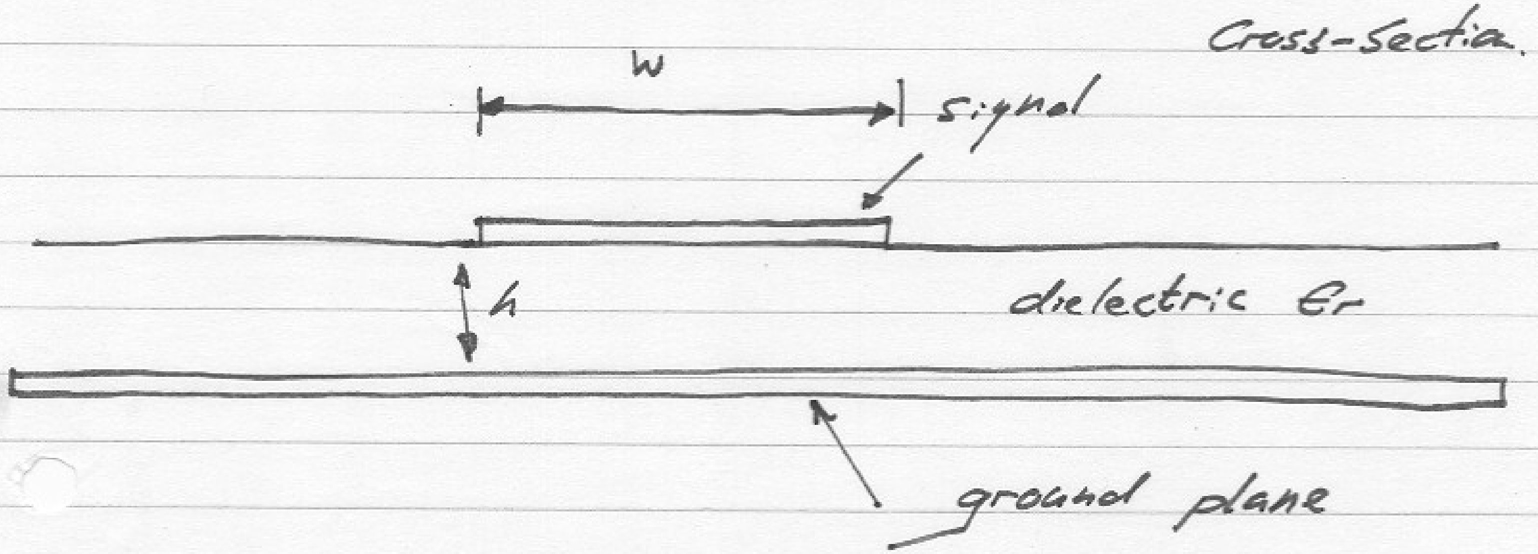
# ECE 202A Notes set 3

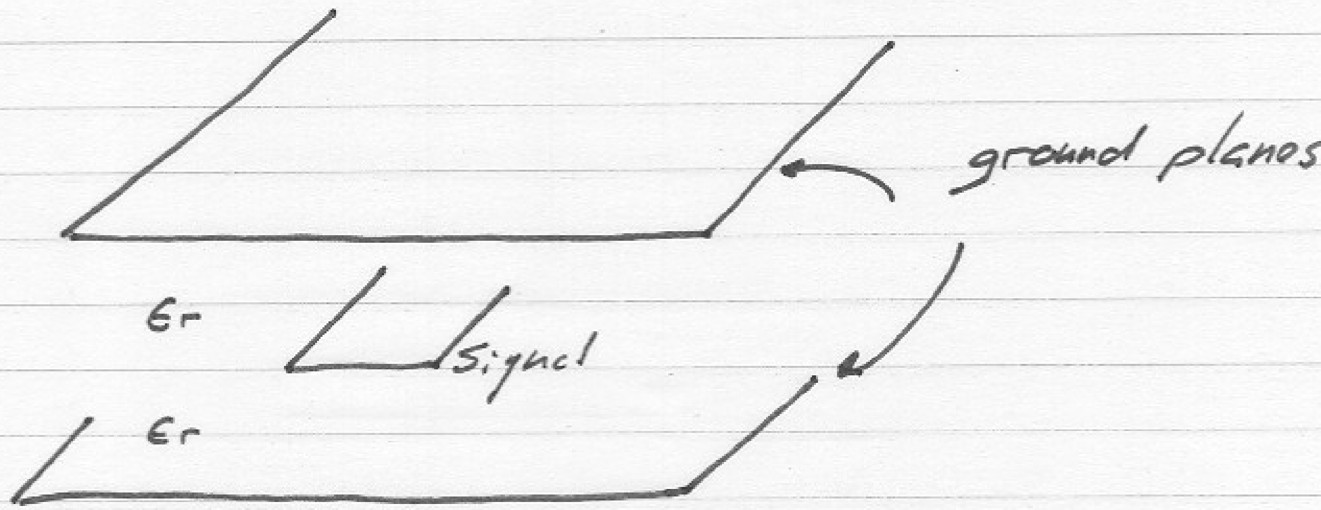
## Planar transmission Lines

Microstrip transmission line:




Some expressions for $Z_0$ & $v$ given in text. The CAD program LINECALC is more useful. We will also develop here some approximate expressions.

Compare microstrip to <u>stripline</u>

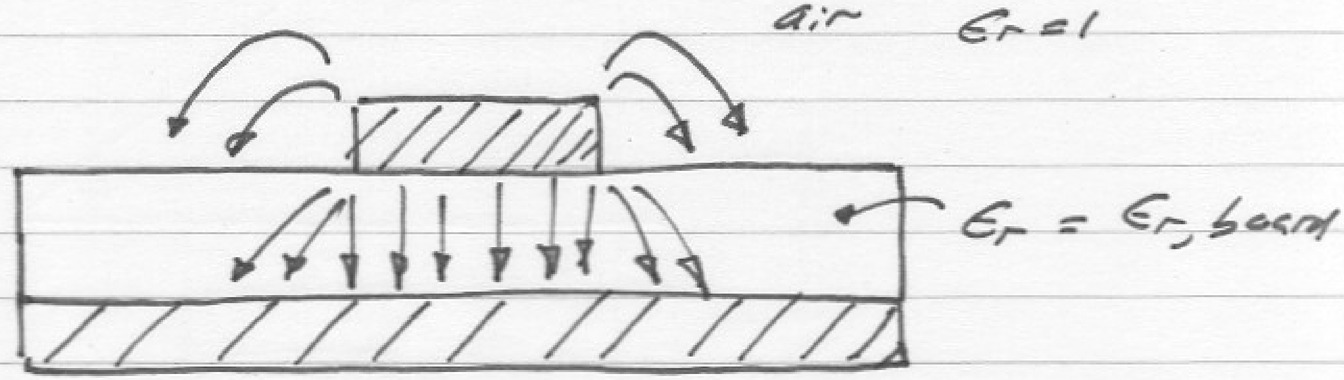



This is made by sandwiching 2 PC boards together.

All fields are within material of $\epsilon_r$.

Hence velocity $= c / \sqrt{\epsilon_r}$

In microstrip, the fields are partly in air




So for wide lines, the fields are almost all in board, while narrower lines will have proportionally more field energy in air.

