E6318 - Microwave Circuit Design

Columbia University

Spring 2006

Yves Baeyens

Outline of Lecture 3

- Recap of lecture 2 (transmission line theory)
- Transients in Transmission lines
  - demonstrate Bounce program
- Practical transmission lines:
  - Losses in transmission lines
  - Propagation: TEM, TE, TM (quasi-TEM)
    - Waveguide components
    - Coaxial lines
    - Stripline
    - Microstrip (Thin-Film microstrip)
    - Coplanar Waveguide
- Conclusions

Some announcements!

- Our class is taped for CVN, on-campus access to the CVN lecture is available through the following URL: http://ocstream.cvn.columbia.edu/index.asp?cn=ELENE6318
- Need to make up for last weeks lecture, please let Austin know your availability for make-up Class over next couple weeks

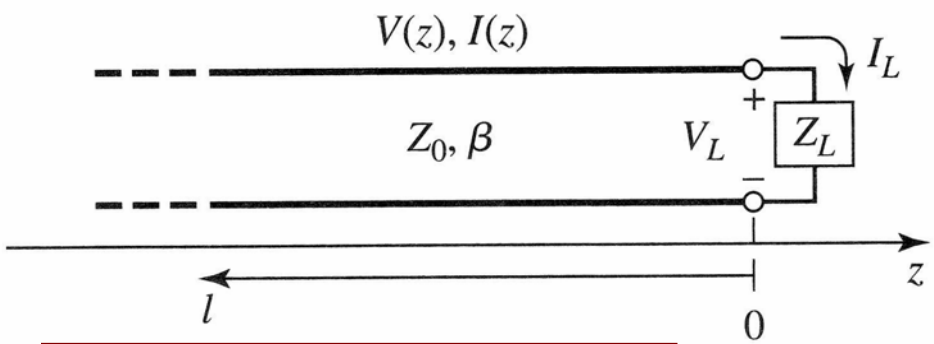
Calendar

- Course: Th 4:10-6:40 PM, 1127 Mudd
  - 01/19
  - 01/26
  - 02/02 to be rescheduled
  - 02/09
  - 02/16
  - 02/23
  - 03/02
  - 03/09 **Midterm**
  - 03/16 Spring Holidays
  - 03/23
  - 03/30
  - 04/06
  - 04/13
  - 04/20
  - 04/27
  - **Final** (05/11)

Recap transmission line theory (1)

$$V(z) = V_o^+ e^{-\gamma z} + V_o^- e^{+\gamma z}$$

$$I(z) = \frac{V_o^+}{Z_0} e^{-\gamma z} - \frac{V_o^-}{Z_0} e^{+\gamma z}$$



$$\boxed{Z_{in} = Z_0 \frac{Z_L + jZ_0 \tan \beta l}{Z_0 + jZ_L \tan \beta l}}$$

$$\boxed{\gamma = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)}}$$

$$\boxed{Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} = \frac{V_o^+}{I_0^+} = -\frac{V_o^-}{I_0^-}}$$

$$\boxed{\lambda = \frac{2\pi}{\beta}} \qquad \boxed{\Gamma = \frac{V_o^-}{V_o^+} = \frac{Z_L - Z_0}{Z_L + Z_0}}$$

$$\boxed{P_{av} = \frac{1}{2} \frac{|V_o^+|^2}{Z_0} \left(1 - |\Gamma|^2\right)}$$

$$\boxed{|V(z)| = |V_o^+| \left|1 + |\Gamma| e^{j(\theta - 2\beta l)}\right|}$$

$Z_{in} = jZ_0 \tan \beta l$ for short circuit, so inductive for $l<\lambda/4$

$Z_{in} = -jZ_0 \cot \beta l$ for open circuit, capacitive for $l<\lambda/4$

For lines with small loss ($R<<\omega L$, $G<<\omega C$)

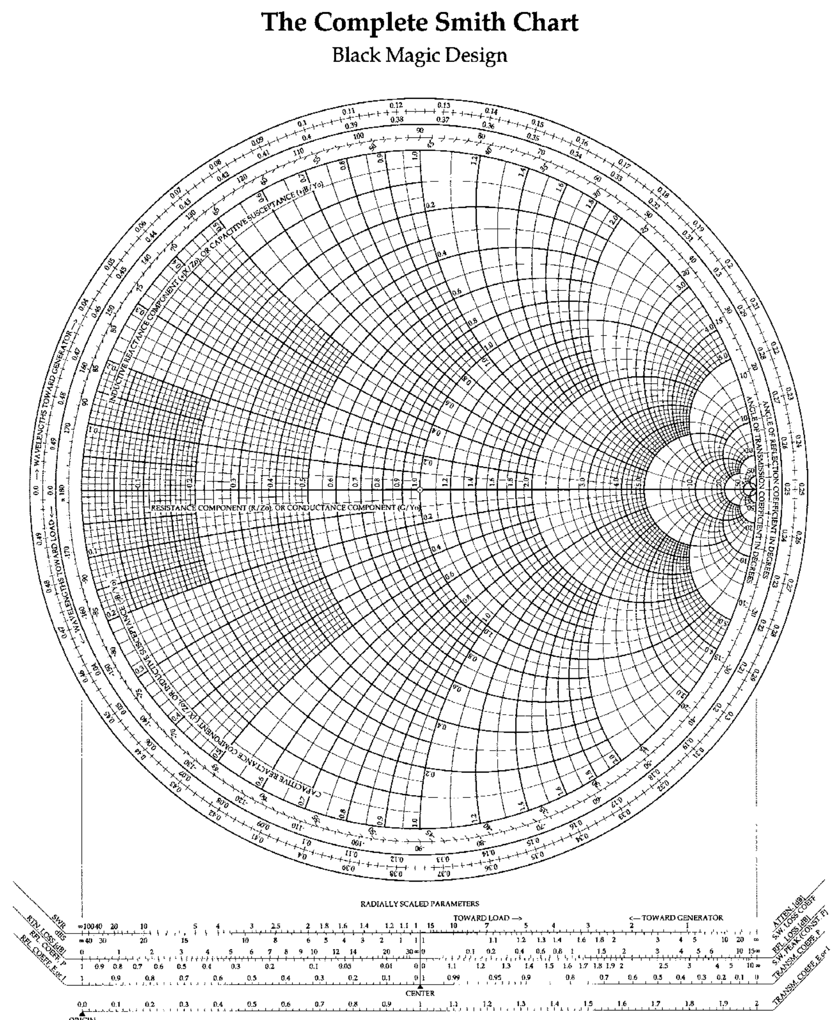
$$\boxed{\begin{aligned} \alpha &\cong \frac{1}{2}\left(R\sqrt{\frac{C}{L}} + G\sqrt{\frac{L}{C}}\right) = \frac{1}{2}\left(\frac{R}{Z_0} + GZ_0\right) \\ \beta &\cong \omega\sqrt{LC} \end{aligned}}$$

Recap transmission line theory (2)

Smith Chart: Polar plot of voltage reflection coefficient Γ

- Easy conversion from Γ to normalized impedances using impedance (or admittance) circles printed on chart
- Scales for VSWR, RL, etc..
- Graphical aid TL problems:
  - TL impedance equation: rotation
  - Conversion to admittance: imaging
- Only valid for lossless lines
- For lossy lines:

$$\boxed{\Gamma(l) = \Gamma(0) e^{-2j\beta l} e^{-2\alpha l}}$$



Recap transmission line theory (3)

Couple other observations in transmission lines:

- Lossless transmission lines will not absorb any power!
- Impedance transformation using transmission lines can help to increase power transfer from generator to load.
- Optimal power transfer when generator sees complex conjugate of its generator impedance
- Mismatched generator causes multiple reflections, can result in internal resonances
- Lossy lines will absorb power, both for incoming and for reflected wave
- Zero-reflection at input does not necessarily mean no internal standing waves, multiple reflection will cancel at interface

Lossy TL's – The distortionless TL (FYI)

General expression for γ implies that β may become a non-linear function of frequency. This results in frequency dependence for the phase velocity v_p, so for wideband signals different frequency components might travel at different velocities causing dispersion and distortion of signal.

Special case are lines with linear phase factor: distortionless, TL parameters satisfy: $\frac{R}{L} = \frac{G}{C}$

$$\gamma = j\omega\sqrt{LC}\sqrt{1 - 2j\frac{R}{\omega L} - \left(\frac{R}{\omega L}\right)^2} = j\omega\sqrt{LC}\left(1 - j\frac{R}{\omega L}\right) = \frac{R}{Z_0} + j\omega\sqrt{LC}$$

Result is line with β a linear function of frequency + attenuation is frequency independent: no distortion on pulses

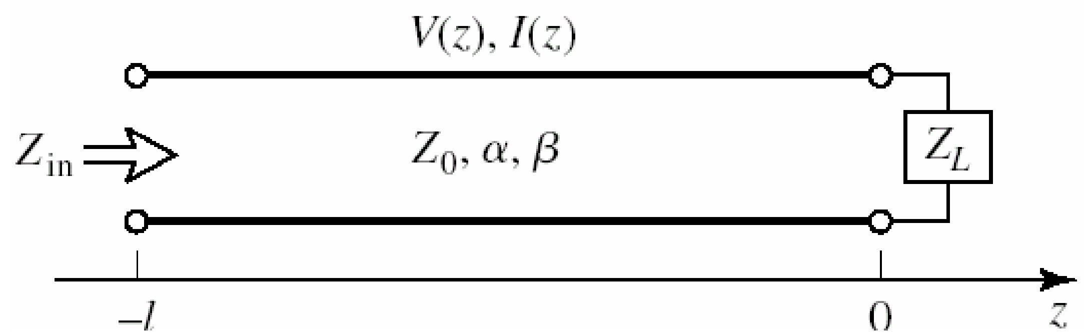
R/L=G/C often requires adding series inductance

The terminated lossy transmission line (FYI)

For lossy line with ~ real Z_0, expressions for voltage and current:

$$V(z) = V_o^+ \left[e^{-\gamma z} + \Gamma e^{+\gamma z} \right]$$

$$I(z) = \frac{V_o^+}{Z_0} \left[e^{-\gamma z} - \Gamma e^{+\gamma z} \right]$$



Z_{in} at distance l from load (in negative z-direction):

$$Z_{in} = \frac{V(-l)}{I(-l)} = Z_0 \frac{1 + \Gamma e^{-2\gamma l}}{1 - \Gamma e^{-2\gamma l}} = Z_0 \frac{Z_L + Z_0 \tanh \gamma l}{Z_0 + Z_L \tanh \gamma l}$$

Power lost in line:

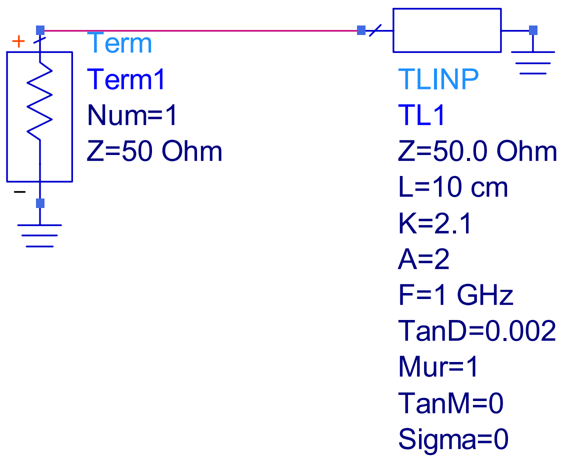
incident reflected

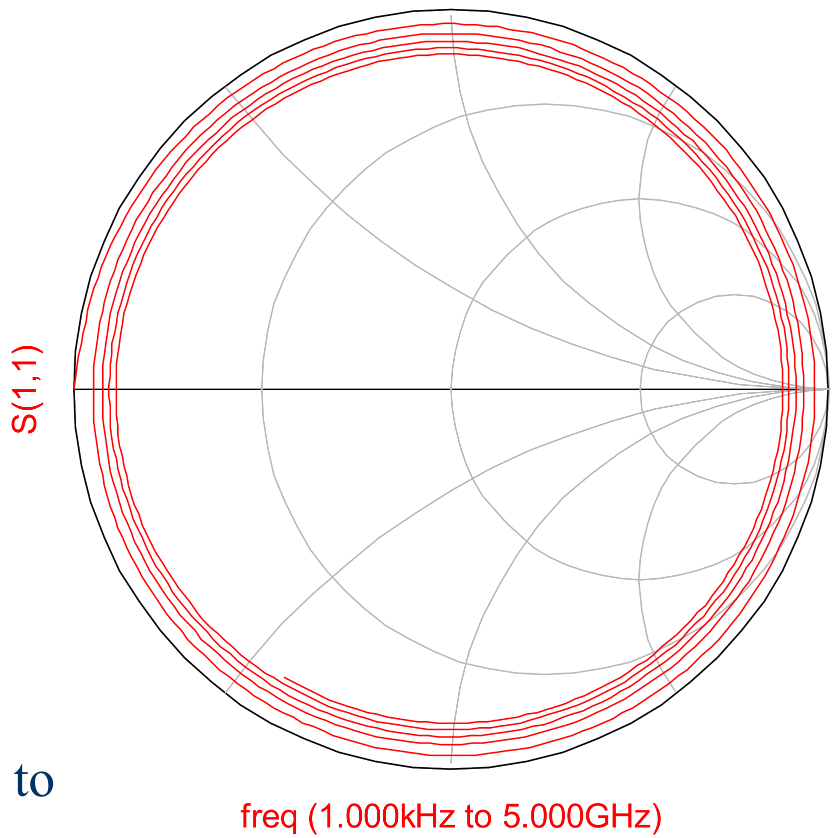
$$P_{loss} = P_{in} - P_L = \frac{\left|V_0^+\right|^2}{2 Z_0} \left(\left(e^{2\alpha l} - 1 \right) + |\Gamma|^2 \left(1 - e^{-2\alpha l} \right) \right)$$

