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Triple corner reflector antenna and its performance in H-plane

K.T. Mathew, J. Jacob, S. Mathew and U. Raveendranath

An experimental investigation of a novel corner reflector antenna with improved performance is reported. The corner reflector antenna has been structurally modified by attaching two more reflector elements. The parameters such as the primary corner angle, position, width and angle of the secondary elements have significant effects on the H-plane radiation characteristics of the antenna. Certain symmetric configurations of this triple corner reflector (TCR) antenna provide sharp beams with a gain of the order of 3 dB over that of the conventional corner reflector (CR) antenna.

Introduction: Among different types of reflector antennas, the corner reflector antenna is remarkable due to its simple structure. The beam modification of corner reflector antennas is well described in literature [1, 2]. Joseph et al. [3] reported a dual corner reflector (DCR) antenna, with a third element attached to one of the primary elements of the corner reflector (CR) antenna, with beam shaping and tilting properties. In this Letter, the design and performance of a triple corner reflector (TCR) antenna having higher directivity and enhanced axial power compared with those of a CR antenna is presented.

The TCR antenna is presented.

Antenna design and experimental setup: The schematic diagram of the TCR antenna is shown in Fig. 1. The antenna is constructed by attaching two additional conducting elements C and D to either side of a CR antenna (Fig. 1). The distance l, angle β and width w of the secondary elements can be varied. In the experimental we used an HP 8350B sweep oscillator, HP 8410C network analyser and a plotter. A standard horn was used as the transmitter. The antenna under test (AUT) was used as the receiver. The experiment was conducted in the X-band.

Experimental details and results: The secondary elements C and D of width w were attached symmetrically at a distance l from the apex of the antenna. Keeping the distance l constant the radiation patterns for various angles β and width w of the secondary elements were analysed. This was repeated for various values of l. The values of β, w and l were optimised for maximum on-axis power. A TCR antenna can also be designed by folding the primary elements of a conventional CR antenna for optimum l, w and β. The experiment was repeated for primary corner angles 60°, 90° and 120°. The experimental iteration shows that the TCR antenna with the configuration α = 120°, β = 80° and w = 1.5 A provides the highest gain and the HPBW is half the value of the corresponding conventional 90° CR antenna. A comparison of the H-plane radiation patterns of the optimum TCR antenna and the conventional 90° CR antenna is given in Fig. 2. The sidelobe level for the TCR antenna is below −13 dB. Table 1 shows a comparison of the antenna parameters of the prominent TCR antenna configurations with a 90° CR antenna. Configurations other than the optimum show comparatively higher sidelobe levels.

![Fig. 1 Schematic diagram of TCR antenna](image)

Fig. 1 Schematic diagram of TCR antenna

<table>
<thead>
<tr>
<th>Antenna configuration</th>
<th>α = 90°</th>
<th>α = 120°</th>
<th>β = 60°</th>
<th>β = 45°</th>
<th>β = 80°</th>
<th>l = 1A</th>
<th>l = 1A</th>
<th>l = 1A</th>
<th>w = 1.5A</th>
<th>w = 1.5A</th>
<th>w = 1.5A</th>
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</thead>
<tbody>
<tr>
<td>Relative gain [dB]</td>
<td>2.5</td>
<td>2.6</td>
<td>3.4</td>
<td>0</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>VSWR</td>
<td>1.27</td>
<td>1.19</td>
<td>1.33</td>
<td>1.31</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>HPBW</td>
<td>16°</td>
<td>23°</td>
<td>19°</td>
<td>38°</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table 1: Comparison of relative gain, half power beam width and VSWR of 90° CR antenna and various TCR antenna configurations

Conclusion: A novel corner reflector antenna has been designed and its characteristics compared with the conventional 90° CR antenna. An attractive feature of this antenna is the enhanced axial power and reduced HPBW. In the H-plane it has a gain 3 dB greater than that of a conventional antenna. Thus a TCR antenna can effectively replace a conventional CR antenna.

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References

Fig. 2 Rate-1/2 systematic binary convolutional encoder with feedback

code with the smallest error path multiplicity was chosen. The code search results for the modulation index of 0.5 are listed and compared to rate-2/3 codes for 2-PSK/4-PSK [1] and two-level codes [4] for 8-PSK.

