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LA THÈSE A ÉTÉ MICROFILMÉE TELLE QUE NOUS L'AVONS REÇUE
STUDY OF MICROSTRIP TRIANGULAR RESONATORS

by

Misel Cuhaci

A thesis submitted to the School of Graduate Studies, University of Ottawa, in partial fulfilment of the requirements for the degree of Master of Applied Science.

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ABSTRACT

The radiation properties of microstrip triangular resonators are investigated. Both the radiation pattern and the radiation quality factor \( Q_r \) are derived analytically utilizing the current distribution on the resonator and numerical techniques. The different loss mechanisms of the microstrip resonators are described. The radiation losses were determined experimentally. Computed and experimental results are in reasonable agreement. The equivalent lumped circuit parameters of the resonator as well as some possible applications are also discussed.
LIST OF CONTENTS

ACKNOWLEDGEMENTS ........................................ ii
ABSTRACT ............................................... iii

CHAPTER I  Introduction

1.1  Introduction .......................................... 1
1.2  Literature survey of the theoretical and experimental studies on the losses in microstrip resonators 4
1.3  Thesis summary ..................................... 7

CHAPTER II  Field and Current Distribution in The Triangular Resonator

2.1  Introduction .......................................... 8
2.2  Electrical and magnetic properties of the triangular resonator ........................................ 8
2.3  Electromagnetic fields in terms of the Hertz vector .... 12
2.4  The current distribution in the triangular resonator .... 14

CHAPTER III  Radiation Properties of The Triangular Resonator

3.1  Introduction .......................................... 19
3.2  Far zone electric field components .......................... 19
3.3  Radiated power ....................................... 21
3.4  Numerical results ................................... 22
3.5  Conclusion of the theoretical analysis .................. 23

- iv -
APPENDIX C  Computer program to calculate the radiation loss of a triangular resonator .......................... 76
APPENDIX D  Computer program to find the radiation pattern of a triangular resonator .......................... 81
REFERENCES .................................................................................................................. 85

List of Tables .............................................................................................................. vii
List of Figures ........................................................................................................... viii
## LIST OF TABLES

<table>
<thead>
<tr>
<th>TABLE</th>
<th>Description</th>
<th>PAGE</th>
</tr>
</thead>
<tbody>
<tr>
<td>3.1</td>
<td>Computed radiation data for the triangular resonator</td>
<td>24</td>
</tr>
<tr>
<td>5.1</td>
<td>Experimental results for the triangular resonator ($\varepsilon_r=10$)</td>
<td>47</td>
</tr>
<tr>
<td>5.2</td>
<td>Experimental results for the triangular resonator ($\varepsilon_r=2.5$)</td>
<td>47</td>
</tr>
<tr>
<td>5.3</td>
<td>Experimental results for the disc resonator ($\varepsilon_r=10$)</td>
<td>48</td>
</tr>
</tbody>
</table>
LIST OF FIGURES

FIGURE                                      PAGE
1.1  Microstrip Structure                    2
1.2  Triangular Microstrip Resonator         2
2.1  Magnitude of the electric field of the fundamental resonant mode (TM_{0,1,-1}) in a triangular resonator. The dotted lines in the triangular plane indicate magnetic field lines. (from ref.2) 11
2.2  Resonator and Image Configuration       15
3.1  Q^{-1}_r versus (h/f), computed results 25
3.2  Computed values of Q^{-1}_r versus both relative and effective permittivity 27
4.1  Set-up for radiation pattern measurement 29
4.2  Triangular resonator theoretical and experimental E-plane radiation pattern 32
4.3  E-plane cross-polarization pattern      33
4.4  Triangular resonator theoretical and experimental H-plane pattern 34
4.5  H-plane cross-polarization pattern      35
5.1  Jig and its cover used in Q_r measurements 43
5.2  Set-up to measure both the reflected and the transmitted power 45
5.3  Measured Q^{-1}_r versus h . f for circular and triangular resonators 49
5.4  Measured Q^{-1}_r versus h . f for triangular resonator 50
<table>
<thead>
<tr>
<th>FIGURE</th>
<th>Description</th>
<th>PAGE</th>
</tr>
</thead>
<tbody>
<tr>
<td>6.1</td>
<td>Possible Microstrip coupling configurations</td>
<td>52</td>
</tr>
<tr>
<td>6.2</td>
<td>Lumped element representation of a one port cavity</td>
<td>54</td>
</tr>
<tr>
<td>6.3</td>
<td>Lumped element representation of a two port cavity</td>
<td>57</td>
</tr>
<tr>
<td>6.4</td>
<td>(Gap Capacitance/substrate thickness) versus (Gap width/substrate thickness)</td>
<td>61</td>
</tr>
<tr>
<td>7.1</td>
<td>Experimental and computed values of $Q_r^{-1}$ versus (h,f)</td>
<td>64</td>
</tr>
<tr>
<td>7.2</td>
<td>Experimental and computed values of $Q_r^{-1}$ versus ($\varepsilon_e$) and ($\varepsilon_r$)</td>
<td>65</td>
</tr>
<tr>
<td>A.1</td>
<td>Rectangular and circular coordinate system used in the theoretical analysis</td>
<td>67</td>
</tr>
</tbody>
</table>
CHAPTER I

INTRODUCTION

1.1 INTRODUCTION

Stripline or microstrip structures are being used more and more at microwave and millimeter wave frequencies. The structure consists of a conductive strip deposited on a dielectric substrate backed by a conductive ground plane Fig. 1.1. The electromagnetic properties depend on the geometry, such as strip width $w$, strip thickness $t$, and substrate thickness $h$, the dielectric and magnetic properties of the substrate and the medium (usually air) above it.

This planar configuration is small and consequently lighter; circuits fabricated with this technique are cheaper than waveguide or coaxial systems. Also, beam-lead active or passive devices can be bonded directly to the conductor strip on the dielectric substrate [1]. Ultimately, the planar configuration lends itself to the fabrication of integrated circuits.

However, the applications of Microwave Integrated Circuits (MIC) as transmission medium at microwave frequencies are limited by a number of factors such as line attenuation due to dielectric and resistive losses, breakdown effects at high power levels and radiation from discontinuities. The last effect which constrains the microwave circuit
Fig. 1.1 - Microstrip Structure

Fig. 1.2 - Triangular Microstrip Resonator
designer but represents a desired property for an antenna engineer, constitutes one of the major problems in MIC circuit design. The analysis of the power radiated from these structures is complicated by the fact that two different dielectric media (air and substrate) are present simultaneously.

Resonators are one of the basic building blocks for the microwave engineer, as they are used to design filters, circulators, oscillators etc. A resonant element in MIC generally presents a discontinuity to the transmitted signal. The poor quality factor of the microstrip resonators, which is due to dielectric and resistive losses, is further aggravated by the power radiated from the discontinuities.

There are numerous papers analysing the performance of different resonators, e.g. the rectangular, disc, ring, etc. The equilateral triangle and the disc resonators are of special interest because of the 120° symmetry. For the triangular resonator Fig. 1.2, some studies have been carried out by Hoefer, Helszajn and James [2,3] to determine the electric and magnetic fields and to calculate the resonant frequencies of the different modes.

In this thesis, the triangular microstrip resonator will be analysed. The theoretical expression for the radiation will be derived and compared with experimental results.
1.2 LITERATURE SURVEY OF THE THEORETICAL AND EXPERIMENTAL STUDIES ON THE LOSSES IN MICROSTRIP RESONATORS.

In February 1960, Lewin [4] published a paper describing the "Radiation From Discontinuities In Strip-line". In this article a method to derive the theoretical radiation expressions from the known current distribution is presented. A rigorous analysis of the radiation from microstrip was found to be difficult because of the complex transmission medium and therefore the author of the paper made some simplifying assumptions. The method so developed was used to study transmission line configurations like the open-circuit, matched co-axial termination, mismatched termination, parallel post resonator formed by two parallel posts and the right-angle corner. His results were approximate but it was shown later by other authors [2,5,6,9] that there was reasonable agreement with measurements.

In April 1969, Denlinger [5] gave some experimental results showing the relation between microstrip parameters and the fractional amount of power radiated from a resonant line. He verified that the percentage of the power radiated depends on the thickness of the substrate. The measurements agreed with the values Lewin calculated for the case of an open-circuit microstrip line. He also made radiation measurements on the disc resonator and derived some empirical conclusions.

radiation loss from open-circuit resonators appeared in October 1973. In this article he discussed the conductor, dielectric and radiation losses of a rectangular resonator and showed that an extension to Lewin's analysis could yield the radiation loss of this geometry. He also suggested using the structure as an antenna element after considering the radiation patterns of different modes.

In September 1970, Easter and Roberts [9] described the calculations of the radiation loss of a half-wavelength open-circuit resonator and gave results for a range of different effective permittivities. They used Lewin's method [4] to derive the theoretical expressions which were then checked experimentally. The same authors published another paper [10] in April 1971 where they compared different types of resonators, namely the ring, open-circuit linear, short-circuit linear and the "hairpin" or "U" shaped resonator. They showed that the hairpin resonator radiates less than the straight resonator.

Sobol's study of the radiation conductance of open-circuit microstrip [11] appeared in November 1971. He calculated the radiation conductance at the open-circuit end of a microstrip stub and used the result to find the total loss of that configuration. He found good agreement between his results and Denlinger's results [5].

In June 1975, Belohoubek and Denlinger [12] studied the loss mechanism of microstrip resonators. They derived formulae for the radiation Q of open-circuit microstrip resonators based on Lewin's work [4]. They described the effect of radiation on the overall circuit Q as a function of the characteristic impedance, frequency, dielectric constant
and substrate thickness. These results showed that radiation losses at higher frequencies dominate over the conductor and dielectric losses; they also showed that microstrip lines on substrates with a low dielectric constant radiate more.

There are also papers published on microstrip antennas using the rectangular or circular resonators as basic elements. Howell [13] discussed these two configurations in his paper which appeared in January 1975. He gave design procedures for linearly and circularly polarized antennas and also presented the measured radiation patterns.

The work of Morel et al. [15] on the theoretical investigation of the circular disc antenna appeared in April 1976. The far-field total radiated power and power losses were calculated. These expressions were derived considering air as the dielectric separating the conducting plates. Also, approximate results were derived for the various dielectric materials. At the same time Walton et al [16] investigated experimentally the radiation characteristics of circular discs. They studied the antenna elements with regard to their driving point impedances and far-field radiation patterns.

In November 1976, Derneryd [14] studied linearly polarized microstrip antennas. He characterized the square and rectangular elements by their radiation pattern, directivity and equivalent admittance. Also, he gave a design procedure for open-circuit halfwave resonators and for arrays of resonators.
1.3 THESIS SUMMARY

The object of this thesis is to study theoretically and experimentally the radiation loss of an equilateral triangular microstrip resonator. An approach similar to Lewin's [14] was taken to derive the theoretical expressions for the radiation pattern and the radiation quality factor, $Q_r$. Numerical methods were used to obtain the theoretical results which were then compared to the measurements. The measured radiation Q-factor of the triangular structure was also compared with that of a circular resonator.

The resonators were studied experimentally as lossy microstrip circuit elements and their unloaded Q-factor, $Q_0$, was measured in each sample for both the open and shielded case. Neglecting the losses due to imperfect shielding, the discrepancy in the two measurements was attributed to the radiation loss.