Wide lines: $v \cong \dfrac{c}{\sqrt{\epsilon_r}}$

For lines of non-infinite width (w/H)

$v \neq c / \sqrt{\epsilon_r}$. This leads to the idea of

an effective dielectric constant

$$\epsilon_{r,eff} \triangleq c^2 / v^2$$

so that $v = c / \sqrt{\epsilon_{r\,eff}}$

The text gives, for example, that

$$\epsilon_{r,eff} \cong \frac{\epsilon_r + 1}{2} - \frac{\epsilon_r - 1}{2} \left(1 + 12 \frac{h}{w}\right)^{-1/2}$$

very roughly.

Advice: Ignore above formula, and use

LINECALC.

## Characteristic Impedance:

For wide lines ($w/h >> 1$), there is a uniform plane wave between the conductors and negligible fringing field.

$$\Rightarrow \quad Z_0 = \sqrt{\frac{\mu_0}{\epsilon_0}} \sqrt{\frac{1}{\epsilon_r}} \cdot \frac{h}{w} \quad \text{as } W/H \rightarrow \infty$$

this is not applicable to lines of reasonable width.

Generally $Z_0$ & $v$ are determined by electromagnetic calculations. Results are effectively tabulated in common CAD programs.

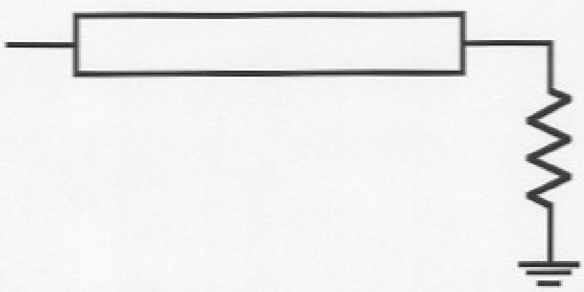
Some typical material parameters

Line parameters are for W=W

| Dielectric | $\varepsilon_r$ | $\varepsilon_{r,eff}$ | vphase | Zo |
|---|---|---|---|---|
| Duriod | 2.2 | 1.77 | 3/4 c | 85Ω |
| epoxy-glass | 4.8 | 3.28 | 0.55 c | 66Ω |
| Alumina (Al2O3) | 9.8 | 4.18 | 0.5 c | 40Ω |
| GaAs | 13 | 5.3 | 0.43 c | 35 Ω |
| Quartz (SiO2) | 3.8 | | | |

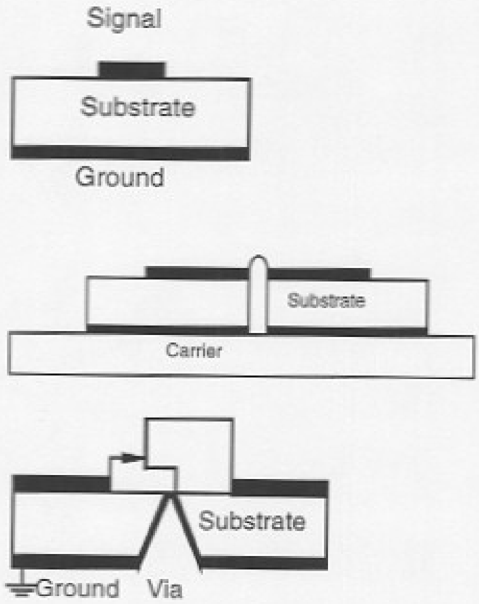
duriod is available in a variety of dielectric constants. Sapphire (crystalline Al2O3) is sometimes used, but is anisotropic. SiO2 is used both in its crystalline form (Quartz) and in amorphous form (fused silica)

**Vias in microstrip:**



We need a home in the board. This is fairly easy in 202a (drill a hole and pass a wire through it) but some wiring inductance will result. This will be a key problem in circuit construction. Vias are a very serious difficulty for microwave ICs both because of fabrication difficulties (etch a hole through the substrate, electrooplate it) and because of the ≈12 pH inductance. The higher the frequency, the more of a problem this inductance will become.

## Microstrip Line



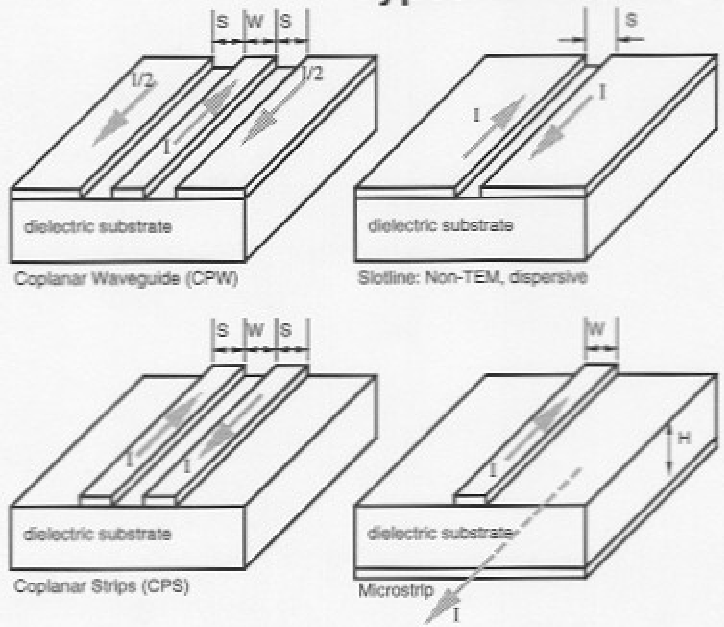

Dominant transmission medium in microwave IC's

Key advantage: IC interconnections have very low ground lead inductance- more important than signal line inductive parasitics in amplifiers

Key problem: through-wafer grounding holes (vias).
Via inductance forces progressively thinner wafers at higher frequencies. Microstrip is used with good performance in 65 GHz monolithic circuits

# Coplanar Lines, Coplanar Strip

**Types of Planar Transmission Lines**




Well designed CPW and CPS have (S+2W)<<h (better than 2:1 ratio)

For these structures, if H>>S+2W

$$\varepsilon_{r,eff} = \frac{\varepsilon_r + 1}{2}$$

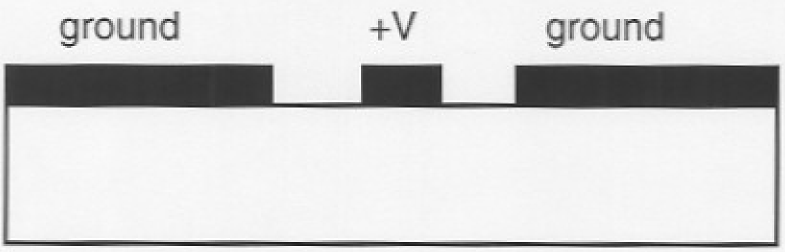
Advantages:

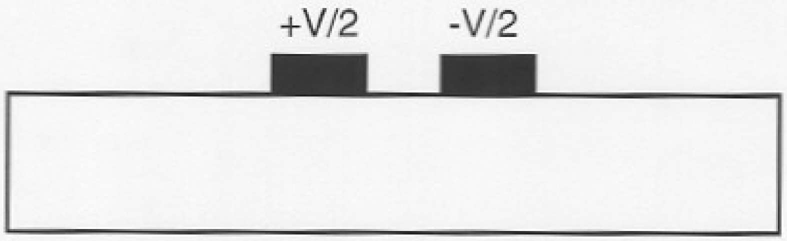
Don't need backside ground plane, don't need to etch vias, easy ground connection without inductance

disadvantages:

need to tie the two ground planes together periodically, difficulty in maintaining the ground-plane connection on one IC the same as on the IC or package to which it is connected (more on this later)

