Module-2
RF PCB Design
Microstrip Discontinuities & Simulation in Momentum

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Today's Topics

- Modeling &n EM Simulations
  - Schematic Simulation & Layout Simulation
  - Momentum Simulation
- Microstrip Discontinuities
- GND path Breaks
- AC resistance and Skin Effect
- Differential Signaling
- Odd and Even Mode Impedances
- Decoupling Capacitors

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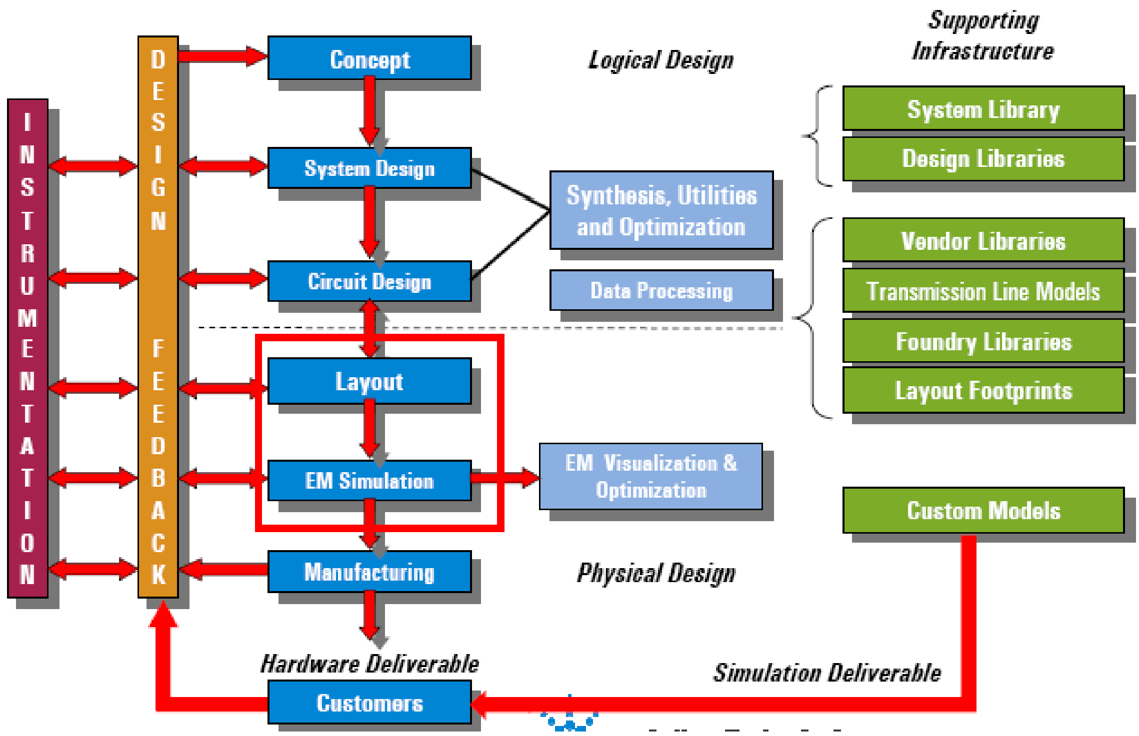
Design Flow



RF PCB …and….Maxwell

Maxwell's Equations

Faraday's Law:

$$\nabla \times \mathbf{E} = -\frac{\partial}{\partial t}\mathbf{B}$$

Ampère's Circuital Law:

$$\nabla \times \mathbf{H} = \mathbf{J} + \frac{\partial}{\partial t}\mathbf{D}$$

Gauss' Laws:

$$\nabla \cdot \mathbf{B} = 0 \qquad \nabla \cdot \mathbf{D} = \rho$$

Constitutive Equations:

$$\mathbf{B} = \mu \mathbf{H} \qquad \mathbf{D} = \varepsilon \mathbf{E}$$

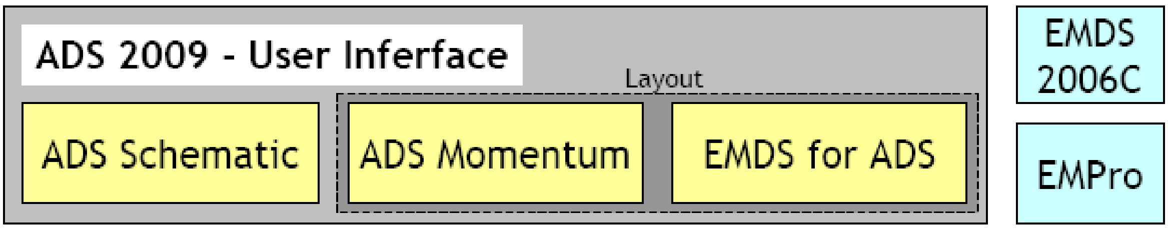
Actual solution complex and for realistic problems require approximations

Solutions to Maxwell's equations using numerical approximations of is known as the study of computational electromagnetic (CEM)

CEM Techniques

- Examples of full-wave computational electromagnetic (CEM) techniques include:
  - Finite difference time domain (FDTD) Method
  - Method of Moments (MoM) Method
  - Finite Element (FE) Method
  - Transmission Line Matrix (TLM) Method
  - The Method of Lines (MoL)
  - The Generalized Multipole Technique (GMT)

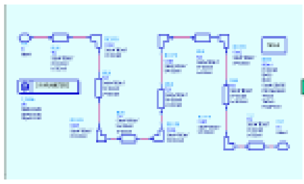
ADS Software Overview

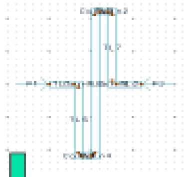


- **ADS Schematic:** Circuit-theory based simulator
  - Analytical models determine input-output behaviour of components
- **ADS Momentum:** 2.5D-Method-of-Moments (MoM) simulator
  - Full wave solver for layered media, arbitrary planar geometry, vias
- **EMDS for ADS:** 3D-Finite Element Method (FEM) simulator
  - Full wave solver that can handle finite substrates and include 3D models
- **EMDS** (Stand alone): 3D-Finite Element Method (FEM) simulator
  EMPro 3D-Finite Difference Time Domain (FDTD) simulator
  - EMDS and EMPro have their own 3D-CAD-GUI

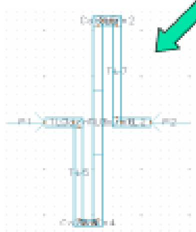
Design Steps ….ADS & Momentum

* Substrate Is computed In Momentum or Can be read from a pre-computed database.

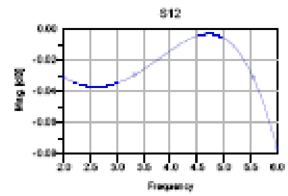




ADS interface
- 1a) Schematic with artwork
- 1b) Layout artwork

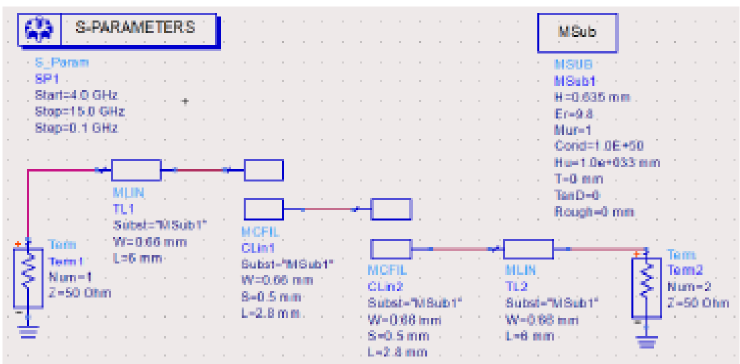


MOM interface
- 2) Define the substrate*
- 3) Mesh and solve
- 4) Display Results

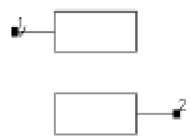


ADS Schematic

- Standard components; e.g. Microstrip filter sections



Microstrip filter section



Range of Usage

$0.1 \le \frac{W}{H} \le 10$

$0.1 \le \frac{S}{H} \le 10$

$1 \le Er \le 18$

Simulation frequency $\le \frac{25}{H(mm)}$ (GHz)

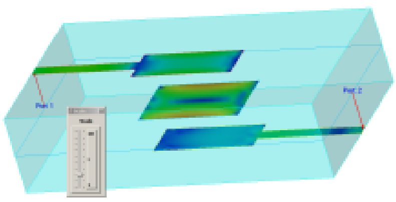
- (+) Fast and accurate results
- (+) Design Guides available
- (+) Real time tuning capabilities
- !Make sure that models are within their ***Range of Usage!***

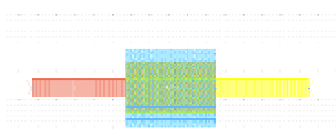
ADS Momentum

- 2.5D (vias) simulation tool for passive circuits.
- Layout driven (accepts arbitrary geometry).
- Method of Moments technique as planar solver.
- Green s functions used to solve integral equations.
- Unlimited substrate database for computation.
- Full microwave or RF mode S-Parameter solutions.
- Layout look-alike components for ADS simulation.
- Co-simulation with ADS and optimization.
- Visualization of current and far-field patterns.

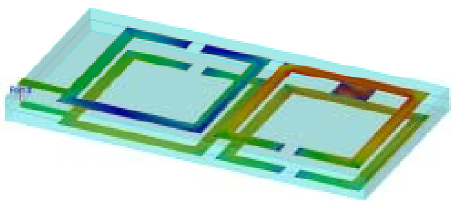
ADS Momentum

- Arbitrary 2D metallic structures and vias in <u>layered media</u>
- <u>Substrates</u> are <u>infinite</u> in lateral direction!
- Slots in metallic sheets
- Multilayer designs
- Planar antennas
- Visualisation of Current density