Different coupling configurations to the microstrip resonators were described and the equivalent circuit for the resonator was studied.

To conclude, the use of a triangular resonator as a circuit and antenna element were briefly discussed.
CHAPTER II

FIELD AND CURRENT DISTRIBUTION IN THE TRIANGULAR RESONATOR

2.1 INTRODUCTION

Microstrip structures are studied, in general, by replacing the planar configuration with an equivalent waveguide model which consists of electric walls to replace the conductors, and of magnetic walls on the sides to contain the electric field. Solutions for the fields and cut-off frequencies for a waveguide with an equilateral triangular cross-section are described by Schelkunoff [20]. These results were applied to predict the resonances of triangularly shaped microstrip resonators by Hoefer, Helszajn and James [2,3]. Their work, as summarized in the first section below, provides a basis for the radiation analysis. To simplify the analysis, the Hertz vector will also be introduced in this Chapter.

2.2 ELECTRICAL AND MAGNETIC PROPERTIES OF THE TRIANGULAR RESONATOR

The TM\textsubscript{m,n,λ} \textsuperscript{*} mode field patterns in the triangular resonator filled with a dielectric material of permittivity \( \varepsilon \), permeability \( \mu \) and with no variation of the field along the thickness of the structure are given by:

\(*\) The TM-modes are defined such that the magnetic field possesses components in the xy-plane only, since the field equations have been derived originally in a waveguide of triangular cross-section.
\[ E_z = A_{m,n,\lambda} T(x,y)_{m,n,\lambda} \quad \ldots \quad (2.1) \]

\[ H_x = \frac{j}{\omega \mu} \frac{\delta E_z}{\delta y} \quad \ldots \quad (2.2) \]

\[ H_y = -\frac{j}{\omega \mu} \frac{\delta E_z}{\delta x} \quad \ldots \quad (2.3) \]

\[ H_z = E_x = E_y = 0 \quad \ldots \quad (2.4) \]

where

i) \( A_{m,n,\lambda} \) is an arbitrary amplitude factor

\[ T(x,y)_{m,n,\lambda} = \cos \left( \frac{2\pi x}{\sqrt{3}a} + \frac{2\pi}{3} \lambda \right) \cos \left( \frac{2\pi (m-n)y}{3a} \right) \]

\[ + \cos \left( \frac{2\pi x}{\sqrt{3}a} + \frac{2\pi}{3} m \right) \cos \left( \frac{2\pi (n-\lambda)y}{3a} \right) \]

\[ + \cos \left( \frac{2\pi x}{\sqrt{3}a} + \frac{2\pi}{3} n \right) \cos \left( \frac{2\pi (\lambda-m)y}{3a} \right) \quad (2.5) \]

where "a" is the side of the triangle

ii) \( m, n, \lambda \) are integers which are never zero simultaneously and satisfy the condition

\[ m + n + \lambda = 0 \quad (2.6) \]
The resonant frequencies are:

\[ f_{m,n,l} = \frac{1}{2} \left( \varepsilon \mu \right)^{-\frac{1}{2}} \left[ \left( \frac{4}{3a} \right)^2 \left( m^2 + mn + n^2 \right) \right]^{\frac{1}{2}} \] ....(2.7)

with \( \varepsilon = \varepsilon_0 \varepsilon_r \) and \( \mu = \mu_0 \mu_r \)

where \( \varepsilon_r \) and \( \mu_r \) are the relative dielectric constant and the relative permeability.

Fig. 2.1 shows the \( TM_{0,1,-1} \) mode which is associated with the lowest resonant frequency in a triangular resonator.

The energy stored in the structure is expressed by:

\[ W_{\text{stored}} = \frac{c}{2} A^2 \sum_{m,n,l} \int_S \left[ T(x,y) \right]_m^n \text{d}S \] ....(2.8)

where \( c \) is the microstrip thickness and \( S \) is the surface area of the triangle.

The above expressions are general. For the particular case analysed here, \( m = 1, n = 0 \) and \( \lambda = -1 \). This is the case of the dominant \( TM_{1,0,-1} \) mode for which the expressions (2.1) to (2.8) can be written as:

\[ E_z(x,y,z) = A_{1,0,-1} \left[ 2 \cos \left( \frac{2\pi x}{\sqrt{3a}} + \frac{2\pi}{3} \right) \cos \frac{2\pi y}{3a} + \cos \frac{4\pi y}{3a} \right] \] ....(2.9)

\[ H_x(x,y,z) = -j A_{1,0,-1} \varepsilon_0 \left[ \cos \left( \frac{2\pi x}{\sqrt{3a}} + \frac{2\pi}{3} \right) \cos \frac{2\pi y}{3a} + \sin \frac{4\pi y}{3a} \right] \] ....(2.10)
Fig. 2.1 - Magnitude of the electric field of the fundamental resonant mode $\text{TH}_{0,1,-1}$ in a triangular resonator. The dotted lines in the triangular plane indicate magnetic field lines. (from ref.2)
\[ H_y(x,y,z) = j\sqrt{3} A_{1,0,-1} \xi_o \left[ \sin \left( \frac{2\pi x}{3a} \right) + \frac{2\pi}{3} \cos \frac{2\pi}{3a} y \right] \quad \ldots(2.11) \]

with \[ \xi_o = \sqrt{\frac{\epsilon_o}{\mu_o}} = \frac{1}{120\pi} S \quad \ldots(2.12) \]

The fundamental frequency of resonance is:

\[ f_{o,1,0,-1} = \frac{2c}{3a \sqrt{\epsilon_r}} \quad \ldots(2.13) \]

where \( c = 3 \times 10^8 \) m/s

The stored energy, after carrying out the integration becomes:

\[ W_{\text{stored}} = \frac{3\sqrt{3}}{16} \epsilon_0 \lambda^2 A_{r,0,-1}^2 \quad \ldots(2.14) \]

2.3 ELECTROMAGNETIC FIELDS IN TERMS OF THE HERTZ VECTOR

The Hertz vector, \( \vec{H} \), is a mathematical quantity, which can be used to express other field quantities by means of simple operators. The derivation for the following equation can be found in the literature [21].

\[ \vec{H} = \frac{-i}{4\pi} \sqrt{\frac{1}{\epsilon_o}} \frac{1}{k\epsilon_r} \int \frac{\exp(-jk\sqrt{\mu_r\epsilon_r} r)}{r} \hat{J} \, dr \quad \ldots(2.15) \]
where \( \mathbf{j} \) is the vector representing the current distribution on the radiating element, in A/m².

\( \tau \) is the volume surrounding the source, in m³.

\( r \) is the radial distance in meter from the source to a point in space.

\( \varepsilon_r \) and \( \mu_r \) are respectively the relative permittivity and permeability of the space.

\[ k = \frac{2\pi}{\lambda} \text{ in m}^{-1} \quad \text{(wave number)} \]

\( \lambda \) is the wavelength in the free space.

In terms of the Hertz vector, the electric and magnetic fields, at any point in the space containing the source, can be written as

\[ \mathbf{E} = \nabla (\mathbf{E} \cdot \mathbf{n}) + k^2 \varepsilon_r \mu_r \mathbf{H} \quad \quad \ldots (2.16) \]

\[ \mathbf{H} = jk \sqrt{\frac{\varepsilon_0}{\mu_0}} \varepsilon_r (\nabla \times \mathbf{E}) \quad \quad \ldots (2.17) \]

It can be seen from the above expressions that the knowledge of the complete or entire current distribution in a structure, with a defined geometry, is sufficient to derive the electric and magnetic fields radiating into space.
2.4 THE CURRENT DISTRIBUTION IN THE TRIANGULAR RESONATOR

The current in a microstrip structure is the vectorial sum of two currents, namely the conduction and the dielectric polarization current.

a) The conduction current density;

In perfect conductors, the current flow at high frequencies can be considered as a true surface current due to the "skin effect". Inside the conductor there is no displacement current, and a time varying magnetic field does not exist. On the surface, the conduction current density is,

\[ \mathbf{J}_c = \mathbf{n} \times \mathbf{H} \]  \hspace{1cm} (2.18)

where \( \mathbf{n} \) is the unit vector orthogonal to \( \mathbf{H} \), the magnetic field.

To ensure low resistive losses, for all practical microstrip configurations, the conductors are a few skin depths thick. This implies that the displacement current is entirely under the metallization and the conductors can be replaced by a current sheet for analytical purposes. Also the ground plane can be replaced by the image of the top conductor Fig. 2.2. The latter is a valid transformation since generally the ground plane is much larger in surface area than the triangular conductor.
Fig. 2.2 - Resonator and Image Configuration
The expression for the magnetic field vector, assuming \( \vec{H} \) is constant along the z-axis is:

\[
\vec{H} = H_x(x,y,h)\hat{a}_x + H_y(x,y,h)\hat{a}_y.
\]  

and \( \vec{n} = \hat{a}_z \)

where \( \hat{a}_x, \hat{a}_y, \hat{a}_z \) are unit direction vectors.

The error introduced by the above approximation is negligible as long as the thickness \( h \), of the substrate is much smaller than the guided wavelength \( \lambda_g \). The conduction current density \( \vec{J}_{cr} \), on the resonator, at \( z = h \), is obtained from (2.18).

\[
\vec{J}_{cr} = -H_y(x,y,h)\hat{a}_x + H_x(x,y,h)\hat{a}_y \quad \ldots (2.20)
\]

The image of the current density \( \vec{J}_{cr} \), at \( z = -h \), is,

\[
\vec{J}_{ci} = H_y(x,y,-h)\hat{a}_x - H_x(x,y,-h)\hat{a}_y \quad \ldots (2.21)
\]

b) The dielectric polarization current density;

Maxwell's equation can be modified to emphasize the contribution of the polarization current to the total current (4).
\[ \mathbf{\nabla} \times \mathbf{\hat{H}} = j \frac{k}{120\pi} \mathbf{\hat{E}} + (j \frac{k}{120\pi} (\varepsilon - 1) \mathbf{\hat{E}}) \]  \hspace{1cm} (2.22)

where \[ \frac{1}{120\pi} = \sqrt{\frac{\varepsilon_0}{\mu_0}} \]

The above expression shows that the dielectric polarization acts as an impressed current density, \( \mathbf{\hat{J}}_P \):

\[ \mathbf{\hat{J}}_P = j \frac{k}{120\pi} (\varepsilon - 1) \mathbf{\hat{E}} \]  \hspace{1cm} (2.23)

In a microstrip structure, \( \varepsilon \) is a function of \( z \), for \( z > h \), \( \varepsilon = \varepsilon_{\text{air}} \), and for \( z \leq h \), \( \varepsilon = \varepsilon_r \) of the dielectric. The analysis will be simplified by taking \( \varepsilon \) as the effective permittivity, \( \varepsilon_e^* \), of the substrate material. Due to this approximation, as \( \varepsilon_{\text{air}} < \varepsilon_e < \varepsilon_r \), the contribution of the polarization current is under-valued in between the conductors and over-valued above. The polarization current density is independent of \( z \), and for simplicity, it is taken at \( z = 0 \).

\[ \mathbf{\hat{J}}_P(x,y,z) = \mathbf{\hat{J}}_P(x,y,0) = j \frac{k}{120\pi} (\varepsilon_e - 1) E_z(x,y,0) \mathbf{\hat{a}}_z \]  \hspace{1cm} (2.24)

* The expression used to evaluate \( \varepsilon_e \) is given on page 22.
c) The total current density is:

\[ J_T = J_{CR} + J_{CI} + J_P \]  \hspace{1cm} (2.25)

\[ J_T = \left\{ \begin{array}{c}
- H_y(x,y,h) + H_y(x,y,-h) \\
H_x(x,y,h) - H_x(x,y,-h) \\
\frac{jk}{120 \pi} (\varepsilon_e - 1) E_z(x,y,0)
\end{array} \right\} \hat{a}_x + \hat{a}_y + \hat{a}_z \]  \hspace{1cm} (2.27)

The above expression will be used to derive the radiation properties of the triangular resonator.
CHAPTER III

RADIATION PROPERTIES OF THE TRIANGULAR RESONATOR

3.1 INTRODUCTION

Two different methods can be used to derive the expression for the radiated power from a microstrip structure. The first is the waveguide approach where the geometry of the structure defines the aperture which radiates into the hemisphere above the ground plane. In the above case the field distribution at the aperture must be known. The second method considers the structure as an antenna with a known current distribution. Either approach requires certain simplifying assumptions, the validity of which determines the accuracy of the theoretical results. In the following analysis, knowing the current distribution in the triangular structure, the antenna approach is used.