Coplanar Waveguide is a unbalanced transmission line, while coplanar strips is balanced:




<u>If</u> the substrate is thick and the metal thin $Z_0$ depends only on the ratio $s/w$

## Parasitic Modes on transmission lines:

2 Classes of modes:

Quasi-TEM: circuit-like, require 2 wires

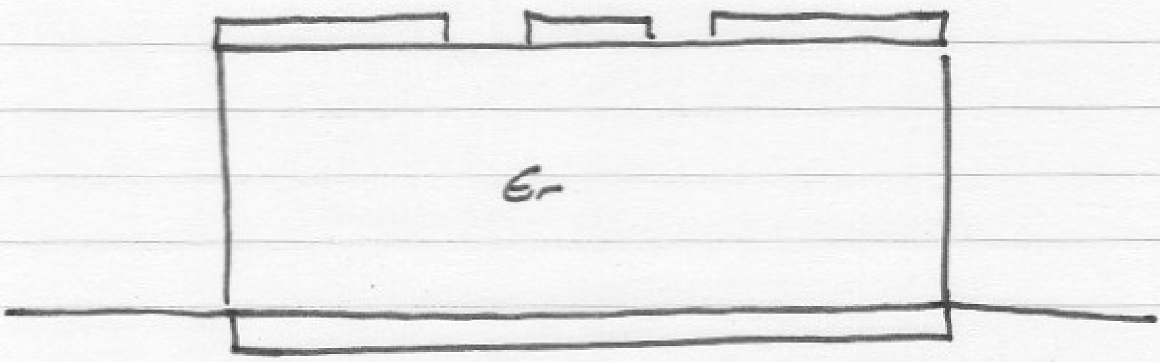
Non-TEM: modes similar to waveguide, optical fiber, etc.

**Lets consider circuit-like modes first.**

A system with N conductors will propagate (N-1) Quasi-TEM modes.

Lets consider an example:

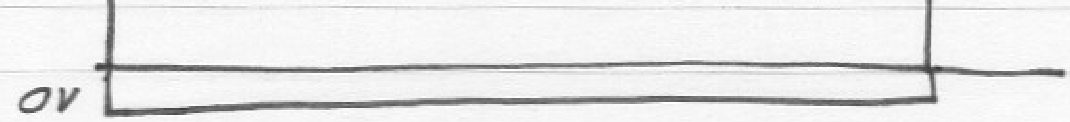
# Parasitic Modes on Coplanar Waveguide




Ground (brass carrier

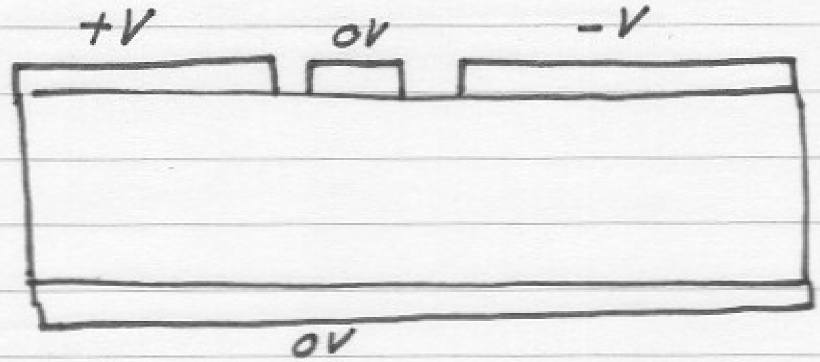
There are 4 conductors, hence 3 modes!

CPW mode (desired)



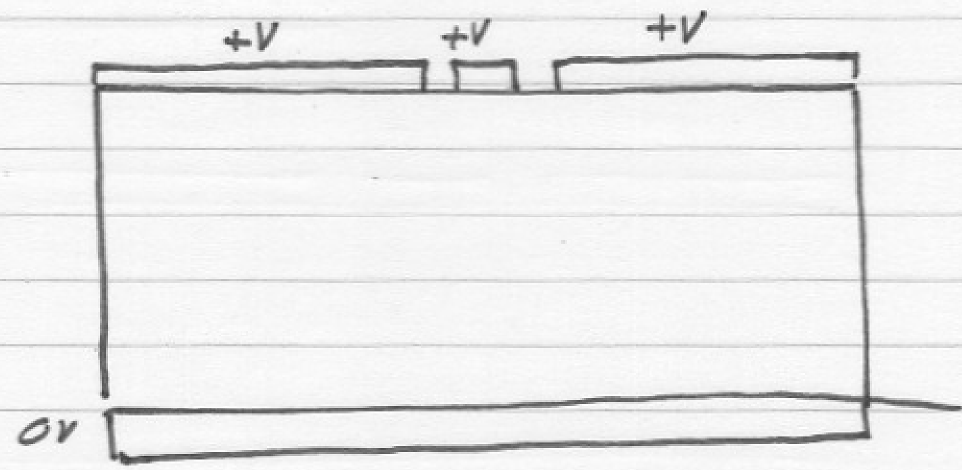

(waves propagate in both directions)

CPS mode:



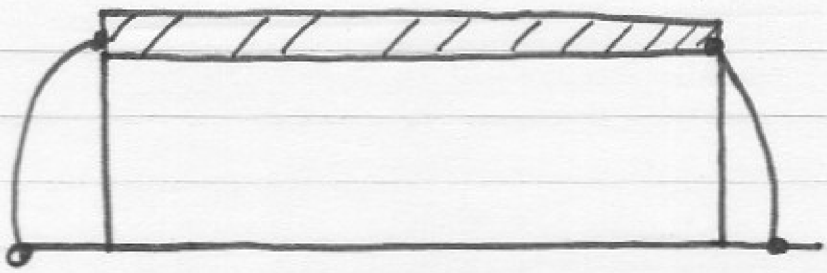

This is suppressed by connecting the 2 grounds together at distances < λ/8.

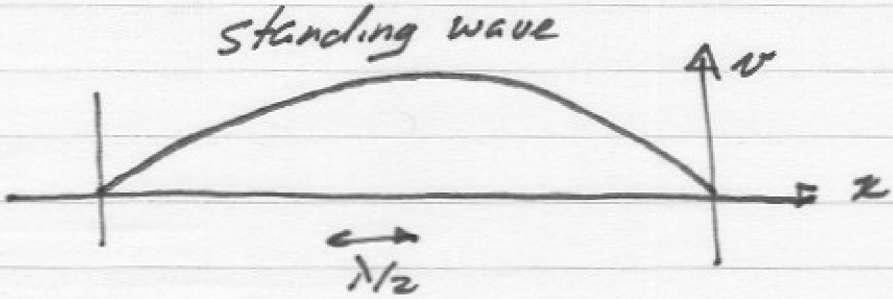
Microstrip mode




This is [possibly] suppressed by connecting the CPW ground planes ~~tk~~ to the brass carrier.

