THE CIRCULAR MICROSTRIP PATCH ANTENNA – DESIGN AND IMPLEMENTATION

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ABSTRACT
A FORTRAN program was developed to simulate the basic parameters of a microwave circular patch antenna. These parameters are the actual radius of patch, effective radius of patch, conductance due to radiation, conductance due to conduction, conductance due to dielectric loss, directivity, input resistance and quality factors due to conduction, dielectric loss, and radiation. Alongside, these parameters were manually computed. Four substrates were selected - Galium Arsenide, Duroid, Indium Phosphide and Silicon. Deductions made from the results showed that Galium Arsenide is suitable when smaller antenna size and low power handling capability is apriori. However, when size is not a constraint, Duroid is exceptional especially in directivity and high power radiation. Patch radius decreases as the resonant frequency increases (0.2374cm at 10.0GHz and 0.05079 cm at 45.0GHz for GaAs). The results obtained in this design compare favorably with results obtained from manual computation of the same parameters and these agree with other designs such as the rectangular patch.

Keywords: *Circular Microstrip Patch Antenna Design.*

1. INTRODUCTION
Microstrip antennas basically consist of a radiating patch on one side of a dielectric substrate, which has a ground plane on the other side. The patch is generally made of conducting material such as copper and gold (Wikipedia, 2010). The patch is very thin ($t \ll \lambda_o$ where λ_o is free space wavelength) and is placed a small fraction of a wavelength ($h \ll \lambda_o$ usually $0.003\ \lambda_o \leq h \leq 0.05\ \lambda_o$) above the ground plane. The microstrip patch is designed so its pattern maximum is normal to the patch (broadside radiator). This is accomplished by properly choosing the mode (field configuration) of excitation beneath the patch.
There are numerous substrates that can be used for the design of microstrip patch antennas and their dielectric constants are usually in the range of $2.2 \leq \varepsilon_r \leq 12$. Those desirable for antenna performance are thick substrates whose dielectric constant are in the lower end of the range due to better efficiency, larger bandwidth, and loosely bound fields for radiation into space but at the expense of larger element size. Microstrip patch antennas radiate primarily because of the fringing fields between the patch edge and the ground plane. The radiation increases with frequency, thicker substrates, lower permittivity, and originates mostly at discontinuities (Lewin, 1960)
Since microstrip antennas are often integrated with other microwave circuitry, a compromise has to be reached between good antenna performance and circuit design. The radiating element and the feed lines are usually photo etched on the dielectric substrate. The radiating patch may be square, rectangle, thin strip (dipole), circular, elliptical, triangle or any other configuration. A microstrip antenna is very versatile and made for a wide range of resonant frequencies, polarization patterns and impedances. Due to its operational features viz low efficiency, low power, high quality factor, poor polarization purity, poor scan performance and very narrow frequency bandwidth, it is suitable for mobile and government security systems where narrow bandwidth are priority. They are also used on laptops, microcomputers, mobile phones etc (Wikipedia, 2010).

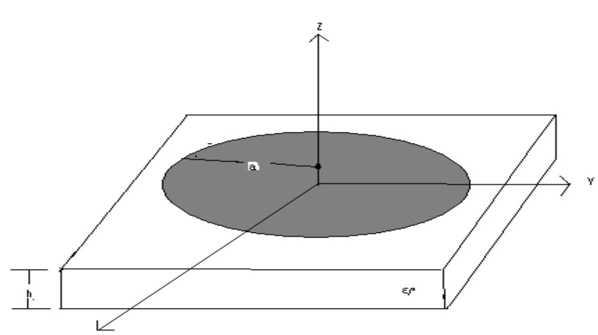

Fig 1 Circular patch antenna

2. METHODS OF ANALYSIS

There are three popular models for the analysis of microstrip antennas - viz transmission line model, cavity model, and full wave model. The transmission line model is the simplest. It gives a good physical insight but is less accurate. The cavity model, which is used in this work, is quite complex but gives good physical insight and is more accurate. The full wave model is the most complex. It is very accurate in the design of finite and infinite arrays or stacked structures.

