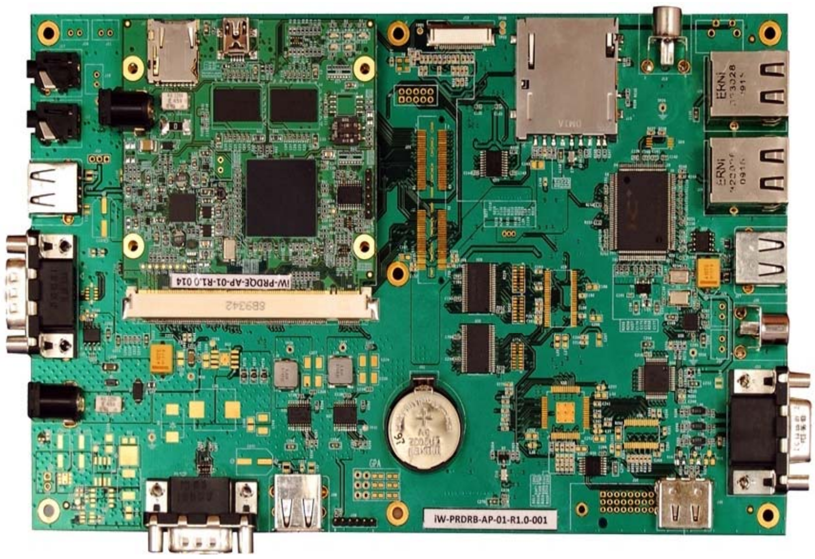
Module-2
RF PCB Design
Basics Concepts &Techniques

Rashad.M.Ramzan, Ph.D
FAST-NU, Islamabad

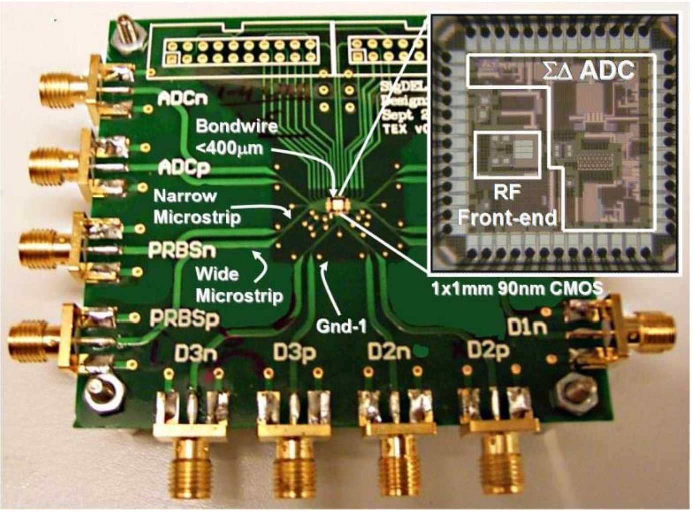
Objective of RF PCB Design Module

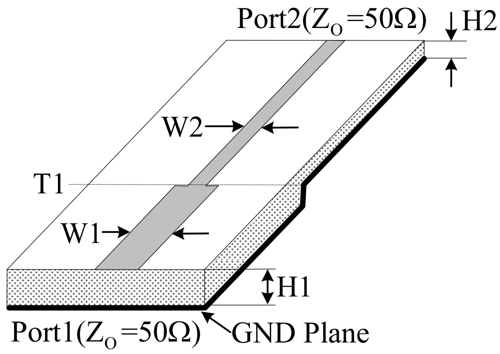


- Design of Digital PCB with RF commutation Chips modules on it!!
- Tools
  - Protel or Altium
  - Used together with ADS

- Approach is to use the RF tool like ADS to implement the layout in Altium.
- Advantage: Productivity and accuracy at same time

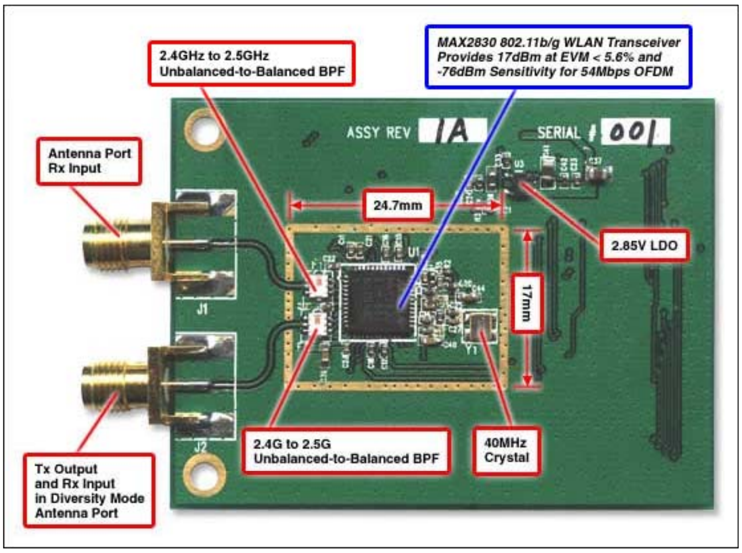
Objective of Module-2





- 90nm Wideband RF Frontend Test Bed (1-6GHz)
  - 90nm Chip and 4-layer FR4 PCB Designed By Rashad
- Protel was used to design the board and ADS to design the transmission lines

Objective of RF PCB Design Module



- Complete RF PCB on High Speed Substrates using ADS tool suite
- Need strong theory and background of Electromagnetic
  - Impedance matching, TDR, Smith chart, S-Parameters, Transmission Lines, Transceiver Architecture , LNA, Mixer Design, PLL, DLL, Pas etc

Outline of Today's Lecture

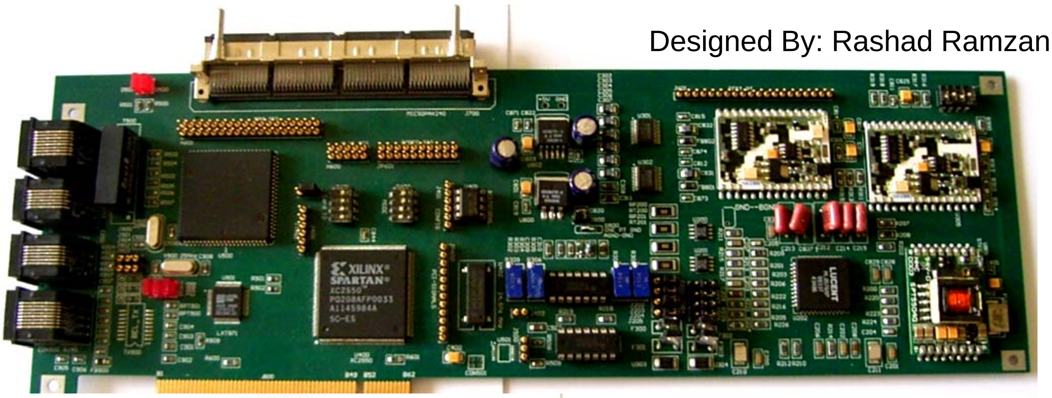
- Objective of this three day module
  - RF PCB and RF as a part of High Speed Digital PCB
  - Tools for both applications
- Difference in RF and Digital PCB
  - Frequency Range
    - Sine vs. Square (Trapezoidal)
    - Narrow Band vs Wide Band
  - Termination types & 50Ω matching
  - Impedance Matching Criteria
  - PCB Material, Layer Stack

Outline of Today's Lecture

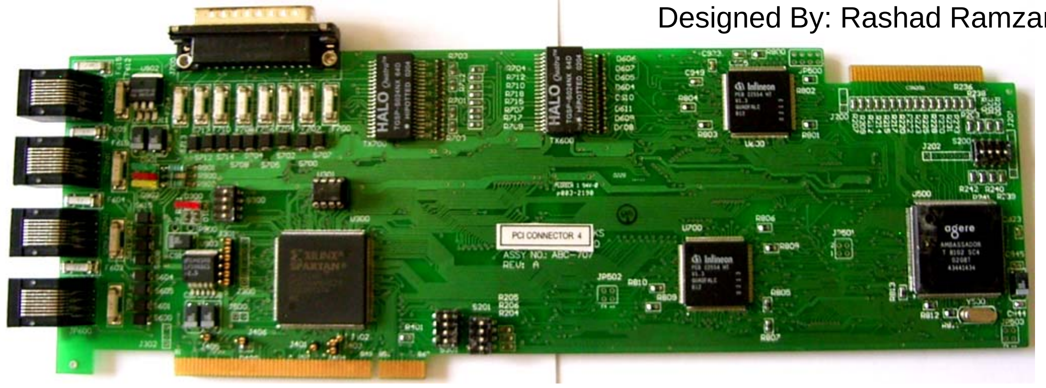
- Transmission Lines in RF PCBs and Digital PCBs
  - Basic Transmission Modes in PCB Traces
- Impedance Matching
  - Smith Chart …. A Necessary Tool
- S-Parameters

Module-1: Analog and Digital PCBs

Designed By: Rashad Ramzan

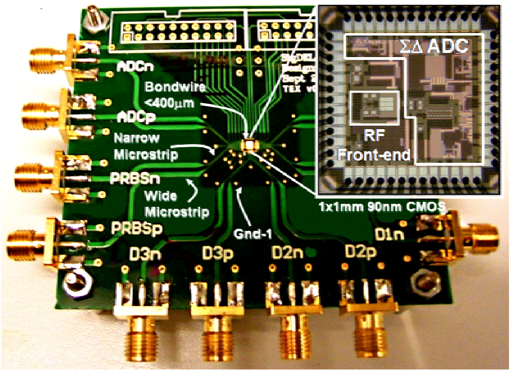


Designed By: Rashad Ramzan

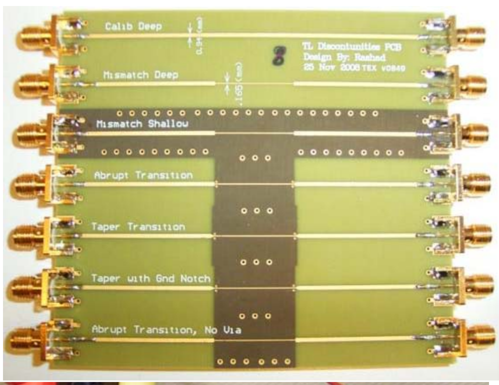


Module-2: RF PCBs

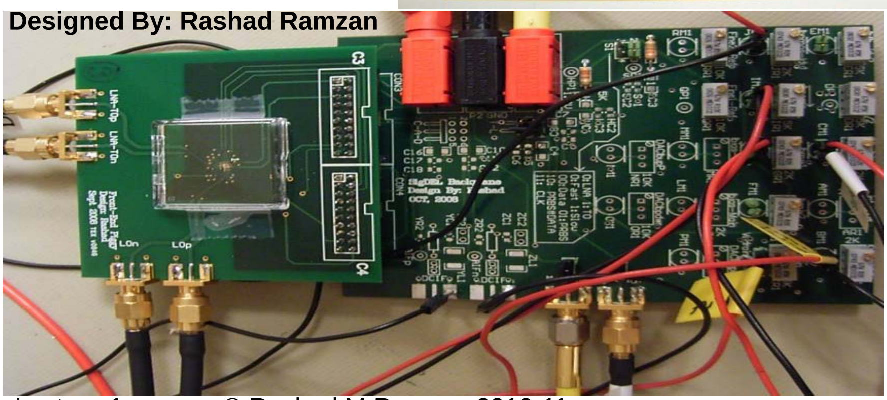
Designed By: Rashad Ramzan



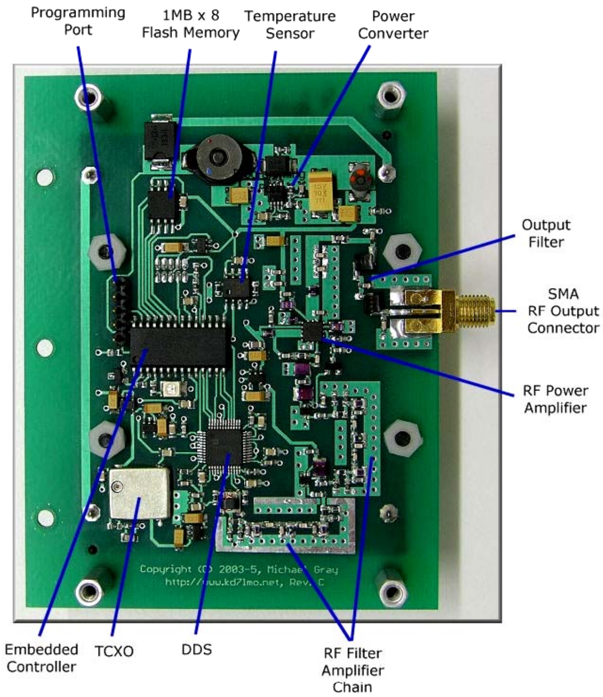
Designed By: Rashad Ramzan

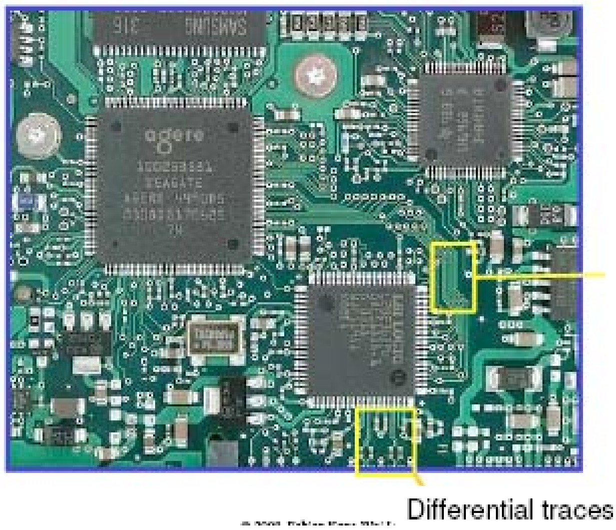


Designed By: Rashad Ramzan

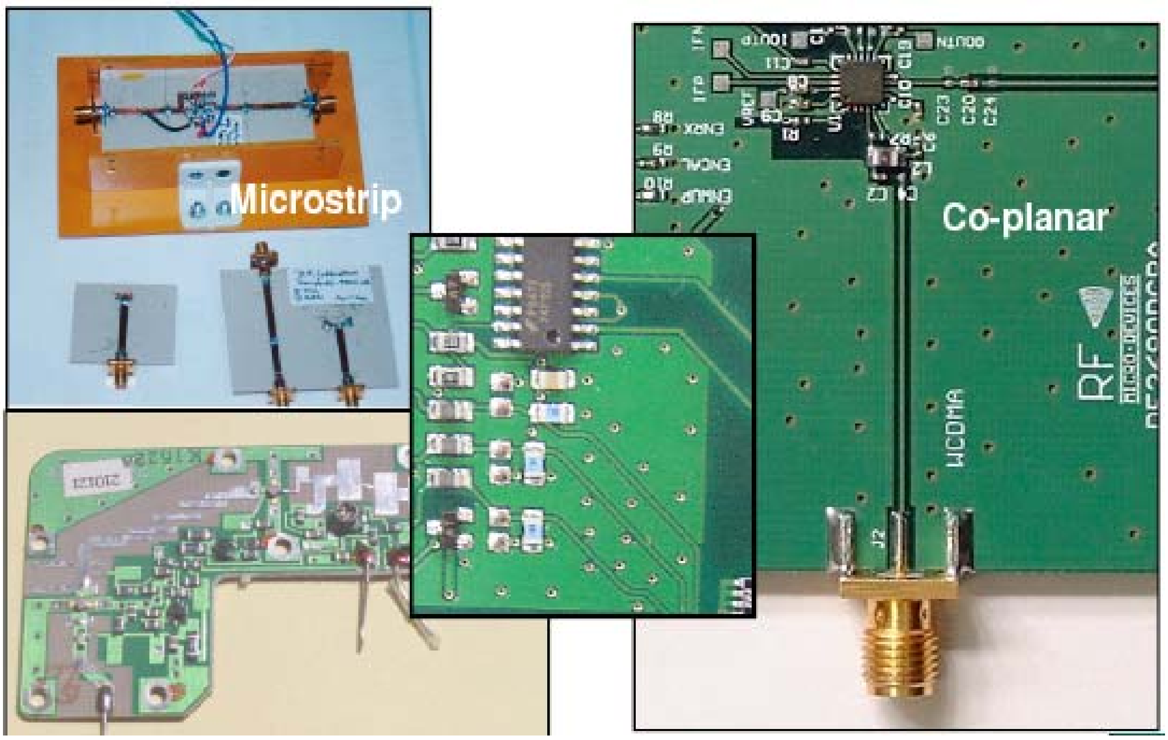


Digital & RF PCBs: Are they Same?

