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Mixed-mode S-parameters and Conversion Techniques

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1. Introduction

Differential signaling in analog circuits is an old technique that has been utilized for more than 50 years. During the last decades, it has also been becoming popular in digital circuit design, when low voltage differential signaling (LVDS) became common in high-speed digital systems. Today LVDS is widely used in advanced electronics such as laptop computers, test and measurement instrument, medical equipment and automotive. The reason is that with increased clock frequencies and short edge rise/fall times, crosstalk and electromagnetic interferences (EMI) appear to be critical problems in high-speed digital systems. Differential signaling is aimed to reduce EMI and noise issues in order to improve the signal quality. However, in traditional microwave theory, electric current and voltage are treated as single-ended and the S-parameters are used to describe single-ended signaling. This makes advanced microwave and RF circuit design and analysis difficult, when differential signaling is utilized in modern communication circuits and systems. This chapter introduces the technique to deal with differential signaling in microwave and millimeter wave circuits.

2. Differential Signal

Differential signaling is a signal transmission method where the transmitting signal is sent in pairs with the same amplitude but with mutual opposite phases. The main advantage with the differential signaling is that any introduced noise equally affects both the differential transmission lines if the two lines are tightly coupled together. Since only the difference between the lines is considered, the introduced common-mode noise can be rejected at the receiver device. However, due to manufacturing imperfections, signal unbalance will occur resulting in that the energy will convert from differential-mode to common-mode and vice versa, which is known as cross-mode conversion. To damp the common-mode currents, a common-mode choke can be used (without any noticeable effect on the differential currents) to prevent radiated emissions from the differential lines. To produce the electrical field strength from microamperes of common-mode current, milliamperes of differential current are needed (Clayton, 2006). Moreover, the generated electric and magnetic fields from a differential line pair are more localized compared to

those from single-ended lines. Owing to the ability of noise rejection, the signal swing can be decreased compared to a single-ended design and thereby the power can be saved.
When the signal on one line is independent of the signal on the adjacent line, i.e., an uncoupled differential pair, the structure does not utilize the full potential of a differential design. To fully utilize the differential design, it is beneficial to start by minimizing the spacing between two lines to create the coupling as strong as possible. Thereafter, the conductors width is adjusted to obtain the desired differential impedance. By doing this, the coupling between the differential line pair is maximized to give a better common-mode rejection.
S-parameters are very commonly used when designing and verifying linear RF and microwave designs for impedance matching to optimize gain and minimize noise. Although, traditional S-parameter representation is a very powerful tool in circuit analysis and measurement, it is limited to single-ended RF and microwave designs. In 1995, Bockelman and Einsenstadt introduced the mixed-mode S-parameters to extend the theory to include differential circuits. However, owing to the coupling effects between the coupled differential transmission lines, the odd- and even-mode impedances are not equal to the unique characteristic impedance. This leads to the fact that a modified mixed-mode S-parameters representation is needed. In this chapter, by starting with the familiar concepts of coupling, crosstalk and terminations, mixed-mode S-parameters will be introduced. Furthermore, conversion techniques between different modes of S-parameters will be described.

