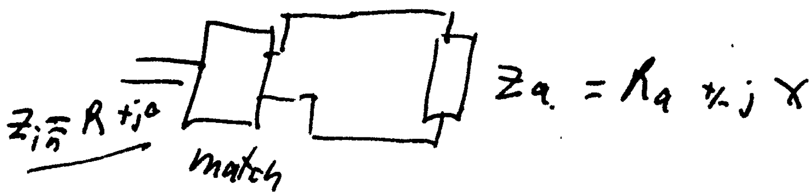
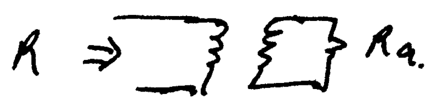
~~[illegible]~~: other matching techiques:



The match 1) transforms R_a to R

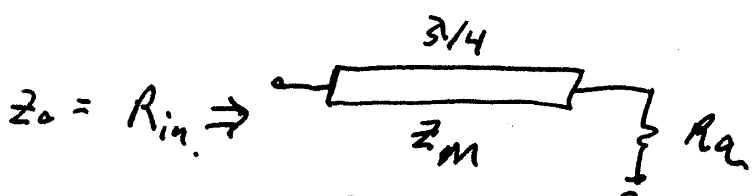


2) cancels the reactive component X

simplest:

the antenna is designed to provide $Z_a = R_a \pm \underline{j0}$ i.e no reactance,

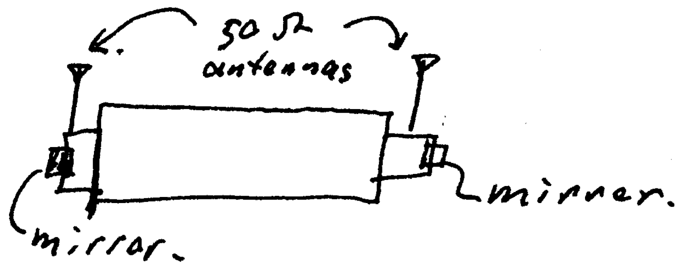
we can then use a quarter wave transmission line:



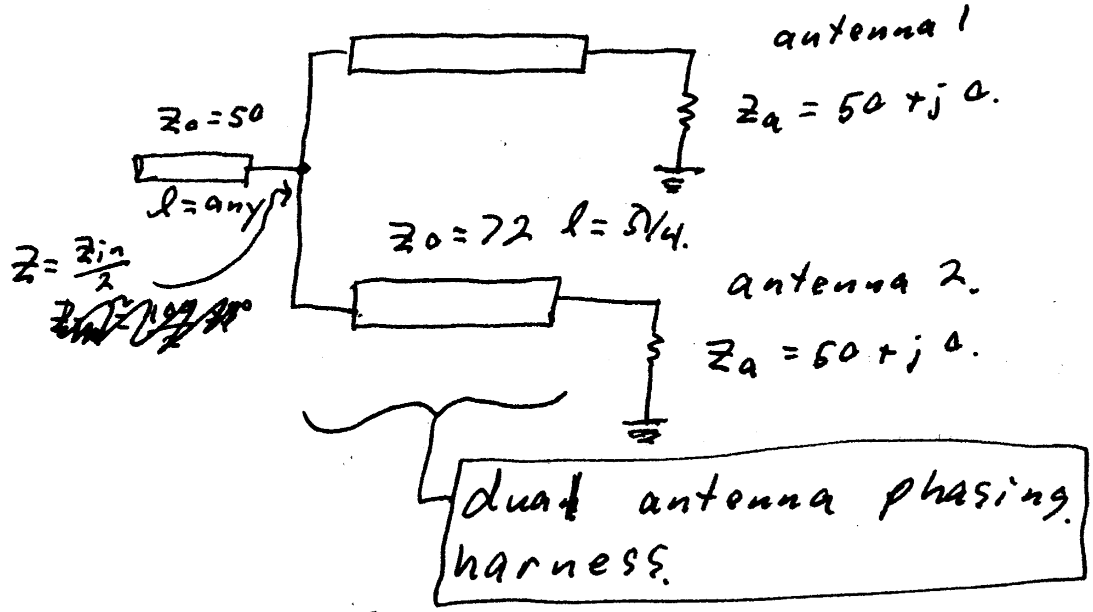
$$R_{in} = \frac{Z_m^2}{R_a}$$

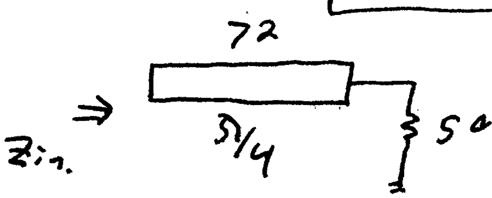
533 ~~54~~ 55

Example → dual CB radio antennas on Semi trucks. Each antenna is ≈ 50 Ω Rin.



model. $Z_o = 72$ $l = \lambda/4$





$$Z_{in} = \frac{72^2}{50} = 103.7\ \Omega.$$

antennas are in parallel so $Z = \frac{103.7}{2} = 51.84$

$$VSWR = \frac{51.84}{50} = 1.04:1$$

533-19.

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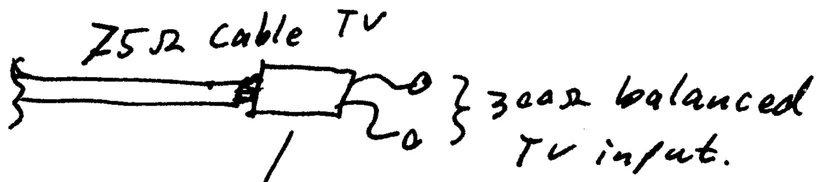
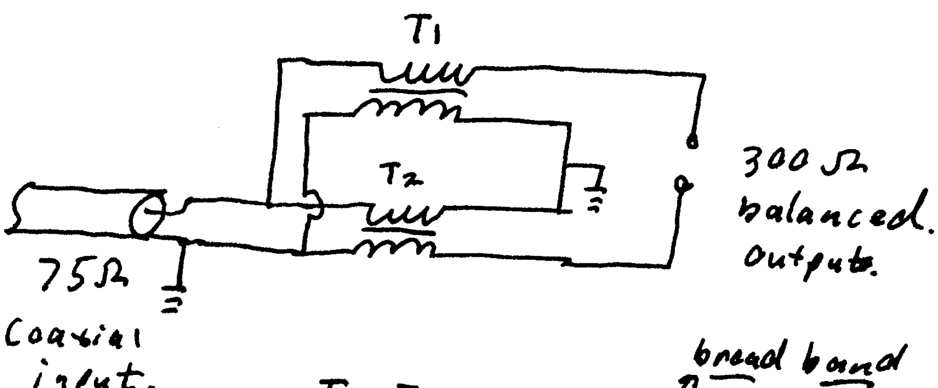
TV balun;



