

A 32-GHz Microstrip Array Antenna for Microspacecraft Application

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JPL/NASA is currently developing microspacecraft systems for future deep space applications. One of the frequency bands being investigated for microspacecraft is the Ka-band (32 GHz), which can be used with smaller equipment and provides a larger bandwidth. This article describes the successful development of a circularly polarized microstrip array with 28 dBic of gain at 32 GHz. This antenna, which is thin, flat, and small, can be surface-mounted onto the microspacecraft and, hence, takes very little volume and mass of the spacecraft. The challenges in developing this antenna are minimizing the microstrip antenna's insertion loss and maintaining a reasonable frequency bandwidth.

I. Introduction

JPL is currently considering the development of microspacecraft for future deep space missions in order to meet NASA's goal of having small and inexpensive spacecraft. The microspacecraft,¹ having sizes on the order of 0.5 m, will certainly require components that are small both in size and mass. One of the frequency bands under investigation for use with microspacecraft is the Ka-band (32 GHz). In this frequency range, components are smaller and larger bandwidths can be achieved as compared with a lower frequency range. A larger bandwidth will result in a higher data rate.

In microspacecraft development, antennas are one of the major components that warrant attention since they generally require significant amounts of real estate and mass. Conventional antennas such as horns and parabolic reflectors, although efficient radiators, are generally bulky in size and large in mass. As a result, three different types of flat antennas with very thin profiles have been proposed for the microspacecraft. Since they are flat, the antennas can be surface-mounted onto the spacecraft without additional supporting structures, such as those needed for curved parabolic reflectors. Therefore, they require very little volume and mass.

The first type is the microstrip array [1], in which many printed microstrip patch elements are combined by microstrip transmission lines. Due to the relatively high in-

¹ C. Salvo and C. Nunez, "Code R Microspacecraft Subsystem Technology Task," JPL Interoffice Memorandum 3132-92-099 (internal document), Jet Propulsion Laboratory, Pasadena, California, March 23, 1992.

sertion loss that generally exists in microstrip transmission lines, the microstrip array cannot be very large; otherwise the antenna will be very inefficient. At Ka-band, the microstrip array will have an acceptable insertion loss if its aperture dimension is less than 15 cm. The second type is the planar slotted waveguide array [2], which is generally bulky at low microwave frequencies, but which can be made low profile with a small mass at high frequencies, such as Ka-band. Because the slotted waveguide encounters very little insertion loss, its aperture size can be large. The upper limit of its size is governed by its metallic mass. At Ka-band, a reasonable dimension range for the planar slotted waveguide array is from 5 to 35 cm.

For antennas with dimensions larger than 35 cm, a third type of flat antenna, called the microstrip reflectarray or flat reflector [3], can be used. This antenna has similarities to the conventional parabolic reflector, in which both a feed and a reflecting surface are needed. However, for the microstrip reflectarray, the reflecting surface can be flat rather than curved. Many identical but isolated microstrip patches (which are not connected to each other by power dividers) with phase delay lines of different lengths are printed on this flat reflecting surface. It is these phase delay lines that will cause the reradiated energy originating from the feed to become coherent in the far field. Without power-division transmission lines, the insertion loss of the antenna will not increase as the size of the antenna increases. Consequently, the aperture size of the reflectarray can be very large while still maintaining good efficiency. Since they are flat, these reflecting surfaces can be surface-mounted onto the spacecraft without additional structure support, such as is needed for a curved parabola. Therefore, the spacecraft can have a lower mass and a lower volume by utilizing the reflectarray antenna.

Due to limited resources in fiscal year 1993, only the first type of antenna, the microstrip array, was developed. The other two types of antennas are planned to be developed in the following years. This article documents the development of the Ka-band microstrip array, which must be circularly polarized with 28 dBic of peak gain. The array employs a series feed technique [4] for its power-division transmission lines in order to reduce insertion loss, and it uses sequentially arranged [5] elements for its subarrays to improve the quality of the circular polarization (CP).

II. Antenna Requirements

The antenna technologies studied here are intended for a series of future JPL/NASA microspacecraft applications. The set of antenna requirements generated here, however,

are for a particular application called Asteroid Comet Microspacecraft Explorer (ACME-II).² The ACME-II spacecraft, as illustrated in Fig. 1, has a hexagonal shape with dimensions on the order of 0.5 m. The bottom of the spacecraft is used for the propulsion adapter unit and does not have enough room for the antenna. The top is allocated for an optical camera, louvers, etc., and also does not have adequate space for the antenna. The only likely location for the antenna is on one of the six side panels, which are allocated for the solar arrays. It seems appropriate to mount a small and flat array antenna flush with the solar array. The antenna requirements for the ACME-II spacecraft are given in Table 1.