2.1 Coupling and Crosstalk

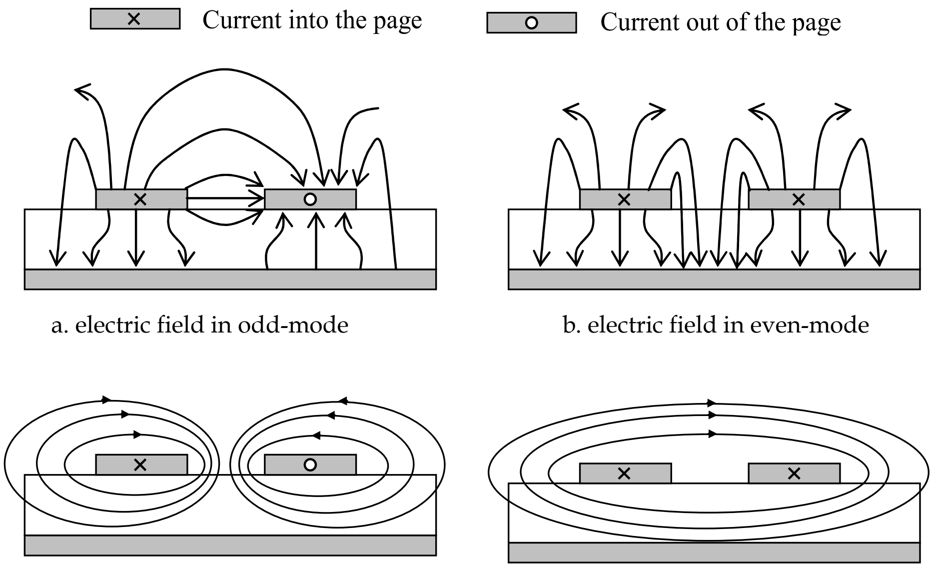
Like in single-ended signaling, differential transmission lines need to be correctly terminated, otherwise reflections arise and distortions are introduced into the system. In a system where parallel transmission lines exist, either in differential signaling or in parallel single-ended lines, line-to-line coupling arises and it will cause characteristic impedance variations. The coupling between the parallel single-ended lines is also known as crosstalk and it is related to the mutual inductance (L_m) and capacitance (C_m) existing between the lines. The induced crosstalk or noise can be described with a simple approximation as following

$$V_{noise} = \mathrm{L_m}\frac{\mathrm{dI_{driver}}}{\mathrm{dt}} \tag{1}$$

$$I_{noise} = C_m\frac{dV_{driver}}{dt} \tag{2}$$

where V_{noise} and I_{noise} are the induced voltage and current noises on the adjacent line and $\mathrm{V_{driver}}$ and $\mathrm{I_{driver}}$ are the driving voltage and current on the active line. Since both the voltage and current noises are induced by the rate of current and voltage changes, extra care is needed for high-speed applications.
The coupling between the parallel lines depends firstly on the spacing between the lines and secondly on the signal pattern sent on the parallel lines. Two signal modes are defined, i.e., odd- and even-modes. The odd-mode is defined such that the driven signals in the two adjacent lines have the same amplitude but a 180 degree of relative phase, which can be related to differential signal. The even-mode is defined such that the driven signals in the two adjacent lines have the same amplitude and phase, which can be related to common-

mode noise for a differential pair of signal. Fig. 1 shows the electric and magnetic field lines in the odd- and even-mode transmissions on the two parallel microstrips. Fig. 1a shows that the odd-mode signaling causes coupling due to the electric field between the microstrips, while in the even-mode shown in Fig 1b, there is no direct electric coupling between the lines. Fig. 1c shows that the magnetic field in the odd-mode has no coupling between the two lines while, as shown in Fig. 1d, in the even-mode the magnetic field is coupled between the two lines.



c. magnetic field in odd-mode d. magnetic field in even-mode

Fig. 1. Odd- and even-mode electric and magnetic fields for two parallel microstrips.

2.2 Odd-mode

The induced crosstalk or voltage noise in a pair of parallel transmission lines can be approximated with Equation 1. For the case of two parallel transmission lines the equation can be rewritten as following

$$V_1 = L_0 \frac{dI_1}{dt} + L_m \frac{dI_2}{dt} \tag{3}$$

$$V_2 = L_0 \frac{dI_2}{dt} + L_m \frac{dI_1}{dt} \tag{4}$$

where L_0 is the equivalent lumped-self-inductance in the transmission line and L_m is the mutual inductance arisen due to the coupling between the lines. Signal propagation in the odd-mode results in $I_1 = -I_2$, since the current is always driven with equal magnitude but in opposite directions. Substituting it into Equations 3 and 4 yeilds

$$V_1 = (L_0 - L_m) \frac{dI_1}{dt} \tag{5}$$

$$V_2 = (L_0 - L_m)\frac{dI_2}{dt} \tag{6}$$

This shows that, due to the crosstalk, the total inductance in the transmission lines reduces with the mutual inductance (L_m).
Similarly, the current noise in the parallel transmission lines can be estimated with Equation 2. For two parallel transmission lines the equation can be rewritten as following

$$I_1 = C_0\frac{dV_1}{dt} + C_m\frac{d(V_1 - V_2)}{dt} \tag{7}$$

$$I_2 = C_0\frac{dV_2}{dt} + C_m\frac{d(V_2 - V_1)}{dt} \tag{8}$$

where C_0 is the equivalent lumped-capacitance between the line and ground, and C_m is the mutual capacitance between the transmission lines arisen due to the coupling between the lines. Signal propagation in odd-mode results in $V_1 = -V_2$. Substituting it into Equations 7 and 8 yields

