3 Analysis Methods and CAD Programs

The design of even a simple rectangular microstrip antenna can be challenging. For more complex antennas, the design process becomes very complicated and time consuming. Only a few design equations exist for microstrip antennas, and they are limited in their accuracy and generally apply only to simple, single layered rectangular and circular patch geometries. Moreover, once fabricated, the antennas must be tuned by the cut and try method. There are complicated models and full wave solutions to some of the complex patch geometries, but they are often very difficult to apply in a practical design. A more realistic approach for designing complicated geometries is to begin using the simple equations to estimate the patch dimensions and then use CAD programs to optimize the design. The antenna is then fabricated and tuned by cut and try methods.

3.1 Microstrip Antenna Analysis Techniques

There are several common methods to analyze microstrip antennas. The simplest is to treat the microstrip antenna as a transmission line. This technique is generally limited to basic geometries and has only limited accuracy. Another technique is to treat the patch as a resonant cavity. This model is limited to geometries having substrate thicknesses much less than a wavelength. The final method is to apply fullwave techniques to the analysis of the antenna. While this method is usually very accurate and can be applied to complex geometries, it is much more difficult to use than the simpler models.

3.1.1 Transmission Line Model

The transmission line model [48-53] is the simplest method used to analyze microstrip antennas. Antenna design parameters can be estimated quickly. This model is generally applicable to simple antennas; however, recently it has been modified for more complicated geometries. The model represents a microstrip antenna as a very wide microstrip transmission line of limited length. The two open ends at the radiating sides of the patch, between the patch and the ground plane, can be thought of radiating apertures or slots. These slots are treated as two high impedance loads separated by a low impedance transmission line, as shown in figure 3.1. Using this model, relatively simple equations have been derived for the patch dimensions, input impedance, quality factor, and radiation fields for rectangular patches.

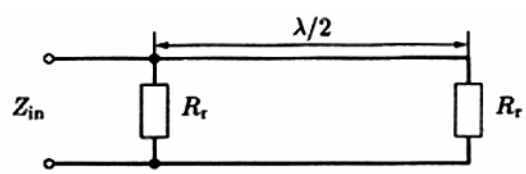


Figure 3.1. Equivalent transmission line model [54]

3.1.2 Cavity Model

The cavity model [55-59] provides an intuitive understanding to the patch operation and usually provides more accurate solutions than the transmission line model. The main assumption for this model is that the substrate thickness is much less than the wavelength ($h < 0.05\lambda$). Therefore, the electric fields are assumed to exist only perpendicular to the ground plane and the patch element, resulting in only TM modes. Under these conditions, the patch is modeled as a resonant cavity. The patch and ground plane are represented as electric walls, while the sides of the cavity represented as magnetic walls, as shown in figure 3.2.

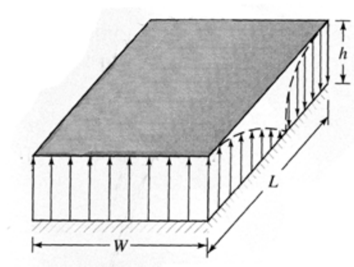


Figure 3.2. Resonant cavity having magnetic walls [60].

3.1.3 Fullwave Solutions

Fullwave solutions [61-85] use Maxwell's equations and enforce the boundary conditions to account for all relevant wave mechanisms in the microstrip antenna. Their accuracy depends on the accuracy of the numerical methods used to derive the solution. Their formulation and final field solutions are too lengthy to include here. References [86-88] cover these in sufficient detail. By far, the most common numerical analysis technique used in the analysis of microstrip antennas is the Method of Moments (MoM). However, during the past decade, the Finite Difference Time Domain (FDTD) analysis has been successfully applied to the analysis of microstrip antennas and appears more frequently in the literature. These methods are complicated to develop and require intense computational resources. Fullwave solutions lack generality, requiring that the integral equations be derived and numerical methods reapplied for each different antenna geometry investigated.

The general fullwave procedure is to derive electric field equations for both the feed and the patch. Often, these are very complicated electric field integral equations.

