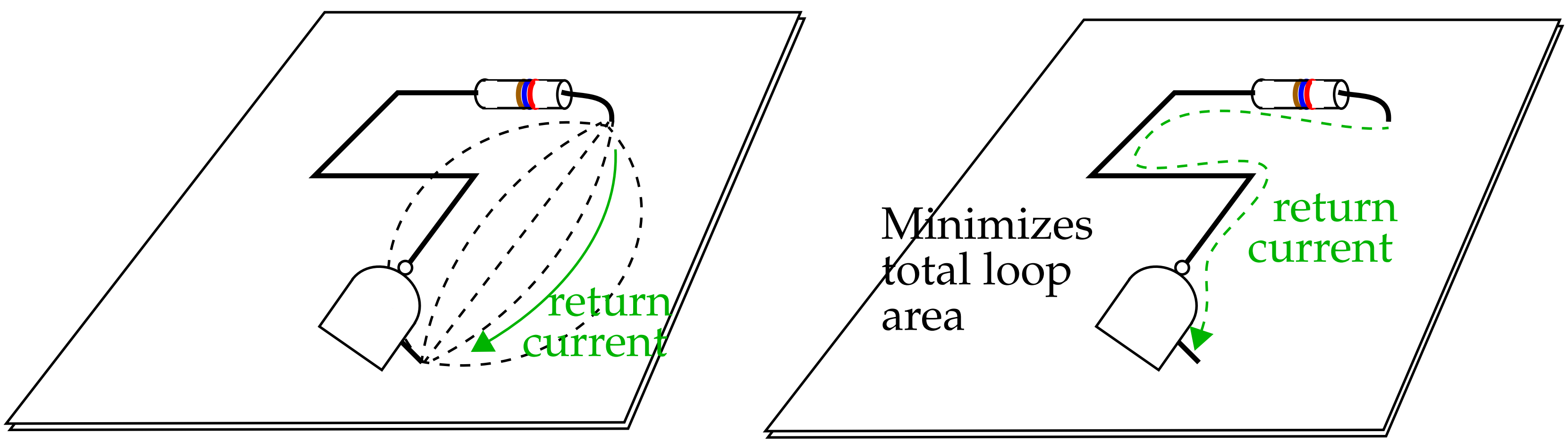
Crosstalk

Ground and power planes serve to:

- Provide stable reference voltages
- Distribute power to logic devices
- Control crosstalk

Here, we derive formulas for computing crosstalk and show how to reduce it using well designed PCB layer stacks.

At low speeds, current follows the path of least resistance



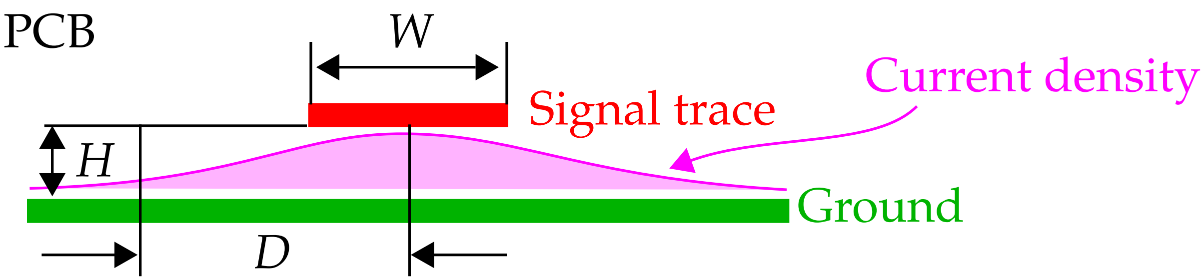
Low frequency High frequency

For high frequency, it follows the path of least inductance.

Crosstalk

The return current distribution:

Cross section of PCB



Current density for a point D inches away from trace.

$$i(D) = \frac{I_0}{\pi H} \frac{1}{1 + (D/H)^2} A/\text{in.}$$

Here, we see a balance between two opposing forces.

Too narrow a distribution increases inductance (skinny wires have more inductance than broad ones) and *too broad* a distribution increases inductance (by increasing loop area).

The real distribution minimizes inductance.

The current distribution given by this equation also minimizes the total energy stored in the magnetic field surrounding the signal wire.

Crosstalk in Solid Ground Planes

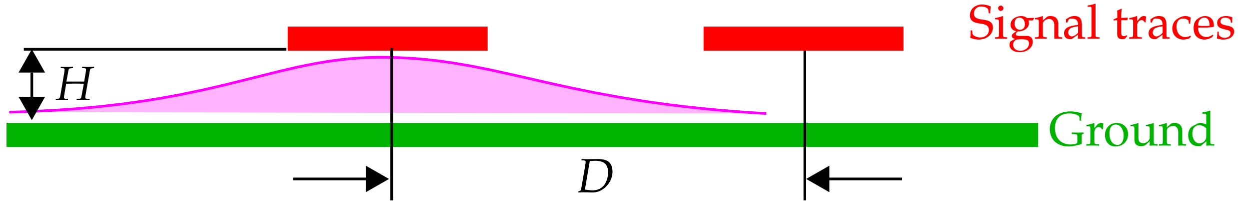
Crosstalk depends on the amount of **mutual** inductance and capacitance.

We focus on *inductive crosstalk* since it is usually as big or bigger than capacitive crosstalk in digital systems.

We covered the theory of mutual inductance coupling earlier.

Returning signal currents generate magnetic fields, which induce voltages in other circuit traces.

The induced voltages are proportional to the *derivative* of the driving signal, e.g., faster rise times produce a more significant effect.



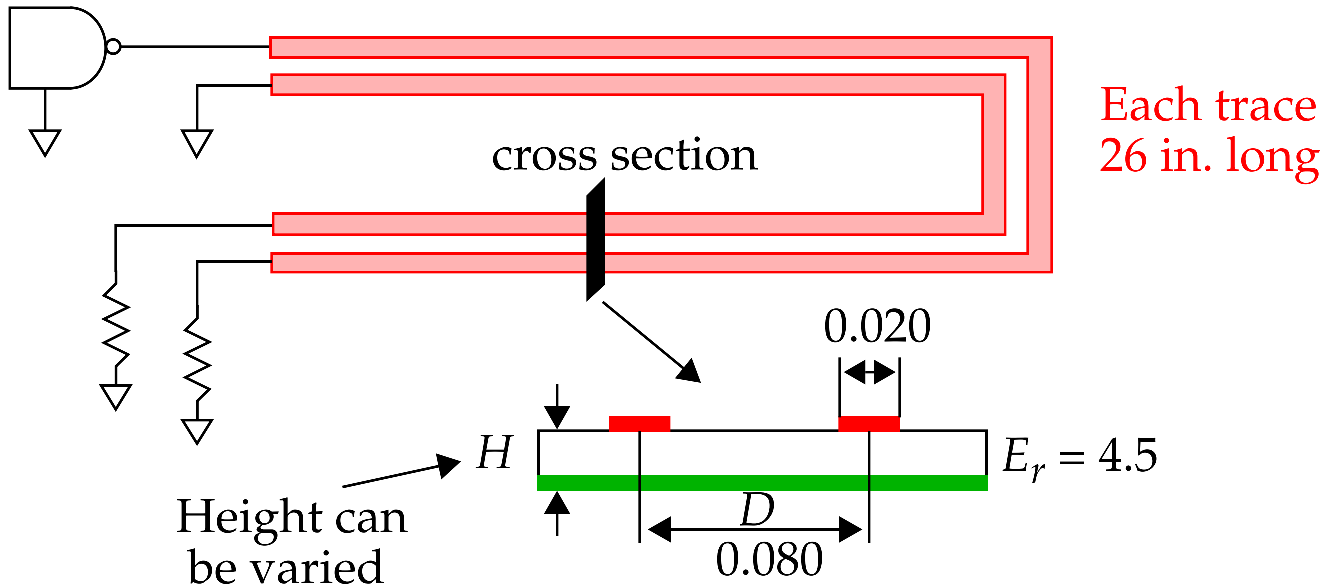
Ratio of measured noise voltage and driving step size

$$\text{Crosstalk} = \frac{K}{1 + (D/H)^2}$$

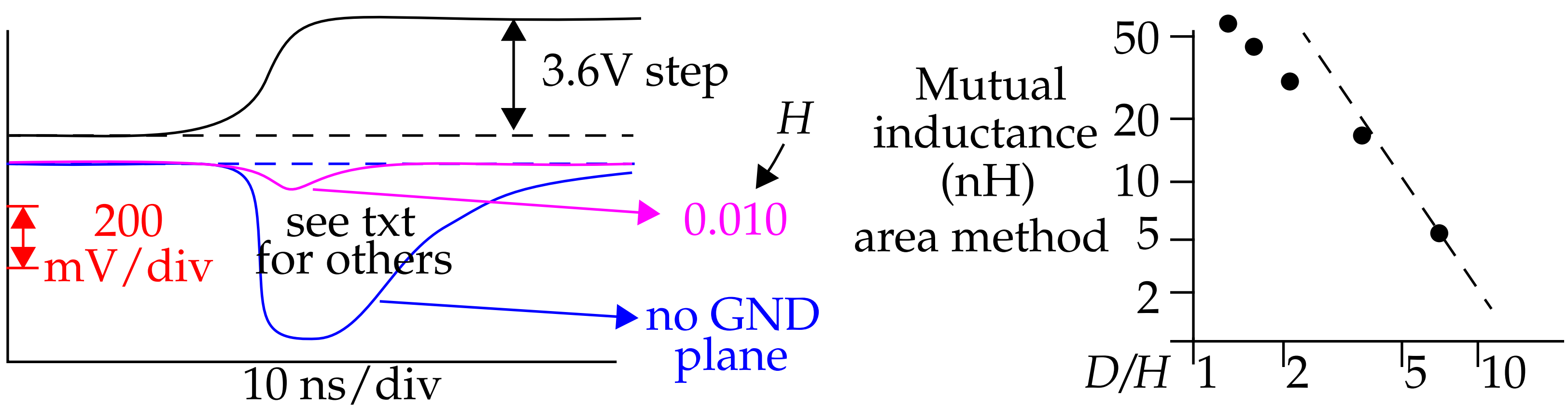
where K depends on the rise time and length of interfering traces (<1).

Crosstalk in Solid Ground Planes

Experiment to test the *D/H* dependency in this equation.

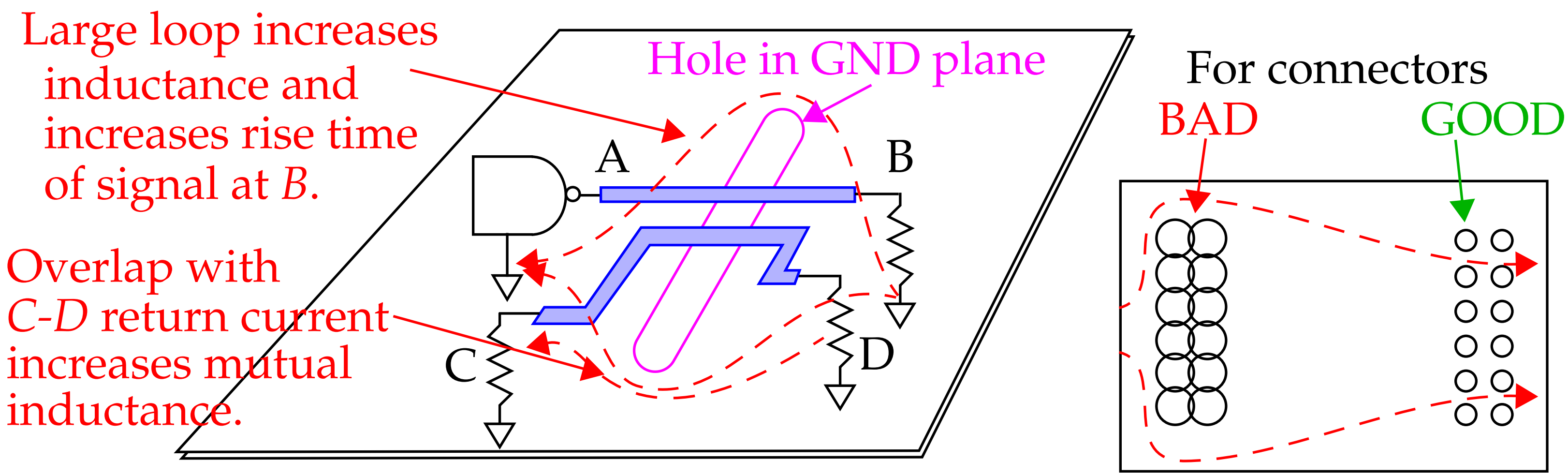


Here, the height *H* can be changed by adding dielectric plates.



Crosstalk in Slotted Ground Planes

Slotted GND planes are a big mistake:



You may be tempted to do this when you run out of room on the regular routing layers and *cram* a trace in on the GND plane.

The effective inductance in series with *A*-*B* is approximated by:

$$L = 5D\ln\left(\frac{D}{W}\right)$$

D = perpendicular extent of current diversion away from signal trace, in.
W = trace width, in.

Inductance is almost completely independent of slot *width*, i.e., making the slot narrow does NOT help.

Crosstalk in Slotted Ground Planes

Rise time degradation depends on the termination conditions.

The *worse case* is a long line with an apparent src resistance on either side of the inductance of Z_0.

The resulting 10-90% rise time of the L/R filter (weighted in with the natural signal rise time) is:

$$T_{\text{10-90 L/R}} = 2.2\frac{L}{2Z_0} \longrightarrow T_{\text{composite}} = \sqrt{(T_{\text{10-90 L/R}})^2 + (T_{\text{10-90 signal}})^2}$$

For a short line driving a heavy capacitive load C, the 10-90% rise time (assuming critical damping) is:

$$T_{10-90} = 3.4\sqrt{LC}$$

The Q of this circuit depends on R_S.

If $Q > 1$, it rings, if Q is near 1, this equation approximates rise time well, if $Q < 1$, the rise time is actually slower than this approximation.

Crosstalk in Slotted Ground Planes

If a second trace, e.g. *C-D* also intersects the slot, the mutual inductance between the two traces is given by the previous inductance formula.

The cross coupling voltage is derived by the mutual inductance and the time rate of change in current in the driver.

$$V_{\text{crosstalk}} = \frac{\Delta I}{T_{10\text{-}90}} L_M$$

For a long line, the ΔI is the drive voltage/*characteristic impedance*.

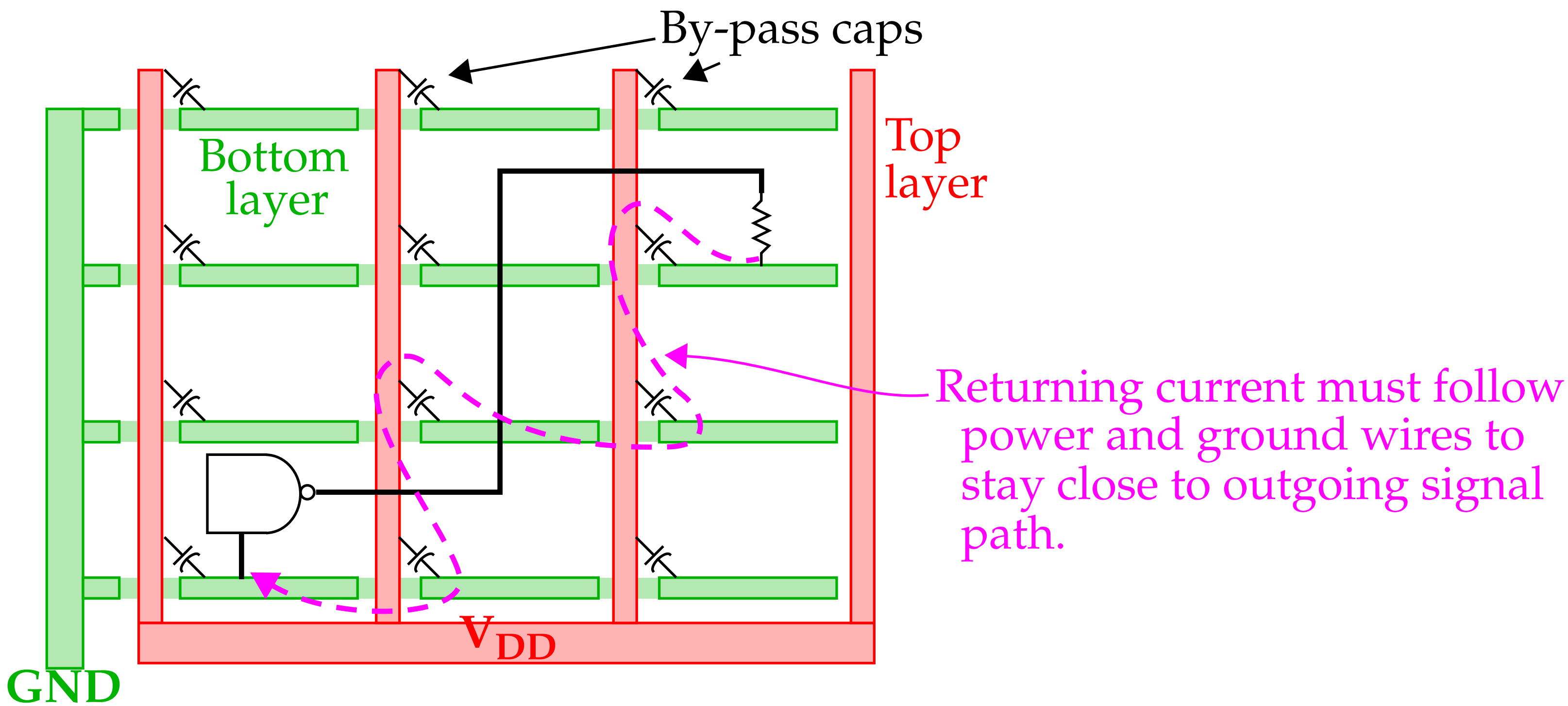
$$V_{\text{crosstalk}} = \frac{\Delta V}{T_{10\text{-}90} Z_0} L_M$$

For short lines driving a large capacitive load, the time rate of change in current is the second derivative of voltage:

$$V_{\text{crosstalk}} = \frac{1.52 \Delta V C}{(T_{10\text{-}90})^2} L_M$$

Crosstalk in Cross-Hatched Ground Planes

If you are forced to use a two layer board:



Note that signals are interleaved with the power and GND routing.

This layout increases *mutual inductance*.

Appropriate for low-speed CMOS and TTL design -- not high-speed logic.

Current returns equally well through the GND and power wires.

Crosstalk in Cross-Hatched Ground Planes

Will your circuit work with the increased mutual inductance (over *solid GND planes*)?

First estimate the **self-inductance**:

$$L \approx 5Y\ln\left(\frac{X}{W}\right)$$

X = hatch width, in.
W = trace width, in.
Y = trace length, in.

Here, *trace length* is the directed distance between the src and driver.

This equation can also be used to estimate the **mutual inductance**, L_M, for a 2nd trace that runs closely to the first.

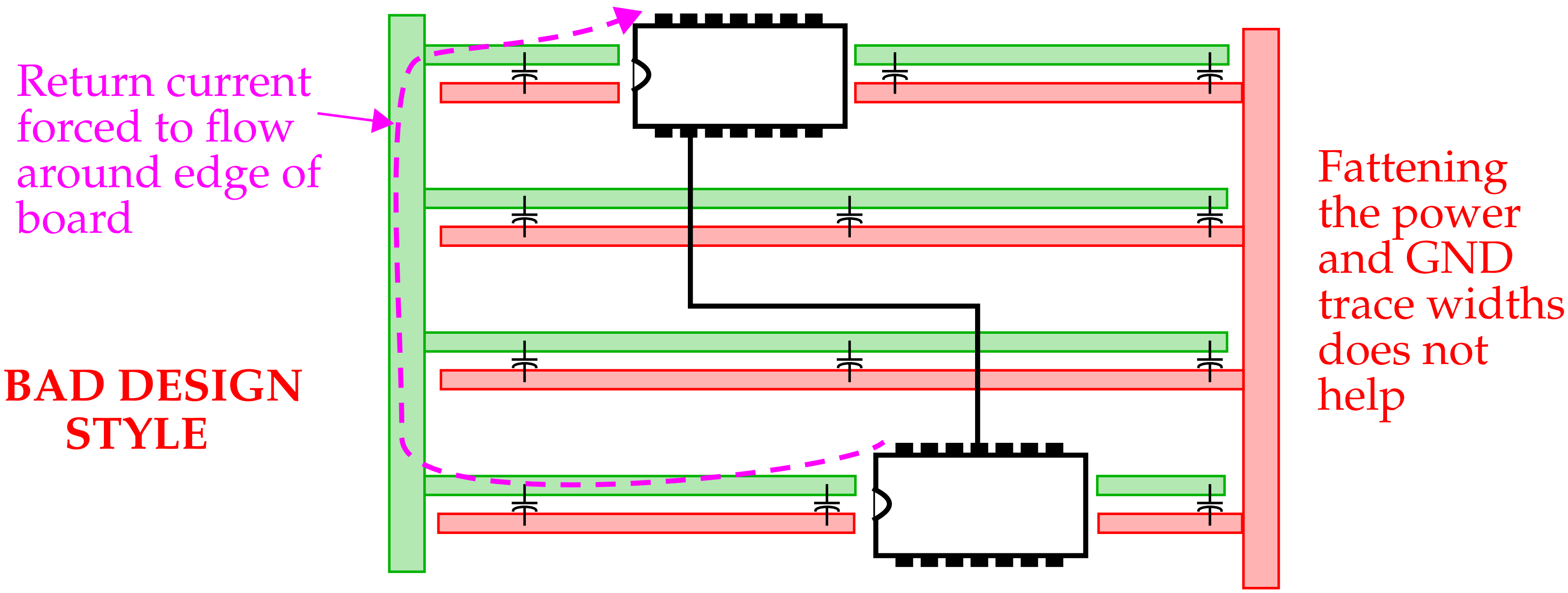
If two traces are separated by a larger distance D, the *mutual inductance* decreases, modeled with D in the denominator as before:

$$L_M = \frac{5Y\ln(X/W)}{1+(D/X)^2}$$

Rise time degradation and cross-talk voltage computed as before.

Crosstalk with Power and Ground Fingers

This scheme saves even more board area:



Both power and GND are routed on the same layer.

The diversion of the current introduces a huge amount of *self* and *mutual inductance*, both modeled as.

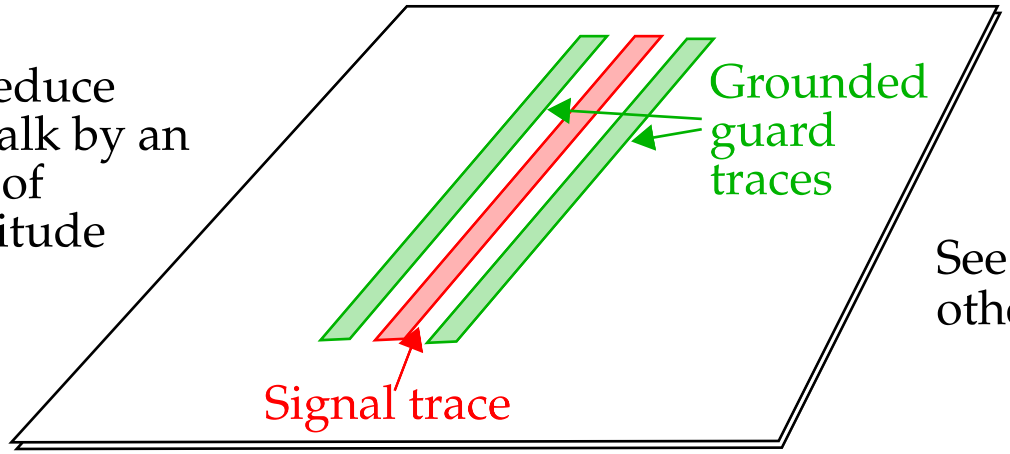
$$L \approx 5Y \ln\left(\frac{X}{W}\right)$$

X = board width!
W = trace width
Y = trace length

Guard Traces

These are used a lot in analog design:

Can reduce crosstalk by an order of magnitude



See text for other illustrations

However, for digital, *solid GND planes* provide most of the benefits of guard traces (guard traces help very little).

As a rule of thumb, coupling between microstrip signal lines is *cut in half* by inserting a third line (GNDed at both ends) between them.

Coupling is halved yet again if the third line is *frequently* tied (through vias) to the GND plane.

In digital with a solid GND plane, the reduction in crosstalk is not significant even if it's possible to insert a *guard* trace between two traces.

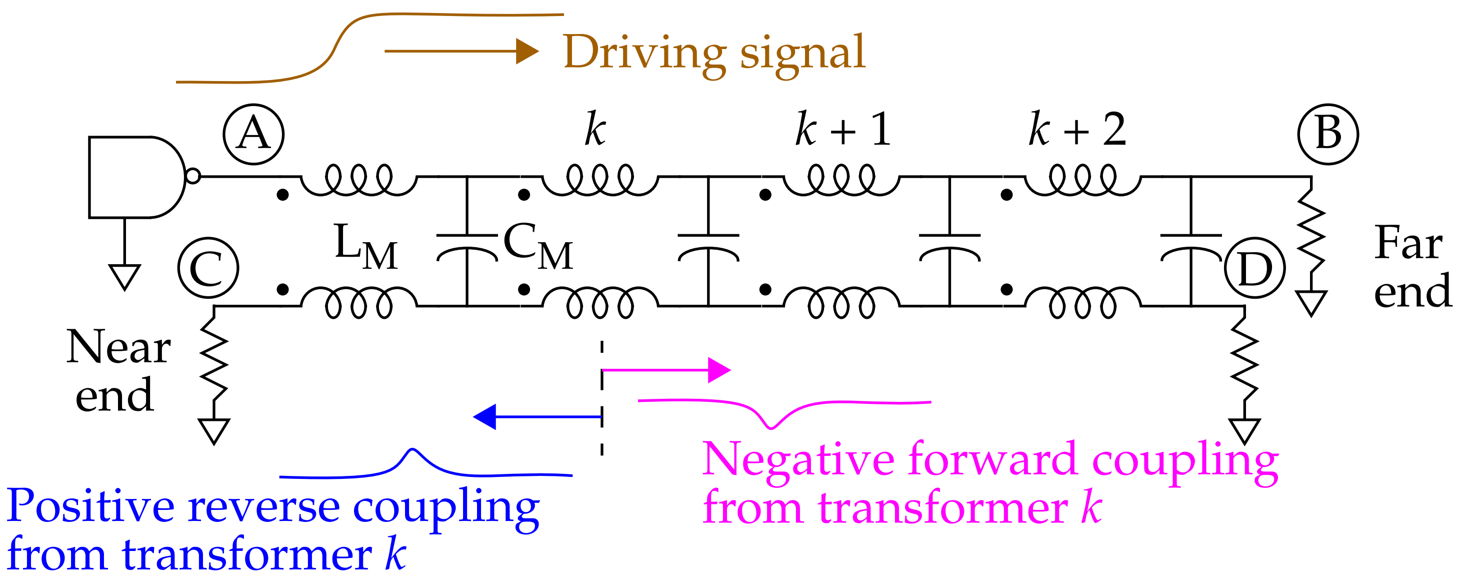
Near-End and Far-End Crosstalk

The crosstalk examples considered thus far use a *lumped* circuit model.

This inductive coupling model does not work well for long lines.

Coupling between long, distributed transmission lines involves both *mutual inductance* and *mutual capacitance*.

Let's consider *mutual inductive* coupling first.



Wire *A-B* carries the signal whose magnetic fields induce voltages on wire *C-D*.

Near-End and Far-End Crosstalk

Mutual inductance normally acts like a *transformer*.

The **distributed** nature of the inductance makes it appear as a sequence of transformers.

Under the assumption that the coupling is small, the effect on the propagation of signal *A*-*B* is small.

At each transformer, a small blip is produced on the adjacent line that propagates both forward and backwards.

Consider transformer *k*. Upon arrival of the step, the inductor on line *A*-*B* produces:

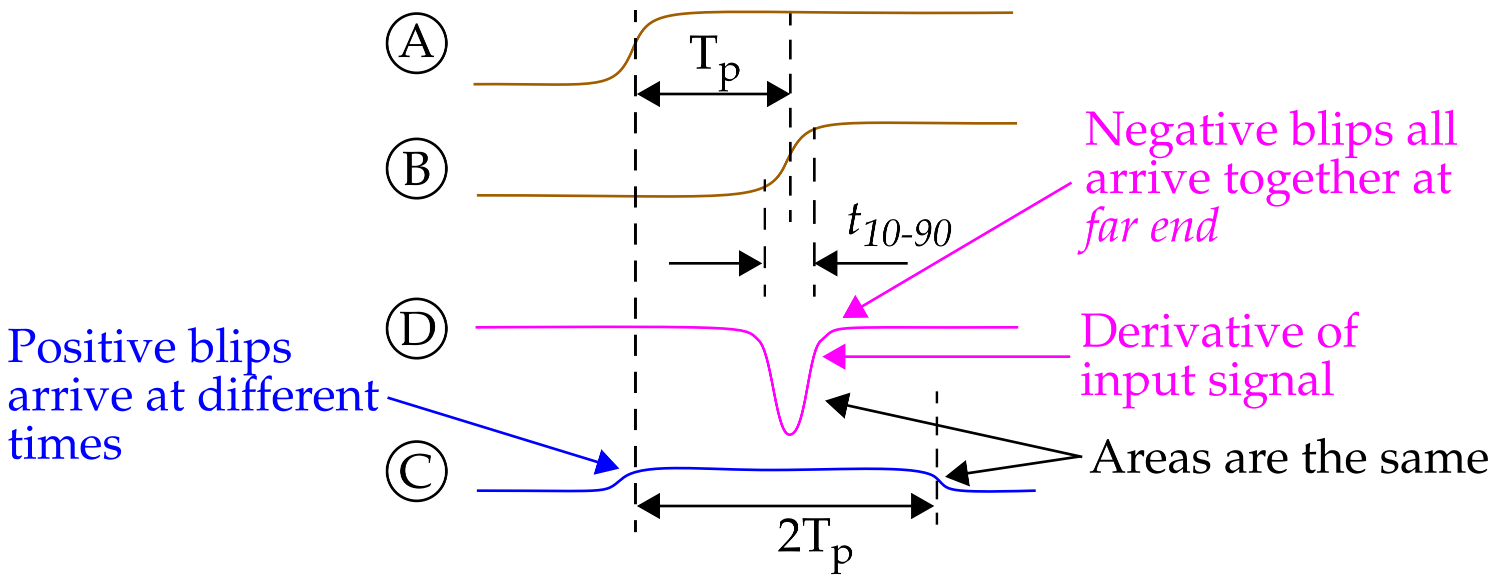
$$V_{\text{reverse}}(t) = L_M \frac{di}{dt}$$

The transformer reproduces this blip on line *C*-*D* with the polarity indicated.

Positive blip to the left, negative blip to the right.

Near-End and Far-End Crosstalk

Forward and reverse mutual inductance coupling:



The *negative* blip size at the far end is proportional to the **total** mutual inductance (sum of L_Ms).

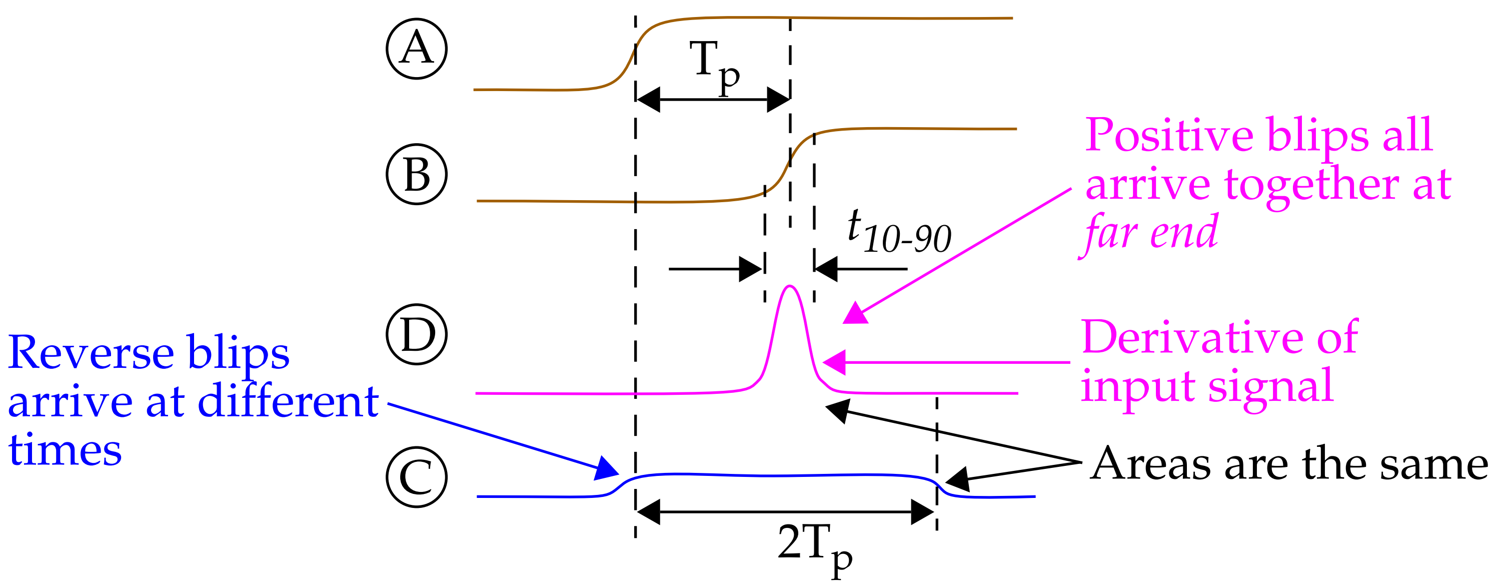
Lengthening the line increases the size of the blip at *D*.

