# Analyzing Crosstalk and Ground Bounce in Multiconductor Systems Using SPICE Circuit Simulations

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## ABSTRACT

**Unwanted coupling within electronic systems in the form of crosstalk (electric or magnetic coupling between circuits) and ground bounce (common impedance coupling due to shared ground conductors) can threaten the operation of high-speed digital systems by causing false triggering. This paper investigates the mechanisms causing crosstalk and ground bounce, derives a practical equivalent circuit that can be used to model their behavior in multiconductor systems, and demonstrates a useful guideline for applying the circuit model to real systems. The equivalent circuit derived can be used to investigate the behavior of coupled multiconductor transmission systems with a circuit simulation tool such as SPICE. Such analysis is applied to a three-pin ground-signal-victim arrangement in order to gain a qualitative understanding of the factors influencing crosstalk and ground bounce. A 3 x 3 matrix of pins is also simulated to demonstrate the typical behavior of a multiconductor system in which multiple lines may be driven simultaneously. Some general design advice on minimizing crosstalk and ground bounce is offered on the basis of these simulations.**

## INTRODUCTION

As electronic technology continues to evolve, a major trend is evident: manufacturers are continuously striving to build denser electronic systems while at the same time trying to drive them at higher switching speeds. One of the implications of this trend is that the possibilities are increased for unwanted coupling between circuits within the system. Such coupling manifests itself in the form of crosstalk (electric or magnetic coupling between circuits) and ground bounce (common impedance coupling due to shared grounds) and can have a significant effect on the noise margin and, thus, the integrity of the system.

This paper presents an explanation of the mechanisms causing crosstalk and ground bounce. While both phenomena are distributed effects rooted in field behavior, it is often possible to model them from a circuit perspective using a circuit-analysis tool such as SPICE. Since coupling between pins within a connector is often a concern, two different conductor geometries will be examined in order to gain some qualitative understanding about crosstalk and ground bounce within multiconductor systems. The simplest arrangement evidencing coupling to a quiet circuit is three parallel pins, where one is assigned to be the signal conductor, a second one acts as the return path, and the third pin is the inactive victim. This geometry will be modeled with specific attention paid to the effects of conductor spacing and dielectric variation. To examine the behavior of a more realistic representation of a typical multiconductor system, a 3 by 3 matrix of pins will be modeled with various conductors being driven simultaneously. The effects of different drive schemes on crosstalk, ground bounce, and the current distribution among the ground conductors will be explored.

## DEFINITION OF CROSSTALK AND GROUND BOUNCE

Figure 1 shows the simplest structure where coupling to an inactive circuit can be examined. Three conductors are present. One is designated as the signal conductor because a potential is applied between this wire and the second conductor, the ground plane or return path. It is assumed that this voltage is time-dependent so that it has some rate of change, The third conductor is designated as the victim wire and shares the same return conductor with the signal wire. These conductors need not be wires or planes.
For application of the technique under discussion, each

conductor merely needs to maintain consistent geometry and spacing over the region being considered.

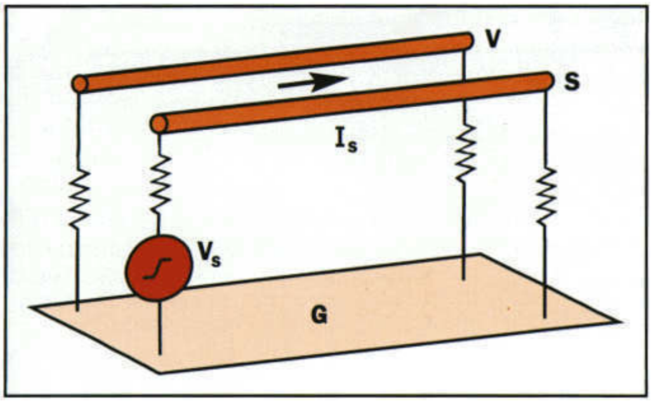


**Figure 1.** Two conductors sharing the same ground. A time-varying voltage $V_S$ excites the signal conductor causing current $I_S$.

As a time-changing voltage propagates down the signal conductor, the electric and magnetic fields associated with this propagation extend through the space occupied by the victim line. These time-changing fields induce energy to propagate along the victim conductor. The mutual inductance between the victim conductor and the signal conductor acts as a coupling mechanism that induces a longitudinal potential across the victim line. Likewise, the capacitance between the victim and signal conductors acts as a vehicle to induce current in the victim-ground circuit. This inductive (magnetic) and capacitive (electric) coupling is called crosstalk and is defined by[1]

$$V_V = k_m M_{SV} \frac{dI_S}{dt} \qquad (1)$$

$$I_V = k_c C_{SV} \frac{dV_S}{dt} \qquad (2)$$

where $M_{SV}$ is the mutual inductance per unit length between the signal and victim lines, $C_{SV}$ is the mutual capacitance per unit length between the lines, and $k_m$ and $k_c$ are constants of proportionality that will be determined by the terminating impedances and the inductance and capacitance from the victim to ground.

One must remember that this crosstalk is a distributed phenomenon: for each differential section of the lines, capacitive and inductive crosstalk is inducing energy on the victim line. If the conductors are electrically long—that is, if the wavelength of the energy they carry is short compared to their physical lengths—this behavior can only be correctly modeled by multiconductor transmission line theory. Such theory says that the arrangement displayed in Figure 1, where there are three conductors and one is defined to be at reference potential, must be treated as two coupled transmission lines. The circuit diagram for one differential section of the coupled lines is shown in Figure 2. It can be seen that the coupling between the lines is represented in this diagram by $M_{SV}$ and $C_{SV}$. $L_{SG}$ represents the combined effects of the self-inductance of the signal conductor, the self-inductance of the ground conductor, and the mutual inductance between the signal and the ground. $L_{VG}$ includes the self-inductance of the victim conductor, the self-inductance of the ground, and the mutual inductance between the victim and ground. For simplicity, conductor and dielectric losses are assumed to be negligible and the lines are considered lossless.