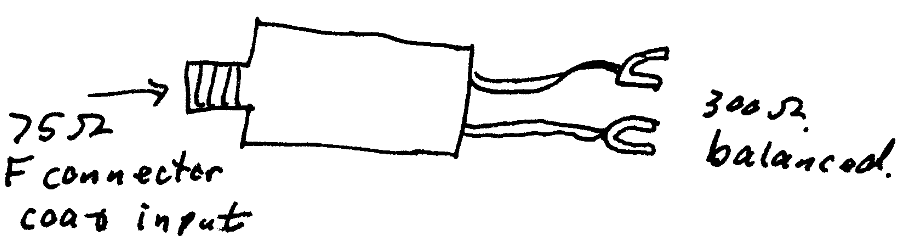
~~Eg.~~ Balun - what is in this?

for TV the Balun must cover a frequency range of ≈ 50 mHz to over 800 mHz.

Broad band!



T_1, T_2 are 1:1 broad band transformers



see section 9.7 in text for other methods

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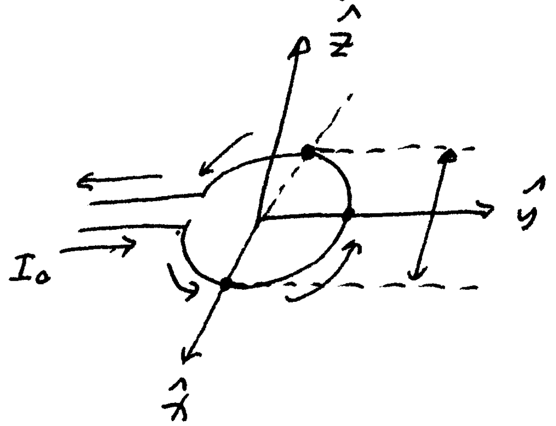
Small Loop Antenna in x-y plane.



assume

$|\bar{K}| = \frac{2\pi}{\lambda} << d$

& $kd << 1$

vector potential (no mag potential)

$$\vec{A} = \frac{e^{-jkr}}{4\pi r} \iiint \vec{J}(r')\, e^{j(\vec{K}\cdot\vec{r}')}\, dv'$$

(remember r' is source vector)

$$E_\theta = -\eta H_\phi = -jk\eta A_\theta \qquad E_\phi = -\eta H_\theta = -jk\eta A_\phi$$

$e^{j\vec{k}\cdot\vec{r}'} \qquad |\vec{k}\cdot\vec{r}'| << 1$

$$e^{jkr} \cong 1 + j\vec{k}\cdot\vec{r}' + \dots H.O.T$$

$$\vec{A} = \frac{e^{-jkr}}{4\pi r}\, jk\sin\theta I_o S\, \hat{\phi}$$

where S is the area inclosed by the loop.

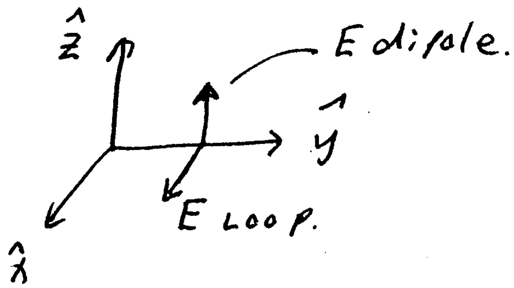
$$S = \pi\left(\frac{d}{2}\right)^2$$

$$\vec{A} = A_\phi\, \hat{\phi}$$

$$E_\theta = 0, \quad E_\phi = -jk\eta A_\phi = \frac{-1}{\eta} H_\theta$$

The loop $\bar{E}$ is // to xy plane.

dipole is $\perp$ to xy plane.



Consider the tota radiated power.

$$P_r = \iint d\Omega \, P(\theta, \phi)$$

$$= \int_0^{2\pi} \int_0^{\pi} \sin\theta \, d\theta \, d\phi \, \frac{1}{2} \frac{|E|^2}{\eta}$$

$$P_r = 20 I_o^2 (K^2 S)^2 \text{ watts.}$$

so the radiation. resistance is.

$$R_r = \frac{2P_r}{I_o^2} = 20 (K^2 S)^2 \ \underline{\Omega}$$

note: factor of 1/2

533-9.20

note; there was a question in class about R_r in the book vs ~~my~~ form: (eq 2-54 p 238.)

are they the same?

eq 5-24 $$R_r = \frac{\eta 2\pi}{3}\left(\frac{KS}{\lambda}\right)^2 = \frac{\eta}{3\cdot 2\pi}\left(\frac{KS\cdot 2\pi}{\lambda}\right)^2$$

$$= \frac{\eta}{3\cdot 2\pi}\left(\frac{K^2 S}{\cancel{\lambda}}\right)^2 = \frac{120\pi}{3\cdot 2\pi}\left(\frac{K^2 S}{\cancel{\lambda}}\right)^2 = \frac{60}{3}\left(\frac{K^2 S}{\cancel{\lambda}}\right)^2 = 20\left(K^2 S\right)^2 \therefore$$

they are the same!

I like the form $R_r = f(\text{wave vector, Area})$

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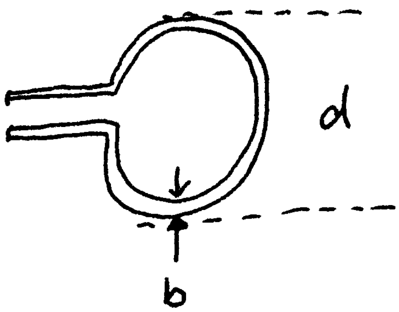
EE533-21

Example - single loop.