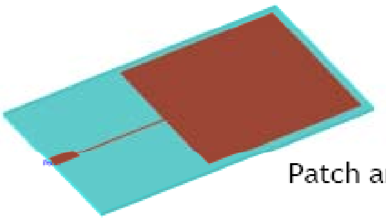


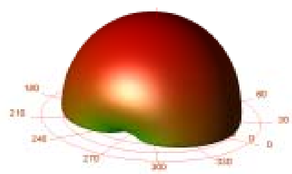


Aperture coupled microstrip transition



Multilayer open loop filter

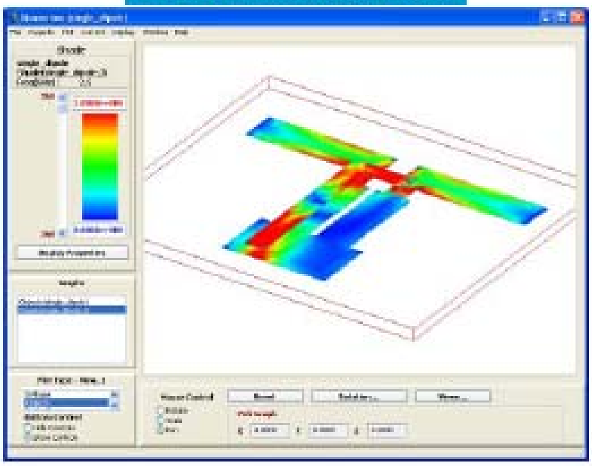




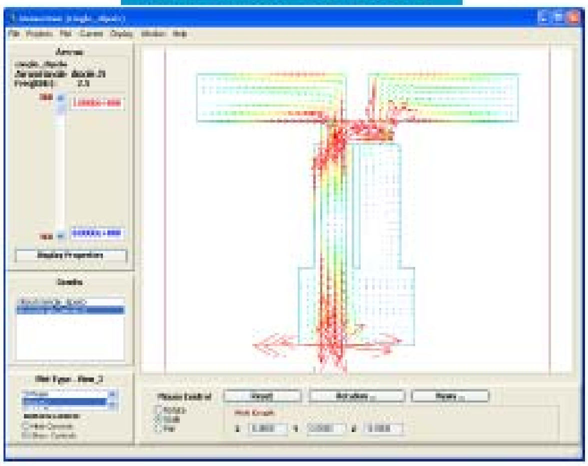
Patch antenna

ADS Momentum Current Visualization

Shaded current visualization



Arrow current visualization



When to use Momentum

- When no analytical model exists- co-simulate with ADS.
- Arbitrary geometry on a uniform planar substrate.
- To determine coupling effects-thin layers & close proximity of metal.
- To find narrow resonances not found using analytical models.
- To display radiation patterns plot far-fields (antennas, etc.)
- To display current patterns and density.
- For CPW: coplanar waveguide results with no slot mode.
- When 3D solvers take too long or are not required.
- To optimize (modify geometry) a passive layout to achieve desired results.

How Momentum Works?

Fullwave EM

Maxwell's Equations

⇩

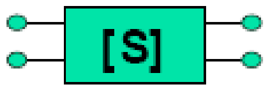
Matrix Equation

[Z][I]=[V]

⇳

Equivalent Circuit

$[Z]=[R]+j\omega[L]+1/j\omega[C]^{-1}$



- Fullwave electric & magnetic Green's functions.
- includes space and surface radiation.
- L(w)&C(w) are complex and frequency dependent
- Z(w) matrix reload CPU intensive.

$$Z_{i,j} = j\omega L_{i,j} + \frac{1}{j\omega C_{i,j}}$$

$$L_{i,j} = \iint_S dS\,\mathbf{B}_i(\mathbf{r}) \cdot \iint_{S'} dS'\,G^A(\mathbf{r},\mathbf{r}')\mathbf{B}_j(\mathbf{r}')$$

$$\frac{1}{C_{i,j}} = \iint_S dS\,\nabla\cdot\mathbf{B}_i(\mathbf{r}) \iint_{S'} dS'\,G^V(\mathbf{r},\mathbf{r}')\nabla\cdot\mathbf{B}_j(\mathbf{r}')$$

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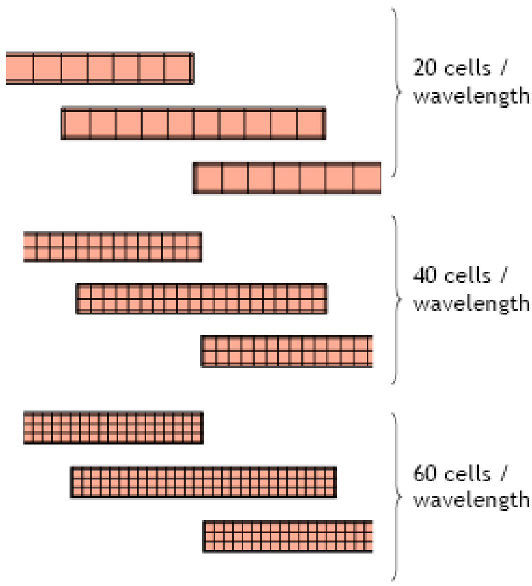
ADS Momentum

- !Mesh quality is crucial!
  - Subdivide metallic strips in basic 2D cells
  - Cell size based on a cells/wavelength-criterion
  - Different mesh types: Transmission line and edge mesh
  - Trade-off: Cell size / numerical effort
- Momentum vs. Circuit simulator
  - More modeling flexibility
  - More accurate results
    - Including radiation, coupling
  - **But:** increased <u>numerical effort</u> compared to circuit-theory based solver
  - limited to <u>infinite substrates</u>

Different edge meshes
(central section of edge coupled line filter)



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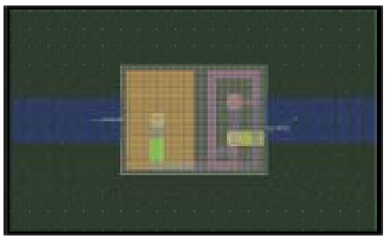
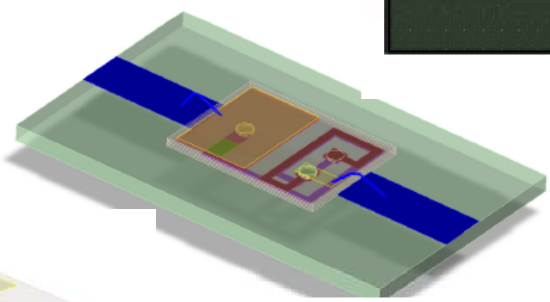
SPICE & Momentum

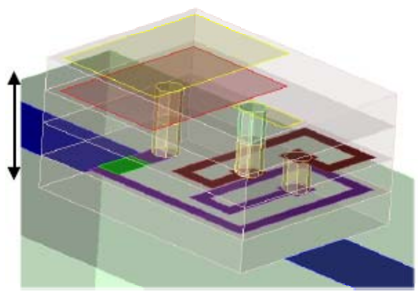
	SPICE	Momentum RF	Momentum MW
	Spice model	S parameters RF mode	S parameters MW mode
• Quasi-static inductance	V	V	V
• Quasi-static capacitance	V	V	V
• DC conductor loss (s)	V	V	V
• DC substrate loss (s)	V	V	V
• dielectric loss (tgd)		V	V
• skin effect loss		V	V
• substrate wave radiation			V
• space wave radiation			V

ADS EMDS

- Finite substrates
  - Defined as 3D vias in dielectric layers
  - Fringing effects
  - Broadside radiation of planar antennas
- Import of 3D components from EMPro
- Visualisation of EM fields in the computational volume
- But: increased computational burden

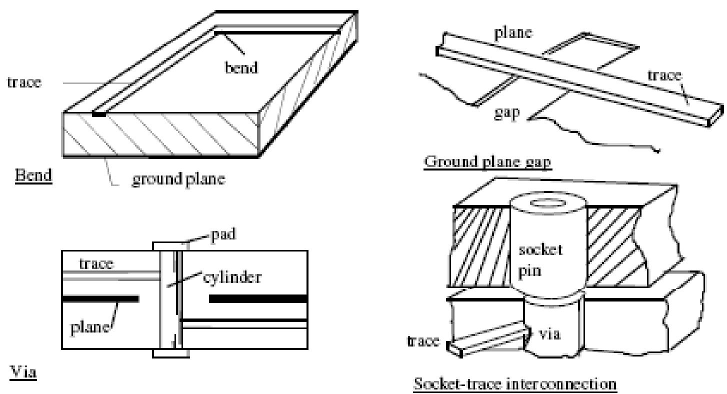




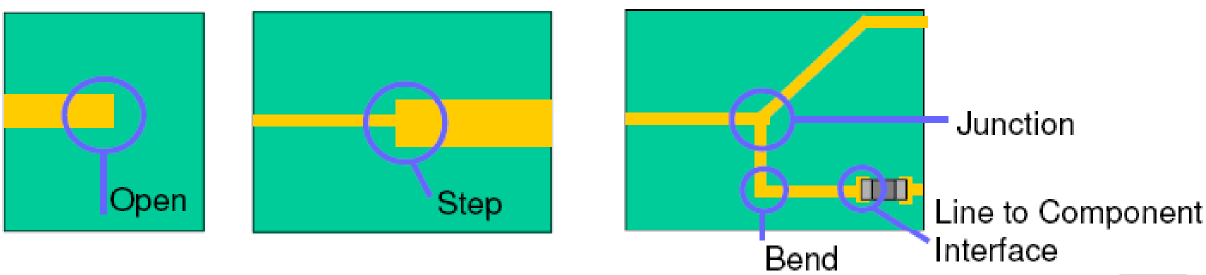


Transmission Line Discontinuities &Corrections

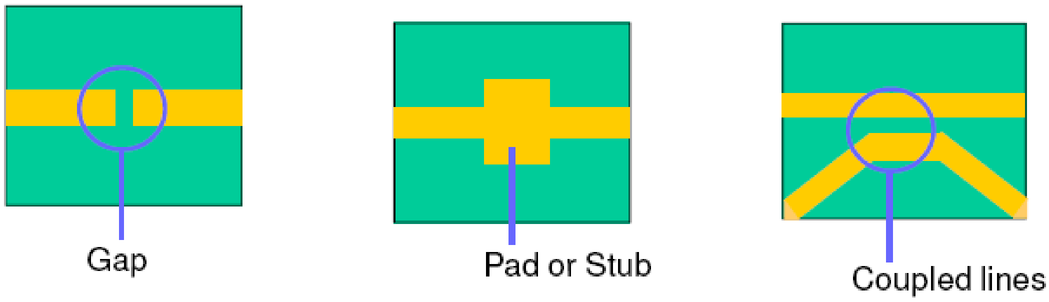
Transmission Line Discontinuities

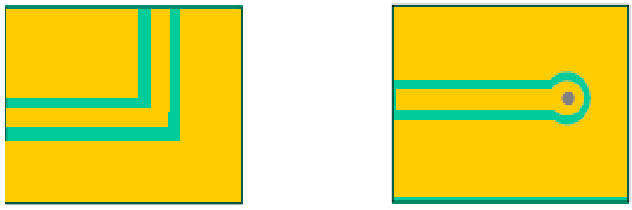


Here we illustrate the discontinuities using microstripline. Similar structures apply to other transmission line configuration as well.