The antenna gain of 28 dBic is set for the microspacecraft's near-Earth encounter with an asteroid or a comet at a range of approximately 1 AU. This gain yields an antenna beamwidth of about 5 deg. With this relatively wide beamwidth, fine beam pointing at the Earth is not required. Therefore, an automatic antenna tracking capability is not required; rather, a fixed broadside beam is adequate for the antenna system. Coarse antenna pointing is to be done by the spacecraft's onboard system, such as the star tracker. The challenge in developing the proposed microstrip array antenna is to achieve the 28 dBic of gain with a minimal amount of insertion loss over a relatively wide bandwidth of 1 GHz. In other words, the efficiency of the microstrip array may need some attention. Low profile, small mass, and low cost are certainly the attractive features of the microstrip array antenna. Other antennas, such as horns and parabolic reflectors, although known to be efficient radiators, may have very poor efficiency in terms of antenna volume and mass for microspacecraft applications. For example, for a horn antenna to provide 28 dBic of gain, it will need approximately 9 cm × 9 cm of aperture with 20 cm of length. A significant amount of the spacecraft's volume is needed to accommodate the horn antenna since the spacecraft's dimension is only 50 cm across. Consequently, the horn antenna will have a volume at least 50 times and a mass at least 10 times as large as the microstrip array. To summarize, for microspacecraft applications, the salient physical features of the microstrip array can compensate for the antenna's relative inefficiency at Ka-band.

III. Antenna Development Background

To develop a high-gain microstrip array antenna, three major steps are generally required: the single patch el-

² Ibid.

ement development, a building-block subarray development, and the full array development. Since the microstrip patch is a narrow-band device (generally less than three-percent bandwidth), the development of the single patch is necessary to ensure its correct resonant frequency, polarization purity, bandwidth, etc. Currently, no mathematical tool can directly design a complete microstrip array to operate at the desired frequency. Although some mathematical tools are very accurate, one cannot ensure that the fabrication tolerance and the manufacturer's material specifications are all within the required accuracy. For example, the dielectric constant of the substrate is generally 1 percent off from the manufacturer's specification, which can cause a similar percentage of change in resonant frequency. The second step is the development of the building-block subarray, which means that the full array consists of many identical subarrays. This subarray should be developed prior to the full array to ensure good circular polarization and to minimize overall development risk. Once the single subarray is developed, many of these identical subarrays can then be combined by microstrip transmission lines to form the final array.

A. Single-Patch Element Development

For an array to generate CP, its element is generally required to be circularly polarized. However, when the radiation efficiency is not a concern, a CP array can be composed of linearly polarized (LP) elements [6,7]. Since the radiation efficiency is of concern, the patch element must be circularly polarized. There are basically two techniques for a single microstrip to generate CP. One is to have a square or circular patch excited by two orthogonally located feeds in time quadrature, as shown in Fig. 2(a). The other is to have a truncated-corner square patch, as shown in Fig. 2(b), excited by a single feed. Since the insertion loss of a microstrip array is mostly incurred in the microstrip power-distribution transmission lines, the lengths and complexity of these transmission lines should be minimized. As a result, the truncated-corner, single-feed patch rather than the orthogonally fed patch is selected for the array. There are several other techniques [8] to generate CP by a single-feed patch, such as using a tilted slot at the center of a square patch and a circular patch with two indented edges.

All these single-feed CP methods work as a result of the fields underneath the patch being perturbed by the small truncations. The perturbation splits the field into two orthogonal modes that are degenerate in frequency and in-phase quadrature. The amount of perturbation must be just right so that two orthogonally polarized fields with

equal amplitudes and orthogonal phases can be formed. Consequently, the CP bandwidth is very narrow and generally is less than 1 percent. This narrow bandwidth will cause the resonant CP frequency to be very sensitive to fabrication tolerance, temperature change, etc. One technique for increasing the CP bandwidth is to arrange four neighboring elements sequentially in orientations and in phases. This technique is discussed in the next subsection.

The design of the truncated-corner square patch is assisted by both the cavity modal theory [9] and the variational method [10]. The cavity modal theory, used here to design an unperturbed square patch, assumes that the fields underneath the patch can be expanded in terms of trigonometric functions (modal functions). By solving for the unknown coefficients of these modal functions through known boundary conditions, the resonant frequency and input impedance can be determined. The amount of corner truncation for CP generation is then determined by the variational method with the simple close-form equation given below [10]:

$$f' = f \left(1 + \frac{S'}{2S} \right) \quad (1)$$

with

$$\frac{S'}{S} = \frac{1}{2Q}$$

where S' is the sum of the two truncated areas, S is the area of the unperturbed square patch, f is the resonant frequency of the unperturbed patch, f' is the frequency for the optimal axial ratio, and Q is the quality factor of the unperturbed patch.

The quality factor, Q , is a function of the substrate dielectric constant, thickness, and loss tangent. It can be obtained simply through a curve given in [10] or calculated from [8]. It can also be determined through the impedance bandwidth (in percent for a 2:1 voltage standing wave ratio [VSWR]) either calculated or measured for the unperturbed patch by the following simple relationship [11]:

$$Q = \frac{1}{\sqrt{2}(\text{bandwidth})} \quad (2)$$

With the above formulations, the truncated-corner square patch was designed with the following dimensions (see

Fig. 2): $a = 0.038$ cm, $b = 0.297$ cm, the relative dielectric constant = 2.2, and the dielectric thickness = 0.025 cm. The measured input return loss at 32.0 GHz is -21 dB, which corresponds to a VSWR of 1.22:1. The measured axial ratio is best at 32 GHz but with a relatively large magnitude of 3 dB. This relatively large axial ratio is the result of the narrow-band characteristic with a high sensitivity to fabrication tolerance and the manufacturer's material specification tolerance. Nevertheless, it is expected that this large axial ratio can be reduced by the sequentially arranged subarray technique to be discussed in the next subsection. A photograph of this single patch with truncated corners is shown in Fig. 3, where the patch is fed by an impedance-matched microstrip transmission line and an end-launched OS-50 coax connector. The printed antenna circuit is etched on Duroid 5880 substrate material. Due to its extremely thin profile, the antenna is mounted on a brass plate for mechanical support.

