ECE 145A / 218 C, notes set 2: Transmission Line Parasitics

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Transmission Lines

Approximate properties of microstrip line.

Skin Effect Losses

substrate modes and loss by coupling into these.

Lateral modes on lines

Excitation of unwanted circuit-like modes, ground continuity

Packaging and power supply resonances

Skin Loss

Skin effect losses I

Wave equation inside metal (ignore x variation)

$k_y^2 + k_z^2 = k^2$ where $k^2 = (j\omega\mu)(j\omega\varepsilon + \sigma)$

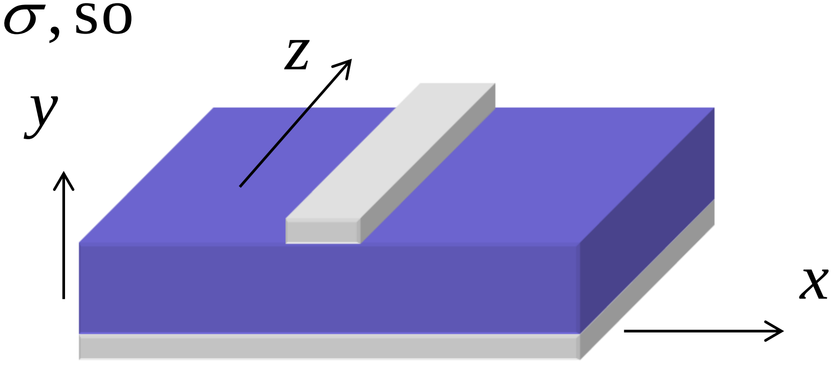
The wavelength along the transmission-line is long,

and the penetration distance of current into the metal is small,

so $k_y^2 >> k_z^2 \rightarrow k_y^2 = k^2 = (j\omega\mu)(j\omega\varepsilon + \sigma)$

In a metal, at low frequencies $j\omega\varepsilon << \sigma$, so

$$\boxed{k_y^2 = j\omega\mu\sigma}$$



Skin effect losses II

In a metal, at low frequencies $j\omega\varepsilon << \sigma$, so

$$k_y = \pm\sqrt{j\omega\mu\sigma} = \pm(1+j)\left(\frac{\omega\mu\sigma}{2}\right)$$

Defining the *skin depth* as $\delta = \sqrt{2/\omega\mu\sigma}$:

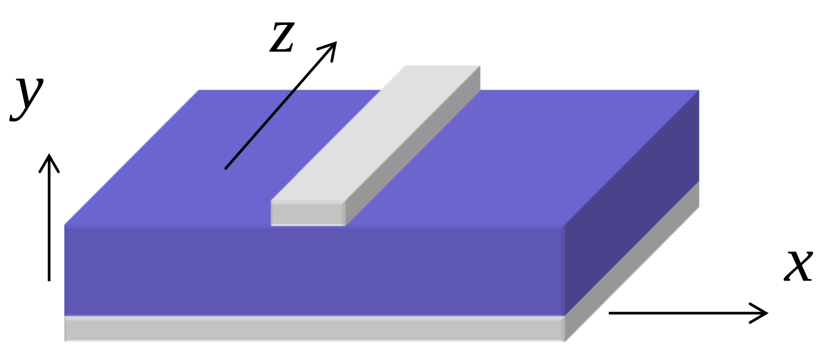
$$E(z) = E_o e^{-y/\delta} e^{-jy/\delta}$$

The field dies down exponentially with distance into the metal.

The $(1/e)$ penetration depth is the skin depth δ

δ varies as $\omega^{-1/2}$

At 100GHz in Gold, $\delta \cong 200$ nm



Skin effect losses III

Let us treat this approximately :

The conductor only carries current in a layer of thickness δ.

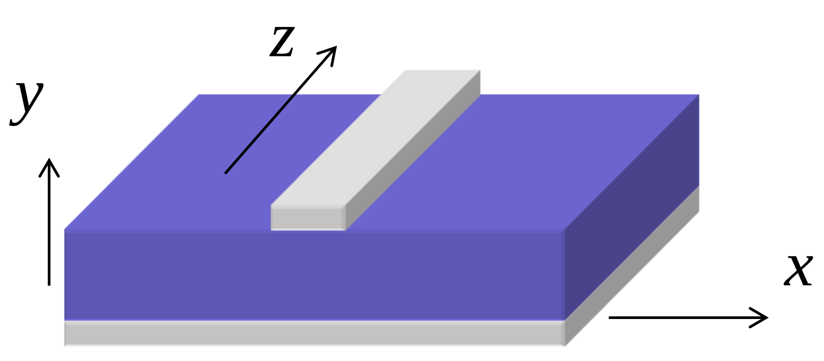
With conductivity σ and width W,

the conductor has resistance per unit length

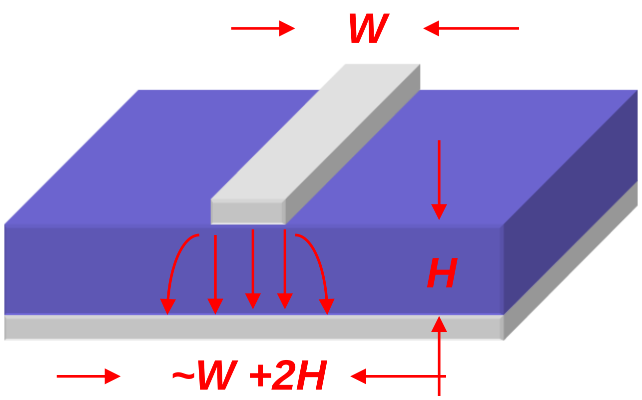
$$R_{series} / L = 1/\sigma\delta W = (1/\sigma W)\cdot\sqrt{\omega\mu\sigma/2} = (1/W)\cdot\sqrt{\omega\mu/2\sigma}$$

A more careful treatment develops the concept of surface impedance.

See the appendix.



Skin effect losses IV



Transmission-lines have skin effect in both the signal and ground conductors. For microstrip:

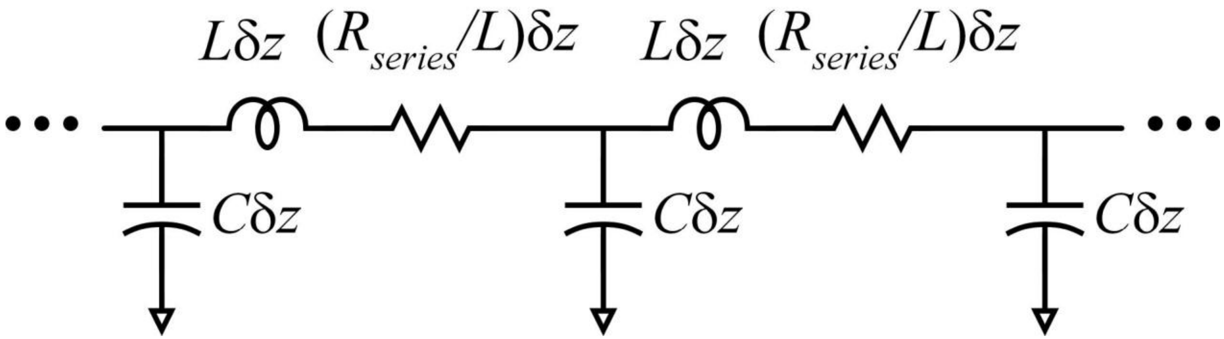
$$R_{series}/L \approx \frac{1}{\delta\sigma}\frac{1}{W} + \frac{1}{\delta\sigma}\frac{1}{W+2H} = \left(\frac{1}{W} + \frac{1}{W+2H}\right)\frac{1}{\delta\sigma} = \frac{1}{P\delta\sigma}$$

In general, we can write this as

$$R_{series}/L = 1/\delta\sigma P = (1/P)\cdot\sqrt{\omega\mu/2\sigma}$$

where P is the effective current-carrying periphery.

Skin effect losses IV



$$R_{series} / L = (1/P) \cdot \sqrt{\omega\mu / 2\sigma}$$

From our earlier transmission - line analysis,
this introduces attenuation per unit distance

$$\alpha \cong \frac{R_{series} / L}{2Z_0} \qquad \alpha \propto \sqrt{\omega}$$

This is called skin loss

Loss Tangent

Loss Tangent

Common dielectrics also introduce high-frequency attenuation.

This effect is quantified by a *loss tangent*

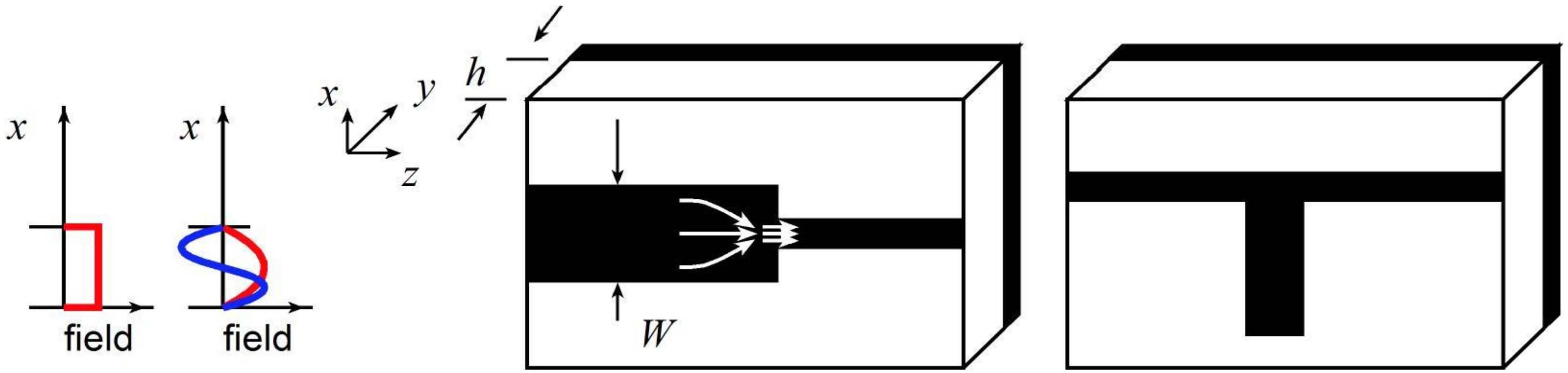
$$\varepsilon_r = \varepsilon_{r,\text{real}} + j\varepsilon_{r,\text{imaginary}} = \varepsilon_{r,\text{real}}(1 + j\tan(\delta))$$

We should be aware of dielectric losses,

but we will not discuss these further in this class.

transverse transmission-line modes

Lateral Modes (1)



In dielectric : waves of form $\vec{E}_0 e^{j\omega t} e^{\pm jk_x x} e^{\pm jk_y y} e^{\pm jk_z z}$

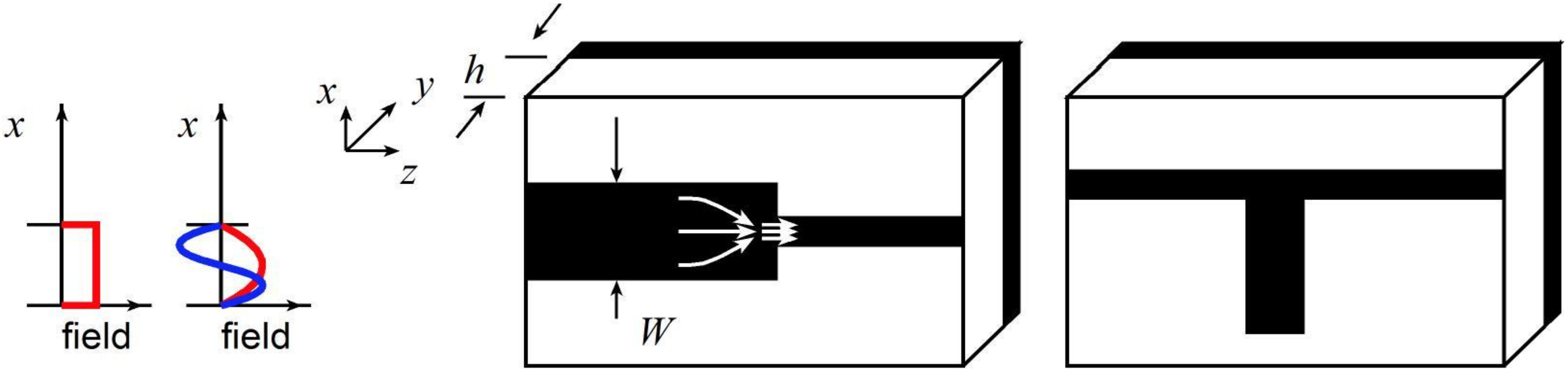
$$k_x^2 + k_y^2 + k_z^2 = k^2 = \varepsilon_r \omega^2 / c^2 = (2\pi/\lambda_d)^2$$

