Design of A Dielectric-rod Fed Cassegrain Reflector Antenna
공학석사학위논문

유전체봉 피드를 이용한 카세그레인 반사경 안테나 설계

Design of A Dielectric-rod Fed Cassegrain Reflector Antenna

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Abstract

In this thesis, design a monopulse reflector antenna operating in w-band. A sum pattern is realized using a dielectric rod placed at feed center, an elevation difference pattern is realized using two dielectric rods placed at top and bottom, and an azimuth difference pattern is realized using two dielectric rods placed at right and left. The monopulse comparator consists of a power divider and an E/H-plane bend, respectively, when implementing the elevation channel and the azimuth channel. The sub- and main reflectors are designed for the Cassegrain optics. The gain of the antenna is also changed through the shaping and expansion of the reflector. The monopulse feed is over 20% bandwidth centered around the operating frequency in the W-band, the designed feed
has the sum pattern, -10 dB beamwidth of 46° -48°, gain of 16.4-17.4 dB, cross polarization ratio of 25-27 dB, phase error of ±41° withing -10dB beamwidth. In the case of the difference pattern, the designed feed has a gain of 15.2 dB, a null depth greater than -50 dB, and a gain taper of 4.5 dB over -10 dB. In the synthesis of the monopulse feed and the monopulse comparator, the reflection coefficient of the azimuthal direction channel, the reflection coefficient of the high-order directional channel, and the reflection coefficient of the sum channel were checked and satisfied -20 dB or less. In the case of using the reflector antenna, the sum channel gain pattern had a gain of 39.2dB and an Side love level of -20dB. The difference channel gain pattern had a gain of 36.4dB and a null depth of -49.8dB.

* A thesis for the degree of Master in February 2018.
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I. Introduction

A Prime-Focus reflector and a cassegrain reflector monopulse antenna are widely used to track targets such as rockets, satellites, and missiles [1]. In the design of a monopulse antenna using a reflector, the size of the feed opening must be minimized, and the sub-reflector must be irradiated so that the sum pattern of the reflector and the difference pattern are optimally realized at the same time [2]-[3]. As a cassegrain monopulse reflector feed structure, a multimode single aperture feed [4]-[5], four horn feeds [6], four dielectric rod feeds [7], five dielectric rod feeds [8]-[9] Has been used. The complexity of the monopulse feedforward selection scheme and the monopulse comparator for the difference pattern formation must be considered. In the case of the monopulse feed using five dielectric rods, the monopulse comparator is advantageous in that it can be easily implemented because the sum pattern feeder and the difference pattern feeder are separated. In addition, since the sum and difference patterns of the feed can be independently optimized, the monopulse antenna characteristic can be finally obtained. It is difficult to design a 5-element monopulse feed with low-gain elements such as dipoles, waveguide openings, and patches so that the sum of the pattern beam width and the difference pattern beam width is optimized when irradiating the reflector of the cassette. In this case, the sum pattern beam width is too wide and the car pattern side lobe is too large. This problem can be solved by using a small, high gain dielectric rod emitter. The existing results on 5-element dielectric rod monopulse feeds are very limited and are not presented in
detail. In this thesis, I designed a catagrain reflector monopulse feed using five dielectric rods and analyzed the frequency characteristics and the detailed characteristics at the center frequency. We have employed the widely-used commercial simulation software Microwave Studio™ by CST, the accuracy of which has been proven in numerous publications worldwide.

![Fig. 1.1 Operation concept of monopulse radar](image)

The main structure of such a monopulse antenna is composed of a cassegrain structure (double reflector), a feed and a comparator. In this paper, the optimum design of each component is made according to the target design specification. The Cassegrain antenna was simulated by deriving the appropriate analytical method based on optical theory and by using the feed pattern as it is. The composition of this paper is as follows. In Chapter II, III and IV sections, design process and design results of monopulse feed, monopulse comparator and reflector antenna are presented. Finally, Chapter V summarizes the design results and presents conclusions.
II. Design of 5 Dielectric-rod Feed

2.1 monopulse feed structure

Figure 2.1 (a) shows the structure of a monopulse feed designed in this thesis. It consists of a dielectric rod emitter and a square waveguide block for feeding it. One dielectric rod at the center for generating the sum pattern, one dielectric rod at the top and bottom for forming the high angle pattern, and one dielectric rod at the left and one side respectively for forming the azimuthal difference pattern. The dielectric rod emitter was fed into a square waveguide. A square waveguide was chosen because square waveguide rather than circular waveguide is convenient to implement orthogonal mode converter, magic T junction, power divider, and bend, which are monopulse comparator parts. Figure 2.1 (b) is a longitudinal cross-sectional view of the feed, showing the dielectric taper of the square waveguide insert, the taper of the radially-sectioned dielectric rod (a linearly decreasing diameter).

