Millimeter-Wave Waveguide Slot-Array Antenna Covered by a Dielectric Slab and Arrayed Patches

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***Abstract*—A high-performance waveguide, 16 × 16 slot array antenna with a cover structure at 38 GHz is presented. The protective cover structure, which consists of a dielectric slab and arrayed conductor patches, offers an advantage for designing a highly directive array antenna because it lowers the resonant conductance of the slot set such that a large number of slots can be arrayed. The widths of the slots in this antenna are fixed to 1 mm for easy production with a common computer-controlled milling machine. Moreover, the manufacturing process for this antenna is simplified by positioning the waveguide corrugations and slots together on a plate, which eliminates a welding process. Measurements show an antenna gain of 30.8 dBi, a depolarization level below −30 dB, and an efficiency level of 71% at the frequency band of 37.6–38.6 GHz.**

***Index Terms*—Arrayed patches, dielectric cover, waveguide slot array antenna, wide slot.**

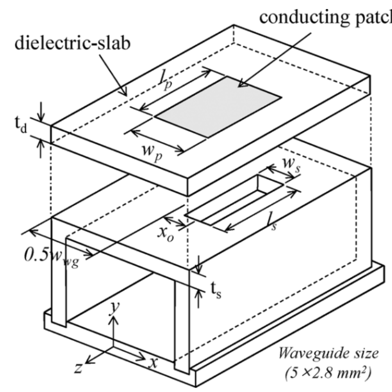

Fig. 1. Geometry of a single-slot set consisting of a radiating slot and a cover structure.

I. Introduction

A WAVEGUIDE slot array antenna is a good candidate for use in a broadband wireless local loop (BWLL) or a fixed wireless access (FWA) at the 38-GHz band because of its ability to control amplitude and phase distributions and polarization as well as its low power-loss characteristics and its single-layered flat structure. A serial-feed system using the T- or Π-junction makes it possible to have a single-layer waveguide slot array antenna [1]. The usual configuration for a single-layered waveguide slot antenna consists of a thin top plate with slots and a thick base plate with waveguide corrugations. Those two plates need to be welded for a reliable electrical contact because the top plate is generally too thin to make a sufficient mechanical contact. A simple manufacturing technique was introduced in [2], which showed a good contact between these plates without welding by putting the slots and waveguide corrugations together on a single plate. In another advance, the alternating-phase feeding technique was introduced in [3] and [4] to minimize the power spill between adjacent waveguides caused by imperfect contacts. In this letter, we combine the techniques in [2] and [3] to get rid of the unwanted power flows between adjacent waveguides.

Most theoretical analysis for slot antenna has been based on the assumption that the width of slot is infinitesimally narrow [4]–[8]. However, it is very difficult to manufacture the infinitesimally narrow slot with the normal milling technique. To overcome these limitations on the application and manufacture of waveguide slot antennas, we need a relatively wide slot for easy manufacture. A dielectric slab can be used as a cover for a waveguide slot antenna to protect it from water and other intrusion. However, this dielectric slab significantly affects the performance of these antennas because of the impedance change and the surface wave in the dielectric layer.

We propose an analytical initial-design procedure for a millimeter-wave waveguide slot array antenna with a dielectric cover structure and relatively wide slots. The degradation of the antenna characteristics by the dielectric cover and the wide slots has been overcome using conducting patches over the dielectric cover. The design goal of this study is to design a practical waveguide slot array antenna for a point-to-point FWA system at 38 GHz with 450 MHz bandwidth.

II. Single-Slot Antenna

Fig. 1 shows a waveguide single-slot antenna with a dielectric-slab cover and a rectangular patch on it. The slot has a relatively wide width ($w_s = 1$ mm) for easy production. The dielectric constant of the dielectric slab is $\varepsilon_r = 2.2$, and its thickness is $t_d = 0.5$ mm. The radiation patch on the dielectric slab is added over the slot to lower the admittance of the single slot set so that a large number of slots can be arrayed. The self-admittances of a single slot set with and without a cover structure were simulated using a full-wave analysis with the high-frequency structure simulator (HFSS).