$b = 0.0812\,mm$ (20 AWG wire)

$f = 1\,MHz \Rightarrow \lambda = 300\,m$

$d = 2\,cm$

$\Rightarrow R_r = 20(k^2 S)^2$

$S = \pi\left(\frac{d}{2}\right)^2 \quad \Rightarrow R_r = 3.8 \times 10^{-13}\,\Omega.$

ohmic Loss:

$R_s = \sqrt{\frac{\mu_0 \pi f}{\sigma}}$ $\qquad \sigma = 5.7 \times 10^{-7}\,℧/m$ copper

$= 2.63 \times 10^{-4}\,\Omega.$

$R_\Omega = R_s \dfrac{L = \text{length}}{\text{circumference}} = R_s \dfrac{\pi b}{\pi d}$

$R_\Omega = 0.0648\,\Omega.$

The effeciency $\varepsilon = \dfrac{R_r}{R_\Omega + R_r} \cong 0\%$!

If we repeat the example with ferrite and multiple turns:

$N = 100$ turns $\quad \mu_r = \mu_{eff} = 20{,}000$

now $R_{r_N} = 1.52\,\Omega \quad R_\Omega = 6.48\,\Omega$.

$\mathcal{E}$fficiency $= 19\% \Rightarrow$ Low, but usable.

the Directivity & far field patterns remain the same.

but Gain↑ because η_{eff}↑

Reactance of the Loop $\quad Z_a = R_a + jX$

d = diameter of Loop
b = diameter of wire.
it may be shown:

$$L_o = \frac{d}{2}\mu\left[\ln\left(\frac{8d}{b}\right) - 1.75\right]$$

for our single loop example

$$L_o = 4.48\times10^{-8}\,H = 44.8\,nH$$

$$X_o = j\omega_o L_o = \underline{j\,0.279\,\Omega}.$$

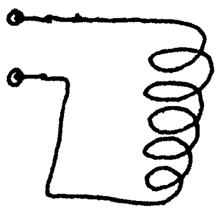
for multiple loops. $\quad L = \mu_{eff} N L_o$

for the example $X = 5.58\times10^{5}\,\Omega$

EE 533-21

multi-turn Loops

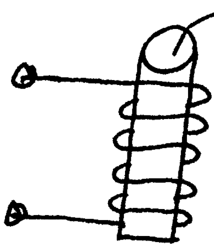
we usually use multi-turn loops at low f



$$R_{rN} = 20(K^2 S N)^2$$

N = # of Loops.

we often Load the Antenna with ferrite. (example - AM radio Antenna)



$\mu = \mu_{eff}\,\mu_0$

↳ effective μ_r dependent upon physical dimension and material μ_r.

the material is ferrite $20 < \mu_r < 20{,}000$

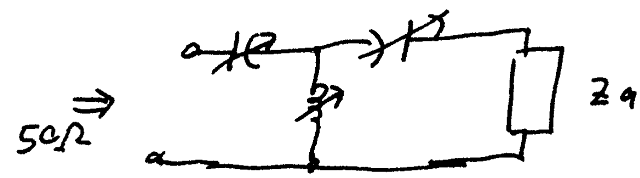
$$R_{rN}\Big|_{\substack{\text{ferrite}\\ \text{load}}} = 20\left(N \mu_{eff} K^2 S\right)^2$$

EE533-22

Low frequency Lumped element matching.

for antennas < 30 mHz we use lumped element matching networks.

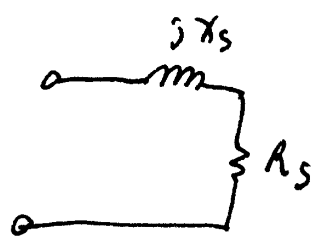
example - the antenna matcher we looked at in class.



a simple method is the Q transform.

it uses 2-elements to perform the match. A variation of this is used in AM radios.

Q - transform.



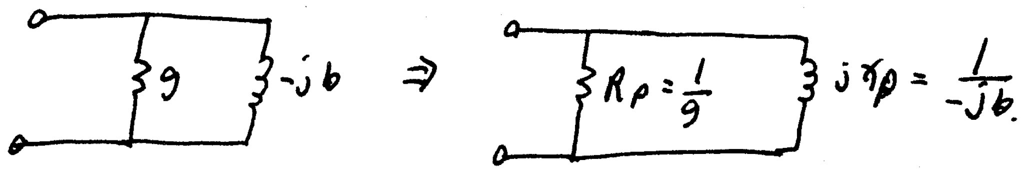
$$Q \triangleq \frac{|X_s|}{R_s}$$

$$Z = R_s + jX_s$$

$$Y = \frac{1}{Z} = \frac{1}{R_s + jX_s} = \frac{R_s}{R_s^2 + X_s^2} - \frac{jX_s}{R_s^2 + X_s^2} = g + jb.$$

EE 533-21

So the parallel equivalent is:



$$R_p = \frac{1}{g} = \frac{R_s^2 + X_s^2}{R_s} = R_s\left(1 + \frac{X_s^2}{R_s^2}\right) = \boxed{R_s\left(1+Q^2\right) = R_p}$$

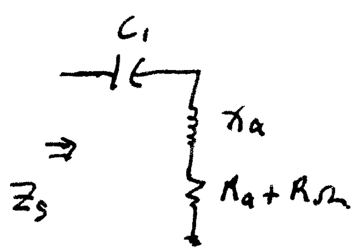
it can also be shown that

$$\boxed{X_p = X_s\left(1 + \frac{1}{Q^2}\right)}$$ (capacitive or inductive X)

note that $Q = \frac{|X_s|}{R_s} = \frac{R_p}{|X_p|}$.

apply this to our multiLoop example;
use capacitors for the match;

terminal Z
$Z_A = R_a + R_\Omega + j X_a = 8 + j\, 5.58 \times 10^5\ \Omega.$



$$Z_s = 8 + j\left(558K - X_{C_1}\right)$$

EE533-21

we want $R_p = 50 = 8(1+Q^2)$

$\Rightarrow Q = 2.29.$

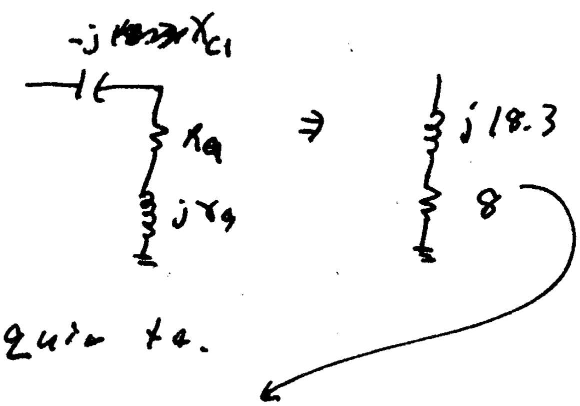
$Q = \frac{|X_a - X_{c1}|}{R_S}$ $\qquad |X_a - X_{c1}| = Q \cdot R_s = 2.29 \cdot 8$

$|X_a - X_{c1}| = 18.3\,\Omega.$

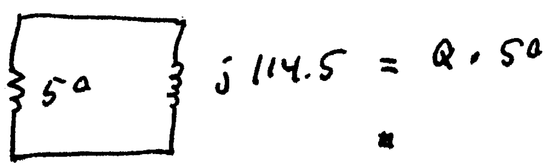
$X_{c1} = 558K - X_{c1} \Rightarrow X_{c1} \cong 558k.$

$X_{c1} = \frac{1}{2\pi C_1} \Rightarrow C_1 \cong .2pf \rightarrow$ (tough todo!)