Table 1: Comparison of code search results

<table>
<thead>
<tr>
<th>Proposed rate-2/3 codes</th>
<th>Rate-2/3 codes [1]</th>
<th>Rate-2/3 codes for 8-PSK</th>
</tr>
</thead>
<tbody>
<tr>
<td>dree</td>
<td>dree</td>
<td>dree</td>
</tr>
<tr>
<td>gamma</td>
<td>gamma</td>
<td>gamma</td>
</tr>
<tr>
<td>2</td>
<td>4</td>
<td>4</td>
</tr>
<tr>
<td>3</td>
<td>02</td>
<td>4</td>
</tr>
<tr>
<td>4</td>
<td>2</td>
<td>4</td>
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<td>5</td>
<td>3</td>
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<tr>
<td>6</td>
<td>35</td>
<td>4</td>
</tr>
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<td>7</td>
<td>155</td>
<td>4</td>
</tr>
<tr>
<td>8</td>
<td>231</td>
<td>4</td>
</tr>
<tr>
<td>9</td>
<td>367</td>
<td>4</td>
</tr>
</tbody>
</table>

In Table 1, the error path multiplicity $N_{err}$ was calculated by counting error paths from the all-zeros path and $H^*$ denotes the coefficient vector $(h_1,h_2,...,h_s)$ of the systematic binary convolutional encoder of Fig. 2. The coefficients are written in octal form. In Table 1, the ACGs were calculated by the equation

$$\gamma = 10 \log_{10} \frac{d_{ree}}{d_{ref}} \text{[dB]} \quad (3)$$

where $d_{ree}$ is the MSE of the reference scheme, uncoded QPSK, which is 2. As shown in Table 1, the proposed rate-1/2 codes require smaller constraint length to obtain the same value of $d_{ree}$ which reduces the decoder complexity. Also, the lower code rate reduces the decoder complexity.

In calculating the ACGs in Table 1, the spectral efficiency of the proposed codes was assumed to be 2 bit/symbol. However, considering the slight decrease in spectral efficiency owing to $C_s$, small differences arise between the real ACGs and those listed in Table 1, e.g., if a rate-7/8 punctured binary convolutional code [5] of which $d_{ref}$ is 3 is used as $C_s$, a spectral efficiency of 1.875 bit/symbol results. If the constraint length of $C_s$ is 6, the ACG is 6.8 dB compared to 7.1 dB listed in Table 1. In spite of the slight decrease n ACG, larger code gain is still guaranteed.

**Conclusion:** Two-level codes for 2-FSK/4-PSK for $h = 0.5$ are presented. The encoding and decoding structures are based on multilevel encoding and multistage decoding. The results show additional ACGs of 0.8-2.2 dB and lower decoder complexity than adavon and Wolf’s rate-2/3 trellis codes for 2-FSK/4-PSK, and also show additional ACGs of 1.1-2.1 dB to Pottie and Taylor’s 2-level codes for 8-PSK.

**References**

1. IEEE 1996 8 March 1996
2. 3. Y. Raveendranath, J. Jacob and K. T, Mathew
3. Complex permittivity measurement of liquids with coaxial cavity resonators using a perturbation technique

**Index terms:** Cylindrical waveguide, Resonators

Open-ended coaxial cavity resonators for measuring the complex permittivity of liquids are presented. Measurements are based on the cavity perturbation technique. Resonator I operates in the frequency range 0.6-7 GHz, and resonator II operates in the frequency range 5-14 GHz. The complex permittivity of water is measured. Results are compared with those obtained using other methods.

**Introduction:** A survey of the literature reveals different techniques adopted to measure the complex permittivity of materials. Among the available methods, the cavity perturbation technique has a unique advantage owing to its many attractive features [1, 2]. A rectangular waveguide cavity resonator with capillary insert [3] is used for measuring the dielectric parameters of liquids, but in that particular band only. Although the broadband coaxial probe techniques [4, 5] give excellent results, they require elaborate calibration procedures and significant errors may occur owing to improper contact at the probe-material interface. In the present method, open-ended coaxial resonators are used for measuring the complex permittivity of liquids.

**Design and theoretical considerations:** A section of a coaxial line with one end open acts as a resonator when its length is an odd multiple of a quarter wavelength. The open end is a potential source of radiation loss resulting in a low quality factor. To overcome this difficulty, the outer conductor is elongated beyond the end of the inner conductor such that it forms a cylindrical waveguide operating below cutoff for a given resonance frequency. Along the axis of the waveguide there is a removable centre conductor. For a given length of the centre conductor a series of resonance frequencies can be obtained. The diameter of the centre conductor is properly selected for impedance matching. Design parameters of the resonators used are shown in Table 1.

