antenna is about 0.95λg. It is also noted that λg is the wavelength in the slot.

Figure 2 shows the measured return loss for the proposed dual-frequency slot antenna. For comparison, the calculated (by IE3D™ [5]) return loss is also expressed in Figure 2. From the results, the 10-dB return-loss impedance bandwidths of the two resonant frequency are about 1.6% (1595–1620 MHz) and 17% (2159–2567 MHz), respectively. Also, the 10-dB return-loss impedance bandwidth of the first resonant frequency excited by the slot loop antenna is very small, which is inconvenient for some applications. Thus, the broadband operation is especially important for this antenna.

Figure 3 shows the measured and simulated return losses for a constructed prototype operating within the broadband range. It is clearly seen that for the proposed broadband antenna, large operating bandwidths formed by the excitation of two adjacent resonant modes at 1974 (about 0.91λg) and 2426 MHz (about 0.98λg) is observed. And, the impedance bandwidth is 797 MHz, or about 35% with respect to the center frequency at 2307 MHz. Radiation characteristics for the prototype are also measured and plotted in Figure 4. These two resonant modes are seen to be of the same polarization planes and similar radiation patterns. Measured antenna gain for frequencies within the impedance bandwidth is presented in Figure 5. The antenna gain variation is less than 1.5 dBi, with the peak antenna gain at about 6.4 dBi.

4. CONCLUSIONS

A novel, simple design of a CPW-fed hybrid coupled slot antenna with dual-frequency or broadband operation has been proposed and experimentally studied. From the results, the first and second frequencies can be easily controlled by the dimensions of the slot loop antenna and the dipole slot antenna, respectively. The operating frequencies of the proposed antenna are of same polarization planes and have similar radiation patterns. The broadband operation is obtained by varying the two resonant frequencies close to each other. The proposed design’s broadband antenna bandwidth is about 35%.

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ABSTRACT: A multi-pole gap-coupled unidirectional dielectric radiator (UDR) is proposed for broadband applications in the millimeter-wave band. The advantages of gap-coupled UDR include ease of planar integration, low loss, and low cost. Design parameters of the gap-coupled UDR, such as resonator length and gap distance, are determined using an equivalent circuit model of an evanescent nonradiative dielectric (NRD) guide. Prototypes of the gap-coupled UDR are designed and measured at 38 GHz. The measured results show good agreement with simulated data. © 2003 Wiley Periodicals, Inc. Microwave Opt Technol Lett 38: 498–501, 2003; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.11102

Key words: unidirectional dielectric radiator (UDR); gap-coupled antenna; nonradiative dielectric (NRD) guide; equivalent circuit model

1. INTRODUCTION

The development of broadband, low-loss, low-cost antennas at millimeter-wave frequencies for ultra-fast communication systems is extremely important. The previously reported unidirectional dielectric radiator (UDR) [1], which utilizes the nonradiative dielectric (NRD) waveguide [2] structure, has the advantage of low cost, low loss, and ease of fabrication at millimeter-wave frequencies. However, it suffers from narrow bandwidth because it utilizes a high-Q resonant element. Furthermore, the feeding methods for the previously reported UDR are aperture coupling [3] and excitation using coaxial cable [1]. These feeding methods may introduce additional transition loss and are not suitable for multi-beam feed, such as NRD Rotman lens [4].

