Transmission Lines

At high frequencies, transmission lines are superior to *point-to-point* wiring for several reasons:

- Less distortion
- Less radiation (EMI)
- Less crosstalk

However, this comes at a cost: they consume **more power**.

The improved signal integrity and performance makes this worthwhile.

Let's analyze an example *point-to-point* wiring using **wire-wrap**.

The parameters of the system are:

- The average length of a signal net is 4 in.
- The average height above GND of the nets is 0.2 in.
- The wire size is 0.01 in. diameter (AWG 30)
- The signal rise time is 2.0 ns
- F_{knee} is 250 MHz (0.5/2.0 ns).

Signal Distortion in Point-to-Point Wiring

A 2 ns rise time has an electrical length of:

$$l = \frac{\text{rise time (ps)}}{\text{propagation delay (ps/in)}} = \frac{2000 \text{ ps}}{85 \text{ ps/in}} = 23.5 \text{ in.}$$

Is this system **lumped** or **distributed**?

Critical dimension is $l/6 = 3.9$ in.

Is it true that this system will **not** experience ringing (since the average net length is ~4 in.)?

Note that **distributed** systems ALWAYS ring, i.e., have *overshoot* and *undershoot*, unless terminated.

The Q determines if **lumped** systems ring: it measures how quickly signals die out.

Low-Q circuits damp quickly while high-Q circuits cause signals to bounce around.

Signal Distortion in Point-to-Point Wiring

Remember Q definition from chapter 3: ratio of energy stored to energy lost per radian of oscillation.

$$Q \approx \frac{\sqrt{L/C}}{R_S}$$

The following can be used to approximate the maximum overshoot voltage (under assumption that $Q > 0.5$):

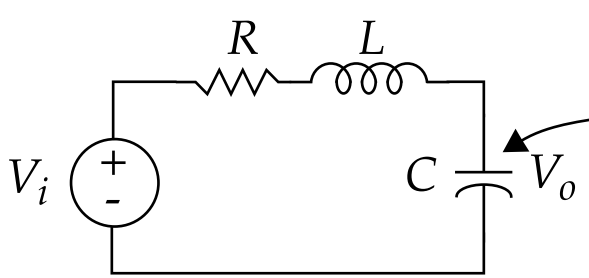
$$\frac{V_{overshoot}}{V_{step}} = e^{-\left[\frac{\pi}{\sqrt{(4Q^2 - 1)}}\right]}$$

