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)

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

The wavelength along the transmission-line is long, and the penetration distance of current into the metal is small,

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

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

$$k_y^2 = j\omega\mu\sigma$$
Skin effect losses II

In a metal, at low frequencies \( j\omega\varepsilon \ll \sigma \), 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 \(\delta\)

\(\delta\) varies as \(\omega^{-1/2}\)

At 100GHz in Gold, \(\delta \approx 200\) nm
Skin effect losses III

Let us treat this approximately:

The conductor only carries current in a layer of thickness $\delta$. With conductivity $\sigma$ and width $W$, the conductor has resistance per unit length

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

A more careful treatment develops the concept of surface impedance. See the appendix.
Transmission - lines have skin effect in both the signal and ground conductors. For microstrip:

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

In general, we can write this as

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

where \( P \) is the effective current - carrying periphery.
Skin effect losses IV

\[ R_{\text{series}} / L = \left(1 / P\right) \cdot \sqrt{\omega \mu / 2\sigma} \]

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

\[ \alpha \approx \frac{R_{\text{series}} / L}{2Z_0} \quad \alpha \propto \sqrt{\omega} \]

This is called skin loss
Loss Tangent
Loss Tangent

Common dielectrics also introduce high-frequency attenuation.

This effect is quantified by a *loss tangent*

\[ \varepsilon_r = \varepsilon_{r,\text{real}} + j\varepsilon_{r,\text{imagine}} = \varepsilon_{r,\text{real}} (1 + j \tan(\delta)) \]

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:

\( k_y = 0 \) and \( k_x = n\pi / W \) for \( n = 0, 1, 2, ... \)

\[
\rightarrow k_z^2 = \varepsilon_r \omega^2 / c^2 - (n\pi / W)^2
\]
Lateral Modes (2)

\[ k_z^2 = \varepsilon_r \omega^2 / c^2 - \left( n\pi / W \right)^2 \]

1) Multi-mode propagation if \( W > \lambda_d / 2 \):

\[ \beta_z = \sqrt{\varepsilon_r \omega^2 / c^2 - \left( n\pi / W \right)^2} \]

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

\[ \alpha_z = \sqrt{\left( n\pi / W \right)^2 - \varepsilon_r \omega^2 / c^2} \]
Lateral Modes---and Junction Parasitics (3)

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

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 $E_0 e^{j\omega t} e^{\pm jk_x x} e^{\pm jk_y y} e^{\pm jk_z z}$

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

We can have standing waves across the substrate thickness:

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

Substrate with top, bottom metal surfaces
→ modes with \( h = \lambda_d / 2, \lambda_d, 3\lambda_d / 2 \ldots \)

Substrate with no top metal → tranverse \( E \)-mode;
strongly confined as \( \lambda_d / 4 \rightarrow T \); weakly confined at low frequencies.
These dielectric slab modes can propagate in $x$ and in $z$.

Nonzero mode coupling ("radiation") loss at all frequencies. Very strong mode coupling when $h \geq \frac{\lambda_d}{4}$
Substrate modes are allowed when $\lambda_d \leq 2 \cdot (\text{substrate thickness})$

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 $k_y = (n\pi / h)$
Modes couple strongly when $k_{y,CPW} = k_{y,substrate mode}$

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

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

"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 substrateradiation will be major problems.
Loss of Coaxial Cable

Single-mode propagation requires

\[ f \leq c \cdot \frac{2}{\pi} \varepsilon_r^{-1/2} \left( D_{\text{inner}} + D_{\text{outer}} \right)^{-1} \]

Skin loss \( \alpha_{\text{skin}} \propto f^{1/2} / D_{\text{inner}} \quad \rightarrow \quad \text{Loss} \quad \alpha_{\text{skin}} \propto f^{3/2} \]
"circuit-type"
parasitic modes
Transmission-Line Parasitic Modes

Nominal Coplanar Waveguide

Nominal Coplanar Strips

- 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
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.

