Overview of Microstrip Antennas

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Overview of Microstrip Antennas

Also called “patch antennas”

- One of the most useful antennas at microwave frequencies (f > 1 GHz).
- It consists of a metal “patch” on top of a grounded dielectric substrate.
- The patch may be in a variety of shapes, but rectangular and circular are the most common.

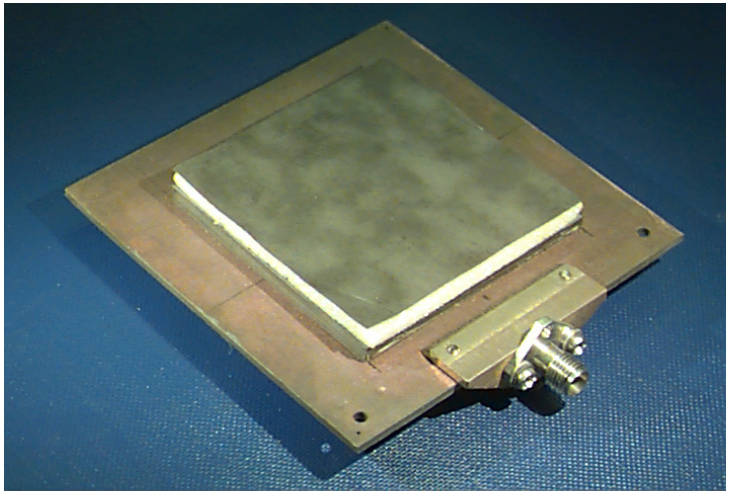
History of Microstrip Antennas

- Invented by Bob Munson in 1972.
- Became popular starting in the 1970s.

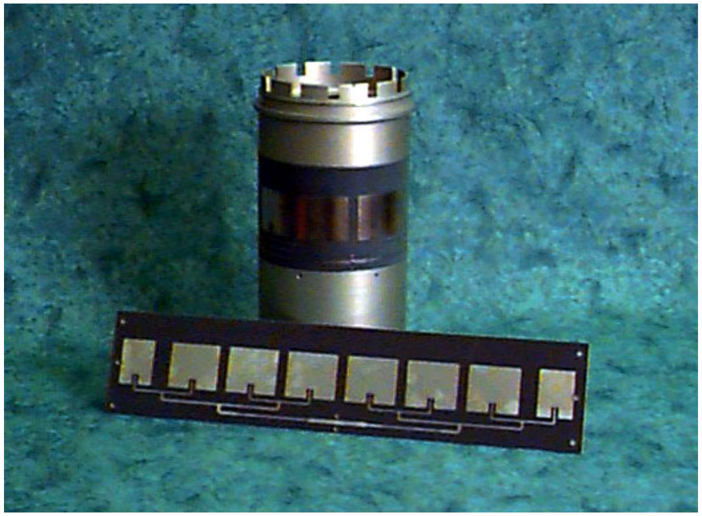
R. E. Munson, "Microstrip Phased Array Antennas," *Proc. of Twenty-Second Symp. on USAF Antenna Research and Development Program,* October 1972.

R. E. Munson, "Conformal Microstrip Antennas and Microstrip Phased Arrays," *IEEE Trans. Antennas Propagat.*, vol. AP-22, no. 1 (January 1974): 74–78.

Typical Applications



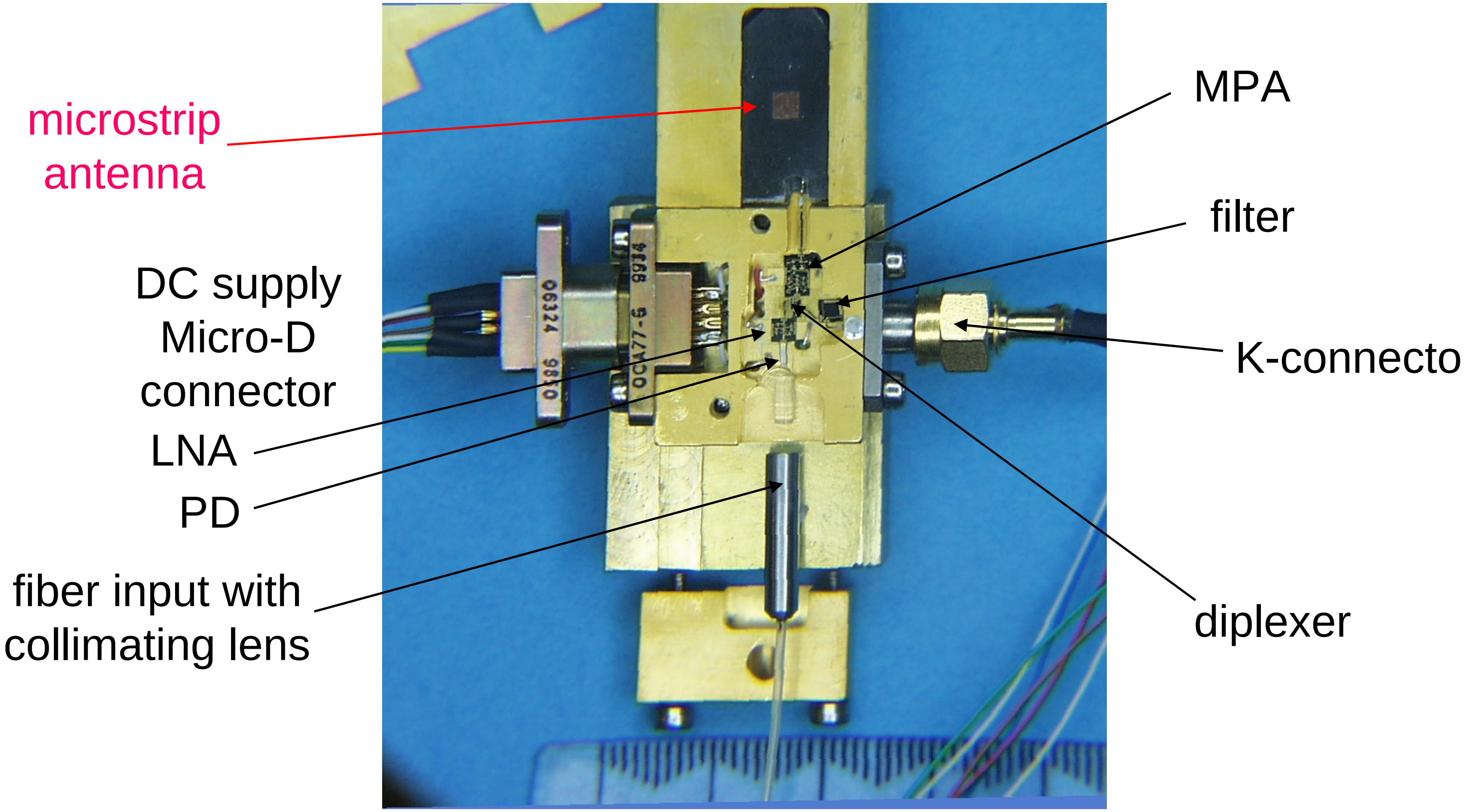
single element



array

(Photos courtesy of Dr. Rodney B. Waterhouse)

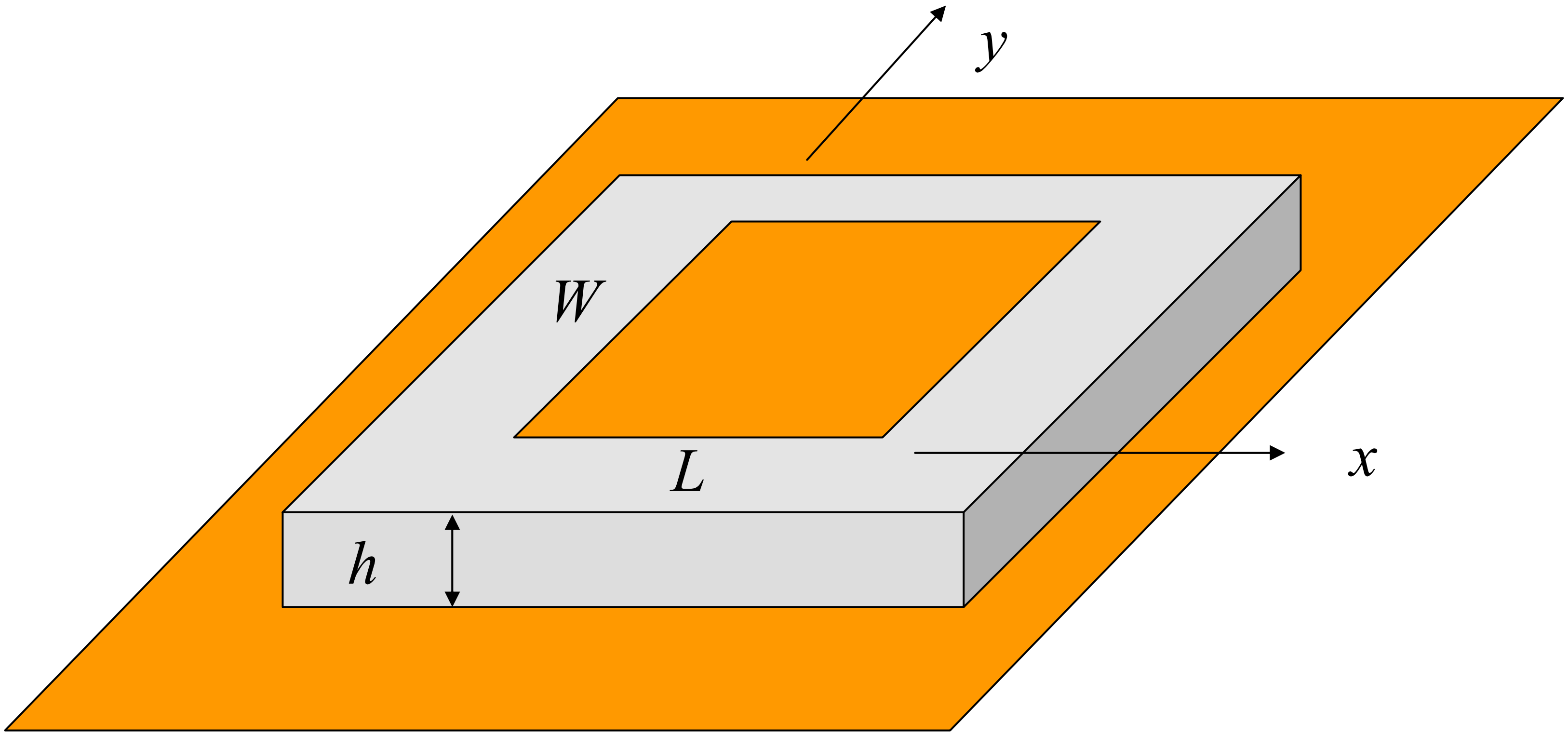
Typical Applications (cont.)



Microstrip Antenna Integrated into a System: HIC Antenna Base-Station for 28-43 GHz

(Photo courtesy of Dr. Rodney B. Waterhouse)

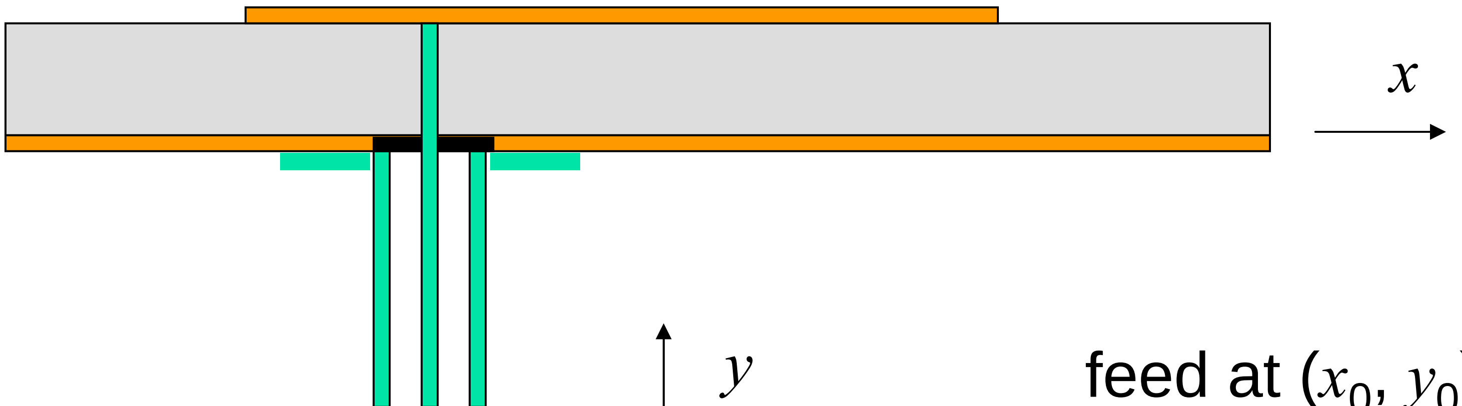
Geometry of Rectangular Patch

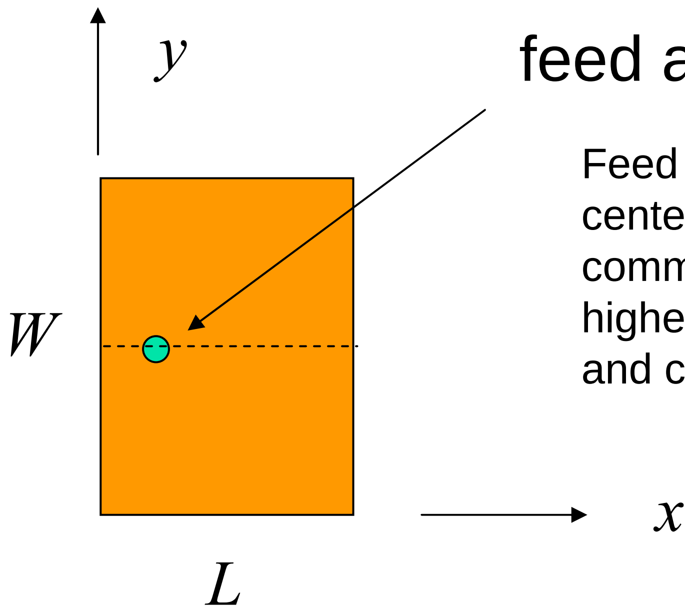


Note: L is the resonant dimension. The width W is usually chosen to be larger than L (to get higher bandwidth). However, usually $W < 2\,L$. $W = 1.5\,L$ is typical.

Geometry of Rectangular Patch (cont.)

view showing coaxial feed





Feed along the centerline is the most common (minimizes higher-order modes and cross-pol)

Advantages of Microstrip Antennas

- Low profile (can even be “conformal”).
- Easy to fabricate (use etching and phototlithography).
- Easy to feed (coaxial cable, microstrip line, etc.) .
- Easy to use in an array or incorporate with other microstrip circuit elements.
- Patterns are somewhat hemispherical, with a moderate directivity (about 6-8 dB is typical).

Disadvantages of Microstrip Antennas

- Low bandwidth (but can be improved by a variety of techniques). Bandwidths of a few percent are typical.
- Efficiency may be lower than with other antennas. Efficiency is limited by conductor and dielectric losses*, and by surface-wave loss**.

* Conductor and dielectric losses become more severe for thinner substrates.

** Surface-wave losses become more severe for thicker substrates (unless air or foam is used).

Basic Principles of Operation

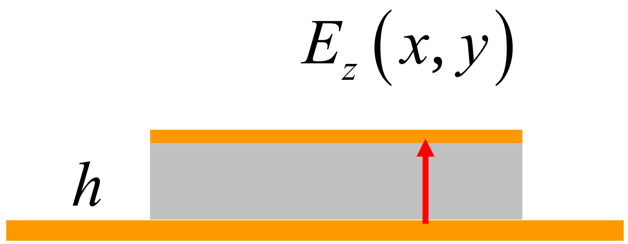
- The patch acts approximately as a resonant cavity (short circuit walls on top and bottom, open-circuit walls on the sides).
- In a cavity, only certain modes are allowed to exist, at different resonant frequencies.
- If the antenna is excited at a resonant frequency, a strong field is set up inside the cavity, and a strong current on the (bottom) surface of the patch. This produces significant radiation (a good antenna).

Thin Substrate Approximation

On patch and ground plane, $\underline{E}_t = \underline{0} \Rightarrow \underline{E} = \underline{\hat{z}}\, E_z(x, y)$

Inside the patch cavity, because of the thin substrate, the electric field vector is approximately independent of z.

Hence $$\underline{E} \approx \underline{\hat{z}}\, E_z(x, y)$$



Thin Substrate Approximation

Magnetic field inside patch cavity:

$$\underline{H} = -\frac{1}{j\omega\mu}\nabla\times\underline{E}$$

$$= -\frac{1}{j\omega\mu}\nabla\times\left(\underline{\hat{z}}E_z\left(x,y\right)\right)$$

$$= -\frac{1}{j\omega\mu}\left(-\underline{\hat{z}}\times\nabla E_z\left(x,y\right)\right)$$

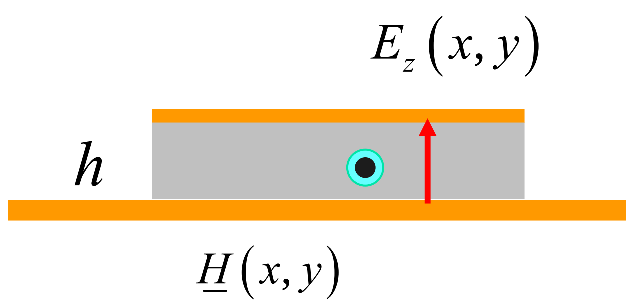
Hence

$$\underline{H}\left(x,y\right) = \frac{1}{j\omega\mu}\left(\underline{\hat{z}}\times\nabla E_z\left(x,y\right)\right)$$

Thin Substrate Approximation (cont.)

$$\underline{H}(x,y) = \frac{1}{j\omega\mu}\left(\underline{\hat{z}} \times \nabla E_z(x,y)\right)$$

Note: the magnetic field is purely horizontal.
(The mode is TM_z.)



Magnetic Wall Approximation

On edges of patch,

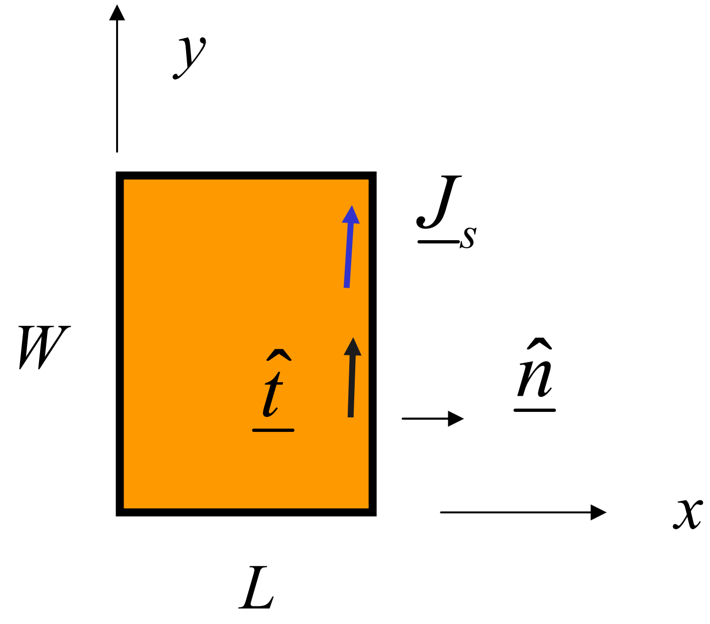
$$\underline{J}_s \cdot \underline{\hat{n}} = 0$$

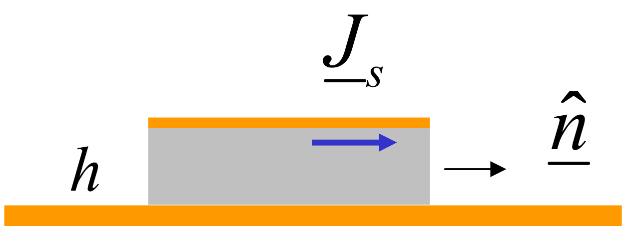
Also, on lower surface of patch conductor we have

$$\underline{J}_s = \left(-\underline{\hat{z}} \times \underline{H}\right)$$

Hence,

$$\underline{H}_t = \underline{0}$$

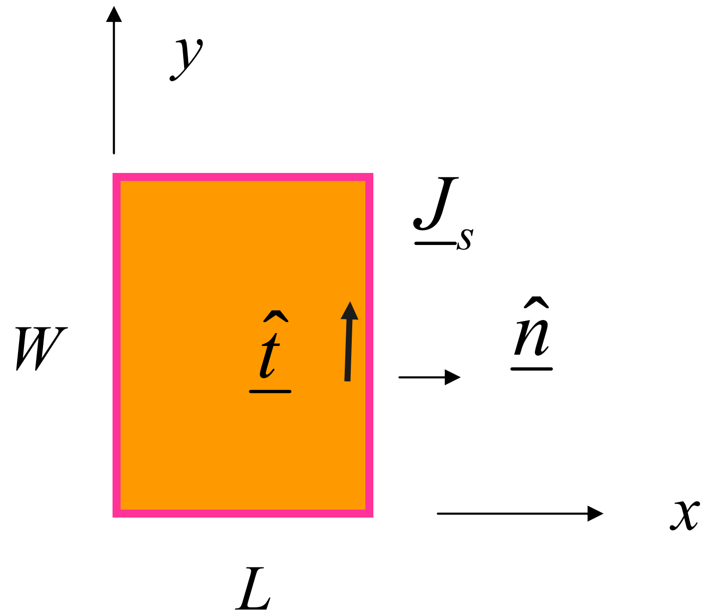




Magnetic Wall Approximation (cont.)