3.2 FAR ZONE ELECTRIC FIELD COMPONENTS:

The expression for the Hertz vector (2.15), in the case of a structure radiating into the free space is,

\[ \mathbf{\hat{H}} = \frac{-j}{4\pi} \sqrt{\frac{\mu_0}{\varepsilon_0}} \frac{1}{k} \int_{\Gamma} \frac{e^{jkr}}{r} \mathbf{\hat{r}} d\tau \]  \hspace{1cm} \text{...(3.1)}


- 19 -
After substituting $r$ with the approximate expression derived for the far-field zone (Appendix A):

$$
\tilde{\mathbf{n}} = \frac{-j\sqrt{\mu_0}}{4\pi} \frac{1}{k} \int_T e^{-jk\left(\frac{r_o}{r_o} - \frac{x_o}{r_o} x - \frac{y_o}{r_o} y - \frac{z_o}{r_o} z\right)} \mathbf{J} \, dt 
$$

...(3.2)

Neglecting the second order terms in the denominator and substituting $\frac{1}{120\pi}$ instead of the intrinsic wave impedance $\sqrt{\frac{\varepsilon_o}{\mu_o}}$, the Hertz vector is further simplified.

$$
\tilde{\mathbf{n}} = -j\frac{30}{k} \frac{e^{-jk\rho}}{r_o} \int_T e^{jk\left(\frac{x_o}{r_o} x + \frac{y_o}{r_o} y + \frac{z_o}{r_o} z\right)} \mathbf{J} \, dt
$$

...(3.3)

This integration is carried out in Appendix B. The result of the integration being independent of $x$, $y$ and $z$, the electric field vector (2.16) can be simplified:

$$
\ddot{\mathbf{E}} = k^2 \mu_r \varepsilon_r \tilde{\mathbf{n}}
$$

...(3.4)

where

$$
\mu_r = \varepsilon_r = 1
$$

The coordinate transformation described in Appendix A is then applied to (3.4) and after equating to zero the radial component of the
electric field vector as the definition of the far-field approximation requires, the following expressions are obtained.

\[
\begin{align*}
E_r &= 0 \\
E_\theta &= k^2 \left( \Pi_x \cos \theta \cos \phi + \Pi_y \cos \theta \sin \phi - \Pi_z \sin \theta \right) \\
E_\phi &= k^2 \left( -\Pi_x \sin \phi + \Pi_y \cos \phi \right)
\end{align*}
\] ...

(3.5)

3.3 RADIATED POWER:

The time rate of energy flow per unit area, for a radiating element, is the Poynting vector or power density. The average Poynting vector is

\[
P = \frac{1}{2} \text{Re}(\mathbf{E} \times \mathbf{H}^*)
\] ...

(3.6)

where \( E, H \) are peak values

and \( \mathbf{H}^* \) is the complex conjugate of \( \mathbf{H} \). The far-zone radiation assumes that the field vectors have only real components, therefore noticing that

\[
\mathbf{H}^* = \sqrt{\frac{\varepsilon_0}{\mu_0}} \mathbf{E}^*
\] ...

(3.7)

the average radiated power is expressed as:

\[
P_r = \frac{1}{2} \sqrt{\frac{\varepsilon_0}{\mu_0}} \iint |\mathbf{E}|^2 \, ds
\] ...

(3.8)

The above expression is a scalar quantity and can be rewritten
for this particular case as:

\[ P_r = \frac{1}{2} \sqrt{\frac{\varepsilon_0}{\mu_0}} \int_0^{2\pi} \int_0^{\pi/2} \left( E_\theta^2 + E_\phi^2 \right) r^2 \sin \theta \, d\theta \, d\phi \] ... (3.9)

Numerical method can be used to evaluate (3.9) but they do not lead to an absolute result since the arbitrary amplitude factor \((A_{1,0,-1})^2\) will be involved. This is avoided by calculating the radiation quality factor, \(Q_r\). This factor is defined as follows:

\[ Q_r = \frac{\omega_0 \cdot \text{Energy stored in the circuit}}{\text{Average power radiated}} \] ... (3.10)

Another quantity which is often used is the fraction of the stored power which is radiated.

\[ P_{\text{radiated}} \left( \text{in } \% \right) = \frac{1}{Q_r} \times 2\pi \times 100\% \] ... (3.11)

### 3.4 NUMERICAL RESULTS:

A computer program was written (Appendix C) to calculate the radiation \(Q\) using the expression (3.10). Some of the computed results are given in Table 3.1. The effective dielectric constant used in the program was calculated from Schneider's [1] expression.

\[ \varepsilon_e \approx \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \cdot \left(1 + \frac{10h}{w}\right)^{1/2} \] with \(t = 0\) ... (3.12)
where \( w \) is the width of the microstrip
\( h \) is the thickness of the microstrip
\( t \) is the thickness of the metallization

Although the width of the structure is variable, the effective permittivity was assumed to be constant in order to simplify the calculations. \( \varepsilon_r \) was taken to be the value corresponding to the width of the triangle at the origin of the x-y coordinates system.

In Fig. 3.1, the computed values for \( Q_r^{-1} \) are plotted against \( h.f \), where \( f \) is the resonant frequency associated with the fundamental mode. As it is shown below, the relationship between the inverse of the radiation \( Q \) and the product \( h.f \) is linear.

From expression (2.13),
\[
a \propto f^{-1}
\]
and \( k = \frac{2\pi}{\lambda} \) implies,
\[
k \propto f
\]
from Appendix C, for small
\[
\left( \frac{k Z_o h}{r_o} \right), \quad \tilde{P} \propto h.f^{-1}
\]
thus from (2.31),
\[
\tilde{E} \propto h
\]
and from (2.14),
\[
W_{\text{stored}} \propto h.f^{-2}
\]
therefore from (3.8),
\[
Q_r \propto (h.f)^{-1}
\]

3.5 CONCLUSION OF THE THEORETICAL ANALYSIS

The results of the theoretical analysis show that the power radiated from the triangular structure is proportional to the product of thickness of the dielectric material and the resonance frequency.
<table>
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<th>A (mm)</th>
<th>h (mm)</th>
<th>$\varepsilon_r$</th>
<th>$f$ (GHz)</th>
<th>$\varepsilon_e$</th>
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<th>Stored Energy $x \times 10^9$ (w)</th>
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Table 3.1 - Computed radiation data for the triangular resonator
Fig. 3.1 - $Q_r^{-1}$ versus the product (h.f), computed results

- $Q_r$ is the radiation quality factor,
- $Q_r$ is the substrate thickness,
- $f$ is the resonance frequency.
Another factor affecting the radiation is the permittivity of the substrate; a decrease in the value of the relative dielectric constant corresponds to an increase in the radiated power. The latter is shown on Fig. (3.2) where theoretically calculated values of the radiation quality factor are plotted against the relative permittivity for constant frequency of resonance and thickness of the substrate.
Fig. 3.2 - Computed values of $Q_{r}^{-1}$ versus both relative and effective permittivity. ($h=0.64\text{mm}, f=10\text{GHz}$)
CHAPTER IV

RADIATION PATTERN MEASUREMENTS

4.1 ANTENNA TEST MEASUREMENTS

The theoretical expressions and results for the power radiated from the triangular resonator, were derived in the previous Chapters, by using some approximations. A comparison of the measured field patterns with the theoretical results may be used as a first verification of the correctness of the analysis.

The experiment involved the E- and H-field pattern measurements of the triangular resonator in an anechoic chamber. The set-up, Fig. (4.1), consisted of (i) a transmitting horn antenna, (ii) the device under test (DUT) as the receiving antenna, mounted on a table whose plane can rotate around the horizontal and the vertical axis and (iii) a standard gain horn. The E- and H-plane far-field patterns were measured by rotating the DUT around an axis orthogonal and parallel to the E-field of the radiating element respectively.

The power gain of the DUT was measured by comparison with the standard gain horn. The reference gain of the standard gain horn was computed from the structures geometry.
Fig. 4.1 - Set-up for radiation pattern measurement.
4.2 THEORETICAL RESULTS OF THE E- AND H-PLANE PATTERNS

The graph of the modulus of the Poynting vector, $|\vec{P}|$, at a constant distance from the radiating element, as a function of $\Theta$ or $\phi$ is called the power pattern. This expression was derived in section (3.2) as:

$$|\vec{P}| = \frac{1}{2} \sqrt{\frac{e_0}{\mu_0}} |\vec{E}|^2 \quad ... (4.1)$$

This can be written as:

$$|\vec{P}| = \frac{1}{2} \times \frac{1}{120\pi} \left( E_\Theta^2 + E_\phi^2 \right) \quad ... (4.2)$$

The expression (4.2) is solved using numerical methods by replacing the arbitrary amplitude factor $(A_1, 0, -1)^2$ with unity. The computer program is given in Appendix D. To compute the results of the E-plane pattern, the value of $\Theta$ is varied from $0^\circ$ to $90^\circ$ in steps of $10^\circ$ and $\phi$ is kept constant respectively at $0^\circ$ and $180^\circ$. The H-plane pattern results are calculated for $\phi = 90^\circ$ and $270^\circ$.

4.3 THE TESTED STRUCTURES

A singly loaded triangular resonator was designed on a 1.27 mm thick alumina substrate with $\varepsilon_r = 10$. The frequency of resonance was measured to be 9.6 GHz and the return loss was 15 dB which corresponds
to a VSWR of 1.43. The substrate was mounted on a 1.2 m diameter ground plane for the radiation pattern measurements. In order to lower the effect of the feed radiation, the substrate dielectric was extended around the coax to microstrip transition. The maximum measured power gain of the element was about 4.2 dB.

A second circuit was designed, this time on a 1.52 mm RT/Duroid substrate, $\varepsilon_r = 2.2$. The measured bore sight power gain was 6.5 dB.