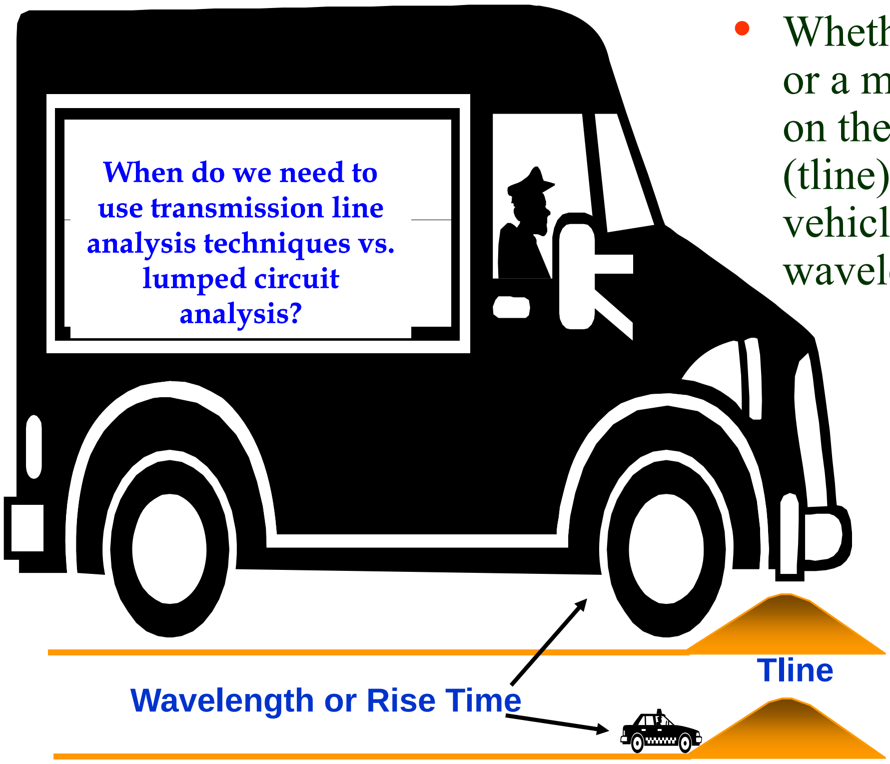




Digital & RF PCBs: Are they Same?



When a Trace is Transmission Line?



- Whether it is a bump or a mountain depends on the ratio of its size (tline) to the size of the vehicle (signal wavelength)

$t_{of} \geq 0.5 t_{rise}$

For squrewave

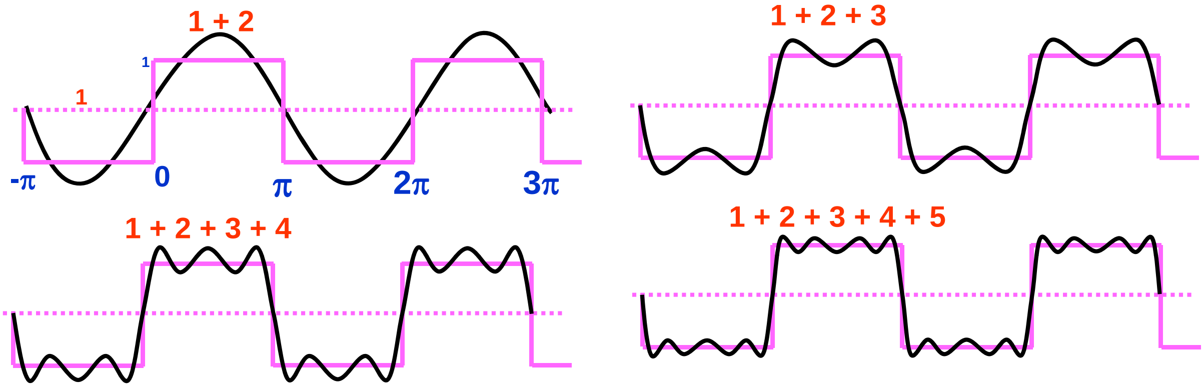
$l_{trans-line} \geq \lambda / 8$

For Sinewave

RF vs. Digital PCB: Sine vs. Square

Digital signals are composed of an infinite number of sinusoidal functions – the Fourier series

The Fourier series is shown in its progression to approximate a square wave:



Square wave: Y = 0 for -π < x < 0 and Y=1 for 0 < x < π

Y = 1/2 + 2/pi(sinx + sin3x/3 + sin5x/5 + sin7x/7 … + sin(2m+1)x/(2m+1) + …

1 2 3 4 5

May do with sum of cosines too.

RF vs. Digital PCB: Sine vs. Square

Where does that famous equation $F = \frac{0.35}{Tr}$ come from?

- It can be derived from the response of a step function into a filter with time constant tau

$$V = V_{input}(1 - e^{-t/\tau})$$

- Setting $V=0.1V_{input}$ and $V=0.9V_{input}$ allows the calculation of the 10-90% risetime in terms of the time constant

$$t_{10-90\%} = t_{90\%} - t_{10\%} = 2.3\tau - 0.105\tau = 2.195\tau$$

- The frequency response of a 1 pole network is

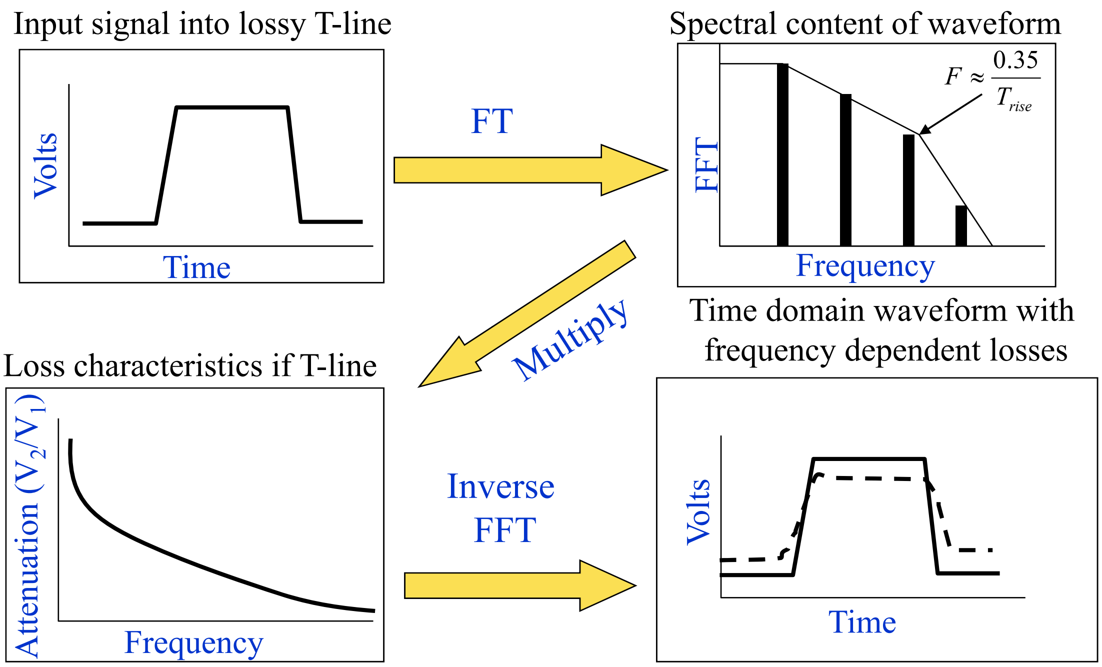
$$F_{3dB} = \frac{1}{2\pi\tau} \rightarrow \tau = \frac{1}{2\pi F_{3dB}}$$

- Substituting into the step response yields

$$t_{10-90\%} = \frac{1.09}{\pi F_{3dB}} = \frac{0.35}{F_{3dB}}$$

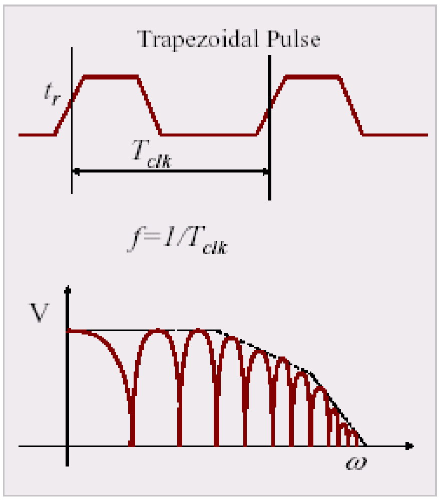
$$t_{10-90\%} = \frac{0.35}{F_{3dB}}$$

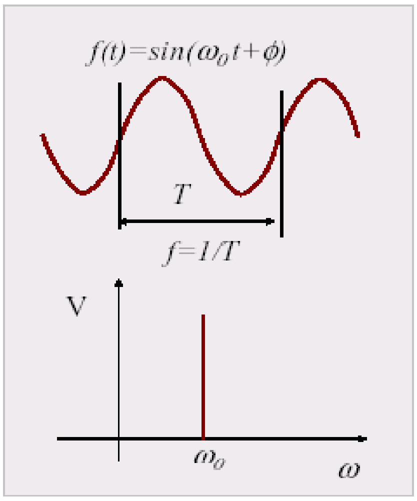
RF vs. Digital PCB: Sine vs. Square



RF vs. Digital PCB: Sine vs. Square

Digital PCB Tracks Carriers trapezoidal waves





RF PCB Tracks Carriers Sine waves

$$f_{3dB} = \frac{0.35}{t_{10-90\%}} = \frac{0.35}{t_r}$$

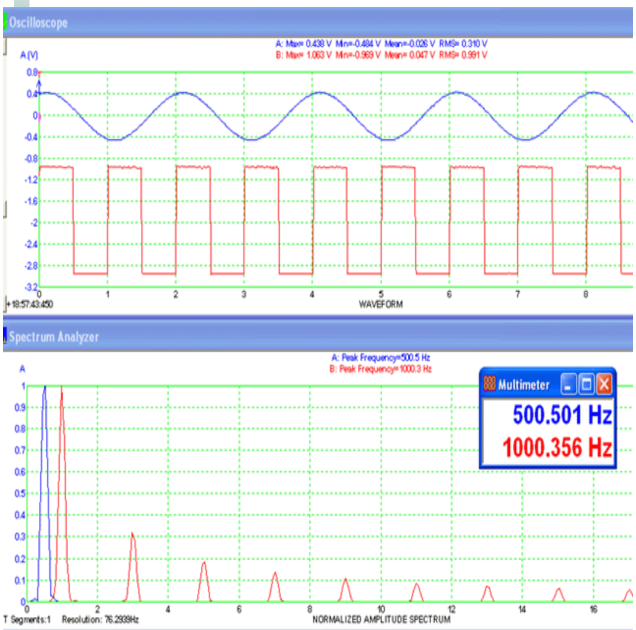
$$f_{BW} = \frac{4 \times 0.35}{t_r} = \frac{1.4}{t_r}$$