They are constructed using Green's functions, having the basic form,

$$\vec{E}_{\tan} = \vec{E}_{inc} + \int g\left(\vec{r}, \vec{r}\,'\right)\vec{J}\left(\vec{r}\right) d\vec{S} = 0, \tag{3.1}$$

where g is the Green's function, $\vec{J}$ is the current on the metallic surfaces, $\vec{E}_{\tan}$ is the tangential field on the metallic surface, and $\vec{E}_{inc}$ is the impressed incident field (generally from the source signal). The only unknown in the equation is the surface current on the patch. Often, these equations can only be solved numerically. The most common numerical method applied to microstrip antennas is the method of moments. Applying numerical methods are complicated and require the use of high-speed computers having significant resources of RAM. Once the surface currents are determined, then the far field radiation pattern and other antenna parameters can be determined.

While many journal articles present fullwave solutions to a variety of patch geometries and parameters, applying these results requires an understanding of advanced electromagnetics and applied mathematics. Moreover, the theory presented is usually very specific and cannot easily be applied to other general cases. In most cases it is more practical to design an antenna using the simplified models and existing CAD programs.

3.2 Patch Dimensions, Input Impedance, and Quality Factor

3.2.1 Dimensions for Rectangular Patch

The width (dimension b in figure 2.2a) of the rectangular patch is arbitrary. Narrow widths become less efficient and wide widths can create high order modes. As the width is increased, the input impedance decreases. A recommended practical width, having good efficiency, is given by [89]

$$W = \frac{1}{2f_r\sqrt{\mu_0\varepsilon_0}}\sqrt{\frac{2}{\varepsilon_r + 1}}, \tag{3.2}$$

where f_r is the resonant frequency of the patch and ε_r is the relative permittivity of the substrate. For a circularly polarized patch, the width is designed using the same design technique used in determining the resonating length (dimension *a* in figure 2.2a).

Because of the fringing fields, the effective electrical length of the patch is slightly longer than the physical length. To account for this fringing effect, an effective dielectric constant is used for frequencies less than 10 GHz.

$$\varepsilon_{eff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left[1 + 12\frac{h}{W}\right]^{-1/2}. \tag{3.3}$$

where h is the thickness of the substrate. Taking into account of the fringing fields, the resonant length of the patch is determined by

$$L = \frac{\lambda_g}{2} - 2L_{fringing}, \tag{3.4}$$

where

$$\lambda_g = \frac{1}{f_r\sqrt{\mu_0\varepsilon_0\varepsilon_{eff}}}, \tag{3.5}$$

and

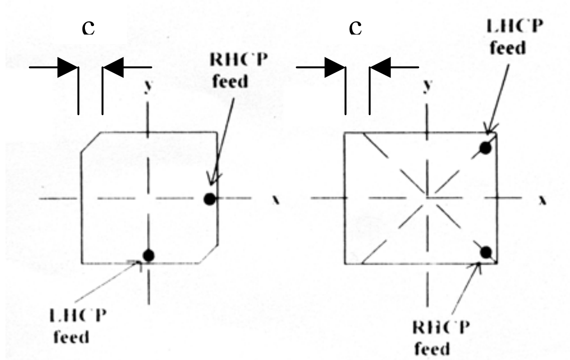
$$L_{fringing} = \left[0.412t\frac{\left(\varepsilon_{eff} + 0.3\right)\left(\frac{W}{h} + 0.264\right)}{\left(\varepsilon_{eff} + 0.258\right)\left(\frac{W}{h} + 0.8\right)}\right]. \tag{3.6}$$

The resonant frequency of a given patch is determined by

$$f_r = \frac{c}{2\sqrt{\varepsilon_{eff}}\,(L + 2L_{fringing})}. \tag{3.7}$$

3.2.2 Dimensions for Single Circularly Polarized Patch

The two most common single fed circularly polarized patches are shown in figure 3.3. For the truncated corner patch shown in figure 3.3a, the required perturbation is produced by trimming of the opposite corners on one diagonal of a square patch. The patch is fed on either side, resulting in either right hand or left hand circular polarization. For the nearly square patch shown figure 3.3b, the perturbation is produced by making one side of the patch longer than the other. The dimensions are derived using the variational method [90, 91].



a. truncated **b.** nearly square

Figure 3.3. Truncated and nearly square patch elements [92].

