Robust patch antenna design

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Abstract

The research in this report is done in the framework of the 60 GHz band research at the Eindhoven University of Technology. The 60 GHz band is of great interest since there is a massive spectral space (5 GHz) allocated worldwide for dense wireless communication. To use the spectral space a reliable and low-cost interconnection is needed. An important part of this interconnection is the antenna. Smart antenna structures are needed to optimize signal transmission and reliability. The antenna focused on in this report is the square patch antenna. To excite the antenna a coaxial probe feed is used. To realize a low-cost solution, the antenna design must be robust for tolerances in production. To give an insight into patch antenna technology and to derive the relevant parameters of a patch antenna the cavity model is used. The obtained parameters are: the frequency, the dielectric constant, the loss tangent, the thickness of the substrate, the width and length of the patch and the probe position. The tolerance of the parameters are calculated using a MATLAB program based on the cavity model and the software tool FEKO, which uses the method of moments. After simulations an optimal height of the patch and an optimal thickness of the coaxial probe have been found which enhance patch antenna efficiency and robustness. Also it is found to be very important to make the width of the antenna as precise as possible, since this is the most sensitive parameter of the patch. After the simulations it has become clear that the cavity model can only be used within a stringent parameter range.
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Chapter 1

Introduction

The research in this report is done in the framework of the 60 GHz band research at the Eindhoven University of Technology. The 60 GHz band is of great interest since there is a massive spectral space (5 GHz) allocated worldwide for dense wireless communication. To use the spectral space a reliable and low-cost interconnection is needed. An important part of this interconnection is the antenna. Smart antenna structures are needed to optimize signal transmission and reliability.

There are several options for 60 GHz antennas. The antenna focused on in this report is the square patch antenna [1, 2, 3, 4, 5, 6]. To excite the antenna a coaxial probe feed is used. To realize a low-cost solution, the antenna design must be robust for tolerances. Therefore the report focuses on the effect of tolerances on the patch antenna and tries to make the antenna more robust and efficient for the 60 GHz band. To give an insight into patch antenna technology an introduction is given in Chapter 2. Simulations to find the tolerances of the patch antenna are in Chapter 3. Finally Chapter 4 gives the conclusions of this report.
Chapter 2

Introduction to patch antenna technology

2.1 Introduction

To get an insight in the physical behavior of a patch antenna an analytical model will be derived. The model will be used to find the parameters that influence antenna performance.

2.2 Physical description of a patch antenna

A patch antenna is constructed out of a dielectric with a conductive ground plane underneath. On top of the dielectric is a conductive strip. The conductive strip on top of the dielectric is called the patch (Figure 2.1). The width $a$ and length $b$ of the antenna typically lies around $\lambda/2$, where $\lambda$ is the wavelength of the operating frequency. The height $h$ of the antenna is a small fraction of the wavelength. The patch conductor is normally copper and can take any shape, but generally simple geometries are used. This simplifies the analysis and performance prediction. Usually the patches are photo etched on the dielectric substrate.

There are different techniques to feed the patch antenna. The method used in this study is a coaxial feed line. A coaxial probe feeds the patch on the backside of the antenna. The inner conductor of the coaxial line is connected to the conductive strip and the outer conductor is connected to the conductive ground-plane (Figure 2.1). Other feed technologies are for example a coplanar feed line, where the feed line is on the same substrate as the patch and is directly connected to the patch. Another alternative is an electromagnetically coupled
feed line, where the feed line is not directly connected to the patch, but is electromagnetically coupled to the patch [7].

Between the patch and the ground plane the electric fields are perpendicular to the conductive planes. Near the edges of the patch the field bends outwards, these are called the "fringe fields". The bending of the fields near the edges increases the effective width and length of the antenna, as shown in Figure 2.2. Because of the bending of the fields the patch antenna radiates.

2.3 Modelling

2.3.1 Introduction

To determine the resonance frequency of the patch the cavity model is used. In this model the antenna is considered as a resonating cavity. This implies that the patch antenna is viewed as a box, with perfect electric conducting ground and top plane equal in size to the patch [8]. The side walls are assumed to be magnetic conductors. Because of the boundary conditions only electric fields exist which are perpendicular to the conducting planes. The magnetic field has a vanishing tangential component at the four side walls. The fields of the lowest resonating mode (assuming $a \geq b$) are given by:
\[ E_z(x) = -E_0 \cos\left(\frac{\pi x}{a}\right) \quad \text{for} \quad 0 \leq x \leq a \]
\[ H_y(x) = -H_0 \sin\left(\frac{\pi x}{a}\right) \quad \text{for} \quad 0 \leq x \leq a \]  

(2.1)

This means that the dimensions of the patch dictate the modes that can exist in this perfect cavity. The lowest possible frequency is often the frequency at which the patch is used.

### 2.3.2 Cavity model

Consider a rectangular patch of width \(a\) and length \(b\) over an infinite ground plane with a substrate of thickness \(h\) and a dielectric constant \(\varepsilon_r\), as shown in Figure 2.1. As long as the substrate is electrically thin, the electric field will be \(z\)-directed and the interior modes will be \(TM_{mn}\) to \(z\) so that according to the cavity model [6]:

\[ E_z(x, y) = \sum_m \sum_n A_{mn} \psi_{mn}(x, y), \]  

(2.2)

where \(A_{mn}\) are the mode amplitude coefficients and \(\psi_{mn}\) are the \(z\)-directed orthonormalized electric field mode vectors. For the elementary case of a non-radiating cavity,

\[ \psi_{mn}(x, y) = \cos k_m x \cos k_n y, \]  

(2.3)

with

\[ k_m = \frac{m\pi}{a} \quad \text{and} \quad k_n = \frac{n\pi}{b}. \]  

(2.4)

The resonance condition is given by:

\[ k^2 = k_m^2 + k_n^2 = k_{mn}^2, \]  

(2.5)

where the propagation constant \(k = \frac{\sqrt{\varepsilon_r} \lambda_0}{\lambda_0}\). Here \(\lambda_0\) represents the wavelength of the fields in free space. The amplitude coefficients are given by:

\[ A_{mn} = \frac{j \lambda_{mn}^2}{\omega \mu abh \left( k^2 - k_{mn}^2 \right)} \int_V J_z \psi_{mn}(x, y) dV, \]  