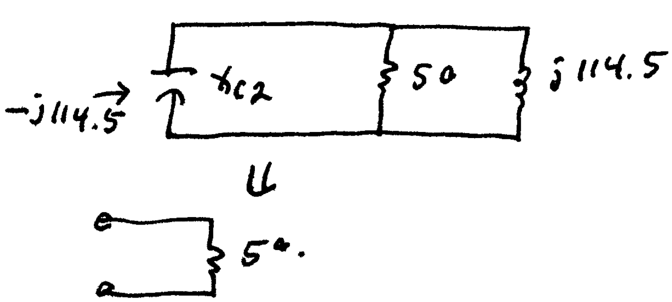
So this:



is now equiv to.



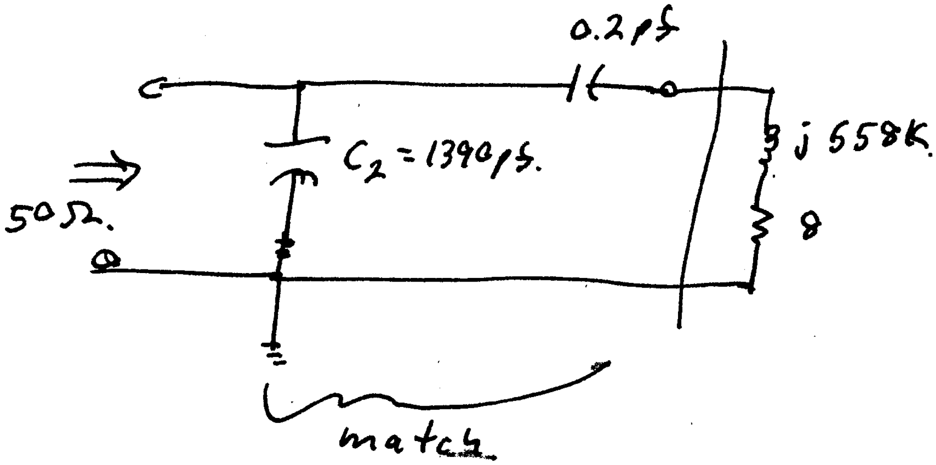
add the shunt C_2.



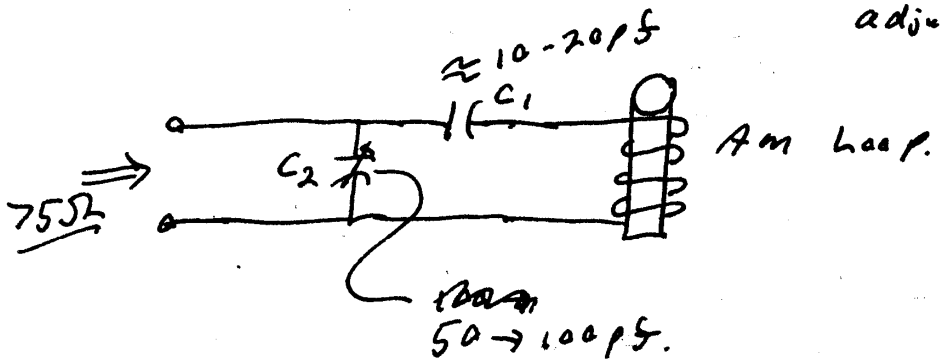
EE 533-21

$$114.5 = \frac{1}{2\pi f C_2} \Rightarrow \quad C_2 = 1.39\,nf = 1390\,pf$$

final match.



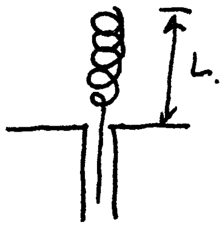
note: typical AM match with tuning adjustment.



- C_2 is often a "peaking" tune cap in the radio.

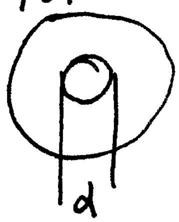
EE 533-22

Helical Antennas.

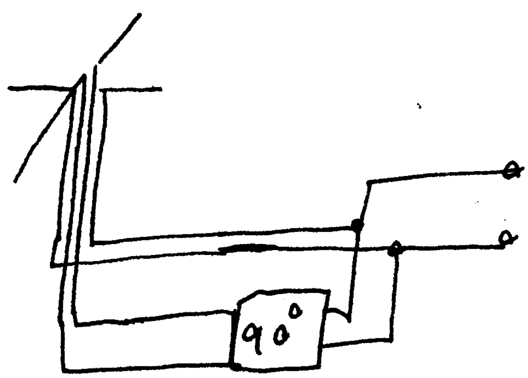


These are often used for satellite antennas as they are capable of producing C.P. radiation.

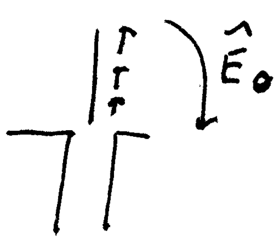
Top view.

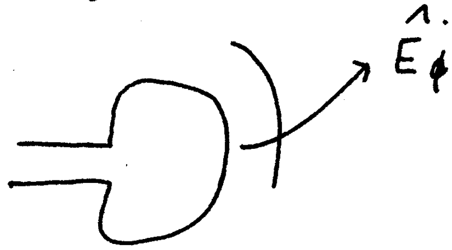


C.P. may be done with dipoles - but requires a phasing harness.