The size of the perturbation in both cases is related to the size and quality factor of the patch. For the truncated corner patch,

$$\frac{\Delta s}{S} = \frac{1}{2Q} \tag{3.8}$$

and for the nearly square patch,

$$\frac{\Delta s}{S} = \frac{1}{Q}, \tag{3.9}$$

where Δs is the area of the perturbation segment and S is the area of the patch. From equations 3.8 and 3.9, the perturbation length 'c' for the truncated patch is given by

$$c = \sqrt{\Delta s} \tag{3.10}$$

and for the nearly square patch it is given by

$$c = \frac{\Delta s}{L}. \tag{3.11}$$

The resonant frequencies of the new modes are functions of the perturbation segments. For the truncated patch,

$$f_1 = f_0\left[1 - \frac{2\Delta s}{S}\right] \tag{3.12}$$

and

$$f_2 = f_0. \tag{3.13}$$

For the nearly square patch,

$$f_1 = f_0\left[1 - \frac{\Delta s}{S}\right] \tag{3.14}$$

and

$$f_2 = f_0. \tag{3.15}$$

Because f_o is the resonant frequency of the patch without perturbation, the length of the single fed square patch should be slightly smaller than for the linearly polarized case.

3.2.3 Input Impedance for Rectangular Patch

The input impedance of a rectangular patch can be determined using the transmission line model where each radiating edge is represented by an equivalent parallel admittance $Y_1 = G_1 + jB_1$ and $Y_2 = G_2 - jB_2$ at resonance. For a rectangular patch at resonance, $G_1 = G_2$ and $B_1 = B_2$. Neglecting the mutual effects of each slot, the resonant input resistance is given by

$$R_{in} \approx \frac{1}{G_1 + G_2} = \frac{1}{2G_1}, \tag{3.16}$$

where

$$G_1 = G_2 = \begin{cases} \dfrac{1}{90}\left(\dfrac{W}{\lambda_0}\right)^2 & \text{for } W \le 0.35\lambda_0 \\ \dfrac{1}{90}\left(\dfrac{W}{\lambda_0}\right)^2 & \text{for } 0.35\lambda_0 \le W \le 2\lambda_0. \\ \dfrac{1}{120}\left(\dfrac{W}{\lambda_0}\right)^2 & \text{for } 2\lambda_0 < W \end{cases} \tag{3.17}$$

3.2.4 Input Impedance for Wraparound Patch

The total input resistance, for the entire band, is calculated using the transmission line model is

$$R_{in} = \frac{1}{2W\,G_s}, \tag{3.18}$$

where the conductance G_s is

$$G_s = \frac{\pi}{\lambda\eta_0}\left(1 - \frac{(k_0 h)^2}{24}\right). \tag{3.19}$$

The input impedance of the wraparound patch at each feed point is

$$R_F = N_F \cdot R_{in}, \tag{3.20}$$

where N_F is the number of feeds used in the wraparound patch design.

3.2.5 Quality Factor for Rectangular Element

The quality factor (Q) is a figure of merit that is a measure of antenna losses. For single fed circularly polarized elements, the perturbation sections are proportional to this factor. A microstrip antenna, often modeled as a parallel resonant circuit or a resonant cavity, has a large quality factor. The losses are results of radiation, resistance of the copper, the dielectric losses in the substrate, and surface wave losses. For thin substrates, surface wave losses are assumed negligible, and the total quality factor is written as

$$\frac{1}{Q_t} = \frac{1}{Q_c} + \frac{1}{Q_d} + \frac{1}{Q_{rad}}, \tag{3.21}$$

where the quality factor due to the dielectric, the conductor, and radiation are

$$Q_d = \frac{1}{\tan\delta}, \tag{3.22}$$

$$Q_c = \frac{1}{2}\frac{\eta\mu_r k_0 h}{R_s}, \tag{3.23}$$

and

$$Q_{rad} = \frac{2ck_{mn}W_E}{\sqrt{\varepsilon_r}P_{rad}}, \tag{3.24}$$

where

$$P_{rad} = \frac{1}{\eta_0}\int_0^{2\pi}\int_0^{\pi/2}\left(\left|E_\theta\right|^2 + \left|E_\phi\right|^2\right) r^2 \sin\theta d\theta d\phi \,, \tag{3.25}$$

and

$$W_E = \frac{1}{2}\varepsilon_r h \,. \tag{3.26}$$

Using the transmission line model, the quality factor can be approximated easier by [93],

$$Q \cong \frac{\pi}{4Z_m\left[G_1 + G_{12}\right]}\,, \tag{3.27}$$

where Z_m is the characteristic impedance found by

$$Z_m = \frac{60}{\sqrt{\varepsilon_{eff}}}\left[\frac{W}{2h} + 0.441 + \frac{1}{\pi}1.451 + \ln\left(\frac{W}{2h}\right) + 0.94\right]^{-1} \tag{3.28}$$

and the mutual conductance is

$$G_{12} = \frac{1}{120\pi^2}\int_0^{\pi}\left[\frac{\sin\left(\frac{k_0 W}{2}\cos\theta\right)}{\cos\theta}\right] J_0(k_0 L\sin\theta)\sin^3\theta \, d\theta \,. \tag{3.29}$$