Example of lossy TL terminated by short

$$\Gamma(l) = \Gamma e^{-2j\beta l} e^{-2\alpha l} = \Gamma e^{-2\gamma l}$$

ADS schematic lossy line terminated by short

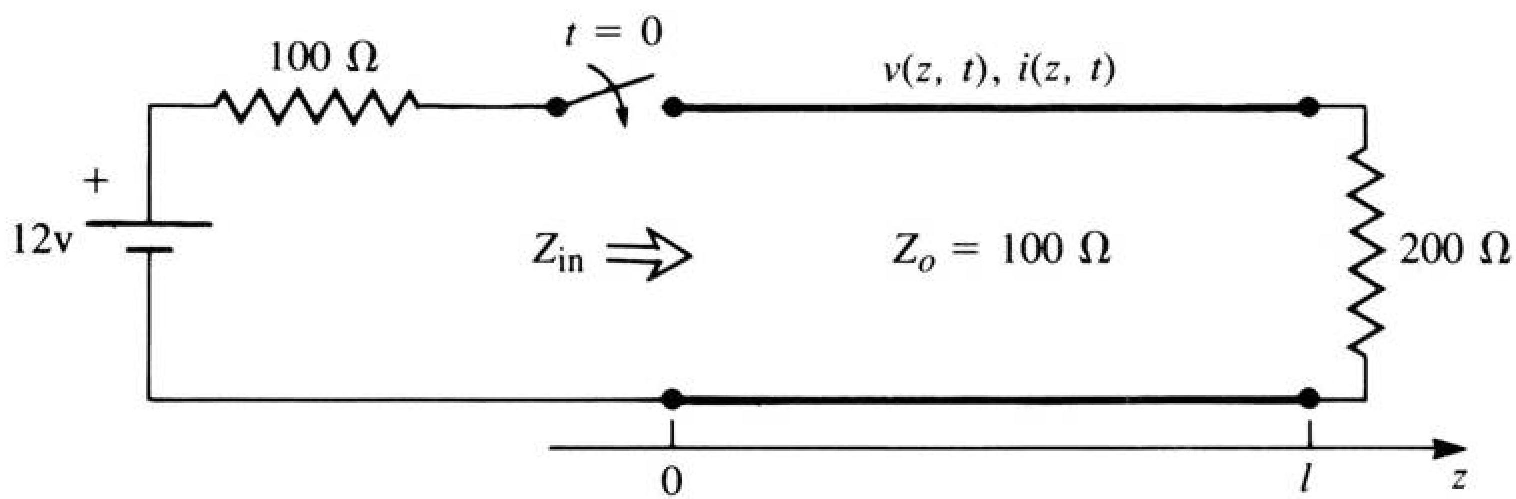




Result as function of frequency is circle spiraling towards center due to increased loss at higher frequency

Transients in Transmission Lines

- So far all calculations for sinusoidal signal at one fixed frequency, satisfactory for lot of microwave applications, but not for short pulse transmission
- Intuitive or rigorous (Laplace transform) approach for transient solution
- Example: 12-V DC source switched on at t=0



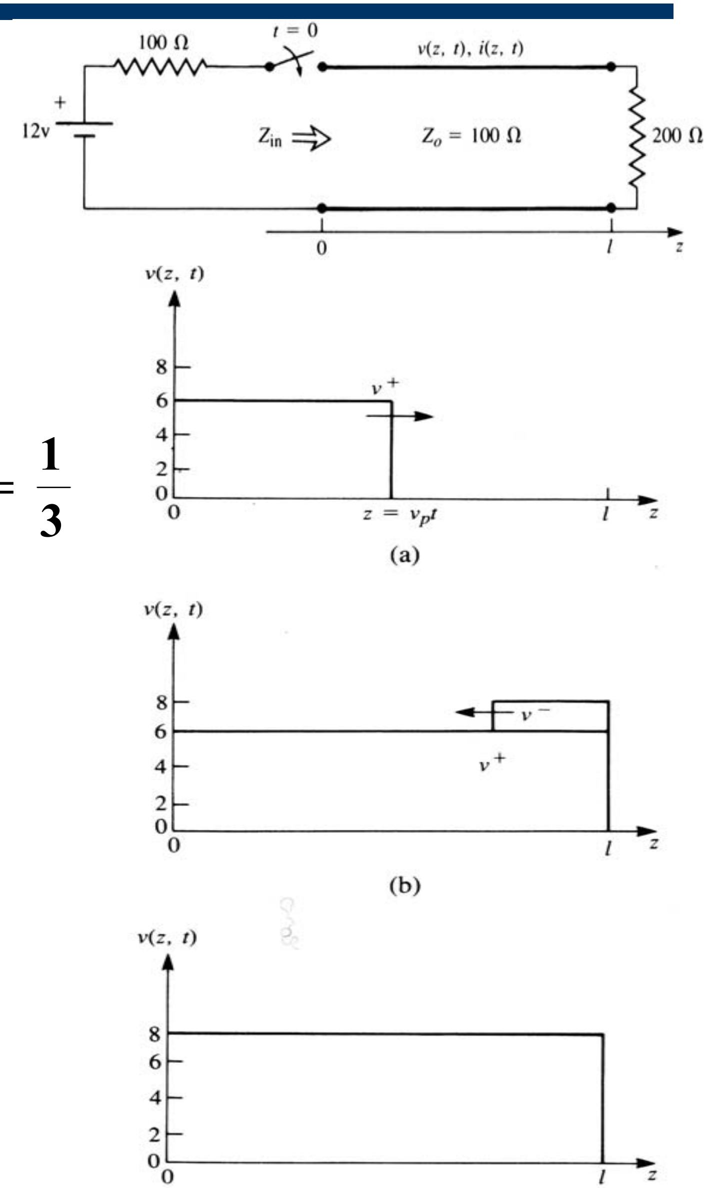
Transients in Transmission Lines: example

- ! At t=0, the generator does not see yet the 200 Ω load at end TL, looks at Z_0 of line, initially voltage division: 12V*(100/200) = 6V propagates with velocity v_p
- At $t=l/v_p$, pulse reaches load and gets partly reflected with coefficient of: $\Gamma_l = \dfrac{200 - 100}{200 + 100} = \dfrac{1}{3}$
- Superposition incident and reflected voltage:

$$v^+ + v^- = 6 + 6\Gamma_l = 8 \text{ V}$$

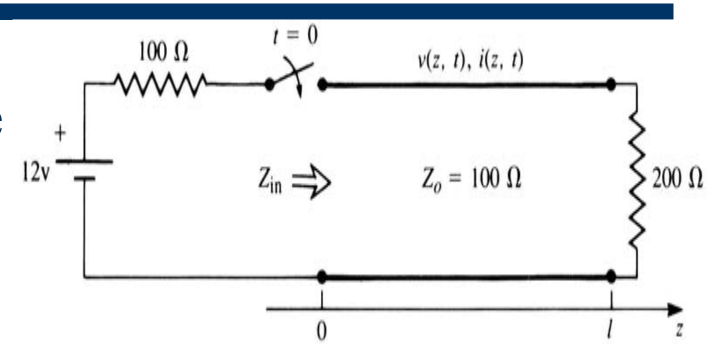
- At $t=2l/v_p$, pulse has made round trip absorbed by generator: 8V everywhere as can be expected from DC voltage division with 100 Ω generator and 200 Ω load:

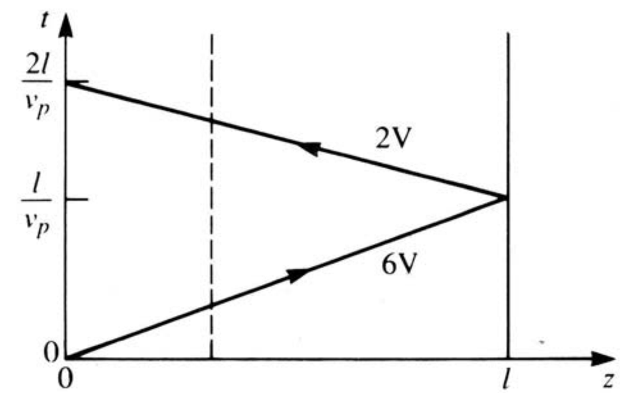
$$12 \frac{200}{200 + 100} = 8 \text{ V}$$



Transients in TL's: bounce diagram (FYI)

- Bounce diagram: way of viewing pulse propagating in time and position
  - X-axis: position along TL
  - Y-axis time
- Total voltage: draw line at position from $t=0..t_0$, total voltage adding voltages each component wave present (waves which intersect vertical line)

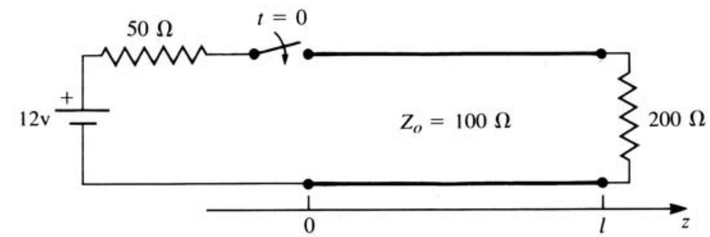




Transients in TL's: multiple reflection

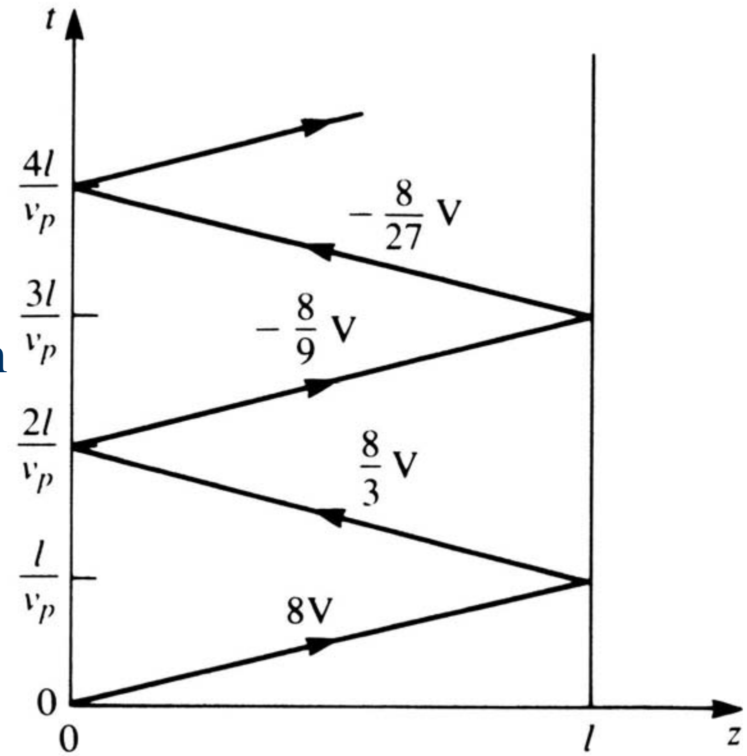
- Also generator mismatched to line, multiple reflections occur at load and generator

$$\Gamma_l = \frac{200 - 100}{200 + 100} = \frac{1}{3} \qquad \Gamma_g = \frac{50 - 100}{50 + 100} = -\frac{1}{3}$$



- At $t=l/v_p$, pulse reaches load and gets partly reflected with coefficient of 1/3
- At $t=2l/v_p$, pulse has made round trip gets reflected by generator:
- Finally, will settle at 9.6V everywhere as can be expected from DC voltage division with 50 Ω generator and 200 Ω load:

$$12 \frac{200}{200 + 50} = 9.6 \text{ V}$$



Transient and steady–state behaviour demo

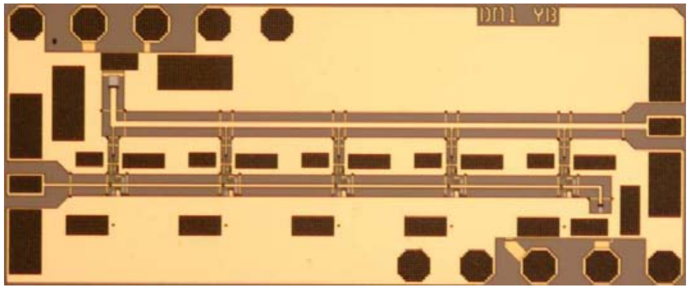
- A very nice program to **visualize transients in transmission lines,** called Bounce can be found on the website of Professor C.W. Trueman at Concordia University, Montreal. Please download the program bounce.exe on your PC and take a look at the examples. We'll do small demo of this program.
- Interesting set-ups:
  - Step-function along transmission line
  - Sine-wave on line with mismatched load: standing waves
  - Sine-wave on quarterwave-length transformer: internal reflections, but toal reflection dies out
  - Sine-wave in open or shorted transmission line with mismatched (low-impedance) generator: resonant behaviour!

Why using transmission lines at high-speed?

- Make longer connections which:
  - Are low-loss: low **ohmic** or **dielectric** losses + minimal **radiation**
  - Have well-known characteristics allowing to avoid degradation in analog or digital systems due to reflections, ringing, limited bandwidth, …
- Use the transmission lines as circuit element:
  - Reactive matching element to optimize power transfer, maximize gain of active elements, etc…
  - Frequency dependent properties: resonators, filters
  - Multiports with interesting power combining properties: couplers, hybrid-T's. etc…

Couple examples transmission line applications

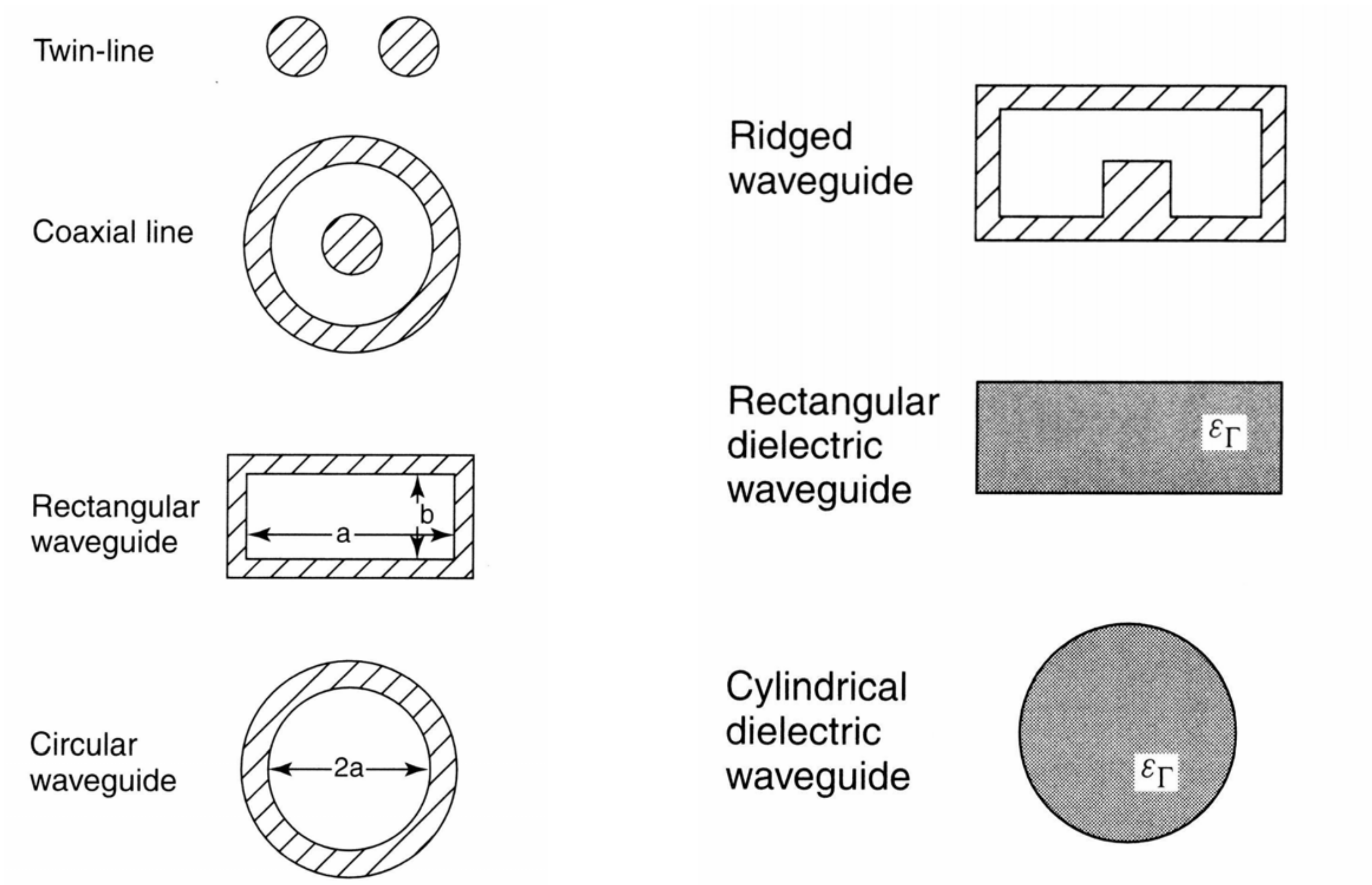
- For connections:
  - Telecom (phone, datacom, 1G ethernet) and cable TV networks: twisted pair, coax
  - High-speed backplanes, data-transmission, PCI-Express…
- Whole class of microwave circuits:
  - TL Filters (eg. couple line bandpass) and resonators (VCO)
  - Power Combiners/Dividers, hybrids, quadrature
  - Matching networks in µ-wave & mm-wave circuits
  - Distributed Amplifiers, mixers, analog circuits, etc…