$$I_1 = (C_0 + 2C_m)\frac{dV_1}{dt} \tag{9}$$

$$I_2 = (C_0 + 2C_m)\frac{dV_2}{dt} \tag{10}$$

Equations 9 and 10 show that, in opposite to the inductance, the total capacitance increases with the mutual capacitance.
The addition of mutual inductance and capacitance shows that the characteristic impedance as well as the phase velocity is directly dependant of the mutual coupling, as shown with the following equations

$$Z_{oo} = \sqrt{\frac{R + j\omega L}{G + j\omega C}} = \sqrt{\frac{R + j\omega(L_0 - L_m)}{G + j\omega(C_0 + 2C_m)}} \tag{11}$$

$$\vartheta_{po} = \frac{1}{\sqrt{LC}} = \frac{1}{\sqrt{(L_0 - L_m)(C_0 + 2C_m)}} \tag{12}$$

where Z_{oo} and v_{po} are the odd-mode impedance and phase velocity, respectively.
Consequently, the total characteristic impedance in the odd-mode reduces due to the coupling or crosstalk between the parallel transmission lines and the phase velocity changes as well.

2.3 Even-mode

In the case of even-mode where the signals are driven with equal magnitude and phase, $V_1 = V_2$ and $I_1 = I_2$, Equations 3, 4, 7 and 8 can be rewritten to the following:

$$V_1 = (L_0 + L_m)\frac{dI_1}{dt} \tag{13}$$

$$V_2 = (L_0 + L_m)\frac{dI_2}{dt} \tag{14}$$

$$I_1 = C_0 \frac{dV_1}{dt} \tag{15}$$

$$I_2 = C_0 \frac{dV_2}{dt} \tag{16}$$

Consequently, in opposite to the odd-mode case, the even-mode wave propagation changes the even-mode impedance (Z_{oe}) and phase velocity (v_{pe}) as shown below:

$$Z_{oe} = \sqrt{\frac{R+j\omega L}{G+j\omega C}} = \sqrt{\frac{R+j\omega(L_0+L_m)}{G+j\omega C_0}} \tag{17}$$

$$\vartheta_{pe} = \frac{1}{\sqrt{LC}} = \frac{1}{\sqrt{(L_0+L_m)(C_0)}} \tag{18}$$

2.4 Terminations

As shown in the previous section, the impedance varies due to the odd- and even-mode transmissions and the coupling between the transmission lines. Fig. 2 shows a graph of the odd- and even-mode impedance change as a function of the spacing between two specific parallel-microstrips. If the loads connected to the parallel lines have a simple termination as commonly used in the single-ended case, reflections will occur due to $Z_{oo} \neq Z_{oe} \neq Z_0$. Fig. 3 shows two termination configurations, i.e., Pi- and T-terminations, which can terminate both the odd- and even-mode signals in coupled parallel transmission lines.

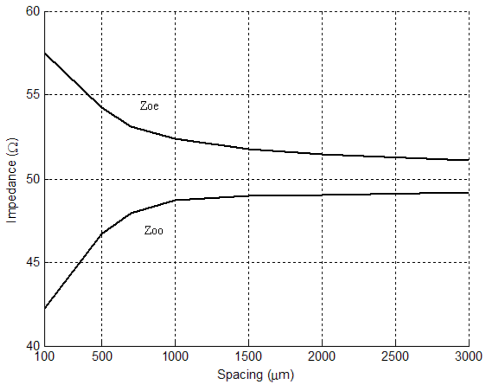
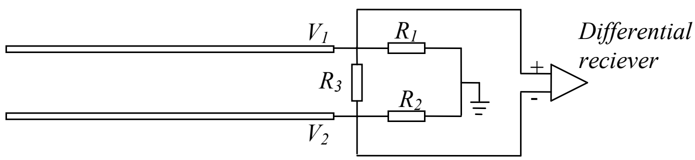
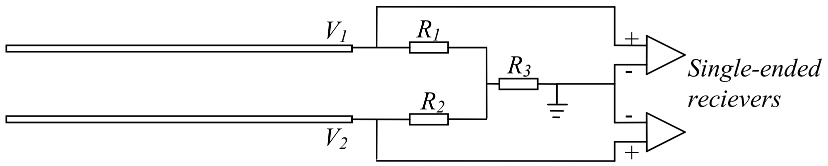


Fig. 2. Variation of the odd- and even-mode impedances as a function of the spacing between two parallel microstrips.



a. Pi-termination



b. T-termination

Fig. 3. Termination configurations for coupled transmission lines.

