ECE 202A Notes set 1

Course content - what are we trying to accomplish?

- Refer to course syllabus -

* specifically -

High frequency transistor circuit design.

use of high frequency instruments.

* Key to course content will be systematically treating the effects of conductors (transmission lines) on circuit performance.

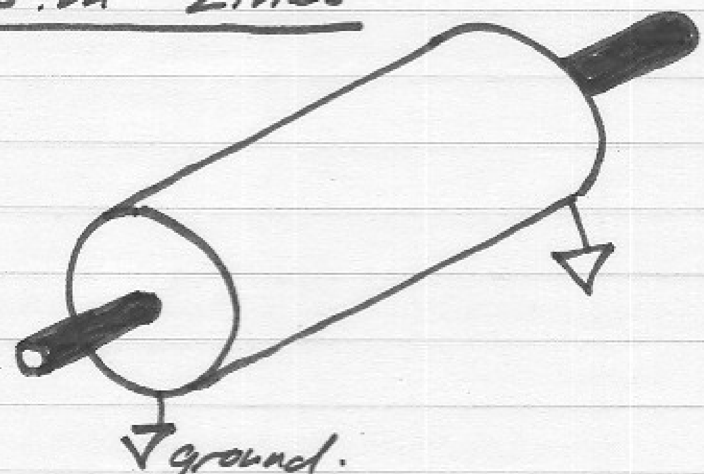
Course content will involve

* significant review of transmission line theory - with emphasis on circuit-element analogs.
* significant review of transistor models for high frequency design.

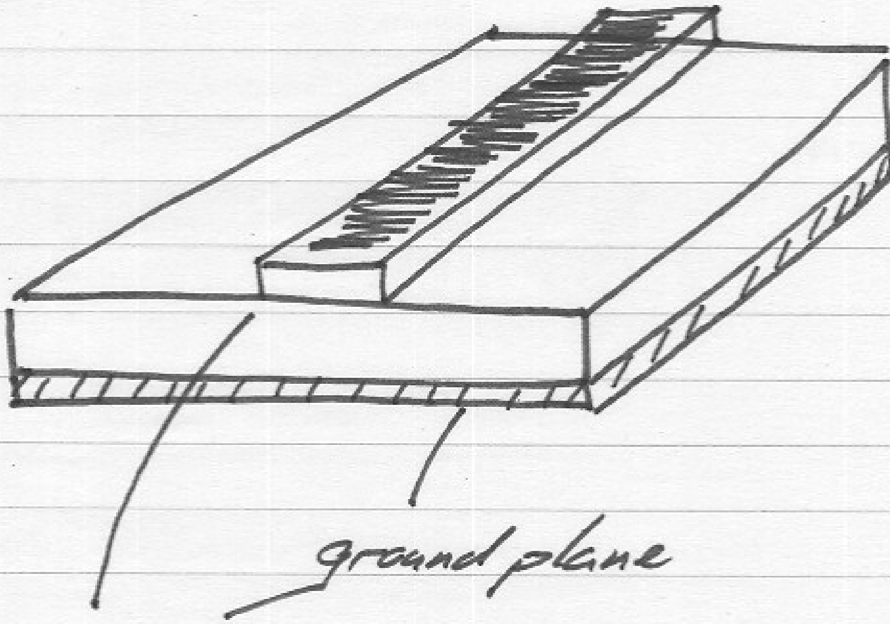
We will try to cover this <u>review</u> material in 3-4 weeks. Experience has shown that the material cannot be skipped

Transmission Lines

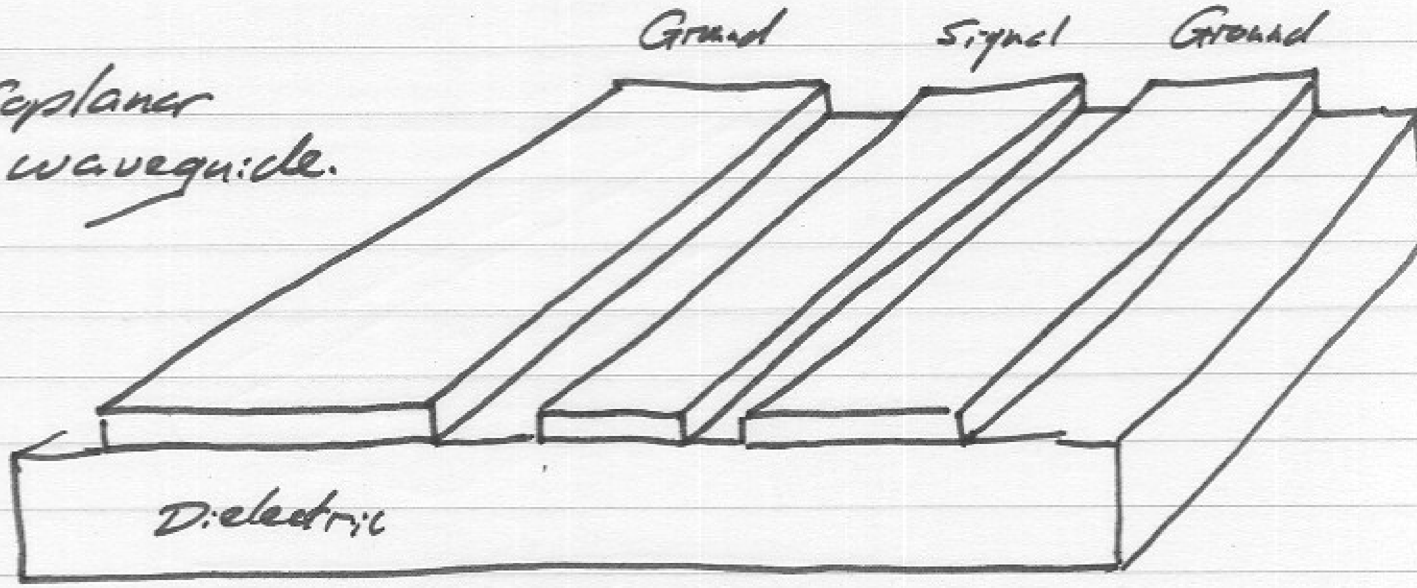
Types:



Microstrip:



Coplanar waveguide.



... and many others ...

the common features are:

1) a pair of conductors

2) a geometry which does not change with distance

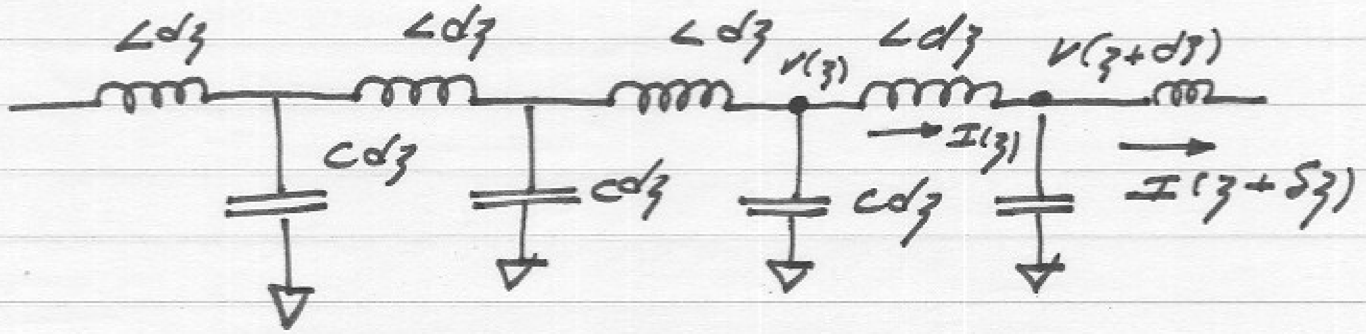
- These characteristics are sufficient for the propagation of a guided wave.

Review of transmission-line theory.

We will approach this from a strictly circuit point of view - ignoring the normal electromagnetic approach.

Justification: develop models for transmission lines which strongly parallel lumped electrical elements. This will aid us in circuit design.

