1. Introduction

In the past few decades, as the growing demands of ground-satellite communications, many researches focusing on the satellite onboard antennas have been conducted. A high gain is always required for a satellite onboard antenna since its high EIRP means a high quality link and a size and cost reduction of receiving dishes on earth stations. The most common antennas used in satellite communications are parabolic reflectors. They offer high gain, wide band, and low fabrication cost. However, conventional parabolic reflectors (paraboloids or dishes) suffer from a large volume of the antenna system including the reflector, the feed horn and sometimes the subreflector, which results in a heavy weight and difficulties in manufacture and installment on a satellite.

To deal with this issue, planar antennas, which allow size and cost reductions, have been considered as the best candidate to replace the reflectors [1]. Moreover, with the development of advanced technologies such as printed circuit board (PCB) and high precision etching techniques in the past few years, planar slotted array antennas have become an effective choice for many satellite and millimeter-wave applications.

Radial Line Slot Antenna (RLSA) is a planar slot array antenna that utilizes the oversized radial waveguide as the feeding network and narrow slot as the radiating element. RLSA is well known for its high gain, high efficiency, and compact structure. The first RLSA was designed and commercialized for DBS reception at 12 GHz [2], and followed by a number of applications such as RLSAs for power transmission [3], plasma excitation in semiconductor processing [4], radars for collision avoidance [5], and radio access networks in the millimeter waveband [6]. Recently, RLSAs that can realize shaped beam [7], and high gain - lightweight RLSA [8] were studied for satellite onboard uses.

In 2010, a lightweight RLSA with a honeycomb structure for X-band operation was designed, manufactured, and successfully mounted on the Japanese satellite called “Planet-C-Akatsuki” [8]–[10]. The key improvement in this antenna was the use of a honeycomb structure as the dielectric spacer to reduce the antenna weight. As result, a 900 mm diameter RLSA weights only about 1 kg, which is very attractive for a satellite onboard application. Motivated by this success, we continued to develop a 32 GHz RLSA with honeycomb structure in part of the “Hayabusa 2” project of JAXA [11]. Different from Planet-C, the honeycomb structure fabricated by NoMEX® fiber, which has cell sizes small enough for mm-wave, suffers a great loss of about 0.014–0.018 dB/mm, and it degrades the antenna performance considerably [12]–[14]. Nevertheless, demands for high gain and lightweight planar antennas in mm-wave are rapidly increasing. Therefore, researches to make use of lightweight, lossy materials such as NoMEX® honeycomb seem unavoidable, and become urgent tasks for satellite antenna designers.

This paper proposes a new RLSA design that takes losses of the dielectric waveguide into account, or put differently, the lossy design of RLSA. The basic idea for improving the antenna gain is to adopt strongly coupled slots in the inner part of the aperture so that the energy is radiated early during radially outward travel. This early radiation can theoretically suppress the energy absorbed by the lossy waveguide, however, it also compromise the uniform radiation of RLSA. This “tradeoff” is mathematically optimized to maximize the antenna gain, and is detailed in Sect. 2. Section 3 shows numerical results for various cases with different antenna sizes, loss factors, and slot coupling distributions. Section 4 report measured performances of several prototypical RLSAs, which are designed and fabricated using a very lossy material at 60 GHz. Finally, conclusions are drawn in Sect. 5.
2. Gain Optimization by the Slot Coupling Control

In RLSA operation, the use of an oversized radial waveguide filled with low loss dielectric materials is the main advantage for high antenna gain and efficiency, given the low conductor loss of the metal slot plate and reflection canceling effect of the unit radiating element \[15\]. Figure 1 shows a typical configuration of RLSA, and the antenna operation is illustrated in Fig. 2. A simple coaxial feeder attached to the bottom metal plate of the antenna at the center and creates a cylindrical wave travelling inside a radial waveguide filled with dielectrics. This travelling wave then excites the slots etched on the top plate and determines the uniformity of the radiated electrical field.

Takahashi et al. \[16\]–\[18\] have established a procedure to realize uniform amplitude and phase of the field radiated from the slots, which could assure high antenna gain if the dielectric loss was negligible. The key techniques of this procedure are:

- the slot coupling distribution to control the amplitude of the electrical field radiated from slots;
- the slot positioning to have an in-phase excitation and radiation.