Transmission Line Discontinuities

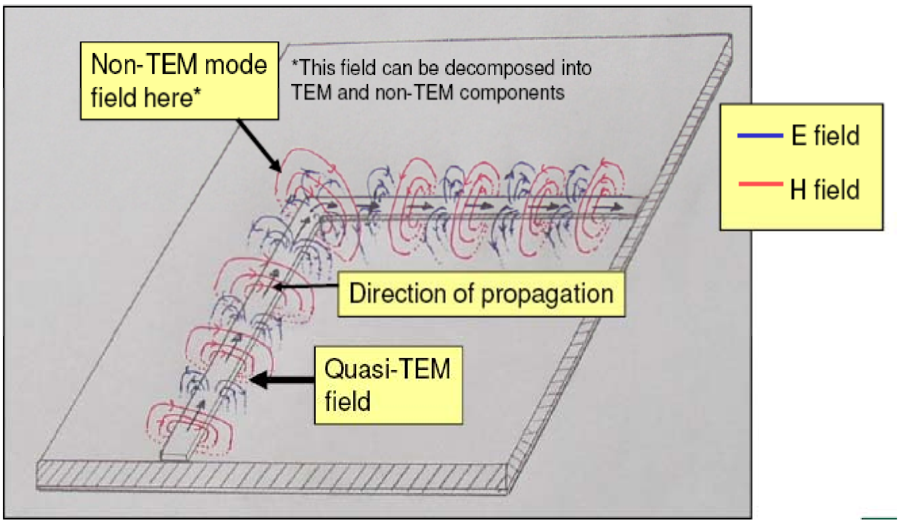




Examples of bend and via on co-planar Tline.

Transmission Line Discontinuities

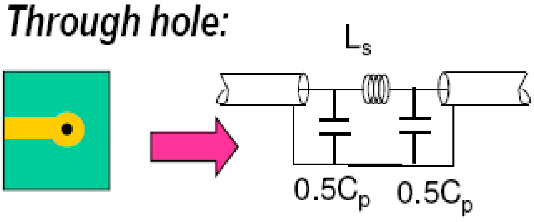
- Introduction of discontinuities will distort the uniform EM!!
- Most of these induced higher order modes are non-propagating



- The effect of discontinuity is usually reactive (the energy stored in the local fields is returned back to the system
- The effect of reactive system to the voltage and current can be modeled using LC circuits (which are reactive elements).

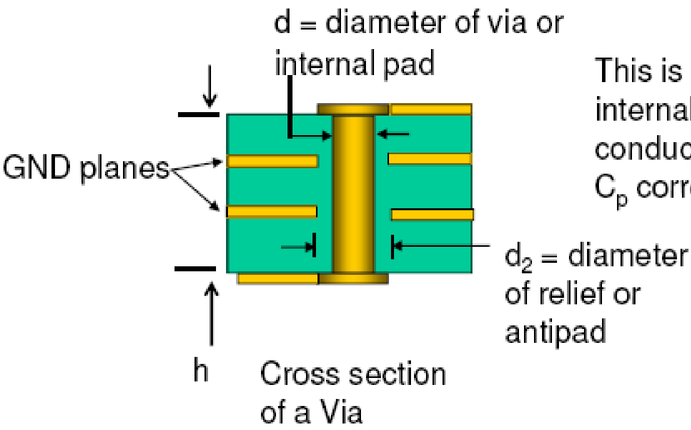
Discontinuities: Through Hole Via

Through hole:



$$L_s \cong 0.2h\left(\ln\left(\frac{4h}{d}\right)+1\right) \quad \textbf{(2.1a)}$$

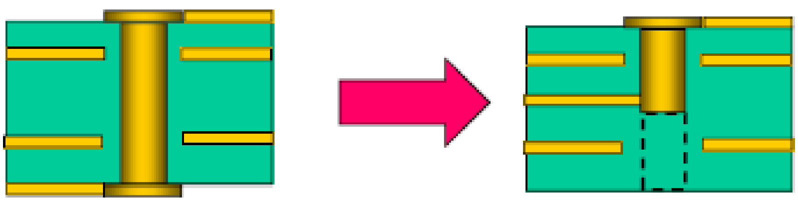
$$C_p \cong 0.056\frac{\varepsilon_r hd}{d_2-d}N \quad \textbf{(2.1b)}$$



This is the capacitance between the via and internal plane. If there are multiple internal conducting planes, then there should be one C_p corresponding to each internal plane.

L_s in nH
C_p in pF
h in mm
d and d_2 in mm
ε_r = dielectric constant of PCB
N = number of GND planes

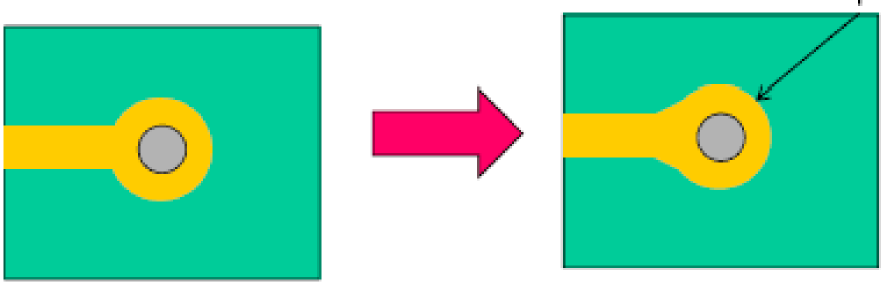
Discontinuities: Through Hole Via



Side view of a Via

Reduce the length of the via by using inner trace layer and remove the stub (e.g. use buried or blind via).

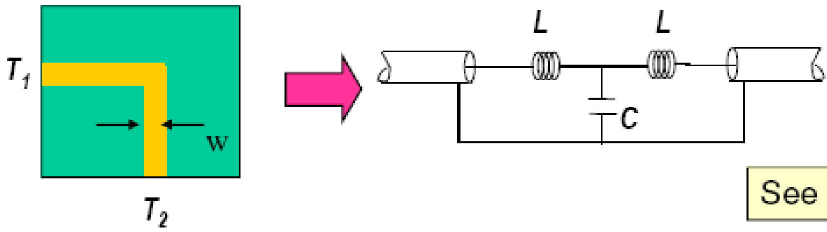
A tear-drop shape pad (typically only recommended for frequency above 1 GHz).



Top view of a Via

Discontinuities: Bends

90° Bend:



See Edwards [4], Chapter 5

Approximate quasi-static expressions for L_1, L_2 and C:

$$\frac{C}{w}=\frac{(14\varepsilon_r+12.5)\frac{w}{d}-(1.83\varepsilon_r-2.25)}{\sqrt{w/d}}\ \text{pF/m} \quad \text{for} \quad \frac{w}{d}<1$$

$$\frac{C}{w}=(9.5\varepsilon_r+1.25)\frac{w}{d}+5.2\varepsilon_r+7.0\ \text{pF/m} \quad \text{for} \quad \frac{w}{d}>1 \qquad \textit{(2.2a)}$$

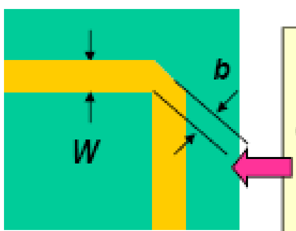
$$\frac{L}{d}=100\left[4\sqrt{\frac{w}{d}}-4.21\right]\ \text{nH/m} \qquad \textit{(2.2b)}$$

ε_r = dielectric constant of substrate, assume non-magnetic.
d = thickness of dielectric in meter.

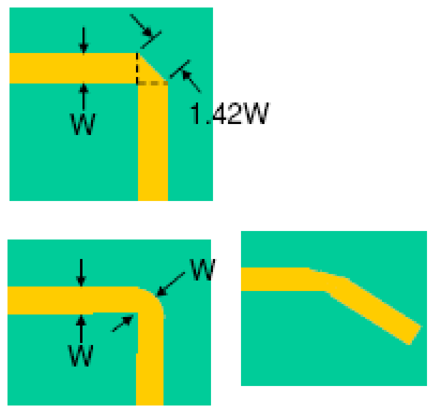
Reducing the Effects of Bends

- Generally not necessary under 300MHz frequencies

Chamfering of bends



For 90° bend:
It is seen that the optimum chamfering is b=0.57W (see Chapter 5, Edwards [4])
For further examples see Chapter 2, Bahl [7].



Discontinuities: Bends

- For a 90º microstrip line bend, with w = 0.92mm, d = 0.51mm, ε_r = 4.6 (non-magnetic substrate). Find the value of L and C for the bend.

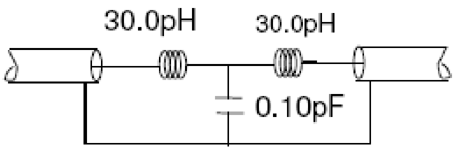
$$\frac{C}{w} = (9.5\times 4.6 + 1.25)1.8 + 5.2\times 4.6 + 7.0$$

$$= 111.83\ \text{pF/m} \qquad \frac{w}{d} = 1.834$$

$$\Rightarrow \text{C} = 111.83\times 0.00092 = 0.10\text{pF}$$

$$\frac{L}{d} = 100\left[4\sqrt{1.8} - 4.21\right] = 115.66\ \text{nH/m}$$

$$\Rightarrow \text{L} \cong 60\text{pH}$$



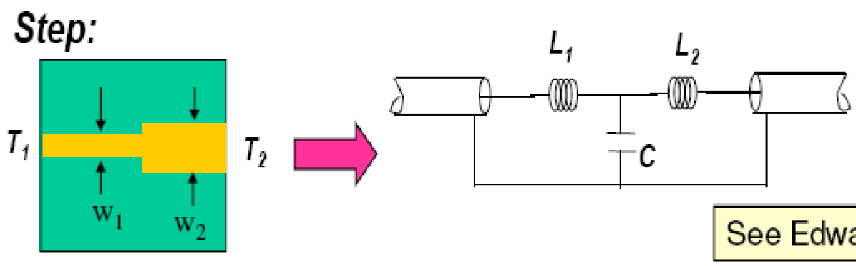
At 1GHz:

Reactance of C $\quad X_c = \frac{1}{2\pi f C} \cong 1592$

Reactance of L $\quad X_L = 2\pi f L \cong 0.38$

Typically the effect of bend is not important for frequency below 1 GHz. This is also true for discontinuities like step and T-junction.