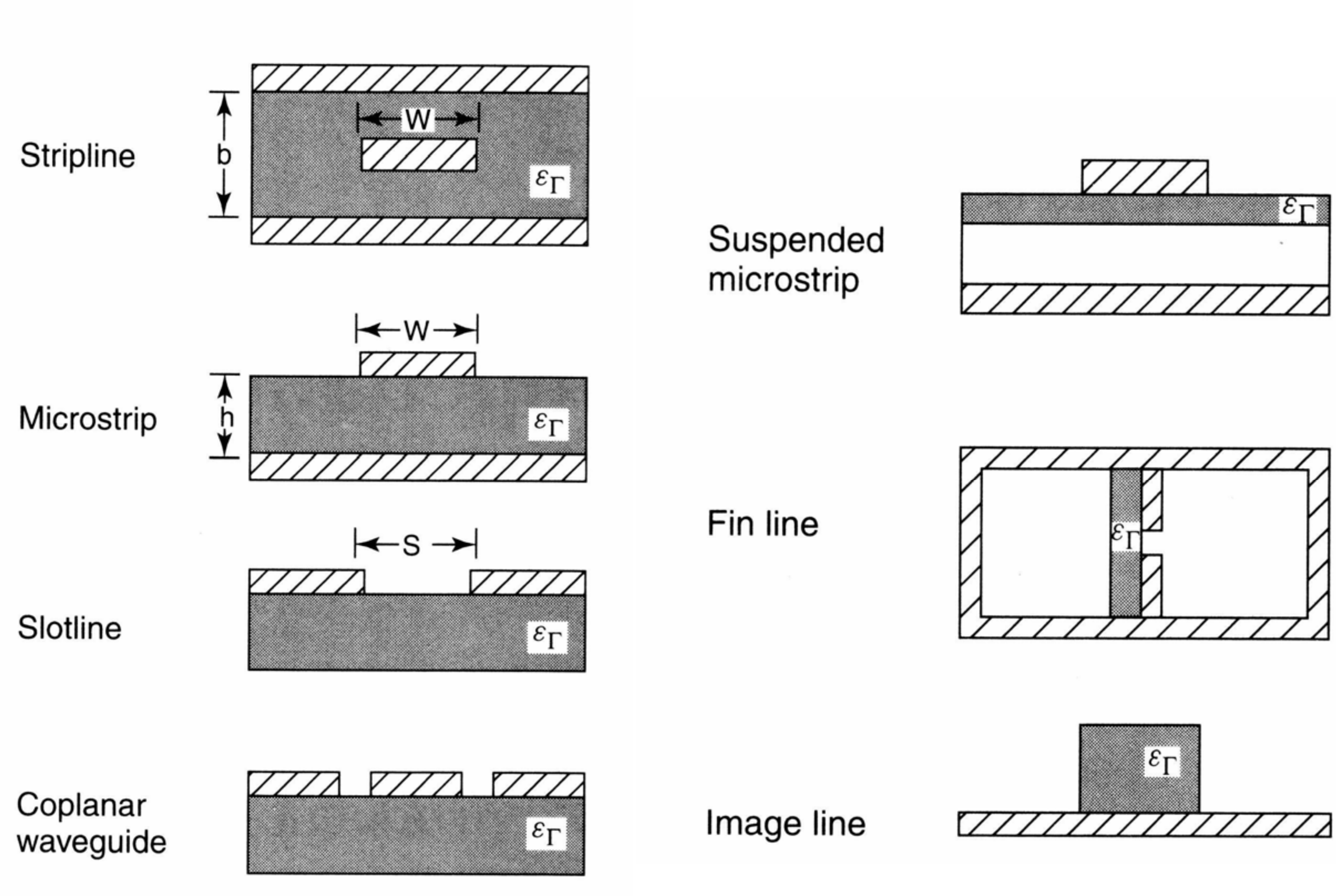
Trade-offs for choice suitable TL medium

- Electrical Properties
  - Impedance range
  - Line Loss (Q)
  - Frequency range (higher order modes)
  - Shielding performance (radiation losses, coupling)
  - Dispersion
  - Power Capability
  - Ease of inserting active or passive component
  - Balanced versus unbalanced lines
- Mechanical properties
  - Ease of fabrication, tolerance
  - Stability vs. environment
  - Flexibility
  - Weight, size and COST

Conventional transmission lines



Integrated planar transmission lines



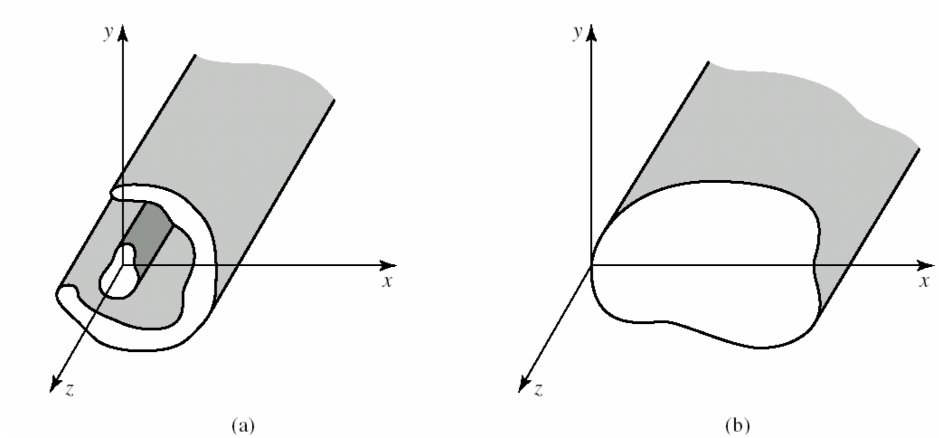
General solutions TEM, TE and TM

$$H_x = \frac{j}{k_c^2}\left(\omega\epsilon\frac{\partial E_z}{\partial y} - \beta\frac{\partial H_z}{\partial x}\right),$$

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

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

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



2-conductor closed waveguide

General solutions of Maxwell equations for 4 transverse fields in terms of E_z and H_z

$$k_c^2 = k^2 - \beta^2$$ **k_c=cutoff wavenumber**

$$k = \omega\sqrt{\mu\varepsilon} = 2\pi/\lambda$$ **k=wavenumber dielectric**

$$\varepsilon = \varepsilon_0\varepsilon_r(1 - j\tan\delta)$$ **tan δ=loss tangent dielectric**

TEM or TE-TM??

- TEM: $\beta = \omega\sqrt{\pi\varepsilon} = k$ $\quad k_c = 0$
  - no longitudinal field components ($E_z = H_z = 0$): k_c needs to be 0
  - from Maxwell's equations: for TEM, transverse E and H fields satisfy Laplace's equation: the transverse fields of TEM wave are same as static fields; no low frequency cut-off
  - 2 or more conductors needed, no TEM in closed conductor
  - voltage, current and impedance well-defined

$$V_{12} = \Phi_1 - \Phi_2 = \int_1^2 \overline{E} \cdot d\bar{l}$$

$$I = \oint_C \overline{H} \cdot d\bar{l} \qquad \text{Ampere's law}$$

  - examples TEM: parallel plate waveguide, coaxial, stripline,…

TE and TM

- TE (H-waves) or TM (E-waves):
  - Only longitudinal H-field component (E_z=0 & $H_z \neq 0$) for TE
  - Only longitudinal E-field component (H_z=0 & $E_z \neq 0$) for TM
  - $k_c \neq 0$ and propagation constant β dependent frequency and TL geometry

$$\beta = \sqrt{k^2 - k_c^2}$$

  - supported in closed conductors or with 2 or more conductors
  - examples TE/TM: rectangular or circular waveguide, higher order modes in coaxial, stripline,…

Attenuation in TL's

$$\boxed{\alpha = \alpha_d + \alpha_c}$$

α_d=attenuation due to dielectric loss and α_c due to conductor loss

- α_c using perturbation method (not seen here), depends on field distribution in guide, needs to be calculated for each geometry, typically expressed in function of surface resistance: R_s, frequency dependent ($\sim\sqrt{f}$), related with skin effect

$$\boxed{R_s = \frac{1}{\sigma\delta_s} = \sqrt{\frac{\omega\mu}{2\sigma}}}$$

- α_d for line filled with homogeneous dielectric can be calculated from propagation constant (see p. 97), is typically $\sim f$

$$\boxed{\begin{aligned}\alpha_d &\cong \frac{k^2 \tan\delta}{2\beta}\ \text{Np/m} \quad (\text{TE or TM waves}) \\ &\cong \frac{k \tan\delta}{2}\ \text{Np/m} \quad (\text{TEM waves})\end{aligned}}$$

Skin effect

- At LF’s, most of conductor’s cross section carries current.
- As frequency increases, current moves to skin of conductor.
- Skin depth is depth at which conductor’s current is reduced to 1/e = 37% of surface value: $\boxed{\delta_s = 1/\sqrt{\pi f \mu \sigma}}$

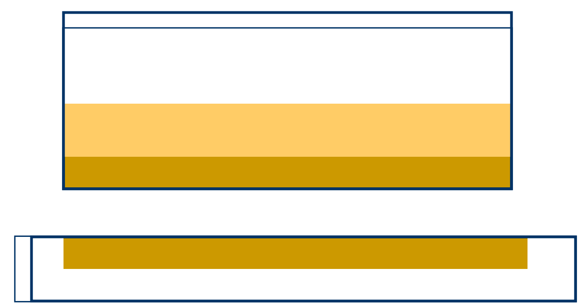
Isolated conductor:

Low frequency

High frequency

Conductor and ground:

Low frequency



High frequency

Waveguides (TE or TM)

- Conventional transmission lines can have either too much loss or too little power capability at microwave frequencies.
- Hollow waveguides present an alternative.
- Electromagnetic waves reflect from the walls of the wave guide as it travels its length.
- Waveguide typically made out of brass, Al, can be Ag plated,
- Rectangular, elliptical & circular topology
- No radiation losses as E and H fields are contained.
- Dielectric losses are small (air)
- Minimal losses in conductive walls

Modes in rectangular waveguides

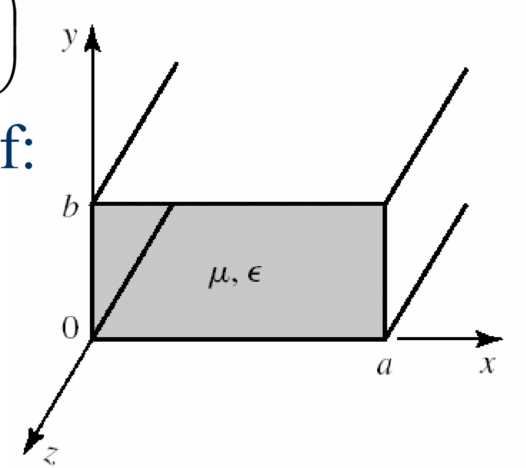
- Both TE and TM, no TEM
- More then one mode propagating: overmoded waveguide
- Single-mode operation is achieved by using the mode with the lowest cutoff frequency. (dominant mode)
- Waveguide is used between its cutoff frequency and that of the mode of the next lowest cutoff frequency.
- For rectangular waveguide propagation constant:

$$\beta = \sqrt{k^2 - k_c^2} = \sqrt{k^2 - \left(\frac{m\pi}{a}\right)^2 - \left(\frac{n\pi}{b}\right)^2}$$

- Is real, corresponding to propagating wave, if:

$$k > k_c = \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2}$$

m=# variations in x, n:# variations y



Modes in rectangular waveguides

- Each mode has cut-off frequency $f_{cm,n}$ given by:

$$f_{c\,mn} = \frac{k_c}{2\pi\sqrt{\mu\varepsilon}} = \frac{1}{2\pi\sqrt{\mu\varepsilon}}\sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2}$$

- For rectangular waveguide (a>b) dominant mode is TE_{10}:

$$f_{c_{10}} = \frac{1}{2a\sqrt{\mu\varepsilon}} \qquad \beta = \sqrt{k^2 - (\pi/a)^2}$$

- For $f<f_c$, β is imaginary, all field components will decay exponentially: cut-off or evanescent modes
- Wave impedance (relates transverse electric & magnetic fields)

