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Effect of Feed Location on Rectangular Microstrip Antenna at TM_{11} Mode

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ABSTRACT

The variation of 2:1 VSWR impedance band width with the change of feed location, at TM_{11} mode excitation is studied experimentally. Experimental investigations revealed that the 2:1 VSWR impedance band width of a rectangular microstrip patch antenna at TM_{11} mode could be improved by properly choosing the location of the feed point along one of the radiating edges; and this is verified theoretically. This investigation may be useful for low gain microstrip antennas with conical radiation patterns.

1. INTRODUCTION

A microstrip antenna with higher order mode excitation can be used as a land mobile antenna or for satellite communication[1]. The 2:1 VSWR band width of a microstrip antenna is 2-3 per cent of its resonant frequency. Experimentally, an appreciable improvement in 2:1 VSWR impedance band width is achieved[2] by exciting the antenna at TM_{11} mode and properly locating the feed point along the edge of the antenna. In this communication, the variation of 2:1 VSWR impedance band width with the change of feed location, at TM_{11} mode excitation, is studied experimentally. At TM_{11} mode, by shifting the feed location along one of the edges of the antenna which corresponds to the non-radiating edges for fundamental mode, an improved 2:1 VSWR impedance band width (maximum of 6.5 per cent) was achieved. The measured results are confirmed by the use of both the transmission line model and the cavity model. The theoretical and measured conical radiation patterns at TM_{11} mode are also presented.

2. THEORY

2.1 Transmission Line Model

The transmission line model for microstrip antenna, developed by Pues and Van de Capelle[3], was further developed by Dearnley and Barel[4] by incorporating change in equivalent length and feed offset to determine the input characteristics over a large band of frequencies. In the Transmission line model, the rectangular microstrip antenna is represented by an open-ended transmission line with characteristic admittance, propagation constant and physical length.

The input characteristic of the rectangular patch[4] is given by

$$Y_{patch} \doteq 2Y_o \frac{Y_o^2 + Y_t^2 + 2Y_s Y_o \coth(\gamma L) - 2Y_m Y_o \,\mathrm{csch}(\gamma L)}{(Y_o^2 + Y_t^2)\coth(\gamma L) + (Y_o^2 - Y_t^2)\cosh(2\gamma\Delta_{off})} \times \frac{1}{\mathrm{csch}(\gamma L) + Y_s Y_o} \quad (1)$$

where

$$Y_t^2 = Y_s^2 + Y_m^2 \text{ and } \Delta_{off} = \left| \frac{L}{2} - x_f \right| \quad (2)$$

Y_o = the characteristic admittance of the patch,

$\gamma = \alpha + j\beta$ is the complex propagation constant,

where α is the attenuation constant (the copper and dielectric loss is accounted by α) Y_s, the self-admittance, and Y_m, the mutual admittance, are determined from[3] and coth(z) and csch(z) are the complex hyperbolic cotangent and cosecant functions

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of argument z, respectively. Δ_{off} will be termed as the feed offset relative to the center of the rectangular microstrip patch.

The resonant length L_{mn} is given by

$$L_{mn} = \frac{1}{\sqrt{\left[\frac{m}{L + 2\Delta L}\right]^2 + \left[\frac{n}{W + 2\Delta W}\right]^2}} \quad (3)$$

where m and n are integers and $\Delta L, \Delta W$ are two line extension terms accounting fringing effects.

Thus the resonant length L_{mn} also must have line extension term, and so, it is defined in terms of an equivalent physical length L_{eq} as:

$$L_{mn} = L_{eq} + 2\Delta L_{eq} \quad (4)$$

from Eqn (3), we have

$$L_{mn} = \frac{LW + 2(L\,\Delta W + W\,\Delta L) + 4\,\Delta L\,\Delta W}{\sqrt{m^2\,[W + 2\,\Delta W]^2 + n^2\,[L + 2\Delta L]^2}} \quad (5)$$

Thus, to get an expression for L_{eq} and also to determine an equivalent width W_{eq} (which is also a function of line extension term ΔL_{eq}), the following boundary conditions have to be met:

(a) When $m = 1$ and $n = 0$, then $L_{mn} = L + 2\,\Delta L$, so $L_{eq} = L$ and $W_{eq} = W$

(b) When $m = 0$ and $n = 1$, then $L_{mn} = W + 2\,\Delta W$, so $L_{eq} = W$ and $W_{eq} = L$

(c) The Eqns (4) and (5) should be equal for other conditions.