In this paper, we propose a multi-pole gap-coupled UDR that has the merits of broadband, low loss, and low cost at millimeter-wave frequencies. In addition, gap-coupled UDR is suitable for an integrated end-front using an NRD guide. For instance, gap-coupled UDR can be applied to the antenna of an NRD Rotman lens of a multi-beam feed without any additional transition. A gap-coupled UDR is concisely designed with an equivalent circuit model. This equivalent circuit model employs an evanescent waveguide K-inverter circuit, incorporating the fact that the structure of a gap-coupled UDR is similar to that of an NRD gap-coupled filter. Design parameters of a gap-coupled UDR, such as resonator length and a gap distance, are determined using the equivalent circuit model. The position of the resonator from the conductor end plate is further optimized using the finite-element
A good union between the antenna and NRD guide. Coupled UDR’s planar integration will not be difficult due to the size. From a structural viewpoint, the proposed multi-pole gap-resonator controls the matching characteristics of antenna. The position of a virtual dielectric strip is not the same as that of a UDR, the condition, a virtual dielectric strip is inserted, as shown in Figure 2(a). If a UDR is well matched, all input power will go similar to that of an NRD gap-coupled filter. Figure 2(a) shows a model employs an evanescent waveguide method (FEM) commercial software Ansoft-HFSS. Experimental prototypes, one-pole and three-pole UDR, are designed and measured at the operation frequency of 38 GHz. For experimental launching of the LSM11 mode in the NRD guide feed, the waveguide-to-NRD guide transition has been also fabricated in the Q band [4]. The measured return loss and far-field radiation patterns of both one-pole and three-pole gap-coupled UDR are discussed.

2. SCHEMATIC OF MULTI-POLE GAP-COUPLED UDR

A conceptual structure of a multi-pole gap-coupled UDR is depicted in Figure 1. The overall UDR is composed of dielectric resonators, gaps, and an NRD guide of the same thickness. The dielectric resonator and NRD guide are sandwiched by two metal plates. As illustrated in Figure 1, resonator length, gap distance, and the resonator’s position are briefly denoted as \( R_n \), \( G_n \), and \( P \) in our study, respectively. The length of resonator \( R_n \) determines the operation frequency of antenna. The distance of gap \( G_n \) controls the matching characteristics of antenna. The position of a resonator \( P \) is especially related to an effective antenna aperture size. From a structural viewpoint, the proposed multi-pole gap-coupled UDR’s planar integration will not be difficult due to the good union between the antenna and NRD guide.

3. EQUIVALENT CIRCUIT MODEL OF A GAP-COUPLED UDR

An equivalent circuit model of a gap-coupled UDR is developed in this paper. The model employs an evanescent waveguide K-inverter circuit because the structure of a gap-coupled UDR is similar to that of an NRD gap-coupled filter. Figure 2(a) shows a schematic diagram of multi-pole gap-coupled UDR with a virtual dielectric. If a UDR is well matched, all input power will go through air without any reflection. To model the above matched condition, a virtual dielectric strip is inserted, as shown in Figure 2(a). Then, a gap-coupled UDR can be considered as a gap-coupled NRD filter. However, because the impedance of an inserted virtual dielectric strip is not the same as that of a UDR, the position \( P \) is further optimized using Ansoft-HFSS to complete a perfect impedance matching.

Since distance of two parallel conductor plates is smaller than half a wavelength in free space [2], the height and width of an NRD guide are calculated using the following equations:

\[
a / \lambda_0 = 0.45, \quad \text{(1a)}
\]

\[
(b / \lambda_0) \sqrt{\varepsilon_r - 1} = 0.4 \sim 0.6, \quad \text{(1b)}
\]

where \( a, b \), and \( \varepsilon_r \) are dielectric height, dielectric width, and relative dielectric constant, respectively.

The air gap region in Figure 2(a) can be represented by an impedance inverter \( K \) of an evanescent waveguide since the NRD guide has a similar structure with a dielectric-filled metal waveguide. The resulting impedance inverter \( K \) constructed by a gap is shown in Figure 2(a). Then, the values of \( K \) and \( \phi \) are given by Eqs. (2), (3), and (4) [5]:

\[
K = \tan \left( -\frac{1}{2} \tan^{-1} \left( \frac{2X_0}{Z_g} + \frac{X_0}{Z_g} \right) + \frac{1}{2} \tan^{-1} \left( \frac{X_0}{Z_g} \right) \right), \quad \text{(2)}
\]

\[
\phi = -\pi + \tan^{-1} \left( \frac{2X_0}{Z_g} + \frac{X_0}{Z_g} \right) + \tan^{-1} \left( \frac{X_0}{Z_g} \right), \quad \text{(3)}
\]