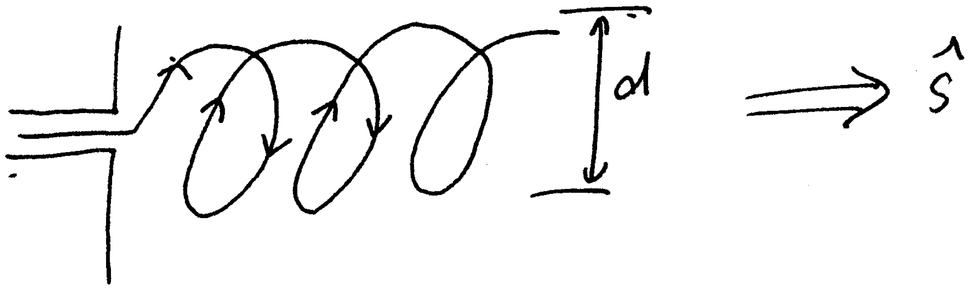
C.P. may also be generated using Loops - $\hat{E}_\phi$ vs E_θ of dipole.





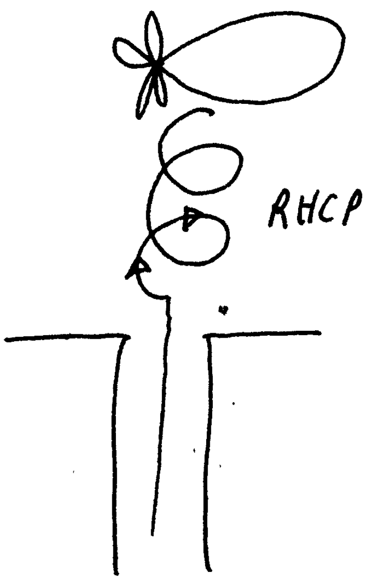
The Helical with proper pitch and d has both $\hat{E}_\theta + \hat{E}_\phi \Rightarrow$ C.P. !

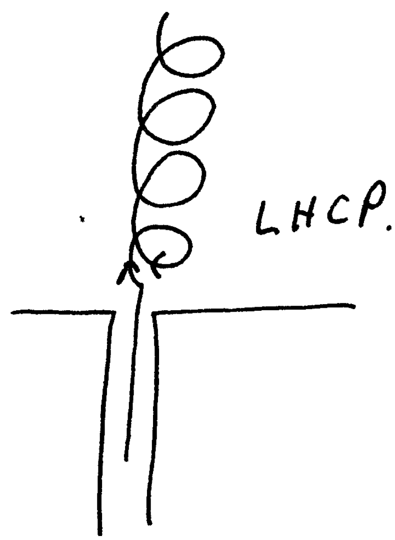
Normal mode ; $d \ll \lambda$ $\Rightarrow$ end fire helical.
or axial mode.
winding for LHCP.



Axial mode Large $\frac{3\lambda}{4} \le d \le \frac{4\lambda}{3}$

can give gain > 10dB !



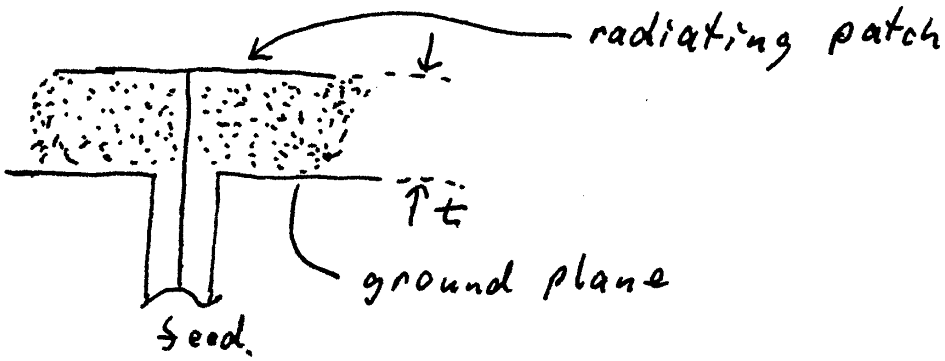


general discussion of other antenna types.

EE533-24

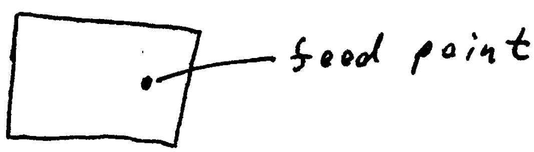
Microstrip Antennas

Advantages
- easy to fabricate
- inexpensive
- conformal
- easy to Z match.
- rugged.

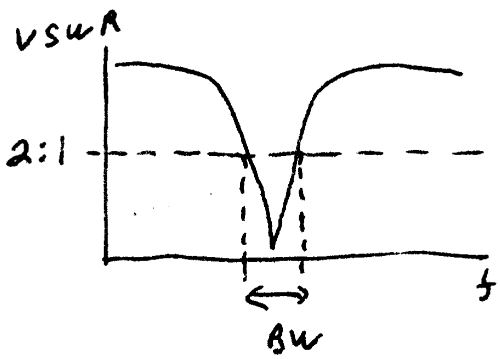


$0.3\% \lambda \leq t \leq 5\% \lambda.$

Top



Disadvantage → narrow bandwidth.



2:1 VSWR
Bandwidth < 5%.

$$BW\% = \frac{BW}{f_c} \times 100.$$

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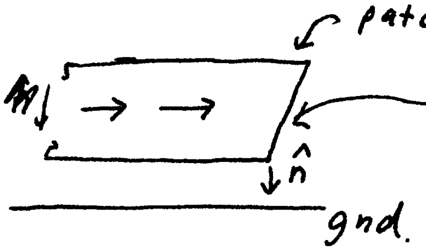


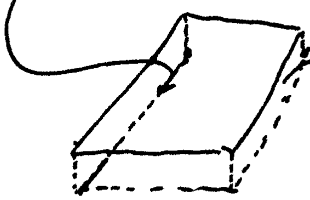
at end $\vec{J}_n = 0$

surface current

$\vec{J} = \vec{n} \times \vec{H}$

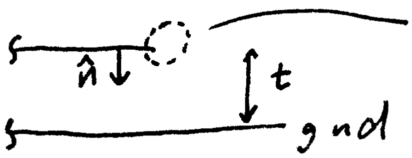
└ normal to surface.

top - perfect conductor



around edge $\vec{H}_{\parallel} = 0$

⇒ perfect magnetic conductor.