4.4 RESULTS AND DISCUSSION

The results of the measured and theoretical radiation patterns for the resonator, on an alumina substrate, are compared in Figures 4.2, 4.3, 4.4 and 4.5. The theoretical results are normalized to the corresponding maximums of the measured radiation patterns. It can be seen that the theory underestimates the radiation. As an antenna element, the resonator has a wide half-power beamwidth which is the case for most of the microstrip single antenna structures. Also, because the cross-polarization is low, the triangular element is essentially linearly polarized.

The power gain measurements show that the lower the relative dielectric constant of the substrate, the more the antenna element radiates, as was expected from the theory.
Fig. 4.2 - Triangular resonator theoretical and experimental E-plane radiation pattern. \((\varepsilon_r = 10, h = 1.27 \text{ mm}, f = 9.6 \text{ GHz})\)

- - measured pattern
- - - calculated pattern
Fig. 4.3 - E-plane cross-polarization pattern
(for the same substrate as in Fig. 4.2)
Fig. 4.4 - Triangular resonator theoretical and experimental H-plane pattern. (for the same substrate as in Fig. 4.2).

---

measured pattern

---

theoretical pattern
Fig. 4.5 - H-plane cross-polarization pattern.
(for the same substrate as in Fig. 4.2).
CHAPTER V

RADIATION QUALITY FACTOR MEASUREMENTS

5.1 DEFINITION OF THE QUALITY FACTOR

The losses in a resonant cavity determine its quality factor, Q. A perfect resonator, without any coupling, dielectric, resistive or radiation losses would have a Q of infinity. Mainly three different Q's - the unloaded, the loaded and the external Q - are used in the analysis of a cavity and its peripheral circuits.

For the unloaded Q, \( Q_o \), consider the energy dissipated at resonance in the cavity alone [17]:

\[
Q_o = \frac{\text{Energy stored in cavity}}{\text{Energy dissipated in cavity per radian}} \quad \text{at resonance}
\]

The dissipated energy depends, besides the cavity mode, on one or more of the following losses, (i) the resistivity of the walls, (ii) the dielectric losses and (iii) the radiation. The unloaded Q of a 4 GHz waveguide cavity is greater than 10,000 but in the case of microstrip resonators, \( Q_o \) is only of the order of a few hundreds.

A cavity is usually coupled to a matched transmission line, and because of the loading effect of the line, the overall Q of the circuit...
is lower than $Q_o$. The loaded $Q$, $Q_L$, is defined as:

$$Q_L = \frac{\text{Energy stored in cavity}}{\text{Energy dissipated in both cavity and external circuit per radian at resonance}}$$

The loaded and unloaded $Q$ are related to each other by the means of the external $Q$, $Q_e$. This relation may be expressed as:

$$\frac{1}{Q_L} = \frac{1}{Q_o} + \frac{1}{Q_e} \quad \cdots (5.1)$$

The definition of $Q_e$ is:

$$Q_e = \frac{\text{Energy stored in cavity}}{\text{Energy dissipated in external circuit per radian at resonance}}$$

If the cavity is loaded at two points, then expression (5.1) can be modified to:

$$\frac{1}{Q_L} = \frac{1}{Q_o} + \frac{1}{Q_{e1}} + \frac{1}{Q_{e2}} \quad \cdots (5.2)$$

5.2 LOSSES IN MICROSTRIP RESONATORS

Microstrip structures have low $Q$ factors due to the high losses. The unloaded $Q$ of a microstrip is expressed by the sum of three main losses.

$$\frac{1}{Q_o} = \frac{1}{Q_e} + \frac{1}{Q_c} + \frac{1}{Q_r} \quad \cdots (5.3)$$
where:  
\( Q_e \) is a measure of the dielectric losses
\( Q_c \) is a measure of the conductor losses
\( Q_r \) is a measure of the radiation losses

\( Q_e \) is inversely proportional to the loss tangent of the dielectric material. For alumina substrates, with relative dielectric constant, \( \varepsilon_r \), of approximately 10, the loss tangent is of the order of 10^{-4} up to 10 GHz. For RT/Duroid 5880, \( \varepsilon_r = 2.2 \), the loss tangent is of the order of 10^{-3}. The surface roughness of the dielectric also contributes to the dielectric losses. In order to account for these losses, \( Q_e \) must be divided by a factor of 1.04 in the case of alumina substrates at 8 GHz, and by 1.8 in the case of RT/Duroid 5880 [12].

The conductor losses can be calculated using a simple analysis which does not include the effects of fringing fields and radiation [6]. \( Q_c \) can be expressed as:

\[
Q_c = (\pi \mu_0 f \sigma)^{\frac{1}{2}} h \tag{5.4}
\]

where:
\( \sigma \) is the conductivity in \( \text{S/m} \)
\( h \) is the substrate thickness in m
\( f \) is the frequency in Hz
\( \mu_0 = 4\pi \times 10^{-7} \text{ H/m} \)

For copper (\( \sigma = 5.9 \times 10^7 \text{ S/m} \)), at 10 GHz, on a 0.635 mm thick substrate,

\( Q_c \sim 969 \)
For bulk gold, \( \sigma = 4.55 \times 10^7 \, \text{S/m} \)

\[ Q_c = 851 \]

From literature it is known that radiation losses depend on the geometry of the structure \([8,10]\). Therefore, neglecting dielectric and conductor losses and assuming that radiation is the dominant source of losses cannot be justified for all cases, since \( Q_r \) is of the order of a few hundreds.

5.3 THE QUALITY FACTOR MEASUREMENT

The \( Q \) of a resonator can be found by measuring the bandwidth between the half-power frequencies when the circuit is excited by a constant current source.

\[ Q = \frac{f_0}{\Delta f} \] \hspace{1cm} (5.5)

where:

\( f_0 \) is the center frequency,

\( \Delta f = f_2 - f_1 \), is the half-power bandwidth

The measurement will give the loaded \( Q, Q_L \), which will lead to \( Q_o \) after a simple calculation, provided the input and output coupling are made equal.

If the input and output coupling coefficients between the cavity and the external circuit are \( \beta_1 \) and \( \beta_2 \) respectively, then (18)

\[ Q_L = \frac{Q_o}{1 + \beta_1 + \beta_2} \] \hspace{1cm} (5.6)
and the transmitted power factor at resonance is:

$$\tau_{res} = \frac{4\beta_1\beta_2}{(1 + \beta_1 + \beta_2)^2} \quad \ldots \ldots (5.7)$$

For \( \beta_1 = \beta_2 = \beta \), the expressions (5.6) and (5.7) become,

$$Q_L = \frac{Q_o}{(1 + 2\beta)} \quad \ldots \ldots (5.8)$$

$$\tau_{res} = \frac{4\beta^2}{(1 + 2\beta)^2} \quad \ldots \ldots (5.9)$$

Respectively, it is also interesting to note that the transmitted power is maximum for \( \beta_1 = \beta_2 \).

The expression (5.8) can be rewritten as:

$$1 - \frac{Q_L}{Q_o} = \frac{2\beta}{1 + 2\beta} \quad \ldots \ldots (5.10)$$

Which leads to:

$$\tau_{res} = \left(1 - \frac{Q_L}{Q_o}\right)^2 \quad \ldots \ldots (5.11)$$

$$Q_o = \frac{Q_L}{1 - \sqrt{\tau_{res}}} \quad \ldots \ldots (5.12)$$
Thus it can be seen from expression (5.12) that a transmission loss measurement at the resonant frequency and a knowledge of the half-power bandwidth of the resonator will lead to the unloaded $Q$. It should also be noted that $Q_L$ approaches $Q_0$ as the coupling between the cavity and the external circuit becomes looser.

5.4 THE RADIATION $Q$ MEASUREMENT TECHNIQUE

A direct measurement of $Q_r$, as can be seen from the expression (5.3), may involve quantities such as the resistivity of the microstrip deposition, the loss tangent of the dielectric etc. which are difficult to measure. A practical method that overcomes the above problem consists of two identical measurements; in the first measurement, the resonator is in the open, i.e. radiating. Then the resonator is placed into a shielding box with highly conducting walls. The dielectric and conductor losses being the same, it follows that,

$$\frac{1}{Q_r} = \frac{1}{Q_0} - \frac{1}{Q_s} \quad \ldots (5.13)$$

Where, $Q_s$ is the shielded $Q$ which assumes no radiation.

The dimensions of the shielding box which can be considered as a rectangular waveguide cavity should be such that it will not sustain any box modes at the frequencies of interest.
For the fundamental mode in a rectangular cavity, $TE_{101}$, the resonant frequency is (19):

$$f = \frac{c \sqrt{a^2 + d^2}}{2 ad} \quad \ldots (5.14)$$

where:
- $a$ is the width of the box
- $d$ is the length of the box
- $c$ is the speed of light in the vacuum

Fig. (5.1) shows a jig and a shield arrangement used for the experiments. They are made of solid brass blocks and held together by four screws.

5.5 EXPERIMENTAL ARRANGEMENT

Circular and triangular resonators were fabricated on alumina with $\varepsilon_r \approx 10$, $h = .635$ mm and $h = 1.27$ mm; Custom with $\varepsilon_r \approx 10$, $h = 1.52$ mm; and Rexolite with $\varepsilon_r \approx 2.5$, $h = .76$ mm and $h = 1.52$ mm substrates. The conductor layer is 4 $\mu$m thick gold in the case of alumina substrates and 25.4 $\mu$m thick copper for the other two substrates.

The designs were made such that the resonant frequency of the elements would be in the range 6 to 12 GHz. The theoretical expressions used to determine the center frequency are:

a) For the circular disc [6],

$$f = 1.84118 \frac{c}{\pi \sqrt{\varepsilon_r} D} \quad \ldots (5.15)$$
Fig. 5.1 - Jig and its cover used in $Q_r$ measurements
b) For the triangle \[22\], \( f = \frac{2c}{3A \sqrt{\varepsilon}} \) \((5.16)\)

where: 
- \(D\) is the diameter of disc
- \(A\) is the side of triangle

The resonators were coupled through a gap to a 50 \(\Omega\) line. An interesting point in the above expressions is that, if in (5.15), \(D\) is replaced by \(\frac{A \sqrt{3}}{2}\), the height of the triangle, then the expressions (5.15) and (5.16) are almost equal. The difference is around 3%.

The experiment consisted of a transmission loss measurement Fig. 5.2 between two frequencies swept by a generator. A reference level was set by connecting a semi-rigid coaxial cable between the detector head 'B' of the amplitude analyser and the output of the directional coupler. Then, the resonator jig, without its cover, was inserted between the coax and the directional coupler to measure the insertion loss. This technique was repeated for the shielded jig and \(Q_r\) was derived using the relations (5.5), (5.12) and (5.13). For all the measurements, the transmission loss at resonance was greater than 12 dB. Also, in order to make the simplifying assumption \(B_1 = B_2\) explained in section 5.3, the return loss of each port was measured and made equal.

5.6 RESULTS AND DISCUSSION

The measured and the computed resonant frequencies are listed in Tables 5.1, 5.2 and 5.3. It can be seen that the difference between the theoretical and measured frequencies is greater for alumina than for
Fig. 5.2 - Set-up to measure both the reflected and the transmitted power
Custom-K even though both substrates have an $\varepsilon_r = 10$. This may be partially caused by tolerances in the dielectric constants which can be as high as $\pm 20\%$ for the Custom-K material. Another possible reason for this discrepancy is that the theoretical expression for the resonant frequency does not include the effects of the fringing field. Therefore, the measured frequencies will differ if the thicknesses of the used substrates are different. For the low dielectric constant material, the theoretical frequency is closer to the measured one.