![Fig. 2.1 Structure of monopulse feed (a) 3D view (b)Side view](image)
Table 2.1 Design specifications of the monopulse antenna

<table>
<thead>
<tr>
<th>Specification</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency</td>
<td>W-Band</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>10%</td>
</tr>
<tr>
<td>Reflection coefficient</td>
<td>&gt; -20dB</td>
</tr>
<tr>
<td>Polarization</td>
<td>Linear</td>
</tr>
<tr>
<td>Sum pattern -10dB beamwidth</td>
<td>±23°</td>
</tr>
<tr>
<td>Sum pattern cross-pol level</td>
<td>&gt; -25dB</td>
</tr>
<tr>
<td>Difference pattern taper</td>
<td>-5dB</td>
</tr>
<tr>
<td>Null depth</td>
<td>&gt; -50dB</td>
</tr>
</tbody>
</table>

2.2 monopulse feed design

Figure 2.2 shows the optimal design dimensions of the monopulse feed designed in this thesis. The diameter of the portion exposed outside the waveguide in the dielectric rod was linearly decreased to 0.249 wavelength at the end. By adjusting the length of the radiation part, the beam width design goal of the sum pattern was satisfied. In the 5-element monopulse feed design, the spacing of the two elements for the car pattern is made as small as possible so that the lattice lobe is generated at a large angle from the feed center axis. The size of the lattice is designed to be smaller than the size of the main lobe by 10 dB or more by the gain pattern of the radiating element. The gain pattern of the dielectric rod for the central sum pattern is affected by the height of the dielectric rod for the surrounding car pattern. Therefore, the height of the dielectric pattern for the car pattern is appropriately made smaller than the height of the dielectric pattern for the patterned pattern, so that the sum pattern beam width and the difference pattern beam width are simultaneously designed to be close to the optimum value.
A monopulse comparator waveguide connected to a monopulse feed is energized with air. Therefore, the diameter of the waveguide insertion portion of the dielectric rod is linearly reduced to reduce the reflection coefficient.

Figure 2.2 shows the direction of the electric field vector at the excitation of the input waveguide port (numbered in the figure) for phono pulse pattern implementation. By combining the second and third ports in the reverse direction using a monopulse comparator, the high angle pattern is synthesized and the fourth and fifth ports are combined in the reverse direction to synthesize the azimuth difference pattern. Table 2.2 shows the design specifications of the

---

**Fig. 2.2** Structure dimension

**Table 2.2** Optimum dimensions of the proposed feed

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>a1</td>
<td>3.97(\lambda_0)</td>
<td>b1</td>
<td>0.413(\lambda_0)</td>
</tr>
<tr>
<td>a2</td>
<td>1.32(\lambda_0)</td>
<td>b2</td>
<td>0.658(\lambda_0)</td>
</tr>
<tr>
<td>a3</td>
<td>2.65(\lambda_0)</td>
<td>b3</td>
<td>2.11(\lambda_0)</td>
</tr>
<tr>
<td>a4</td>
<td>0.627(\lambda_0)</td>
<td>b4</td>
<td>0.156(\lambda_0)</td>
</tr>
<tr>
<td>a5</td>
<td>5.31(\lambda_0)</td>
<td>b5</td>
<td>0.188(\lambda_0)</td>
</tr>
<tr>
<td></td>
<td></td>
<td>b6</td>
<td>0.249(\lambda_0)</td>
</tr>
</tbody>
</table>
proposed monopulse feed. The proposed monopulse feed is designed in the following order. First, Rexolite (dielectric constant of 2.53, loss tangent of 0.0007), which is widely used as a dielectric rod material, was selected. The well-known method is applied to the design of the dielectric rod antenna [10]. The diameter of the dielectric rod (equal to the square waveguide width) was 0.658 in consideration of the propagation of the HE\textsubscript{11} mode in the dielectric rod and the square waveguide size suitable for monopulse comparator design.