Since the magnetic field is approximately independent of z, we have an approximate PMC condition on the edge.

$$\underline{H}_t = \underline{0} \quad \text{(PMC)}$$



PMC

Magnetic Wall Approximation (cont.)

$$\underline{\hat{n}} \times \underline{H}(x, y) = \underline{0}$$

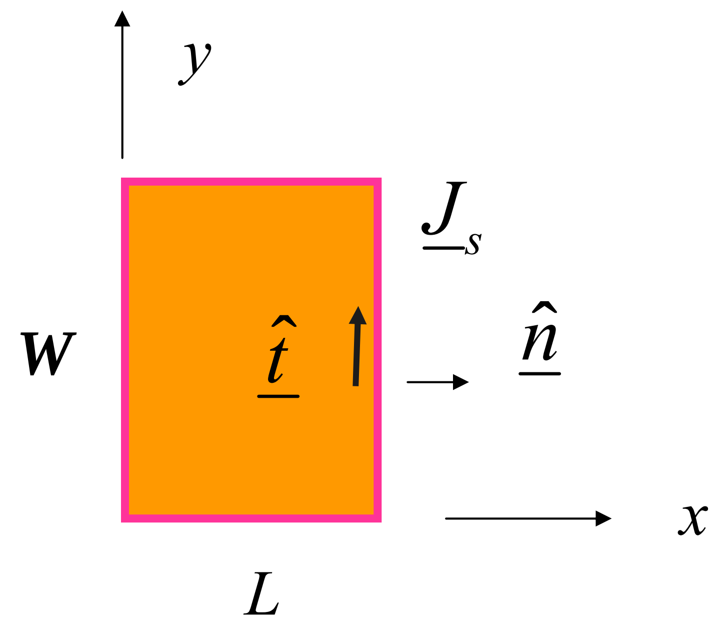
$$\underline{H}(x, y) = \frac{1}{j\omega\mu}\left(\underline{\hat{z}} \times \nabla E_z(x, y)\right)$$

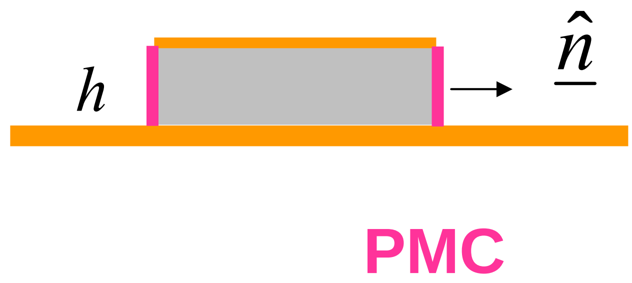
Hence,

$$\underline{\hat{n}} \times \left(\underline{\hat{z}} \times \nabla E_z(x, y)\right) = \underline{0}$$

$$\Rightarrow \quad \underline{\hat{z}}\left(\underline{\hat{n}} \cdot \nabla E_z(x, y)\right) = \underline{0}$$

$$\frac{\partial E_z}{\partial n} = 0$$





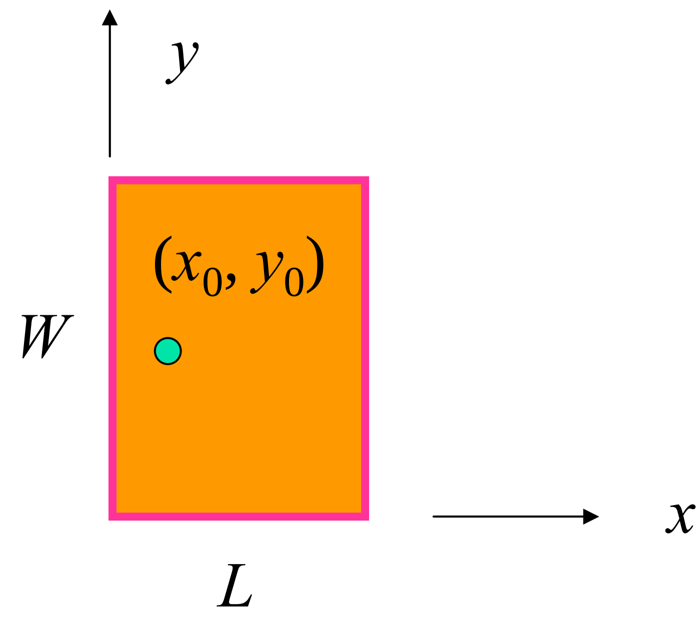
Resonance Frequencies

$$\nabla^2 E_z + k^2 E_z = 0$$

From separation of variables:

$$E_z = \cos\left(\frac{m\pi x}{L}\right)\cos\left(\frac{n\pi y}{W}\right)$$

(TM_{mn} mode)

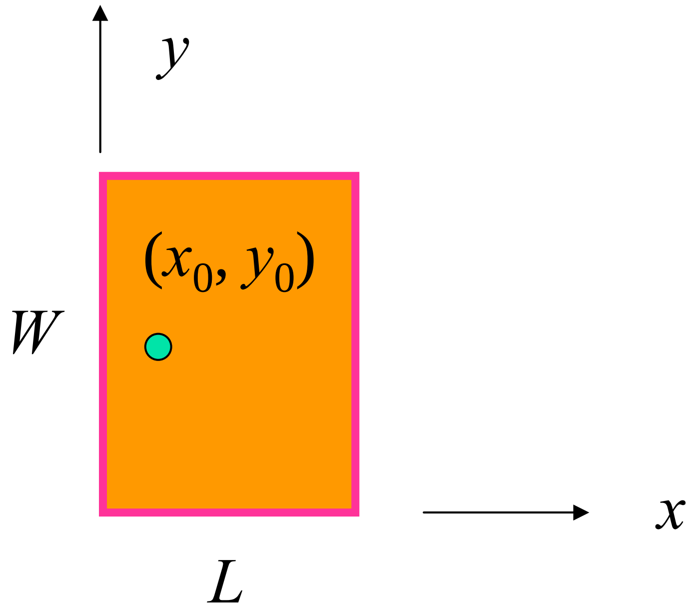


$$\left[-\left(\frac{m\pi}{L}\right)^2 - \left(\frac{n\pi}{W}\right)^2 + k^2\right] E_z = 0$$

Hence
$$\left[-\left(\frac{m\pi}{L}\right)^2 - \left(\frac{n\pi}{W}\right)^2 + k^2\right] = 0$$

Resonance Frequencies (cont.)

$$k^2 = \left(\frac{m\pi}{L}\right)^2 + \left(\frac{n\pi}{W}\right)^2$$



Recall that

$$k = \omega\sqrt{\mu_0\varepsilon_0}\sqrt{\varepsilon_r}$$

$$\omega = 2\pi f$$

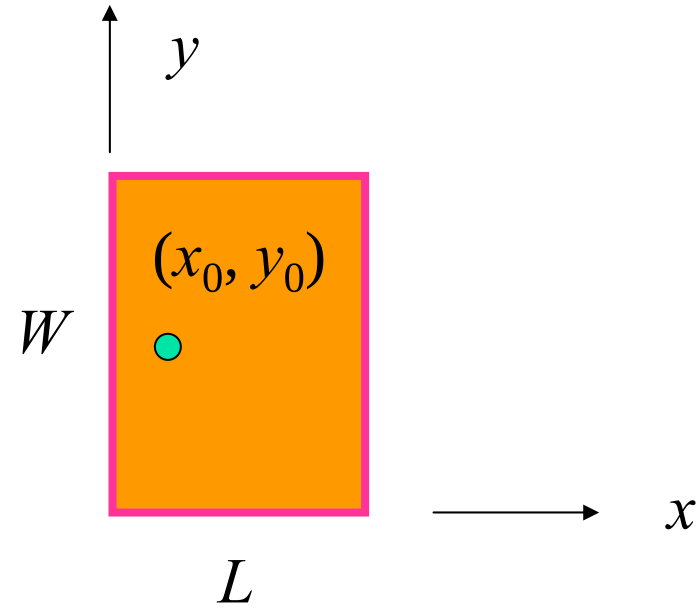
Hence

$$f = \frac{c}{2\pi\sqrt{\varepsilon_r}}\sqrt{\left(\frac{m\pi}{L}\right)^2 + \left(\frac{n\pi}{W}\right)^2} \qquad c = 1/\sqrt{\mu_0\varepsilon_0}$$

Resonance Frequencies (cont.)

Hence $f = f_{mn}$

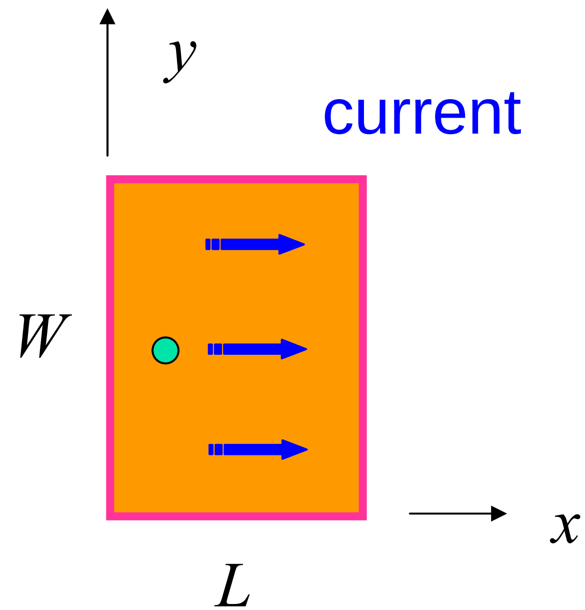
(resonance frequency of
(m, n) mode)



$$f_{mn} = \frac{c}{2\pi\sqrt{\varepsilon_r}} \sqrt{\left(\frac{m\pi}{L}\right)^2 + \left(\frac{n\pi}{W}\right)^2}$$

(1,0) Mode

This mode is usually used because the radiation pattern has a broadside beam.



$$E_z = \cos\left(\frac{\pi x}{L}\right)$$

$$f_{10} = \frac{c}{2\sqrt{\varepsilon_r}}\left(\frac{1}{L}\right)$$

$$\underline{J}_s = \underline{\hat{x}}\left(\frac{-1}{j\omega\mu_0}\right)\left(\frac{\pi}{L}\right)\sin\left(\frac{\pi x}{L}\right)$$

This mode acts as a wide microstrip line (width W) that has a resonant length of 0.5 guided wavelengths in the x direction.

Basic Properties of Microstrip Antennas

Resonance Frequency

The resonance frequency is controlled by the patch length L and the substrate permittivity.

Approximately,

$$f_{10} = \frac{c}{\sqrt{\varepsilon_r}}\left(\frac{1}{2L}\right)$$

Note: this is equivalent to saying that the length L is one-half of a wavelength in the dielectric:

$$kL = \pi \quad \Longrightarrow \quad L = \lambda_d / 2 = \frac{\lambda_0 / 2}{\sqrt{\varepsilon_r}}$$

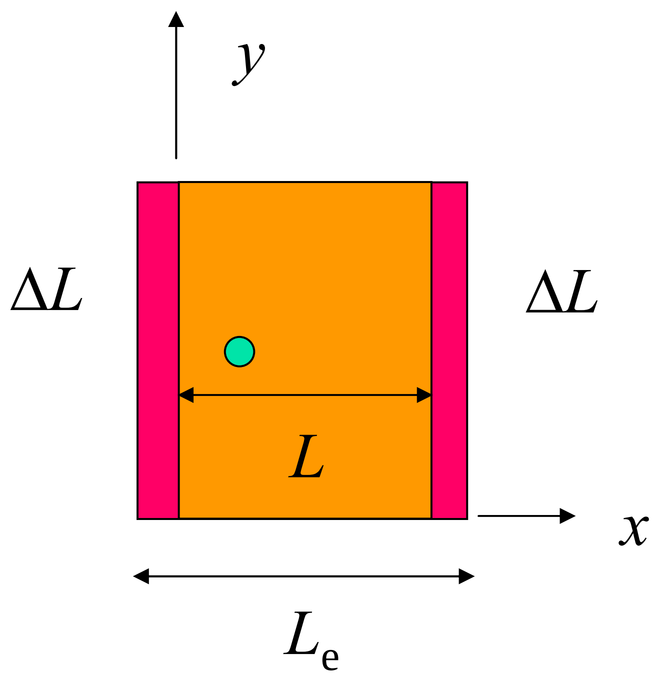
Note: a higher substrate permittivity allows for a smaller antenna (miniaturization) – but lower bandwidth.

Resonance Frequency (cont.)

The calculation can be improved by adding a "fringing length extension" ΔL to each edge of the patch to get an "effective length" L_e .

$$L_e = L + 2\Delta L$$

$$f_{10} = \frac{c}{2\sqrt{\varepsilon_r}}\left(\frac{1}{L_e}\right)$$



Resonance Frequency (cont.)

Hammerstad formula:

$$\Delta L / h = 0.412 \left[\frac{\left(\varepsilon_r^{eff} + 0.3\right)\left(\frac{W}{h} + 0.264\right)}{\left(\varepsilon_r^{eff} - 0.258\right)\left(\frac{W}{h} + 0.8\right)} \right]$$

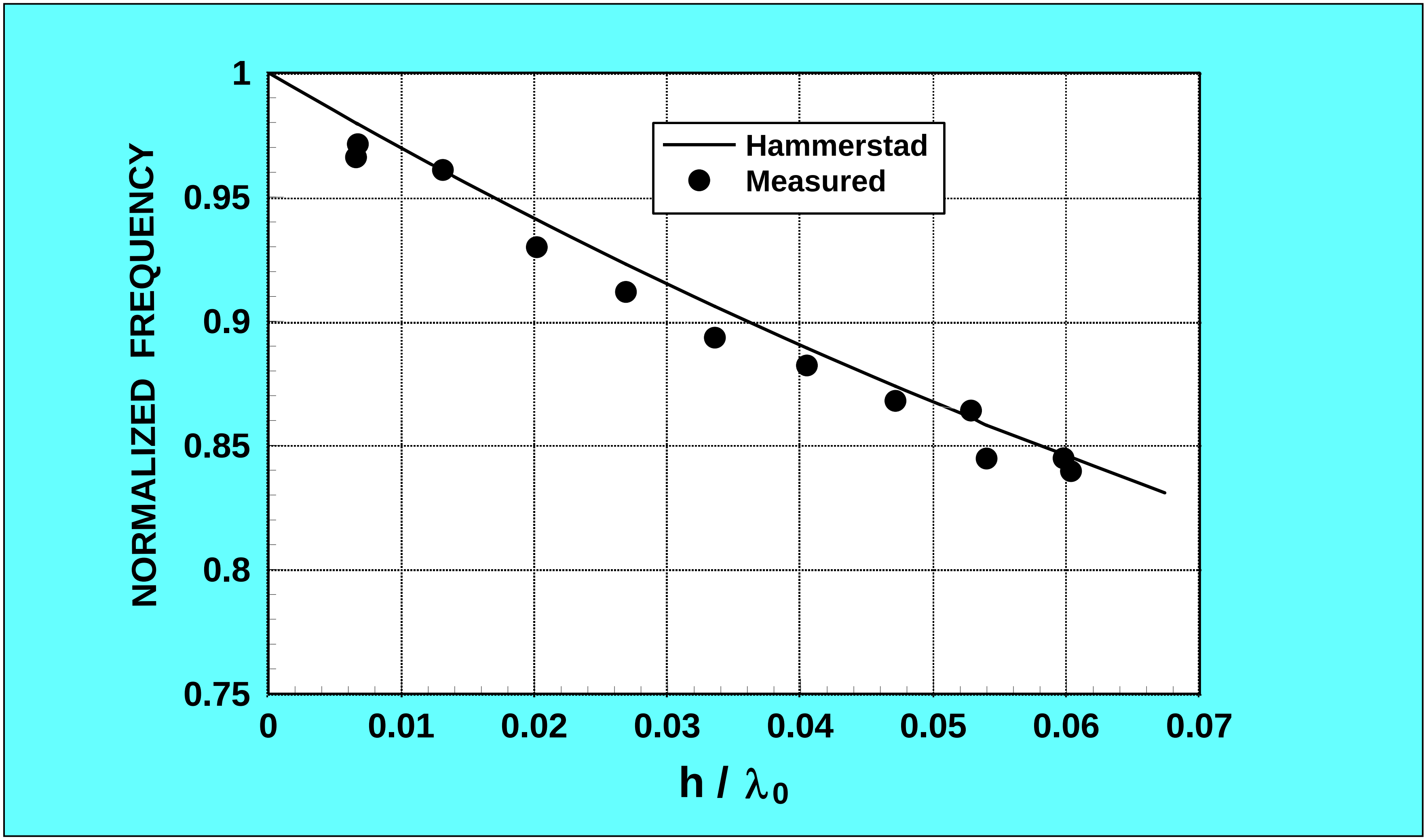
$$\varepsilon_r^{eff} = \frac{\varepsilon_r + 1}{2} + \left(\frac{\varepsilon_r - 1}{2}\right)\left[1 + 12\left(\frac{h}{W}\right)\right]^{-1/2}$$

Resonance Frequency (cont.)

Note: $\Delta L \approx 0.5\ h$

This is a good “rule of thumb.”

Results: resonance frequency



$\varepsilon_r = 2.2$

$W/L = 1.5$

The resonance frequency has been normalized by the zero-order value (without fringing):

$$f_N = f / f_0$$

Basic Properties of Microstrip Antennas

Bandwidth: substrate effects

- The bandwidth is directly proportional to substrate thickness h.
- However, if h is greater than about 0.05 λ_0 , the probe inductance becomes large enough so that matching is difficult.
- The bandwidth is inversely proportional to ε_r (a foam substrate gives a high bandwidth).

Basic Properties of Microstrip Antennas

Bandwidth: patch geometry

- The bandwidth is directly proportional to the width W.

Normally $W < 2L$ because of geometry constraints:

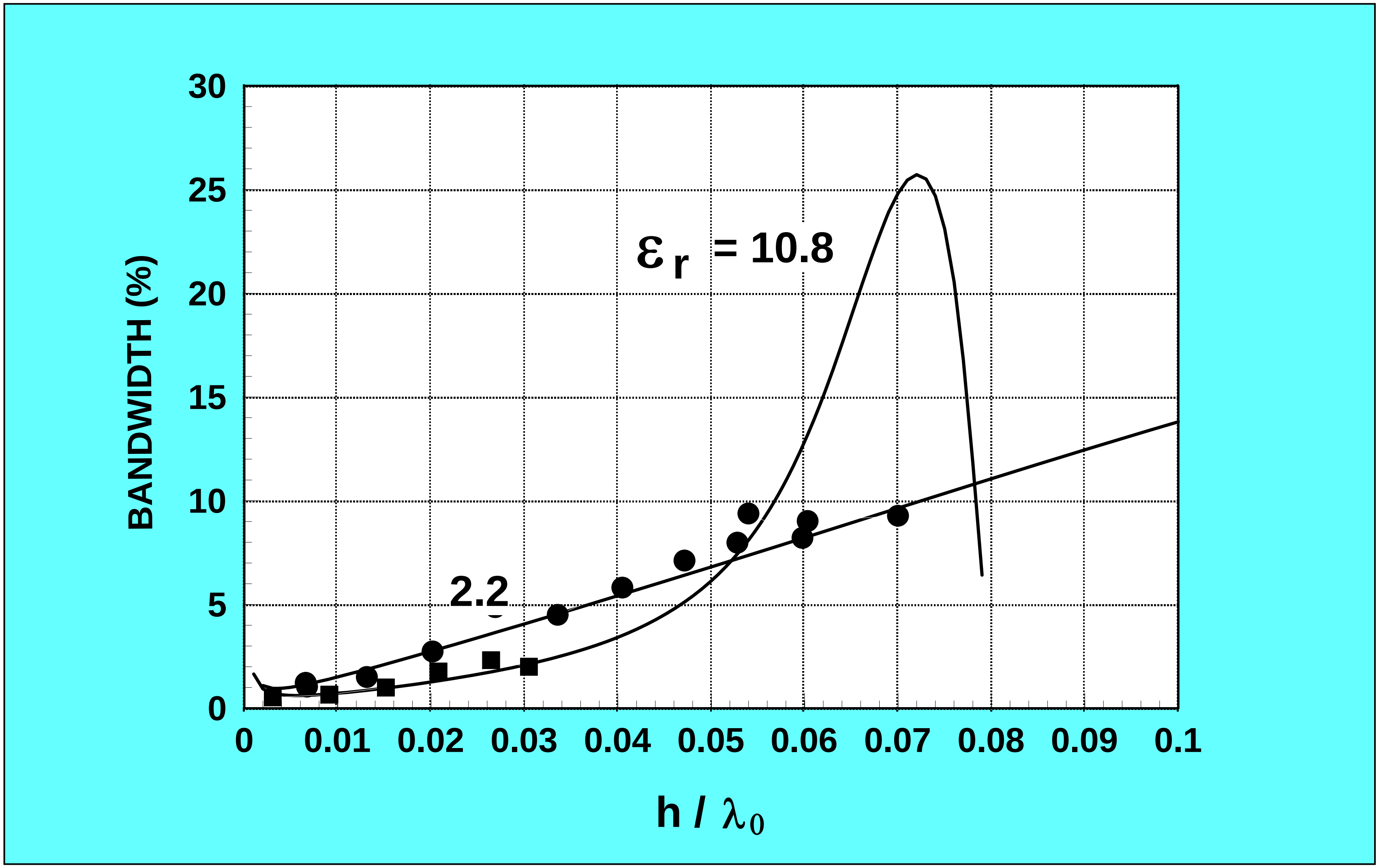
$W = 1.5\,L$ is typical.

Basic Properties of Microstrip Antennas

Bandwidth: typical results

- For a typical substrate thickness (h / λ_0 = 0.02), and a typical substrate permittivity (ε_r = 2.2) the bandwidth is about 3%.
- By using a thick foam substrate, bandwidth of about 10% can be achieved.
- By using special feeding techniques (aperture coupling) and stacked patches, bandwidth of over 50% have been achieved.

Results: bandwidth



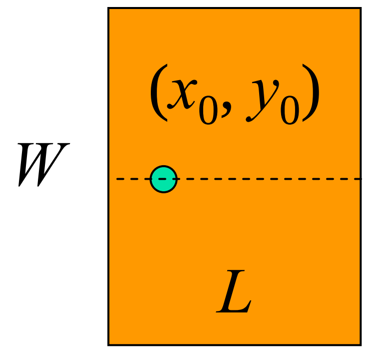
The discrete data points are measured values. The solid curves are from a CAD formula.

$$\varepsilon_r = 2.2 \text{ or } 10.8 \qquad W/L = 1.5$$

Basic Properties of Microstrip Antennas

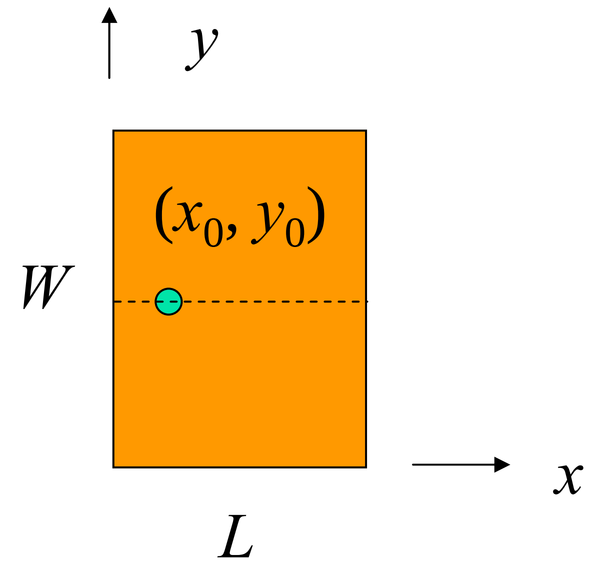
Resonant Input Resistance

- The resonant input resistance is almost independent of the substrate thickness h.
- The resonant input resistance is proportional to ε_r.
- The resonant input resistance is directly controlled by the location of the fed point. (maximum at edges $x = 0$ or $x = L$, zero at center of patch.



Resonant Input Resistance (cont.)

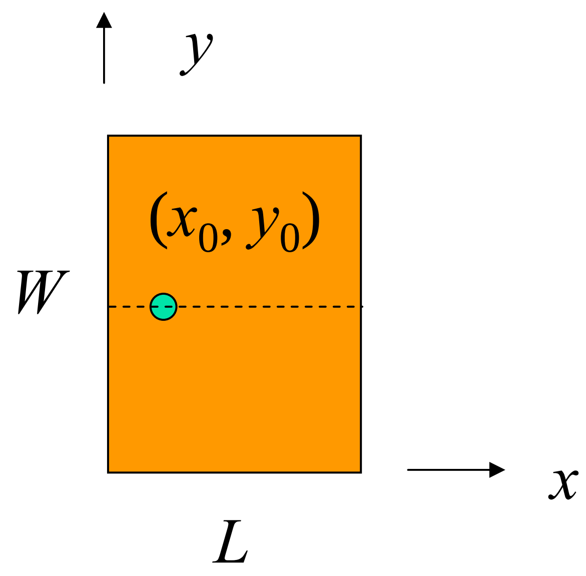
Note: patch is usually fed along the centerline ($y = W / 2$) to maintain symmetry and thus minimize excitation of undesirable modes.



Resonant Input Resistance (cont.)

For a given mode, it can be shown that the resonant input resistance is proportional to the square of the cavity-mode field at the feed point.

$$R_{in} \propto E_z^2 \left(x_0 , y_0 \right)$$



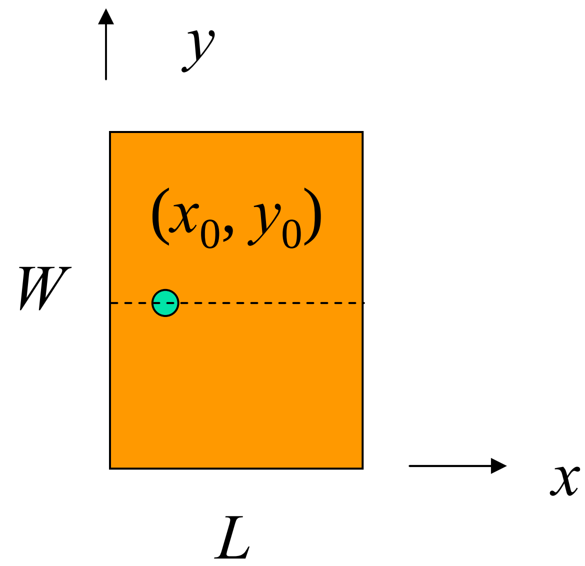
For (1,0) mode:

$$R_{in} \propto \cos^2 \left(\frac{\pi x_0}{L} \right)$$

Resonant Input Resistance (cont.)

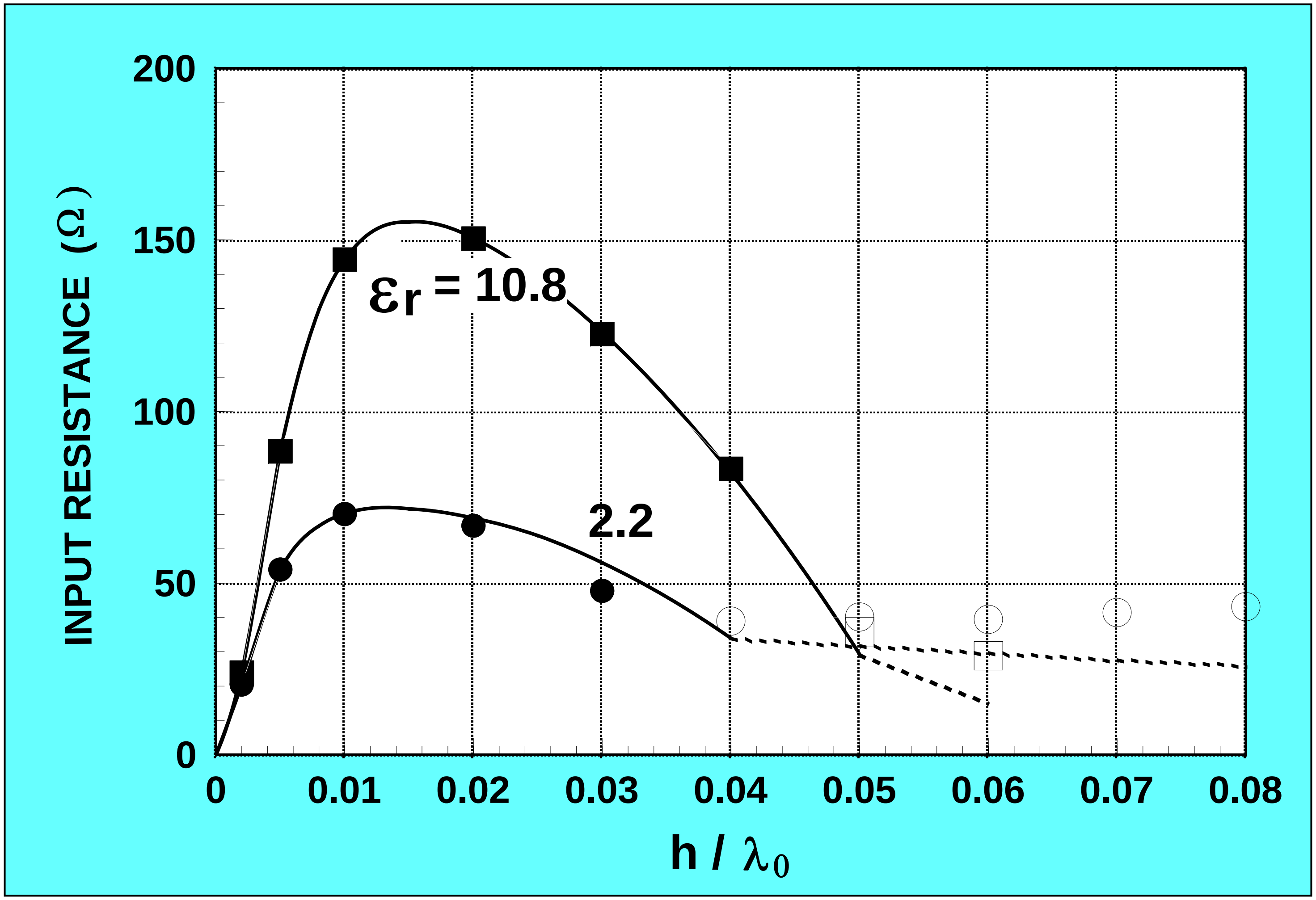
Hence, for (1,0) mode:

$$R_{in} = R_{edge}\cos^2\left(\frac{\pi x_0}{L}\right)$$



The value of R_{edge} depends strongly on the substrate permittivity. For a typical patch, it may be about 100-200 Ohms.