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Ground bounce noise

$\Delta V_{in}$
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

- Brass carrier and assembly ground interconnect substrate
- IC with backside ground plane & vias
- Near-zero ground-ground inductance
- IC vias eliminate on-wafer ground loops
power-supply resonance
Power Supply Resonance

Resonates at \( f = \frac{1}{2\pi} \sqrt{\frac{L_{\text{bond}}}{C_{\text{on}}}} \)

gain peak / suckout, oscillation, etc.

Active (AC) supply regulation

Passive filter synthesis

\[ R = \sqrt{\frac{L_1}{C_1}} \]
\[ \sqrt{\frac{L_1}{C_1}} = \sqrt{\frac{L_2}{C_2}} = \cdots \]

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
Parasitic slot mode

Hard to ground IC to package

Parasitic microstrip mode

ground plane breaks → loss of ground integrity

substrate mode coupling or substrate 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.

poor ground integrity

loss of impedance control

ground bounce

coupling, EMI, oscillation
**Classic Substrate Microstrip: Summary**

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**Thick Substrate** → low skin loss

\[ \alpha_{\text{skin}} \propto \frac{1}{\varepsilon_r^{1/2} H} \]

**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 \( \approx \lambda/4 \) thickness

**lines must be widely spaced**

**ground vias must be widely spaced**

Line spacings must be \( \approx 3 \times \) (substrate thickness)

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all factors require very thin substrates for >100 GHz ICs

→ lapping to \( \approx 50 \) 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-$Z_0$ lines**

**Expression:**

$$Z_o \sim \frac{n_0}{\varepsilon_r^{1/2}} \left( \frac{H}{W + H} \right)$$
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-$Z_0$ 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 placement**

- **still have problem with package grounding**

  ...need to flip-chip bond

- **thin dielectrics** → narrow lines
  → high line losses
  → low current capability
  → no high-$Z_0$ lines

- **no substrate radiation, no substrate losses**
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
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 \( \gamma \approx \sqrt{j \omega \mu \sigma} = \sqrt{\omega \mu \sigma / 2} + j \sqrt{\omega \mu \sigma / 2} \)

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:

\[
\eta_{metal} = \frac{E}{H} = \sqrt{\frac{j \omega \mu}{j \omega \varepsilon + \sigma}} \approx \sqrt{\frac{j \omega \mu}{\sigma}} = \sqrt{\frac{\omega \mu}{2 \sigma}} + j \sqrt{\frac{\omega \mu}{2 \sigma}}
\]

hence,

\[
\eta_{metal} = \frac{1}{\sigma \delta} + j \frac{1}{\delta \sigma} \quad \text{← note the resistive and inductive terms}
\]

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

\[ Z_{\text{series}} = \frac{1}{P} \left( 1 + \frac{1}{\delta \sigma} \right), \]

where P is the effective current-carrying periphery.

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

\[ Z_{\text{series}} \approx \left( \frac{1}{W} + \frac{1}{W + 2H} \right) \left( 1 + \frac{1}{\delta \sigma} \right) \]
Skin effect losses III

This then introduces both loss and dispersion

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

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

...some secondary change in characteristic impedance

\[ V(z) = V_o e^{-\gamma_{\text{line}} z}, \text{ where} \]

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

\[ \approx j \omega \sqrt{LC} (1 + Z_{\text{series}}/j 2 \omega L) = j \omega \sqrt{LC} + Z_{\text{series}} \left(\sqrt{C/L}\right)/2 \]

\[ \gamma_{\text{line}} = j \omega \sqrt{LC} + Z_{\text{series}}/2Z_o = j \omega \sqrt{LC} + \frac{1}{2Z_0 P \delta \sigma} + \frac{j}{2Z_0 P \delta \sigma} \]

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)

Skin effect causes pulse broadening proportional to distance

\[ h(t) \equiv C \ast U(t/\tau) \left(\frac{t/\tau}{\tau} \right)^{-3/2} \exp\left(-\frac{\tau}{t}\right) \]

where \( \tau = \left[ l \sqrt{\frac{\mu}{\sigma}} / 4Z_0 P \right]^2 \)
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.

![Skin Effect Step Response](image)

still hasn't reached 0.9 Volts!