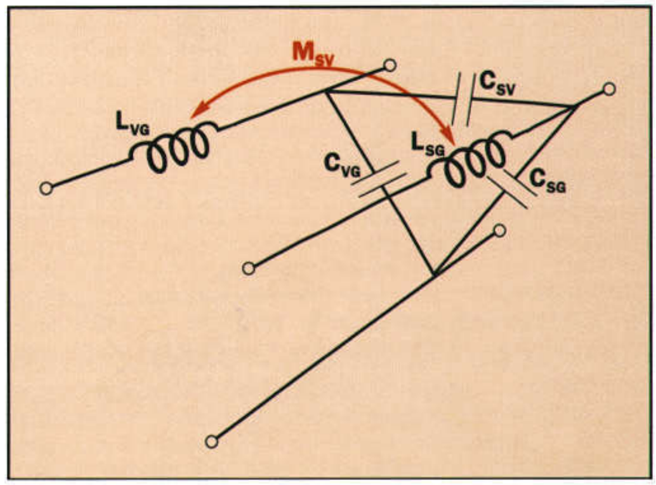


**Figure 2.** Differential section of two coupled transmission lines.

A subtle characteristic of the traditional transmission line representation is the lack of a separate inductor in the return path to represent the self-inductance of the ground. The reason for depicting the equivalent circuit this way ii that allowing one conductor to be at reference potential throughout its length makes a solution to the wave equations for the lines analytically tractable. The practical implication of this representation is that no longitudinal potential difference can occur along the ground conductor. Physically speaking, real ground paths will have some finite impedance corresponding to their resistance and self-inductance. Furthermore, mutual inductances and mutual capacitances between the ground conductor and the various signal conductors will influence this impedance. The voltage difference developed across this return conductor impedance is commonly called ground bounce. The ground-bounce voltage is defined by[2]

$$V_{GB} = Z_G I_G \qquad (3)$$

where the impedance $Z_G$ accounts for the combined effects of the ground path's resistance and self-inductance as well as mutual inductance and mutual capacitance to other conductors. Because ground bounce causes the reference potential to change along the length of the ground, it becomes a possible source for coupling noise into circuits that share that ground, particularly if many signal lines are active simultaneously. When many lines are active at the same time, more current traverses the common return path impedance, causing an increase in the ground-bounce voltage.

Figure 3 illustrates a practical equivalent circuit for representing a differential section of two coupled transmission lines. In this revised circuit, $L_S$, $L_G$, and $L_v$ only represent the self-inductance of the various individual conductors. As before, the mutual inductances are explicitly stated with the addition of $M_{VG}$ and $M_{SG}$ to represent mutual inductance from the signal and victim to the ground. Because the circuit shown in Figure 3 contains an impedance in the shared ground path, the contributions of both crosstalk and ground bounce will be included.

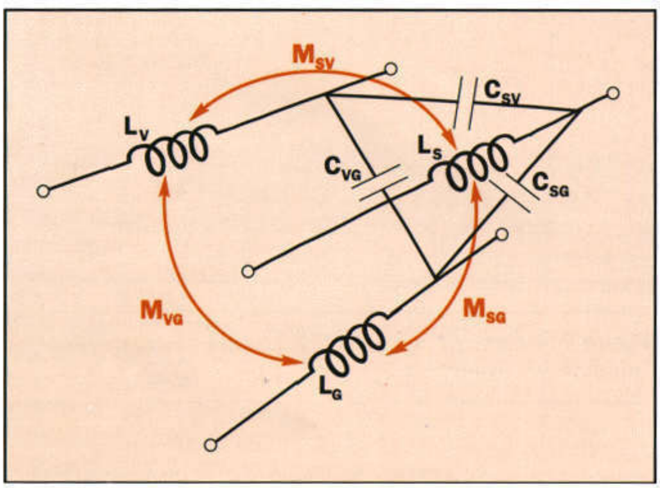


**Figure 3.** Practical equivalent circuit for differential section of two coupled transmission lines allowing crosstalk and ground bounce.

## MODELING CROSSTALK AND GROUND BOUNCE WITH SPICE

Due to the complexity of the circuit illustrated in Figure 3, an analytical solution to the practical crosstalk and ground bounce model would be difficult, if not impossible. Consequently, the only pragmatic means to attack the problem is through numerical methods. One of the better ways of modeling such a problem is to develop an equivalent circuit that considers the distributed nature of crosstalk and ground bounce and then to analyze the circuit in a software simulation. The large family of available SPICE programs are well suited to this task because they permit the representation of resistors, capacitors, inductors, mutual coupling between inductors, and even active devices if so desired.

A reasonable circuit model maybe developed by considering how many differential sections the system must be divided into in order to reasonably represent the distributed nature of the coupling. In reality, these sections (or stages as they will be referred to from now on) do not really represent a differential length of the system but rather a length over which it can practically be assumed that the voltage and current are nearly constant. Assuming a finite risetime step voltage or trapezoidal pulse input, a good rule of thumb governing the stage length says that it should have a propagation delay about one-tenth of the risetime. The method for calculating the number of stages can be shown as follows:

$$\frac{\text{structure length}}{v_p} = \text{structure propagation delay}$$

$$\frac{\text{structure propagation delay}}{\text{risetime}} \times 10 = \text{number of stages} \quad (4)$$

where $v_p$ is the propagation velocity down the lines. In air this value will be 11.81 inches/ns. For example, if a 1-ns risetime step is sent propagating in air down a coupled line system that is 1 inch long, the structure propagation delay will be 84.7 ps. Dividing this value by 1 ns and multiplying by 10 gives 0.847 stages. According to the guideline, only one stage is needed to represent this network.

## CROSSTALK AND GROUND BOUNCE IN A THREE-CONDUCTOR SYSTEM

In the previous sections, a practical equivalent circuit to represent crosstalk and ground bounce was developed as well as a guideline for using this circuit to build a system model based on the structure's electrical length and the signal risetime. This approach will now be applied to a three-pin system in order to gain some qualitative understanding of the factors that affect crosstalk and ground bounce.

The system to be examined consists of three circular pins, each 0.020 inch in diameter and 0.765 inch long. The pins were arranged parallel to each other and placed in a row on 0.100-inch centerlines. The signal and victim pins were terminated to the ground pin via 50-Ω resistors to simulate the placement of this structure into a 50-Ω system. Figure 4 illustrates the setup; the ground, signal, and victim pins are arranged as marked.

A lumped-element circuit model for this configuration, assuming a single stage, is shown in Figure 5. The mutual inductances have been eliminated from the model and have been replaced by dependent voltage sources in the appropriate lines. This model is instructive for understanding the qualitative effects caused by the crosstalk and ground bounce. In fact, according to the guideline presented in the last section, this model would be suitable to use for rise-times greater than 650 ps. Also shown in Figure 5 are the measurement points for looking at the backward crosstalk (BXT) and forward crosstalk (FXT) on the victim line. These voltages are actually caused by a combination of the field coupling and the common impedance coupling, despite the fact that their names would imply otherwise.