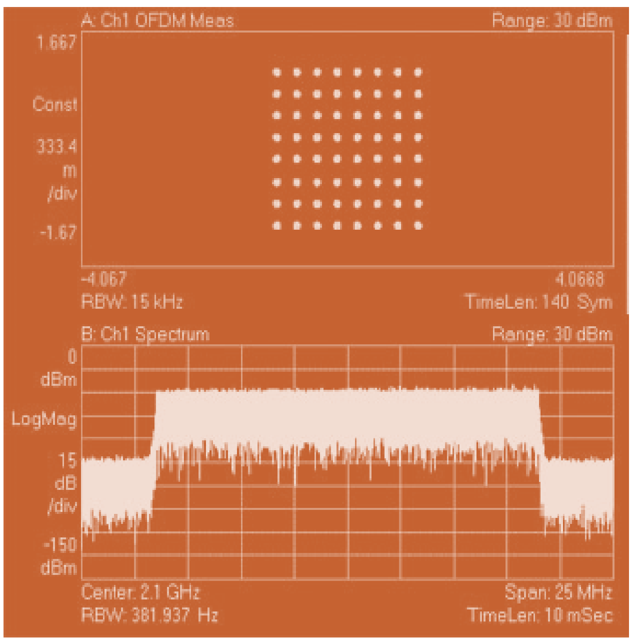
$$f_{BW} = f_O + \Delta f \quad or$$

$$\omega_{BW} = \omega_O + \Delta\omega$$

$\Delta\omega$ *is BW in case of mudulation*

Real Sine vs. Square





PCB Material: Dielectric Constant

- Is measure of how much charge two conductors can hold at a certain fixed voltage. Low Dk hold less charge and high Dk more charge. Its also measure of the ratio of velocity in conductor and free space.
  - High Dk → Small width for same Zo
  - High Dk → Large propagation delay

$$Z_o = \left(\frac{79}{\sqrt{\varepsilon_r + 1.41}}\right) ln \left(\frac{5.98\,H}{0.8\,W + T}\right) \Omega \quad \text{Valid for } 5 < W < 15 \text{ mils}$$

$$C_o = \frac{0.67(\varepsilon_r + 1.41)}{ln\left(\frac{5.98H}{0.8W + T}\right)} \text{ pf/in.}$$

$$t_{pd} = 1.017\sqrt{0.475\,\varepsilon_r + 0.67} \quad \text{(ns/ft)}$$

PCB Material: Loss Tangent

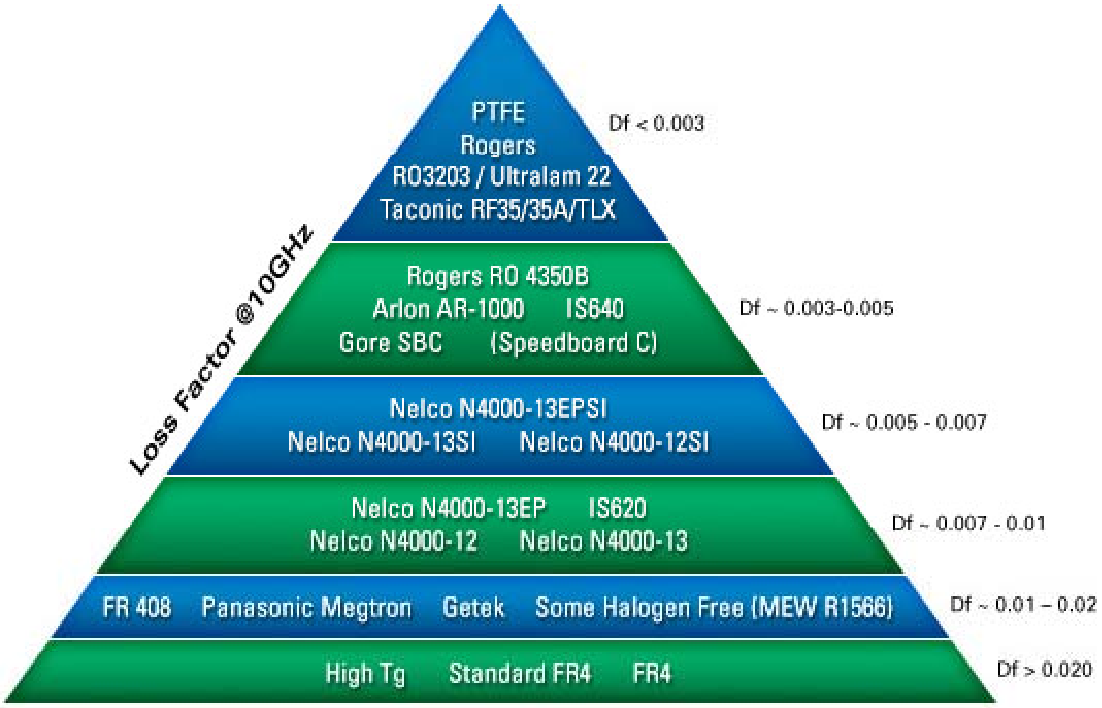
$$\alpha = 2.3\,f\,\tan(\delta)\cdot\sqrt{\varepsilon eff}$$

Where : α - Attenuation in dB / inch.
f - Frequency in GHz
tan (δ) - Loss tangent of material
εeff - Effective relative Er of material

- Is measure of how much electromagnetic energy is absorbed by dielectric material. Like microwave oven, things that heat up quickly has high loss tangent. Glass and ceramic are low Df materials.
  - Loss is frequency dependent, increases with frequency.
  - Low loss improves signal integrity--- Very Important for RF applications

RF vs. Digital: PCB Materials

Advanced Laminate Material



RF vs. Digital: PCB Materials

Materials intended for digital applications

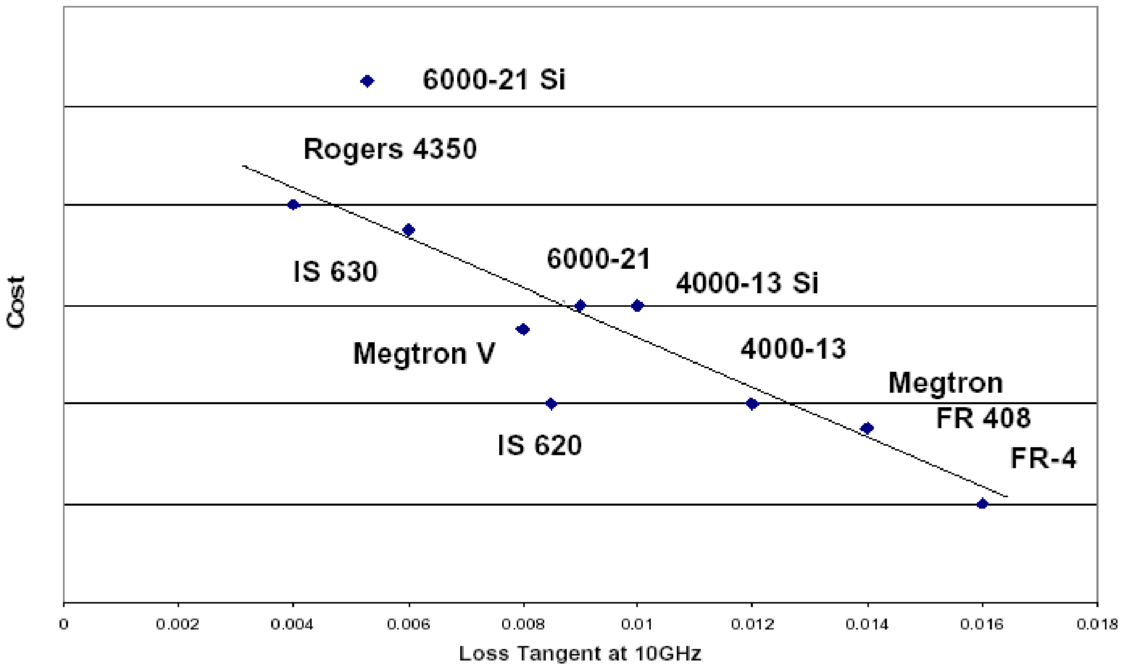
Material	Er (* at 1.0 MHz)	Thickness tolerance	Copper style	Multilayer compatible	Loss tangent
FR4	3.9 – 4.6*	+/- 1-2 mils	ED only	Yes	.02 - .03
FR408	3.4 – 4.1*	+/- 1-2 mils	ED only	Yes	.01 - .015
BT Epoxy	3.9 – 4.6*	+/- 1-2 mils	ED only	Yes	.015 - .02
Cyanate Ester	3.5 – 3.9*	+/- 1-2 mils	ED only	Yes	.009
Polyimide	4.0 – 4.5*	+/- 1-2 mils	ED only	Yes	.01
GETEK	3.5 – 4.2*	+/- 1-2 mils	ED only	Yes	.012
Nelco 4000-13	3.7 (1GHz)	+/- 1 mil	ED only	Yes	.01
Nelco 4000-13SI	3.5 (1GHz)	+/- 1 mil	ED only	Yes	.009
Nelco 6000	3.5 (1GHz)	+/- 1 mil	ED only	Yes	.008
Nelco 6000SI	3.2 (1GHz)	+/- 1 mil	ED only	Yes	.005
Speedboard N	3.0 *	+/- 1 mil	Prepreg	Yes	.02
Speedboard C	2.6 – 2.7*	+/- 1 mil	Prepreg	Yes	.004
Arlon 25 / Rogers 4003	3.4 (10GHz)	+/- 1 mil	ED only	Yes	.0027

RF vs. Digital: PCB Materials

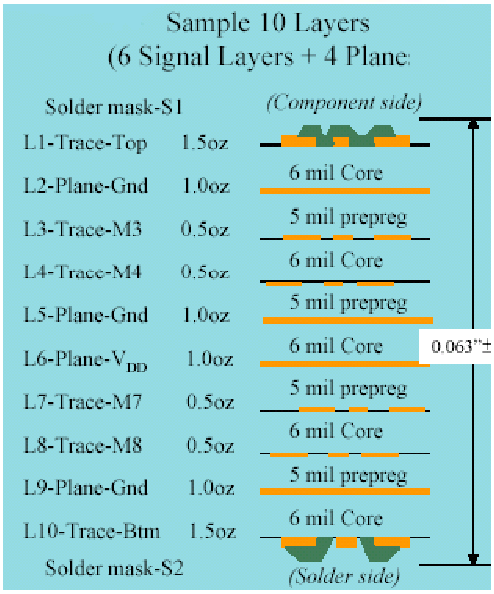
Materials for analog applications

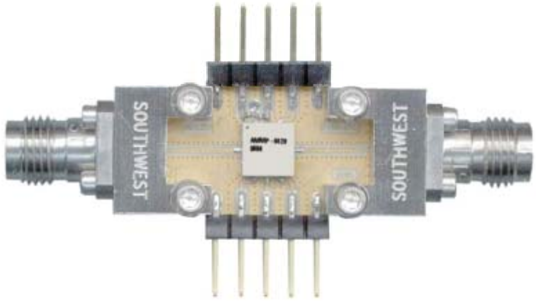
Material	Er (10.0 GHz)	Thickness tolerance	Copper style	Multilayer compatible	Loss tangent
Rogers Ultralam 2000	2.4 – 2.6	+/- .5 mil	ED / rolled	No	.0019
Rogers 5870	2.3	+/- .5 mil	ED / rolled	No	.0012
Rogers 5880	2.2	+/- .5 mil	ED / rolled	No	.0009
Rogers 6002	2.94	+/- .5 mil	ED / rolled	Yes	.0012
Rogers 3003	3.0	+/- 1 mil	ED / rolled	Yes	.0013
Rogers 6006	6.15	+/- .5 mil	ED / rolled	No	.0019
Rogers 6010	10.2	+/- .5 mil	ED / rolled	No	.0023
Rogers 3006	6.15	+/- 1 mil	ED / rolled	Yes	.0025
Rogers 3010	10.2	+/- 1 mil	ED / rolled	Yes	.0035

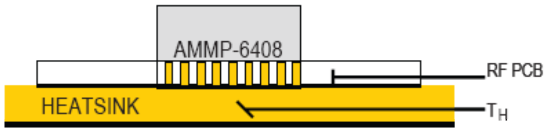
RF vs. Digital: PCB Materials



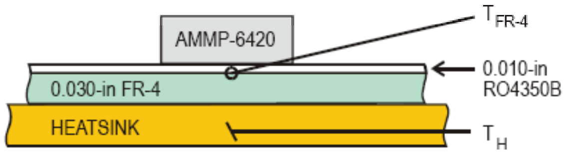
RF vs. Digital: Layer Stack







Single Layer



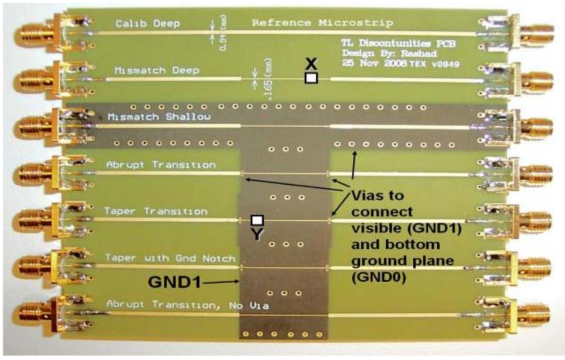
Multi Layer

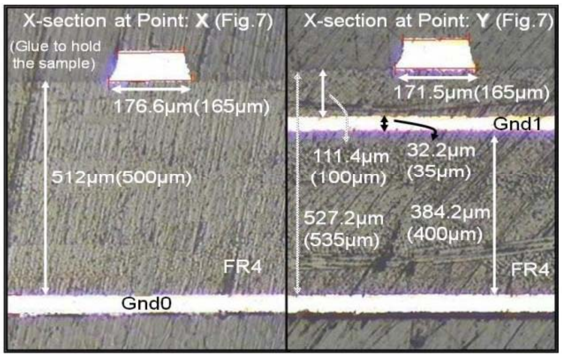
RF vs. Digital: 50Ω matching

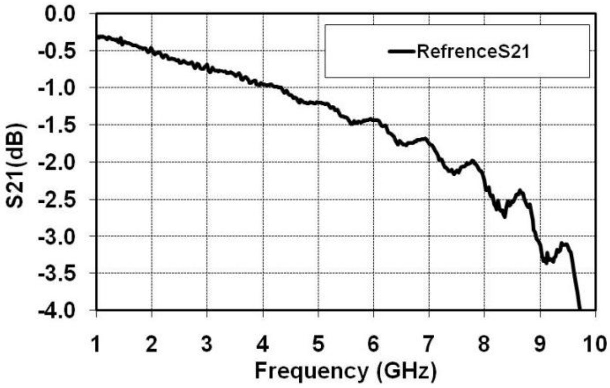
- Digital PCBs
  - Buses and CLK signals are laid out on controlled impedance traces of preferably 60 Ω and above with no load matching.
    - Reason: although reflections occurs, still we want to use the voltage divider rule and control the reflection by source series terminations
    - We can make the digital IC $Z_{in} \geq$ (few) KΩ at frequencies a high as ~800MHz
- RF PCBs
  - Its always 50Ω impedance matching
    - We can not make the RF IC $Z_{in} \geq$ (few)KΩ at GHz frequencies
    - 3dB (Half RF Power) loss in every connection