B. Four-Element Subarray Development

It has been known that the CP quality and bandwidth can be improved if four elements, each being a narrow-band CP element, can be arranged sequentially [5,6] in orientation and in phase, as shown in Fig. 4. The four elements are arranged in 0-, 90-, 180-, and 270-deg fashion to achieve symmetry and to cancel undesirable higher-order modes (explained in [6]) and, thus, to achieve purer polarization. Figure 5 presents the photograph of the sequentially arranged four-element subarray with each single element having dimensions identical to those presented in Fig. 3. The element spacing is 0.73 free-space wavelength, and the required differential phases between elements are achieved by designing transmission lines of different lengths. A vertically launched OS-50 coax connector is used to feed the microstrip-transmission-line power-dividing circuit. The measured input return loss at 32 GHz is -17.4 dB (VSWR = 1.31:1). Its 1.5:1 VSWR bandwidth is about 0.5 GHz. The radiation pattern measured with a spinning dipole is shown in Fig. 6, where an axial ratio of less than 0.25 dB is demonstrated in the broadside direction. This axial ratio is significantly better than that of the single element. This four-element subarray is used as the building block in the full array, where all the subarrays are identical and are combined by microstrip power distributing lines, as discussed in the next section.

IV. Full Array Design

To provide the required antenna gain of 28 dBic, an aperture size of approximately 10 cm \times 10 cm is needed at

32 GHz. The number of radiating elements needed is about 200. When combining such a relatively large number of elements by microstrip transmission lines at Ka-band, the insertion loss becomes significant. One technique to minimize this insertion loss is to employ the series-feed method in which all the four-element subarrays are combined serially by the power distributing circuit. This series-feed method is illustrated in Fig. 7. The input/output coax connector, located at the center of the array, feeds the microstrip transmission line with a two-way power division that splits the power parallel to the left and right identical halves. Each half array is then fed serially in columns by the microstrip line. The upper and lower halves of the array are fed in parallel at the array's horizontal center line and then fed serially at each subarray location.

Figure 8 illustrates the difference between the series-feed and the parallel-feed techniques for a four-element array having uniform power distribution. A series-fed array means that a certain amount of power is used to feed the first element; then a certain amount from the remaining power is used to feed the second element at a delayed time; and this continues until the fourth element is reached. At the fourth element, only one-quarter of the total power remains to be transmitted. In other words, full power goes through the transmission line at the first element location, and the power diminishes as it travels toward the last element. Consequently, the largest loss in total power for the transmission line occurs at the first element, and this power loss reduces as the power travels down the series-fed line. For this reason, most of the insertion loss occurs at the center region of the array, shown in Fig. 7. For a parallel-fed array, all elements are fed equally at the same time by parallel transmission lines. Therefore, the insertion loss exists almost uniformly from the beginning of the transmission line all the way to the end of the line. Thus, the series-fed array will encounter significantly less insertion loss than will the parallel-fed array.

Note that at each power division point in Fig. 7 the microstrip line changes width. This was uniquely designed to ensure uniform power distribution with proper impedance matches throughout the array. It should be noted that the microstrip lines are impedance matched at every junction point throughout the array so that multiple reflections of the signal are minimized to reduce insertion loss. Furthermore, the bandwidth of this array is slightly larger than the conventional "resonant" array, in which impedances are not matched and multiply reflected signals are significant. For the array in Fig. 7, the spacings between adjacent columns in the horizontal direction and between adjacent subarrays in the vertical direction are all integral multiples of the "guide" wavelength, so that the far-field

radiation from all patches is in phase in the broadside direction. Here the guide wavelength is the effective wavelength of the transmission line in a dielectric substrate. Two guide wavelengths are used here so that they are long enough to physically accommodate the subarrays and yet short enough to avoid excess insertion loss. Because of this guide wavelength requirement for phase coherence, the spacing between two adjacent patch elements becomes approximately 0.74 free-space wavelength. Also, because the guide wavelengths are slightly different for different transmission line widths, the spacing between adjacent patches is slightly different throughout the array (not all exactly equal to 0.74 free-space wavelength). This is why not all the patch elements in any given vertical column or horizontal row are precisely aligned. In the array, there are a total of 8 vertical columns with 6 subarrays in each column, and, therefore, there are a total of 48 sequentially arranged subarrays with a total of 192 microstrip patch elements.