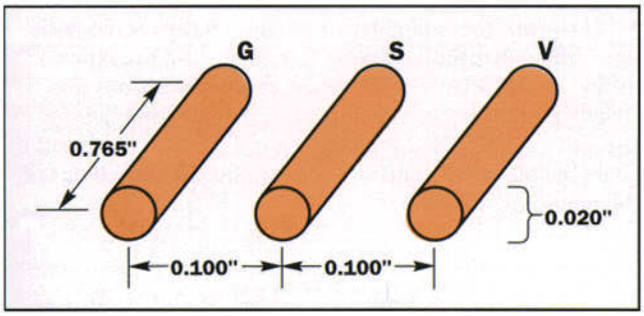


**Figure 4.** Three pin setup for analyzing crosstalk and ground bounce.

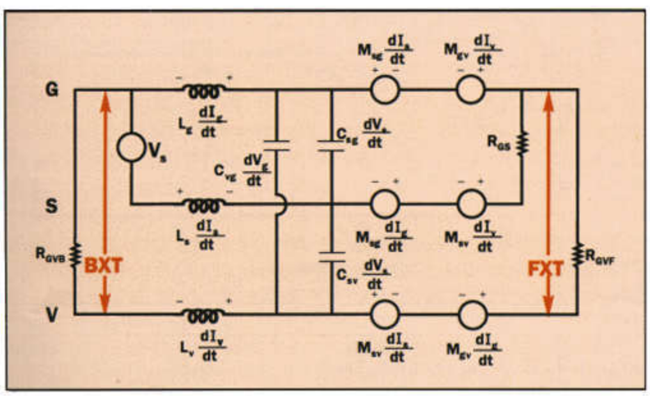


**Figure 5.** Single stage circuit model for three parallel pins allowing crosstalk and ground bounce. Measurement points for backward crosstalk (BXT) and forward crosstalk (FXT) voltages are shown.

If one examines Figure 5 and imagines, for a moment, that the capacitors are not present (implying the absence of capacitive coupling), it can be seen that the signal current will travel down the signal conductor to the termination, $R_{GS}$, connecting it with the ground line. At this node, two paths are available for returning to the voltage source: through the ground conductor and through the victim conductor and its terminations, $R_{GVF}$ and $R_{GVB}$. Consequently, parallel paths back to ground are established and some current will flow through the victim line. Because of the direction of current flow relative to ground, a positive voltage for the backward crosstalk will be observed across $R_{GVB}$. However, a negative voltage for the forward crosstalk will be seen across $R_{GVF}$. If $R_{GVB} = R_{GVF}$, then the backward and forward crosstalk will have the same amplitude but opposite signs. If capacitive coupling is now allowed to enter the picture, the capacitors will act as current sources by inducing current in the victim conductor. This current will travel in both directions along the victim line. At the near end the voltage drop produced across $R_{GVB}$ will add with that caused by the inductive coupling. The result will bean increase in the backward crosstalk. At the far end this current flows opposite that produced by the inductive coupling. Therefore, it will reduce the negative amplitude of the forward crosstalk seen across $R_{GVF}$. If the capacitive coupling is strong enough, it is even possible for the amplitude of the forward crosstalk to turn positive. Consequently, relative to what would be observed if only inductive coupling were present, the capacitive coupling is seen to increase the backward crosstalk amplitude while decreasing the negative forward crosstalk amplitude.

Another important aspect of Figure 5 is that it clearly shows the sources for ground bounce. A voltage drop across the self-inductance of the ground conductor $L_g$ will be present as well as the induced voltage caused by the changing currents in the signal and victim lines. The impedance associated with $L_g$ and the induced voltage associated with $M_{gv}$ act as a source helping to drive a return current through the victim line. Theoretically, this voltage could be measured by probing across the ground line. Realistically, ground bounce measurements are difficult and error-prone. Because good ground paths typically have a low but finite impedance, the impedance of the probe and its ground line tend to skew the measured voltage. Fortunately, these difficulties do not occur in a simulation. One can be fairly certain that if a model accurately predicts the backward and forward crosstalk, it will give a fair assessment of the ground-bounce voltage also.

With these thoughts in mind, the two-stage SPICE model shown in Figure 6 was developed for the three-pin setup. The equivalent circuit used for each stage differs slightly from that shown in Figure 3. Instead of using an L-C network as the basis for each stage, an L-C-L circuit was used because of its symmetry. The resulting model actually has the inductance for each pin broken into four stages while the capacitance is distributed over two. Mutual inductances between the inductors are not shown in Figure 3 because of the complexity they would introduce. The mutual inductance between each pair of inductors is entered in the model by specifying the coupling coefficient defined as[3]

$$K_{12} = \frac{M_{12}}{\sqrt{L_1 L_2}} \qquad (5)$$

This coefficient assumes values ranging from 0 to 1 since the mutual inductance can never be as great as the larger of the two self-inductances. The self-inductances and mutual inductances were calculated with the proprietary program PATH_L. By using Grover's formulas[4] and numerical implementations of the Neumann equation,[5] PATH_L can compute the self-inductance of individual wire segments as well as the mutual inductance between pairs of segments.

Each wire was divided into four segments of equal length. The self-inductance of a segment and the mutual inductances between all the segments were computed. The capacitances were computed using the proprietary program CAPRAY. This program uses a method-of-moments calculation[6] to compute the capacitance matrix for up to ten parallel wires. The capacitances supplied by CAPRAY were halved to obtain the capacitor values seen in Figure 6. The zero-valued voltage sources VIM, V2M, and V3M were placed in the circuit for the purpose of monitoring the current through each pin. Source and termination impedances used (RS, RT, RVI, and RVO) were 50 Ω.

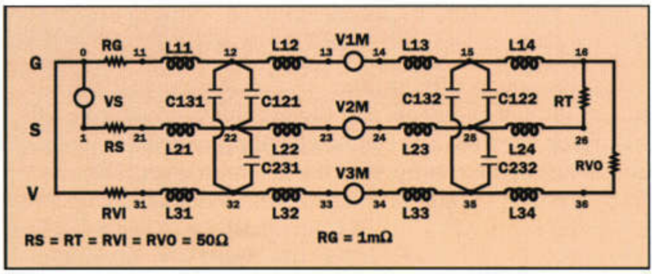


**Figure 6.** Two stage SPICE model for three parallel pins allowing for crosstalk and ground bounce. For simplicity, the coupling between the inductors is not shown.

In order to verify the validity of this model, backward and forward crosstalk were simulated using a 1-ns risetime step. The results of the SPICE simulation were compared with actual measurements made using a Tektronix 7854 Digitizing Oscilloscope with an S-54 Pulse Generator head.[7] This head produces a smooth 250-mV step that has a l-ns risetime. In order to closely simulate the S-54's step shape, the step used in" the SPICE analysis was a piecewise linear approximation of an integrated Gaussian curve. To produce the same step amplitude as the S-54, a source step of 500 mV was used in the simulation. Because the impedance of the signal pin is small compared with the 50-Ω terminations, half of the SPICE source voltage will be dropped across the source 50-Ω resistor and the input voltage to the signal pin will be 250 mV. Backward crosstalk in the SPICE model was measured between node 31 and node 0. Forward crosstalk was observed between node 36 and node 16. The crosstalk voltages (VBXT and VFXT) are shown in Figure 7 along with the computed ground-bounce voltage (VGB). Peak crosstalk was found to be 25.1 mV in the backward direction and –23.0 mV in the forward direction. A calculated ground-bounce voltage of 39.8 mV was seen. The measured values for these voltages were 25.5 mV for the backward crosstalk and – 21.7 mV for the forward crosstalk.[8] Correlation seen between the measured model and the simulation was within 1.6 percent in the backward case and 6.1 percent in the forward case. Clearly, the equivalent circuit model works well for a 1-ns risetime.