Discontinuities: Step in Width



See Edwards [4] Chapter 5

Approximate quasi-static expressions for L_1, L_2 and C:

$$\frac{C}{\sqrt{w_1 w_2}} = (10.1\log\varepsilon_r + 2.33)\frac{w_1}{w_2} - 12.6\log\varepsilon_r - 3.17\ \text{pF/m} \quad \text{for} \quad \varepsilon_r \le 10;\ 1.5 \le \frac{w_2}{w_1} \le 10$$

$$\frac{C}{\sqrt{w_1 w_2}} = 130\log\left(\frac{w_2}{w_1}\right) - 44\ \text{pF/m} \quad \text{for} \quad \varepsilon_r = 9.6;\ 3.5 \le \frac{w_2}{w_1} \le 10$$

(2.3a)

$$\frac{L}{d} = 40.5\left(\frac{w_1}{w_2} - 1.0\right) - 75\frac{w_1}{w_2} + 0.2\left(\frac{w_1}{w_2} - 1.0\right)^2\ \text{nH/m}$$

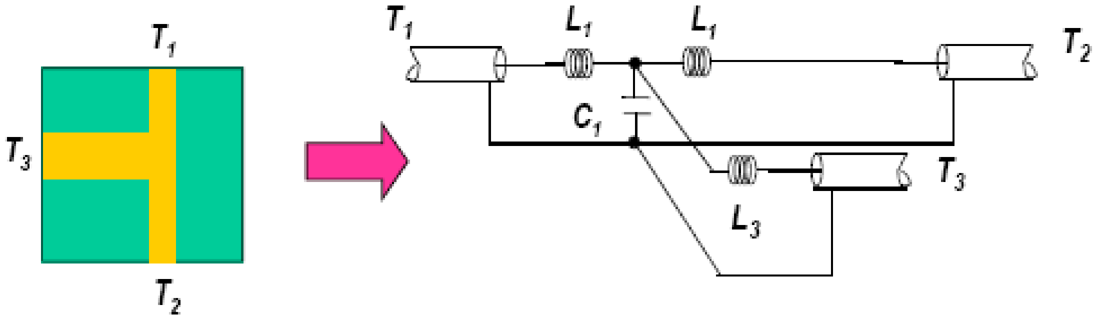
(2.3b)

$$L_1 = \frac{L_{m1}}{L_{m1} + L_{m2}}L \qquad L_2 = \frac{L_{m2}}{L_{m1} + L_{m2}}L$$

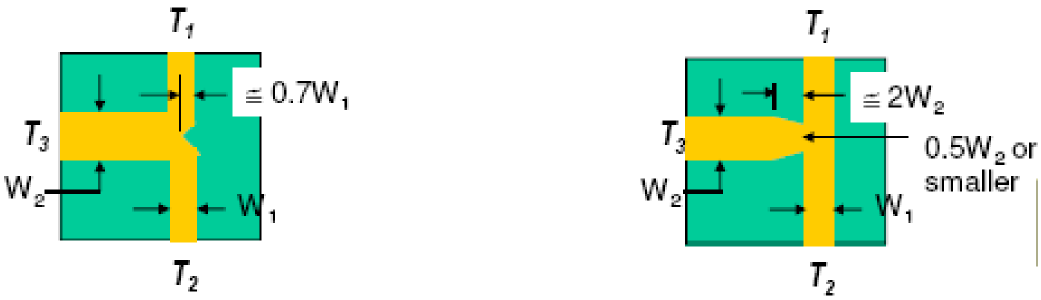
L_{m1} and L_{m2} are the per unit length inductance of T_1 and T_2 respectively.

Discontinuities: T-Junction

T-Junction:

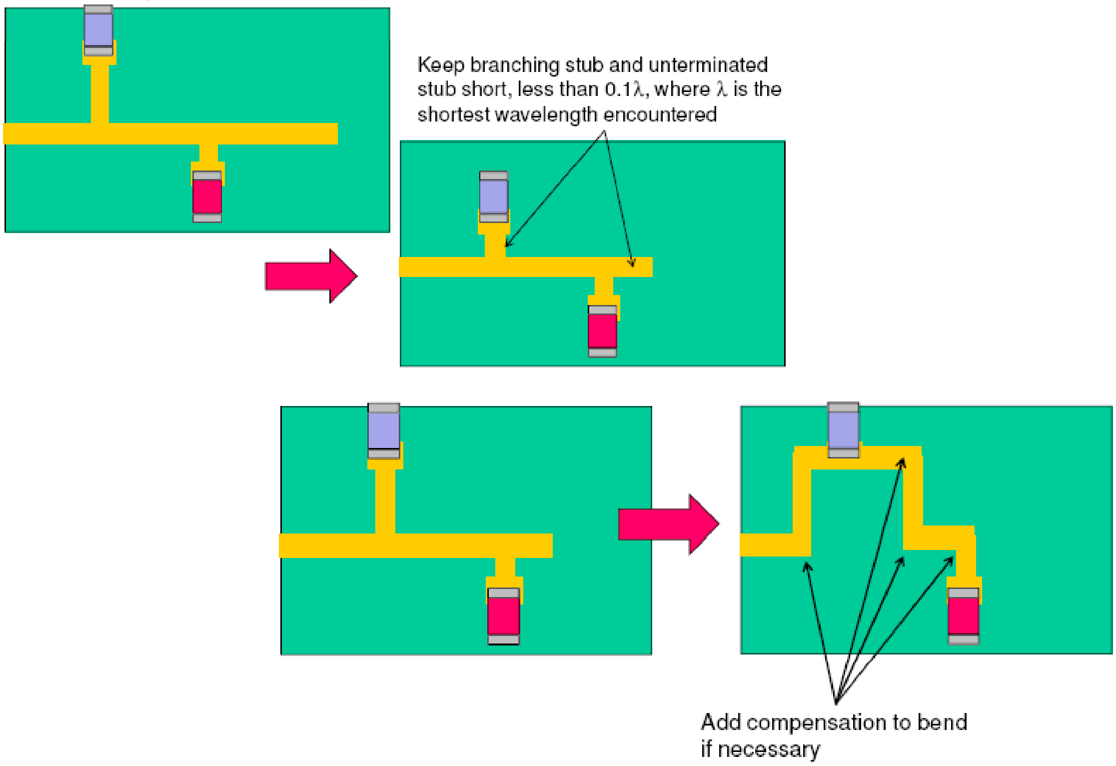


Mitering of junction



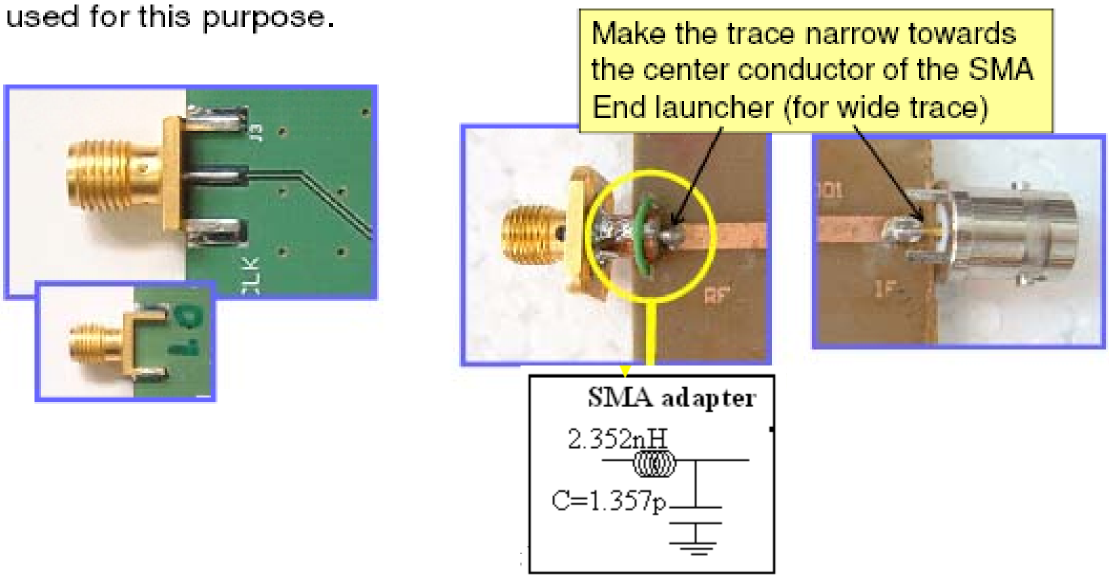
Discontinuities: Stubs

- Stubs are short trace segment that extend from the main trace. Most often they connect to a solder pad and a component.

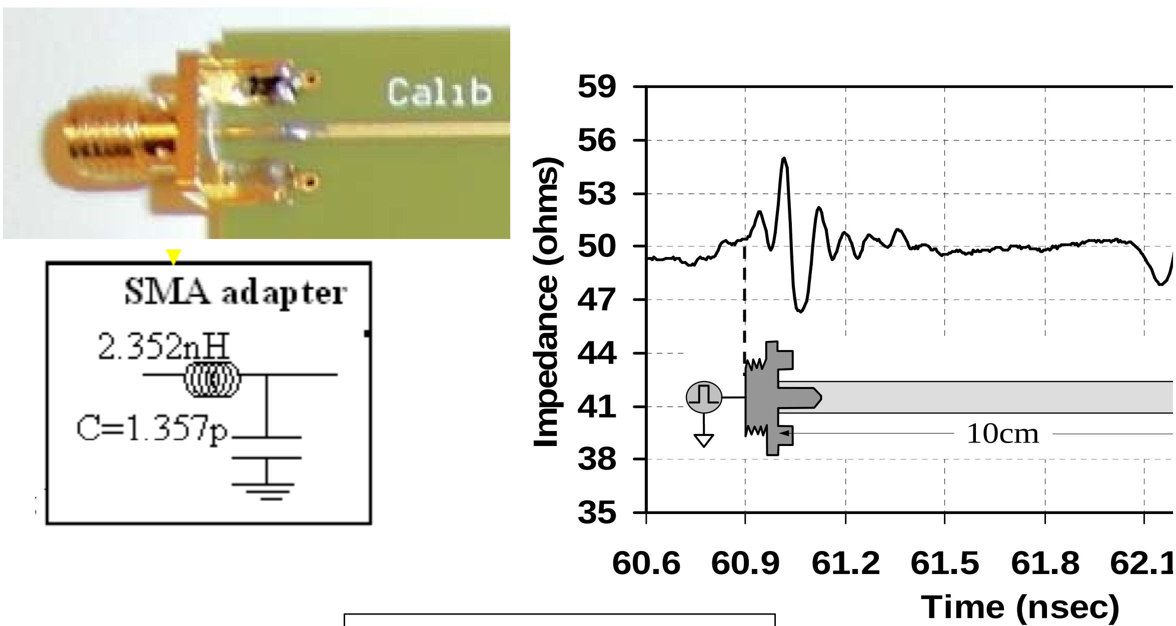


Discontinuities: SMA Launch

- Since most microstrip/stripline line invariably leads to external connection from the PCB, an interface is needed.
- Usually the microstrip/strip line is connected to a co-axial cable.
- The SMA to PCB adapter, also called the SMA End-launcher is usually used for this purpose.