The latter is normally realized by spacing slots in the radial direction a distance \(S_\rho\) of about one guided wavelength \(\lambda_g\), and some small adjustments of the slot angles to fix the radiating phase. On the other hand, the slot coupling distribution to control the radiated amplitude is theoretically calculated through an energy conservation model as shown in Fig. 3.

2.1 Energy Conservation Models

Figures 3(a), (b), and (c) demonstrate the energy conservation models for a lossless case, a general lossy case, and a lossy case in an incremental distance \(\Delta \rho\), respectively. These figures suggest that if the dielectric losses are high, an early radiation is recommended to reduce the amount of energy absorbed by the lossy dielectric materials. However, doing so would sacrifice the uniformity of the radiated field, which means losing the directivity. Therefore, we have to look for a reasonable compromise between early radiation and uniformity for a gain optimization.

Considering a travelling wave in an incremental distance \(\Delta \rho\) inside a lossy waveguide as illustrated in Fig. 3(c), the energy conservation can be expressed as,

\[
P_{\text{in}}(\rho) \frac{2\pi \rho d - P(\rho + \Delta \rho) 2\pi (\rho + \Delta \rho) d}{P_{\text{out}}} = 2\pi \rho P(\rho) \tau_{\rho} \Delta \rho + 2\pi \rho P(\rho) \tau_\rho \Delta \rho \]

(1)

Fig. 1 Structure of a typical radial line slot antenna.

Fig. 2 Operation of a typical radial line slot antenna.

Fig. 3 Energy conservation models in lossless/lossy radial waveguides.
where $d$ is the waveguide height and $\tau_i/\tau_j$ is the ratio of the radiated/absorbed power to the inner power per unit length,

$$
\tau_i(\rho) = \frac{P_{rad}(\rho)}{\tau_j} = \frac{P_{rad}(\rho)}{P_{in}(\rho)}
$$

$$
\tau_j = \frac{P_{loss}(\rho)}{P_{in}(\rho)}
$$

with $P_{in}$, $P_{out}$, $P_{rad}$, and $P_{loss}$ denoted in Fig. 3(c).

In this paper, as we focus on the design of high gain RLSA, the residual power at the antenna rim is assumed to be negligible. For example, in the practical design of Hayabusa 2 antenna [13], the residual power $t$ is estimated to be less than 1%. For an RLSA with small aperture, the limitation of the coupling factor results in an considerably small than its length, a slot coupling or coupling factor width. And for a longitudinal slot whose width is much smaller than its length, a slot coupling or coupling factor is mainly determined by the slot length. In term of the coupling factor distribution

Theoretically, if the phase is uniform over the aperture, the antenna gain at the bore-sight of an RLSA can be calculated as a function of coupling factor $\alpha_r(\rho)$ in straight-forward manner by,

$$
G[\alpha(\rho)] = \frac{\left(\int_{\rho_{\min}}^{\rho_{\max}} E(\rho) d\rho d\phi \right)^2}{2\pi \rho_{\min} P(\rho_{\min})}
$$

with $E(\rho)$ is the amplitude of the radiated field at a radial distance $\rho$.

$$
E(\rho) = \sqrt{P(\rho)\alpha_r(\rho)}
$$

Given the normalized incident power $2\pi \rho_{\min} P(\rho_{\min}) = 1$, (7) and (8) is simplified to (9),

$$
G[\alpha(\rho)] = \left(\int_{\rho_{\min}}^{\rho_{\max}} \sqrt{P(\rho)\alpha_r(\rho)} d\rho \right)^2
$$

where $\rho_{\max}$ and $\rho_{\min}$ stand for the position of the outermost and innermost slots, respectively. Substituting (6) into (9), we obtain the gain proportional to a function of coupling factor $\alpha_r(\rho)$.

$$
f(\alpha_r, \rho) = \left(\int_{\rho_{\min}}^{\rho_{\max}} \sqrt{\alpha_r(\rho)} \cdot e^{-(\alpha_r\rho + \rho \int_0^\rho \alpha_r(\rho) d\rho)} d\rho \right)^2
$$

