Circuit Modeling and Noise Reduction for Bent Differential Transmission Lines

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Abstract

Differential signaling has become a popular choice for multi-gigabit digital applications, due to its low noise generation and high common-mode noise immunity. This paper describes a methodology to extract the equivalent circuits of the discontinuities for strongly coupled differential lines from the calculated full-wave S-parameters. The signal integrity effects of the bent differential transmission lines in a high-speed digital circuit is then simulated in time domain. A new way to reduce the common mode noise at the receiver by shunt compensation capacitances is also proposed.

1. Model Setup

Differential signaling is frequently employed in PC board designs for high-speed digital systems. It is quite common that the signal traces are bent for practical layout consideration. To investigate the signal integrity issues of these bends, it is necessary to develop lumped and frequency-dependent equivalent circuits for the discontinuities [1]

For example, consider a bent differential line shown in Fig.1(a). Since the size of the bends is usually much smaller than the wavelength of the considered frequency range, the structure can be modelled as Fig.1(b) where the bend can be approximated by a simple lumped circuit shown in Fig.1(c). This lumped circuit model is an extension of the π model used for bend of a single transmission line. Though this model can include radiation, conductor and dielectric losses as frequency dependent resistance, these losses are small and negligible for practical applications up to Gbps rates.

2. De-embedding Technique for Strongly Coupled Structures

Most of the present commercial EM field solvers assume isolated output ports in de-embedding the transmission line effects. When the input/output transmission lines are strongly coupled, a more general approach need be developed. For simplicity, consider a system of N bent coupled transmission lines. By using field solvers, its full wave characteristics can be modeled in terms of a 2N-port ABCD matrix. Based on the cascade property, the ABCD matrix can be divided into three parts: two sections of coupled transmission lines and a bend structure in between. As show in Fig.1(b), the ABCD matrix can be formulated as

$$\begin{bmatrix}\vec{V}_D\\ \vec{I}_D\end{bmatrix}=\begin{bmatrix}\vec{A}_\ell & -\vec{B}_\ell\\ -\vec{C}_\ell & \vec{D}_\ell\end{bmatrix}^{-1}\cdot\begin{bmatrix}\vec{A}_D & \vec{B}_D\\ \vec{C}_D & \vec{D}_D\end{bmatrix}\cdot\begin{bmatrix}\vec{A}_\ell & \vec{B}_\ell\\ \vec{C}_\ell & \vec{D}_\ell\end{bmatrix}\cdot\begin{bmatrix}\vec{V}_A\\ \vec{I}_A\end{bmatrix}=\begin{bmatrix}\vec{A}_0 & \vec{B}_0\\ \vec{C}_0 & \vec{D}_0\end{bmatrix}\cdot\begin{bmatrix}\vec{V}_A\\ \vec{I}_A\end{bmatrix} \tag{1}$$

where the subscript "D" denotes for the bend and "ℓ" for a section of the coupled transmission lines with length ℓ.

To de-embed the transmission lines, consider a section of coupled transmission lines of length 2ℓ. Its ABCD matrix can be formulated as

$$\begin{bmatrix}\vec{V}_D\\ \vec{I}_D\end{bmatrix}=\begin{bmatrix}\vec{A}_\ell & -\vec{B}_\ell\\ -\vec{C}_\ell & \vec{D}_\ell\end{bmatrix}^{-1}\cdot\begin{bmatrix}\vec{A}_\ell & \vec{B}_\ell\\ \vec{C}_\ell & \vec{D}_\ell\end{bmatrix}\cdot\begin{bmatrix}\vec{V}_A\\ \vec{I}_A\end{bmatrix}=\begin{bmatrix}\vec{A}_0 & \vec{B}_0\\ \vec{C}_0 & \vec{D}_0\end{bmatrix}\cdot\begin{bmatrix}\vec{V}_A\\ \vec{I}_A\end{bmatrix} \tag{2}$$

It is readily available that the ABCD matrix of a single transmission line assuming quasi-TEM mode propagation satisfies the property

$$\begin{bmatrix}A_\ell & B_\ell\\ C_\ell & D_\ell\end{bmatrix}=\begin{bmatrix}\cos\beta\ell & -jZ_0\sin\beta\ell\\ -jY_0\sin\beta\ell & \cos\beta\ell\end{bmatrix}=\begin{bmatrix}A_\ell & -B_\ell\\ -C_\ell & D_\ell\end{bmatrix}^{-1} \tag{3}$$

The ABCD_ℓ matrix can be solved from (2) and then substituted into (1) to extract the desired

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$ABCD_D$ for the coupled bend.