In this simulation, the cross section of the waveguide is 5×2.8 mm^2, and its longitudinal length is about one guided wavelength. The input and output ports are assumed to be perfectly matched, and a radiation box has been positioned over the waveguide slot with the height of about 0.6 λ_0.

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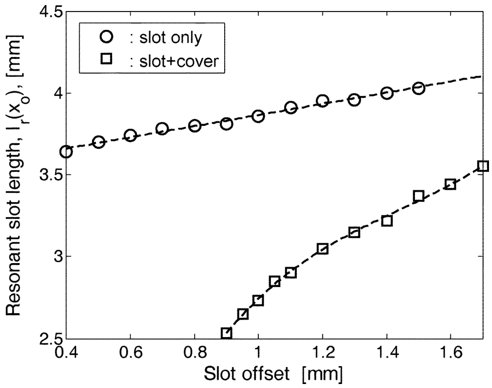

Fig. 2. Resonant slot lengths versus slot offset for the single-slot sets.

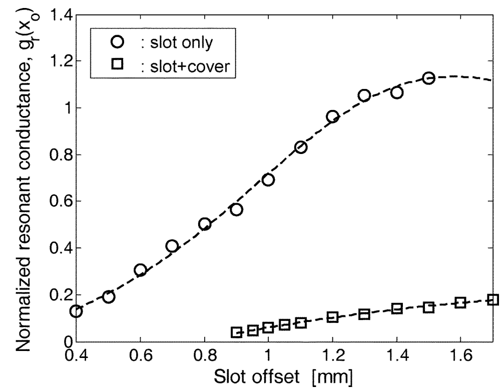

Fig. 3. Normalized resonant conductance versus slot offset for the single-slot sets.

A. Uncovered Single-Slot Set

We first examined the normalized self-admittance ($y = Y/Y_0$) of an uncovered single-slot set as well as the normalized resonant conductance (g_r) and the resonant slot lengths (l_r) for various slot offsets (x_o), which ranged from 0.4 to 1.5 mm for a fixed slot width ($w_s = 1$ mm; i.e., $0.13\lambda_0$) and a fixed waveguide thickness ($t_s = 0.5$ mm). After computing the reflection coefficient ($= S_{11}$) at the input port of the slot with the matched output port for various slot lengths and slot offsets, we then computed the normalized self-admittances (y) from the reflection coefficients as

$$y = g + jb = -\frac{2S_{11}}{1 + S_{11}}. \quad (1)$$

For each slot offset (x_o), the corresponding resonant slot length (l_r) and the normalized resonant conductance (g_r) were determined by searching the zero-crossing of the normalized self-susceptance (b). Figs. 2 and 3 show the resonant slot lengths (l_r) and the normalized resonant conductance (g_r), respectively, for the uncovered and covered single-slot sets. The lines with circles in Figs. 2 and 3 represent the uncovered single-slot set. For offsets of less than 0.4 mm, the resonant characteristics were deteriorated so much that we were not able to select the appropriate resonant slot length and resonant conductance from the simulation because the slot with a wide width is positioned close to the center of the waveguide broad-wall. As shown in Figs. 2 and 3, both the resonant slot length (l_r) and the resonant conductance (g_r) increase as the slot offset increases.

As shown in Fig. 4, the data sets of the normalized self-admittance were renormalized and juxtaposed for viewing the resonance characteristics of the antenna [9], [10] for the normalized slot length ($0.7 \leq l_n \leq 1.3$) at five different slot offsets ranging from 0.7 to 1.5 mm. Fig. 4 shows that the renormalized self-admittance (y_n) components for the uncovered slot were not affected much by the slot offset values even with the wide slot width.

B. Covered Single Slot Set

We have two ways for the impedance matching at the input port of a waveguide slot-array antenna. One way is to properly design the T- or pi-type junctions of the feed in each slot-array branch [3] or to properly use short-pins [2]. Another way is to properly design the slot arrays to have appropriate input impedance as in this study. As shown in Fig. 3, the minimum normalized resonant conductance without the cover structure is about 0.135 at a slot offset of $x_o = 0.4$ mm because of the relatively large slot width, which causes a low resistance and consequently a low radiation efficiency.