(2.6)
with $\mu$ the permeability of the dielectric, $J_z$ the current source, $V$ the volume of the patch and $\omega = 2\pi f$ with $f$ the frequency. Here $\chi_{mn}$ is given by:

$$
\chi_{mn} = \begin{cases} 
1, & m = 0 \text{ and } n = 0 \\
\sqrt{2}, & m = 0 \text{ or } n = 0 \\
2, & m \neq 0 \text{ and } n \neq 0 
\end{cases}.
$$

(2.7)

If it is assumed the patch antenna has a probe feed at position $x_p$ and $y_p$ with negligible diameter, the current source $J_z$ can be modelled as $J_z = I_0 \delta(x - x_p)\delta(y - y_p)$. Here $I_0$ can be chosen constant because of the limited height of the substrate. The amplitude coefficients are then given by:

$$
A_{mn} = jI_0 \eta \chi_{mn}^2 \frac{\kappa}{k^2 - k_{mn}^2} \psi_{mn}(x_p, y_p),
$$

(2.8)

where $\eta = \sqrt{\frac{\mu}{\varepsilon}}$ represents the intrinsic impedance.

In practice the dielectric material is not perfectly isolating, the patch and ground plane are not perfectly conducting and the cavity has no side walls which are perfect magnetic conductors. The absence of the side walls means that the electric field on the edge bends outwards this can be accounted for in a cavity which is slightly larger than the dimensions of the patch (Figure 2.1).

The absence of the side walls means that the antenna radiates from these sides. Parameters that influence the radiation properties of the antenna, next to the dimensions of the cavity, are the frequency, the conductivity of the patch and ground plane, the dielectric constant of the dielectric and the loss tangent of the dielectric.

### 2.3.3 Radiation pattern

In the previous section the electric field inside the patch has been presented. For the $TM_{10}$ mode the field distribution, according to the cavity model, is shown in Figure 2.3. From this calculated electric field it is possible to derive the far field pattern. This is done by using the equivalent magnetic current density $\vec{M} = \vec{E} \times \vec{n}$ [9] on the side walls of the antenna according to Love’s equivalence principle [10]. This is illustrated in Figure 2.4.

From the equivalent magnetic current density on the side walls the electric potential vector for a side wall can be derived [11]

$$
\vec{F} = \frac{\varepsilon_0}{4\pi} \int_S \frac{\vec{M}_S e^{-j\kappa_s |\vec{R}|}}{|\vec{R}|} dS, 
$$

(2.9)
Figure 2.3: electric field distribution of a standard patch for $TM_{10}$ mode

Figure 2.4: electric field distribution (a) and equivalent magnetic current density (b)
where $\varepsilon_0$ represents the permittivity of free space and $k_0 = \frac{2\pi}{\lambda_0}$ the propagation constant. Here $\lambda_0$ represents the wavelength of the fields in free space. The surface $S$ represents the side wall and $|\vec{R}|$ the distance between the source and observation point.

To determine the radiation pattern of a rectangular patch antenna each side is assumed to be an independent radiating aperture. The expression for the radiated electric field in the far field region for the side $x = 0$ given in spherical coordinates (Figure 2.5) is:

$$E_{rad} = -j\frac{k_0}{\varepsilon_0} F_y (\cos(\theta) \sin(\phi) \vec{a}_\phi + \cos(\phi) \vec{a}_\theta),$$  \hspace{1cm} (2.10)

where

$$F_y = \frac{\varepsilon_0 b h \vec{a}_y e^{-jk_0 r}}{\pi r} A_{mn} \left[ (-1)^n e^{-j2Y} - 1 \right] \frac{Y}{4Y^2 - n^2 \pi^2} \sin(Z) e^{-jZ},$$  \hspace{1cm} (2.11)

and

$$Y = \frac{b}{2} k_0 \sin(\theta) \sin(\phi),$$  \hspace{1cm} (2.12)

$$Z = \frac{h}{2} k_0 \cos(\theta).$$

Every opposite side has an equal radiation pattern with a phase difference due to the mode and the distance between the sides. The total radiated field can be calculated by considering the antenna as two arrays of two aperture antennas. The array factor for two identical antennas with phases $\alpha_1$, $\alpha_2$ and a vectorial spacing $\vec{d}$ is given by [12]
\[ AF = 2 \cos \left( \frac{k_0 \vec{d} \cdot \hat{r}}{2} + \frac{\alpha_1 - \alpha_2}{2} \right) e^{j \frac{\alpha_1 - \alpha_2}{2}}. \] (2.13)

Thus the total radiated field is given by

\[ \vec{E}_{\text{rad}} = \vec{E}_{\text{rad}}^a AF_a + \vec{E}_{\text{rad}}^b AF_b, \] (2.14)

with

\[ \vec{E}_{\text{rad}}^a = k_0 \frac{b h}{\pi r} e^{-jk_0 r} A_{mn} Y \left[ \frac{(-1)^n e^{-j2Y} - 1}{4Y^2 - n^2 \pi^2} \right] \sin(Z) e^{-jZ} (\cos(\theta) \sin(\phi) \vec{a}_\phi + \cos(\phi) \vec{a}_\theta), \]
\[ \vec{E}_{\text{rad}}^b = k_0 \frac{a h}{\pi r} e^{-jk_0 r} A_{mn} X \left[ \frac{(-1)^n e^{-j2X} - 1}{4X^2 - n^2 \pi^2} \right] \sin(Z) e^{-jZ} (\cos(\theta) \sin(\phi) \vec{a}_\phi - \sin(\phi) \vec{a}_\theta), \] (2.15)

and

\[ AF_a = 2 \cos(X - \alpha_m) e^{j\alpha_m}, \]
\[ AF_b = 2 \cos(Y - \alpha_n) e^{j\alpha_n}, \] (2.16)

where

\[ X = a \frac{k_0}{2} \sin(\theta) \cos(\phi), \]
\[ Y = b \frac{k_0}{2} \sin(\theta) \sin(\phi), \]
\[ Z = h \frac{k_0}{2} \cos(\theta), \] (2.17)

and

\[ \alpha_i = \begin{cases} \frac{\pi}{2} & \text{for } i = 2 \\ 0 & \text{for } i = 1 \end{cases}. \] (2.18)

The radiated field for multiple modes is found by summing the radiated fields of the independent modes.