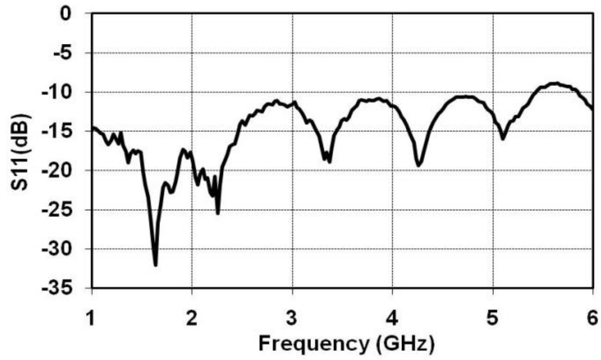
Microstrips on FR4 PCB

Designed and measured by: Rashad





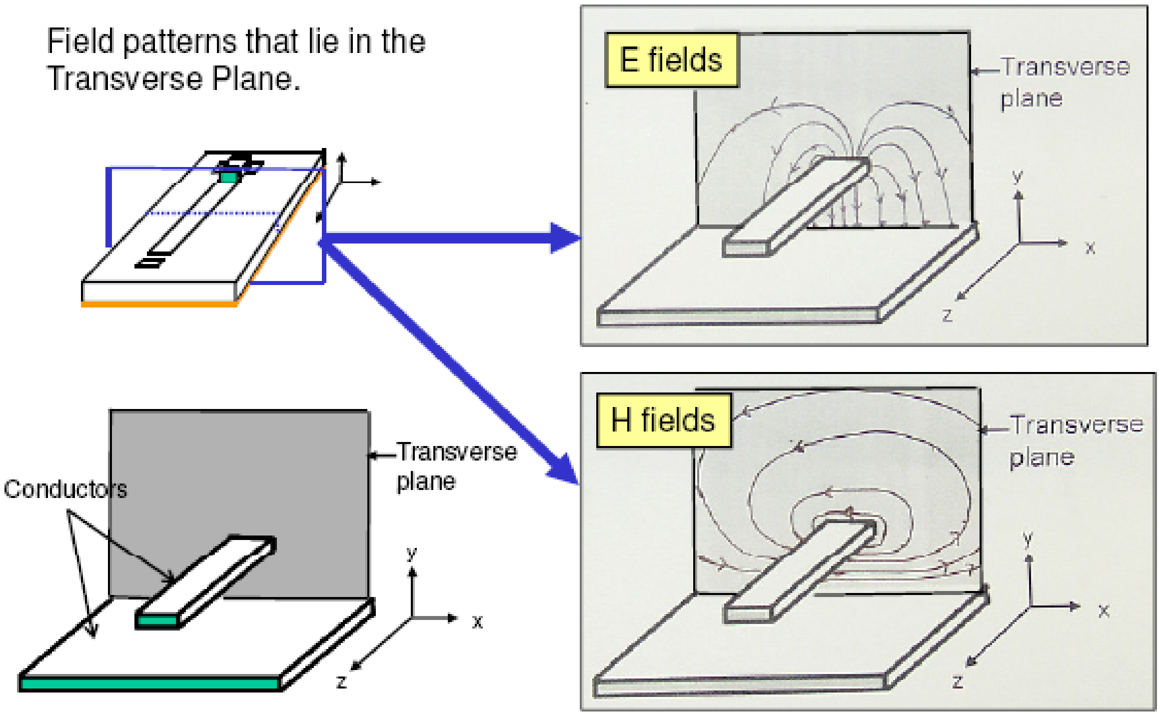




Propagation in Transmission Lines

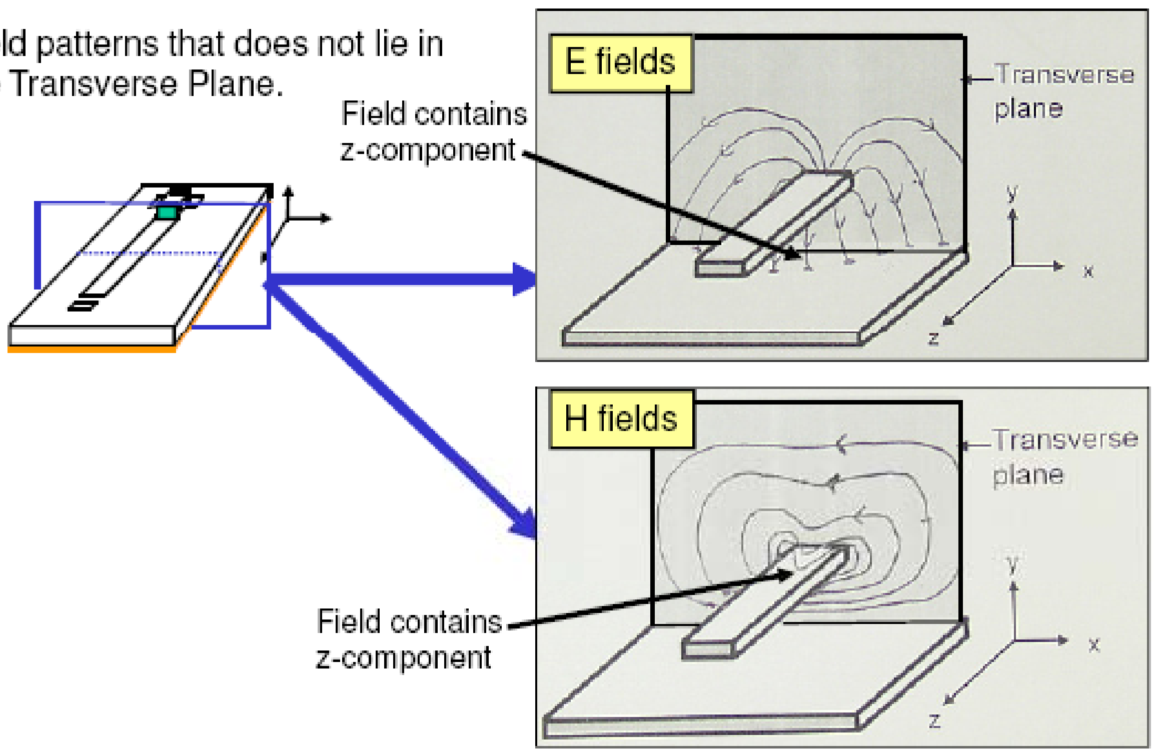
Transverse E & H Field Pattern

Field patterns that lie in the Transverse Plane.

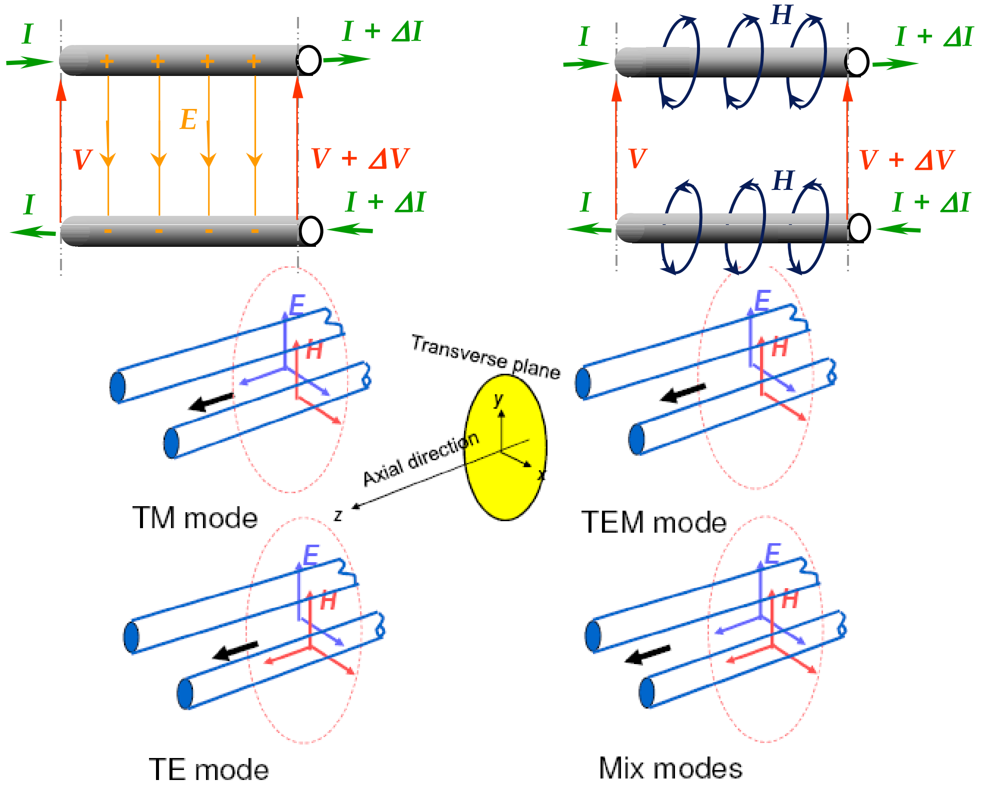


Non-transverse E & H Field Pattern

Field patterns that does not lie in the Transverse Plane.



Possible Propagation Modes

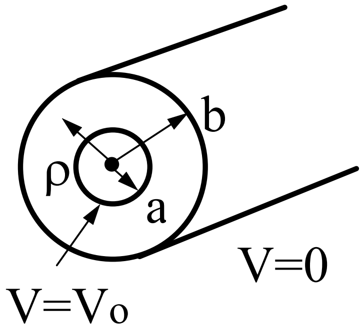


Why TEM is desirable?

- Cutoff frequency is zero – Therefore wideband transmission is possible like co-axial cables
- No dispersion, signals of different frequencies travel at the same speed, no distortion of signals
- <u>Sometime we deliberately want to have a cutoff frequency so that a microwave filter can be designed</u>

Coaxial Cables -TEM

- TEM exists in co-axial cable
- Higher-order modes exist in coaxial line but is usually suppressed
- Dimension of the coaxial line is controlled so that these higher-order modes are cutoff
- Dominate higher-order mode $\mathbf{TE_{11}}$ is mode, the cutoff wavenumber ($\mathbf{k_c}$) can only be obtained by solving a transcendental equation, approximate $\mathbf{k_c = 2/(a+b)}$ is often used in practice



$$H_y = \frac{-j}{k_c^2}\left(\omega\varepsilon \frac{\partial E_z}{\partial x} + \beta \frac{\partial H_z}{\partial y}\right)$$

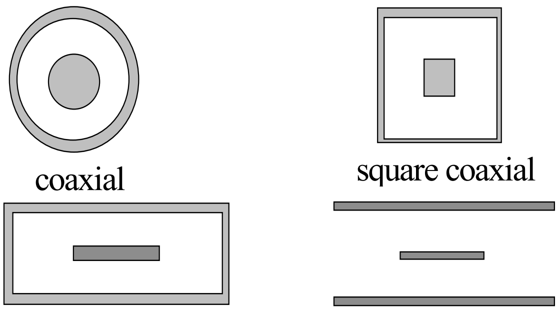
$$E_x = \frac{-j}{k_c^2}\left(\beta \frac{\partial E_z}{\partial x} + \omega\mu \frac{\partial H_z}{\partial y}\right)$$

$$Z_{TEM} = \frac{E_x}{H_y} = \frac{\omega\mu}{\beta} = \sqrt{\frac{\mu}{\varepsilon}} = \eta$$

$$Z_o = \frac{V_o}{I_a} = \frac{\eta \ln(b/a)}{2\pi}$$

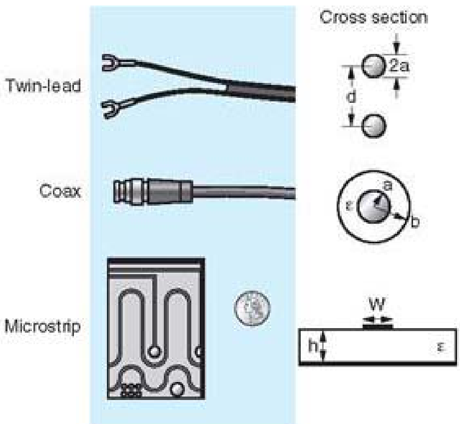
Microstrip and Striplines

Strip line was developed from the square coaxial

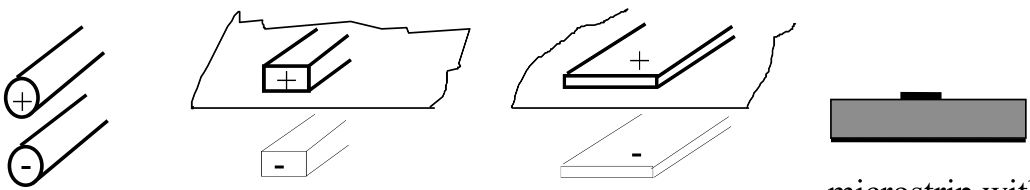


coaxial   square coaxial

rectangular line   flat stripline



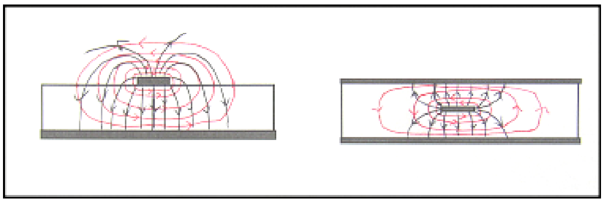
Evolution of microstrip



two-wire line   single-wire above ground (with image)   microstrip in air (with image)   microstrip with grounded slab

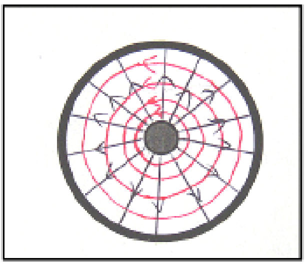
Modes in Microstrip

- A microstrip line suspended in air can support TEM wave
- A PCB microstrip does not support TEM wave
- A PCB microstrip fields constitute a hybrid TM-TE wave



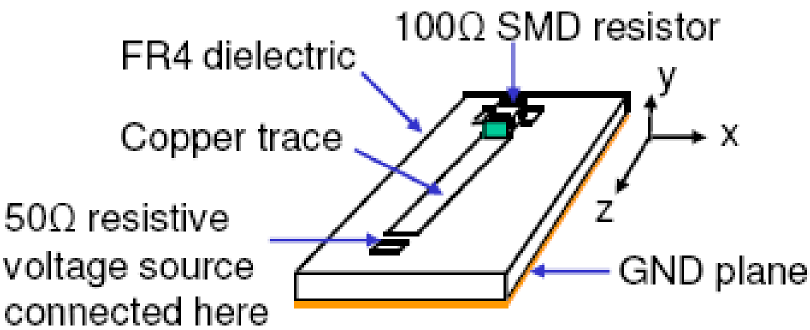
TEM or quasi-TEM mode for microstrip and stripline