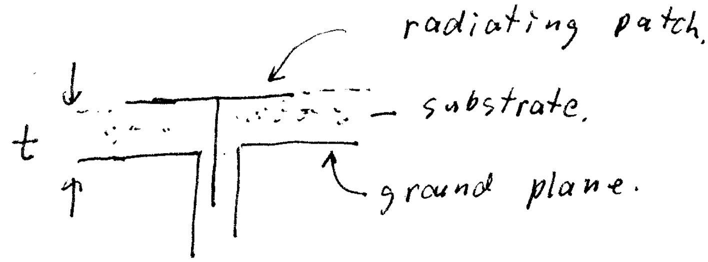


$\vec{H}_{\parallel} = 0$

on the conducting surface - $\vec{E}_{\parallel} = 0$

perfect electric conductor. ⌊ tangential component

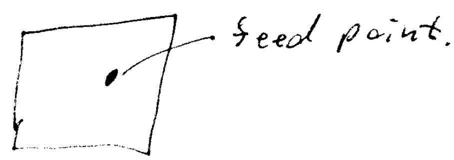
3/1/00 Microstrip Antennas



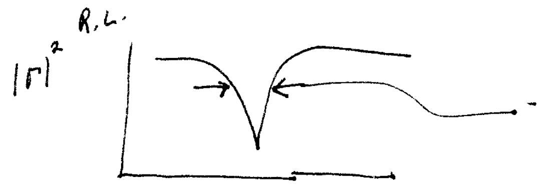
advantages:
- easy to fabricate
- inexpensive, conformal
- easy to Z match
- rugged

$0.3\% \le t \le 5\% \lambda$

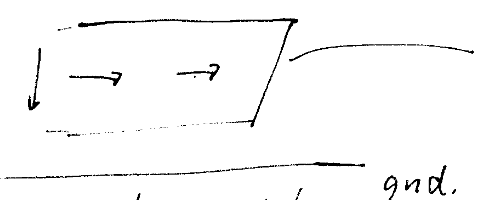
Top



Disadvantage → narrow bandwidth.



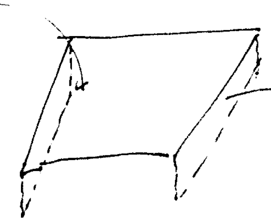
bandwidth < 5%



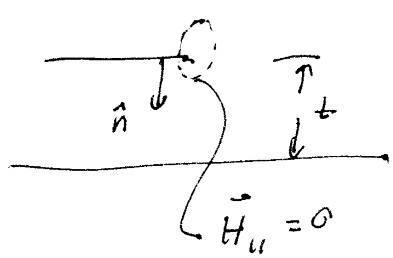
at end $\vec{J}_n = 0$

top - perfect electric conductor.

surface current
$\vec{J} = \hat{n} \times \vec{H}$
└ normal.

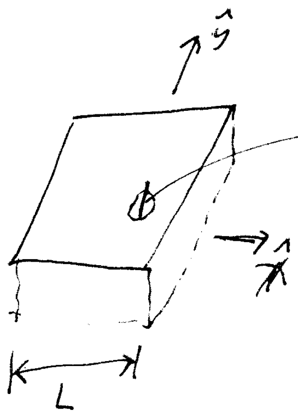


around edge.
$\vec{H}_{\parallel} = 0$
└ perfect magnetic conductor.



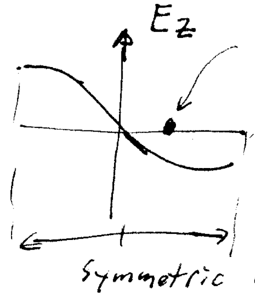
on conducting surface $\vec{E}_{\parallel} = 0 \Rightarrow$ perfect electric conductor.
└ tangental component.

45



feedpoint

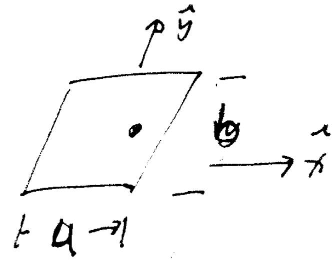
standing wave is formed.



$max = \frac{\lambda_\epsilon}{2}$ - in the medium.

$\lambda_\epsilon = \frac{\lambda_o}{\sqrt{\epsilon_r}}$

max radiation normal to surface. - Bell-shaped.



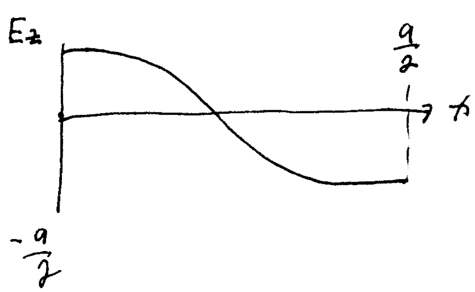
E-plane is $\hat{x}-\hat{z}$ plane.

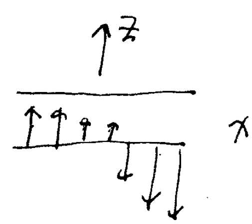
H-plane $\hat{y}-\hat{z}$ plane.

9/29/04 wed. 10/4 mod viewed

$\bar{E} = E_z \hat{z}$ $\quad E_{zmn} = \cos\left(\frac{m\pi}{a}x\right)\cos\left(\frac{n\pi y}{b}\right)$

dominant mode. $m=1,\ n=1$

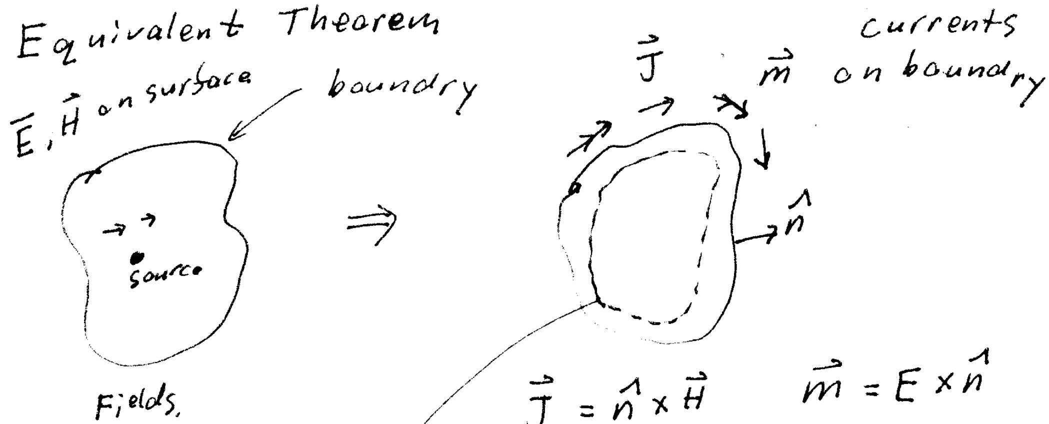




....