Fig. 3a shows the Pi-termination configuration. In the odd-mode transmission, i.e., $V_1 = -V_2$ a virtual ground can be imaginarily seen in the middle of R_3 and this forces $R_3/2$ in parallel with R_1 or R_2 equal to Z_{oo}. Since no current flows between the two transmission lines in the even-mode, i.e., $V_1 = V_2$, R_1 and R_2 must thus be equal to Z_{oe}. For a matched differential system with optimized gain and noise, the following expressions need to be fulfilled for a Pi-termination configuration.

$$R_1 = R_2 = Z_{oe} \tag{19}$$

$$R_3 = 2\frac{Z_{oe}Z_{oo}}{Z_{oe}-Z_{oo}} \tag{20}$$

Fig. 3b shows the T-termination configuration. In the odd-mode transmission, i.e., $V_1 = -V_2$, a virtual ground can be seen between R_1 and R_2 and this makes R_1 and R_2 equal to Z_{oo}. In the even-mode transmission, i.e., $V_1 = V_2$ no current flows between the two transmission lines. This makes R_3 to be seen as two $2R_3$ in parallel, as illustrated in Fig. 4. This leads to the conclusion that Z_{oe} must be equal to R_1 or R_2 in serial with $2R_3$. Equations 21 and 22 show the required values of the termination resistors needed for the T-termination configuration in order to get a perfect matched system (Hall et al., 2000).

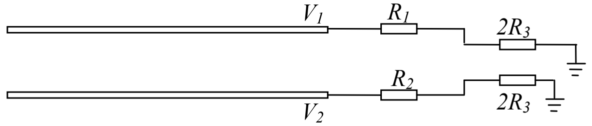


Fig. 4. Equivalent network for T-network termination in even-mode.

$$R_1 = R_2 = Z_{oo} \tag{21}$$

$$R_3 = \frac{1}{2}(Z_{oe} - Z_{oo}) \tag{22}$$

3. S-parameters

Scattering parameters or S-parameters are commonly used to describe an n-port network operating at high frequencies like RF and microwave frequencies. Other well-known parameters often used for describing an n-port network are *Z* (impedance), *Y* (admittance), *h* (hybrid) and *ABCD* parameters. The main difference between the S-parameters and other parameter representations lies in the fact that S-parameters describe the normalized power waves when the input and output ports are properly terminated, while other parameters describe voltage and current with open or short ports. S-parameters can also be used to express other electrical properties like gain, insertion loss, return loss, voltage standing wave ratio, reflection coefficient and amplifier stability.

3.1 Single-ended

The travelling waves used in the transmission line theory are defined with incident normalized power wave a_n and reflected normalized power wave b_n.

$$a_n = \frac{1}{2\sqrt{Z_0}}(V_n + Z_0 I_n) \tag{23}$$

$$b_n = \frac{1}{2\sqrt{Z_0}}(V_n - Z_0 I_n) \tag{24}$$

where index n refers to a port number and Z_0 is the characteristic impedance at that specific port. The normalized power waves are used for the definition of single-ended S-parameters. Fig. 5 shows a sketch of a two-port network with the normalized power wave definitions (Kurokawa, 1965).

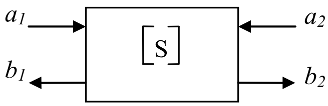


Fig. 5. S-parameters with normalized power wave definition of a two-port network.