3.3 Microstrip Patch Radiation

3.3.1 Electric Field Equations

The far field radiation pattern of an antenna is described by the antenna's electric field in spherical coordinates, as shown in figure 3.4. Using the cavity model, the far-field radiating fields for a planar rectangular patch, determined from equivalent magnetic current densities on the cavities' magnetic walls, are [94],

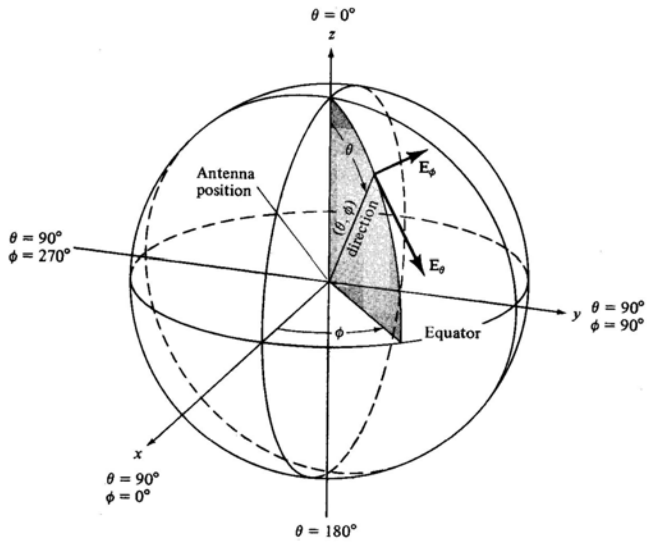


Figure 3.4. Radiated field coordinate system geometry [95].

$$E_\theta = j\frac{k_0 W V_0 e^{-jk_0 r}}{\pi r}\left\{\frac{\sin\left(\frac{k_0 h}{2}\cos\phi\right)}{\frac{k_0 h}{2}\cos\phi}\right\}\cos\left(\frac{k_0 L_e}{2}\sin\phi\right) \tag{3.30}$$

and

$$E\phi = j\frac{k_0 W V_0 e^{-jk_0 r}}{\pi r}\left\{\sin\theta\frac{\sin\left(\frac{k_0 h}{2}\sin\theta\right)}{\frac{k_0 h}{2}\sin\theta}\frac{\sin\left(\frac{k_0 W}{2}\cos\theta\right)}{\frac{k_0 W}{2}\cos\theta}\right\}, \tag{3.31}$$

where V_o is the voltage across each radiating slot, E_θ is limited to $0 \le \theta < \pi/2$ because of the ground plane, and *h* is the substrate thickness. The electric field equations get much more complicated for the cylindrical case [96-101].

3.3.2 Losses Due to Polarization Mismatch

The polarization of a radiated field is defined as the orientation of the electric field vector. The general case is when the field is elliptically polarized, as shown in figure 3.5. If the polarizations of a transmitting and receiving antenna are equal, there will be no losses due to polarization mismatch. However, for linear polarized antennas, the loss is infinite when the transmitting and receiving antennas are orientated orthogonally. For circular polarized antennas, the loss is infinite when their electric fields are rotating in opposite directions, or different by 180 degrees. Therefore, to minimize or eliminate any losses that result from polarization mismatch, it is important to match the polarizations of the transmitting and receiving antennas. In special cases, it may be impossible to guarantee the orientation of an antenna. This is the case for an antenna on a sounding rocket. A sounding rocket's orientation and position is not static, but continuously changing. This movement results in changes in the polarization of the signal received by the ground station. Therefore, to ensure signal reception, it is necessary that the ground station's antenna be circularly polarized.