The condition to maximize the function $f(\alpha_r, \rho)$ is,

$$
J[u] = \int_{\rho_{\min}}^{\rho_{\max}} \sqrt{\alpha_r(\rho)} \cdot e^{-(\alpha_r\rho + \rho \int_0^\rho \alpha_r(\rho) d\rho)} \to \text{maximum}
$$

with

$$
\alpha_r(\rho) = \int_{\rho_{\min}}^{\rho} \alpha_r(r) dr
$$

Eq. (11) is then solved by the calculus of variations with the augmented function is

$$
F[u] = -\sqrt{\rho u e^{-u}} e^{-\alpha_u}\rho
$$

Consequently, the solution of the coupling factor can be found by solving the following Euler-Lagrange equation,

$$
\frac{\partial F}{\partial u} - \frac{d}{d\rho} \left( \frac{\partial F}{\partial \alpha_u} \right) = 0
$$

and that is:

$$
\alpha_r(\rho) = \frac{2\rho^2\rho}{1-2\rho \alpha_u - \exp[2\alpha_u(\rho - \rho_{\max})]}(1 + 2\alpha_u\rho_{\max})
$$

Substituting (15) in (8), we have a tapered aperture field distribution which embodies an early radiation to cope with the power dissipation by the lossy waveguide. Without further mathematical complication, the coupling factor to realize uniform aperture field distribution in the lossy cases can also be derived by solving (8) with the uniform condition of the amplitude $E(\rho)$.

$$
\alpha_r(\rho) = \frac{2\rho^2\rho}{1 - 2\rho \alpha_u + \exp[2\alpha_u(\rho - \rho_{\max})](2\alpha_u \rho_{\max} - 1)}
$$

To avoid ambiguity, we define the coupling factor given by (15) as “maximum gain coupling” and (16) as “maximum directivity coupling”.

By taking the limit of $\alpha_u \to 0$ in (15) and (16), we recover “no loss design coupling” $\alpha_r(\rho)$ as given in (4),

$$
\lim_{\alpha_u \to 0} \alpha_r(\rho) = \frac{\rho}{\rho_{\max}^2 - \rho^2}
$$
It should be noted that in the process of arriving at (15) and (16), the residual power $t$ is neglected for simplicity. This neglect is quite acceptable for practical RLSA with the medium or higher gain since most of the power would be radiated by slots and further absorbed by lossy dielectrics before reaching the antenna rim.

3. Numerical Results

Some numerical examples are presented here to demonstrate the improvements of designs using the newly derived coupling factor given in (15) for lossy cases. At 60 GHz, a 400 mm ($=80\lambda_0$) diameter RLSA is analyzed for a lossy case with the loss factor $\alpha_l = 0.105$ dB/mm.

Figure 4 compares the coupling distributions given by (4), (15), and (16). The maximum coupling factor $\alpha_{\text{max}}$ is properly chosen as not to disturb the rotational symmetry of the inner field of RLSA [18]. It can be observed that, the maximum gain coupling (15) is stronger than no loss design coupling (4) and maximum directivity coupling (16), and all three curves are converged to a limitation $\alpha_{\text{max}}$ while the slots position approaches the antenna rim. On the other hand, a maximum directivity coupling (16) is the weakest among these three.

The relationship between slot coupling and slot length is established by applying the Method of Moments (MoM) for a unit slot pair model with the periodic boundary condition [19]. Figure 5 presents the slot length distributions associated with the coupling factor distributions $\alpha_r(\rho)$ in Fig. 4. As results of strong coupled slots in maximum gain design, longer slots are adopted in the inner part of the aperture, and the curves are converged to a resonant length $SL_{\text{max}}$ associated with the maximum coupling factor $\alpha_{\text{max}}$ as the slots are positioned near the antenna rim.

