

A BROADBAND BEAM-FIXED PLANAR-ARRAY ANTENNA USING SLOT PAIR AND CONDUCTIVE BAR AT MILLIMETER-WAVE FREQUENCIES

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Received 22 September 2003

ABSTRACT: This paper presents a broadband millimeter-wave planar-array antenna that uses a slot pair and crossing conductive bar. The antenna is fed by a waveguide feed network in the E-plane to reduce the feed-line loss and a microstrip-feed network in the H-plane in order to decrease fabrication costs. The waveguide and microstrip are coupled through the slot pair, which is placed a quarter guided-wavelength distance apart, so that the reflected waves from the slots cancel each other. As a result, the antenna exhibits very broadband characteristics. The conductive bar is laid above the slots to increase the coupling. Results from simulation and measurement of the fabricated antenna show the reflection bandwidth to be more than 7.1% of the 40.5–43.5-GHz frequency range. The antenna also has fixed beams, low side-lobe levels (SLLs) at all frequencies of interest, and a high gain of 28.1 dBi at 41.5 GHz. © 2004 Wiley Periodicals, Inc. *Microwave Opt Technol Lett* 41: 150–154, 2004; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.20076

Key words: planar antenna; slot; microstrip; waveguide; feed; millimeter-wave

1. INTRODUCTION

Due to the growth of communications in the millimeter-wave-length range, the demand for low-cost, broadband, planar-array antennas has increased. Among the many types of millimeter-wave antennas available, a microstrip patch antenna fed only with a microstrip-feed network has the lowest fabrication cost. However, in millimeter-wave applications, it also has significant feed-line loss [1], and thus is inappropriate for use in large array designs. In contrast, a slotted-waveguide array antenna, which uses slots with a waveguide feed network, has low losses but is expensive to fabricate [2]. A compromise between feed-line loss and fabrication cost is an antenna that utilizes both the waveguide- and microstrip-feed network [3, 4]. However, the standing-wave feed network in [3, 4] results in a very narrow bandwidth and the frequency-dependent main-beam direction.

This paper presents a beam-fixed and broadband 30×30 array antenna that operates in the 40.5–43.5 GHz frequency range. A waveguide-feed network is used in the E-plane direction to reduce feed-line loss, and a microstrip-feed network is

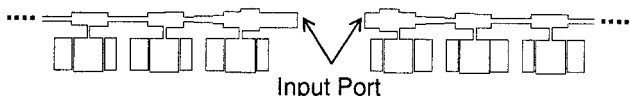
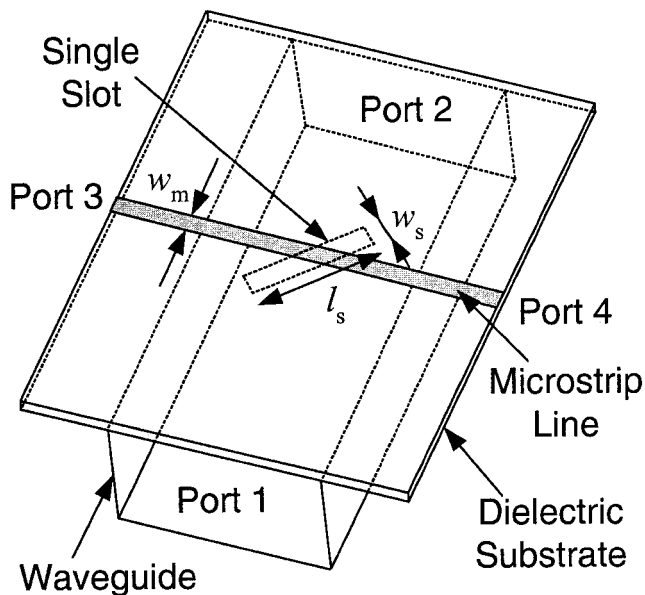
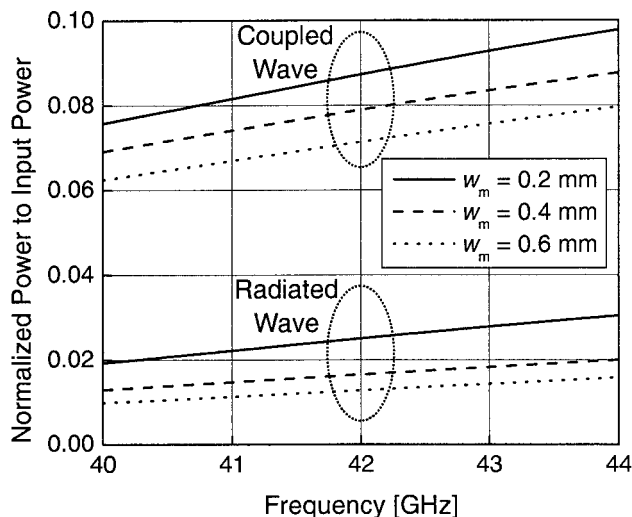