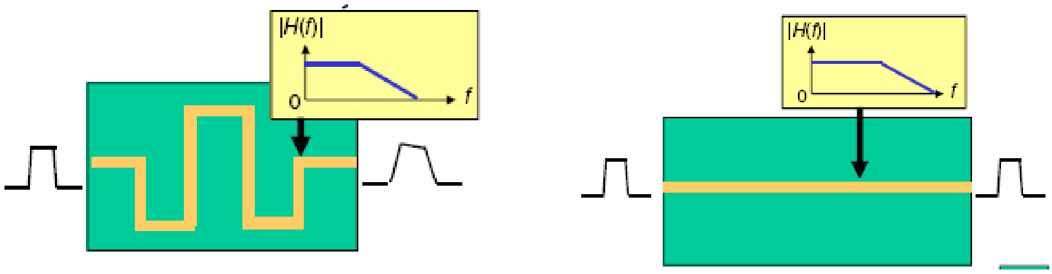


Discontinuities: SMA Launch



Measured and Design By Rashad

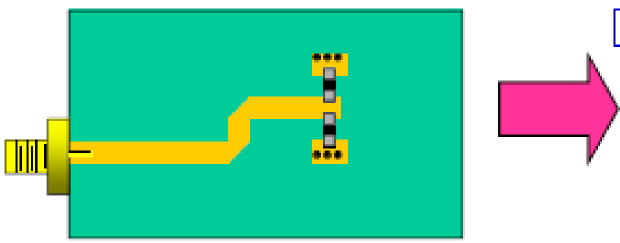
BW Reduction due to Bends

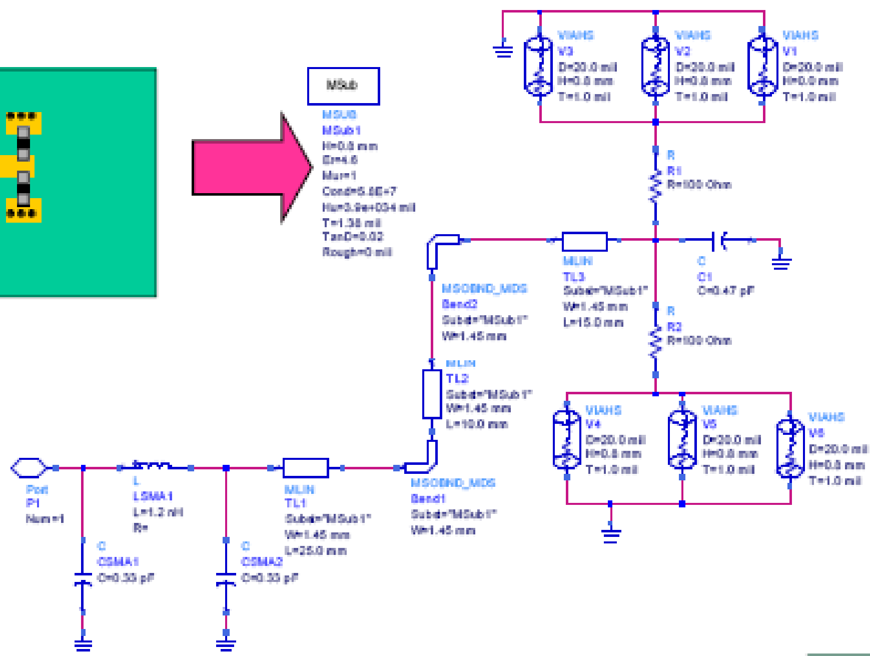


•Networks can be interpreted as Low-Pass Filters.
•The inductor
•attenuates the electrical signal at high frequency while the capacitor shunts electrical energy at high frequency.
•Thus the effect of having too many discontinuities in a high-frequency circuit reduces the overall bandwidth of the interconnection.
•Another consequence of discontinuities is attenuation due to radiation from the discontinuity.

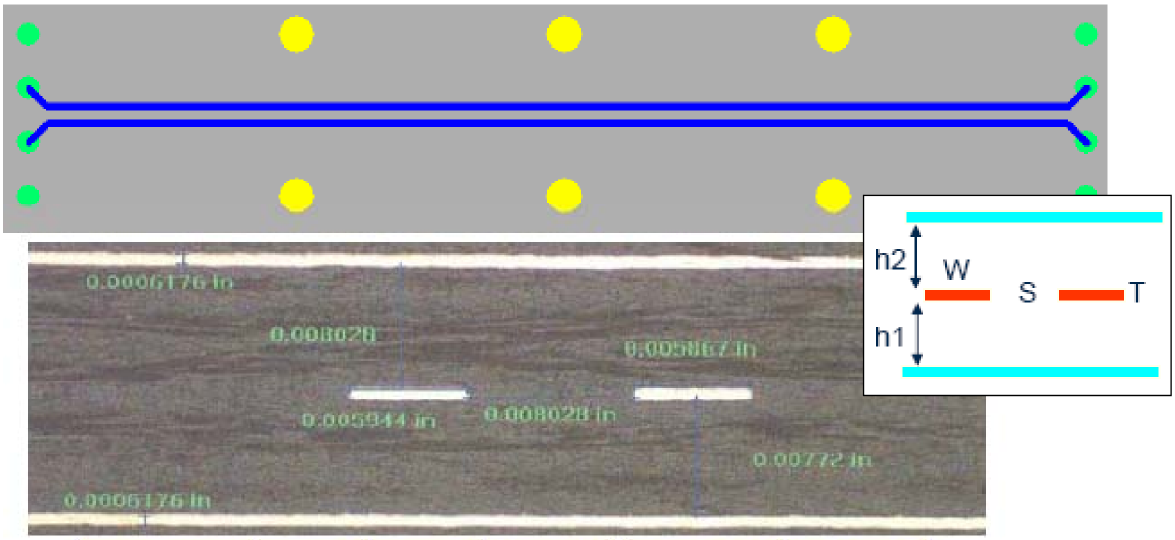
Discontinuities Modeling and Simulation

- A $Z_c = 50\Omega$ microstrip Tline is used to drive a resistive termination as shown.



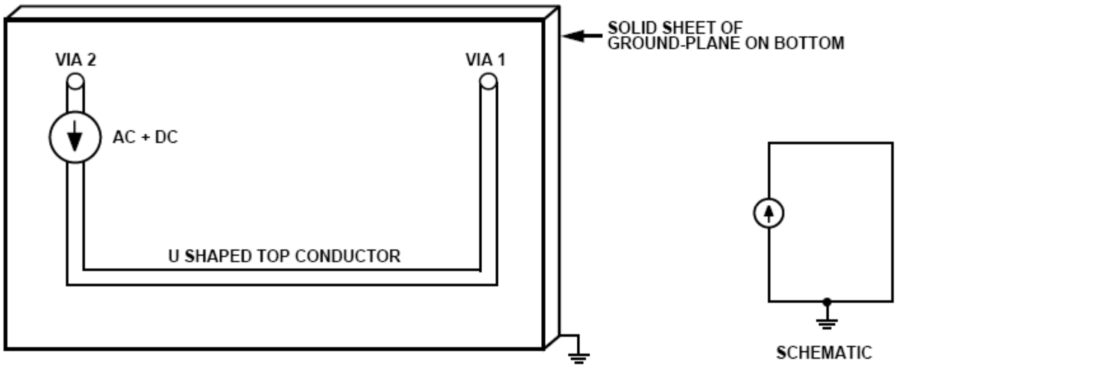


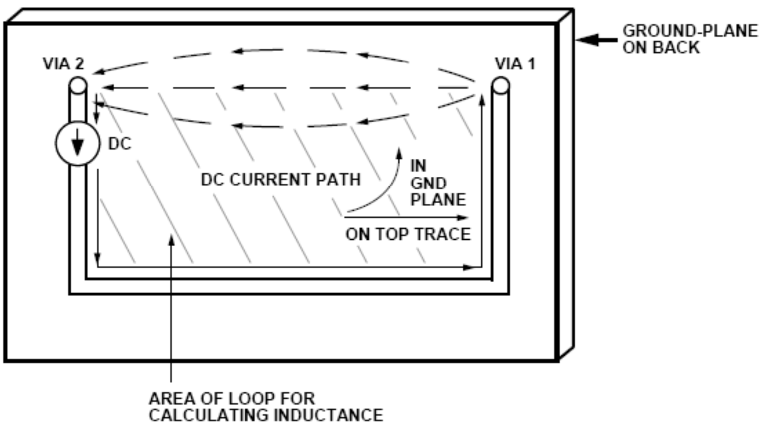
Manufacturer Tolerances



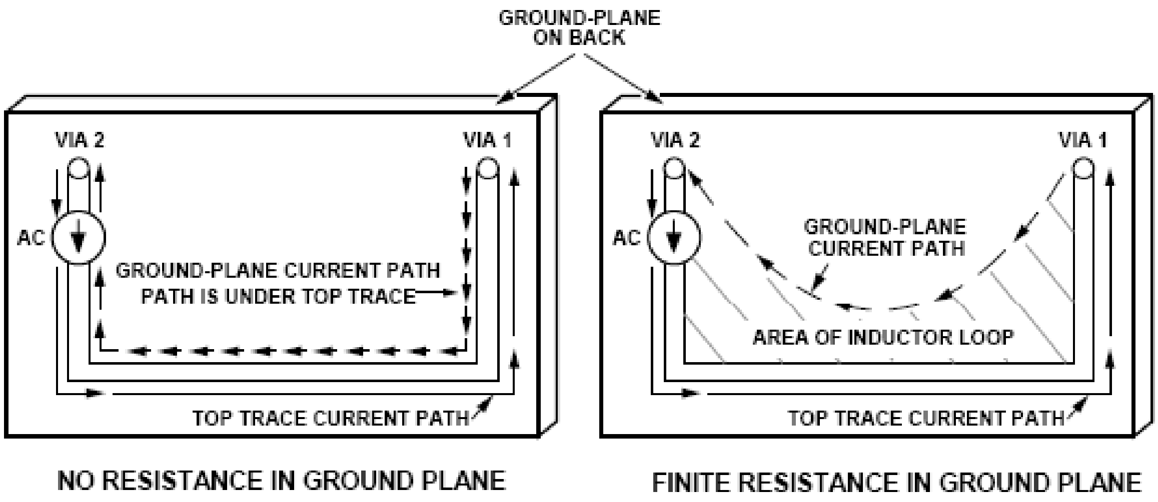
Parameter	Width W	Spacing S	Thickness T	Bot. Height h1	Top Height h2
Designed	6.0 mil	8.0 mil	0.7 mil	8.0 mil	9.0 mil
Modified by fab	6.7 mil	7.3 mil	0.5 Oz	8.0 mil	9.2 mil
Measured	5.9 mil	8.0 mil	0.62 mil	7.7 mil	8.0 mil