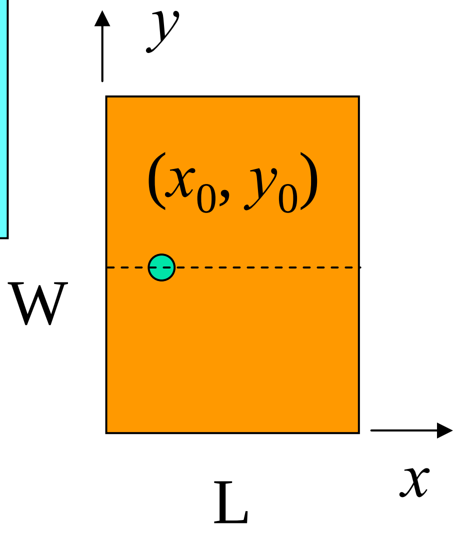
Results: resonant input resistance



The discrete data points are from a CAD formula.

$\varepsilon_r = 2.2$ or 10.8

$W/L = 1.5 \qquad x_0 = L/4,\ \ y_0 = W/2$



Basic Properties of Microstrip Antennas

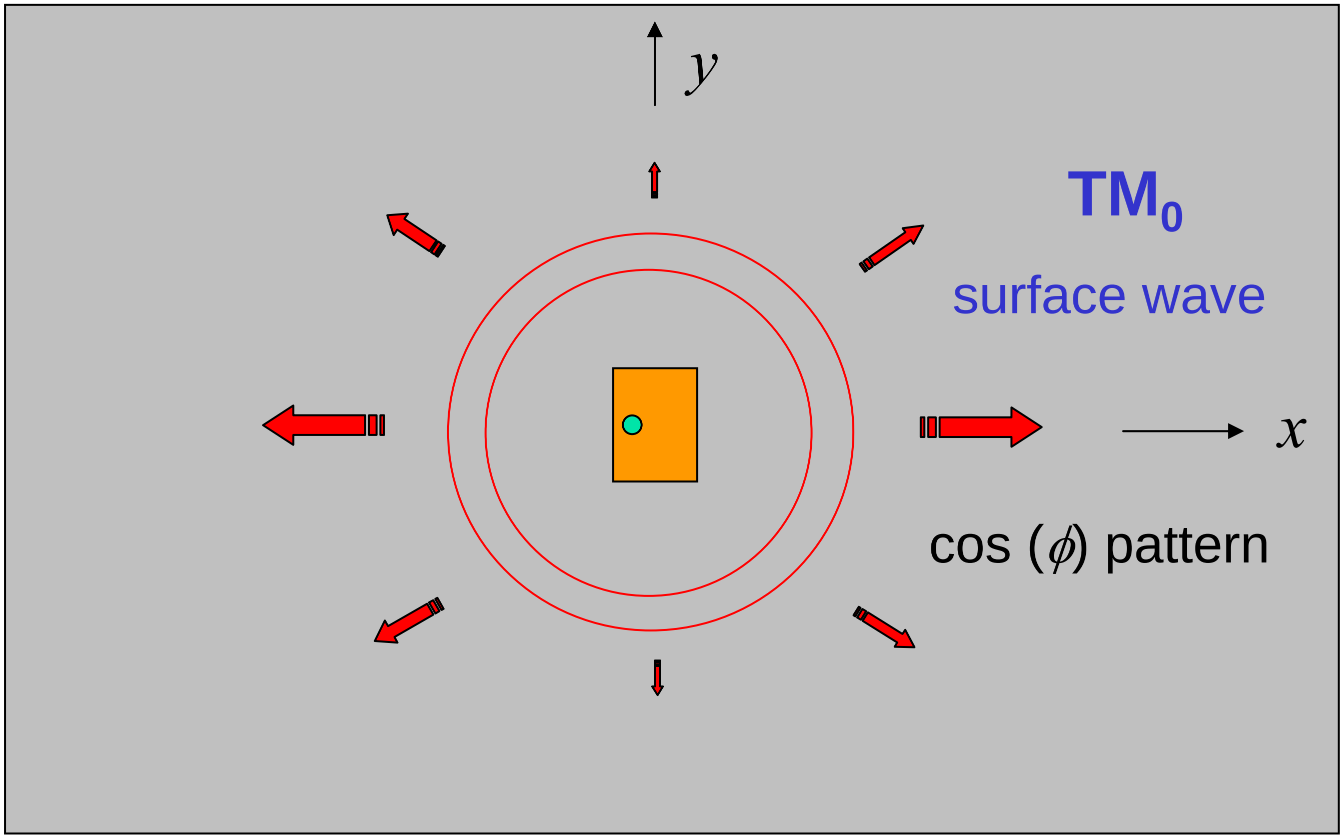
Radiation Efficiency

- Radiation efficiency is the ratio of power radiated into space, to the total input power.

$$e_r = \frac{P_r}{P_{tot}}$$

- The radiation efficiency is less than 100% due to
  - conductor loss
  - dielectric loss
  - surface-wave power

Radiation Efficiency (cont.)



Radiation Efficiency (cont.)

Hence,

$$e_r = \frac{P_r}{P_{tot}} = \frac{P_r}{P_r + \left(P_c + P_d + P_{sw}\right)}$$

P_r = radiated power

P_{tot} = total input power

P_c = power dissipated by conductors

P_d = power dissipated by dielectric

P_{sw} = power launched into surface wave

Radiation Efficiency (cont.)

- Conductor and dielectric loss is more important for thinner substrates.
- Conductor loss increases with frequency (proportional to $f^{1/2}$) due to the skin effect. Conductor loss is usually more important than dielectric loss.

$$R_s = \frac{1}{\sigma\delta} \qquad \delta = \sqrt{\frac{2}{\omega\mu\sigma}}$$

R_s is the surface resistance of the metal. The skin depth of the metal is δ.

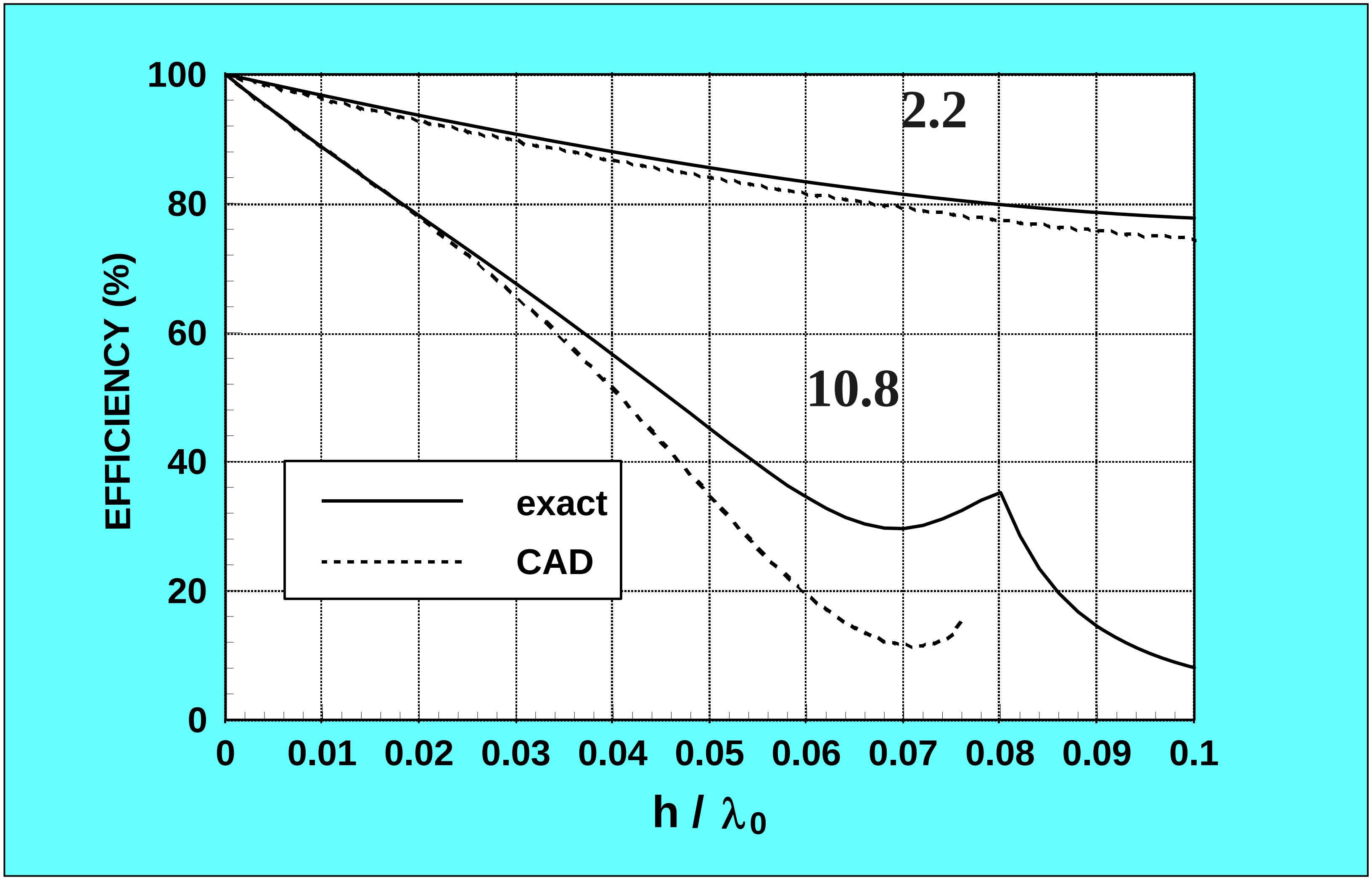
Radiation Efficiency (cont.)

- Surface-wave power is more important for thicker substrates or for higher substrate permittivities. (The surface-wave power can be minimized by using a foam substrate.)

Radiation Efficiency (cont.)

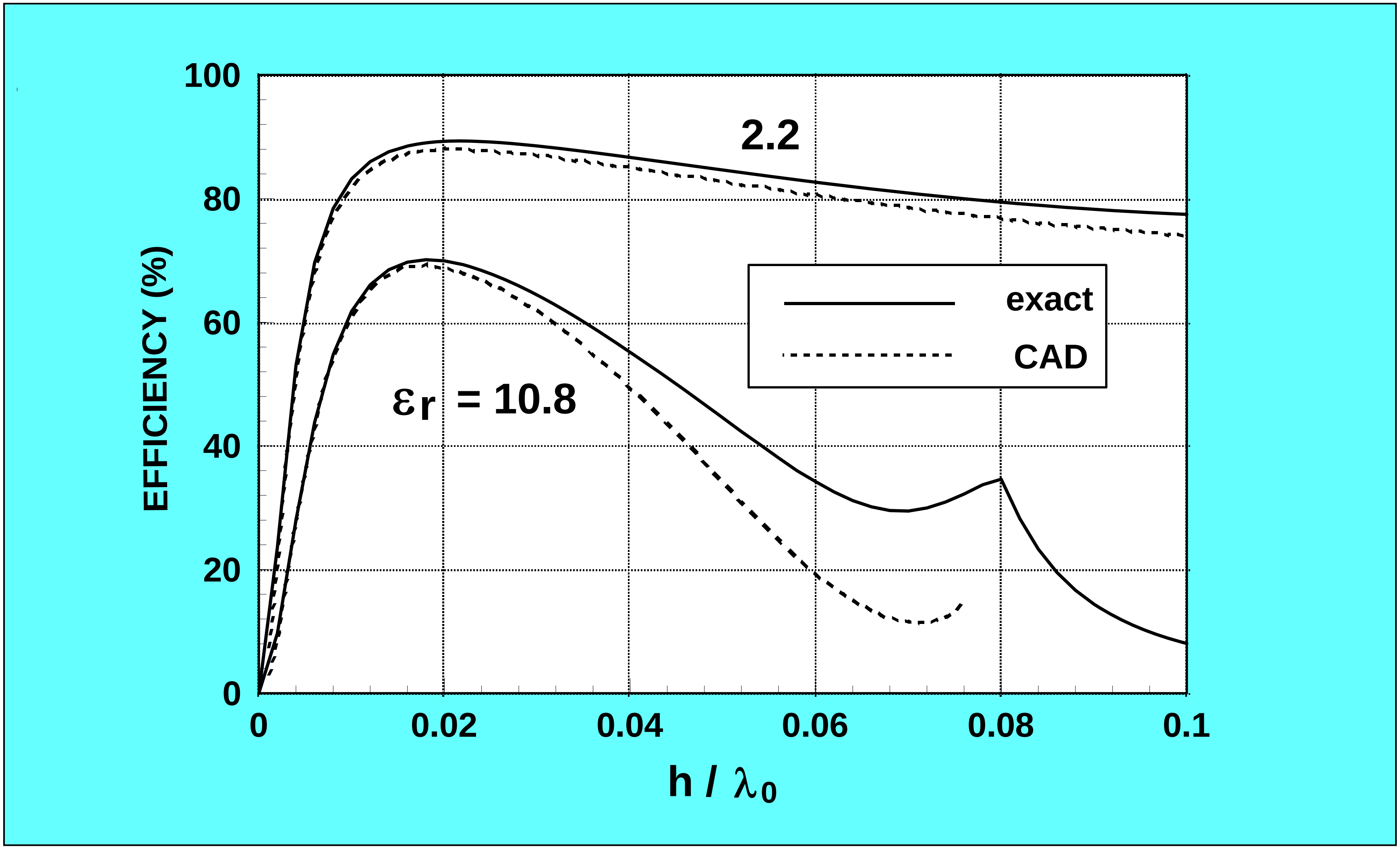
- For a foam substrate, higher radiation efficiency is obtained by making the substrate thicker (minimizing the conductor and dielectric losses). The thicker the better!
- For a typical substrate such as $\varepsilon_r = 2.2$, the radiation efficiency is maximum for $h / \lambda_0 \approx 0.02$.

Results: conductor and dielectric losses are neglected



$\varepsilon_r = 2.2$ or 10.8 $\quad W/L = 1.5$ $\quad$ Note: CAD plot uses Pozar formulas

Results: accounting for all losses

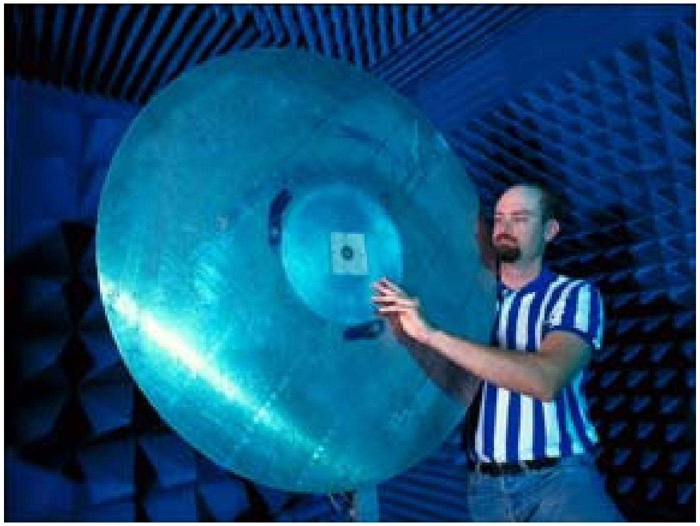


$\varepsilon_r = 2.2$ or 10.8 $\quad W/L = 1.5$ Note: CAD plot uses Pozar formulas

Basic Properties of Microstrip Antenna

Radiation Patterns

- The E-plane pattern is typically broader than the H-plane pattern.
- The truncation of the ground plane will cause edge diffraction, which tends to degrade the pattern by introducing:
  - rippling in the forward direction
  - back-radiation



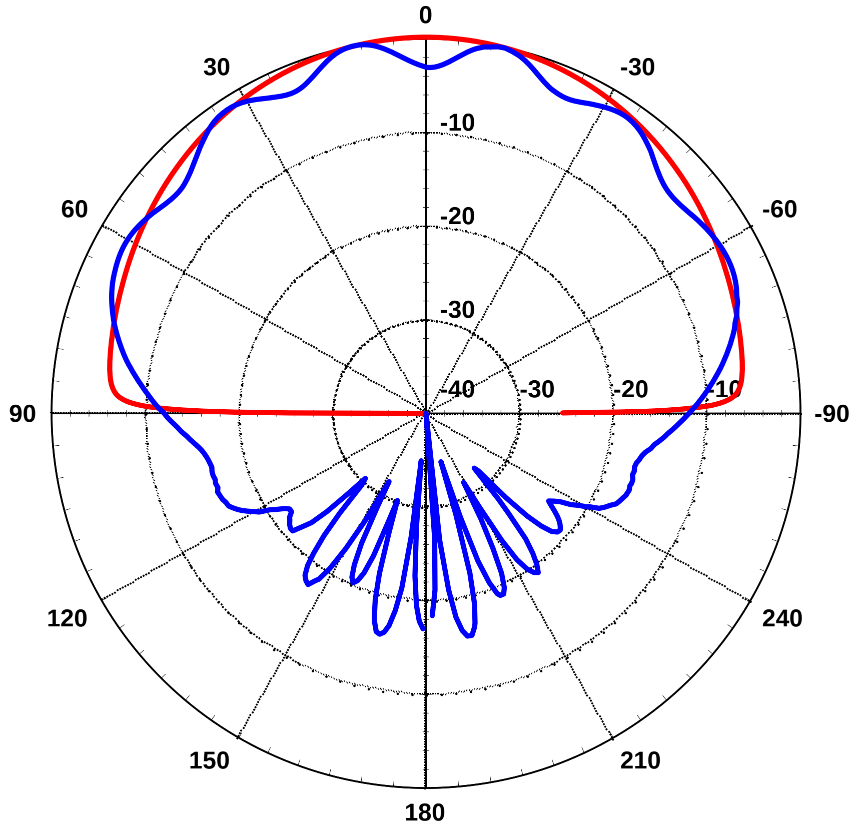
Note: pattern distortion is more severe in the E-plane, due to the angle dependence of the vertical polarization E_θ and the SW pattern. Both vary as cos (ϕ).

Radiation Patterns (cont.)

E-plane pattern

Red: infinite substrate and ground plane

Blue: 1 meter ground plane

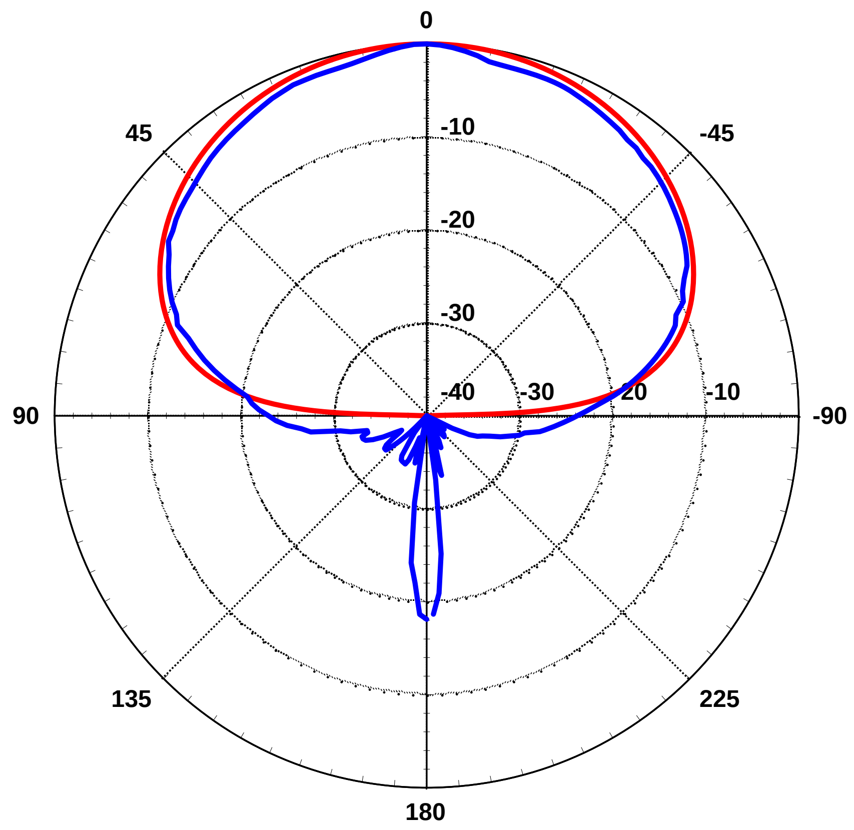


Radiation Patterns (cont.)

H-plane pattern

Red: infinite substrate and ground plane

Blue: 1 meter ground plane

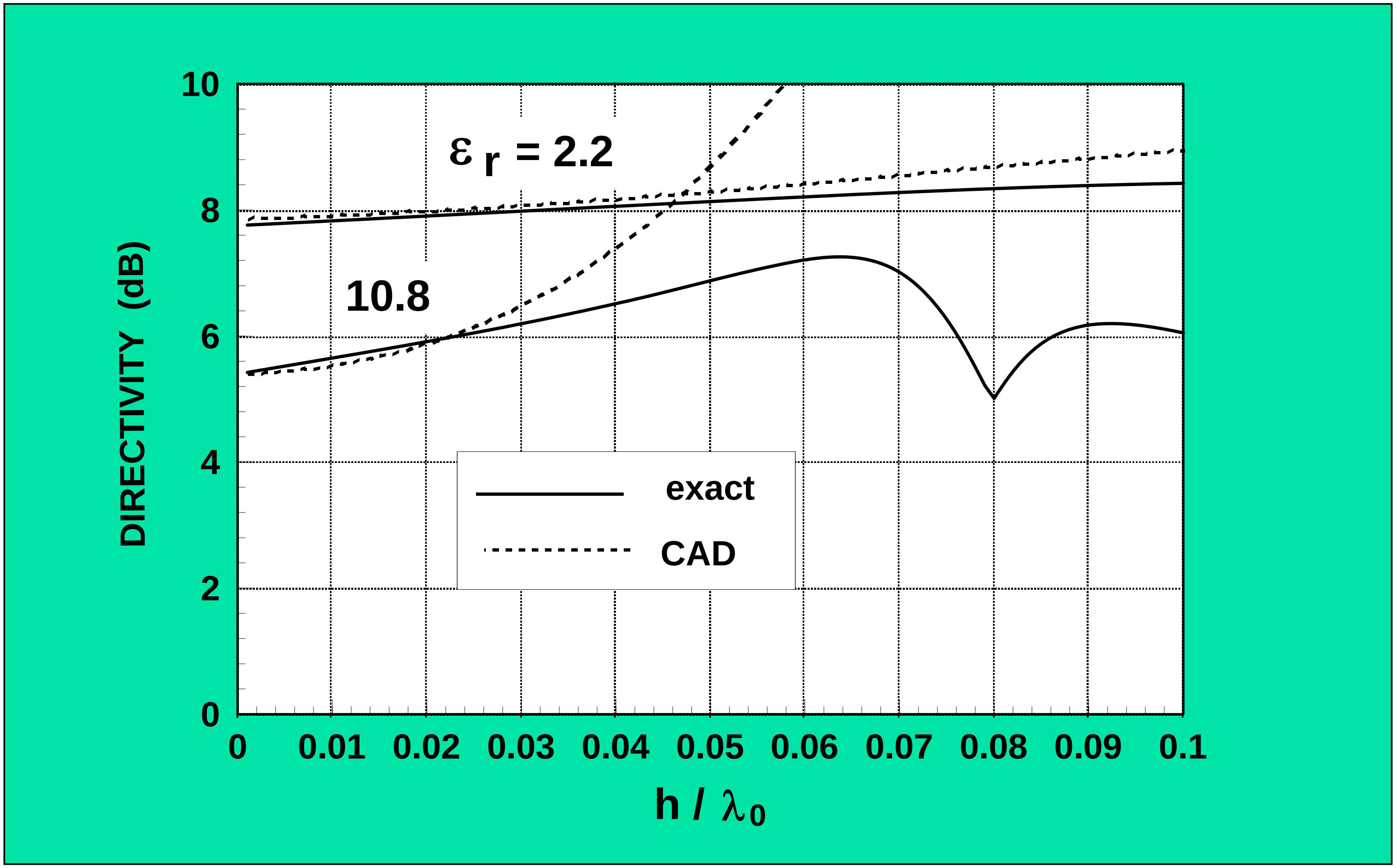


Basic Properties of Microstrip Antennas

Directivity

- The directivity is fairly insensitive to the substrate thickness.
- The directivity is higher for lower permittivity, because the patch is larger.