Waves can propagate *laterally* on transmission-line :

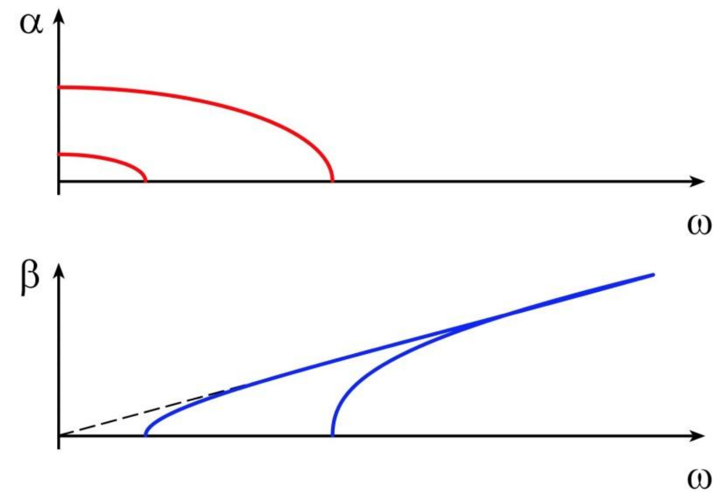
$$k_y = 0 \text{ and } k_x = n\pi / W \quad \text{for } n = 0, 1, 2, \ldots$$

$$\rightarrow k_z^2 = \varepsilon_r \omega^2 / c^2 - (n\pi / W)^2$$

Lateral Modes (2)



$$k_z^2 = \varepsilon_r \omega^2 / c^2 - (n\pi / W)^2$$



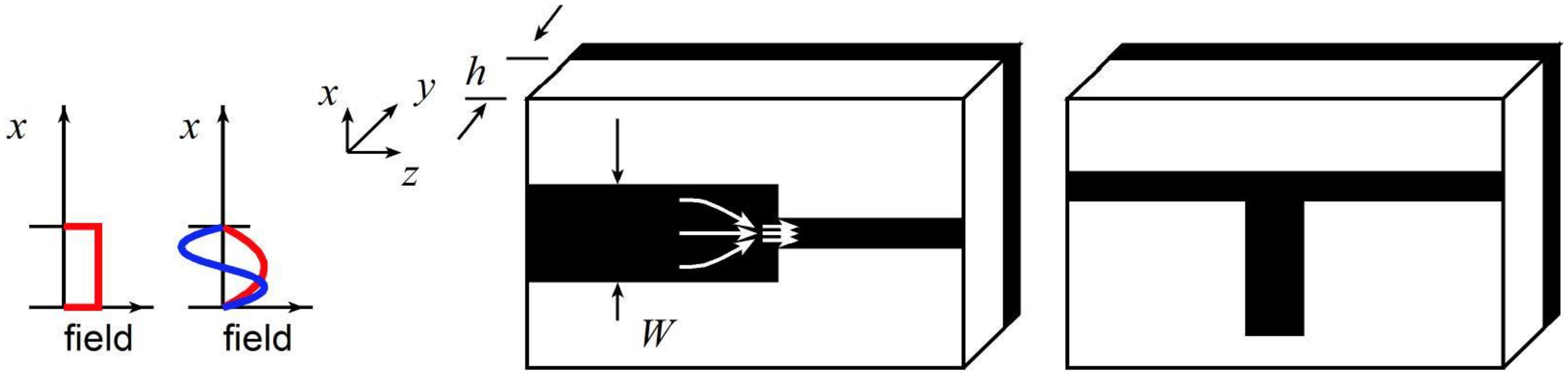
1) Multi-mode propagation if $W > \lambda_d/2$.

$$\beta_z = \sqrt{\varepsilon_r \omega^2 / c^2 - (n\pi / W)^2}$$

2) Evanescent propagation $e^{-\alpha_z z}$ if $W < \lambda_d/2$:

$$\alpha_z = \sqrt{(n\pi / W)^2 - \varepsilon_r \omega^2 / c^2}$$

Lateral Modes---and Junction Parasitics (3)



Evanescent propagation $e^{-\alpha_z z}$ if $W \cong \lambda_d/2$:

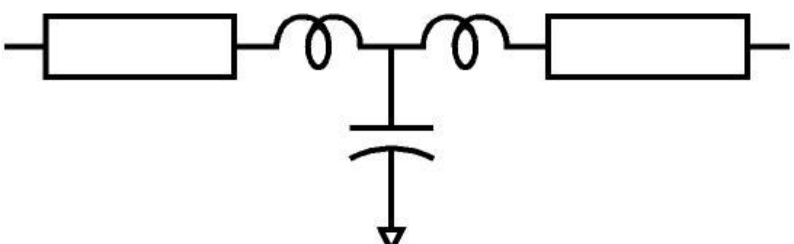
Reactive power in evanescent modes $\rightarrow$ junction parasitics

ADS library junction models, or electromagnetic simulation.

Lessons:

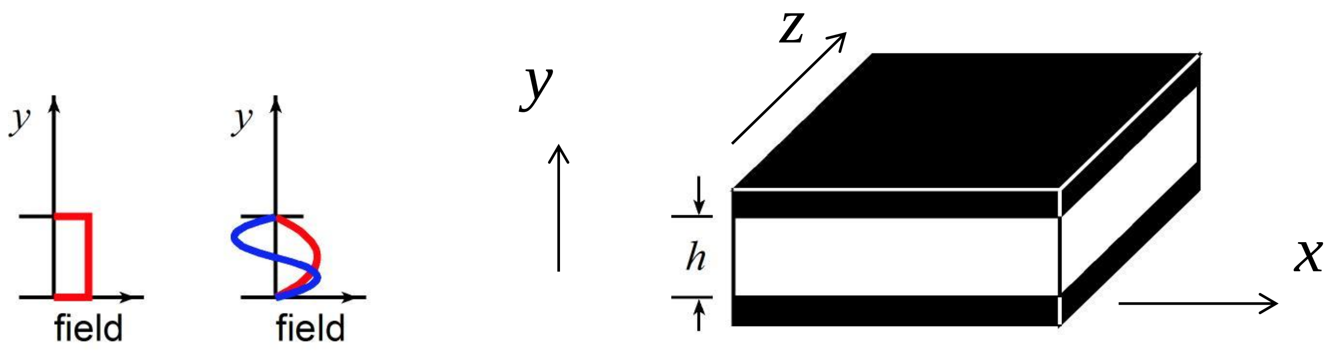
lines must be much narrower than a half - wavelength.

must model junction parasitics



Substrate Modes and Radiation Loss

Substrate Modes



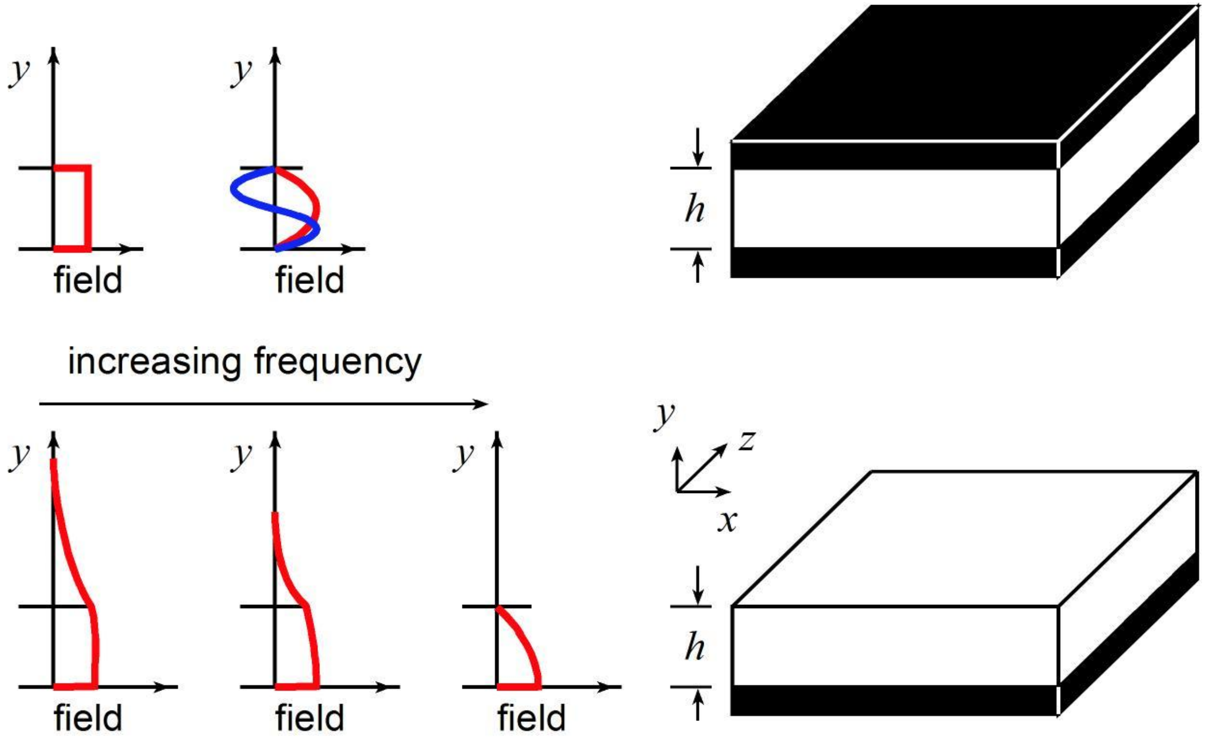
In dielectric : waves of form $E_0 e^{j\omega t} e^{\pm jk_x x} e^{\pm jk_y y} e^{\pm jk_z z}$

$$k_x^2 + k_y^2 + k_z^2 = k^2 = \varepsilon_r \omega^2 / c^2 = (2\pi / \lambda_d)^2$$

We can have standing waves across the substrate thickness :

$$k_y = n\pi / h \text{ for } n = 0, 1, 2, \ldots$$

Substrate Modes



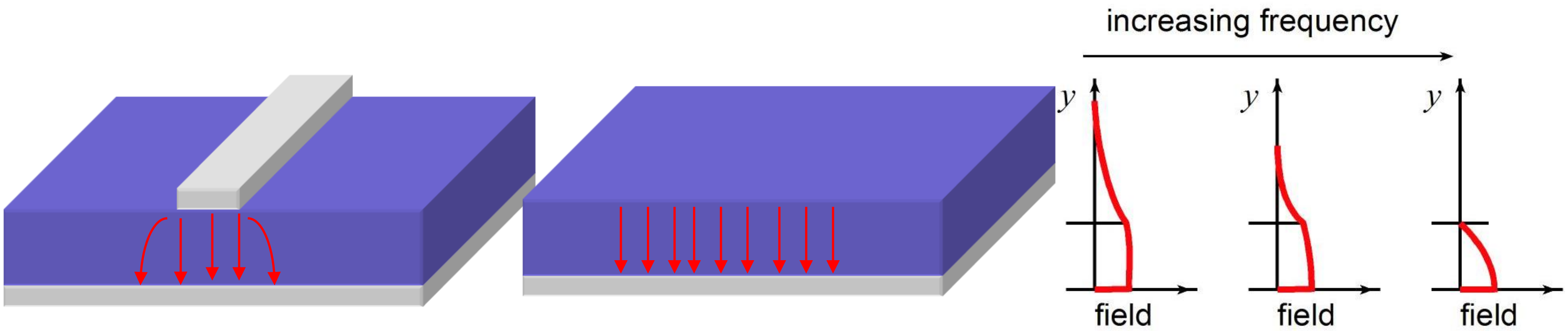
Substrate with top, bottom metal surfaces

$\rightarrow$ modes with $h = \lambda_d/2,\ \lambda_d,\ 3\lambda_d/2 \ldots$

Substrate with no top metal $\rightarrow$ tranverse E - mode;

strongly confined as $\lambda_d/4 \rightarrow T$; weakly confined at low frequencie s.