The two-port S-parameters are defined as follows.

$$S_{11} = \left.\frac{b_1}{a_1}\right|_{a_2=0} \equiv reflection\ at\ port\ 1 \tag{25}$$

$$S_{12} = \left.\frac{b_1}{a_2}\right|_{a_1=0} \equiv reverse\ voltage\ gain \tag{26}$$

$$S_{21} = \left.\frac{b_2}{a_1}\right|_{a_2=0} \equiv forward\ voltage\ gain \tag{27}$$

$$S_{22} = \left.\frac{b_2}{a_2}\right|_{a_1=0} \equiv reflection\ at\ port\ 2 \tag{28}$$

and in matrix form

$$\begin{Bmatrix} b_1 \\ b_2 \end{Bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{Bmatrix} a_1 \\ a_2 \end{Bmatrix} \tag{29}$$

When measuring S-parameters, it is important not to have any power wave reflected at port 1 or 2, i.e., $a_1 = 0$ or $a_2 = 0$, as shown in Equations 25-28. Otherwise errors are included in the results. For an n-port network, Equation 29 can be extended to the following expression (Ludwig & Bretchko, 2000):

$$\begin{pmatrix} b_1 \\ b_2 \\ \vdots \\ b_n \end{pmatrix} = \begin{bmatrix} S_{11} & S_{12} & \cdots & S_{1n} \\ S_{21} & S_{22} & \cdots & S_{2n} \\ \vdots & \vdots & \ddots & \vdots \\ S_{n1} & S_{n2} & \cdots & S_{nn} \end{bmatrix} \begin{pmatrix} a_1 \\ a_2 \\ \vdots \\ a_n \end{pmatrix} \tag{30}$$

3.2 Mixed-mode

A two-port single-ended network can be described by a 2x2 S-parameter matrix as shown by Equation 29. However, to describe a two-port differential-network a 4x4 S-parameter matrix is needed, since there exists a signal pair at each differential port. Fig. 6 shows a sketch of the power wave definitions of a two-port differential-network, i.e., a four-port network.

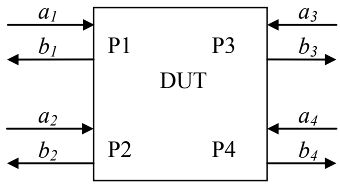


Fig. 6. Power wave definition of a differential two-port network.

Since a real-world differential-signal is composed of both differential- and common-mode signals in general, the single-ended four-port S-parameter matrix does not provide much insight information about the differential- and common-mode matching and transmission. Therefore, the mixed-mode S-parameters must be used. The differential two-port and mixed-mode S-parameters are defined by Equation 31 (Bockelman & Eisenstadt, 1995).

$$\begin{bmatrix} b_{d1} \\ b_{d2} \\ b_{c1} \\ b_{c2} \end{bmatrix} = \begin{bmatrix} \begin{bmatrix} S_{dd11} & S_{dd12} \\ S_{dd21} & S_{dd22} \end{bmatrix} & \begin{bmatrix} S_{dc11} & S_{dc12} \\ S_{dc21} & S_{dc22} \end{bmatrix} \\ \begin{bmatrix} S_{cd11} & S_{cd12} \\ S_{cd21} & S_{cd22} \end{bmatrix} & \begin{bmatrix} S_{cc11} & S_{cc12} \\ S_{cc21} & S_{cc22} \end{bmatrix} \end{bmatrix} \begin{bmatrix} a_{d1} \\ a_{d2} \\ a_{c1} \\ a_{c2} \end{bmatrix} \tag{31}$$

where a_{dn}, a_{cn}, b_{dn} *and* b_{cn} are normalized differential-mode incident-power, common-mode incident-power, differential-mode reflected-power, and common-mode reflected-power at port n. The mixed-mode S matrix is divided into 4 sub-matrixes, where each of the sub-matrixes provides information for different transmission modes.

- S_{dd} sub-matrix: differential-mode S-parameters
- S_{dc} sub-matrix: mode conversion of common- to differential-mode waves
- S_{cd} sub-matrix: mode conversion of differential- to common-mode waves
- S_{cc} sub-matrix: common-mode S-parameters

With mixed-mode S-parameters, characteristics about the differential- and common-mode transmissions and conversions between differential- and common-modes can be found (Bockelman & Eisenstadt, 1995).