3. Lumped Circuit Model Extraction

Given the ABCD matrix of bend discontinuity, the corresponding S matrix can be obtained and written as

$$\begin{bmatrix} b_1 \\ b_2 \\ b_3 \\ b_4 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{21} & S_{22} & S_{23} & S_{24} \\ S_{31} & S_{32} & S_{33} & S_{34} \\ S_{41} & S_{42} & S_{43} & S_{44} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \\ a_3 \\ a_4 \end{bmatrix} = \begin{bmatrix} \alpha_{11} & \alpha_{12} & 1+\alpha_{13} & \alpha_{14} \\ \alpha_{21} & \alpha_{22} & \alpha_{23} & 1+\alpha_{24} \\ 1+\alpha_{31} & \alpha_{32} & \alpha_{33} & \alpha_{34} \\ \alpha_{41} & 1+\alpha_{42} & \alpha_{43} & \alpha_{44} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \\ a_3 \\ a_4 \end{bmatrix} \tag{4}$$

where α's are small quantities if only the size of discontinuity is much smaller than one wavelength of the frequency range.

By KCL law while neglecting all the terms corresponding to higher orders of α's, it is not difficult to relate α's with the lumped circuit elements in Fig.1(c) by

$$\alpha_{11} - \alpha_{13} = \frac{(R_{13} + j\omega L_{13,s})}{Z_0}, \quad \alpha_{12} - \alpha_{14} = \frac{(R_{12} + j\omega L_{12,m})}{Z_0}$$

$$-\frac{(\alpha_{11}+\alpha_{13}+\alpha_{12}+\alpha_{14})}{2} = (G_{11}+j\omega C_{11,s})\times Z_0 = (G_{33}+j\omega C_{33,s})\times Z_0$$

$$\frac{(\alpha_{12}+\alpha_{14})}{2} = (G_{12}+j\omega C_{12,m})\times Z_0 = (G_{34}+j\omega C_{34,m})\times Z_0 \tag{5}$$

4. Transient Analysis

Let the differential transmission lines structure in Fig.1(a) be with substrate dielectric constant $\varepsilon_r = 4.3$, height 1.5mm, microstrip width 1.75mm, thickness 0.1mm, and center to center separation 2.5mm. The structure is analysed by full-wave solver IE3D at 2GHz frequency. Following the aforementioned procedure, equivalent circuit of the coupled right-angle bends is extracted and the lumped circuit elements are L_{13}=0.187nH, L_{24}=2.10nH, K_L=0.129, C_{11}=C_{33}=0.062pF, C_{22}=C_{44}= 0.282pF, C_{12}=C_{34}=0.023pF, R_{13}=0.08mΩ, and R_{24}=0.39mΩ. The S parameters obtained directly from the field solver and those deduced from the extracted lumped circuit model are in good agreement, although not shown here. Similar procedure has been employed for coupled round-corner bends and Fig 2. shows the structure and the extracted equivalent circuit values.

Given the equivalent circuit model, it is easy to simulate the signal integrity effects due to the bend by circuit simulation, e.g., Spice. Consider a right-angle bent differential signal lines with $\ell = 26$ mm. and a ramped step signal of ±0.5 volts is incident to its one end. Fig.3 shows the simulated waveforms of the common mode signal at the receiver end and the reflected differential mode at the sending end with rising time of 100ps and 50ps as a parameter.

Figure 4 shows the resultant waveforms when the right-angle bend is replaced by the round corner bend. It is found that the round corner bend can reduce the reflection at the sending end, but can hardly improve the common mode noise at the receiver.