Substrate Mode Coupling: Microstrip

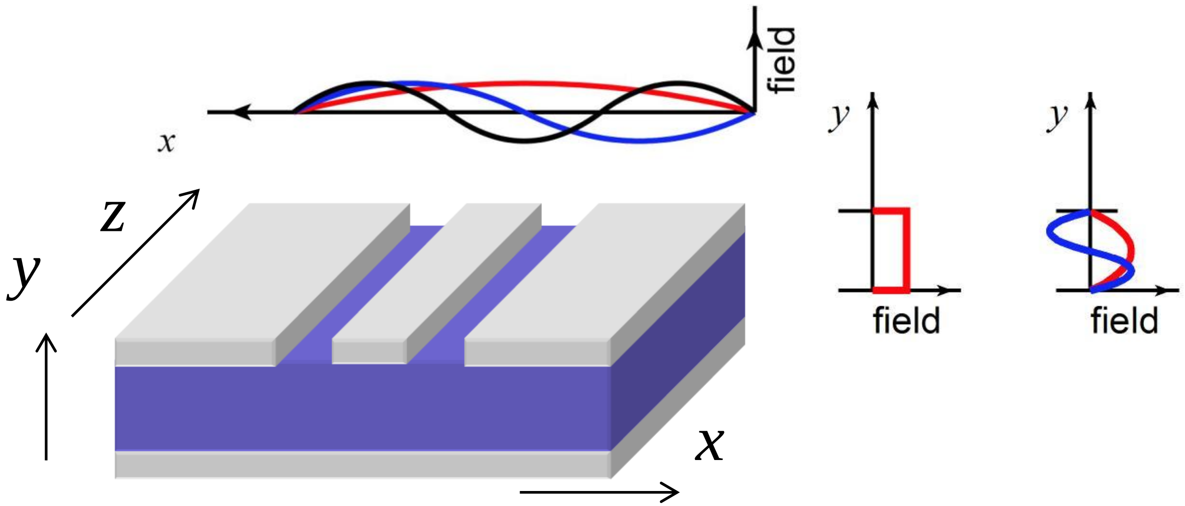


These dielectric slab modes can propage in x and in z.

Nonzero mode coupling ("radiation") loss at all frequencies.

Very strong mode coupling when $h \geq \lambda_d / 4$

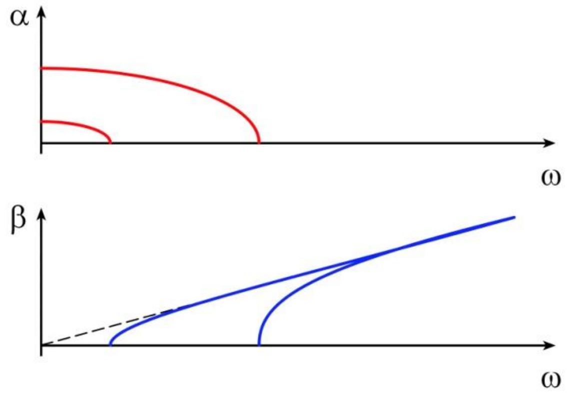
Substrate Mode Coupling: CPW



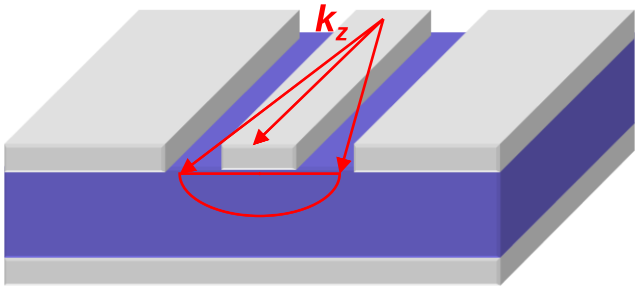
Substrate modes are allowed when $\lambda_{\mathrm{d}} \le 2 \cdot (\text{substrate thickness})$

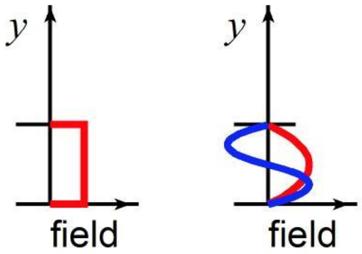
Waves are of form $\vec{E}_0 e^{j\omega t} e^{\pm jk_x x} e^{\pm jk_y y} e^{\pm jk_z z}$

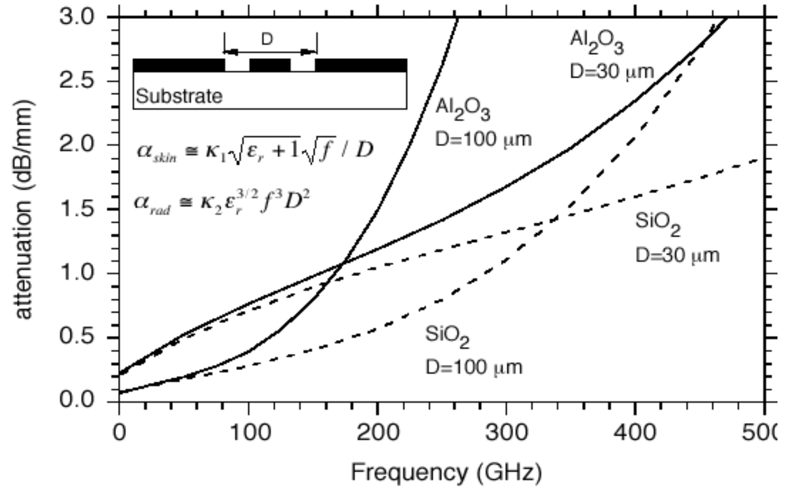
$\rightarrow k_x^2 + k_y^2 + k_z^2 = k^2$ where $k^2 = c^2/\varepsilon_r \omega^2 = (2\pi/\lambda_d)^2$ and $k_y = (n\pi/h)$



Substrate Mode Coupling: CPW







Modes couple strongly when $k_{y,\text{CPW}} = k_{y,\text{substrate mode}}$

Given thick substrate, $H >> \lambda_d$:

mode coupling loss, dB/mm $\propto (\text{line transverse dimensions})^2 \cdot \text{frequency}^2$

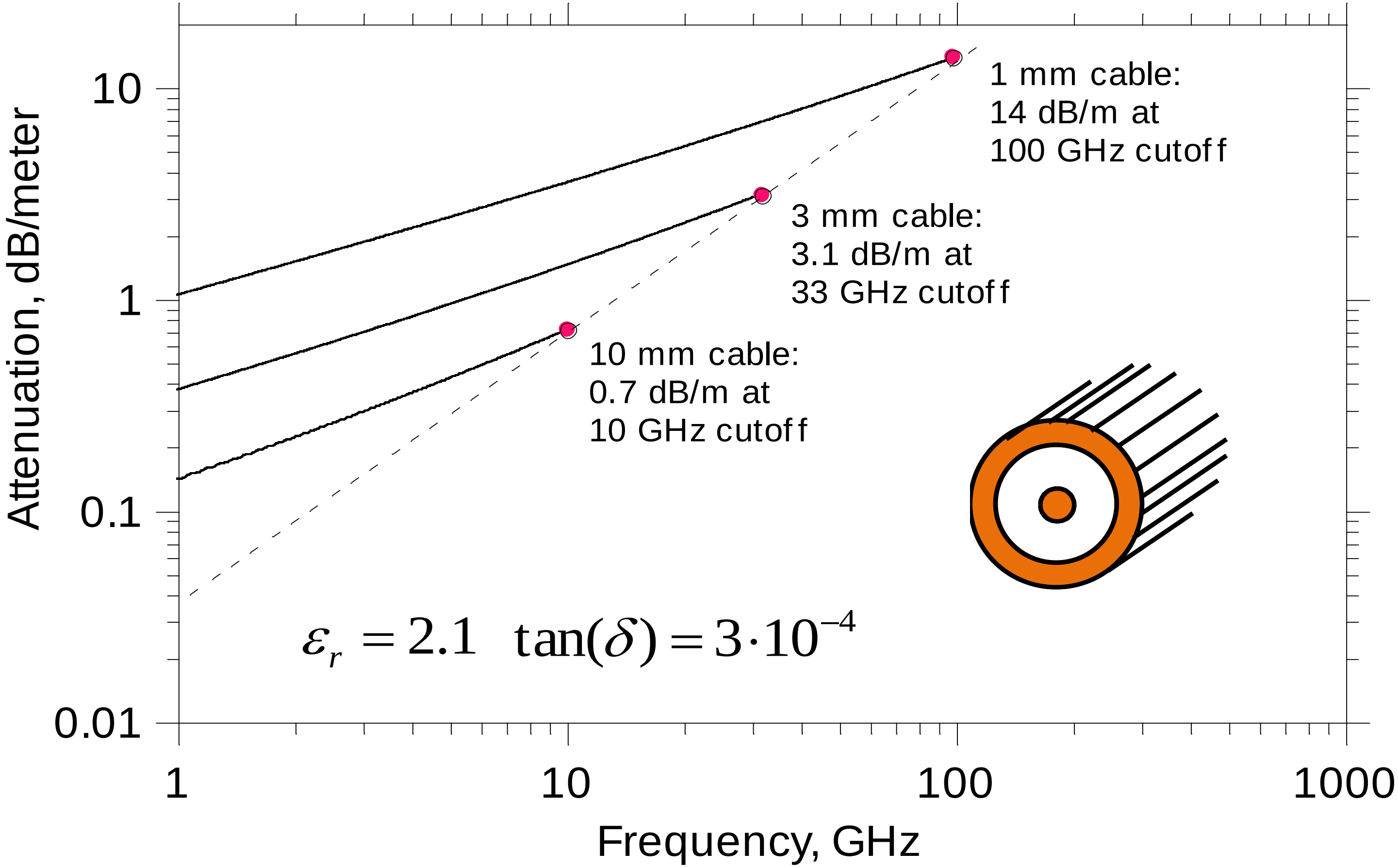
"radiation loss"

Transmission-Line Losses

If we use narrow lines and thin substrates
then skin-effect losses will be large.

If we use wide lines and thick substrates
then lateral modes and substrate radiation
will be major problems.

Loss of Coaxial Cable

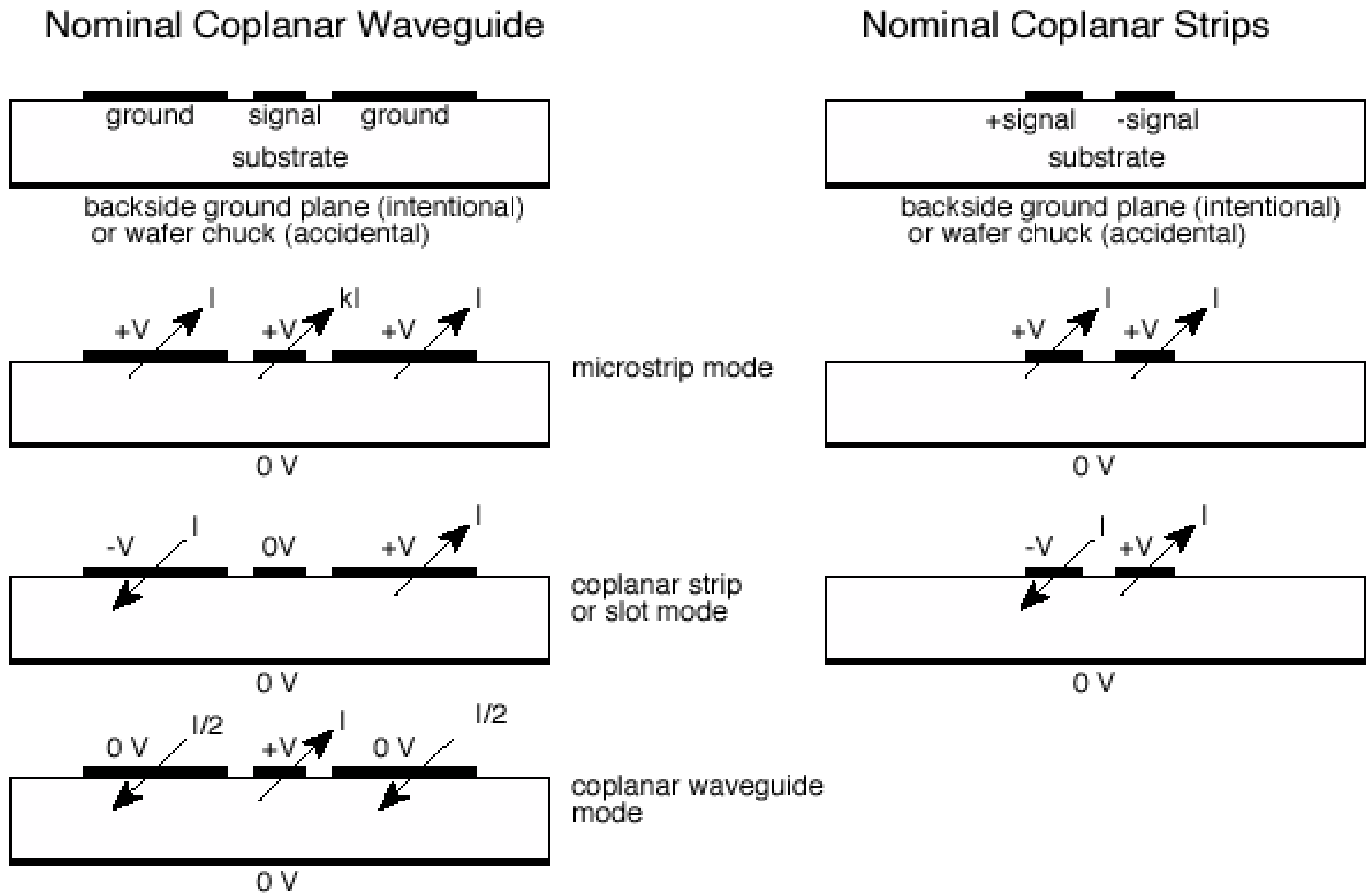


Single-mode propagation requires $f \le c \cdot (2/\pi)\varepsilon_r^{-1/2}\left(D_{inner} + D_{outer}\right)^{-1}$

Skin loss $\alpha_{skin} \propto f^{1/2} / D_{inner} \longrightarrow$ Loss $\alpha_{skin} \propto f^{3/2}$

"circuit-type" parasitic modes

Transmission-Line Parasitic Modes



- Total number of quasi-TEM modes is one less than # of conductors
- Care must be taken to avoid excitation of parasitic modes
- unexpected results will otherwise arise...