The remainder of the analysis undertaken in this section uses a generator producing a 1-V step with a risetime of 200 ps. For the previously stated reasons, this source introduces a 500-mV step onto the signal pin. This particular risetime was chosen because today's fastest logic families are using edge rates of this speed. The question remains as to whether the model shown in Figure 6 will be accurate at such a fast risetime. Referring back to Equation 4, a two-stage model would be reasonable down to a risetime of 324 ps. However, the symmetry of the model causes the inductance to be distributed across four stages—a breakdown usable down to 162 ps. While the capacitance does not strictly meet the recommended breakdown for a 200-ps risetime, the SPICE circuit shown in Figure 6 was found to give sensible results and will be used for the remainder of this section.

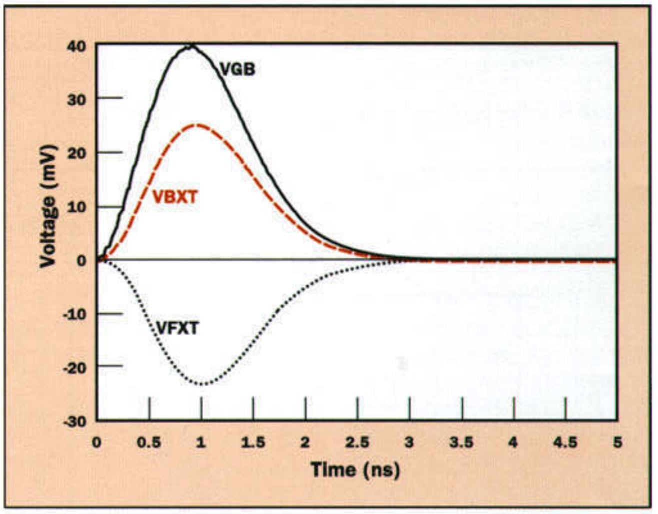


**Figure 7.** Ground bounce, backward crosstalk, and forward crosstalk for 1-ns risetime.

In Figure 8 the computed currents through each of the three conductors are shown. Examination of the current plot shows that, just as predicted, the victim pin functions as a second ground return. This fact can be surmised by noting that the victim current is negative like the ground current. At its peak around 300 ps, the return current through the victim pin is approximately half as large (4 mA) as the current in the ground pin (8 mA). Under these transient conditions the amount of energy returning through the victim line is certainly nonnegligible. It may also be noted that as the time gets near 1 ns, the transient excitation has passed and the victim current is decaying toward zero.

Figure 9 shows SPICE's computed waveforms for the ground-bounce voltage, backward crosstalk, forward crosstalk, and the voltage across the victim pin. One may note that the peak ground-bounce voltage is about 360 mV—almost twice the peak amplitude of the crosstalk voltages. Because Figure 8 shows the current in the victim pin always flowing in the same direction as the ground current, the backward and forward crosstalk must have opposite signs. The fact that their amplitude is nearly identical seems to indicate that the pins are spaced far enough apart that the capacitive coupling is quite small compared to the inductive coupling. It is worthwhile to observe that the peak-to-peak amplitude of the voltage drop across the victim pin is about the same amplitude as the crosstalk voltages and about two-thirds the amplitude of the ground-bounce voltage.

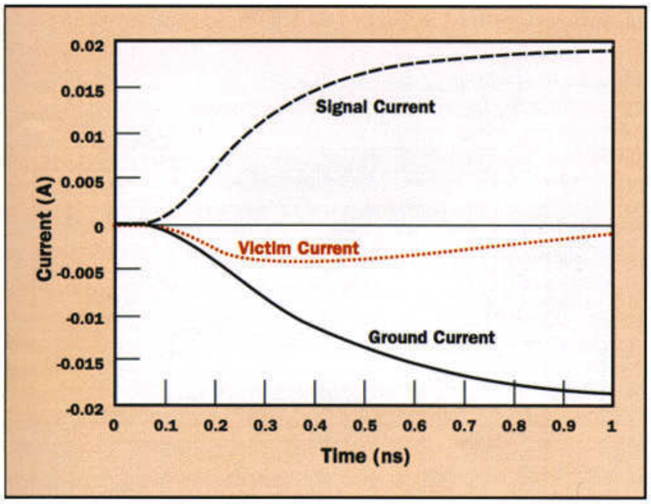


**Figure 8.** Currents generated across the three pins by the 200-Ps risetime step.

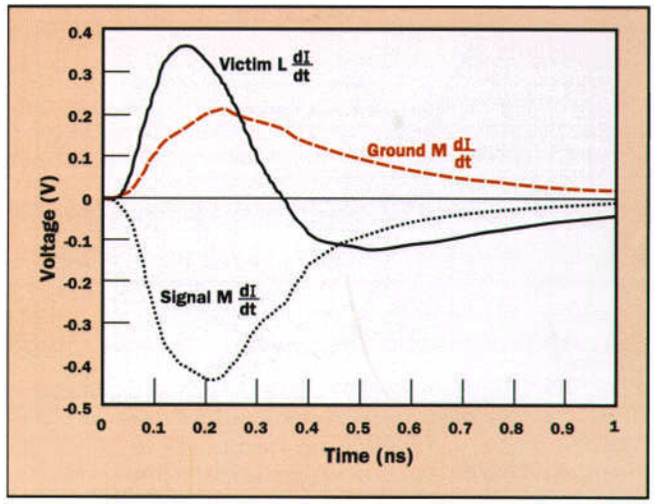


**Figure 10.** Voltages across the victim pin due to its self-inductance, mutual inductance with the signal pin, and mutual inductance with the ground pin.