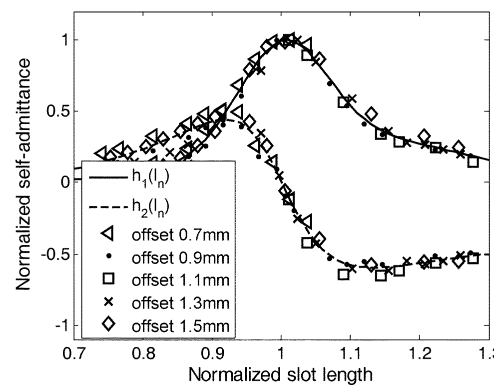

Fig. 4. Normalized self-admittance versus normalized slot length for the uncovered single-slot sets.

We propose a cover structure consisting of a dielectric slab and a conducting patch, as shown in Fig. 1, to reduce the resonant conductance of the antenna instead of designing the complicated feeder structure. In addition to the performance improvement, the cover structure blocks the slotted holes from rainfalls and other intrusion. The slim-look antenna without an additional radome will work at certain FWA systems.

The length of the conducting patch (l_p) was selected to be equal to the length of the waveguide slot (l_s) in order to avoid higher mode radiation and to reduce the complexity of the antenna simulation. The conducting patch, which has a width $w_p = 2$ mm, is aligned so that its center is located directly over the longitudinal side of the radiating slot in order to get an optimum field distribution between the slot and the conducting patch. The characteristic functions of the covered slot set, including the normalized self-admittance, the resonant slot length, and the normalized resonant conductance, were obtained using the procedure shown in the previous section as in Figs. 2 and 3 (lines with squares).

We were not able to select the proper resonant slot length and resonant conductance for offsets less than 0.9 mm from the simulation of the covered slot set. The resonant slot lengths of the covered slot are much smaller than those of the uncovered slot, as shown in Fig. 2, because of the effect of the dielectric

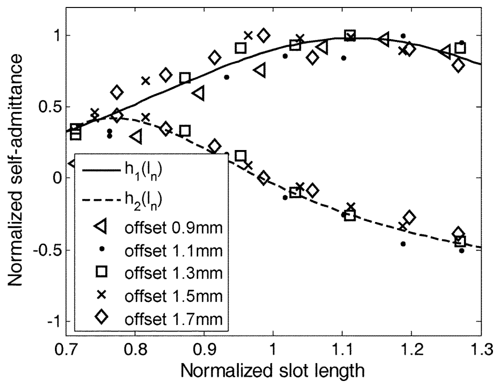

Fig. 5. Normalized self-admittance *vs.* normalized slot length for the dielectric-covered single-slot sets.

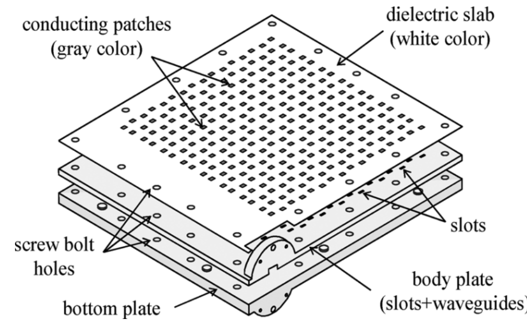

Fig. 6. Geometry of the 16 × 16 slot-array antenna with the cover structure.

slab. The normalized resonant conductance of the covered slot set is much lower than that of the uncovered slot set, as shown in Fig. 3. The normalized resonant conductance for a 0.9-mm slot offset is only about 0.037, which is low enough for a 16×1 slot array. That is, it is less than $0.0625\ (= 1/16)$. The resonant slot length (l_r) and the normalized resonant conductance (g_r) of the covered slot set were data-fitted as functions of the slot offset (x_o) over the range, 0.9 mm $< x_o <$ 1.7 mm, with the third- and second-order polynomials as

$$l_r(x_o) = 1.5926x_o^3 - 6.887x_o^2 + 10.852x_o - 2.8155 \quad (2)$$
$$g_r(x_o) = -0.09026x_o^2 + 0.4109x_o - 0.2606. \quad (3)$$