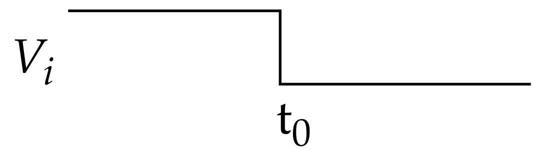
V_{step} = nominal steady-state level

$V_{overshoot}$ = voltage rise above V_{step}

The ideal 2nd order circuit decays with time constant 2L/R

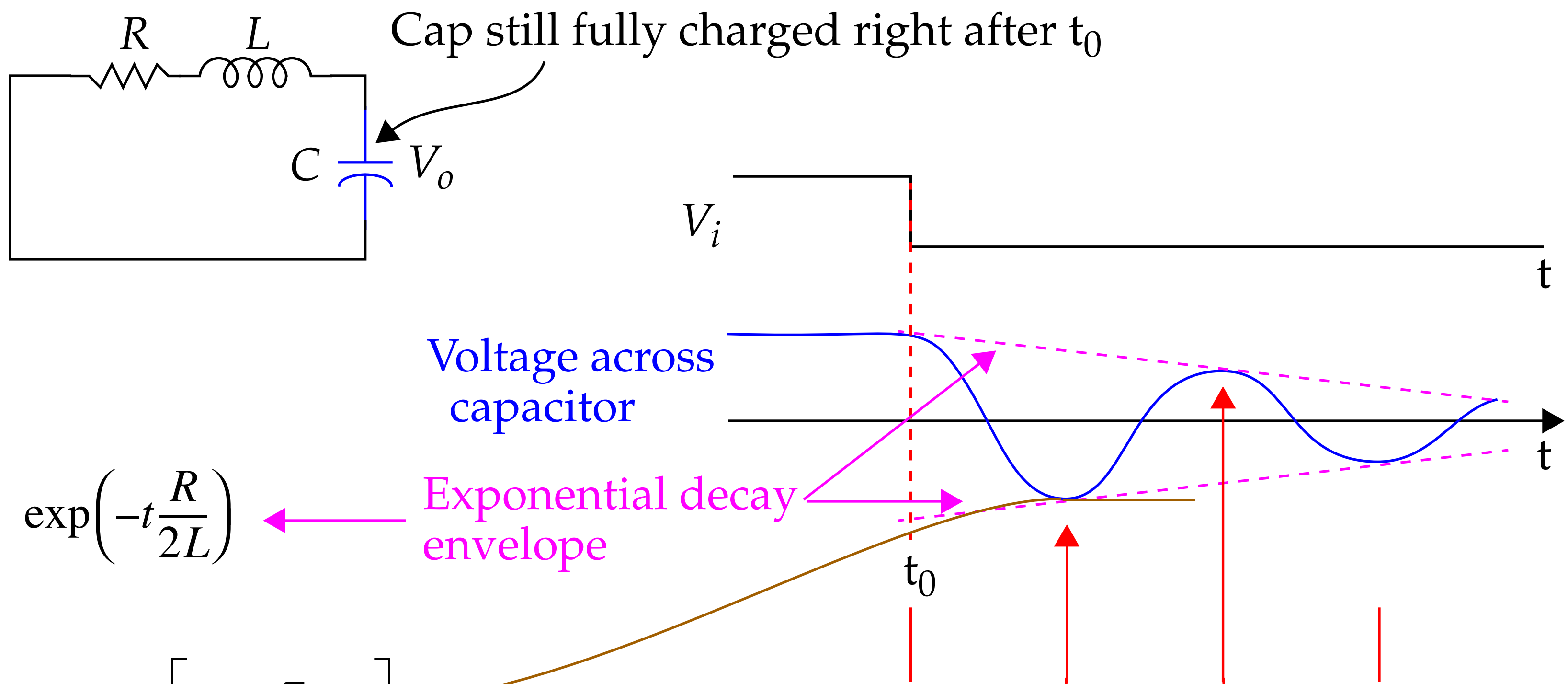
Cap fully charged before t_0





Signal Distortion in Point-to-Point Wiring

After t_0, driver is 0 V.



$$\exp\left(-t\frac{R}{2L}\right)$$

Exponential decay envelope

$$-V_o \exp\left[\frac{-\pi}{\sqrt{4Q^2-1}}\right]$$

Max overshoot

Time of max overshoot

$$\frac{\pi\sqrt{LC}}{\sqrt{1-\frac{1}{Q^2}}}$$

max undershoot

$$\frac{2\pi\sqrt{LC}}{\sqrt{1-\frac{1}{Q^2}}}$$

Ringing period: $2\pi\sqrt{LC}$

Signal Distortion in Point-to-Point Wiring

Rule of thumb:

A digital circuit with a Q of 1, in response to a *perfect step input*, produces a 16% overshoot.

A Q value of 2 displays a 44% overshoot.

A Q less than 0.5 has no overshoot or ringing.

Also note that ringing is related to the natural ringing frequency of the circuit **and the rise time of the driver**.

The basic problem with the example circuit is the high inductance associated with the *point-to-point* wiring.

Large wiring inductance working into a heavy cap load yields a high-Q circuit.

For the example system, the wire inductance is approximated by:

$$L = X(5.08 \times 10^{-9}) \ln\left(\frac{4H}{D}\right) = 4(5.08 \times 10^{-9}) \ln\left(\frac{4 \times 0.2}{0.01}\right) = 89nH$$

Signal Distortion in Point-to-Point Wiring

Then Q can be computed assuming a TTL driver R of 30 Ω:

$$Q \approx \frac{\sqrt{L/C}}{R_S} = \frac{\sqrt{89 \times 10^{-9} / 15 \times 10^{-12}}}{30} = 2.6$$

The expected worst case overshoot voltage with V_{step} 3.7 V (TTL step output) is:

$$V_{overshoot} = V_{step} \exp\left(\frac{-\pi}{\sqrt{4Q^2 - 1}}\right) = 3.7e^{\frac{-3.14159}{\sqrt{4(2.6)^2 - 1}}} = 3.7e^{-0.616} = 2.0V$$

Note that this worst case overshoot **only occurs** if the logic drivers can transmit significant energy at frequencies above the ringing frequency:

$$F_{ring} = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi\sqrt{[89 \times 10^{-9} 15 \times 10^{-12}]}} = 138 MHz$$

The system spec. was 250MHz so full amplitude ringing occurs.