Results: Directivity

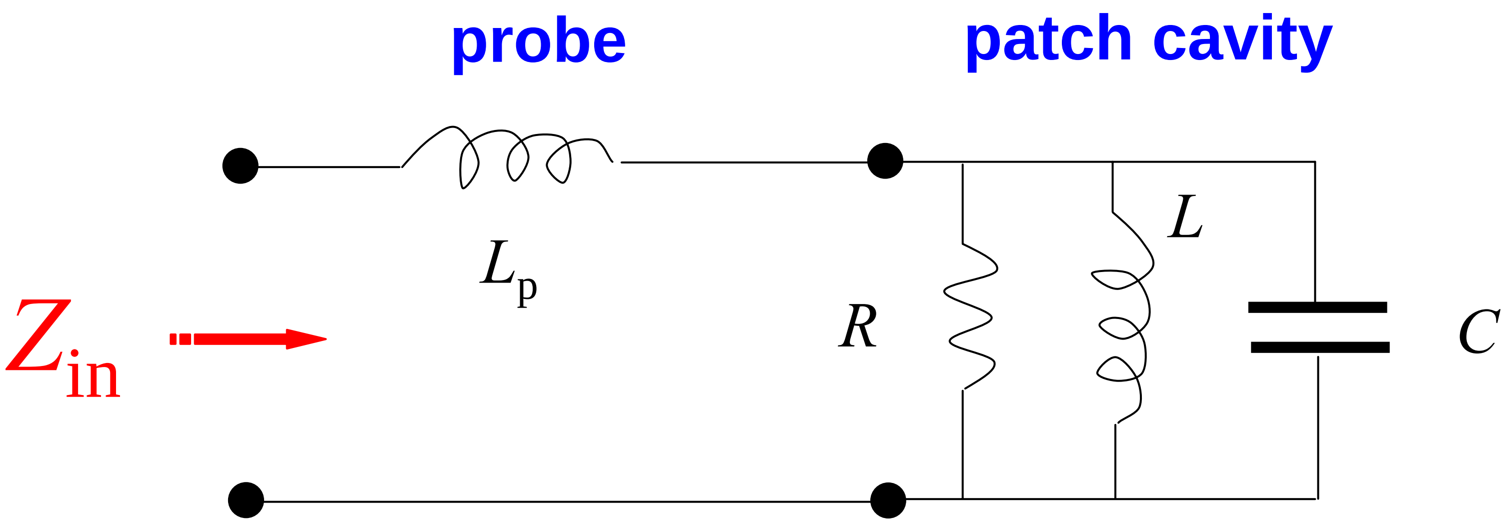


ε_r = 2.2 or 10.8

$W/L = 1.5$

Approximate CAD Model for Z_{in}

- Near the resonance frequency, the patch cavity can be approximately modeled as an RLC circuit.
- A probe inductance L_p is added in series, to account for the "probe inductance".



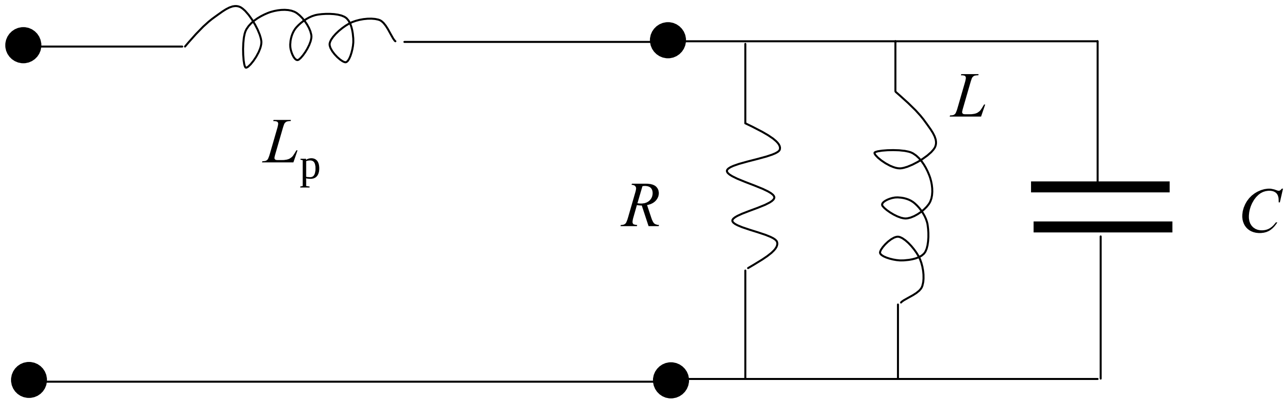
Approximate CAD Model (cont.)

$$Z_{in} \approx j\omega L_p + \frac{R}{1 + j2Q\left(f / f_0 - 1\right)}$$

$$Q = \frac{R}{\omega_0 L} \qquad BW = \frac{1}{\sqrt{2}\,Q}$$

BW is defined here by SWR < 2.0.

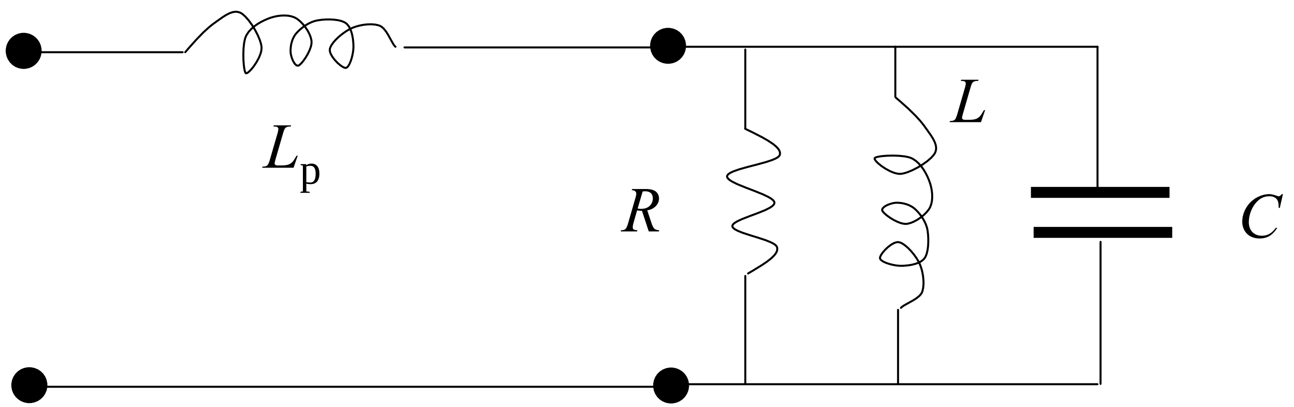
$$\omega_0 = 2\pi f_0 = \frac{1}{\sqrt{LC}}$$



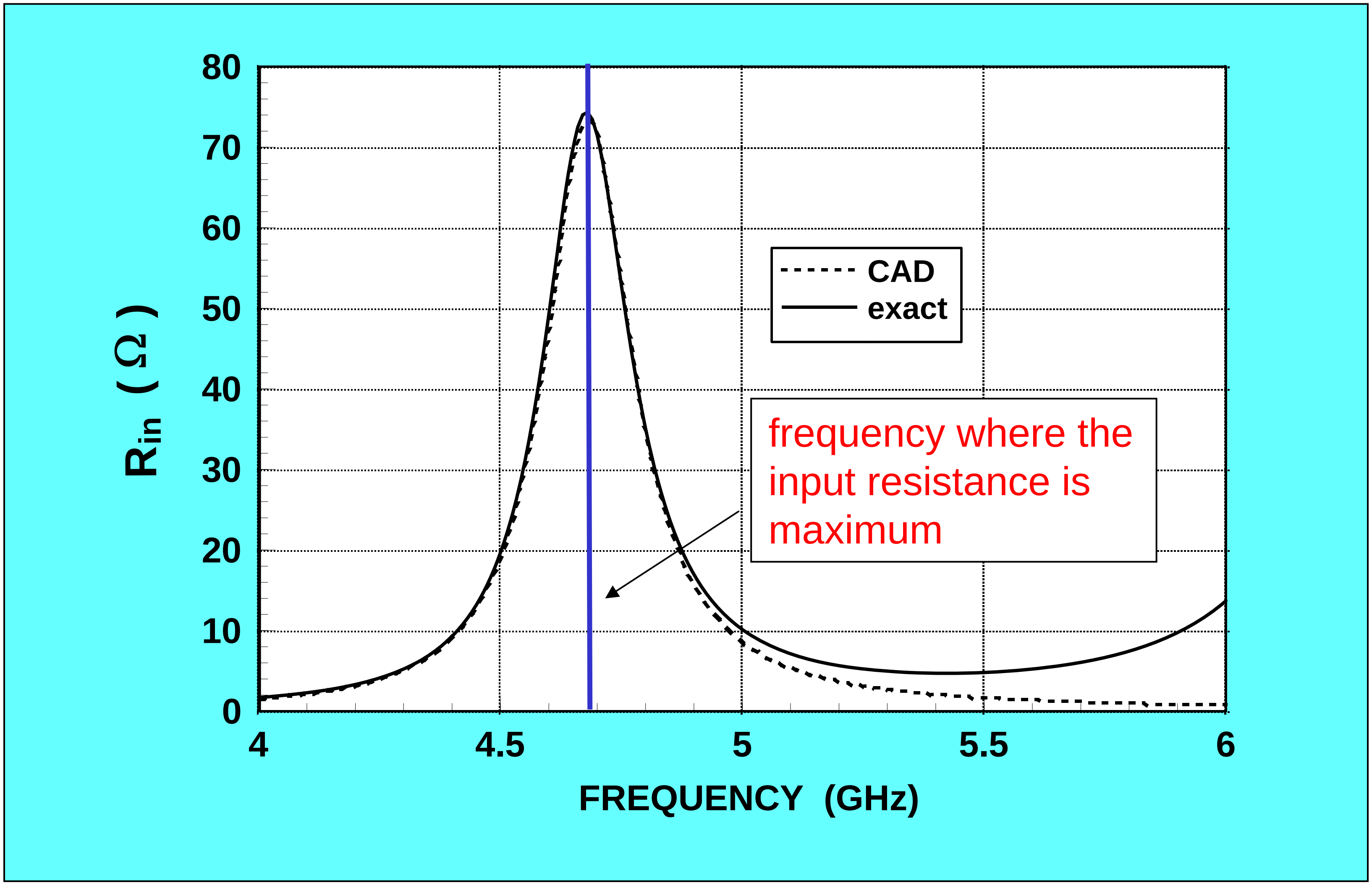
Approximate CAD Model (cont.)

$$R = R_{\text{in max}}$$

$R_{\text{in max}}$ is the input resistance at the resonance of the patch cavity (the frequency that maximizes R_{in}).

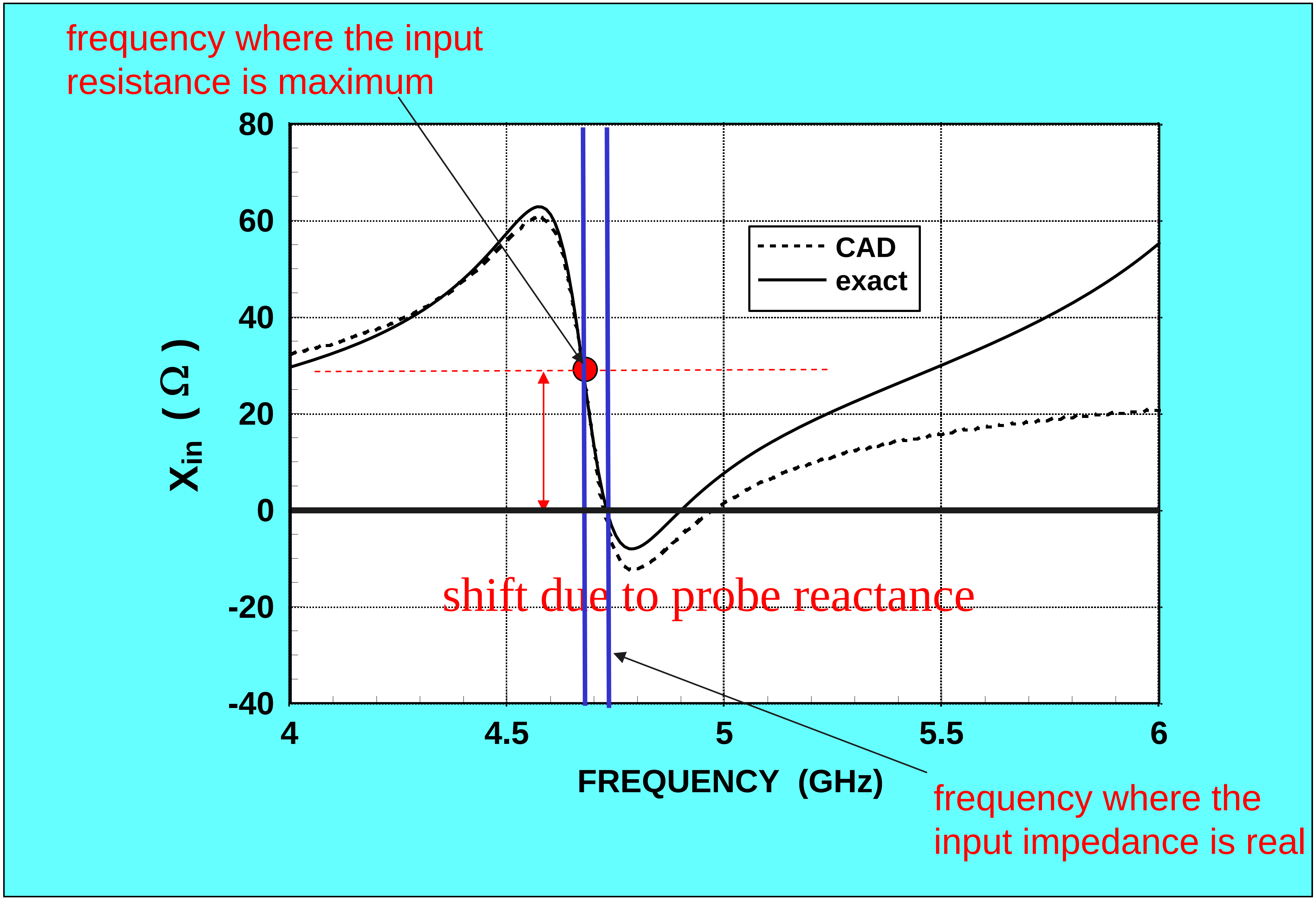


Results : input resistance vs. frequency



$\varepsilon_r = 2.2$ $W/L = 1.5$ $L = 3.0$ cm

Results: input reactance vs. frequency



$\varepsilon_r = 2.2$ $\quad W/L = 1.5$ $\quad L = 3.0$ cm

Approximate CAD Model (cont.)

Approximate CAD formula for feed (probe) reactance (in Ohms)

a = probe radius h = probe height

$$X_f = \frac{\eta_0}{2\pi}(k_0 h)\left[-\gamma + \ln\left(\frac{2}{\sqrt{\varepsilon_r}\,(k_0 a)}\right)\right]$$

$$X_f = \omega L_p$$

$\gamma \doteq 0.577216$ (Euler's constant)

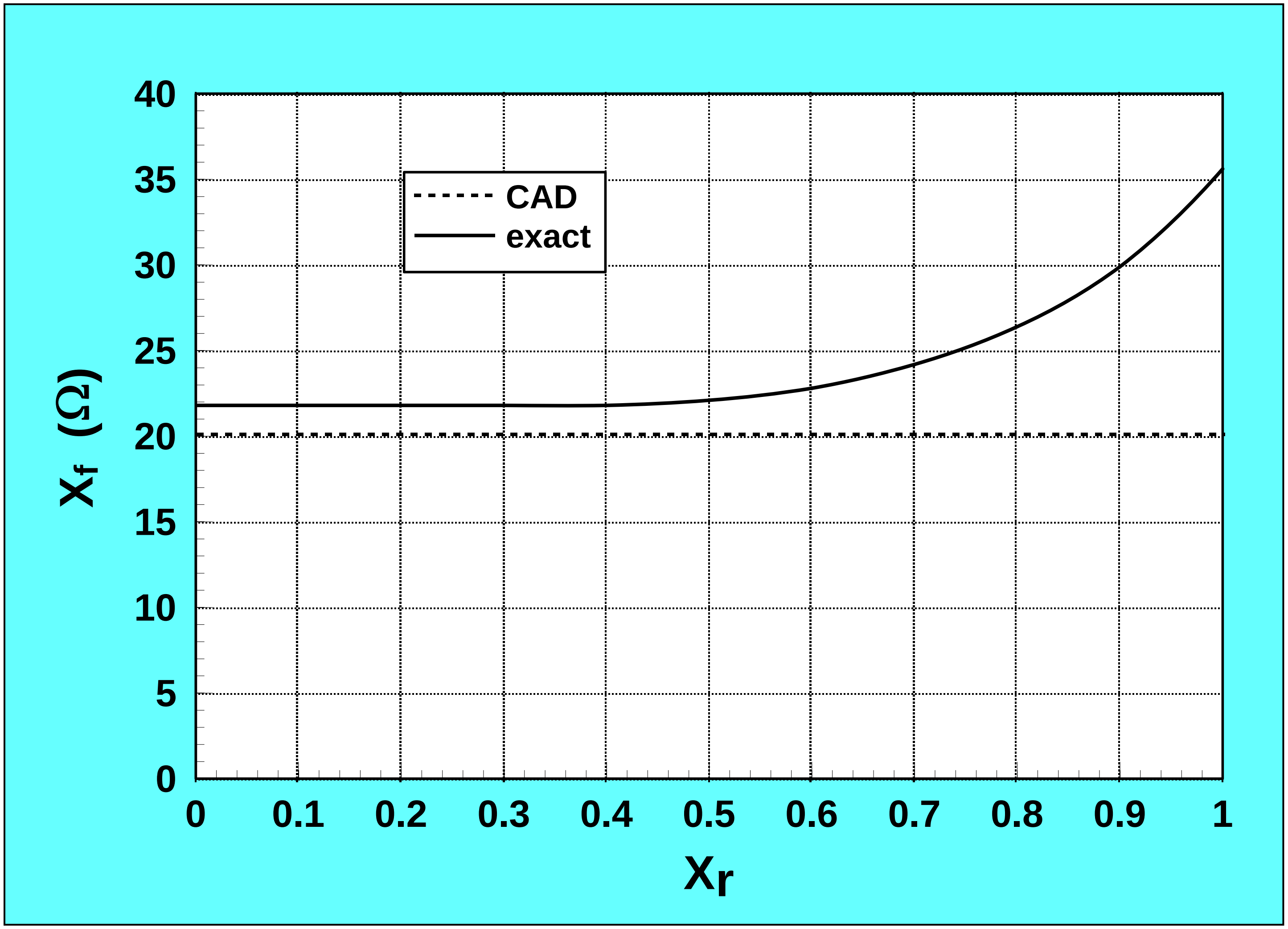
$$\eta_0 = \sqrt{\mu_0 / \varepsilon_0} = 376.73\Omega$$

Approximate CAD Model (cont.)

- Feed (probe) reactance increases proportionally with substrate thickness h.
- Feed reactance increases for smaller probe radius.

$$X_f = \frac{\eta_0}{2\pi}\left(k_0 h\right)\left[-\gamma + \ln\left(\frac{2}{\sqrt{\varepsilon_r}\left(k_0 a\right)}\right)\right]$$

Results: probe reactance ($X_f = X_p = \omega L_p$)



$\varepsilon_r = 2.2$

$W/L = 1.5$

$h = 0.0254\ \lambda_0$

$a = 0.5$ mm

$x_r = 2\ (x_0 / L) - 1$

x_r is zero at the center of the patch, and is 1.0 at the patch edge.

CAD Formulas

In the following viewgraphs, CAD formulas for the important properties of the rectangular microstrip antenna will be shown.

CAD Formula: Radiation Efficiency

$$e_r = \frac{e_r^{hed}}{1 + e_r^{hed}\left[\ell_d + \left(\frac{R_s}{\pi \eta_0}\right)\left(\frac{1}{h/\lambda_0}\right)\right]\left[\left(\frac{3}{16}\right)\left(\frac{\varepsilon_r}{p\,c_1}\right)\left(\frac{L}{W}\right)\left(\frac{1}{h/\lambda_0}\right)\right]}$$

where

$$\ell_d = \tan\delta = \text{loss tangent of substrate}$$

$$R_s = \text{surface resistance of metal} = \frac{1}{\sigma\delta} = \sqrt{\frac{\omega\mu}{2\sigma}}$$

CAD Formula: Radiation Efficiency (cont.)

$$e_r^{hed} = \frac{P_{sp}^{hed}}{P_{sp}^{hed} + P_{sw}^{hed}} = \frac{1}{1 + \frac{P_{sw}^{hed}}{P_{sp}^{hed}}}$$

where

$$P_{sp}^{hed} = \frac{1}{\lambda_0^2}\left(k_0 h\right)^2 \left(80\pi^2 c_1\right)$$

$$P_{sw}^{hed} = \frac{1}{\lambda_0^2}\left(k_0 h\right)^3 \left[60\pi^3 c_1 \left(1 - \frac{1}{\varepsilon_r}\right)^3\right]$$

Note: "hed" refers to a unit-amplitude horizontal electric dipole.

CAD Formula: Radiation Efficiency (cont.)

Hence we have

$$e_r^{hed} = \frac{1}{1 + \frac{3}{4}\pi\left(k_0 h\right)\left(\frac{1}{c_1}\right)\left(1 - \frac{1}{\varepsilon_r}\right)^3}$$

(Physically, this term is the radiation efficiency of a horizontal electric dipole (hed) on top of the substrate.)

CAD Formula: Radiation Efficiency (cont.)

The constants are defined as

$$c_1 = 1 - \frac{1}{\varepsilon_r} + \frac{2/5}{\varepsilon_r^2}$$

$$p = 1 + \frac{a_2}{10}\left(k_0 W\right)^2 + \left(a_2^2 + 2a_4\right)\left(\frac{3}{560}\right)\left(k_0 W\right)^4 + c_2\left(\frac{1}{5}\right)\left(k_0 L\right)^2$$
$$+ a_2 c_2\left(\frac{1}{70}\right)\left(k_0 W\right)^2\left(k_0 L\right)^2$$

$$c_2 = -0.0914153$$

$$a_2 = -0.16605$$

$$a_4 = 0.00761$$

CAD Formula: Radiation Efficiency (cont.)

Improved formula (due to Pozar)

$$e_r^{hed} = \frac{1}{1 + \frac{P_{sw}^{hed}}{P_{sp}^{hed}}}$$

$$P_{sp}^{hed} = \frac{1}{\lambda_0^2}\left(k_0 h\right)^2\left(80\pi^2 c_1\right)$$

$$P_{sw}^{hed} = \frac{\eta_0 k_0^2}{4}\frac{\varepsilon_r\left(x_0^2-1\right)^{3/2}}{\varepsilon_r\left(1+x_1\right)+(k_0 h)\sqrt{x_0^2-1}\left(1+\varepsilon_r^2 x_1\right)}$$

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

$$x_0 = 1 + \frac{-\varepsilon_r^2 + \alpha_0\alpha_1 + \varepsilon_r\sqrt{\varepsilon_r^2 - 2\alpha_0\alpha_1 + \alpha_0^2}}{\varepsilon_r^2 - \alpha_1^2}$$

CAD Formula: Radiation Efficiency (cont.)