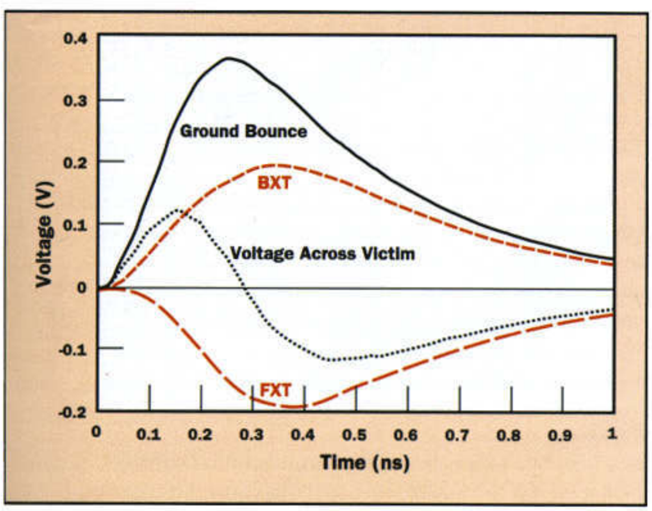


**Figure 9.** Ground bounce, backward crosstalk, forward crosstalk, and the voltage drop across the victim pin.

Figure 10 displays the various inductive contributions to the voltage across the victim pin. While the waveforms shown in Figures 7, 8, and 9 were directly obtainable from SPICE, the curves in this figure were generated by differentiating the current waveforms shown in Figure 8 and multiplying them by the appropriate inductance values. The self-inductance used was the self-inductance for the entire victim pin. Likewise, the mutual inductances used were the total mutual inductance between the signal and victim pins and the total mutual inductance between the ground and victim pins. Each of these inductances was calculated using PATH_L.

In Figure 11 the three waveforms plotted in Figure 10 are summed and displayed with the total voltage across the victim pin first seen in Figure 9. It is immediately obvious that the total voltage across the victim pin is virtually identical to the sum of the inductive voltage contributions. This result is direct proof of the phenomenon first suggested through comparison of the crosstalk voltages in Figure 9, that is, the energy induced on the victim pin is almost solely attributable to the inductive coupling. When these pins are on 0.100-inch centerlines, the capacitive coupling is indeed negligible.

Since it has been established that the separation distance between the wires is too large for capacitive coupling to be a significant contribution to the crosstalk, it is reasonable to expect that capacitive coupling would also have a negligible contribution to the ground-bounce voltage. This expectation is examined in Figure 12 by comparing the ground-bounce voltage with the sum of the inductive contributions to the ground-pin voltage. The inductances used were again computed by PATH_L and reflect the self-inductance for the entire ground pin as well as the total mutual inductances between the signal pin and ground and the victim pin and ground. A critique of Figure 12 shows that the wave shape and amplitude of the sum of the inductive voltage contributions across the ground pin are virtually identical to the ground-bounce voltage. Therefore, it is reasonable to conclude that given a far-spaced pin configuration where capacitive coupling is small, the ground-bounce voltage can be attributed almost exclusively to the self-inductance of the ground and mutual inductances between the ground and the other current-carrying conductors in the system.

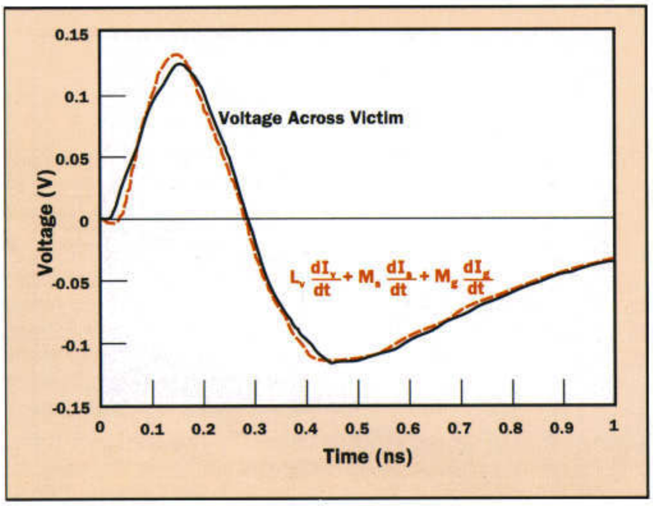


**Figure 11.** Comparison between the victim pin voltage drop and the sum of the inductive contributions to this voltage. It is obvious that capacitive coupling to the victim pin is negligible.

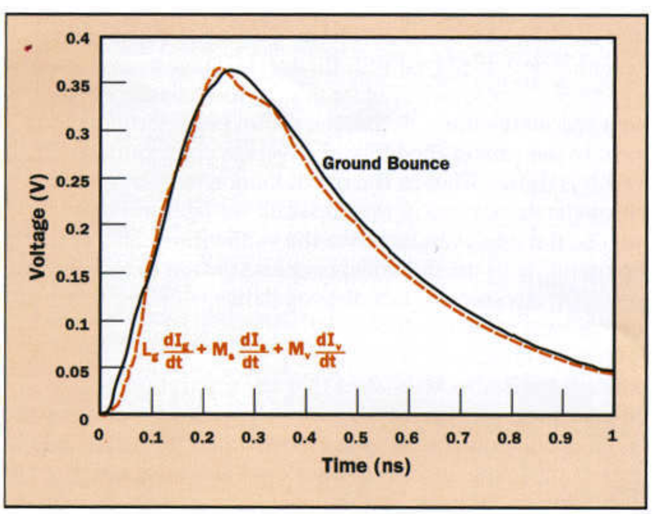


**Figure 12.** Comparison between the ground-bounce voltage and the sum of the inductive contributions to the ground pin voltage. It is apparent that the ground bounce can be attributed almost entirely to inductive effects in the ground conductor.

The ramifications of limited capacitive coupling on a multiconductor system have now been seen: equal forward and backward crosstalk, and common impedance coupling attributable to inductive contributions only. It remains to be seen what impact will be evident if the capacitive coupling is increased. To test this situation the centerline spacing of the pins was reduced to 0.025 inch, the inductances and capacitances were recalculated using PATH_L and CAPRAY, and the behavior of the model was recomputed by SPICE.

Figure 13 illustrates, for the new model, the same four waveforms plotted in Figure 9. A comparison between the two figures indicates several differences attributable to the change in spacing.

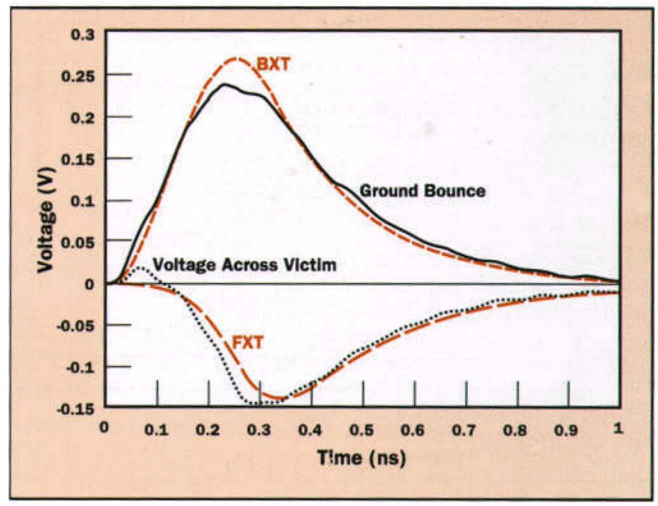


**Figure 13.** Ground bounce, backward crosstalk, forward crosstalk, and the voltage drop across the victim pin. Centerline spacings were reduced to 0.025 inch.