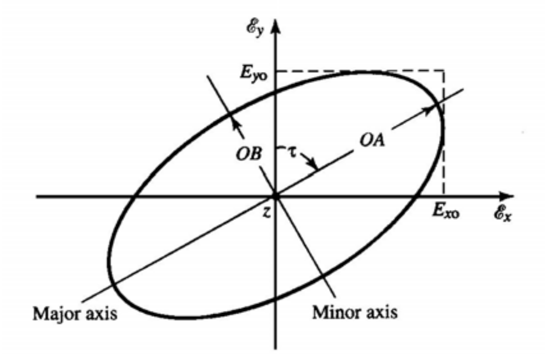


Figure 3.5. Polarization ellipse at z = 0 [102].

With the assumption that the ground station's antenna is circularly polarized, this section will investigate the losses due to polarization mismatch between the transmit and receive antennas. The two extreme cases are when the polarization of the sounding rocket's transmit antenna is either linearly or circularly polarized.

The first case is when the sounding rocket uses a linearly polarized antenna and the ground station uses a circularly polarized antenna. Since the rocket moves over time, the orientation of linearly polarized antenna with respect to the ground station will vary between vertical and horizontal. Because the receiving antenna is circularly polarized, the resulting mismatch losses are constant over time, irrespective of the orientation of the linearly polarized transmit antenna. The magnitude of this constant mismatch loss is 3 dB, assuming zero tilt angle. In contrast, if a linearly polarized antenna replaced the ground station's circularly polarized antenna, the losses would not be constant, but would vary between 0 dB and infinity over time. Therefore, when the rocket has a linearly polarized antenna, to guarantee that the ground station receives the signal, the 3 dB mismatch loss resulting from the linearly to circularly polarized antennas is acceptable.

The second case is when the sounding rocket uses a circularly polarized microstrip antenna and the ground station uses a circularly polarized antenna, both having the same polarization sense. As the rocket is in motion, the orientation of the rocket will change with respect to the ground station. Because the rocket's circularly polarized antenna is only circularly polarized broadside to the patch, and only over a narrow frequency range, there will still be polarization losses between the rocket and the ground station. Off from broadside, the polarization becomes elliptical, ultimately becoming linear along the longitudinal axis of the rocket. The mismatch in this second case results in a 0 to 3 dB loss in signal strength over time received at the ground station.

Generally, a circularly polarized antenna is only circularly polarized for limited bandwidth and beamwidth. For the case of a microstrip antenna on a sounding rocket, it is only circularly polarized broadside to the cylinder. The degree of circular polarization is described by the axial ratio, which is determined by,

$$AR = \left| 20\log\left(\frac{\text{Major Axis Electric Field}}{\text{Minor Axis Electric Field}}\right)\right|. \tag{3.32}$$

When an antenna is circularly polarized, the axial ratio magnitude is 0 dB. As the antenna polarization becomes elliptical, the axial ratio increases in magnitude. If the axial ratio continues to increase in magnitude, the antenna's polarization will eventually become linear.

In the first case of a linearly polarized antenna on the rocket ($AR_{LP} = \infty$) and a circularly polarized antenna at the ground station ($AR_{CP} = 0$ dB), the mismatch loss is a constant 3 dB. In the second case of a circularly polarized microstrip antenna on the rocket (0 dB > AR_{EP} > ∞) and a circularly polarized antenna on the ground ($AR_{CP} = 0$ dB), the mismatch loss varies from 0 to 3 dB.

To quantify the loss resulting from polarization mismatch as a circularly polarized antenna becomes elliptical, the polarization loss factor can be calculated using the following equation,

$$\begin{aligned} e_{pol} &= \left|\hat{e}_{rocket} \bullet \hat{e}_{ground}\right|^2 \\ &= \left| \frac{\hat{x} + \dfrac{j}{AR_{linear}}\hat{y}}{\sqrt{1 + \dfrac{1}{AR^2{}_{linear}}}} \bullet \frac{\hat{x} + j\hat{y}}{\sqrt{2}} \right|^2 = \frac{\left(1 + \dfrac{1}{AR_{linear}}\right)^2}{2\left(1 + \dfrac{1}{AR^2{}_{linear}}\right)}. \end{aligned} \tag{3.33}$$

The unit vectors $\hat{e}_{rocket}$ and $\hat{e}_{ground}$ represent the sounding rocket and ground station electric fields having the general form,

$$\vec{E} = (E_{0x}\hat{x} + E_{0y}\hat{y})e^{-jk_0 z}. \tag{3.34}$$

A plot of the polarization loss factor is shown in figure 3.6. The plot relates the axial ratio to the mismatch loss assuming one of the two antennas is perfectly circularly polarized. In our case, the ground station's antenna is assumed ideally circularly polarized. If the sounding rocket has an antenna that is also perfectly circularly polarized, then its axial ratio will be 0 dB. This corresponds to a polarization mismatch loss of 0 dB. If the sounding rocket has an elliptically polarized antenna, having an axial ratio of 10 dB, then the polarization loss will be approximately 1 dB. Finally, as seen in the plot, an axial ratio of greater than 40 dB results in a loss of approximately 3 dB. Therefore, a rough guideline is that an axial ratio of 30 dB equates to the performance of a linearly polarized antenna.