Due to the fact that the array must have integral multiple numbers of subarrays and preferably be symmetrically fed from the center, it is not possible to have a square aperture and yet still fully utilize the required 10-cm \times 10-cm area (see Fig. 7). Consequently, a rectangular aperture is formed. The overall antenna physical size is 11.4 cm \times 8.9 cm. The microstrip patches and lines are etched on a 0.025-cm-thick dielectric substrate (Duroid 5880), which in turn is mounted on a 0.16-cm-thick supporting aluminum plate. The total antenna thickness is 0.185 cm. The array's actual radiating aperture size (10.8 cm \times 8.3 cm) is slightly smaller than its physical size. This is because additional ground planes on the edges of the array are needed to accommodate the transmission lines.

The power distribution circuit for the full array is a combination of parallel- and series-feed [4] transmission lines. It will not be a good design if a sole series-feed technique is used. This is because a sole series-fed array will cause its main beam to be scanned away from the intended direction as frequency is changed due to accumulated phase change. With a relatively wide band signal, the composite transmit beam of all the frequencies within the band will appear to be wider and the gain at the receiver direction will thus be lower. Viewed another way, the spectrum of the antenna filter response will be distorted. For the array shown in Fig. 7, the right and left halves of the array are fed in parallel, while within each half array, the columns are fed serially. As frequency changes, the beam of each half array will scan. However, the two half arrays will scan in opposite directions, leaving the composite beam pointed in the broadside direction

with a lower gain. Because the beamwidth of each half array is wider than the complete array, its beam scan due to frequency changes will result in smaller gain changes than that of a narrowbeam, solely series-fed array. To summarize, this combined parallel and series feed technique will result in a stable beam direction over a relatively wide bandwidth with a relatively low insertion loss.

V. Experimental Results

The full array, as sketched in Fig. 7, has been fabricated and tested and is shown in Fig. 9. The input is a perpendicularly launched OS-50 coax connector. Its measured input return loss is given in Fig. 10, where a return loss of -23.6 dB (1.14:1 VSWR) is noted at 32 GHz. The 1.5:1 VSWR bandwidth is about 2 GHz, which is significantly wider than the four-element subarray's bandwidth of 0.5 GHz. This is partly because the full array's input match is better designed and partly due to the fact that the full array has a much larger resistive insertion loss in the transmission lines to lower the returned energy. One can note that in Fig. 9 there are two tuning stubs located a quarter-wavelength away from the input connector and that there is none for the four-element subarray. These tuning stubs can eliminate a large portion of the mismatch over a significant bandwidth.

Both the measured narrowbeam and broadbeam principal plane patterns of the rectangular aperture array are shown in Figs. 11 and 12, respectively. The 3-dB beamwidths are 4.5 deg in one principal plane and 5.7 deg in the orthogonal plane. These patterns, measured at 32 GHz, show a peak side-lobe level of -12 dB and a peak cross-pol level of -18 dB (corresponding to a 2-dB axial ratio). It is believed that the 2-dB axial ratio, which is worse than that of the four-element subarray, is mostly caused by the accumulated phase errors in the series-fed microstrip lines and is partly due to the leakage radiation from these lines. The measured 3-dB axial ratio bandwidth of the full array is 1.3 GHz. The array's peak gain is 28.4 dBic, which was measured using a standard gain horn by the comparison method. Based on the radiating aperture size of 10.8 cm \times 8.3 cm, the calculated directivity of a uniformly distributed array should be 31.0 dBic. This implies that the array has an overall efficiency of 55 percent with a total loss of 2.6 dB. The loss is mostly the result of the microstrip transmission line's ohmic losses (1.0 to 1.5 dB), which occurred in the substrate dielectric and in the copper conductor. The remaining portions of the loss are mismatch loss, cross-pol loss, side-lobe loss, patch element loss, and input connector loss. The cross-pol and side-lobe losses are partly due to imperfect design and

partly caused by radiation leakage from the transmission lines. It is expected that the overall loss would be significantly higher if the series-fed technique and the single-fed CP patch were not used. The antenna gain was also measured across a frequency bandwidth, which demonstrated that the array has a bandwidth of 0.95 GHz with 1 dB of maximum gain drop. The antenna's total mass, including the aluminum supporting plate, is 0.053 kg (1.88 oz). The characteristics and performance of the array antenna are summarized in Table 2.

VI. Conclusions

A 32-GHz circularly polarized microstrip array with 192 patch elements has been successfully developed for the microspacecraft application. It is intended as a downlink

telecommunication spacecraft antenna. Each patch element is circularly polarized with a single feed and two truncated corners.

The stability and quality of the circular polarization is improved by arranging four adjacent elements in a sequential fashion. Thus, the full array consists of 48 identical four-element subarrays. To minimize the insertion loss, these subarrays are combined primarily by series-feed microstrip transmission lines, and impedances are matched throughout the array. The antenna achieved the required gain of 28 dBic with a physical size of 11.4 cm × 8.9 cm and a thickness of 0.18 cm. Due to its low profile, the array antenna is to be surface-mounted along with the solar panels. The total antenna mass is only 0.053 kg (1.88 oz). The antenna has achieved the goal of occupying very little of the volume and mass of the spacecraft.

Acknowledgments

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Table 1. ACME-II antenna requirements.