The total radiated power for the mode with indices \( mn \) can be found by integrating the square of the radiated field over half a hemisphere, since the antenna does not radiate below the infinite ground plane,

\[ P_{\text{rad}} = \int_{\theta=0}^{\pi/2} \int_{\phi=0}^{2\pi} \frac{1}{2\mu_0} |\vec{E}_{\text{rad}}|^2 r^2 \sin(\theta) d\phi d\theta, \] (2.19)
with $\eta_0 = \sqrt{\frac{\mu_0}{\varepsilon_0}}$ the wave impedance for vacuum.

### 2.3.4 Input impedance

#### Lossless input impedance

The input impedance of the antenna is defined as

$$Z_{\text{in}} = -\frac{1}{|I_0|^2} \int_V \vec{E} \cdot \vec{J}^* dV,$$

where $I_0$ is the injected current. If the field distribution of equation (2.2) and a volume $V$ equal to the inner space of the cavity is used, the input impedance will be imaginary because there are no losses in the cavity model. The input impedance then can be modelled as a sum of reactances $X_{mn}$ [6].

$$Z_{\text{in}} = -j \sum_m \sum_n \eta \chi_{mn} h \frac{k}{k^2 - k_{mn}^2} \psi_{mn}^2(x_p, y_p) = -j \sum_m \sum_n \chi_{mn}. \quad (2.21)$$

This is equivalent to a series of $LC$ circuits with network parameters [6]

$$L = \sqrt{\mu_0 \eta \chi_{mn}^2 h} \frac{1}{\psi_{mn}^2(x_p, y_p)},$$

$$C = \sqrt{\frac{1}{\eta \chi_{mn}^2 h \psi_{mn}(x_p, y_p)}}. \quad (2.22)$$

#### Losses

In Equation (2.21) losses due to radiation, non-perfect isolation of the dielectric and non-perfect conduction of the conducting planes are not considered. To account for these losses an extra resistance term is added for each mode.

The input impedance of a single mode is given by

$$Z_{\text{in}} = \frac{j\omega RL}{R - \omega^2 RLC + j\omega L} = -\frac{j\omega \frac{1}{C}}{\omega^2 - \omega_0^2(1 + \frac{1}{Q})}, \quad (2.23)$$

where the resonance frequency equals $\omega_r = \frac{1}{\sqrt{LC}}$ and the quality factor is given by $Q = \frac{R}{2L}$. The total quality factor can be calculated from the quality factors that are related to the separate losses [10].
\[
\frac{1}{Q} = \frac{1}{Q_{rad}} + \frac{1}{Q_c} + \frac{1}{Q_d},
\]  
\tag{2.24}

with

\[
Q_{rad} = \frac{2\omega W_E}{P_{rad}}. \tag{2.25}
\]

Here \( P_{rad} \) is the total radiated power as defined in Equation (2.19) and the stored electric energy \( W_E \) is given by the volume integral over the squared electric field within the cavity

\[
W_E = \int_V \varepsilon |\mathbf{E}|^2 dV = |A_{mn}|^2 \frac{\varepsilon_0 h^2}{4\lambda_{mn}^2}. \tag{2.26}
\]

The quality factor losses in the conducting planes are given by

\[
Q_c = \frac{\hbar}{\sqrt{\frac{\omega}{\pi}} \mu_0 \sigma}, \tag{2.27}
\]

where \( \sigma \) is the conductivity of the patch and the ground plane. The losses in the dielectric are given by

\[
Q_d = \frac{1}{\tan \delta} = \frac{\omega \varepsilon}{\sigma_d}, \tag{2.28}
\]

where \( \tan \delta \) is the loss tangent of the dielectric and \( \sigma_d \) represents the conductivity of the dielectric.

**Input impedance model**

The input impedance can be modelled as a cascade of \( RLC \) circuits. This approach will result in a slowly converging series that can take long computation time. Therefore the model is simplified. The \( TM_{00} \) mode has no inductance \( L \) and a capacitance of \( C = \frac{ab\varepsilon}{h} \). The impedance of the \( TM_{00} \) mode is therefore modelled as a parallel \( RC \) circuit

\[
Z_{00} = \frac{R}{1 + j\omega C} = -\frac{j}{Q^2}, \tag{2.29}
\]
where $Q$ is calculated by equation (2.24). The non-radiating modes, higher order modes, have no radiation losses. The inductive reactance of the higher order modes can be approximated by [8]

$$X_L = \sqrt{\frac{\mu}{\varepsilon}} \tan(\omega \sqrt{\mu\varepsilon} h).$$ (2.30)

The higher order modes are modelled as a parallel $RL$ circuit

$$Z_{RL} = \frac{j\omega RL}{R + j\omega L} = \frac{j\omega L}{1 + \frac{L}{Q}},$$ (2.31)

with $\frac{1}{Q} = \frac{1}{Q_c} + \frac{1}{Q_d}$. Another method to account for the higher order modes is to approximate the converging series of $RLC$ circuits with a limited number of higher order modes. Good approximation can be achieved by limiting the modes until $TM_{10,10}$ if the $TM_{10}$, $TM_{01}$ or $TM_{11}$ mode is excited.

To excite the antenna optimal and minimize reflections a matched input impedance is required. The input impedance of the antenna is maximal when it is excited on the edges and minimal if the patch is excited in the center, as can be derived by combining equation (2.3) and equation (2.8). Therefore the antenna will be excited somewhere between the edges and the center, depending on the impedance of the feed line.

### 2.4 Patch antenna parameters

From the cavity model the parameters that determine the characteristics of the patch antenna are derived: the frequency, the dielectric constant, the loss tangent, the thickness of the substrate, the width and length of the patch and the probe position. In the next chapters the tolerance of the parameters will be calculated using a MATLAB program based on the cavity model and the software tool FEKO, which uses the method of moments. The parameters that will be tested are:

- frequency
- dielectric constant
- loss tangent
- thickness of the substrate
- width and length of the patch
- probe position
2.5 Conclusions

The parameters which influence antenna performance have been derived using the cavity model. In the next chapter the effects of these parameters on antenna performance will be calculated with a MATLAB program and FEKO, a program based on the method of moments.
Chapter 3

Simulating a patch antenna

3.1 Introduction

The effects of the parameters on antenna performance will be calculated with a MATLAB program based on the cavity model and FEKO, a program based on the method of moments. For simulations a standard square patch antenna will be used. The results of the simulations will be used to simulate an optimized square patch antenna at 60 GHz.