The quantity associated with radiated EM wave is the Poynting vector given as:(Balanis, 1982)

$$S = E \times H \qquad (1)$$

where S is instantaneous Poynting vector, E is instantaneous electric field intensity and H is instantaneous magnetic field intensity. The complex fields E and H are related to their instantaneous counterparts by (Richards, 1988; Gonca, 2005):

$$\left.\begin{aligned} E(x,y,z,t) &= \mathrm{Re}\left[E(x,y,z)e^{iwt}\right] \\ H(x,y,z,t) &= \mathrm{Re}\left[H(x,y,z)e^{iwt}\right] \end{aligned}\right\} \qquad (2)$$

Using and the identity $\mathrm{Re}\left[Xe^{iwt}\right] = \frac{1}{2}\left[Xe^{iwt} + X^{*}e^{iwt}\right]$ equation (1) can be rewritten as;

$$S = \tfrac{1}{2}\mathrm{Re}\left[E \times H^{*}\right] + \tfrac{1}{2}\mathrm{Re}\left[E \times H\, e^{iwt}\right] \qquad (3)$$

Hence, the time average Poynting vector can be written as (Gonca, 2005; Akande, 2003)

$$S_{av} = \tfrac{1}{2}\mathrm{Re}\left[E \times H\right] \quad Wm^{-2} \qquad (4)$$

The factor ½ appears because the E and H fields are peak values and not rms.

This research aims to design and implement a circular microstrip patch antenna suitable for use at microwave frequencies. Its performance parameters will be determined through manual computation and also a computer program to be developed in *FORTRAN*. Both results obtained will be compared for accuracy.

3. CIRCULAR PATCH AND FIELD CONFIGURATION

The mode supported by the circular patch antenna can be found by treating the patch, ground plane and the material between the two as a circular cavity. The radius of the patch is the only degree of freedom to control the modes of the antenna (Balanis, 1982). The antenna can be conveniently analyzed using the cavity model (Richards, 1988; Gonca, 2005). The cavity is composed of two electric conductors at the top and the bottom to represent the patch and the ground plane and by a cylindrical perfect magnetic conductor around the circular periphery of the cavity. The dielectric material of the substrate is assumed to be truncated beyond the extent of the patch (Richards, 1988).

The field configuration within the cavity can be found using the vector potential. The magnetic vector potential A_z must satisfy, the homogeneous wave equation (Balanis, 1982)

$$\nabla^2 A_z(\varphi,\phi,z) \quad + \quad k^2 A_z(\varphi,\phi,z) \quad = \quad 0 \qquad (5)$$

whose solution is written as

$$A_x = B_{mnp} J_m(k_\rho \rho')\left[A_2 \cos(m\phi') + B_2 \sin(m\phi')\right]\cos(k_z z') \qquad (6)$$

with constraint equation of

$$(k_\rho)^2 + (k_z) = k_r^2 = \omega_r^2 \mu\varepsilon \qquad (7)$$

The electric and magnetic fields are related to the vector potential A_z by (Balanis, 1982)

$$E_\rho = -j\frac{1}{\omega\mu\varepsilon}\frac{\partial^2 A_{zz}}{\partial\rho\partial z} \qquad H_\rho = \frac{1}{\mu}\frac{1}{\rho}\frac{\partial A_z}{\partial\phi} \qquad (8a)$$

$$E_\phi = -j\frac{1}{\omega\mu\varepsilon}\frac{1}{\rho}\frac{\partial^2 A_z}{\partial\rho\partial z} \qquad H_\phi = -\frac{1}{\mu}\frac{\partial A_z}{\partial\rho}_x \qquad (8b)$$

$$E_\rho = -j\frac{1}{\omega\mu\varepsilon}\left(\frac{\partial^2}{\partial z^2} + k^2\right)A_z \qquad H_z = 0 \qquad (8c)$$

These are subjected to the boundary conditions of (Balanis, 1982; Richards, 1988; Gonca, 2005)

$$E_\rho(0 \le \rho' \le a, 0 \le \phi' \le 2\pi, z' = 0) = 0$$
$$E_\rho(0 \le \rho' \le a, 0 \le \phi' \le 2\pi, z' = h) = 0 \quad (9)$$
$$H_\phi(\rho' = a, 0 \le a, 0 \le \phi' \le 2\pi, 0 \le z' \le h) = 0$$

The primed coordinates ρ', ϕ', z' are used to represent the fields within the cavity while $J_m(x)$ is the Bessel function of the first order.