Fig. 5 shows the renormalized self-admittance (y_n) for the slot set with the cover structure, which can be data-fitted as

$$y_n = \frac{y(l_n)}{g_r} = g_n(l_n) + jb_n(l_n) \quad (4)$$

where

$$g_n(l_n) = 0.9825\ \exp\left[-\left(\frac{l_n - 1.118}{0.3975}\right)^2\right]$$
$$b_n(l_n) = \frac{-0.9302l_n^2 + 1.485l_n - 0.5607}{l_n^2 - 1.42l_n + 0.5744}. \quad (5)$$

As the slot length changes, the renormalized conductance (g_n) and susceptance (b_n) of the covered slot set vary more slowly than those of the uncovered slot, which makes it easy to determine an initial value for designing the slot arrays. An appropriate normalized self-admittance value of the covered single slot in a slot-array antenna can be calculated for given values of a slot offset and a slot length by using the characteristic functions given in (2)–(5).

III. 16 × 16 Slot-Array Antenna

The waveguide 16×16 slot-array antenna of this letter has two compartments: a radiation part and a power-dividing part. The radiation part consists of 16 16×1 slot arrays, as shown in Fig. 6, and the power-dividing part consists of a waveguide input port and 16 T-junctions for the alternating-phase feed. For this antenna, the 20-dB Chebyshev power distribution is applied in designing both the 16×1 slot arrays and the 1×16 power dividers.

We first searched the slot offset from (3) for the necessary normalized resonant conductance (g_r) with the Chebyshev power ratios. Then, working from (2), we used the calculated offset value to seek the resonant slot length (l_r). The acquired slot offset and slot length became the initial values of the 16×1 array antenna. Therefore, we were able to accurately determine the optimum slot sizes for this 16×1 slot array without any tedious tuning procedure. The optimum slot offsets and lengths of the entire structure of the 16×1 slot array can be verified using a full-wave analysis.

Next, we used the waveguide T-junction to construct the 1×16 power divider. The alternating-phase feed system was employed in this power divider to minimize the mutual coupling and the power leakage between adjacent radiating slots [3], [4]. The input and output ports of each T-junction were designed as perfectly matched ports with a reflection less than -30 dB, while the numerical simulation was used to design the branch ports with Chebyshev distribution ratios within a tolerance of ± 0.2 dB.

With the 16 16×1 slot arrays and the 1×16 power divider, we optimized the performance of the 16×16 array using a circuit analysis with the MWoffice instead of the time-consuming full-wave analysis tool because we had an appropriate initial design based on the design procedure in the previous section. We obtained the optimum performance of the array antenna with slight adjustments for the distance between the last slot and the short-ended wall and the distance between the first slot and the branch port of the power divider. The optimized 16×16 slot-array antenna, as shown in Fig. 6, was manufactured using a computer-numerical-controlled (CNC) milling technique. The dielectric slab and the arrayed conductor patches were fabricated by etching a printed circuit board with RT/Duroid 5880. The exterior dimensions of this antenna were $123.5 \times 139.2 \times 8.8\ \text{mm}^3$, and it had an input flange for a WR22 ($5.7 \times 2.8\ \text{mm}^2$) waveguide connection.

We used a vector network analyzer HP8510C to measure the return loss of the slot array antenna, and we used a computer-controlled positioning and data acquisition system to measure the radiation patterns of the antenna with a 1° resolution [11]. Fig. 7 shows the measured return loss and the radiation efficiency of the slot array antenna with the cover structure in the frequency range of 37.6 $<$ f $<$ 38.6 GHz. The measured return loss at 38 GHz was about 30 dB, and the bandwidth for $|S_{11}| < -10$ dB was about 500 MHz. The radiation efficiency using the wheeler-cap method [12] was about 71%.