Improved formula (cont.)

$$\alpha_0 = s \tan\left[\left(k_0 h \right) s \right]$$

$$\alpha_1 = -\frac{1}{s} \left[\tan\left[\left(k_0 h \right) s \right] + \frac{\left(k_0 h \right) s}{\cos^2\left[\left(k_0 h \right) s \right]} \right]$$

$$s = \sqrt{\varepsilon_r - 1}$$

CAD Formula: Bandwidth

$$BW = \frac{1}{\sqrt{2}}\left[\ell_d + \left(\frac{R_s}{\pi \eta_0}\right)\left(\frac{1}{h/\lambda_0}\right) + \left(\frac{16}{3}\right)\left(\frac{p\,c_1}{\varepsilon_r}\right)\left(\frac{h}{\lambda_0}\right)\left(\frac{W}{L}\right)\left(\frac{1}{e_r^{hed}}\right)\right]$$

BW is defined from the frequency limits f_1 and f_2 at which SWR = 2.0:

$$BW = \frac{f_2 - f_1}{f_0}$$

(multiply by 100 if you want to get %)

CAD Formula: Resonant Input Resistance

(probe-feed)

$$R = R_{edge} \cos^2\left(\frac{\pi x_0}{L}\right)$$

$$R_{edge} = \frac{\left(\frac{4}{\pi}\right)(\eta_0)\left(\frac{L}{W}\right)\left(\frac{h}{\lambda_0}\right)}{\ell_d + \left(\frac{R_s}{\pi \eta_0}\right)\left(\frac{1}{h/\lambda_0}\right) + \left(\frac{16}{3}\right)\left(\frac{p c_1}{\varepsilon_r}\right)\left(\frac{W}{L}\right)\left(\frac{h}{\lambda_0}\right)\left(\frac{1}{e_r^{hed}}\right)}$$

CAD Formula: Directivity

$$D = \left(\frac{3}{pc_1} \right) \left[\frac{\varepsilon_r}{\varepsilon_r + \tan^2(k_1 h)} \right] \left(\operatorname{tanc}^2(k_1 h) \right)$$

where

$$\operatorname{tanc}(x) \equiv \tan(x) / x$$

CAD Formula: Directivity (cont.)

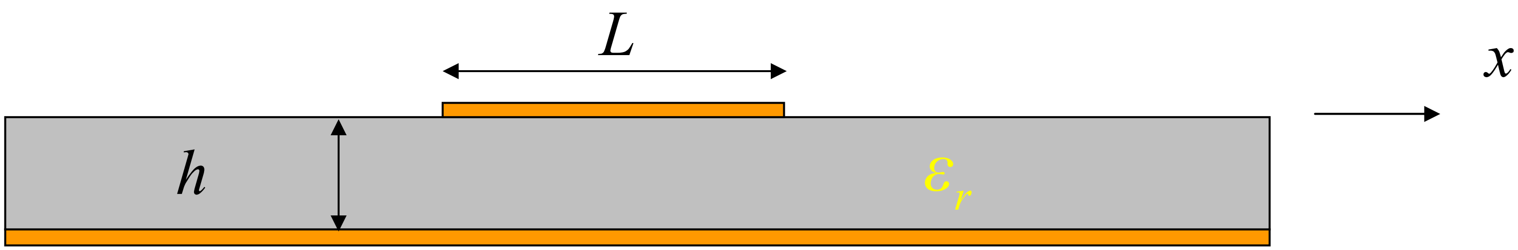
For thin substrates:

$$D \approx \frac{3}{p\, c_1}$$

(The directivity is essentially independent of the substrate thickness.)

CAD Formula: Radiation Patterns

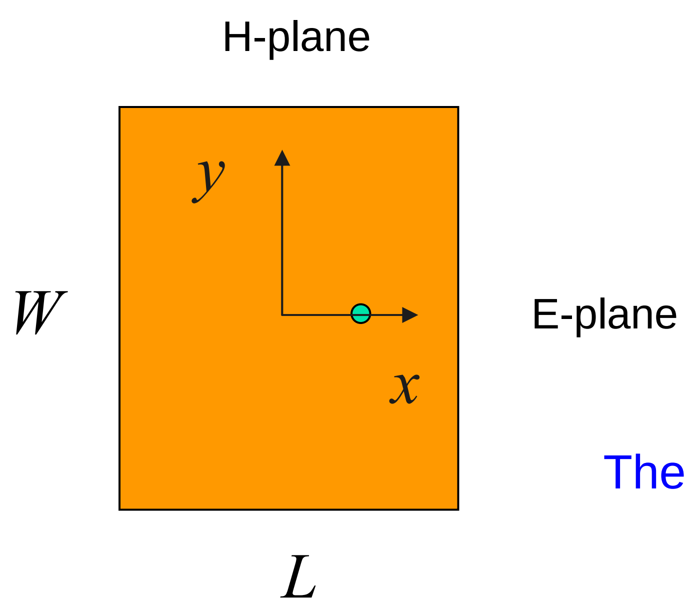
(based on electric current model)



infinite GP and substrate

The origin is at the center of the patch.

$$\underline{J}_s = \underline{\hat{x}} \cos\left(\frac{\pi x}{L}\right)$$



The probe is on the x axis.

CAD Formula: Radiation Patterns (cont.)

The far-field pattern can be determined by reciprocity.

$$E_i(r,\theta,\phi) = E_i^{hex}(r,\theta,\phi)\left(\frac{\pi WL}{2}\right)\left[\frac{\sin\left(\frac{k_y W}{2}\right)}{\frac{k_y W}{2}}\right]\left[\frac{\cos\left(\frac{k_x L}{2}\right)}{\left(\frac{\pi}{2}\right)^2-\left(\frac{k_x L}{2}\right)^2}\right]$$

$$i=\theta \text{ or } \phi$$

$$k_x = k_0 \sin\theta \cos\phi$$

$$k_y = k_0 \sin\theta \sin\phi$$

The “hex” pattern is for a horizontal electric dipole in the x direction, sitting on top of the substrate.

CAD Formula: Radiation Patterns (cont.)

$$E_{\phi}^{hex}\left(r,\theta,\phi\right) = -E_0 \sin\phi \; F\left(\theta\right)$$

$$E_{\theta}^{hex}\left(r,\theta,\phi\right) = E_0 \cos\phi \, G\left(\theta\right)$$

where

$$E_0 = \left(\frac{-j\omega\mu_0}{4\pi r}\right) e^{-jk_0 r}$$

$$F\left(\theta\right) = 1 + \Gamma^{TE}\left(\theta\right) = \frac{2\tan\left(k_0 h \, N\left(\theta\right)\right)}{\tan\left(k_0 h \, N\left(\theta\right)\right) - j \, N\left(\theta\right)\sec\theta}$$

$$G\left(\theta\right) = \cos\theta\left(1 + \Gamma^{TM}\left(\theta\right)\right) = \frac{2\tan\left(k_0 h \, N\left(\theta\right)\right)\cos\theta}{\tan\left(k_0 h \, N\left(\theta\right)\right) - j \dfrac{\varepsilon_r}{N\left(\theta\right)}\cos\theta}$$

$$N\left(\theta\right) = \sqrt{\varepsilon_r - \sin^2\left(\theta\right)}$$

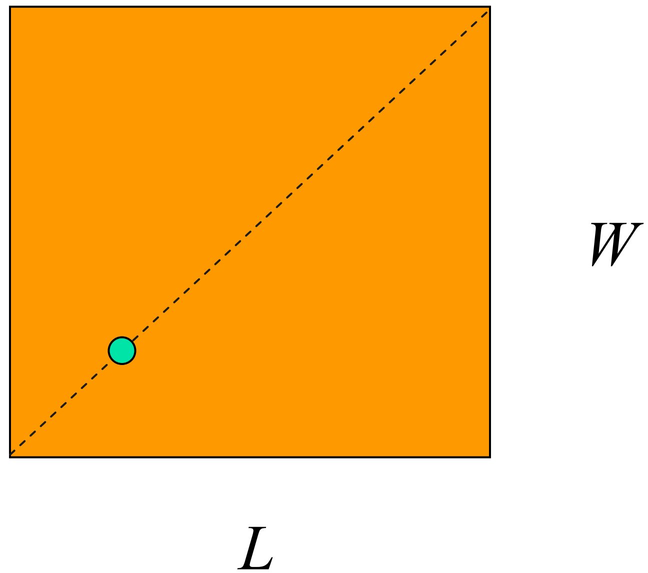
Circular Polarization

Three main techniques:

1) Single feed with “nearly degenerate” eigenmodes.
2) Dual feed with delay line or 90^o hybrid phase shifter.
3) Synchronous subarray technique.

Circular Polarization: Single Feed

The feed is on the diagonal.
The patch is nearly (but not exactly) square.



Basic principle: the two modes are excited with equal amplitude, but with a ±45° phase.

Circular Polarization: Single Feed

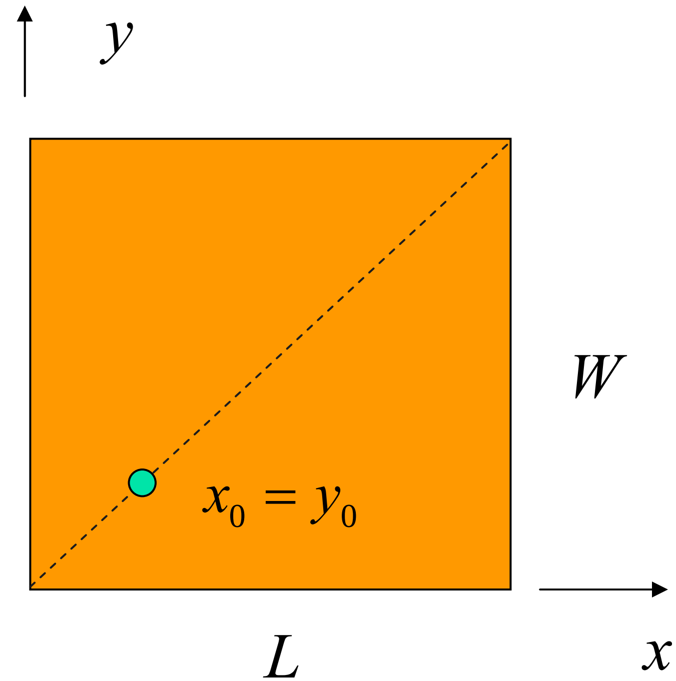
Design equations:

$$f_0 = f_{CP}$$

Resonant frequency is the optimum CP frequency

$$BW = \frac{1}{\sqrt{2}Q}$$

(SWR < 2)



$$f_x = f_0\left(1 \mp \frac{1}{2Q}\right)$$

$$f_y = f_0\left(1 \pm \frac{1}{2Q}\right)$$

Top sign for LHCP, bottom sign for RHCP.

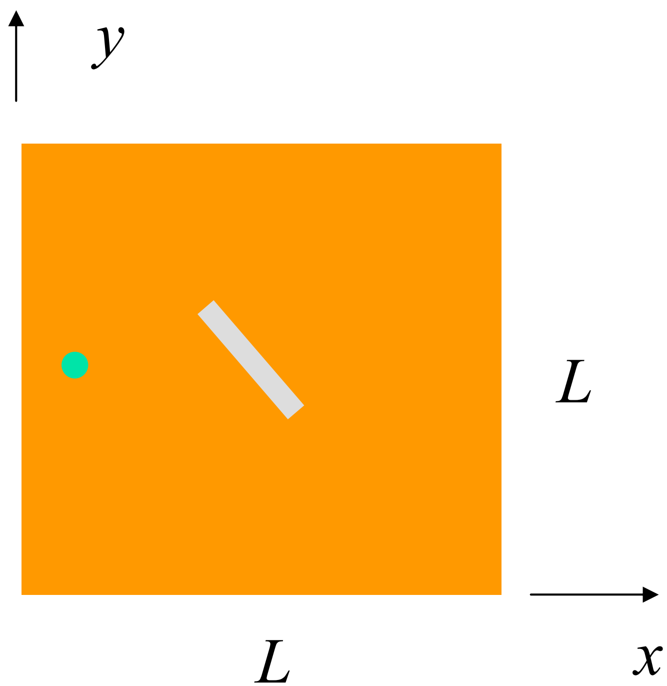
$$R = R_x = R_y$$

R_x and R_y are the resonant input resistances of the two LP (x and y) modes, for the same feed position as in the CP patch.

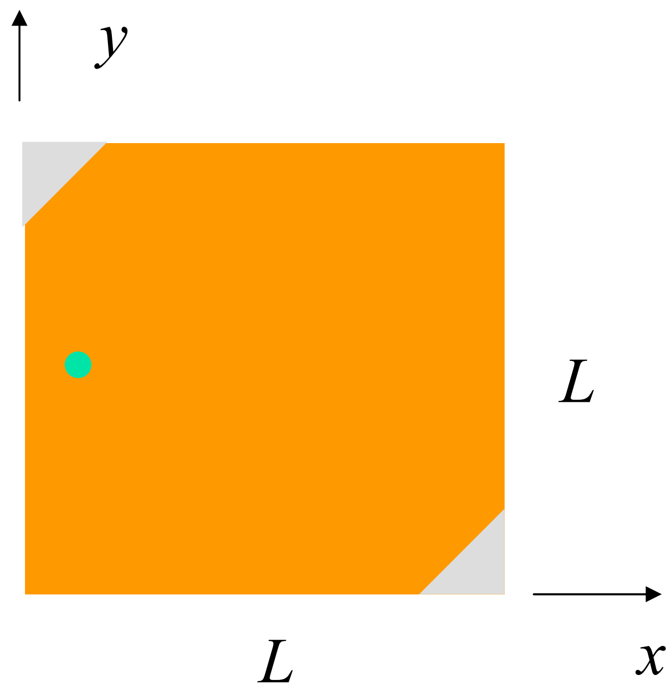
Circular Polarization: Single Feed (cont.)

Other variations

Note: diagonal modes are used as degenerate modes



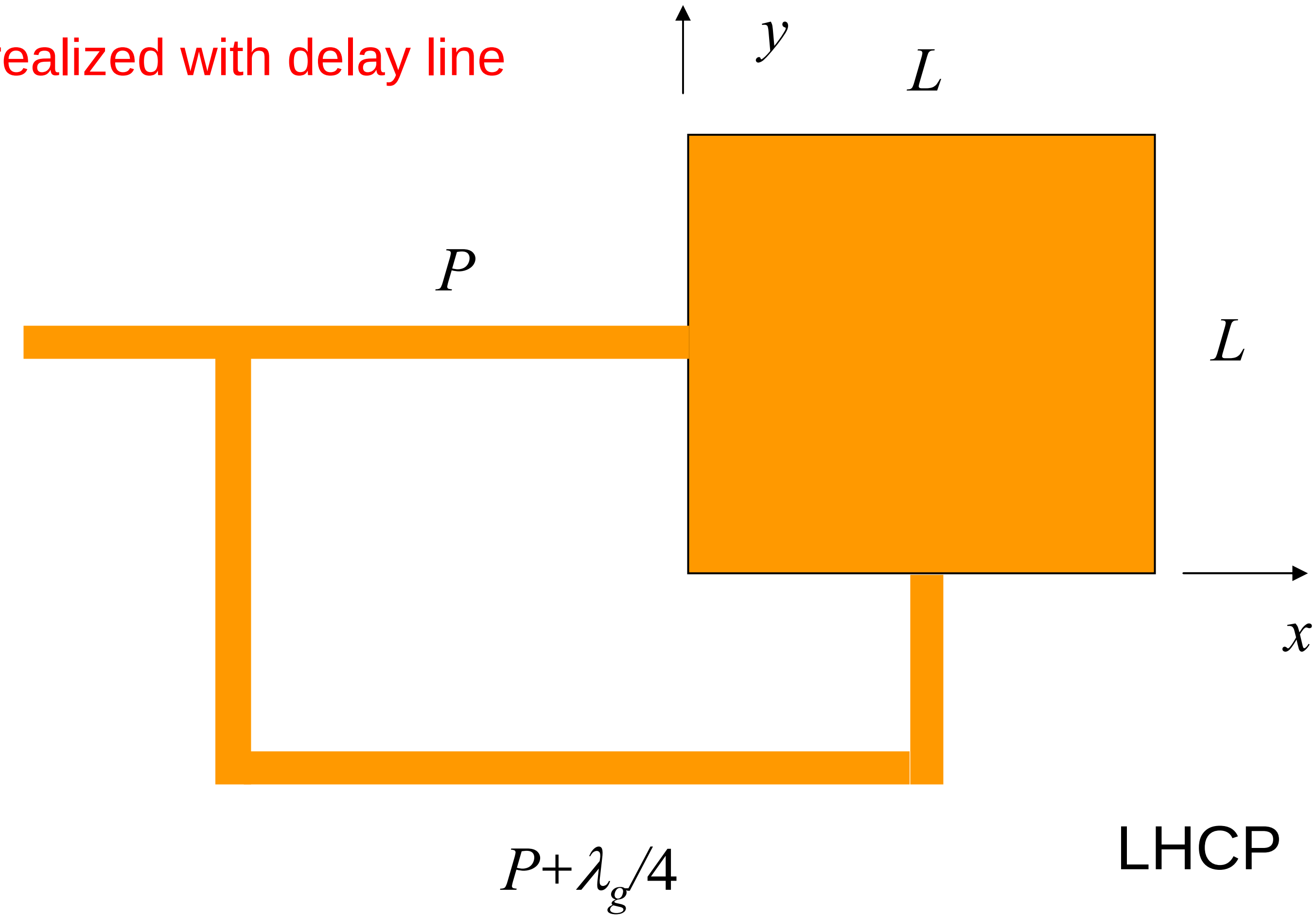
Patch with slot



Patch with truncated corners

Circular Polarization: Dual Feed

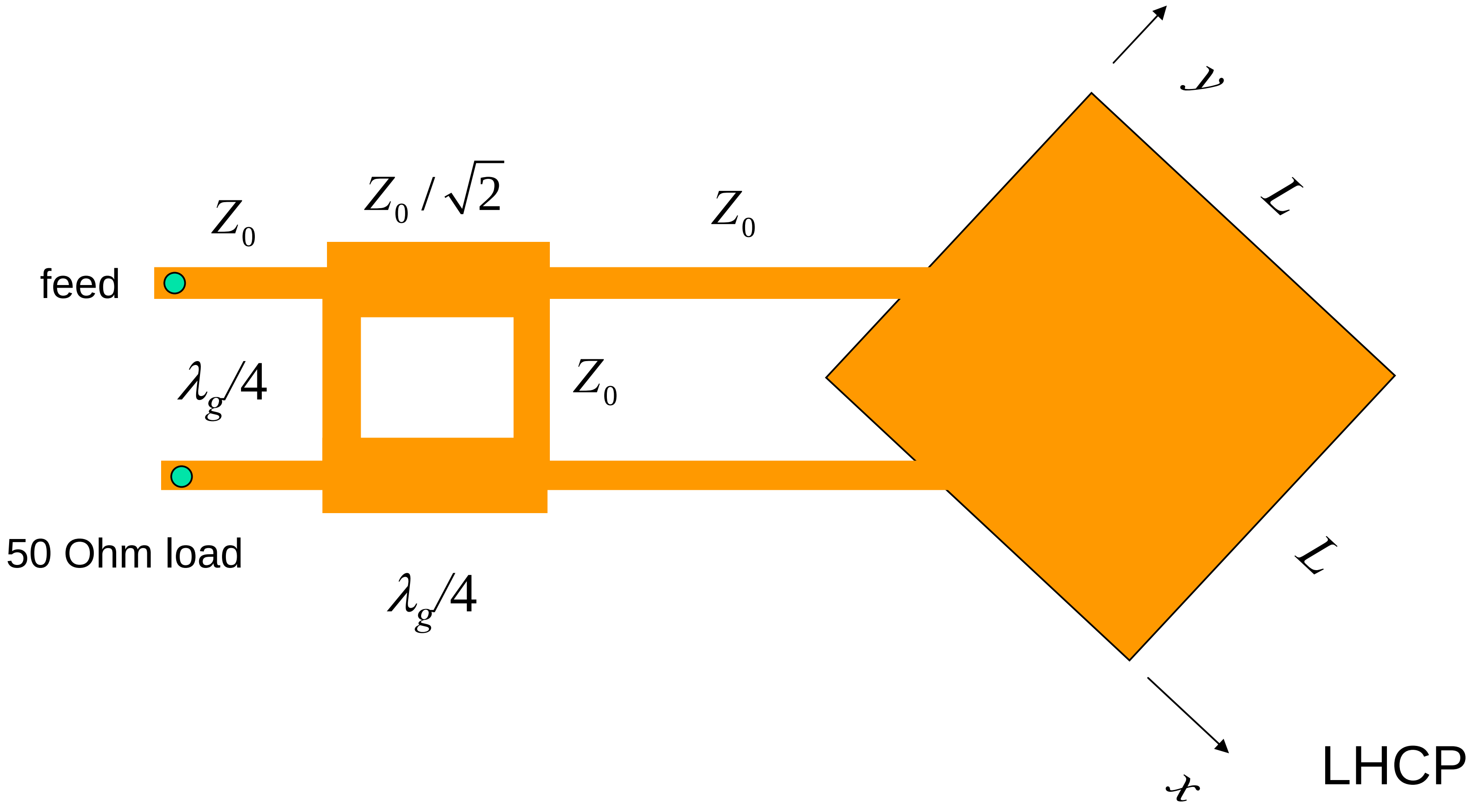
Phase shift realized with delay line



LHCP

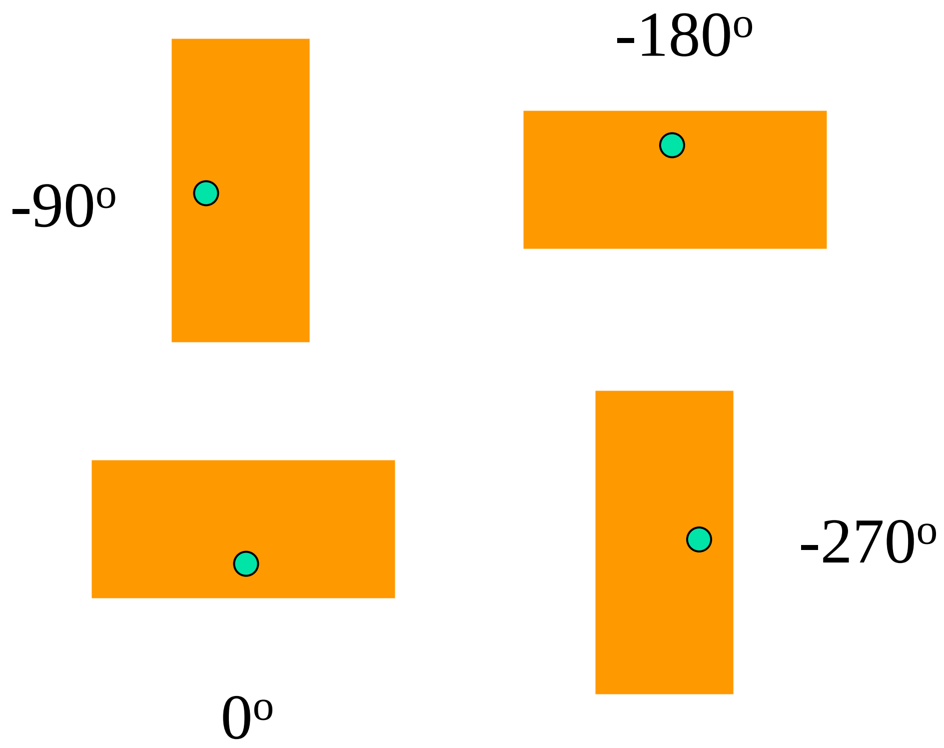
Circular Polarization: Dual Feed

Phase shift realized with 90º hybrid (branchline coupler)



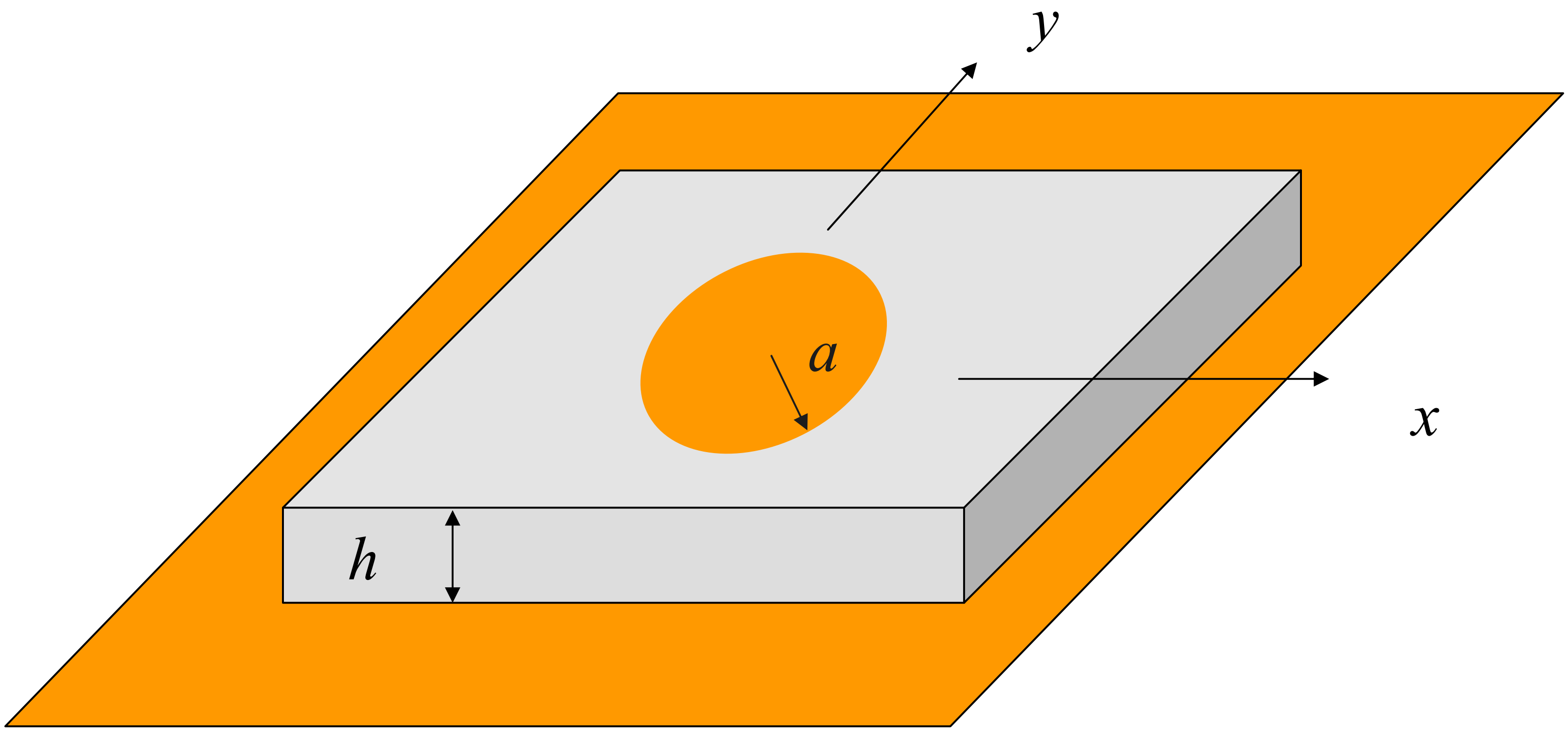
Circular Polarization: Synchronous Rotation

Elements are rotated in space and fed with phase shifts



Because of symmetry, radiation from higher-order modes tends to be reduced, resulting in good cross-pol.