$$k_\rho = X'_{mn} / a \quad (10a)$$

$$k_z = \frac{p\pi}{h} \quad (10b)$$

where $m = 0,1,2,..........$, $n = 1,2,3,..........$ and $p = 0,1,2,..........$

X'_{mn} represents the zeroes of the derivative of the Bessel function $J_m(x)$ and they determine the order of the resonant frequencies. The resonant frequencies of the cavity, and thus the microstrip antenna are found using (7) and (10). For typical microstrip antennas, the substrate height is very small, the fields along z are essentially constant (p = 0 and k_z = 0). Therefore the resonant frequencies for the TM_{mn0} modes can be written as (Richards, 1988; Gonca, 2005)

$$(f_r)_{mn0} = \frac{1}{2\pi\sqrt{\mu\varepsilon}}\left(\frac{X'_{mn}}{a}\right) \quad (11)$$

The field radiated by the circular patch can be found by using the Equivalence principle whereby the circumferential wall of the cavity is replaced by an equivalent magnetic current density radiating in free space. Based on cavity model and assuming a TM^z_{110} mode field's distribution beneath the patch, the normalized electric and magnetic fields within the cavity for the cosine azimuthal variations can be written as (Balanis, 1982; Burkholder and Lundin 2006)

$$E_p = E_\phi = H_z = 0, \; E_z = E_o j_1(kp^1)\cos\phi^1 \quad (12a)$$

$$H_p = j\frac{E_o}{\omega\mu^o} = \frac{1}{p} j_1(kp^1)\sin\phi^1, \; H_\phi = j\frac{E_o}{\omega\mu^o} j_1(kp^1)\cos\phi^1 \quad (12b)$$

Where $` = \partial / \partial p \; and \; \phi^1$ is the azimuthal angle along the perimeter of the patch.

Based on (12a) evaluated at the electrical equivalent edge of the disk ($\rho^1 = a_e$), the magnetic current density can be written as (Balanis, 1982; Burkholder and Lundin 2006)

$$M_s = -2\hat{n} \; X \; E_a /_{p^1 = ae} = a_\phi 2E_o J_1(ka_e)\cos\phi^1 \quad (13)$$

Since the height of the substrate is very small and the current density of (6) is uniform along the z direction, we can approximate (13) by a filamentary magnetic current of (Balanis, 1982)

$$I_m = hM_s = \partial_\phi 2hE_o J_1(ka_e)\cos\phi^1 = \partial_\phi 2V_o \cos\phi^1 \quad (14)$$

where $V_o = hE_o J_1(ka_e) \; at \; \phi^1 = 0$. The microstrip antenna can now be treated as a circular loop. And using the radiation equations we can write (Balanis, 1982; Burkholder and Lundin 2006)

$$E_r = 0 \quad (15a)$$

$$E_\theta = -j\frac{k_o a^e V_o e^{-jk_o{}^r}}{2r}\{\cos\phi \; j^1{}_{02}\} \quad (15b)$$

$$E_\phi = j\frac{k_o a^e V_o e^{-jk_o{}^r}}{2r}\{\cos\theta\sin\phi \; j_{02}\} \quad (15c)$$

$$J^1{}_{o2} = J_o(k_o a^e \sin\theta) - (k_o a^e \sin\theta) \quad (15d)$$

$$J_{o2} = J_o(k_o a^e \sin\theta) + (k_o a^e \sin\theta) \quad (15e)$$

where a_e is the effective radius. The fields in the principal planes reduces to (Balanis, 1982)

$$E - plane(\phi = 0^o, 180^o, \; 0^o \le \theta \le 90^o)$$

$$E_\theta = j\frac{k_o a_e V_o e^{-jk_o r}}{2r}\left\{j^1{}_{02}\right\} \tag{16a}$$

$$E_\theta = 0 \tag{16b}$$

$$E_\phi = j\frac{k_o a_e V_o e^{-jk_o r}}{2r}\left\{\cos\theta j_{02}\right\} \tag{16c}$$

4. DESIGN ANALYSIS OF CIRCULAR PATCH MICROSTRIP ANTENNA

4.1 Circular Patch Radius and Effective Radius

Since the dimension of the patch is treated a circular loop, the actual radius of the patch is given by (Balanis, 1982)

$$a = \frac{F}{\left\{1+\frac{2h}{\pi\varepsilon_r F}\left[\ln\left(\frac{\pi F}{2h}\right)+1.7726\right]\right\}^{1/2}} \tag{17}$$

$$F = \frac{8.791\times10^9}{f_r\sqrt{\varepsilon_r}}$$

Equation (17) does not take into consideration the fringing effect. Since fringing makes the patch electrically larger, the effective radius of patch is used and is given by (Balanis, 1982)

$$a_e = a\left\{1+\frac{2h}{\pi\varepsilon_r a}\left[\ln\left(\frac{\pi a}{2h}\right)+1.7726\right]\right\}^{1/2} \tag{18}$$

Hence, the resonant frequency for the dominant $TM^z{}_{110}$ is given by (Balanis, 1982)

$$(f_r)_{110} = \frac{1.8412\nu_0}{2\pi a_e\sqrt{\varepsilon_r}} \tag{19}$$

where v_0 is the free space speed of light.