The amplitudes corresponding to the couplings in Fig. 4 are then calculated by (8) together with (6), assuming the loss factor $\alpha_l = 0.105$ dB/mm, and their normalized distributions are plotted in Fig. 6. The amplitudes corresponding to maximum gain coupling (15) and no loss design coupling (4) are both linearly tapered, and the sooner start at higher level than the latter one as a result of adopting stronger coupled slots in the inner part of the aperture. This steep tapered distribution confirms our desired early radiation as proposed in Fig. 3(b). Moreover, a uniform field corresponding to the maximum directivity coupling distribution consolidates the fidelity of (16) as well as the accuracy of our mathematical calculations in Sect. 2. Nonetheless, the low magnitude level ($\sim -11$ dB) is the consequence of the excessively suppressed coupling to realize the uniformity against the existence of a considerable loss, and it suggests the gain degradation. We also include a uniform amplitude distribution normalized at 0 dB level for the lossless case (black line) as reference.

Finally, we calculate the antenna gain using (7) for a number of cases with different antenna sizes, loss factors, and coupling distributions, and plot it in Fig. 7. The antenna radius $R_{\text{max}}$ is varied from $10\lambda_0$ to $40\lambda_0$ by a $10\lambda_0$ step, and the loss factors $\alpha_l = 0$, 0.06, or 0.105 dB/mm is considered here. Three designs corresponding to maximum gain coupling (15), maximum directivity coupling (16), and
**no loss design coupling** (4) are distinguished by color, i.e., red, blue, and dark yellow, respectively. For the lossless case ($\alpha_l = 0 \, \text{dB/mm}$), these lines are identical and indicated by the black line, as explained by (17).

Figure 7 suggests that loss factor $\alpha_l$ is the dominant parameter for the antenna gain though the new design (15) suppresses the degradation especially for larger radius and higher loss. For example, in case the loss factor $\alpha_l = 0.06 \, \text{dB/mm}$ and antenna radius $R = 200 \, \text{mm}$), more than about 7 dB gain degradation occurs. For lossy cases, the improvements of the red lines over the blue and dark yellow lines substantiate our idea, i.e., *early radiation* would prevent the dissipated power in lossy waveguides, hence increase the antenna gain. More importantly, the gain enhancement seriously depends on the loss factor and the antenna size, and is significant if the loss factor is very high (e.g., $\alpha_l = 0.105 \, \text{dB/mm}$) and the antenna size is sufficiently large. For instant, a remarkable gain enhancement ($\approx 1.5 \, \text{dB}$) is achieved for $R_{\text{max}} = 200 \, \text{mm}$ and $\alpha_l = 0.105 \, \text{dB/mm}$.

From another view point, if the gain of 37 dBi is specified with the lossy material of $\alpha_l = 0.105 \, \text{dB/mm}$, for example, only the *maximum gain design* (15) satisfies it by about 300 mm RLSA, while other design could not meet this specification even if the aperture size increases.

Figure 8 clarifies the advantage of the *maximum gain coupling* (15) over the *no loss design coupling* (4) and the *maximum directivity coupling* (16), by characterizing the gain of a 400 mm diameter RLSA as functions of the loss factor. Two observations can be drawn from this figure:
- the *maximum directivity design* suffers from the dielectric losses most seriously;
- the *maximum gain design* is indispensable to minimize the degradation caused by the dielectric losses, especially in case of very lossy dielectric materials and/or very large apertures.

---

**Fig. 7** Predicted gain for different antenna sizes and loss factors of the dielectric materials.

**Fig. 8** Predicted gain for a $\phi = 400 \, \text{mm}$ antenna with different loss factors of the dielectric materials and different coupling distributions.

---

### 4. Designs and Measurements of 60 GHz RLSAs Having Waveguides Filled with RO4003CTM

In order to verify our theory proposed in Sect. 2 and confirm the numerical results calculated in Sect. 3, we designed and fabricated three 60 GHz RLSAs having 400 mm diameter and waveguides filled with a lossy material — Roger4003C (RO4003CTM), a product of Rogers Corporation. RO4003CTM belongs to RO4000® series high frequency circuit materials, which provide stable electrical characteristics (dielectric constant $\varepsilon_r$, loss tangent $\delta$) over a wide range in high frequency bands. Detail of RO4003CTM electrical properties are reported in RO4000® Laminates data sheet for 0.5 GHz up to 10 GHz band [20].