Parameter	Requirement
Frequency, GHz	32
Operation	Downlink transmit only
Gain, dBic	≥ 28
Polarization	Right-hand circular
Side-lobe level, dB	≤ -13
Cross-pol level, dB	≤ -15
Beam direction	Fixed broadside
Input VSWR at 32 GHz	$\leq 1.5:1$
Bandwidth, GHz	≥ 1
Power handling, W	5
Antenna mass, kg	≤ 0.5
Aperture size, cm	10 x 10

Table 2. Characteristics and performance of the array antenna.

Parameter	Characteristic or performance	Requirement
Center frequency, GHz	32	32
Input VSWR at 32 GHz	1.14:1	1.5:1
1.5:1 VSWR bandwidth, GHz	2	1
Peak gain, dBic	28.4	28.0
-1-dB gain bandwidth, GHz	0.95	1.0
Beamwidth, deg	5.7 x 4.5	$\sim 5 \times 5$
Peak side-lobe level, dB	-12	-13
Polarization	Right-hand CP	Right-hand CP
3-dB axial ratio bandwidth, GHz	1.3	1
Radiating aperture size, cm	10.8 x 8.3	$\leq 10 \times 10$
Physical size, cm	11.4 x 8.9	$\leq 10 \times 10$
Antenna thickness, cm	0.18	Not specified
Antenna mass, kg	0.053	≤ 0.5
Power handling, W	Expected 10	5

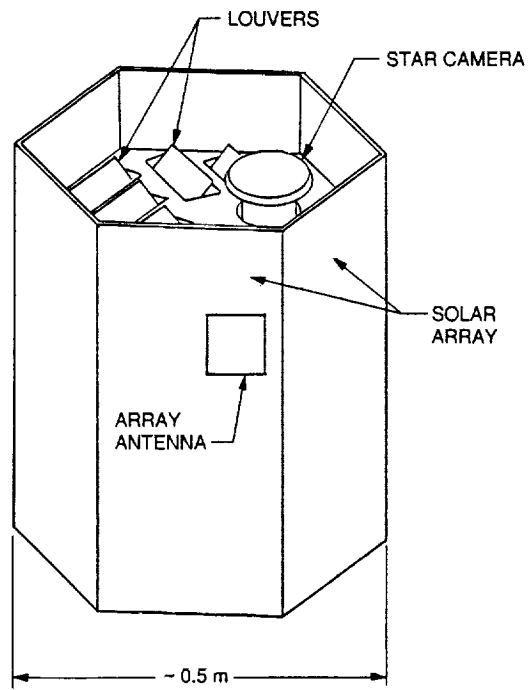


Fig. 1. ACME-II microspacecraft configuration.

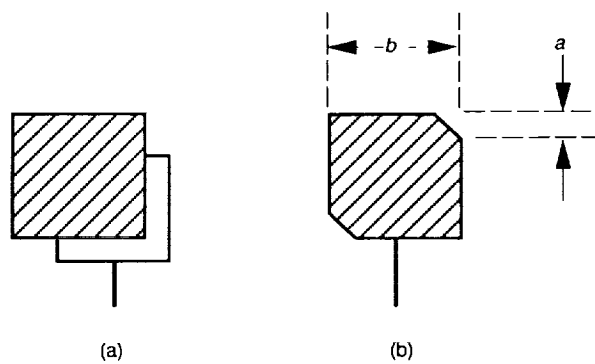


Fig. 2. Circularly polarized single-patch antenna with (a) orthogonal dual feeds and (b) single feed with truncated corners.



Fig. 3. The circularly polarized single-feed patch.

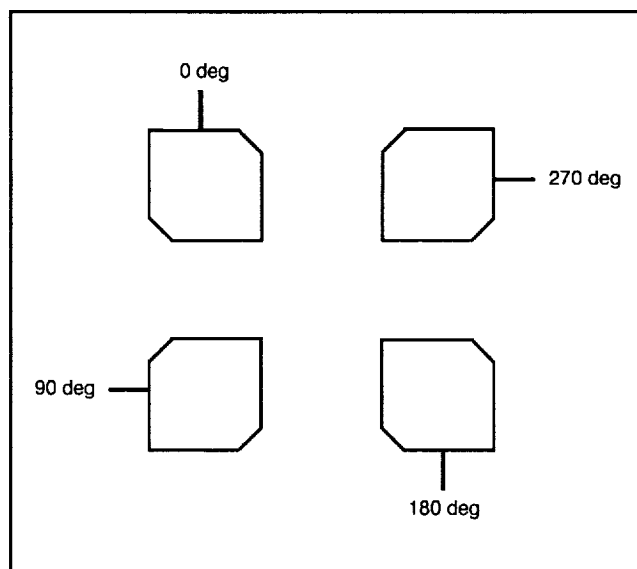


Fig. 4. The circularly polarized four-element subarray with sequential arrangement.