E field
H field



TEM field pattern for coaxial cable

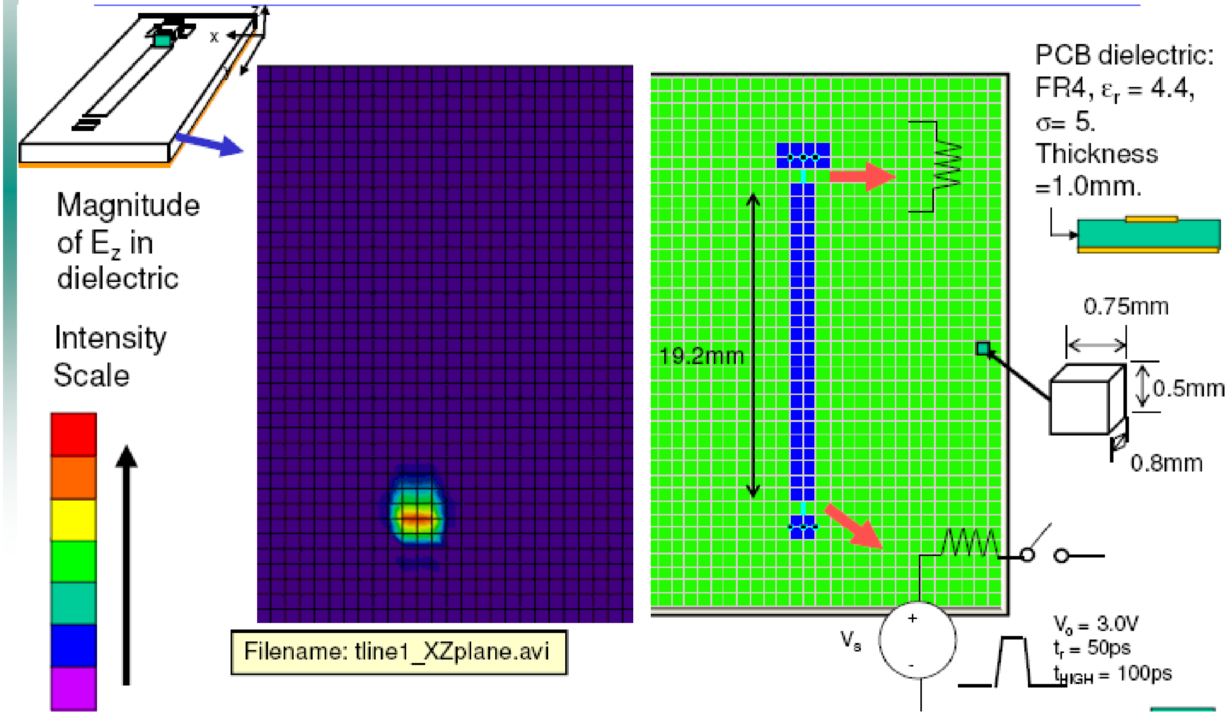
Demo EM Field in Microstrip



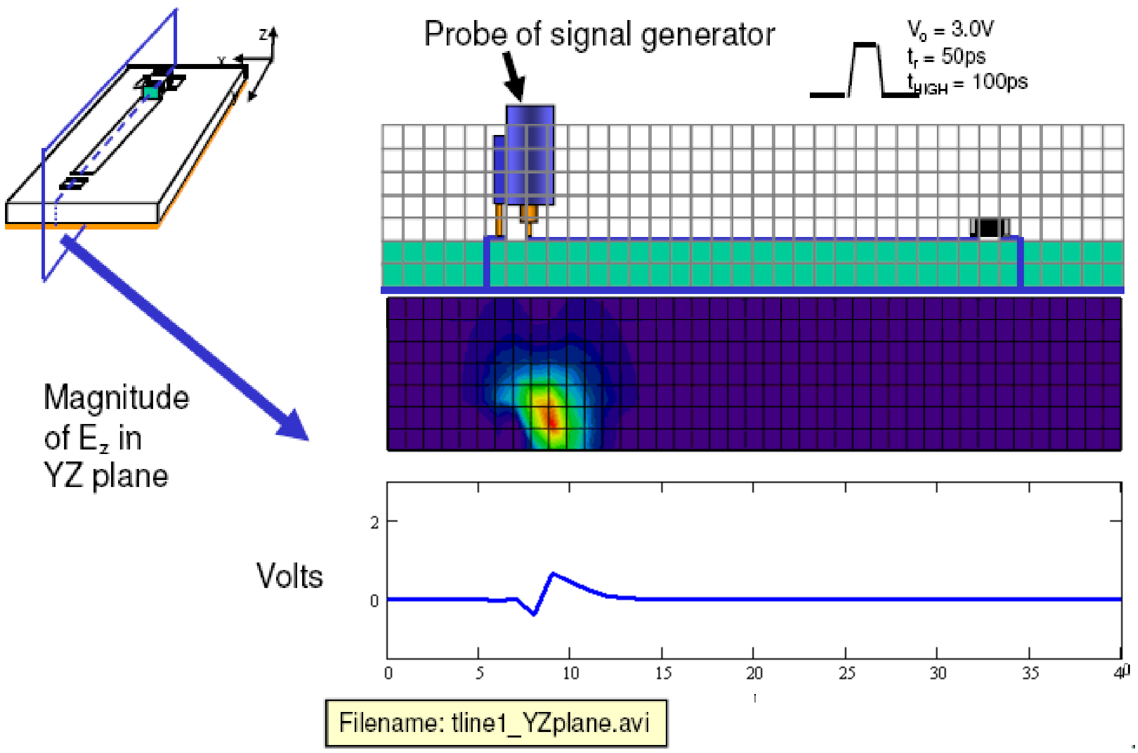
A numerical method, known as Finite-Difference Time-Domain (FDTD) is applied to Maxwell's Equations, to provide the approximate values of E and H fields at selected points on the model at every 1.0 picoseconds

Ref: http://pesona.mmu.edu.my/~wlkung/Phd/phdthesis.htm)

Magnitude of E_z in Dielectric



Magnitude of E_z in YZ Plane



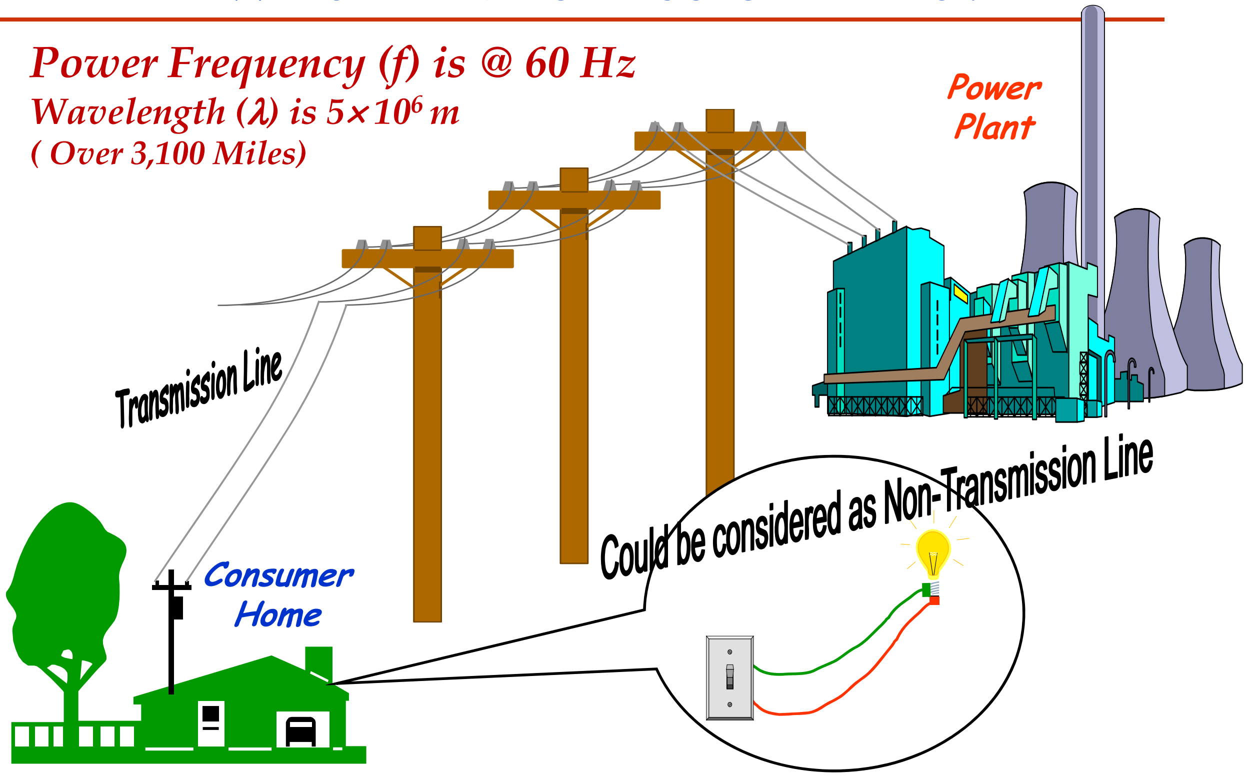
Comparison: PCB Transmission Lines

Microstrip line	Stripline	Co-planar line
Suffers from dispersion and non-TEM modes	Pure TEM mode	Suffers from dispersion and non-TEM modes
Easy to fabricate	Difficult to fabricate	Fairly difficult to fabricate
High density trace	Mid density trace	Low density trace
Fair for coupled line structures	Good for coupled line structures	Not suitable for coupled line structures
Need through holes to connect to ground	Need through holes to connect to ground	No through hole required to connect to ground

Impedance Matching

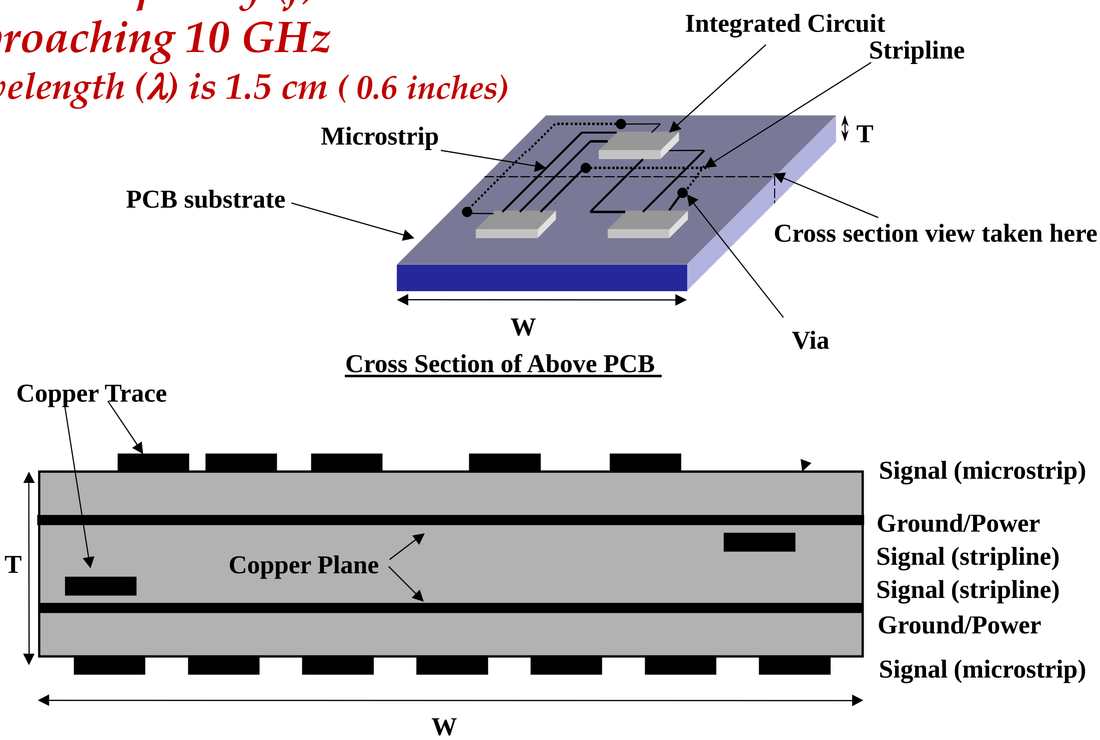
When Transmission Line?

Power Frequency (f) is @ 60 Hz
Wavelength (λ) is 5×10^6 m
(Over 3,100 Miles)



PCB Transmission Lines?

Signal Frequency (f) is approaching 10 GHz
Wavelength (λ) is 1.5 cm (0.6 inches)



When Transmission Lines in RF PCB?

$$t_{of} > 0.5 t_{rise} \quad For\ squrewave \quad (Digital\ PCBs)$$

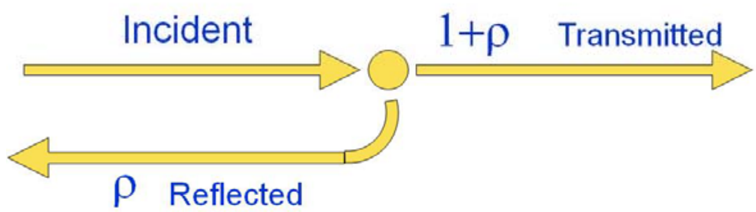
$$where\ t_{of} = \frac{length}{Phase\ Velocity}, called\ time\ of\ flight$$

$$l_{trans-line} > \lambda/10\ (For\ Sine\ in\ Digital\ PCBs)$$

$$l_{trans-line} > \lambda/20\ (For\ RF\ PCBs)$$

- In Digital PCBs we treat the signal in voltage & use voltage divisor rules
- In RF we deal with RF power, not the voltage & current, then the reflection becomes important

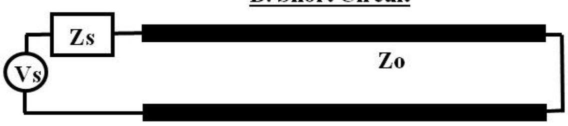
Reflection Constant



A: Terminated in Zo

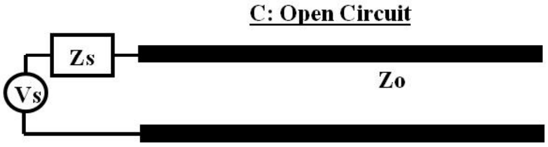
$$\rho = \frac{Zo - Zo}{Zo + Zo} = 0$$

B: Short Circuit



$$\rho = \frac{0 - Zo}{0 + Zo} = -1$$

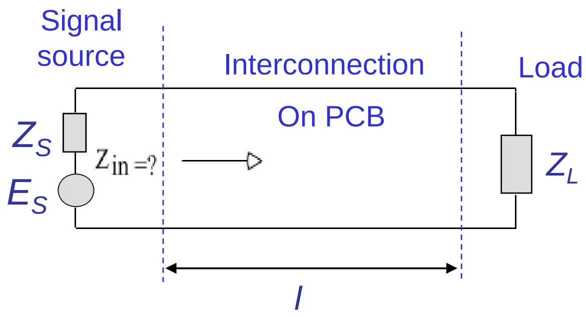
C: Open Circuit



$$\rho = \frac{\infty - Zo}{\infty + Zo} = 1$$

Why Impedance Matching?

- Repetition of Lecture on Transmission Lines



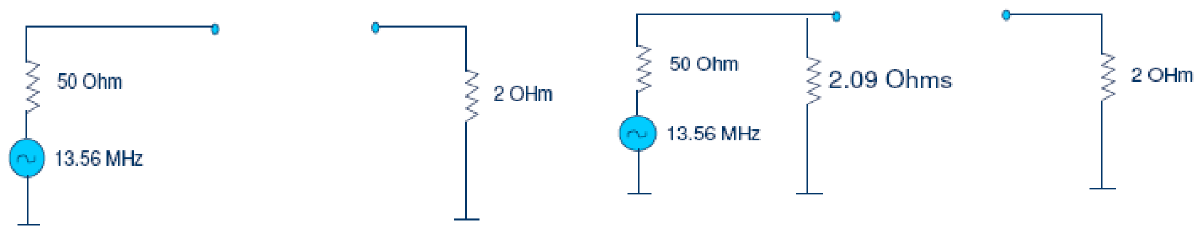
$$Z = Z_0 \frac{Z_L + jZ_0 \tan\theta}{Z_0 + jZ_L \tan\theta} \quad (1)$$

Z_L	l/λ	θ	
∞	1/6	60°	−28.87j
0	1/6	60°	86.6j
∞	1/4	90°	0
0	1/4	90°	∞
∞	1/2	180°	∞
0	1/2	180°	0
∞	1/3	120°	28.87j
0	1/3	120°	−86.60j

- When matched $Z_0 = Z_L$ then Z_{in} is not dependent upon length of line.