![Fig. 2.3 Feed port number and electric field direction of monopulse feed](image)

Fig. 2.3 Feed port number and electric field direction of monopulse feed

Figure 2.4 shows the correlation between the center element and the peripheral element. The reflection coefficient when there is only the center element for forming the sum channel and the case where there are no remaining elements and the case where the center element for forming the sum channel and the surrounding four elements are arranged in the same manner are compared. It is confirmed that the reflection coefficient increases to -20dB or more in the low
band and the high band based on the center frequency when the center element only exists and it is confirmed that the frequencies of the low band and the high band are lower when both the center element and the peripheral element are lowered. It is confirmed that the peripheral device affects the reflection coefficient of the central element.

Fig. 2.4 Reflection coefficient according to presence or absence of the difference channel element

Figure 2.5 shows the reflection coefficient according to the dielectric taper of the sum channel forming device. The tapered dielectric rods were inserted in the waveguide, and the dielectric coils were positioned to the opening of the waveguide, and the reflection coefficients of the tapered dielectric rods filled in the waveguide were compared. It was confirmed that the taper of the dielectric rod affects the impedance matching of the reflection coefficient.
Fig. 2.5 Characteristics of the impedance taper in the designed feed

Figure 2.6 shows the transfer coefficients for each device. It was confirmed that all of them satisfied -30 dB or less in the required frequency range.

Fig. 2.6 Transmission coefficients of the designed feed

Figure 2.7 shows the reflection coefficient of an optimized monopulse feed. Is an optimum designed reflection coefficient of the device for forming a sum channel and the devices for forming a neighboring channel.
Figure 2.7 Reflection coefficients of the designed feed

Figure 2.8 shows the result of changing the gain pattern depending on the size and the presence of the dielectric rod element of the monopulse feed. Figure 2.8 shows the gain pattern of the sum channel device only after removing the channel device. Figure 2.8 shows the result of removing the sum channel element and showing the gain pattern of the next channel element. It was confirmed that the gain of the sum channel changes depending on the presence or absence of the difference channel element and the SLL of the difference channel element changes depending on the presence or absence of the sum channel element. Figure 2.9 shows the case where the length of the secondary channel element is equal to that of the sum channel element. Figure 2.10 shows the case where the sum channel element length is equal to the length of the secondary channel element. It is confirmed that the gain of the sum channel decreases. It was confirmed that the length of the sum channel element determines the gain. Figure 2.11 shows the case where the length of all elements is 1.5 times. All devices were lengthened and the gain of the sum channel increased. Figure 2.12 shows that the gain of
the sum channel is reduced when the length of all elements is 0.5 times. Figure 2.13 shows the gain pattern when the length of the subchannel is 0.5 times.

**Fig. 2.8** Gain pattern according to presence or absence of sum channel and difference channel element

**Fig. 2.9** Increasing the differential channel element
Fig. 2.10 Reduction of sum channel elements

Fig. 2.11 Increase sum channel and difference channel

Fig. 2.12 Decrease sum channel and difference channel
Figure 2.13 0.5 times reduction of the difference channel element

Figure 2.15 shows the gain and gain patterns at the center and at the center frequency $f_0$ of the designed feed. The maximum gain of the sum channel at the total interface and the magnetic interface is 17.4dB, and the sidelobe level at -19.4dB is -18.5dB at the interface. The gain of the secondary channel is 15.2dB, and the zero point depth is less than -70dB.

(a) 
(b)

Fig. 2.14 E-plane pattern (a) full angle (b) Accuracy
Figure 2.15 H-plane pattern (a) full angle (b) Accuracy

Figure 2.16 shows the three-dimensional gain pattern at the center frequency $f_0$ and presents the sum channel, the azimuthal azimuthal pattern of the main channel, and the azimuthal pattern of the main channel. The gain pattern of the sum channel is shown as a round shape. The symmetrical gain pattern is shown. The elevation angle and azimuth pattern also show a symmetrical gain pattern.
Fig. 2.16 3D radiation pattern (a) Sum (b) E-plane (b) H-plane
Table 2.3 Summary of results