3.3 Single-Ended to Mixed-Mode conversion

The best way to measure the mixed-mode S-parameters is to use a four-port mixed-mode vector network analyzer (VNA). In the case where the mixed-mode S-parameters cannot directly be simulated or measured, the single-ended results can first be obtained and then converted into mixed-mode S-parameters by a mathematical conversion. This section shows how it is done for a differential two-port network shown in Fig. 6.
The differential- and common-mode voltages, currents and impedances can be expressed as below, where n is the port number,

$$V_{dn} = V_{2n-1} - V_{2n} \qquad V_{cn} = \frac{V_{2n-1}+V_{2n}}{2} \tag{32}$$

$$I_{dn} = \frac{I_{2n-1}-I_{2n}}{2} \qquad I_{cn} = I_{2n-1} + I_{2n} \tag{33}$$

$$Z_d = \frac{V_d}{I_d} = 2Z_{oo} \qquad Z_c = \frac{V_c}{I_c} = \frac{Z_{oe}}{2} \tag{34}$$

Similar to the single-ended incident- and reflected-powers, the differential- and common-mode incident- and reflected-powers are defined as follows

$$a_{dn} = \frac{1}{2\sqrt{Z_{dn}}}(V_{dn} + Z_{dn}I_{dn}) \tag{35}$$

$$a_{cn} = \frac{1}{2\sqrt{Z_{cn}}}(V_{cn} + Z_{cn}I_{cn}) \tag{36}$$

$$b_{dn} = \frac{1}{2\sqrt{Z_{dn}}}(V_{dn} - Z_{dn}I_{dn}) \tag{37}$$

$$b_{cn} = \frac{1}{2\sqrt{Z_{cn}}}(V_{cn} - Z_{cn}I_{cn}) \tag{38}$$

where a_{dn}, a_{cn}, b_{dn} and b_{cn} are normalized differential-mode incident-power, common-mode incident-power, differential-mode reflected-power, and common-mode reflected-power at port n, respectively. The voltage and current at port n can be expressed by rewriting Equations 35-38 to as follow

$$V_n = \sqrt{Z_0}(a_n + b_n) \tag{39}$$

$$I_n = \frac{1}{\sqrt{Z_0}}(a_n + b_n) \tag{40}$$

Inserting Equations 32-34 and Equations 39-40 into Equations 35-38 and assuming that $Z_{oo} = Z_{oe} = Z_0$, yields the following results

$$a_{dn} = \frac{a_{2n-1} - a_{2n}}{\sqrt{2}} \qquad a_{cn} = \frac{a_{2n-1} + a_{2n}}{\sqrt{2}} \tag{41}$$

$$b_{dn} = \frac{b_{2n-1} - b_{2n}}{\sqrt{2}} \qquad b_{cn} = \frac{b_{2n-1} + b_{2n}}{\sqrt{2}} \tag{42}$$

As shown by Equations 41 and 42 the differential incident and reflected waves can be described by the single-ended waves. Inserting Equations 41 and 42 into Equation 31, the following expression is obtained.

$$[S^{mm}] = [M][S][M]^{-1} \tag{43}$$

$$where\ [S^{mm}] = \begin{bmatrix} S_{dd11} & S_{dd12} & S_{dc11} & S_{dc12} \\ S_{dd21} & S_{dd22} & S_{dc21} & S_{dc22} \\ S_{cd11} & S_{cd12} & S_{cc11} & S_{cc12} \\ S_{cd21} & S_{cd22} & S_{cc21} & S_{cc22} \end{bmatrix},$$

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{21} & S_{22} & S_{23} & S_{24} \\ S_{31} & S_{32} & S_{33} & S_{34} \\ S_{41} & S_{42} & S_{43} & S_{44} \end{bmatrix} and\ [M] = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & -1 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 1 & 1 & 0 & -1 \\ 0 & 0 & 1 & 1 \end{bmatrix}$$

As shown by Equation 43, single-ended S-parameters can be converted into mixed-mode S-parameters with the [M]-matrix (Bockelman & Eisenstadt, 1995).
Note that the conversion method assumes that $Z_{oo} = Z_{oe} = Z_0$. This assumption is only true if the coupling between the differential signals does not exist. Figure 2 clearly shows that if there is a coupling between the transmission lines then $Z_0 \neq Z_{oo} \neq Z_{oe}$. Although this conversion method is widely used today, the weakness of the conversion method has been noticed by people working in the area (Vaz & Caggiano, 2004).
Based on this observation, two new parameters k_{oo} and k_{oe} depending on the coupling between the transmission lines are introduced (Huynh et al., 2007). By this extension, the effect of differential- and common-mode impedances are included in the conversion. Inserting $Z_{oo} = k_{oo}Z_0$ and $Z_{oe} = k_{oe}Z_0$ into Equations 35-38, the following equations can be obtained