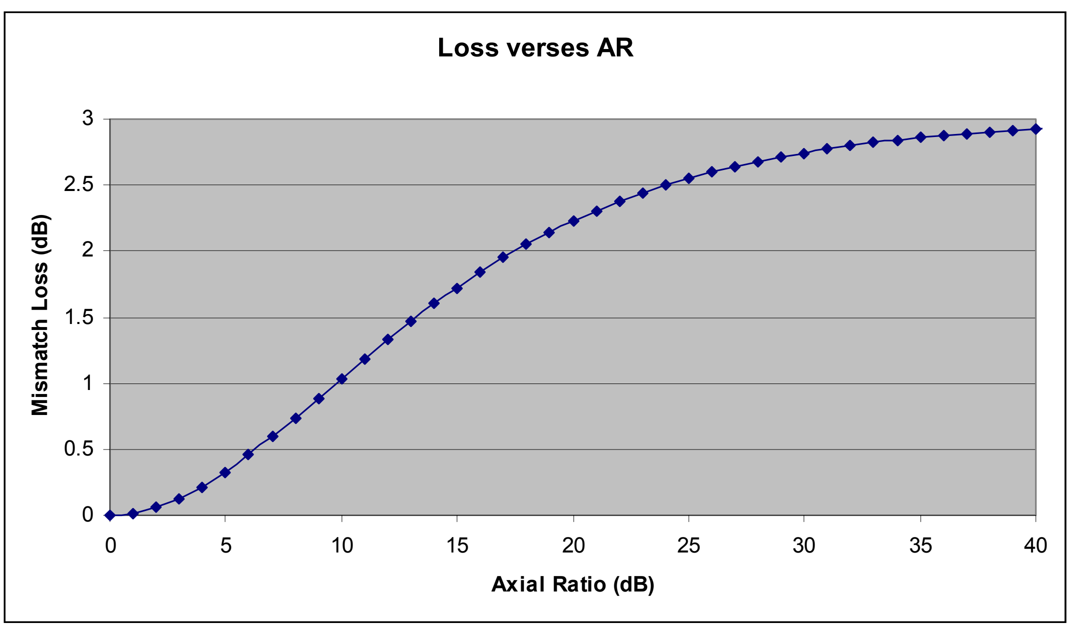


Figure 3.6. Loss versus AR for sounding rocket antenna.

3.4 Hertzian Electric Dipole (HED) Formulas

The following analysis formulas for the rectangular microstrip antenna are derived from analytical approximations of exact formulas, not empirical formulas [103]. They become more accurate as the substrate thickness decreases.

3.4.1 Radiation Quality Factor

Assuming no conductor loss and no dielectric loss, the radiation efficiency that accounts only for the power loss due to surface waves is

$$e_r^0 = \frac{Q_r}{Q_{sp}}, \tag{3.35}$$

where the radiation quality factor is defined as

$$\frac{1}{Q_r} = \frac{1}{Q_{sp}} + \frac{1}{Q_{sw}}. \tag{3.36}$$

The space and surface wave quality factors can be approximated using the radiation efficiency of a horizontal electric dipole, e_r^{hed}, on the top of a lossless substrate. The analysis formula for the space wave quality factor is

$$Q_{sp} = \frac{3}{16}\left(\frac{\varepsilon_r}{p \cdot c_1}\right)\left(\frac{L_e}{W_e}\right)\left(\frac{\lambda_0}{h}\right), \tag{3.37}$$

where

$$L_e = L + 2L_{fringing}, \tag{3.38}$$

$$W_e = W + 2\frac{h\ln(4)}{\pi}, \tag{3.39}$$

$$c_1 = \frac{1}{n_1^2} + \frac{2}{5n_1^4}, \tag{3.40}$$

and

$$n_1 = \sqrt{\varepsilon_r \mu_r} \ . \tag{3.41}$$

The array factor p, which relates the efficiency of the HED to the efficiency of a rectangular patch, is defined as