First, the backward crosstalk amplitude is now more than twice as large as the forward crosstalk—270 mV versus 130 mV. This change can be traced to the currents being driven through the near and far end terminations by the increased capacitive coupling. In comparison with Figure 9, the backward crosstalk amplitude has increased by about 70 mV while the forward crosstalk amplitude has decreased by 70 mV.

Second, by comparing the ground-bounce voltages, it can be seen that the decreased spacing caused the ground-bounce voltage peak to drop from 360 mV to about 240 mV. It has yet to be ascertained whether this drop should be attributed to increases in the capacitance, the inductance, or both.

Third, it can be observed that the decreased spacing has changed the wave shape and amplitude of the voltage across the victim pin. Increases in both the capacitance and the inductance are responsible for this change.

In Figure 14, the inductive contributions to the voltage across the victim pin are shown. These waveforms resemble those seen in Figure 10, except that the amplitudes have all increased. This increase is. of course. due to the decreased spacing between the pins. These curves were computed in the same manner as those in Figure 10: the currents calculated for each pin by SPICE were differentiated and multiplied by the appropriate inductances computed by PATH_L.

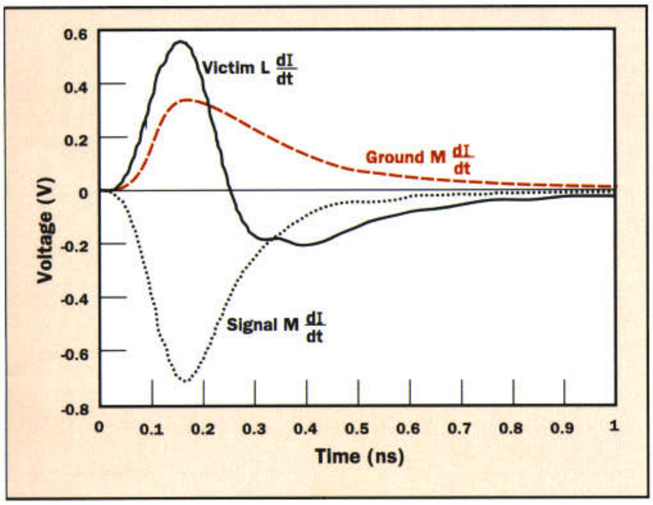


**Figure 14.** Voltages across the victim pin due to its self-inductance, mutual inductance with the signal pin, and mutual inductance with the ground pin. Centerline spacing is 0.025 inch.

Similar to Figure 11, Figure 15 displays the sum of the inductive voltage contributions as well as the voltage across the victim pin computed by SPICE. Comparison of the two waveforms proves that the voltage across the victim pin is no longer attributable to inductive effects alone. The victim pin voltage must contain a significant capacitive contribution. This observation agrees with the assessment given to account for the difference in the crosstalk voltages seen in Figure 13.

The impact that increased capacitive coupling has on the observed crosstalk has been shown conclusively. However, it remains to be seen how the increased capacitance will influence the ground-bounce voltage. To make this determination the self-inductance of the ground and its mutual inductances with the signal and victim pins are again multiplied by the appropriate differentiated currents. In Figure 16 these inductive contributions to the voltage across the ground pin are summed and plotted with the ground-bounce voltage computed by SPICE. Comparison of the two waveforms shows them to be very similar in shape and nearly identical in amplitude—much like Figure 12. Consequently, it can be concluded that even when the capacitive coupling within the system is significant, ground bounce is still almost completely attributable to inductive effects.

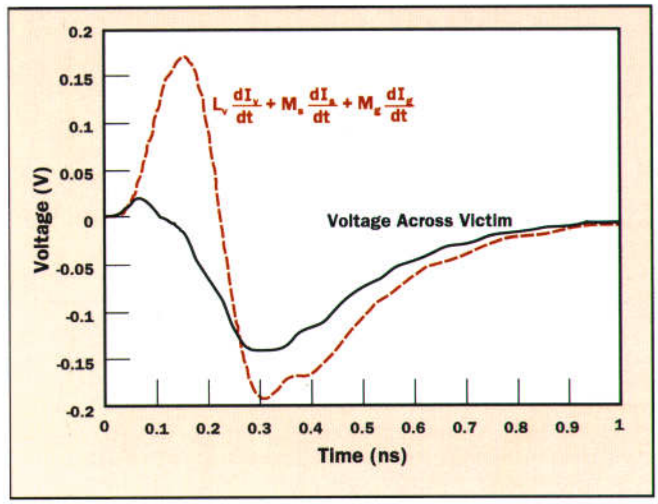


**Figure 15.** Comparison between the victim pin voltage drop and the sum of the inductive contributions to this voltage. Reducing the centerline spacing has caused a significant increase in capacitive coupling, which can no longer be ignored.

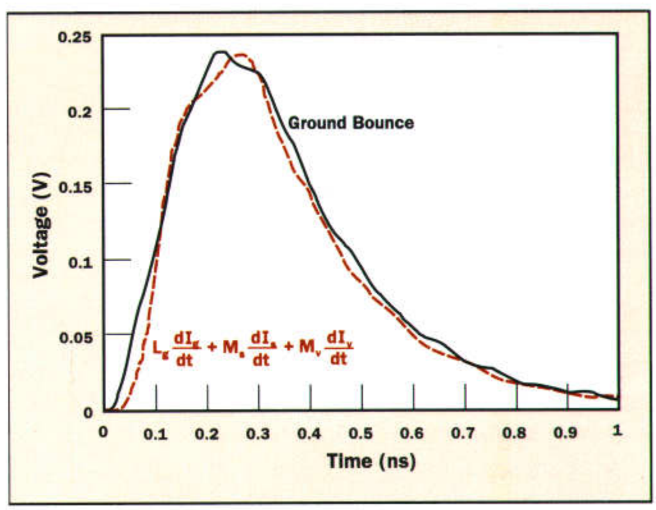


**Figure 16.** Comparison between the ground-bounce voltage and the sum of the inductive contributions to the ground pin voltage. Even with the reduced centerline spacing it is still apparent that the ground bounce can be attributed almost entirely to inductive effects.

Common impedance coupling through shared conductors is virtually independent of capacitive effects.