**Table 1:** Design parameters of cavity resonators

<table>
<thead>
<tr>
<th>Dimension, mm</th>
<th>Resonator I (Cutoff frequency 7 GHz)</th>
<th>Resonator II (Cutoff frequency 15 GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Length of resonator, $L$</td>
<td>250</td>
<td>140</td>
</tr>
<tr>
<td>Inner radius of resonator, $b$</td>
<td>16.25</td>
<td>6.64</td>
</tr>
<tr>
<td>Radius of centre conductor, $a$</td>
<td>3.0</td>
<td>1.36</td>
</tr>
<tr>
<td>Distance from centre conductor to feed loop, $c$</td>
<td>3.0</td>
<td>0.75</td>
</tr>
<tr>
<td>Feed loop inner dimension, $\Delta_l \times \Delta f$</td>
<td>2 x 8</td>
<td>1 x 2</td>
</tr>
<tr>
<td>Lengths of centre conductor, $L$</td>
<td>168,113.88</td>
<td>120,100</td>
</tr>
</tbody>
</table>

Beginning with Maxwell’s equations, the standing wave field components of the resonant TEM mode are obtained by combining the forward and backward propagating waves as

$$E_0^r = \frac{G}{\rho} e^{j\beta s} + \frac{B}{\rho} e^{-j\beta s} \quad (1a)$$

$$H_0^r = \frac{A}{\eta} e^{j\beta s} + \frac{B}{\eta} e^{-j\beta s} \quad (1b)$$

**References**


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where η is the free space impedance and A and B are constants. The boundary conditions require that $E_x^0$ must vanish at $Z = 0$ and $Z = L$, where $L$ is the length of the centre conductor. The first condition gives $A = B$, and the second gives the resonance condition.

When the sample is introduced into the cavity, the complex relative frequency shift is given by Waldron [6] as

$$\frac{\partial \Omega}{\Omega} = \frac{(\varepsilon_r - 1) V_0 E_0}{\pi L (\varepsilon_r - \varepsilon_0) V_0}$$

where $E_0$, $H_0$, and $E_r$ are the fields in the unperturbed cavity, $E$ and $H$ are the fields in the interior of the sample, $\varepsilon_r = \varepsilon_r' + j\varepsilon_r''$, and $\mu_r = \mu_r' - j\mu_r''$, and $V_0$ and $V$ are the volumes of the cavity and sample, respectively.

When the dielectric sample is introduced at the position of maximum electric field, the second term of the numerator of eqn. 2 is insignificant. The denominator of eqn. 2 represents the total electromagnetic energy stored in the cavity. Substituting the value of the $E$ field from eqn. 1 into eqn. 2 and solving this, we obtain

$$\frac{\delta \Omega}{\Omega} \approx \frac{(\varepsilon_r - 1) V_0}{\pi L (\varepsilon_r - \varepsilon_0) V_0}$$

The volume of the sample $V = \pi r^2 (b - a)$, where $r$ is the radius of the capillary tube, $a$ is the radius of the inner conductor and $b$ is the inner radius of the outer conductor. The real and imaginary parts of the complex frequency shift are given by

$$\frac{\delta \Omega}{\Omega} \approx \frac{(f_2 - f_1) L (b + a)}{2\pi r^2}$$

Equating eqns. 3 and 4, we obtain

$$\frac{(\varepsilon_r' - 1) = \frac{L (b + a)}{2\pi r^2}}{f_2}$$

$$\varepsilon_r'' = \frac{L (b + a)}{2\pi r^2} \left( \frac{1}{Q_1} - \frac{1}{Q_2} \right)$$

where $f_1$ and $f_2$ are the resonance frequencies of the cavity loaded with an empty capillary tube and a tube with sample liquid, respectively. $Q_1$ and $Q_2$ are the corresponding quality factors.

Fig. 1 Schematic diagram of open-ended coaxial resonator

Measurement and results: The experimental setup consists of an HP 8510B network analyser, an S-parameter test set, a sweep oscillator and an instrument computer. The resonator is connected to one port of the S-parameter test set. The liquid under study is taken in a capillary tube, made of low loss fused silica. The diameter of the tube is properly selected to satisfy the perturbation condition [3]. The capillary tube is introduced into the cavity through the slot on its outer wall and held at the position of maximum electric field corresponding to each resonance frequency, by sliding it along the slot, as shown in Fig. 1. The resonance frequency and $Q$-factor of the cavity loaded with an empty capillary tube are measured. Then the capillary tube is filled with the sample liquid and sealed. The measurements are repeated. For resonator I sharp resonance peaks are available from 0.6 to 7GHz. However, resonance peaks are not very sharp at low frequencies for resonator II. So resonator II is used for the measurements in the frequency range 7 to 17GHz.

Fig. 2 Real part of complex permittivity of water against frequency

X-X Wei and Sridhar [4]
A-A Mathew et al. [3]
O-O present method

Fig. 3 Imaginary part of complex permittivity of water against frequency

X-X Wei and Sridhar [4]
A-A Mathew et al. [3]
O-O present method

The real and imaginary parts of the complex permittivity of water are measured at 27°C (Figs. 2 and 3). The results for water are compared with those obtained by Mathew et al. [3] and Wei et al. [4].

Conclusion: In this Letter, a new broadband technique for the measurement of complex permittivity of liquids is presented. It is particularly useful for liquids available in small quantities. This method is comparatively simple and easy to operate. Measurement can also be extended to solids.

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References