where \( Z_g = \omega_\mu / \alpha \) and \( Z_\alpha = \omega_\mu / \beta \);

\[
\frac{K_{ji}}{Z_0} = \frac{\pi \omega_\alpha}{2 g_0 g_i \omega_i}, \quad \text{(4a)}
\]

\[
\frac{K_{j, j+1}}{Z_0} = \frac{\pi \omega_\alpha}{2 \omega_i \sqrt{g_{s,j+1} g_j}}, \quad \text{(4b)}
\]

\[
\lambda_s = \frac{\lambda_s + \lambda_{s1}}{2} \quad \text{(4c)}
\]

where \( g_0, g_1, \ldots, g_{n+1} = \) element values of low pass filter, \( \omega_\alpha = \) normalized cutoff frequency, \( \omega_0 = \) center frequency of filter, \( \omega_1 \) and \( \omega_2 = \) respective lower and upper passband frequencies, \( \lambda_0 = \) free-space wavelength, \( \lambda_{s1}, \lambda_s, \lambda_{s2} = \) guide wavelength at frequencies \( \omega_0, \omega_1, \omega_2 \).

By comparing Eq. (2) with Eq. (4), gap distance \( G \) can be calculated. Resonator length \( R \) is also calculated using Eqs. (3) and (5):

\[
R = \frac{[ \pi + 0.5(\phi_1 + \phi_2)]}{\beta_0}. \quad \text{(5)}
\]

Table 1 lists parameters of one-pole and three-pole gap-coupled UDR calculated by the above procedure. Note that \( G_{11} \) is changed to \( P \), which is the resonator’s position from the conductor end plate, as shown in Figure 2(a).

### TABLE 1  Designed Parameters of One-Pole and Three-Pole Gap-Coupled UDR

<table>
<thead>
<tr>
<th></th>
<th>( G_2 ) (mm)</th>
<th>( G_3 ) (mm)</th>
<th>( G_4 ) (mm)</th>
<th>( R_1 ) (mm)</th>
<th>( R_2 ) (mm)</th>
<th>( R_1 ) (mm)</th>
<th>( P ) (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>One-pole UDR ( a = 3.55 ; \text{mm}, ; b = 4.175 ; \text{mm} )</td>
<td>4.59</td>
<td>—</td>
<td>—</td>
<td>2.89</td>
<td>—</td>
<td>—</td>
<td>3.6</td>
</tr>
<tr>
<td>Three-pole UDR ( a = 3.55 ; \text{mm}, ; b = 4.175 ; \text{mm} )</td>
<td>2.37</td>
<td>4.9</td>
<td>4.9</td>
<td>2.92</td>
<td>2.87</td>
<td>2.92</td>
<td>1.5</td>
</tr>
</tbody>
</table>

Operation frequency = 38 GHz, relative dielectric constant = 2.08.
4. MEASUREMENT RESULTS AND DISCUSSION

Prototypes of both one-pole and three-pole gap-coupled UDR, with Q-band waveguide-to-NRD-guide transition for excitation of the LSM$_{11}$ mode, were fabricated in order to compare their bandwidth and far-field radiation patterns. The principle of waveguide-to-NRD-guide transition can be described as follows. A dielectric-filled waveguide operating at TE$_{10}$ mode is used as the input port of the transition. The electric field parallel to the flared parallel plate is then launched to the UDR’s input port as a cross section of the rectangular waveguide rotates 90°. Thus, the parallel electric field is converted with the LSM$_{11}$ mode at the UDR’s input port, since the electric field of the LSM$_{11}$ mode is parallel to the NRD guide’s conducting plate. To measure the characteristic of the proposed UDR, we utilized a 25-dBi Q-band standard horn antenna and vector network analyzer (HP8510C). The measurement of far-field radiation patterns was performed with 5° steps over $-90°$ to $90°$.