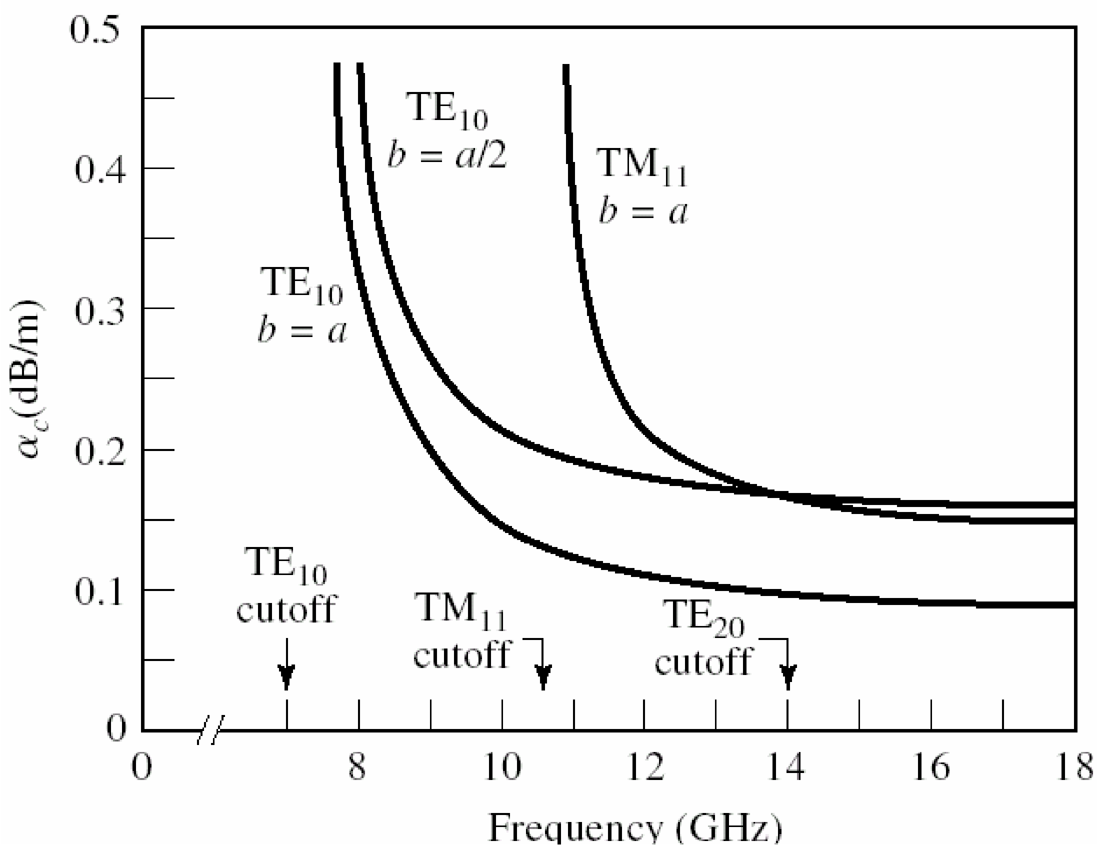
$$Z_{TE} = \frac{E_x}{H_y} = -\frac{E_y}{H_x} = \frac{k}{\beta}\sqrt{\frac{\mu}{\varepsilon}} = \frac{k}{\beta}\eta$$

- TM modes: same formulas for β, however lowest possible TM_{11}

$$f_{c_{11}} = \frac{1}{2\pi\sqrt{\mu\varepsilon}}\sqrt{\left(\frac{\pi}{a}\right)^2 + \left(\frac{\pi}{b}\right)^2}$$

Attenuation in rectangular waveguides

Attenuation various modes rectangular brass waveguide (a=2cm)



Attenuation due to conductor loss $$\alpha_c = \frac{R_s}{a^3 b \beta k \eta}\left(2b\pi^2 + a^3 k^2\right)$$

Example: X-band waveguide*

Consider a length of air-filled copper X-band waveguide, with dimensions $a = 2.286$ cm, $b = 1.016$ cm. Find the cutoff frequencies of the first four propagating modes. What is the attenuation in dB of a 1 m length of this guide when operating at $f = 10$ GHz?

Solution

From (3.84), the cutoff frequencies are given by

$$f_{c_{mn}} = \frac{c}{2\pi}\sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2}.$$

Computing f_c for the first few values of m and n gives:

Mode	m	n	$f_{c_{mn}}$(GHz)
TE	1	0	6.562
TE	2	0	13.123
TE	0	1	14.764
TE, TM	1	1	16.156
TE, TM	1	2	30.248
TE, TM	2	1	19.753

** You will find different example, Teflon filled, in Pozar, Ed. III*

Example: X-band waveguide

Thus the TE_{10}, TE_{20}, TE_{01}, and TE_{11} modes will be the first four modes to propagate (the TE_{11} and TM_{11} modes have the same cutoff frequency).

At 10 GHz, $k = 209.44\ \text{m}^{-1}$, and the propagation constant of the TE_{10} mode (the only propagating mode) is

$$\beta = \sqrt{k^2 - \left(\frac{\pi}{a}\right)^2} = \sqrt{\left(\frac{2\pi f}{c}\right)^2 - \left(\frac{\pi}{a}\right)^2} = 158.05\ \text{m}^{-1}.$$

The surface resistivity of the copper walls is ($\sigma = 5.8 \times 10^7 \text{S/m}$)

$$R_s = \sqrt{\frac{\omega\mu}{2\sigma}} = 0.026\ \Omega,$$

so the attenuation constant, from (3.96), is

$$\alpha_c = \frac{R_s}{a^3 b\beta k\eta}(2b\pi^2 + a^3k^2) = 0.0125\ \text{Np/m},$$

$$\alpha_c(\text{dB}) = -20\log e^{-\alpha_c} = 0.11\ \text{dB/m},$$

Rectangular Waveguide data (appendix I)

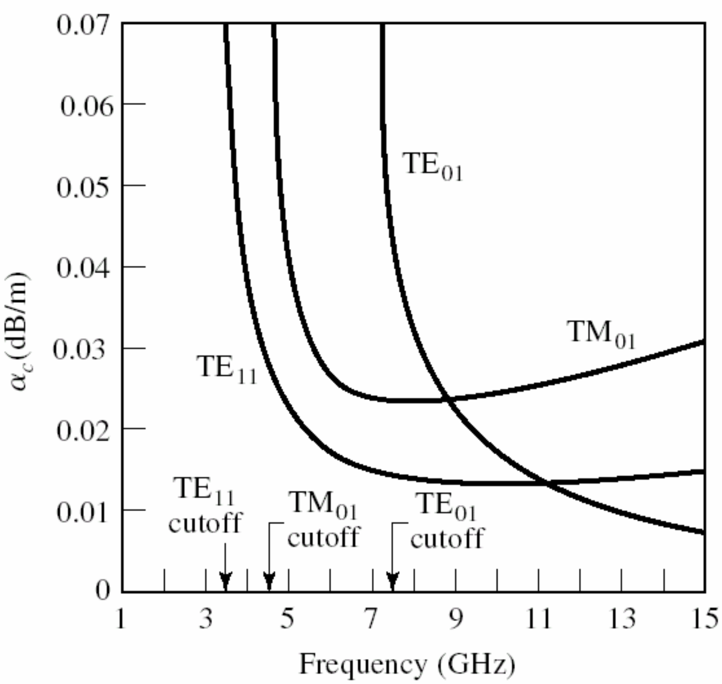
Band*	Recommended Frequency Range (GHz)	TE_{10} Cutoff Frequency (GHz)	EIA Designation WR-XX	Inside Dimensions Inches (cm)
L	1.12–1.70	0.908	WR-650	6.500×3.250 (16.51×8.255)
R	1.70–2.60	1.372	WR-430	4.300×2.150 (10.922×5.461)
S	2.60–3.95	2.078	WR-284	2.840×1.340 (7.214×3.404)
H (G)	3.95–5.85	3.152	WR-187	1.872×0.872 (4.755×2.215)
C (J)	5.85–8.20	4.301	WR-137	1.372×0.622 (3.485×1.580)
W (H)	7.05–10.0	5.259	WR-112	1.122×0.497 (2.850×1.262)
X	8.20–12.4	6.557	WR-90	0.900×0.400 (2.286×1.016)
Ku (P)	12.4–18.0	9.486	WR-62	0.622×0.311 (1.580×0.790)
K	18.0–26.5	14.047	WR-42	0.420×0.170 (1.07 ×0.43)
Ka (R)	26.5–40.0	21.081	WR-28	0.280×0.140 (0.711×0.356)
Q	33.0–50.5	26.342	WR-22	0.224×0.112 (0.57 ×0.28)
U	40.0–60.0	31.357	WR-19	0.188×0.094 (0.48 ×0.24)
V	50.0–75.0	39.863	WR-15	0.148×0.074 (0.38 ×0.19)
E	60.0–90.0	48.350	WR-12	0.122×0.061 (0.31 ×0.015)
W	75.0–110.0	59.010	WR-10	0.100×0.050 (0.254×0.127)
F	90.0–140.0	73.840	WR-8	0.080×0.040 (0.203×0.102)
D	110.0–170.0	90.854	WR-6	0.065×0.0325 (0.170×0.083)
G	140.0–220.0	115.750	WR-5	0.051×0.0255 (0.130×0.0648)

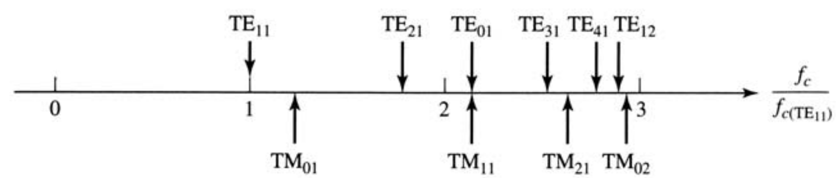
Especially at lower frequencies, waveguides have large size & weight + difficult in integration: replaced by microstrip or coaxial, except for very high Q passives or very high power: e.g. WR-90 (X-band): CW power rating 200-290 kW!, loss: 5.49-3.383 dB/100 ft (Al)

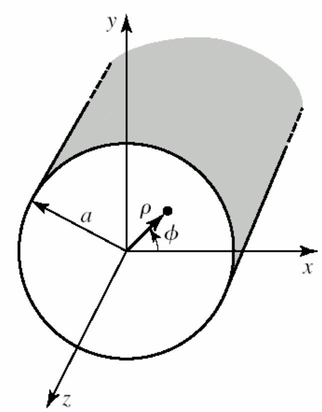
Other application: ultra-high-frequency: available up to 2.2 THz

Circular Waveguide

- Again TM and TE possible, calculation done analytically using Bessel functions…
- Circular waveguide can be very low loss
- Main problem is the close separation of different modes, reducing usable bandwidth (~25%)

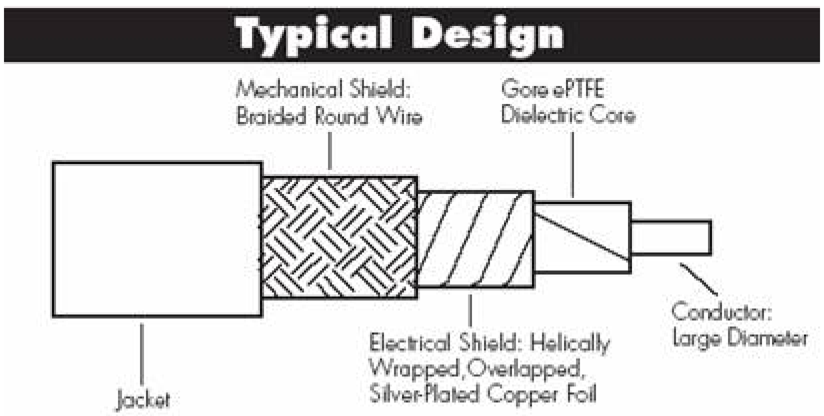






Coaxial Lines

- Shielded line common for high-frequency signal transmission
- Applications in cable TV, video transmission cables, interconnection, measurements
- Normally Z_0=50 Ohm, except for some lowest loss applications with Z_0=75 Ohm

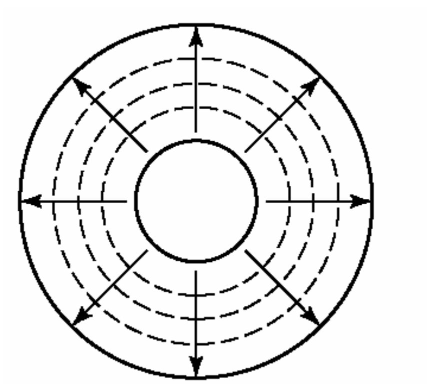


TEM & higher order modes in coaxial lines

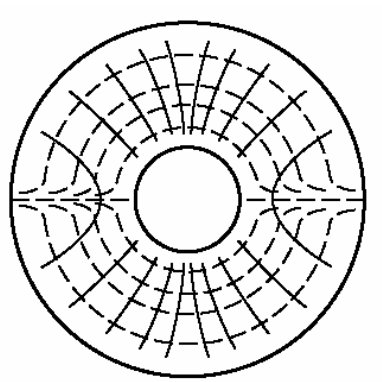
- Coaxial line normally operates in TEM mode
- No cut-off frequency TEM: use from DC-to mm-waves
- EM-fields from static fields using cylindrical coordinates
- Can support waveguide modes, dominant is TE_{11}, cutoff frequency calculation numerically, approximation

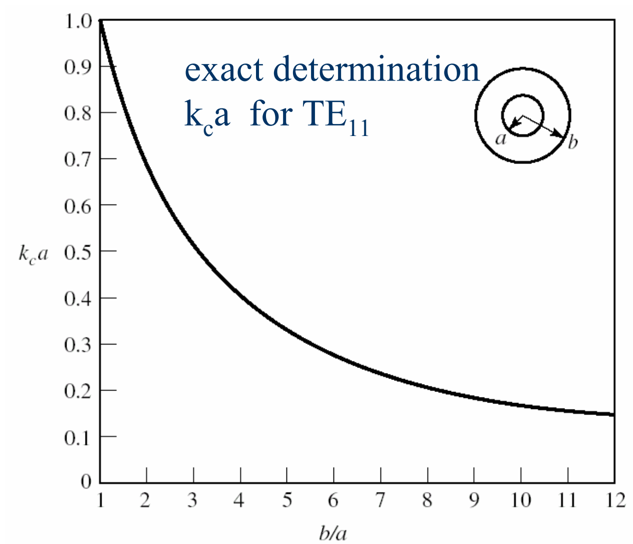
$$k_c \cong \frac{2}{a+b} \qquad f_c = \frac{c k_c}{2\pi\sqrt{\varepsilon_r}}$$

TEM



TE_{11}





Coaxial line connectors

Overview of coaxial connectors

Connector	Metrology	Instrument	Production	Cutoff Freq (GHz)	Sexed	Precision Slotted Connector
Type F(75)	N	N	Y	1	Y	N
BNC (50 & 75)	N	N	Y	2	Y	N
SMC	N	Y	N	7	Y	N
Type-N (50 & 75)	Y	Y	Y	18	Y	Y
APC-7 or 7 mm	Y	Y	Y	18	N	N
SMA (4.14mm)	N	N	Y	22	Y	N
3.55 mm	Y	Y	Y	34	Y	Y
2.92 mm or "K" [1]	N	Y	Y	44	Y	N
2.4 mm [2]	Y	Y	Y	52	Y	Y
1.85 mm [2, 3]	N	Y	Y	70	Y	N
1.0 mm	N	Y	Y	110	Y	N

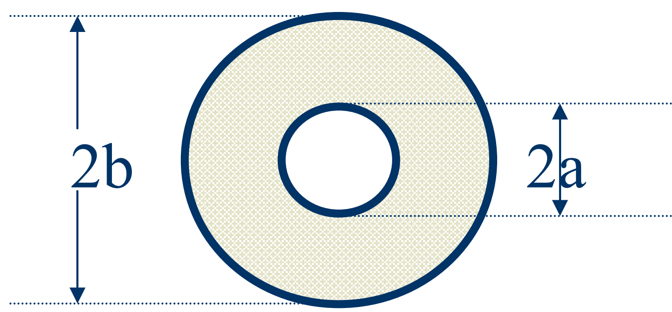
Paul Neill

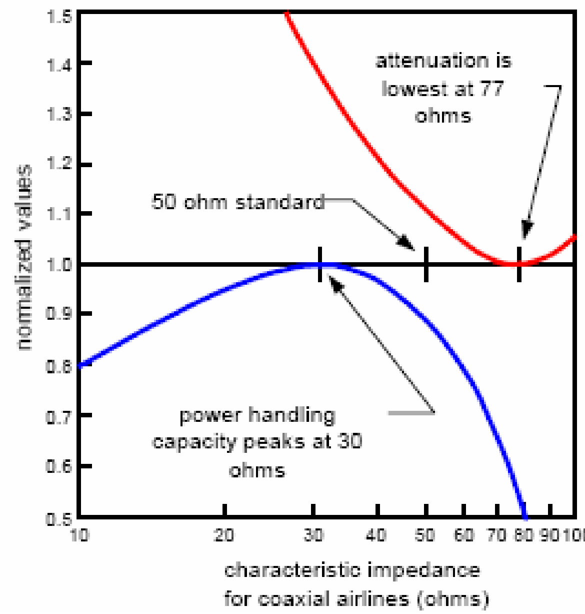
1. Compatible with SMA and 3.5 mm connectors.
2. Not compatible with SMA, 3.5 or 2.92 mm connectors
3. Compatible with 2.4 mm connector

Why 50 Ω ??