AC Current Return Path

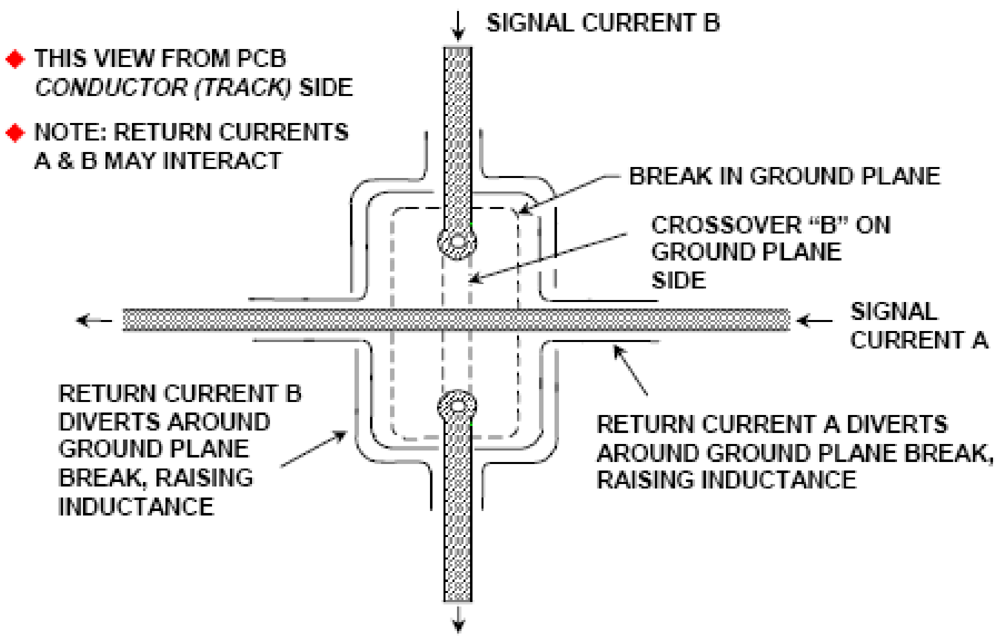




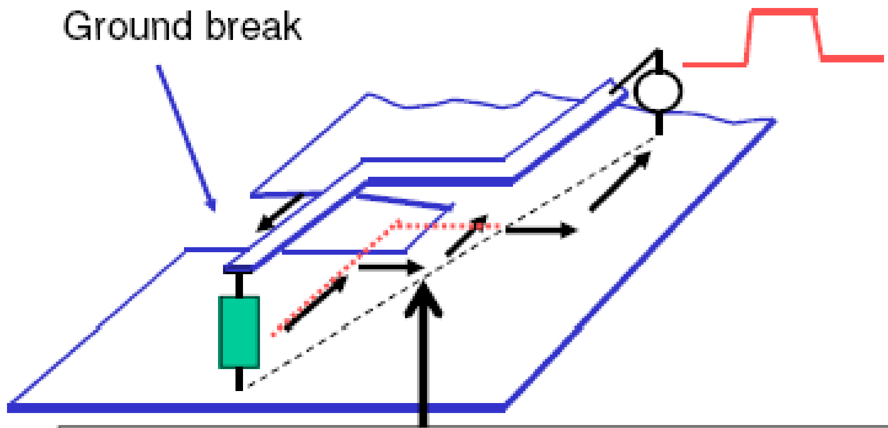
AC Current Return Path



GND Plane Break



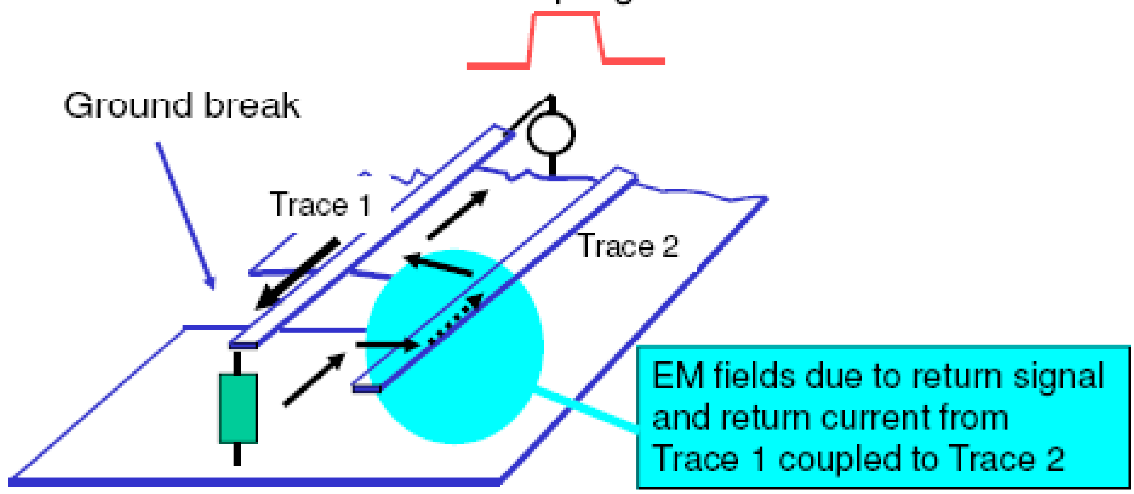
Partial GND Break



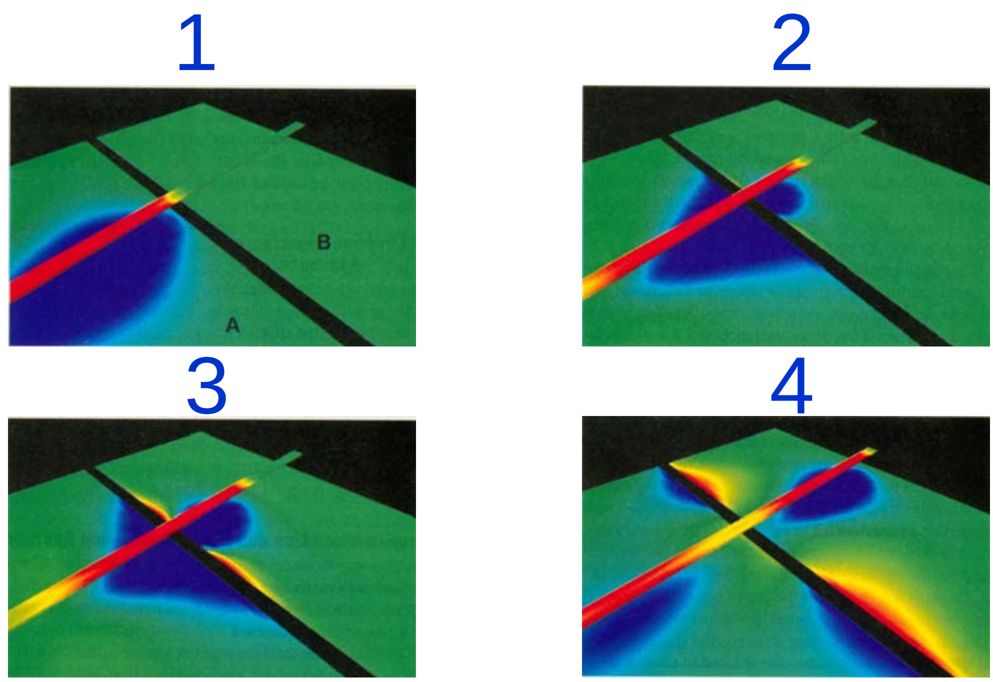
Detour for return current, more magnetic flux linkage. This will manifest as extra inductance in the interconnect circuit.
Furthermore the E and H fields will 'extend' further from the circuit, making it easier for this circuit to interfere and be interfered.

GND Break increases Coupling

Ground break also increases coupling between traces.



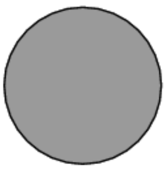
Discontinuities: Through Hole Via



IEEE Circuits & Devices Nov. 1997

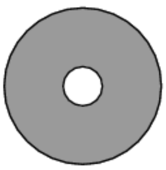
Skin Depth – AC Resistance

Skin effect confines 63% (e^{-1}) of the current to 1 skin depth – the current density will decease exponentially into the thickness of the conductor



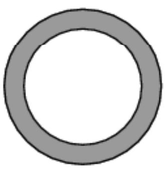
Cross-sectional area of a round conductor available for conducting DC current

"DC resistance"



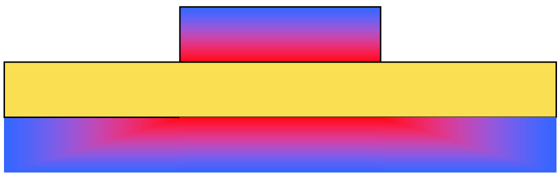
Cross-sectional area of the same conductor available for conducting low-frequency AC

"AC resistance"



Cross-sectional area of the same conductor available for conducting high-frequency AC

"AC resistance"



$$\delta = \sqrt{\frac{2}{\mu\,\omega\,\sigma}} = \sqrt{\frac{\rho}{f\,\pi\,\mu}}$$

μ = permeability ($4\pi \times 10^{-7}$ H/m), note: H = henries = Ω·s

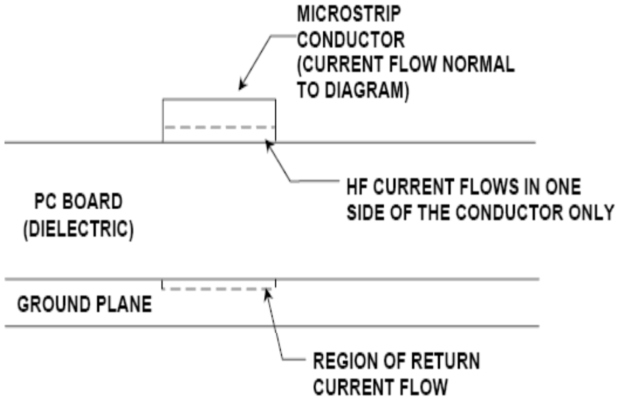
δ = skin depth (m)

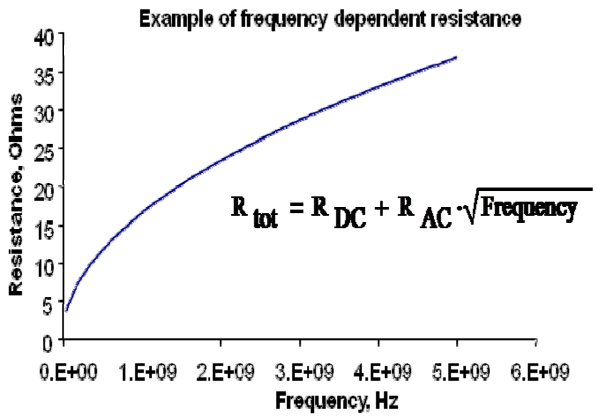
ρ = resistivity (Ω·m)

ω = radian frequency = $2\pi f$ (Hz)

σ = conductivity (mho/m), note: mho [℧] = siemen [S]