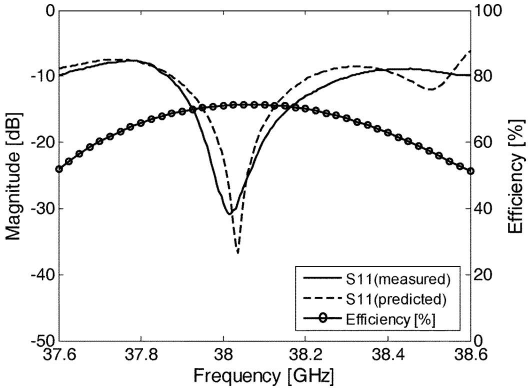

Fig. 7. (left-side) Measured $|S_{11}|$ and (right-side) radiation efficiency of the 16×16 slot-array antenna with the cover structure.

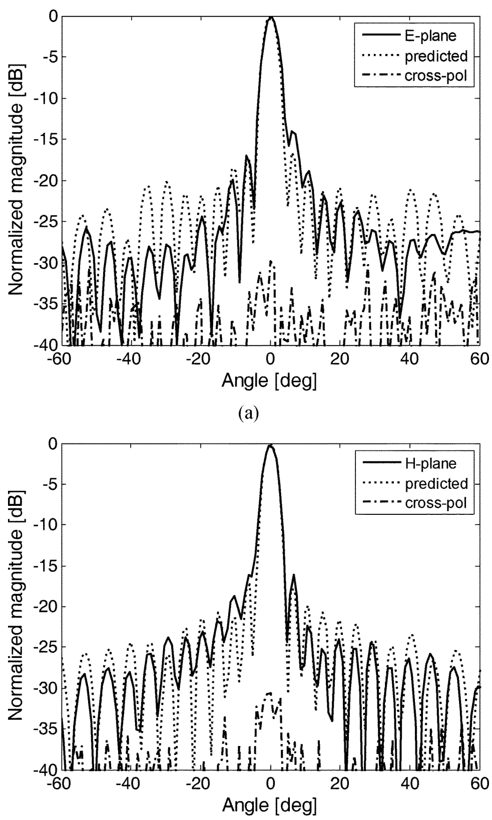

Fig. 8. Measured radiation patterns of the 16×16 slot-array antenna with the cover structure: (a) principal E-plane and (b) principal H-plane.

Fig. 8(a)-(b) show the measured radiation patterns along the principal E- and H-planes. The half-power beamwidth (HPBW) of the measured E- and H-plane patterns was 4.4° and 4.1°, respectively, which agrees well with the theoretically predicted HPBW of 4.2°.

In case of the E-plane pattern, the alternating-phase-fed arrays influenced the degradation of the antenna pattern (HPBW of 4.4° or sidelobe levels) because, as shown in Fig. 6, the radiating slot sets were not evenly placed in the horizontal direction. The relative power level of the maximum first sidelobes on the E- and H-plane patterns were about -14.1 (right side) and -16.1 dB (left side), respectively, and most of others were below -20 dB. The depolarization levels were below -30 dB in both radiation patterns. The measured antenna gain was 30.8 dBi.

IV. Conclusion

We have proposed a high-performance waveguide, 16×16 slot-array antenna that is covered with a dielectric slab and 16×16-array conductor patches. The cover structure leads to high performance from the slot-array antenna because it lowers the resonant conductance of the slot sets. We have also provided an initial design rule that makes it easier to design the slot-array antenna. The antenna performance predicted in our simulation agrees well with the measured results. The manufactured antenna has demonstrated a gain of 30.8 dBi; an efficiency of 71%; HPBWs of 4.4° and 4.1° on the principal E- and H-planes, respectively; and depolarization levels below -30 dB at 38 GHz. The proposed antenna will be cost-effective because its wide slot width allows it to be easily manufactured with a common CNC milling technique, and it does not need welding for the slot plane. The proposed antenna can be used in an outdoor environment because of the dielectric cover.

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