A Broadband Transmitarray Using Double Square Ring Elements

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Abstract— **Dual-resonant double square ring elements are applied to the design of a four-layer transmitarray operating at 30 GHz. The design procedure is described and results of a 7.5% –1 dB gain bandwidth and 47% radiation efficiency are reported.**

Keywords- Transmitarray, broadband, double square ring

I. INTRODUCTION

Antennas for satellite-based telecommunication systems should be high-gain, broadband, lightweight, and inexpensive to manufacture. Among the possible options, the transmitarray antenna is a promising potential technology to meet these requirements. It is relatively easy to fabricate, and does not suffer the insertion loss of a phased array's feed network at millimeter wave frequencies [1]. Comprised of a planar array of printed patch elements, the transmitarray avoids the size and weight disadvantages of reflector antennas or shaped lenses. Furthermore, the feed can be placed directly in front of the aperture without incurring the blockage losses of a reflectarray configuration. The main drawback of the transmitarray is its limited bandwidth which is usually around 5% or less. To overcome this bandwidth limitation while minimizing antenna thickness, this paper presents a novel transmitarray operating at 30 GHz which uses a double square ring as the unit cell element.

To date, transmitarrays have been manufactured using stub-loaded patches or using elements connected through multiple layers by a delay line [2,3]; however, these long lengths of line can make element placement difficult and can result in spurious cross-polarized radiation [1] or unwanted modes in the layered transmitarray [4]. Additionally, the use of amplifier stages by some authors increases the cost and complexity and makes the antenna performance susceptible to failure of the active devices [5,6]. In an approach that simplifies the fabrication process, a third scheme has been implemented by Chaharmir *et al* in which four unconnected layers are cascaded and the phase change is accomplished not through delay lines, but by varying the cross-dipole element lengths [7]. They noted the impracticality of achieving the full 360° phase shift using a single layer, and consequently the disadvantage of this approach is that the required multiple layers increases the antenna's thickness. This paper will focus on increasing the bandwidth of the transmitarray while maintaining the same number of layers as this latter method.

II. DOUBLE SQUARE RING ELEMENT ANALYSIS AND DESIGN

A. Background

To collimate the feed signal incident wave, the transmitarray uses the antenna elements on its surface to re-phase the incoming spherical wave and then re-transmits the signal as a plane wave. The amount of phase adjustment needed at each antenna element depends on how much phase an incident ray has accumulated travelling between the feed horn and the transmitarray surface. The necessary phase compensation value Φ_i at each element is given by [7]

$$\Phi_i + k[\boldsymbol{R_i} + \boldsymbol{r_i} \bullet \boldsymbol{u_0}] = 2\pi n,\ n=0,1,2,\ldots \qquad (1)$$

where, as shown in Fig. 1, $\boldsymbol{R_i}$ is the vector to the i^{th} element from the feed's phase centre, $\boldsymbol{r_i}$ is the position vector to the i^{th} element from the transmitarray centre, k is the propagation constant, and $\boldsymbol{u_0}$ is the intended direction of the transmitted main beam. Since the transmitarray diameter will be many wavelengths, each element must be capable of producing at least 360° of phase shift. In this design, the desired phase shift from each element is obtained by changing the element's dimensions around its resonant dimension; the impedance seen by the incident wave changes with the cell geometry and so the required phase compensation value can be specified.

B. Choosing the Double Square Ring Element

The first goal of this work is to increase the transmitarray bandwidth, which is primarily limited by the bandwidth of the element itself [5]. Therefore, a loop element, known to have a wide bandwidth is chosen [8].

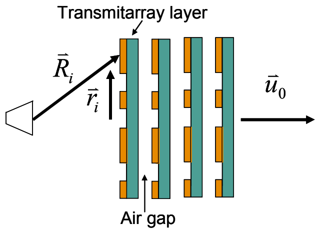

Figure 1. Transmitarray geometry

To increase the phase change that can be obtained from this element, a second, concentric ring is added, as shown in Fig. 2. The addition of the second ring introduces a second resonance in the frequency response and thereby allows a greater total phase change to be obtained from this element by increasing the slope of the phase-versus-frequency curve between the two resonances [9]. This increased phase change is important since fewer layers are then required to achieve the full 360º coverage and the antenna thickness, weight, cost, and fabrication complexity are correspondingly decreased. However, the inner ring also introduces greater complexity to the design since there are now two resonant frequencies where the transmission magnitude is poor, and which must be avoided when designing the unit cell. In optimizing the position of the two resonances, therefore, the extra degree of freedom can be used to find the best balance between the phase change and the transmission magnitude bandwidth.