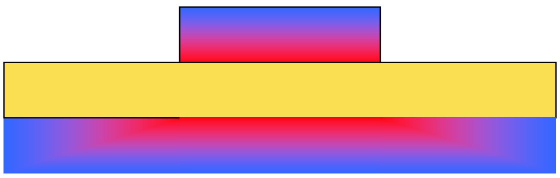
Skin Depth- AC Resistance

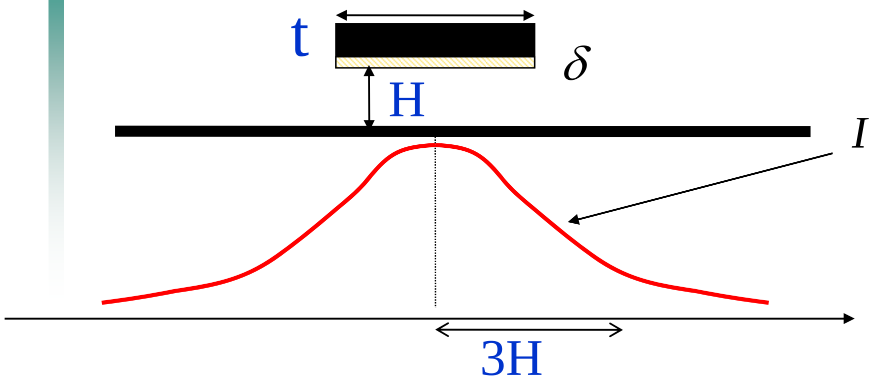




- Skin Depth $\approx$ $6.61 / \sqrt{f}$ cm, f in Hz
- Skin Resistance $\approx$ 2.6×10^{-7} $\sqrt{f}$ ohms per square, f in Hz
- Since skin currents flow in both sides of a PC track, the value of skin resistance in PCBs must take account of this

GND Current AC Resistance





$$I(D) = \frac{I_O}{\pi H} \cdot \frac{1}{1 + (D/H)^2}$$

(Current Density in plane)

GND (Current) AC Resistance

- The current density formulae can be integrated to get the total current contained within chosen bounds

$$\int_{-3H}^{3H} \frac{I_o}{\pi H} \cdot \frac{1}{1+(D/H)^2} dD = \frac{2I_o}{\pi} \cdot \tan^{-1}(3) = 0.795 I_o$$

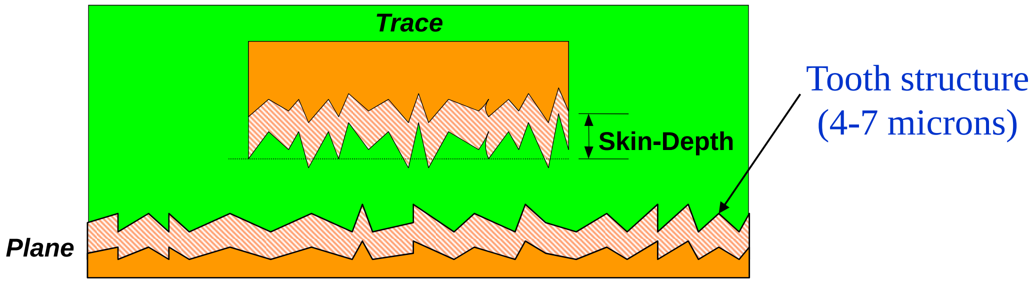
- This shows that 79.5% of the current is contained in a distance +/- 3H (W of 6H) from the conductor center
- Assuming a penetration of 1 skin depth, the ground return resistance can be approximated as follows

$$R_{ac\_ground} \approx \frac{\rho}{A_{ground}} = \frac{\rho}{6H\delta} = \frac{\sqrt{\rho \pi \mu F}}{6H} \quad \Omega / length$$

$$\boxed{R_{AC\_total} = R_{signal} + R_{ground}}$$

Surface Roughness

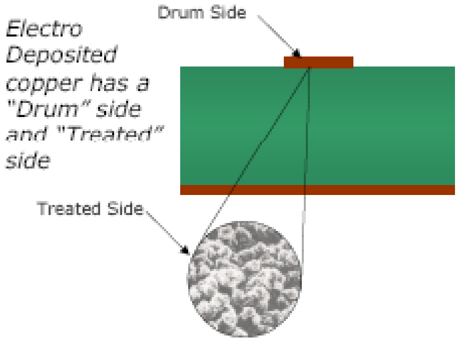
- The formulae presented assumes a perfectly smooth surface
- The copper must be rough so it will adhere to the laminate
- Surface roughness can increase the calculated resistance 10-50% as well as frequency dependence proportions
- Increase the effective path length and decreases the area



Copper Options

- Common Thickness Options

Weight	Thickness
0.25 oz	0.4 mil (9 µm)
0.5 oz	0.7 mil (18 µm)
1.0 oz	1.4 mil (35 µm)
2.0 oz	2.8 mil (70 µm)

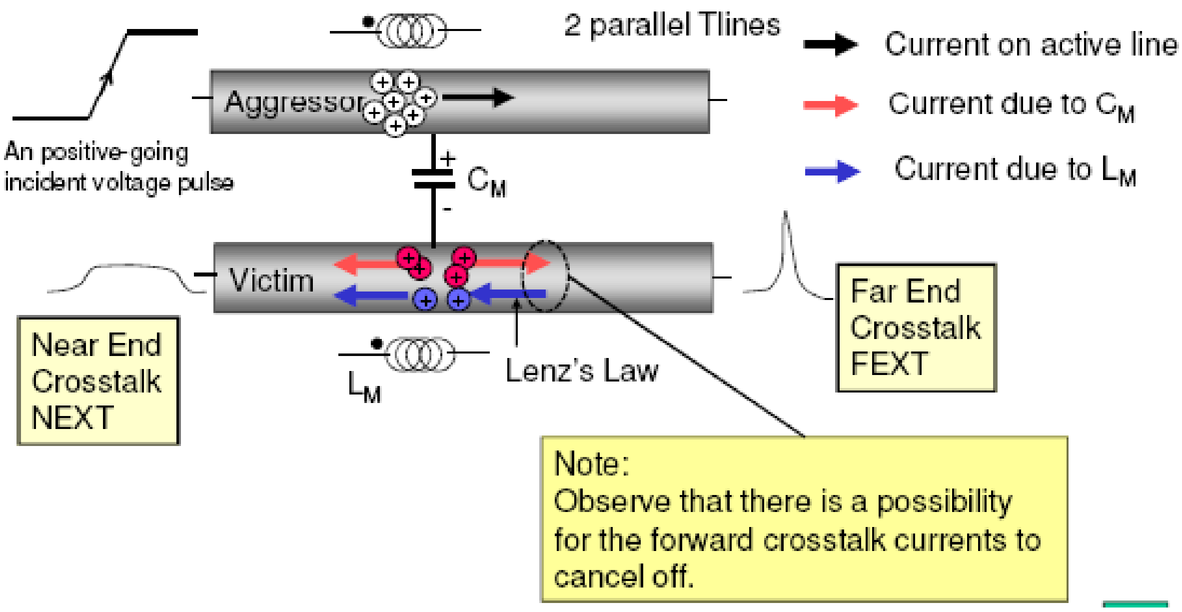


- Plating Options

Option	Surface Roughness	Notes
Rolled	55 µ in	Lower loss at high frequency (>1 GHz) More precise geometries for critical applications (couplers, distributed filters)
Electro Deposited	75 µ in (0.5 oz) 94 µ in (1 oz) 120 µ in (2 oz)	For Treated (Dendritic) Side Untreated (Drum) Side is 55 µ in Less prone to pealing

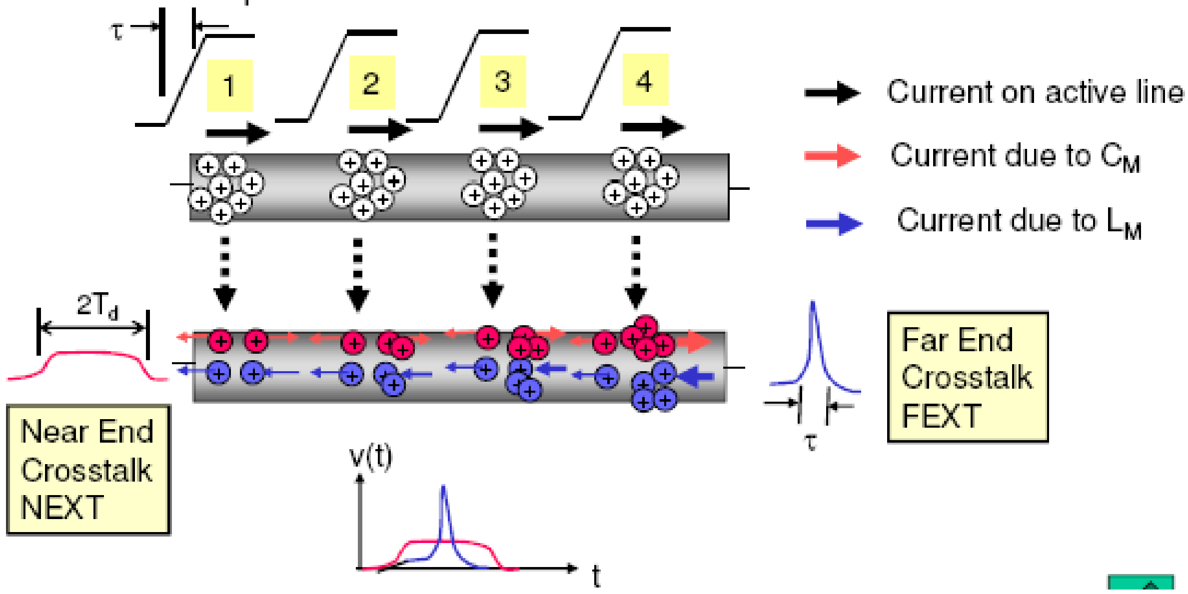
Mutual Coupling and Crosstalk

- The current components due to distributed mutual capacitance and inductance.



Mutual Coupling and Crosstalk

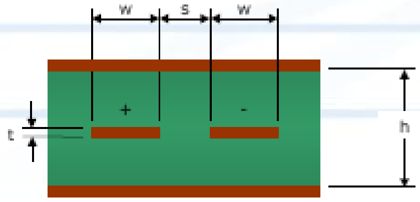
- As the positive-going pulse travels from left to right, it will induce current on the adjacent line. The induced current components travel out from the point of induction to both directions.