Two simple solutions can easily be derived so that the first two conditions are satisfied, i.e.,

$$L_{eq} = \frac{LW}{\sqrt{m^2 W^2 + n^2 L^2}} \quad (6)$$

$$W_{eq} = \frac{LW}{\sqrt{n^2 W^2 + m^2 L^2}} \quad (7)$$

or

$$L_{eq} = \frac{LW + 2[mL\,\Delta W + nW\,\Delta L]}{\sqrt{m^2\,[W + 2\,\Delta W]^2 + n^2\,[L + 2\Delta L]^2}} \quad (8)$$

$$W_{eq} = \frac{LW + 2[nL\,\Delta W + mW\,\Delta L]}{\sqrt{n^2\,[W + 2\,\Delta W]^2 + m^2\,[L + 2\Delta L]^2}} \quad (9)$$

The relations for equivalent feed position for fundamental modes (TM_{10} and TM_{01}), are given by[4]

$$\Delta\,x_{off} = \left|\frac{L}{2} - x_f\right|, \text{for } TM_{10} \quad (10)$$

$$\Delta\,y_{off} = \left|\frac{W}{2} - y_f\right|, \text{for } TM_{01} \quad (11)$$

Thus, for non-zero modes, the equivalent feed offset terms can be written in a manner similar to that of L_{eq} and W_{eq}

$$\Delta_{eq\,off} = \frac{\Delta\,x_{off}\,\Delta\,y_{off}}{\sqrt{m^2\,\Delta\,x_{off}^2 + n^2\,\Delta\,y_{off}^2}} \quad (12)$$

Another possibility used to determine equivalent offset term ($\Delta_{eq\,off}$) from an equivalent feed position Z_{eq} is,

$$\Delta_{eq\,off} = \left[\frac{L_{eq}}{2} - Z_{eq}\right] \quad (13)$$

where

$$Z_{eq} = \frac{x_f\,y_f}{\sqrt{m^2\,x_f^2 + n^2\,y_f^2}} \quad (14)$$

where L_{eq} is the equivalent mode length.

From these two relations for the determination of equivalent offset term, it has been found that Eqn (12) gives better results than Eqn (13)[4].

Then, the total input admittance of a rectangular patch is

$$Y_{patch} = \left[\frac{1}{Y_{10}} + \frac{1}{Y_{01}} + \frac{1}{Y_{11}} + \frac{1}{Y_{21}} + \frac{1}{Y_{12}}\right]^{-1} \quad (15)$$

where

$Y_{mn} =$

$$2Y_0 \frac{Y_0^2 = Y_t^2 + 2Y_s\,Y_0 \coth\,(L_{eq}) - 2Y_m Y_0\,\mathrm{csch}\,(\gamma L_{eq})}{(Y_0^2 + Y_t^2)\coth(\gamma L_{eq}) + (Y_0^2 - Y_t^2)\cosh(2\gamma\Delta_{eq})\mathrm{csch}(\gamma l_{eq}) + 2Y_s Y_0} \quad (16)$$

Now for a microstrip antenna, fed coaxially, the inductive reactance of the feed (X_L) is added up in series with the patch impedance. Thus the input impedance Z_{patch} would be

$$Z_{patch} = \frac{1}{Y_{patch}} + j\,X_{feed} \quad (17)$$

The probe reactance is given by[4]

$$X_{feed} = \frac{\eta_0\,w\,h}{2\,\pi\,c} \ln\left[\frac{4\,C}{\zeta\,\omega\,d\,\sqrt{\varepsilon_r}}\right] \quad (18)$$

where $\zeta = 1.781072$.