C. *Designing the Unit Cell*

Extensive simulations were performed in HFSS to optimize both the transmission magnitude and phase response of the double-square ring element. In each simulation, lossless materials were assumed and only normal incidence angles were considered; furthermore, perfect electric- and magnetic-wall boundary conditions were enforced around a single unit cell to result in image planes and the simulation of an infinite layer of antenna elements. By placing the two resonant frequencies of the double square ring element close together, the designer can increase the slope of the phase-frequency curve thereby increasing the rate of phase change versus element width. Unfortunately, two closely spaced resonances produce narrow-band operation, or equivalently, limit the range of geometric variation of the element. This approach, which therefore results in large phase increments per variation of the unit geometrical dimensions, leads to phase errors and decreased gain. A better strategy is to accept a slower rate of phase change and to use a wider range of widths to cover the same total phase range. This method, which requires separating the two resonant frequencies, is best accomplished by increasing the gap between the inner and outer rings. Two variations on this strategy were investigated: first, the element width is varied for a large gap size, and second, the outer ring width is fixed and the gap itself is varied. This second approach actually combines two different geometries which together cover the full 360º phase range. These two geometries differ only in the size of their outer rings, and both implement the phase compensation by varying the gap dimension. The two sets cover different, non-overlapping phase ranges and both are used on each transmitarray layer depending upon what phase compensation value is required. The two methods of variable-width and variable-gap are illustrated in Fig. 2. This figure also shows the dimensions that are varied under each scheme to produce the phase shift. For ease of reference, each geometric variation will be referred to as an "element index" (*i.e.,* the first geometric variation of each configuration will be referred as index 1, while the second variation of each is index 2, and so on).

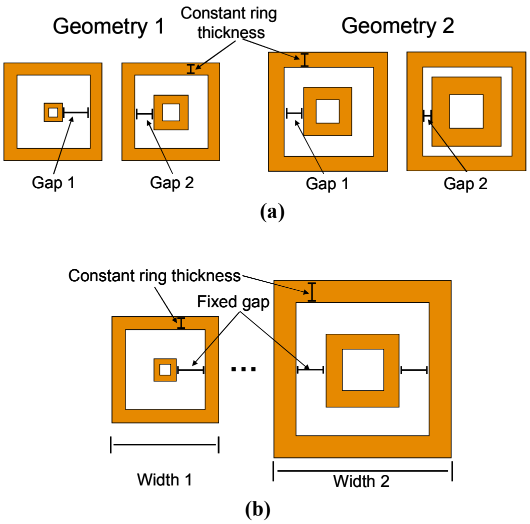

Figure 2. Comparison of (a) variable-gap and (b) variable-width designs

The geometries for both types of unit cells were optimized in HFSS to provide the largest single-layer phase change while maintaining transmission above –3 dB over a 28-31 GHz bandwidth. Multiple layers of each of the two unit cells were also simulated, and it was determined that four cascaded layers were required in each case to obtain the full 360º coverage while still exceeding the specified minimum transmission level. The optimized dimensions of the unit cell and multilayer structures are given in Table 1. Fig. 3 shows the transmission magnitude and phase at 30 GHz versus element index for a four-layer structure. The variable-width geometry uses a narrower range of width values since its large gap size limits the size of elements that can fit within the unit cell boundaries. For the four-layer variable-width design, over the range of 8 quantized element widths from 4 mm to 5.4 mm, it was found that the amplitude of the transmitted wave is greater than –2 dB and usually close to 0 dB for a frequency band of 28-30 GHz, while the phase change at 30 GHz is 329º. For the four-layer variable-gap design, 15 quantized inner ring widths from 0.6 mm to 2.4 mm and from to 4.2 mm to 5 mm result in transmission amplitude greater than –3 dB over a frequency band of 28-30.5 GHz. The phase change at 30 GHz is 312º, which although slightly smaller than that of the variable-width design, is accomplished using nearly twice as many element widths and consequently, the quantization error of the phase response is decreased.