Circuit model of transmission lines:



C = capacitance / length

L = inductance / length

Circuit analysis gives us

$\partial V/\partial z = -L\, \partial I/\partial t$ and $\partial I/\partial z = -C\, dV/dt$

these are called the telegrapher's equations

from which the wave equation is found:

$$\partial^2 V/\partial z^2 = LC\, \partial^2 V/\partial t^2$$

The solutions are

$$V(z,t) = V^{+}(t - z/v) + V^{-}(t + z/v)$$

$$I(z,t) = \frac{V^{+}(t - z/v)}{Z_0} - \frac{V^{-}(t + z/v)}{Z_0} + I_0$$

where V^{+} and V^{-} are the forward and reverse voltage waves, the velocity is

$v = 1/\sqrt{LC}$, the characteristic impedance $Z_0 = \sqrt{L/C}$, and I_0 is a D.C. current.

These are simple and familiar relationships;

key point:

<u>voltage at any point:</u>

$$V(z,t) = V^+(z,t) + V^-(z,t)$$

<u>current at any point.</u>

$$I(z,t) = \frac{V^+(z,t) - V^-(z,t)}{Z_0} + I_0$$

It is standard in microwave texts to define <u>normalized</u> <u>wave</u> <u>amplitudes</u>:

a = forward wave amplitude = $v^+/\sqrt{Z_0}$

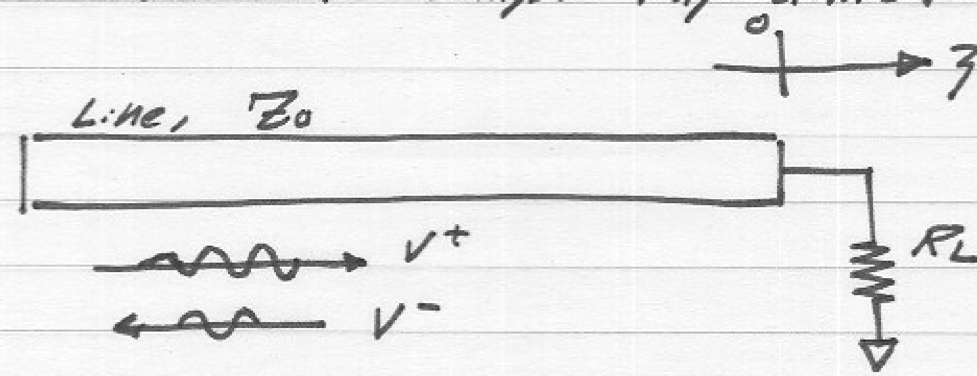
b = reverse wave amplitude = $v^-/\sqrt{Z_0}$

power in forward wave = $a^2(t)$

power in reverse wave = $b^2(t)$

cautions: Power will be aa^* (bb^*) once we deal with phasors. Also, definitions must be generalized if we are to deal with <u>complex</u> characteristic impedances.

Reflections off simple impedances:



boundary conditions:

Line		load	
$(V^+ - V^-)/Z_0$	$=$	I	Current
$V^+ + V^-$	$=$	V	voltage
		$V = R_L I$	

Combining these:

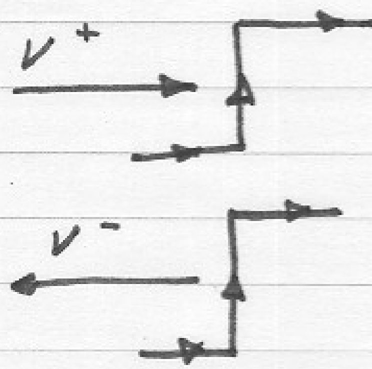
$$V^-(z=0,t) = V^+(z=0,t) \cdot \Gamma_L$$

$$\Gamma_L = \frac{R_L/Z_0 - 1}{R_L/Z_0 + 1} = \text{reflection coefficient}$$

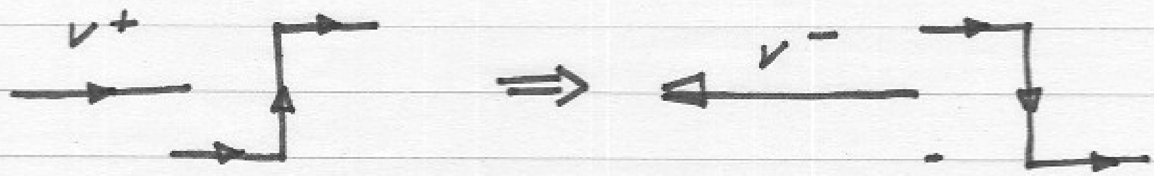
Cases:

open circuit ($R_L \to \infty$)

$\Gamma_L = +1$



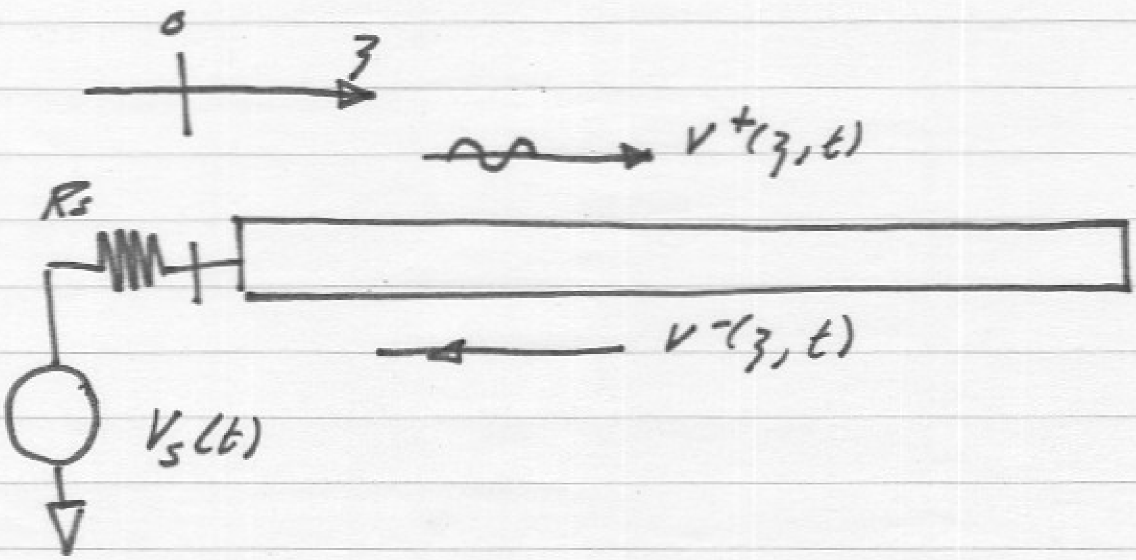
short circuit ($R_L \to 0$) $\Rightarrow$ $\Gamma_L = -1$



Matched Load: $R_L = Z_0 \Rightarrow \Gamma_L = 0$

No reflection.

Similar Considerations can be applied at the sending end of a transmission line:



again, we apply the boundary conditions:

$$v(z=0) = V_s - R_s\, I(z=0)$$

$$V = v^+ + v^- \qquad \text{and} \qquad I = (v^+ - v^-)/Z_0$$

To obtain:

$$v^+(z=0,t) = \Gamma_s\, v^-(z=0, t) + V_s(t)\, T_s$$

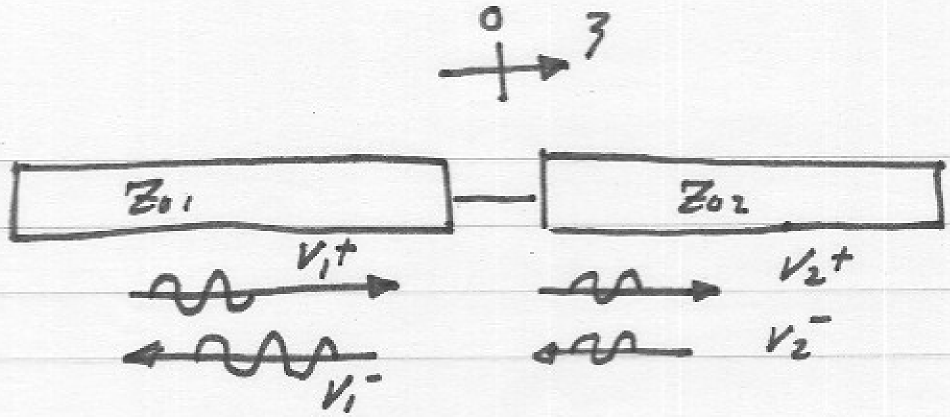
T_s = "source transmission coefficient"

$$= \frac{Z_0}{Z_0 + R_s}$$

Γ_s = source reflection coefficient

$$= \frac{R_s/Z_0 - 1}{R_s/Z_0 + 1}$$

Junctions between lines of dissimilar impedance can be similarly treated, starting with the boundary condition of the equality of voltages & currents.



$$V_2^+ = T_{12} V_1^+ + \Gamma_{22} V_2^-$$

where $T_{12} = \dfrac{2Z_{02}}{Z_{01} + Z_{02}}$

$$\Gamma_{22} = \frac{Z_{01}/Z_{02} - 1}{Z_{01}/Z_{02} + 1}$$

More comments about transmission lines:

Lines fall into two classes

unbalanced Lines

* Have a signal conductor and ground
* ground is at zero potential relative to distant objects.
* True unbalanced: coaxial line.