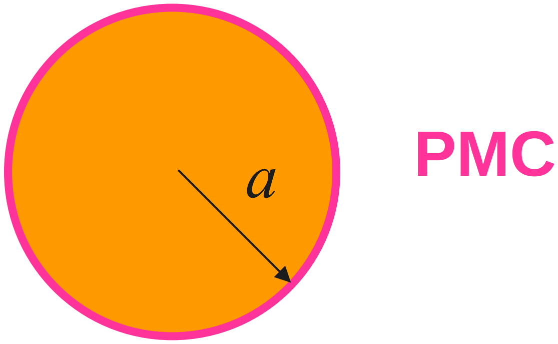
Circular Patch



Circular Patch: Resonance Frequency

From separation of variables:

$$E_z = \cos(m\phi) J_m(k\rho)$$



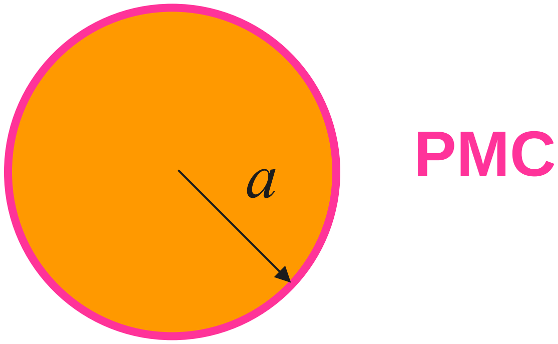
J_m = Bessel function of first kind, order m.

$$\left. \frac{\partial E_z}{\partial \rho} \right|_{\rho=a} = 0 \quad \Rightarrow \quad J'_m(ka) = 0$$

Circular Patch: Resonance Frequency (cont.)

$$ka = x'_{mn}$$

(nth root of $J_m{}'$ Bessel function)



$$f_{mn} = \frac{c}{2\pi\sqrt{\varepsilon_r}}\, x'_{mn}$$

Dominant mode: TM_{11}

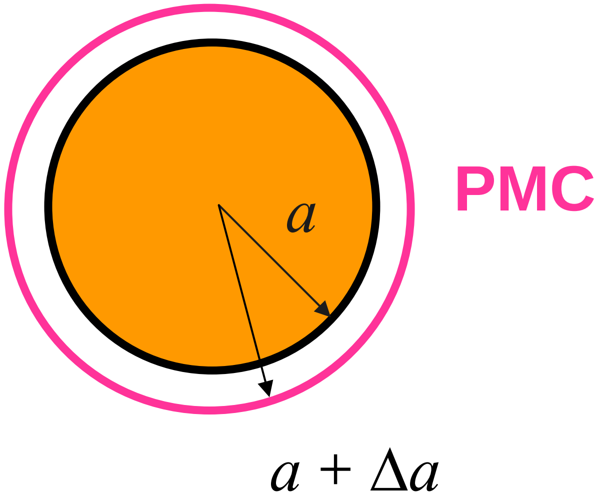
$$f_{11} = \frac{c}{2\pi a\sqrt{\varepsilon_r}}\, x'_{11}$$

$$x'_{11} \approx 1.842$$

Circular Patch: Resonance Frequency (cont.)

Fringing extension: $a_e = a + \Delta a$

$$f_{11} = \frac{c}{2\pi a_e \sqrt{\varepsilon_r}}\, x'_{11}$$



"Long/Shen Formula" :

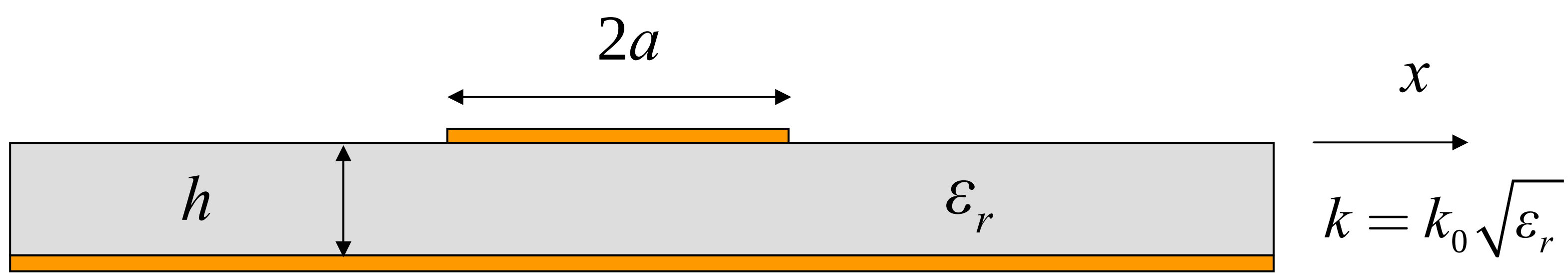
$$a_e = a\sqrt{1 + \frac{2h}{\pi a \varepsilon_r}\left[\ln\left(\frac{\pi a}{2h}\right) + 1.7726\right]}$$

or

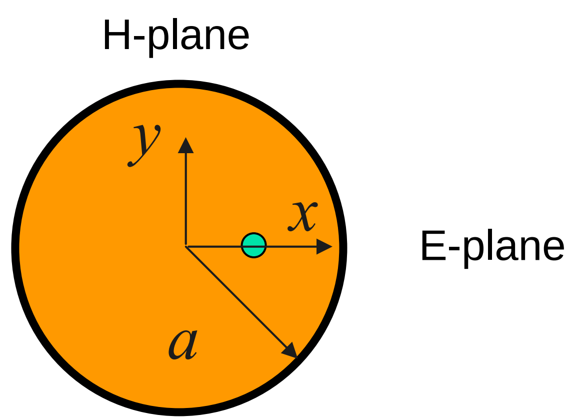
$$\Delta a = \frac{h}{\pi \varepsilon_r}\left[\ln\left(\frac{\pi a}{2h}\right) + 1.7726\right]$$

Circular Patch: Patterns

(based on magnetic current model)



infinite GP and substrate



The origin is at the center of the patch.

The probe is on the x axis.

In patch cavity:

$$E_z(\rho,\varphi) = \cos\varphi\left(\frac{J_1(k\rho)}{J_1(ka)}\right)\left(\frac{1}{h}\right)$$

(The edge voltage has a maximum of one volt.)

Circular Patch: Patterns (cont.)

$$E_{\theta}^{R}(r,\theta,\varphi) = 2\pi a \frac{E_0}{\eta_0} \,\mathrm{tanc}(k_{z1}h)\, \cos\varphi \; J'_1(k_0 a \,\sin\theta) Q(\theta)$$

$$E_{\varphi}^{R}(r,\theta,\varphi) = -2\pi a \frac{E_0}{\eta_0} \,\mathrm{tanc}(k_{z1}h)\, \sin\varphi \left(\frac{J_1(k_0 a \,\sin\theta)}{k_0 a \,\sin\theta} \right) P(\theta)$$

where

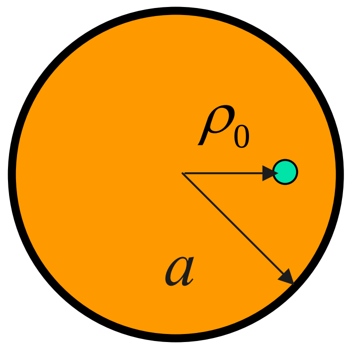
$$\mathrm{tanc}(x) = \tan(x)/x$$

$$P(\theta) = \cos\theta \left(1 - \Gamma^{TE}(\theta)\right) = \cos\theta \left[\frac{-2jN(\theta)}{\tan(k_0 h N(\theta)) - jN(\theta)\sec\theta} \right]$$

$$Q(\theta) = 1 - \Gamma^{TM}(\theta) = \frac{-2j\left(\dfrac{\varepsilon_r}{N(\theta)}\right)\cos\theta}{\tan(k_0 h\, N(\theta)) - j\dfrac{\varepsilon_r}{N(\theta)}\cos\theta}$$

$$N(\theta) = \sqrt{\varepsilon_r - \sin^2(\theta)}$$

Circular Patch: Input Resistance



$$R_{in} \approx R_{edge} \left[\frac{J_1^2\left(k\rho_0\right)}{J_1^2\left(ka\right)} \right]$$

Circular Patch: Input Resistance (cont.)

$$R_{edge} = \left[\frac{1}{2P_{sp}}\right] e_r$$

e_{r} = radiation efficiency

where

$$P_{sp} = \frac{\pi}{8\eta_0}(k_0 a)^2 \int_0^{\pi/2} \mathrm{tanc}^2\left(k_0 h N(\theta)\right) \cdot \left[\left|Q(\theta)\right|^2 J_1'^{\,2}(k_0 a \sin\theta) + \left|P(\theta)\right|^2 J_{inc}^2(k_0 a \sin\theta) \right] \sin\theta \, d\theta$$

$$J_{inc}(x) = J_1(x)/x$$

P_{sp} = power radiated into space by circular patch with maximum edge voltage of one volt.

Circular Patch: Input Resistance (cont.)

CAD Formula:

$$P_{sp} = \frac{\pi}{8\eta_0}(k_0 a)^2 I_c$$

$$I_c = \frac{4}{3} p_c$$

$$p_c = \sum_{k=0}^{6} \left(k_0 a\right)^{2k} e_{2k}$$

$$e_0 = 1$$

$$e_2 = -0.400000$$

$$e_4 = 0.0785710$$

$$e_6 = -7.27509 \times 10^{-3}$$

$$e_8 = 3.81786 \times 10^{-4}$$

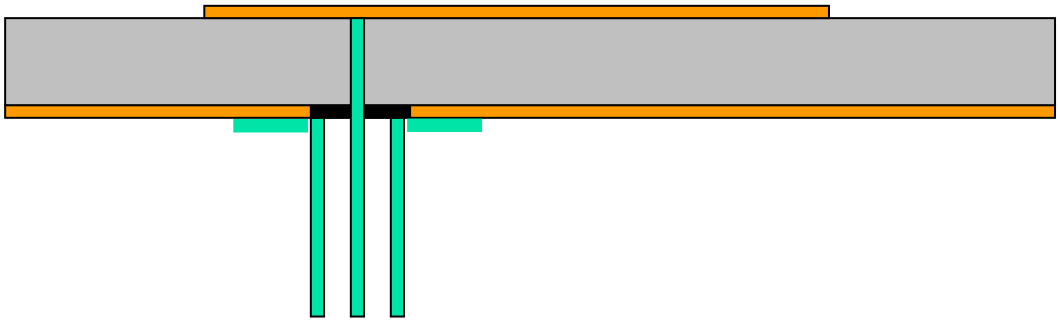
$$e_{10} = -1.09839 \times 10^{-5}$$

$$e_{12} = 1.47731 \times 10^{-7}$$

Feeding Methods

Some of the more common methods for feeding microstrip antennas are shown.

Feeding Methods: Coaxial Feed



$$R = R_{edge} \cos^2\left(\frac{\pi x_0}{L}\right)$$

Advantages:

- simple
- easy to obtain input match

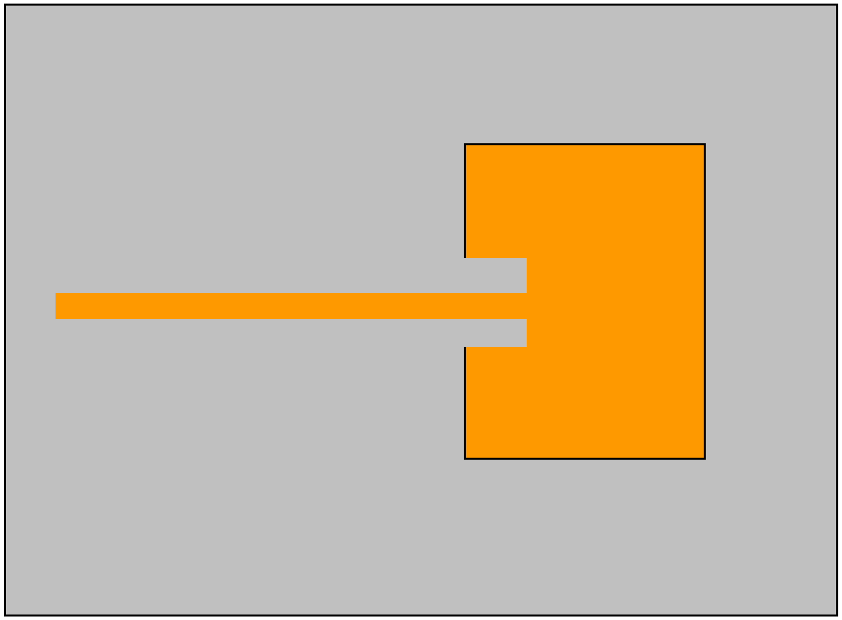
Disadvantages:

- difficult to obtain input match for thicker substrates, due to probe inductance.
- significant probe radiation for thicker substrates

Feeding Methods: Inset-Feed

Advantages:

- simple
- allows for planar feeding
- easy to obtain input match



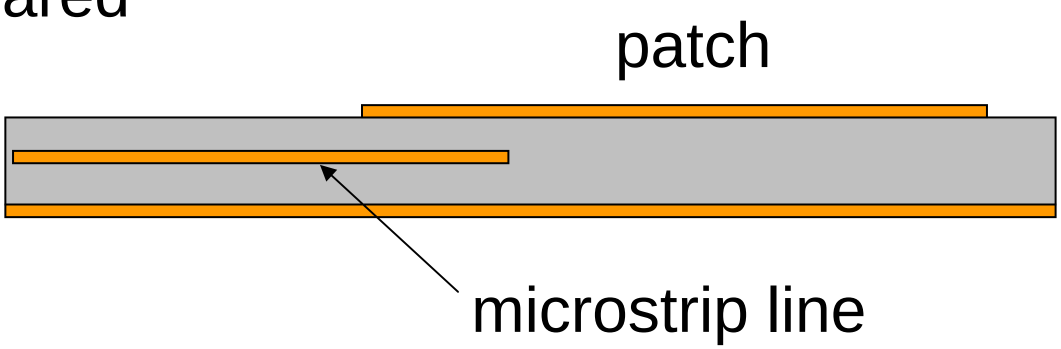
Disadvantages:

- significant line radiation for thicker substrates
- for deep notches, pattern may shown distortion.

Feeding Methods: Proximity (EMC) Coupling

Advantages:

- allows for planar feeding
- less line radiation compared to microstrip feed



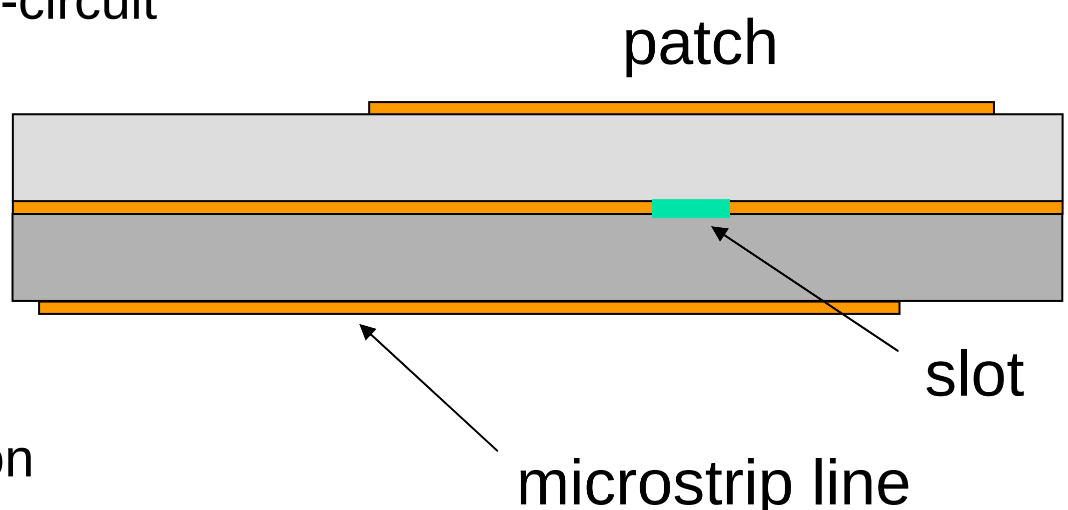
Disadvantages:

- requires multilayer fabrication
- alignment is important for input match

Feeding Methods: Aperture Coupled Patch (ACP)

Advantages:

- allows for planar feeding
- feed radiation is isolated from patch radiation
- higher bandwidth, since probe inductance problem restriction is eliminated and a double-resonance can be created.
- allows for use of different substrates to optimize antenna and feed-circuit performance



Disadvantages:

- requires multilayer fabrication
- alignment is important for input match

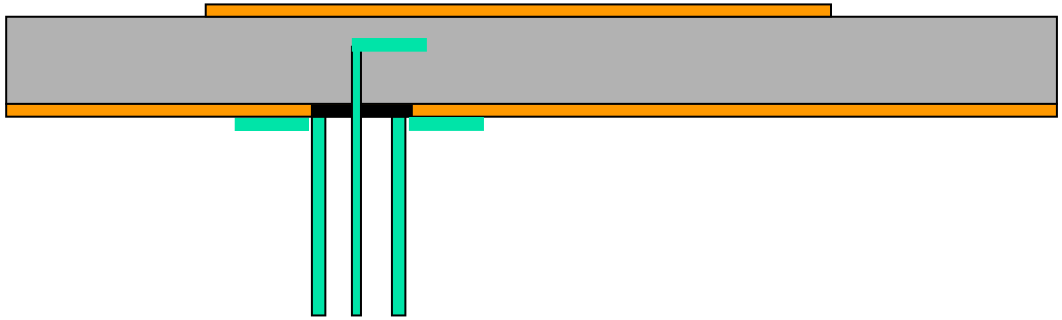
Improving Bandwidth

Some of the techniques that has been successfully developed are illustrated here.

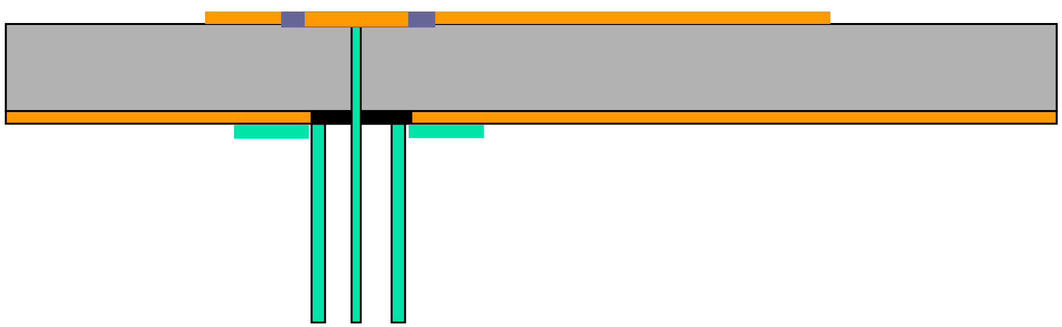
(The literature may be consulted for additional designs and modifications.)

Improving Bandwidth: Probe Compensation

L-shaped probe:



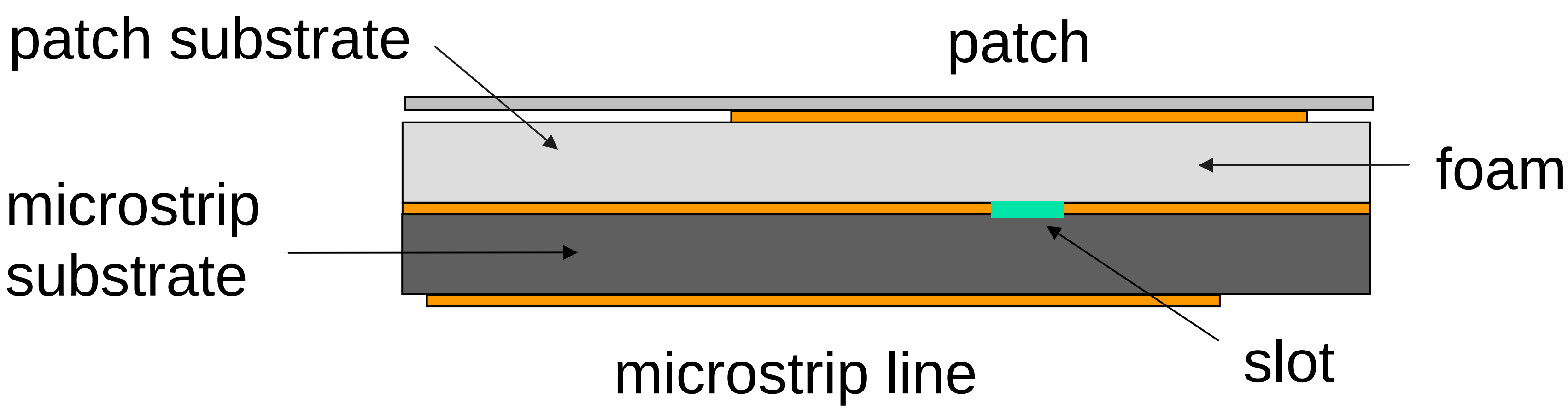
capacitive “top hat” on probe:



Improving Bandwidth: SSFIP

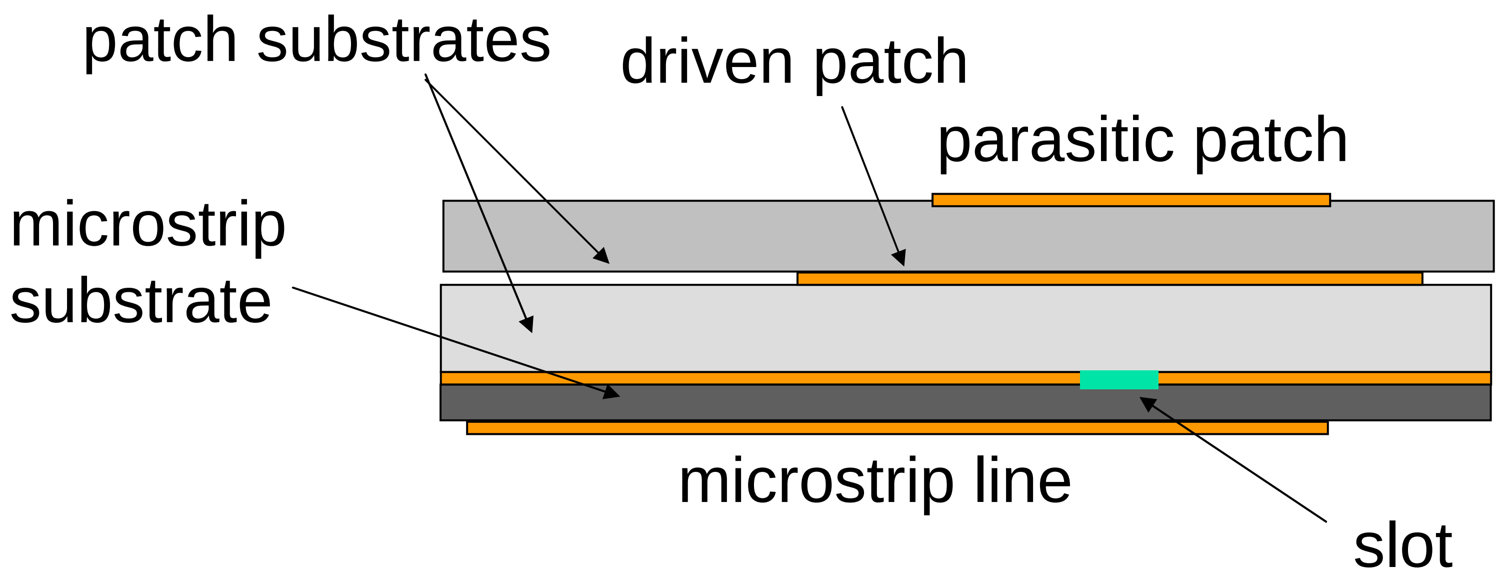
SSFIP: Strip Slot Foam Inverted Patch (a version of the ACP).