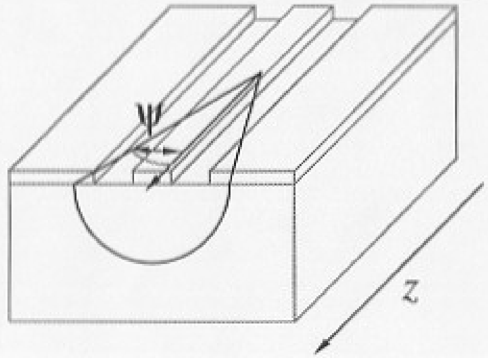
The problem is that if the chip is $> \lambda/2$ wide we will not suppress these modes by grounding at the chip boundary:

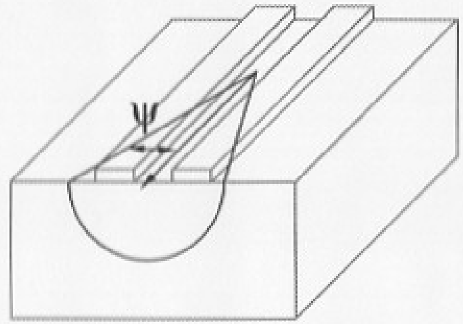





This is *the* problem with coplanar waveguide: maintaining the same ground potential on chips connecting together.

# Transmission-Line Radiation Losses



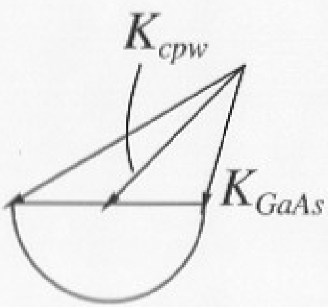

Transmission line velocity is

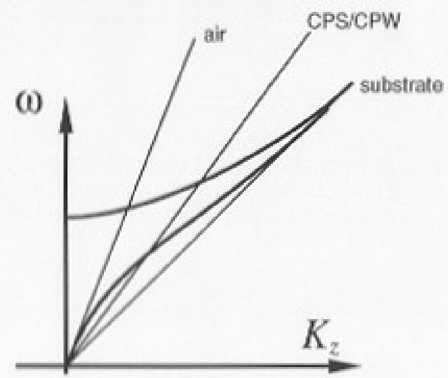
$$v = c\Big/\sqrt{(1+\varepsilon_r)/2}$$

Velocity of a plane wave in the substate is $v = c/\sqrt{\varepsilon_r}$ , *which is slower.*

Power radiates at angle $\psi$ determined by matching $K_z$.

WIth substrate of finite thickness, radiation shows frequency structure due to substrate modes



Radiation generally refers to the other class of problems: coupling energy into non-TEM modes propagating in the substrate

radiation loss is radiation into substrate modes

To radiate we must

- be able to match wavelengths in direction of propagation ($k_z$ matching)
- have significant field overlap.
- substrate modes must exist.

[ $k_z$ matching is possible because of the wave speed

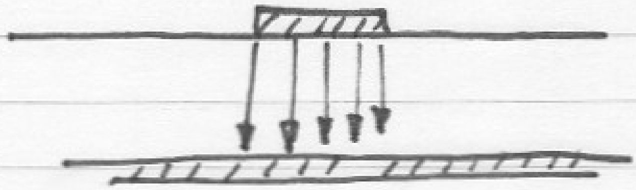
## Several Comments regarding radiation:

1) Radiation not important when substrate is thin compared to $\lambda_d$.

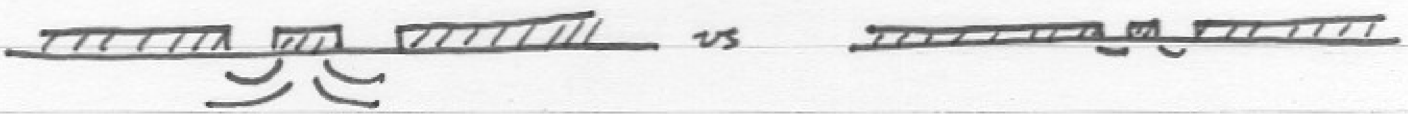
In this case the modes are either forbidden or have velocities approaching c (hence $K_z$ can't match)

2) Radiation can be eliminated by slowing. Lines can be designed with a slow wave velocity. Radiation then not possible. But slow-wave lines have other problems.

3) Coupling depends on geometry.



Microstrip has fields throughout the substrates and couples strongly with many modes.




Coplanar waveguide (and coplanar strip) can be scaled to fine dimensions which result in small field penetration into the substrate and therefore small radiation.

For CPS/CPW:

$$\alpha_{rad} \propto (s+2W)^2 \cdot f^3$$

For Microstrip lines, radiation problems will demand that the substrate become thinner as the operating frequency increases.

For Integrated circuits, this becomes a problem with breakage.

## Dispersion

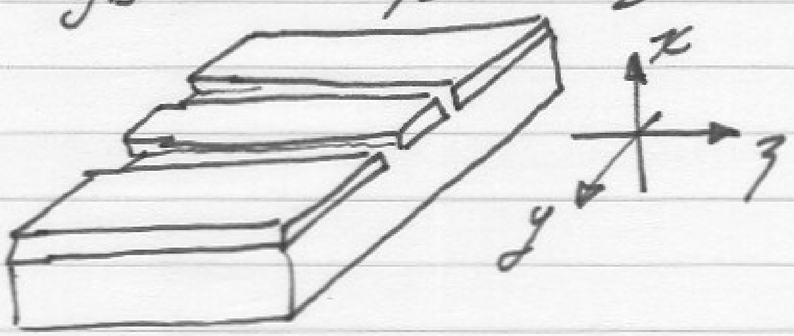
Modes are in a mixed-dielectric medium.

Quasi-TEM mode = sum of spatial harmonics

Each spatial harmonic must satisfy:

in air $\quad K_{xA}^2 + K_{yA}^2 + K_{zA}^2 = K_A^2 = \omega^2/c^2$

in board $\quad K_{xB}^2 + K_{yB}^2 + K_{zB}^2 = K_B^2 = \epsilon_r \omega^2/c^2$




we must also have $K_{zA} = K_{zB} = K_z$ and

$K_{yA} = K_{yB} = K_y$.