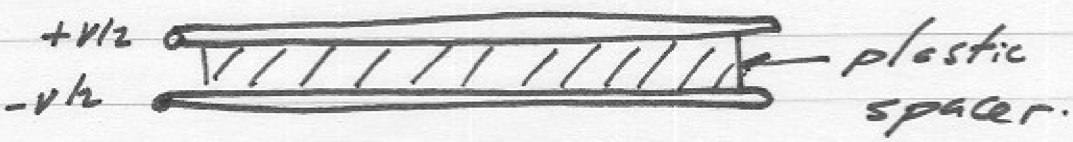
almost unbalanced: Microstrip

coplanar waveguide

with these there is a small potential on the nominal ground plane.

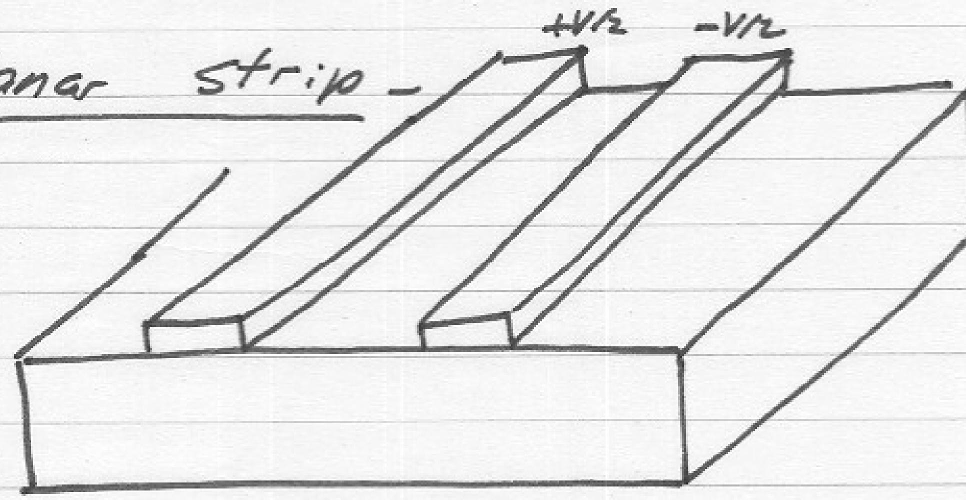
balanced lines

* have 2 symmetric conductors
* conductor potentials are symmetric with respect to distant objects.
* Twin-lead (TV) antenna cable



Twisted-pair (wires twisted by drill)

Coplanar strip



Simple but important transmission line laws:

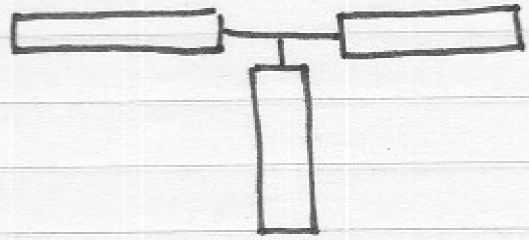
1) Source and load impedances should be equal to the characteristic impedance if reflections are to be avoided.

2) Think about the voltages on transmission line conductors before connecting them

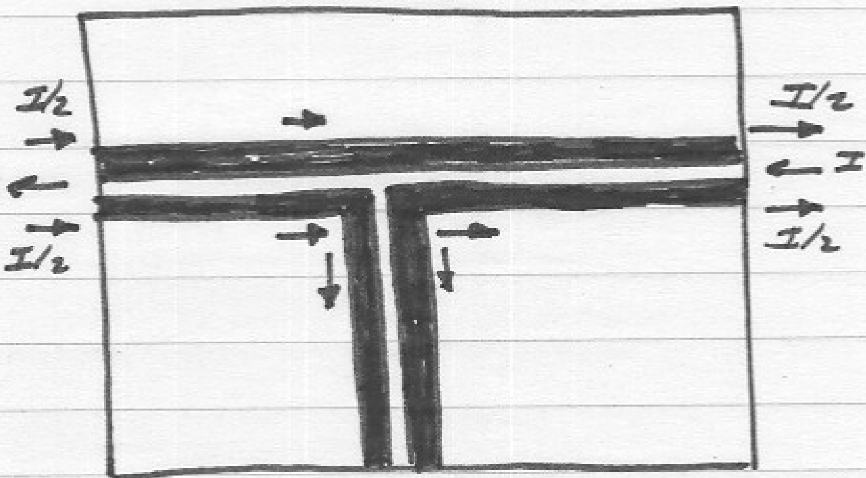
3) Think about the currents on transmission-line conductors before connecting them.

Example T-junction in coplanar waveguide

circuit diagram

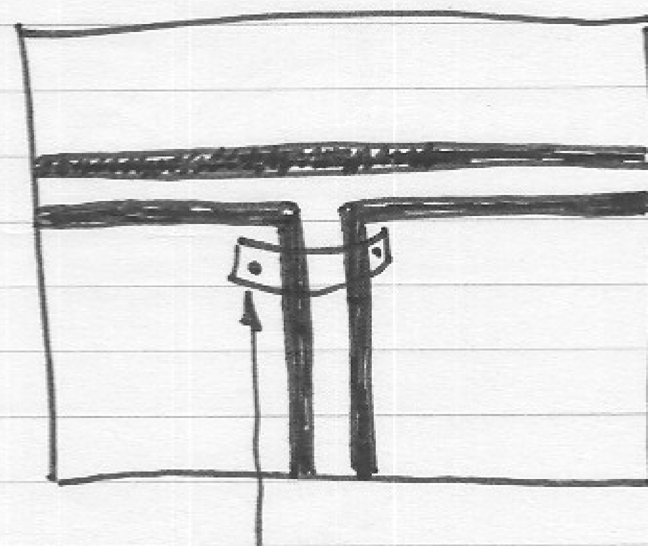


[Incorrect IMPLEMENTATION.]