<table>
<thead>
<tr>
<th>E-plane sum pattern</th>
<th>Gain</th>
<th>17.4dB</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Side lobe level</td>
<td>-18.5dB</td>
</tr>
<tr>
<td>E-plane diff pattern</td>
<td>Gain</td>
<td>15.2dB</td>
</tr>
<tr>
<td></td>
<td>Side lobe level</td>
<td>-10.0dB</td>
</tr>
<tr>
<td></td>
<td>Null depth</td>
<td>-74.6dB</td>
</tr>
<tr>
<td>H-plane sum pattern</td>
<td>Gain</td>
<td>17.4dB</td>
</tr>
<tr>
<td></td>
<td>Side lobe level</td>
<td>-19.4dB</td>
</tr>
<tr>
<td></td>
<td>Null depth</td>
<td>-80.5dB</td>
</tr>
<tr>
<td>H-plane sum pattern</td>
<td>Gain</td>
<td>15.2dB</td>
</tr>
<tr>
<td></td>
<td>Side lobe level</td>
<td>-11.6dB</td>
</tr>
</tbody>
</table>

Figure 2.17 shows the phase center of the dielectric feed at center frequency $f_0$. The center of the phase at the center frequency $f_0$ is located at $5.10\lambda_0$ in the $Z^+$ direction from the opening of the waveguide horn. Table 2.4 presents the phase center of the frequency-dependent feed. The higher the frequency, the more the phase center of the feed is away from the opening of the waveguide horn to the $Z^+$ axis.

Table 2.4 Phased center of each frequency

<table>
<thead>
<tr>
<th>Frequency</th>
<th>$Zc$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.9$f_0$</td>
<td>3.85$\lambda_0$</td>
</tr>
<tr>
<td>0.95$f_0$</td>
<td>4.44$\lambda_0$</td>
</tr>
<tr>
<td>$f_0$</td>
<td>5.10$\lambda_0$</td>
</tr>
<tr>
<td>1.05$f_0$</td>
<td>5.69$\lambda_0$</td>
</tr>
<tr>
<td>1.1$f_0$</td>
<td>6.21$\lambda_0$</td>
</tr>
</tbody>
</table>
Fig. 2.17 Phased center of a $f_0$ designed feed

Figure 2.18 shows the gain pattern of the main polarized wave and the crossed polarized wave by frequency. The phase center is a frequency-dependent pattern based on $f_0$. Figure 2.19 shows the phase pattern for each frequency. Table 2.5 summarizes the results of the gain and phase patterns. Table 2.5 shows the gain and phase patterns for each frequency. It was confirmed that $\theta_E$ $\theta_H$ per frequency was the same at $0.9f_0$ and $1.1f_0$ at $48^\circ$ and the same at $0.95f_0$, $f_0$, and $1.05f_0$ at $48^\circ$. It is confirmed that $\Delta \phi_E$ and $\Delta \phi_H$ per frequency are the smallest at the center frequency and become larger as the distance from the center frequency increases.
Fig. 2.18 Gain patterns versus frequencies of the designed feed. Frequency from 0.9 times to 1.1 times the center frequency by 0.05 times the center frequency

(a) $0.9f_0$ (b) $0.95f_0$ (c) $f_0$ (d) $1.05f_0$ (e) $1.1f_0$
Fig. 2.19 Phase patterns versus frequencies of the designed feed. Frequency from 0.9 times to 1.1 times the center frequency by 0.05 times the center frequency

(a) 0.9\(f_0\) (b) 0.95\(f_0\) (c) \(f_0\) (d) 1.05\(f_0\) (e) 1.1\(f_0\)
Table 2.5 Summarizes the performance of the monopulse feed

<table>
<thead>
<tr>
<th>$f$(GHz)</th>
<th>$G_{\text{max}}$ (co-p ol) (dB)</th>
<th>$G_{\text{max}}$ (X-pol) (dB)</th>
<th>$\theta_E$ (deg)</th>
<th>$\theta_H$ (deg)</th>
<th>$\Delta \phi_E$ (deg)</th>
<th>$\Delta \phi_H$ (deg)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.9$f_0$</td>
<td>16.4</td>
<td>-9.2</td>
<td>48</td>
<td>48</td>
<td>41</td>
<td>26</td>
</tr>
<tr>
<td>0.95$f_0$</td>
<td>17.0</td>
<td>-10.4</td>
<td>46</td>
<td>46</td>
<td>20</td>
<td>13</td>
</tr>
<tr>
<td>$f_0$</td>
<td>17.4</td>
<td>-10.3</td>
<td>46</td>
<td>46</td>
<td>4.3</td>
<td>-5</td>
</tr>
<tr>
<td>1.05$f_0$</td>
<td>17.6</td>
<td>-10.9</td>
<td>46</td>
<td>46</td>
<td>-18.4</td>
<td>-21</td>
</tr>
<tr>
<td>1.1$f_0$</td>
<td>17.4</td>
<td>-7.99</td>
<td>46</td>
<td>48</td>
<td>-35.7</td>
<td>-37.3</td>
</tr>
</tbody>
</table>
Ⅲ. Design of Monopulse Comparator