4.2 Conductance

The conductance due to the radiated power of the circular microstrip patch antenna can be computed based on the the radiated power expressed as; (Balanis, 1982)

$$P_{rad} = |V_0|\frac{(k_0a_e)^2}{960}\int_0^{\pi/2}\left[J'_{02}{}^2+\cos^2\theta J_{02}{}^2\right]\sin\theta d\theta \tag{20a}$$

equation (20a) can be further broken down to

$$P_{rad} = |V_0|^2\frac{(k_0a_e)^2}{960}\left\{\begin{aligned}&\left[1-\frac{3}{2^3}\alpha^2\sin^2\theta+\frac{1}{3\times2^3}\alpha^4\sin^4\theta-\frac{7}{12\times2^{10}}\alpha^6\sin^6\theta\right]+\\&0.333\left[1-\frac{3}{2^3}\alpha^2\sin^2\theta+\frac{1}{3\times2^3}\alpha^4\sin^4\theta-\frac{7}{12\times2^{10}}\alpha^6\sin^6\theta\right]\end{aligned}\right\} \tag{20b}$$

Where $V_0 = hE_0J_1(ka_e)$, $\alpha = k_0a_e$ and $k_0 = \frac{2\pi f_r}{\nu_0}$ (21)

k_0 is the free space phase constant. The conductance across the gap between the patch and the ground plane at $\varphi' = 0^o$ is given as (Balanis, 1982)

$$G_{rad} = \frac{(k_0a_e)^2}{480}\int_0^{\pi/2}\left[J'_{02}{}^2+\cos^2\theta J_{02}{}^2\right]\sin\theta d\theta \tag{22}$$

Equation (22) further reduces to

$$G_{rad} = \frac{(k_0 a_e)^2}{480}\left\{\begin{array}{l}\left[1-\frac{3}{2^3}\alpha^2\sin^2\theta+\frac{1}{3\times 2^3}\alpha^4\sin^4\theta-\frac{7}{12\times 2^{10}}\alpha^6\sin^6\theta\right] + \\ 0.333\left[1-\frac{3}{2^3}\alpha^2\sin^2\theta+\frac{1}{3\times 2^3}\alpha^4\sin^4\theta-\frac{7}{12\times 2^{10}}\alpha^6\sin^6\theta\right]\end{array}\right\} \quad (23)$$

G_{rad} accounts for radiation and dielectric losses and are expressed as (Balanis, 1982)

$$G_c = \frac{\varepsilon_{mo}\pi\left(\pi\mu_0 (f_r)_{10}\right)^{-3/2}}{4h^2\sqrt{\sigma}}\left[(ka_e)^2 - m^2\right] \quad (24)$$

$$G_d = \frac{\varepsilon_{mo}\tan\delta}{4\mu_0 h (f_r)_{10}}\left[(ka_e)^2 - m^2\right] \quad (25)$$

where G_c is the conductance due to conduction losses, G_d is the conductance due to dielectric losses and f_r is the resonant frequency of the dominant mode. The total conductance can be expressed as

$G_t = G_{rad} + G_c + G_d$ (26)

The conductivity of the substrate, $\sigma = \frac{G_t l}{A}$, and skin dept, $\delta = \sqrt{2/(\omega\mu_0\sigma)}$ (27)

4.3 Directivity

For all antennas the directivity is one of the most important figures -of –merit, and is given as (Balanis, 1982);

$$D_0 = \frac{U_{max}}{U_0} = \frac{4\pi U_{max}}{P_{rad}} \quad (28a)$$

Considering the electric field from (15), the maximum radiation intensity and radiated power in (20), the directivity of a single slot may be written as (Balanis, 1982).

$$D_0 = \frac{(k_0 a_e)^2}{120 G_{rad}} \quad (28b)$$

This directivity is not strongly influenced by height of substrate as long as it is maintained electrically small. It is a function of patch radius.

4.4 Resonant Input Impedance

The input impedance of a circular patch at resonance is real and the input power is independent of the feed point position on the circumference (Balanis, 1982). Taking the reference of the feed point at $\phi' = 0^\circ$, the input resistance at any radial distance $\rho' = \rho_0$ from the center of the patch can be written as (Balanis, 1982)

$$R_{in}(\rho' = \rho_0) = \frac{1}{G_t}\frac{J^2{}_m(k\rho_0)}{J^2{}_m(ka_e)} \quad (29)$$

For the circular patch antenna, the resonant input resistance with an inset feed is (Balanis, 1982).

$$R_{in}(\rho' = \rho_0) = R_{in}(\rho' = a_e)\frac{J^2{}_m(k\rho_0)}{J^2{}_m(ka_e)}$$

$$R_{in}(\rho' = a_e) = \frac{1}{G_t} \quad (30)$$

where the radial distance is given as

$$\rho = \frac{2(2a)}{\lambda} \quad (31)$$

For thick elements (substrate), reactance may be significant and needs to be taken into account in impedance matching and in determining the resonant frequency of a loaded element. (Richards, 1988)

4.5 Quality Factor

The quality factor is a figure of merit that is representative of the antenna losses. Typically there are radiation, conduction (ohmic),dielectric and surface wave losses. The summation of which gives the total quality factor Q_t, which is influenced by all these losses .

$$Q_t^{-1} = Q_{rad}^{-1} + Q_c^{-1} + Q_d^{-1} + Q_{sw}^{-1} \quad (32)$$

For very thin substrates, losses due to surface waves (Q^{-1}_{sw}) are very small and can be neglected.
For such substrates ($h << \lambda_0$) and may be represented as

$$Q_c^{-1} = h\sqrt{\pi f \mu \sigma} \quad (33)$$

$$Q_d^{-1} = \frac{1}{\tan\delta} \quad (34)$$

$$Q_{rad}^{-1} = \frac{2\omega\varepsilon_r}{h\left(G_{rad}/2a\right)}\left(\frac{a}{2}\right) \quad (35)$$

Tan δ is the loss tangent of the substrate material, σ is the conductivity of the conductors associated with the patch and ground plane and $G_t/2a$ is the total conductance per unit length of the radiating aperture. G_{rad} is usually the dominant factor since it is inversely proportional to the patch height.

5. PROGRAM DESIGN AND SIMULATION

The program written in FORTRAN using WATFORT g77 compiler was developed based on equations (17) to (35). The program was run on the DOS mode and results exported to Microsoft word. The main program reads in the microstrip parameters then determines the ideal radiation characteristics.

5.1 Input Parameters:

$\varepsilon_r = dielectric\ constant\ of\ substrate$

$f_r = resonant\ frequency\ of\ substrate$

$h = height\ of\ substrate$

$v_0 = Speed\ due\ to\ free\ space = 3\times10^{10} cm/s$

$\mu_0 = Permitivity\ due\ to\ free\ space = 4\pi\times10^{-3} H/cm$

5.2 Output Parameters

$a = The\ radius\ of\ patch$

$a_e = The\ effective\ radius\ of\ patch$

$G_{rad} = Conductance\ between\ gap\ and\ ground$

$G_d = Conductance\ due\ to\ dielectric$

$G_t = Total\ conductance$

$D_0 = Directivity\ of\ slot(\phi = 0^0)$

$R_{in}(\rho' = \rho_0) = Resonant\ input\ resistance\ (\phi = 0^0)$

$Q_c^{-1} = Quality\ factor\ due\ to\ conduction\ (Ohmic)\ losses$

$Q_d^{-1} = Quality\ factor\ due\ to\ dielectric\ losses$

$Q_{rad}^{-1} = Quality\ factor\ due\ to\ radiation\ (space\ wave)\ losses$

The radius of patch is obtained from equation 17. The effective patch radius is given in equation 18. Radiation power is given in equation 20. Radiation conductance is given in equation 23. The voltage across gap is $V_0 = hE_0J_1(ka_e)$ (for $\theta = 90^0$). Conductivity of the substrate and the skin depth are given in equation 27. The resonant frequency is given in equation 19. Conductance due to radiation loss of equatio3.90 is $G_c = \frac{\varepsilon_{mo}\pi\left(\pi\mu_0 (f_r)_{10}\right)^{-3/2}}{4h^2\sqrt{\sigma}}\left[(ka_e)^2 - m^2\right]$. The total conductance is given in equation 26. The directivity of a slot is given in equation 28. The resonant input impedance is given in equation 30. The total radial distance is given in equation 31 and the quality factors are given in equations 33, 34 and 35.

6. MANUAL COMPUTATION OF ANTENNA PARAMETERS

Substrate 1: Silicon.

Resonant frequency of 30GHz, dielectric constant of 11.8 and a height of 0.20cm, the following were obtained.

$$F = \frac{8.791\times10^{9}}{30\times10^{9}\sqrt{11.8}} = 0.08531$$

The radius is found using equation 17

$$a = \frac{0.08531}{\left\{1+\frac{2\times0.20}{\pi\times11.8\times0.08531}\left[\ln\left(\frac{\pi\times0.08531}{2\times0.20}\right)+1.7726\right]\right\}^{1/2}} = 0.07875$$

The effective radius is found using equation 18

$$a_e = 0.07875\left\{1+\frac{2\times0.20}{\pi\times11.8\times0.07875}\left[\ln\left(\frac{\pi\times0.07875}{2\times0.1588}\right)+1.7726\right]\right\}^{1/2} = 0.08543$$

The free space phase constant is obtained using equation 21

$$k_0 = \frac{2\pi\times30\times10^{9}}{3\times10^{10}} = 6.2832$$

Using equation 21 and the Bessel function, we can obtain the voltage across gap.

$$E_0 \approx 3.9383\,;\; V_0 = 0.20\times3.9383\times0.04994 = 0.03934$$

The constant α is obtained using equation 21; $\alpha = 6.2832\times0.08543 = 0.5368$

The radiated power is given by equation 20b

$$P_{rad} = |0.03934|^2\frac{(0.5368)^2}{960}\left\{\begin{array}{l}\left[1-\frac{3}{2^3}\alpha^2\sin^2\theta+\frac{1}{3\times2^3}\alpha^4\sin^4\theta-\frac{7}{12\times2^{10}}\alpha^6\sin^6\theta\right]+\\ 0.333\left[1-\frac{3}{2^3}\alpha^2\sin^2\theta+\frac{1}{3\times2^3}\alpha^4\sin^4\theta-\frac{7}{12\times2^{10}}\alpha^6\sin^6\theta\right]\end{array}\right\} = 0.000000554$$

The conductance between gap and ground is gotten using equation 23.

$$G_{rad} = \frac{(0.5368)^2}{480}\left\{\begin{array}{l}\left[1-\frac{3}{2^3}\alpha^2\sin^2\theta+\frac{1}{3\times2^3}\alpha^4\sin^4\theta-\frac{7}{12\times2^{10}}\alpha^6\sin^6\theta\right]+\\ 0.333\left[1-\frac{3}{2^3}\alpha^2\sin^2\theta+\frac{1}{3\times2^3}\alpha^4\sin^4\theta-\frac{7}{12\times2^{10}}\alpha^6\sin^6\theta\right]\end{array}\right\} = 0.0007165\,Siemens$$

The conductance, angular frequency and skin depth are obtained from equation 26.

$$\sigma = \frac{0.0007161\times2\times0.07875}{\pi(0.07875)^2} = 0.005789$$

$$\omega = 2\pi\times30\times10^{9} = 1.88496\times10^{10}$$

$$\delta = \sqrt{2/\left(1.8849\times10^{10}\times4\pi\times10^{-3}\times0.005789\right)} = 0.000381899$$

The resonant frequency is obtained using equation (19); $(f_r)_{110} = \frac{1.8412\times3\times10^{10}}{2\pi\times0.08543\sqrt{11.8}} = 2.99557\times10^{10}$

The conductance due to conduction is given by equation (24).

$$G_c = \frac{11.8\times\pi\left(\pi\times4\pi\times10^{-2}\times2.99557\times10^{10}\right)^{-3/2}}{4(0.20)^2\sqrt{0.005789}}\left[(0.08543)^2-m^2\right] = 5.4649\times10^{-13}$$

Conductance due to dielectric is also gotten using (25)

$$G_d = \frac{11.8\tan0.000381899}{4\times4\pi\times10^{-3}\times0.2\times2.99557\times10^{10}}\left[(0.08543)^2-m^2\right] = 1.9\times10^{-15}$$

Then, the total conductance is obtained using equation 27.

$$G_t = 0.0007165 + 5.4649 \times 10^{-13} + 1.9 \times 10^{-15} = 0.0007165$$

The directivity of slot is obtained from equation (28), $D_0 = \dfrac{(0.5368)^2}{120 \times 0.0007165} = 3.3513$

The radial distance is obtained using equation (31); $\rho = \dfrac{2(2 \times 0.07875)}{3} = 0.04961$

The resonant input resistance is also found using equation (30)

$$R_{in} = \frac{1}{0.0007165} \frac{J^2{}_0(0.04961)}{J^2{}_0(0.08543)} = 1395.7026\Omega$$

Finally, the quality factor due to conduction; dielectric and radiation are computed using equations (33), (34) and (35) respectively.

$$Q_c^{-1} = 0.2\sqrt{\pi \times 30 \times 10^9 \times 4\pi \times 10^{-3} \times 0.005789} = 0.0019109$$

$$Q_d^{-1} = \frac{1}{\tan 0.00038189} = 6.665 \times 10^{-6}$$

$$Q_{rad}^{-1} = \frac{2 \times 1.88496 \times 10^{11} \times 11.8}{0.2\left(0.0007165 / 2 \times 0.07875\right)} \left(\frac{0.07875}{2}\right) = 5.1949 \times 10^{-15}$$

The same method was used to compute the parameters for the other three substrates with baseline parameters (Resonant frequency, dielectric constant and height) stated on each Table. Computed values of the radius, effective radius, conductance of slot, conductance between gap and ground, conductance due to conduction, conductance due to dielectric, total conductance, directivity of slot, resonant input resistance, quality factor due to conduction, quality factor due to dielectric, quality factor due to radiation of a circular patch microstrip patch antenna using different substrates are shown on tables 1 to 4 on the Appendix

7. RESULTS

Table 1 Silicon f_r =30e9Hz ε_r =11.8 h =0.20cm

Parameters	a	a_e	G_1	G_{rad}	G_c	G_d	G_t	D_0	$R_{in}(\rho'=\rho$	Q_c^{-1}	Q_d^{-1}	Q_{rad}^{-1}
Manual computation	0.0788	0.0854	7.161e-4	7.165e-4	5.465e-13	1.91e-15	7.165e-4	3.3513	1395.7026	1.911e-3	6.664e-6	5.195e-15
Program computation	0.0788	0.0854	7.165e-4	7.165e-4	5.463e-13	1.906e-15	7.165e-4	3.3513	1395.7020	1.909e-3	6.664e-6	5.195e-15

Table 2 RT/Duroid 5880 (low index) f_r =10e9Hz ε_r =2.2 h =0.1588cm

Parameters	a	a_e	G_1	G_{rad}	G_c	G_d	G_t	D_0	$R_{in}(\rho'=\rho$	Q_c^{-1}	Q_d^{-1}	Q_{rad}^{-1}
Manual computation	0.5249	0.5984	0.00223	0.00223	6.101e-11	1.68e-13	0.00223	5.8662	418.3568	6.093e-3	1.689e-5	4.651e-15
Program computation	0.5249	0.5984	0.00223	0.00223	6.101e-11	1.68e-13	0.00223	5.8662	418.3567	6.093e-3	1.689e-5	4.651e-15

Table 3 Gallium Arsenide(Static) f_r =45e9Hz ε_r =12.90 h =0.10cm

Parameters	a	a_e	G_1	G_{rad}	G_c	G_d	G_t	D_0	$R_{in}(\rho'=\rho$	Q_c^{-1}	Q_d^{-1}	Q_{rad}^{-1}
Manual computation	0.0508	0.0545	6.618e-4	6.619e-4	4.420e-13	7.709e-16	6.619e-4	3.3193	1518.4191	2.605e-3	4.546e-6	3.516e-15
Program computation	0.0508	0.0545	6.619e-4	6.619e-4	4.420e-13	7.708e-16	6.619e-4	3.3193	1518.4185	2.605e-3	4.547e-6	3.516e-15

Table 4 Indium Phosphide(Static) f_r =45e9 Hz ε_r =12.4 h =0.10cm

Parameters	a	a_e	G_1	G_{rad}	G_c	G_d	G_t	D_0	$R_{in}(\rho'=\rho_0$	Q_c^{-1}	Q_d^{-1}	Q_{rad}^{-1}
Manual computation	0.0517	0.0556	6.877e-4	6.859e-4	4.377e-13	7.633e-16	6.859e-4	3.3333	1465.3005	2.581e-3	4.500e-6	3.659e-15
Program computation	0.0517	0.0556	6.859e-4	6.859e-4	4.383e-13	7.643e-16	6.859e-4	3.3333	1465.2991	2.582e-3	4.506e-6	3.659e-15