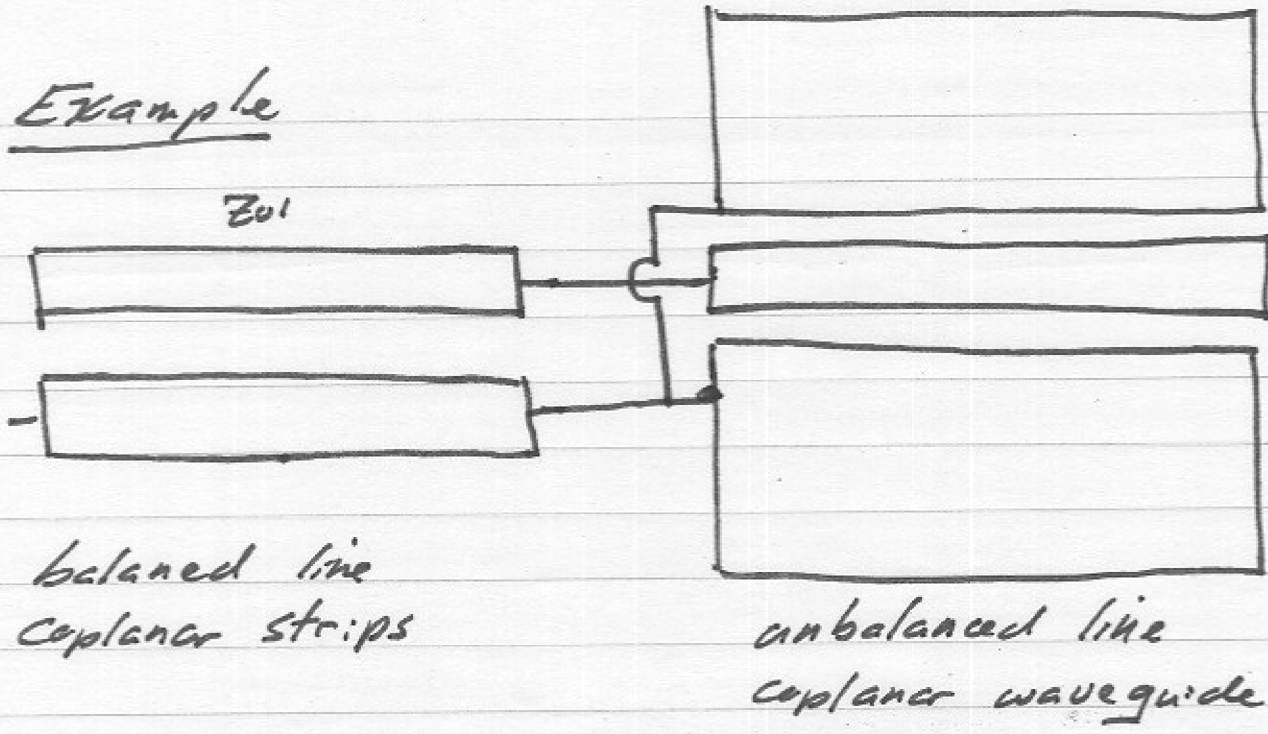
T-junction interrupts the current in the ground plane!

[Correct Implementation]



air-bridge connection between grounds.

Example



This is a problem because the balanced line has potentials ($+V/2$ & $-V/2$) on both conductors while the coplanar line has potential = zero on the ground conductor.

* There are a <u>large</u> <u>number</u> of ways to go wrong - and a simple rule to avoid them: think: "where do the currents go, and where are the conductor voltages?"

* If we do make such a mistake - what happens? Generally there are more conductors involved than shown in the above examples. sets of propagating voltage waves between the various conductors will be excited. The solution involves the use of an <u>impedance</u> <u>matrix</u> <u>method</u>

Viewing transmission lines as approximate lumped elements:

We want to be able to think of transmission lines with the same level of visualization which (we hope) we have developed for lumped circuits.

For any transmission line:

Z_0 = impedance = $\sqrt{L/C}$

v = velocity = $1/\sqrt{LC}$

Suppose the line has length ℓ

Then the total inductance is

$$L_T = \frac{Z_0 \ell}{v} = Z_0 \tau$$

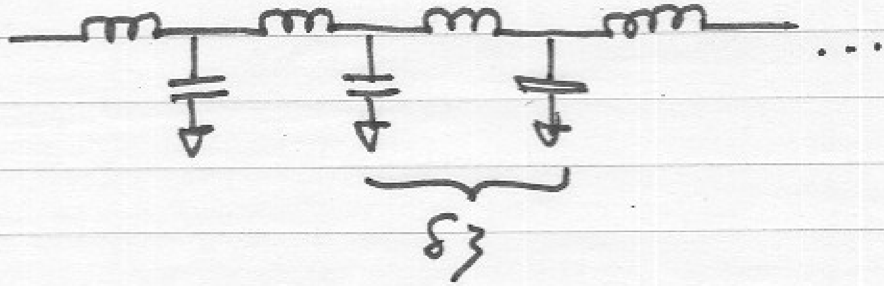
and the total capacitance is

$$C_T = \frac{\ell}{Z_0 v} = \tau / Z_0$$

where $\tau = \ell / v$ is the time delay or time-of-flight on the line.

Note that in terms of electrical properties that the physical line length is irrelevant; properties depend upon impedance and delay time

A true transmission line is distributed:



But:

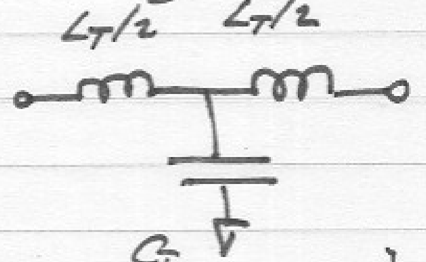
* For lines much shorter than a wavelength ($\ell \ll \lambda$)

  - or equivalently -

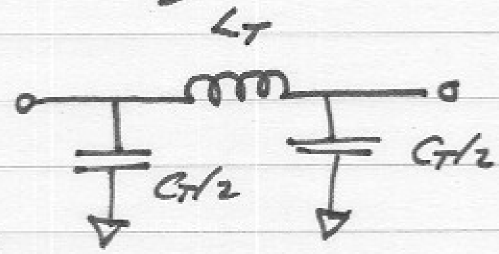
* For lines whose delay times are much shorter than a pulse-waveform risetime ($\tau \ll T_{Rise}$)

* The transmission line can be replaced by a lumped-element equivalent

T-equivalent

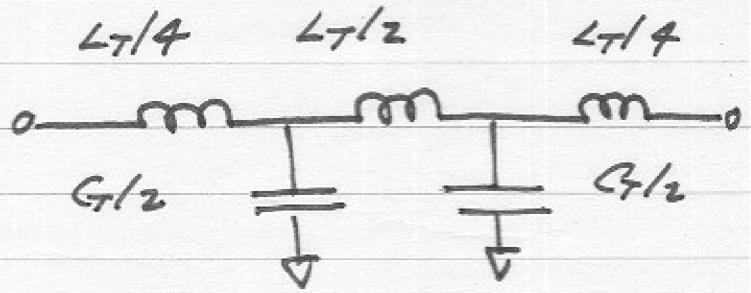


Π-equivalent



$$\boxed{\begin{aligned} L_T &= T Z_0 \\ C_T &= T/Z_0 \end{aligned}}$$

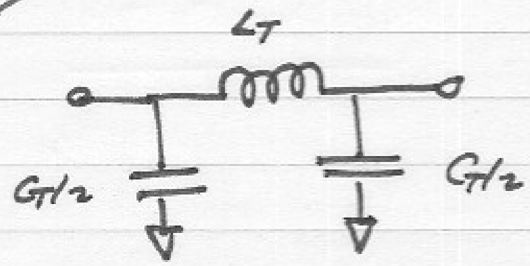
For longer lines, just break into N sections before using the T or Π equivalents:



<u>etc.</u>

The parallel between lumped and distributed elements is helpful:

a high-impedance line - if short - is approximately an inductor.