To Avoid "Circuit-Type" Parasitic Modes

1) Where do the currents flow?

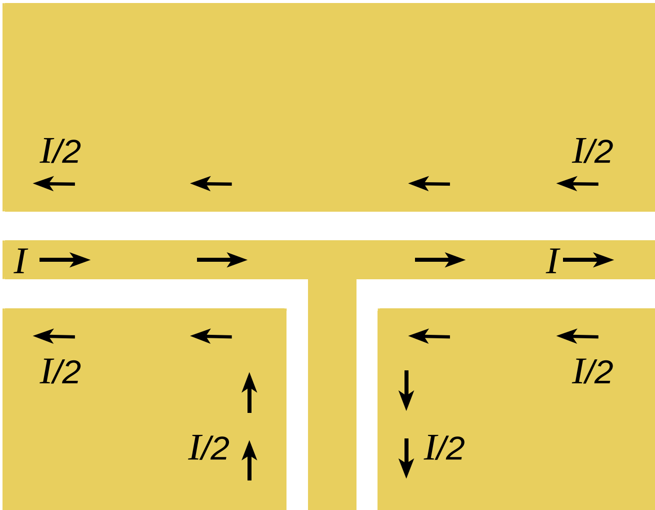
2) Which conductors have what voltages for which modes?

Be aware that:

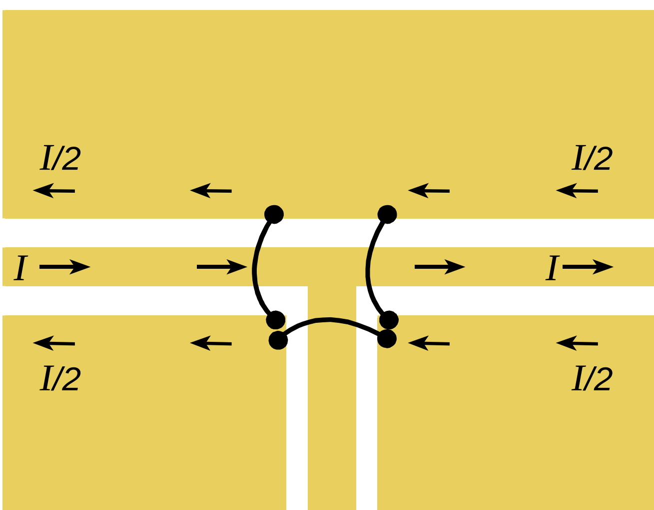
• currents must flow in the ground planes of unbalanced transmission lines. The currents flow close to the edge of the ground plane nearest the signal conductor.

• there are equal and opposite voltages on the 2 conductors of balanced transmission lines. This seriously restricts the types of junctions allowable.

Example of Parasitic Mode Excitation



A slot-line mode is excited at a CPW junction

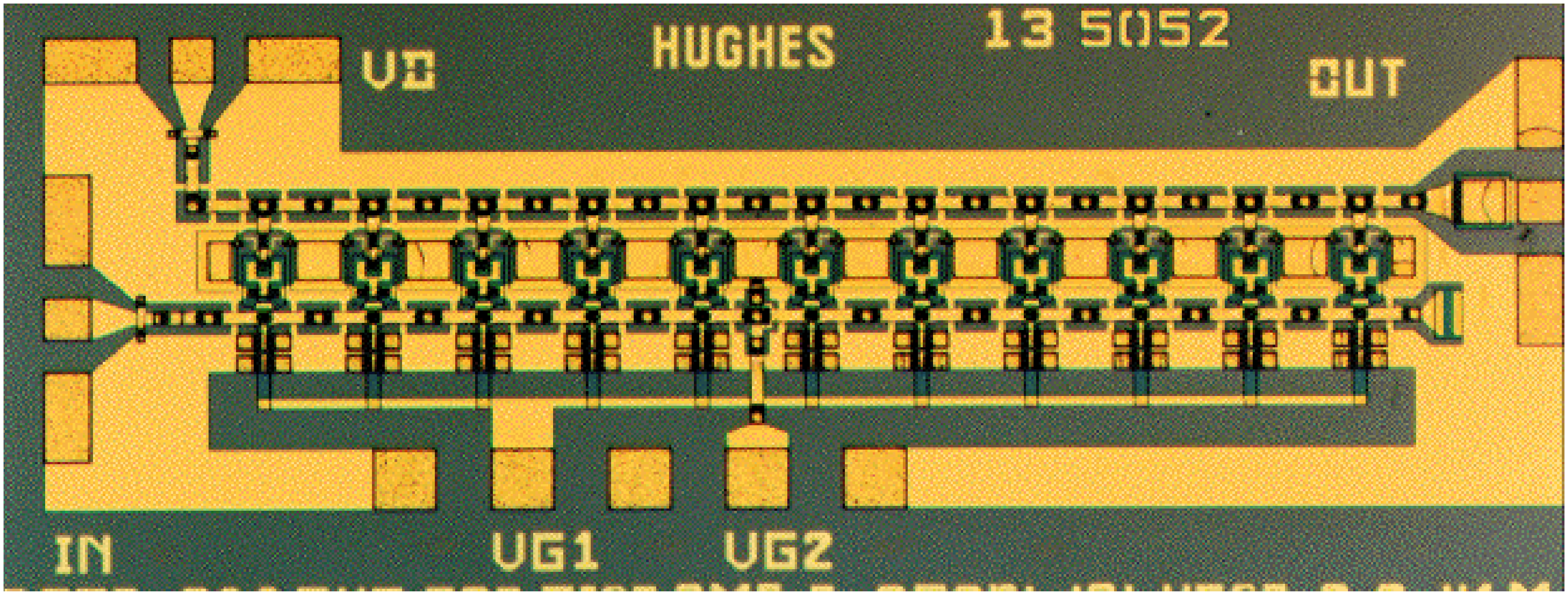


The fix...

this is one of many possible examples...

Example of IC using CPW wiring

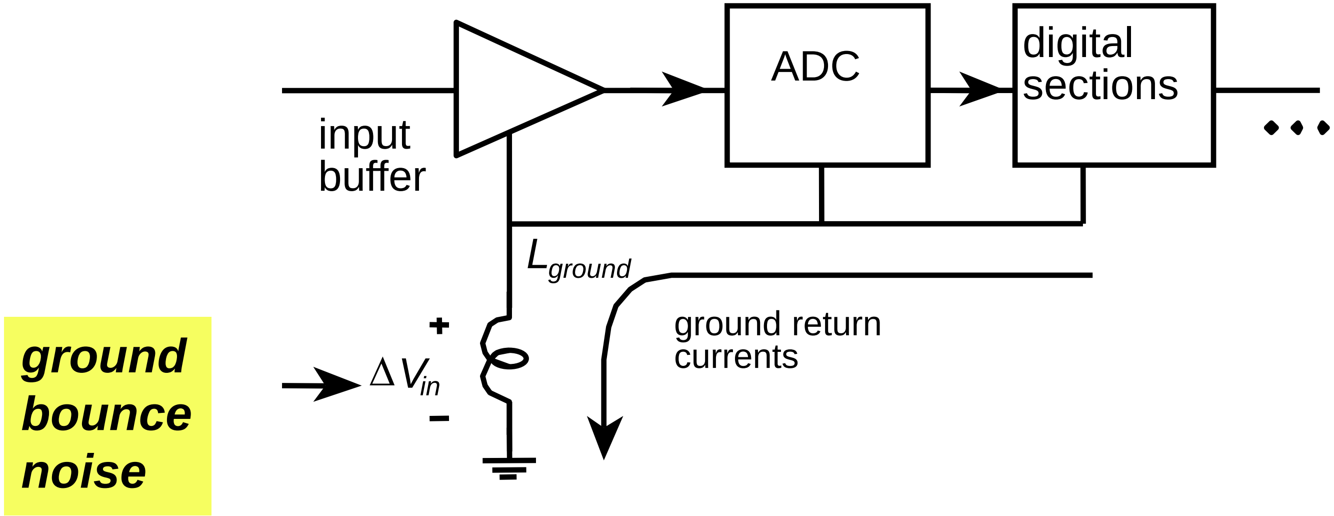
1-180 GHz HEMT amplifier (UCSB / HRL)
Note the ground bridges



B. Agarwal UCSB, M. Matloubian, HRL

package resonance and grounding

What is Ground Bounce ?

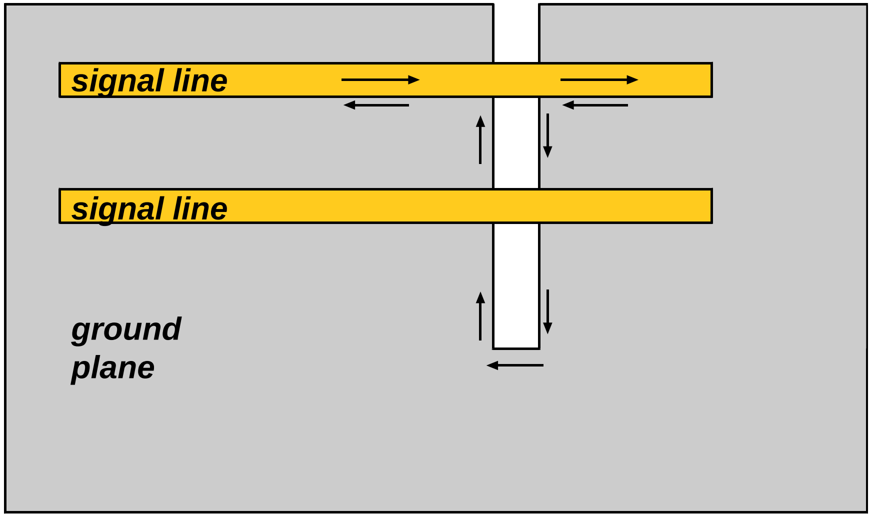


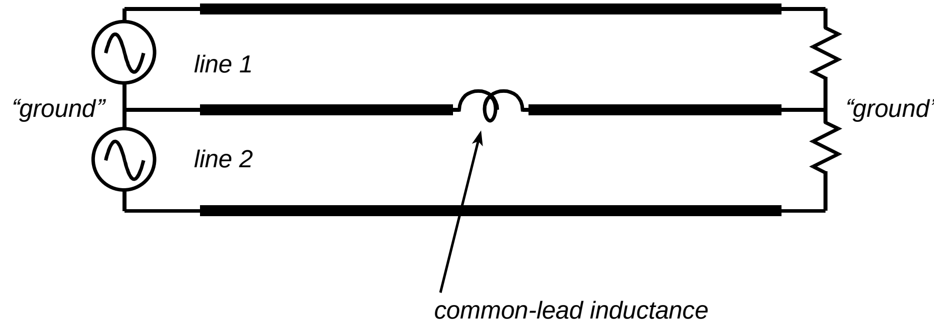
"Ground" simply means a reference potential shared between many circuit paths.