Note that the "height" of the *positive* reverse coupling peak does not increase with a longer line, only its duration.

Near-End and Far-End Crosstalk

Now let's consider *mutual capacitance* coupling.

It is very similar to *mutual inductance* (blips propagate in both directions along *C*-*D*) except for the polarity of the forward blip:



The two *forward* effects cancel (*D*) while the *reverse* effects reinforce (*C*).

Striplines are well balanced between inductive and capacitive couplings.

Microstrip electric field lines travel through air (capacitive coupling smaller), yielding a small negative *forward coupling coefficient*.

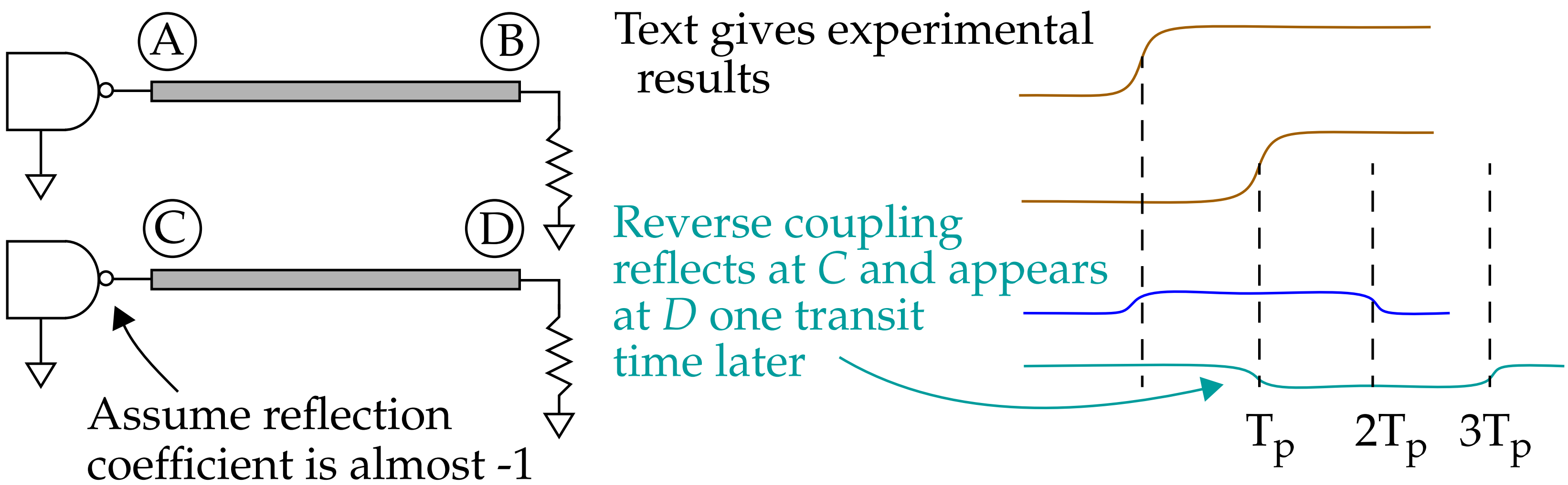
Near-End and Far-End Crosstalk

The inductive component is much **larger** over slotted, hatched or finger type GND plane arrangements.

The forward crosstalk component is never larger than the reverse crosstalk.

The model just analyzed assumes the forward and reverse crosstalk propagate and then terminate at *C* and *D*.

In digital applications without src terminations, the driver connected on the left end of the wire is usually low impedance.



Characterizing Crosstalk Between Two Lines

Forward crosstalk is proportional to the *derivative* of the driving signal and to the line length.

The constant of proportion depends on the balance between inductive and capacitive coupling.

Reverse crosstalk looks like a square pulse, with rise/fall times comparable to the input signal and height proportional to the driving signal amplitude.

The amplitude is **independent** of the line length and depends on the line parameters.

The duration of the pulse is $2T_p$.

This model holds for fast edges, but once known for fast edges, a model for slow edges (or any signal) can be derived:

$$\text{Reverse coupling}(t) = \alpha_R[V(t) - V(t - 2T_p)]$$

with $V(t)$ representing the driving wfm, α_R the reverse coupling coefficient for fast-edged signal, and T_p is the propagation delay of the line.

Characterizing Crosstalk Between Two Lines

For lines longer than half the signal rise time, the reverse coupling has time to build up to its full value.

In this case, the reverse coupling coefficient equals:

$$\alpha_R = \frac{1}{1 + (D/H)^2}$$

D = separation between lines
H = line height above GND plane

For shorter lines, the reverse coupling ramps up and then back down, never reaching the maximum value.

A **series termination** eliminates reverse-coupled crosstalk while an **end termination** attenuates the returning refl ection of the main signal.

The use of both types of terminations is the best approach.