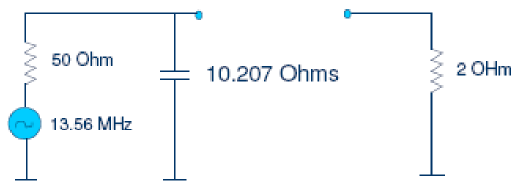
How Matching Works?

We want to match Zs of 50Ω with 2 Ω Load???



$$Z = \frac{R1 \times R2}{R1 + R2} = 2 \text{ Ohm}$$

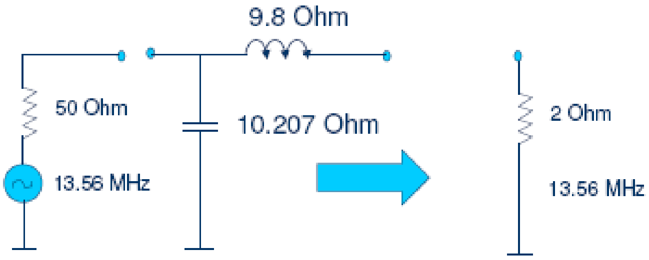
Now lets see what will happen if we replace the resistor with a capacitor of 1150 pF which has impedance X=1/2ΠfC =-j10.207 Ohm @ 13.56 MHz



$$Z = \frac{R.(-jX)}{R+(-jX)} = 2 - j\,9.8 \text{ Ohms}$$

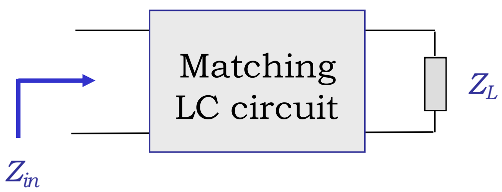
How Matching Works?

$$Z = \frac{R.(-jX)}{R+(-jX)} = 2 - j\,9.8 \text{ Ohms}$$



So if we add an inductor which impedance value @ 13.56 MHz is 9.8 Ohm Inductor and capacitance will cancel each other and our match will be complete.
L=115 nH @13.56 MHz

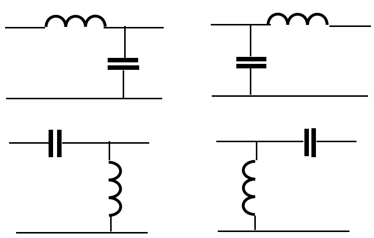
LC Impedance Matching



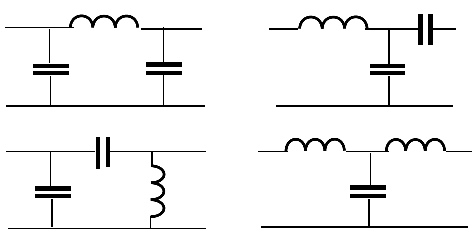
In most cases Z_{in} is expected to be real

Impedance must be matched:
$X_{in} = -X_s$ and $R_{in} = R_s$

L-match circuits:

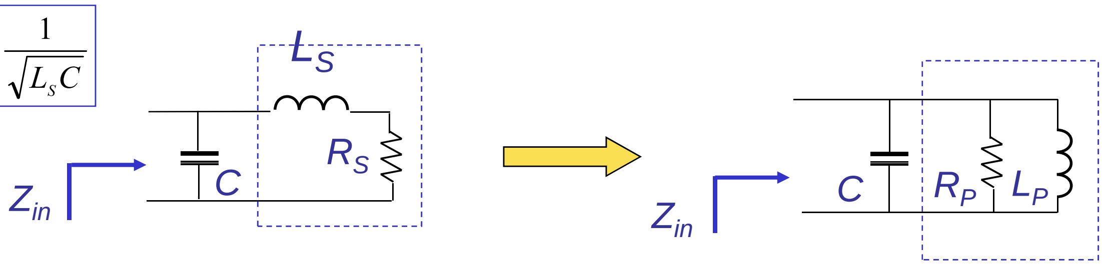


π-match and T-match circuits:



Narrow Band Transformation

$$\omega_0 \cong \frac{1}{\sqrt{L_S C}}$$



$$R_S + j\omega L_S = \frac{R_P j\omega L_P}{R_P + j\omega L_P}$$

$$R_S = \left.\frac{R_P}{R_P^2/(\omega L_P)^2 + 1}\right|_{\omega\approx\omega_0} = \frac{R_P}{Q^2+1} \approx \frac{R_P}{Q^2}$$

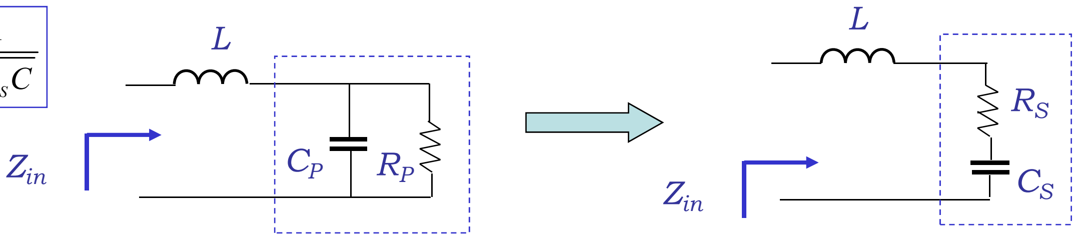
$$L_S = \left.\frac{L_P \cdot R_P^2/(\omega L_P)^2}{R_P^2/(\omega L_P)^2 + 1}\right|_{\omega\approx\omega_0} = \frac{L_P \cdot Q^2}{Q^2+1} \approx L_P$$

$$Z_{in}\Big|_{\omega\approx\omega_0} = R_P \approx Q^2 R_S$$

Upwards Resistance Transformer

Pi & T match Circuits

$$\omega_0 \cong \frac{1}{\sqrt{L_S C}}$$



$$\frac{1}{R_P} + j\omega C_P = \frac{j\omega C_S / R_S}{1/R_S + j\omega C_S}$$

$$R_S = \left.\frac{R_P}{R_P^2(\omega C_P)^2 + 1}\right|_{\omega\approx\omega_0} = \frac{R_P}{Q^2+1} \approx \frac{R_P}{Q^2}$$

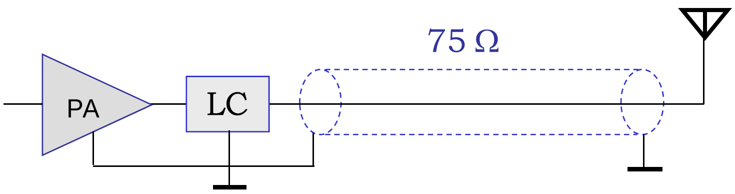
$$C_S = C_P \cdot \left.\frac{R_P^2(\omega C_P)^2 + 1}{R_P^2(\omega C_P)^2}\right|_{\omega\approx\omega_0} = C_P \cdot \frac{Q^2+1}{Q^2} \approx C_P$$

$$Z_{in}\Big|_{\omega\approx\omega_0} = R_S \approx \frac{R_P}{Q^2}$$

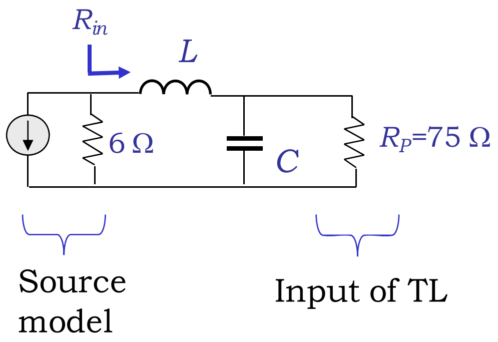
As $C_S \approx C$ the equivalent circuit has also a resonance at ω_0

Note that once R_P and R_S are related, Q is defined and it cannot be improved by those simple L-match circuits

Example LC Matching



Match PA of output resistance 6 Ω to 75 Ω transmission line, also matched to antenna.
f=100MHz



$$R_{in} = \frac{R_P}{Q^2+1} \rightarrow Q^2 = \frac{R_P}{R_{in}} - 1 = 11.5$$

$$Q^2 = \frac{(R_P)^2}{L/C} \rightarrow L/C = \frac{75^2}{11.5} = 489$$

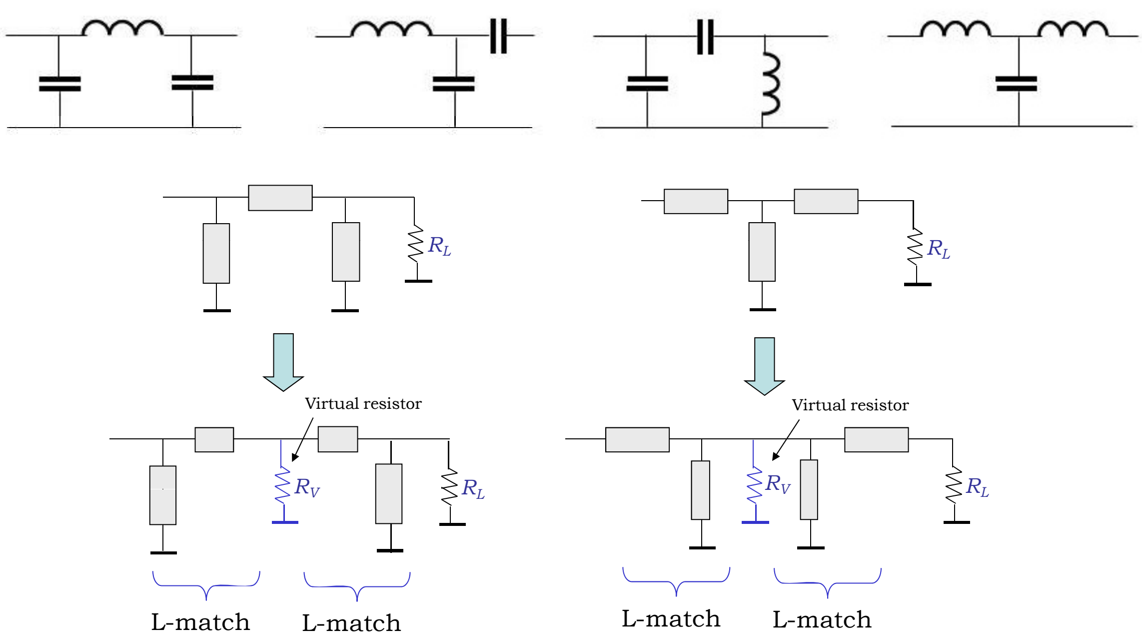
$$(2\pi f)^2 = \frac{1}{LC} \rightarrow LC = \frac{1}{(2\pi \cdot 10^8)^2} = 254 \cdot 10^{-20}$$

$$L = 35.2 \text{ nH}, \quad C = 72 \text{ pF}$$

Observe that the matching circuit also works as LPF that attenuates higher harmonic components of PA

Pi & T Match Circuits

π-match and T-match circuits:



There is one degree of freedom more, we can also choose Q

Smith charts revisited

$$\rho = \frac{Z_L - Z_0}{Z_L + Z_0} = \frac{Z_L/Z_0 - 1}{Z_L/Z_0 + 1} = \frac{Z - 1}{Z + 1}$$

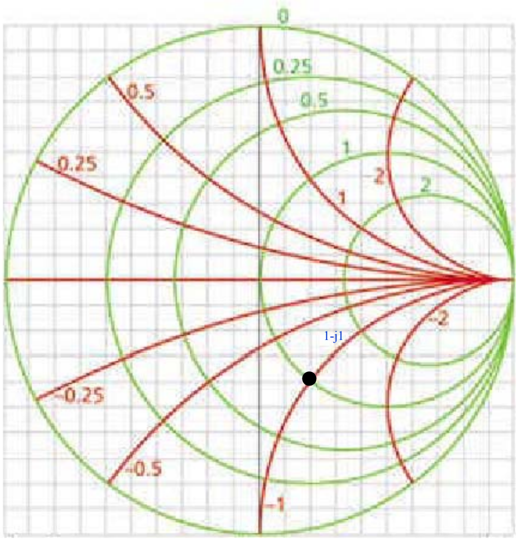
$$Z = a + jb, \quad \rho = c + jd$$

Normalized loading impedance

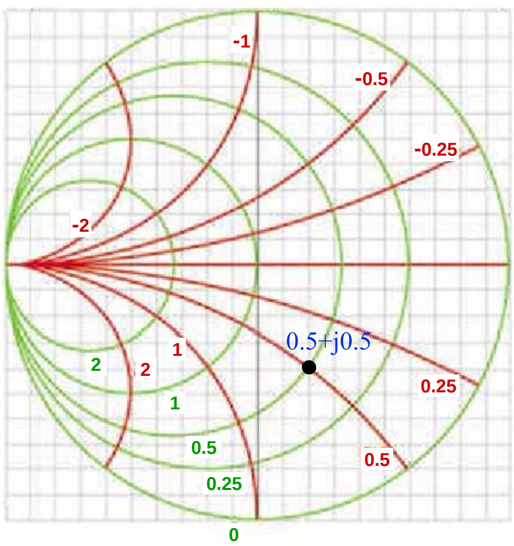
$$\rho = c + jd = \frac{a^2 + b^2 - 1}{(a+1)^2 + b^2} + j\left(\frac{2b}{(a+1)^2 + b^2}\right)$$

Impedance – Admittance Conversion

Constant R and X circles

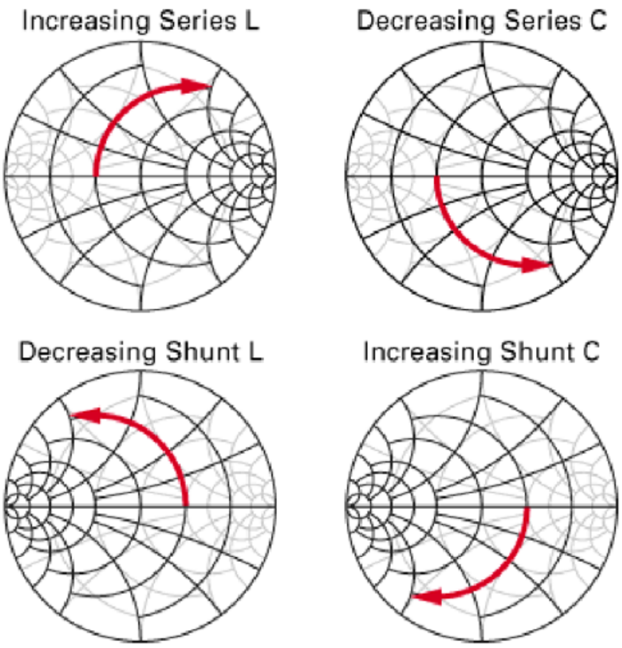


Constant G and B circles



Z = R+jX Same load Y = 1/Z = G+jB

Modifying Admittance or Impedance



$\Delta X>0$ $\Delta X<0$

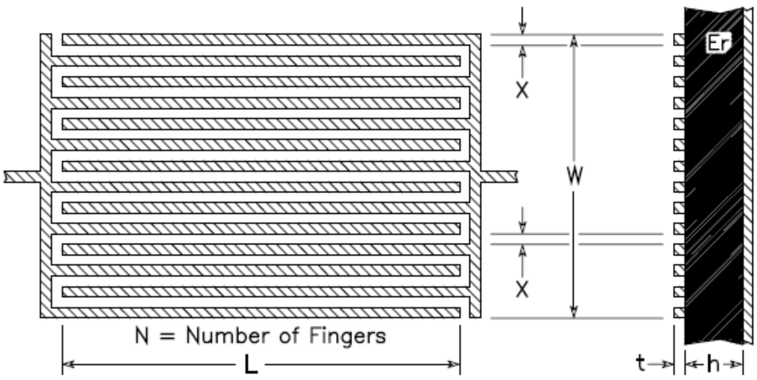
$\Delta B<0$ $\Delta B>0$

Impedance Matching: Smith Chart

- Show the Simulation of the Impedance matching…..You can also see these simulation at….
- http://www.amanogawa.com/archive/transmissionA.html