- Bandwidths greater than 25% have been achieved.
- Increased bandwidth is due to the thick foam substrate and also a dual-tuned resonance (patch+slot).

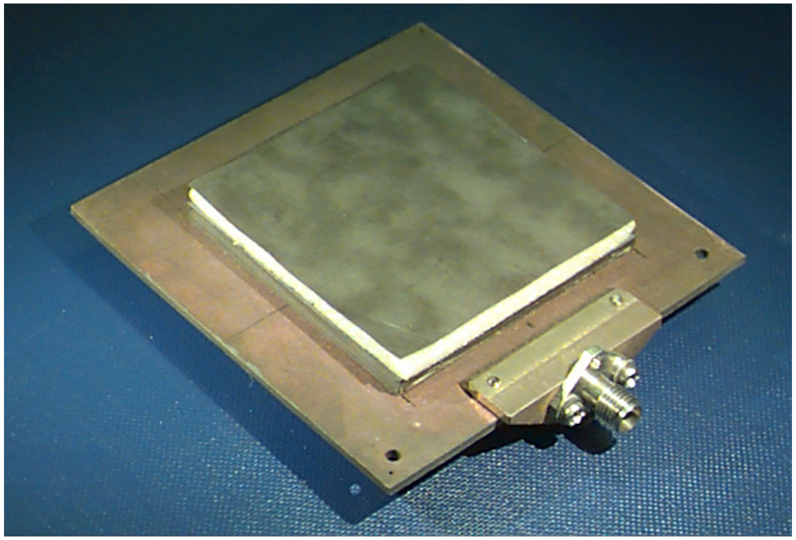


Improving Bandwidth: Stacked Patches

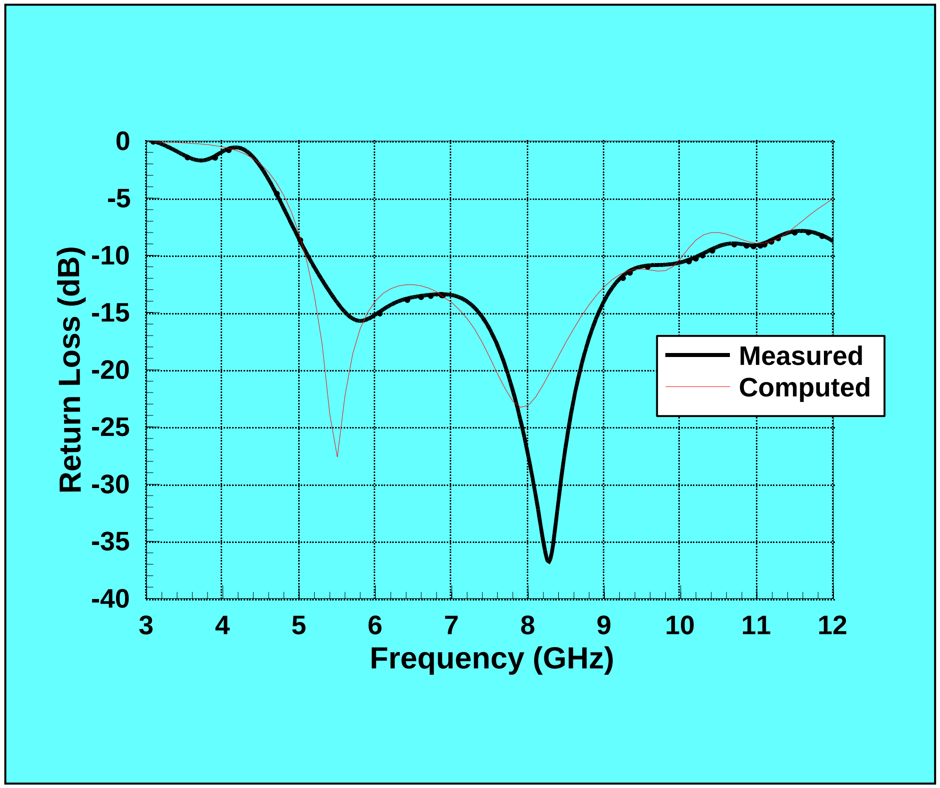
- Bandwidth increase is due to thick low-permittivity antenna substrates and a dual or triple-tuned resonance.
- Bandwidths of 25% have been achieved using a probe feed.
- Bandwidths of 100% have been achieved using an ACP feed.



Improving Bandwidth: Stacked Patches (cont.)

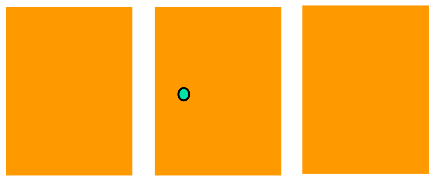


stacked patch with ACP feed

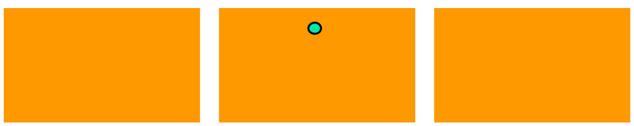


-10 dB S_{11} bandwidth is about 100%

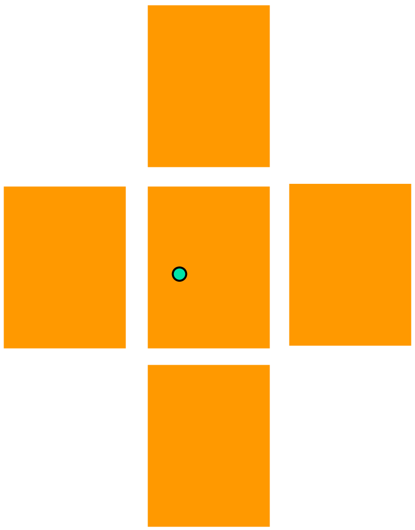
Improving Bandwidth: Parasitic Patches



Radiating Edges Gap Coupled Microstrip Antennas (REGCOMA).



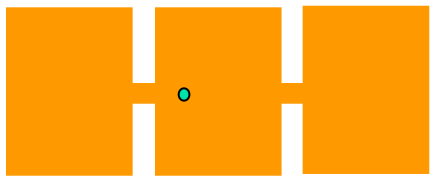
Non-Radiating Edges Gap Coupled Microstrip Antennas (NEGCOMA)



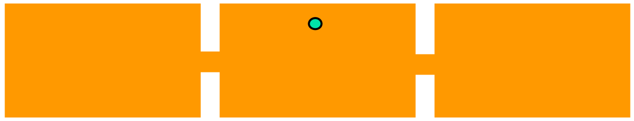
Four-Edges Gap Coupled Microstrip Antennas (FEGCOMA)

Bandwidth improvement factor:
REGCOMA: 3.0, NEGCOMA: 3.0, FEGCOMA: 5.0?

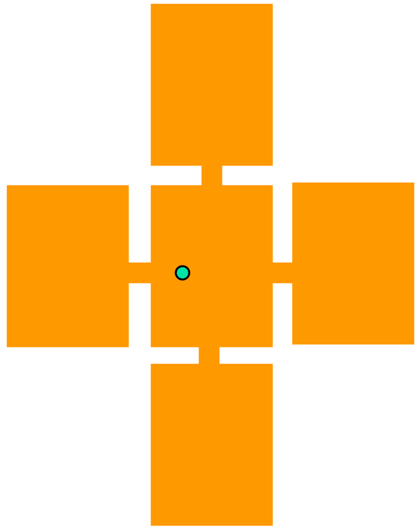
Improving Bandwidth: Direct-Coupled Patches



Radiating Edges Direct Coupled Microstrip Antennas (REDCOMA).



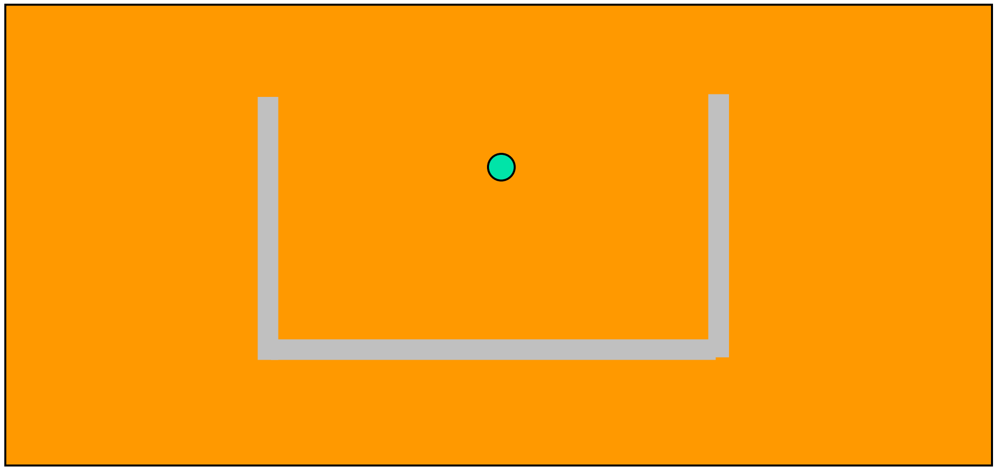
Non-Radiating Edges Direct Coupled Microstrip Antennas (NEDCOMA)



Four-Edges Direct Coupled Microstrip Antennas (FEDCOMA)

Bandwidth improvement factor:
REDCOMA: 5.0, NEDCOMA: 5.0, FEDCOMA: 7.0

Improving Bandwidth: U-shaped slot

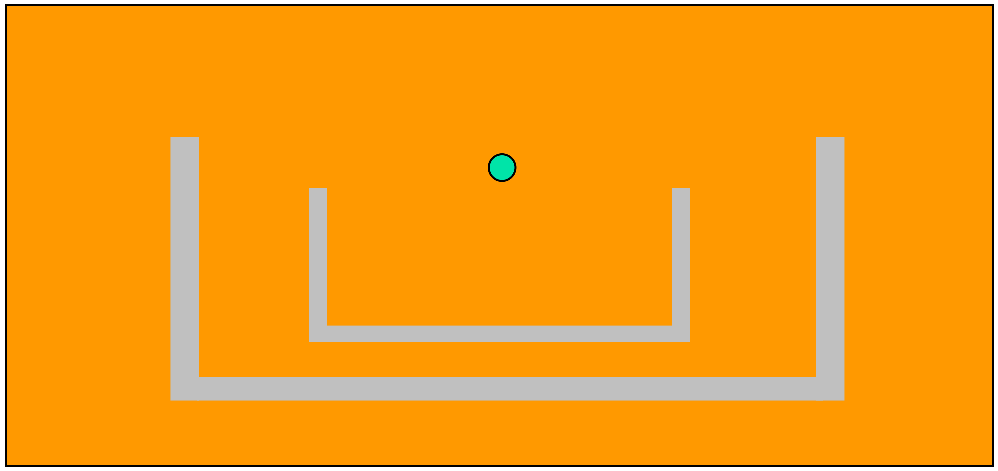


The introduction of a U-shaped slot can give a significant bandwidth (10%-40%).

(This is partly due to a double resonance effect.)

“Single Layer Single Patch Wideband Microstrip Antenna,” T. Huynh and K. F. Lee, Electronics Letters, Vol. 31, No. 16, pp. 1310-1312, 1986.

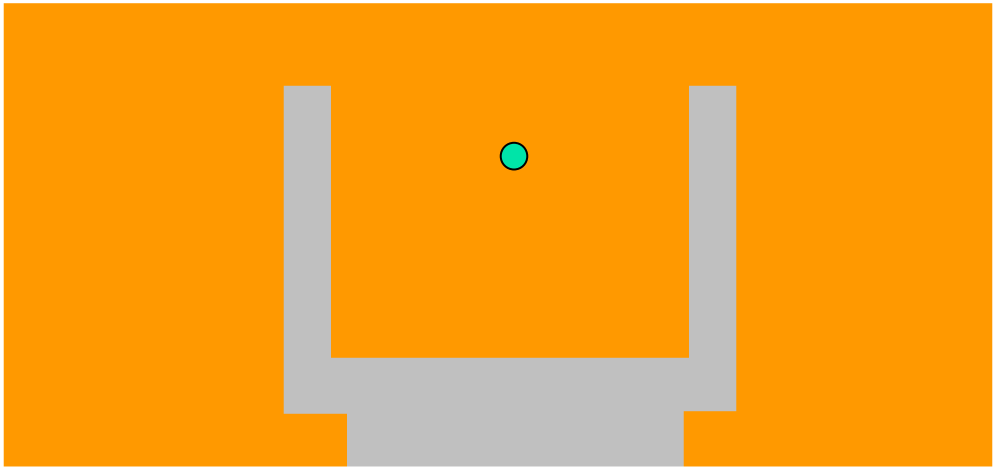
Improving Bandwidth: Double U-Slot



A 44% bandwidth was achieved.

"Double U-Slot Rectangular Patch Antenna," Y. X. Guo, K. M. Luk, and Y. L. Chow, Electronics Letters, Vol. 34, No. 19, pp. 1805-1806, 1998.

Improving Bandwidth: E-Patch



A modification of the U-slot patch.

A bandwidth of 34% was achieved (40% using a capacitive "washer" to compensate for the probe inductance).

"A Novel E-shaped Broadband Microstrip Patch Antenna," B. L. Ooi and Q. Shen, Microwave and Optical Technology Letters, Vol. 27, No. 5, pp. 348-352, 2000.

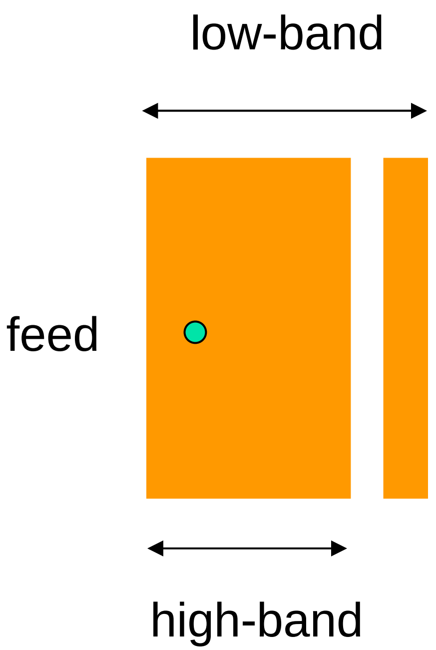
Multi-Band Antennas

A multi-band antenna is often more desirable than a broad-band antenna, if multiple narrow-band channels are to be covered.

General Principle:

Introduce multiple resonance paths into the antenna. (The same technique can be used to increase bandwidth via multiple resonances, if the resonances are closely spaced.)

Multi-Band Antennas: Examples



Dual-Band E patch

Dual-Band Patch with Parasitic Strip

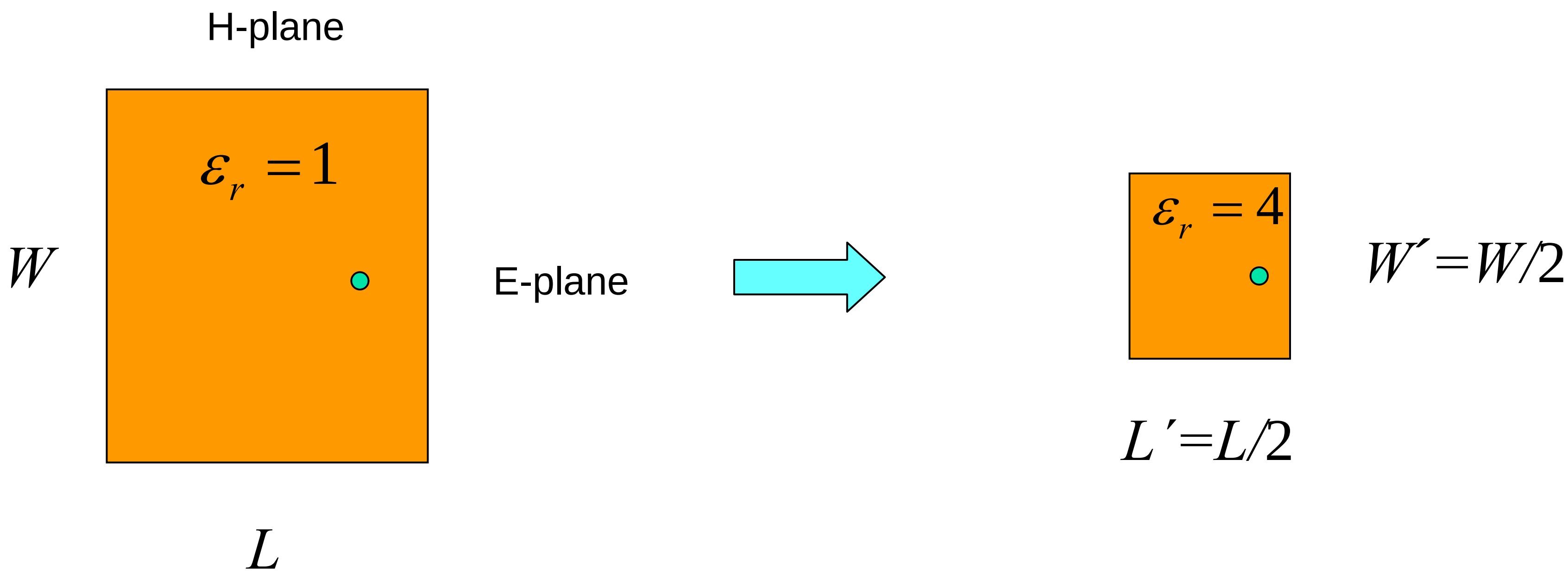
Miniaturization

- High Permittivity
- Quarter-Wave Patch
- PIFA
- Capacitive Loading
- Slots
- Meandering

Note: miniaturization usually comes at a price of reduced bandwidth.

General rule: maximum obtainable bandwidth is proportional to the volume of the patch (based on the Chu limit.)

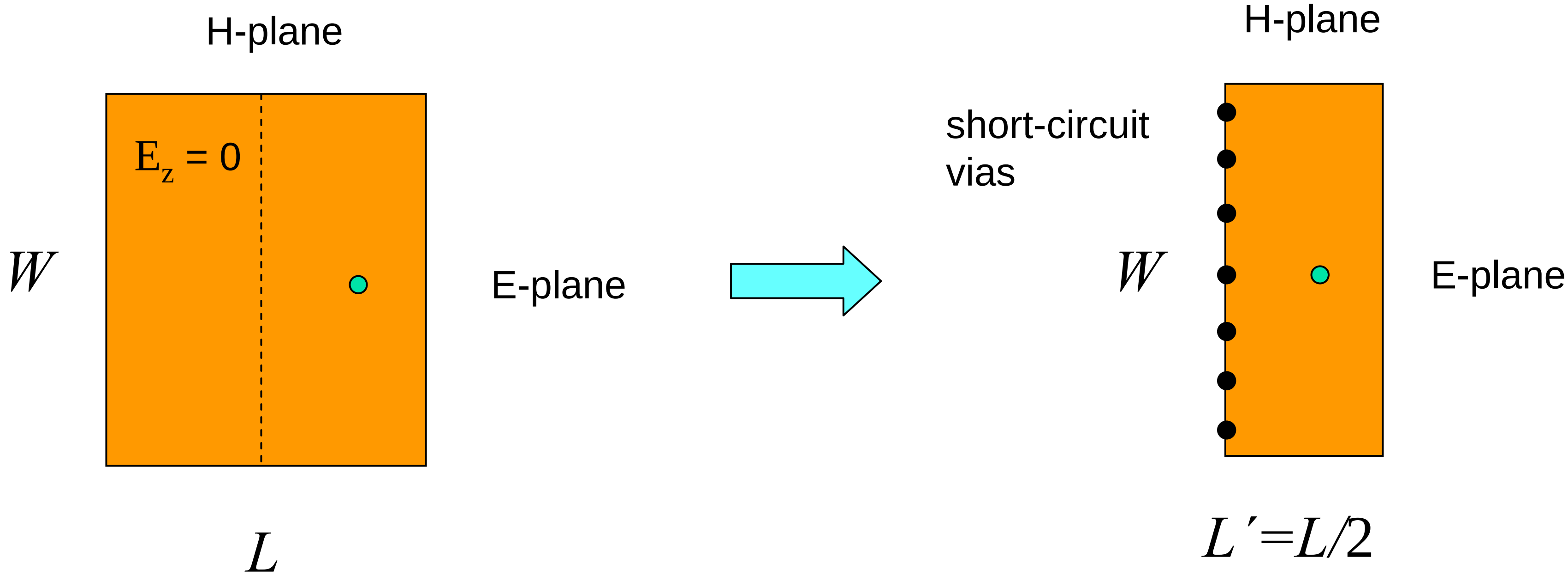
Miniaturization: High Permittivity



It has about one-fourth the bandwidth of the regular patch.

(Bandwidth is inversely proportional to the permittivity.)

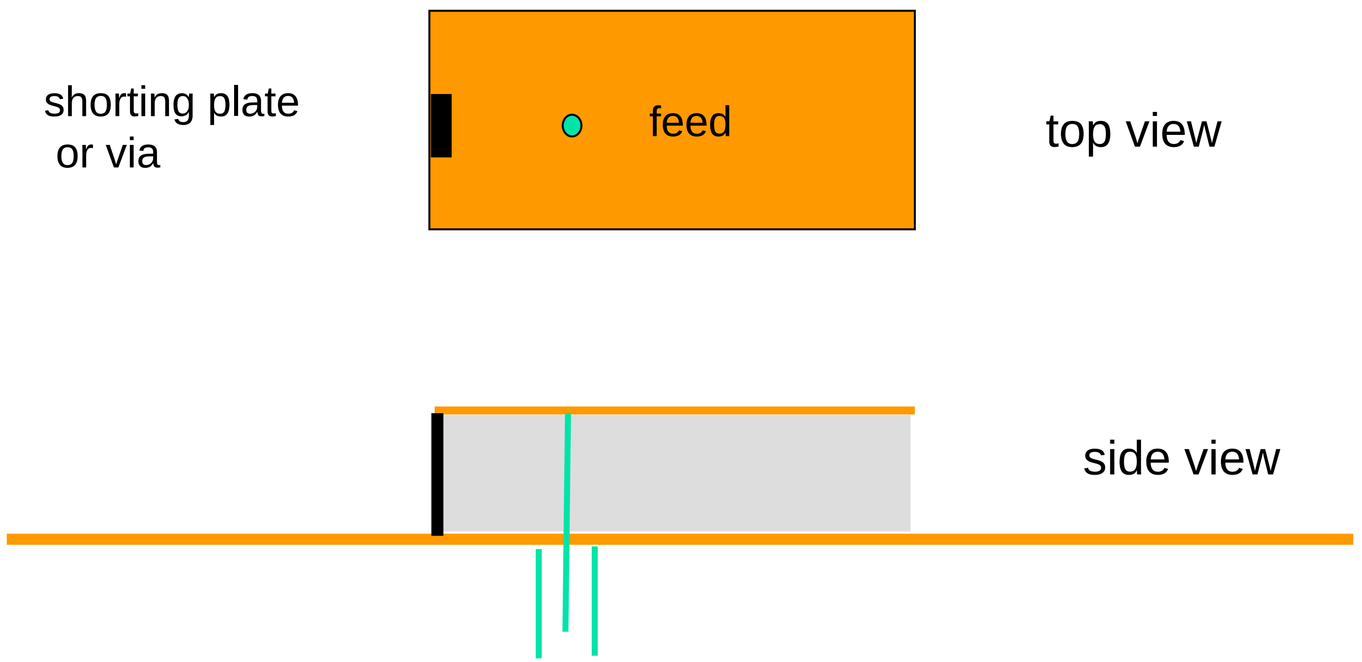
Miniaturization: Quarter-Wave Patch



It has about one-half the bandwidth of the regular patch.