To the extent that it has nonzero impedance, circuits will couple in unexpected ways

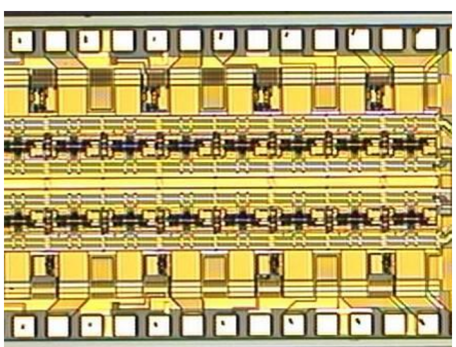
RFI, resonance, oscillation, frequently result from poor ground systems

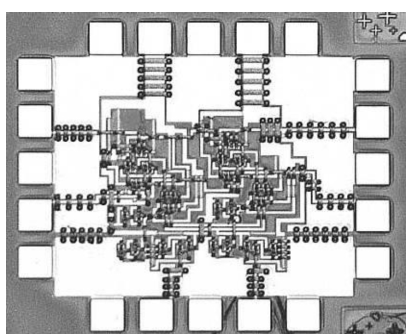
Ground Bounce on an IC: break in a ground plane

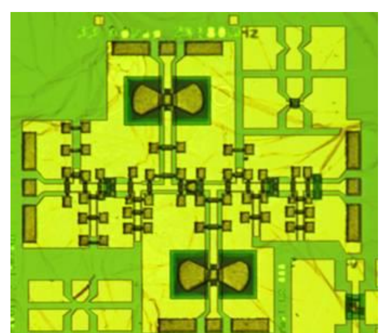




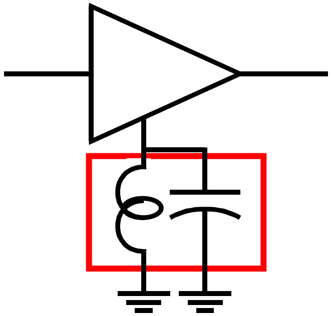
coupling / EMI due to poor ground system integrity is common in high-frequency systems
whether on PC boards
...or on ICs.

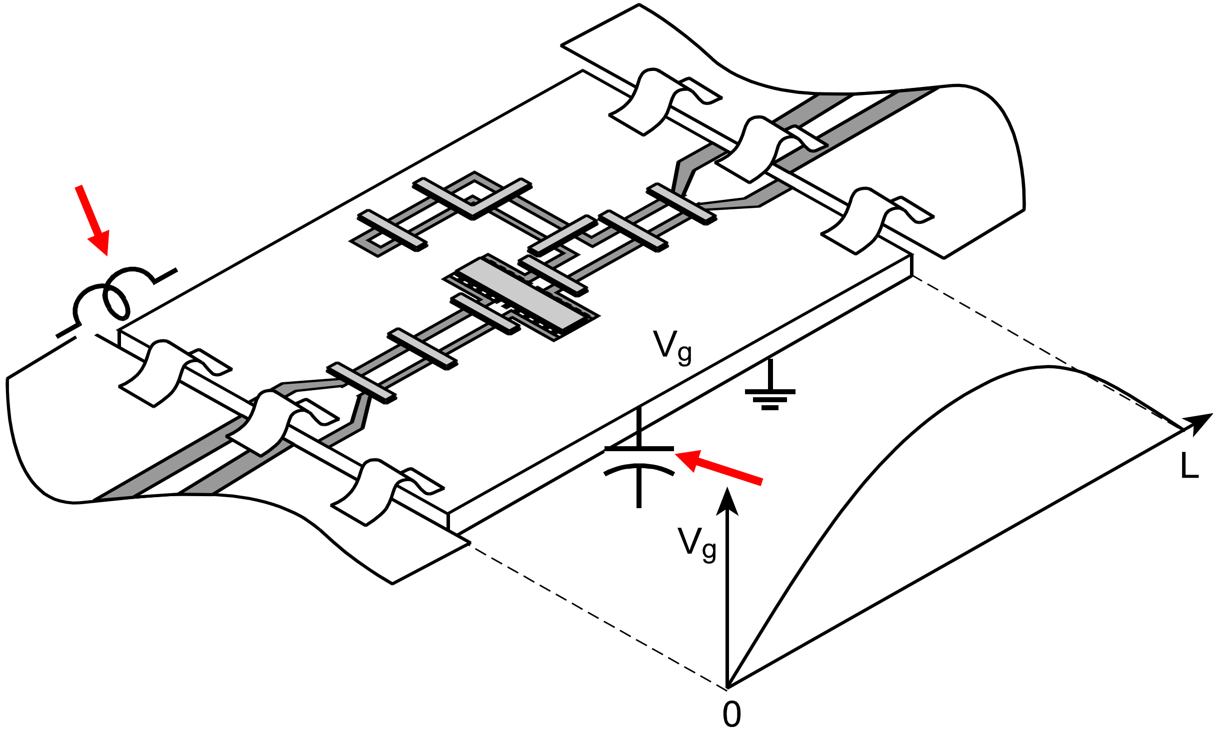






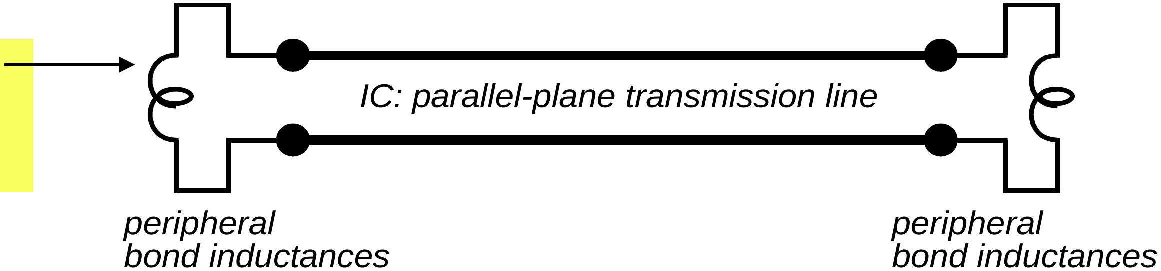
Ground Bounce: IC Packaging with Top-Surface-Only Ground



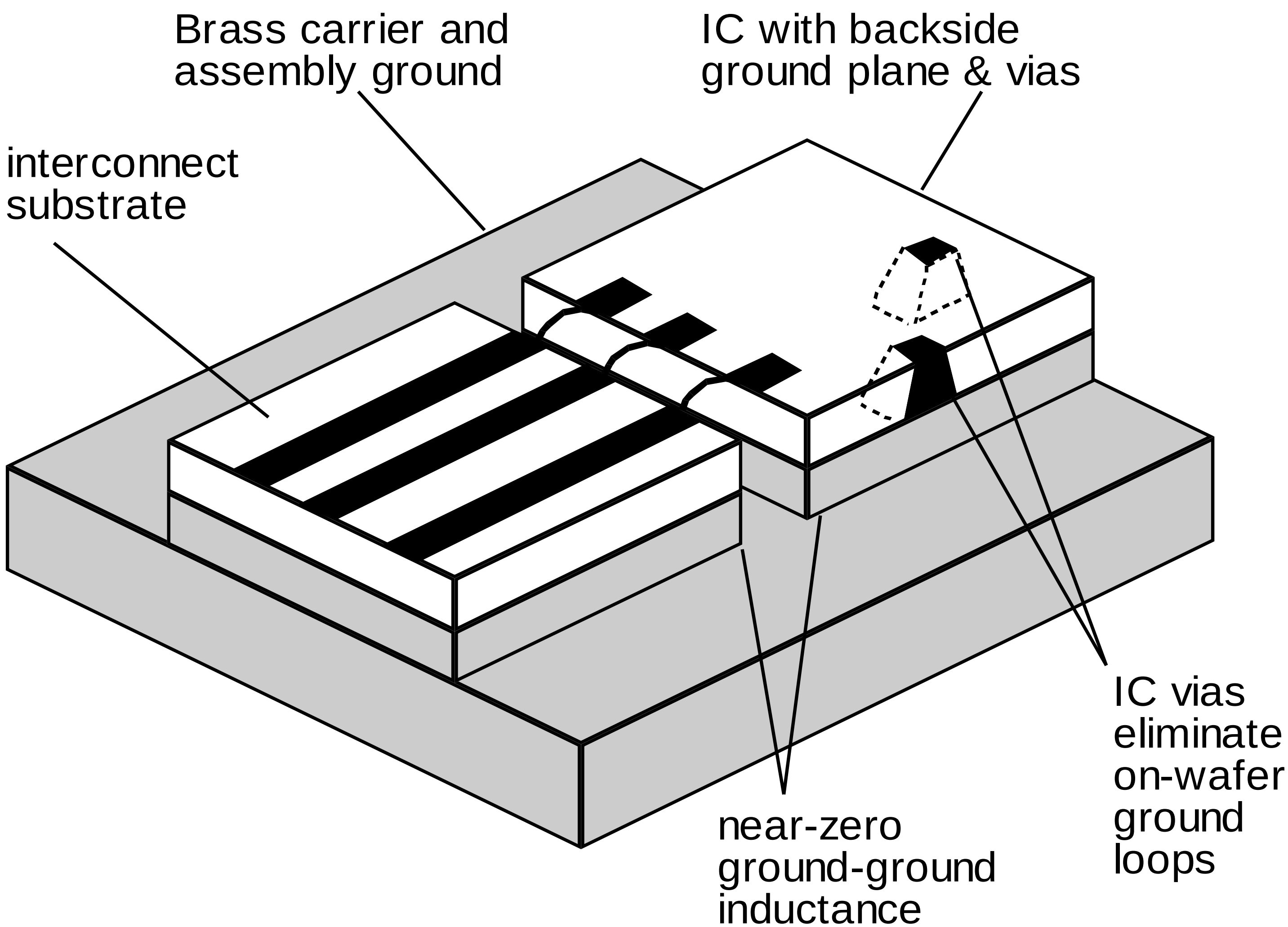


Peripheral grounding allows parallel plate mode resonance
die dimensions must be <0.4mm at 100GHz

Bond wire inductance aggravates the effect: resonates with through-wafer capacitance at 5-20 GHz



Substrate Microstrip: Eliminates Ground Return Problems

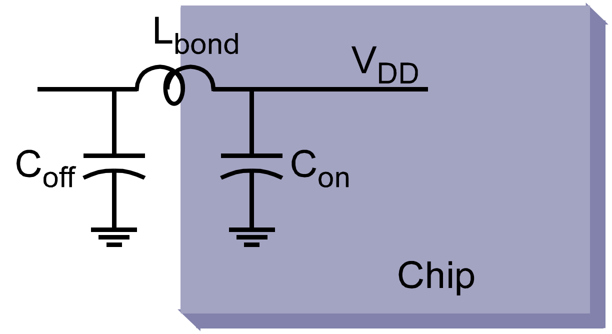


power-supply resonance

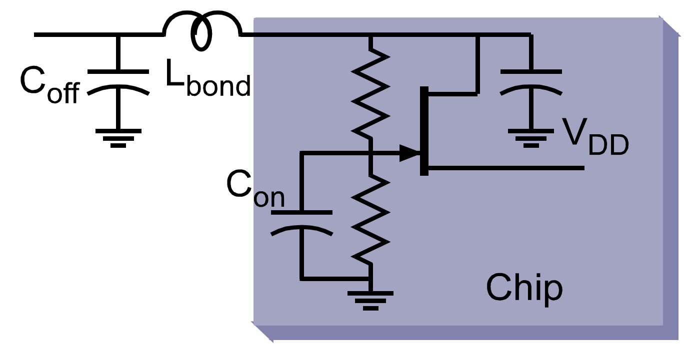
Power Supply Resonance

Resonates at $f = 1/2\pi\sqrt{L_{bond}C_{on}}$

gain peak / suckout, oscillation, etc.