3.1 Orth-Mode Transducer Design

For this monopulse comparator, a square waveguide OMT operating at W-band is designed. After an optimum design using a commercial software (CST's Microwave Studio), The following sections provide detailed descriptions of the proposed OMT. The OMT is electrically a four-port. The physical structure, however, corresponds to that of a three-way controller since the physically common gate combines the two electrically independent polarization directions. The ideal OMT is a lossless system. The OMT is a passive and thus reversible (reciprocal) system. This means that an OMT can be used both in the transmission branch of an antenna system and in the receiving branch of an antenna system. The ideal OMT is adapted to all gates. From this property of the ideal OMT, an important requirement arises for the design of a real OMT. In the case of a poor adaptation of the gates over the bandwidth considered, a large part of the energy is reflected into the coupling-in port. This reflected energy is referred to as a reflection loss (Return Loss) and is expressed in dB as a ratio to the coupled energy. Figure 3.1 shows the shape of the proposed broadband orthogonal mode converter (OMT) in this paper. The proposed orthogonal mode transducer is composed of a common port composed of a square waveguide, a linear port having a horizontal polarization component and a straight line disposed in a straight line with the common port, and a vertical port and a side port formed in a vertical direction.
The OMT as we shown in Fig. 3.1 has 4 ports, port 1 is input witch has two mode, mode 1 is vertical polarization and mode 2 horizontal polarization. Port 3 is output for vertical polarization, port 1 is output for horizontal polarization. Simulated result was shown by Fig. 3.2.

Fig. 3.1 Orth-Mode Transducer

Fig. 3.2 OMT simulation result (a) Reflection coefficient (b) Transmission coefficient
3.2 E-plane Power Divider design

The power divider is one of the building blocks of waveguide components such as multiplexers, transmitters and receiver modules and array feed networks. Wideband power dividers with high output power ratio are often required in the feed network of the array of waveguide radiators. Full waveguide-band 1:1 power dividers have been investigated for H-plane and E-plane geometries. Literature on wideband power dividers with high output ratio is quite limited. Power dividers operating at the full waveguide band can be designed using a Y-junction and impedance matching steps. Figure 3.3 shows the structure of the proposed power divider. The input waveguide port 1 and two output waveguide port 2, 3 have the same dimension of a and b.

Fig. 3.3 E-Plane Power Divider structure
3.3 E/H-plane bend design

The E- and H-plane right-angle bends are designed. To reduce discontinuity reactance, the bend is mitered, where part of the bend region is removed. We used two-step mitering. Figure 3.5 shows the structure of E- and H-plane right-angle bends. The best result are shown in the figure 4.14.

![Fig. 3.5 H-Plane bend structure](image)
Fig. 3.6 H-Plane bend reflection coefficient

Fig. 3.7 E-Plane bend structure

Fig. 3.8 E-Plane bend reflection coefficient
3.4 Monopulse comparator synthesis

Figure 3.9 shows the structure of the comparator in the azimuthal direction of the difference channel. Port 3 is input port 1 and port 2 is output port. Ports 1 and 2 have a phase difference of 90 degrees and are composed of one E-plane divider, four E-plane bend, and two H-plane bend. Figure 3.10 shows the reflection coefficient of port 3 and confirmed that it satisfies -20dB or less. Figure 3.11 shows the transmission coefficients of port 3, ports 1 and 2, and shows a transmission coefficient of -3 dB. Figure 3.12 shows the transfer phase difference between port 1 and port 2, and the phase difference between the two ports is 90 degrees.