Figure 1 Part of the simulation model of the microstrip-feed network with patch element



(a)



(b)

Figure 2 Single slot-to-microstrip transition: (a) geometry; (b) normalized power of coupled and radiated wave to input power for various microstrip line width w_m

distributed in the H-plane direction to reduce the fabrication cost. The two feed networks are coupled through a center-inclined slot pair that both cancels the reflected wave and increases the coupled wave. For an additional increase in coupling, a narrow, thin conductive bar is installed across and above the slot.

Section 2 describes the antenna elements and the feed network. In section 3, we introduce our proposed slot pair and crossed conductive-bar scheme. Section 4 discusses and compares the performance differences noted between simulated and measured results. The paper ends with a short summary of our conclusions.

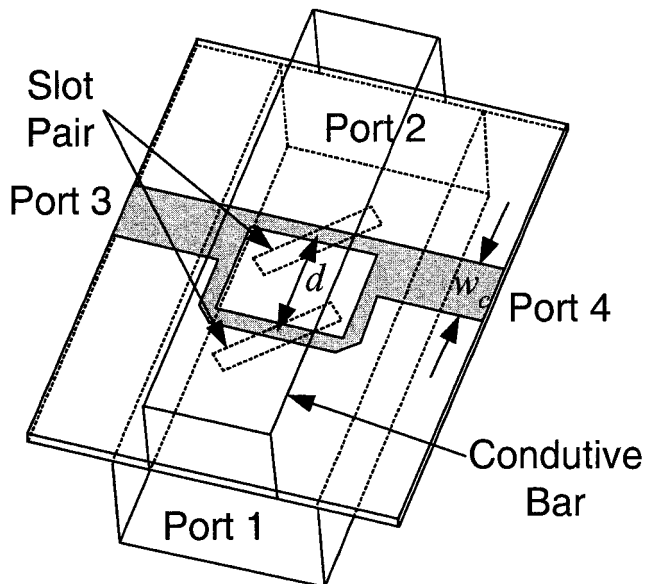


Figure 3 Proposed slot-to-microstrip transition with a slot pair and a conductive bar

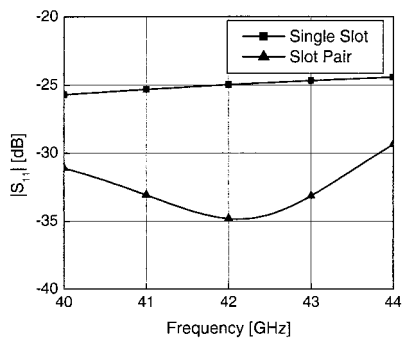


Figure 4 Magnitude of return loss S_{11} , of a single slot and a slot pair

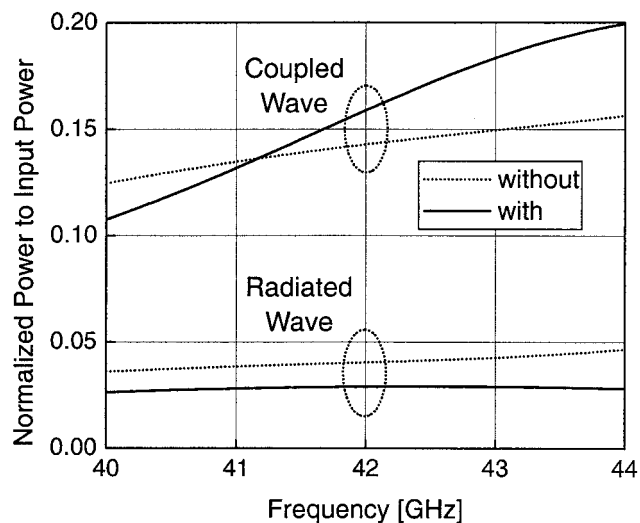


Figure 5 Normalized power of coupled and radiated waves to input power for the slot pair with or without a conductive bar

