commun. in Japan, Wiley InterSci., pt.1, vol.76, no.12, pp.93–102, Dec. 1993.
[103] M. Ogawa, T. Watanabe, K. Nishikawa, T. Harada, E. Teramoto and M. Morita; "Mobile Antenna System for Direct Broadcasting Satellite," Proc. of ISAP '96, vol.4, pp.1197–1200, Sept. 1996.
[104] M. Takahashi, M. Ando, N. Goto, N. Ishii and K. Imahashi "A planar antenna for ECR plasma system," IEICE Spring Natl. Conf., B-65, March 1993. (in Japanese)
[105] T. Ohmi, "Preface in new era of semiconductor manufacturing I," Ultra Clean Tech., vol.9, Suppl. 1., 1997.
[106] T. Yamamoto, M. Ando, N. Goto, N. Ishii, M. Takahashi and Y. Horiike, "Plasma etching system using a radial line slot antenna," IEICE Commum. Conf., B-27, Sept. 1995.
[107] Y. Yasaka, D. Nozaki, K. Koga, M. Ando, T. Yamamoto, N. Goto, N. Ishii and T. Morimoto, "Production of large-diameter uniform plasma in mTorr range using microwave discharge," Jpn. J. Appl. Phys., vol.38, pt.1, no.7B, pp.4309–4312, July 1999.
[108] A. Akiyama, T. Yamamoto, J. Hirokawa, M. Ando, E. Takeda and Y. Arai, "High gain radial line slot antennas for millimeter wave applications," IEE Proc. Microwaves, Antennas Propag. vol.147, no.2, pp.134–138 April 2000.

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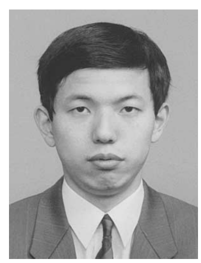

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