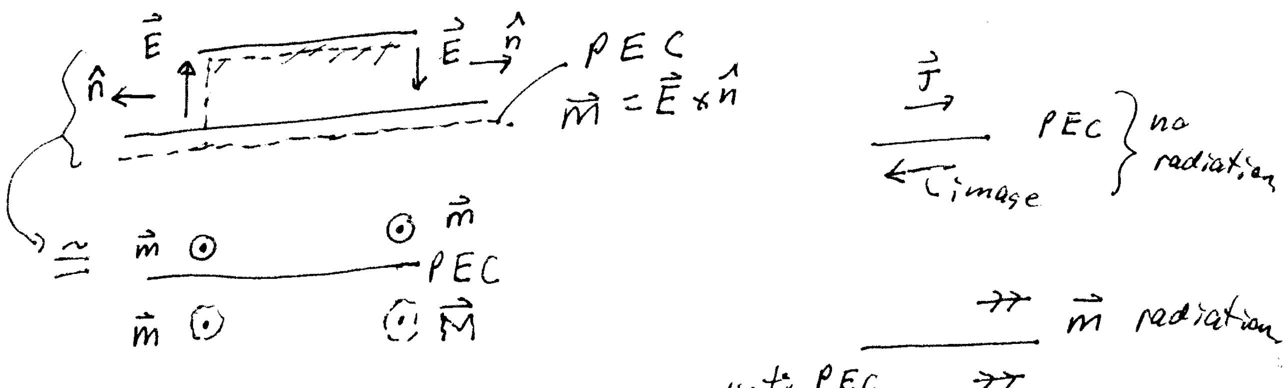
Cavity model - small amount of Energy leaks from edge

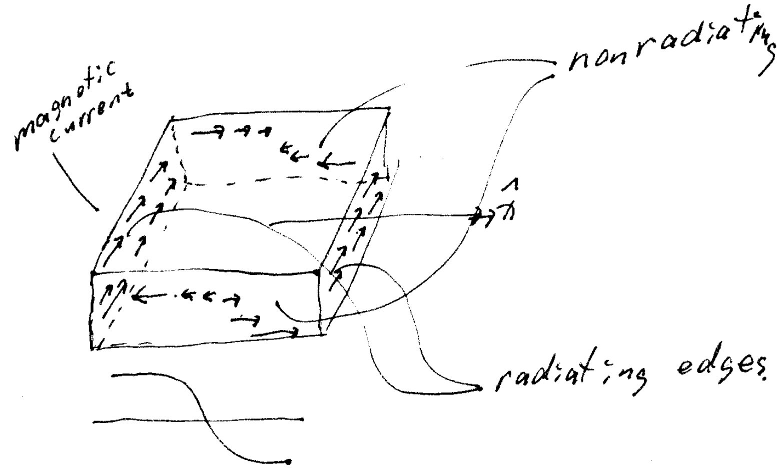
Equivalent Theorem



$\vec{J} = \hat{n} \times \vec{H}$ $\qquad \vec{m} = E \times \hat{n}$

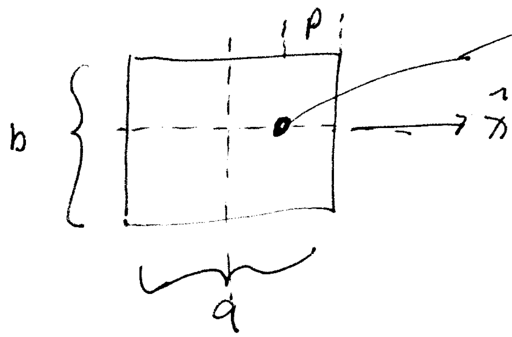
3 choices
none, perfect electric conductor.
" magnetic conductor.





23:58 Useful formulas for microstrip Antennas.

feedpoint offset in one axis.



a = length.

b = width.

due to fringing.

$2a = \lambda_\epsilon$ - in dielectric. $\lambda_\epsilon = \dfrac{\lambda_0}{\sqrt{\varepsilon_{eff}}}$ $\varepsilon_{eff} < \varepsilon_r$

$$\varepsilon_{reff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left[1 + \frac{10t}{b}\right]^{1/2}$$

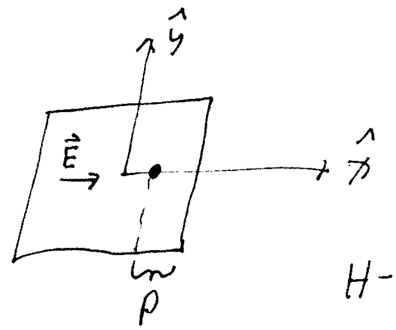
fringing

$a_{eff} > a$ $\quad a_{eff} = a + \Delta a.$

$$\frac{\Delta a}{t} = 0.412 \cdot \frac{(\varepsilon_{reff} + 0.3)\left(\frac{b}{t} + 0.264\right)}{(\varepsilon_{reff} - 0.258)\left(\frac{b}{t} + 0.813\right)}$$

$$f_r = \frac{c}{\lambda_\epsilon} = \frac{c}{\sqrt{\varepsilon_{eff}}\; 2a_{eff}} = \frac{c}{2}\,\frac{1}{\sqrt{\varepsilon_{reff}}}\,\frac{1}{a + \Delta a}$$

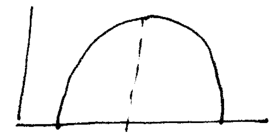
$$\boxed{f_r \approx \frac{c}{2}\,\frac{1}{\sqrt{\varepsilon_r}}\,\frac{1}{a}}$$ within a few %



$t \ll a, b.$

H-plane is $\hat{y}$-$\hat{z}$ plane.

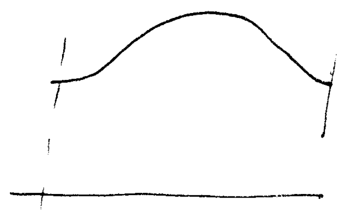
$$F(\theta) = \frac{\sin\left(\frac{\pi b}{\lambda_\varepsilon}\sin\theta\right)}{\frac{\pi b}{\lambda_\varepsilon}\sin\theta}\cos\theta$$



H-plane pattern.