3.2 Simulating using MATLAB

3.2.1 Introduction

To simulate a patch antenna the cavity model is implemented in MATLAB. The cavity model will be used to determine the effect of the different parameters on patch antenna performance. For the simulations a standard patch was used. The dimensions of the antenna are: width $a = 29.00\, \text{mm}$, length $b = 29.00\, \text{mm}$ and height $h = 1.6\, \text{mm}$. The probe position is $x_p = 8.0\, \text{mm}$, $y_p = 14.5\, \text{mm}$ and the probe diameter $d = 0.8\, \text{mm}$. The conductance of the copper ground-plane and patch is $\sigma = 5.8 \cdot 10^{-7}\, \Omega^{-1}\, \text{m}^{-1}$. The dielectric constant of the dielectric equals $\varepsilon_r = 4.28$ and the loss tangent is given by $\tan \delta = 1.6 \cdot 10^{-2}$. This corresponds to a patch antenna etched on FR4 material with a metal layer made of copper. The operating frequency corresponds to 2.45 GHz. This frequency lies within an ISM-band (Industrial, Scientific and Medical band), where license-free operation is permitted [6].
3.2.2 Simulation results

Radiation pattern

The cavity model was used to create a three dimensional model of the radiation pattern of the standard patch antenna. The result can be seen in Figure 3.1. The pattern is symmetrical for the $\varphi = 0^\circ$ plane and the $\varphi = 90^\circ$ plane and is dominated by the $TM_{10}$ mode due to the probe position.

Input impedance

The cavity model is used to calculate the input impedance of the standard patch. The result is compared to a full wave simulation for verification (Figure 3.2). Figure 3.2 shows the cavity model has a mismatch of 1\% in the resonating frequency and a 10\% mismatch of the absolute value of the input impedance at the resonance frequency compared to the full wave simulation.

The effect of parameters on patch performance

To excite the antenna a coaxial probe feed is used. To limit reflections the feed line is matched to the input impedance at the resonating frequency. To compare the effects of varying parameters it is necessary to cancel out the effect of mismatch of the probe on the other frequencies. Therefore the effective
radiated power is used. The effective radiated power $P_{\text{effrad}}$ is the ratio between the radiated power $P_{\text{rad}}$ and the power received by the patch antenna $P_{\text{received}}$. With the received power being equal to the total power to the patch minus the power loss due to reflections $P_{\text{received}} = P_{\text{total}} - P_{\text{reflections}}$. The received power also equals the radiated power combined with the losses in the patch $P_{\text{received}} = P_{\text{rad}} + P_{\text{losspatch}}$. Therefore the effective radiated power $P_{\text{effrad}}$ equals

$$P_{\text{effrad}} = \frac{P_{\text{rad}}}{P_{\text{received}}}.$$  \hfill (3.1)

A comparison will be made between the cavity model implemented in MATLAB and a full wave simulation (Figure 3.3). As Figure 3.3 shows the cavity model can not calculate the effective power accurately for non-resonating frequencies. This is because the field distribution on the radiating edges as defined in Chapter 2.3.2 is an estimation which only holds near the resonating frequencies.

Next to this limitation of the cavity model is another problem as demonstrated in Figure 3.4. In Figure 3.4 the effective radiated power is shown as a function of the height $h$ and the frequency. The height is varied over a distance $0.1\,\text{mm} < h < 8\,\text{mm}$, which corresponds to $0.0017\lambda_{\text{res}} < h < 0.1379\lambda_{\text{res}}$. The height of the standard patch is $h = 1.6\,\text{mm}$, which corresponds to $h = 0.028\lambda_{\text{res}}$. Figure 3.4 suggests the patch could generate energy. This is clearly wrong. The effect is due to the assumption in the cavity model the height $h$ can be neglected when calculating the field in the cavity. This is because the patch antenna is considered to be electrically thin. When the height $h$ becomes larger the effective aperture of the patch becomes larger, since the side walls grow in size. Next to the gain in aperture is the extra loss in the dielectric due to the increased
thickness. In practise eventually the losses in the dielectric will overcome the extra gain and the effective radiated power of the antenna is reduced to a point where the patch is no longer considered to be an effective antenna. In the cavity model the losses are only accurately calculated when the height $h$ is electrically thin $h < 0.05 \cdot \lambda_{\text{res}}$, with $\lambda_{\text{res}}$ the wavelength of the resonance frequency [13]. When the height is increased $h = 0.05 \cdot \lambda_{\text{res}}$ no longer holds and the cavity model fails.

Figure 3.5 shows the model is not suitable for calculating the change in the effective radiated power due to deviations in the permittivity of the dielectric, $\varepsilon_r$. This is because frequencies beside the resonance frequency cannot be calculated accurately, as described above. A change in the permittivity results in a change of the resonance frequency since $f_{\text{res}} = \frac{c_0}{\sqrt{\varepsilon_r \lambda_{\text{res}}}}$, with $c_0$ the speed of light, $\varepsilon_r$ the relative permittivity and $\lambda_{\text{res}}$ the wavelength of the resonance frequency. Since the wavelength is determined by the dimensions of the patch (Equation 2.1) a change in $\varepsilon_r$ results in a shift in the resonance frequency.

Because of these reasons the cavity model is not suitable to calculate the antenna gain for the full spectrum also not at the resonance frequency. As a result the cavity model is unsuitable to be used as a tool for calculating the effect of parameters on antenna performance.