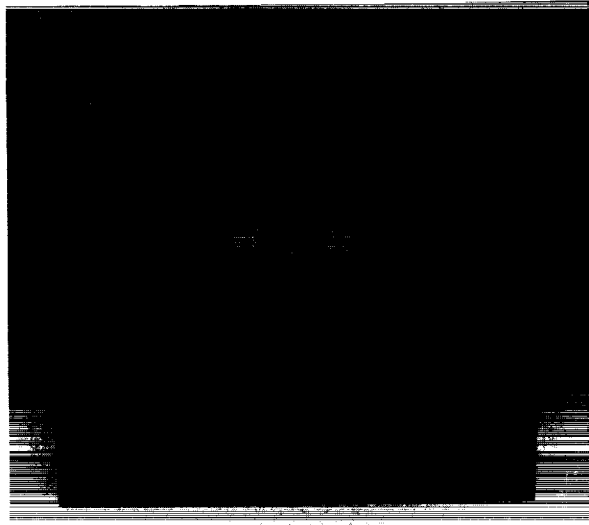


Fig. 5. The sequentially arranged four-element subarray.

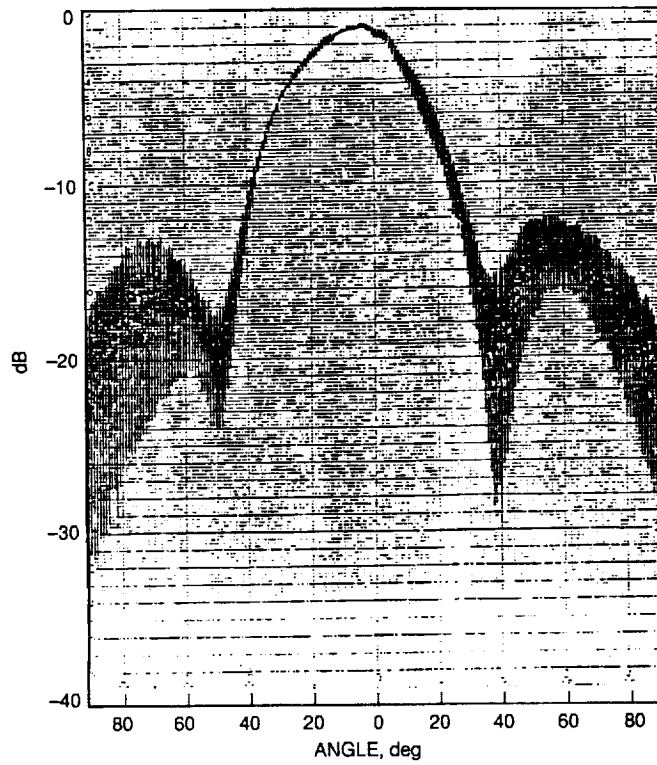


Fig. 6. Measured spinning-linear pattern of the four-element subarray.

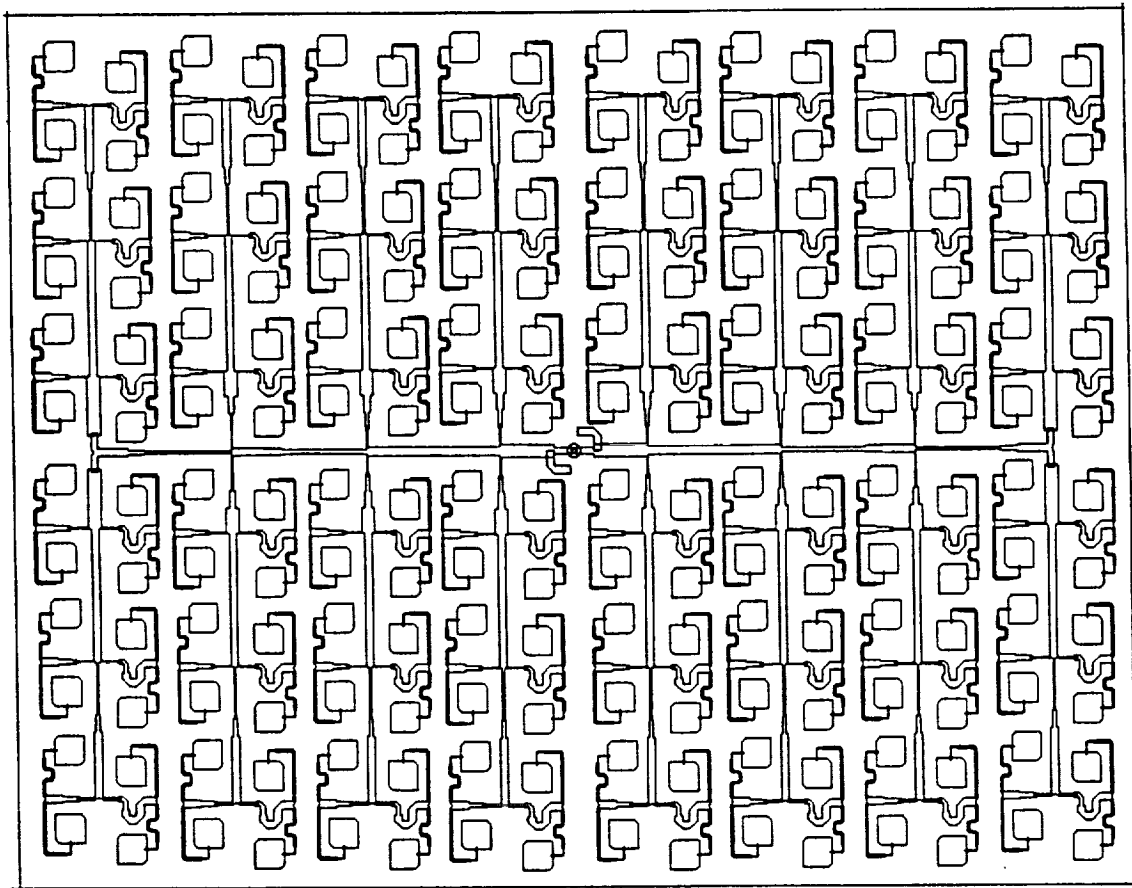


Fig. 7. The full array design.