E-plane ($\hat{x}$-$\hat{z}$ plane)

$$F(\theta) = \cos\left(\frac{\pi a}{\lambda_e}\sin\theta\right)$$

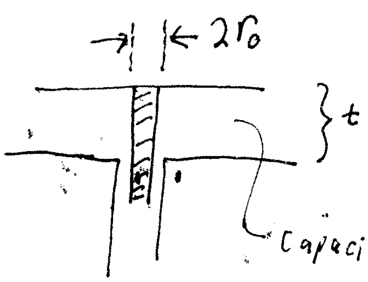


broader

E-plane pattern.

Input impedance.

$$Z_a = \underbrace{-j\omega\mu t\,\frac{\frac{2}{ab}\cos^2\left(\frac{\pi p}{a}\right)}{\varepsilon_r k^2\left[1-\frac{j}{Q}\right]-\left(\frac{\pi}{a}\right)^2}}_{\text{from patch.}} + \underbrace{j\omega L}_{\text{feed structure.}}$$



$$L = \frac{-\mu t}{2\pi}\ln(k r_0)$$

capacitance use to cancel

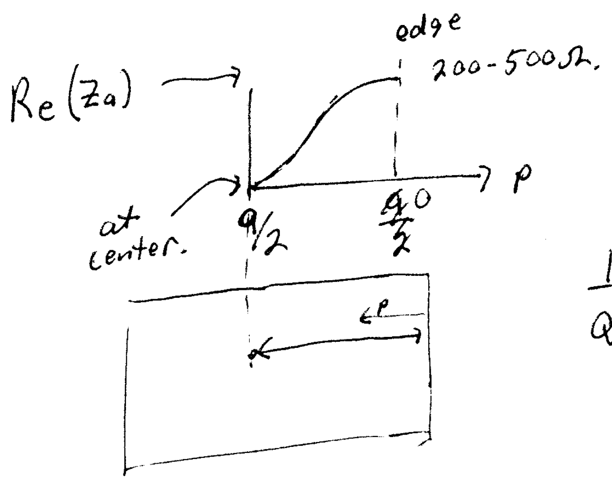
near the resonant frequency:

$$\epsilon K^2 \approx \left(\frac{\pi}{a}\right)^2 \quad \text{in denominator.}$$

then $Z_a \approx Q \dfrac{\omega \mu t}{\epsilon_r K^2} \psi^2 \qquad \psi = \sqrt{\dfrac{2}{ab}} \cos\left(\dfrac{\pi \rho}{a}\right)$

$$\eta = \sqrt{\frac{\mu}{\epsilon}} \;, \quad K = \omega\sqrt{\mu\epsilon} = \frac{2\pi}{\lambda_0}\sqrt{\epsilon_r}$$

$$Z_a = Q \frac{\eta t}{\epsilon_r K} \psi^2 \quad \propto \cos\left(\frac{\pi \rho}{a}\right)^2$$



$\rho = \dfrac{a}{2}$ at center, $z=0$

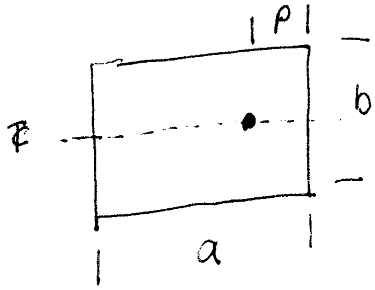
$$\frac{1}{Q} = \frac{1}{Q_r} + \frac{1}{Q_\epsilon} + \frac{1}{Q_c}$$

Q_r: radiation loss. Q_ϵ: dielectric loss. Q_c: conduction loss

$$Q_r = \frac{2\omega W_E}{P_r} \qquad 2W_E = W_T = \text{Total stored energy}$$

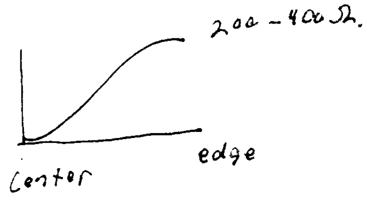
$$P_r = \frac{1}{2\eta}\iint \sin\theta \, d\theta \, d\phi \left[|E_\theta|^2 + |E_\phi|^2 \right]$$

microstrip patch.



at resonant frequency

$$Z_a = \frac{Q \eta t}{\varepsilon_r k} \psi^2 \qquad \psi = \sqrt{\frac{2}{ab}} \cos\left(\frac{\pi \rho}{a}\right)$$



Loss factors.

$$\frac{1}{Q_\varepsilon} = \tan\delta_\varepsilon$$

δ_ε loss tangent $\begin{cases} \text{FR-4} \approx 0.01 \\ \text{Rogers RO8000 series} \\ \tan\delta_\varepsilon \approx 0.0009 \\ \text{one of the best} \end{cases}$

$$\frac{1}{Q_c} = \frac{\Delta}{t}$$

Δ is skin depth $\quad \Delta = 0.029\left(\frac{\lambda_0}{\sigma}\right)^{1/2}$ (in meters)

σ — conductivity

$$Q_r = \frac{2\omega W_E}{P_r}$$

W_E — electrical Energy stored.

$$P_r = \frac{1}{2\eta} \int_{-\pi}^{\pi} \int_{0}^{2\pi} \sin\theta \, d\theta \, d\phi \, |\vec{E}|^2$$

time average.

$$W_E = \frac{1}{2}\frac{1}{2} \varepsilon_r \varepsilon_0 \int |\vec{E}|^2 dv$$

Example. $\varepsilon_r = 2.2$ @ 2 GHz.

for Antennas, want low ϵ_r + $\tan\delta_\varepsilon$

$t = 62$ mils $= 0.062'' = 0.157$ cm.

$\lambda_o = \frac{30}{2} = 15$ cm. $\quad \frac{t}{\lambda} = \frac{0.157}{15} = 1\%$ want .3% to .5%

$$a = \frac{\lambda_\epsilon}{2} = \frac{\lambda_0/\sqrt{\epsilon_r}}{2} \approx 5.06 \text{ cm.}$$

$$b = \frac{3a}{2} \approx 7.59 \text{ cm.}$$

Large b: better rad. eff. (antenna) but small enough to suppress higer-order modes.

$\theta_{BWE} = 87.5^\circ$ $\quad \theta_{BWH} = 127.9^\circ$.

Pin location:

$\leftarrow \frac{1}{3}a$ approx