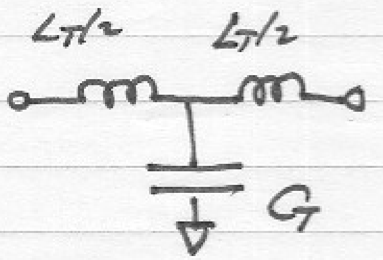


$L_T = Z_0 T$ big

$C_T = T/Z_0$ small

a low impedance line - if short - is approximately a capacitor

~~It is often said~~

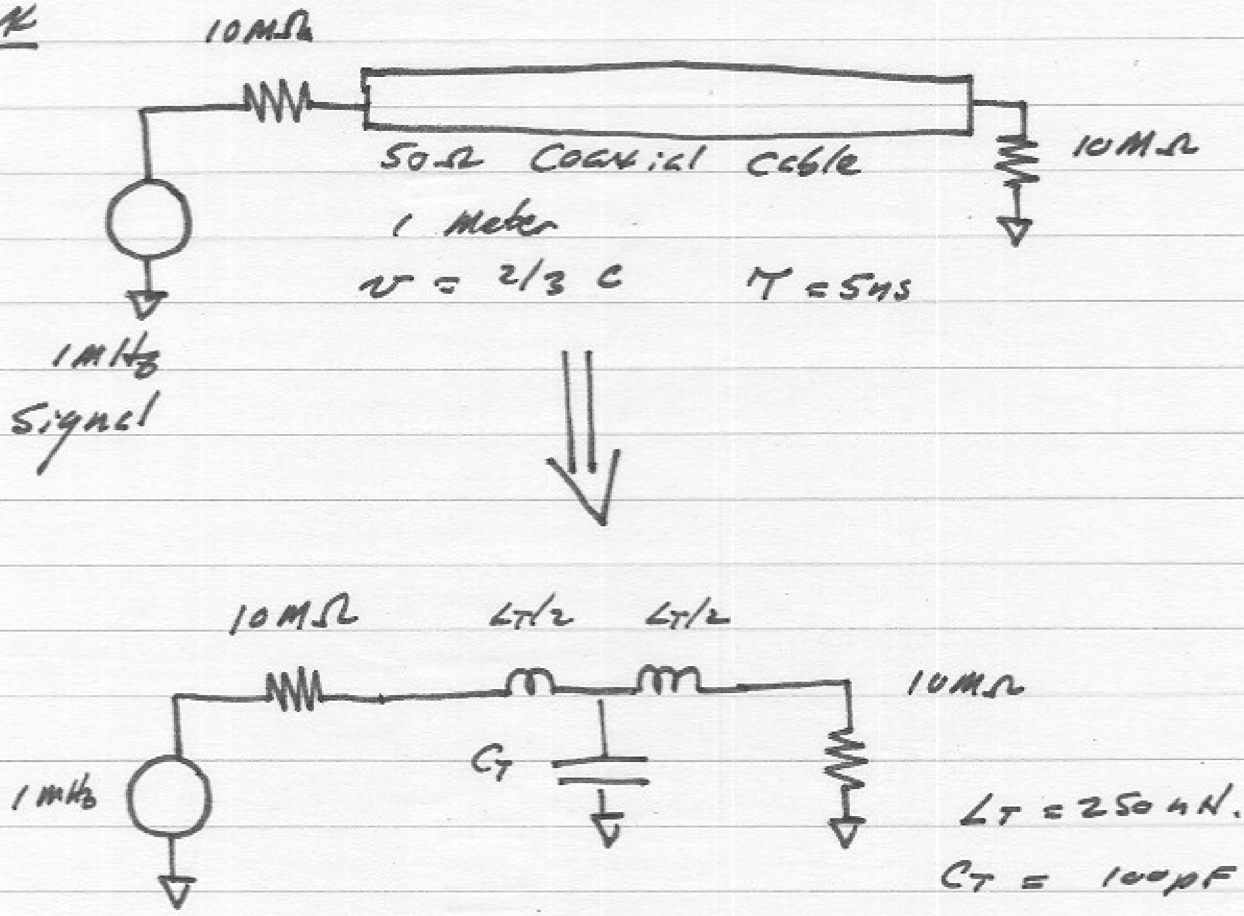


$L_T = Z_0 T$ small

$C_T = T/Z_0$ big.

It is often said that if a line is much smaller than a wavelength that it can be <u>neglected</u> — This is <u>not true</u>

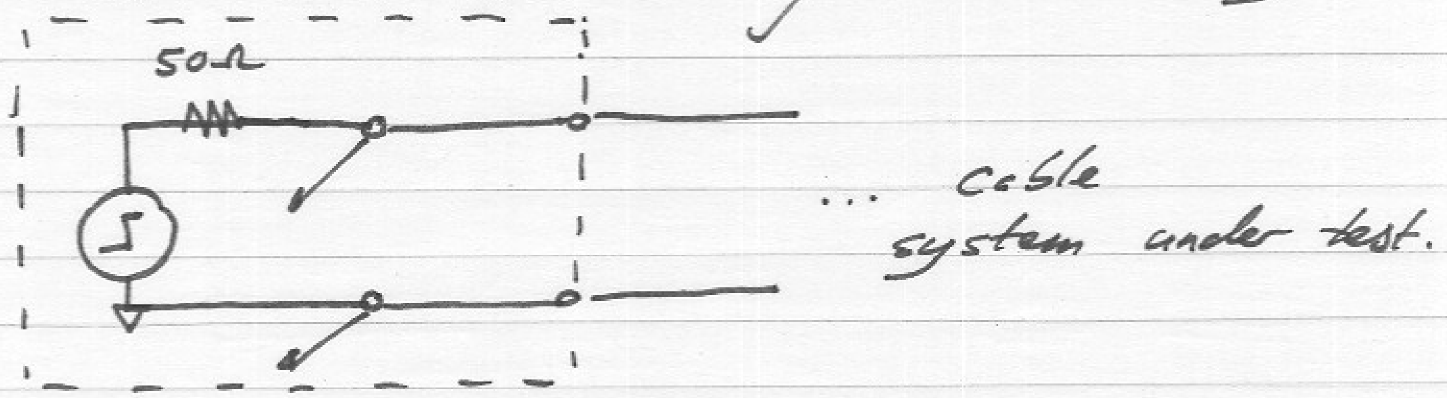
<u>Ex</u>



while L_T is negligible ($L_T/20M\Omega$ is small)

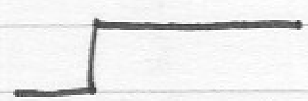
C_T is significant ($C_T \cdot 5M\Omega = 500\mu s$) !

Time-domain reflectometry - Echo testing:



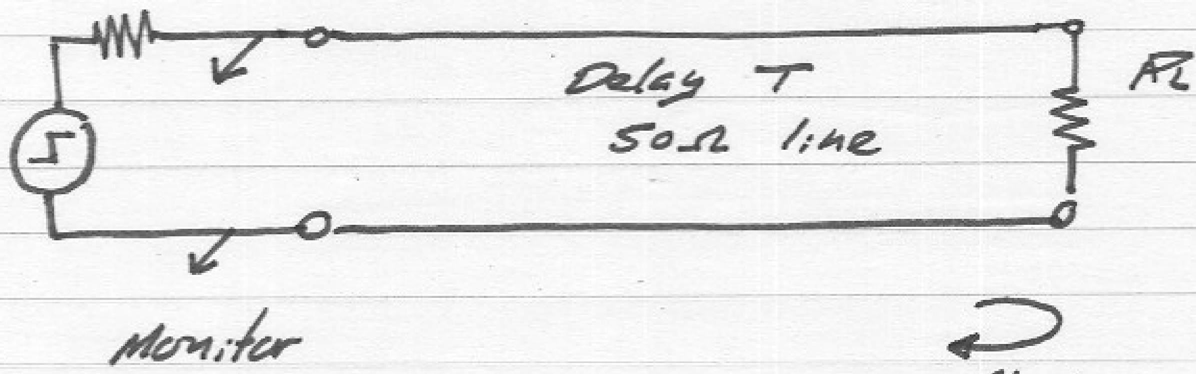
To monitor points for oscilloscope

The voltage observed at the monitor point is the sum of the incident & reflected waves...

since the incident signal is a ,

the reflected signal is readily identified.

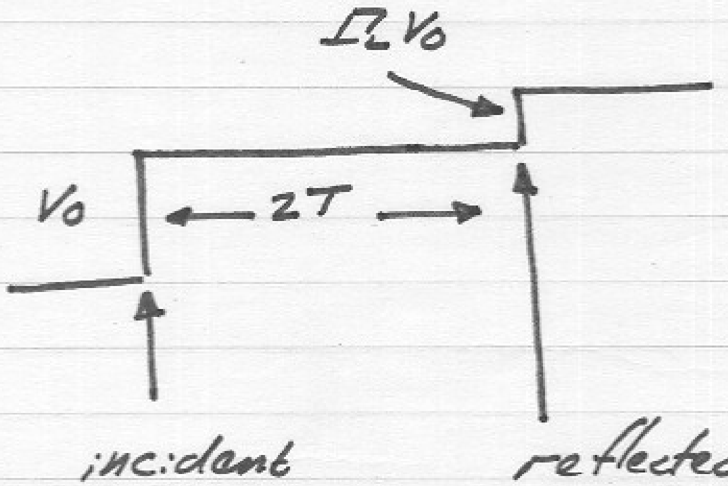
<u>Example</u> Load resistor at end of line



reflection

magnitude $\Gamma_L = \dfrac{R_L/Z_0 - 1}{R_L/Z_0 + 1}$

Voltage monitored:



From the magnitude of Γ, R_L can be determined.