3.2.3 Conclusions

The cavity model is easy to implement in MATLAB or any other software tool. The model is flexible and gives good insight in the physical behavior of a patch
Figure 3.4: effective radiated power to frequency and height

Figure 3.5: effective radiated power to frequency and $\varepsilon_r$
antenna. The model can be used within a stringent parameter range to calculate the input impedance \( h < 0.05\lambda_{res} \). Due to the nature of the model its range of accuracy is limited and it is unsuitable for studying the effects of different parameters on antenna performance.
3.3 Simulating using FEKO

3.3.1 Introduction

To simulate a patch antenna the software tool FEKO, which uses the method of moments, is used. FEKO will be used to determine the effect of the different parameters on patch performance. For the simulations a standard patch was used. The dimensions of the antenna are: width $a = 29.00\,\text{mm}$, length $b = 29.00\,\text{mm}$ and height $h = 1.6\,\text{mm}$. The probe position is $x_p = 8.00\,\text{mm}$, $y_p = 14.5\,\text{mm}$ and the probe diameter $d = 0.8\,\text{mm}$. The conductance of the copper ground-plane and patch is $5.8 \cdot 10^{-7}\,\text{S}$. The dielectric constant of the dielectric equals $\varepsilon_r = 4.28$ and the loss tangent is given by $\tan\delta = 1.6 \cdot 10^{-2}$. This corresponds to a patch antenna etched on FR4 material with a metal layer made of copper. The operating frequency corresponds to 2.45 GHz. This frequency lies within an ISM-band (Industrial, Scientific and Medical band), where license-free operation is permitted [6].

3.3.2 Simulation results

Thickness of the substrate

The thickness of the substrate is varied over a height $h$: $0.5\,\text{mm} \leq h \leq 9.0\,\text{mm}$. This can also be written as: $0.0083\lambda_{res} \leq h \leq 0.151\lambda_{res}$, where $\lambda_{res} = \frac{c_0}{\varepsilon_r f_{res}}$, with $c_0$ the speed of light, $f_{res}$ the resonance frequency of the standard patch according to FEKO (Figure 3.2) and $\varepsilon_r$ the permittivity of the dielectric.

In Figure 3.6 the effective radiated power to frequency and height is shown. Figure 3.6 shows the effective radiated power has an optimum with respect to the height. The optimal height lies in the range $3\,\text{mm} \leq h \leq 5\,\text{mm}$, which can also be written as: $0.05\lambda_{res} \leq h \leq 0.0838\lambda_{res}$. At these values for the height $h$ the effective radiated power is nearly constant.

Also the shift in the resonance frequency due to a shift in height is calculated, $\frac{\partial f_{res}}{\partial h}$. In Figure 3.7 the imaginary input impedance to frequency and height is shown. The resonance frequency is defined as

$$\text{Im}(Z_{in}) = 0$$
$$\frac{\partial \text{Im}(Z_{in})}{\partial f} < 0. \quad (3.2)$$

This means that at the resonance frequency the imaginary input impedance equals zero and has a negative gradient. Due to linear interpolation in the $x$- and $y$-direction the zero line in Figure 3.7 is not smooth. The value of $\frac{\partial f_{res}}{\partial h} = -0.054\,\text{GHz/mm}$. To compare different parameters the gradient constant $G$ is
introduced, which represents the percentage shift of the resonance frequency due to a percentage shift in the value of the parameter. For the height the gradient constant is:

\[
G_h = \frac{\partial f_{res}}{\partial h} f_0 = \frac{\% \text{ shift in } f_{res}}{\% \text{ shift in height}} = -0.036,
\]  

(3.3)

where \( f_0 \) is the resonance frequency of the standard patch and \( h_0 \) is the height of the standard patch. This means that if the height \( h \) shifts 8\% the resonance frequency shifts 8\% \cdot G_h = -0.288\%. This constant holds for 0.0083\( \lambda_{res} \leq h \leq 0.0755\lambda_{res} \). This corresponds to 0.5\text{mm} \leq h \leq 4.5\text{mm} for the standard patch antenna.

The most robust height is the height between 0.05\( \lambda_{res} \leq h \leq 0.0755\lambda_{res} \), since there the effective radiated power is nearly constant and there is a small shift in resonance frequency. This corresponds to 3\text{mm} \leq h \leq 4.5\text{mm} for the standard patch antenna. The advised height therefore is:

\[
h_{advised} = 0.06275\lambda_{res}.
\]  

(3.4)

Note that the resonance frequency is now shifted from 2.43\text{GHz} at \( h = 0.0268\lambda_{res} \) to 2.32\text{GHz} at \( h = 0.06275\lambda_{res} \), with \( \lambda_{res} \) the resonance wavelength of the standard patch antenna. Therefore \( h = 0.06275\lambda_{res} \) corresponds to \( h = 3.74\text{mm} \) for the standard patch antenna.
Dielectric constant

The dielectric constant of the standard patch antenna is varied over: $3 \leq \varepsilon_r \leq 5$. The results of the simulations are shown in Figure 3.8 and Figure 3.9. The lines in Figure 3.9 are not smooth due to linear interpolation. Figure 3.8 shows that for lower values of $\varepsilon_r$ the effective radiated power increases most for higher frequencies. Figure 3.9 shows the resonance frequency rises for lower values of $\varepsilon_r$, as was to be expected since $f_{res} = \frac{c_0}{\sqrt{\varepsilon_r \lambda_{res}}}$. The shift in resonance frequency due to a shift in $\varepsilon_r$ is: $\frac{\partial f_{res}}{\partial \varepsilon_r} = -0.25 GHz / (\Delta \varepsilon_r = 1)$. The gradient constant $G$ for $\varepsilon_r$ equals:

$$
G_{\varepsilon_r} = \frac{\partial f_{res}}{\partial \varepsilon_r} \frac{\varepsilon_{r0}}{f_0} = \frac{\% \text{ shift in } f_{res}}{\% \text{ shift in } \varepsilon_r} = -0.446. \quad (3.5)
$$

where $f_0$ is the resonance frequency of the standard patch and $\varepsilon_{r0}$ is the relative permittivity of the standard patch. Note that,

$$
|G_{\varepsilon_r}| > |G_h|, \quad (3.6)
$$

which indicates the patch is more sensitive to variations in the relative dielectric constant than to variations in height in respect to the resonance frequency. A low value for $\varepsilon_r$ is advised, since a low value of $\varepsilon_r$ improves the effective radiated power. The value of the improvement in the effective radiated power at the shifting resonance frequency due to a change in $\varepsilon_r$ can be seen in Figure 3.8.
Figure 3.8: effective radiated power to frequency and $\varepsilon_r$

Figure 3.9: imaginary input impedance to frequency and $\varepsilon_r$
Figure 3.10: effective radiated power to frequency and tan δ