Expect amplitude of ringing to be about 1/2 that predicted if F_{knee} was 138 MHz (i.e., if rise time is about 1/2 the ringing period).

EMI in Point-to-Point Wiring

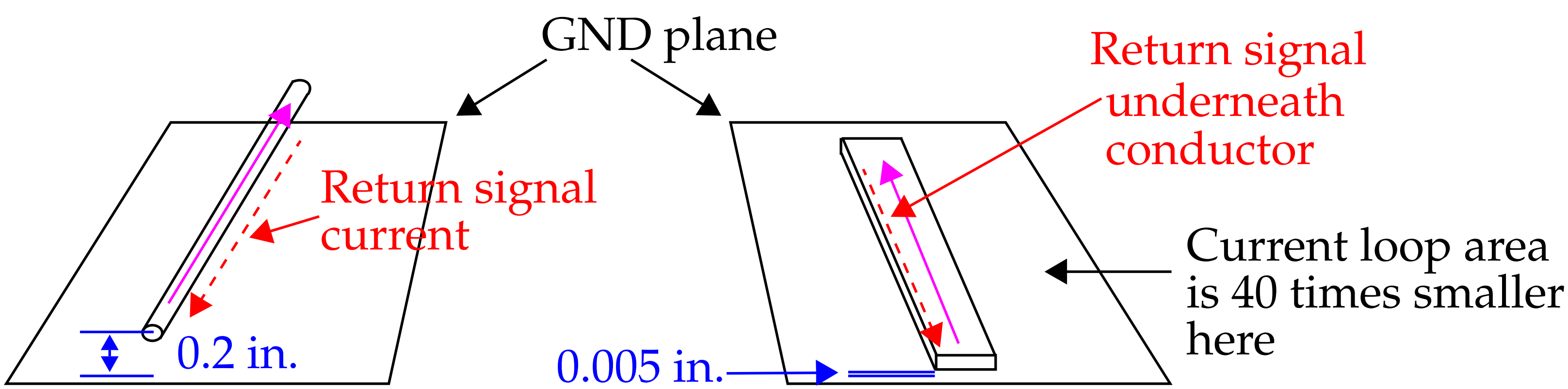
Large *EMI* (electromagnetic interference) fields are generated from large, fast current loops.

The magnetic fields from these loops radiate into space.

Transmission lines dramatically reduce EMI by keeping the return currents **close** to the signal path (magnetic fields cancel).

Wire-wrap current loops return currents at some distance from the signal line, increasing the total loop area.

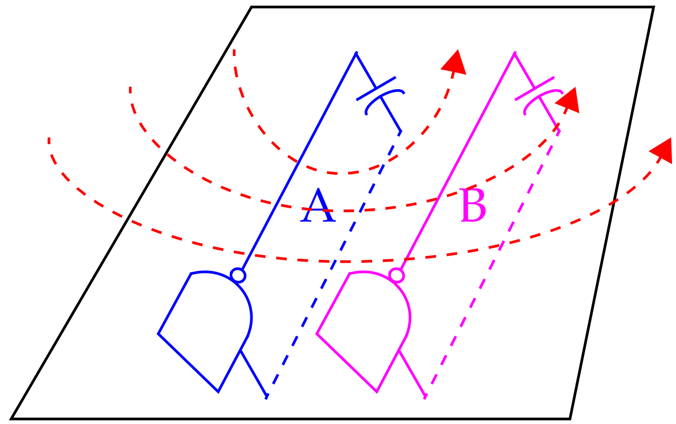
The magnetic fields are directly proportional to the loop area.