TABLE I. DIMENSIONS OF VARIABLE-GAP AND VARIABLE-WIDTH DESIGNS

Parameter	**Variable Gap**	**Variable Width**
Conductor Thickness	0.2 mm	0.2 mm
Gap Size	1.3-2.2, 0.2-0.6 mm	1.3 mm
Outer Ring Width	5.4 mm, 5.8 mm	4-5.4 mm
Cell Size	6.0 mm	6.0 mm
Substrate Thickness	0.127 mm	0.127 mm
Relative Permittivity	2.2	3.0
Layer Separation	3,3,3 mm	3,2,3 mm

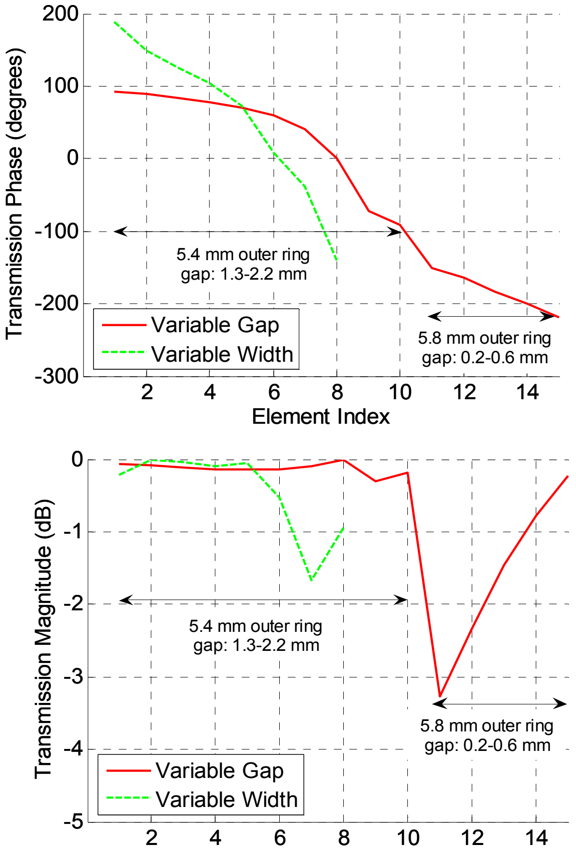

Figure 3. Comparison of multilayer transmission magnitude (top) and phase (bottom) versus element width index for both variable-width and variable-gap designs

Using the multilayer transmission magnitude and phase properties simulated in HFSS, array theory was used to calculate the radiation patterns at 30 GHz of a transmitarray of dimensions 12.6 x 12.6 cm for both the variable-gap and fixed-gap designs, and the results are plotted in Fig. 4 (top). The directivity of an ideal aperture at 30 GHz of equal dimensions and with the feed pattern modeled by a $cos^{9}(\theta)$ distribution is calculated to be 31.9 dB. The theoretically predicted directivity for the two competing designs is 31.26 dBi and 30.59 dBi for the variable-gap and variable-width transmitarrays, respectively; thus, the simulated aperture efficiencies are approximately 85% and 73%. The gain versus frequency is given in Fig. 4 (bottom), and the 1 dB gain bandwidths are 3.25 GHz (11%) for the variable-gap and 3 GHz (10%) for the variable-width design. Higher efficiency and larger bandwidth make the variable-gap unit cell the best choice, and it is this design that was used in the fabricated transmitarray.

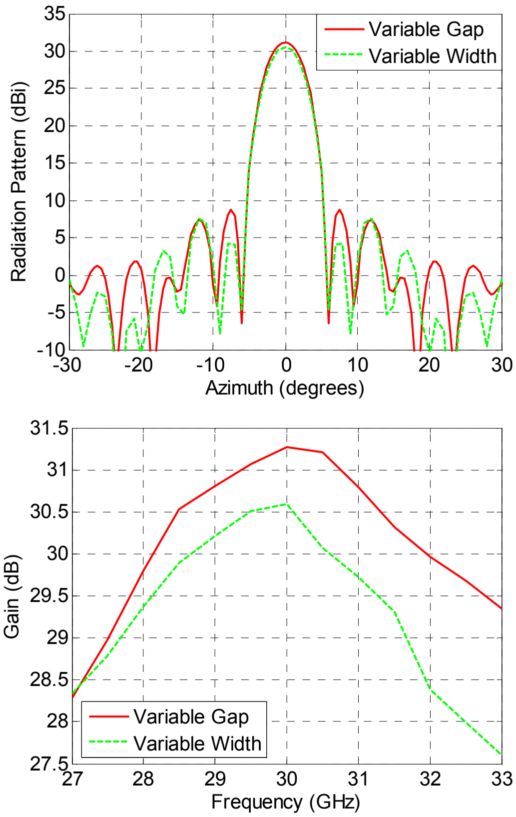

Figure 4. Comparison of simulated radiation patterns at 30 GHz (top) and simulated gain versus frequency (bottom) for variable-gap and variable-width designs