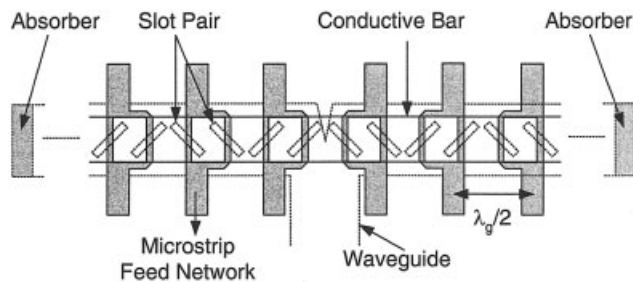


Figure 6 Part of the waveguide-feed network around the T-junction

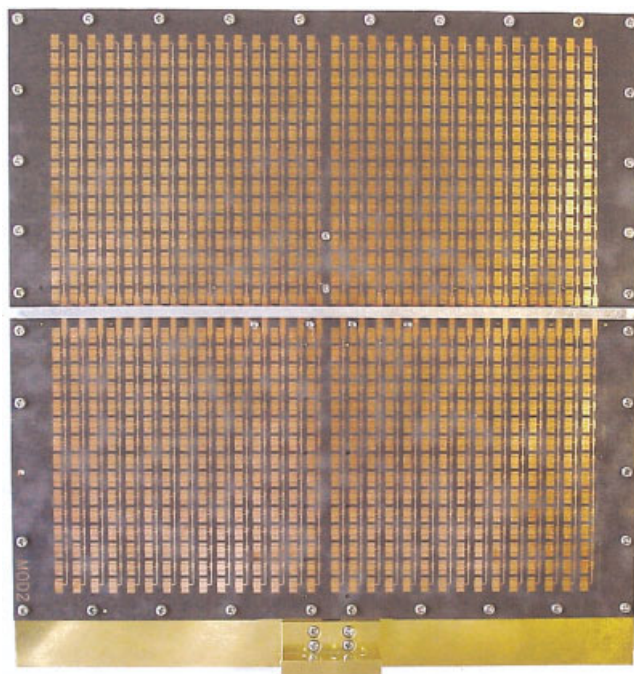


Figure 7 Photograph of the fabricated 30×30 array antenna. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com.]

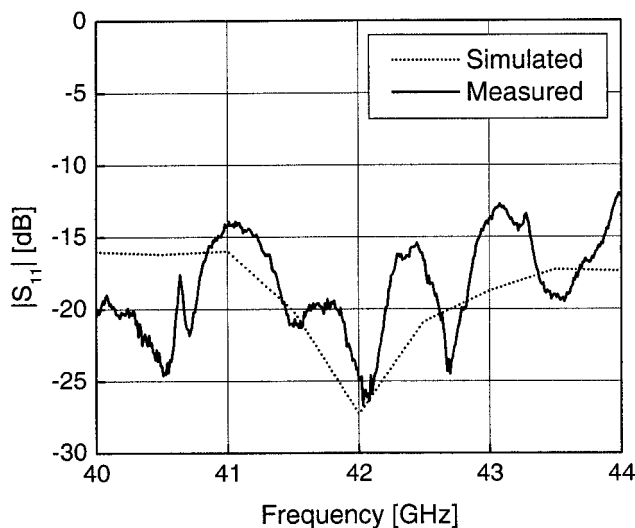


Figure 8 Magnitude of simulated and measured return losses S_{11} , of the array antenna