EMI is proportional to wire height above GND

Crosstalk in Point-to-Point Wiring

Crosstalk arises from changing magnetic fields, e.g., loop *A* fields penetrate loop *B*:



dI/dt in loop *A* change magnetic flux encircled by loop *B*, introducing a noise voltage in loop *B* called **crosstalk**.

For our example system, assume we have two adjacent loops, 4 in. x 0.2 in. high (h), running parallel at a separation, s, of 0.1 in.

$$L_M = L\left[\frac{1}{1+(s/h)^2}\right] = 71nH$$

The *self-inductance* of each net, L, was computed earlier as 89 nH.

The closeness of these numbers indicates a highly coupled condition.

Crosstalk in Point-to-Point Wiring

The crosstalk is computed by finding the maximum *dI/dt* in loop *A* and multiplying by the mutual inductance, L_M, to obtain a crosstalk voltage.

We analyzed the ringing in loop *A* after the driving gate launches the transition.

Here, we indicated that the maximum overshoot at the **load cap *C*** occurs at one-half the ringing period (7.2 ns/2 = 3.6 ns in our example) after t_0.

Plugging this value for *rise time* (our best quess), into:

$$\frac{dI}{dt} = \frac{1.52 \times \Delta V}{T_r^2} C = \frac{(1.52)(3.7)}{(3.6 \times 10^{-9})^2} 15 \times 10^{-12} = 6.5 \times 10^{6}\ \text{A/s}$$

Yields a crosstalk of about 12%:

$$\text{Crosstalk} = \frac{dI}{dt}(max)L_M = (6.5 \times 10^{6})(71 \times 10^{-9}) = 0.46\ \text{V}$$

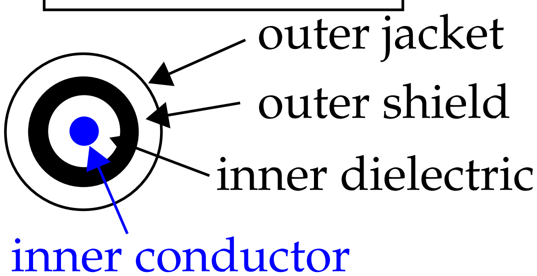
Since crosstalk adds linearly, bundling together 10 or 20 wires to form a bus is a really bad idea.

10 wires easily fits in 1/10 in. -- yields a 50% value for crosstalk!

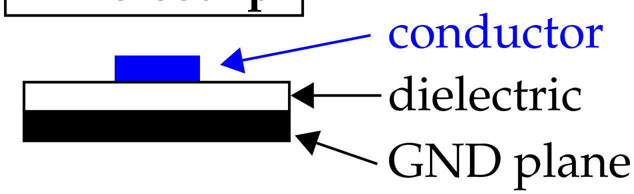
Infinite Uniform Transmission Lines

The transmission line forms that we will study include:

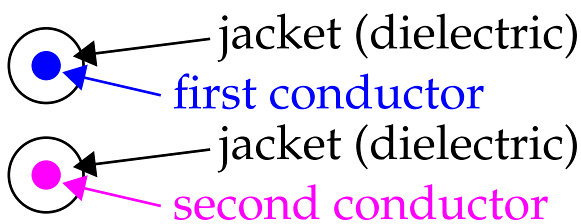
Coaxial cable



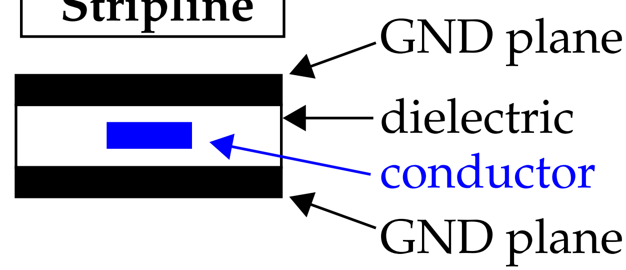
Microstrip



Twisted pair



Stripline



Twisted pair is called balanced, while the others are called *single-ended* or *unbalanced*.

For unbalanced, signal current flows out the signal wire and back along the GND connection.

The GND connection is larger and can be shared with other signal wires.

Infinite Uniform Transmission Lines

An ideal transmission line consists of two **perfect** conductors (zero resistance, uniform in cross-section and extend forever).