Fig. 3.9 Structure of difference channel azimuth pattern
Fig. 3.10 Azimuth channel reflection coefficient

Fig. 3.11 Azimuth channel transmission coefficient

Fig. 3.12 Azimuth channel phase
Figure 3.13 shows the structure of the comparator in the elevation direction of the difference channel. Port 3 is input port 1 and port 2 is output port. Ports 1 and 2 have a phase difference of 90 degrees and are composed of one E-plane divider, four E-plane bend, and two H-plane bend. Figure 3.14 shows the reflection coefficient of port 3 and confirmed that it satisfies -20dB or less. Figure 3.15 shows the transmission coefficients of port 3, ports 1 and 2, and shows a transmission coefficient of -3 dB. Figure 3.16 shows the transfer phase difference between port 1 and port 2, and the phase difference between the two ports is 90 degrees.

Fig. 3.13 Structure of difference channel elevation pattern
Fig. 3.14 Elevation channel reflection coefficient

Fig. 3.15 Elevation channel transmission coefficient

Fig. 3.16 Elevation channel phase
This section deals with a combination of the monopulse comparator and the monopulse feed discussed in previous chapter. The monopulse feed and a monopulse comparator to be combined to give the phase difference. As the calculation of data too much, computer configuration problems can not use The Microwave Studio™ 2015 by CST to calculate, so use of The Design Studio™ 2015 by CST to simulation.

Fig. 3.17 Monopulse feed composition
Fig. 3.18 Difference channel elevation reflection coefficient

Fig. 3.19 Difference channel azimuth reflection coefficient
Fig. 3.20 Sum channel horizontal reflection coefficient

Fig. 3.21 Sum channel vertical reflection coefficient
IV. Monopulse Reflector Antenna

4.1 Reflector Antenna Design

The structure of the proposed Cassegrain antenna is shown in Figure 4.1. The proposed figure has a circular symmetry structure with respect to the z-axis. In general, a Cassegrain antenna consists of two focal points, a main reflector and a sub-reflector. The main reflector forms a parabolic surface by a virtual focal point. The sub-reflector consists of a hyperboloid formed from a mathematical relationship between the focal point coincident with the phase center of the actual feed horn and the virtual focal point. A double reflector structure in which the convex portions of the main reflector and the sub-reflector are arranged in parallel in the same direction is called a cassegrain structure. In the cassegrain design, each parameter is determined by considering the antenna target design specification and the characteristics of the feed.

![Fig. 4.1 Cassegrain antenna shape](image)

- 34 -
The main reflector can be represented by the distance from Dm to the focal point, and the angle from the feed axis (z axis) of the main reflector to the edge of the reflector. The focal distance (Fm) of the reflector can be easily calculated using the diameter and the depth of the reflector (H0).

![Diagram of main reflector section](image)

**Fig. 4.2** Main reflector section

The sub-reflector serves to make the spherical wave of the narrow beam width emitted from the actual focal point look like a spherical wave of the wide beam width generated at the virtual focal point. When the line dividing the distance between the actual focal point and the virtual focal point by the same length is drawn and the x-axis is viewed, the surface of the sub-reflecting plate can be obtained by rotating the hyperbola along the z-axis.
The reflector antenna can be drawn using the following equations 4.1 to 4.11.

\[
\phi_v = 2\tan^{-1}\left(\frac{D_m}{4F_m}\right) \tag{4.1}
\]

\[
\frac{1}{\tan\phi_v} + \frac{1}{\tan\phi_r} = 2\frac{F_c}{D_s} \tag{4.2}
\]

\[
1 - \frac{\sin\frac{1}{2}(\phi_v - \phi_r)}{\sin\frac{1}{2}(\phi_v + \phi_r)} = 2\frac{L_v}{F_c} \tag{4.3}
\]
The vertical distance from the focal point to an arbitrary point on the paraboloid of the reflection plate and back to the axis perpendicular to the focal point is always constant regardless of the point on the paraboloid, and the spherical wave radiated from the focal point is reflected on the parabolic surface and is converted into a plane wave. In addition, all path lengths from the focal point to the reflector and to the aperture are the same and twice the fringe distance (Fm). Under ideal conditions, the maximum gain of the reflector antenna is determined by the aperture area Ap of the reflector as shown in equation (4.11).
However, in case of ideal conditions, it is necessary to consider radiation efficiency, opening taper efficiency, overflow efficiency, reach efficiency and so on, since it is a value that does not consider overflow or resistance loss at all.

\[ G_{\text{max}} = \frac{4\pi}{\lambda^2} A_p \]  

(4.12)

\[ G = \epsilon_{ap} D_0 = \epsilon_{ap} \frac{4\pi}{\lambda^2} A_p \]  

(4.13)

The effective efficiency \((\epsilon_{ap})\) can be expressed as a product of the aperture distribution efficiency \((\epsilon_t)\), the overflow efficiency \((\epsilon_s)\), the shielding efficiency \((\epsilon_b)\) by the sub-reflecting mirror, and other accompanying efficiencies \((\epsilon_e)\). Other efficiencies include irregular surface errors due to dimensional errors in actual fabrication, reflector phase errors, feed phase errors and scattering at the corner of the sub-reflector, and other efficiency losses are typically less than 1dB.