- Optimum between lowest loss and power handling airline

$$Z_0 = 138\log_{10}\left(\frac{b}{a}\right)$$



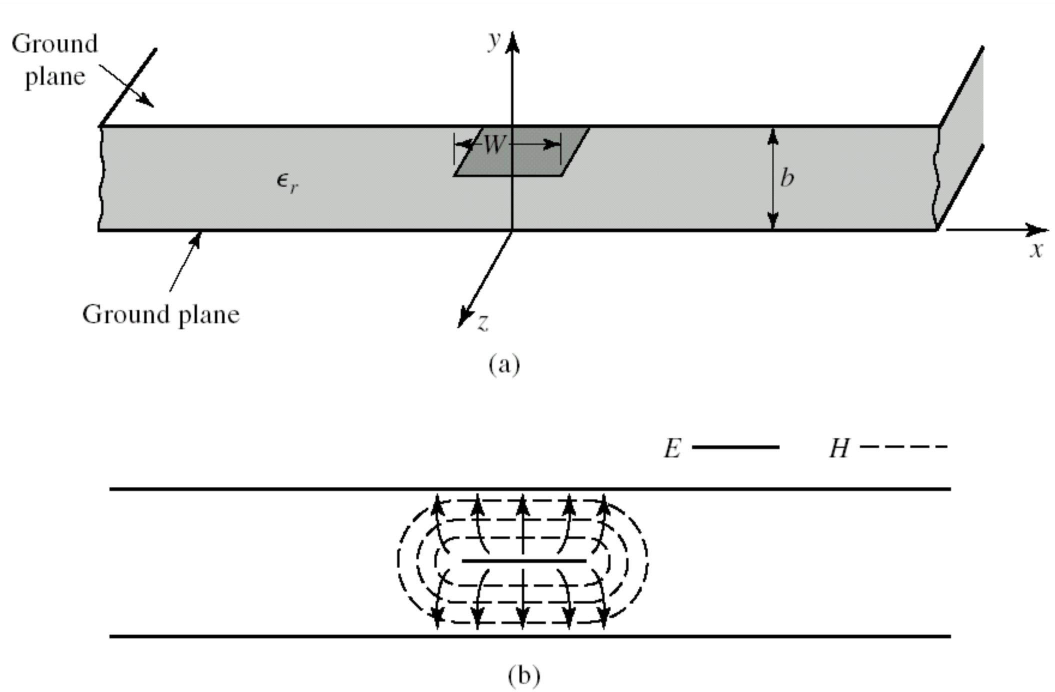


$$\textbf{lowest attenuation } \frac{b}{a} = 3.6\,; \quad Z_0 = 77.6\Omega$$

$$\textbf{optimum power } \frac{b}{a} = 1.65\,; \quad Z_0 = 30\Omega$$

Strip-lines

- Planar transmission line supporting TEM mode (usual mode of operation)



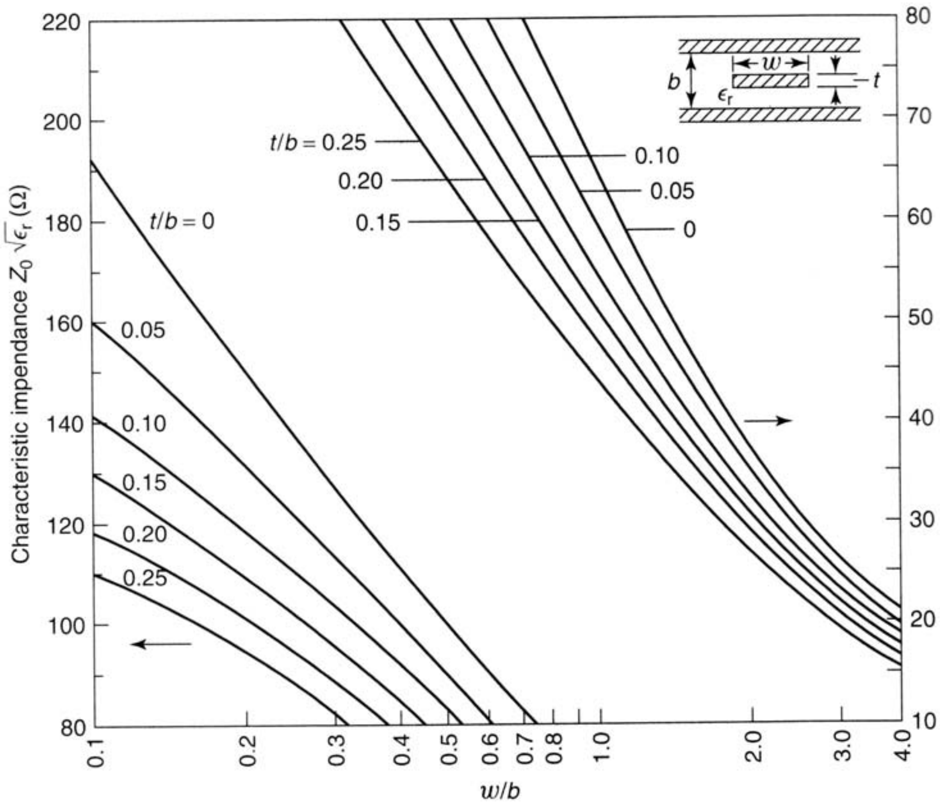
thin strip with width W between conducting grounds with separation b

- Also TE and TM modes are possible suppressed with shorting screws between ground and restricting $b<\lambda/4$

Characteristics of Striplines

TEM line: $v_p = 1/\sqrt{\mu_0\varepsilon_0\varepsilon_r} = c/\sqrt{\varepsilon_r}$ $\quad \beta = \dfrac{\omega}{v_p} = \omega\sqrt{\mu_0\varepsilon_0\varepsilon_r} = \sqrt{\varepsilon_r}\,k_0$

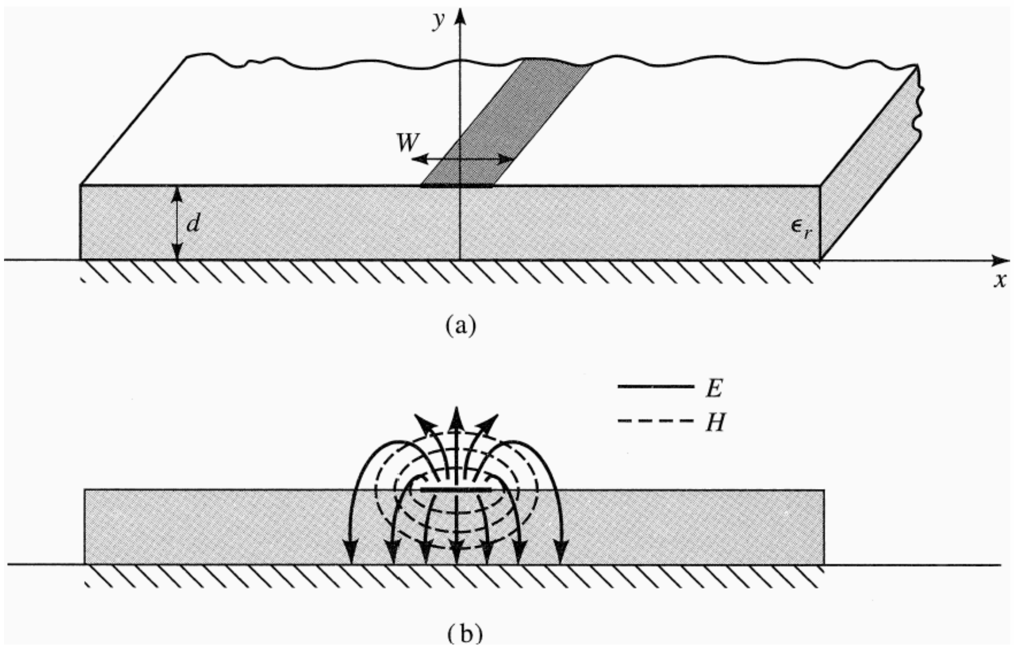
Other characteristics: conformal mapping needed, results fitted to simple formulas (see Pozar, p.139)



K. Chang, I. Bahl and V. Nair, "RF and Microwave Circuit and Component Design for Wireless Systems", Wiley Series in Microwave and Optical Engineering

Microstrip lines

- most common TL in MIC's & MMIC's, easily defined using lithography, easily integrates passive & active components
- does not support pure TEM (different ε_r under and above line), still E and H-field in direction propagation negligible compared with TEM fields: quasi-TEM



thin strip with width W printed on thin grounded dielectric substrate with height d

Characteristics of Microstrip lines

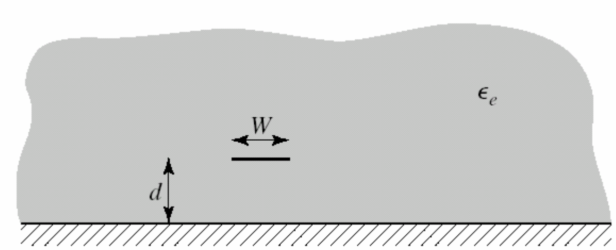
- quasi TEM line: most field lines in dielectric region concentrated between strip and ground, some in air
- phase velocity different in air and dielectric, no phase match possible for TEM: hybrid TE-TM, quasi-TEM for d<<λ, good approximations v_p, γ,Z_0 from static case

$$\beta = \frac{\omega}{v_p} = \omega\sqrt{\mu_0 \varepsilon_0 \varepsilon_e} = \sqrt{\varepsilon_e} k_0 \quad v_p = c / \sqrt{\varepsilon_e}$$

- ε_e is effective dielectric constant, dependent on d & W, t

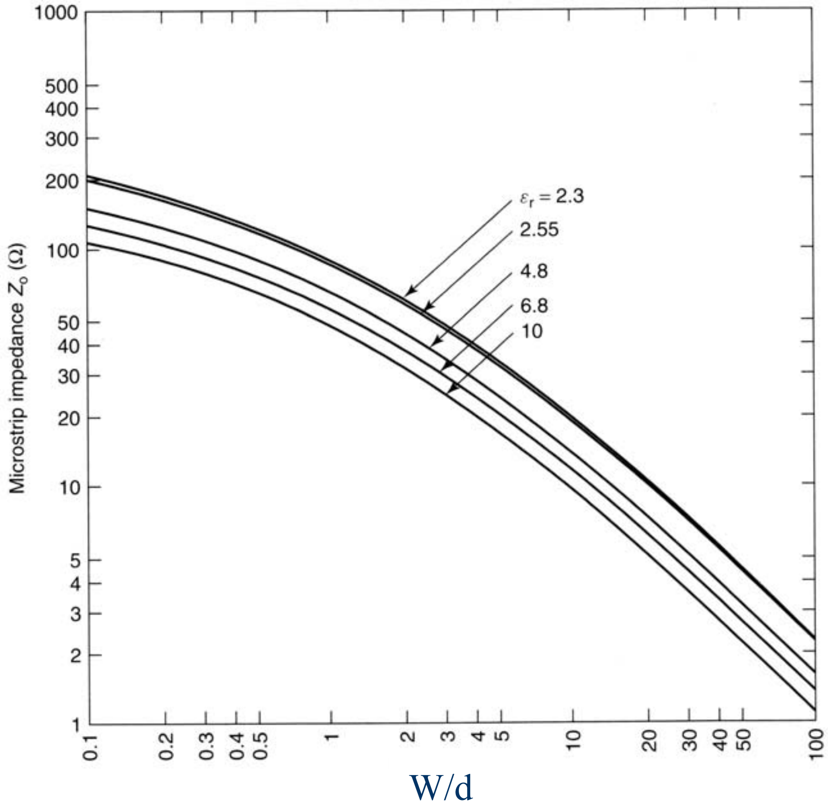
$$1 < \varepsilon_e < \varepsilon_r$$

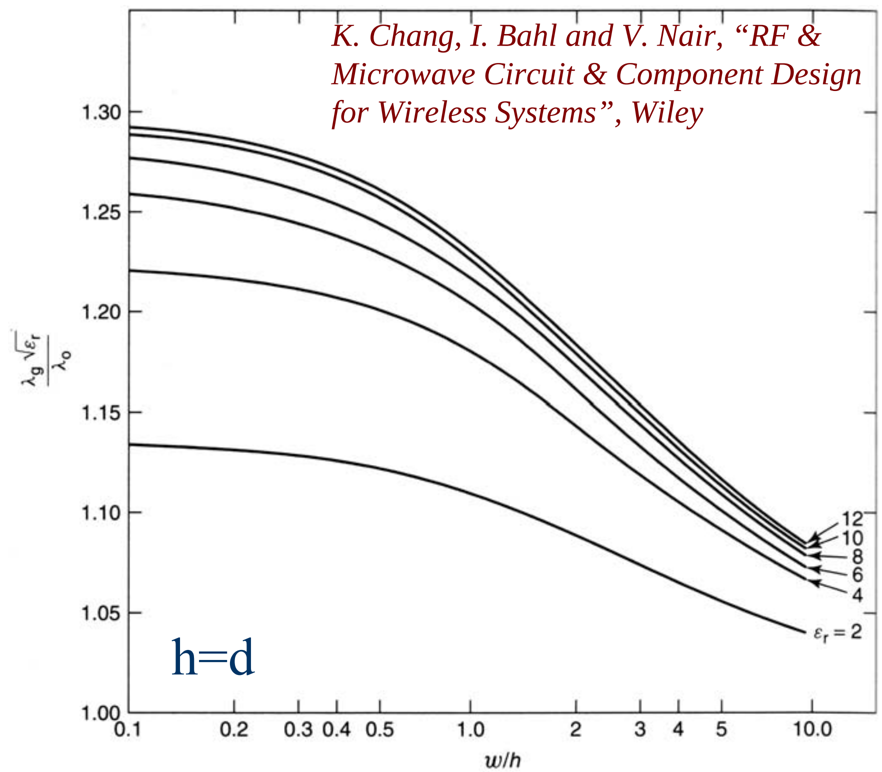
$$\varepsilon_e \cong \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \frac{1}{\sqrt{1 + 12\, d / W}}$$



Using ε_e approximate formulas for Z_0 & α_d can be calculated (see Pozar p. 145)

Calculated characteristic impedance





- Typical microwave simulators (ADS) have very rigorous formulas for μ-strip characteristics built-in (also effect t)
- Multilayer stacks in real IC's might complicate calculation

Example of microstrip calculations

EXAMPLE 3.7 **Microstrip Design**

Calculate the width and length of a microstrip line for a 50 Ω characteristic impedance and a 90° phase shift at 2.5 GHz. The substrate thickness is $d = 0.127$ cm, with $\epsilon_r = 2.20$.