Figure 5(a) considers two bent structures back to back connected and Fig.5(b) show the simulated common mode noise at the receiver. It is found that the dual coupled bend can be helpful for reducing the common mode noise, however the compensation becomes less effective if the length of coupled lines between the two coupled bends increases.

An efficient way to reduce the common mode noise at the receiver is to shunt compensation capacitances at the bend as shown in Fig.6(a). From the simulated common mode noise at the receiver in Fig.6(b), the two compensation capacitances can add some extra time delay for the signal along the inner short path. To minimize the mode conversion parameter S_{c2d1} [5] (ratio of common mode noise at receiving port to differential mode input signal), the optimal value of compensating capacitance can be decided from (4) and (5). It is given by

$$C_C = C_{22} + \frac{L_{24} - L_{13}}{2Z_o} - C_{11} \tag{6}$$

where Z_o denotes the odd mode impedance of differential lines. For the aforementioned right-angle bend structure, the optimal compensating capacitance can be calculated to be about 0.6pF. The simulated results in Fig.6 validate the correctness of (6). Also, the maximum common-mode noise versus the compensation capacitance is listed in Table 1. For the present example, the employment of compensation capacitances reduces the common mode noise by 53%.

5. Conclusions

Based on the cascade property of the ABCD matrix, a systematic procedure has been established to extract lumped circuit model for discontinuities of differential lines. The circuit model can be employed to investigate the signal integrity issues of the discontinuities on high-speed digital signals. The round corner bend can reduce the reflection at the sending end, but can hardly improve the common mode noise at the receiver. Two bent structures back to back connected closely is one way to reduce the common mode noise. Insertion of compensation capacitances is another effective way.

References

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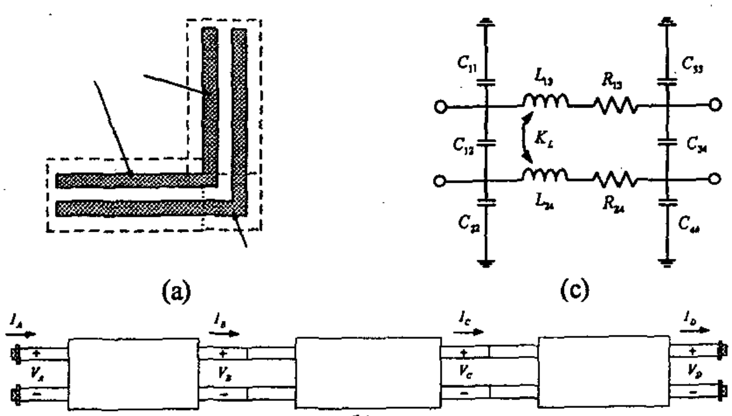


Fig 1.Bent coupled transmission line. (a) layout, (b) block diagram, and (c) equivalent circuits

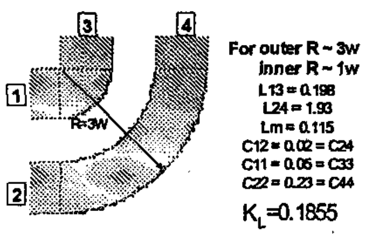
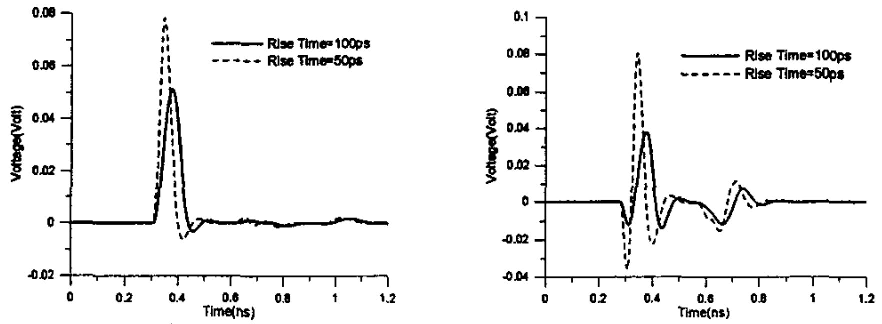