Differential Pairs

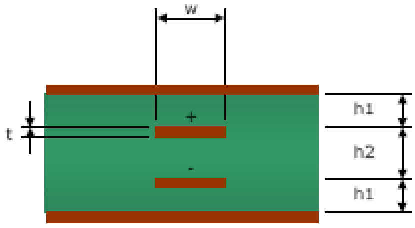
- Lower Cross-talk, Lower Radiation
- Common mode noise rejection
- Reduces ground reference problems
- High dynamic range analog applications
  - Log Amplifiers
- High Resolution ADC/DAC
  - Low noise (small signal) analog applications
- Transducers
  - Critical High Speed Digital Applications
- Low amplitude clocks
- Low jitter requirements
- Almost ALL RF applications

Differential Pairs



Edge Coupled

- Geometry and spacing defined by artwork
- High differential impedance easily achievable
- Impedance reduced as "s" is reduced
- As "s" is increased, impedance approaches 2x single ended impedance
- Difficult to rout through fine pitch holes



Broadside Coupled

- Geometry is effected by layer registration
- Low differential impedance easily achievable
- Easy to route, easy to maintain matched lengths

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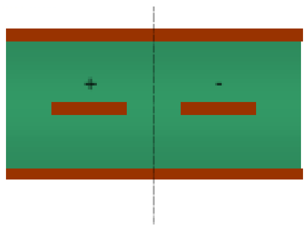
Differential Impedance Definitions



- Single-Ended Impedance (Z_O)
  Impedance on a single line with respect to ground when not coupled to another line
  $(Z_0 = \sqrt{Z_{Odd} Z_{Even}})$
- Differential Impedance (Z_{DIF})
  The impedance on one line with respect to the coupled line, when the lines are driven by equal and opposite signals
- Odd Mode Impedance (Z_{Odd})
  Impedance on a single line with respect to ground when the other coupled line is driven by equal and opposite signals ($Z_{DIF} = 2Z_{Odd}$)
- Common Mode Impedance (Z_{CM})
  Impedance of the two lines combined with respect to ground
- Even Mode Impedance (Z_{Even})
  The impedance on one line with respect to ground when the coupled line is driven by an equal and in-phase signal ($Z_{Even} = 2Z_{CM}$)

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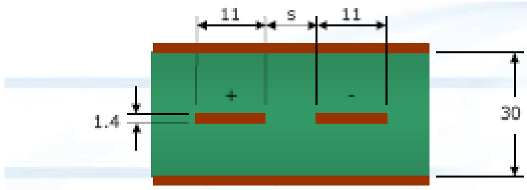
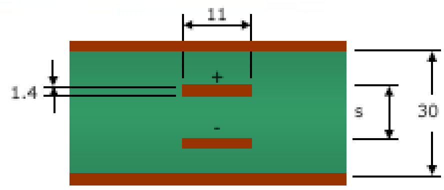
Differential Impedance Examples



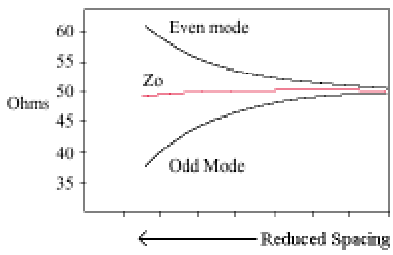
Edge Coupled

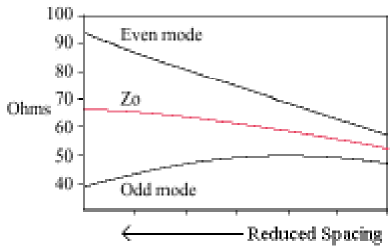


Broadside Coupled

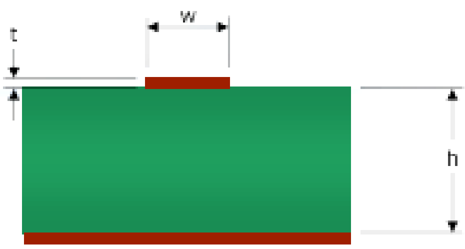
S	Z_{Even}	Z_{Odd}	Z_o	Z_{Diff}
100	49.4	49.4	49.4	98.8
25	50.9	47.9	49.4	95.8
10	56.1	41.9	48.6	83.8

S	Z_{Even}	Z_{Odd}	Z_o	Z_{Diff}
20	41.3	34.7	37.8	69.4
10	68.6	34.7	48.8	69.4

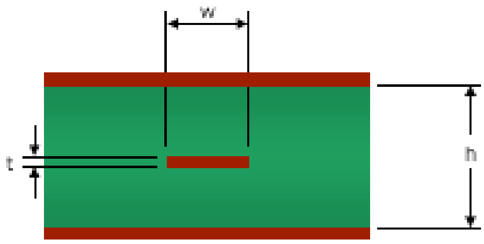




Impedance of Transmission Lines



Micro-Strip



Strip-Line

$$Z_o = \frac{87}{\sqrt{\varepsilon_r + 1.41}} \ln\left(\frac{5.98h}{0.8w + t}\right)$$

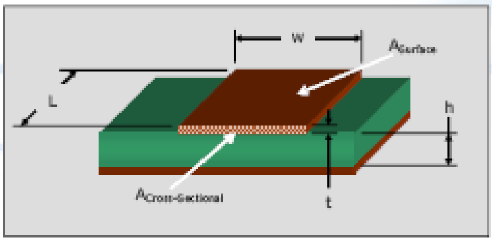
$$Z_o = \frac{60}{\sqrt{\varepsilon_r}} \ln\left(\frac{1.9b}{0.8w + t}\right)$$

when $0.1 < \frac{w}{h} < 2.0$ and $1 < \varepsilon_r < 15$

when $\frac{w}{h} < 0.35$ and $\frac{t}{h} < 0.25$

Simulations

Decoupling Capacitors



Resistance

$$R_{DC} = \rho \frac{l}{A_{Cross-Sectional}} = \rho \frac{l}{tw} \qquad \rho = 0.679\ \mu\Omega - inch$$

Allowing l=w permits calculation of Resistance per Square.

$$\text{Resistance/Square} = \rho \frac{1}{t} = (0.679\ \mu\Omega - in)\frac{1}{0.0014} = 485\ \mu\Omega\ /\ Square$$

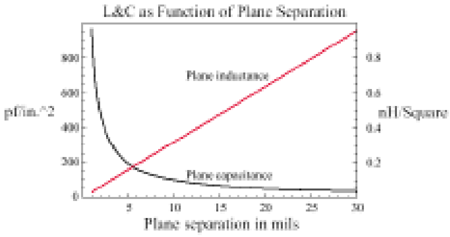
(For 1 ounce copper which is 0.0014" thick)

Inductance

$$L = \mu_0 h \frac{l}{w} \qquad \mu_0 = 4\pi x 10^{-7} \frac{H}{m} = 0.32 \frac{nH}{inch}$$

Capacitance

$$C = \varepsilon_0 \varepsilon_r \frac{LW}{h} = \varepsilon_0 \varepsilon_r \frac{A_{Surface}}{h} \qquad \varepsilon_r = 225\ x10^{-15} \frac{F}{in}$$



FR4 Dielectric Thickness (mils)	Inductance (pH/square)	Capacitance (pF/inch²)
8	260	127
4	130	253
2	65	506

ESR and ESL of Capacitors

- Physical capacitors have parasitic elements that limit their ability to stabilize supply lines
- Equivalent series inductance (ESL)
- Equivalent Series Resistance (ESR)

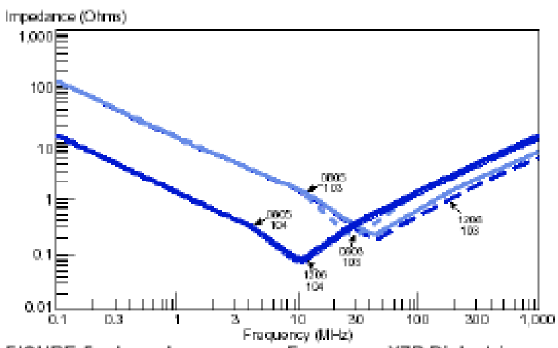


FIGURE 5 Impedance versus Frequency X7R Dielectric

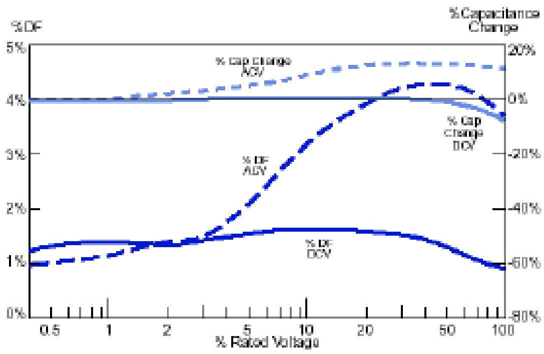


FIGURE 7. X7R Capacitance & DF versus Applied AC/DC Voltages

Plots courtesy of Kemet

Decoupling Caps Packages

Part Number	Package	Capacitance (uF)	ESL (nH)	ESR (Ω)	SRF (MHz)
C0603C103K5RAC, Kemet	**EIA 0603**	0.01	1.8	0.25	38
C0805C104K5RAC, Kemet	**EIA 0805**	0.10	1.9	0.10	12
T491B685K010AS, Kemet	**EIA 3528-21**	6.8	1.9	0.3	1.4
T494C476K010AS, Kemet	**EIA 6032-28**	47	2.2	0.2	0.5

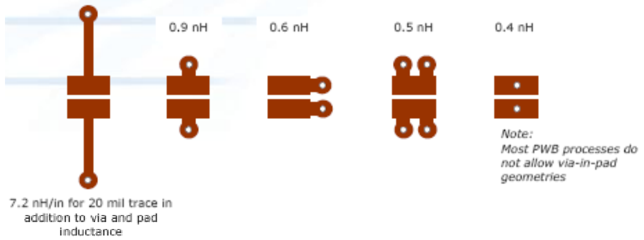
$$f_{SRF} = \frac{1}{2\pi\sqrt{LC}}$$

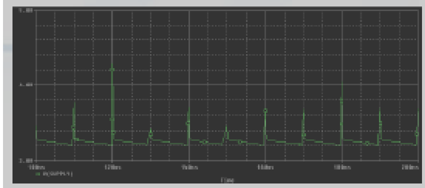
Self Resonant Frequency

$$Q = \frac{1}{DF} = \frac{X_C}{R} = \frac{1}{2\pi\ f\ C\ R}$$

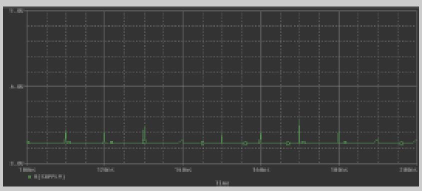
Quality Factor, Dissipation Factor

Decoupling Connection Geometry

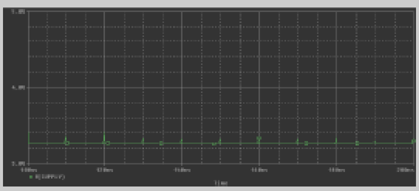




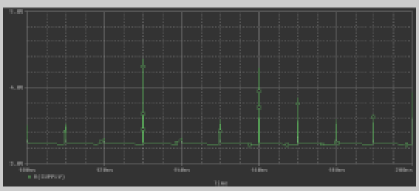
6.8 uF, Ripple: 1 Vpp



6.8 uF + 0.1 uF, Ripple: 0.3 Vpp



6.8 uF + 0.1 uF + 0.01 uF, Ripple: 0.1 Vpp



6.8 uF + 0.1 uF + 0.01 uF w/ long traces, Ripple: 1.1 Vpp

Summary

- Modeling &n EM Simulations
  - Schematic Simulation & Layout Simulation
  - Momentum Simulation
- Microstrip Discontinuities
- GND path Breaks
- AC resistance and Skin Effect
- Differential Signaling
- Odd and Even Mode Impedances
- Decoupling Capacitors