Dependence of the radiation on the substrate thickness and frequency is shown in the graphs, Fig. 5.3 and Fig. 5.4, where $Q^{-1}$ is plotted against h.f. The radiation losses increase as the product h.f increases. Also as a basis for comparison, radiation from a 50 $\Omega$ quarter wave resonator as calculated by Belohoubek and Denlinger [12] is shown in Fig. 5.3. The lines drawn through the measured points have a slope close to unity, while, for the quarter wave resonator, the slope of the line is approximately 2. The latter result is expected since Lewin's analysis [4] shows that radiation from a microstrip line is proportional to $(h.f)^2$. The graphs also demonstrate that the radiation loss for circular resonators is higher than for triangular ones. The quarter wave, 50 $\Omega$ resonator radiates least among the measured structures. In general, the losses may be lowered by using thin and high dielectric constant materials.
<table>
<thead>
<tr>
<th>A(mm)</th>
<th>h(mm)</th>
<th>Measured f(GHz)</th>
<th>Calculated f(GHz)</th>
<th>$Q_s$</th>
<th>$Q_o$</th>
<th>$Q_r$</th>
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<td>9.40</td>
<td>0.64</td>
<td>6.6</td>
<td>6.7</td>
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<td>8.0</td>
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<td>589</td>
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<td>18</td>
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Table 5.1 - Experimental results for the triangular resonator ($\varepsilon_r = 10$)

<table>
<thead>
<tr>
<th>A(mm)</th>
<th>h(mm)</th>
<th>Measured f(GHz)</th>
<th>Calculated f(GHz)</th>
<th>$Q_s$</th>
<th>$Q_o$</th>
<th>$Q_r$</th>
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<td>15.24</td>
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<td>8.3</td>
<td>8.3</td>
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<tr>
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<td>10.0</td>
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<td>38</td>
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<td>0.25</td>
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<td>8.3</td>
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<td>63</td>
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<td>128</td>
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<td>12.5</td>
<td>12.4</td>
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<td>97</td>
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Table 5.2 - Experimental results for the triangular resonator ($\varepsilon_r = 2.5$)
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<th>h (mm)</th>
<th>Measured f (GHz)</th>
<th>Calculated f (GHz)</th>
<th>Q&lt;sub&gt;s&lt;/sub&gt;</th>
<th>Q&lt;sub&gt;o&lt;/sub&gt;</th>
<th>Q&lt;sub&gt;r&lt;/sub&gt;</th>
</tr>
</thead>
<tbody>
<tr>
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<td>91</td>
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<td>1.27</td>
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<td>6.8</td>
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<td>11.5</td>
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<td>4.32</td>
<td>1.27</td>
<td>10.4</td>
<td>12.9</td>
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<td>8.8</td>
<td>445</td>
<td>19</td>
<td>20</td>
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</table>

Table 5.3 - Experimental results for the disc resonator (ε<sub>r</sub> = 10)
Fig. 5.3 - Measured $Q_r^{-1}$ versus h.f for circular and triangular resonators
Fig. 5.4 - Measured $Q_r^{-1}$ versus h.f for triangular resonator

$\Delta$ Triangular resonator

$h = 0.25 \text{ mm}, 0.76 \text{ mm}$

$\varepsilon_r = 2.5$
CHAPTER VI

COUPLING TO A MICROSTRIP RESONATOR

6.1 COUPLING TO MICROSTRIP STRUCTURES:

The radiation properties of the triangular resonator were derived assuming a very loose coupling between the resonant element and the external circuitry. Other applications, e.g. antenna measurements, may require a tight coupling. Therefore, the understanding of and the control of the energy coupled to the resonator is important.

Three types of coupling, namely capacitive, direct and hole coupling Fig. 6.1 (a,b,c) may be used with the microstrip configuration. In this Chapter, the equivalent circuit of a resonator and the coupling mechanism are studied. Some experimental gap capacitance values are also given.

6.2 EQUIVALENT CIRCUIT OF ONE PORT RESONATOR:

A one-port resonator or a reflection cavity can be represented by any of the configuration given in Fig. 6.1. The equivalent circuit of an unloaded cavity using lumped elements in parallel is shown in Fig. 6.2a. The same circuit can be modified to take into account the coupling which can be assumed to be in the form of an ideal transformer of n:1
a) Capacitive gap coupling

b) Direct coupling

c) Hole coupling

Fig. 6.1 - Possible Microstrip coupling configurations
can be assumed to be in the form of an ideal transformer of n:1 turns ratio, the losses are incorporated in $R_{\text{eq}}$ Fig. 6.2b [17].

The expressions,

$$R_o = n^2 R_{\text{eq}}$$

$$L_o = n^2 L_{\text{eq}}$$

$$C_o = \frac{C_{\text{eq}}}{n^2}$$

represent the relationship between Fig. 6.2a and Fig. 6.2b, whereas Fig. 6.2c is obtained by normalizing the above expressions to $Z_o$.

The admittance of the normalized equivalent circuit is:

$$\bar{Y} = \frac{1}{\bar{R}} + j \left( \omega C - \frac{1}{\omega L} \right)$$

$$\bar{V} = \bar{S} + j \bar{S}$$

with

$$\omega_o = \frac{1}{\sqrt{LC}}$$

Experimentally, $n$ can be found from VSWR measurements.

$$|S_{11}| = \text{antilog} \left( - \frac{\text{Return loss}}{20} \right)$$
Fig. 6.2 - Lumped element representation of a one port cavity
and

\[ n^2 = \frac{1 - \frac{S_{in}}{S_{in}}} {1 + \frac{S_{in}}{S_{in}}} \]  \hspace{1cm} \ldots(6.8)

\( \bar{G} \) can be found from the expression (6.5) after a VSWR measurement at resonance, where \( \bar{B} = 0 \). A second VSWR measurement close to \( \omega_o \) will give \( \bar{L} \) and \( \bar{C} \). This can be shown after rearranging \( \bar{B} \) as,

\[ \bar{B} = \frac{\sqrt{\frac{C}{L}}}{C} \left( \frac{\omega}{\omega_o} - \frac{\omega}{\omega} \right) \]  \hspace{1cm} \ldots(6.9)

At \( \omega = \omega_o + \Delta \omega \),

\[ \bar{B} = \frac{\sqrt{\frac{C}{L}}}{L} \left( \frac{\omega_o + \Delta \omega}{\omega_o} - \frac{\omega}{\omega_o} \Delta \omega \right) \]  \hspace{1cm} \ldots(6.10)

this expression can be represented approximately by:

\[ \bar{B} \approx \frac{\sqrt{\frac{C}{L}}}{L} \left( \frac{2 \Delta \omega}{\omega_o} \right) \]  \hspace{1cm} \ldots(6.11)

Therefore,

\[ \bar{L} \approx \frac{2 \Delta \omega}{\bar{B} \omega_o^2} \]  \hspace{1cm} \ldots(6.12)

and,

\[ \bar{C} \approx \frac{\bar{B}}{2 \Delta \omega} \]  \hspace{1cm} \ldots(6.13)
6.3 THE TWO-PORT (DOUBLE LOADED) RESONATOR EQUIVALENT CIRCUIT

The two-port resonator is a cavity with an input and output port. It is also called a transmission cavity. The equivalent circuit with ideal transformers is represented in Fig. 6.3a. The analysis of this circuit is similar to that of the one-port cavity. The output transformer is first eliminated Fig. 6.3b. Then the normalized circuit is as in Fig. 6.3c with,

\[ \bar{R}' = \frac{n^2}{n_{12}} \cdot \frac{Z_o'}{Z_o} \]  
\[ \bar{R} = \frac{n^2}{Z_o} \cdot R_{eg} \]  
\[ \bar{L} = \frac{n^2}{Z_o} \cdot L_{eg} \]  
\[ \bar{C} = \frac{Z_o}{n^2} \cdot C_{eg} \]

The admittance of the normalized circuit is

\[ \bar{Y} = \frac{1}{\bar{R}'//\bar{R}} + j \left( \omega \bar{C} - \frac{1}{\omega \bar{L}} \right) \]
a) Two port equivalent circuit

b) Without output coupling

c) Normalized equivalent circuit

Fig. 6.3 - Lumped element representation of a two port cavity.
This expression differs from (6.4) only in the resistive term which is $\overline{R}'$ in parallel with $\overline{R}$.

$$\overline{R}' \parallel \overline{R} = \frac{\overline{R}}{\overline{R} + \overline{R}'}$$  \hspace{1cm} (6.19)

From this it can be seen that if the input and output coupling are identical and $Z_0 = Z_0'$, then $\overline{R}' = 1$ and therefore $\overline{R}' \parallel \overline{R}$ is always less than 1 which means the cavity is undercoupled.

In the case of the two port cavity, $R_{eg}$, $C_{eg}$ and $L_{eg}$ can be found from $Q_o$ and $Q_1$ measurements using the relations given by Altman (17),

$$\overline{R} = \frac{Q_o}{Q_{e1}}$$  \hspace{1cm} (6.20)

$$\overline{R}' = \frac{Q_{e2}}{Q_{e1}}$$  \hspace{1cm} (6.21)

$$Q_o = \overline{C} \overline{R} \omega_o$$  \hspace{1cm} (6.22)

$$\omega_o = \frac{1}{\sqrt{L \overline{C}}}$$  \hspace{1cm} (6.23)

and the expression (5.2),

$$\frac{1}{Q_1} = \frac{1}{Q_o} + \frac{1}{Q_{e1}} + \frac{1}{Q_{e2}}$$
For identical input and output coupling,

\[ Q_{e1} = Q_{e2} = Q_e \]

and with substitution of the expression (6.20), (5.2) becomes:

\[ \frac{1}{Q_1} = \frac{1}{Q_0} + \frac{2R}{Q_0} \quad \ldots (6.24) \]

Therefore,

\[ R = \frac{1}{2} \left( \frac{Q_0}{Q_1} - 1 \right) \quad \ldots (6.25) \]

and

\[ R' = 1 \]

6.4) GAP CAPACITANCE MEASUREMENTS OF THE CAPACITIVE COUPLED TRIANGULAR RESONATOR

A common coupling structure for microstrip is the gap coupling. An equal input and output coupling coefficient for symmetrical structure like a disk or rectangle, can be realized by two identical gaps. In the case of the triangular resonator, in order to have equal coupling coefficients, the input and output gaps should be different. For the latter structure, the relation between the gap width and the capacitance is shown in Fig. 6.4. This graph was plotted by measuring the gap capacitance of a x10 model triangular resonator on a HP 4270A automatic...
capacitance bridge. As a comparison, the capacitance of gaps in a 50 Ω transmission lines is also plotted. The later graph was published in the Microwave Engineers Handbook and Buyer's Guide 1969.
Fig. 6.4 - (Gap Capacitance/substrate thickness) versus (Gap width/substrate thickness)
CHAPTER VII

CONCLUSION

The objective of the thesis was to derive an expression for the radiation loss from triangular microstrip resonators and to verify the theoretical results by measurements. Theoretical results for the radiated power were obtained by numerically integrating the electric far field of the resonator.