Active (AC) supply regulation

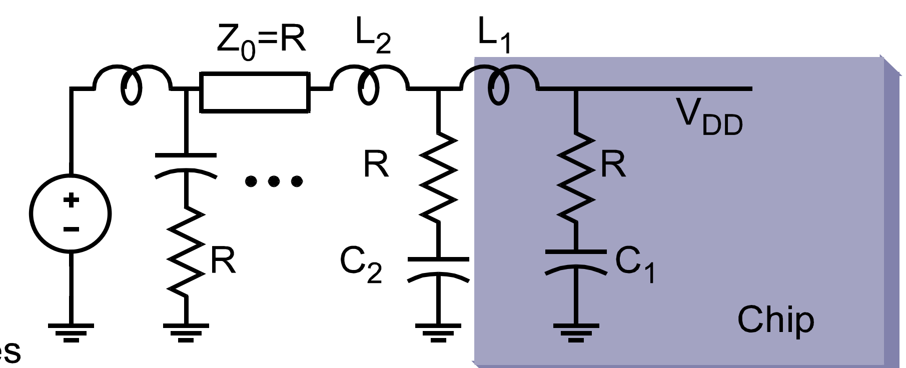


Passive filter synthesis

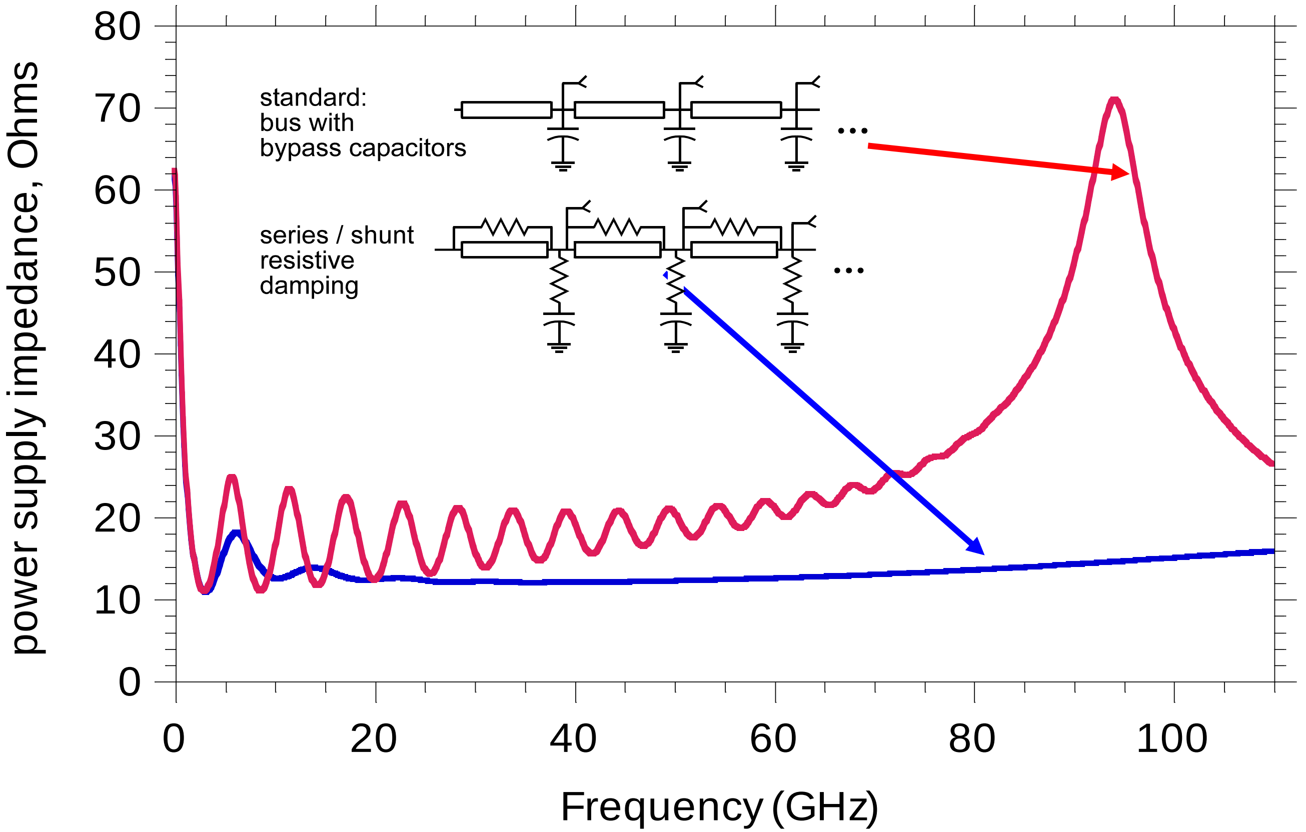
$$R = \sqrt{L_1/C_1}$$

$$\sqrt{L_1/C_1} = \sqrt{L_2/C_2} = \cdots$$

supply impedance is R at all frequencies



Power Supply Resonances; Power Supply Damping

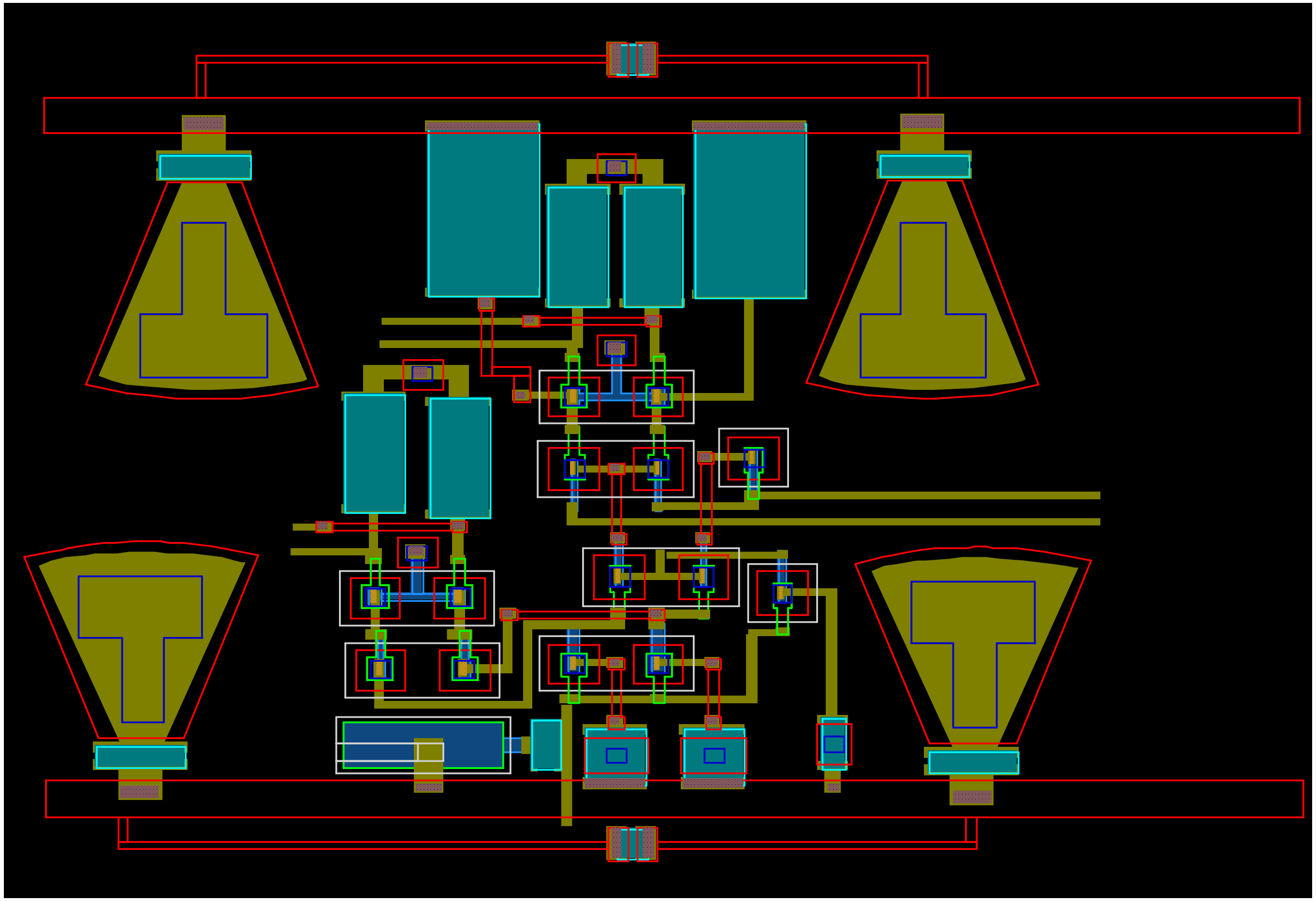


90 GHz--local resonance between power supply capacitance and supply lead inductance

*~N*5GHz resonances--global standing wave on power supply bus*

Power supply is certain to resonate: we must model, simulate, and add damping during design.

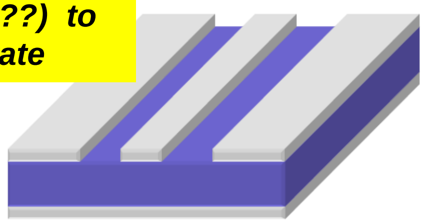
Standard cell showing power busses



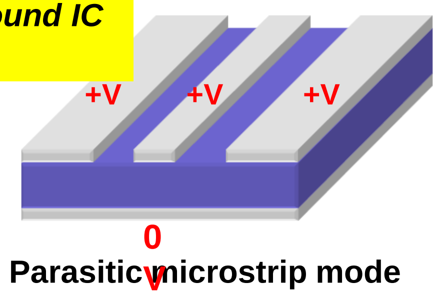
Interconnects:
Summary,
Design Strategy

Coplanar Waveguide: Summary

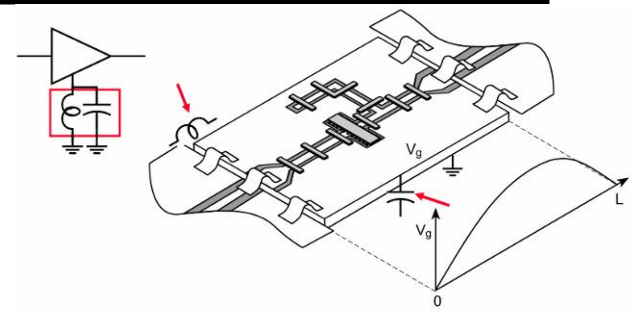
No ground vias
No need (???) to thin substrate



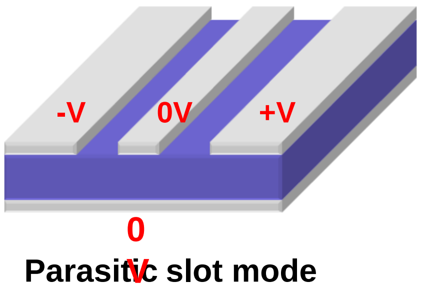
Hard to ground IC to package



Parasitic microstrip mode

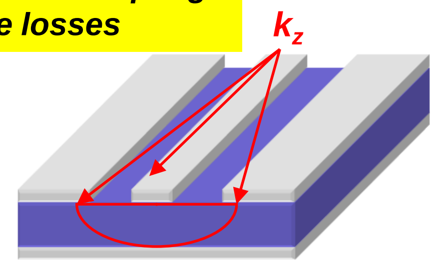


ground plane breaks → loss of ground integrity



Parasitic slot mode

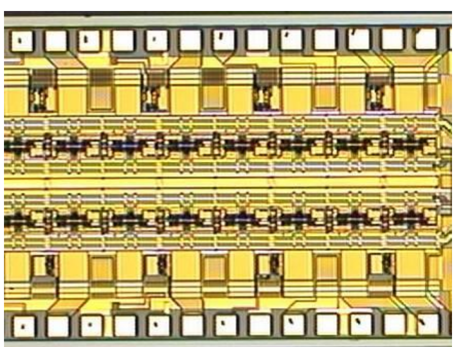
substrate mode coupling or substrate losses



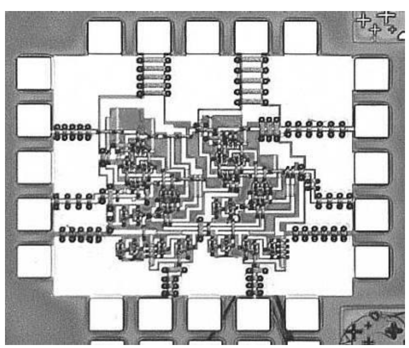
III-V: semi-insulating substrate→ substrate mode coupling

Silicon conducting substrate → substrate conductivity losses

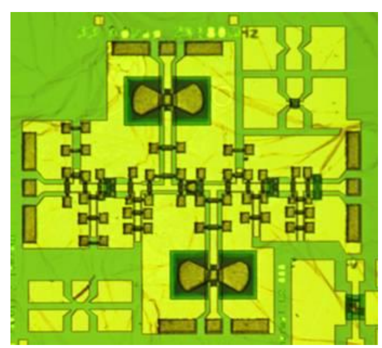
Repairing ground plane with ground straps is effective only in simple ICs
In more complex CPW ICs, ground plane rapidly vanishes
→ common-lead inductance → strong circuit-circuit coupling



40 Gb/s differential TWA modulator driver note CPW lines, fragmented ground plane



35 GHz master-slave latch in CPW note fragmented ground plane



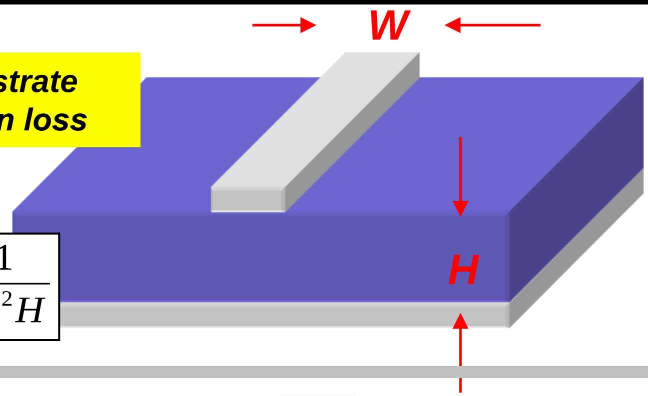
175 GHz tuned amplifier in CPW note fragmented ground plane

poor ground integrity
↓
loss of impedance control
↓
ground bounce
↓
coupling, EMI, oscillation