Loss tangent

The loss tangent is varied over the values: $0 \leq \tan \delta \leq 0.1$. The results of the simulation can be found in Figure 3.10 and Figure 3.11. Figure 3.10 shows that the effective radiated power is maximal when $\tan \delta$ is as low as possible. Figure 3.11 shows the imaginary input impedance with respect to frequency and $\tan \delta$. Due to the low values of the effective radiated power for values of $\tan \delta > 0.03$ (Figure 3.10) the patch no longer functions as an effective antenna for these values of the loss tangent. For the gradient $\frac{\partial f_{res}}{\partial \tan \delta}$ to be low it is advised to keep the value of $\tan \delta \leq 0.03$ as can be seen in Figure 3.11. For these values the gradient constant $G$ for $\tan \delta$ equals:

$$G_{\tan \delta} = \frac{\partial f_{res}}{\partial \tan \delta} \frac{\tan \delta_0}{f_0} = \frac{\% \text{ shift in } f_{res}}{\% \text{ shift in } \tan \delta} = 0.0035. \quad (3.7)$$

with $0 \leq \tan \delta \leq 0.03$, $\tan \delta_0$ the loss tangent of the standard patch and $f_0$ the resonance frequency of the standard patch. In practice the value of the loss tangent will never be zero but values well below $\tan \delta = 0.03$ are possible.

Width and length of the patch

The width $a$ and length $b$ of the patch are varied over a distance $-2.0\,mm \leq \partial x \leq 2.0\,mm$ and $-2.0\,mm \leq \partial y \leq 2.0\,mm$ as shown in Figure 3.12.

First the width $a$ is varied over a distance $\partial x$. The results of the simulation can be found in Figure 3.13 and Figure 3.14. Figure 3.13 shows the effective
Figure 3.11: imaginary input impedance to frequency and $\tan \delta$

Figure 3.12: varying the patch antenna’s width over $\partial x$ and length over $\partial y$
radiated power is at its maximum if the frequency is higher and width \( a \) is smaller. Figure 3.14 shows the resonance frequency is higher when width \( a \) is smaller. The lines in Figure 3.14 are not smooth due to linear interpolation. The new width \( a_{\text{new}} \) of the patch is:

\[
a_{\text{new}} = a_0 + 2\partial x,
\]

with \( a_0 \) the width of the standard patch and \( \partial x \) the variation of one side of the patch in the \( x \) direction. As shown in Figure 3.13 and Figure 3.14 the effective radiated power at the resonance frequency improves when \( a_{\text{new}} \) becomes smaller. This is due to the effect of the optimal height \((0.05\lambda_{\text{res}} \leq h \leq 0.0755\lambda_{\text{res}})\) in respect to the wavelength of the resonance frequency \( \lambda_{\text{res}} \) as shown in Figure 3.6. The dimension of the width \( a \) for the \( TM_{10} \) mode directly determines the wavelength of the resonance frequency as Equation 2.1 shows. For this Equation the width \( a = \frac{1}{2}\lambda_{\text{res}} \). Note that in Equation 2.1 the effect of the fringe fields, which cause an increase in the effective width, as discussed in Chapter 2.2 aren’t accounted for. Now we can write the effective height \( h_{\text{effective}} \) as:

\[
h_{\text{effective}} = \frac{h_0}{\lambda_{\text{res}}},
\]

where \( h_0 \) is the height of the standard patch and \( \lambda_{\text{res}} \) is the wavelength of the resonance frequency of the patch. By using Equation 3.9 we can derive that for \(-1.0 \text{mm} \leq \partial x \leq 1.0 \text{mm} \) the height in respect to the wavelength varies over \( 0.025\lambda_{\text{res}} \leq h \leq 0.029\lambda_{\text{res}} \). Which as Figure 3.6 shows is at the region for which the effective radiated power varies substantially. Thus varying the width \( a \) over a distance \( \partial x \) when the height is not in the stable region \((0.05\lambda_{\text{res}} \leq h \leq 0.0755\lambda_{\text{res}})\) causes a substantial shift, in this case from 0.41 to 0.47, in effective radiated power at the resonance frequency.

Because the width of the antenna directly determines the resonance frequency this parameter is sensitive to changes since \( \frac{\partial f_{\text{res}}}{\partial a} = -0.16 \text{GHz/mm} \). The value of the gradient constant \( G \) in respect to the width \( a \) is:

\[
G_a = \frac{\partial f_{\text{res}}}{\partial a} \frac{a_0}{f_0} = \frac{\% \text{ shift in } f_{\text{res}}}{\% \text{ shift in } a} = -1.909,
\]

with \( a_0 \) the width of the standard patch and \( f_0 \) the resonance frequency of the standard patch. The value of the gradient constant indicates that even small deviations in the width can lead to large deviations in the resonance frequency. The value of the gradient constant for the width is the highest found so far. Emphasizing the importance of this variable to be as precise as possible in production. In respect to the effective radiated power it is advised the width \( a \) corresponds to the optimal height \( h = 0.06275\lambda_{\text{res}} \) as is derived above.

Secondly the length \( b \) is varied over a distance \( \partial y \). The results of the simulations
Figure 3.13: effective radiated power to frequency and shift in width $\partial x$

Figure 3.14: imaginary input impedance to frequency and shift in width $\partial x$
Figure 3.15: Effective radiated power to frequency and shift in length $\partial y$

can be found in Figure 3.15 and Figure 3.16. Figure 3.15 shows that the effective radiated power is higher if the length $b$ is larger. This is because for the $TM_{10}$ mode the antenna radiates from the side walls in the $y$ direction (Figure 2.1). So an enlargement of the side walls in the $y$ direction is an enlargement in antenna aperture and thus an increase in effective radiated power.

Figure 3.16 shows the resonating frequency is higher if the length $b$ is smaller. The gradient of the resonating frequency to the length $b$ is $\frac{\partial f_{res}}{\partial b} = -0.003 GHz/mm$. The gradient constant $G$ in respect to the length $b$ equals:

$$G_b = \frac{\partial f_{res}}{\partial b} \frac{b_0}{f_0} = \frac{\% \text{ shift in } f_{res}}{\% \text{ shift in } b} = -0.036,$$

(3.11)

with $b_0$ the width of the standard patch and $f_0$ the resonating frequency of the patch. The gradient constant $G_b << G_a$ indicating variations in the length have far less influence than equal variations in the width on the resonating frequency.