The measurement of the radiated power was made in an anechoic chamber. The measured radiation pattern followed qualitatively the theoretically predicted pattern. Since the theoretical derivation contained an arbitrary amplitude factor, \( A_{1,0,-1} \) an absolute comparison with the measured results could not be made. The radiation loss measurements overcame the above problem. The theoretical results proved to be in good agreement with measurements for substrates with low dielectric constant, but for high dielectric constant values they differed by a factor of about two as shown in Fig. 7.1 and Fig. 7.2. This discrepancy could be due to the two assumptions made about the dielectric media viz.

i) the permittivity, \( \varepsilon \), in section 2.4 was assumed independent of the 'z' direction and taken to be equal to the effective permittivity, \( \varepsilon_e \).

ii) \( \varepsilon_e \), which is a function of 'x' according to the expression (3.12), was assumed constant.
The results show that for large values of $\varepsilon_r$, the theory underestimates the radiation losses.

Resonators in general have wide circuit applications, but the triangular structure with its $120^\circ$ symmetry is particularly suited to the fabrication of 3 port ferrite circulators. The experimental comparison between circular and triangular resonators showed that the triangular configuration radiates the least Fig. 5.3. Although the difference is small, it provides useful information for the design of a low loss microstrip circulator. For filters, the rectangular resonator, which is nothing but a resonant section of transmission line, is the simplest structure to use. It also radiates less than the circular or triangular resonators [5,7,8,9]. The radiation pattern of triangular resonators show a large beamwidth, which is one of the characteristics of microstrip structures [16]. The measured E-plane and H-plane 3dB beamwidth is respectively $180^\circ$ and $100^\circ$. However, a more directional antenna can be designed with an array of single elements [14]. The present study also shows that the circular resonator is a better candidate for antenna elements since it radiates more than the triangular one.
Fig. 7.1 - Experimental and computed values of $Q_r^{-1}$ versus (h.f)
Fig. 7.2 - Experimental and computed values of $Q^{-1}$ versus $(\varepsilon_e)$ and $(\varepsilon_r)$

- Computed curve
- Experimental points
APPENDIX A

1. THE COORDINATE SYSTEM

The coordinate system for the triangular resonator is defined as in Fig. A.1.

A point \( P \) in space can be described as:

\[
\begin{align*}
  x_o &= r_o \sin \theta \cos \phi \\
  y_o &= r_o \sin \theta \cos \phi \\
  z_o &= r_o \cos \theta
\end{align*}
\] ...

2. APPROXIMATE EXPRESSION FOR THE DISTANCE \( r \):

From Fig. A.1, the distance between the field point \( P \) and any point \( M \) on the triangle is:

\[
r = \sqrt{(x_o - x)^2 + (y_o - y)^2 + (z_o - z)^2}
\] ...

Using the far-field considerations, \( x_o \gg x, y_o \gg y, z_o \gg z \), we can neglect the second order terms in \( x, y, z \).

\[
r \equiv \sqrt{x_o^2 + y_o^2 + z_o^2 - 2(x_o x + y_o y + z_o z)}
\] ...

- 66 -
Fig. A.1 - Rectangular and circular coordinate system used in the theoretical analysis.
The above expression can be simplified to:

$$ r = r_0 - \frac{x_0}{r_0} x - \frac{y_0}{r_0} y - \frac{z_0}{r_0} z $$

...A.6

using the binomial expansion and neglecting the higher order terms.

3. TRANSFORMATION FROM CARTESIAN TO SPHERICAL COORDINATES

This transformation can be achieved through multiplying the cartesian coordinate system matrix by the transformation matrix $[T]$.

$$ [T] = \begin{bmatrix}
\sin \theta \cos \phi & \sin \theta \sin \phi & -\cos \theta \\
\cos \theta \cos \phi & \cos \theta \sin \phi & -\sin \theta \\
-\sin \phi & \cos \phi & 0
\end{bmatrix} $$

...A.7

The spherical coordinate system components in terms of the rectangular components are expressed as:

$$
C_r = C_x \sin \theta \cos \phi + C_y \sin \theta \sin \phi + C_z \cos \theta \\
C_\theta = C_x \cos \theta \cos \phi + C_y \cos \theta \sin \phi - C_z \sin \theta \\
C_\phi = -C_x \sin \phi + C_y \cos \phi
$$

...A.8

4. CHANGE OF VARIABLES

The following change of variables is used in Appendix B to simplify calculations.

Let,

$$\alpha = \frac{2\pi}{\sqrt{3} a} x + \frac{2\pi}{3}$$

...A.9
And
\[ \beta = \frac{2\pi}{3a} \gamma \] ....A.10

Thus
\[ x = -\frac{a\sqrt{3}}{3} \] corresponds to \( \alpha = 0 \) ....A.11

\[ x = \frac{a\sqrt{3}}{6} \] corresponds to \( \alpha = \pi \) ....A.12

\[ y = -\left( \frac{x}{\sqrt{3}} + \frac{a}{3} \right) \] corresponds to \( \beta = \frac{-\alpha}{3} \) ....A.13

\[ y = \left( \frac{x}{\sqrt{3}} + \frac{a}{3} \right) \] corresponds to \( \beta = \frac{\alpha}{3} \) ....A.14

Also
\[ y = \frac{3a}{2\pi} \beta \], \[ x = \frac{3a}{2\pi} \alpha - 3 \] ....A.15

\[ dy = \frac{3a}{2\pi} d\beta \], \[ dx = \frac{\sqrt{3}a}{2\pi} d\alpha \] ....A.16
APPENDIX B.

DERIVATION OF THE HERTZ VECTOR FOR THE TRIANGULAR RESONATOR

In section 3.2, the hertzian vector was simplified to:

\[ \mathbf{\hat{H}} = -j \frac{30}{k} \frac{e^{-jkr_0}}{r_0} \int e^{jk \left( \frac{x_0}{r_0} x + \frac{y_0}{r_0} y + \frac{z_0}{r_0} z \right)} \mathbf{j} \, dT \]

The expression for the current density was:

\[ \mathbf{\hat{J}} = \mathbf{\hat{J}}_T = \begin{cases} -H_y(x, y, h) + H_y(x, y, -h) \end{cases} \hat{a}_x 
+ \begin{cases} H_x(x, y, h) - H_x(x, y, -h) \end{cases} \hat{a}_y 
+ \begin{cases} j \frac{k}{120\pi} (\varepsilon_e - 1) E_z(x, y, 0) \end{cases} \hat{a}_z \]

The above integration is carried out separately for each component of the vector \( \mathbf{\hat{H}} \). The limits of integration in the \( x \) and \( y \) directions are defined in Appendix A, Fig. A.1. The integration in the \( z \) direction is carried out first and is evaluated at \( z = h \) and \( z = -h \) since the fields are assumed constant in the \( z \) direction.
\[ \Pi_x = -\frac{j 30}{k} e^{-jkr_0} \int e^{jk\left(\frac{x_0}{r_0}x + \frac{y_0}{r_0}y + \frac{z_0}{r_0}z\right)} \left(\mathcal{H}_y(x, y, h) + H_y(x, y, -h)\right) dzdydx \]

\[ \Pi_x = -\frac{j 30}{k} e^{-jkr_0} \left\{ -e^{-jkr_0} \int e^{jk\left(\frac{x_0}{r_0}x + \frac{y_0}{r_0}y\right)} H_y(x, y, h) dydx \right\} \]

as \( H_y(x, y, h) = H_y(x, y, -h) \),

\[ \Pi_x = -\frac{30}{k} e^{-jkr_0} \left\{ -2j \left( e^{\frac{jkr_0}{2r_0}} - e^{-\frac{jkr_0}{2r_0}} \right) \int e^{jk\left(\frac{x_0}{r_0}x + \frac{y_0}{r_0}y\right)} H_y(x, y, h) dydx \right\} \]

\[ \Pi_x = -\frac{60}{k} e^{-jkr_0} \sin\left(\frac{kz_0 h}{r_0}\right) \int e^{jk\left(\frac{x_0}{r_0}x + \frac{y_0}{r_0}y\right)} H_y(x, y, h) dydx \]

Using the same approach for \( \Pi_y \) and \( \Pi_z \)

\[ \Pi_y = \frac{60}{k} e^{-jkr_0} \sin\left(\frac{kz_0 h}{r_0}\right) \int e^{jk\left(\frac{x_0}{r_0}x + \frac{y_0}{r_0}y\right)} H_x(x, y, h) dydx \]

\[ \Pi_z = \frac{30}{r_0} \left(\epsilon - 1\right) \left[ 2\frac{r_0}{kz_0} \sin\left(\frac{kz_0 h}{r_0}\right) \right] \int e^{jk\left(\frac{x_0}{r_0}x + \frac{y_0}{r_0}y\right)} E_z(x, y, 0) dydx \]
The limits of integration are:

\[- \frac{a \sqrt{3}}{3} \leq x \leq \frac{a \sqrt{3}}{6}\]

and

\[-\left(\frac{x}{\sqrt{3}} + \frac{a}{3}\right) \leq y \leq \left(\frac{x}{\sqrt{3}} + \frac{a}{3}\right)\]

The above integrations are carried out after performing the change of variables described in Appendix A.

The x component of the vector \( \Pi \) becomes:

\[
\Pi_x = -\frac{60 e^{-jk \theta}}{k r_o} \sin \left(\frac{k z_o h}{r_o}\right) \int_0^\Pi \int_{-\frac{\alpha}{3}}^{\frac{\alpha}{3}} e^{jk x_0} \left(\frac{\sqrt{3} a}{2 \pi} \alpha - 3\right) e^{j k y_o} \frac{3a}{2 \pi} \beta
\]

\[
\left\{j \sqrt{3} A_{1,0,-1} e^{-j \theta} (\sin \alpha \cos \beta)\right\} \frac{3 \sqrt{3} a^2}{4 \pi^2} d\beta d\alpha
\]

Let \( C_x = -\frac{60 e^{-jk \theta}}{k} A_{1,0,-1} \left(\frac{3a}{2 \pi}\right)^2 \sin \left(\frac{k z_o h}{r_o}\right) e^{-jk} \left(r_o + \frac{3k x_o}{r_o}\right) \)

\[
m = \frac{3k x_o}{r_o} \frac{\sqrt{3} a}{2 \pi} \quad n = \frac{k y_o}{r_o} \frac{3a}{2 \pi}
\]

Then,

\[
\Pi_x = C_x \int_0^\Pi \int_{-\frac{\alpha}{3}}^{\frac{\alpha}{3}} e^{j \frac{m \alpha}{3}} e^{j \frac{n \beta}{3}} (\sin \alpha \cos \beta) d\alpha d\beta
\]

From integral tables we have,

\[
\int e^{ay} \cos by dy = \frac{e^{ay}}{b^2 + a^2} (a \cos by + b \sin by)
\]
Therefore,

\[ \Pi_\chi = C_\chi \int_0^\pi \left( e^{j\frac{\alpha}{3}} \sin\alpha \right) \left[ \frac{e^{jn\beta}}{1 - n^2} (jn \cos\beta + \sin\beta) \right] \frac{\alpha}{3} \, d\alpha \]

\[ \Pi_\chi = C_\chi \int_0^\pi \left( e^{j\frac{\alpha}{3}} \sin\alpha \right) \left[ \frac{jn \cos\alpha}{3} \left( e^{j\frac{n\alpha}{3}} - e^{-j\frac{n\alpha}{3}} \right) + \frac{\sin\alpha}{3} \frac{1}{1 - n^2} \right] \left( e^{j\frac{\alpha}{3}} + e^{-j\frac{\alpha}{3}} \right) \, d\alpha \]

The above expression can be rearranged to:

\[ \Pi_\chi = \frac{C_\chi}{4(1 - n^2)} \int_0^\pi \left[ (n - 1) e^{j\frac{\alpha}{3}} P_1 + (n + 1) e^{j\frac{\alpha}{3}} P_2 - (n + 1) e^{j\frac{\alpha}{3}} P_3 \right. \]

\[- (n - 1) e^{j\frac{\alpha}{3}} P_4 - (n - 1) e^{j\frac{\alpha}{3}} P_5 - (n + 1) e^{j\frac{\alpha}{3}} P_6 \]

\[ + (n + 1) e^{j\frac{\alpha}{3}} P_7 + (n - 1) e^{j\frac{\alpha}{3}} P_8 \] \, d\alpha

with \( P_1 = m + n + 4 \), \( P_2 = m + n + 2 \), \( P_3 = m - n + 4 \),

\( P_4 = m - n + 2 \), \( P_5 = m + n - 2 \), \( P_6 = m + n - 4 \),

\( P_7 = m - n - 2 \), \( P_8 = m - n - 4 \).