Classic Substrate Microstrip: Summary

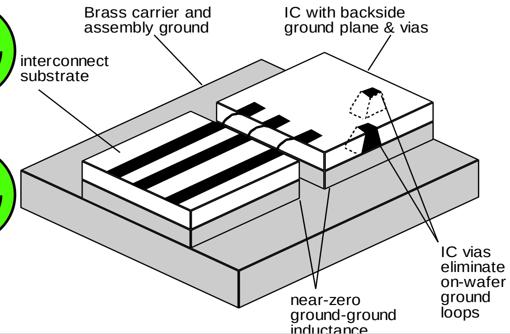
Thick Substrate
→ low skin loss

$$\alpha_{skin} \propto \frac{1}{\varepsilon_r^{1/2} H}$$

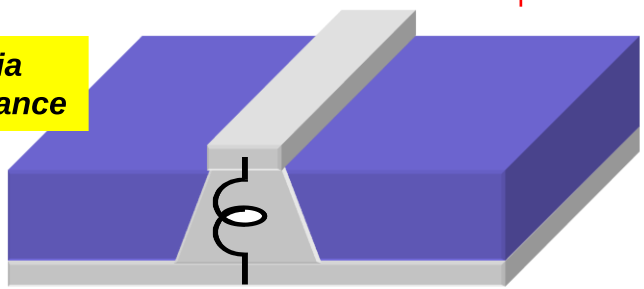


Zero ground inductance in package

No ground plane breaks in IC

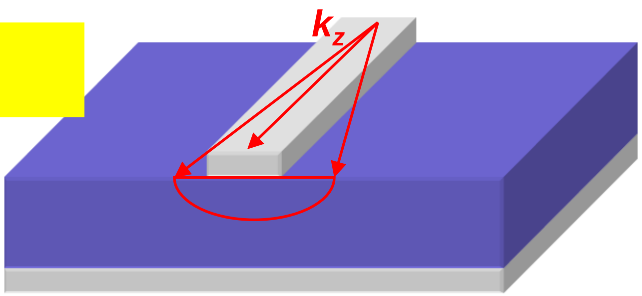


High via inductance



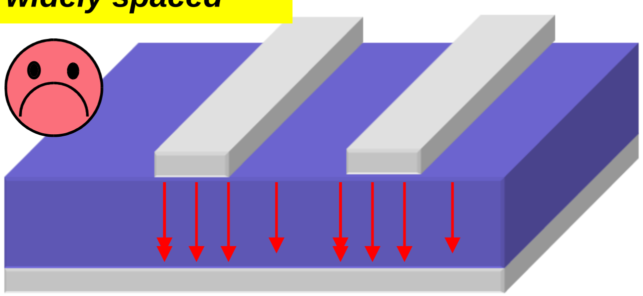
12 pH for 100 μm substrate -- 7.5 Ω @ 100 GHz

TM substrate mode coupling



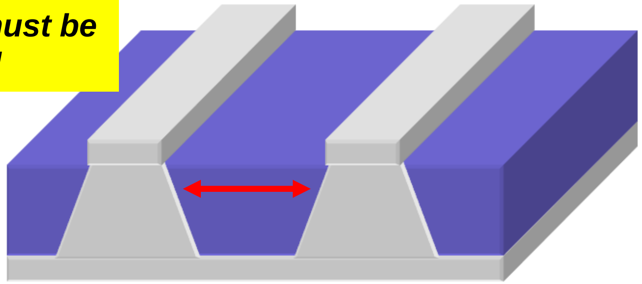
Strong coupling when substrate approaches ~λ_d/ 4 thickness

lines must be widely spaced



Line spacings must be ~3*(substrate thickness)

ground vias must be widely spaced



all factors require very thin substrates for >100 GHz ICs
→ lapping to ~50 μm substrate thickness typical for 100+ GHz

III-V MIMIC Interconnects -- Thin-Film Microstrip

narrow line spacing → IC density

no substrate radiation, no substrate losses

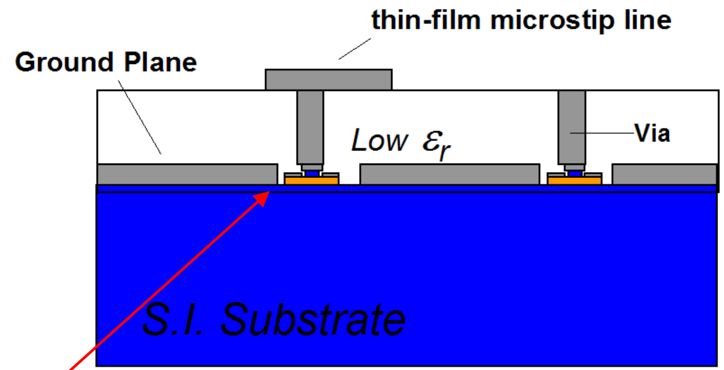
fewer breaks in ground plane than CPW

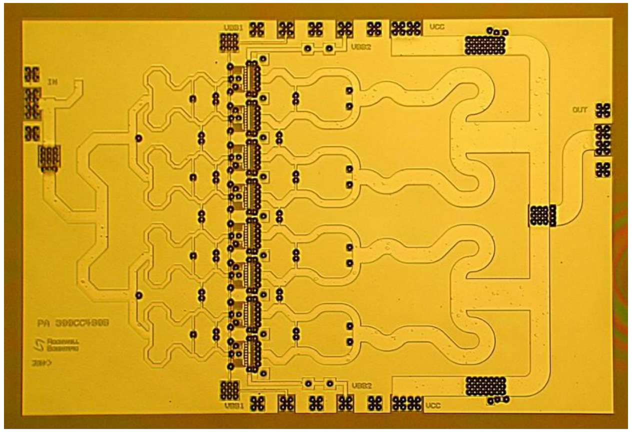
... but ground breaks at device placements

still have problem with package grounding

...need to flip-chip bond

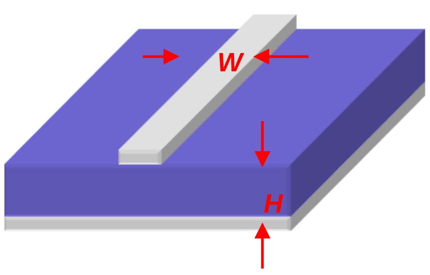
thin dielectrics → narrow lines
→ high line losses
→ low current capability
→ no high-Z_o lines





InP mm-wave PA (Rockwell)

$$Z_o \sim \frac{\eta_o}{\varepsilon_r^{1/2}}\left(\frac{H}{W+H}\right)$$



III-V MIMIC Interconnects -- Inverted Thin-Film Microstrip

narrow line spacing → IC density

Some substrate radiation / substrate losses

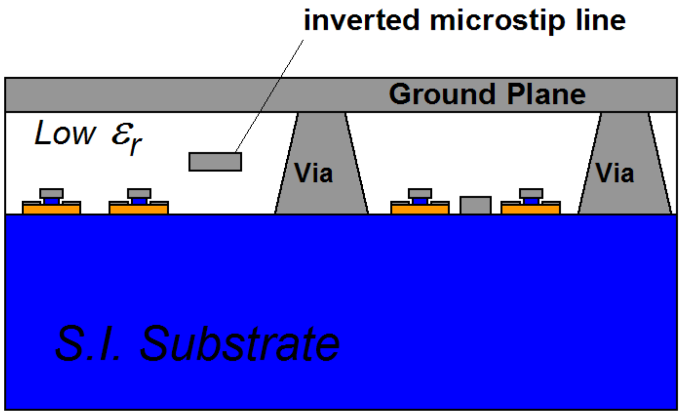
No breaks in ground plane

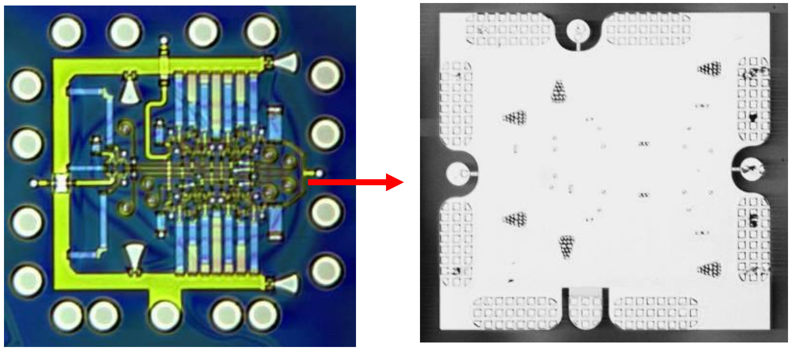
... no ground breaks at device placements

still have problem with package grounding

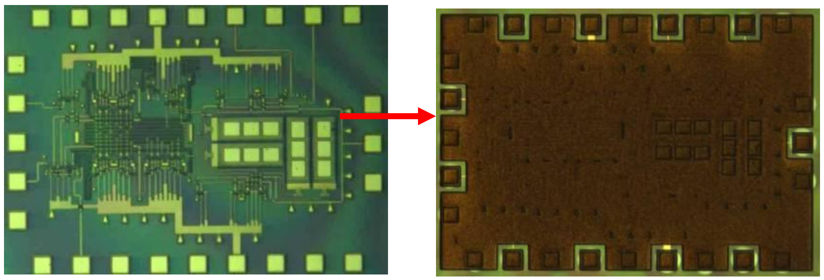
...need to flip-chip bond

thin dielectrics → narrow lines
→ high line losses
→ low current capability
→ no high-Z_o lines





InP 150 GHz master-slave latch



InP 8 GHz clock rate delta-sigma ADC

VLSI Interconnects with Ground Integrity & Controlled Z_0

narrow line spacing → IC density

no substrate radiation, no substrate losses

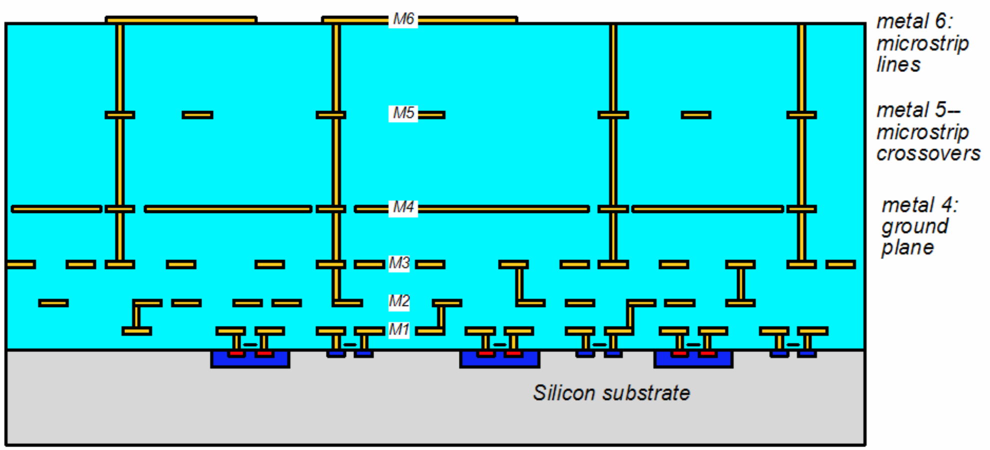
negligible breaks in ground plane

negligible ground breaks @ device placemen

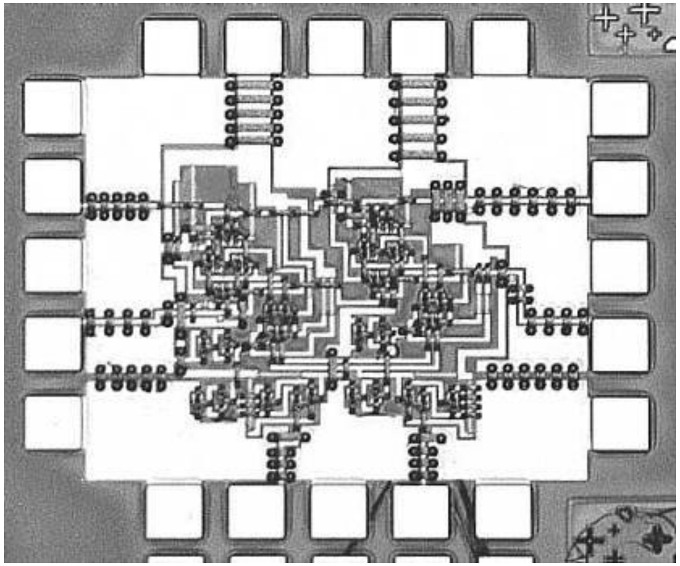
still have problem with package grounding