Fig 2.Equivalent circuit of round corner bends

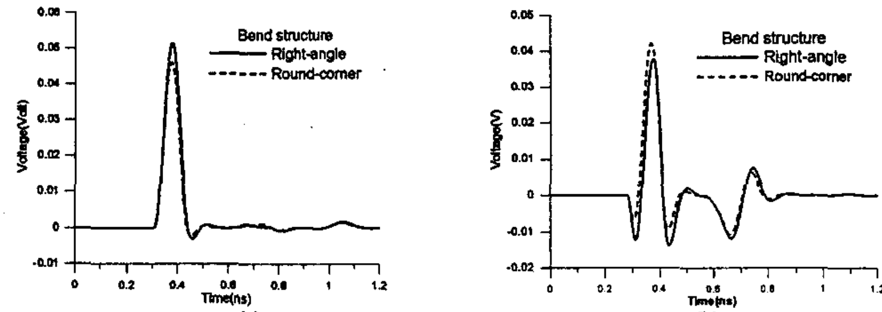
compensating capacitance	common mode noise, $V_{peak-to-peak}$
No Cc	0.05456V
with Cc=0.4pF	0.02667V
with Cc=0.5pF	0.02561V
with Cc=0.6pF	0.02550V
with Cc=0.7pF	0.02960V
with Cc=0.8pF	0.03400V

Table 1. Maximum common mode noise versus compensation capacitance



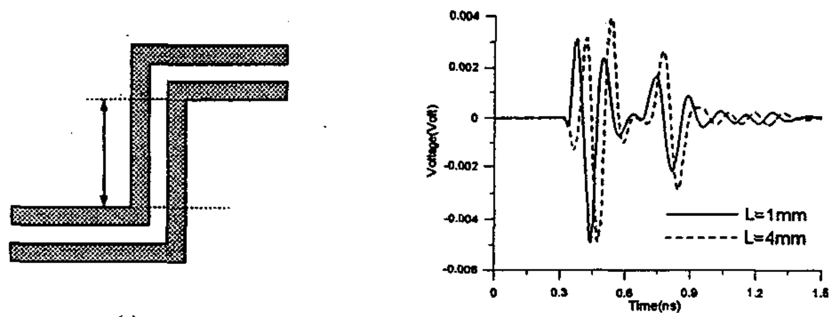
(a) (b)

Fig. 3 (a) Common mode noise at receiver and (b) reflected differential mode noise at sending end for right-angle bent differential line



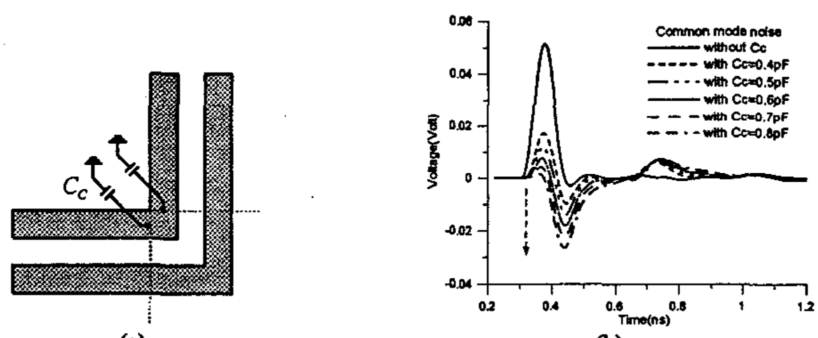
(a) (b)

Fig 4. Signal integrity effect of round corner differential bend in (a) common mode noise at receiver and (b) reflected differential mode noise at sending end for right-angle bent differential line



(a) (b)

Fig 5. (a) A dual differential bends structure and (b) simulated common mode noise at receiver



(a) (b)

Fig 6. (a) A differential bend with compensation capacitances and (b) simulated common mode noise at receiver