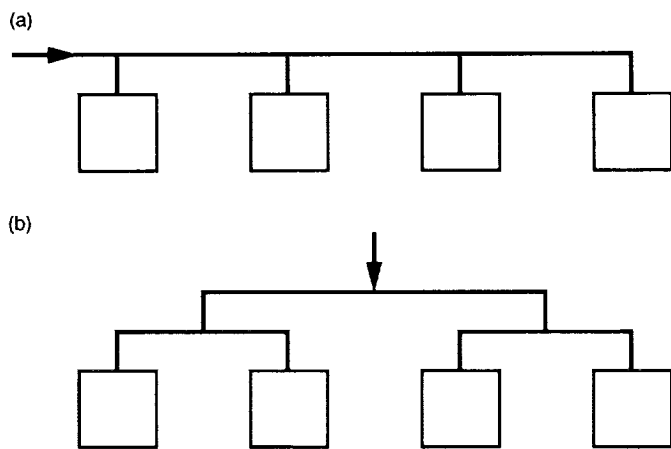


Fig. 8. Comparison of the microstrip (a) series-feed and (b) parallel-feed techniques with a four-element array.

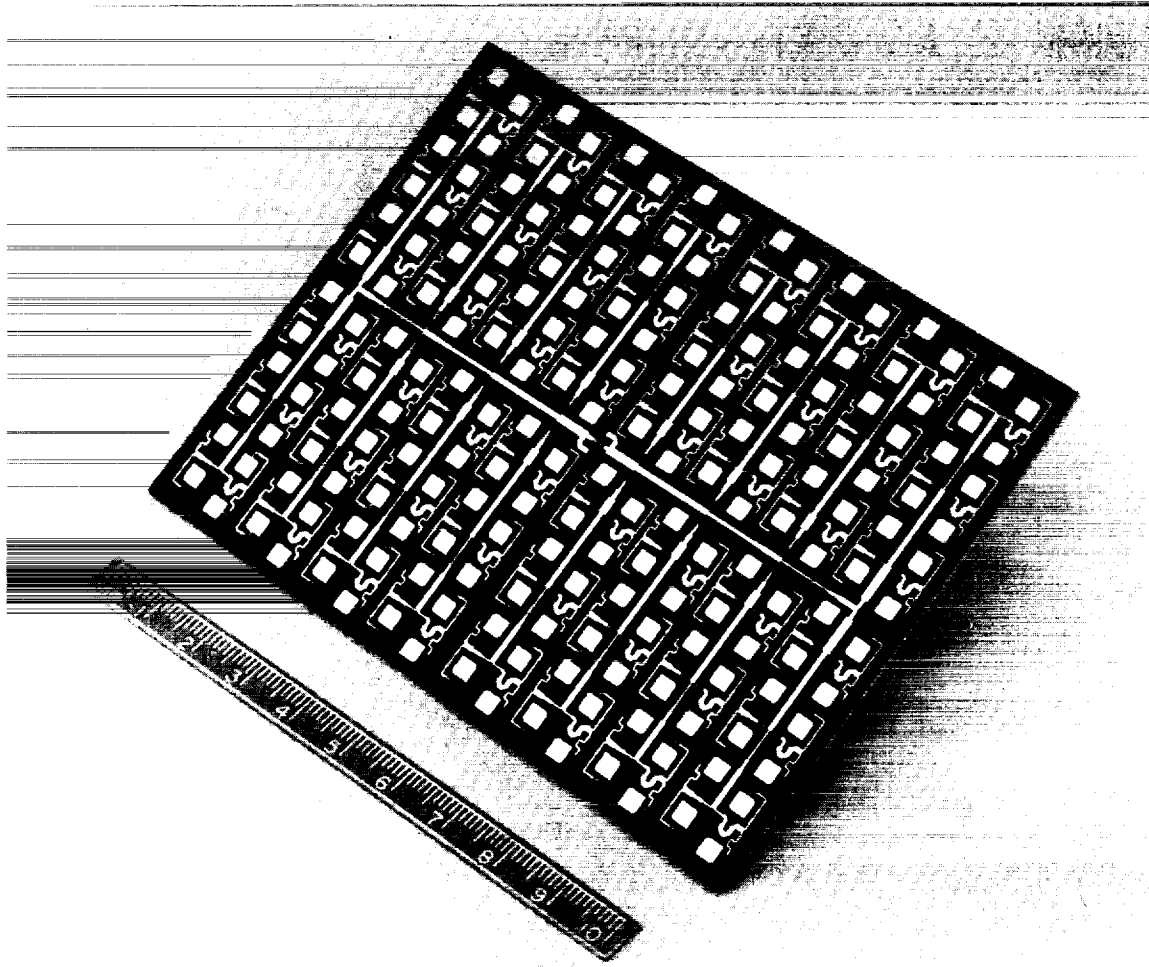


Fig. 9. The 32-GHz full array with 192 patch elements.