$$p = 1 + \frac{a_2(k_0W_e)^2}{10} + \frac{3(a_2^2 + 2a_4)(k_0W_e)^4}{560} + \frac{c_2(k_0L_e)^2}{5} + \frac{a_2c_2(k_0W_e)^2(k_0L_e)^2}{70} \tag{3.42}$$

where

$$k_0 = \frac{2\pi}{\lambda_0}, \tag{3.43}$$

and

$$a_2 = -0.16605,$$
$$a_4 = 0.00761,$$
$$c_2 = -0.0914153.$$

3.4.2 Radiation Efficiency

The radiation efficiency of the horizontal electric dipole is,

$$e_r^{hed} = \frac{P_{sp}^{hed}}{P_{sp}^{hed} + P_{sw}^{hed}}, \tag{3.44}$$

where the power of the HED space wave is

$$P_{sp}^{hed} = \frac{1}{\lambda_0^2}(k_0h)^2(80\pi^2\mu_r^2c_1), \tag{3.45}$$

and the power of the HED surface wave is

$$P_{sw}^{hed} = \frac{\eta_0 k_0^2}{8} \frac{\varepsilon_r (x_0^2 - 1)^{3/2}}{\varepsilon_r(1 + x_1) + (k_0h)(1 + \varepsilon_r^2 x_1)\sqrt{x_0^2 - 1}}, \tag{3.46}$$

with

$$x_1 = \frac{x_0^2 - 1}{\varepsilon_r - x_0^2}\,, \tag{3.47}$$

$$x_0 = 1 + \frac{-\varepsilon_r^2 + \alpha_0\alpha_1 + \varepsilon_r\sqrt{\varepsilon_r - 2\alpha_0\alpha_1 + \alpha_0^2}}{(\varepsilon_r^2 - \alpha_1^2)}\,, \tag{3.48}$$

$$\alpha_0 = s\tan(sk_0h)\,, \tag{3.49}$$

$$\alpha_1 = -\frac{1}{s}\left[\tan(sk_0h) + \frac{sk_0h}{\cos^2(sk_0h)}\right], \tag{3.50}$$

and

$$s = \sqrt{\varepsilon_r - 1}\,. \tag{3.51}$$

The rectangular patch radiation efficiency is

$$e_r = \frac{e_r^{hed}}{1 + e_r^{hed}\left(\tan\delta + \frac{R_s}{\pi\eta_0\mu_0}\cdot\frac{\lambda_0}{h}\right)\left(\frac{3}{16}\cdot\frac{\varepsilon_r}{p\cdot c_1}\cdot\frac{L_e}{W_e}\cdot\frac{\lambda_0}{h}\right)} \tag{3.52}$$

where the surface resistivity is

$$R_s = \sqrt{\frac{\omega\mu_0}{2\sigma}}\,. \tag{3.53}$$

3.4.3 Input Resistance, Bandwidth, and Directivity

Using methods based on the cavity model, the input resistance of a rectangular patch is

$$R_{in} = \frac{4\mu_r\eta_0}{\pi\cdot\tan\delta_{eff}}\cdot\frac{L_e}{W_e}\cdot\frac{h}{\lambda_0}\cos^2\left(\frac{\pi x_0}{L_e}\right), \tag{3.54}$$

where

$$\tan\delta_{eff} = \frac{1}{Q_{total}}. \tag{3.55}$$

The bandwidth formula of the patch is

$$BW = \frac{1}{\sqrt{2}}\left(\tan\delta_{eff} + \frac{R_s}{\pi\eta_0}\cdot\frac{\lambda_0}{h} + \frac{16}{3}\cdot\frac{p\cdot c_1}{\varepsilon_r}\cdot\frac{h}{\lambda_0}\cdot\frac{W_e}{L_e}\cdot\frac{1}{e_r^{hed}}\right). \tag{3.56}$$

The gain and directivity of the patch is

$$D = \frac{\eta}{40\pi}\cdot\frac{1}{p\cdot c_1} \tag{3.57}$$

and

$$G = e_r D. \tag{3.58}$$

3.5 Description of Relevant CAD Programs

There are several CAD programs available for microstrip antennas. They range from simple and inexpensive DOS programs to complicated and expensive fullwave analysis programs. The ones that will be used in the subsequent chapters are described in this section. The DOS based CAD programs are available with Sainati's microstrip antenna text [104]. An accurate, fullwave CAD program from Ansoft is Clementine, which analyzes microstrip patches on a cylinder. While these patches can be of any dimension, they are restricted to only a single substrate layer. In other words, Clementine is unable to analyze patches with aperture coupled feeds or superstrate overlays.