The sub-reflector serves to make the spherical wave of the narrow beam width emitted from the actual focal point look like a spherical wave of the wide beam width generated at the virtual focal point. When we look at the x axis by dividing the distance between the actual focus and the virtual focus by the same length, we can obtain the surface of the sub reflector by rotating the hyperbola, which can be represented by the equation (4.9), along the z axis.
Table 4.1 Optimum dimensions of the proposed reflector

<table>
<thead>
<tr>
<th>Dimension</th>
<th>Value</th>
<th>Dimension</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$D_m$</td>
<td>$37.6\lambda$</td>
<td>$L_r$</td>
<td>$6.57\lambda$</td>
</tr>
<tr>
<td>$D_s$</td>
<td>$6.26\lambda$</td>
<td>$L_v$</td>
<td>$4.64\lambda$</td>
</tr>
<tr>
<td>$F_m$</td>
<td>$29.8\lambda$</td>
<td>$\theta_v$</td>
<td>$25.0^\circ$</td>
</tr>
<tr>
<td>$F_c$</td>
<td>$11.2\lambda$</td>
<td>$\theta_r$</td>
<td>$34.84^\circ$</td>
</tr>
<tr>
<td>$F_s$</td>
<td>$18.5\lambda$</td>
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<td></td>
</tr>
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</table>

Table 4.2 Summary of results

<table>
<thead>
<tr>
<th></th>
<th>Gain</th>
<th></th>
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</thead>
<tbody>
<tr>
<td>Sum channel</td>
<td>39.2dB</td>
<td></td>
</tr>
<tr>
<td>Side lobe level</td>
<td>-19.9dB</td>
<td></td>
</tr>
<tr>
<td>Del channel</td>
<td>36.4dB</td>
<td></td>
</tr>
<tr>
<td>Side lobe level</td>
<td>-25.8dB</td>
<td></td>
</tr>
<tr>
<td>Null depth</td>
<td>-60dB</td>
<td></td>
</tr>
</tbody>
</table>

Fig4.4. Reflector application structure
Fig. 4.5 Reflector gain pattern

Fig. 4.6 Co-pol and x-pol sum pattern

Fig. 4.7 Co-pol and x-pol del pattern
4.2 Main/Sub reflector shaping

Figure 4.4 shows the method of forming the sub-reflector curved surface. Figure 4.6 shows the gain pattern of the sum channel and the difference channel through the process of forming the sub-reflector curved surface. The gain pattern of the difference channel showed a maximum gain of 0.2 dB and a side love level (SLL) increased from -25.4 dB to -28.2 dB. Figure 3 shows the gain pattern of the sum channel. The molding gain decreased from 39.8dB to 39dB and the SLL (Side Love Level) changed from -19.9dB to –20.1dB. Shaping by changing the value of $\phi_r$ from 34.84° to 38°.

\[\text{Fig. 4.8 Structure of sub reflector shaping}\]
Fig. 4.9 In MWS Structure of sub reflector shaping

Fig. 4.10 Gain pattern (a) Sum (b) Difference
Figure 4.7 shows a method of increasing the gain by enlarging the sub-reflector size. Figure 4.8 shows the gain pattern of the sum channel and the difference channel after the extension of the sub-reflector surface. The gain pattern of the difference channel showed 0.6dB increase in maximum gain from 36.3dB to 36.9dB and a change in side love level (SLL) from -25.4dB to -17.0dB. Figure 3 shows the gain pattern of the sum channel. The gain at the time of expansion decreased from 39.1dB to 38.7dB and the SLL (Side Love Level) changed from -20.1dB to -16.5dB.