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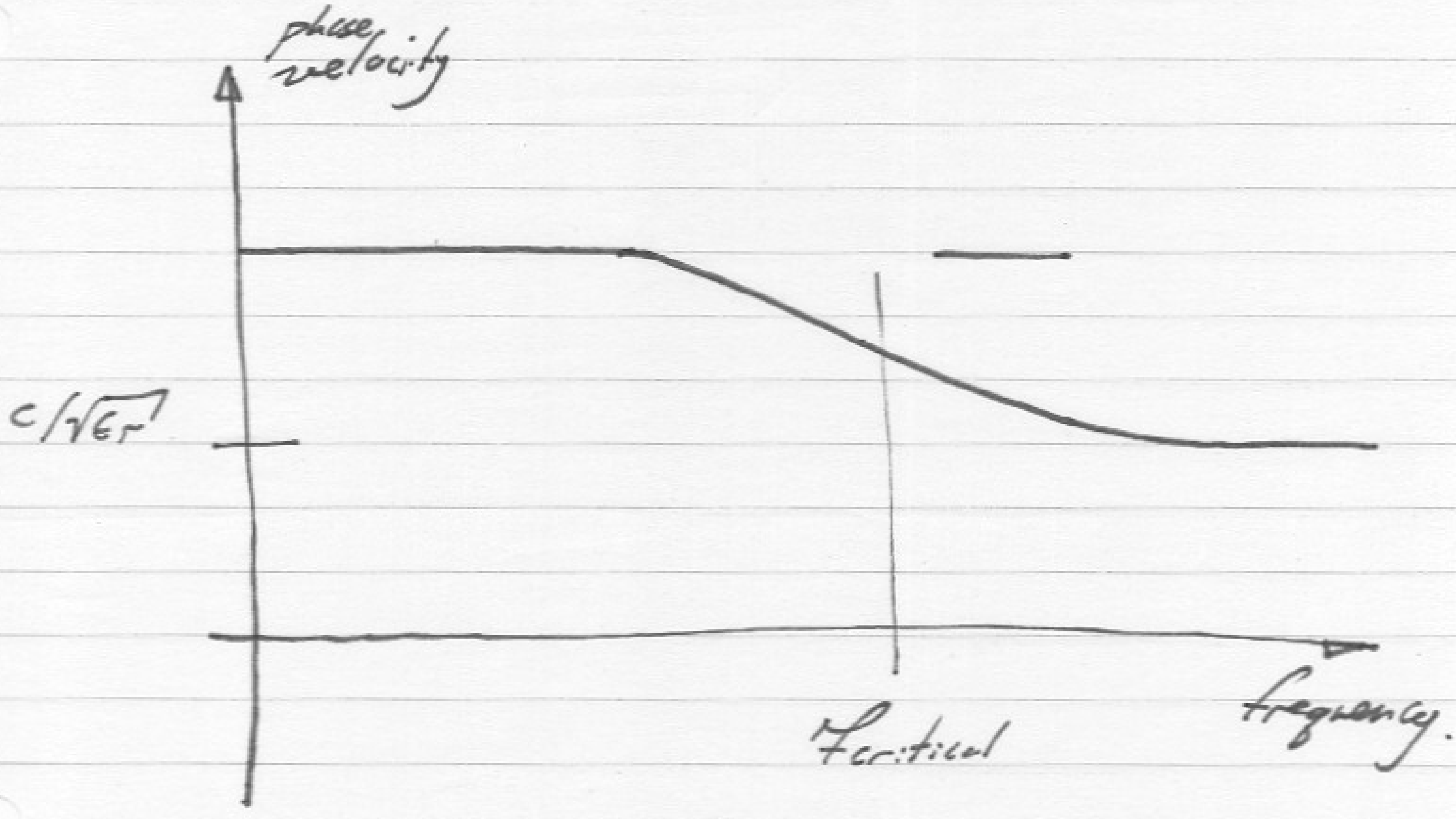
This is a boundary-value problem. Lets not bother solving its but just observe some general points:

- As frequencies increase, fields will concentrate themselves in the high-$\epsilon$ material
- velocity will decrease at higher frequencies.
- Frequency at which strong changes are observed is inversely proportional to the lines' lateral dimensions.




at $f_{critical}$, transverse dimensions $\cong \lambda_d$.

## Engineering Observation:

Near $f_{critical}$, change in velocity is large

line is also but radiating strongly

– So line is of little value anyway.

At frequencies of $f_{critical}/10$, or so,

there will be small changes in velocity.

This will be important for those systems

using transmission lines as high-Q elements.

Skin-effect losses

in metal, waves propagate as

$$E(z) = E_0 e^{-\gamma z}, \quad \gamma = \alpha + j\beta$$

where the propagation constant $\gamma$

$$\gamma = \sqrt{j\omega\mu(j\omega\epsilon + \sigma)}$$

$\sigma$ is the conductivity.

at microwave frequencies in metal, $\sigma \gg \omega\epsilon$

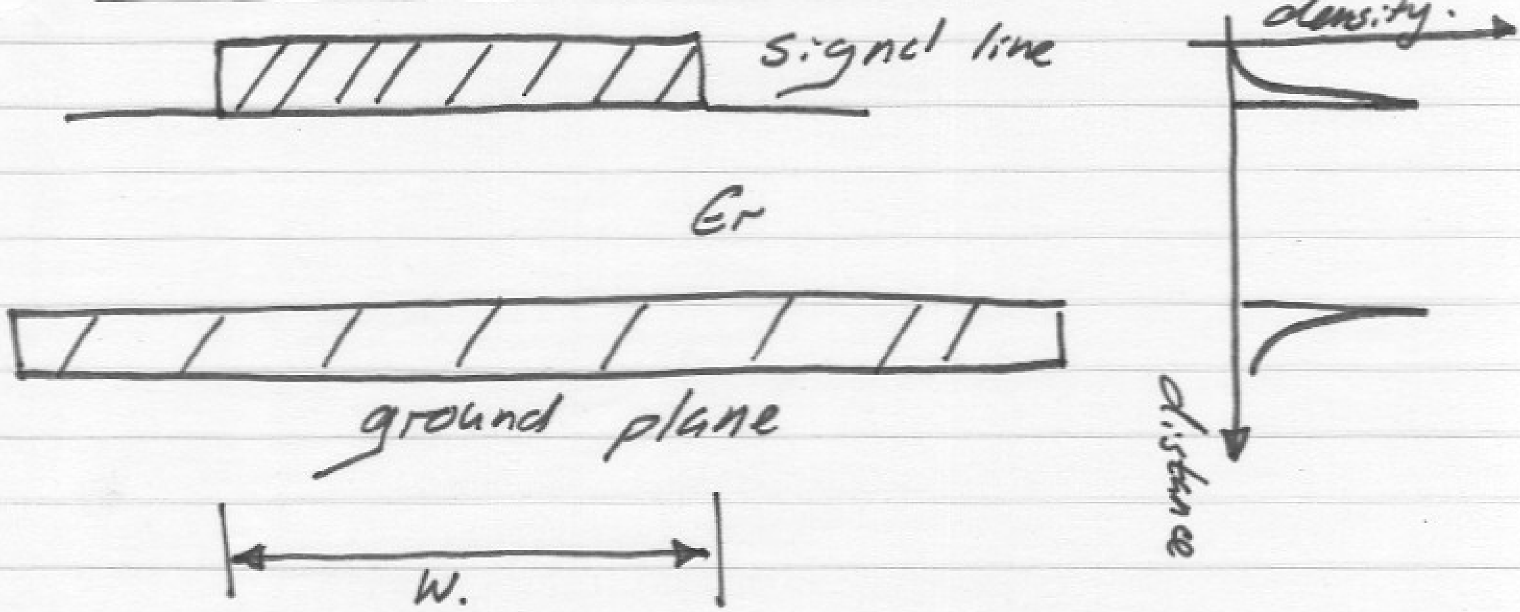
and

$$\gamma = \sqrt{j\omega\mu\sigma}$$

→ current penetrates in metal for a depth of

$$\delta = \frac{1}{\alpha_{metal}} = \sqrt{\frac{2}{\omega\mu\sigma}}$$

for a wide microstrip Line:




... the current has an exponential distribution near the 2 conductor surfaces of "thickness" (exponential decay length) $\delta$.

metal sheet resistance = $\rho_{sheet}$

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

resistance per unit length

$$R = \frac{\rho_{sheet}}{W} \cdot 2 = 2 \cdot \frac{1}{W}\sqrt{\frac{\omega\mu}{2\sigma}}$$

signal and ground

This gives rise to attenuation per unit length along the transmission line of

$$\alpha = \frac{\text{series resistance per unit length}}{2Z_0}$$

$$= \frac{1}{2Z_0} \frac{1}{W} \sqrt{\frac{\omega\mu}{2\sigma}}$$

Note that we have assumed a wide microstrip line - This gives a uniform lateral distribution of current and a $1/W$ dependence. Narrow Microstrip lines and coplanar lines have an uneven lateral current distribution and the term $1/W$ is wrong. Loss still varies as $\sqrt{\omega}$