The result of the second integration is:
\[ \Pi_x = \frac{3 C_x}{4(1 - n^2)} \left[ \frac{(n - 1)}{P_1} \left( e^{\frac{j\pi}{3} P_1} - 1 \right) + \frac{(n + 1)}{P_2} \left( e^{\frac{j\pi}{3} P_2} - 1 \right) - \frac{(n + 1)}{P_3} \left( e^{\frac{j\pi}{3} P_3} - 1 \right) - \frac{(n - 1)}{P_4} \left( e^{\frac{j\pi}{3} P_4} - 1 \right) - \frac{(n - 1)}{P_5} \left( e^{\frac{j\pi}{3} P_5} - 1 \right) - \frac{(n + 1)}{P_6} \left( e^{\frac{j\pi}{3} P_6} - 1 \right) + \frac{(n + 1)}{P_7} \left( e^{\frac{j\pi}{3} P_7} - 1 \right) + \frac{(n - 1)}{P_8} \left( e^{\frac{j\pi}{3} P_8} - 1 \right) \right] \]

Using the same approach \( \Pi_y \) and \( \Pi_z \) is found to be:

\[ \Pi_y = \frac{3 C_y}{4(1 - n^2)} \left[ \frac{(n - 1)}{P_1} \left( e^{\frac{j\pi}{3} P_1} - 1 \right) + \frac{(n + 1)}{P_2} \left( e^{\frac{j\pi}{3} P_2} - 1 \right) - \frac{(n + 1)}{P_3} \left( e^{\frac{j\pi}{3} P_3} - 1 \right) - \frac{(n - 1)}{P_4} \left( e^{\frac{j\pi}{3} P_4} - 1 \right) + \frac{(n - 1)}{P_5} \left( e^{\frac{j\pi}{3} P_5} - 1 \right) + \frac{(n + 1)}{P_6} \left( e^{\frac{j\pi}{3} P_6} - 1 \right) - \frac{(n + 1)}{P_7} \left( e^{\frac{j\pi}{3} P_7} - 1 \right) - \frac{(n - 1)}{P_8} \left( e^{\frac{j\pi}{3} P_8} - 1 \right) \right] \]

\[ -j \frac{3 C_y}{2(4 - n^2)} \left[ \frac{(n - 2)}{P_2} \left( e^{\frac{j\pi}{3} P_2} - 1 \right) + \frac{(n + 2)}{P_4} \left( e^{\frac{j\pi}{3} P_4} - 1 \right) + \frac{(n + 2)}{P_5} \left( e^{\frac{j\pi}{3} P_5} - 1 \right) - \frac{(n - 2)}{P_7} \left( e^{\frac{j\pi}{3} P_7} - 1 \right) \right] \]
\[ \Pi_z = \frac{3 \, C_z}{2(1 - n^2)} \left[ \frac{(n - 1)}{P_1} \left( e^{j \frac{\pi}{3}} P_1 \cdot 1 \right) + \frac{(n + 1)}{P_2} \left( e^{j \frac{\pi}{3}} P_2 \cdot 1 \right) \right. \\
- \frac{(n + 1)}{P_3} \left( e^{j \frac{\pi}{3}} P_3 \cdot 1 \right) - \frac{(n - 1)}{P_4} \left( e^{j \frac{\pi}{3}} P_4 \cdot 1 \right) \\
+ \frac{(n - 1)}{P_5} \left( e^{j \frac{\pi}{3}} P_5 \cdot 1 \right) + \frac{(n + 1)}{P_6} \left( e^{j \frac{\pi}{3}} P_6 \cdot 1 \right) \right] \\
+ \frac{3 \, C_z}{2(4 - n^2)} \left[ \frac{(n - 2)}{P_2} \left( e^{j \frac{\pi}{3}} P_2 \cdot 1 \right) - \frac{(n + 2)}{P_4} \left( e^{j \frac{\pi}{3}} P_4 \cdot 1 \right) \right. \\
+ \frac{(n + 2)}{P_5} \left( e^{j \frac{\pi}{3}} P_5 \cdot 1 \right) - \frac{(n - 2)}{P_7} \left( e^{j \frac{\pi}{3}} P_7 \cdot 1 \right) \right] \\
\]

where

\[ C_y = C_x \sqrt{\frac{3}{\sqrt{3}}} \]

\[ C_z = \frac{30(e - 1)}{120\pi} \sqrt{\frac{2\pi}{kz_0}} \sin \left( \frac{kz_0 h}{r_0} \right) \left( \frac{3a}{2\pi} \right)^2 e^{-jk \left( r_0 + \frac{3x_0}{r_0} \right)} A_{1,0,1} \]
APPENDIX C

COMPUTER PROGRAM TO CALCULATE THE RADIATION LOSS OF A TRIANGULAR RESONATOR

The input to the program consists of the length of the triangle's side (mm), the substrate thickness (mm) and the relative dielectric constant. The print-out contains the resonance frequency (Hz), the effective dielectric constant, the radiated power (W), the stored energy times the frequency (ω), the percentage radiated power, the radiation Q and the product h . f (m·Hz). The program uses a computer library subroutine to calculate a double integration.
THIS PROGRAM CALCULATES THE POWER RADIATED
FROM A TRIANGULAR SHAPED MICROSTRIP RESONATOR

5000 C
THIS IS THE MAIN FROG.

6000 C
EXTERNAL F,INTPOW,FX1,FY2

7000 COMMON A,H,EPSSREL,EPSEFF,RO,A101,F1,F2,C1,CY,CZ

8000 I MPPLICIT DOUBLE PRECISION (A-Z)

9000 C
INTEGER YES

10000 WRITE(108,10)

11000 10 FORMAT(/'RADIATION ANALYSIS OF THE TRINGULAR RESONATOR'/)

12000 10 WRITE(108,20)

13000 20 FORMAT(/'ENTER TRIANGLE SIDE LENGTH IN METRIC UNITS'/)

14000 WRITE(108,51)

15000 51 FORMAT(/'RELATIVE DIELECTRIC CONST.'/)

16000 WRITE(108,52)

17000 52 FORMAT(ALL NUMBERS IN *8.2 FORM')

38000 1 READ(105,1) A,H,EPSSREL

19000 10 FORMAT(3E15.8)

20000 P=3.1415922535D0

21000 N=1.0D0

22000 H=1.0D0

23000 A101=1.0D0

24000 RO=1.0D0

25000 W=ABSORT(3.0D0)/3.0D0

26000 HOVER=1.0D0/((10.0D0)/W)

27000 EPSEFF=(EPSSREL+1.0D0)/2.0D0+((EPSSREL-1.0D0)/2.0D0)

28000 1X=1.0D0/DSORT(HOVERW))

29000 EPS=1.0-9/(F136.0)

30000 STENE=(EPSSREL*EPS**((A101)**2)*9.0D0/DSORT(3.0D0))/30.0D0

31000 RFREQ=2.0D0/(A101*DSORT(EPSSREL))

32000 STPWW=STPWW*FREQ

33000 WRITE(109,10)

34000 10 FORMAT('THE RESONANCE FREQUENCY IN Hz., IS';)

35000 WRITE(108,9) FREQ

36000 9 FORMAT(2X,D16.8)

37000 WRITE(109,48)

38000 48 FORMAT(2X,'THE EFFECTIVE DIELECTRIC CONSTANT IS';)

39000 WRITE(108,47) EPSEFF

40000 47 FORMAT(2X,D16.8)'/'

41000 CALL DINTO(DINTO,0.0,1.570796327,1.0,F1,F2)

42000 1.1+F1*INTPOW)

44000 C
THE INTEGRATION LIMITS ARE PHI=0 TO 2*PI

45000 C
PHI=0 TO 2*PI

46000 C

47000 ANSWER=DINTG/(2.0D0*120.0D0)

48000 CONDR:+((2.0D0*RFREQ*STPWW)*ANSW)/ANSWER

49000 PFRAD=ANSW*100.0D0/STPWW

5000 C
WRITE(109,20)
71.000 20 FORMAT('-------- THE RADIATED POWER IS --------
52.000 21 WRITE(10,31) ANSWER
53.000 22 FORMAT(3X,B12.3)
54.000 23 WRITE(103,60)
55.000 60 FORMAT('------ STORED ENERGY & "F" ------
56.000 61 WRITE(108,11) STOP
57.000 62 WRITE(109,70)
58.000 70 FORMAT(2X,'PERCENTAGE RADIATED POWER ------
59.000 71 WRITE(108,71) PPRAD
60.000 72 FORMAT(3X,B12.3)
61.000 73 WRITE(103,72)
62.000 74 FORMAT(2X,'RADIATION Q ------
63.000 75 WRITE(109,71) NORM
64.000 76 IF(NRPPFV.EQ.
65.000 77 WRITE(103,61)
66.000 61 FORMAT('-------CONNECTED RADIATION Q ------
67.000 62 WRITE(109,71) CORGR
68.000 63 WRITE(109,41)
69.000 41 FORMAT('--------- dWF ---------
70.000 42 WRITE(10,71) HF
71.000 43 WRITE(109,42)
72.000 44 FORMAT(2X,'IF YOU WANT TO CONTINUE ENTER 1, IF NOT ENTER 0."
73.000 45 READ(103,24) ES
74.000 46 FORMAT(10)
75.000 47 IF(YES.EQ.1.) GO TO 100
76.000 48 STOP
77.000 END
78.000 C
79.000 C
80.000 C
81.000 C
82.000 C
83.000 SECOND INTEGRAL LOWER LIMIT
84.000 C
85.000 C
86.000 C
87.000 C
88.000 C
89.000 C
90.000 C
91.000 C
92.000 C
93.000 C
94.000 C
95.000 C
96.000 C
97.000 C
98.000 C
99.000 C
100.000 C