**Probe position and diameter**

The probe position in the $x$ direction $x_p$ is varied over a distance $x_p + \partial x$ ($-2.0 mm \leq \partial x \leq 2.0 mm$) and in the $y$ direction $y_p$ over a distance $y_p + \partial y$ ($-2.0 mm \leq \partial y \leq 2.0 mm$).

First the position is changed in the $x$ direction. The results of the simulations can be found in Figure 3.17 and Figure 3.18. Figure 3.17 shows the effective radiated power improves for higher frequencies. This is because for those frequencies the
height to wavelength ratio improves towards the optimal height to wavelength ratio as discussed in the previous section. Figure 3.18 shows the resonance frequency is higher if the probe position is more to the center of the patch. The gradient of the resonance frequency to a shift of the probe position in the $x$ direction is $\frac{\partial f_{\text{res}}}{\partial x_p} = -0.0066 \text{GHz/mm}$. The gradient constant $G$ to the probe position $x_p$ equals:

$$G_{x_p} = \frac{\frac{\partial f_{\text{res}}}{\partial x_p}}{f_0} = \frac{\% \text{ shift in } f_{\text{res}}}{\% \text{ shift in } x_p} = -0.0198,$$

(3.12)

with $x_0$ the probe position of the standard patch and $f_0$ the resonance frequency of the standard patch. The effective radiated power stays near constant for the resonating frequency at a changing position $x_p$ in the $x$ direction.

Secondly the position in the $y$ direction $y_p$ is changed. The results of the simulations can be found in Figure 3.19 and Figure 3.20. Figure 3.19 shows that the effective radiated power increases for higher frequencies. This is because for those frequencies the height to wavelength ratio is improved towards the optimal height to wavelength ratio. Figure 3.20 shows that the resonance frequency is independent for the position of the probe in the $y$ direction for the region $6.0\text{mm} \leq y_p \leq 10.0\text{mm}$ which can also be written as $0.1\lambda_{\text{res}} \leq y_p \leq 0.13\lambda_{\text{res}}$. Therefore the gradient constant $G$ in respect to position $y_p$ equals:

$$G_{y_p} = \frac{\frac{\partial f_{\text{res}}}{\partial y_p}}{f_0} = \frac{\% \text{ shift in } f_{\text{res}}}{\% \text{ shift in } y_p} = 0,$$

(3.13)
Figure 3.17: effective radiated power to frequency and shift in probe position $\partial x$

Figure 3.18: imaginary input impedance to frequency and shift in probe position $\partial x$

33
in the region \(0.1\lambda_{res} \leq y_p \leq 0.13\lambda_{res}\), with \(y_0\) the probe position of the standard patch and \(f_0\) the resonance frequency of the standard patch. The effected radiated power is nearly constant for the resonating frequency at the region \(0.1\lambda_{res} \leq y_p \leq 0.13\lambda_{res}\).

In the third simulation the diameter of the probe \(d\) is varied over a distance \(0.1mm \leq d \leq 2.0mm\). The results of the simulations can be found in Figure 3.21 and Figure 3.22. Figure 3.21 shows that the effective radiated power has an optimum when the diameter is between \(0.3mm \leq d \leq 1.0mm\). This can also be written in terms of wavelengths, \(0.005\lambda_{res} \leq d \leq 0.016\lambda_{res}\), with \(\lambda_{res}\) the wavelength of the resonance frequency. Figure 3.22 shows the resonating frequency is higher for smaller diameters. The gradient of the resonating frequency in respect to the diameter of the probe is \(\frac{\partial f_{res}}{\partial d} = -0.013GHz/mm\). The gradient constant \(G\) in respect to the diameter \(d\) equals:

\[
G_d = \frac{\partial f_{res}}{\partial d} \cdot \frac{d_0}{f_0} = \frac{\% \text{ shift in } f_{res}}{\% \text{ shift in } d} = 0.0042,
\]  

(3.14)

with \(d_0\) the diameter of the standard patch and \(f_{res}\) the resonance frequency of the standard patch. The gradient constant shows the diameter has a small influence on the resonance frequency in respect to the other parameters. Combining this with the effect on effective radiated power the advised diameter \(d_{advised}\) equals:

\[
d_{advised} = 0.01\lambda_{res}.
\]  

(3.15)
Figure 3.20: imaginary input impedance to frequency and shift in probe position $\partial y$

Since this diameter is in the middle of the stable area $0.005\lambda_{res} \leq d \leq 0.016\lambda_{res}$ for the effective radiated power and has the smallest gradient.

### 3.3.3 Conclusions

The simulations in FEKO gave good results which can be used for optimizing a square patch antenna with a coaxial probe feed. After simulations the gradient constant $G$ was introduced to compare the sensitivity of the resonance frequency in respect to the chosen parameters. The gradient constant $G$ gives the shift in percentages of the resonance frequency due to a percentage shift in the parameter. The results of the simulations can be seen in table 3.1.
Figure 3.21: Effective radiated power to frequency and diameter

Figure 3.22: Imaginary input impedance to frequency and diameter
### Table 3.1: found parameters with optimization and gradient constant

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Optimization</th>
<th>Gradient constant</th>
</tr>
</thead>
<tbody>
<tr>
<td>height $h$</td>
<td>$h_{optimal} = 0.06275 \lambda_{res}$</td>
<td>$G_{h} = -0.036$</td>
</tr>
<tr>
<td>dielectric constant $\varepsilon_r$</td>
<td>should be as small as possible</td>
<td>$G_{\varepsilon_r} = -0.446$</td>
</tr>
<tr>
<td>loss tangent $\tan \delta$</td>
<td>$\tan \delta$ should be as small as possible</td>
<td>$G_{\tan \delta} = 0.0035$</td>
</tr>
<tr>
<td>width $a$</td>
<td>$a$ should be as precise as possible</td>
<td>$G_{a} = -1.909$</td>
</tr>
<tr>
<td>length $b$</td>
<td>$b$ should be as large as possible</td>
<td>$G_{b} = -0.036$</td>
</tr>
<tr>
<td>probe position $x_p$</td>
<td>can be shifted between 0 and $\frac{1}{2}a$</td>
<td>$G_{x_p} = -0.0198$</td>
</tr>
<tr>
<td>probe position $y_p$</td>
<td>advised at $\frac{1}{2}b$</td>
<td>$G_{y_p} = 0$</td>
</tr>
<tr>
<td>probe diameter $d$</td>
<td>$d_{optimal} = 0.01\lambda_{res}$</td>
<td>$G_{d} = 0.0042$</td>
</tr>
</tbody>
</table>

#### 3.4 Simulating a patch antenna at 60GHz

##### 3.4.1 Introduction

After simulations with MATLAB and FEKO optimal parameters for a square patch antenna with a coaxial probe feed have been derived. In this section the obtained parameters will be used to simulate an optimal square patch antenna at 60GHz.