Fig. 4.11 Structure of sub reflector expansion
Figure 4.9 shows the molding method of the main reflector curved surface. Figure 4.10 shows the gain pattern of the sum channel and the difference channel through the process of forming the sub-reflector curved surface. The gain pattern of the difference channel showed a -3 dB decrease in maximum gain from 33.5 dB to 36.5 dB and a change in side love level (SLL) from -25.3 dB to -17.8 dB.
Figure 3 shows the gain pattern of the sum channel. The molding gain increased from 39.0dB to 40.7dB and the SLL (Side Love Level) changed from -20.0dB to -20.3dB.

Fig. 4.14 Structure of main reflector shaping

Fig. 4.15 In MWS structure of main reflector shaping
Fig. 4.16 Gain pattern (a) Sum (b) Difference

Table 4.3 Coordinates of the point to be shifted

<table>
<thead>
<tr>
<th>$x$</th>
<th>$Z_{old}$</th>
<th>$Z_{new}$</th>
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</thead>
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<tr>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>$1.56 \lambda_0$</td>
<td>0.18$\lambda_0$</td>
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<td>$3.13 \lambda_0$</td>
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<td>0.156$\lambda_0$</td>
</tr>
<tr>
<td>$4.72 \lambda_0$</td>
<td>0.184$\lambda_0$</td>
<td>0.278$\lambda_0$</td>
</tr>
<tr>
<td>$6.27 \lambda_0$</td>
<td>0.329$\lambda_0$</td>
<td>0.429$\lambda_0$</td>
</tr>
<tr>
<td>$7.83 \lambda_0$</td>
<td>0.514$\lambda_0$</td>
<td>0.611$\lambda_0$</td>
</tr>
<tr>
<td>$9.40 \lambda_0$</td>
<td>0.742$\lambda_0$</td>
<td>0.830$\lambda_0$</td>
</tr>
<tr>
<td>$10.97 \lambda_0$</td>
<td>1.009$\lambda_0$</td>
<td>1.081$\lambda_0$</td>
</tr>
<tr>
<td>$12.53 \lambda_0$</td>
<td>1.319$\lambda_0$</td>
<td>1.373$\lambda_0$</td>
</tr>
<tr>
<td>$14.10 \lambda_0$</td>
<td>1.67$\lambda_0$</td>
<td>1.705$\lambda_0$</td>
</tr>
<tr>
<td>$15.67 \lambda_0$</td>
<td>2.06$\lambda_0$</td>
<td>2.073$\lambda_0$</td>
</tr>
<tr>
<td>$17.24 \lambda_0$</td>
<td>2.49$\lambda_0$</td>
<td>2.498$\lambda_0$</td>
</tr>
<tr>
<td>$18.80 \lambda_0$</td>
<td>2.96$\lambda_0$</td>
<td>2.968$\lambda_0$</td>
</tr>
</tbody>
</table>
V. Conclusion

In this thesis, I designed a monopulse reflector antenna that operates in W-band. In the monopulse feed implementation using a dielectric rod, the monopulse comparator can be easily implemented because the feed pattern portion and the feed pattern portion are separated. In addition, since the sum and difference patterns of the feeds can be optimized independently, the characteristics of a monopulse antenna can be finally obtained. Monopulse feeds using five dielectric rods have not been well studied. In the case of monopulse comparator configuration, the synthesis of the high angle and azimuth direction circuit is symmetrically used and the preparation is excellent. The designed feeds have a -10dB beamwidth of 46°-48°, a gain of 16.4-17.4dB, a cross-polarization rate of 25-27dB, a beam width of -10dB within a 20% bandwidth around the W-band has a phase error characteristic of ±41°. The gain pattern has a gain of 15.2dB at the center frequency, a zero depth of 50dB or more, and a gain taper of 4.5dB in the sum pattern-10dB beam width. In the synthesis of the monopulse feed and the monopulse comparator, the reflection coefficient of the azimuthal direction channel, the reflection coefficient of the high-order directional channel, and the reflection coefficient of the sum channel were checked and satisfied -20 dB or less. In the case of using the reflector antenna, the sum channel gain pattern had a gain of 39.2dB and an Side love level of -20dB. The difference channel gain pattern had a gain of 36.4dB and a null depth of -49.8dB. The antenna proposed in this thesis is considered to be useful in the w-band tracking system.
참고 문헌


[8] C. Kumar, V. S. Kumar, and V. V. Srinivasan, “Design aspects of a compact dual band feed using dielectric rod antennas with multiple element