Miniaturization: Planar Inverted F Antenna (PIFA)

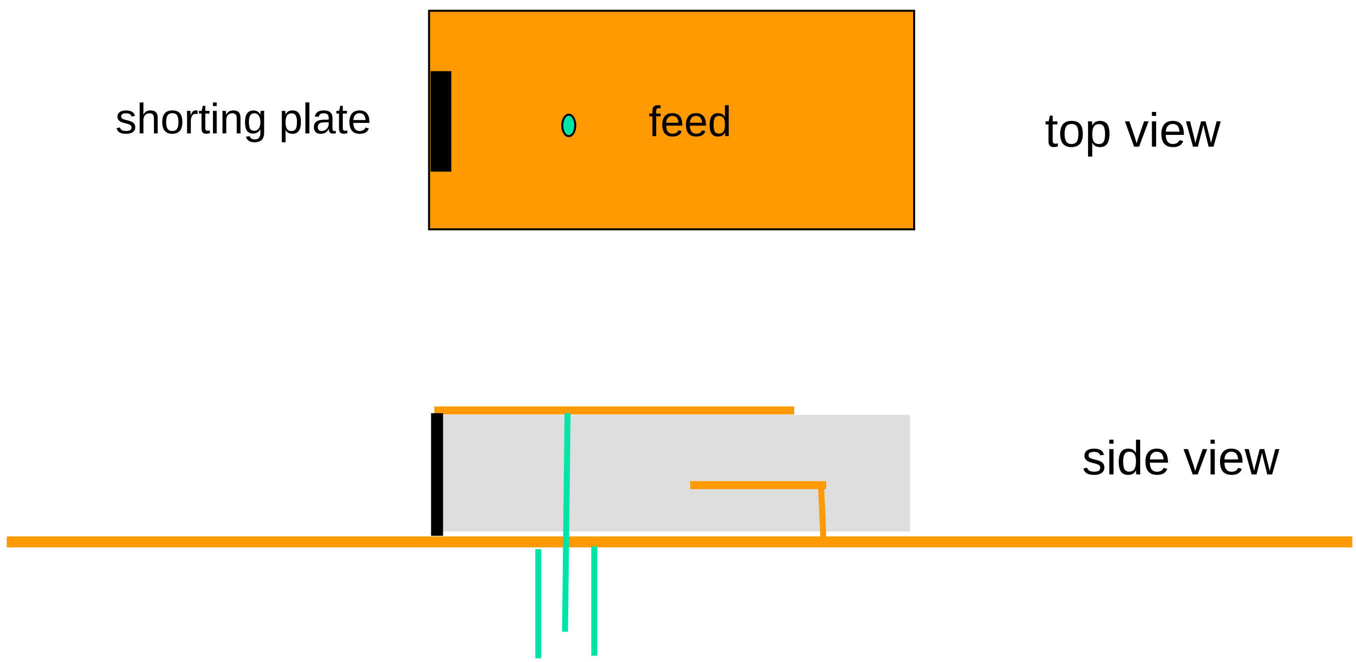
$$L < \lambda_d / 4$$



A single shorting plate or via is used.

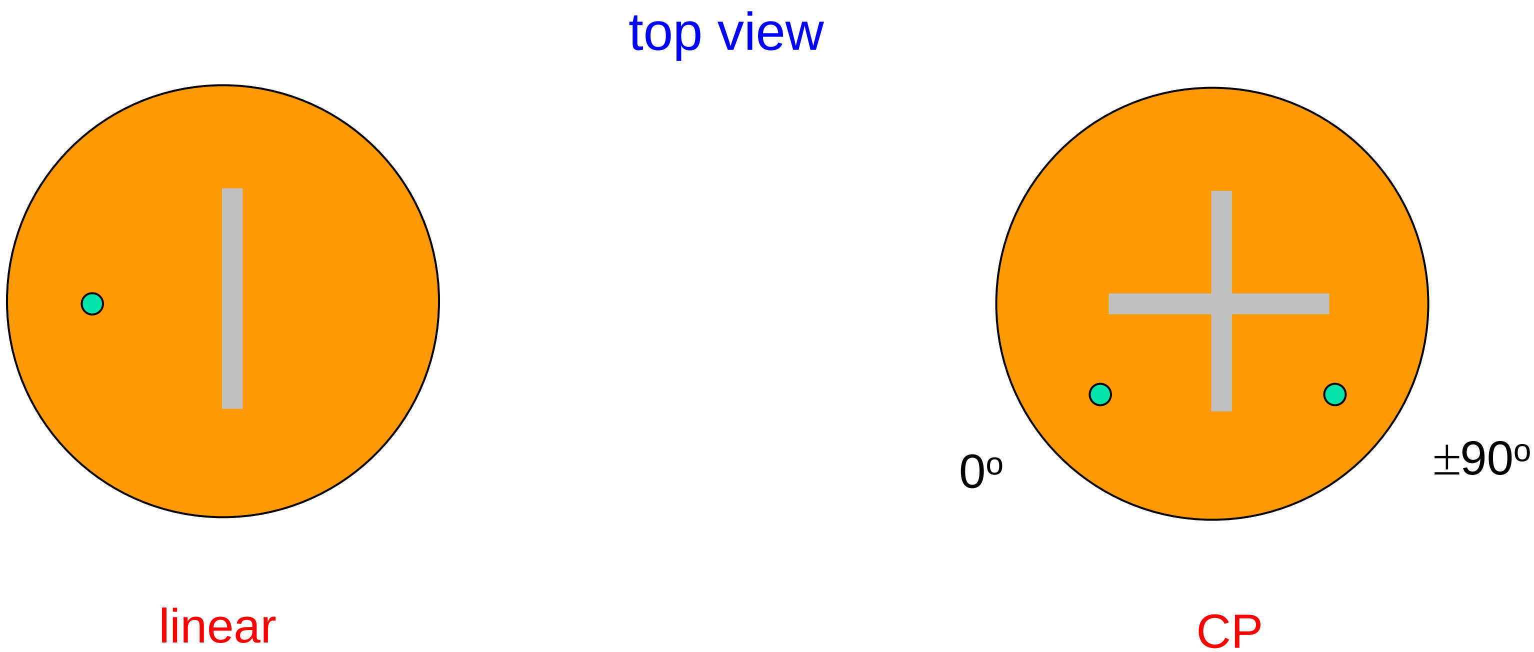
This antenna can be viewed as a limiting case of the quarter-wave patch, or as an LC resonator.

PIFA with Capacitive Loading



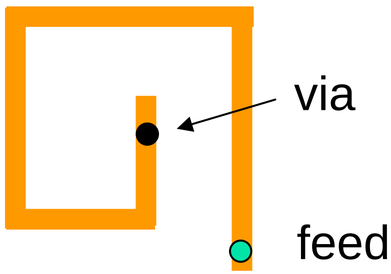
The capacitive loading allows for the length of the PIFA to be reduced.

Miniaturization: Slotted Patch

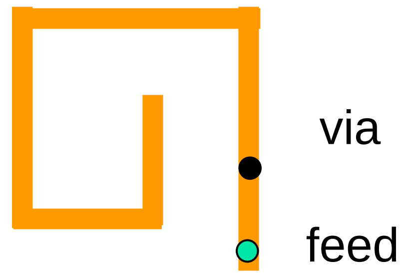


The slot forces the current to flow through a longer path, increasing the effective dimensions of the patch.

Miniaturization: Meandering



meandered quarter-wave patch

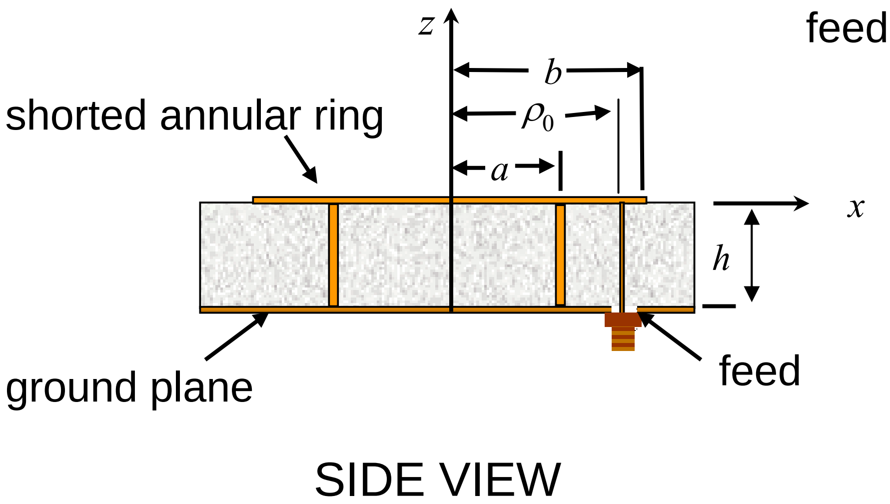


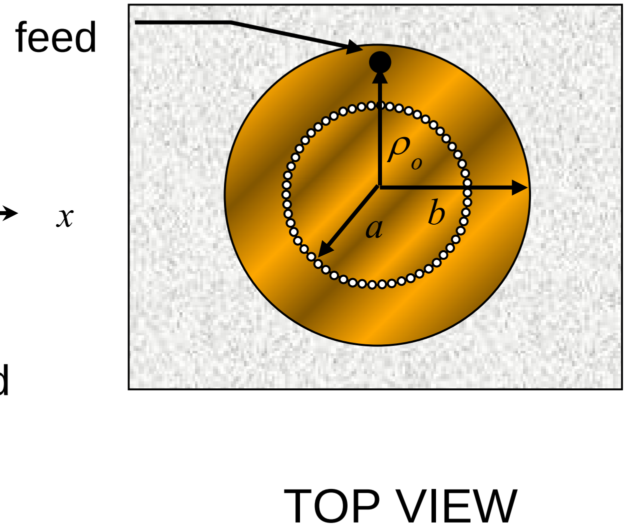
meandered PIFA

Meandering forces the current to flow through a longer path, increasing the effective dimensions of the patch.

Improving Performance:

Reducing Surface-Wave Excitation and Lateral Radiation



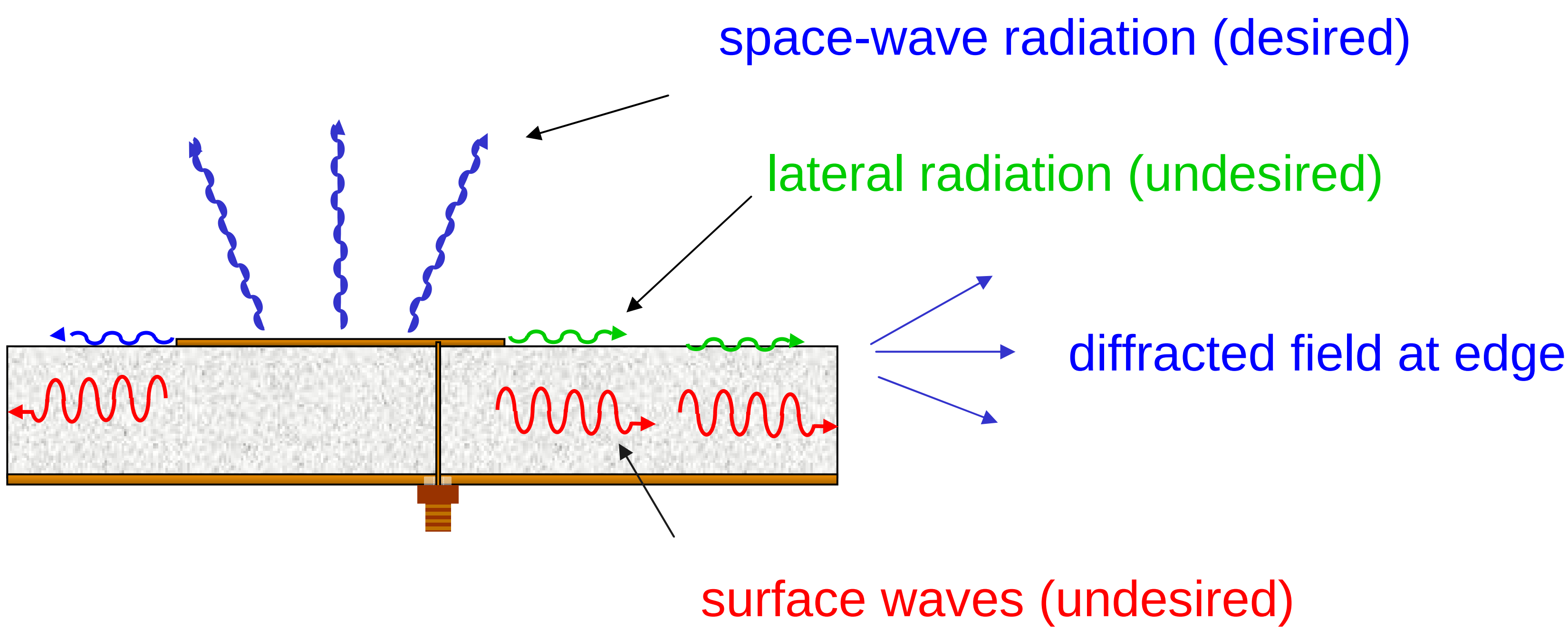


Reduced Surface Wave (RSW) Antenna

D. R. Jackson, J. T. Williams, A. K. Bhattacharyya, R. Smith, S. J. Buchheit, and S. A. Long, “Microstrip Patch Designs that do Not Excite Surface Waves,” IEEE Trans. Antennas Propagat., vol. 41, No 8, pp. 1026-1037, August 1993.

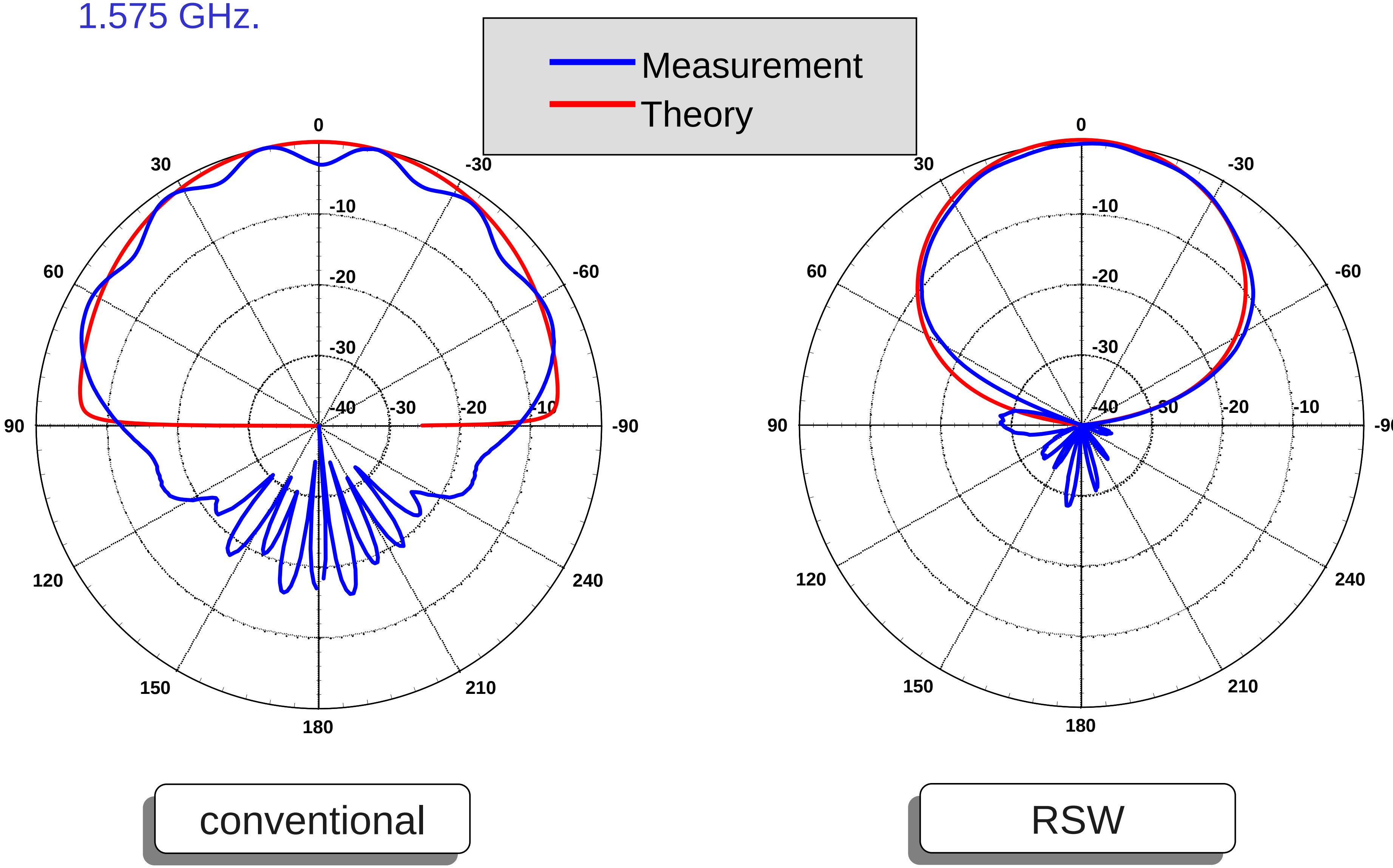
RSW: Improved Patterns

Reducing surface-wave excitation and lateral radiation reduces edge diffraction.



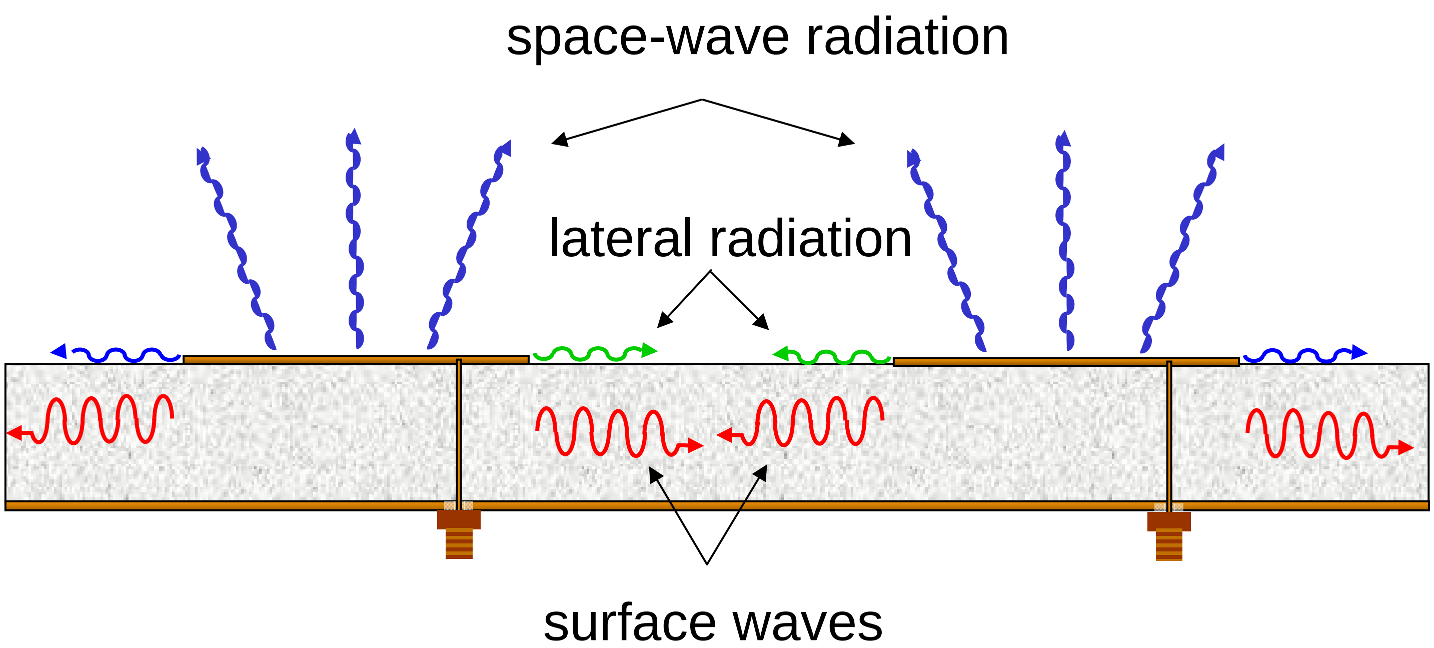
RSW: E-plane Radiation Patterns

Measurements were taken on a 1 m diameter circular ground plane at 1.575 GHz.



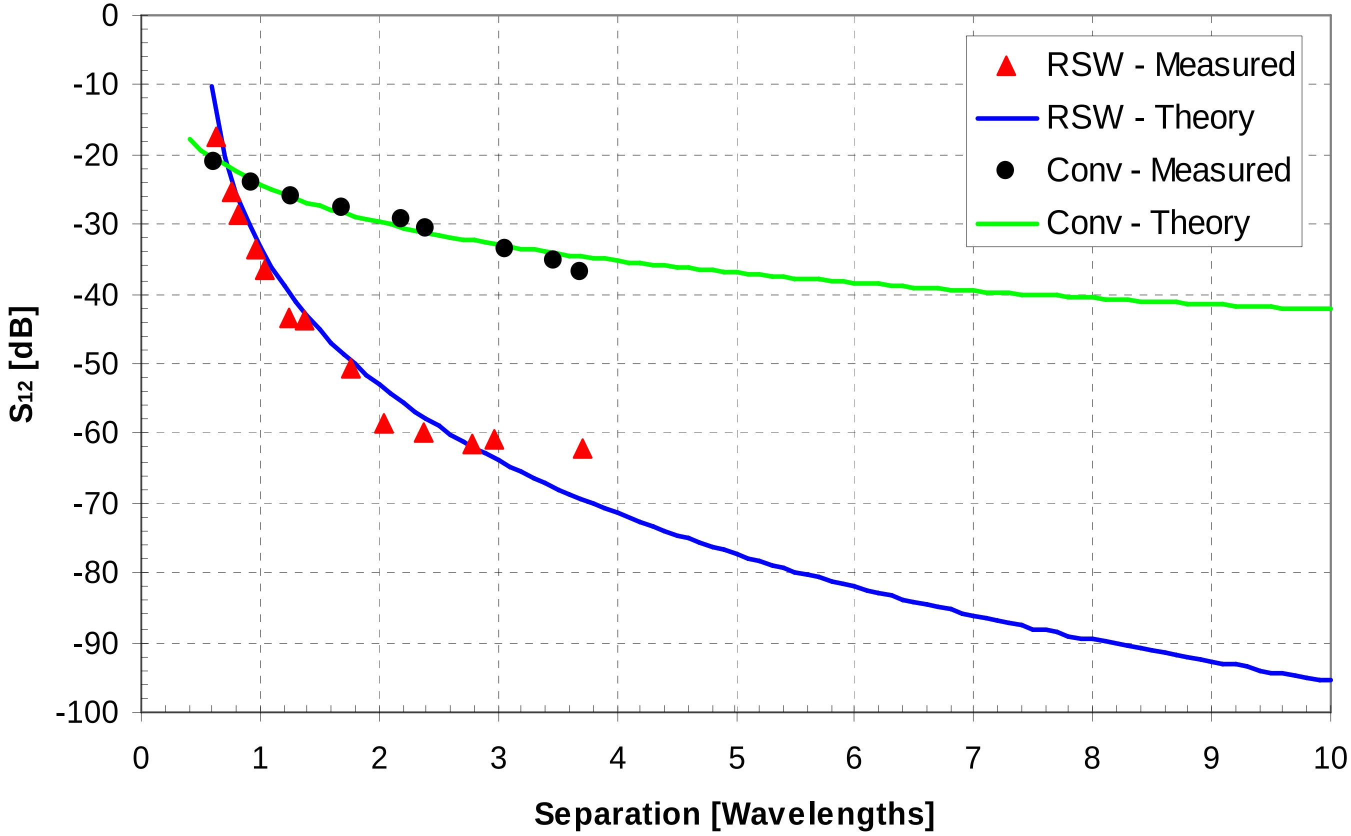
RSW: Mutual Coupling

Reducing surface-wave excitation and lateral radiation reduces mutual coupling.



RSW: Mutual Coupling (cont.)

Reducing surface-wave excitation and lateral radiation reduces mutual coupling.



"Mutual Coupling Between Reduced Surface-Wave Microstrip Antennas," M. A. Khayat, J. T. Williams, D. R. Jackson, and S. A. Long, IEEE Trans. Antennas and Propagation, Vol. 48, pp. 1581-1593, Oct. 2000.

References

General references about microstrip antennas:

Microstrip and Printed Antenna Design, Randy Bancroft, Noble Publishers, 2004.

Microstrip Patch Antennas: A Designer's Guide, Rodney B. Waterhouse, Kluwer Academic Publishers, 2003.

Microstrip Antenna Design Handbook, R. Garg, P. Bhartia, I. J. Bahl, and A. Ittipiboon, Editors, Artech House, 2001.

Advances in Microstrip and Printed Antennas, K. F. Lee, Editor, John Wiley, 1997.

Microstrip Antennas: The Analysis and Design of Microstrip Antennas and Arrays, David M. Pozar and Daniel H. Schaubert, Editors, Wiley/IEEE Press, 1995.

References (cont.)

General references about microstrip antennas (cont.):

Millimeter-Wave Microstrip and Printed Circuit Antennas, P. Bhartia, Artech House, 1991.

The Handbook of Microstrip Antennas (two volume set), J. R. James and P. S. Hall, INSPEC, 1989.

Microstrip Antenna Theory and Design, J. R. James, P. S. Hall, and C. Wood, INSPEC/IEE, 1981.

References (cont.)

More information about the CAD formulas presented here for the rectangular patch may be found in:

Computer-Aided Design of Rectangular Microstrip Antennas, D. R. Jackson, S. A. Long, J. T. Williams, and V. B. Davis, Ch. 5 of *Advances in Microstrip and Printed Antennas*, K. F. Lee, Editor, John Wiley, 1997.

References (cont.)

References devoted to broadband microstrip antennas:

Compact and Broadband Microstrip Antennas, Kin-Lu Wong, John Wiley, 2003.

Broadband Microstrip Antennas, Girish Kumar and K. P. Ray, Artech House, 2002.

Broadband Patch Antennas, Jean-Francois Zurcher and Fred E. Gardiol, Artech House, 1995.