##### 3.4.2 Simulation results

To simulate the optimal patch antenna the software tool FEKO, which uses the method of moments, is used. For the simulations a patch antenna with a coaxial feed is used. The dimensions of the antenna are: width $a = 1.107\, \text{mm}$, length $b = 1.107\, \text{mm}$ and height $h = 0.15\, \text{mm}$. The probe position is $x_p = 0.305\, \text{mm}$, $y_p = 0.5535\, \text{mm}$ and the probe diameter $d = 0.024\, \text{mm}$. The conductance of the copper ground-plane and patch is $\sigma = 5.8 \cdot 10^{-7} \Omega^{-1} \text{m}^{-1}$. The dielectric constant of the dielectric equals $\varepsilon_r = 4.28$ and the loss tangent is given by $\tan \delta = 1.6 \cdot 10^{-2}$.

The results of the simulations can be found in Figure 3.23 and Figure 3.24. Figure 3.23 shows the patch has a resonating frequency at $f_{res} = 60\, \text{GHz}$. Figure 3.24 shows the effective radiated power is 0.565, which is well above the value of the standard patch used in the previous section, which was around 0.44. So even on 60GHz the rules of thumb which were derived in the previous section with respect to optimal height ($h = 0.06275\lambda_{res}$) and optimal probe diameter ($d = 0.01\lambda_{res}$) hold.

The effects of a small dielectric constant $\varepsilon_r < 3$ need more study, since it is not yet clear if the rules for optimal height an optimal diameter hold for values of $\varepsilon_r < 3$. Therefore simulations are done on the 60GHz patch antenna where $\varepsilon_r$ is
Figure 3.23: input impedance to frequency

Figure 3.24: effective radiated power to frequency
Figure 3.25: imaginary input impedance to frequency and $\varepsilon_r$. Three dimensional (a) and contour line (b)

varied over $1.1 \leq \varepsilon_r \leq 5.0$. The resonance wavelength is, $\lambda_{res} = \frac{c_0}{\sqrt{\varepsilon_r f_{res}}}$, with $c_0$ the speed of light. The dimensions of the patch are a function of $\lambda_{res}$: the width $a = 0.458\lambda_{res}$, the length $b = 0.458\lambda_{res}$, the height $h = 0.06275\lambda_{res}$, the probe position $x_p = \frac{a}{2\lambda_{res}}$, $y_p = \frac{b}{2}$ and the probe diameter $d = 0.01\lambda_{res}$. Therefore for every change in $\varepsilon_r$ the dimensions of the patch are adjusted. Thus making it possible to verify if the rules of thumb for optimal height and diameter hold for low values of $\varepsilon_r$.

The results of the simulations can be found in Figure 3.25, Figure 3.26 and Figure 3.27. As seen in Figure 3.25 and Figure 3.26 the resonance frequency is lower for lower values of $\varepsilon_r$. This is because the fringe fields (Chapter 2.2) become larger for lower values of the relative dielectric constant $\varepsilon_r$ and thus the effective width becomes larger. Which in turn leads to a lower resonance frequency. Therefore the width $a$ of the antenna needs to become smaller to make the effective width equal to $\frac{1}{2}\lambda_{res}$.

The effective radiated power improves when the dielectric constant $\varepsilon_r$ has a lower value. Figure 3.27 shows the effective radiated power is relatively flat over the considered spectrum area ($55GHz \leq f \leq 65GHz$). This indicates the rules of thumb which were derived in the previous section with respect to optimal height ($h = 0.06275\lambda_{res}$) and optimal probe diameter ($d = 0.01\lambda_{res}$) hold for low values of $\varepsilon_r$. But as shown in Figure 3.25 and Figure 3.26 the imaginary part of the input impedance does not equal zero at the maximum of the real part of the input impedance for values of the dielectric constant $\varepsilon_r < 3$. This needs consideration when the coaxial probe feed impedance is matched to the input impedance of the patch antenna.
Figure 3.26: real input impedance to frequency and $\varepsilon_r$

Figure 3.27: effective radiated power to frequency and $\varepsilon_r$
3.4.3 Conclusions

The optimization of the height and the diameter still apply at 60GHz. Optimizing the dielectric constant to $\varepsilon_r \approx 1$ needs further investigation, due to effects on the input impedance of the patch.

3.5 Conclusions

The cavity model can be used within a stringent parameter range to calculate the input impedance. Due to the nature of the model its range of accuracy is limited and it is unsuitable for studying the effects of different parameters on antenna performance.

The effects of the parameters on antenna performance that have been derived using FEKO can be found in Table 3.1. The optimization of the height and the probe diameter, that result in a rise in effective radiated power and makes the patch antenna more robust for variations caused in production, still apply at 60GHz. Optimizing the dielectric constant to $\varepsilon_r \approx 1$ needs further investigation, due to effects on the input impedance of the patch.
Chapter 4

Conclusions

According to the simulations the standard square patch antenna can be made more robust and efficient by implementing the recommendations made in this report (Table 3.1). An optimal height and probe diameter are found which enhance efficiency and robustness. Also it is very important to make the width of the antenna as precise as possible since this is the most sensitive parameter of the patch antenna.

Another interesting conclusion is that the cavity model can only be used within a stringent parameter range. Therefore it is unsuitable for calculating antenna gain and cannot be used to calculate the effect of different parameters on antenna performance.

Because basing conclusions on only one model, in this case FEKO, is risky it is advised to verify the recommendations of this report with other models and measurements.
Bibliography