...need to flip-chip bond

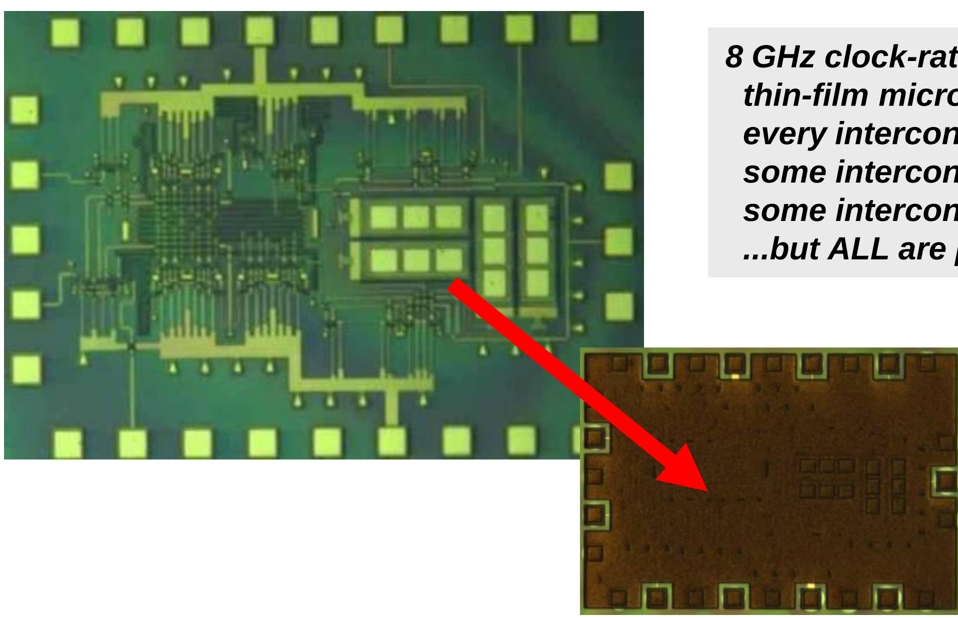
thin dielectrics → narrow lines
→ high line losses
→ low current capability
→ no high-Z_o lines



No clean ground return ? → interconnects can't be modeled !



35 GHz static divider
interconnects have no clear local ground return
interconnect inductance is non-local
interconnect inductance has no compact model



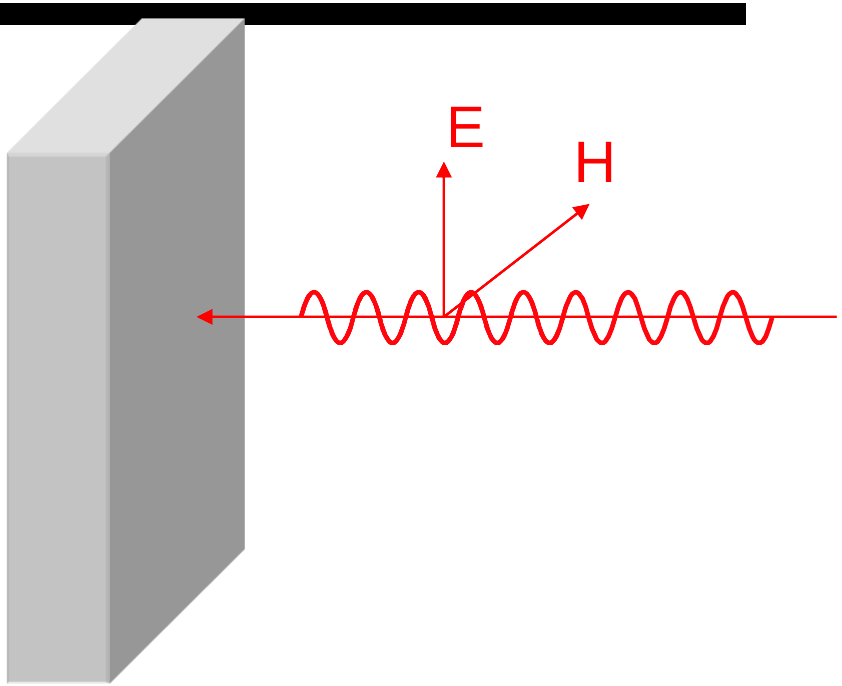
8 GHz clock-rate delta-sigma ADC
thin-film microstrip wiring
every interconnect can be modeled as microstrip
some interconnects are terminated in their Zo
some interconnects are not terminated
...but ALL are precisely modeled

InP 8 GHz clock rate delta-sigma ADC

End

Appendix (optional)

Skin effect losses I



Given a plane wave perpendicularly incident in direction z onto a sheet of metal :

$$\frac{\partial E}{\partial z} = -j\omega\mu H \quad \text{and} \quad \frac{\partial H}{\partial z} = -(j\omega\varepsilon + \sigma)E$$

Hence $E(z) = E_o e^{-\gamma z}$ where $\gamma = \sqrt{j\omega\mu(j\omega\varepsilon + \sigma)}$

If $\omega\varepsilon << \sigma$, then $\gamma \cong \sqrt{j\omega\mu\sigma} = \sqrt{\omega\mu\sigma/2} + j\sqrt{\omega\mu\sigma/2}$

Defining the skin depth as $\delta = \sqrt{2/\omega\mu\sigma}$ we find that

$$E(z) = E_o e^{-z/\delta} e^{-jz/\delta}$$

...the field dies down exponentially with distance into the metal.

Wave impedance in the metal is :

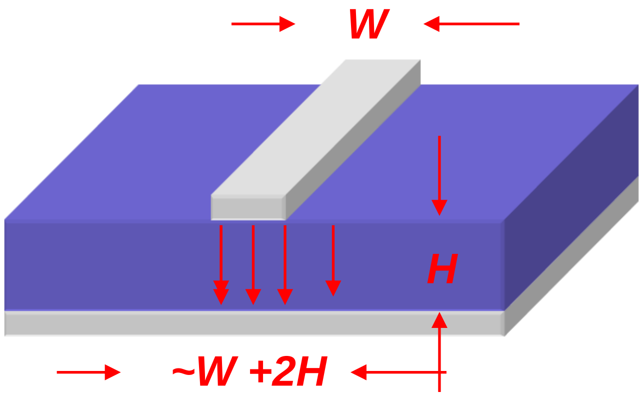
$$\eta_{metal} = \frac{E}{H} = \sqrt{\frac{j\omega\mu}{j\omega\varepsilon + \sigma}} \cong \sqrt{\frac{j\omega\mu}{\sigma}} = \sqrt{\frac{\omega\mu}{2\sigma}} + j\sqrt{\frac{\omega\mu}{2\sigma}}$$

hence,

$$\eta_{metal} = \frac{1}{\sigma\delta} + j\frac{1}{\sigma\delta} \leftarrow \text{note the resistive and inductive terms}$$

This is the SURFACE IMPEDANCE.

Skin effect losses II



In a transmission line, the wave travels parallel, not perpendicular to the metal surface, but the same surface impedance is seen, provided that the transmission - line wavelength is much larger than the skin depth

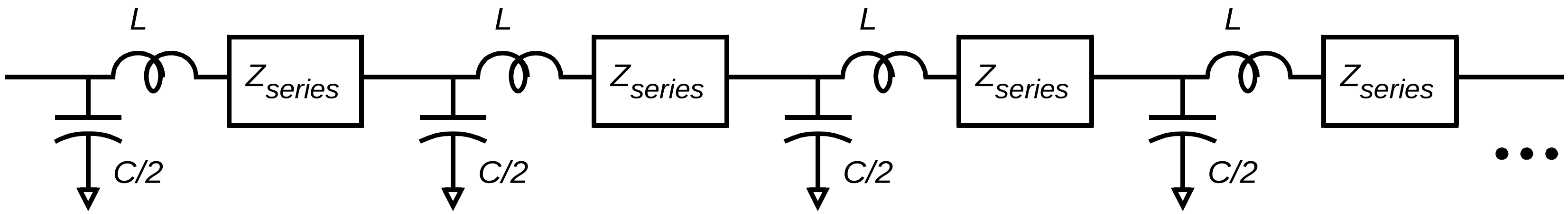
The transmission - line then has an added series impedance per unit distance of

$Z_{series} = \dfrac{1}{P}\dfrac{(1+j)}{\delta\sigma}$, where P is the effective current - carrying periphery.

For this microstrip line, there is surface impedance both in the signal and ground lines

$$Z_{series} \approx \left(\frac{1}{W} + \frac{1}{W+2H}\right)\frac{(1+j)}{\delta\sigma}$$

Skin effect losses III



This then introduces both loss and dispersion

$$Z_{series}=\frac{1}{P}\frac{(1+j)}{\delta\sigma}\quad\rightarrow\frac{\partial V}{\partial z}=-(j\omega L+Z_{series})I\text{ and }\frac{\partial I}{\partial z}=-j\omega CV$$

$$\rightarrow Z_o=\frac{V^+(z)}{I^+(z)}=\sqrt{\frac{j\omega L+Z_{series}}{j\omega C}}=\sqrt{\frac{j\omega L+1/P\delta\sigma+j/P\delta\sigma}{j\omega C}}$$

....some secondary change in characteristic impedance

$$\rightarrow V(z)=V_oe^{-\gamma_{line}z}\text{, where}$$

$$\gamma_{\text{line}}=\sqrt{(j\omega L+Z_{series})j\omega C}=j\omega\sqrt{LC}\sqrt{1+Z_{series}/j\omega L}$$

$$\cong j\omega\sqrt{LC}(1+Z_{series}/j2\omega L)=j\omega\sqrt{LC}+Z_{series}\left(\sqrt{C/L}\right)/2$$

$$\gamma_{\text{line}}=j\omega\sqrt{LC}+Z_{series}/2Z_o=j\omega\sqrt{LC}+\frac{1}{2Z_0P\delta\sigma}+\frac{j}{2Z_0P\delta\sigma}$$

Skin Loss dispersion

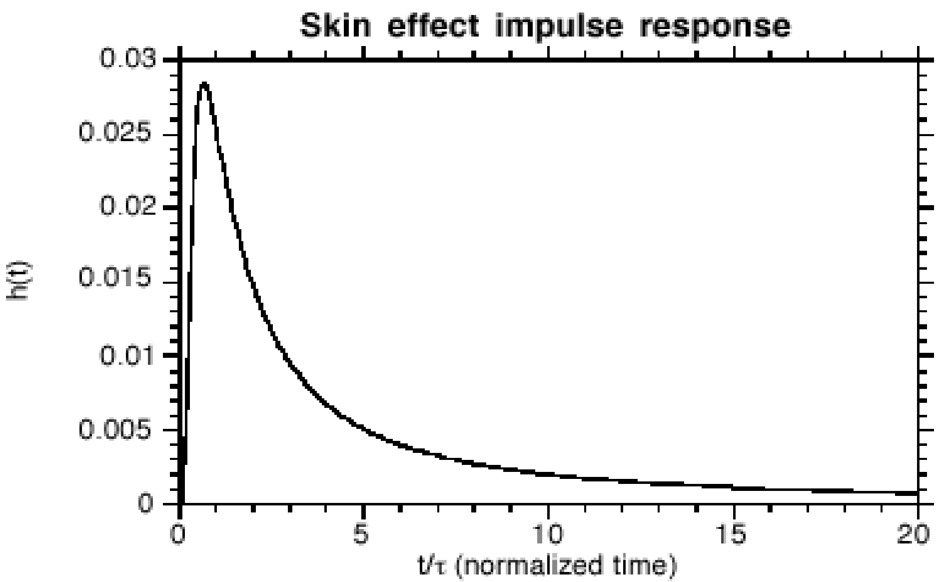
Skin Effect losses, IV

The impulse response of the transmission line can then be found.
(Wiginton and Nahman, Proc. IRE, February 1957)

$$h(t) \cong C * U(t/\tau)\frac{(t/\tau)^{-3/2}\exp(-\tau/t)}{\tau\sqrt{\pi}}$$

$$where \quad \tau = \left[l\sqrt{\frac{\mu}{\sigma}} \Big/ 4Z_0 P \right]^2$$

Skin effect causes pulse broadening proportional to distance^2



Skin effect losses V

The step response is the integral of the impulse response.
Note the initial fast rise and the subsequent "dribble-up" characteristic of skin effect losses.