$$a_{dn} = \frac{(1+k_{oo})(a_{2n-1} - a_{2n}) + (1-k_{oo})(b_{2n-1} - b_{2n})}{2\sqrt{2k_{oo}}} \tag{44}$$

$$a_{cn} = \frac{(1+k_{oe})(a_{2n-1} + a_{2n}) + (1-k_{oe})(b_{2n-1} + b_{2n})}{2\sqrt{2k_{oe}}} \tag{45}$$

$$b_{dn} = \frac{(1+k_{oo})(b_{2n-1} - b_{2n}) + (1-k_{oo})(a_{2n-1} - a_{2n})}{2\sqrt{2k_{oo}}} \tag{46}$$

$$b_{cn} = \frac{(1+k_{oe})(b_{2n-1} + b_{2n}) + (1-k_{oe})(a_{2n-1} + a_{2n})}{2\sqrt{2k_{oe}}} \tag{47}$$

Inserting Equations 44-47 into Equations 31 results in the following equation

$$[S^{mm}] = ([M_1][S] + [M_2])([M_1] + [M_2][S])^{-1} \tag{48}$$

$$where\ [M_1] = \begin{bmatrix} \frac{1+k_{oo}}{2\sqrt{2k_{oo}}} & -\frac{1+k_{oo}}{2\sqrt{2k_{oo}}} & 0 & 0 \\ 0 & 0 & \frac{1+k_{oo}}{2\sqrt{2k_{oo}}} & -\frac{1+k_{oo}}{2\sqrt{2k_{oo}}} \\ \frac{1+k_{oe}}{2\sqrt{2k_{oe}}} & \frac{1+k_{oe}}{2\sqrt{2k_{oe}}} & 0 & 0 \\ 0 & 0 & \frac{1+k_{oe}}{2\sqrt{2k_{oe}}} & \frac{1+k_{oe}}{2\sqrt{2k_{oe}}} \end{bmatrix}$$

$$and\ [M_2] = \begin{bmatrix} \frac{1-k_{oo}}{2\sqrt{2k_{oo}}} & -\frac{1-k_{oo}}{2\sqrt{2k_{oo}}} & 0 & 0 \\ 0 & 0 & \frac{1-k_{oo}}{2\sqrt{2k_{oo}}} & -\frac{1-k_{oo}}{2\sqrt{2k_{oo}}} \\ \frac{1-k_{oe}}{2\sqrt{2k_{oe}}} & \frac{1-k_{oe}}{2\sqrt{2k_{oe}}} & 0 & 0 \\ 0 & 0 & \frac{1-k_{oe}}{2\sqrt{2k_{oe}}} & \frac{1-k_{oe}}{2\sqrt{2k_{oe}}} \end{bmatrix}$$

As shown by Equation 48, the single-ended S-parameter representation can be converted to mixed-mode S-parameters with the [M_1] and [M_2] matrixes for coupled transmission lines.

4. Conclusions

Owing to the existence of coupling between two parallel transmission lines, the characteristic impedance of the transmission line will change depending on the line spacing and the signal pattern transmitting on the adjacent lines. The defined odd- and even-mode signals can be related to differential- and common-mode signals in the differential transmission technique. Consequently, this leads to the fact that working with highly coupled differential transmission lines one must take the odd- and even-mode impedance variations into account. Otherwise, mismatching will occur and distortions will be introduced into the system.

Furthermore, mixed-mode S-parameters were introduced and how to convert single-ended to mixed-mode S-parameters was also presented. The impedance changes due to the odd- and even-mode signal pattern are included to give a high accuracy when using mixed-mode S-parameters.

What is not included in this chapter is how to find the odd- and even mode impedances in practice. However, one can do it by using a differential time domain reflectormeter (TDR) to find the odd- and even-mode impedances. In fact, companies providing vector network analyzers can provide TDR as an embedded module, so only one vector network analyzer is needed for real measurements.

Moreover, complex odd- and even-mode impedance have not been taken into consideration in this chapter. Further studies can be done for verification of the theory.

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