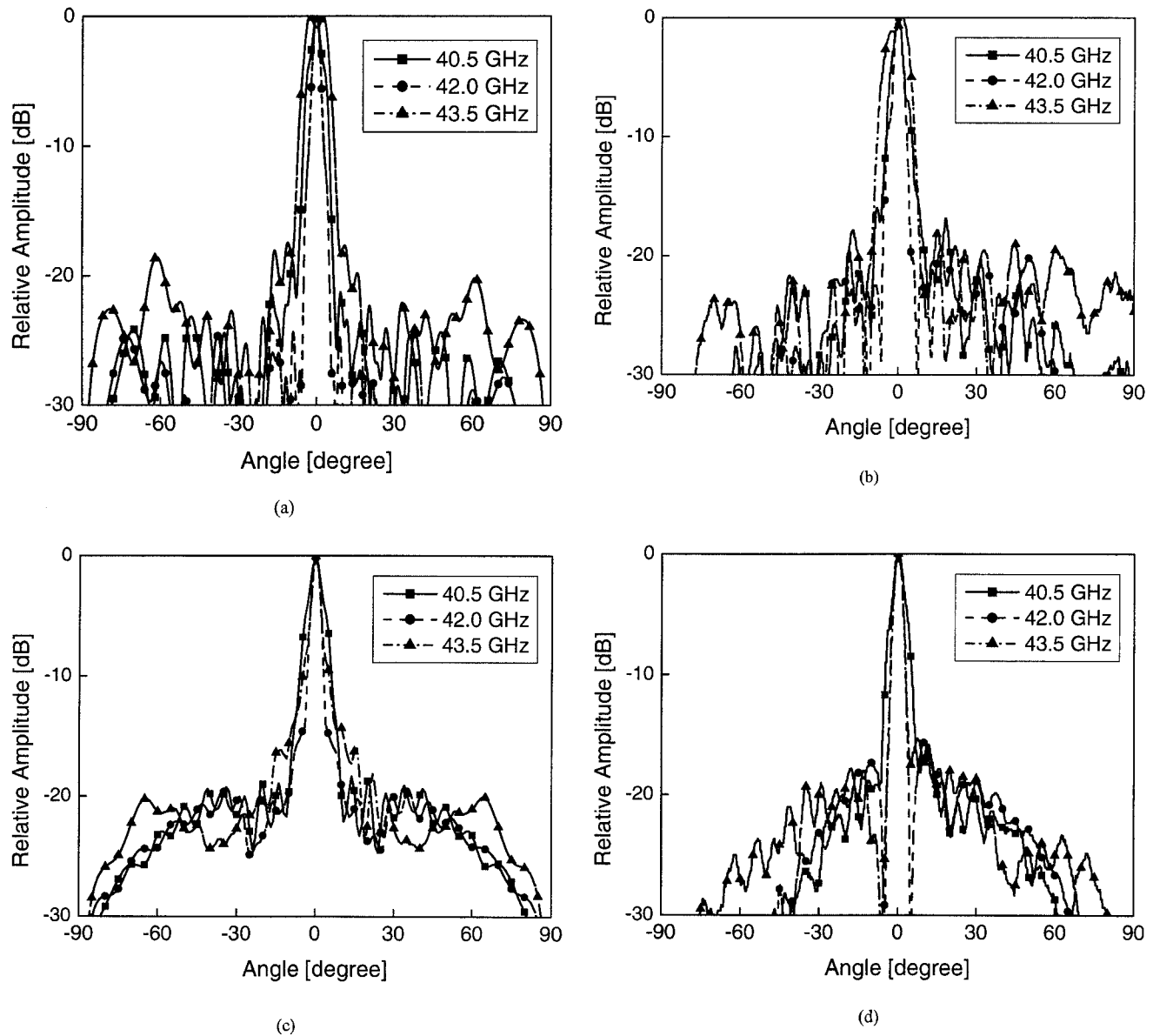


Figure 9 (a) Simulated and (b) measured E-plane radiation patterns; (c) simulated and (d) measured H-plane radiation patterns

2. MICROSTRIP-FEED NETWORK WITH ANTENNA ELEMENT

The microstrip patch element with the parasitic patch near the nonradiating edge described in [5] is adopted as an antenna element. The dielectric constant of the substrate is 2.2 and the height is 10 mil. Analysis of the performance of this element by the method of moments (MoM) tool IE3D [6] shows a bandwidth of 3.0 GHz ($VSWR \leq 2.1$) and well shaped radiation patterns.

The antenna is fed by a waveguide and microstrip-feed network. The microstrip-feed network is designed to have -30 dB of SLL by use of the Dolph–Chebyshev method [7]. Figure 1 shows the part of the simulation model that contains the microstrip-feed network with the patch element. It is not possible to simulate the entire patch and microstrip-feed network simultaneously, so only the 1×30 array antenna has been analyzed and the H-plane radiation pattern determined [7]. The spacing between elements is 5.1 mm. However, the central element spacing is 10.5 mm due to the placement of the inclined slot. This broad spacing increases the side-lobe level (SLL) of the H-plane radiation pattern, which will be depicted in section 4.