Solution

We first find W/d for $Z_0 = 50\ \Omega$, and initially guess that $W/d > 2$. From (3.197),

$$B = 7.985, \qquad W/d = 3.081.$$

So $W/d > 2$; otherwise we would use the expression for $W/d < 2$. Then $W = 3.081d = 0.391$ cm. From (3.195) the effective dielectric constant is

$$\epsilon_e = 1.87.$$

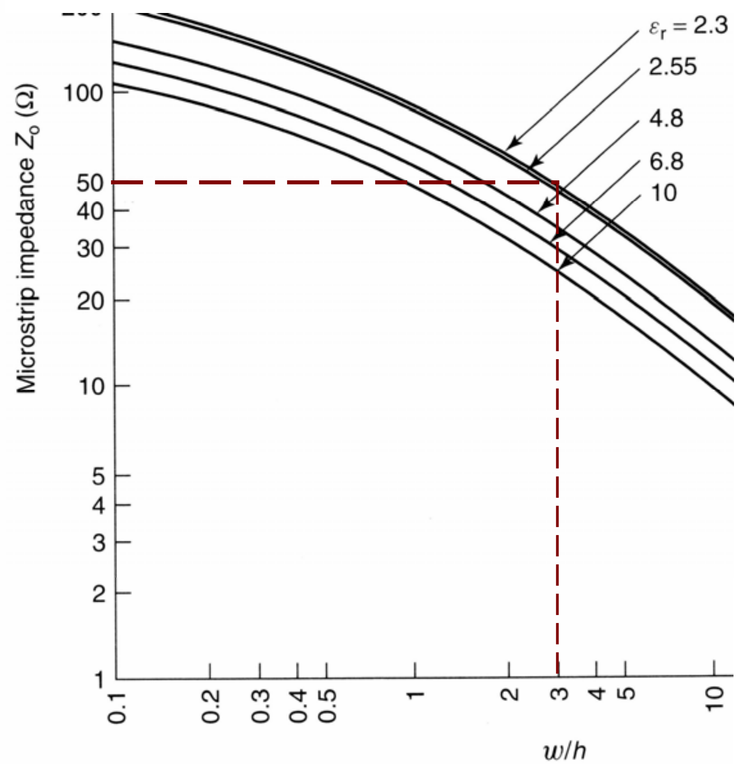
The line length, ℓ, for a 90° phase shift is found as

$$\phi = 90° = \beta\ell = \sqrt{\epsilon_e}k_0\ell,$$

$$k_0 = \frac{2\pi f}{c} = 52.35\ \mathrm{m}^{-1},$$

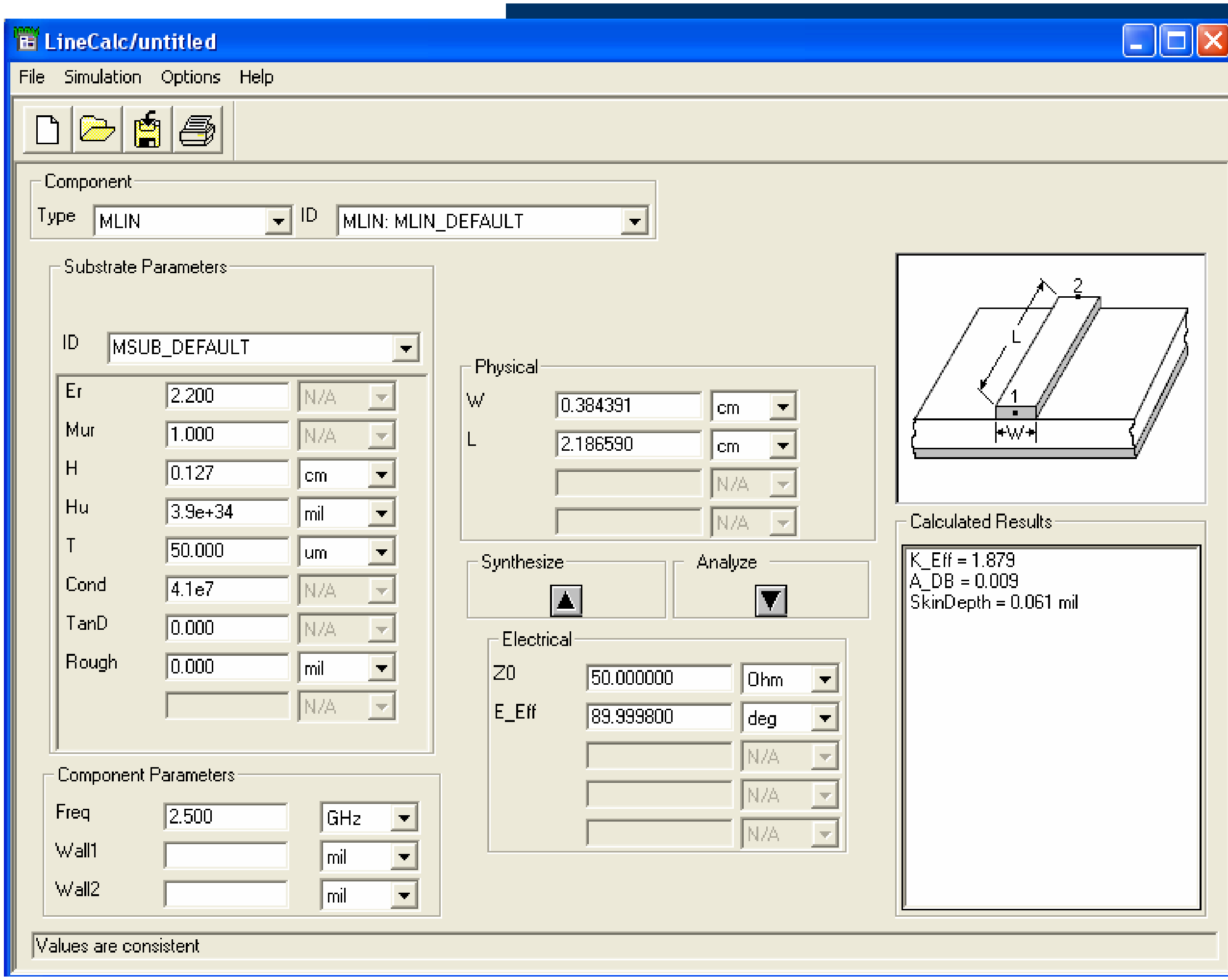
$$\ell = \frac{90°(\pi/180°)}{\sqrt{\epsilon_e}k_0} = 2.19\ \mathrm{cm}.$$

○



Using microwave simulator, characteristics can be calculated much faster and more accurate (taking into account thickness of conductor, disperion, etc…) Example: Agilents Linecalc, for multi-layer however: EM sim….

Example numerical calculation: Agilent linecalc



Difference is mainly effect thickness (W=0.391 for t<<H)

However at 10G:W=0.40, 25G:0.46, 100G:0.54cm

Microstrip substrate materials

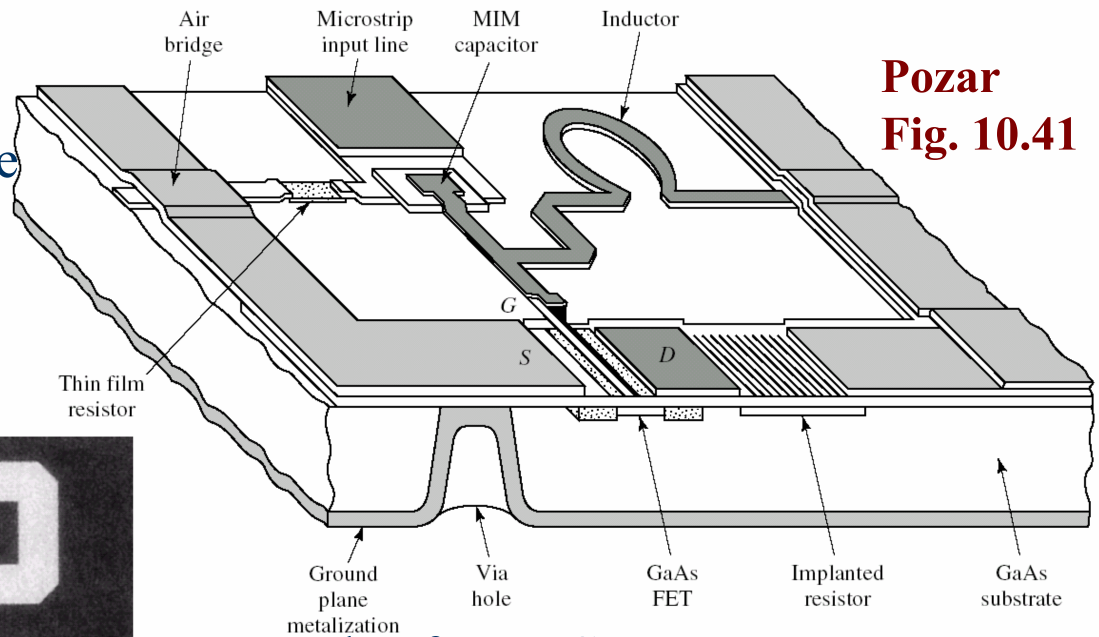
Material	Type of Material	Dielectric Constant	Loss Tangent	Other Characteristics
Fused Silica	Amorphous form of quartz (SiO_2)	3.78	< 0.0001 to at least 20 GHz	Expensive, brittle; difficult to obtain good metal adhesion.
Alumina	Ceramic form of alumina (Al_2O_4)	9.0 – 10.0	<0.0015 to 25 GHz	Characteristics depend on manufacture; $k = 9.8$ is most common.
Sapphire	Crystalline alumina (Al_2O_4)	8.6 horizontal, 10.55 vertical	<0.0015 in all directions	Electrically anisotropic.
RT Duroid® 5880[a]	Composite; PTFE[b]-fiber-glass	2.20	0.0009 at 10 GHz	Low-cost "soft" substrate; widely used.
RT Duroid® 5870[a]	Composite; PTFE-fiber-glass	2.33	0.0012 at 10 GHz	Low-cost "soft" substrate; widely used.
RT Duroid® 6006[a]	Composite; ceramic-PTFE	6.15	0.0019 at 10 GHz	Not mechanically as good as other materials
Silicon	Crystal (Si)	11.9	Very lossy	Dielectric loss is a problem for RF/MW circuits.
Gallium Arsenide	Crystal (GaAs)	12.9	Typically 0.001	Used for monolithic circuits only.
Indium Phosphide	Crystal (InP)	12.4	Typically 0.001; depends on purity	If you're using this exotic monolithic-circuit material, presumably you know something about it!

a. Rogers Corp., Chandler, Arizona
b. Polytetrafluoroethylene, or Teflon®

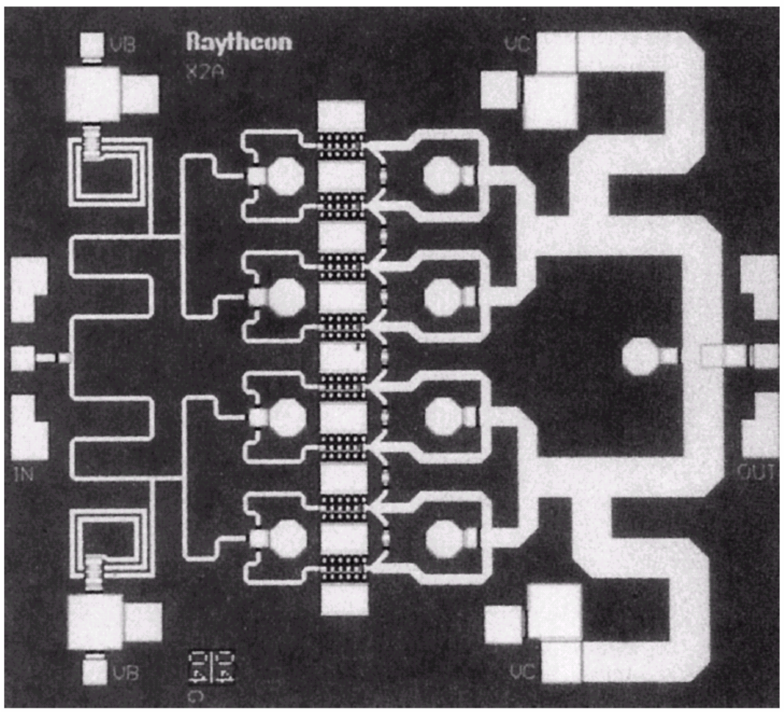
K. Chang, I. Bahl and V. Nair, "RF and Microwave Circuit and Component Design for Wireless Systems", Wiley Series in Microwave and Optical Engineering