3.5.1 DOS Based CAD Programs

Several DOS programs are available with Saintia's text, which quickly determine basic parameters and radiation patterns for microstrip patch elements and arrays. The more relevant ones include PATCHD and PATCH9 for analyzing rectangular patches, WRAPRND for analyzing a wraparound patch, and CPPATCH for analyzing either a nearly square or a truncated corner circularly polarized patch element.

3.5.1.1 PATCHD

PATCHD calculates the basic parameters for both rectangular and circular patch elements. These parameters include the resonant dimensions, input resistance, efficiency, and quality factor. The calculations are based on closed form expressions derived from full wave solutions, which include the surface wave effects of the patch but not the feed. The accuracy of the results is expected to be within a few percent and is valid for a limited range of permittivity, substrate height, and patch width.

3.5.1.2 PATCH9

PATCH9 calculates basic parameters for rectangular patch elements. It has two modes, one used for design and the other for analysis. In the design mode, the length and input impedance is determined as a function of frequency. In the analysis mode, the input impedance is determined based on the patch dimensions and parameters entered. Based on the transmission line model, PATCH9 is expected to have an accuracy of 2 to 3 percent over a limited range of permittivity, substrate height, and patch width. In contrast to PATCHD, PATCH9 includes the feed in its calculations, including edge, probe, and inset feeds.

3.5.1.3. WRAPRND

WRAPRND calculates the length, resonant frequency, number of feeds for a wrap around patch. The input impedance is not calculated. The accuracy of the results is expected to be better than five percent.

3.5.1.4 CPPATCH

CPPATCH calculates the resonant frequency and input impedance of nearly square and truncated corner circularly polarized patch elements. CPPATCH is based on both PATCHD and PATCH9. It uses the curve fit equations based on fullwave analysis from PATCHD and uses the feed models from PATCH9. The accuracy is expected to be about three percent, which is significant since the axial ratio bandwidth of a single feed, circularly polarized element is typically less than one percent.

3.5.2 Clementine

Ansoft's Clementine is a fullwave analysis CAD program that analyzes single layered microstrip geometries on a cylinder. Once the geometry is entered, which may be of any dimension, it can be analyzed. The analysis results include S-parameters, input impedance, surface currents, and far-field radiation patterns. In all these cases, the results can be viewed in text or graphic form.

Although Clementine can accurately analyze microstrip antennas, it cannot directly design an antenna. A microstrip patch or array of patches must be designed through an iterative process. In order to determine the initial patch dimensions to begin this iterative design process, Clementine provides an estimation tool, which quickly provides estimated dimensions for microstrip feeds, rectangular patch elements, and circularly polarized elements. These estimated patch dimensions are determined using the cavity model for planar surfaces. The number of iterations required to have an accurately designed antenna depends on the complexity of the patch and its feed network.

Clementine is uses mixed-potential integral equations and the method of moments. Its results are accurate and take into account all mutual coupling effects. The microstrip geometry is subdivided into meshed sections automatically before the simulation can occur. The sizes of the mesh can either be entered or automatically determined. If the meshed sections consist of many triangular sections that are very narrow, there is a chance that the results may be inaccurate.

Fullwave solutions that are based on the method of moments can be computationally intensive, especially when the patch dimensions are large. The size of the structure that can be simulated depends on the memory capacity of the computer (RAM). Ansoft recommends a computer with 64 MB of RAM to simulate arrays having 12 elements (3000 unknowns). For example, a 500 MHz Pentium III with 256 MB of RAM took between 10 to 20 hours to analyze a wraparound patch designed for an operating frequency of 2.2155 GHz. In contrast, a 366 MHz Pentium III with only 64 MB of RAM was unable perform the same simulation.

While the design of the antenna cannot be based solely on a CAD tool, they are extremely helpful to the designer. They can enable the designer to complete an antenna design much faster, while the old “cut and try” method may take several months and be significantly more expensive.