3. WAVEGUIDE-FEED NETWORK WITH SLOT PAIR AND CONDUCTIVE BAR

To have a broad reflection bandwidth, the traveling-type feed network is adopted. Figure 2(a) shows the geometry of slot-to-microstrip transition. The waveguide network is milled and placed below the dielectric substrate. Thus, the ground plane of the substrate serves as the upper conductor plate of the waveguide. The center-inclined slot is etched on the ground plane of the substrate. The slot length l_s and width w_s are 2.8 and 0.5 mm, respectively. In Figure 2(b), as the width of the microstrip line w_m decreases from 0.6 to 0.2 mm, the power of the coupled wave increases and, at the same time, that of the undesirable radiated wave also increases. Based on those results, w_m is determined to be 0.4 mm. The analysis of this configuration has been performed with the aid of Ansoft's finite-element method simulator, HFSS [8]. It is well known that the magnitudes of S_{31} and S_{41} are almost equal, and their phases are about 180° out of phase [2]. To have sufficient spacing between the antenna elements in the E-plane direction, the width of the broad wall of the standard WR-22 waveguide is reduced to 4.8 mm, which results in an interelement

spacing, 5.3 mm of a half-guided wavelength. However, if several slots are positioned a half-guided-wavelength distance $\lambda_g/2$ apart [see Fig. 2(a)], the reflected waves from each slot accumulate and a high reflection coefficient at the input port results. To overcome that inherent characteristic, we propose the slot pair described in the following paragraphs.

Figure 3 illustrates the proposed slot-to-microstrip transition in this antenna design, which consists of two inclined slots, a conductive bar, and the microstrip line. The incident wave from port 1 couples through the slot pair and then propagates toward ports 3 and 4, which are connected to the microstrip feed network, as shown in Figure 1. The slot length, slot width, and width of coupling microstrip line are 2.8, 0.5, and 0.4 mm, respectively, which are equal to those of Figure 2(a). The width of combining microstrip line w_c is adjusted to have maximum coupling and determined to be 1.5 mm. The distance between the two slots d is $\lambda_g/4$ (2.65 mm), which results in the reflected waves from the two slots being offset. This achieves a very low reflection coefficient. Figure 4 shows the return loss of a single slot and the slot pair with $d = 2.65$ mm. The return loss of slot pair is about -35 dB at 42.0 GHz, which is much lower than the -25 dB obtained with the single slot. Another advantage of the slot pair is that the coupled wave increases. This can be easily found by comparing the coupled power of the slot pair, as shown in Figure 5, to that of Figure 2(b) of the single slot in Figure 2(b).

To increase the amount of wave coupling, we place a grounded narrow and thin conductive bar 2.0-mm above the slot pair. The width of the bar is 3.0 mm and the thickness is 0.5 mm. The conductive bar functions as a reflector that redirects undesirable radiating waves to the microstrip line. To verify the conductive bar, the coupled and undesirable radiated powers without and with a conductive bar are examined, as plotted in Figure 5. As expected, the coupled wave of slot pair with a bar is larger than that of slot pair without a bar. The slot pair with a bar radiates a smaller power than that without a bar. Although this difference is small, it becomes large in a large array with a number of slot pairs.

The geometry of the waveguide-feed network with the slot pair and conductive bar is shown (in part) in Figure 6. We adopted a center-fed feed network to achieve the beam-fixed characteristics [5]. The waveguide is divided into two paths by using a T-junction. A T-junction with a triangular ridge, which provides a low reflection coefficient, divides the incident power into two branch waveguides. Two ends of waveguide are terminated with the absorbing material to implement a traveling-type feed network. The proposed slot pairs are placed a distance of a half-guided-wavelength $\lambda_g/2$ apart (5.3 mm) and their orientations are rotated alternately to achieve in-phase coupling. The 15 slot pairs are placed along one waveguide-feed network and thus the total number of slot pairs becomes 30.