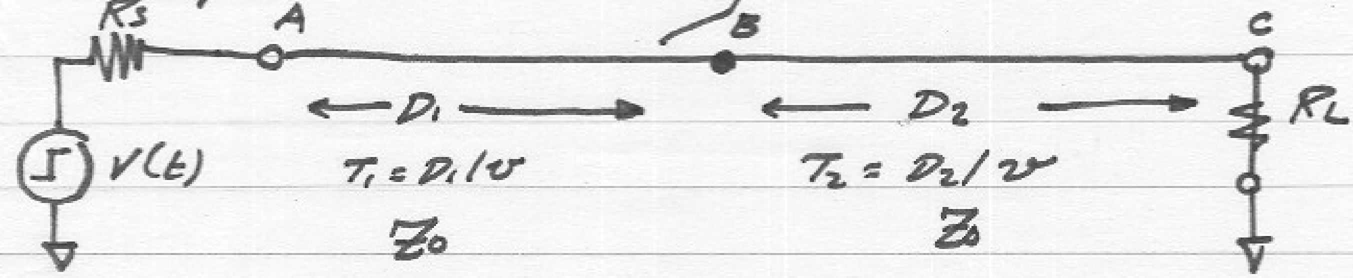
Before Continuing with the TDR, lets discuss lattice diagrams.

* Handy tool for analysis of transmission line systems with pulsed excitation and resistive source & load terminations.

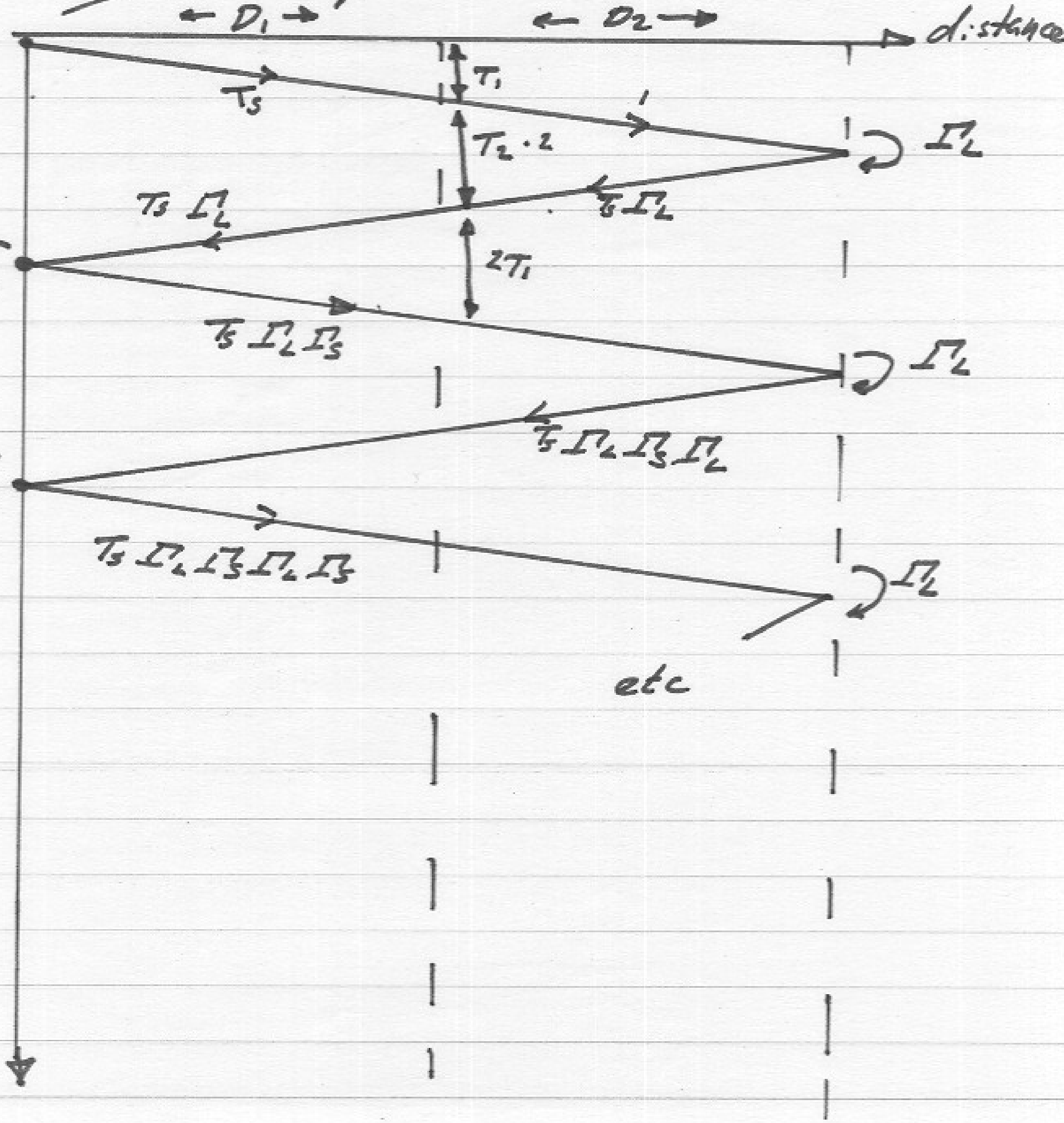
*. Particular applicability for analysis of data bus & interconnects in high speed digital systems.

Best to ~~analyze~~ present by an example.

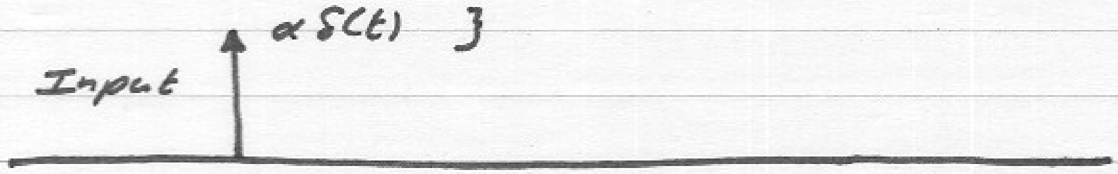
example: lattice diagram:

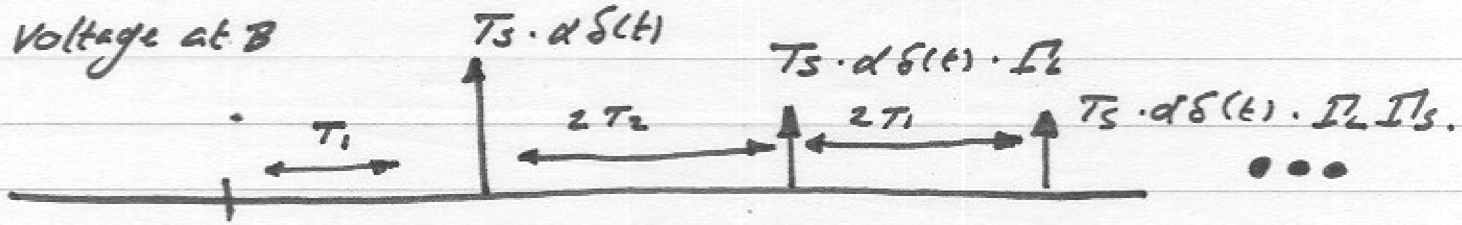


analyze for impulse response first - then use convolution for general response.