Ideal transmission lines have three properties:

- The line is infinite in extent.
- Signals propagate without distortion.
- Signals propagate without attenuation.

The voltage at any point along an ideal transmission line is a perfect delayed copy of the input wfm.

Propagation delay: The delay per unit length along a transmission line in ps/in.

Propagation velocity: Inverse of propagation delay -> in./ps.

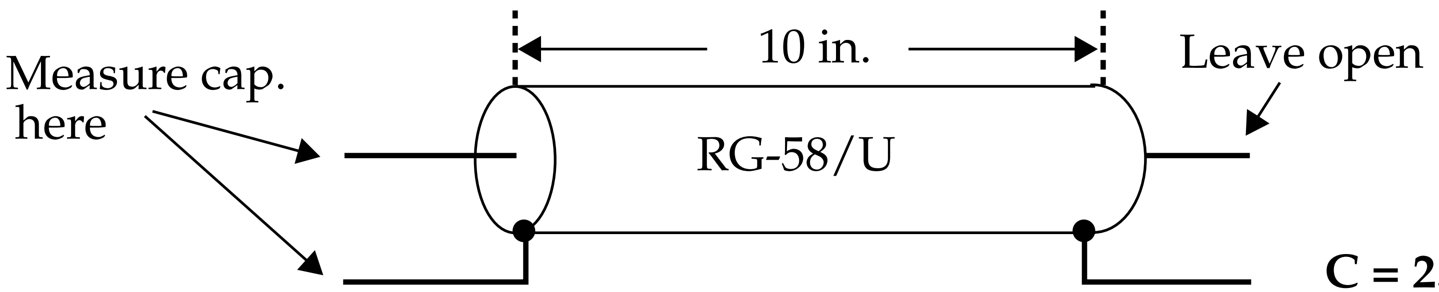
Sometimes expressed as a percentage of the velocity of light in a vacuum, given as 0.0118 in./ps or as delay, 84.7 ps/in.

Infinite Uniform Transmission Lines

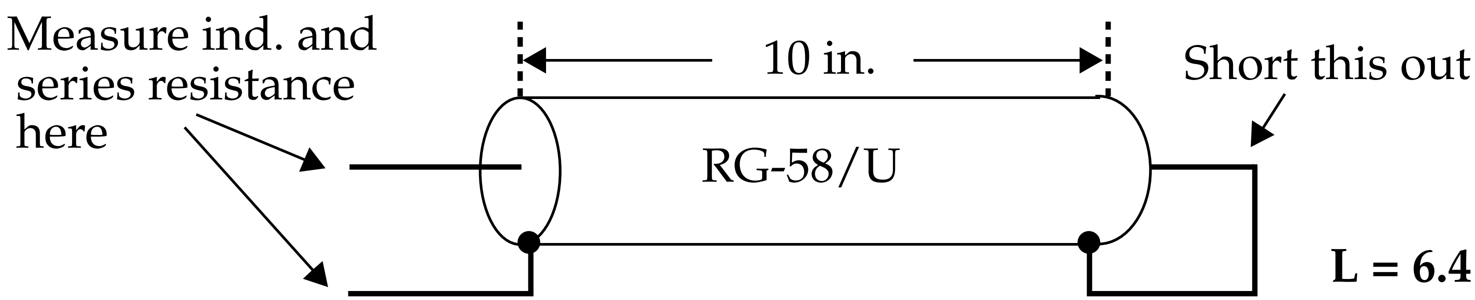
The propagation delay of any transmission line is related to its *series inductance* and *parallel capacitance* per unit length.

The fine balance between these is responsible for distortionless signal propagation.

Measuring transmission line cap. and ind. (need an impedance meter).



C = 2.6 pF/in.



L = 6.4 nH/in.

The series resistance is very small (but not zero), 0.9 mΩ/in.