4. RESULTS

Figure 7 shows a photograph of a fabricated 30×30 array antenna. The substrate is attached to the base plate and secured in place by a large number of metal screws. A narrow conductive bar is placed horizontally along the centerline of the antenna. The entire patch area is $16.0 \text{ cm} \times 16.0 \text{ cm}$. Figure 8 compares the magnitude of return loss S_{11} from simulation and measurements taken from the fabricated antenna. The simulated results are obtained from the antenna geometry shown in Figure 6. The simulation and actual performance results both have a broadband bandwidth of more than 3 GHz ($\text{VSWR} \leq 2.0$), which meets the design goal. The simulated and measured radiation patterns are plotted in Figure 9. The E-plane simulated patterns are calculated using the

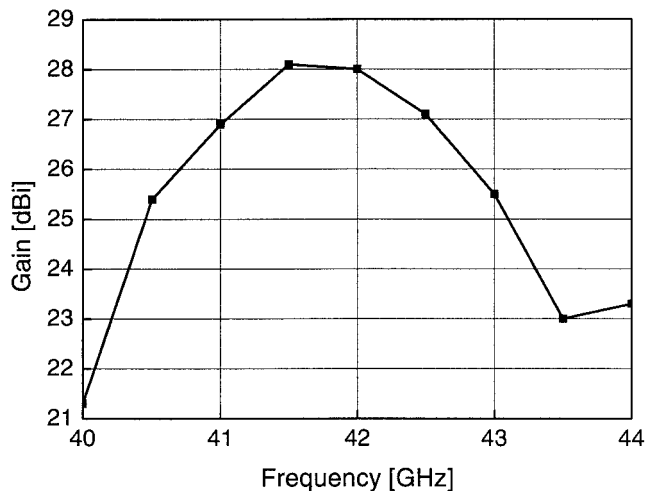


Figure 10 Measured antenna gain

magnitude and phase values obtained from simulation of the antenna geometry (Fig. 6) and from the antenna element pattern [7]. The main-beam directions are fixed for varying operating frequencies. The SLLs of the simulated and measured patterns are less than approximately -18 dB at all frequencies. The SLL of the H-plane network is designed to be -30 dB, but the simulated and measured values are only about -20 dB, due to broad spacing between the central two patches, fabrication tolerances, and mutual coupling between the antenna elements and the microstrip-feed network. At frequencies far from the center frequency, the beam widths of the E- and H-planes gradually broaden because of the center-feed configuration of both networks. The measured antenna gain is plotted in Figure 10. The highest gain was 28.1 dBi, at a frequency of 41.5 GHz. The gain at this frequency is very low as compared to the aperture gain. This is considered to be due mainly to the inherent conductor loss of the microstrip lines [1]. The imperfect electrical contact between the ground plate of the substrate and the waveguide may be another factor. This can be overcome by using a choke [3, 9].

5. CONCLUSION

This paper has presented the design of a 30×30 millimeter-wave planar array antenna. The antenna is fed by a microstrip and waveguide-feed network, which are combined through the use of a slot pair and conductive bar. The proposed slot-to-microstrip scheme provides a low reflection and a high coupling. The measured reflection bandwidth of array antenna is broad as over 3.0 GHz ($\text{VSWR} \leq 2.0$). The main beams at all design frequencies have a fixed direction. The SLLs at all design frequencies are below -18 dB in E-/H-plane. The measured maximum gain is about 28.1 dBi. The aperture antenna efficiency is low, but this could be improved by achieving a better electrical contact between the ground plate of the substrate and the waveguide, or by the use of a choke.

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HARMONIC SUPPRESSION BY PHOTONIC BANDGAP ON CPW-FED LOOP-SLOT ANTENNA

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Received 18 September 2003

ABSTRACT: A conventional CPW-fed loop-slot antenna, incorporating a PBG structure in the feeding network, is presented. Experimental results show that the PBG structure with cross-shaped lattices exhibits a well behaved band-stop characteristic; furthermore, the impedance bandwidth of the proposed antennas also becomes wider than that of the conventional CPW-fed slot antenna. In this design, the impedance bandwidth can reach 1320 MHz, which is about 60% with respect to the central frequency of 2200 MHz. © 2004 Wiley Periodicals, Inc. *Microwave Opt Technol Lett* 41: 154–156, 2004; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.20077

Key words: harmonic suppression; CPW-fed loop-slot antenna; photonic bandgap; broadband

1. INTRODUCTION

CPW-Fed slot antennas have been widely used for wireless applications because they are compatible with monolithic integrated circuits and active solid-state devices. In addition, CPW-fed slot antennas exhibit a large bandwidth with bidirectional radiation patterns [1]. A CPW-fed square-slot antenna with a tuning stub to enhance the impedance bandwidth was proposed in [2]. By adjusting the location of a widened tuning stub, good impedance matching can be easily obtained. It is seen that the larger the spacing between the ground plane and the widened tuning stub, the wider impedance bandwidth. However, the bandwidth of the higher-order mode also increases with that of the dominant mode, which is a potential problem of electromagnetic interference and compatibility.