μ-wave Components

μ-wave Capacitors



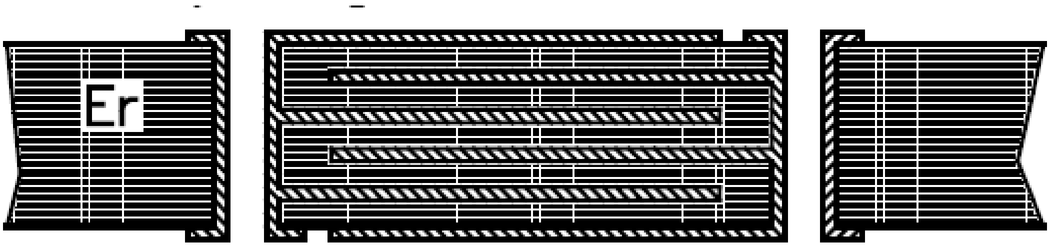
$$C_2 = \frac{\varepsilon_r + 1.0}{w} L\left[(N - 3.0)A1 + A2\right] \quad \text{(pF / in)}$$

$$A1 = \left[0.3349057 - 0.15287116\left(\frac{t}{X}\right)\right]^2$$

$$A2 = \left[0.50133101 - 0.22820444\left(\frac{t}{X}\right)\right]^2$$

(Equation valid for h>w/N)

µ-wave Multilayer Capacitor



$$C = \frac{0.229\varepsilon_r A(n-1.0)}{d} \quad \text{(pF)}$$

where:

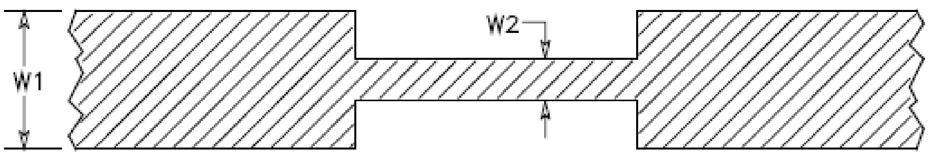
A = area of planes in square inches

n = number of conductor layers

d = plate spacing

µ-wave Inductor

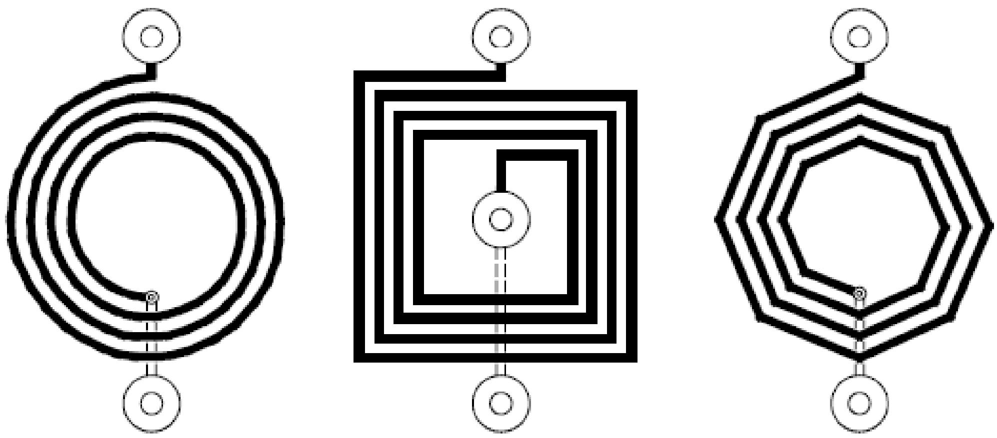
Inductor- Inline Inductor is formed by a Very Thin, High Impedance Trace.



- Length Must be Shorter than Critical Length to Prevent Reflections. Can Remove Plane(s) to Boost Inductance.
- L = Zo^2 x C or Tpd x Zo (Many Equations available. This is Extremely Accurate.)

µ-wave Inductors

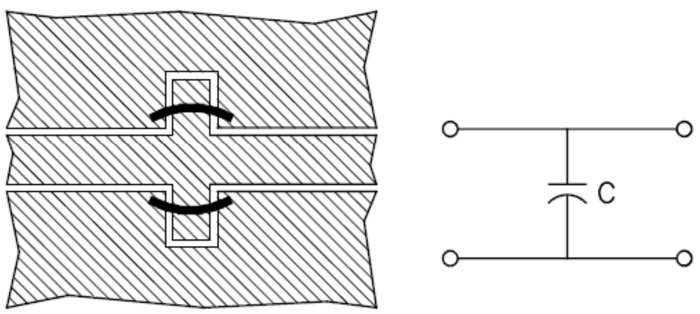
✦ Spiral Inductors -



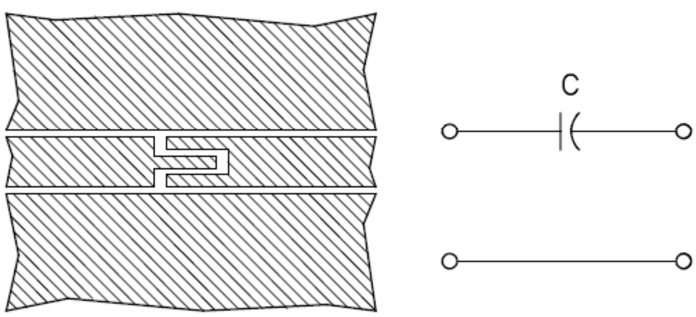
(Equations in Wadell- Pages 392-406)

µ-wave Capacitors

CPW & CPWG Shunt Capacitor -

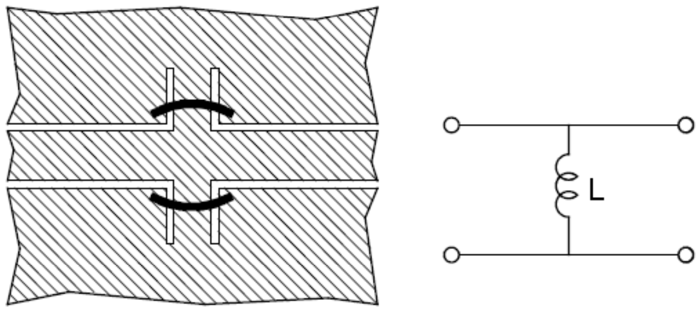


CPW & CPWG Series Capacitor -

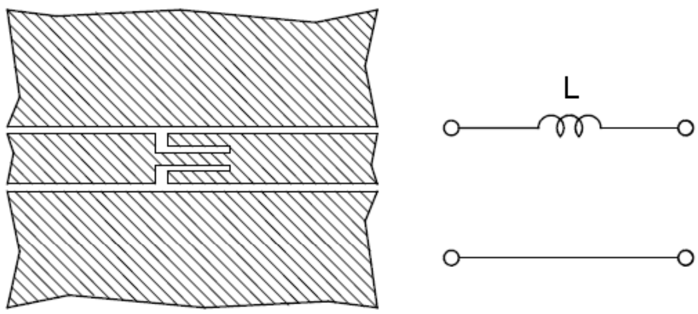


µ-wave Inductors

CPW & CPWG Shunt Inductor -

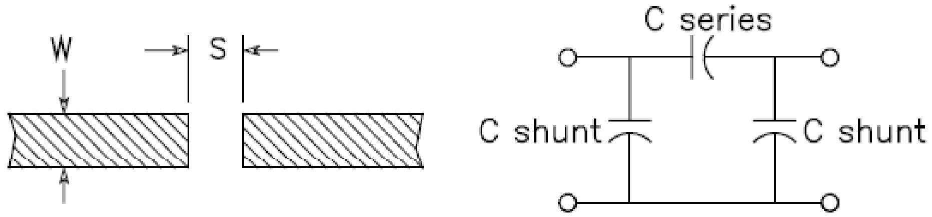


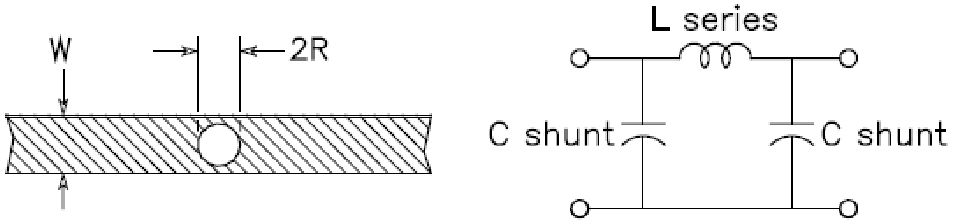
CPW & CPWG Series Inductor -



µ-wave Capacitors & Inductors

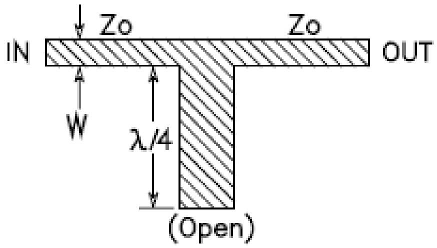
Gap in Centered Stripline Conductor-



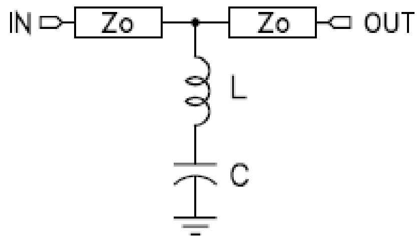


μ-wave Capacitors & Inductors

- λ/4 Stub is Series Resonant Circuit at Frequency.
- Circuit Shorts to Ground at λ/4, 3/4λ, etc.
- Open Circuit at DC, λ/2, λ, etc.



Microstrip Open–Stub

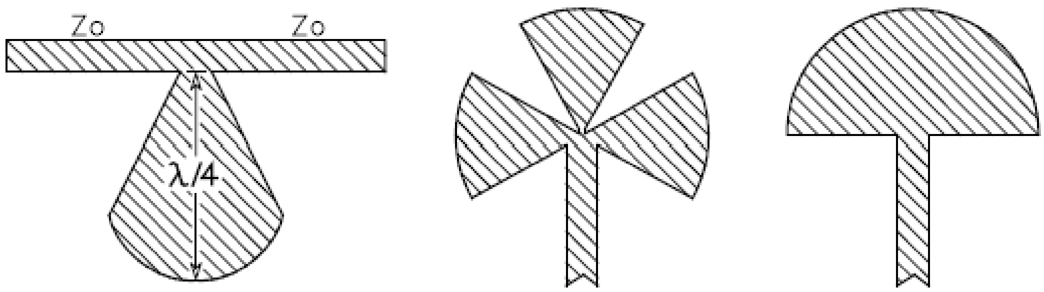


Microstrip Open–Stub Equivalent Circuit at Resonant Frequency

- 2W Wide for High Q and to Prevent Reflections.

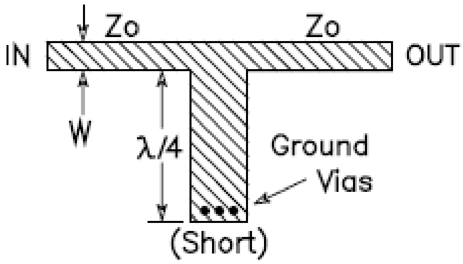
μ-wave Capacitors & Inductors

- Open Stubs (one just shown) have Narrow Frequency Over which they Short to Ground.
- Flaring the Stub Increases Frequency Response.

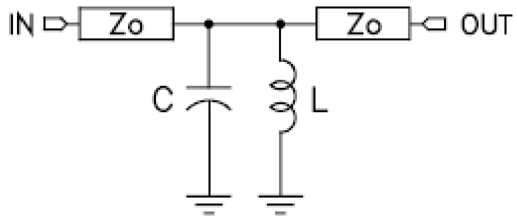


μ-wave Capacitors & Inductors

- λ/4 Stub, Shorted to Ground, is Parallel Resonant Filter at Frequency of Interest.
- Circuit Shorts to Ground at DC, λ/2, λ, etc.
- Open Circuit at λ/4, 3/4 λ, etc.