Characteristic Impedance of Infinite Uniform Transmission Lines

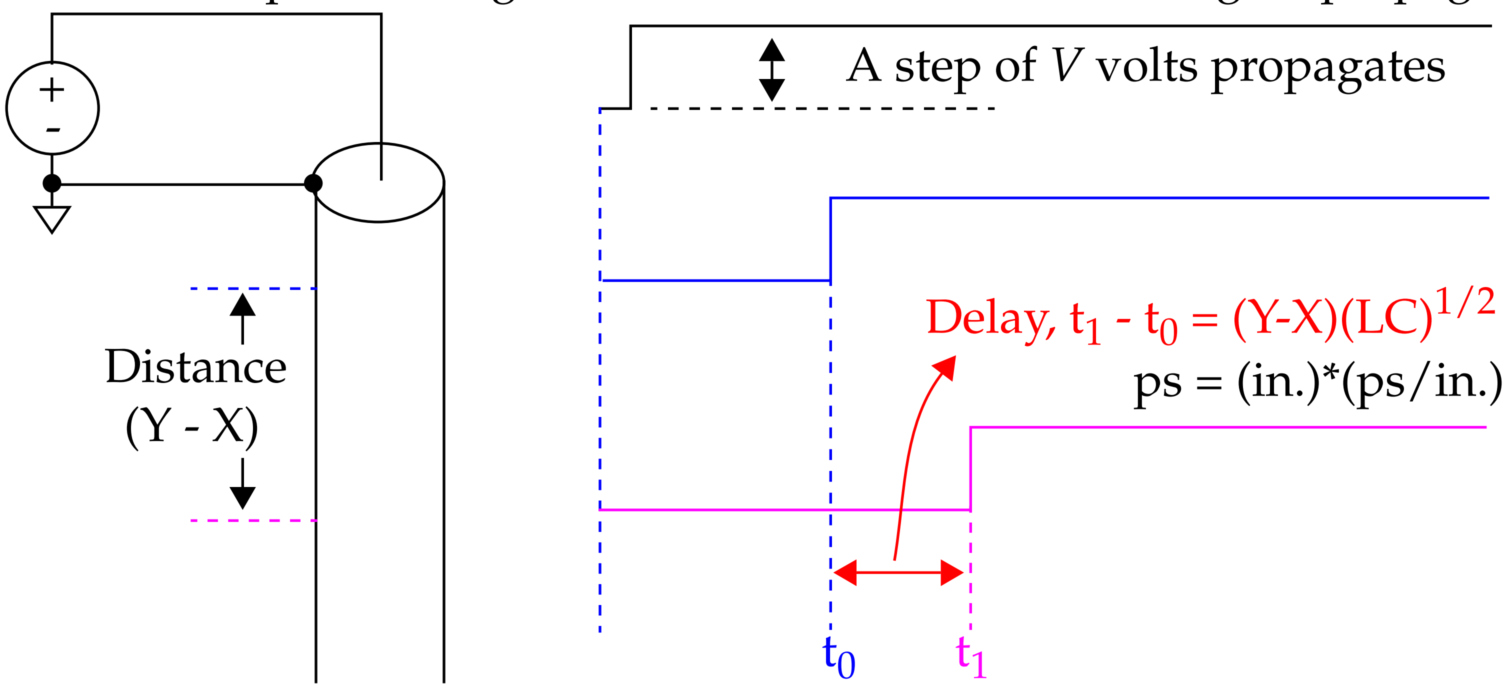
The input impedance is computed from the C/in. and propagation delay.

EM wave theory gives propagation delay as:

$$\text{Delay} = \sqrt{(L/\text{in.})(C/\text{in.})} \times 10^{12} \text{ ps/in.}$$

Given C/in. and prop. delay, we can compute input impedance.

First compute average current needed to sustain the signal propagation:



Characteristic Impedance of Infinite Uniform Transmission Lines

The capacitance between points X and Y charges to voltage V over the time interval T.

To compute the current, first compute the capacitance over this distance:

$$C_{XY} = (C/\text{in.})(Y - X)$$

And then the total charge supplied by the driver is computed from $Q = CV$:

$$Q = C_{XY}V = (C/\text{in.})(Y - X)V$$

The time interval during which C_{XY} must be charged:

$$T = (Y - X)\sqrt{(L/\text{in.})(C/\text{in.})}$$

Then average current is:

$$I = \frac{\text{charge}}{T} = \frac{(C/\text{in.})(Y - X)V}{(Y - X)\sqrt{(L/\text{in.})(C/\text{in.})}}$$

Characteristic Impedance of Infinite Uniform Transmission Lines

This gives us the current flow required to sustain a propagating step edge of V volts.