In the case of one-pole gap-coupled UDR, a return loss of about $-30$ dB is measured at 38.6 GHz. However, it is observed that the magnitude of $S_{11}$ has some noisy pattern, as shown in Figure 3. This phenomenon is considered to be due to leakage loss resulting from some imperfection during fabrication and contact between the dielectric resonator and parallel plate. The half-power beam width (HPBW) is measured to be $30°$ in the $E$-plane pattern and $60°$ in the $H$-plane pattern, respectively. Measured results show good agreement with those of FEM simulation, as illustrated in Figures 3(a), 4(a), and 5(a). Also, isolation between co-polar and cross-polar is measured at 25 dB or more in both $E$-plane and $H$-plane patterns. Thus, we know that UDR has a good linear polarization and no side-lobe. An absolute gain of 13.1 dBi has been obtained by comparing the difference when using a Q-band standard horn antenna.

The characteristics of the three-pole gap-coupled UDR are measured using the same method used with the one-pole gap-coupled UDR. As shown in Figure 3(b), the bandwidth of the three-pole UDR is much broader, as expected, than that of one-pole UDR. The HPBW is measured to be $30°$ in the $E$-plane patterns and $70°$ in the $H$-plane pattern, as shown in Figures 4(b) and 5(b), respectively. The measured isolation between co-polar/ cross-polar and absolute gain are 25 dB or more and about 13 dBi, respectively. The characteristics of the three-pole UDR are almost the same as the characteristics of the one-pole UDR, except for broader bandwidth.
5. CONCLUSION

A multi-pole gap-coupled UDR has been proposed for ultra-fast communication systems in the millimeter-wave frequency. The dimensions of UDR, such as resonator length and gap distance, were determined using an equivalent circuit model of an evanescent nonradiative dielectric (NRD) guide. This procedure considerably reduced the time to design UDR. The resonator’s position from the conductor end plate is optimized with an FEM simulator for perfect matching. The experimental results agree well with those of the simulation. In particular, multi-pole UDR has much broader bandwidth than that of one-pole UDR, indicating that it is suitable for ultra-fast communication systems.

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A FOAM-BASE SURFACE-MOUNTABLE SHORTED MONOPOLE ANTENNA FOR WLAN APPLICATION

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ABSTRACT: A surface-mountable shorted monopole antenna, constructed by folding a branch-line metal strip onto a foam base for 2.4/5.2/5.8-GHz WLAN operation, is presented. The branch-line metal strip is comprised of a longer arm, a shorter arm, and a matching arm. The longer and shorter arms generate, respectively, a lower resonant mode covering the 2.4-GHz band (2400 –2484 MHz) and an upper resonant mode covering the 5.2-GHz (5150 –5350 MHz) and 5.8-GHz (5725–5875 MHz) bands, and good impedance matching for frequencies across the two resonant modes is easily achieved by adjusting the distance between the feeding point and the shorting point in the matching arm. Experimental results of a constructed prototype of the proposed antenna are presented. © 2003 Wiley Periodicals, Inc. Microwave Opt Technol Lett 38: 501–503, 2003; Published online in Wiley InterScience (wwwinterscience.wiley.com). DOI 10.1002/mop.11103

Key words: antennas; surface-mountable antennas; shorted monopole antennas; dual-frequency antennas; WLAN antennas

1. INTRODUCTION

Surface-mountable antennas with a foam or air base, instead of a ceramic base [1], have recently been demonstrated [2]. Such foam-base surface-mountable antennas are generally formed by folding a metal strip or metal line onto a foam base, and can be constructed at low cost. In addition, this kind of foam-base antenna in general will not break, which is an advantage over the ceramic chip antenna.

For the foam-base antenna design shown in [2], the antenna is mainly operated as a monopole structure, and an isolation distance between the antenna’s major radiating portion and the system ground plane is required. This isolation distance has a great effect on the impedance matching of the antenna and, in addition, it will increase the total antenna height from the system ground plane. In this paper, we present a new design of a foam-base surface-mountable shorted monopole antenna suitable for wireless local area network (WLAN) operation in the 2.4/5.2/5.8-GHz bands. In

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Figure 5 Simulated and measured H-plane patterns of gap-coupled UDR: (a) one-pole; (b) three-pole

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