One other interesting difference between Figures 9 and 13 still needs to be addressed. Recall that in Figure 9 the peak ground-bounce voltage was 360 mV with the pins spaced on 0.100-inch centerlines. When the separation distance was reduced to 0.025 inch, the ground-bounce voltage decreased to about 240 mV. Why does the ground-bounce voltage change when the separation is reduced? The cause of this behavior lies in the nature of mutual inductance. The mutual inductance between two conductors is directly proportional to the reciprocal of their separation. As the separation is reduced, the self-inductance of the ground conductor remains unchanged. However, the mutual inductance between the signal and the ground increases faster than the mutual inductance between the victim and the ground because of the signal pin's closer proximity to the ground. Since the polarity of the signal current is opposite that of the ground and victim currents, its inductive coupling with the ground has a reducing effect on the ground-bounce voltage. Consequently, as the separation distance shrinks, the canceling effect of the signal coupling grows faster than the additive contribution of the victim coupling, and the ground-bounce voltage diminishes.

The analysis of this simple three-conductor structure has yielded some insight into the behavior of real coupled transmission systems. It has been observed that in the absence of capacitive coupling, the backward and forward crosstalk observed on a given victim line will be of identical amplitude with opposite polarity if the terminations are the same impedance. If capacitive coupling is significant, then its presence will add to the amplitude of the observed backward crosstalk while subtracting from the amplitude of the forward crosstalk. Furthermore, the ground-bounce voltage, which causes the reference potential of the ground conductor to fluctuate, is virtually entirely attributable to inductive effects. Changes in the mutual inductance between the ground conductor and other conductors in the system (either by changes in separation or through the addition of magnetic media) will have an impact on the magnitude of the ground-bounce voltage. Depending upon the signal and ground conductor assignments in a transmission system, the ground-bounce voltage could decrease or increase when the inductive coupling changes.

## CROSSTALK AND GROUND BOUNCE IN A 3 x 3 PIN ARRAY

While the three-pin system analyzed in the previous section was useful for examining the mechanisms responsible for crosstalk and ground bounce, few real systems where crosstalk or ground bounce are a concern are limited to three conductors. Typical coupled multiconductor systems contain many signal lines (which may switch simultaneously) as well as many shared ground returns. To more accurately portray real systems, this section will concentrate on investigating the behavior of a 3 x 3 array of pins that are assigned to a 1:1 signal-to-ground pattern as shown below:

$G_7 \; S_8 \; G_9$

$S_4 \; G_5 \; S_6$

$G_1 \; S_2 \; G_3$

The letter in each position denotes whether the conductor is intended as a signal wire or a ground return. The subscript with each letter denotes the number that will be used to reference the pin in that position.

To construct a model for this problem, the pins were assumed to have the same length and diameter values used in the three-pin analysis. The initial centerline spacing from the last model of 0.100 inch was used for both the vertical and horizontal centerline separations between nearest neighbors. The ground pins were commoned together at each end through l-mΩ resistors leading to a common node. Similarly, all the signal conductors were terminated with 50-Ω resistors at either end. For a particular simulation, signal pins that were to be used as victim lines were tied to the ground node at each end through the 50-Ω resistors. Active signal lines were driven simultaneously through their near end terminations by a 1-V source generating a 200-ps risetime step. The inductances and capacitances were again calculated using PATH-L and CAPRAY. As in the three-pin simulation, a two-stage SPICE model was used to generate the data in this section.

It is worthwhile noting that in real applications the cost factor dictates that customers will prefer to assign as few conductors to ground as possible. Signal-to-ground ratios of 3:1, 5:1, and even 10:1 are not unusual. Compared to such usage, the staggered 1:1 signal-to-ground pattern chosen for simulation in this section is well known to provide the best signal integrity available from an open-conductor transmission system. Any crosstalk or ground bounce data illustrated in this section would certainly be worse if a higher signal-to-ground pattern were used for the same pin array.

Table 1 illustrates the effect on the ground-bounce voltage if multiple pins are driven simultaneously. In a good interconnection system, the source and load impedances will be much larger than the impedance of the interconnects. In such a situation, the terminating impedances act as a voltage divider and the portion of the signal voltage that is not dropped across the source impedance is dissipated across the load. In this 3 x 3 array, where each signal pin is terminated with 50 Ω at each end, half of the 1-V step, or 500 mV, will appear as the input voltage on each driven pin. For each signal/victim arrangement listed in the table, the peak ground-bounce voltages shown are the voltage drops computed by SPICE across the ground pin array. The last column of the table illustrating the ratio of the ground-bounce voltage to the 500-mV input voltage shows that it is possible for the peak ground bounce to be more than one-third the amplitude of the driven voltage. Such a fluctuation in the reference potential of a real digital system could very possibly cause inadvertent logic switching. The more discouraging fact shown by the table, however, is the previously predicted characteristic of the ground bounce increasing as more pins are driven simultaneously.

It has been established that ground bounce may have a significant impact on the performance of this multipin array. How significant will the crosstalk be? Table 2 lists both the peak backward and forward crosstalk values and their ratios to some of the ground-bounce voltages shown in Table 1. These values were computed with as many as three signal lines driven simultaneously. One of the first notable items in this table is the fact that when pin 2 is driven, the crosstalk voltages are higher on pin 4 than pin 8. Pin 4's closer proximity to the active line is responsible for this behavior.

**Table 1.** Ground-bounce voltage increases with the number of driven lines.

| Driven Line(s) | Victim Line(s) | Peak Ground-bounce Voltage (mV) | Peak $V_{GB}/V_{In}$ |
|---|---|---|---|
| 2 | 4,6,8 | 47.3 | 9.5% |
| 2,6 | 4,8 | 94.6 | 19% |
| 2,4,6 | 8 | 141.9 | 28% |
| 2,4,6,8 | — | 189.0 | 38% |

**Table 2.** Crosstalk increases with the number of driven lines and is a function of the victim's proximity to the driven signal line(s).

| Driven Line(s) | Victim Line(s) | Peak $V_{GB}$ (mV) | Peak $V_{BXT}$ (mV) | Peak $V_{BXT}/V_{GB}$ | Peak $V_{FXT}$ (mV) | Peak $V_{FXT}/V_{GB}$ |
|---|---|---|---|---|---|---|
| 2 | 4 | 47.3 | 16.3 | 34% | −15.4 | 33% |
| 2 | 8 | 47.3 | 7.8 | 16% | −7.1 | 15% |
| 2,6 | 4,8 | 94.6 | 23.8 | 25% | −21.9 | 23% |
| 2,4,6 | 8 | 141.9 | 40.0 | 28% | −37.3 | 26% |

Some general trends are also evident in Table 2. For instance, the crosstalk increased almost linearly with each additional driven pin. However, the increase in the crosstalk as a percentage of the ground bounce does not follow the same linear propensity. This difference is a result of the fact that as the number of driven pins was increased, the ground-bounce voltage is increasing nearly linearly along with the crosstalk. Another notable characteristic shown by Table 2 is that the crosstalk in this example never reached more than about one-third the amplitude of the ground-bounce voltage. This data would seem to suggest that in multiconductor systems, ground bounce is a larger threat to signal integrity than crosstalk coupled noise. Such an inference is not necessarily true.