To alleviate the serious symptoms of the conventional CPW-fed loop-slot antenna, the photonic bandgap (PBG) structure is a promising candidate in this regard. As is well known, PBG structures, originating in the optical regime and scaleable the microwave and millimeter-wave applications, are frequency-selective

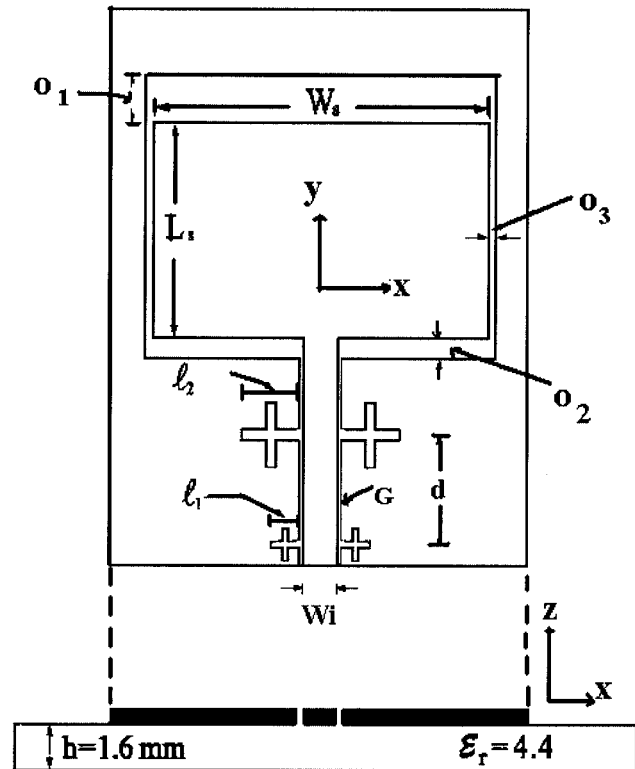


Figure 1 Geometry of a broadband loop-slot antenna with cross-shaped lattices

surfaces capable of restricting the propagation of electromagnetic waves along one or more directions within a certain band of frequencies. PBG structures have been applied to eliminate the harmonic modes in microstrip patch antennas [3] due to their appealing low-pass filter characteristic [4]. In [5], the conventional CPW-fed loop-slot antenna, incorporating a PBG structure in the feeding network, has been realized successfully. The PBG structure with square-shaped lattices has exhibited great performance for impedance bandwidth enhancement and harmonic suppression.

In this contribution, a PBG structure with cross-shaped lattices is utilized for harmonic suppression. The design of antenna and PBG structure with cross-shaped lattices is discussed in detail.

2. ANTENNA DESIGN

In this design, cross-shaped lattices have been used to suppress harmonic modes and the geometry of the proposed antenna, as shown in Figure 1. The slot antenna is fabricated on an inexpensive FR4 substrate with dielectric constant $\epsilon_r = 4.4$ and substrate thickness $h = 1.6$ mm. Its width W_s is 40 mm and length L_s is 21.5 mm, and it has three gap widths of $O_1 = 4$ mm, $O_2 = 2$ mm, and $O_3 = 1$. The conventional CPW-fed line is designed with a strip width W_i of 4 mm and a gap width G of 0.4 mm, corresponding to a characteristic impedance of 50Ω . Furthermore, the cross-shaped lattices are placed on both sides of the ground plane along the feed line. The dimension of the larger cross-shaped lattice is $l_2 \times l_2 (7 \times 7 \text{ mm}^2)$ while that of the smaller cross-shaped lattice is $l_1 \times l_1 (3 \times 3 \text{ mm}^2)$. Moreover, the distance d between two lattices is chosen as 11 mm.

As is known, the distance between two cells, which is equal to half guided wavelength ($\lambda_g/2$) in [4, 6], determines the cutoff frequency of the PBG structure. In this design, the PBG structure used is more compact, since the distance between two cells ap-