# Skin Effect Losses, I

$$\gamma_{metal} = \sqrt{j\omega\mu(j\omega\varepsilon + \sigma)}$$

$$\cong \sqrt{j\omega\mu\sigma}$$

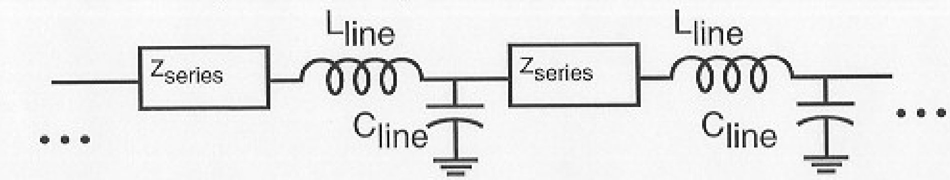
$$\alpha_{metal} + j\beta_{metal} = \sqrt{\omega\mu\sigma/2} + j\sqrt{\omega\mu\sigma/2}$$

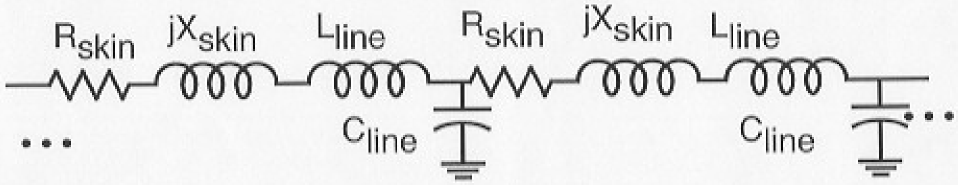
$$= (1/\delta)(1 + j)$$

$$where \quad \delta = \sqrt{2/\omega\mu\sigma}$$

$$Z_{series} = \gamma_{metal}/\sigma P$$

$$= (1/\delta\sigma P) + j(1/\delta\sigma P)$$




Surface impedance of the metal interconnections of a transmission line introduces loss proportional to the square root of frequency.

Dispersion is also introduced, as the skin impedance has equal real and complex parts

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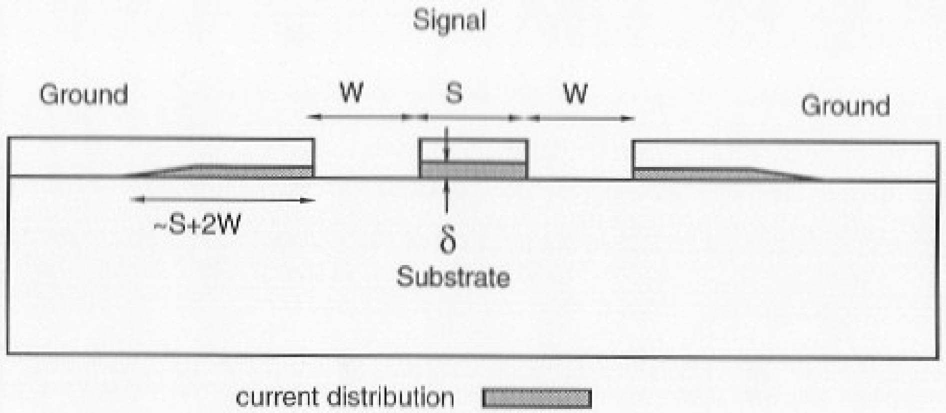
# Skin Effect Losses II




For a coplanar line the effective current carrying periphery P is approximately the width of the center conductor (IF S is relatively small compared to W, a higher-impedance line)

$$\gamma_{line} = \sqrt{j\omega C\left(j\omega L_{line} + Z_{series}\right)}$$
$$= j\omega\sqrt{LC}\sqrt{1+\left(Z_{series}/j\omega L_{line}\right)}$$
$$\cong j\omega\sqrt{LC}\left(1+\left(Z_{series}/2j\omega L_{line}\right)\right)$$
$$= j\omega\sqrt{LC} + \frac{Z_{series}}{2Z_0}$$
$$Z_0 = \sqrt{L/C}$$

Following this, the line propagation constant $\gamma$ can be found, and the transfer function for a line of length $l$ is exp($-\gamma l$)

$$Z_{series} = \left(1/\delta\sigma P\right) + j(1/\delta\sigma P)$$

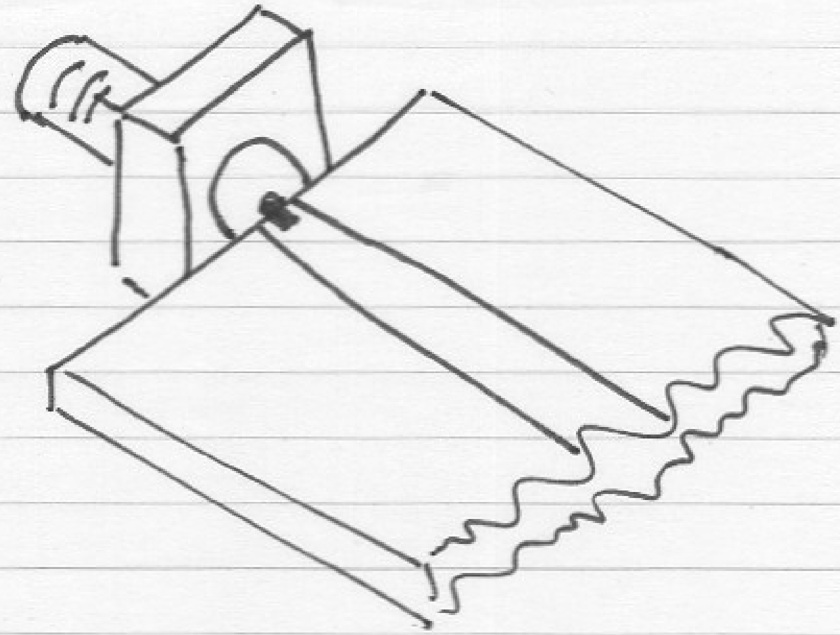


(26)

## Discontinuities:

big problem: launch:

different field configurations in coaxial cable and microstrip.

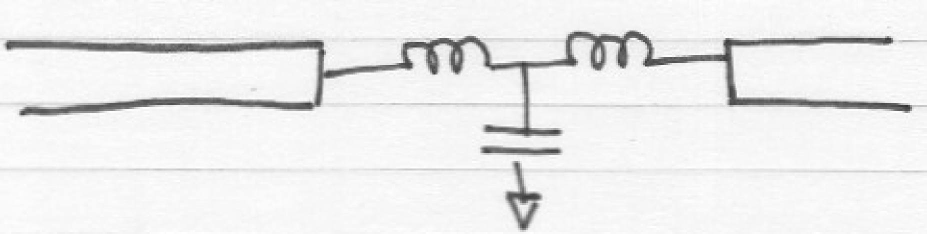


as current distributions and field distributions for the fundamental modes are different on the 2 sides of the junction, higher-order modes ~~that~~ are excited in order to satisfy the boundary conditions.

* If these higher-order modes can propagate we lose power → disasterous.

* If the higher-order modes cannot propagate, they are evanescent, not propagating but storing energy within one decay length of the junction.