#### 4.1 RO4003CTM Characteristics at 60 GHz by Outward Travelling Wave Measurement in a Radial Line

As the first step to prepare for the antenna design, we characterize the RO4003CTM at 60 GHz by outward travelling wave measurement in a radial line using our inner field measurement system. Operating mechanism of the inner field measurement system was explained in one of our recent reports [14]. This system provide us the one dimensional amplitude distribution of the field propagating inside RO4003CTM and the phase constant $\beta$, from which we can obtain the loss factor $\alpha_l$ and the dielectric constant $\varepsilon_r$, respectively.

Figure 9 presents the measured phase of the inner field. The gradient between measured curves and vertical axis defines the phase constant $\beta$, and it is from $-132.7 \, \text{deg/mm}$ to $-134.7 \, \text{deg/mm}$. Substitute $\beta$ into (18),

$$\varepsilon_r = \left(\frac{\lambda_0}{\lambda_g}\right)^2 = \left(\frac{\lambda_0\beta}{2\pi}\right)^2$$

we can identify the dielectric constant $\varepsilon_r$ of RO4003CTM, as about 3.38–3.5.

On the other hand, the measured amplitude of the inner
field is indicated by the dotted lines in Fig. 10. There are some discrepancies at a radial distance of about 25 cm due to the standing wave operation, but in general, we can observe a decaying trend of the measured amplitude in comparison with the theoretical lossless curve of the cylindrical wave (solid black line). And from that, the attenuation or loss factor \( \alpha \) is estimated to be about 0.08 dB/mm – 0.11 dB/mm.

### 4.2 Designs and Measurements of 400 mm φ RLSAs

Three 400 mm diameter RLSAs were designed at 60 GHz with the coupling factor distributed as in (4), (15), and (16), and with the assumptions of the dielectric constant \( \varepsilon_r \) and loss factor \( \alpha_l \) of RO4003C™ are 3.38 and 0.105 dB/mm, respectively. Each antenna has more than 30000 radiating elements, and the fast design/analysis procedure utilizing the Method of Moments (MoM) [13] is applied to predict the antenna performance. For clarification and simplicity, we name the antenna designed following the coupling factor distributed in (4), (15), and (16) as Ant1, Ant2, and Ant3, respectively.

Three antennas were designed at 60 GHz, and fabricated antennas were measured in the near field, with 5\( \lambda_0 \) distance from the WR15 probe, as shown in Fig. 11. The measured gains of three RLSAs are indicated by the solid-dotted lines in Fig. 12. It is observed that the operating frequency of the three is shifted down by about 1 GHz from the prediction (solid lines). This frequency shift corresponds to the slot over-etching by about 27 \( \mu m \), as is usual the case with slot fabrications at 60 GHz band [21], [22]. Regardless of the frequency shift, we confirm that Ant2, the design following the maximum gain coupling (15) has the best gain performance among these three prototypes; its peak is 0.8 dB and 3.2 dB higher than that of Ant1 — the no loss design coupling (4), and Ant3 — the maximum directivity coupling (16), respectively.

The predicted gain for three antennas using the parameters of over etching of 27 \( \mu m \), slightly higher dielectric constant \( \varepsilon_r \) of 3.49 and the loss factor of 0.105, are also included in Fig. 12 by the dash lines, which reasonably agree with the measured gains of three antennas. These parameters are in range of the measurement sensitivity of RO4003C™, as explained in Fig. 9 and Fig. 10. This result substantiates our theory in Sect. 2, i.e. the maximum gain coupling proposed in (15) give us the best antenna performance in case high loss materials are used.
5. Conclusions

In this paper, we have proposed a new design concept for RLSAs having lossy dielectric substrates. Based on the operation of conventional RLSA, a mathematical formula of the new coupling factor was derived in order to maximize the antenna gain. Numerical results were presented to substantiate the accuracy of the new coupling factor for both lossless and lossy cases. Finally, three designs using high loss dielectric material—RO4003C™, and following different coupling factor distributions, were evaluated and their measurement results supported our new design concept. This early radiation design ideology can be applied to improve the performance of Hayabusa 2 RLSA [14], which adopts high loss material-NoMEX® for the honeycomb structure.

Acknowledgments

This work was conducted in part as “the Research and Development for Expansion of Radio Wave Resources” under a contract with the Ministry of Internal Affairs and Communications. It was also supported by BEAM—the Project under Erasmus Mundus Action. The authors thank Mr. Sakurai at Tokyo Tech. for his support with the experiments.