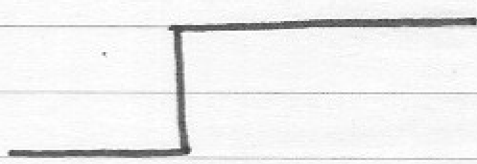


We can now represent the impulse response at point B:

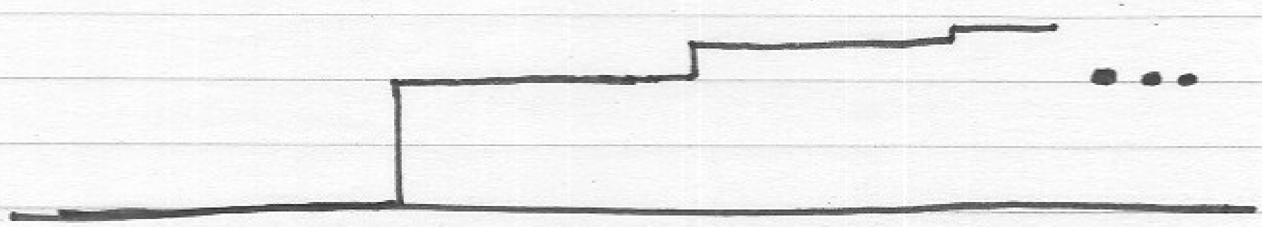




so if the input is a step-function:



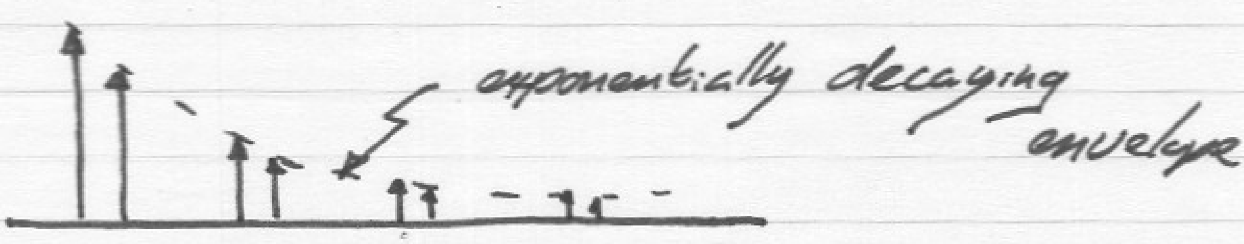
Then the voltage at point b will be:



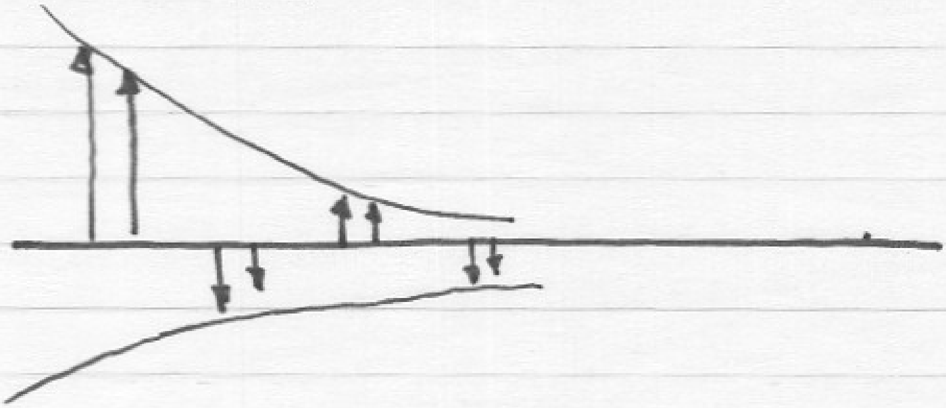
Comment:

depending upon the values of R_S & R_L,

the impulse response may look like so:

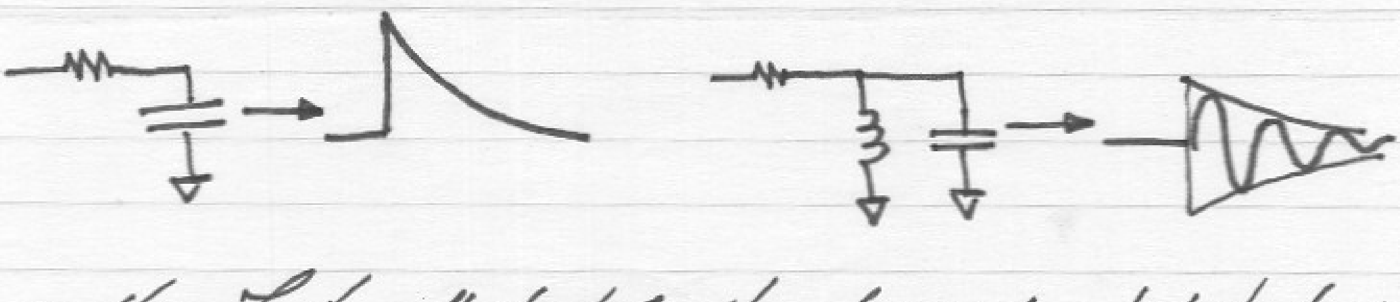


or like so:



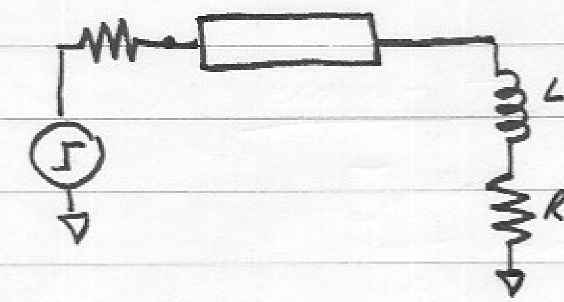
You are advised to compare these to the

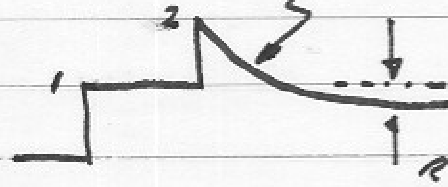
impulse responses of lumped circuits, e.g.



... as this further illustrates the lumped-distributed relationships.