THIS IS THE FUNCTION SUB TO CALCULATE THE INTEGRAL FUN.
101.000 C
102.000 FUNCTION INTPOW(THETA, PHI)
103.000 IMPLICIT DOUBLE PRECISION(A-Z)
104.000 DOUBLE COMPLEX Z, ZX1, ZX2, ZX3, ZX4, ZX5, ZX6, ZX7, ZX8, ZX9,
105.000 IP1, ZY, PI, ZZZ, PI1, ETA, EPSK, EPSF, R, THETA, PHI
106.000 COMMON M, N, EPSK, EPSF, R, A01, PI, CA, CY, C2
107.000 Y0, DSIN(THETA) * DCOS(PHI)
108.000 X0, DSIN(THETA) * DSIN(PHI)
109.000 Z0, DCOS(THETA)
110.000 C = 3.0 * PI / (1.0 * DSORT(EPSK) + 3.0)
111.000 GAMMA = 4.0 * PI
112.000 M = 3.0 * N1 / DSORT(EPSK) + 3.0
113.000 M = 3.0 * N1 / DSORT(EPSK) + 3.0
114.000 PI1 = DSORT(EPSK) + 1.0 / 3.0
115.000 PI1 = DSORT(EPSK) + 1.0 / 3.0
116.000 GAMMA = M + N2 * 2.0
117.000 GAMMA = M + N2 * 2.0
118.000 GAMMA = M + N2 * 2.0
119.000 OMEGA = M + N2 * 2.0
120.000 OMEGA = M + N2 * 2.0
121.000 OMEGA = M + N2 * 2.0
122.000 OMEGA = M + N2 * 2.0
123.000 OMEGA = M + N2 * 2.0
124.000 OMEGA = M + N2 * 2.0
125.000 OMEGA = M + N2 * 2.0
126.000 OMEGA = M + N2 * 2.0
127.000 OMEGA = M + N2 * 2.0
128.000 B1 = M + N2 * 2.0
129.000 B2 = M + N2 * 2.0
130.000 B3 = M + N2 * 2.0
131.000 B4 = M + N2 * 2.0
132.000 B4 = M + N2 * 2.0
133.000 B4 = M + N2 * 2.0
134.000 CX = Z * DSORT(EPSK) + 3.0 * DSORT(EPSK) + 3.0
135.000 CX = Z 
136.000 R = 3.0 * PI1 / DSORT(EPSK) + 3.0
137.000 R = 3.0 * PI1 / DSORT(EPSK) + 3.0
138.000 R = 3.0 * PI1 / DSORT(EPSK) + 3.0
139.000 R = 3.0 * PI1 / DSORT(EPSK) + 3.0
140.000 R = 3.0 * PI1 / DSORT(EPSK) + 3.0
141.000 R = 3.0 * PI1 / DSORT(EPSK) + 3.0
142.000 R = 3.0 * PI1 / DSORT(EPSK) + 3.0
143.000 R = 3.0 * PI1 / DSORT(EPSK) + 3.0
144.000 R = 3.0 * PI1 / DSORT(EPSK) + 3.0
145.000 R = 3.0 * PI1 / DSORT(EPSK) + 3.0
146.000 R = 3.0 * PI1 / DSORT(EPSK) + 3.0
147.000 R = 3.0 * PI1 / DSORT(EPSK) + 3.0
148.000 R = 3.0 * PI1 / DSORT(EPSK) + 3.0
149.000 R = 3.0 * PI1 / DSORT(EPSK) + 3.0
150.000 R = 3.0 * PI1 / DSORT(EPSK) + 3.0
APPENDIX D

COMPUTER PROGRAM TO FIND THE RADIATION PATTERN OF A TRIANGULAR RESONATOR

This program requires the starting point (the angles $\theta$ and $\phi$) in addition to $A$, $h$ and $\varepsilon_r$. In the program, $\theta$ varies from 0 to $90^\circ$ with increments of $10^\circ$ while $\phi$ is held constant. The printed results are $\theta$, the modules of the far-field $E$ vector in the $\theta$ direction, the modules of the far-field $E$ vector in the $\phi$ direction and the poynting vector at that point.
THIS IS THE PROGRAM TO CALCULATE THE RADIATION PATTERN OF A TRIANGULAR RESONATOR.

IMPLICIT DOUBLE PRECISION(A-Z)

INTEGER J

DOUBLE COMPLEX ZZ,ZX1,ZX2,ZX3,ZX4,ZX5,ZX6,ZX7,ZX8,ZXX,

IPX,IPY,IPZ,IPZ,ETHETA,EPHI

WRITE(108,11)

FORMAT(’----- ENTER TRIANGLE SIDE LENGTH, SUBSTRATE ’)

WRITE(108,12)

FORMAT(’ THICKNESS, IN MILS, RELATIVE DIELEC. CONSTANT---’)

READ(105*13) A,H,EPRES

FORMAT(3G)

WRITE(108,14)

FORMAT(’ ENTER THETA AND PHI IN DEGREE ’)

WRITE(108,15) THETA,PHI

WRITE(108,16) THETA,PHI

FORMAT(3G)

WRITE(108,20)

FORMAT(’/ULATE', PHII PHII)

WRITE(108,21) PHI

WRITE(108,22) PHI

WRITE(108,22) PHI

FORMAT(’----- THETA ----- MOD. ETHETA ----- MOD. EPHI

1------ POY. VECTOR------’)

A=2.54D-5*A

H=2.54D-5*H

W=(A*DSQRT(3.0D0))/3.0D0

HOVERM=1.0D0+((&(10.0D0*H)/W)

EPSEFF=((EPRES+1.0D0)/2.0D0)+((EPRES-1.0D0)/2.0D0)

1*(1.0D0/DSQRT(HOVERM))

R0=4.0D0

A101=1.0D0

PI=3.141592654D0

K=4.0D0*PI/(3.0D0*DSQRT(EPRES))

PI1=DSQRT(EPRES)/(1.0D0*PI)

PI2=1.0D0/(1.0D0*PI)

DO 100 J=1,10

THETA=THETA/57.2957793D0

PHI=PHI/57.2957793D0

YORO=DSIN(THETA)*DCOS(PHI)

XORO=DSIN(THETA)*DSIN(PHI)

ZORO=DCOS(THETA)

GAMMA=K*ZORO*H

M=(3.0D0*K*YORO*H*DSQRT(3.0D0))/(2.0D0**PI)

N=(3.0D0*K*YORO**2)/(2.0D0**PI)

GAMMA1=M+N+2.0D0

GAMMA2=M+N-2.0D0

GAMMA3=M-N+2.0D0

GAMMA4=M-N-2.0D0

OMEGA1=M+N+4.0D0

OMEGA2=M+N-2.0D0
51.000 OMEGA3=M+N+2.D0
52.000 OMEGA4=M+N+4.D0
53.000 OMEGA5=M+N+4.D0
54.000 OMEGA6=M-N-2.D0
55.000 OMEGA7=M-N+2.D0
56.000 OMEGA8=M-N-4.D0
57.000 B1=N+1.D0
58.000 B2=N+1.D0
59.000 B3=N-2.D0
60.000 B4=N-2.D0
61.000 
62.000 
63.000 
64.000 
65.000 
66.000 
67.000 
68.000 
69.000 
70.000 QY1=L1B3/GAMMA1
71.000 QY2=L1B4/GAMMA2
72.000 QY3=L1B4/GAMMA3
73.000 QY4=L1B3/GAMMA4
74.000 QY5=L2B1/OMEGA1
75.000 QY6=L2B2/OMEGA2
76.000 QY7=L2B2/OMEGA3
77.000 QY8=L2B2/OMEGA4
78.000 QY9=L2B2/OMEGA5
79.000 QY10=L2B2/OMEGA6
80.000 QY11=L2B2/OMEGA8
81.000 
82.000 
83.000 
84.000 
85.000 
86.000 
87.000 
88.000 
89.000 
90.000 
91.000 
92.000 
93.000 
94.000 
95.000 
96.000 
97.000 
98.000 
99.000 
100.000
2TY1*DCOS(OMEGA1*PI/3.0) - TY2*DCOS(OMEGA2*PI/3.0) + TY2
3+TY1*TY3*DCOS(OMEGA3*PI/3.0) + TY3*TY4*DCOS(OMEGA4*PI/3.0)
4+TY4+TY5*DCOS(OMEGA5*PI/3.0) - TY5+TY6+DCOS(OMEGA6*PI/3.0)
5+TY6+TY7*DCOS(OMEGA7*PI/3.0) - TY7+TY8*DCOS(OMEGA8*PI/3.0)
6-TY8
ZYY=DCMPLX(AYSE, FATOS)
PIY=CY*CDEXP(ZZ)*ZYY
FATHA=TY1*DCOS(OMEGA1*PI/3.0) - TY1+TY2*DCOS(OMEGA2*PI/3.0)
I=TY2*TY3*DCOS(OMEGA3*PI/3.0) - TY3*TY4*DCOS(OMEGA4*PI/3.0)
2+TY4+TY5*DCOS(OMEGA5*PI/3.0) + TY5*TY6*DCOS(OMEGA6*PI/3.0)
3+TY6+TY7*DCOS(OMEGA7*PI/3.0) + TY7+TY8*DCOS(OMEGA8*PI/3.0)
4+TY8+TY9*DCOS(GAMMA1*PI/3.0) - QY1+QY2+QY3*DCOS(GAMMA2*PI/3.0)
5-QY2-QY3*DCOS(GAMMA3*PI/3.0) + QY3-QY4*DCOS(GAMMA4*PI/3.0)
6+QY4)*0.500
INCI=TY1*DSIN(OMEGA1*PI/3.0) + TY2*DSIN(OMEGA2*PI/3.0) +
1+TY3*DSIN(OMEGA3*PI/3.0) + TY4*DSIN(OMEGA4*PI/3.0) -
17.000
2TY5*DSIN(OMEGA5*PI/3.0) - TY6*DSIN(OMEGA6*PI/3.0) -
3+TY7*DSIN(OMEGA7*PI/3.0) - TY8*DSIN(OMEGA8*PI/3.0) +
19.000
4QY1*DSIN(GAMMA1*PI/3.0) - QY2*DSIN(GAMMA2*PI/3.0) - QY3*
5DSIN(GAMMA3*PI/3.0) - QY4*DSIN(GAMMA4*PI/3.0) + QY4*DSIN(PI/3.0) + 0.500
121.000
2ZYY=DCMPLX(FATHA, INCI)
PIZ=CZI*CDEXP(ZZ)*ZZZ
123.000
ETHETA=(K*Z2)*(PIY*DCOS(ETHETA) + DSIN(PHI) + FIX*DCOS(THETA) * DCOS(PI)
124.000
1-ETHETA
125.000
EPHI=(K*Z2)*(PIY*DCOS(PHI) - FIX*DSIN(PHI))
126.000
MODETH=CDABS(ETHETA)
127.000
MODEPH=CDABS(EPI)
128.000
POYVEC=((MODETH*2) + (MODEPH*2))/(1.202*PI)
129.000
THETA=THETA+57.29577930
130.000
PHI=PHI+57.29577930
131.000
WRITE(108, 23) THETA, MODETH, MODEPH, POYVEC
132.000
CONTINUE
133.000
STOP
136.000
END
REFERENCES