Solving for R (or Z) in $V = IR$ gives the **characteristic impedance**:

$$Z_0 = \frac{V}{I} = \sqrt{\frac{L/\mathrm{in.}}{C/\mathrm{in.}}}$$

Note that the input impedance is a **constant**, with no imaginary part and independent of frequency.

The constant ratio is a function of the line's physical geometry.

Triax cable is 10 Ω while TV antenna connections are 300 Ω.

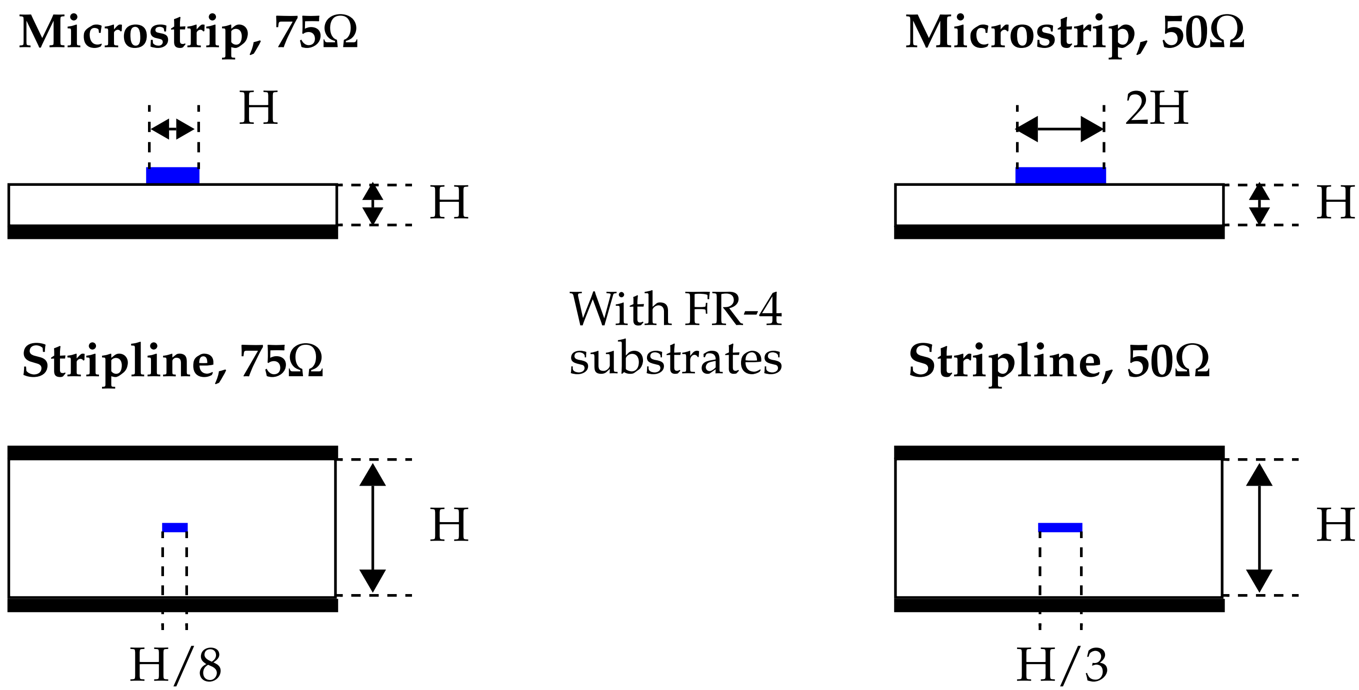
The *characteristic impedance* of the RG-58 cable is:

$$Z_0 = \sqrt{\frac{6.4nH}{2.6pF}} = 50\Omega$$

Characteristic Impedance of Infinite Uniform Transmission Lines

PCB *characteristic impedances* vary from 50 to 75Ω.

The following shows the dimensions needed to build transmission lines on a PCB with characteristic impedances of 50 and 75Ωs.



Note that the tolerance here is +/- 30%!

Characteristic Impedance of Infinite Uniform Transmission Lines

An ideal transmission line looks totally resistive: