EMC Course Notes 2024

Crosstalk

Crosstalk in Transmission Lines

Crosstalk is a term commonly used to describe the unintentional electromagnetic coupling between two circuits or systems in close proximity. We have been using the formula,

$$\text{crosstalk in dB} = 20\log\left|\frac{\text{coupled voltage appearing at receiver in Circuit 2}}{\text{signal voltage in Circuit 1}}\right| \tag{6.1}$$

to quantify this coupling, but it is important to note that there is no standard method for quantifying or calculating crosstalk. The expression above is convenient when describing the crosstalk between two identical linear circuits in the frequency domain. However, it can be problematic when applied to time domain waveforms where very different results are obtained depending on whether the peak, peak-to-peak, or rms voltage is used. Also, the use of decibel notation is questionable when the voltage in the numerator and the voltage in the denominator appear across different impedances.

In this section, which deals with crosstalk in transmission lines, we will continue to use Equation (6.1) to quantify crosstalk between two identical circuits in the frequency domain. For time-domain crosstalk calculations, and for crosstalk in circuits with different impedances, we will simply calculate coupled voltages.

Electrically Short Lines

Transmission lines that are short relative to a quarter-wavelength can generally be modeled using lumped-element parameters. In Chapter 3, the crosstalk for several different circuit configurations was calculated. Here, we'll calculate the crosstalk between the pair of microstrip traces illustrated in Figure 6.1.

In this example, both traces carry signals from a 50-Ω source to a 50-Ω load. The trace width, height and separation are 1.4 mm, 0.8 mm, and 1 mm, respectively. The circuit board plane length and width are 204 mm and 56 mm, respectively. The copper ($\sigma = 5.7 \times 10^7$ S/m) is 17.4 μm thick. The effective relative permittivity of the dielectric is 4.2.

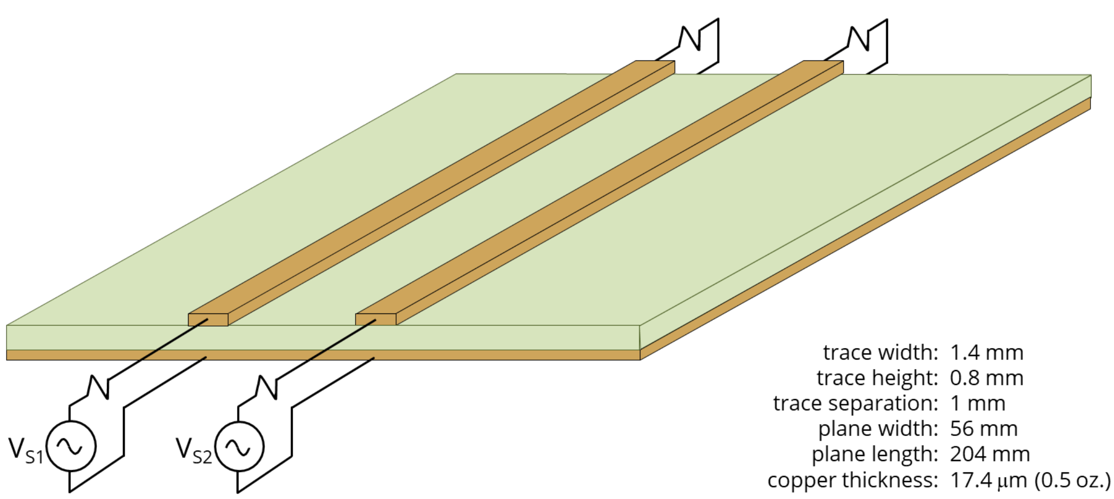

Figure 6.1. Two signal traces sharing a circuit board return plane.

At frequencies where the traces are short relative to a quarter wavelength, the voltage and current on the wires is approximately constant. The traces can be modeled using lumped impedance parameters. For these calculations, we will make a *weak-coupling assumption*. That is, we will assume the coupling to the victim circuit does not significantly affect the voltages and currents in the source circuit. Virtually all practical cases of unintentional crosstalk involve weak coupling. However, in cases where a weak-coupling assumption is made and the calculated crosstalk is greater than about -20 dB, a more precise calculation may be required.

Crosstalk due to Common-Impedance Coupling

Figure 6.2 shows a schematic representation of the two trace circuits and their shared return plane resistance. The crosstalk is the ratio of the induced voltage in the victim circuit (TRACE 2) due to a signal voltage in the source circuit (TRACE 1),

$$\text{XTALK(dB)} = 20\ Log\left(\frac{V_{SIG2}}{V_{SIG1}} \Big|_{V_{S2}=0} \right). \tag{6.2}$$

The voltage dropped across the ground plane is approximately equal to the plane resistance times the Circuit 1 current, $V_G = I_1 R_G = \left(\frac{V_{SIG1}}{R_{L1}} \right) R_G$. A fraction of this voltage $\left(\frac{R_{S2}}{R_{S2}+R_{L2}} \right)$ appears across the Circuit 2 source impedance and the rest $\left(\frac{R_{L2}}{R_{S2}+R_{L2}} \right)$ appears across the Circuit 2 load impedance. The voltage appearing at the load end contributes to the *far-end* crosstalk, which is given by,

$$\text{XTALK}_{\text{FE}} = 20\ \log\left| \frac{V_{RL2}}{V_{RL1}} \right| = 20\ \log\left| \frac{R_G}{R_{S2}+R_{L2}} \frac{R_{L2}}{R_{L1}} \right|. \tag{6.3}$$

The voltage appearing across the source end contributes to the *near-end* crosstalk, which is given by,

$$\text{XTALK}_{\text{NE}} = 20\ \log\left| \frac{V_{RS2}}{V_{RL1}} \right| = 20\ \log\left| \frac{R_G}{R_{S2}+R_{L2}} \frac{R_{S2}}{R_{L1}} \right|. \tag{6.4}$$

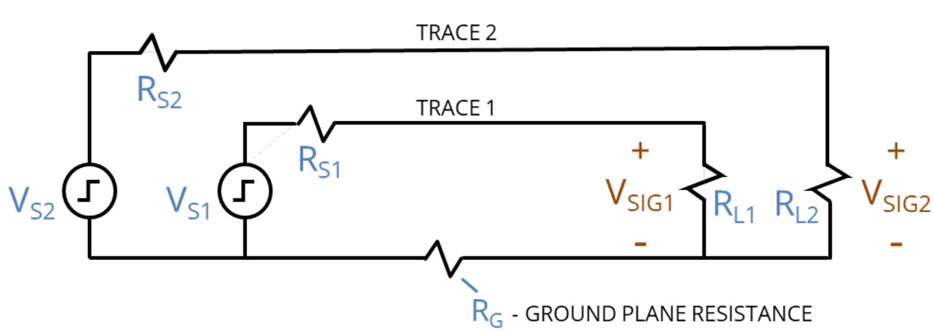

Figure 6.2. Schematic representation of traces with comment return resistance.

In this case, the resistance of the half-ounce copper return plane is,

$$R_G = \frac{\ell}{\sigma A} = \frac{0.204\text{ m}}{\left(5.7\times10^{7}\text{ S/m}\right)\left(0.056\text{ m}\times17.4\times10^{-6}\text{ m}\right)} = 3.7\times10^{-3}\ \Omega. \tag{6.5}$$

The source and load impedances are all 50 Ω, so the calculated crosstalk due to common-impedance coupling is,

$$\text{XTALK} = 20\log\left|\frac{\left(3.7\times10^{-3}\ \Omega\right)}{\left(50\ \Omega+50\ \Omega\right)}\frac{50\ \Omega}{50\ \Omega}\right| \approx -89\text{ dB}. \tag{6.6}$$

The source and load impedances are the same, so the near-end and far-end crosstalk are the same magnitude. However, it's worth noting that the coupled voltages at each end have opposite polarities.

In this example, the amplitude of the crosstalk is independent of frequency. This is generally true for common-impedance coupling except when the shared impedance is a function of frequency. At frequencies where the skin depth determines the value of the shared resistance, common-impedance coupling is generally proportional to the square root of the frequency (i.e., increases at a rate of 10 dB/decade). If the shared impedance is a small, lumped inductance or capacitance, the coupling is generally proportional to frequency. However, in most cases, it's better to treat shared capacitances or inductances as electric-field or magnetic-field coupling, respectively.

A few properties of weak common-impedance coupling between electrically short transmission lines that are worth noting include the following.

- The coupling is proportional to the source circuit current.
- The coupling is usually independent of frequency.
- The near- and far-end coupling are 180^0 out of phase.

Crosstalk due to Electric-Field (Capacitive) Coupling

Figure 6.3 shows a schematic representation of electric-field coupling between the two trace circuits. The electric-field coupling can be quantified using the mutual capacitance between the two traces.

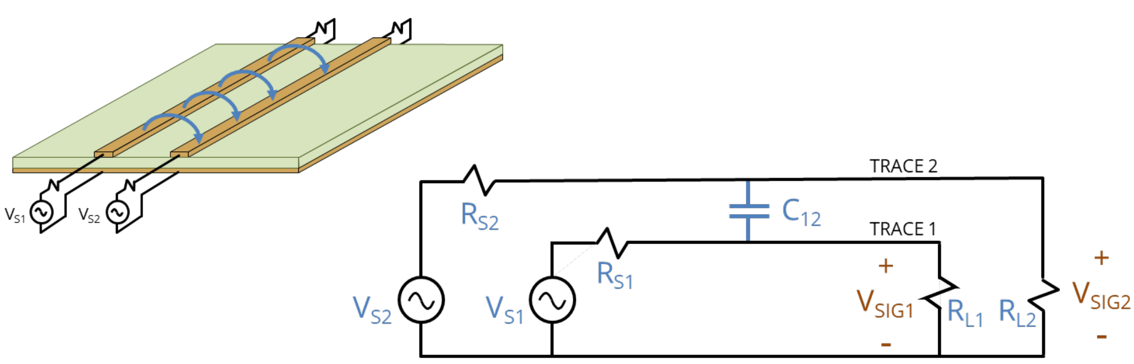

Figure 6.3. Schematic representation of electric-field coupling between traces.

The crosstalk for this circuit was calculated in Chapter 3,

$$\text{XTALK} = 20\log\left|\frac{R_{S2} \parallel R_{L2}}{R_{S2} \parallel R_{L2} + \left(1/j\omega C_{12}\right)}\right|. \tag{6.7}$$

Or, making a weak coupling assumption,

$$\text{XTALK} \simeq 20\log\left|\omega\left(R_{S2} \parallel R_{L2}\right)C_{12}\right|. \tag{6.8}$$

For the two traces illustrated in Figure 6.1, the mutual capacitance is 2.84 pF. Therefore, the calculated crosstalk at 10 MHz is,

$$\begin{aligned}\text{XTALK}_{10\,\text{MHz}} &= 20\log\left[\omega\left(R_{S2} \| R_{L2}\right)C_{12}\right] \\ &= 20\log\left[\left(2\pi\times 10^{7}\ \text{Hz}\right)\left(50\ \Omega \| 50\ \Omega\right)\left(2.84\times 10^{-12}\ \text{F}\right)\right] \\ &= 20\log\left[4.46\times 10^{-3}\right] \\ &= -47\ \text{dB}.\end{aligned} \tag{6.9}$$

Note that the crosstalk is proportional to frequency, so at 20 MHz, it would be 6 dB higher (-41 dB). This is considerably stronger than the common-impedance coupling previously calculated for the same geometry (-89 dB). In this example, the electric-field coupling is higher than the common-impedance coupling at any frequency above about 8 kHz.

A few properties of weak electric-field coupling between electrically short transmission lines that are worth noting include the following.

- The coupling is proportional to the source circuit voltage.
- The coupling is proportional to frequency.
- The near- and far-end coupling are the same.

Crosstalk due to Magnetic-Field (Inductive) Coupling

Figure 6.4 shows a schematic representation of magnetic-field coupling between the two trace circuits. The magnetic-field coupling can be quantified using the mutual inductance between the two trace circuits.

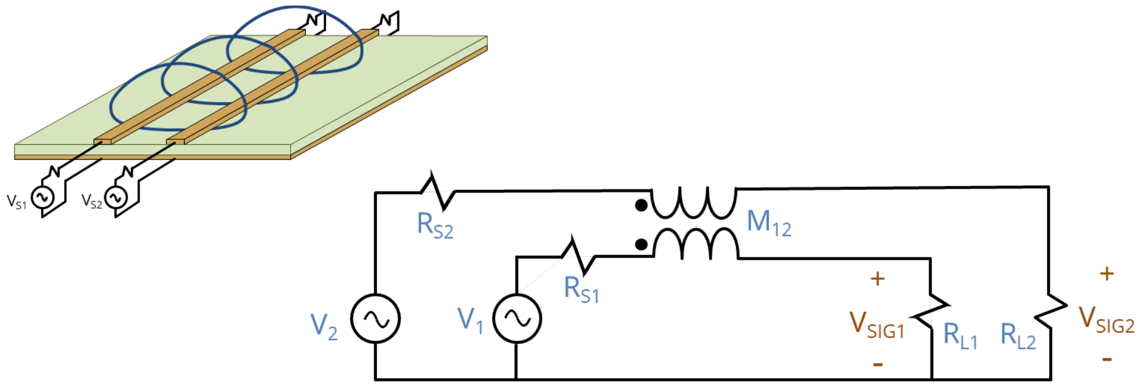

Figure 6.4. Schematic representation of magnetic-field coupling between traces.

The far-end crosstalk for this circuit making a weak-coupling assumption was calculated in Chapter 3,

$$\mathrm{XTALK_{FE}} = 20 \log \left| \frac{\omega M_{12}}{R_{L1}} \left(\frac{R_{L2}}{R_{L2} + R_{S2} + j\omega L_{22}} \right) \right| , \tag{6.10}$$

where M_{12} is the mutual inductance between the two circuits and L_{22} is the self-inductance of the victim circuit. In most practical circuits, $\omega L_{22} \ll \left(R_{L2} + R_{S2} \right)$ and that term can be neglected. Note that the total voltage coupled to the victim loop is $\frac{\omega M_{12}}{R_{L1}}$. The term in parentheses expresses the fraction of the total voltage that appears across R_{L2}. To calculate the near-end crosstalk, R_{S2} would replace R_{L2} in the numerator of that expression.

$$\mathrm{XTALK_{NE}} = 20 \log \left| \frac{\omega M_{12}}{R_{L1}} \left(\frac{R_{S2}}{R_{L2} + R_{S2} + j\omega L_{22}} \right) \right| . \tag{6.11}$$

For the two traces illustrated in Figure 6.1, the mutual inductance is 16.4 nH. Therefore, the calculated far-end crosstalk at 10 MHz is,

$$\begin{aligned} \mathrm{XTALK_{10\,MHz}} &= 20 \log \left[\left(\frac{\omega M_{12}}{R_{L1}} \right) \left(\frac{R_{L2}}{R_{S2} + R_{L2}} \right) \right] \\ &= 20 \log \left[\left(\frac{\left(2\pi \times 10^{7}\ \mathrm{Hz}\right)\left(16.4 \times 10^{-9}\ \mathrm{H}\right)}{50\ \Omega} \right) \left(\frac{50\ \Omega}{50\ \Omega + 50\ \Omega} \right) \right] \\ &= 20 \log \left[1.03 \times 10^{-2} \right] \\ &= -39.7\ \mathrm{dB}. \end{aligned} \tag{6.12}$$

Note that the crosstalk is proportional to frequency, so at 20 MHz, it would be 6 dB higher (-33.7 dB). This is about 7 dB higher than the electric-field coupling. In matched transmission lines with a homogeneous dielectric, the electric-field coupling and the magnetic-field coupling always have the same magnitude. However, in matched microstrip transmission lines such as this, the magnetic field coupling dominates.

A few properties of weak magnetic-field coupling between electrically short transmission lines that are worth noting include the following.

- The coupling is proportional to the source circuit current.
- The coupling is proportional to frequency.
- The near- and far-end coupling are 180^0 out of phase.

Crosstalk due to All Coupling Mechanisms

In unmatched electrically short transmission lines, one coupling mechanism is likely to dominate the overall crosstalk. However, in matched or nearly-matched lines, the magnitude of the electric- and magnetic-field coupling can be similar. At the far-end, the coupled voltages due to the two field-coupling mechanisms are 180^o out of phase. At the near end, the coupled voltages are in phase and the total coupled voltage is their sum.

For example, the total near-end crosstalk for the parallel trace configuration in Figure 6.1 is the crosstalk due to electric-field coupling (4.46×10^{-3}) plus the magnetic-field crosstalk (10.3×10^{-3}). Thus, the overall near-end crosstalk is 14.8×10^{-3} or -36.7 dB. The magnitude of the far-end crosstalk is the magnetic-field crosstalk (10.3×10^{-3}) minus the electric-field crosstalk (4.46×10^{-3}), which is 5.84×10^{-3} or -44.7 dB.

As indicated earlier, common-impedance coupling in circuit board return planes is unlikely to be a concern at frequencies above 100 kHz. The plane resistance is small (less than 1 mΩ per square for half-ounce copper planes). Also, return currents are largely concentrated near their respective signal traces. They will not share the same copper in the return plane unless the traces are vertically stacked above the same region of the plane.

Crosstalk in the Time Domain

Time domain expressions for the crosstalk in electrically short transmission lines are easily obtained by substituting $\partial V/\partial t$ for ωV in the frequency-domain expressions. Note that although these expressions were derived in terms of voltage ratios, common-impedance coupling is proportional to the signal current $\left(I_1 = \frac{V_1}{R_{L1}} \right)$ and magnetic-field coupling is proportional to the time rate of change of the current $\left(\frac{\partial I_1}{\partial t} = \frac{\partial V_1}{\partial t} \cdot \frac{1}{R_{L1}} \right)$. A summary of the time-domain expressions for crosstalk and a visual representation of the crosstalk from a trapezoidal waveform are provided in Figure 6.5.

Time-Domain Coupling in Electrically Short Transmission Lines
(weak coupling assumption)

Common-Impedance Coupling: $V_2 \approx V_1 \frac{R_{Shared}}{R_{L1}} \left(\frac{R_{L2}}{R_{S2} + R_{L2}} \right)$ (same as frequency domain)

Electric-Field Coupling: $V_2 \approx \left(R_{S2} \| R_{L2} \right) C_{12} \left(\frac{\partial V_1}{\partial t} \right)$

Magnetic-Field Coupling: $V_2 \approx -\left(\frac{\partial V_1}{\partial t} \right) \frac{L_{12}}{R_{L1}} \left(\frac{R_{L2}}{R_{S2} + R_{L2}} \right)$ (far end)

$V_2 \approx +\left(\frac{\partial V_1}{\partial t} \right) \frac{L_{12}}{R_{L1}} \left(\frac{R_{S2}}{R_{S2} + R_{L2}} \right)$ (near end)

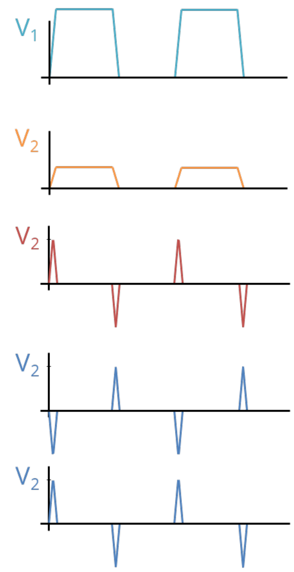

Figure 6.5. Time-domain crosstalk expressions for electrically short lines.

Electrically Long Lines

In short transmission lines, the crosstalk was proportional to the length of the lines. But as transmission lines get longer, the crosstalk cannot increase indefinitely. Two things can happen that limit the amount of crosstalk as the lines get longer.

1. For tightly coupled lines, the weak-coupling assumption may be violated. At this point, the source circuit becomes noticeably loaded. Lumped-element circuit models can still be used to determine the coupled voltage, but the simple equations provided in the previous section no longer apply.
2. When the length of the lines is no longer short relative to a quarter wavelength, voltage and current on the line become a function of position and lumped-element models can no longer be applied.

Most practical transmission lines are not coupled strongly enough to violate the weak-coupling assumption. However, the second situation is common. It is important to be able to calculate the crosstalk between transmission lines that are not electrically short.

Electric-Field Coupling in Long Transmission Lines

Let's start by examining the electric-field coupling in a long transmission line that is matched at both ends. Figure 6.6 shows a schematic illustration of the geometry including lumped-element capacitances that couple the two lines. The orange line illustrates the voltage distribution along the source line at an instant in time.

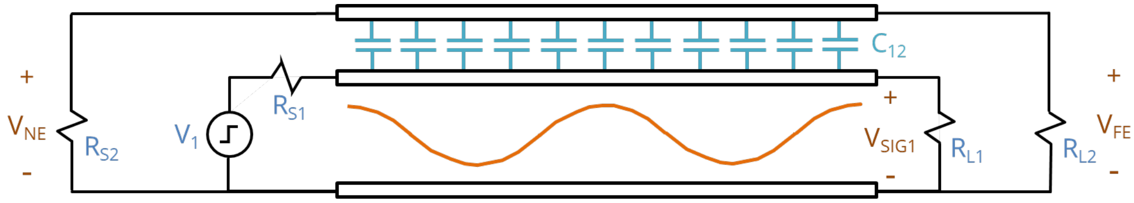

Figure 6.6. Lumped element model of electric field coupling.

To calculate the total far-end electric-field coupling, the contribution of each segment can be added. In the limit as the segment lengths approach zero, this is expressed as the integral,

$$\begin{aligned} \left|V_{SIG2}\right|_{FE} &\approx \left|\int_0^{\ell} V_{SIG1}(x)\left(\frac{Z_{02}}{2}\right)\omega C_{12}\, dx\right| \\ &\approx \left|\left(\frac{Z_{02}}{2}\right)\omega C_{12}\int_0^{\ell} V_1\, e^{-j(\omega t-\beta x)}\, dx\right| \\ &\approx V_1\left(\frac{Z_{02}}{2}\right)\omega C_{12}\left(\frac{1}{\beta}\right)\left|\quad e^{j\beta x}\right|_{x=0}^{x=\ell} \\ &\approx \frac{V_1}{2}\left(\frac{C_{12}}{C_{22}}\right)\left|\, e^{j\beta\ell}-1\,\right|. \end{aligned} \tag{6.13}$$

Note that Z_{02} is the characteristic impedance of the victim transmission line. The result was expressed in terms of C_{22} by noting that $C_{22}=\frac{\beta}{Z_{02}}$. Figure 6.7 shows a plot of the calculated electric-field coupling between two 50-cm transmission lines ($Z_{01} = Z_{02} = 50\ \Omega$, $v = 1.67 \times 10^8$ m/s). Note that the crosstalk reaches its peak value when the length of the lines is an odd multiple of a quarter wavelength. In this case, the mutual capacitance between the lines is, $C_{12} = 12$ pF/m, and the self-capacitance of the victim line is, $C_{22} = 120$ pF/m. Therefore, the peak crosstalk is,

$$\text{XTALK}_{\text{MAX}} = 20\log\left(\frac{C_{12}}{C_{22}}\right) = -20\text{ dB.} \tag{6.14}$$

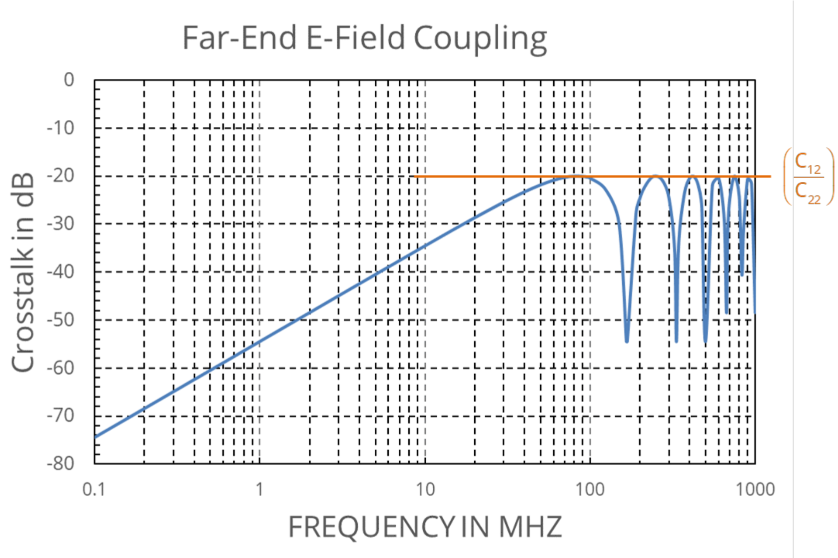

Figure 6.7. Far-end crosstalk between two matched transmission lines due to E-field coupling.

The calculation of the total electric-field coupling at the near-end is a little different. The coupling from each segment does not build uniformly as the signal propagates. Instead, the near-end coupling from each segment propagates back towards the source, so the distance traveled is $2x$ instead of ℓ-x. The expression for the near-end coupling is therefore,

$$\begin{aligned} \left|V_{SIG2}\right|_{NE} &\approx \left| \int_0^\ell V_{SIG1}(x)\left(\frac{Z_{02}}{2}\right)\omega C_{12}\, dx \right| \\ &\approx \left| \left(\frac{Z_{02}}{2}\right)\omega C_{12} \int_0^\ell V_1\, e^{-j(\omega t - 2\beta x)}\, dx \right| \\ &\approx V_1\left(\frac{Z_{02}}{2}\right)\omega C_{12}\left(\frac{1}{2\beta}\right) \left| \; e^{j2\beta x} \Big|_{x=0}^{x=\ell} \right| \\ &\approx \frac{V_1}{4}\left(\frac{C_{12}}{C_{22}}\right) \left| e^{j2\beta\ell} - 1 \right| . \end{aligned} \tag{6.15}$$

Figure 6.8 shows a plot of the calculated near-end electric-field coupling between the same two transmission lines. Note that the near-end crosstalk reaches its peak value when the length of the lines is an odd multiple of a half wavelength. Also, its peak value is half the peak value of the far-end crosstalk.

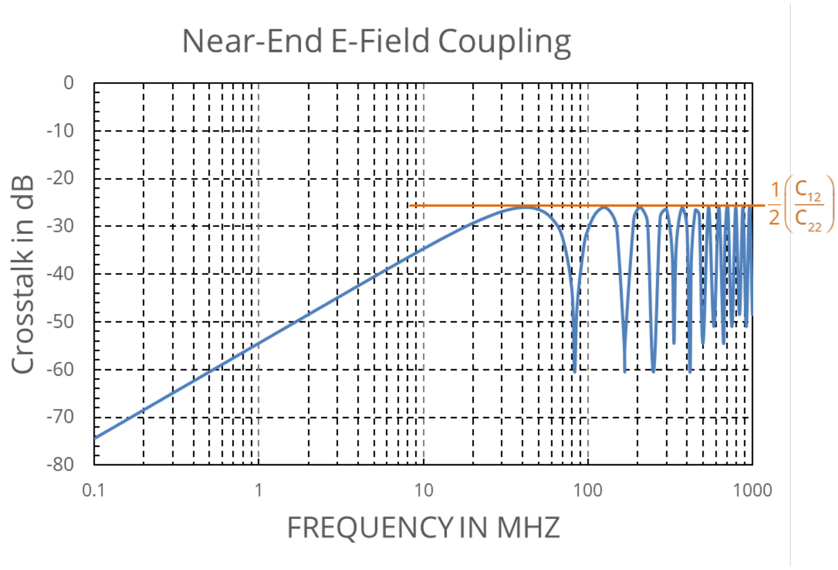

Figure 6.8. Near-end crosstalk between two matched transmission lines due to E-field coupling.

Magnetic-Field Coupling in Long Transmission Lines

A calculation similar to (6.13) can be done to determine the total magnetic-field coupling.

$$
\begin{aligned}
\left|V_{SIG2}\right|_{FE} &\approx \left| \int_0^{\ell} \left(V_{SIG1} \frac{\omega L_{12}}{R_{L1}} \right) \left(\frac{R_{L2}}{R_{S2}+R_{L2}} \right) dx \right| \\
&\approx \left| \left(\frac{\omega L_{12}}{R_{L1}} \right) \left(\frac{R_{L2}}{R_{S2}+R_{L2}} \right) \int_0^{\ell} V_1 \, e^{-j(\omega t-\beta x)} \, dx \right| \\
&\approx V_1 \left(\frac{\omega L_{12}}{R_{L1}} \right) \left(\frac{R_{L2}}{R_{S2}+R_{L2}} \right) \left(\frac{1}{\beta} \right) \left| \quad e^{j\beta x} \right|_{x=0}^{x=\ell} \\
&\approx V_1 \left(\left(\frac{L_{12}}{R_{L1}\sqrt{\mu\varepsilon}} \right) \left(\frac{R_{L2}}{R_{S2}+R_{L2}} \right) \right) \left| \, e^{j\beta\ell} - 1 \, \right| .
\end{aligned}
\tag{6.16}
$$

Figure 6.9 shows a plot of the calculated magnetic-field coupling between two 50-cm transmission lines ($Z_{01} = Z_{02} = 50\ \Omega$, $v = 1.67 \times 10^8$ m/s). Note that the magnitude of the crosstalk in this plot equals the magnitude of the crosstalk due to electric-field coupling. In this case, the mutual inductance between the lines is, $L_{12} = 30$ nH/m, and the self-inductance of the victim line is, $L_{22} = 300$ nH/m.

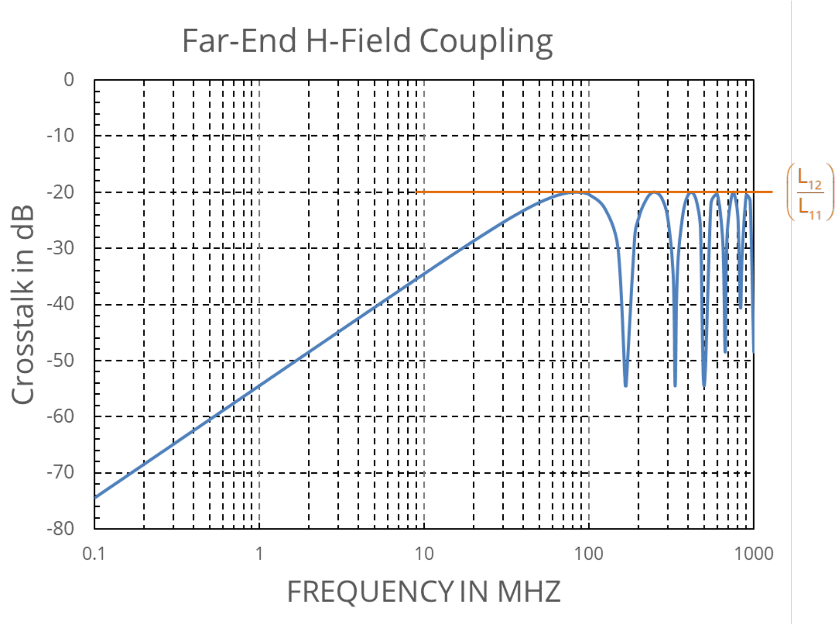

Figure 6.9. Far-end crosstalk between two matched transmission lines due to H-field coupling.

For two transmission lines in a homogeneous dielectric, the magnitude of the electric- and magnetic-field coupling will always be the same. At the far-end, the polarities are opposite, and the electric-field coupling will cancel the magnetic-field coupling. At the near-end, the polarities are the same and the contributions from each coupling mechanism will add.

The calculation of the near-end magnetic field coupling is similar to that in (6.15),

$$\begin{aligned}
\left|V_{SIG2}\right|_{NE} &\approx \left|\int_0^{\ell} V_{SIG1}(x)\left(\frac{1}{2Z_{01}}\right)\omega L_{12}\,dx\right| \\
&\approx \left|\left(\frac{1}{2Z_{01}}\right)\omega L_{12}\int_0^{\ell} V_1\, e^{-j(\omega t-2\beta x)}\,dx\right| \\
&\approx V_1\left(\frac{1}{2Z_{01}}\right)\omega L_{12}\left(\frac{1}{2\beta}\right)\Big|\; e^{j2\beta x}\Big|_{x=0}^{x=\ell} \\
&\approx \frac{V_1}{4}\left(\frac{L_{12}}{L_{11}}\right)\left|\, e^{j2\beta\ell}-1\,\right|.
\end{aligned} \tag{6.17}$$

Figure 6.10 shows a plot of the calculated near-end magnetic-field coupling between the example transmission lines. Note that it is identical to the near-end electric-field coupling. This is because the ratio L_{12}/L_{11} is equal to the ratio C_{12}/C_{22} when the transmission lines are in a homogeneous dielectric.

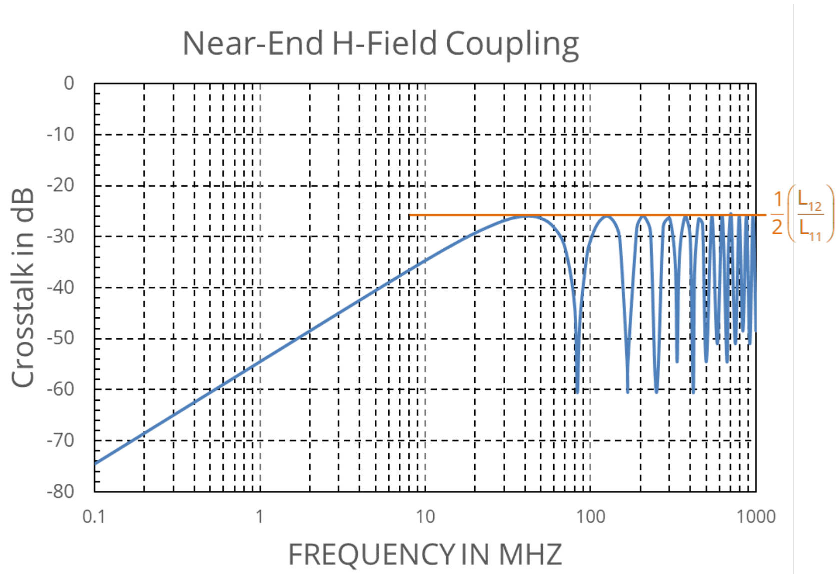

Figure 6.10. Near-end crosstalk between two matched transmission lines due to H-field coupling.

Total-Field Coupling in Long Transmission Lines

In a homogeneous dielectric, the electric- and magnetic-field coupling at the near end are equal in magnitude and add in phase. In this case, the total field coupling is twice the electric- or magnetic-field coupling. For the transmission lines in this example, the total near-end field coupling is plotted in Figure 6.11.

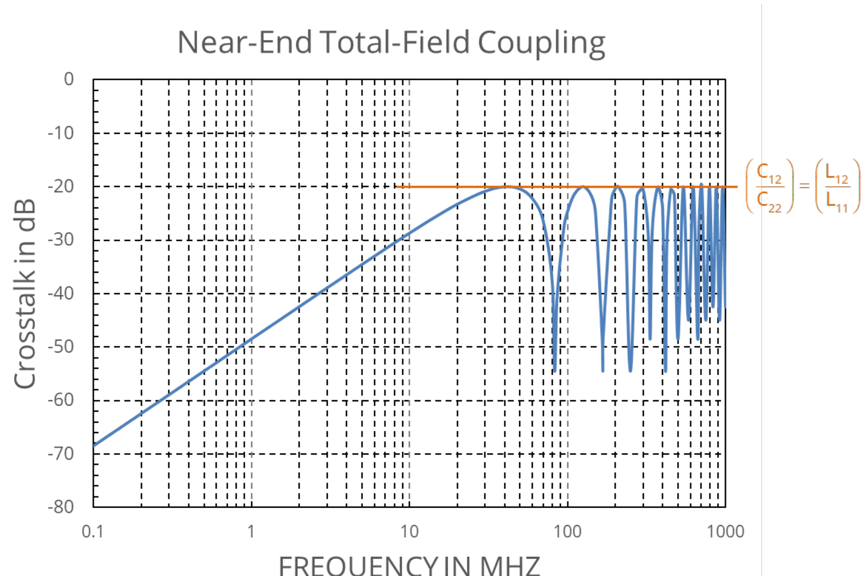

Figure 6.11. Total near-end crosstalk between two matched transmission lines.

In a homogeneous dielectric, the electric-field and magnetic-field coupling at the far end cancel each other. In other words, no voltage is coupled to the far-end. All the coupled voltage appears at the near-end. This is basically how a directional coupler works.

If the dielectric is not homogeneous, for example in a microstrip trace geometry, the two types of field coupling are not necessarily equal. In coupled microstrip traces, the magnetic field coupling will be stronger than the electric field coupling. In this case, the total far-end crosstalk is the crosstalk due to the magnetic-field coupling minus that due to the electric-field coupling.

Crosstalk in the Time Domain

To understand how crosstalk looks in the time domain for electrically long lines, it's helpful to view the lines as a series of electrically short segments. As illustrated in Figure 6.12, a step transition in the first segment of one line induces a spike in the first segment of the coupled line. For the electric-field coupling, the spike is an injected current with an amplitude of $C\frac{\partial V}{\partial t}$, where C is the mutual capacitance per unit length multiplied by the length of the first short segment.

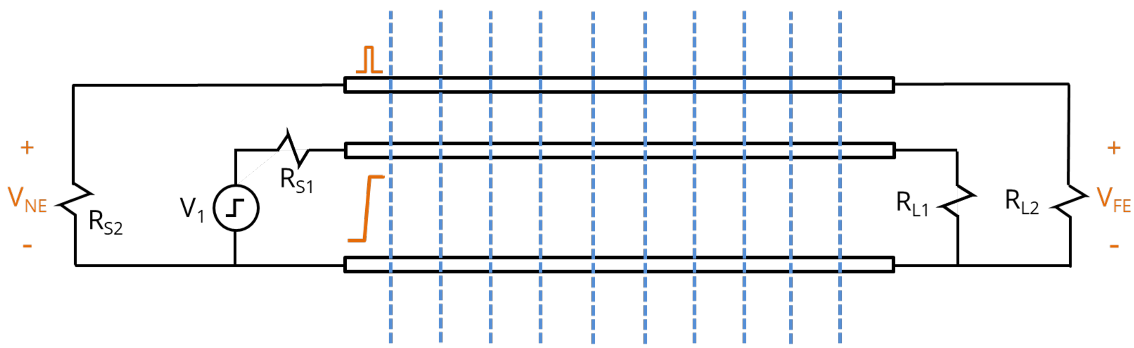

Figure 6.12. Initial crosstalk just after a source transition.

The injected current splits into a component that moves backward towards the near end, and a component that moves forwards toward the far end of the transmission line. A short time later, the transition in the source line has moved to the second segment as illustrated in Figure 6.13. Here, it couples to the second segment of the victim line. But this segment is also receiving the spike that was coupled to segment 1 of the victim trace. The sum of the two produces a spike in the second segment of the victim trace that is about twice as strong as the spike in the first segment.

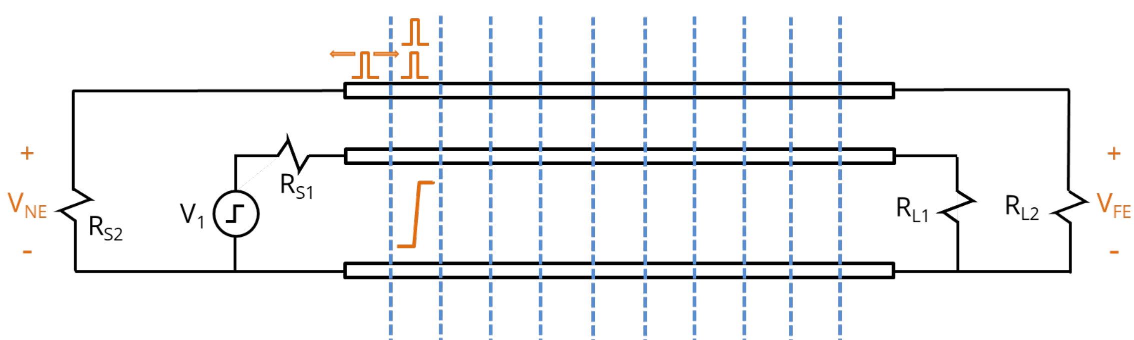

Figure 6.13. Crosstalk as the voltage transition begins to move down the line.

As the step transition continues to move down the source line, it couples more and more voltage to the victim line. As the coupled spike gets larger, its voltage limits the amount of electric-field coupling that can occur. Eventually the spike amplitude reaches a steadystate value and stops growing as indicated in Figure 6.14. The voltage (due to electric-field coupling) observed across the load at the far end is a single spike. For two identical lines matched at both ends, the maximum amplitude of the coupled spike is $V_{\text{2FE-peak}} = \Delta V_1 \left(\frac{C_{12}}{C_{22}} \right)$.

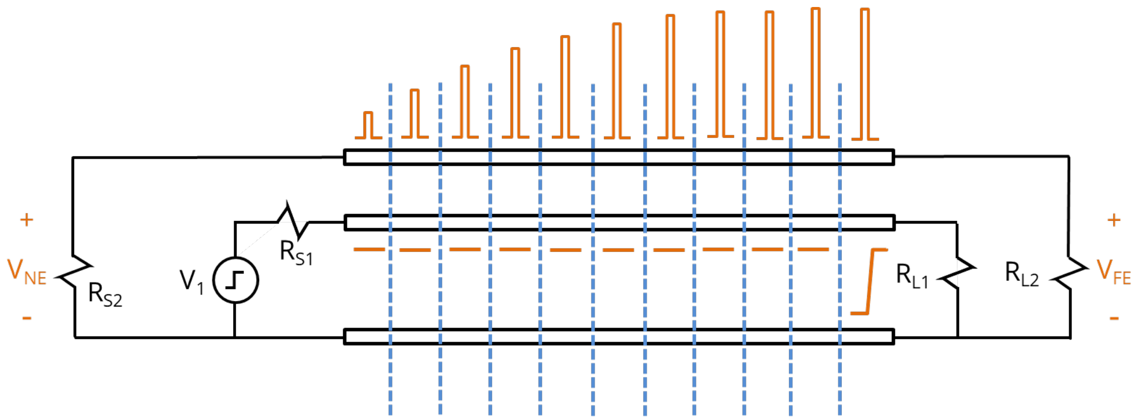

Figure 6.14. Crosstalk as the voltage transition reaches the end of the line.

At the near end, the short spike coupled from the first segment arrived right away, followed by spikes with same amplitude from subsequent segments. In the limit as the segment lengths approach zero, there is a constant stream of new spikes resulting in a coupled voltage waveform that has a constant value. For two identical lines matched at both ends, the maximum amplitude of the coupled spike due to electric-field coupling is $V_{\text{2NE}} = \frac{\Delta V_1}{4} \left(\frac{C_{12}}{C_{22}} \right)$.

The coupled voltage waveforms at the near end, V_{NE}, and the far end, V_{FE}, due to electric-field coupling are shown in the top plot in Figure 6.15. The near end waveform is a constant voltage with a duration of twice the propagation delay. The far-end voltage is a spike arriving at a time equivalent to one propagation delay.

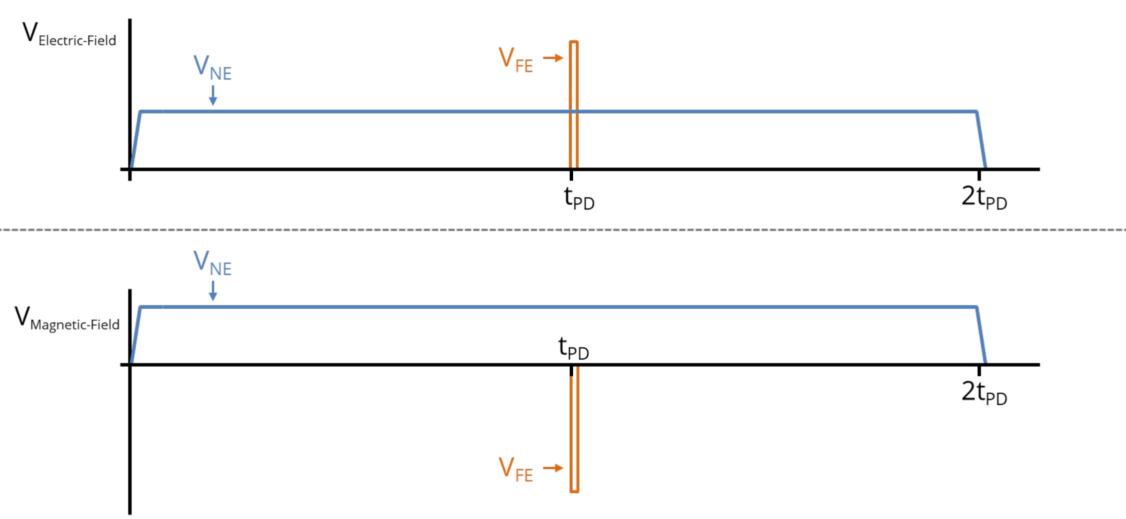

Figure 6.15. Near-end and far-end crosstalk for electric- and magnetic-field coupling.

The magnetic-field coupling between two identical matched lines is plotted in the lower half of Figure 6.15. The near-end crosstalk has the same shape as the electric-field coupling, and an amplitude $V_{2NE} = \frac{\Delta V_1}{4}\left(\frac{L_{12}}{L_{11}}\right)$. The far-end crosstalk is a spike with opposite polarity relative to the electric-field coupled voltage, and an amplitude $V_{2FE\text{-peak}} = \Delta V_1\left(\frac{L_{12}}{L_{11}}\right)$.

The total crosstalk is the sum of the electric- and magnetic-field coupling. These components add in-phase at the near end. At the far end, they have opposite polarities, and the coupled voltage is the electric-field coupled voltage minus the magnetic-field coupled voltage.

In a homogeneous dielectric, the electric and magnetic field coupling have the same magnitude. If the transmission line is matched at both ends, the far-end components cancel and there is no crosstalk at the far end. The near-end crosstalk has the shape shown in Figure 6.15, and an amplitude $V_{2NE} = \frac{\Delta V_1}{4}\left(\frac{C_{12}}{C_{22}} + \frac{L_{12}}{L_{11}}\right)$. If the transmission line is only matched at the far end, some of the near-end crosstalk is reflected, and the reflected waveform appears at the far end.

Modeling Crosstalk in SPICE

Crosstalk can be accurately modeled with circuit simulators if the lumped-element parameters of the transmission lines are known. A common simulation approach is to use 2D field simulators to obtain the per-unit-length RLCG parameters and SPICE to calculate the crosstalk.

Figure 6.16 shows a SPICE model for a pair of coupled, lossless transmission lines. The model is constructed with 20 sub-elements. Each sub-element has 4 lumped element sections yielding a total of 80 lumped-element sections from end to end. In this model, the capacitance per-unit length of both lines is 120 pF/m. The inductance per unit length is 300 nH/m, yielding a characteristic impedance of 50 Ω and a propagation velocity of 1.67×10^8 m/sec. To model a 1-meter length with 80 elements, each element has a capacitance of 1.5 pF and an inductance of 3.75 nH. The mutual capacitance modeled in each section is 1.5 fF and the mutual inductance is 3.75 pH.

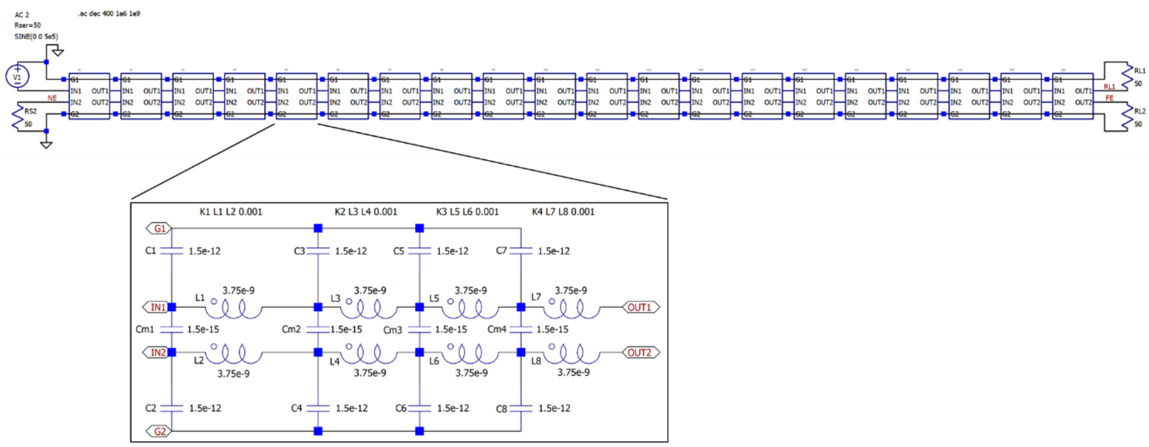

Figure 6.16. SPICE Model of coupled transmission lines with 80 LC sections.

Figure 6.17 shows the calculated crosstalk as a function of frequency from 1 to 1000 MHz. Both transmission lines are matched at both ends. As expected, the near-end crosstalk rises at a rate of 20 dB/decade until the length of the line approaches a quarter wavelength where it reaches its peak value of -60 dB (i.e., C_{12}/C_{22}).

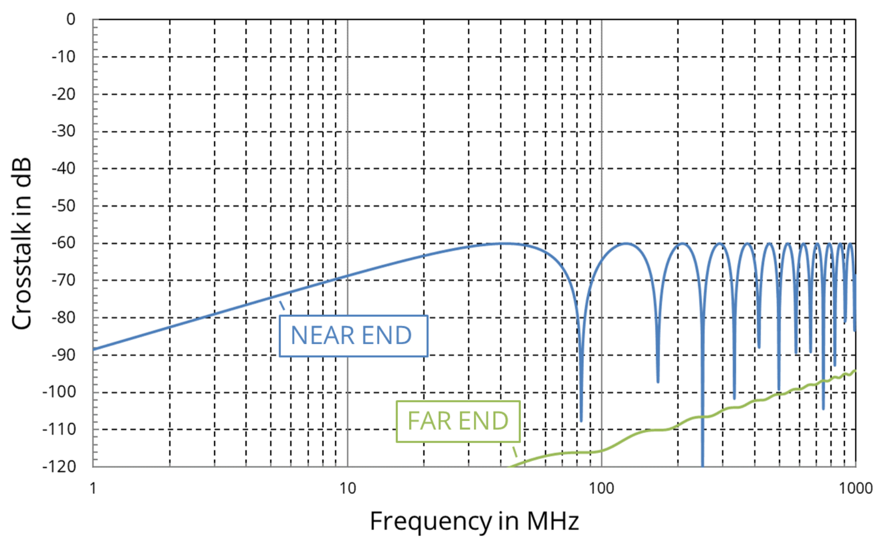

Figure 6.17. Calculated crosstalk between two lines with $C_{12}/C_{22} = L_{12}/L_{11} = 10^{-3}$.

The far-end crosstalk is very small, but it is not zero. This is a consequence of the weak coupling assumption. As the length of the lines starts to become long relative to a wavelength, the power transferred to the victim line starts to measurably reduce the voltage in the source line. This effect becomes more significant when the coupling is stronger. For example, Figure 6.18 shows the calculated crosstalk between the same two transmission lines when the mutual capacitance and mutual inductance are increased by a factor of 10. Note that the near-end crosstalk is 20 dB higher, but the far-end crosstalk is about 40 dB higher.

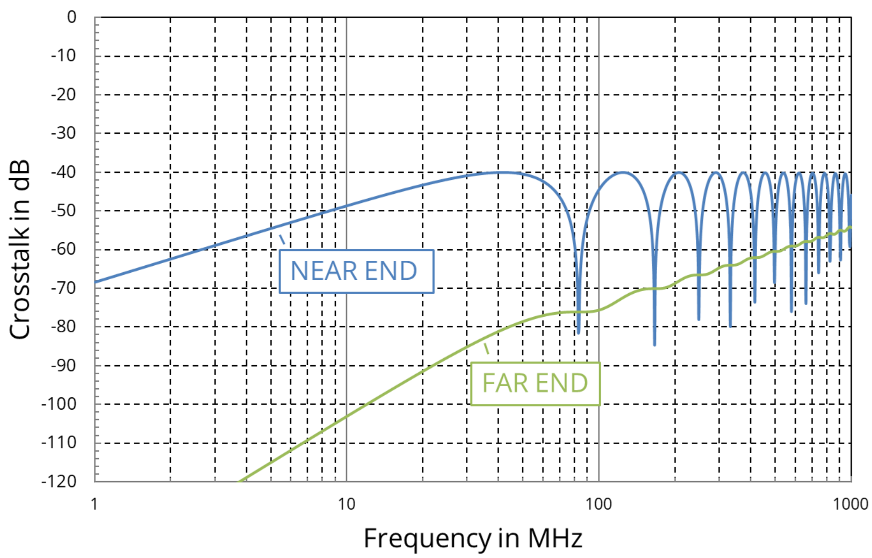

Figure 6.18. Calculated crosstalk between two lines with $C_{12}/C_{22} = L_{12}/L_{11} = 10^{-2}$.

Increasing the mutual capacitance and inductance by another factor of 10 yields the crosstalk shown in Figure 6.19. Note that in this case, the far-end crosstalk exceeds the near-end crosstalk when the lines are about 3 wavelengths or longer. At higher frequencies (or for longer lines) the far-end crosstalk would continue to increase and would eventually (for these lossless lines) approach 0 dB before starting to come down again.

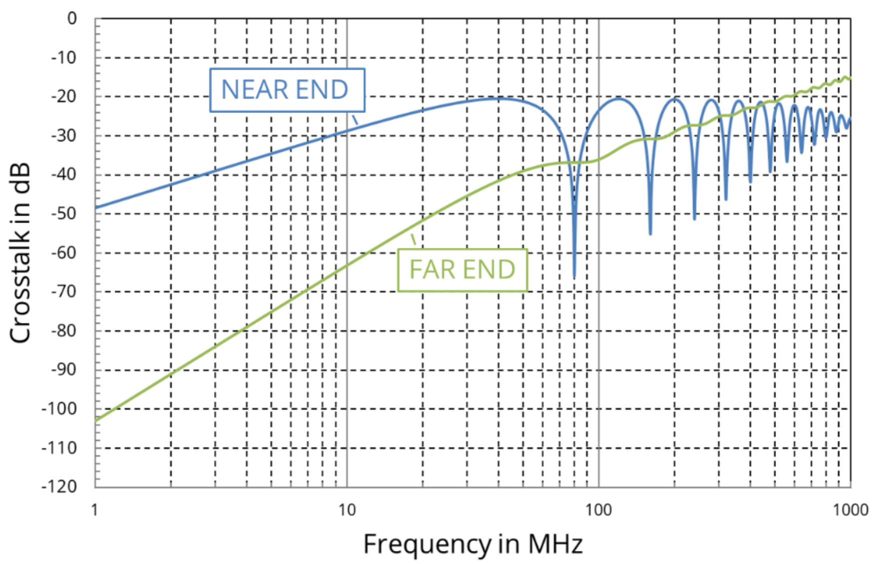

Figure 6.19. Calculated crosstalk between two lines with $C_{12}/C_{22} = L_{12}/L_{11} = 10^{-1}$.

Often, transmission lines are matched at one end or the other, but not both ends. Figure 6.20 shows the calculated crosstalk for the same configuration whose results are shown in Figure 6.18, except the source resistance of the victim circuit has been changed to 1 Ω. In this case, the coupled power that had been flowing to the near end is reflected and now appears at the far end.

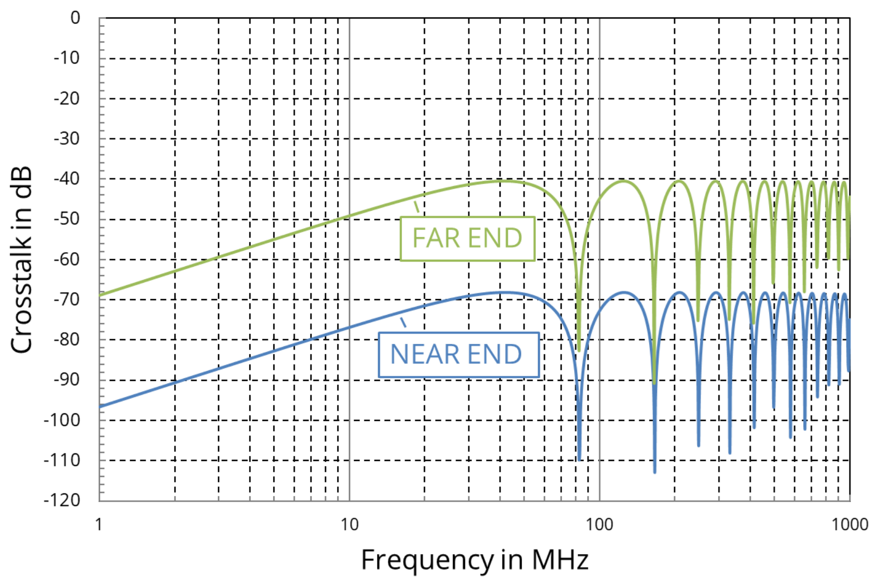

Figure 6.20. Calculated crosstalk with $C_{12}/C_{22} = L_{12}/L_{11} = 10^{-2}$ and R_{NE2}=1 Ω.

When the far-end termination of the source transmission line is set to 5000 Ω, both the near-end and far-end crosstalk are equal. This is illustrated in Figure 6.21.

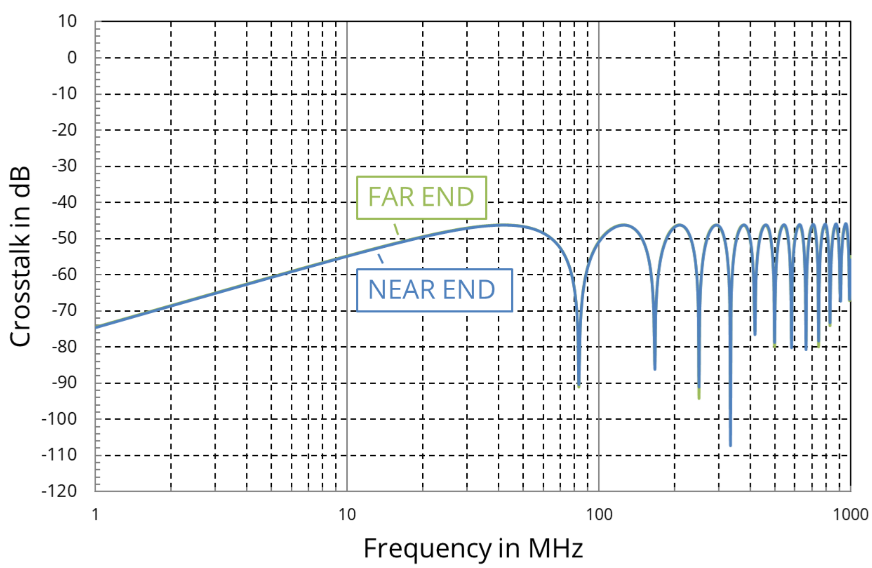

Figure 6.21. Calculated crosstalk with $C_{12}/C_{22} = L_{12}/L_{11} = 10^{-2}$ and R_{FE1}=5000 Ω.

And finally, when the near-end termination of the victim line is 1 Ω and the far-end termination is 5000 Ω, the victim line becomes nearly lossless. In this condition, a lot of crosstalk is observed at resonances of the victim transmission line as illustrated in Figure 6.22. This result helps to emphasize the importance of matching a long transmission line at one or both ends.

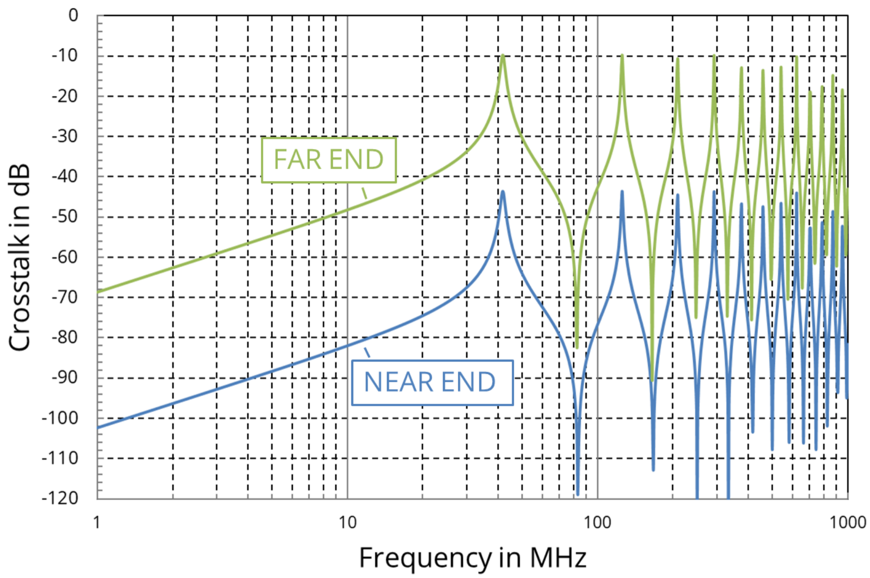

Figure 6.22. Calculated crosstalk with $C_{12}/C_{22} = L_{12}/L_{11} = 10^{-2}$ and R_{NE2}=1 Ω and R_{FE2}=5000 Ω.

The same SPICE model used to perform the frequency domain crosstalk calculations can be used to calculate crosstalk in the time domain. In this case, the excitation appearing on the source transmission line is a 1-volt step function with a 1-nsec transition time. The mutual capacitance and mutual inductance are 10^{-2} times the corresponding self-capacitance and self-inductance. The calculated crosstalk is shown in Figure 6.23. For time-domain results, the crosstalk is not specified in decibels, but rather as a simple ratio.