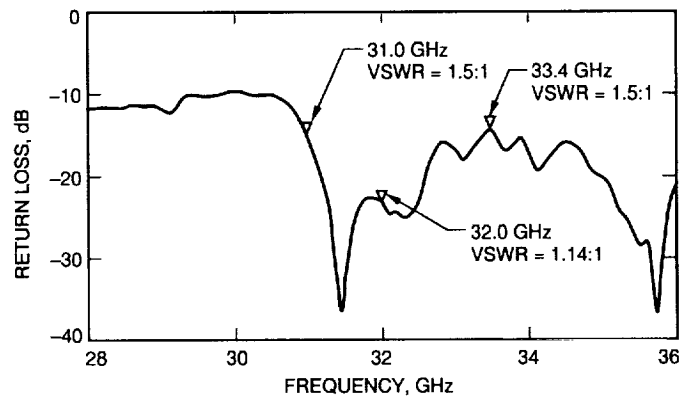


Fig. 10. Measured input return loss of the full array.

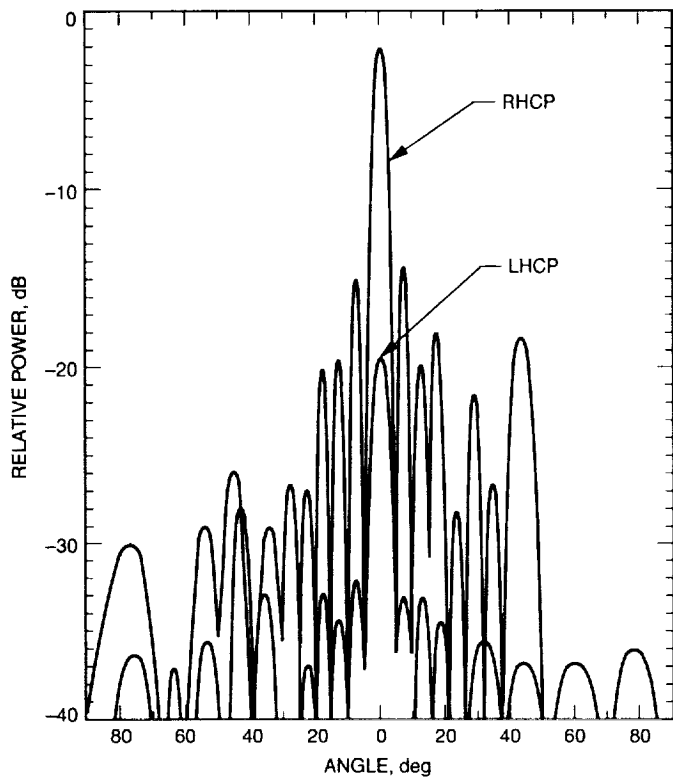


Fig. 11. Measured narrow-beam principal-plane pattern of the full array.

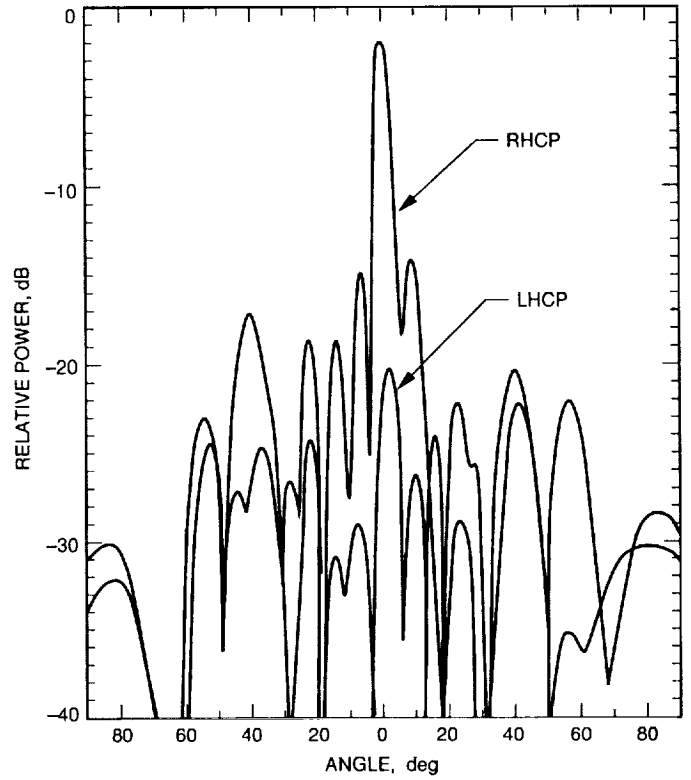


Fig. 12. Measured broad-beam principal-plane pattern of the full array.