2.2 Cavity Model

It is of interest to investigate the variation of input impedance of a rectangular microstrip antenna by cavity model when TM_{11} mode is excited. Modelling the rectangular patch[5,6], bounded at its top and bottom by electric walls and on its sides by magnetic walls, the z-directed electric field inside the cavity can be written as

$$E_z = E_0 \cos\frac{m\pi x}{a}\cos\frac{n\pi x}{b} \tag{19}$$

The resonance frequencies for TM_{11} mode can be computed from the generalised formula[7].

$$f_r = \frac{v_0}{2\varepsilon_{eff}}\sqrt{\left[\frac{m}{a+2\Delta a(b)}\right]^2 + \left[\frac{n}{b+2\Delta b(a)}\right]^2} \tag{20}$$

where,

$$\varepsilon_{eff} = \begin{cases} \varepsilon_{eff}(a) & \text{for } (0,n) \text{ modes} \\ \varepsilon_{eff}(b) & \text{for } (m,0) \text{ modes} \\ \dfrac{\varepsilon_{eff}(a)\,\varepsilon_{eff}(b)}{\varepsilon_r} & \text{for } (m \neq 0\ n \neq 0) \text{ modes} \end{cases}$$

v_0 is the velocity of light in free space. The expression for effective dielectric perittivity ε_{eff} and line extensions $\Delta a(b)$ and $\Delta b(a)$ can be found[8].

Far-field components of a rectangular microstrip antenna may be calculated[9] considering the radiator as a set of four slots of equal width, c. The slot-width may be considered as equal to the substrate thickness, h. Then far-field components for any mode become

$$E_\theta = \left[E_x(\zeta,\eta)\cos\varphi + E_y(\zeta,\eta)\sin\varphi\right]R \tag{21}$$

$$E_\varphi = \left[E_x(\zeta,\eta)\sin\varphi + E_y(\zeta,\eta)\cos\varphi\cos\theta\right]R \tag{22}$$

where

$$R = \frac{j K_o \exp(-j K_o r)}{2\pi r}$$

$$\zeta = k_o \sin\theta\cos\varphi$$

$$\eta = k_o \sin\theta\sin\varphi,\ k_o = \frac{2\pi}{\lambda}$$

$$E_x(\zeta,\eta) = \left[(-1-(-1)^m)\, j \sin\left(\zeta\frac{a'}{2}\right) + (1-(-1)^m)\cos\left(\zeta\frac{a'}{2}\right)\right].h.E.\frac{b'}{2}.\mathrm{sinc}\left(\zeta\frac{c}{2}\right)j^n \left[\mathrm{sinc}\left(\eta.\frac{b'}{2}+\frac{n\pi}{2}\right) + (-1)^n.\mathrm{sinc}\left(\eta\frac{b'}{2}-\frac{n\pi}{2}\right)\right]$$

$$E_y(\zeta,\eta) = \left[(-1-(-1)^n)\, j \sin\left(\eta\frac{b'}{2}\right) + (1-(-1)^n)\cos\left(\eta\frac{b'}{2}\right)\right].h.E.\frac{a'}{2}.\mathrm{sinc}\left(\eta\frac{c}{2}\right)j^m \left[\mathrm{sinc}\left(\zeta.\frac{a'}{2}+\frac{m\pi}{2}\right) + (-1)^m.\mathrm{sinc}\left(\zeta\frac{a'}{2}-\frac{m\pi}{2}\right)\right]$$

$\mathrm{sinc}(x) = \sin(x)/x$ and $b' = b+h,\ a' = a+h$

The input impedance of a rectangular microstrip antenna, when TM_{11} mode is excited, feeding at the point (x_o, y_o), is computed using the relation

$$Z_{in} = \frac{V^2}{P_d + P_c + P_r + 2j\omega(W_e - W_m)} \tag{23}$$

where

$$V = \frac{h E_o}{\sqrt{2}}\cos\left(\frac{m\pi x_o}{a}\right)\cos\left(\frac{n\pi y_o}{b}\right)$$

is the r.m.s. value of excitation voltage.

2.3 Variation of Input Impedance with Feed Position in Fundamental Mode

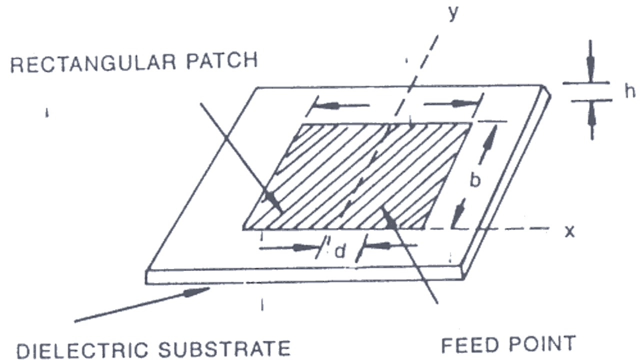

Figure 1. Rectangular microstrip antenna.