III. MEASURED RESULTS

The four-layer transmitarray using the variable-gap method was fabricated according to the dimensions of Table 1. Fig. 5 shows the measured far-field H-plane radiation patterns for both co- and cross-polarised signals at 30 GHz and the measured peak gain versus frequency. The measured gain at 30 GHz of 28 dB corresponds to a radiation efficiency of 41%, and the first sidelobe level is nearly –17 dB below the main peak. The 1 dB gain bandwidth is found to be 2.25 GHz, or 7.5%. The measured results are compared against the simulated performance in Table 2.

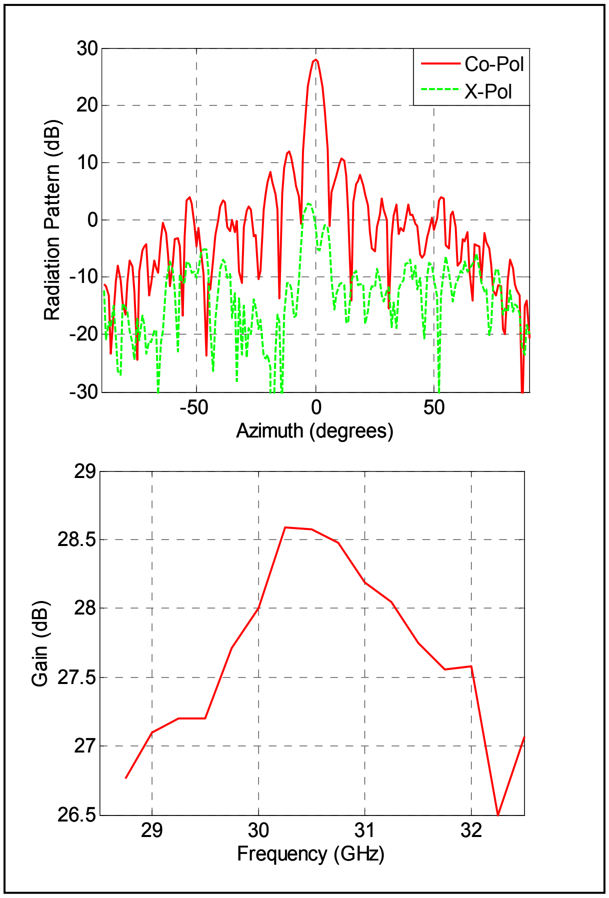

Figure 5. Measured 30 GHz radiation pattern (top) and gain versus frequency (bottom)

TABLE II. COMPARISON OF TRANSMITARRAY MEASURED AND SIMULATED PERFORMANCE

Metric	Measured Value	Theoretical Value
Peak gain	28.59 dB	31.26 dB
Peak radiation efficiency	47% at 30.25 GHz	85% at 30 GHz
Sidelobe level	–17 dB	–23 dB
1 dB gain bandwidth	2.25 GHz (7.5%)	3.25 GHz (10.8%)

Discrepancies between the measured and simulated results can be attributed to a variety of errors. Neglecting to include dielectric and conductor losses likely contributed to the smaller-than-expected radiation efficiency, while assuming the feed signal is normally incident on all elements leads to errors in the magnitude and phase response of edge elements for oblique incidence angles. Furthermore, the infinite array assumption inherent in the boundary conditions set in HFSS does not represent a finite-sized transmitarray layer and consequently does not account for diffraction from the antenna edges. Finally, manufacturing tolerances could have resulted in phase errors in the wavefront compensation, thus decreasing transmitarray gain. Nevertheless, a radiation efficiency of 47% is well within the typical range for transmitarray antennas, and the primary design goal of improving upon antenna bandwidth has been achieved. This combination of relatively high efficiency, a 1 dB gain bandwidth of 7.5%, and simple construction techniques show the potential of double-resonant elements in transmitarray antennas.

IV. CONCLUSIONS

This work has demonstrated the use of double-resonant elements in transmitarray design. Resolving the tradeoff between large phase change and wide bandwidth relies on separating the resonant frequencies as much as possible and implementing the wavefront compensation using small increments in compensating phase values. With these techniques, the double square ring has used the additional degree of freedom of the second ring's resonance to improve the 1 dB gain bandwidth over previous designs without sacrificing other performance metrics.

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