References


Tung Nguyen was born in Haiphong, Vietnam, on March 10, 1987. He received the M.S. degree in electrical electronic engineering from Tokyo Institute of Technology, Japan in 2012. From 2008 to 2009, he joined the Young Science Exchange Program (YSEP), as a researcher at Tokyo Institute of Technology, where he is currently pursuing the M.S degree. His interest has been antenna measurements. Latest the design of high frequency array antennas and planar antennas are his interests. Mr. Nguyen is a Member of IEICE and a Student Member of IEEE. He won the Student Paper Contest in ISAP2012, and a Best Paper Award in ICCE2014. In late 2014, he was named in the URSI Young Scientist Award Scheme.
Jiro Hirokawa was born in Tokyo, Japan, on May 8, 1965. He received the B.S., M.S. and D.E. degrees in electrical and electronic engineering from the Tokyo Institute of Technology (Tokyo Tech), Tokyo, Japan in 1988, 1990 and 1994, respectively. He was a Research Associate from 1990 to 1996, and is currently an Associate Professor at Tokyo Tech. From 1994 to 1995, he was with the antenna group of Chalmers University of Technology, Gothenburg, Sweden, as a Postdoctoral Fellow. His research area has been in slotted waveguide array antennas and millimeter-wave antennas. Prof. Hirokawa is a Fellow IEEE and IEICE. He received an IEEE AP-S Tokyo Chapter Young Engineer Award in 1991, a Young Engineer Award from IEICE in 1996, a Tokyo Tech Award for Challenging Research in 2003, a Young Scientists’ Prize from the Minister of Education, Cultures, Sports, Science and Technology in Japan in 2005, a Best Paper Award in 2007, a Best Letter Award in 2009 from IEICE Communications Society and Asia Pacific Microwave Conference prize in 2011.

Makoto Ando received the D.E. degrees in electrical engineering from Tokyo Institute of Technology, Tokyo, Japan in 1979. From 1979 to 1983, he worked at Yokosuka Electrical Communication Laboratory, NTT, and was engaged in development of antennas for satellite communication. He moved to Tokyo Institute of Technology in 1983 and is currently a Professor. His main interests have been high frequency diffraction theory such as Physical Optics and Geometrical Theory of Diffraction. His research also covers the design of waveguide planar arrays and millimeter-wave antennas. Prof. Ando is the Fellow IEEE and IEICE. He received the Achievement Award and the Paper Awards from IEICE Japan in 1993 and 2009. He also received the 8th Inoue Prize for Science in 1992, the Meritorious Award of the Minister of Internal Affairs and Communications and the Chairman of the Broad of ARIB in 2004 and the Award in Information Promotion Month 2006, the Minister of Internal Affairs and Communications. He served as the guest editor-in-chief of more than six special issues in IEICE, Radio Science and IEEE AP. He was the general chair of the 2004 URSI EMT symposium in Pisa and of the ISAP 2007 in Niigata. He was the Program Officer for engineering science group in Research Center for Science Systems, JSPS in 2006–2009. He served as the member of Scientific Council for Antenna Centre of Excellence - ACE in EU’s 6th framework programme since 2004. He served as the Chair of Commission B of URSI 2002-2005. He was the 2007 President of Electronics Society IEICE, the 2009 President of IEEE Antennas and Propagation Society and is currently the Co-chair of ISAP International Steering Committee and vice-president of URSI and IEICE.

Manuel Sierra Castañer was born in 1970 in Zaragoza (Spain). He obtained a degree in Telecommunication Engineering in 1994 and a Ph.D. in 2000, from the Technical University of Madrid (UPM) in Spain. He worked for Airtel, a cellular company, from 1995 to 1997. Since 1997, he worked in the University “Alfonso X” as an assistant professor. Since 1998 he has worked at the Technical University of Madrid as a research assistant, assistant professor and associate professor. He is a senior member of the IEEE. His current research interests involve planar antennas and antenna measurement systems. Dr. Sierra-Castañer obtained the IEEE APS 2007 Schelkunoff Prize for his paper “Dual-Polarization Dual-Coverage Reflectarray for Space Applications” in 2007.