Microstrip Monolithic Integrated Circuits

Layout monolithic microwave integrated circuit (MMIC) again microstrip topology



Pozar Fig. 10.41



Example of MMIC:
Integrated X-band power amplifier

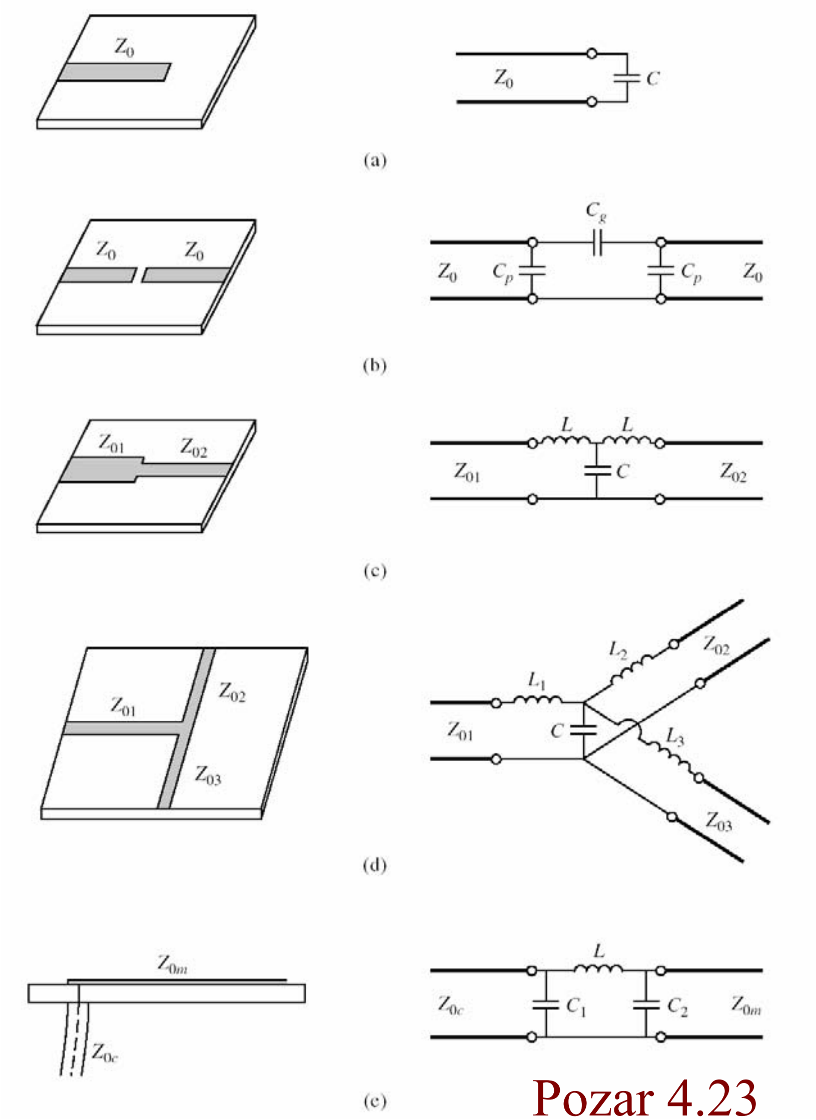
Multiple HBT's combined to deliver 5W

Pozar Fig. 10.42

More recent: **thin-film** μ-strip

Microstrip discontinuities

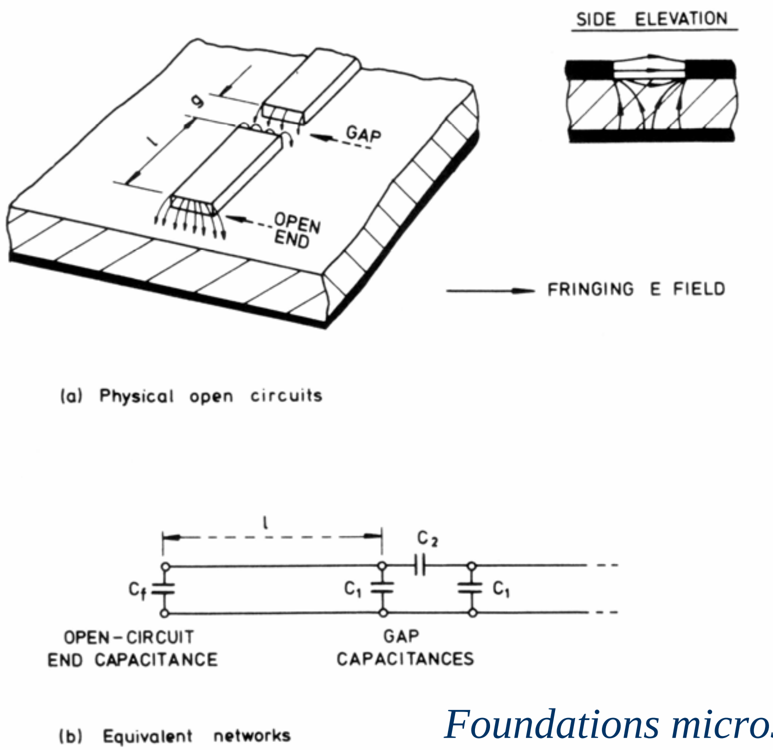
- Compared with ideal TL, additional parasitics associated with discontinuities such as:
  - open end and gap
  - Via-hole to ground
  - change in width (step)
  - T-junction & cross-junction
  - corner or bend
- Local change in E and H-field
  - fringing fields: capacitor
  - change current: inductance

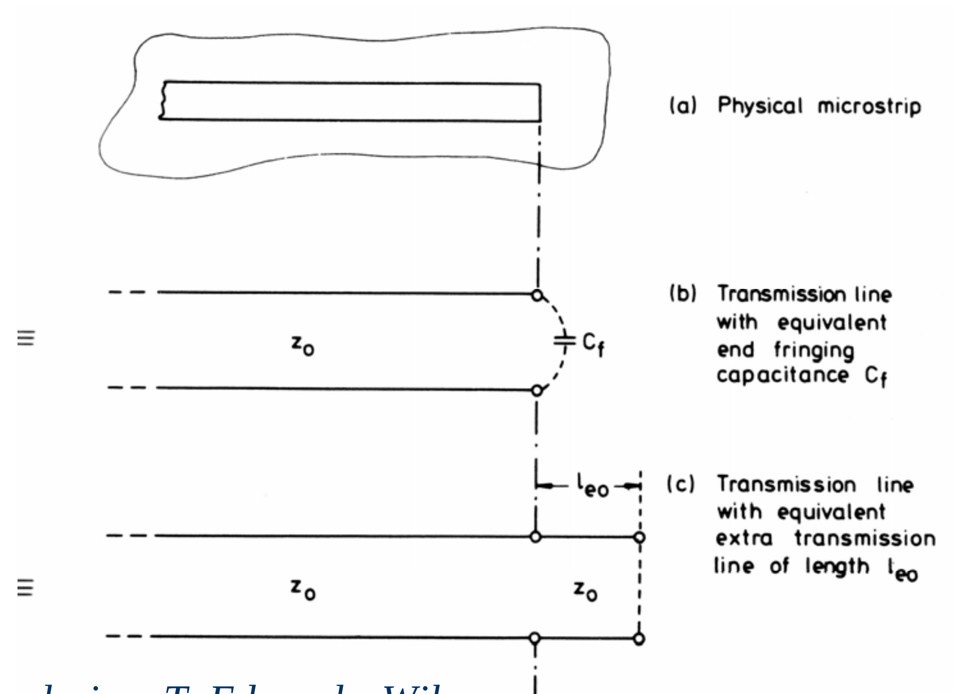


Pozar 4.23

Microstrip discontinuities (2)

- In general, dimensions <<λ: effect approximated by equivalent circuit model, parameters from rigorous EM-simulations
- Alternative : equivalent end effect (only distributed parameters)
- Models available in µwave circuit simulators (ADS), use them!

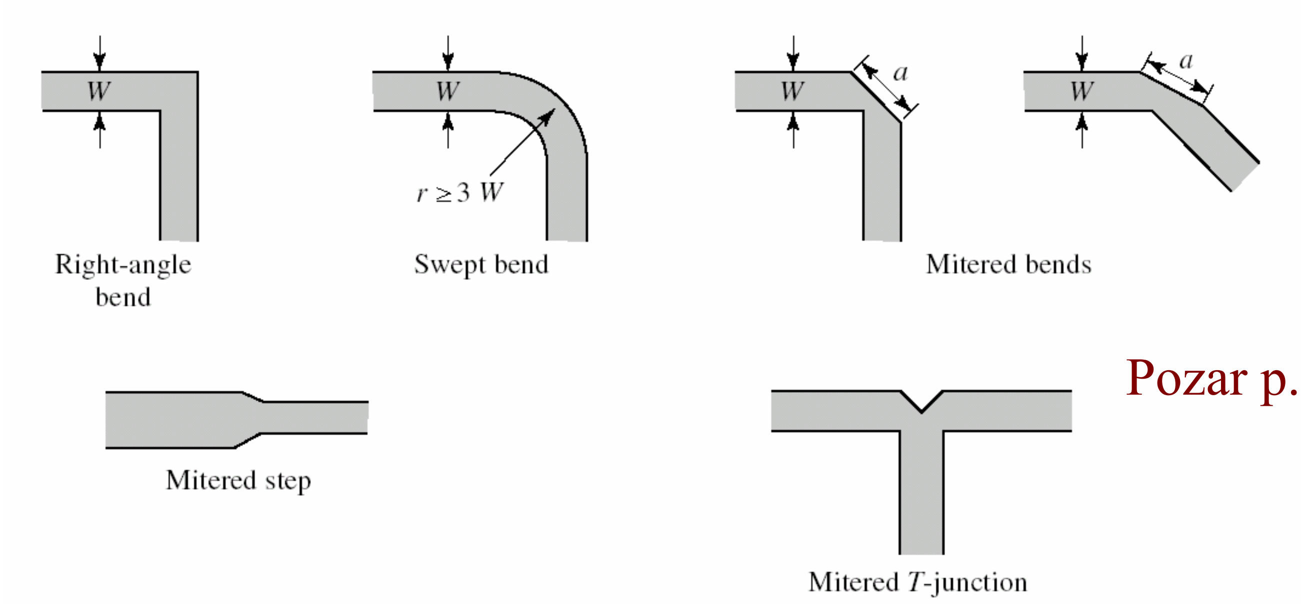




Foundations microstrip design, T. Edwards, Wiley

Microstrip discontinuities compensation

- Effect of discontinuities:
  - additional reactances can cause errors circuit design
  - conversion to other modes (surface-wave mode in µstrip)
  - radiation: loss mechanism, source EMI, coupling
- Whenever possible, effect discontinuity mitigated by making smoother transition or compensation for discontinuity

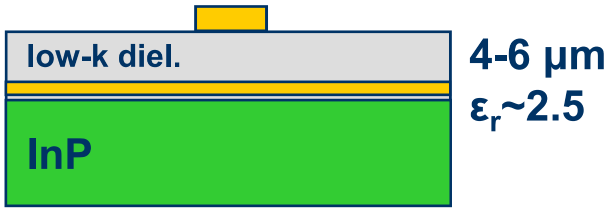


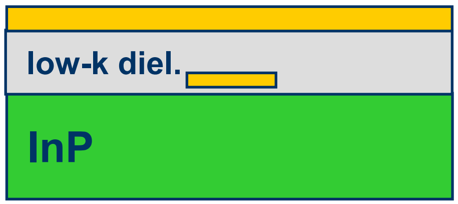
Pozar p.204

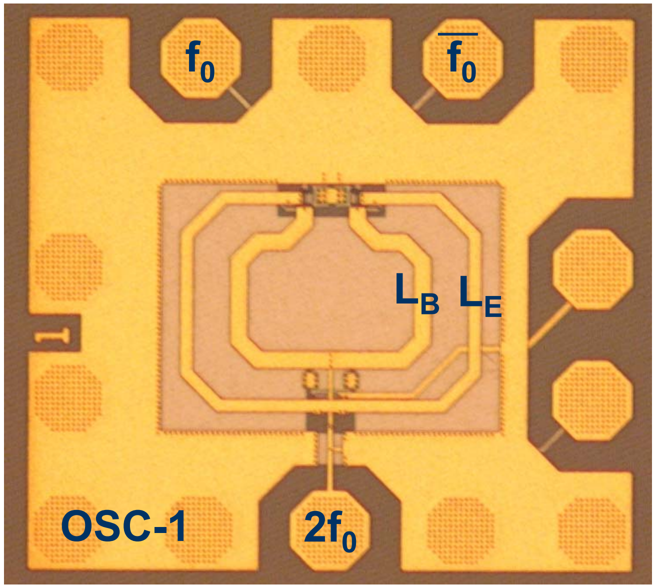
Thin-Film Microstrip (TF-MS)

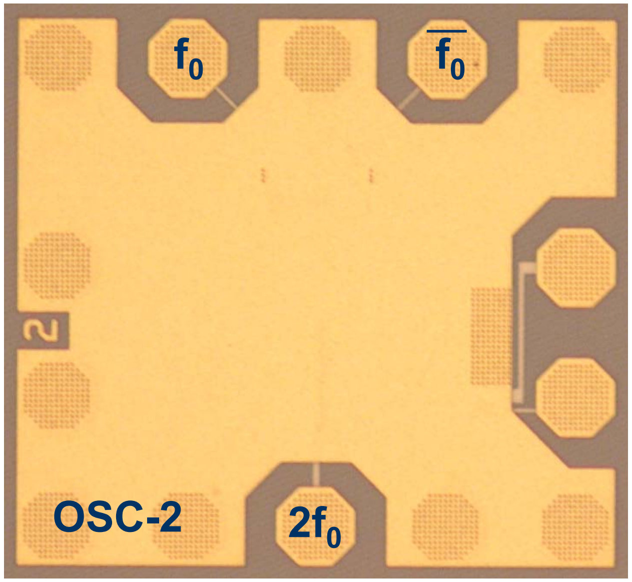
Conventional TFMS:
+ lower loss (fixed Z_O)
+ low $\varepsilon_{r,eff}$ (digital)

Inverted TFMS:
+ good shielding, easy flip-chip
+ high $\varepsilon_{r,eff}$ (delay)



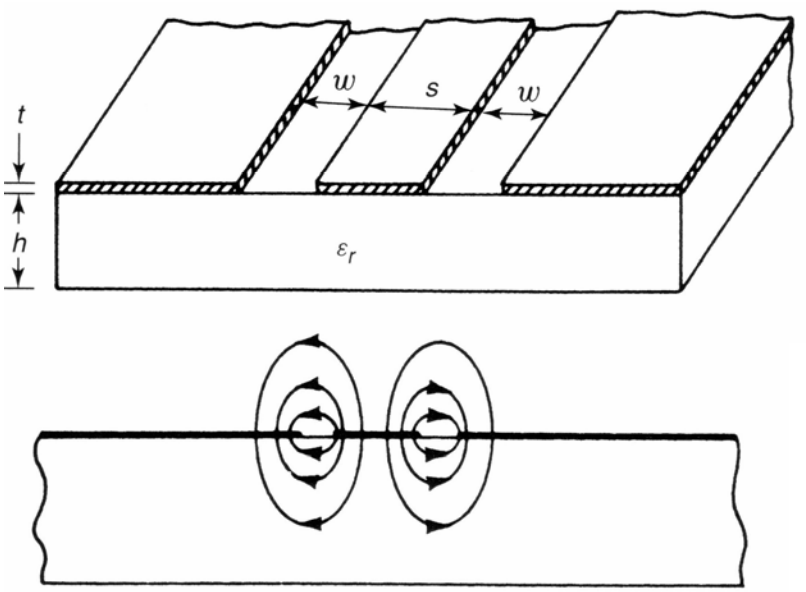






Coplanar waveguide lines (CPW)

- originally introduced by C.P. Wen in 1969
- only more recently (from mid-90's) used in circuits, mainly due to lack of accurate modeling
- also does not support pure TEM (different dielectric constant under and above line) quasi-TEM



Conductor in gap between 2 ground planes (dual microstrip)
Variations:

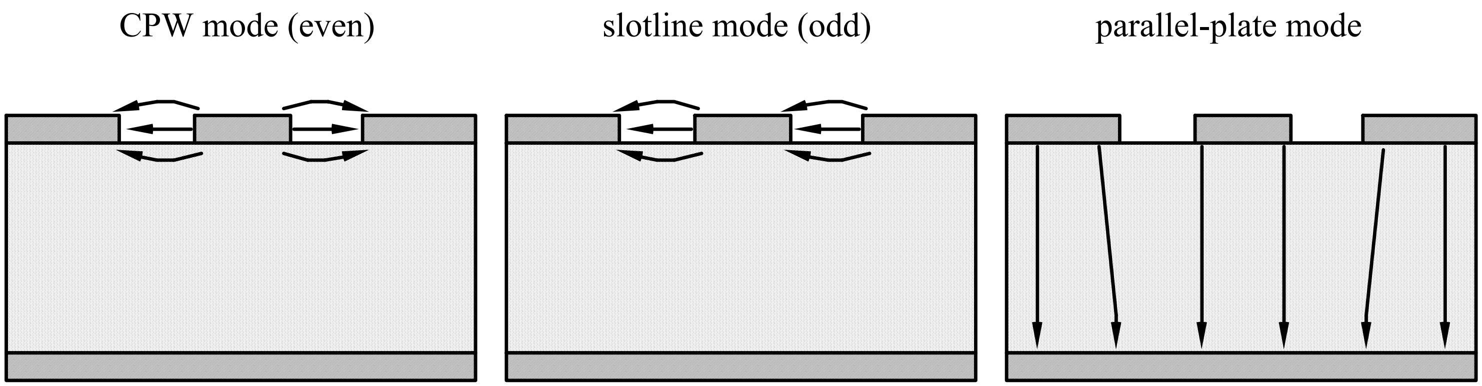
- CPS
- asymmetric CPW
- conductor backed CPW

Advantages CPW versus μ-stripline

- **Uniplanar technology**: signal & ground conductors at same side of substrate; no wafer thinning, via hole etching and backside metallisation needed, this reduces cost
- ground plane is accessible at front side of the wafer, easy implementation of active elements; especially advantageous at very high frequencies due to the **absence of via-hole inductance**;
- correctly designed, CPW lines have **low dispersion** (variation of the effective dielectric constant), important for broad band applications;
- the presence of the ground plane results in a **reduced coupling** between adjacent line, enables a further miniaturisation of the MMIC circuits
- **on-wafer measurement** technique based on coplanar probe tips is commercially available facilitating accurate measurements

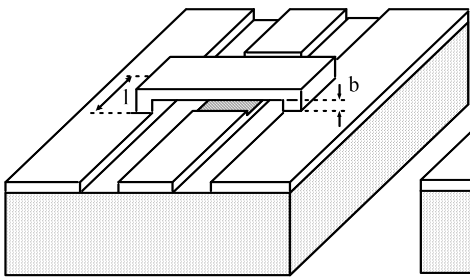
Disadvantages CPW versus μ-stripline

- in CPW the **electrical field is surface-oriented**, to determine accurately the coplanar line characteristics two-dimensional field needs to be solved.
- a coplanar circuit has normally a thick substrate, therefore **heat transfer** can become a problem in high-power applications,
- the CPW line consists of three unconnected conductors, such that both an **even and odd** transmission line mode can propagate, the symmetric CPW mode is the mode commonly used in coplanar circuits. At discontinuities, this mode can be converted into an asymmetric slot-line mode. Such multi-mode propagation should be prevented by the use of air bridges connecting the two ground metallisations to keep them at an equal potential. When a back metallisation is present, additionally a parallel-plate mode can be excited

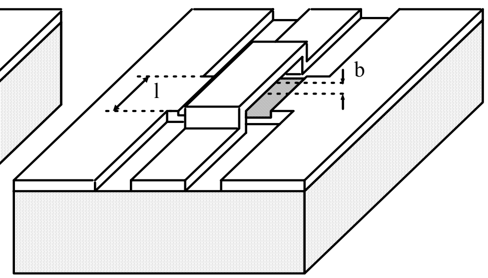


Discontinuities in CPW

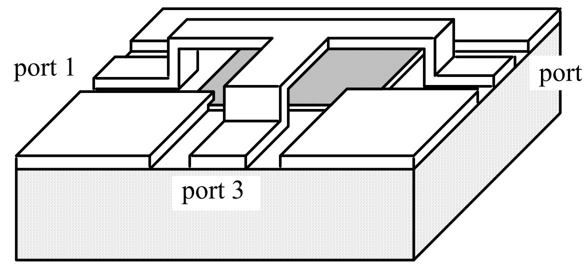
Important to maintain ground continuity at CPW discontinuities: frequent use of airbridges at T- or cross-junctions, bends, at input and output ports lumped elements, etc…

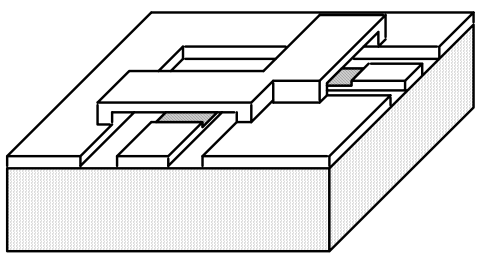


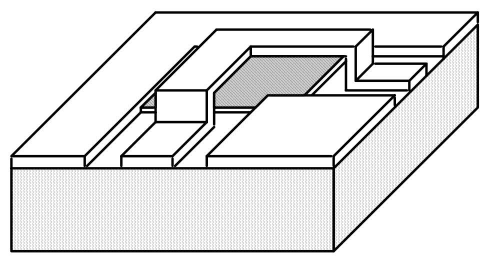
TYPE A

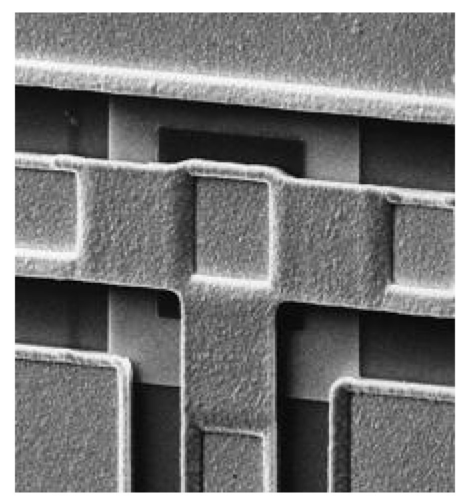


TYPE B

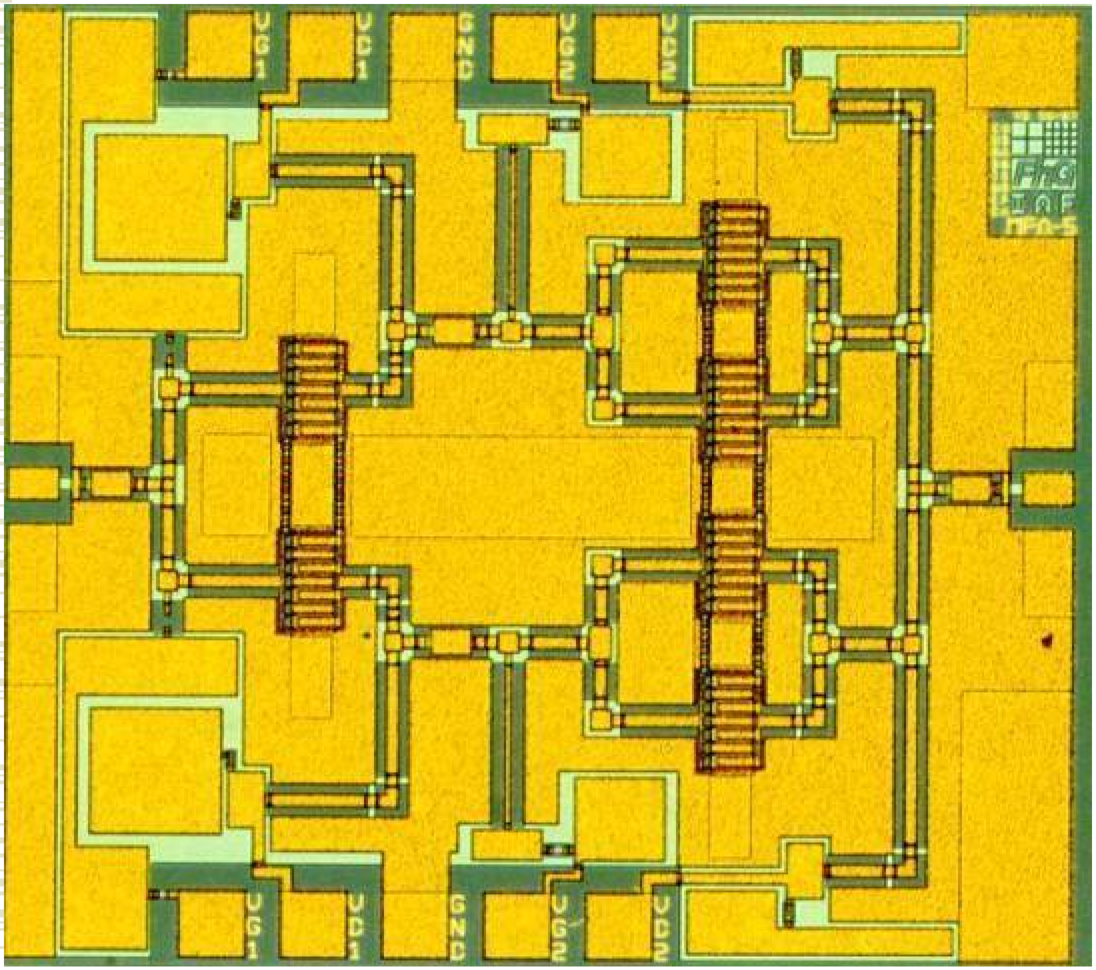








Example of coplanar MMIC (1)



- 1W power amplifier at 42 GHz
- 0.15μm PHEMT

- size: 1.4x1.4 mm^2
- 3.3V power supply
- Power-density 2-3x higher best μ-strip
- C-loaded lines

Summary of common TL's

TABLE 3.6 Comparison of Common Transmission Lines and Waveguides

Characteristic	Coax	Waveguide	Stripline	Microstrip &CPW
Modes: Preferred	TEM	TE_{10}	TEM	Quasi-TEM
Other	TM,TE	TM,TE	TM,TE	Hybrid TM,TE
Dispersion	None	Medium	None	Low
Bandwidth	High	Low	High	High
Loss	Medium	Low	High	High
Power capacity	Medium	High	Low	Low
Physical size	Large	Large	Medium	Small
Ease of fabrication	Medium	Medium	Easy	Easy
Integration with other components	Hard	Hard	Fair	Easy

- Other TL: ridge & dielectric WG, fin-line, balanced lines such as twisted pair, coplanar strip-line, slotline,
- Other important feature: radiation performance

Wave velocities and dispersion in TL

- Different velocities defined in TL:
  - Speed of light in medium: $1/\sqrt{\mu\varepsilon}$
  - Phase velocity: $v_p = \omega/\beta$
  - Group velocity: $v_g = \left(\frac{d\beta}{d\omega}\right)^{-1}\Bigg|_{\omega=\omega_0}$
- Phase velocity: speed at which a constant phase point travels. Dispersion of broadband signals in TL occurs when either phase velocity v_p or attenuation not constant afo frequency, from wave analogy: can be larger then speed of light
- Group velocity: velocity at which a narrow band signal propagates, related with information and power transport, needs to be smaller then c

Example: waveguide wave velocities

EXAMPLE 3.9 Waveguide Wave Velocities

Calculate the group velocity for a waveguide mode propagating in an air-filled guide. Compare this velocity to the phase velocity and speed of light.

Solution

The propagation constant for a mode in an air-filled waveguide is

$$\beta = \sqrt{k_0^2 - k_c^2} = \sqrt{(\omega/c)^2 - k_c^2}.$$

Taking the derivative with respect to frequency gives

$$\frac{d\beta}{d\omega} = \frac{\omega/c^2}{\sqrt{(\omega/c)^2 - k_c^2}} = \frac{k_o}{c\beta},$$

so from (3.234) the group velocity is

$$v_g = \left(\frac{d\beta}{d\omega}\right)^{-1} = \frac{c\beta}{k_0}.$$

The phase velocity is $v_p = \omega/\beta = (k_0 c)/\beta$.

Since $\beta < k_0$, we have that $v_g < c < v_p$, which indicates that the phase velocity of a waveguide mode may be greater than the speed of light, but the group velocity (the velocity of a narrowband signal) will be less than the speed of light. ○

Power capacity of TL's

- Power in TL's is limited by voltage breakdown, for air occurring at breakdown electric field $E_d=3\text{x}10^6$ V/m
- Calculation of capacity requires knowledge of E-field
- For air-filled coax: $E_\rho = V_0/(\rho \ln b/a)$ this is max. for ρ=a

$$V_{\max} = E_d a \ln b/a$$

and maximum power capacity becomes:

$$P_{\max} = \frac{V_{\max}^2}{2Z_0} = \frac{\pi a^2 E_d^2}{\eta_0} \ln \frac{b}{a}$$

- As expected power capacity increases for larger diameter cable, limit is cut-off frequency of higher order mode TE_{11}

$$P_{\max} = \frac{0.025}{\eta_0}\left(\frac{cE_d}{f_{\max}}\right)^2 = 5.8\times10^{12}\left(\frac{E_d}{f_{\max}}\right)^2 \qquad @10\text{GHz} = 520\text{kW}$$

Power capacity of waveguides

In an air-filled rectangular waveguide, the electric field varies as $E_y = E_o \sin(\pi x/a)$, which has a maximum value of E_o at $x = a/2$. Thus the maximum power capacity before breakdown is

$$P_{\max} = \frac{abE_o^2}{4Z_w} = \frac{abE_d^2}{4Z_w} ,$$

which shows that power capacity increases with guide size. For most waveguides, $b \simeq 2a$. To avoid propagation of the TE_{20} mode, we must have $a < c/f_{\max}$, where $f_{\max}$ is the maximum operating frequency. Then the maximum power capacity of the guide can be shown to be

$$\boxed{P_{\max} = \frac{0.11}{\eta_0}\left(\frac{cE_d}{f_{\max}}\right)^2 = 2.6 \times 10^{13}\left(\frac{E_d}{f_{\max}}\right)^2 .}$$

As an example, at 10 GHz the maximum peak power capacity of a rectangular waveguide operating in the TE_{10} mode is about 2300 kW, which is considerably higher than the power capacity of a coaxial cable at the same frequency.

Because arcing and voltage breakdown are very high-speed effects, the above voltage and power limits are peak quantities. In addition, it is good engineering practice to provide a safety factor of at least two, so the maximum powers which can be safely transmitted should be limited to about half of the above values. If there are reflections on the line or guide, the power capacity is further reduced. In the worst case, a reflection coefficient magnitude of unity will double the maximum voltage on the line, so the power capacity will be reduced by a factor of four.

The power capacity of a line can be increased by pressurizing the line with air or an inert gas, or by using a dielectric. The dielectric strength (E_d) of most dielectrics is greater than that of air, but the power capacity may be primarily limited by the heating of the dielectric due to ohmic loss.

Homework & next lecture!!

- Pozar, "Microwave Engineering" (3rd Ed.!) Will put on site!
  - 2.6 & 2.7
  - 3.3
  - 3.19 (optional: see if you can use ADS Linecalc)
  - 3.27

- Due date: 2/16

- Next week, we'll review microwave network analysis, start impedance matching and review HW1