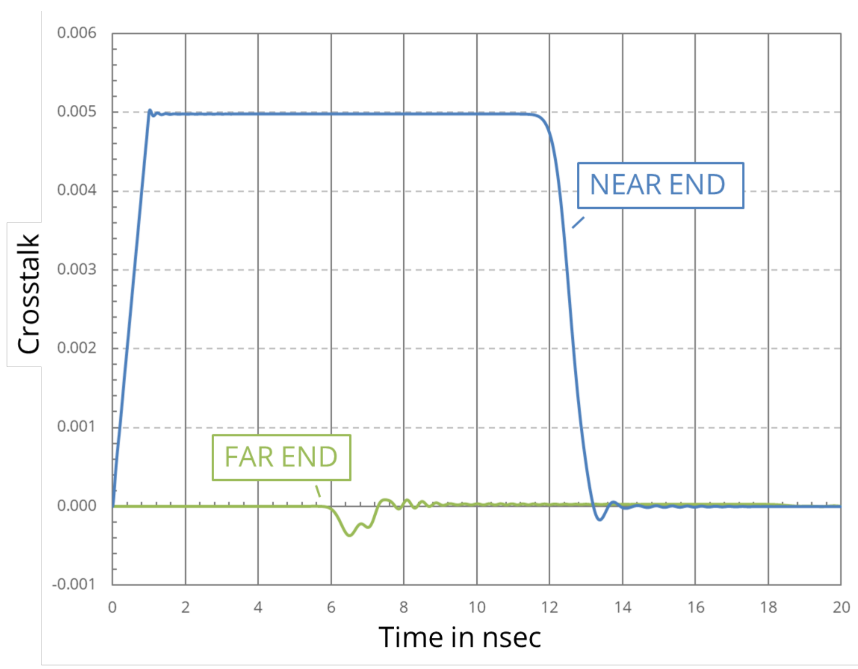

Figure 6.23. Calculated crosstalk between matched lines with $C_{12}/C_{22} = L_{12}/L_{11} = 10^{-2}$.

Note that the propagation delay is 6 nsec. The near end crosstalk shows up immediately and rises to its peak value of $\frac{1}{4}\left(\frac{C_{12}}{C_{22}}+\frac{L_{12}}{L_{11}}\right)$ in 1 nsec. It remains at that value for twice the propagation delay and then falls back to zero in another nanosecond. The far-end crosstalk is nearly zero because the modeled transmission lines are in a homogeneous dielectric (i.e., $C_{12}/C_{22} = L_{12}/L_{11}$).

Figure 6.24 shows the calculated crosstalk of the same lines when the source resistance of the victim line is changed from 50 Ω to 1 Ω. In this case, most of the near-end coupling is reflected and appears at the far end after 6 nsec.

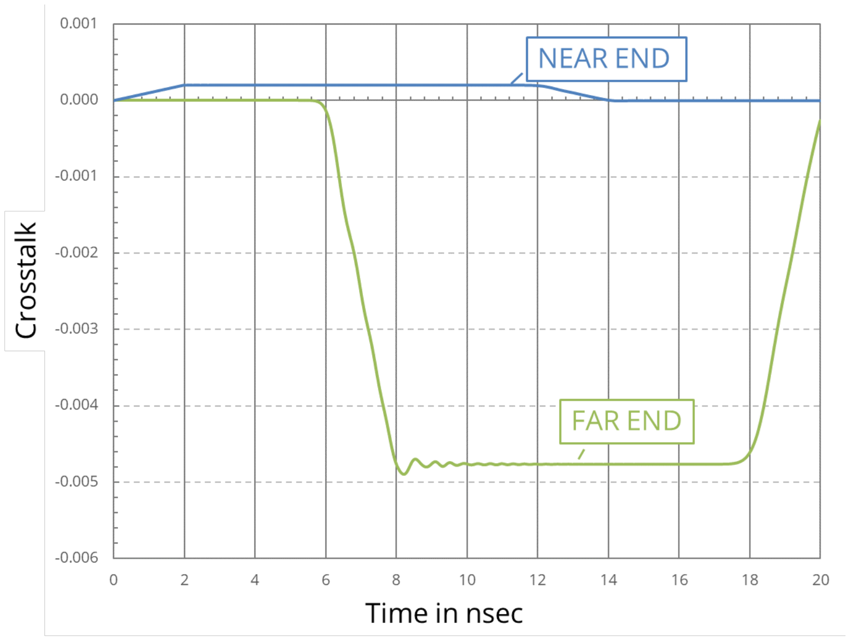

Figure 6.24. Calculated crosstalk with $C_{12}/C_{22} = L_{12}/L_{11} = 10^{-2}$ and R_{NE2}=1 Ω.

Figure 6.25 shows the calculated crosstalk when both lines are matched, but the coupling is stronger (C_{12}/C_{22} = 0.1). This is the same configuration whose frequency-domain results are plotted in Figure 6.19. In this case, because the line is long (T_{PD} = 6 t_r), and the coupling is strong, a spike is observed in the far-end crosstalk.

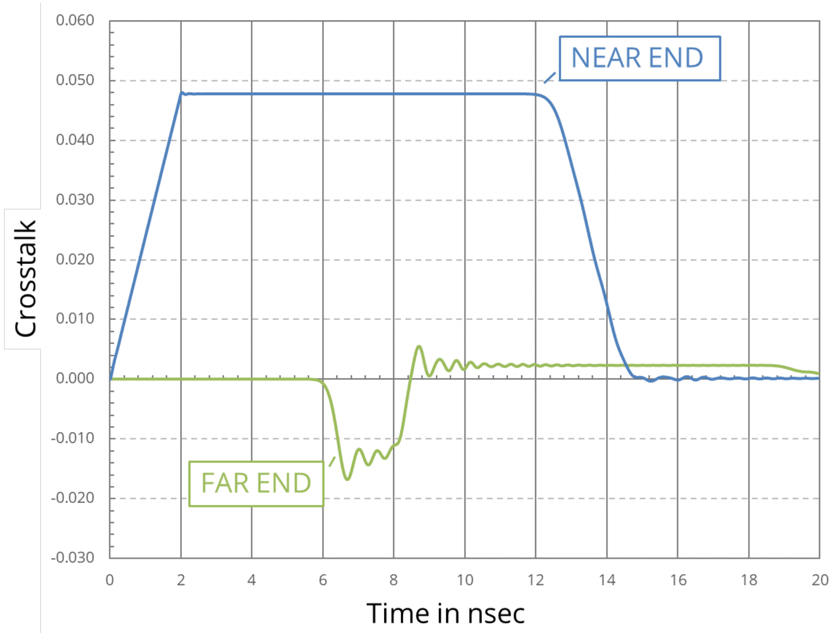

Figure 6.25. Calculated crosstalk between matched lines with $C_{12}/C_{22} = L_{12}/L_{11} = 10^{-1}$.

Finally, Figure 6.26 shows the calculated crosstalk when both lines are matched, but the dielectric is not homogeneous. In this case, C_{12}/C_{22} = 0.5 x 10^{-2}, while $L_{12}/L_{11} = 10^{-2}$. The transmission lines are the same as those whose crosstalk is plotted in Figure 6.23, except that the mutual capacitance is cut in half.

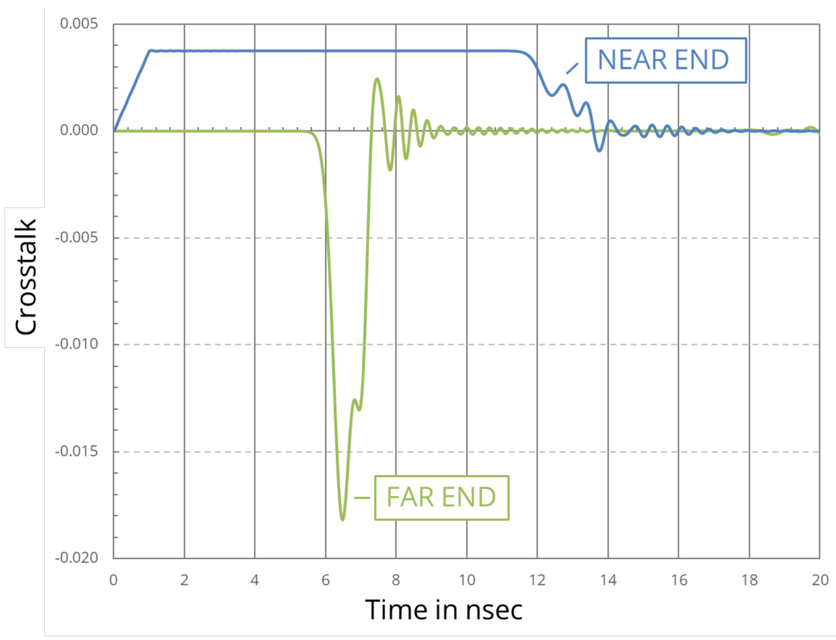

Figure 6.26. Calculated crosstalk between matched lines in a non-homogeneous dielectric.

When the dielectric is not homogeneous, the electric-field coupling and magnetic field coupling do not cancel in the forward direction. As the results in the figure indicate, a spike is observed at the far end with a peak amplitude that exceeds the amplitude of the near-end coupling.