The microstrip antenna configuration is shown in Fig. 1. In the fundamental mode (TM_{10}), the impedance of a rectangular microstrip antenna is independent of feed position along the radiating edges (along y-axis), and this is due to constant vertical electric field along these edges. But for this mode, the impedance varies along the non-radiating edges (along x-axis) because there is a sinusoidal variation in the vertical electric field along these edges. The impedance variation of a rectangular patch antenna along the non-radiating edges is already reported[5,6,11] which show that the impedance increases from zero to a certain maximum value, as the feed point is shifted from the centre to the edge of the patch antenna. Though input impedance changes with feed position along non-radiating edges, the theoretical and experimental results of Fig. 2 reveal that the impedance band width cannot be improved in TM_{10} mode. The design parameters for Fig. 2 were $\varepsilon_r = 2.55$, h = 0.3048 cm., a = 3.65 cm., b = 2.4 cm.

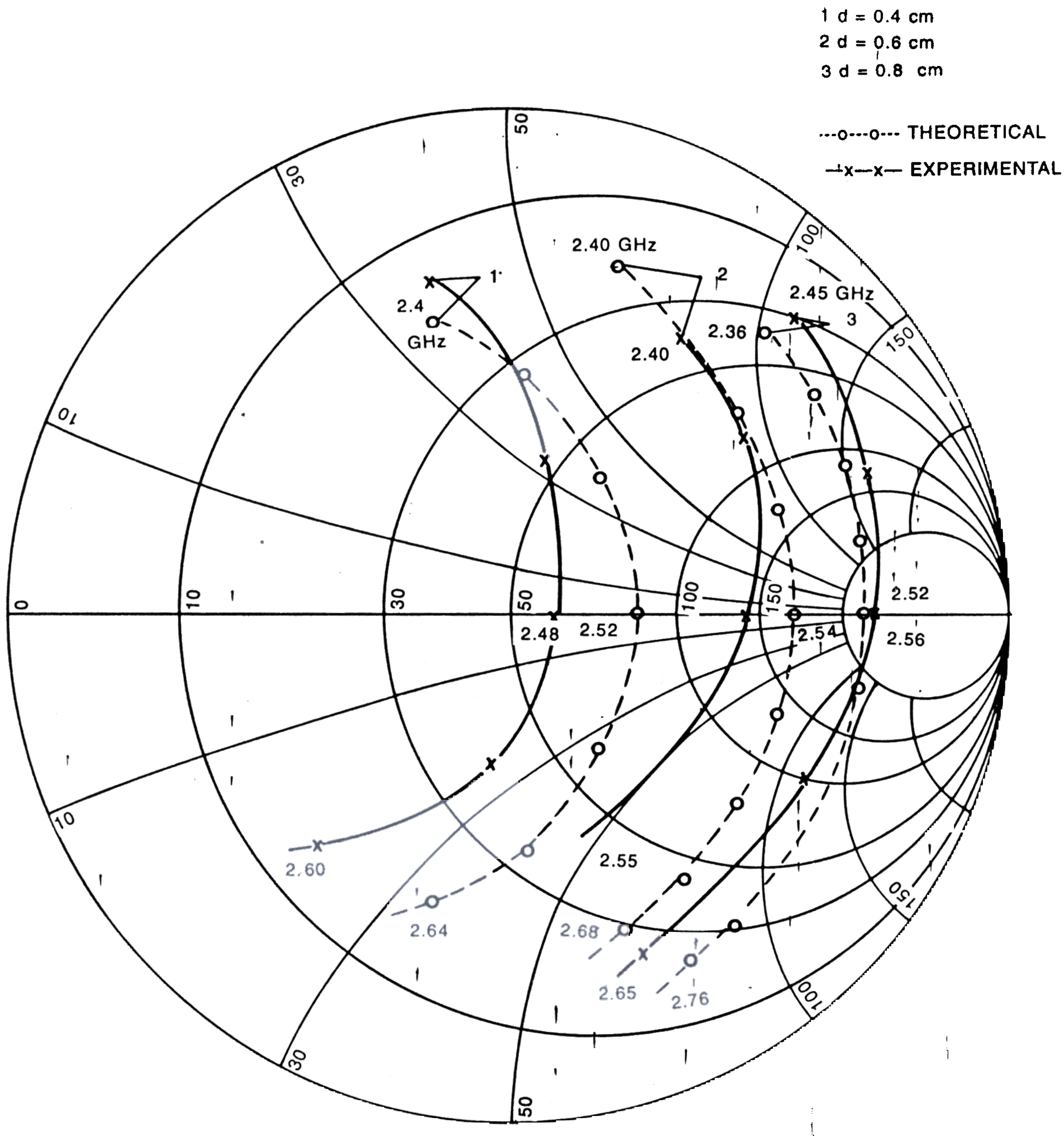

Figure 2. Variation of input impedance with feed location at TM_{10} mode.

2.4 Computed & Measured Date in TM_{11} Mode

A few microstrip antennas were fabricated on PTFE substrate with relative dielectric constant of 2.55 and thickness of 0.3048 cm. The measured results of one of them is presented here. The antenna was fed by a coaxial SMA connector and then TM_{11} mode was excited. The input impedance of such an antenna around the resonant frequency was measured by an HP 8410B network analyser. The antenna configuration is shown in Fig. 1. The patch dimension was 3.65 cm × 2.4 cm. The measured resonance frequency was 4.61 Ghz. At TM_{11} mode, the feed location was varied along x-axis (which corresponds to the non-radiating edge at fundamental mode). The variation of resonance resistance with feed location at TM_{11} mode is shown in Fig. 3. The variation of

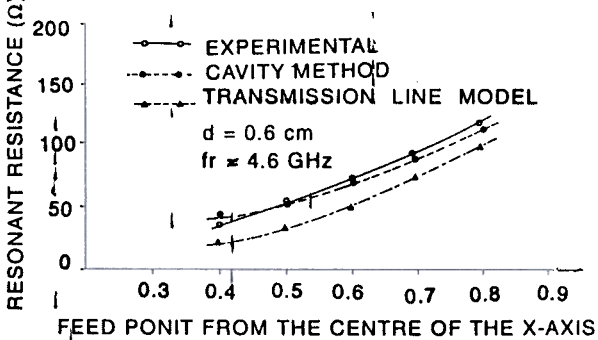

Figure 3. Variation of resonance resistance with feed location along x-direction at TM$_{11}$ mode.

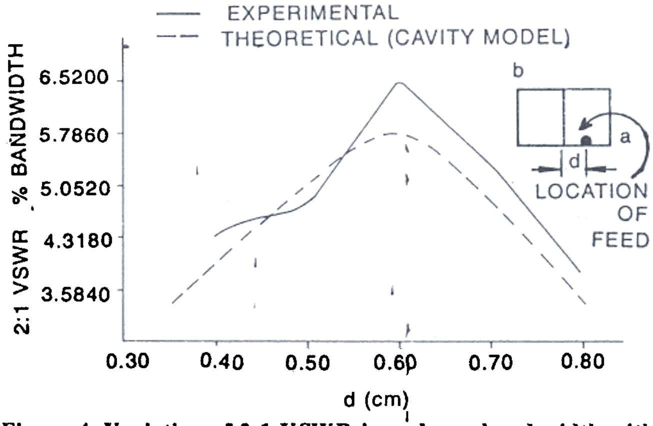

Figure 4. Variation of 2:1 VSWR impedance band width with feed location at TM$_{11}$ mode.

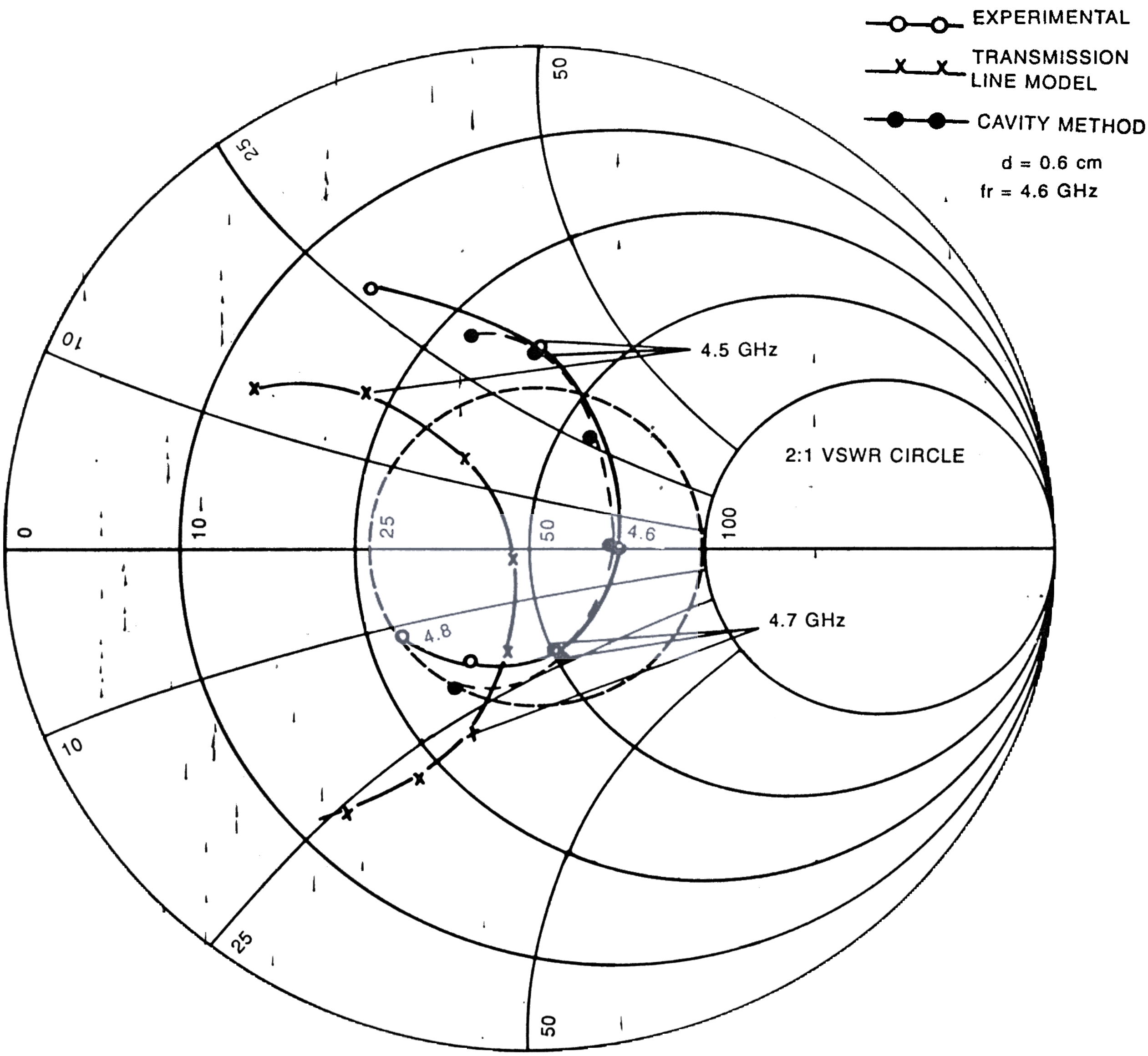

Figure 5. Impedance pattern at TM$_{11}$ mode (feed position d = 0.6 cm, where maximum band width is obtained).

2:1 VSWR impedance band width with feed location is plotted in Fig. 4. Maximum band width of about 6.5 per cent was achieved when feed location, d = 0.6 cm. The measured impedance plot using 2:1 VSWR circle on Smith chart is compared with the theoretical models, as in Fig. 5. Now the feed location is shifted along y-axis (which corresponds to the radiating edge at fundamental mode) and maximum 2:1 VSWR band width of 3.4 per cent was achieved at TM_{11} mode.

Measured conical radiation patterns at TM_{11} mode are compared with theoretical patterns in Fig. 6(a) and Fig. 6(b), for $\varphi = 0^{\circ}$ and $\varphi = 90^{\circ}$ planes, respectively. The ripples in the measured radiation patterns are due to diffraction from finite ground plane (7.1 cm x 4.8 cm).

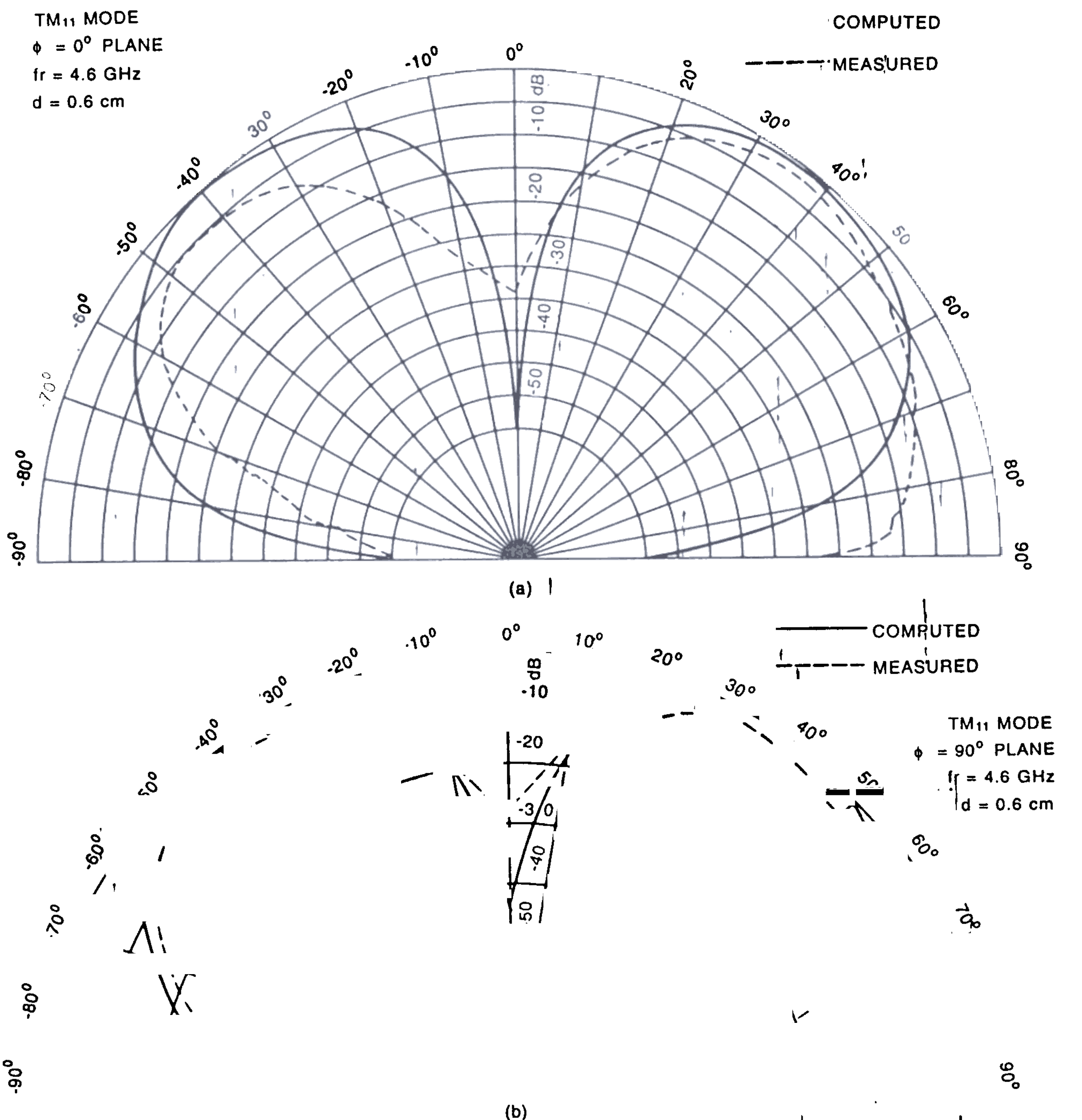

Figure 6(a). Conical radiation pattern at TM_{11} mode ($\varphi = 0^{\circ}$ plane), (b) Conical radiation pattern at TM_{11} mode ($\varphi = 90^{\circ}$ plane).

3. CONCLUSION

A microstrip antenna with higher order mode excitation can be used as a conformal mobile or satellite antenna. The characteristics of a rectangular microstrip antenna in respect of 2:1 VSWR band width at first higher order mode (TM_{11}), are reported here. Both the theory and measurement reveal that by properly choosing the feed location, higher 2:1 VSWR impedance band width can be achieved at TM_{11} mode, as compared to the fundamental mode.

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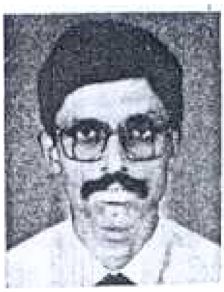
Contributors



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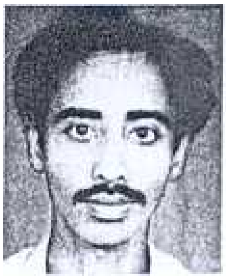

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