Table 5: Comparison of some selected substrates at the same frequency and height. (f_r = 10.0e9, h = 0.1 cm)

Characteristics	**Silicon** **ε_r = 11.80**	**Duroid** **ε_r = 2.20**	**GaAs** **ε_r = 12.90**	**InP** **ε_r = 112**
a	0.2478	0.5421	0.2374	0.2419
a_e	0.2561	0.5960	0.2449	0.2498
G_1	0.0007155	0.002227	0.0006607	0.0006845
G_{rad}	0.0007155	0.002227	0.0006607	0.0006845
G_c	1.8092×10^{-10}	1.5429×10^{-10}	1.8423×10^{-10}	1.8275×10^{-10}
G_d	3.1565×10^{-13}	2.6855×10^{-13}	3.2145×10^{-13}	3.1885×10^{-13}
G_t	0.0007155	0.002227	0.0006607	0.0006845
D_o	3.3507	5.8305	3.3187	3.3325
R_{in}	1433.34	386.62	1552.20	1498.15
Q_c^{-1}	0.01174	0.009841	0.01196	0.01186
Q_d^{-1}	2.0488×10^{-5}	1.7175×10^{-5}	2.0869×10^{-5}	2.0698×10^{-5}
Q_{rad}^{-1}	7.8587×10^{-16}	2.7417×10^{-15}	7.2311×10^{-16}	7.5037×10^{-16}

8. DISCUSSION OF RESULTS

From the results obtained from both the simulation and the manual computation as presented on Tables 1 to 4, it was clear that the outputs of both were very close. The slight differences could be traced to approximations made in manual computations. Table 5 showed that GaAs with a patch radius of 0.2374 cm, has the smallest radius compared to other chosen substrates, hence is best in miniaturization. It also showed a high input resistant (impedance) of 1552.20, which means that loading effect will be minimal. However, it has the least radiation conductance, which implies that its radiation is least when compared to the other substrates. It also showed the least directivity of 3.3187. The reduced radiation can be traced to the fact that it showed the highest losses due to radiation conductance (Q_{rad}), which is the dominant loss for thin substrates. This makes it a prime choice for Bluetooth application. Indium Phosphide is next to Galium Arsenide when reduced size and less loading effect are the major requirements. Duroid has the best conductance (0.002227 Siemens), which means if used to construct an antenna will give a high radiation when compared to the other chosen substrates. The high radiation is due to the high fringing observed. It also showed the best directivity and the least losses due to radiation conductance (Q_{rad}). However, these advantages are at the expense of an enlarged patch radius and a reduced input resistance (increase in loading effect). Therefore, when directivity is apriori, Duroid is the best option. This makes it useful in high frequency mobile telecom. The stability feature of Silicon makes it suitable for patch antenna construction. Patch radius decreases as the required resonant frequency increases. Directivity was seen to increase with resonant frequency, but at the expense of increase in

conduction and dielectric losses. Radiation losses were observed to have reduced with increase in resonant frequency, which translates to the slight increase in the radiation conductance.
It was also deduced that directivity increases with increase in substrate height, but at the expense of loading effect. A lower radiation conductance loss was observed at higher substrate height, but at the expense of increasing losses due to surface wave.

9. SUMMARY AND CONCLUSION

This research was aimed at designing and implementing circular microstrip patch antenna using cavity model. Alongside this, various parameters viz actual and effective patch radii, conductance, directivity, input resistance and quality factor, which dictate the ultimate performance of the antenna were determined by simulation using a program developed in FORTRAN and also by manual computation.
In the course of this project complex design equations were simplified using mathematical formulations and the laws of Physics. Deductions were made on results obtained for different substrates and it was discovered that Gallium Arsenide is suitable when smaller antenna size is required while Duroid is suitable for directivity and radiation power, except for its large size. Also, Silicon was observed to be the most suitable substrate for patch antenna construction due to its moderate radiation characteristics.
The program developed could be used in the design, analysis and manufacture of more complex microstrip antenna configurations to suit different shapes that can be flush mounted on spacecrafts and in microwave circuitry for usage such as in telecommunication, security, aviation, medicine etc.
Results obtained in this design were observed to be in agreement with those obtained from rectangular microstrip patch.

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