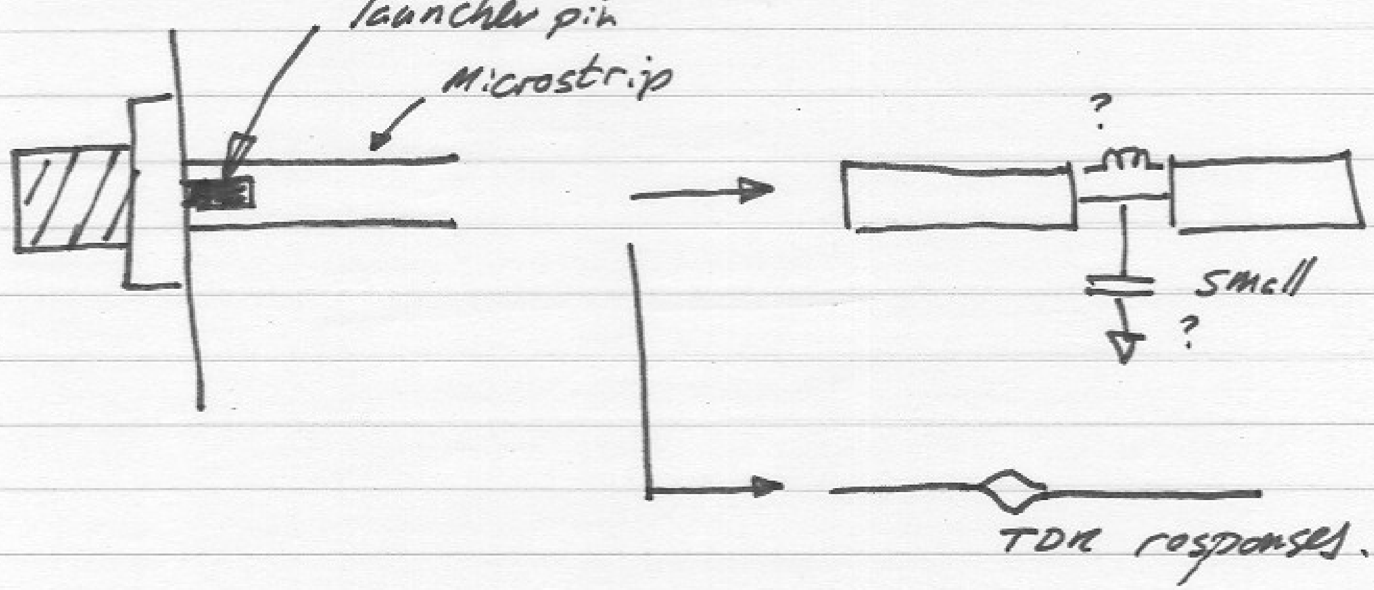
* The evanescent mode energy is modelled by reactive junction parasitics, thus:



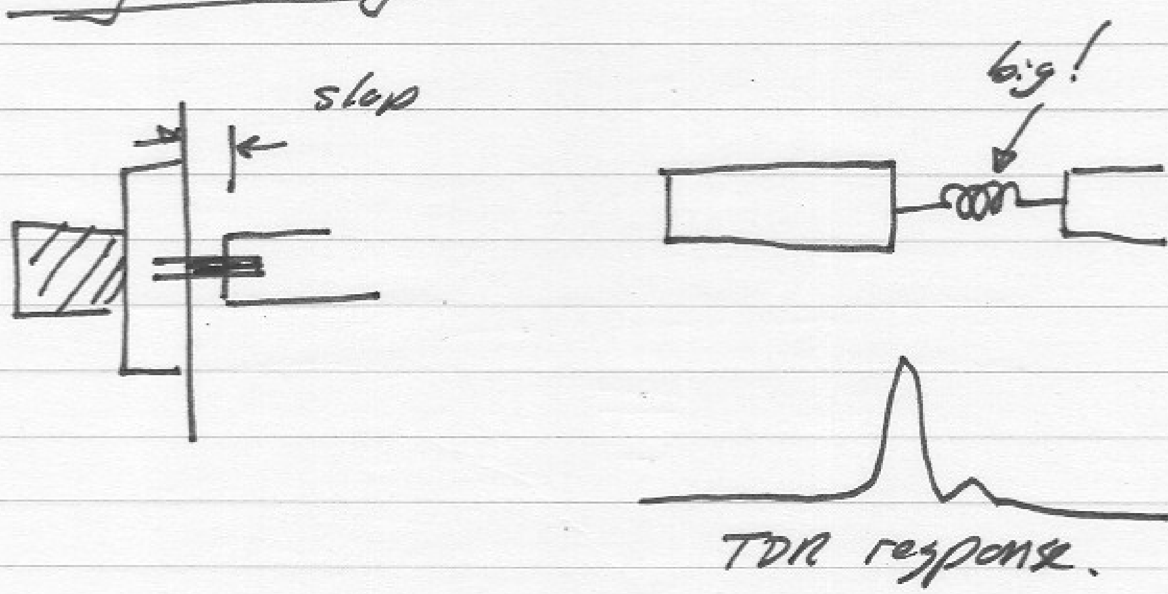
Values are determined by computer simulations; common microwave CAD programs have junction models.

Make sure your launchers are precisely assembled:

Correct:

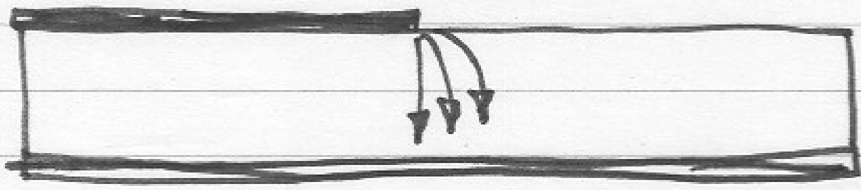



sloppy assembly:



## other discontinuities:

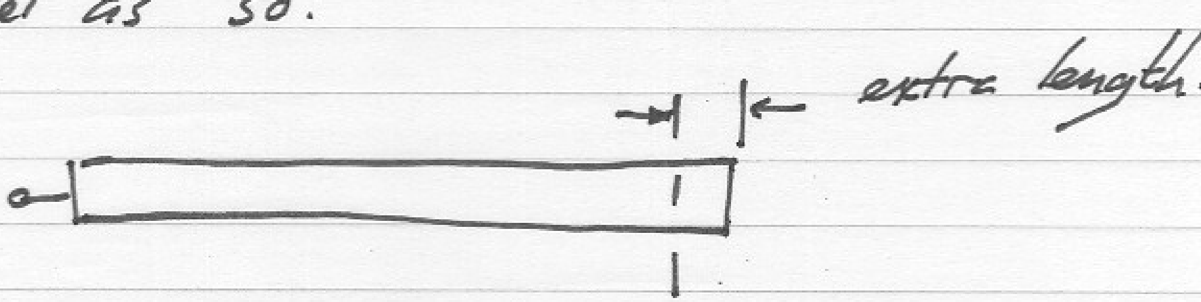
End effect



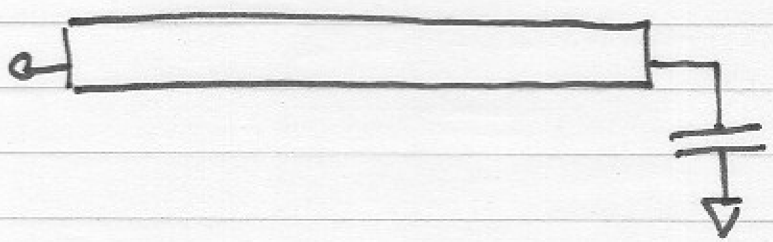
line has Fringing Fields

acts a bit longer:

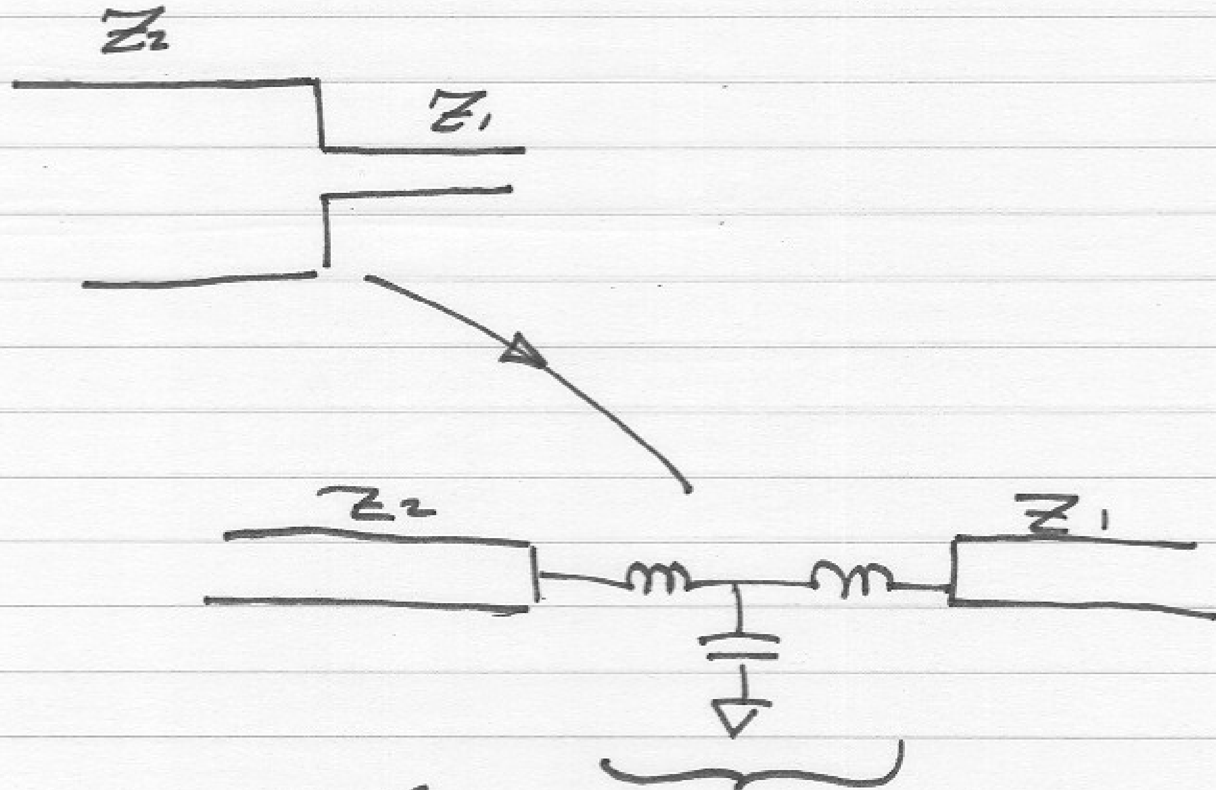
model as so:




or



## Change in Line Geometry




parasitics from junction evanescent modes.

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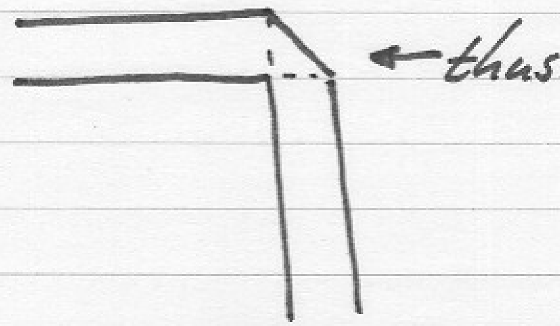
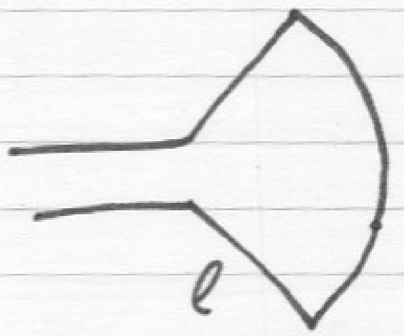
Corners have similar models

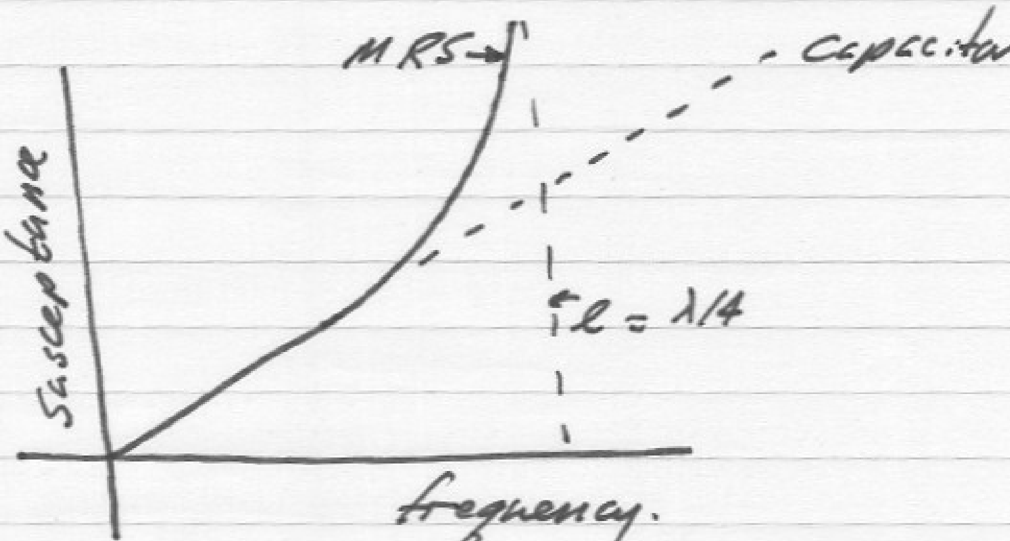
Parasitics are smallest if junctions are mitred:



## Microstrip radial stub

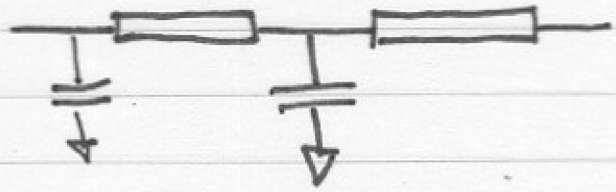



basically an open-circuited line of impedance varying with distance:

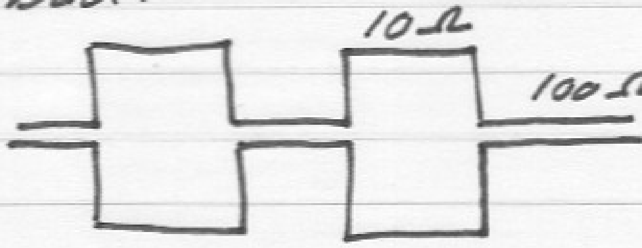


## why use it:

Matching network:

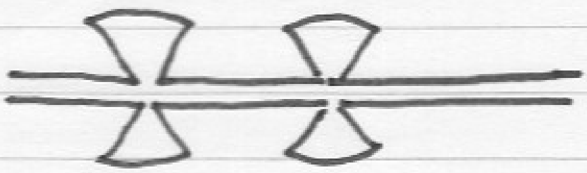


physical realization - bad:




problems with transverse modes on low-Z lines

Alternative:



Comment: no useful analytical model for radial stub. Design matching network using a capacitor, find a stub having the same susceptance over the required bandwidth, and substitute.