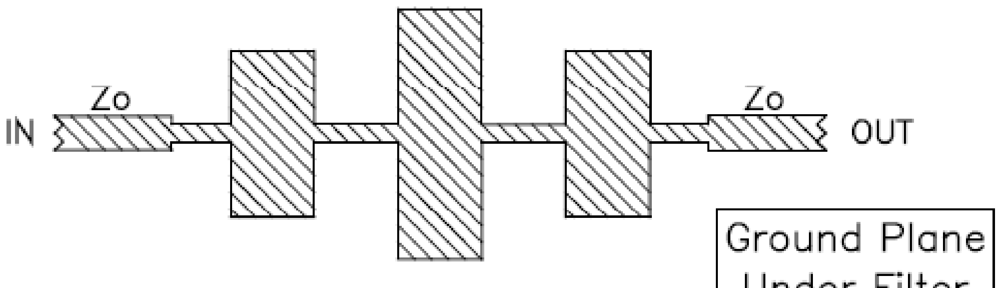
Microstrip Shorted–Stub



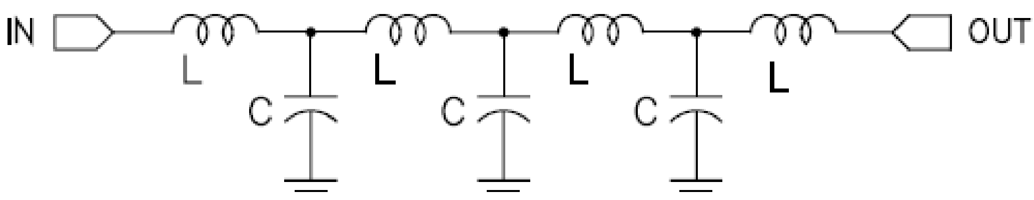
Microstrip Shorted–Stub Equivalent Circuit at Resonant Frequency

μ-wave Low Pass Filter

- Low Pass Filter -

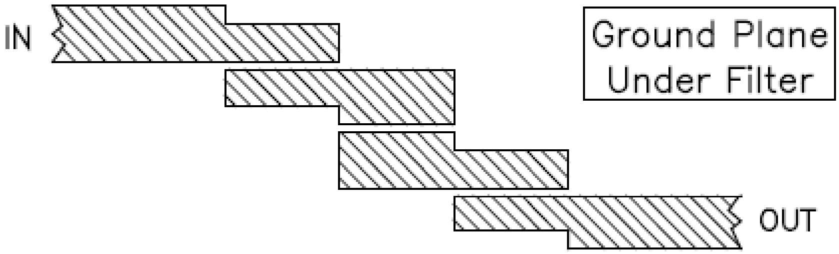


Microstrip Low Pass Filter

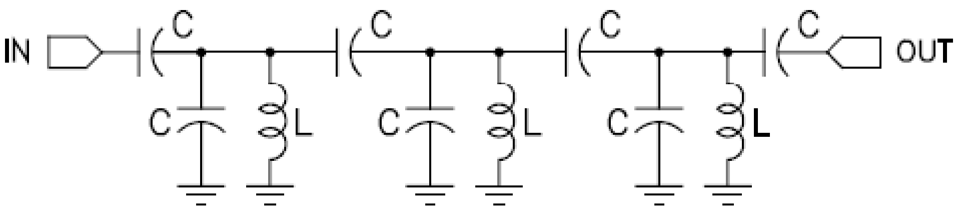


μ-wave Band Pass Filter

- Edge Coupled Band Pass Filter -

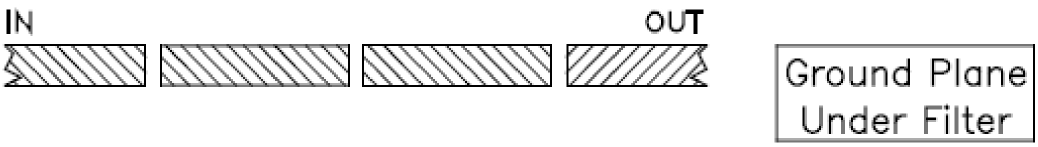


Microstrip Band Pass Filter (Edge Coupled)

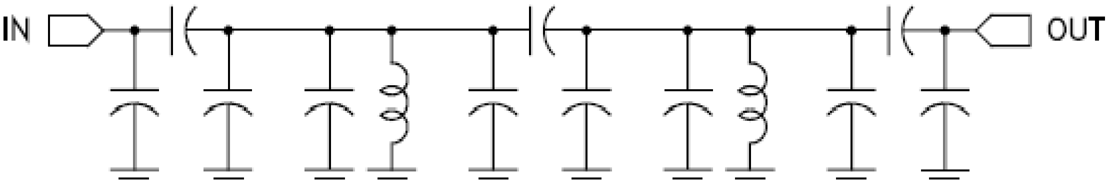


μ-wave Band Pass Filter

- End Coupled Band Pass Filter -

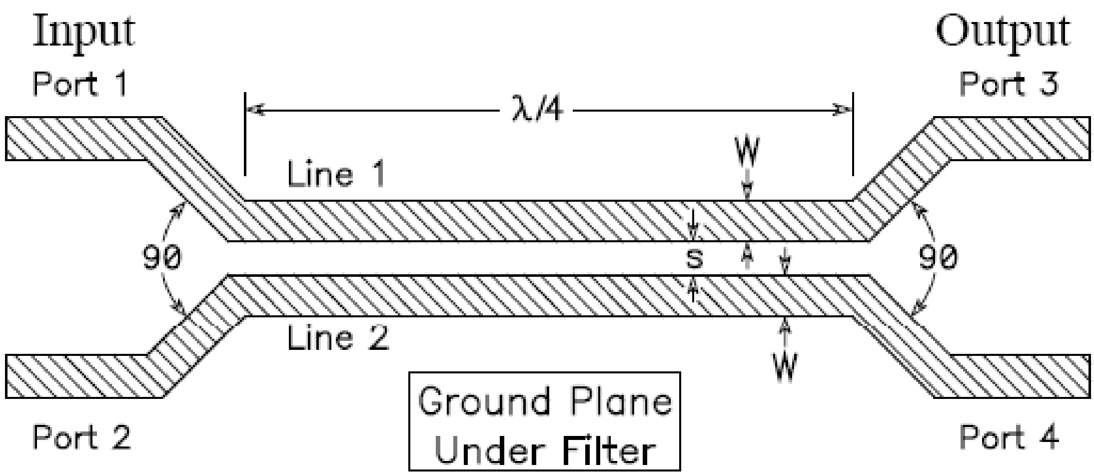


Microstrip Band Pass Filter (End Coupled)



Directional Coupler

- Directional Coupler -

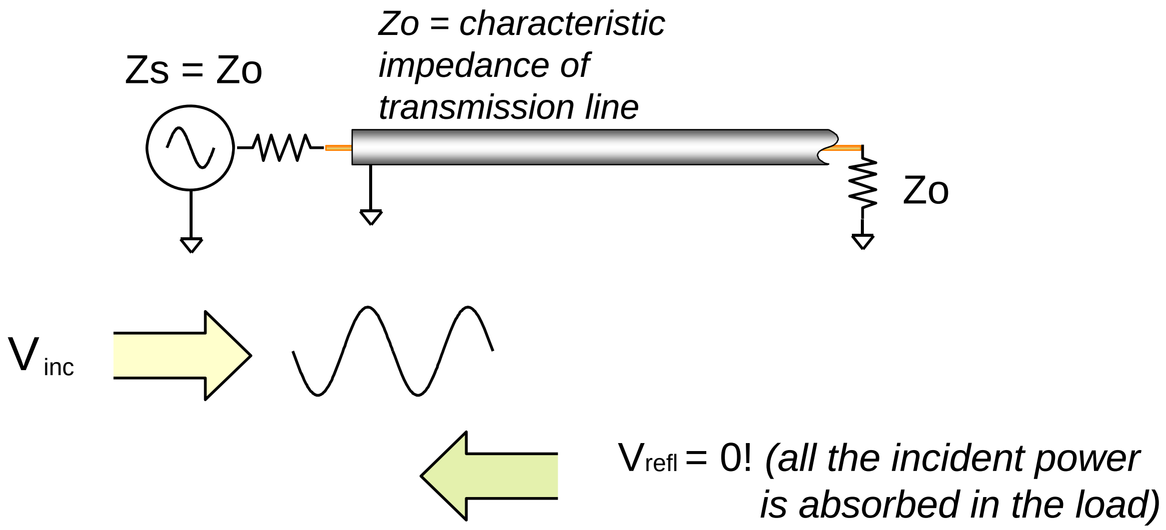


Strongest Coupled Pulse

Weakest Coupled Pulse

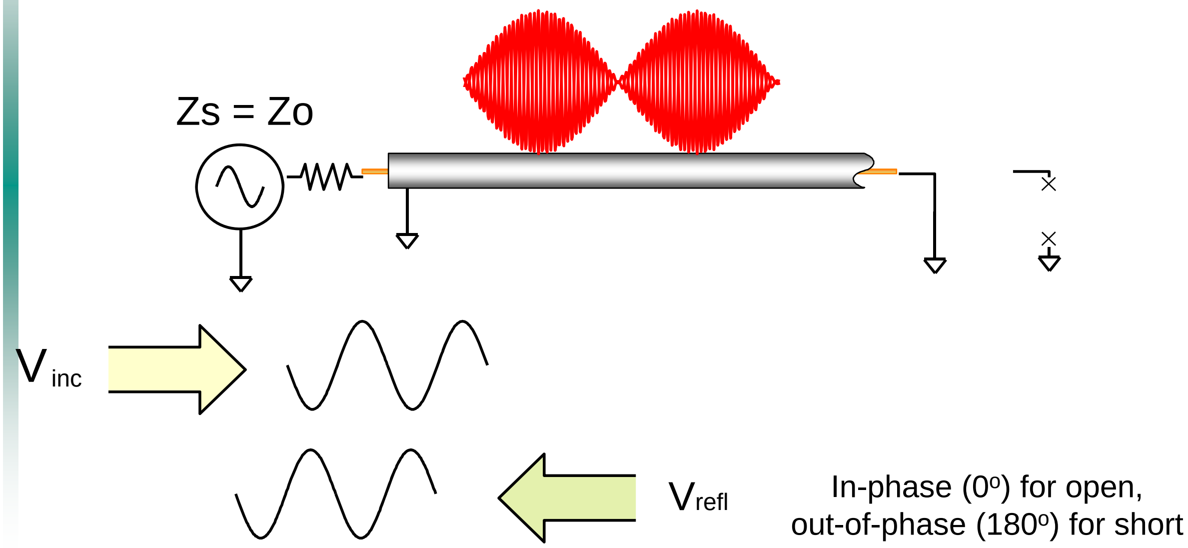
S-Parameters

Transmission Line Terminated with Zo



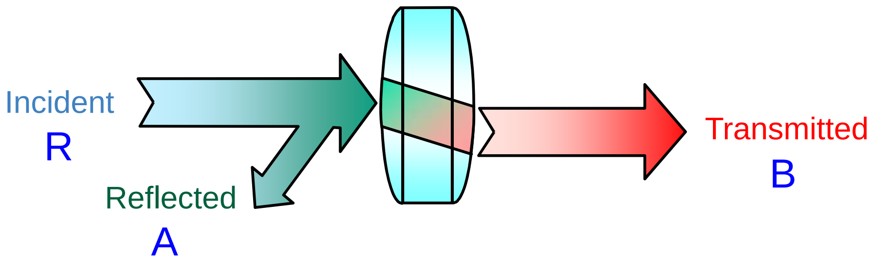
For reflection, a transmission line terminated in Zo behaves like an infinitely long transmission line

Transmission Line with Short & Open



For reflection, a transmission line terminated in a short or open reflects all power back to source

Reflection & Transmission



REFLECTION

$$\frac{\text{Reflected}}{\text{Incident}} = \frac{A}{R}$$

- SWR
- S-Parameters S_{11}, S_{22}
- Reflection Coefficient
- Impedance, Admittance R+jX, G+jB
- Return Loss

TRANSMISSION

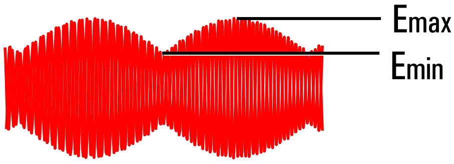
$$\frac{\text{Transmitted}}{\text{Incident}} = \frac{B}{R}$$

- Gain / Loss
- S-Parameters S_{21}, S_{12}
- Transmission Coefficient
- Insertion Phase
- Group Delay

Reflection Parameters

Reflection Coefficient $\Gamma = \frac{V_{reflected}}{V_{incident}} = \rho \angle \Phi = \frac{Z_L - Z_0}{Z_L + Z_0}$

Return loss = $-20 \log(\rho)$, $\rho = |\Gamma|$



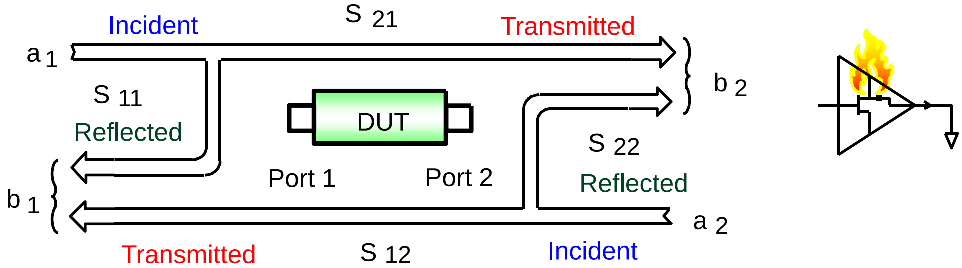
Voltage Standing Wave Ratio

VSWR $= \frac{E_{max}}{E_{min}} = \frac{1+\rho}{1-\rho}$

No reflection ***($Z_L = Zo$)***		***Full reflection*** ***(Z_L = open, short)***
0	**ρ**	1
∞ dB	**RL**	0 dB
1	**VSWR**	∞

Why S-Parameters?

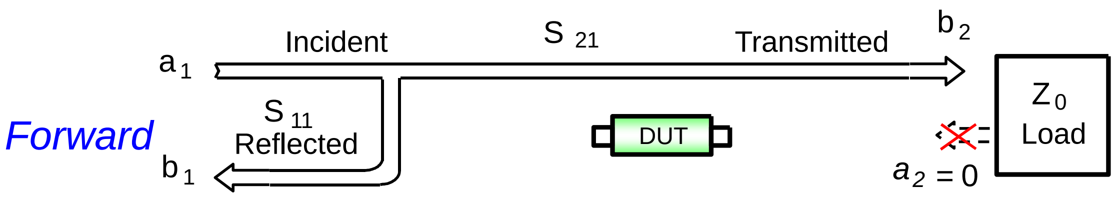
- Relatively easy to obtain at high frequencies
  - Measure voltage traveling waves with a vector network analyzer
  - Don't need shorts/opens which can cause active devices to oscillate or self-destruct
- Relate to familiar measurements (gain, loss, reflection coefficient ...)
- Can cascade S-parameters of multiple devices to predict system performance
- Can compute H, Y, or Z parameters from S-parameters if desired



$$b_1 = S_{11} a_1 + S_{12} a_2$$

$$b_2 = S_{21} a_1 + S_{22} a_2$$

Measuring S-Parameters?

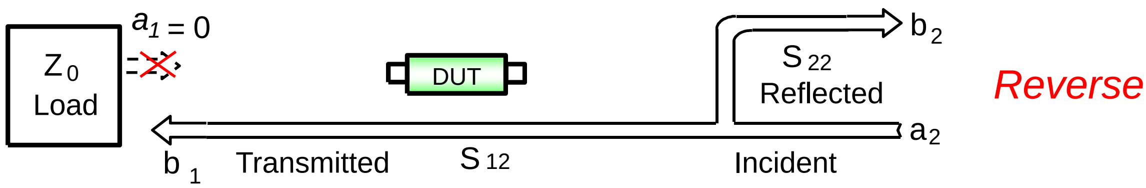


$$S_{11} = \frac{Reflected}{Incident} = \frac{b_1}{a_1}\Bigg|_{a_2 = 0}$$

$$S_{21} = \frac{Transmitted}{Incident} = \frac{b_2}{a_1}\Bigg|_{a_2 = 0}$$

$$S_{22} = \frac{Reflected}{Incident} = \frac{b_2}{a_2}\Bigg|_{a_1 = 0}$$

$$S_{12} = \frac{Transmitted}{Incident} = \frac{b_1}{a_2}\Bigg|_{a_1 = 0}$$



Summary

- Objective of Module-2
- Basic of Transmission Line
- Smith Chart
- Matching
- S-Parameters