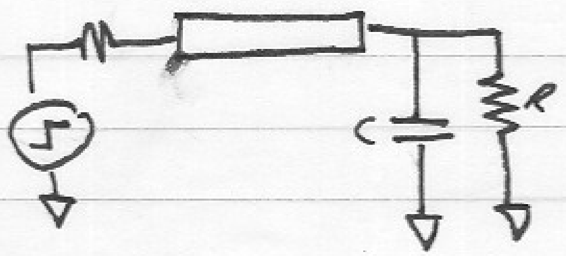
TDR responses with Reactive Loads





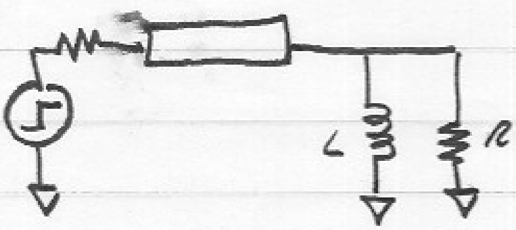
$\tau = \frac{L}{R+Z_0}$

$\Gamma = \frac{\frac{R}{Z_0} - 1}{\frac{R}{Z_0} + 1}$



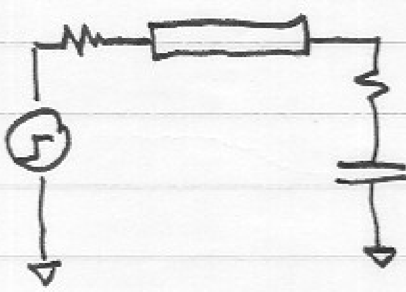
$\Gamma = \frac{R/Z_0 - 1}{R/Z_0 + 1}$

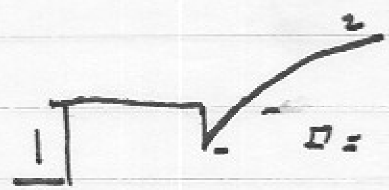
$\tau = (R \| Z_0)\, C$



$\Gamma = \ldots$

$\tau = L / R \| Z_0$

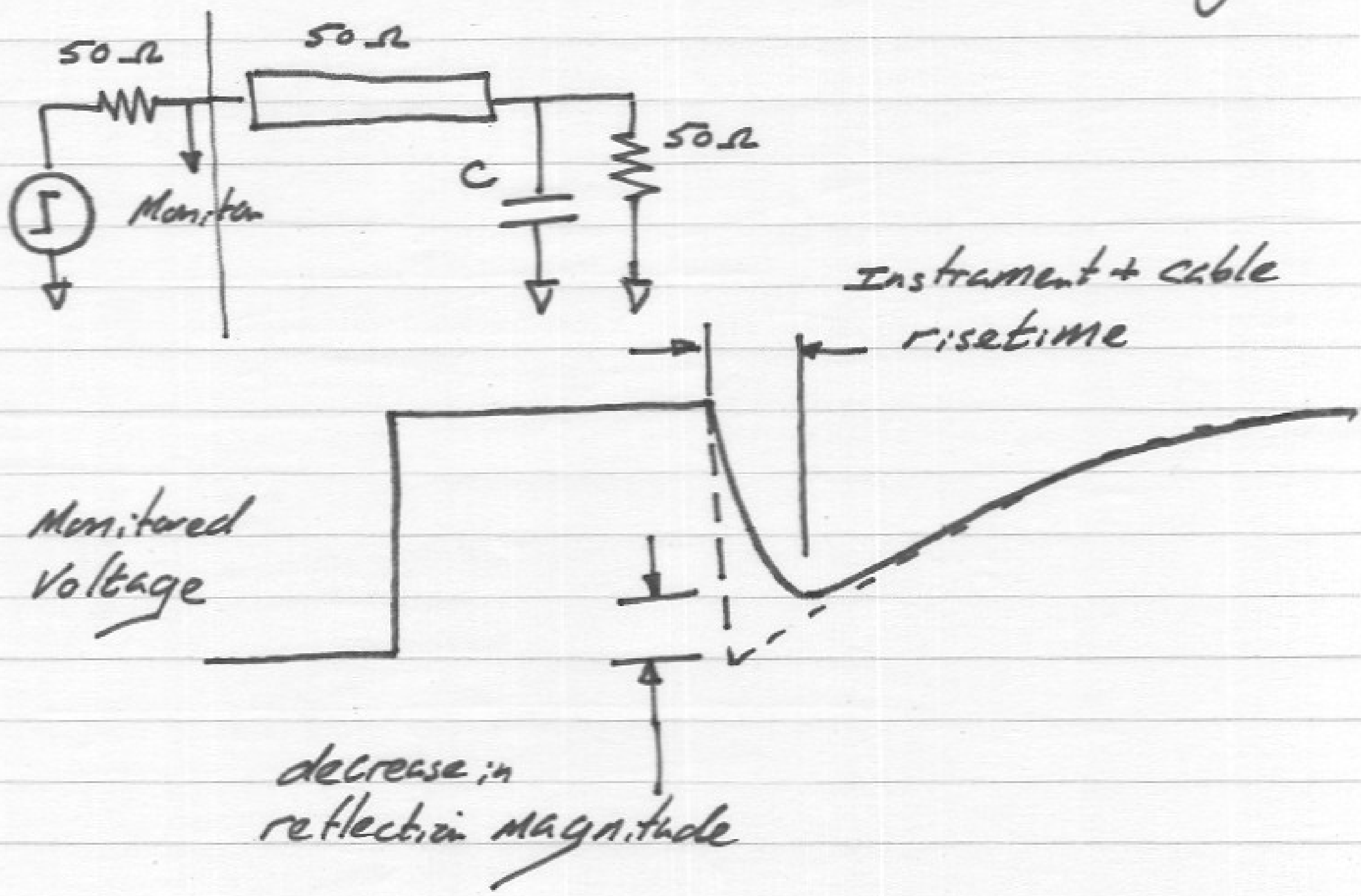




$\Gamma = \ldots$

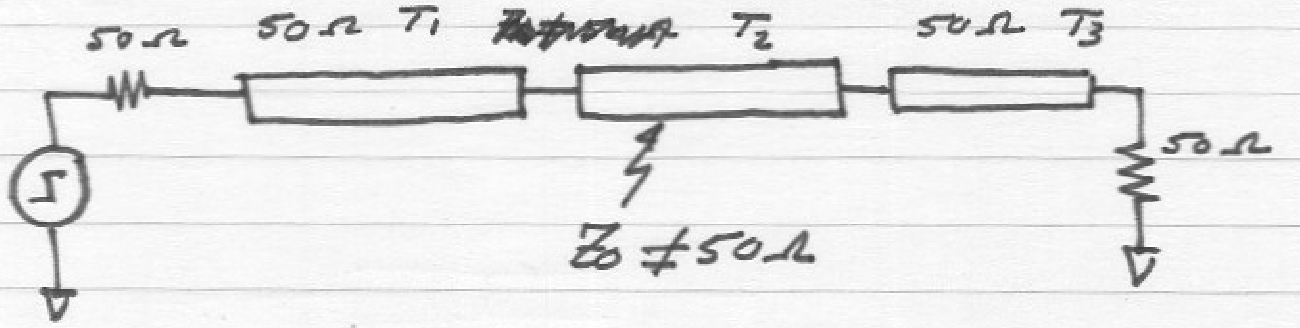
$\tau = (Z_0 + R)\, C$

In using the above responses to characterize circuit components, be warned of the following:



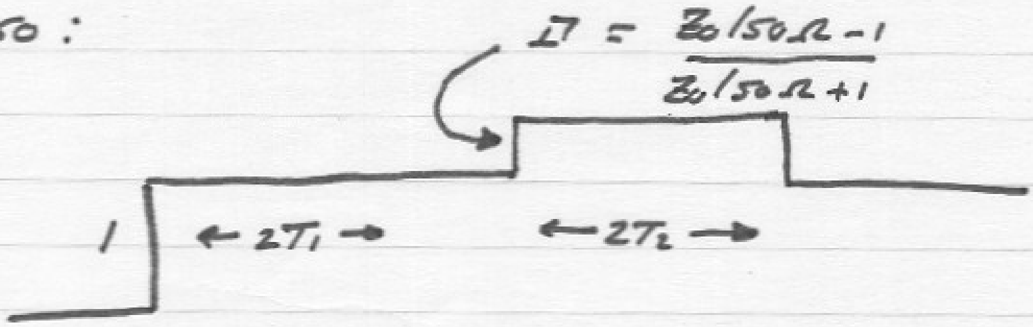
... peak observed reflection magnitude is affected by instrument risetime as well as by the electrical characteristics of the device under test.

TDR of transmission lines:



this can be analyzed *exactly* using a *lattice diagram*. The result is *generally complex*.

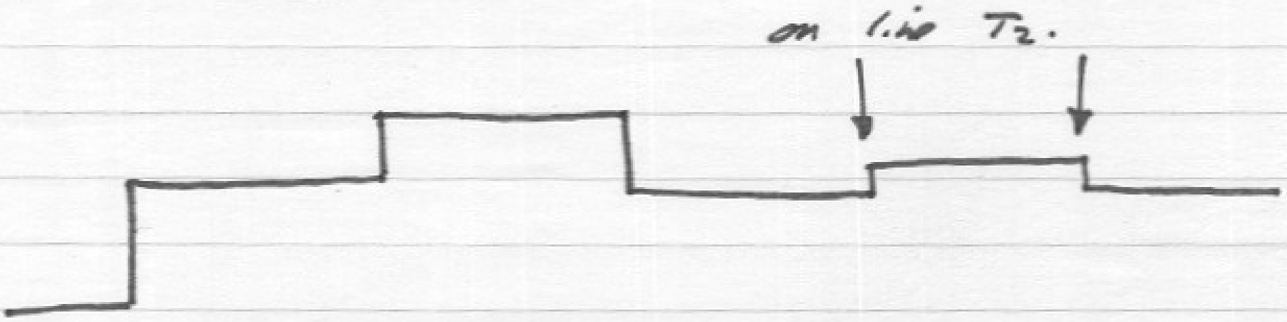
IF Z_o of the second line is reasonably close to 50Ω, then the observed response is like so:



... and the instrument gives us a *plot* *of Z_o as a function of distance*.

If, on the other hand, the line under test has a characteristic impedance far from 50Ω, then the observed response will show significant multiple reflections:

Multiple round trips on line T_2.



.. and the display no longer can be exactly interpreted as a plot of Z_0 vs position.

The key point is that the secondary reflections vary in amplitude to to third order in $(Z_0/50\Omega - 1)$ while the primary reflection varies to first order in $(Z_0/50\Omega - 1)$, so for $Z_0 \approx 50\Omega$, the display is a plot of Z_0 vs position.