Recall that the centerline spacing of this arrangement is too large for capacitive coupling to be significant-as is evident from the similar amplitudes of the backward and forward crosstalk. Therefore, it can be predicted that if the pins were moved closer together, the capacitive coupling would increase and the backward crosstalk would grow substantially. At the same time the forward crosstalk amplitude would decrease and possibly even turn positive. As the backward crosstalk increases, the same spacing reduction would likely produce a decrease in the ground-bounce voltage due to the increased signal-to-ground mutual inductance. The net effect would be that the amplitude of the backward crosstalk could become much larger in comparison to the peak ground-bounce voltage. The lesson to be taken from this scenario is that either the crosstalk, the ground bounce, or both can significantly affect the signal integrity of a system; how these interactions will influence system performance can only be determined under specific circumstances.

One other notable piece of information can be garnered from this simulation. Given a multiconductor system with multiple ground returns, how does the ground current distribute among the conductors and what impact will this distribution have on system performance? Table 3 lists the ground currents flowing through each of the five ground pins in the 3 x 3 model with various numbers of driven lines. Shown below each current amplitude is the percentage of the total ground current flowing through that pin. At low frequencies or slow risetimes, both the capacitive and inductive coupling would be small (due to small dV/dt's and dI/dt's), and one would expect to see the ground current evenly distributed among the five pins. However, at high frequencies and fast risetimes (large dV/dt's and dI/dt's), the coupling between the pins will be significant and the lowest impedance ground paths will be those pins closest to the active line(s). For this reason, when pin 2 was driven alone, 60 percent of the current returned through either pin 1 or pin 3. As more signal lines throughout the array became active, the current tended to distribute among the ground pins closest to those lines. Finally, when all available signal lines were simultaneously active, the ground current was almost evenly distributed across the five ground pins. Consequently, if ground bounce is a problem even when a few lines are active, assigning more ground conductors to the system will probably be of little benefit unless the added grounds are in close proximity to the active signal lines. Conversely, if ground bounce appears to be a problem when many lines are active (as typically happens), adding any low-impedance grounds, regardless of where they are placed in the system, may have some beneficial effect on the performance.

**Table 3.** Ground current distribution as a function of the number of driven lines.

| Driven Line(s) | Current $G_1$ (mA) | Current $G_3$ (mA) | Current $G_5$ (mA) | Current $G_7$ (mA) | Current $G_9$ (mA) |
|---|---|---|---|---|---|
| 2 | 3.0 | 3.0 | 2.2 | 0.9 | 0.9 |
| | (30%) | (30%) | (22%) | (9%) | (9%) |
| 2,6 | 3.8 | 5.9 | 4.7 | 1.8 | 3.8 |
| | (19%) | (30%) | (23%) | (9%) | (19%) |
| 2,4,6 | 6.8 | 6.8 | 7.0 | 4.6 | 4.6 |
| | (23%) | (23%) | (24%) | (15%) | (15%) |
| 2,4,6,8 | 7.6 | 7.6 | 9.4 | 7.6 | 7.6 |
| | (19%) | (19%) | (24%) | (19%) | (19%) |

## CONCLUSION

The concepts of crosstalk and ground bounce have been introduced in this paper. Crosstalk arises through energy being coupled from one conductor to another either capacitively (through electric-field coupling) or inductively (through magnetic-field coupling). Ground bounce, on the other hand, acts as a coupling mechanism to all conductors sharing the same ground return by changing the reference voltage of the ground path. This reference potential is shifted due to a voltage drop across the ground associated with the return path's resistance, self-inductance, and mutual inductance with other active conductors.

An equivalent circuit has been developed that is suitable for modeling both crosstalk and ground bounce using a circuit simulator such as SPICE. A guideline was introduced for determining the number of stages necessary in a circuit model to accurately represent the conductor system of interest. This guideline says that the system should be subdivided into enough sections, or stages, that the propagation delay through each stage is no more than one-tenth of the risetime.

Armed with a practical equivalent circuit for modeling crosstalk and ground bounce and a guideline on how to use it, a three-pin ground-signal-victim arrangement was simulated, The modeled behavior of this system showed that when capacitive crosstalk is negligible, the forward and backward crosstalk will have the same amplitude but opposite polarity. Capacitive crosstalk was seen to inject current onto the victim pin, which moved in both directions from the points of coupling. This capacitive coupling manifested its presence by increasing the amplitude of the backward crosstalk while decreasing the negative amplitude of the forward crosstalk. The ground-bounce voltage was found to principally depend on the self-inductance of the ground conductor and the inductive coupling between the ground and the other conductors. Changes in the capacitive coupling via a decrease in the centerline separation were found to have no impact on the ground-bounce voltage.

In the interest of modeling a more realistic multiconductor system, a 3 x 3 pin array was also analyzed. The presence of multiple ground and signal pins yielded some additional information unavailable from the three-pin simulation. First, the previously predicted increase in ground bounce when multiple lines are driven was readily demonstrated. Second, crosstalk was also seen to increase significantly in the presence of additional driven pins. A qualitative explanation was given to support the notion that with a particular conductor geometry and signal-to-ground pattern, crosstalk and/or ground bounce can significantly hamper system performance, depending upon the coupling present. Third, ground return currents were seen to remain concentrated in the ground conductors nearest the active signal paths. At fast risetimes and high frequencies, the return currents only tend to distribute equally among the return conductors when many lines are driven simultaneously.

The knowledge gained from these simulations can be used to suggest some general design criteria. Crosstalk can usually be minimized by keeping signal lines in close proximity to ground paths. Additionally, surrounding signal lines with ample ground conductors will ensure that signal-to-ground coupling is strong while signal-to-signal crosstalk is minimized. These same guidelines will help keep ground bounce to a minimum if the chosen ground conductor structure provides a low-impedance return path. Such structures as pc board ground planes and multiple parallel lines in a connector or cable fit this criteria; keeping ground paths short also helps. If a multiconductor structure is chosen for the ground path, positive coupling between the ground conductors, which contributes to the ground-bounce voltage, can be minimized by not grouping the ground conductors in close proximity to each other. Finally, minimizing the number of conductors switching simultaneously will reduce the magnitude of the field coupling responsible for crosstalk while concurrently diminishing the return currents responsible for ground bounce. While the number of logic lines changing state at a given time cannot necessarily be arbitrarily dictated, routing such conductors where they are well removed from sensitive inactive lines can at least minimize the crosstalk generated even if the return currents creating ground bounce cannot be diminished.

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